Blind Channel Estimation and Layered Space Frequency Equalisation for MIMO CDMA Uplink Systems

Sonu Punnoose

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List of Acronyms

ACM Autocorrelation Contribution Method

BPSK Binary Phase Shift Keying

BSS Blind Source Separation

CDMA Code Division Multiple Access

CP Cyclic Prefix

CP-OFDM Cyclic Prefix OFDM

DS-CDMA Direct Sequence Code Division Multiple Access

DSP Digital Signal Processor

EGC Equal Gain Combining

EVD Eigen Value Decomposition

FDE Frequency Domain Equalisation

FFT Fast Fourier Transform

GB Guard Band

IBI Inter Block Interference

ICI Inter Carrier Interference

IFFT Inverse Fast Fourier Transform

ISI Inter Symbol Interference

the ZP system [5]. Equalisation in ZP systems is slightly more complicated as described in [4], [5] as a result of the loss of the diagonal matrix structure. The ZP system can be converted to a diagonal matrix form by multiplying the received signal with $\mathbf{R}_{zp}$ [5]. The $P \times$L code matrix for user i in the ZP case, denoted here by $\mathbf{C}_{tz,i}$ is of the form

$$\mathbf{C}_{tz,i} = \begin{bmatrix} c_1 & 0 & \ldots & 0 \\ c_2 & c_1 & 0 & 0 \\ : & : & c_1 & 0 \\ c_N & \ldots & & c_1 \\ 0 & \ldots & \ldots & 0 \\ \vdots & 0 & \ldots & \vdots \\ 0 & \ldots & \ldots & 0 \end{bmatrix} \tag{4.29}$$

The three blind channel estimation approaches can be reformulated as shown below.

1. SS Approach-

   Referring to Eqn. (4.11), the cost function for ZP system is given as

$$\mathbf{h}_{j,i} = \arg \min_{||\mathbf{h}||=1} ||\mathbf{h}_{j,i}^H \mathbf{Qz}_i \mathbf{h}_{j,i}||^2 \tag{4.30}$$

   where the matrix $\mathbf{Qz}_i$ for user i is given by

$$\mathbf{Q}_i = \mathbf{C}_{tz,i}^H \mathbf{U}_{nj} \mathbf{U}_{nj}^H \mathbf{C}_{tz,i} \tag{4.31}$$

   where $\mathbf{h}_{j,i}$ is the channel impulse response between user i and receive antenna j. The eigenvector corresponding to the smallest eigenvalue furnishes the solution.

2. SA Method-

As in the SA approach for the CP system given by Eqn. (4.21), the cost function for ZP case is given as

$$\hat{\mathbf{h}_{\mathbf{j,i}}} = \arg \min_{||\mathbf{h}_{j,i}||=1} \mathbf{h}_{j,i}^H \mathbf{C}_{tz,i}^H \mathbf{R}_{zj}^{-m} \mathbf{C}_{tz,i} \mathbf{h}_{j,i} \tag{4.32}$$

with $\mathbf{C}_{tz,i}$ replacing the $\mathbf{C}_i$ in Eqn. (4.21) and $\mathbf{R}_{zj}$ denoting the auto-correlation matrix at the j^{th} receive antenna for the ZP system.

3. ACM method- Similar to the CP case given by Eqn. (4.28), the cost function to be minimised for the MIMO-ZP-SC-CDMA case is obtained as

$$\hat{\mathbf{h}_1} = \arg \min_{||v||=1} \frac{\mathbf{v}^{\mathrm{H}} \mathbf{C}_{tz,1}^{\mathrm{H}} \mathbf{R}_z^{-2} \mathbf{C}_{tz,1} \mathbf{v}}{\mathbf{v}^{\mathrm{H}} \mathbf{C}_{tz,1}^{\mathrm{H}} \mathbf{Rz}^{-1} \mathbf{C}_{tz,1} \mathbf{v}} \tag{4.33}$$

where as mentioned earlier $\mathbf{R}_{zj}$ is the auto-correlation matrix of the received signal at antenna j while $\mathbf{C}_{tz,i}$ represents the ZP Toeplitz code matrix for the user i.

Figure 4.3 compares the performance of ZP and CP when used for blind channel estimation with SS, SA and ACM methods. As shown, the ZP system performs slightly better as its more robust and amenable to blind channel estimation methods. This is due to the tall Toeplitz nature of the channel matrix in ZP systems as opposed to the diagonal nature of the channel matrix in CP sytems [4].

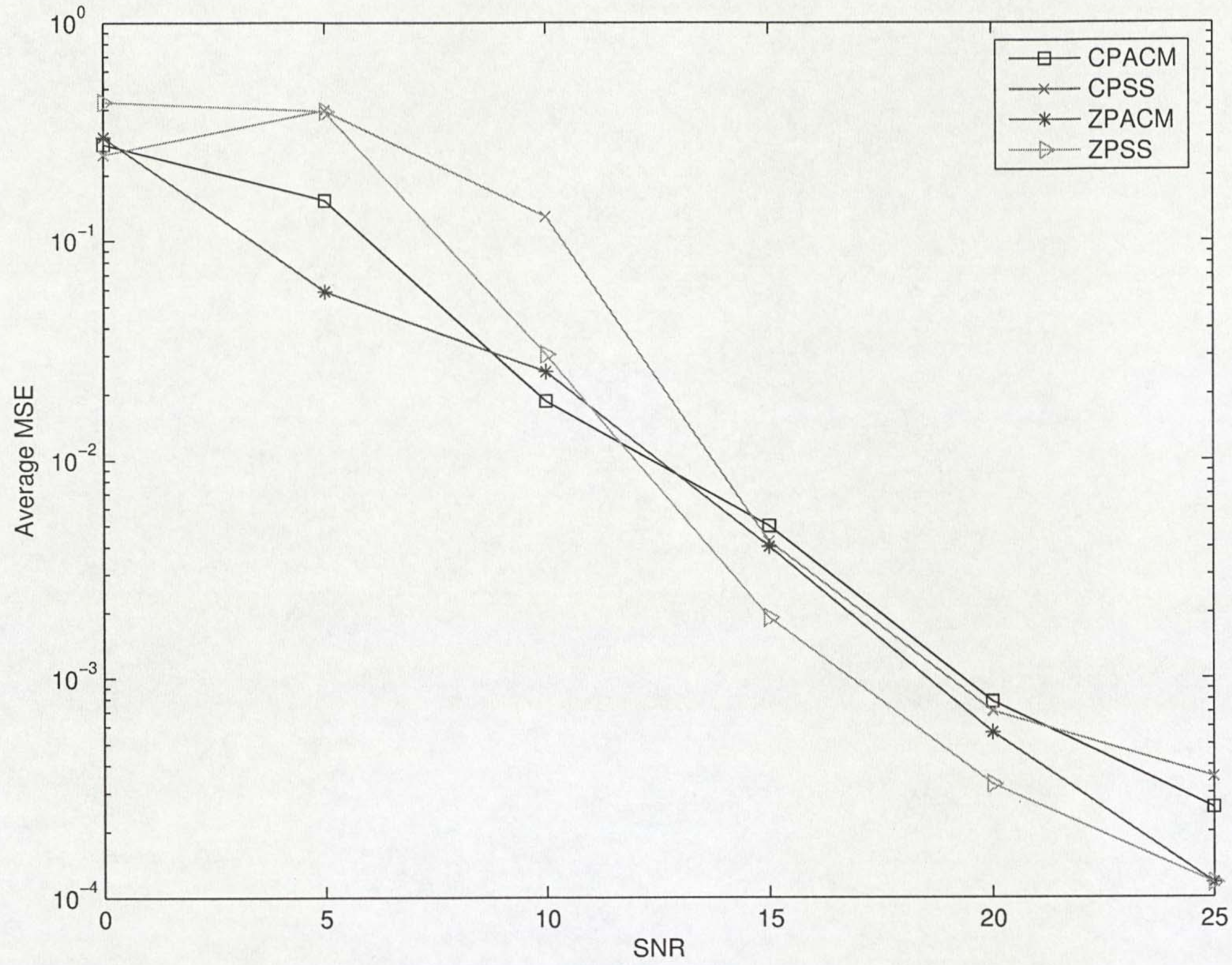

Figure 4.3: Performance of CP vs ZP MIMO SC-CDMA Uplink Channel Estimation

4.4 Equalisation

4.4.1 MMSE Equaliser

The received signal, after the blind channel estimation stage, is stripped of its guard band and then FFT processed. The frequency domain signal is then equalised using a FDE, which is advantageous compared to using a TDE equaliser as mentioned earlier. The FFT processed block of received signal can be equalised by different schemes but we focus on the MMSE equaliser [34] and the LSFE equaliser in the next section. The soft estimate $\hat{\mathbf{d}}_i$ of $\mathbf{d}_i$ for user i is given as

$$\hat{\mathbf{d}}_i = \mathbf{w}_{mmse,i}^H \mathbf{X} \tag{4.34}$$

where $\mathbf{X}$ is the net received signal as defined in Eqn. (4.4). The MMSE equaliser for i^{th} user $\mathbf{w}_{mmse,i}$ is expressed as shown below with $\hat{\mathbf{H}}_i$ denoting the estimate of $\mathbf{H}_i$

$$\mathbf{w}_{mmse,i} = \left(\sum_{i=1}^{N_t} \hat{\mathbf{H}}_i \hat{\mathbf{H}}_i^{\mathrm{H}} + \sigma^2 \mathbf{I} \right)^{-1} \hat{\mathbf{H}}_i \tag{4.35}$$

The equalised signals are then passed through the IFFT stage to transform the signals back to time-domain before despreading. Finally the despread signals are fed into a hard decision device to obtain the final estimate of user's signal.

4.4.2 LSFE Detector

The LSFE detector is similar to the one described for MC-CDMA with one key difference. After using FDE and cancelling the interference in frequency domain, the equalised signal is IFFT processed before being depsread and sent to a hard decision device.

As mentioned previously, the blind channel estimation is done before removal of GB and application of FFT. After ordering the users based on the strength of the received signal i.e. SNR values, the strongest signal is detected first [45]-[46]. After detection, the interference caused by the detected user is removed yielding a modified received signal sans the strongest user. The next strongest user is detected and the process repeated till all the users information are detected.

The block diagram of the LSFE is shown in Figure 4.4, which consists of a blind channel estimation stage followed by N_t LSFE stages performing interference cancellation. The LSFE technique is modified such that the ordering of the streams is done based on the sum of all mean square error (MSE) values calculated for each user. After detection of each of the users,the effects of the detected user is cancelled from the remaining signal. After cancelling the interference caused by the detected signal, the corresponding channel values are also removed to leave the undetected stream with its corresponding channel coefficients.

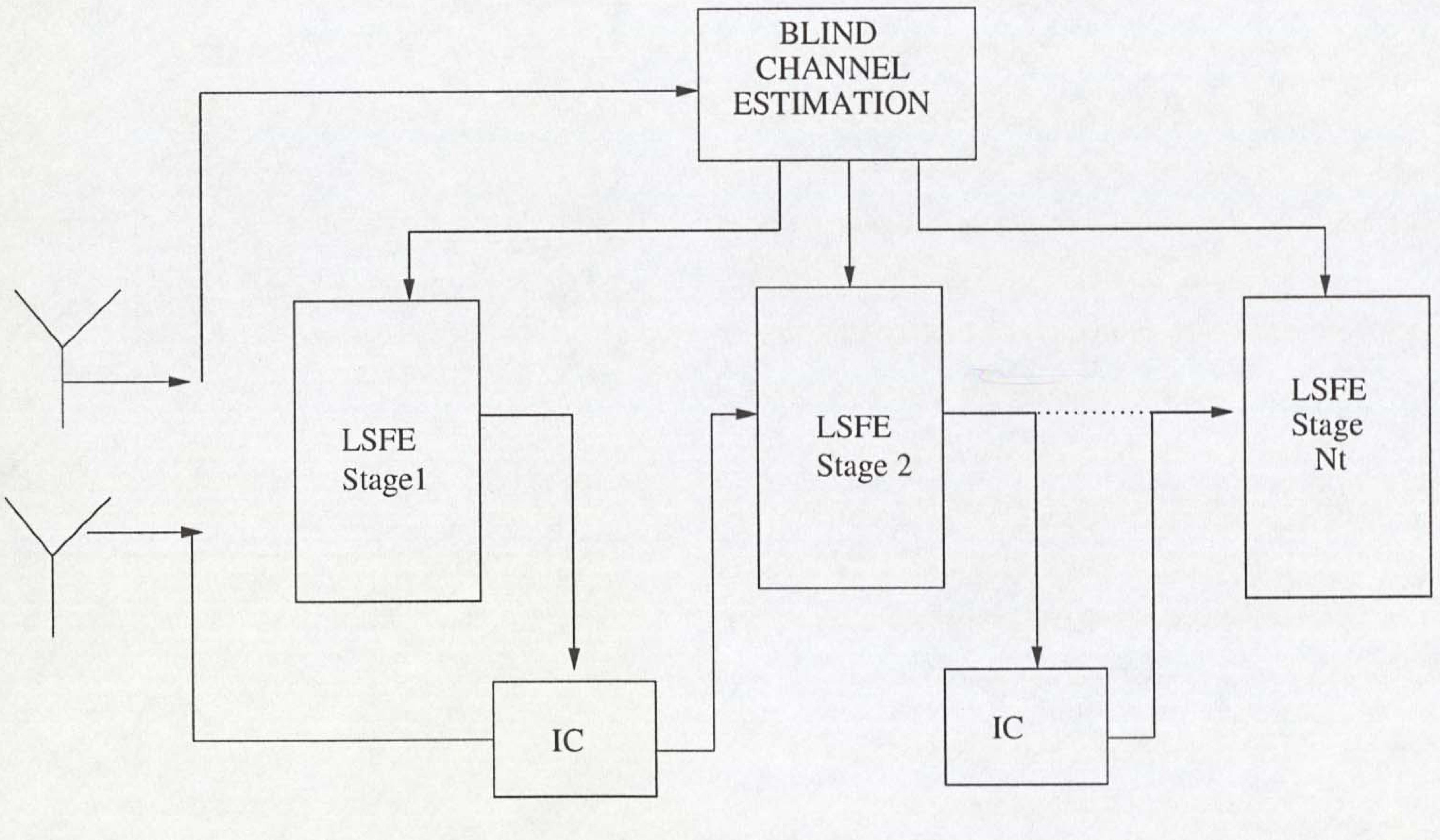

Figure 4.4: Uplink SC-CDMA LSFE Receiver Zhu [45]

Figure 4.5 shows a particular stage of the LSFE equaliser/detector [45], where the FDE block performs equalisation as well as interference cancellation. Then the signal is IFFT processed before the hard decision is made by the decision device. At the end of each stage, the interference from the already detected user(s) is cancelled from the received signals yielding modified received signals with less interference. Thus all users are detected and their corresponding interference are cancelled in a successive manner.

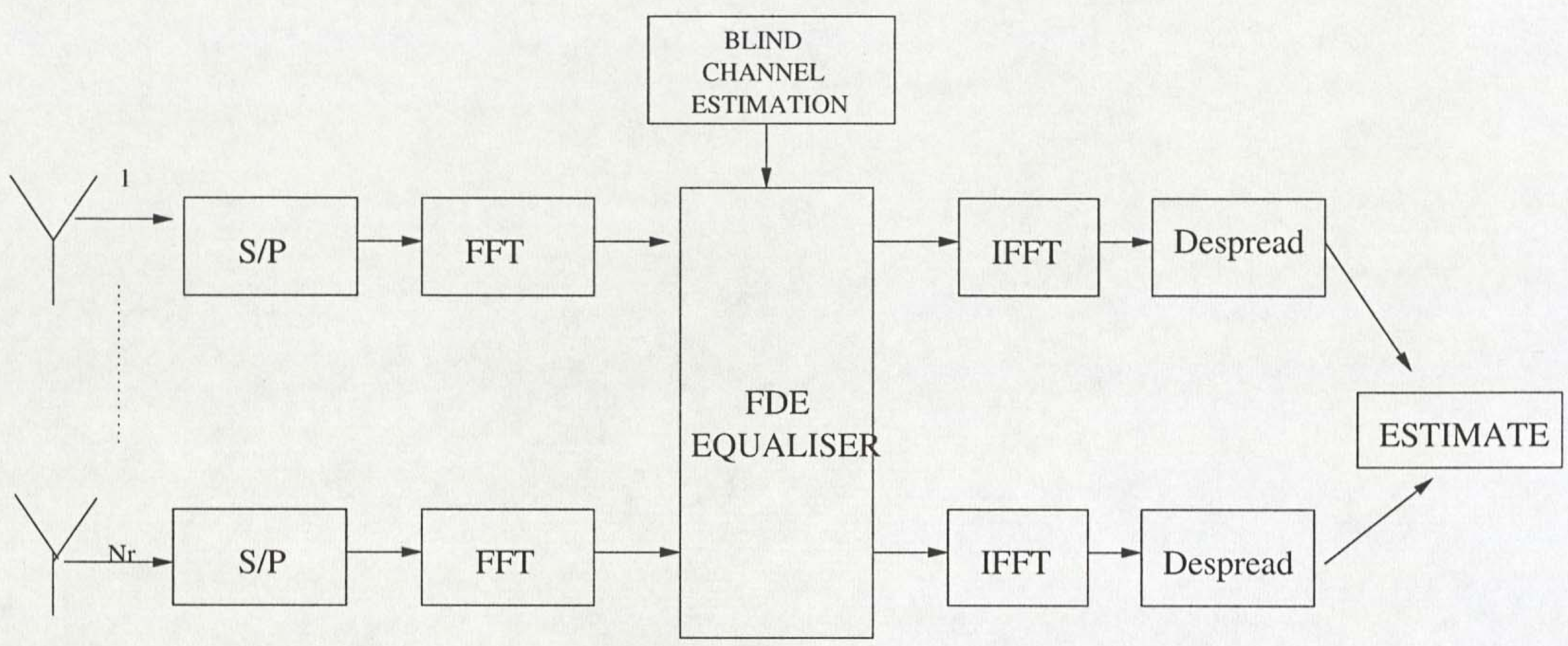

Figure 4.5: A Particular LSFE Stage Zhu [45]

Assuming perfect interference cancellation at each stage, the received signal at a particular stage on the n^{th} subcarrier can be written as

$$\mathbf{X}[n] = \sum_{k} \mathbf{H}_k[n]\mathbf{d}_k[n] + \mathbf{V}[n] \tag{4.36}$$

where k denotes summation over the undetected users. $\mathbf{X}[n]$ is the net received signal on subcarrier n represented as $\mathbf{X}[n] = [\mathbf{X}_1[n]\mathbf{X}_2[n] \ldots \mathbf{X}_{N_r}[n]]^T$. $\mathbf{H}_k[n]$ represents the channel matrix for user k given as $\mathbf{H}_k[n] = [\mathbf{H}_{1,k}[n]\mathbf{H}_{2,k}[n] \ldots \mathbf{H}_{N_r,k}[n]]^T$ and $\mathbf{d}_k[n] = c_k[n]s_k[n]$ denotes the spread signal of user k on subcarrier n. The MSE is defined as follows

$$MSE_{in} = E\left|\hat{s}_i[n] - s_i[n]\right|^2 \tag{4.37}$$

where $\hat{s}_i[n]$ is the soft estimate of s_i. The FDE coefficients are calculated based on the MSE criterion. The soft estimate of $\mathbf{d}_i$ for user i is given as [45]

$$\hat{\mathbf{d}}_i = \mathbf{w}_i^H \mathbf{X} \tag{4.38}$$

where $\mathbf{w}_i$ is the weight vector over all subcarriers for the user i, which can be expressed as $\mathbf{w}_i = [\mathbf{w}_i^T[1]\mathbf{w}_i^T[2]\ldots\mathbf{w}_i^T[N]]^T$. The weight vectors $\mathbf{w}_i[n]$ for the n^{th} subcarrier can be expressed as [45]-[46]

$$\mathbf{w}_i[n] = \mathbf{R}_i^{-1}[n]\hat{\mathbf{H}}_i[n] \tag{4.39}$$

where $\mathbf{R}_i[n]$ is the autocorrelation matrix of the channel response on each subcarrier n for user i and is expressed as

$$\mathbf{R}_i[n] = \sum_k \hat{\mathbf{H}}_k[n]\hat{\mathbf{H}}_k^H[n] + \sigma^2\mathbf{I} \tag{4.40}$$

with k denoting the summation over undetected user signals. The resulting MSE for subcarrier n for user i is given by

$$MSE_{in} = 1 - \frac{1}{N}\sum_{n=0}^{N-1} \hat{\mathbf{H}}_i^{\mathrm{H}}[n]\mathbf{R}_i^{-1}[n]\hat{\mathbf{H}}_i[n] \tag{4.41}$$

The decision variable, which determines the ordering of the streams is given by the sum of all the MSE values for each subcarrier as shown below

$$MSE_i = \sum_{n=1}^{N} MSE_{in} \tag{4.42}$$

The ordering is done from the smallest MSE to the largest and detection of signals is then performed in this order.

4.5 Simulation Results

The simulation setup mainly involved a uplink MIMO SC-CDMA system with 2 transmit antennas and 2 receive antennas unless otherwise specified. The number of subcarriers was set to $N = 32$ with channel modelled as quasi-static block Rayleigh fading channel, which remains constant over the duration of a block. The channel impulse length was set to $L = 4$ with uniform power delay profile and rms delay spread of $1.153\mu s$. A block size of $N_s = 100$ frames was used throughout. As blind channel estimation methods yield estimates upto a scalar factor, a pilot symbol was used to obtain the scaling factor and one pilot symbol was used for each transmit-receive antenna pair. Walsh-Hadamard codes, of size N, i.e. equal to the number of subcarriers, were used throughout because of their orthogonal nature. The BER results are obtained as averaged over 1000 Monte-Carlo runs for the setup described. BPSK modulation was used for in all the simulations. 'SS', 'SA' and 'ACM' represent the Subspace, Subspace Approximation and ACM methods respectively with 'Perfect' denoting the perfect channel case. In practical systems, some form of error-control coding is used to improve the BER and counter the effects of fading and MAI upto an extent.

Figure 4.6 compares the MMSE BER performances of a simple MIMO SC-CDMA uplink system using known channel and blind channel estimates obtained via SS, SA and ACM based methods. As is clearly seen, while the blind channel estimate based performances do converge at high SNR, at lower SNR the ACM based estimate performs the best while the SA has the worst performance.

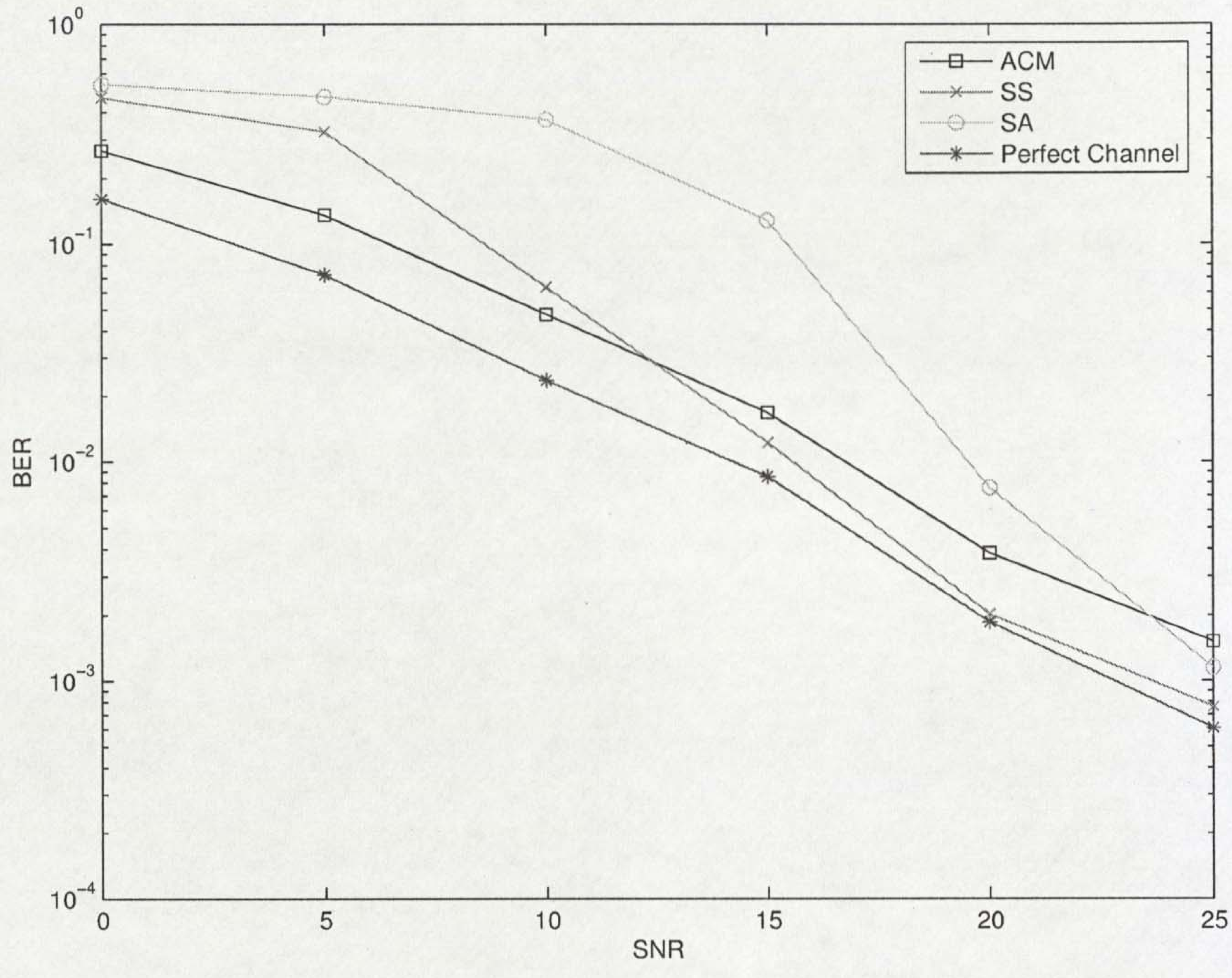

Figure 4.6: Performance of SS,SA and ACM vs known channel for 2x2 MIMO SC-CDMA Uplink System with N=32, L=4

Figure 4.7 shows the improvement in performance obtained as a result of using LSFE over the conventional MMSE equaliser. Here the interference cancellation of the detected users signal at each stage is responsible for the improvement in performance. As shown, at SNR of 20 dB, LSFE provides an improvement of almost an order of magnitude compared to the MMSE equaliser, using the blind channel estimates. The SA method performs the poorest among the three while the SS and ACM methods are quite similar in performance.

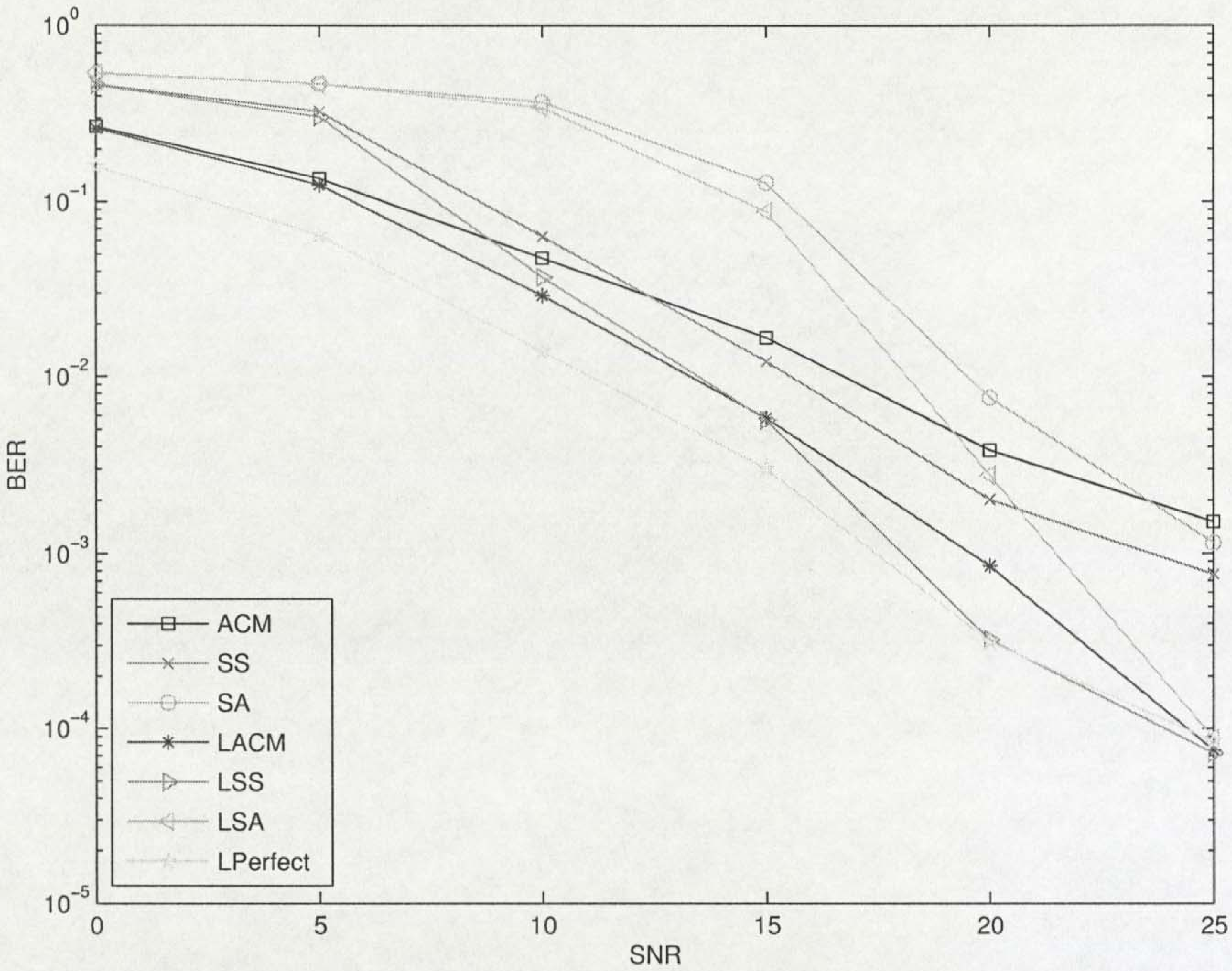

Figure 4.7: Performance of LSFE vs MMSE for 2x2 MIMO SC-CDMA Uplink with N=32,L=4

Figure 4.8 illustrates the gains that are achieved through use of receive diversity at the receiver (base-station) end. The extra diversity afforded by multiple receive antennas provides significant gains. As is seen, the SS, SA and ACM based MMSE equalisers achieve similar performance levels as expected with the ACM performing better at lower SNR. At SNR of 20 dB, the 2 transmit, 4 receive antenna system yields nearly an order of magnitude improvement compared to the 2 transmit, 2 receive antenna system.

Figure 4.9 shows the performance of LSFE and MMSE equalisers using

LSFE Layered Space Frequency Equalisation

MMSE Minimum Mean Square Equalisation

MRC Maximal Ratio Combining

MC-CDMA Multi-Carrier Code Division Multiple Access

MC-TDMA Multi-Carrier Time Division Multiple Access

MIMO Multiple Input Multiple Output

MIMO OFDM Multiple Input Multiple Output OFDM

MIMO MC-CDMA Multiple Input Multiple Output MC-CDMA

MIMO SC-CDMA Multiple Input Multiple Output SC-CDMA

OFDM Orthogonal Frequency Division Multiplexing

OFDMA Orthogonal Frequency Division Multiple Access

ORC Orthogonality Restoring Combining

MAI Multiple Access Interference

MC-DS-CDMA Multi-Carrier Direct Sequence CDMA

MCM MultiCarrier Modulation

MSE Mean Square Error

MT-CDMA Multi Tone CDMA

PIC Parallel Interference Cancellation

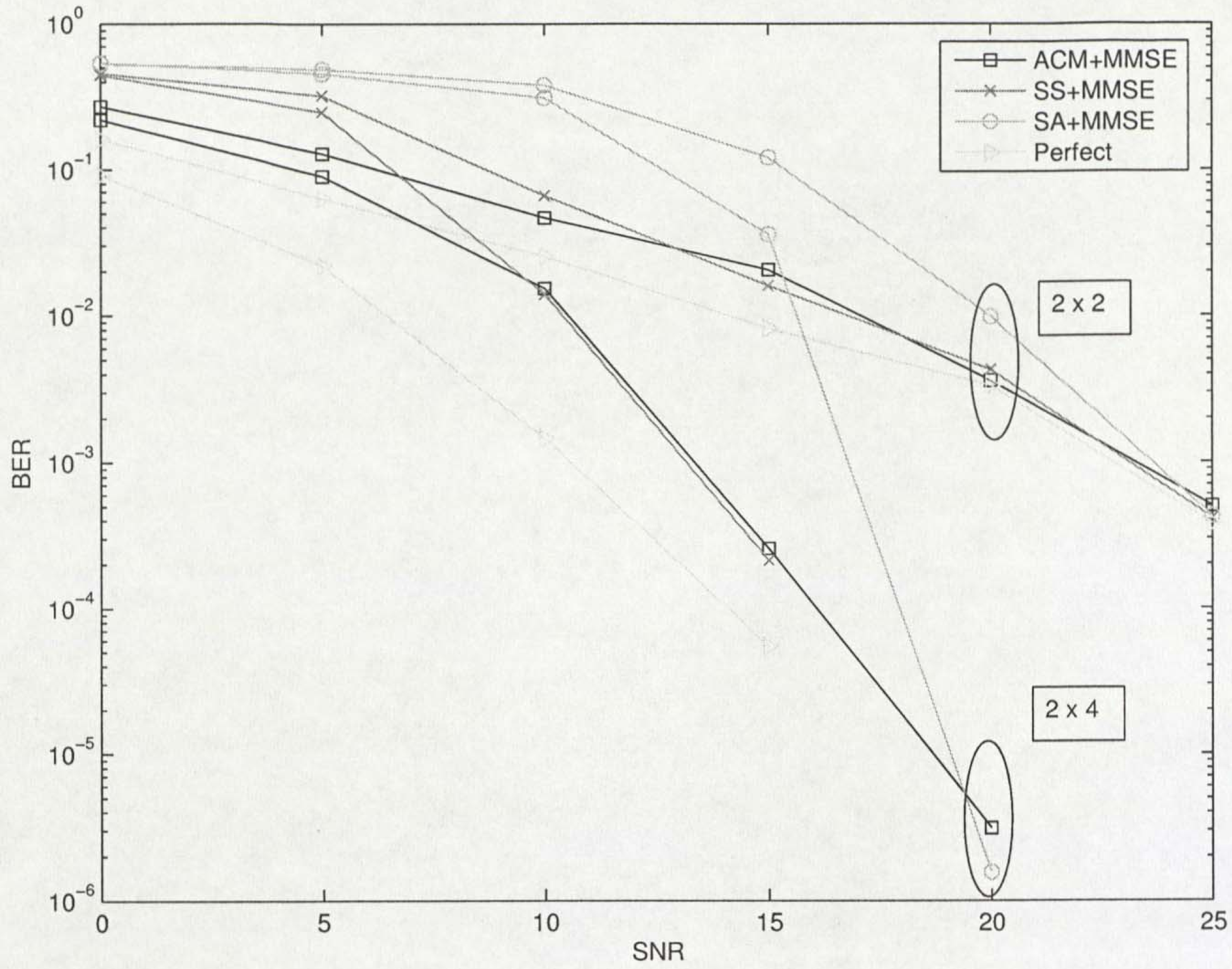

Figure 4.8: Performance of SS+MMSE, SA+MMSE and ACM+MMSE with different receive antennas for 2 transmit antenna MIMO SC-CDMA Uplink with N=32, L=4

the blind estimates as channel length increases at SNR of 20 dB. It is seen that the performance does not vary much as channel length varies. Again the improved performance of LSFE is illustrated as compared to MMSE equaliser. The SS and ACM methods provide similar performance, while the SA method is poor in comparison as illustrated in the plot.

Figure 4.10 illustrates the effect of MAI on a 2 transmit, 2 receive antenna system at the base-station. As the number of users increases, the performance gets worse, but it is clear that the ACM based estimates perform slightly better overall as compared to the subspace based estimates.

Figure 4.11 shows a simple comparison between MIMO MC-CDMA and

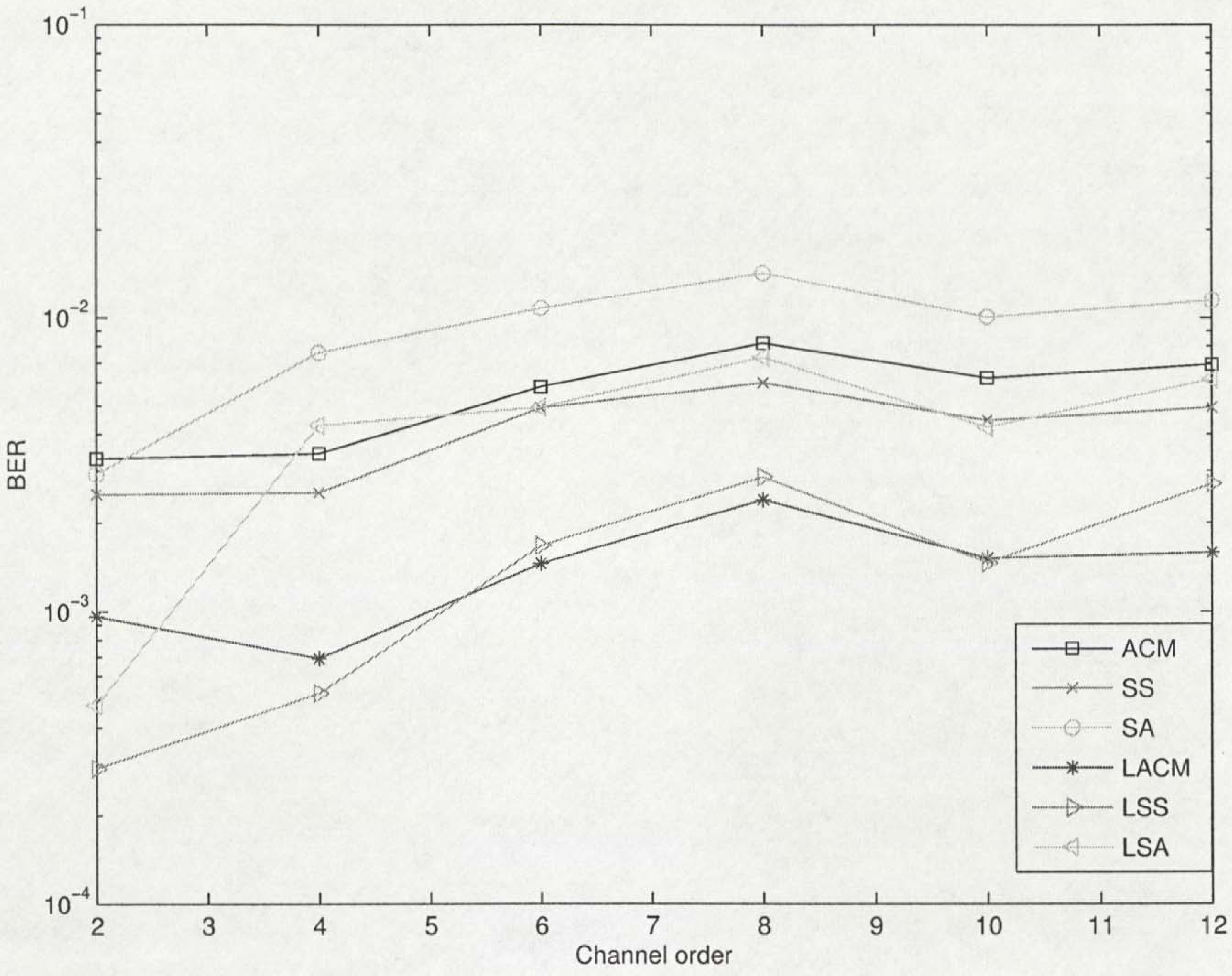

Figure 4.9: Performance of LSFE & MMSE for 2 x 2 MIMO SC-CDMA uplink system with N=32 and varying L

MIMO SC-CDMA uplink systems utilising 2 transmit antenna, 2 receive antenna setup for the same channel conditions described above. It is found that MC-CDMA outperforms SC-CDMA block transmission with FDE for this particular case of Rayleigh fading channel and using Walsh-Hadamard spreading codes. Similar results were obtained for a downlink setup between MC-CDMA using Walsh hadamard codes and DS-CDMA system using Gold codes by Prasad [22], while Giannakis et al [78] provided a comparison between MC-CDMA and DS-CDMA systems using both Gold codes and Walsh hadamard codes in frequency selective Rayleigh fading channel, which destroys code orthogonality. It was shown that performance of MC-CDMA doesn not depend on the spreading code used, while that of DS-CDMA based

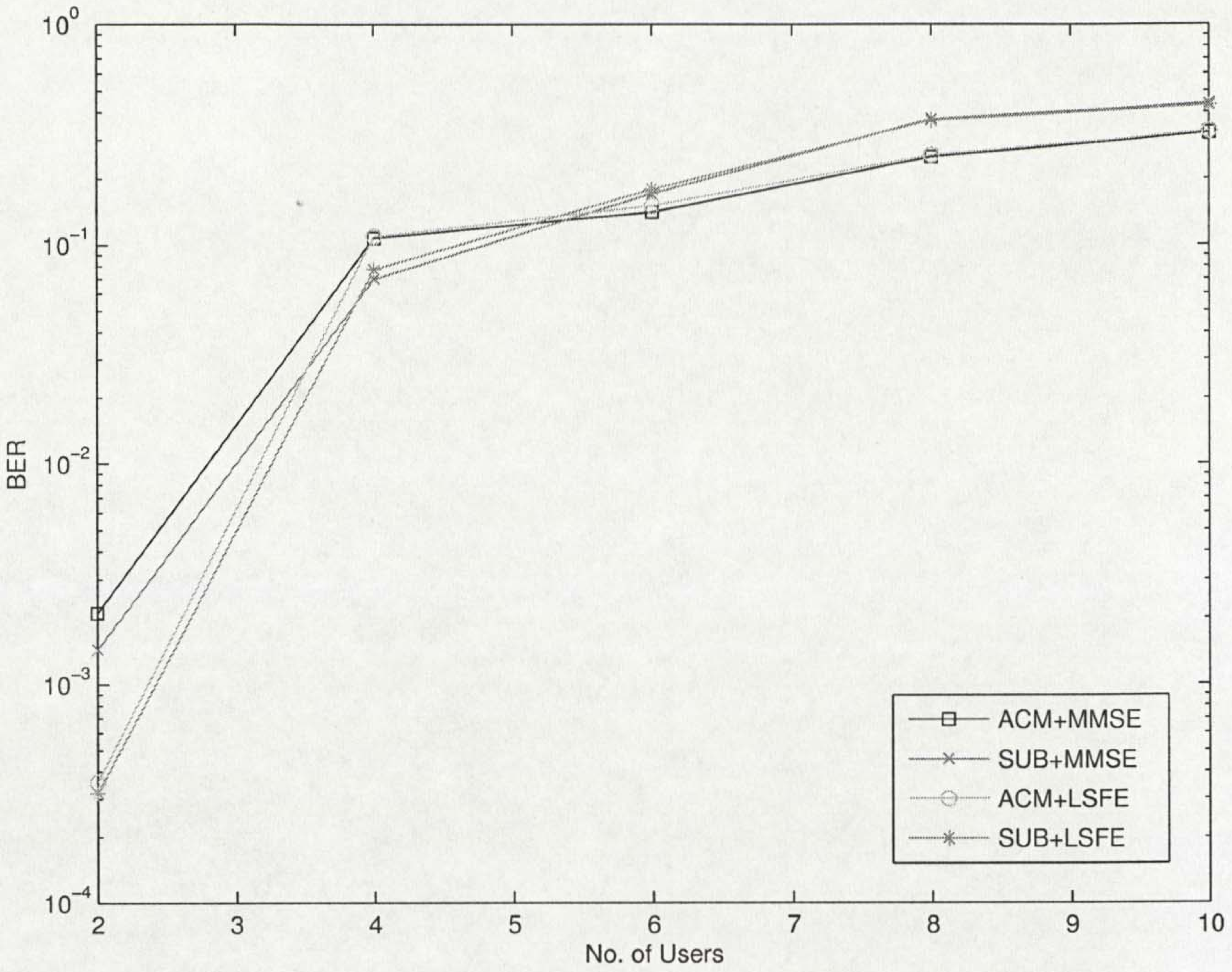

Figure 4.10: Performance of LSFE & MMSE for MIMO SC-CDMA uplink system with N=32,L=4 and varying no. of users(Nu)

schemes does. Other papers such as [34] also provide comparisons between the two using different code setups for MC-CDMA and SC-CDMA in order to obtain similar performance levels.

4.6 Complexity Analysis

The numerical complexity is presented below for the two stages involved namely the Channel estimation and Detection. In this analysis, only matrix multiplications and matrix inverse operations are considered. The notations used are

- N_r - No. of receive antennas

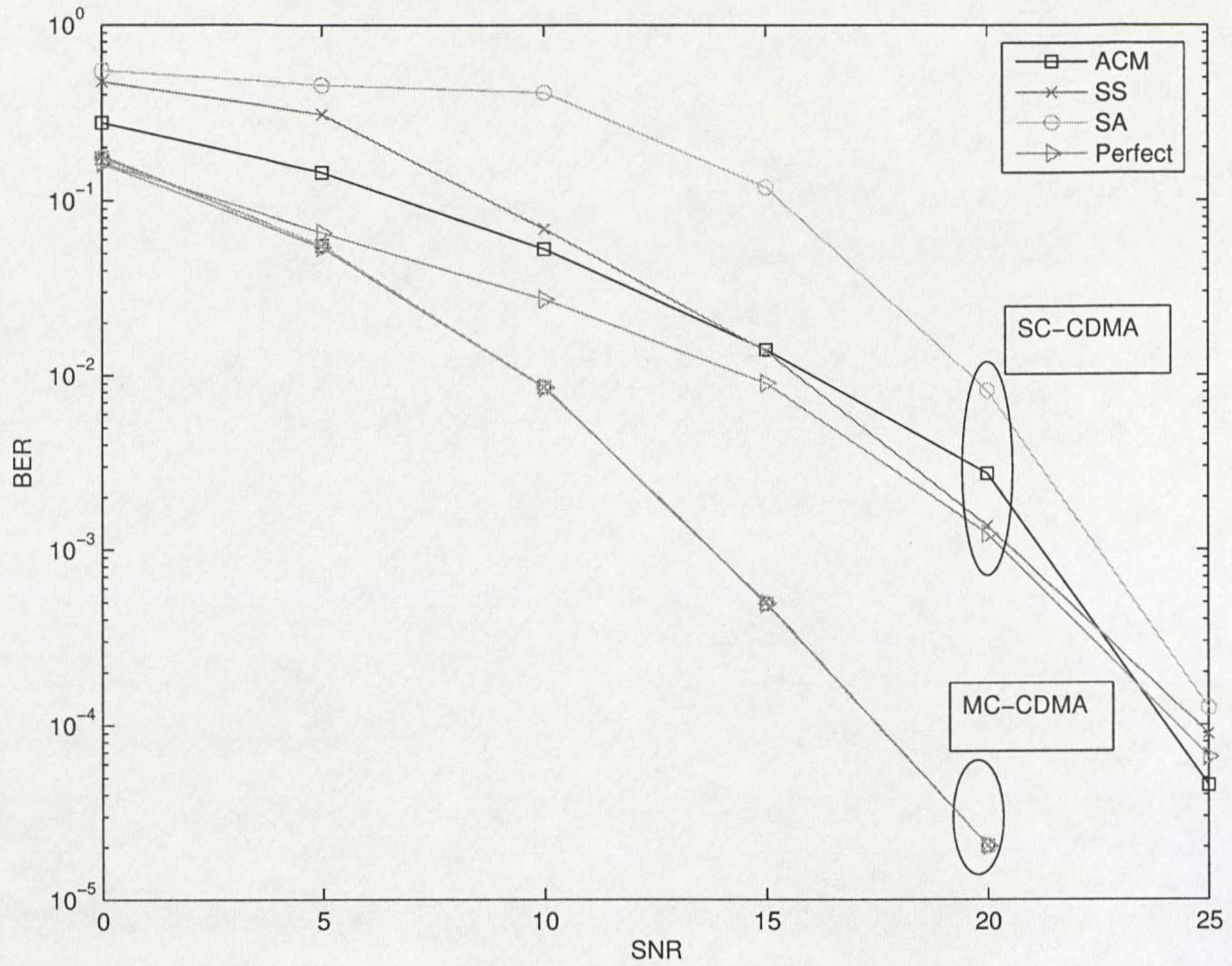

Figure 4.11: Comparison of 2x2 MIMO MC-CDMA vs MIMO SC-CDMA using Walsh-hadamard codes in Rayleigh fading channel with L=4,N=32

- N_t - No. of transmit antennas i.e number of users
- N - No. of subcarriers
- P - Length of the one block i.e. $N + L$
- L - Channel order
- N_s - No. of symbols
- h_i - Channel Impulse response of user i

4.6.1 Subspace Algorithm

The subspace channel estimation method involves

1. Correlation at each receive antenna

   The complexity of which is N^2N_s, where N_s is the number of symbols and N is the number of carriers. Thus the net computations required will be

$$Cost_R = N_rP^2N_s \tag{4.43}$$

2. The Least Squares approach to solving the Eqn. (4.11) using Eqn. (4.12) yields

$$Cost_{ls,i} = P^6L^3 \tag{4.44}$$

   Thus the total cost for all user is given as -

$$Cost_{ls} = N_tP^6L^3 \tag{4.45}$$

Thus the total cost could be written as

$$Cost_{sub} = Cost_R + Cost_{ls} \tag{4.46}$$

which is futher simplified as

$$Cost_{sub} = N_rP^2N_s + N_tP^6L^3 \tag{4.47}$$

4.6.2 SA algorithm

The SA algorithm involves the following steps

1. Correlation at each receive antenna

The complexity of which is $N^2 N_s$, where N_s is the number of symbols and N is the number of carriers. Thus the net computations required will be

$$Cost_R = N_r P^2 N_s \tag{4.48}$$

2. The Least Squares approach to solving the equation Eqn. (4.21) using Eqn. (4.20) yields

$$Cost_{ls,i} = P^6 + P^7 L^2 \tag{4.49}$$

Thus the total cost for all user is given as -

$$Cost_{ls} = N_t P^6 + N_t P^7 L^2 \tag{4.50}$$

Thus the total cost could be written as

$$Cost_{sa} = Cost_R + Cost_{ls} \tag{4.51}$$

which is futher simplified as

$$Cost_{sa} = N_r P^2 N_s + N_t P^6 (1 + PL^2) \tag{4.52}$$

4.6.3 ACM algorithm

Similarly for ACM algorithm, the following steps are involved

1. Auto-correlation for each receive antenna

The same stage as in Subspace which yields a complexity cost of

$$Cost_R = N_r P^2 N_s \tag{4.53}$$

2. Minimisation of the criterion for user 1 as given by Eqn. (4.28) yields the following complexity cost

$$Cost_{ls,1} = P^6 + P^7 L^2 + P^3 + P^7 L^2 \tag{4.54}$$

Thus cost for all users will be

$$Cost_{ls} = N_t P^3 (P^3 + 2P^4 L^2 + 1) \tag{4.55}$$

Thus the total cost can be written as

$$Cost_{acm} = N_r + N_s P^2 + N_t P^3 (P^3 + 2P^4 L^2 + 1) \tag{4.56}$$

4.6.4 Detection

The complexity cost involved for detection is explained below.

MMSE Detector

The MMSE detector is expressed as

$$w_{mmse,i} = R^{-1} e C_i \tag{4.57}$$

with $w_{mmse,i}$ is the equaliser vector for user i and eC_i representing the effective signature for user i.

$$Cost_{wmmse,i} = N_r^2 N^3 + N_r^3 N^3 + N_r^2 N_t N^3 \tag{4.58}$$

$$Cost_{wmmse} = N_t \times Cost_{wmmse,i} \tag{4.59}$$

yielding

$$Cost_{wmmse} = N_t N_r^2 N^3 (N_t + 1 + N_r N) \tag{4.60}$$

LSFE

The stages involved in LSFE are

1. Ordering of users/streams based on mse. The main equation here is

$$Cost_{lsfe1} = N_r^3 + N_r^4 \tag{4.61}$$

2. Detection of ordered stream using MMSE for each user /stream This would be the same value as above but multipleid by N_t times for each user.

$$Cost_{lsfe2} = N_t^2 N_r^2 N^3 (N_t + 1 + N_r N) \tag{4.62}$$

3. Cancellation of detected stream for all $N_t - 1$ stages

$$Cost_{lsfe3} = (N_t - 1) N_r N N_s \tag{4.63}$$

Therefore the total cost of LSFE can be written as

$$Cost_{lsfe} = Cost_{lsfe1} + Cost_{lsfe2} + Cost_{lsfe3} \tag{4.64}$$

Table 4.1: Channel Estimation Complexity for MIMO SC-CDMA Uplink

Cost	Subspace	ACM	SA
$Cost_R$	$N_r P^2 N_s$	$N_r P^2 N_s$	$N_r P^2 N_s$
$Cost_{ls}$	$N_t P^6 L^3$	$N_r + N_s P^2 + N_t P^3 (P^3 + 2P^4 L^2 + 1)$	$N_t P^6 + N_t P^7 L^2$
$Total_{cost}$	$N_r P^2 N_s + N_t P^6 L^3$	$N_r + N_s P^2 + N_t P^3 (P^3 + 2P^4 L^2 + 1)$	$N_r P^2 N_s + N_t P^6 (1 + PL^2)$

Table 4.2: Detection Complexity for MIMO SC-CDMA Uplink

MMSE	LSFE
$N_t N_r^2 N^3 (N_t + 1 + N_r N)$	$N_r (N_r^2 + N_r^3 + N_t^2$ $N_r N^3 (N_t + N_r N + 1) + (N_t - 1) N N_s$

4.7 Summary

Research into SC-CDMA block transmission with GB has grown as a result of the PAPR, frequency offset and synchronisation problems afflicting MCM schemes such as OFDM, MC-CDMA etc. This chapter proposed three blind channel estimation methods for the MIMO SC-CDMA Uplink case, namely the SS based approach, the SA method and the ACM method. The ACM method provides similar performance to the SS approach with the added advantage that it eliminates the need for an initial EVD, rank determination and noise power estimation. The SA approach provided a similar level of performance to the SS method at high SNR but again like the ACM method, required no initial EVD stage though rank order was required. In addition, an LSFE equaliser was integrated with the blind channel estimation stage and its improved performance as compared to that of MMSE equaliser was highlighted.

As illustrated in the simulation results, the LSFE equaliser performs better than the MMSE scheme due to the interference cancellation of the detected signals at each stage. The effect of load (no. of users) as well as varying channel length was also examined and it was shown that the LSFE maintains superior performance compared to MMSE equaliser. Also to be noted is the insensitive nature of the SC-CDMA system to varying channel length. The performance gains obtained via receive diversity are also shown and a complexity analysis of the three blind channel estimation methods was also performed and tabulated. A simple comparison between MIMO SC-CDMA block transmission and MIMO MC-CDMA systems for uplink scenario revealed the dependence of SC-CDMA block schemes on the type of spreading codes [31], [78]. Thus in a frequency selective uplink channel model, the MIMO MC-CDMA scenario provides better performance than the equivalent SC-CDMA block transmission scheme.

PAPR Peak to Average Power Ratio

pMMSE per-carrier MMSE

pMRC per-carrier MRC

SA Subspace Approximation

SC-CDMA Single-Carrier Code Division Multiple Access

SDMA Space Division Multiple Access

SIC Successive Interference Cancellation

SIC VBLAST Successive Interference Cancellation VBLAST

SINR Signal to Interference and Noise Ratio

SNR Signal to Noise Ratio

SOS Second Order Statistics

SS Subspace

STBC Space Time Block Code

TDE Time Domain Equaliser

V-BLAST Vertical Bell Labs Layered Space Time

VLSI Very Large Scale Integrated Circuit

WCDMA Wideband CDMA

ZF Zero Forcing

Chapter 5

Per-Carrier Equalisation

5.1 Introduction

This chapter examines an alternative to the block-wise equalisation schemes considered so far throughout the thesis. The MMSE and LSFE equalisers considered so far equalised the received signal block-wise over all subcarriers. Under the assumptions that fading is uncorrelated between sub-carriers and each subcarrier is attenuated independently of its adjoining subcarriers [82], the distinct frequency bins can be equalised on a per-carrier basis.

The use of Maximal Ratio Combining (MRC), Equal Gain Combining (EGC) and Orthogonality Restoring Combining (ORC) equalisation techniques [31] are examined for both MIMO MC-CDMA and MIMO SC-CDMA block transmission systems in the uplink scenario. Their performances are compared with the conventional MMSE method for both block based and per-carrier based processing. Per-carrier processing has already been applied to OFDM systems [83], where a combined OFDM / SDMA approach was used with per-carrier SIC in conjunction with MMSE equaliser to obtain performance improvements.

In the next section, the MIMO MC-CDMA Uplink scenario, as described in Chapter 3 is equalised using different per-carrier equaliser schemes.

5.2 MIMO MC-CDMA Uplink with Per-Carrier FDE

5.2.1 Algorithm Description

The concept of per-carrier processing was proposed for MC-CDMA downlink by Valkama [84], while Petre et al [85] proposed the use of a two stage multiuser detection scheme for the uplink MC-CDMA systems. Proakis et al [86] dealt with the adaptive MMSE detector for the downlink MC-CDMA system utilising Least Mean Square (LMS) and Recursive Least Squares (RLS) gradient descent algorithms on a per-carrier basis. The use of per-carrier processing was investigated in a MIMO MC-CDMA Uplink scenario by Latva-aho et al [87], where the use of per-subcarrier based processing was done along with PIC to improve performance with the method regarded the subcarriers at different receive antennas as additional subcarriers. It did not group them to yield a MIMO system per subcarrier resulting in degraded performance compared to per-user block based equalisers [87]. In the setup here, each subcarrier yields a $N_t \times N_r$ system, which provides improved performance as well as lower complexity.

The system model setup is exactly the same as in Chapter 3. N_t and N_r

denote the number of transmit and receive antennas respectively and N_t is equal to number of users. L is the channel order, which is assumed to be known at receiver. The spreading codes used are Walsh-Hadamard codes of size N, where N is the number of subcarriers. Here the spreading gain is chosen to be equal to the number of subcarriers and BPSK modulation is used for each user.

At the j^{th} receive antenna, after removal of the guard band and demodulation using FFT, the received signal at the j^{th} receive antenna on the n^{th} subcarrier denoted by $\mathbf{X}_j[n]$ can be expressed as in Chapter 3 Eqn. (3.1)

$$\mathbf{X}_j[n] = \sum_{i=1}^{N_t} \mathbf{H}_{j,i}[n]\mathbf{d}_i[n] + \mathbf{V}_j[n] \tag{5.1}$$

where $\mathbf{H}_{ji}[n]$ is the channel frequency response matrix on the n^{th} subcarrier between the j^{th} and the i^{th} transmit antennas, $\mathbf{V}_j[n]$ is the noise on the n^{th} subcarrier at the j^{th} receive antenna. $\mathbf{d}_i[n]$ is the i^{th} user's spread signal on the n^{th} subcarrier given by $\mathbf{d}_i[n] = c_i[n]s_i[n]$, where $c_i[n]$ is the n^{th} bit of the i^{th} user's spreading code sequence $\mathbf{c}_i = [c_i(1)\ldots c_i(N)]^T$ of size $N \times 1$. $s_i[n]$ is the ith user's data bit transmitted on subcarrier n.

Thus the ith user's net spread signal can be represented as $\mathbf{x}_i = [x_i(1)x_i(2)\ldots x_i(N)]$ $= \mathbf{C}_i\mathbf{s}_i$, where $\mathbf{C}_i = \mathbf{diag}[c_i(1)\ldots c_i(N)]$ is the diagonal spreading code matrix of size $N \times N$. The net received signal $\mathbf{X}[n]$ on subcarrier n can thus be represented as

$$\mathbf{X}[n] = \sum_{i=1}^{N_t} \mathbf{H}_i[n]\mathbf{d}_i[n] + \mathbf{V}[n] \tag{5.2}$$

where $\mathbf{X}[n] = [\mathbf{X}_1^T[n]\mathbf{X}_2^T[n]\ldots\mathbf{X}_{N_r}^T[n]]$. $\mathbf{H}_i[n]$, the $N_r \times 1$ vector of channel frequency response values between the i^{th} user and the N_r receive antennas is defined as

$$\mathbf{H}_i[n] = [H_{1,i}[n] \ldots H_{N_r,i}[n]]^T \tag{5.3}$$

and $\mathbf{V}[n] = [\mathbf{V}_1[n] \ldots \mathbf{V}_{N_r}[n]]$ is the noise on subcarrier n.

In Chapter 3, the use of MMSE equaliser was described on per-block basis utilising the entire channel matrix. Here, the use of three other equaliser schemes namely the MRC (Maximal Ratio Combining), EGC (Equal Gain Combining) and ORC (Orthogonality Restoring Combining) [31] as well as per-carrier MMSE are described for the MIMO uplink scenario using MC-CDMA. These equalisation schemes combine the energy over all subcarriers and thereby provide frequency diversity. The generic form of the per-carrier equalisation can be represented as follows

$$\mathbf{q}[n] = \mathbf{w}[n]\mathbf{X}[n] \tag{5.4}$$

where $\mathbf{q}[n]$ represents the respective equalised signal on subcarrier n using the weight vector $\mathbf{w}$ based on either MRC, EGC,ORC or MMSE scheme. These weight vectors are obtained as described below.

Using the system model Eqns. (5.1) and (5.2), the MRC weight per carrier for user i are given as [31]

$$\mathbf{w}_{mrc,i}[n] = \mathbf{H}_i[n]^H \tag{5.5}$$

where $\mathbf{w}_{mrc,i}[n]$ denotes the MRC weight vector for user i on the n^{th} subcarrier. The corresponding EGC, ORC and MMSE equaliser weights for the n^{th} subcarrier are given as

$$\mathbf{w}_{egc,i}[n] = \frac{\mathbf{H}_i[n]^H}{||\mathbf{H}_i[n]||} \tag{5.6}$$

$$\mathbf{w}_{orc,i}[n] = \frac{\mathbf{H}_i[n]^H}{||\mathbf{H}_i[n]||^2} \tag{5.7}$$

$$\mathbf{w}_{mmse,i}[n] = \mathbf{H}_i[n]^H \left(\sum_{i=1}^{N_t} \mathbf{H}_i[n]\mathbf{H}_i[n]^H + \sigma^2 \mathbf{I}_{N_r} \right)^{-1} \tag{5.8}$$

with $\mathbf{I}_{N_r}$ representing an $N_r \times N_r$ identity matrix. After equalising over all subcarriers and all users, the resultant signal is then despread using the respective user's spreading codes to obtain the estimates of transmitted signal.

5.2.2 Simulation Results

The simulation setup, as in Chapter 3, involved a simple 2 transmit, 2 receive antenna uplink MIMO MC-CDMA system unless otherwise specified. The number of subcarriers was set to $N = 32$ with channel modelled as quasi-static block Rayleigh fading channel, which remains constant over the duration of a block. Also independent fading is assumed between the subcarriers. The channel impulse length was set to $L = 4$ with uniform power delay profile. A block size of $N_s = 100$ frames was used throughout.

A pilot symbol is used for each transmit-receive antenna pair to account for the scalar ambiguity as a result of the blind channel estimation schemes used. Walsh-Hadamard codes, of size N i.e.equal to the number of subcarriers, were used throughout because of their orthogonal nature. The BER results are obtained as averaged over 100 Monte-Carlo runs for the setup described. BPSK modulation was used throughout. 'SS', 'SA' and 'ACM' represent the Subspace, Subspace Approximation and ACM method respectively, while 'Perfect' denotes the case with perfect channel knowledge at the receiver.

Figure 5.1 shows the improvement in performance of per-carrier based equalisation schemes with SS based estimates, obtained as a result of receive diversity . As is seen in figure, MRC based equaliser performs slightly better than MMSE, EGC and ORC for 2 transmit, 2 receive antenna case when using the SS based estimates. The difference in performance is slightly bigger for the 2 transmit, 4 receive antenna case with the MRC being the best of the bunch.

Figure 5.2 illustrates two important points. One is the very good performance of the blind channel estimates using ACM, SS or SA based methods as compared to the Perfect channel case and the other is the improvement in performance obtained as a result of receive diversity as highlighted in the plot earlier.

Figure 5.3 compares the performance of the block based equalisers with the per-carrier version using MRC equaliser. It is seen that at SNR of 15 dB, the per-carrier based equalisation schemes provide nearly 2 orders of magnitude improvement in performance, when compared to block based schemes.

Figure 5.4 compares the performance of the 4 equalisation schemes used versus the number of users in the uplink scenario. As is seen, the MRC, EGC and ORC equalisers provide better performance than MMSE at lower loads,

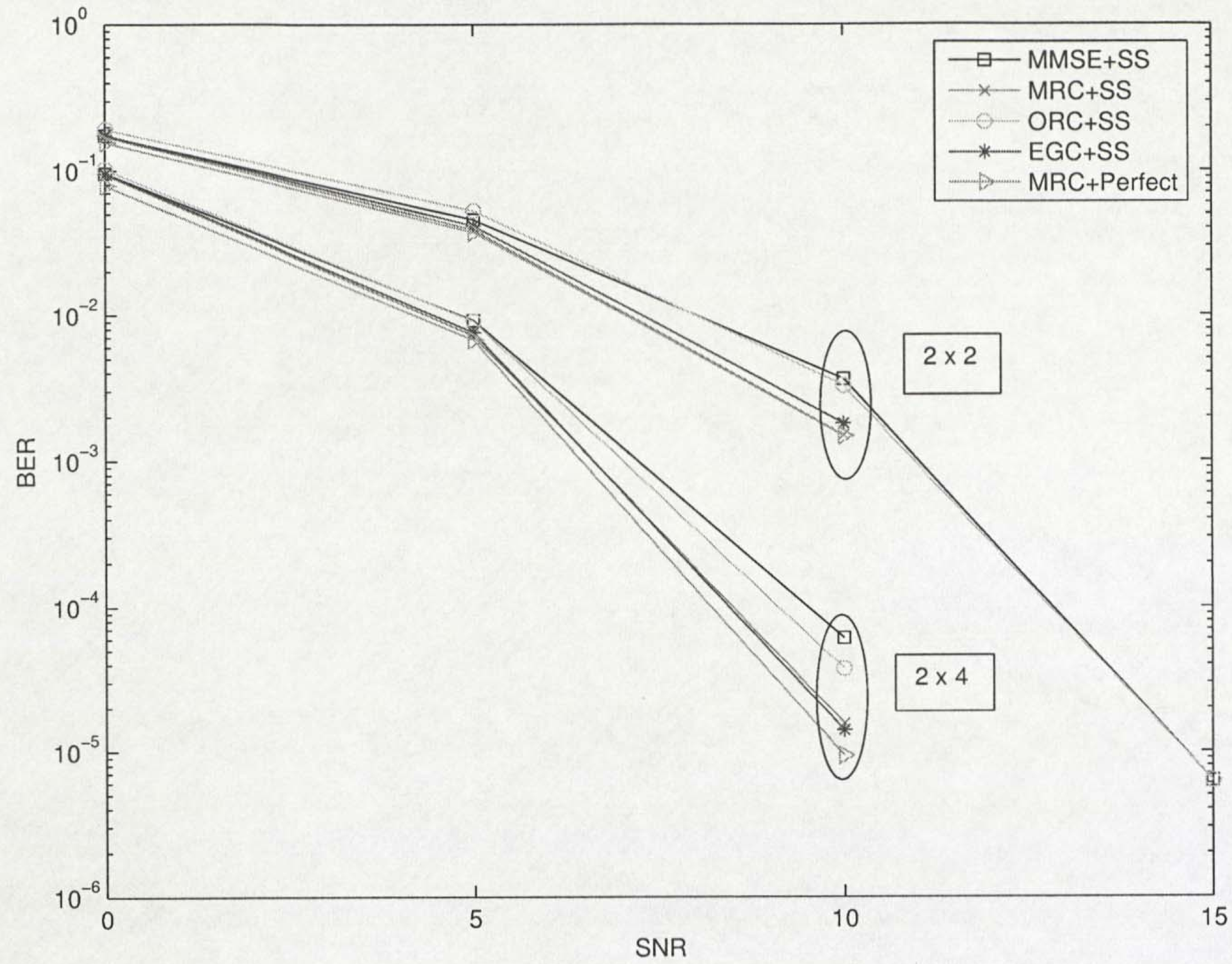

Figure 5.1: Performance of Per-carrier Equalisers using SS estimates for 2 transmit and 2,4 receive antenna case

but as the number of users increases, the MAI increases and the performance degrades equally for all equalisation schemes.

In the next section, per-carrier equalisation is investigated for the MIMO SC-CDMA Uplink block transmission scenario.

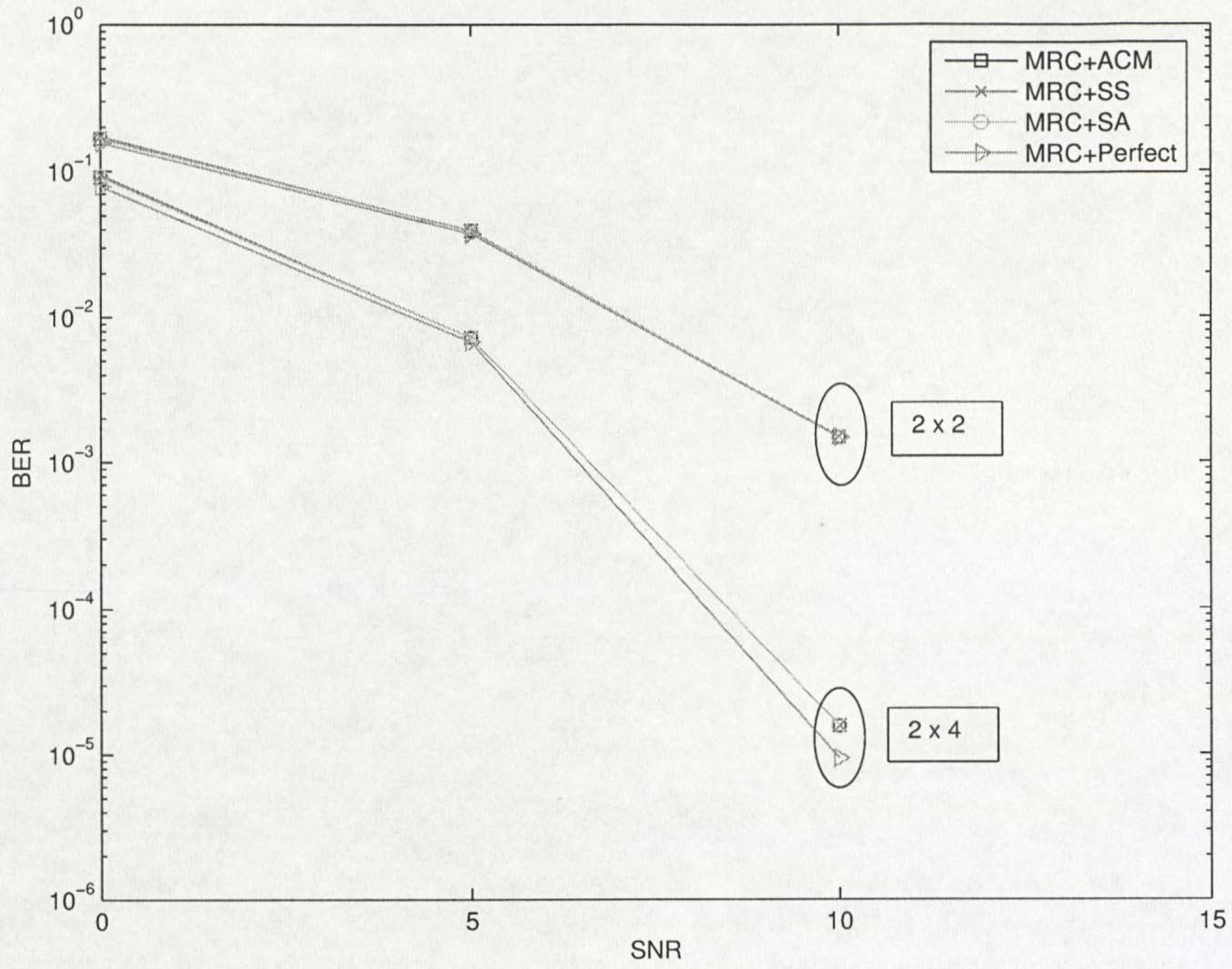

Figure 5.2: Performance of per-carrier MRC equaliser using ACM,SS and SA based estimates for 2 transmit and 2,4 receive anntenna case

5.3 MIMO SC-CDMA block uplink transmission with Per-Carrier FDE

5.3.1 Algorithm Description

In this section, the per-carrier FDE of MIMO SC-CDMA block transmission sytem, described in Chapter 4, is investigated. The MRC, EGC, ORC and MMSE per carrier equalisation schemes are examined and their performances compared using the same simulation setup as before. Using the system model as described in Chapter 4, Eqn. (4.1) the received signal on the n^{th} subcarrier, denoted by $\mathbf{X}_j[n]$, after removal of GB and application of FFT is given as

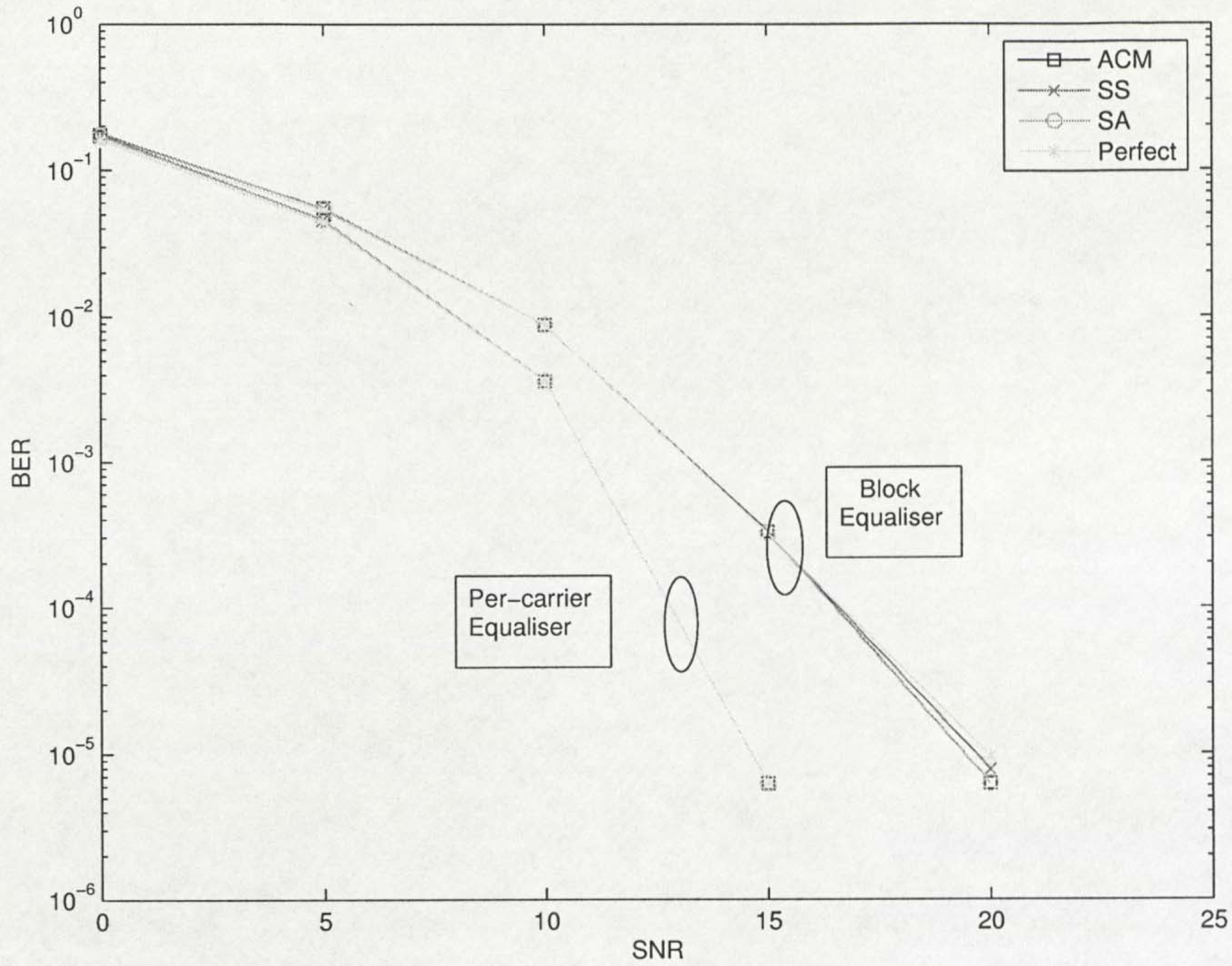

Figure 5.3: Comparison of block based and per-carrier MRC equalisers using ACM, SS and SA based estimates for 2 transmit, 2 receive antenna case

$$\mathbf{X}_j[n] = \sum_{i=1}^{N_t} \mathbf{H}_{j,i}[n] c_i[n] s_i[n] + \mathbf{V}_j[n] \tag{5.9}$$

where $\mathbf{H}_{j,i}[n]$ is the fourier transform coefficient of the channel matrix between transmit antenna i and receive antenna j on sub-carrier n. $c_i[n]$ represents the n^{th} value of the spreading code for user i, while $s_i[n]$ is the copy of the data bit transmitted on subcarrier n and $\mathbf{V}_j[n]$ denotes the noise at the j^{th} receive antenna on the n^{th} subcarrier.

The net received signal $\mathbf{X}[n]$ on the n^{th} subcarrier can be expressed as shown below, with $\mathbf{H}_i[n]$ denoting the channel frequency response matrix between the i^{th} transmit antenna and all receive antennas, expressed as

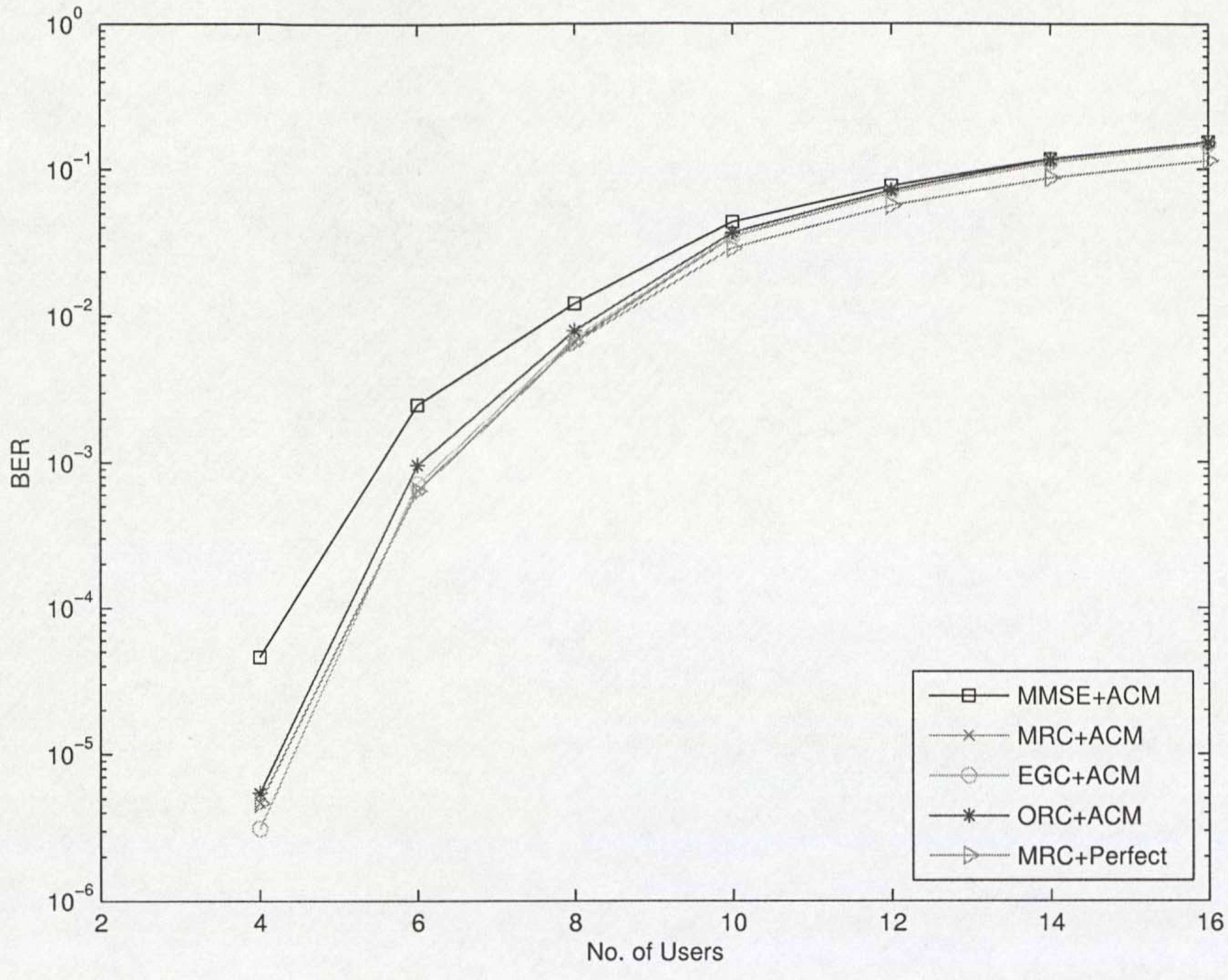

Figure 5.4: Performance of different per-carrier equalisers for varying number of users

$\mathbf{H}_i[n] = [\mathbf{H}_{1,i}[n] \ldots \mathbf{H}_{N_r,i}[n]]^T$.

$$\mathbf{X}[n] = \sum_{i=1}^{N_t} \mathbf{H}_i[n] c_i[n] s_i[n] + \mathbf{V}[n] \tag{5.10}$$

where $\mathbf{V}[n] = [\mathbf{V}_1[n] \ldots \mathbf{V}_{N_r}[n]]^T$ is the noise on the n^{th} subcarrier. As with the MC-CDMA setup earlier, the per-carrier equalisation schemes utilising MRC, EGC and the ORC are considered along with the conventional MMSE [34] for MIMO SC-CDMA uplink block transmission system. The per-carrier equalisation is the same as for MC-CDMA setup except that the IFFT is done after the equalisation in SC-CDMA systems. Thus the equations for the weight vectors corresponding to the different equaliser schemes considered

ZP Zero Padded

ZP-OFDM Zero Padded OFDM

will also be similar to the MC-CDMA case.

The equalised signals are then collected over all subcarriers, which is then IFFT processed as mentioned earlier. After the IFFT stage, the signals are then despread using each user's spreading code and then passed to the hard decision detector to obtain the estimates of the transmitted signal.

5.3.2 Simulation Results

The same simulation setup as in Chapter 4 for MIMO SC-CDMA Uplink block equalisation was used here. It involved an uplink MIMO SC-CDMA system with 2 transmit antennas and 2 receive antennas unless otherwise specified. The number of subcarriers was set to $N = 32$ with channel modelled as quasi-static block Rayleigh fading channel, which remains constant over the duration of a block. The channel impulse length was set to $L = 4$ with uniform power delay profile and rms delay spread of $1.153\mu s$. A block size of $N_s = 100$ frames was used throughout. As blind channel estimation methods yield estimates upto a scalar factor, a pilot symbol was used to obtain the scaling factor and one pilot symbol was used for each transmit-receive antenna pair. Walsh-Hadamard codes, of size N, i.e. equal to the number of subcarriers, were used throughout because of their orthogonal nature. The BER results are obtained as averaged over 1000 Monte-Carlo runs for the setup described. BPSK modulation was used for in all the simulations. 'SS', 'SA' and 'ACM' represent the Subspace, Subspace Approximation and ACM methods respectively with 'Perfect' denoting the perfect channel case.

Figure (5.5) compares the performance of the per-carrier MRC equaliser using ACM,SS and SA based channel estimates for a 2 transmit, 2 receive antenna setup. As illustrated, the ACM performs the best at lower SNR, with the SA based method being the worst. At high SNR, the performance is very near to that obtained using Perfect Channel knowledge.

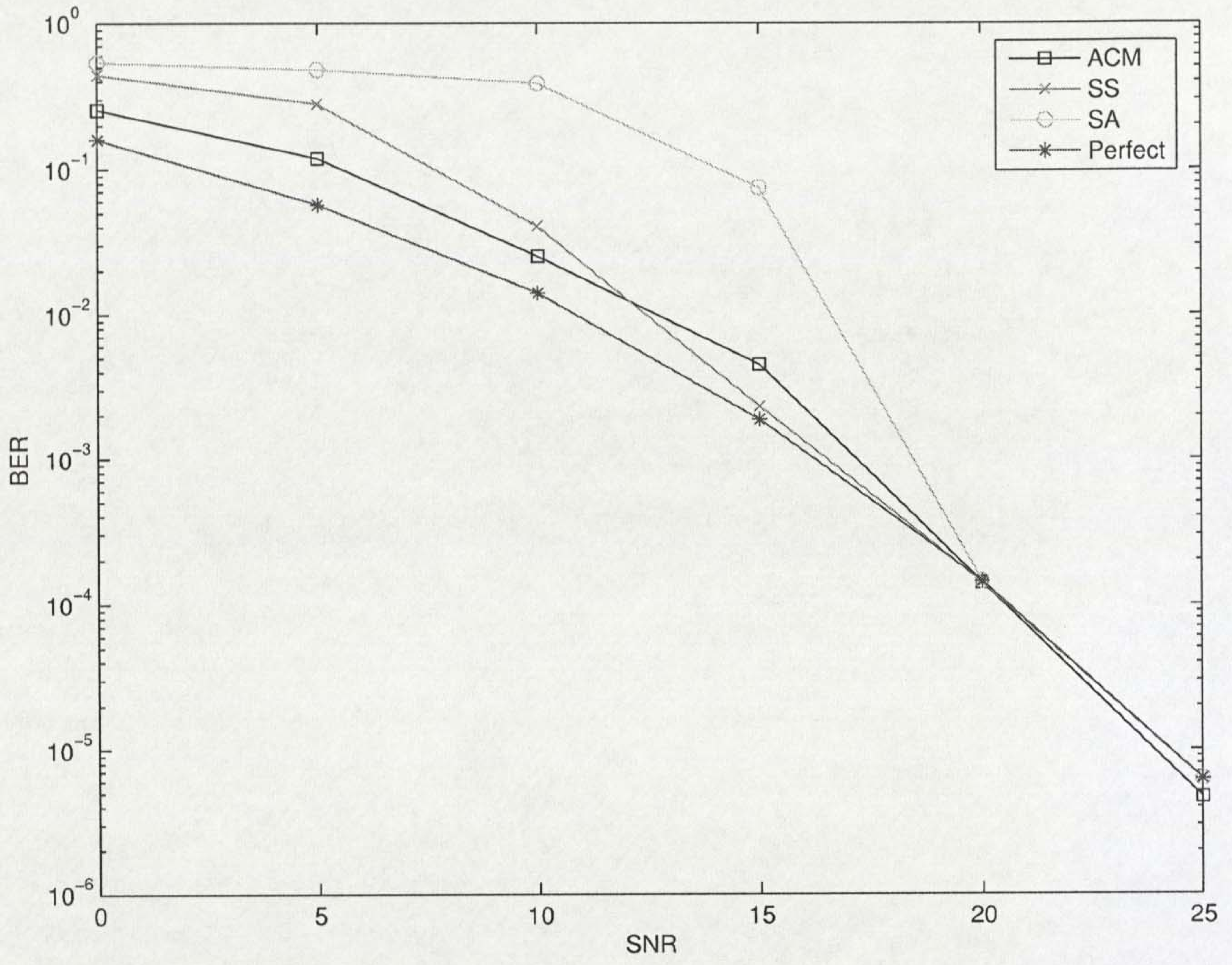

Figure 5.5: Comparison of per-carrier MRC equaliser using ACM, SS and SA based estimates for 2 transmit, 2 receive antenna case

Figure 5.6 shows the performance comparison between block based and per-carrier based MRC equaliser utilising ACM, SS, SA estimates as well as Perfect channel case. At SNR of 20dB, an order of magnitude improvement is obtained using the per-carrier equaliser as compared to the block based scheme. Another key thing to note is that the 3 algorithms perform differently at low SNR, with the SA estimate based method being the worst while the ACM providing the best performance out of the three methods.

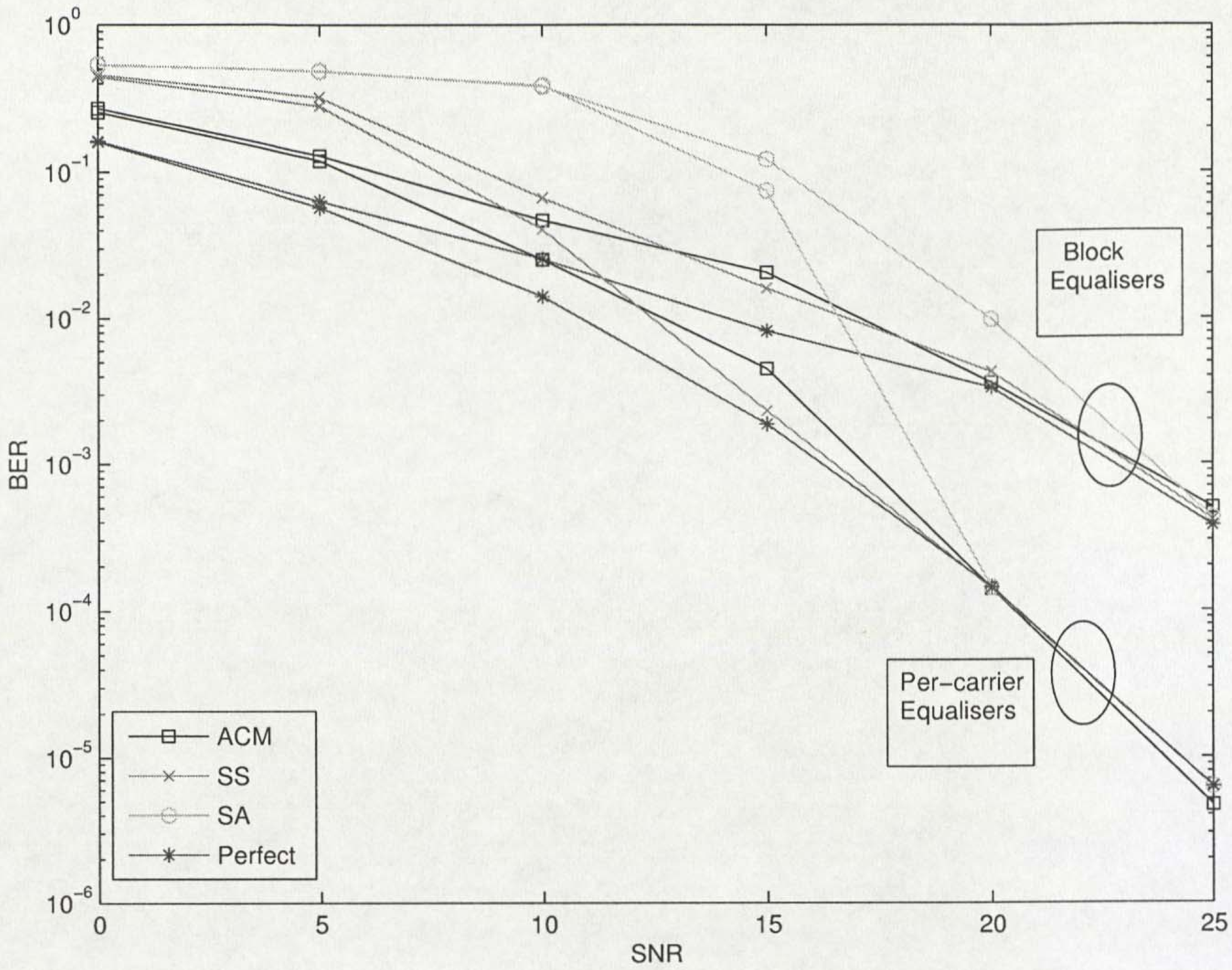

Figure 5.6: Comparison of block based and per-carrier MRC equalisers using ACM, SS and SA based estimates for 2 transmit, 2 receive antenna case

Figure 5.7 illustrates the performance of the different per-carrier equalisers using ACM estimates for varying number of users. As is shown, the performance is worsened as a result of increasing MAI (i.e. increase in number of users) and all the equalisers provide similar performance levels. Also there is quite a difference in performance between the blind channel estimate based per-carrier equalisers and the perfect channel scenario.

Figure 5.8 illustrates the performance improvements obtained via receive diversity. As shown, the various per-carrier equalisers provide similar performance levels. At SNR of 15 dB, the 2 transmit, 4 receive antenna case provides nearly 2 orders of magnitude improvement in performance as compared to the 2 transmit, 2 receive antenna case. The perfect channel case

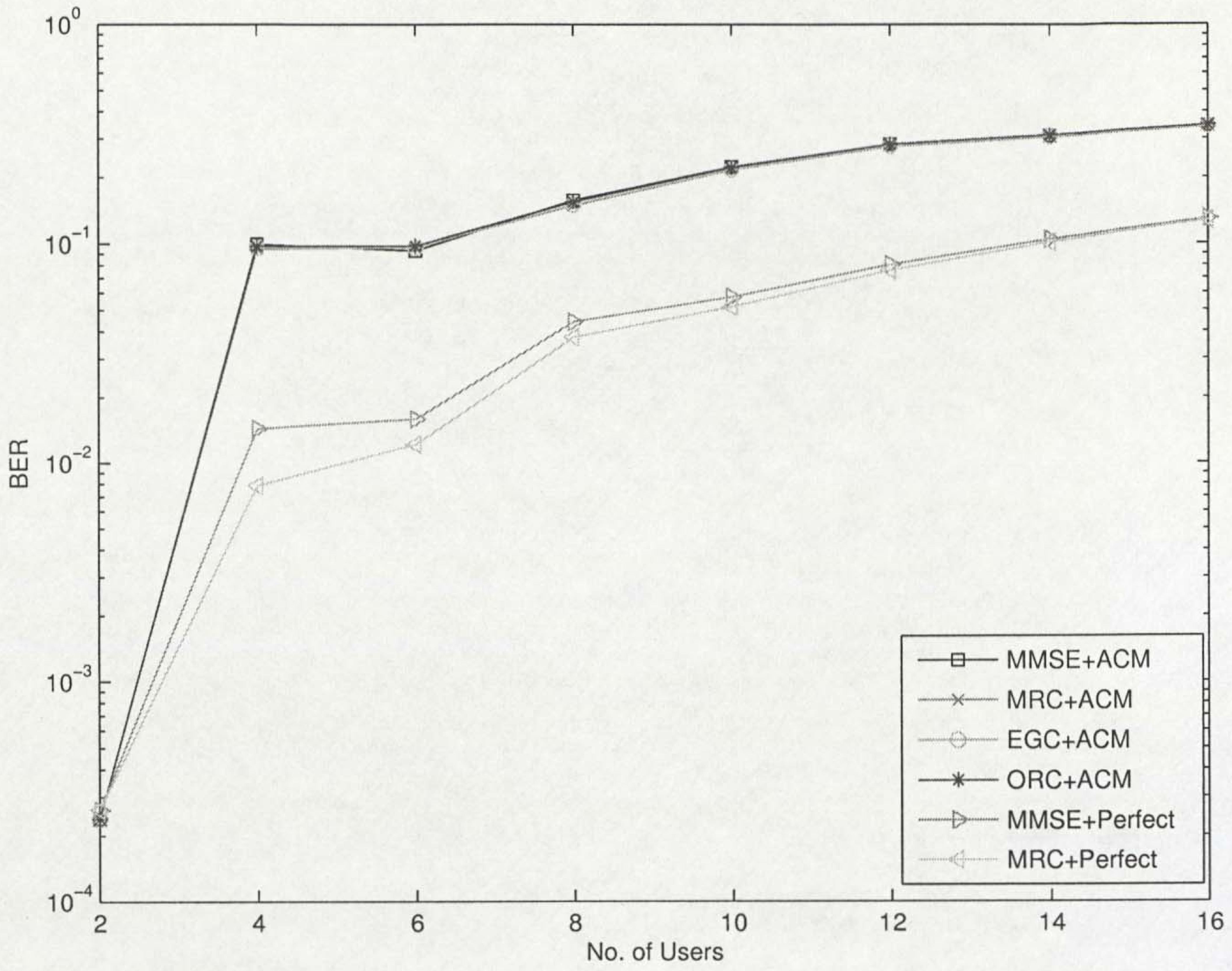

Figure 5.7: Performance of different per-carrier equalisers for varying number of users

is also plotted to provide a comparison between the blind channel estimates based performance and when perfect channel knowledge is available at the receiver.

Figure 5.9 plots the comparison between the per-carrier based MRC and MMSE equalisers for 2 transmit antenna, 2 receive antenna MIMO MC-CDMA and MIMO SC-CDMA uplink schemes described above. The ACM based estimates are used along with per-carrier equalisers and it is seen that as was the case with block-based equalisers in the last chapter, MC-CDMA scheme outperforms SC-CDMA in the Rayleigh fading channel model with Walsh-hadamard spreading codes used. This is because the performance of SC-CDMA with block transmission systems is affected by choice of spreading

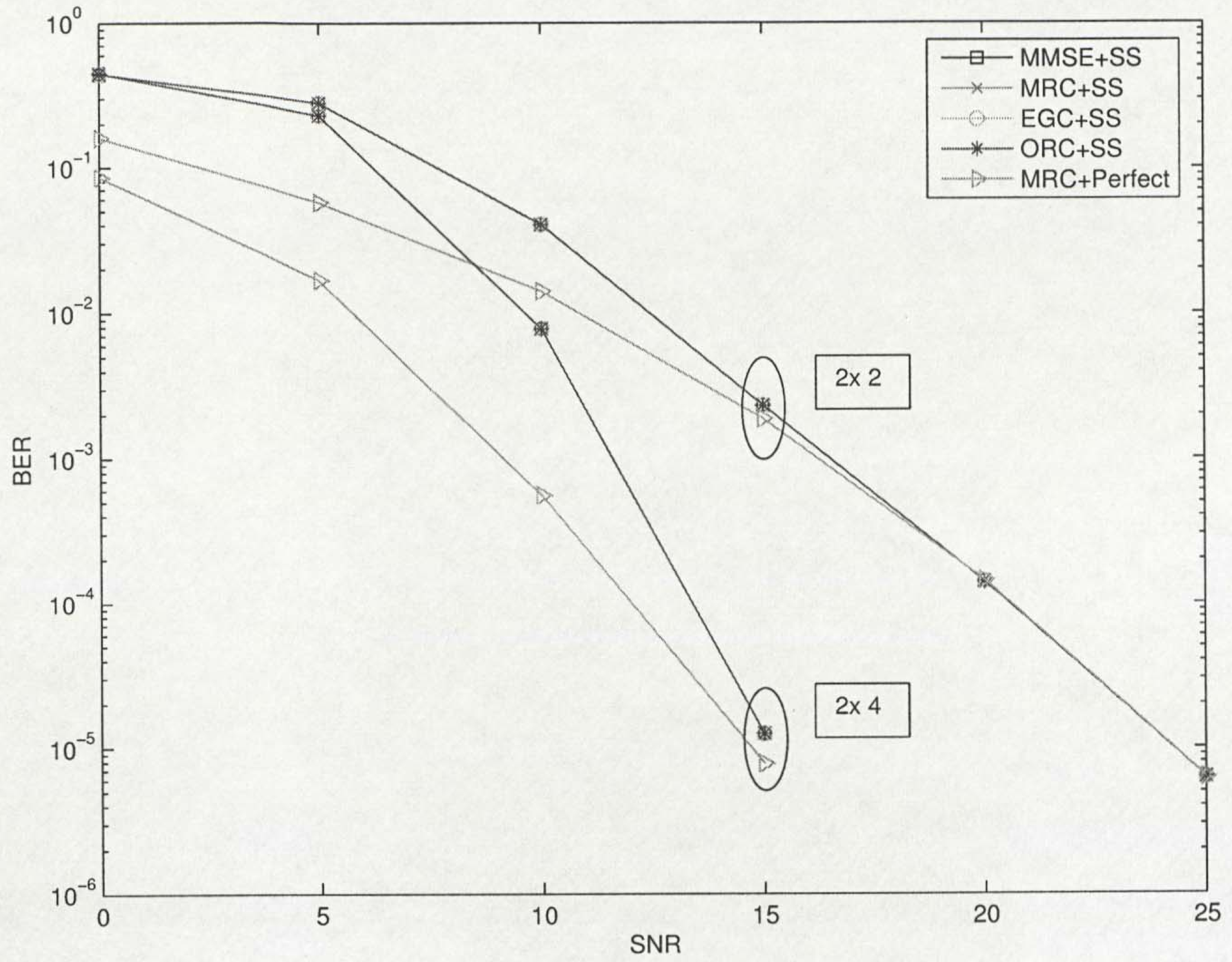

Figure 5.8: Performance of Per-carrier Equalisers using SS estimates for 2 transmit and 2,4 receive antenna case

codes used, while performance of MC-CDMA is independent of spreading code utilised [78].

5.4 Complexity Analysis

As the MIMO MC-CDMA and the MIMO SC-CDMA block transmission with FDE only differ in the IFFT processing stage, the complexity cost is the same for both systems. In the following analysis, the complexity costs involved in computing the MMSE and MRC per-carrier equaliser schemes are given below. The EGC and ORC schemes yield the same matrix computations as the MRC with the addition of a scalar division by the norm and

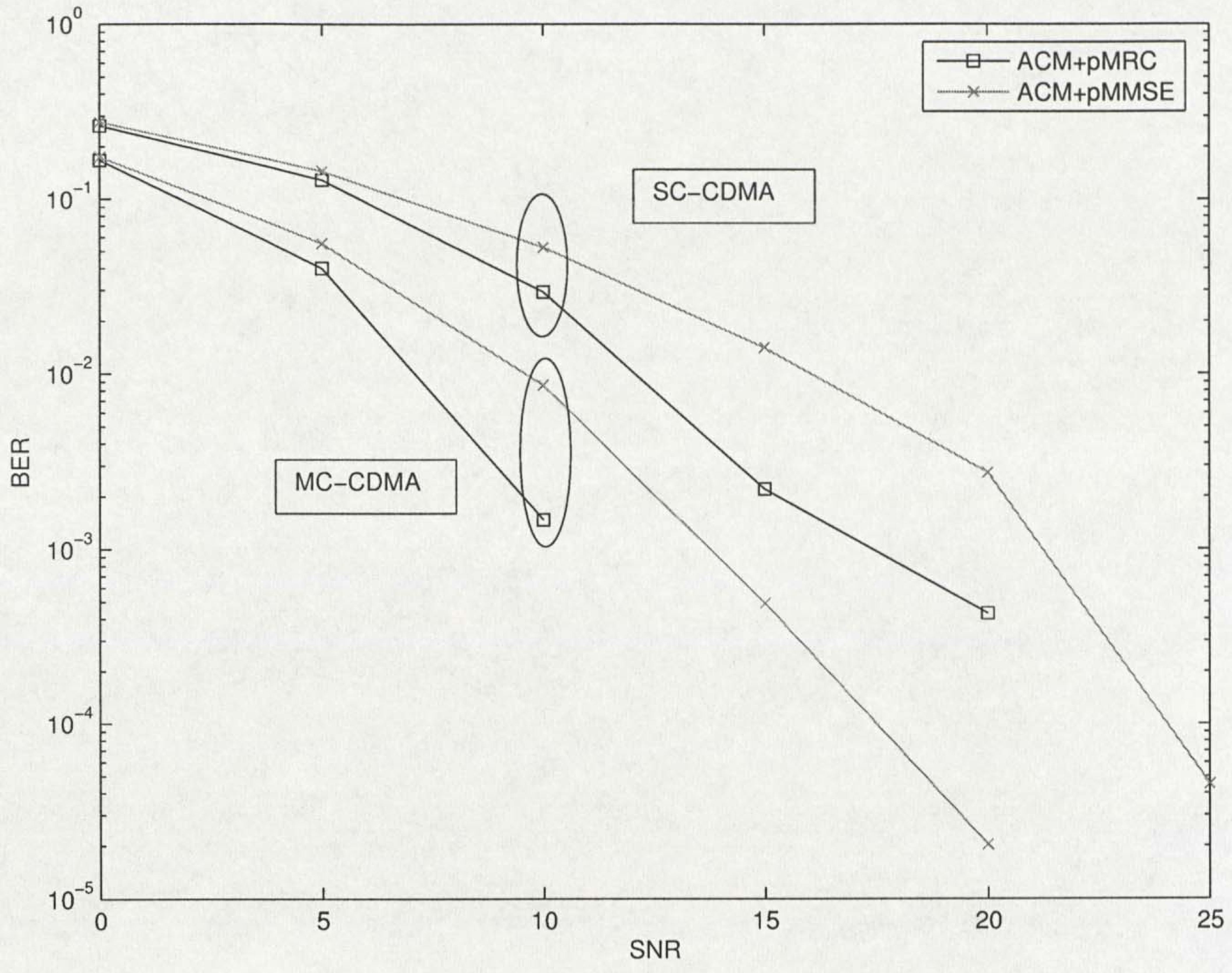

Figure 5.9: Comparison of 2x2 MIMO MC-CDMA and MIMO SC-CDMA per-carrier equalisers using ACM estimates with L=4,N=32

the square of the norm respectively. Only matrix multiplication and matrix inverse operations are considered in this analysis. The same notations were used as in Chapters 3 and 4.

5.4.1 pMMSE Equaliser

Using Eqn. (5.8), the complexity cost for computing the per-carrier MMSE weight vector for the n^{th} subcarrier and user i is given as

$$Cost_{pmmse,i}[n] = N_r^2 N_t + N_r^3 + N_r^2 \tag{5.11}$$

Thus the complexity involved for all users on the n^{th} subcarrier is $N_t \times Cost_{pmmse,i}[n]$ which when simplified yields

$$Cost_{pmmse}[n] = N_t N_r^2 (N_t + N_r + 1) \quad (5.12)$$

Thus the net complexity rendered is obtained by multiplying the above for N subcarriers resulting in

$$Cost_{pmmse} = N_t N_r N (N_t + N_r + 1) \quad (5.13)$$

5.4.2 pMRC Equaliser

The complexity cost incurred when using the per-carrier MRC equaliser, as defined in Eqn. (5.5), for the i^{th} user on the n^{th} subcarrier is given as $Cost_{pmrc,i}[n] = N_r$ and thus for all users on all subcarriers is given by

$$Cost_{pmrc} = N N_t N_r \quad (5.14)$$

Using the simulation setup values, the numerical complexity for each of the above equalisation schemes is obtained as 500% for MMSE assuming 100% for the MRC setup.

5.5 MSE analysis for per-carrier equalisers

A brief MSE analysis of the per-carrier equalisers used is presented here. The general per-carrier equaliser as given by Eqn. (5.4) is used as the starting point to derive the necessary expressions. The corresponding values for

MRC, EGC, ORC and MMSE are obtained by substituting their respective weight vectors denoted by $\mathbf{w}_i$ for user i. The data transmitted is denoted by $\mathbf{d}_i$ having zero mean and unit variance while $\mathbf{X}[n]$ denotes the received signal on subcarrier n and $\mathbf{E}$ represents the expectation operator. $\mathbf{R}[n]$ is the correlation matrix of size $N_r \times N_t$ on subcarrier n. The standard MSE equation is defined as

$$MSE = \mathbf{E}||d_i - \hat{d}_i||^2 \tag{5.15}$$

Substituting $\hat{\mathbf{d}}_i = \mathbf{w}_i\mathbf{X}$, the following equations are obtained for each subcarrier assuming that the fading on different subcarriers is uncorrelated.

$$\begin{aligned} MSE[n] &= \mathbf{E}[(1 - \mathbf{w}_i[n]\mathbf{X}[n])^2] \\ &= \mathbf{E}[(1 - \mathbf{w}_i[n]\mathbf{X}[n])(1 - \mathbf{w}_i[n]\mathbf{X}[n])^H] \\ &= \mathbf{E}(1 - \mathbf{X}^H[n]\mathbf{w}_i^H[n] - \mathbf{w}_i[n]\mathbf{X}[n] + \mathbf{w}_i[n]\mathbf{X}[n]\mathbf{X}^H[n]\mathbf{w}^H[n]) \end{aligned}$$

which can be further simplified by taking the expectation inside the brackets to yield

$$MSE[n] = 1 - \mathbf{E}(\mathbf{X}^H[n]\mathbf{w}_i^H[n]) - \mathbf{E}(\mathbf{w}_i[n]\mathbf{X}[n]) + \mathbf{w}_i^H[n]\mathbf{E}(\mathbf{X}[n]\mathbf{X}^H[n])\mathbf{w}_i[n] \tag{5.16}$$

As the transmitted signal is assumed to be i.i.d with zero mean, the second and the third terms on the right hand side are eliminated. Thus substituting the weight vectors respectively for the different equaliser types, the following equations are obtained.

$$
\begin{aligned}
MSE_{mrc}[n] &= 1 + \mathbf{H}_i[n](\sum_{i}^{N_t} \mathbf{H}_i[n]\mathbf{H}_i[n]^H + \sigma^2 \mathbf{I}_{N_r})\mathbf{H}_i^H \\
MSE_{egc}[n] &= 1 + \frac{\mathbf{H}_i[n]}{||\mathbf{H}_i[n]||}(\sum_{i}^{N_t} \mathbf{H}_i[n]\mathbf{H}_i[n]^H + \sigma^2 \mathbf{I}_{N_r})\frac{\mathbf{H}_i^H}{||\mathbf{H}_i[n]||} \\
MSE_{orc}[n] &= 1 + \frac{\mathbf{H}_i[n]}{||\mathbf{H}_i[n]||^2}(\sum_{i}^{N_t} \mathbf{H}_i[n]\mathbf{H}_i[n]^H + \sigma^2 \mathbf{I}_{N_r})\frac{\mathbf{H}_i^H}{||\mathbf{H}_i[n]||^2} \\
MSE_{mmse}[n] &= 1 + \mathbf{H}_i^H[n]\mathbf{R}^{-1}[n]\mathbf{H}_i[n]
\end{aligned}
$$

5.6 Summary

In this chapter, the per-carrier equalisation schemes were examined for both MIMO MC-CDMA and MIMO SC-CDMA Uplink scenarios. Their performance were compared against the equivalent block-based schemes and it was found that the per-carrier based MMSE and MRC equaliser provided order of magnitude improvements in performance at lower complexity compared to the conventional block-based schemes. These per-carrier equalisers were combined with the blind channel estimates obtained using the SS,SA and ACM methods. It was found that in the MC-CDMA system, the 3 blind channel estimation methods provided similar performance, while in the SC-CDMA case, the ACM provides the best performance with the SA performing the worst at low SNR. Also the performance gains obtained by using receive diversity were shown for both systems. A simple comparison was also done between MIMO MC-CDMA and MIMO SC-CDMA block transmission systems for the uplink scenario utilising per-carrier equalisation. It was found that the use of Walsh-Hadamard spreading codes in a Rayleigh frequency selective fading channel, affects the performance of MIMO SC-CDMA block transmission scheme, while MIMO MC-CDMA remains unaffected providing better performance. as in block-wise equaliser case described in 4.7 in Chapter4.

List of Figures

Chapter 6

Conclusions and Further Work

6.1 Conclusions

MIMO systems have been propelled to the fore as a result of the immense potential gains in performance offered at no extra bandwidth or power in comparison to SISO systems. OFDM [4], which utilise orthogonal subcarriers, thereby providing spectral efficiency, has already been adopted for DAB, DVBs etc as well as for ADSL modems. The combination of CDMA with OFDM is an exciting and viable solution in itself, which when combined with MIMO [39] systems provide even greater benefits. Much research has been done for the downlink channel estimation as well as detection problem using MC-CDMA as well as Single carrier CDMA. Though a few papers have covered the problem for uplink case in MC-CDMA, not much work has been carried out in the MIMO Uplink case.

In this thesis, the MIMO Uplink scenario is examined using both MC-CDMA and SC-CDMA schemes. The MIMO MC-CDMA Uplink scenario with quasi-synchronous channel model is examined in Chapter 3. Three available blind channel estimation methods namely the SS [54], SA[67] and ACM [63] methods are extended to the MIMO MC-CDMA setup. Their performances are simulated for varying SNR, load (number of users) as well as channel order. It was shown that the three channel estimation methods provide a similar level of performance with the ACM and SA methods avoiding the need for the initial EVD stage as in SS method. Also the elimination of rank determination and noise power estimation are other key advantages of the ACM method. In addition, an LSFE equaliser [45],[46] was used in conjunction with these three blind channel estimation algorithms. The combination of MIMO and MC-CDMA was shown to yield very good performance using the three blind channel estimation methods as compared to the perfect channel case. The improved performance of LSFE as compared to MMSE equaliser due to its SIC at each stage as well as the effects of receive diversity were illustrated. A complexity analysis of the three blind channel estimation methods as well as the LSFE and MMSE detectors was also performed.

Research has recently focussed on single carrier schemes combined with FDE as a result of the PAPR and frequency offset problems faced by MC-CDMA. This is the focus of Chapter 4, which examined the MIMO SC-CDMA system with block transmission using FDE. The lack of blind channel estimation for this particular setup motivated this research. The structure of the GB (CP or ZP) used for IBI elimination, was utilised for obtaining blind channel estimates. The same three blind channel estimation methods as for the MIMO MC-CDMA Uplink case, i.e. the SS, SA and the ACM methods were reformulated for the MIMO SC-CDMA Uplink with block transmission scheme. The key differences between the system models used in Chapters 3 and 4 were also highlighted as being **a**.) the displacement of IFFT stage from transmitter to receiver side and **b**.) the spreading being done in frequency and time domains respectively.

The ACM method, while yielding the best performance at low SNR, provided similar performance to the SS at high SNR, with the added advantage that it eliminates the need for an initial SVD, rank determination and noise power estimation. The SA approach performed the worst at lower SNR with similar levels of performance at high SNR, but again like the ACM required no initial SVD stage though rank order was required.

In addition, an LSFE equaliser was integrated with the blind channel estimation stage and its performance compared to that of MMSE equaliser. As illustrated in the simulation results, the LSFE equaliser performs better than the MMSE scheme due to the interference cancellation of the detected signals at each stage. The effect of load (no. of users) as well as varying channel length are also examined and it was shown that the LSFE maintains superior performance compared to MMSE equaliser.

Also to be noted is the insensitive nature of the SC-CDMA system to varying channel length. The performance gains obtained via receive diversity are also shown and a complexity analysis of the three blind channel estimation methods was also performed and tabulated. A simple comparison between MIMO MC-CDMA and MIMO SC-CDMA block transmission systems for the uplink case mentioned revealed the superior performance of the MC-CDMA scheme for the choice of Walsh-Hadamard spreading codes in a Rayleigh frequency selective fading channel. The MIMO SC-CDMA performance was affected by the Walsh codes as well as the fading channel, while the MC-CDMA scheme, through the inherent OFDM equalisation available, was able to counteract the fading well.

Chapters 3 and 4 examined the use of blind channel estimation for MIMO systems with CDMA using block based equalisation schemes. Assuming the uncorrelated and independent fading between the subcarriers, Chapter 5 analysed four per-carrier equalisers namely the MMSE, MRC, ORC and EGC respectively. As illustrated in the simulation results, it was found that the per-carrier setup provided an order of magnitude improvement as compared to the block-based equalisers using blind channel estimates. The effects of varying number of users and receive diversity gains were also examined and are illustrated.

Also a complexity analysis highlighted the fact that these per-carrier schemes provide a much lower complexity as compared to the block based schemes. A simple comparison was performed for the per-carrier equalisers used with both MIMO MC-CDMA and MIMO SC-CDMA and as was the case with block-wise equalisers, the MC-CDMA scenario provided superior performance in the case of Walsh spreading codes in a Rayleigh frequency selective fading channel.

In summary, this thesis proposed the use of MIMO systems combined with MC-CDMA and SC-CDMA block transmission using FDE for the up-link case. As blind channel estimation methods are required for coherent detection, three SOS based blind channel estimation methods were examined namely the SS, SA and the ACM methods. Their performances were examined in the uplink model and it was found that the ACM method provided a similar performance to the SS overall but eliminated the need for rank and noise power estimation at the cost of slightly higher complexity.

Also the LSFE block equaliser was proposed to be used in conjunction with these blind channel estimates and its improved performance noted in comparison to the conventional MMSE block equaliser. Finally per-carrier based equalisation schemes were proposed for MIMO MC-CDMA and MIMO SC-CDMA systems in the uplink scenario yielding better performance than their block-based counterparts, while providing lower complexity. These per-carrier MIMO equalisers were used in conjunction with the same three blind channel estimation methods, to yield blind per-carrier equalisers, which yielded comparable performance to the case when perfect channel knowledge is available at the receiver.

6.2 Future Work

The author would like to conclude by providing a few ideas for future research which are listed below.

- The use of STBC i.e. the use of transmit diversity is one exciting area that has gained a lot of interest recently. With the enormous gains in performance achievable, it would be an interesting challenge to investigate the application of blind channel estimation methods to the STBC MC-CDMA and SC-CDMA setups. A few papers have already applied the SS method to MC-CDMA using space time coding for the uplink with encouraging results. The effects of using different STBC types is another arena open to further investigation in this context.

- As the work throughout the thesis investigated a quasi(synchronous) channel model for the Uplink scenario by assuming that the GB would be longer than the maximum delay, further research could look into adaptive channel equalisation and estimation. The three blind channel estimation algorithms could be implemented adaptively for a time-varying asynchronous channel setup.

- A comprehensive comparative analysis of MC-CDMA versus SC-CDMA could be performed for both MIMO and space time coded systems using different spreading code types.

- The investigation into blind equalisation of both MIMO MC-CDMA and MIMO SC-CDMA without channel estimation stage.

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List of Tables

Chapter 1

Introduction

Mobile communication technology is undergoing dramatic changes in terms of user interaction and services with the arrival of 3G systems. The ever increasing demands for higher data-rates and better spectral efficiency for next generation mobile technology have led researchers to investigate different MultiCarrier Modulation (MCM) schemes such as Orthogonal Frequency Division Multiplexing (OFDM) and Multi-Carrier Code Division Multiple Access (MC-CDMA).

The fundamental idea behind OFDM involves the mixture of a parallel transmission scheme combined with frequency division multiplexing. In 1971, Weinstein and Ebert [1] proposed the idea of using Discrete Fourier Transforms (DFT) for the modulation and demodulation of this parallel frequency division multiplexing scheme lending it its current form today. It is only now that OFDM is coming to the fore because of the vast improvements made in Integrated Circuit (IC) Design and the implementation of Fast Fourier Transform (FFT) and other algorithms on Digital Signal Processors (DSP).

OFDM has found a number of applications since its early days in the military field namely modems and in the recent past in Digital Audio Broadcasting (DAB), Digitial Video Broadcasting (DVB), High-Defintion TV (HDTV), Wireless-LAN (W-LAN) etc. The true potential of this multi-carrier modulation scheme will be in the field of 4G mobile phone with NTT-DoCoMo already running trials in Japan and Korea with some remarkable performances reported [2]. The stage is set for the explosion of this technology onto the world scene with predicted hand-sets set to roll out initially in the Far-East followed by the US and Europe and then to the rest of the world within the next 4 years. The basic idea behind OFDM is the transmission of user information on multiple carriers, which are orthogonal to each other. The key fact here is that that orthogonal nature of individual carriers enables firstly a spectrally efficient transmission scheme and secondly eliminates any interference between the different carriers. The single biggest advantage of this scheme is its ability to transform a frequency selective fading channel into many parallel flat fading channels, which enables simple equalisation. Also the use of an MCM scheme employing block transmission, makes it more robust to Inter-Block Interference (IBI). OFDM has come of age with its acceptance as the standard for DVB, DABs and also recently with its acceptance as transmission technique for High Performance Local Area Networks and becoming part of the IEEE 802.11 WLAN standard [3].

The combination of OFDM with two other technologies namely Code Division Multiple Access (CDMA) and Multiple Input Multiple Output (MIMO) has led to its meteoric rise in status as the harbinger of the next generation technology in mobile communications. The combination of OFDM and

Abstract

Spectrally efficient mobile communication systems are motivated by ever increasing demands of higher data rates to reach the goal of a truly ubiqutious wireless network. Code Division Multiple Access (CDMA), which forms the primary underlying technology of the current 3G mobile transmission phenomenon, has been researched since the late 60's, since it was first used for military applications. In order to push the boundaries further, Multiple Input Multiple Output systems (MIMO), which provide tremendous performance gains with no expense of power or bandwidth, have come to the fore.

In this thesis, the MIMO and CDMA techniques are combined with both Multi-carrier (MC) and Single Carrier (SC) modulation schemes in the uplink scenario, and the results obtained support the proposals of these techniques, as standards for next generation mobile communications.

The first contribution of this thesis is the application of three available blind channel estimation algorithms in the uplink MIMO MC-CDMA scenario. The three blind channel estimation methods applied were the Subspace approach (SS), the Subspace Approximation (SA) approach and the AutoCorrelation Contribution (ACM) methods. It is shown that the three methods provide similar levels of performance with the ACM method eliminating the need for initial Eigen Value Decomposition (EVD) stage as well as rank determination as in the Subspace method. Blind channel estimation was also integrated with the so-called Layered Space Frequency Equaliser (LSFE), which provided better performance than the conventional Minimum Mean Square Error (MMSE) equaliser. Complexity analysis of the three

CDMA termed MC-CDMA is more robust to fading and noise effects, also enabling multiple users to share the same spectrum, while being differentiated by their specific spreading codes. When multiple antennas are used at the transmitter and receiver side, this further enhances performance by making use of the spatial diversity afforded, thus prompting the acceptance of MIMO as the standard for IEEE 802.11n scheme.

This thesis will define and distinguish between the different technologies starting with OFDM and its hybrid MC-CDMA in Chapter 2 before moving onto the MIMO system in Chapter 3. The benefits obtained as well as pitfalls will be discussed for various schemes. Chapter 3 will entail a discussion on the use of MIMO with MC-CDMA in an uplink quasi-synchronous channel. Three blind channel estimation methods namely the SS, SA and the ACM are applied to this setup and their performances compared. In addition a Layered Space Frequency Equalisation (LSFE) equaliser is used in conjunction with these blind channel estimation methods to provide improved performance compared to conventional Minimum Mean Square Equalisation (MMSE) schemes. Complexity analysis of the channel estimation as well as the equalisation schemes used is also presented.

Chapter 4 investigates the MIMO SC-CDMA system with block transmission for uplink scenario. The three blind channel estimation methods are formulated for this setup and their performances compared. The LSFE and the MMSE schemes are also utilised in conjunction with the blind methods and the improved performance of LSFE highlighted. The chapter concludes with a brief complexity analysis of the blind channel estimation algorithms and the equalisation schemes used.

Chapter 5 examines the use of per-carrier equaliser schemes for both MIMO MC-CDMA and MIMO SC-CDMA uplink systems and compares their performance with the block based schemes used in Chapters 3 and 4. A brief complexity analysis of the equaliser schemes used is also provided. Chapter 6 concludes the thesis along with a look to some aspects that could provide interesting avenues for future work.

1.1 Thesis Contributions

A summary of the original work presented in this thesis is listed below along with their corresponding chapters.

- The use of the LSFE equaliser is proposed for the MIMO MC-CDMA Uplink scenario in Chapter 3, which yields an improved performance compared to the conventional MMSE equaliser scheme.

- In Chapter 3, an extension of the Subspace (SS), Subspace Approximation (SA) and the Autocorrelation Contribution Method (ACM) blind channel estimation algorithms to the MIMO MC-CDMA Uplink scenario is proposed.

- In Chapter 4, the application of the LSFE scheme to the MIMO SC-CDMA block transmission scheme for uplink scenario is proposed resulting in improved performance compared to the MMSE scheme.

- Guard Band utilisation for blind channel estimation for MIMO SC-CDMA Uplink block transmission system via the three methods namely SS, SA and ACM method.

- Chapter 5 proposed the use of per-carrier based equalisers in conjunction with the three blind channel estimation methods investigated for both MIMO MC-CDMA and MIMO SC-CDMA block transmission systems in uplink case. The lower complexity involved as well as improved performance compared to block-based equalisers was highlighted.

1.2 List of Publications

The following is the list of publications derived from the original work described in this thesis.

- S. Punnose, X. Zhu, A. K. Nandi, "Layered Space Frequency Equalisation for Multiple Input Multiple Output MC-CDMA Systems in Frequency Selective Fading Channels", Proceedings of Fifth International Conference on Independent Component Analysis and Blind Signal Separation, ICA 2004, September 22-24, Granada, Spain.

- S. Punnose, X. Zhu, A. K. Nandi, "Blind Channel Estimation For MIMO Uplink MC-CDMA Systems with Layered Space Frequency Equalisation", Proceedings of the Second International Conference on Wireless Networking and Computing, WiCom2006, September 22-24 Wuhan, China.

- S. Punnose, X. Zhu, A. K. Nandi, "Blind Channel Estimation For MIMO Uplink Single Carrier CDMA Systems with Layered Space Frequency Equalisation", Submitted to IEE Journal on Communications.

Chapter 2

Background

2.1 OFDM

OFDM is a digital multi-carrier communication scheme that uses multiple orthogonal carriers with overlapping spectra to transmit information [4], [5]. The ease of implementation, when employing Inverse Fast Fourier Transform (IFFT) and Fast Fourier Transform (FFT) for modulation and demodulation respectively is a key advantage of this scheme. The following sections deal with the concept, advantages and disadvantages of OFDM and the wireless Rayleigh fading channel model used througout this thesis.

2.1.1 A Brief History

The original ideas of parallel data transmission were suggested in the late 1960's. The initial idea involved the use of parallel data and frequency division multiplexing with overlapping channels to avoid the need for high speed equalisation [6]. This also provided immunity against distortion and impulsive noise. One of the first papers published was by Chang [7], though there were other eminent contributors to the field too namely Weinstein and Ebert [1], Powers and Zimmerman [8].

The initial applications were in the military field, while other early applications involved the use of this technology for variable data rate modems for use in high frequency radio.

The use of multiple carriers required the condition that these carriers be orthogonal to each other in order to prevent Inter-Carrier Interference(ICI). Initially the major obstacle was the method of generation, which involved the use of sinusoidal generators and coherent demodulators. These became complex and expensive to implement as the number of carriers increased.

The solution to this problem was first suggested by Weinstein and Ebert [1], who suggested the use of DFT (Discrete Fourier Transforms) for the modulation and demodulation stages. This eliminated the need for arrays of generators and coherent demodulators and at the same time provided a completely digital means of implementation using the FFT algorithms. It is important to remember that these were possible only because of the tremendous advances in VLSI chip design and Digital Signal Processor (DSP).

In the 1980's, the use of OFDM was being experimented in high speed modems, digital mobile communications and high density recordings. Hirosaki designed and implemented a 19.2 kbps voice-band data modem using multiplexed QAM [9].

Researchers also realised that the inherent parallel nature and use of multiple carriers in OFDM enabled it to spread out a fade over many symbols, thereby countering the effects of the frequency selective channel as well as the

Doppler effects in mobile channels. In addition to countering the effects of the multi-path phenomenon, the use of guard band was suggested providing resistance against Inter Symbol Interference (ISI). Other applications that were researched included the use of DMT (Discrete Multi-Tone) for magnetic storage channels by Feig et al [10].

More recently i.e. in the last decade, most research regarding OFDM has been concentrated on the fields of wireless communication systems as well as in Digital Audio and Video Broadcasting (DAB/DVB). It is to form the basis of 4G or 'wireless world' technology to be implemented in the coming years as mentioned before and usher in a plethora of new systems and applications.

2.1.2 Concept of OFDM

The OFDM scheme in its essense involves the transmission of blocked data (information) symbols over multiple carriers, which are orthogonal to each other as mentioned earlier. This orthogonality lends the scheme its spectral efficiency as illustrated in Figure 2.1. This figure plots the absolute value of the FFT coefficients (i.e. spectrum) against subcarrier frequency. As can be seen at a particular frequency tone, one subcarrier has a peak while all the others are at zero level, thus eliminating any interference between carriers (ICI). However this orthogonality is often lost as a result of the effects of multi-path fading and is one of the key obstacles to be overcome as explained in the Section 2.1.4.

The ability of OFDM to convert a frequency selective fading channel to parallel flat fading channel simplifies equalisation [4], [5]. This makes it an

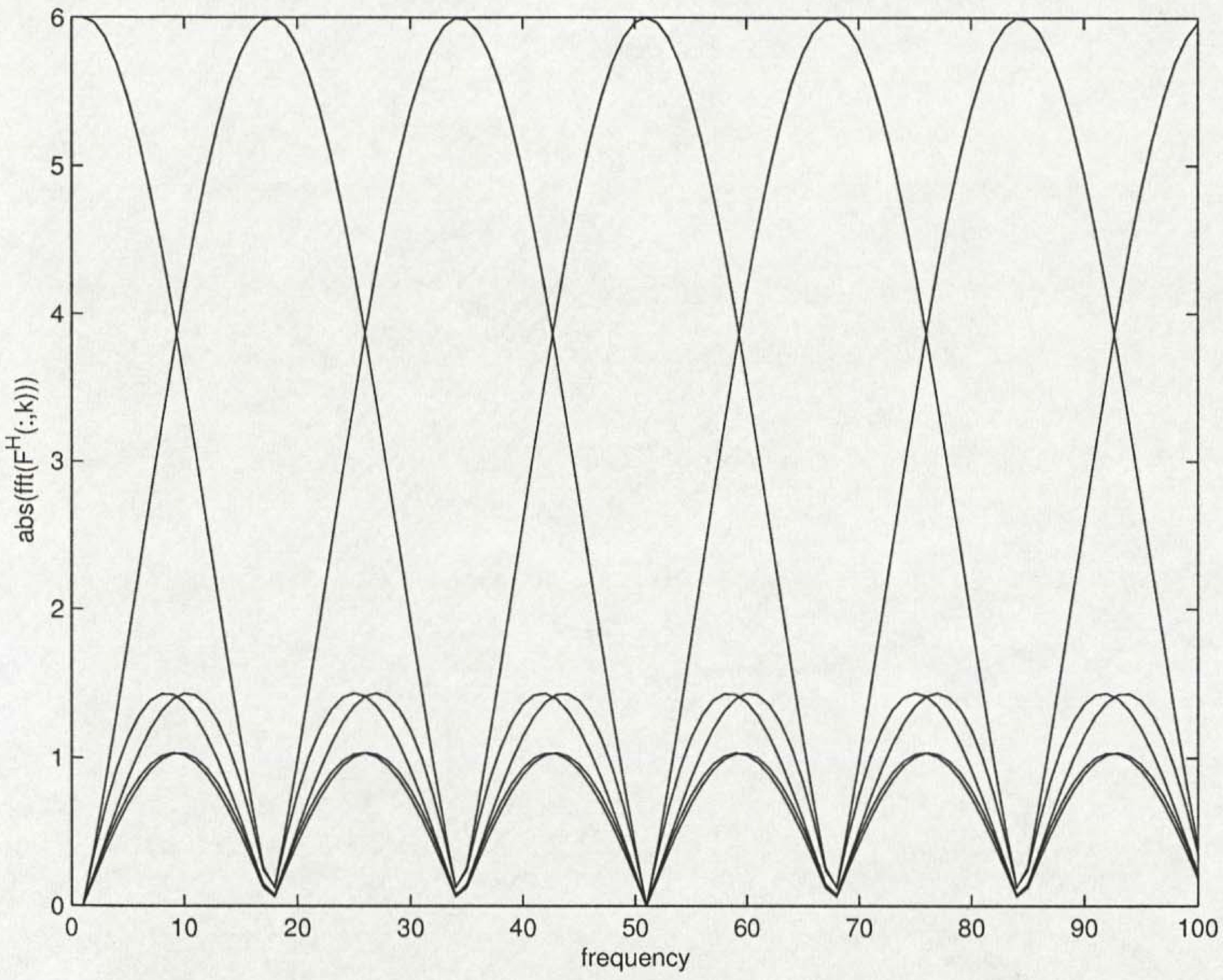

Figure 2.1: Orthogonal Subcarriers Wang [4]

ideal candidate for next-generation mobile technology. The splitting of serial stream into parallel streams also helps to counter the dispersive effects of the multi-path delay spread better as opposed to serial transmission scheme [4].

Advantages and Disadvantages

The key advantageous features of OFDM as compared to serial transmission schemes are enumerated below [4],[10]

- High spectral efficiency due to the orthogonal nature of subcarriers.
- Reduced receiver complexity due to the elimination of ISI by use of guard bands, thus enabling high data rate transmissions in multi-path channels.
- Simplification of implementation made possible by means of FFT al-

gorithms.

- Other hybrid techniques involving combination of OFDM with low - complexity multiple access schemes such as MC-CDMA, MC-TDMA and OFDMA are possible, which combine the advantages of the combining technologies.

- It has a high flexibility in terms of sub-carrier allocation, high data-rate service and user adaptation.

The OFDM scheme is not without its problems [10],[11],[12], the most prominent of which are

- Multi-path effects - As mentioned earlier it causes ISI and also causes echo effects. The use of guard band eliminates this only if the maximum delay spread is less than or equal to the length of the guard band otherwise these effects predominate resulting in loss of performance [4].

- Phase noise - This is caused by the carrier frequency errors. The synchronisation of the transmitter and receiver is a vital factor that affects the working of OFDM. Any frequency offsets result in fast degradation of performance as a result of the loss of orthogonality of the carriers. The performance degrades further with increase in number of carriers [13].

- Equalisation and carrier recovery - One of the biggest problems is that of equalising the frequency selective channel in the presence of fading.

Various techniques such as use of training (pilot) symbols, use of semi-blind and blind algorithms have been proposed to combat this problem [5].

- Peak to Average Power Ratio - Another factor involved in the design of OFDM system is the Peak to Average Power Ratio (PAPR), which depends on the signal constellation and the roll-off factor of the pulse shaping filter used [14]. Its value increases in the case of multiple carrier systems as a function of the number of carriers, causing non-linear distortion in the amplifier. This is due to the non-constant envelope of multi-carrier signals. It also results in amplifying the out of band noise [14].

- Spectral efficiency - The use of guard band to combat ISI further reduces the transmission efficiency of the OFDM system, whereby part of the bandwidth is used by the guard band. In order to improve the efficiency, techniques involving the use of special functions that are localised in both time and frequency domain have been proposed i.e these functions have their Fourier transforms (frequency domain representations) equal to their time domain form. An example of this kind of function is the IOTA (Isotropic Orthogonal Transform Algorithm) function, which is a modified Gaussian stochastic random process [15].

OFDM can also be classified into two types on the basis of type of Guard-Band [5],[11],[12] used to combat ISI. They are -

- Cyclic Prefix OFDM (CP-OFDM) - In this scheme the effects of the time dispersed channel of order 'L' are countered by appending the last

'L' bits of the data to the beginning of the symbol.This is effective only if the cyclic prefix is of length greater than or equal to the maximum delay spread of the channel. It is more widely used technique and equalisation is simple [5].

- Zero Padded OFDM (ZP-OFDM) - This scheme differs from the previous one in that L zeros are appended at the end of the data block to be transmitted as compared to the CP case, where the last L bits are added at the beginning of the block. This notion is relatively new and is growing in popularity [5].

The other types of OFDM that are being investigated include :-

- COFDM (Coded Orthogonal Frequency Division Multiplexing) - This scheme involves the use of different coding and interleaving techniques to improve the efficiency as well as the immunity in a fading environment. Some of the coding techniques that are being used are the Space Time Codes (STC), Trellis Coded Modulation (TCM) etc to name a few, which are used along with time and frequency interleaving methods. Digital broadcasting systems make use of this technology [10], [11].

- Wideband Orthogonal Frequency Division Multiplexing (WOFDM) - In this scheme, the spacing between multiple carriers is such that any frequency errors between the transmitter and receiver is only a small fraction of the spacing and thereby provides immunity against fading [11].

blind channel estimation algorithms and the equalisation schemes used was also performed.

The second contribution of this thesis is the application of the three blind channel estimation algorithms mentioned above to the uplink MIMO SC-CDMA block transmission system. The SS, SA and ACM blind channel estimation methods are applied and compared both for the systems with the CP and the ZP guard band scheme employed. It is shown that while the three methods provide similar performance at high SNR, at lower SNR, the ACM method provides the best performance, while SA performs the worst of the three. These blind channel estimation methods are also combined with LSFE, in comparison with the MMSE scheme. A brief complexity analysis of the three blind channel estimation methods and the equalisation schemes used was also performed.

With block equalisation employed in the first two contributions, the third contribution of this thesis is to investigate the per-carrier based equalisation for both MIMO MC-CDMA and MIMO SC-CDMA with block transmission in the uplink. It is shown that per-carrier based equalisers provide order of magnitude improvement over block based equalisers, while maintaining lower complexity. A brief Mean Square Error analysis was also done for the different per-carrier equalisers utilised.

- Flash Orthogonal Frequeny Division Multiplexing (FOFDM) - This method is also called fast-hopped OFDM and the multiple tones use fast-hopping to spread the data over the given signal spectrum [11].

- Multiple Input Multiple Output OFDM (MIMO OFDM) - This scheme combines the MIMO system with the OFDM transmission scheme. The MIMO system utilises the spatial multiplexing property of multiple transmit antennas, which transmit the data in the same frequency band and are differentiated only by the spatial signatures. It enables improved data-rate transmission scheme proportional to the number of antennas used and is very spectrally efficient [11], [16].

- Vector OFDM (VOFDM) - It is similar to the MIMO OFDM system described above and employs different methods to exploit the spatial diversity using either multiple antennas or a single antenna with multiple array elements [11].

2.1.3 OFDM System Model

A simple block diagram of an OFDM system is shown in Figure 2.2. It highlights the different stages involved in the transmitter and receiver.

The blocks making up the OFDM transmitter shown in Figure 2.2 are described first before presenting a brief mathematical description.

1. Blocking - This is the first step involved in the OFDM transmitter, where the serial data is converted into parallel form for transmission using multiple carriers. The presence of channel induced ISI giving

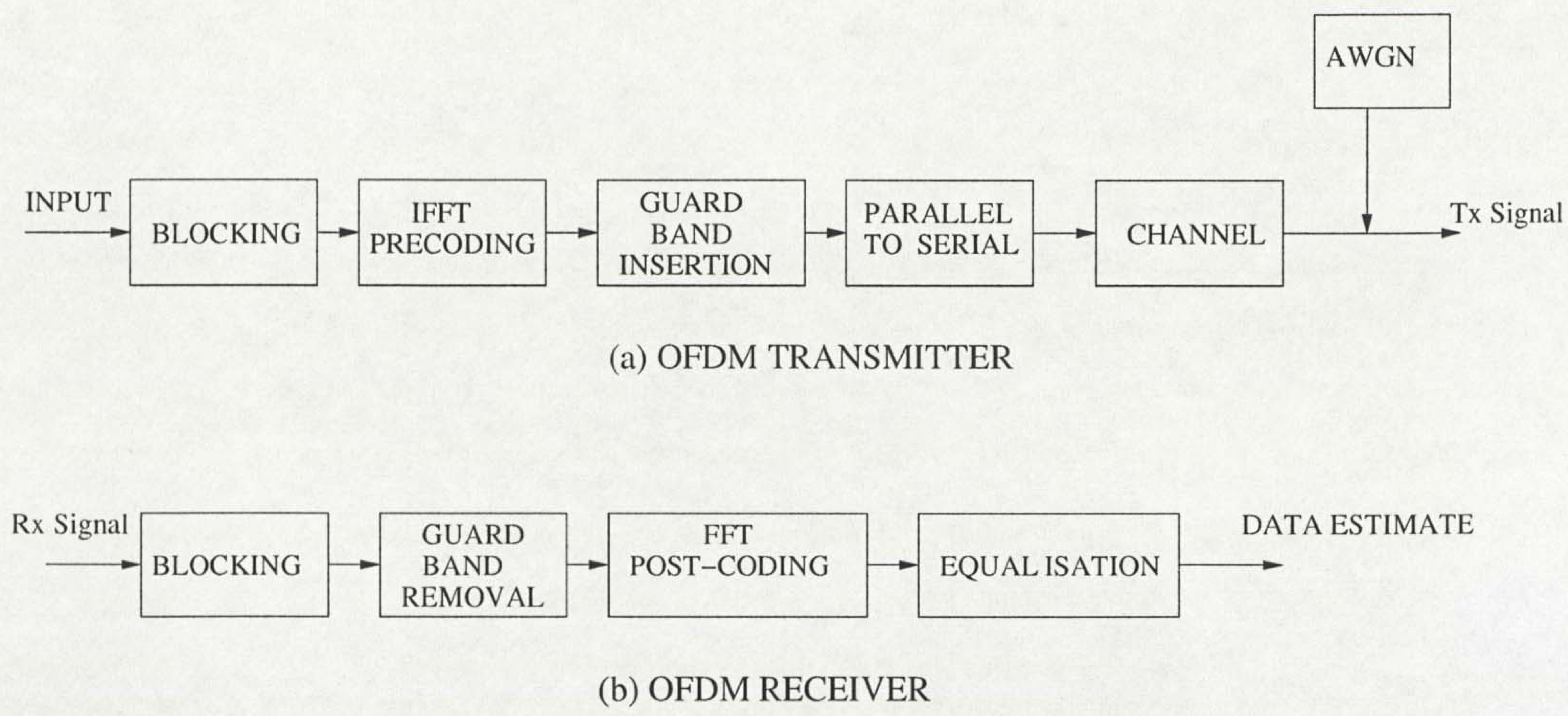

Figure 2.2: OFDM System Model 2.2a. OFDM Transmitter, 2.2b. OFDM Receiver Wang [4]

rise to frequency selectivity is a major factor limiting the performance of any wireless communication. Block transmissions are useful in this scenario and also the use of parallel transmissions enables better transmission efficiency as mentioned earlier.

2. Pre-coding - This stage involves the pre-coding of the parallel blocked data with IFFT matrix, which is used to modulate the carriers with the data. This stage is the basis for this technology, where the use of FFTs enables efficient,digital implementation.

3. Guard Band insertion - This stage is used to insert the guard band, which is done to alleviate the problems caused by ISI, which is as a result of the multi-path phenomenon. It essentially involves the multiplication of the above pre-coded data blocks with a guard inserting matrix.

4. Serialisation - This stage involves the serialisation of the above parallel, precoded and guard band inserted data blocks for transmission through the channel.

2.1.4 Channel

The wireless channel is the main obstacle facing wideband mobile communications [17]. The randomly varying nature of the wireless channel as well as the effects of mobility of the receiver/transmitter also add to the problem. Modelling of the wireless channel thus has been done statistically based on measurements made in the field or in laboratory test environments in order to enable a realistic study of its effects.

The different methods of radio propogation include reflection, diffraction and scattering. There may or may not be a direct line of sight path between transmitter and receiver as well, leading to different channel models. Channel propogation models are grouped into two on the basis of what aspect of the channel they model [18]. These are briefly presented below.

1. Large-Scale Propogation Models- These are used to predict the mean signal strength for an arbitrary transmitter-receiver separation, thus assisting in prediction of coverage area of a transmitter.

2. Small-Scale Propogation Models on the contrary are used to model the rapid changes of the received signal strength over short distances or short time durations. The different fading models namely Rayleigh, Rician etc come under this category.

To summarise, large-scale fading deals with the loss of signal strength as a result of the separation distance between the transmitter and receiver, while small-scale fading deals with the rapid fluctuations of the signal strength brought about by the movement of transmitter/receiver as well as scattering of signals. In this thesis, the channel effects that are modelled and studied are of the small-scale fading kind. The Rayleigh slow fading channel model with uniform power delay profile [18] is used throughout this thesis.

2.1.5 OFDM Receiver

The various blocks involved in the OFDM receiver [4] are explained below

1. Blocking - This step involves the serial to parallel conversion of the received signal in order to enable the block processing of the following stages.

2. Guard Band suppression - This stage involves the suppression/removal of the guard band inserted at the transmitter.

3. Post-Coding - This step involves the multiplication of the guard suppressed received signal blocks with the FFT matrix to retrieve the data that was pre-coded using the IFFT at the transmitter end.

4. Equalisation - This stage performs the equalisation of the received signal blocks in order to account for the channel effects. This is a vast topic in itself and there are various techniques that can be used to perform this step namely pre, post or balanced equalisation [4].

The more common forms, namely the Zero Forcing (ZF) equaliser and the Minimum Mean Square Equalisation (MMSE) equaliser, which is robust in noisy environments, are explained below and will be used throughout this thesis.

2.2 Mathematical Representation of the OFDM System

The simple example case of channel order L and number of subcarriers set to N is considered. The guard band is set to $L_{gb} = L$ in this case.

The data is initially blocked into blocks of length N for conversion to the parallel form. After IFFT modulation, it is then multiplied by the guard inserting matrix $\mathbf{T}$, which can be either $\mathbf{T}_{cp}$ for Cyclic Prefix OFDM (CP-OFDM) or $\mathbf{T}_{zp}$ for Zero Padded OFDM (ZP-OFDM). The received signal $\mathbf{x}(i)$ at time index i is expressed as follows with $\mathbf{u}(i)$ denoting the input blocked symbol of size N at time i and $u(i-1)$ represents the dispersive effect of the channel in the form of the previous block. $\eta(i)$ represents the additive white Gaussian noise at time i [4].

$$\mathbf{x}(i) = \mathbf{H}_0\mathbf{T}\mathbf{u}(i) + \mathbf{H}_1\mathbf{T}\mathbf{u}(i-1) + \eta(i) \tag{2.1}$$

$\mathbf{H}_0$ and $\mathbf{H}_1$ are $P \times P$ Toeplitz matrices [4],[19] given as

$$\mathbf{H}_0 = \begin{bmatrix} h(0) & 0 & 0... & 0 \\ : & h(0) & 0... & 0 \\ h(L). & : & .... & : \\ : & ... & ... & 0 \\ 0 & ... & h(L)... & h(0) \end{bmatrix}$$

$$\mathbf{H}_1 = \begin{bmatrix} 0 & ... & h(L) & ... & h(1) \\ : & . & 0 & ... & : \\ 0 & ... & . & ... & h(L) \\ : & : & ... & . & : \\ 0 & ... & 0 & ... & 0 \end{bmatrix}$$

These are the channel matrices that correspond to the dispersive channel considered, where $P = N + L_{gb}$ is the length of the data block after the guard band insertion stage. This yields blocks of guard inserted data blocks of length $P = N + L_{gb}$

In the case of ZP, L_{gb} zeros are added at the end of the data block of length N to give the same size $P = N + L_{gb}$ blocks as in CP-OFDM, where the L_{gb} data values at the end, are added to the beginning of the block.

2.2.1 CP-OFDM System

The FFT processed and guard band stripped received signal $\widetilde{\mathbf{x}}$ for the CP-OFDM system can be expressed as follows [4]

$$\widetilde{\mathbf{x}}(i) = \widetilde{\mathbf{H}}\mathbf{u}(i) + \eta(i) \tag{2.2}$$

where $\widetilde{\mathbf{H}} = \mathbf{R}_{cp}\mathbf{H}_0\mathbf{T}_{cp}$ is the circulant channel matrix. $\mathbf{R}_{cp}$ is the guard-band removal matrix that performs the guard band suppression with $\mathbf{T}_{cp}$ denoting the CP insertion matrix with $\eta(i)$ representing the additive white Gaussian noise vector.

2.2.2 ZP-OFDM System

The received signal for the ZP-OFDM scheme can be expressed as follows [4]

$$\bar{\mathbf{x}}(i) = \bar{\mathbf{H}}\mathbf{u}(i) + \eta(i) \tag{2.3}$$

where $\bar{\mathbf{H}} = \mathbf{H}_0\mathbf{T}_{zp}$ i.e. $\bar{\mathbf{H}}$ is the tall channel matrix as given above. $\mathbf{T}_{zp}$ is the ZP insertion matrix and $\eta(i)$ denotes the noise vector. Since the last L_{gb} values are all zeros for this case, these are discarded. This matrix $\bar{\mathbf{H}}$ is a tall Toeplitz matrix of size $P \times N$ thereby ensuring its full column rank.

At the receiver end, the received signal is blocked into blocks of length $P = N + L_{gb}$ and this blocked data is then passed to the equaliser stage. after which the FFT is applied to account for the pre-coding stage at the transmitter end done using the IFFT. The equalisation in ZP-OFDM is slightly more complex as a result of the tall Toeplitz nature of the channel model. It can be converted to a circulant model, as in CP-OFDM, by using the overlap and add method. This can be performed by using the R_{zp} matrix, expressed as $R_{zp} = [I_{zp}I]$ with I denoting the identity matrix of size N and I_{zp} denoting the first L columns of I. Thus a distorted signal is obtained as a result which is due to the channel effects.

The equalisation stage is the one of the most important stages and probably one of the components that has been researched the most in OFDM. It involves the equalisation of the distorted blocked received signal and there are many techniques available for performing the equalisation. A brief description of the two commonly used equalisation schemes namely the ZF and MMSE are presented next.

2.3 Equalisation

This section provides a brief explanation of the equalisation process, which is an essential component of any communication system, in the OFDM system [4], [19] described above.

An equaliser can be thought of as a transversal filter, which can be used either at pass-band or base-band [5], [20]. The problem of equalisation can be interpreted as one of deconvolution or inverse modelling in order to reconstruct the original signal. The redundancy obtained from adding the CP or ZP leads to the simplification of the equaliser stage. The problem with serial equalisation is that it eventually simplifies to finding the inverse of the channel, if it exists [4]-[5]. There is also the added problem of noise enhancement caused by non-minimum phase nature of the channel.

The class of block equalisers is considered here for both CP and ZP systems. There are many different kinds of equalisers depending on the system considered. The ZF and the MMSE algorithms are described below for both CP and ZP OFDM systems.

2.3.1 Zero-Forcing Equaliser

The ZF equaliser is basically the MMSE equaliser for the noiseless case. The basic setup for the CP-OFDM ZF equaliser is described next. After the blocked received signal has been been purged of the guard band, it is then passed through a post-processing phase, which is the multiplication with the FFT matrix as explained above. This yields a channel distorted version of the original data, which is then applied to the equaliser stage to obtain as good an estimate of the original data as possible.

CP ZF equaliser

In the CP case, the received signal is the circular convolution of the channel with transmitted block [4]. The circulant channel matrix $\widetilde{\mathbf{H}}$ is not full rank when $\mathbf{H}(z)$ has zeros on the unit circle, which leads to the fact that even if the channel is known, the data cannot be recovered as a result of this condition. The zero-forcing equaliser is obtained as inverse of the circulant matrix if it exists [4]. It is essential to point out that the inverse of $\widetilde{\mathbf{H}}$ exists, if the diagonalised matrix $\mathbf{D}_m$, obtained by pre and post multiplying $\widetilde{\mathbf{H}}$ by the IFFT and FFT matrices is invertible i.e.

$$\mathbf{D}_m = \mathbf{F}\widetilde{\mathbf{H}}\mathbf{F}^{-1} \tag{2.4}$$

where $\mathbf{F}$ and $\mathbf{F}^{-1}$ are the FFT and IFFT matrices respectively. The equation for the ZF equaliser for CP is given as

$$\mathbf{G} = \widetilde{\mathbf{H}}^{-1} \tag{2.5}$$

and the data detected is given by

$$\hat{\mathbf{u}}_{zf}(i) = \mathbf{G}_{zf}\widetilde{\mathbf{x}}(i) \tag{2.6}$$

where $\widetilde{\mathbf{x}}(i)$ is the received signal as obtained in Eqn. (2.2).

ZP-ZF Equaliser

In this case, the only change is that of the channel matrix used and the equation for the ZF equaliser is given as follows

$$\mathbf{G}_{zf} = \bar{\mathbf{H}}^{-1} \tag{2.7}$$

where $\bar{\mathbf{H}}$ is the tall Toeplitz channel matrix as defined in Eqn. (2.3) for ZP-OFDM case. This guarantees the symbol recovery even if channel has zeros on the unit circle [4], [19]. The data detected is given as

$$\hat{\mathbf{u}}_{zf} = \mathbf{G}_{zf}\bar{\mathbf{x}}(i) \tag{2.8}$$

2.3.2 MMSE Equalisation

The process of equalisation or estimation of a data signal mixed with noise is one that is commonly undertaken in all communication systems. Added to it is the problem of the channel dispersion effects and this has lead to the vast amounts of research being done in the field of signal separation or equalisation [21]. The MMSE equaliser or estimator is one of the conventional techniques used. The basic principle of the MMSE equaliser is briefly described next.

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ZP-OFDM MMSE Equaliser

The ZF case considered yields symbol recovery with zero ISI but in the presence of noise the situation changes. In order to trade off the ISI for noise suppression, MMSE detector is used, which minimises the MSE between the data estimated and the actual value. For this scheme, the following assumptions are made [4]-[5]

1. The correlation matrices of signal and noise are assumed known and represented as $\mathbf{R}_u = E[\mathbf{u}(i)\mathbf{u}^H(i)]$ and $\mathbf{R}_\eta = E[\eta(i)\eta^H(i)]$.
2. The transmitted symbols and noise are assumed to be white which yields $\mathbf{R}_u = \mathbf{I}_N$ and $\mathbf{R}_\eta = \sigma^2\mathbf{I}_P$.

The equation for the MMSE detection are given as [4]

$$\mathbf{G}_{mmse,zp} = \mathbf{R}_u\bar{\mathbf{H}}^H(\mathbf{R}_\eta + \bar{\mathbf{H}}\mathbf{R}_u\bar{\mathbf{H}}^H)^{-1} \quad (2.9)$$

which is simplified according to the assumptions above to yield

$$\mathbf{G}_{mmse,zp} = \bar{\mathbf{H}}^H(\bar{\mathbf{H}}\bar{\mathbf{H}}^H + \sigma^2\mathbf{I})^{-1} \quad (2.10)$$

where the term in the brackets is the autocorrelation of the received signal $\bar{\mathbf{x}}(i)$.

An important point to make at this point is that ZP-OFDM is more suitable to blind channel estimation due to its tall Toeplitz channel matrix nature [4], [5]. The equations for the CP-OFDM case are similar except for the differences in the channel matrix and these are briefly illustrated next.

CP-OFDM MMSE Equaliser

The basic setup is the same as before with the same assumptions being made. The resulting equations for the CP case are given with both the original form and then the more simplified version based on the assumptions similar to those for the ZP case. The equation for the MMSE detection are given as

$$\mathbf{G}_{mmse,cp} = \mathbf{R}_u \widetilde{\mathbf{H}}^H (\mathbf{R}_\eta + \widetilde{\mathbf{H}} \mathbf{R}_u \widetilde{\mathbf{H}}^H)^{-1} \tag{2.11}$$

which is simplified according to the assumptions as

$$\mathbf{G}_{mmse,cp} = \widetilde{\mathbf{H}}^H (\widetilde{\mathbf{H}} \widetilde{\mathbf{H}}^H + \sigma^2 \mathbf{I})^{-1} \tag{2.12}$$

2.4 CDMA

Code Division Multiple Access (CDMA) is one type of multiple access scheme where coding is used to provide the multiple access property [22]. Though originally developed for military communications, it has grown in popularity over the last 25 years and Wideband CDMA (WCDMA) is the core technology behind current 3G systems. The basic principle of CDMA involves spreading a user's information signal over a large bandwidth with the help of user-specific code. This unique code can be of different types such as Maximal length sequences, Gold sequences, Kasami sequences, Walsh Hadamard codes etc [22].

The ability to transmit all the users data in the same frequency band and at the same time is what makes CDMA stand out. In order to achieve this, the spreading codes chosen must satisfy certain properties such as low cross-correlation between codes, good auto-correlation properties etc. The choice of codes used depends on the specific application and system requirements. The key features of CDMA are briefly described as follows

1. Multiple access - The use of distinct spreading codes enables the sharing of the same frequency and time slots for transmission of all users data [22].

2. Interference rejection - Spreading codes having low cross-correlation are usually chosen, so that at the receiver, after applying the desired user's spreading code, it despreads the desired user's signal while the interference signal is spread [22].

3. Low Probability of Interception - This is due to the low power density as a result of using pseudo-noise like code sequences thus making it difficult to detect [22].

4. Robustness to fading - The use of CDMA helps in combating the effects of multi-path propogation [22].

CDMA does suffer a few disadvantages and the primary hurdle is the Multiple Access Interference (MAI). This is caused as a result of a large number of users. The despreading process increases the desired users signal power and considers the rest as interference, which grows with the number of users. Another hurdle in use of CDMA systems is the near-far effect.

The near-far effect is caused by the different power levels of received signals of the users depending on the distance from the receiver. The stronger signal often masks the weaker signals and results in increased BER thereby yielding poor performance. The different types of CDMA [22], [23] are

1. Direct Sequence CDMA (DS-CDMA) - The information signal is spread directly via a code signal chosen for its unique properties.

2. Frequency Hopping CDMA (FH-CDMA) - The carrier frequency used for transmission, is varied according to the code signal used.

3. Time Hopping CDMA (TH-CDMA)- The information is transmitted in bursts, the timing of which is determined by spreading code signal.

4. Hybrid Modulation utilises a combination of any two of the above methods.

5. Chirp Modulation- Specific to military radar use, where it involves varying the frequency range of a low power signal over a wide range.

The performance of CDMA systems is constricted by two main forms of interference [22] namely the

1. ISI - This is caused by multi-path propogation effects i.e. the delays caused by diffraction, reflection etc.

2. MAI - As mentioned above, the transmission of multiple users information in the same frequency band results in interference at the receiver

Single user receivers such as matched filter receivers [22],[24], which use individual correlators to detect each user's signal, as well as Rake receivers treat MAI as noise. A Rake receiver [24] counters the effects of multipath fading by using several matched filters or correlators to tune to each of the multiple paths and then combines the outputs to provide improved performance compared to simple matched filter type receivers. Multiuser techniques include primarily two types [22]-

1. Joint Detection (JD) schemes such as Block Linear Equalisers (BLE) namely ZF-BLE, MMSE-BLE as well as Block Decision Feedback Equalisers (BDFE) such as ZF-BDFE and MMSE-BDFE [25],[26].

2. Interference Cancellation schemes such as Successive Interference Cancellation (SIC) receivers, Parallel Interference Cancellation (PIC) receivers [22],[27]

The type of spreading code used determines the performance of CDMA based systems. A wide variety of spreading codes are available with maximal length sequences (m-sequences), Gold codes, Kasami sequences, Orthogonal codes such as Walsh-Hadamard codes [23] etc being the predominant ones. Maximal length sequences are generated by linearly shifting contents of a given shift register of a given length [23]. They have good autocorrelation properties but may suffer large cross-correlation values. Gold codes on the other hand provide bounded cross-correlation values, while providing a large number of possible codes and are generated from maximal length sequences. Kasami sequences have low cross-correlation properties as well as good autocorrelation properties and are quite often used as scrambling codes [23]. Walsh-hadamard codes provide orthogonal code sets and are preferred for synchronous systems. In WCDMA based schemes, the multiple spreading technique is used, which provides flexibility and orthogonality [23]. A short orthogonal channelization code to spread each user's information bit is used, which maintains orthogonality among different users in the same cell. This spread signal is then multipled by a long pseudorandom sequence called scrambling codes, which are cell-specific. In this thesis, only the channelization code is considered.

In the next section, the MC-CDMA system i.e. the combination of CDMA and previously described OFDM scheme, is introduced along with a few of its advantages and key features.

2.4.1 MC-CDMA

This section introduces and describes the concept of Multi-Carrier Code Division Multiple Access (MC-CDMA). It is a hybrid combination of two modulation schemes namely OFDM and CDMA and is gaining popularity due to its advantageous features. The foundations of this hybrid scheme were laid by a few prominent researchers back in 1993, who proposed a variety of schemes to tackle the problem of transmitting high-data rates in a hostile mobile environment. Three main types of hybrid schemes were proposed namely MC-CDMA, Multi-Carrier Direct Sequence CDMA (MC-DS-CDMA) and Multi Tone CDMA (MT-CDMA). The key contributors were Linnartz and Fettweis [28], Lee [29], Brajal and Jourdan [30] for MC-CDMA, DaSilva and Sousa for MC-DS-CDMA and L.Vandendorpe for MT-CDMA respectively.

MC-CDMA is further grouped into two types [31], which are

- Frequency domain spreading - The signals are spread using the corresponding spreading codes followed by modulation of a different subcarrier with each chip.

- Time domain spreading - In this case, the blocked signals are spread using the code and then different subcarriers are modulated with each data block i.e the spreading operation is in time-domain as in DS-CDMA.

This thesis will focus on the use of MC-CDMA with MIMO systems, which is introduced in next section, for the uplink scenario. In the uplink scenario, each user transmits their information from corresponding antenna(s) and the signal received is the superposition of all transmitted user signals. CDMA utilises the spread spectrum property to provide redundancy to the signal and thus combat the effects of fading. It essentially makes the transmission resistant to effects of fading by spreading the original data over a larger bandwidth. But one of the key problems in DS-CDMA is that it suffers from Inter-Chip Interference [22], which is caused as a result of the delay spreads. MC-CDMA on the other hand utilises a frequency domain spreading code that spreads the original data transmitted on multiple carriers.

The use of multiple carriers each with symbol duration larger than the delay spread ensures that it is protected against ICI [28] and ISI, which is combated by means of the guard-band as in OFDM. Also it makes efficient use of the available spectrum, as different users can transmit their data simultaneously, being differentiated by their individual pseudo-noise like signature codes. Employing frequency domain spreading enables MC-CDMA to use all the received signal energy scattered in the frequency domain unlike the DS-CDMA case, where it is very difficult to combine the scattered energy in the time domain.

Some of the advantages of the MC-CDMA scheme [31] are listed below

1. Robustness to MAI- An improvement in performance is obtained due to the frequency diversity accorded as a result of the multiple carriers. It performs relatively better in comparison with DS-CDMA as a result of the frequency diverse transmission scheme used.

2. ISI mitigation- This property is as a result of the OFDM part of MC-CDMA scheme, whereby the guard-band employed ensures ISI cancellation.

3. Spectral efficiency- This feature is accorded as a result of the orthogonal sub-carriers used for transmission (as in OFDM), which enable each subcarrier to transmit multiple user signals differentiated by their respective spreading codes.

In this thesis, Walsh Hadamard codes were used throughout. Sharing a common multiple carrier transmission scheme with OFDM leaves MC-CDMA open to the same vulnerabilities as in OFDM, which were outlined in previous section.

2.4.2 SC-CDMA with block transmission

As mentioned earlier, MC-CDMA schemes are hindered by PAPR [14] as well as carrier frequency sensitivity [32]. The choice of spreading codes used as well as the constellation used cause non-linear distortion in the amplifiers. Single-Carrier Code Division Multiple Access (SC-CDMA) with block transmission and utilising CP [33] combines the advantages of CP with those of

CDMA. M. Sabbaghian [14] utilised a Selective Mapping algorithm, which yielded significantly lower PAPR levels than the MC-CDMA case.

Madhukumar et al [34] explored the SC-CDMA block transmission scheme with CP, investigating the use of different pilot symbols for channel estimation as well as different Frequency Domain Equalisation (FDE) equaliser structures. Peng et al [35] proposed a simple tranceiver structure for CP-CDMA with FDE. The lower complexity, the improved performance compared to conventional RAKE [24] receivers or time domain equalisers makes SC-CDMA block transmissions with GB a very attractive scheme [33],[35]. In Ref. [36] the use of transmit and receive diversity along with CP-CDMA systems and their potential gains was examined. Interference Cancellation based receivers can also be utilised for these systems as in DS-CDMA to yield improved performance [37]. A selective PIC based algorithm was also proposed for uplink CP-CDMA schemes in [38].

In this thesis, MIMO SC-CDMA with block transmission and using CP is investigated and a brief description of the ZP system also provided in the blind channel estimation context. Next a brief introduction of MIMO systems is provided followed by a brief description of its potential channel capacity benefits.

2.5 MIMO

Winters and Seshadri were one of the first people to propose the use of Multiple Input Multiple Output (MIMO) systems in their seminal paper [39]. Other prominent contributions included [40], [41] and references therein.

A MIMO system can also be considered as a extended smart-antenna system [21]. The key difference between the two being that in a smart antenna system, multiple antennas are used only at one end of the transmission scheme i.e either at the transmitter or at the receiver end. The basic idea of a MIMO system involves using multiple antennas at both the transmit and receive ends so as to provide diversity and enhance the performance of the transmission system. The fundamental notion behind diversity is that if multiple replicas or copies of the transmitted signal are received, then there is a high probability that atleast one or more of these replicas will not be in a fade at a given instant [41]. There are three main forms of diversity [17], [42] that are utilised, namely

- Temporal Diversity - It is utilised for time selective fading channels by means of Forward Error Correction Coding (FEC), interleaving etc. The signal is spread over a time that is larger than the coherence time i.e. the minimum time separation between independent channel fades.

- Frequency Diversity - Spread spectrum methods, FEC with MCM, interleaving etc are a few of the methods that exploit the frequency diversity, which is used for a frequency selective fading channel. In this case the signal is spread in frequency domain over a bandwidth larger than the coherence bandwidth i.e the minimum frequency separation between independent channel fades. The coherence bandwidth is the inverse of the delay spread of the channel.

- Spatial Diversity - Utilise the spatial distance greater than coherence distance between multiple transmit and/or receive antennas. The codefinitely kept me on my toes when it came to printer related queries and other sys admin and matrix related queries.

And lastly and most importantly, I would like to thank my parents and sister for their love, support and guidance, without whom this thesis would not have been possible. I dedicate this thesis to them.

herence distance is defined as the minimum spatial separation between antennas for independent fading. It can be either in the form of transmit diversity or receive diversity or both.

The essence of MIMO technology involves space-time processing utilising the spatial as well as temporal diversity provided by multiple antennas. Possibly the most important property that makes MIMO so attractive, is the fact that it has the ability to utilise the usually destructive multi-path propogation effects of wireless channels, to its advantage.

MIMO systems utilising the transmit diversity [17], [21], [42] can be used to provide either

- Data Rate Maximisation -Use of spatial diversity such as Spatial Multiplexing systems
- Performance Maximisation - Use of Space Time Block Code (STBC) or Space Time Trellis Codes (STTC), which provide improved performance

In the next section, the channel capacity equations will illustrate the gains that are obtained as a result of the spatial diversity. The vast improvements in performance obtained at no extra power or bandwidth cost, have resulted in MIMO systems being one of the most hotly researched topics over the last two decades leading to its adoption in IEEE 802.11n standard [3].

The temporal diversity can be utilised by means of some form of coding at the transmitter end. It could be space-time code schemes such as STBC or STTC [17], [21], [42]. This enables a great improvement in performance at the cost of redundant transmission of the coded information. Alamouti's scheme [43] is one of the simplest and most popular forms of STBC. A key advantage of STBC's is the simplified linear processing involved at the receiver as compared to the exponential complexity dependent on number of trellis states for STTCs.

Two of the key terms often used to describe MIMO [17], [21], [42] are briefly explained below

- Diversity order or diversity gain - Defined as the number of independent spatial streams available at the transmitter or receiver. The diversity order is equal to the product of number of transmit and receive antennas assuming an independent fading channel between each transmit-receive pair.

- Array gain -It is defined as the average increase in SNR at the receiver as a result of the coherent combining of multiple antennas at the receiver or transmitter or both.

A mathematical representation of a simple MIMO system is followed by a brief explanation of the channel capacity gains offered by MIMO systems.

2.5.1 MIMO Channel Model

Figure 2.3 shows a simple 2×2 MIMO system, where N_t and N_r represent the number of transmit and receive antennas respectively. The data transmitted is represented by $\mathbf{s}_i$ for $i = 1, 2, \ldots N_t$. The received signal $\mathbf{x}_j$ at the j^{th} antenna can then be expressed as

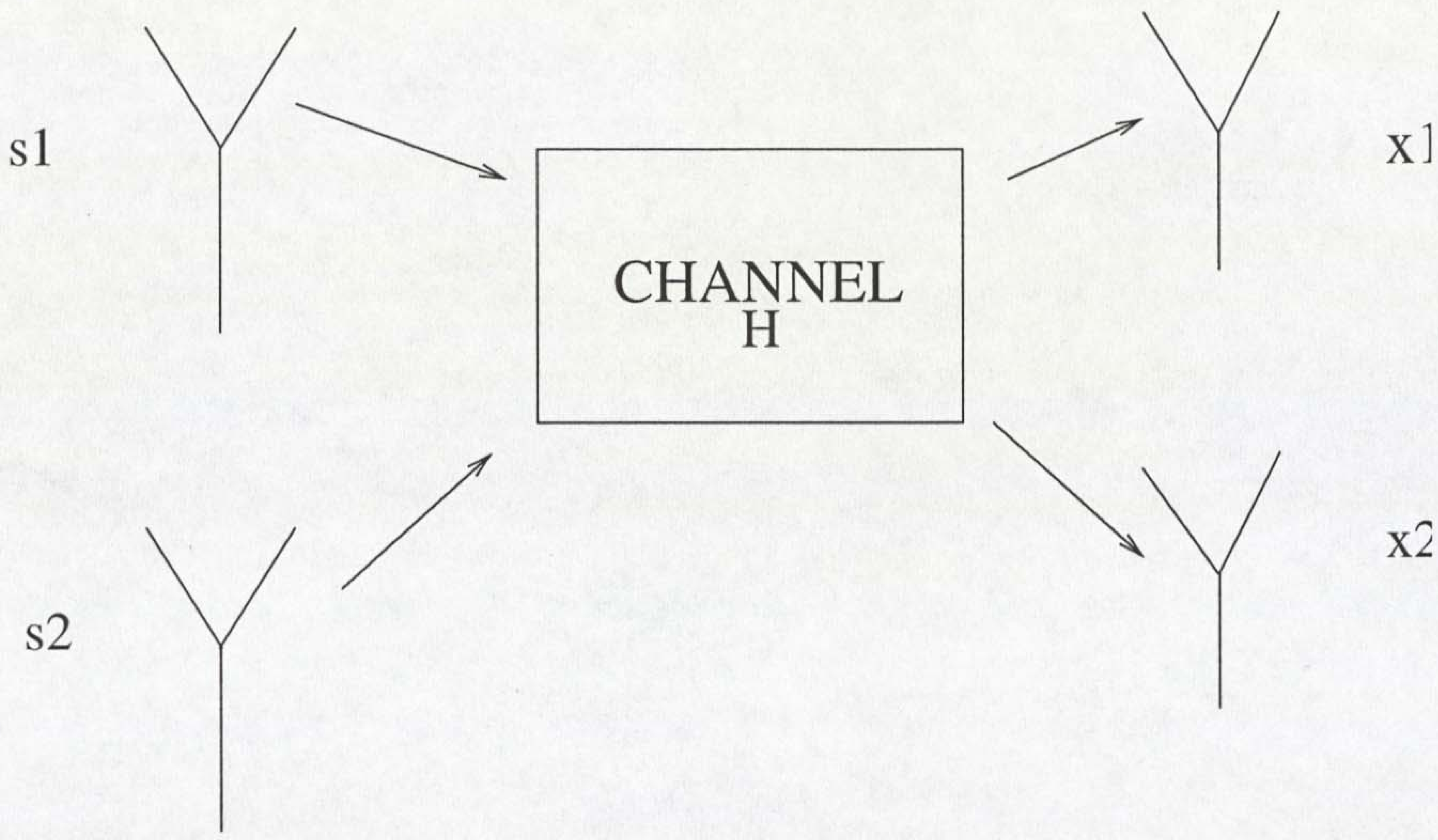

Figure 2.3: Simple 2 × 2 system Paulraj [17], Barbarossa [42]

$$\mathbf{x}_j = \sum_{i=1}^{N_t} \mathbf{H}_{ji}\mathbf{s}_i + \mathbf{n}_j \tag{2.13}$$

$\mathbf{H}_{ji}$ is the channel convolution matrix for the i, j^{th} transmit-receive antenna pair with $\mathbf{H}_{j,i}$ consisting of L paths, where L is the upper bound on the maximum delay spread or the channel order. $\mathbf{n}_j$ denotes the noise at the j^{th} receive antenna.

2.5.2 Channel Capacity

The following assumptions are made for simplifying the expressions for the MIMO channel capacity equations [17],[42] below.

- A Rayleigh flat fading channel model is assumed.
- The data covariance matrix $\mathbf{R}_{ss}$ assumed to be of unit variance and independent of noise $\mathbf{n}$.
- The case of the Channel State Information (CSI) unknown at transmitter is used throughout this section and the thesis.

Thus the covariance matrix of the net received signal $\mathbf{x} = [\mathbf{x}_1^T \mathbf{x}_2^T \ldots \mathbf{x}_{N_r}^T]^T$ can be expressed as shown in equation below.

$$\mathbf{R}_{xx} = \frac{E_s}{N_t}\mathbf{H}\mathbf{R}_{ss}\mathbf{H}^H + N_0\mathbf{I}_{N_r} \tag{2.14}$$

where E_s is the total average energy at the transmitter and N_0 is the noise power. $\mathbf{H}$ denotes the net channel matrix of size $N_r \times N_t$. $\mathbf{I}_{N_r}$ represents the identity matrix of size $N_r \times N_r$. The channel capacity for the flat fading channel is expressed as [17], [42], [44]

$$C = \max_{Tr(\mathbf{R}_{xx})=N_t} log_2\ det\left(\mathbf{I}_{N_r} + \frac{E_s}{N_t N_0}\mathbf{H}\mathbf{R}_{ss}\mathbf{H}^H\right) \tag{2.15}$$

Applying the assumptions stated above, the capacity equation is further simplified as

$$C = log_2\ det\left(\mathbf{I}_{N_r} + \frac{E_s}{N_t N_0}\mathbf{H}\mathbf{H}^H\right) \tag{2.16}$$

$\mathbf{HH}^H = \mathbf{Q}\lambda\mathbf{Q}^H$ represents the Eigen Value Decomposition (EVD) of $\mathbf{HH}^H$ with λ denoting the diagonal matrix of eigenvalues [17]. Utilising the identity that $det(\mathbf{I}_m + \mathbf{AB}) = det(\mathbf{I}_n + \mathbf{BA})$ [17], where $\mathbf{A}$ $(m \times n)$ and $\mathbf{B}$ $(n \times m)$, Eqn. (2.16) is further simplified as

$$C = log_2 \ det \left(\mathbf{I}_{N_r} + \frac{E_s}{N_t N_0} \lambda \right) \tag{2.17}$$

The above equation can be expressed equivalently as follows [17] with r denoting the rank of the channel.

$$C = \sum_{i=1}^{r} log_2 \left(1 + \frac{E_s}{N_t N_0} \lambda_i \right) \tag{2.18}$$

which expresses the MIMO channel capacity as sum of the capacities of r SISO channels, with λ_i denoting the positive eigen values of $\mathbf{HH}^H$ and $\frac{E_s}{N_t}$ representing the transmit power. In the case of $N_t = N_r = N_a$, the capacity equation simplifies as follows with ρ being the average SNR at the receiver.

$$C \cong N_a log_2(1 + \rho) \tag{2.19}$$

In the next chapter, MIMO MC-CDMA uplink system is presented coupled with three blind channel estimation methods and equalisation is performed using MMSE and LSFE schemes.

Chapter 3

MIMO MC-CDMA

3.1 Introduction

Spectral efficiency and higher data demands are driving the mobile research fraternity to further improve existing techniques. MIMO systems [41] have been shown to be an effective solution to improve the spectral efficiency of wireless communications. Much research has been carried out on MIMO techniques, such as the Vertical Bell Labs Layered Space Time (V-BLAST) receiver [40], which employs successive interference cancellation. Zhu and Murch [45] proposed a Layered Space Frequency Equalisation structure for single carrier MIMO systems, which is an extension of V-BLAST in frequency selective channels.

MIMO MC-CDMA systems, which take the advantages of spatial diversity and the MC-CDMA scheme [31], enable different users to share the same spectrum and thereby make utmost use of the available spectrum. Each user is differentiated by means of unique signature (spreading) codes, which spreads the information bits in the frequency domain. Punnoose et al [46] utilised an LSFE structure for MIMO MC-CDMA systems in the downlink, which was shown to yield significant performance improvement over the conventional MMSE approach.

Channel estimation techniques using pilot or training symbols reduces the bandwidth efficiency, which has led to researchers investigating various blind channel estimation approaches. Blind Source Separation (BSS) [47] or Higher Order Statistics (HOS) [48] based methods are not easily applicable for the uplink MC-CDMA signal, where each subcarrier involves a sum of the spread signals of all users leading to a nearly Gaussian mixture.

A few Second Order Statistics (SOS) based methods have been proposed for MC-CDMA namely Turelli's method [49], and others [50],[51], which can be classified as Masked Correlation Difference methods and its variants. These methods are simpler to implement (have lower complexity) compared to the subspace based and subspace related approaches but yield poorer performance. The primary notion behind these approaches is the masking of the received signal with each user's spreading code and obtaining the post-masked autocorrelation. This is then followed by the taking the difference of the pre and post masked autocorrelation matrices to yield the estimate of the desired channel frequency response.

Turelli [49] proposed a correlation difference matrix approach for uplink scenario using aperiodic spreading codes. It computes the difference in the correlations between pre-masked and post masked received signals. A simplified LS approach to solving the cost function was also proposed in the paper. This method was then extended to the case for both periodic as well as aperiodic spreading codes by Dongming et al [50]. A further improvement by using an additional PIC stage was proposed by Chongsen et al [51] for the MC-CDMA uplink scenario. All the above papers assumed a quasi-synchronous model. Sand et al [52] proposed an iterative channel estimtion scheme for MC-CDMA systems using pilot symbols for adaptive channel estimation in a time varying channel. A simple subspace based approach was proposed by Jun Wu et al [53] for a synchronous downlink scenario while References [54] and [55] proposed a blind multi-user detection for MC-CDMA systems using the subspace approach to obtain the equaliser weights directly. A similar approach for an MC-CDMA system with virtual carriers was proposed by Roy [56].

MMSE based detection schemes have already been proposed for MC-CDMA systems [57]-[58]. Interference cancellation schemes proposed to aid in reducing the effects of MAI, include a SIC VBLAST technique proposed for coded MIMO MC-CDMA systems as in [59] as well as for PIC based interference cancellation [60] in MC-CDMA systems. A review of the parallel and successive IC schemes available for MC-CDMA systems was done in [61], while [62] proposed a hybrid IC scheme for MC-CDMA systems.

In this chapter, three blind channel estimation methods for MIMO MC-CDMA uplink systems are investigated. An Autocorrelation Contribution Method (ACM) based method is proposed, in comparison with the Subspace (SS) method and the Subspace Approximation (SA) method. ACM was first proposed in [63] for DS-CDMA systems. ACM provides performance similar to that obtained by subspace based approach, with the added advantage of eliminating the need for rank estimation (i.e an EVD stage) as in the subspace method. The SA method too eliminates the need for an initial EVD stage. Also LSFE [45], [46] in conjunction with the blind channel estimation methods for this scenario was examined, which yielded significant performance enhancement compared to the conventional MMSE method. The effect of receive diversity on performance of LSFE is also shown and a brief complexity analysis of both the blind channel estimation methods as well as the equalisers used concludes the chapter.

3.2 System Model

The block diagram of an uplink MC-CDMA system is shown in Figures 3.1 and 3.2 [31]. The use of multiple antennas at the receiver (base station) is feasible and practical as against the mobile user, where power and complexity issues become paramount. For simplifying matters a simple 2 user Uplink MC-CDMA system is illustrated with each user sending its information using an antenna. At the base-station (receiver) end, receive diversity can be utilised by introducing multiple antennas.

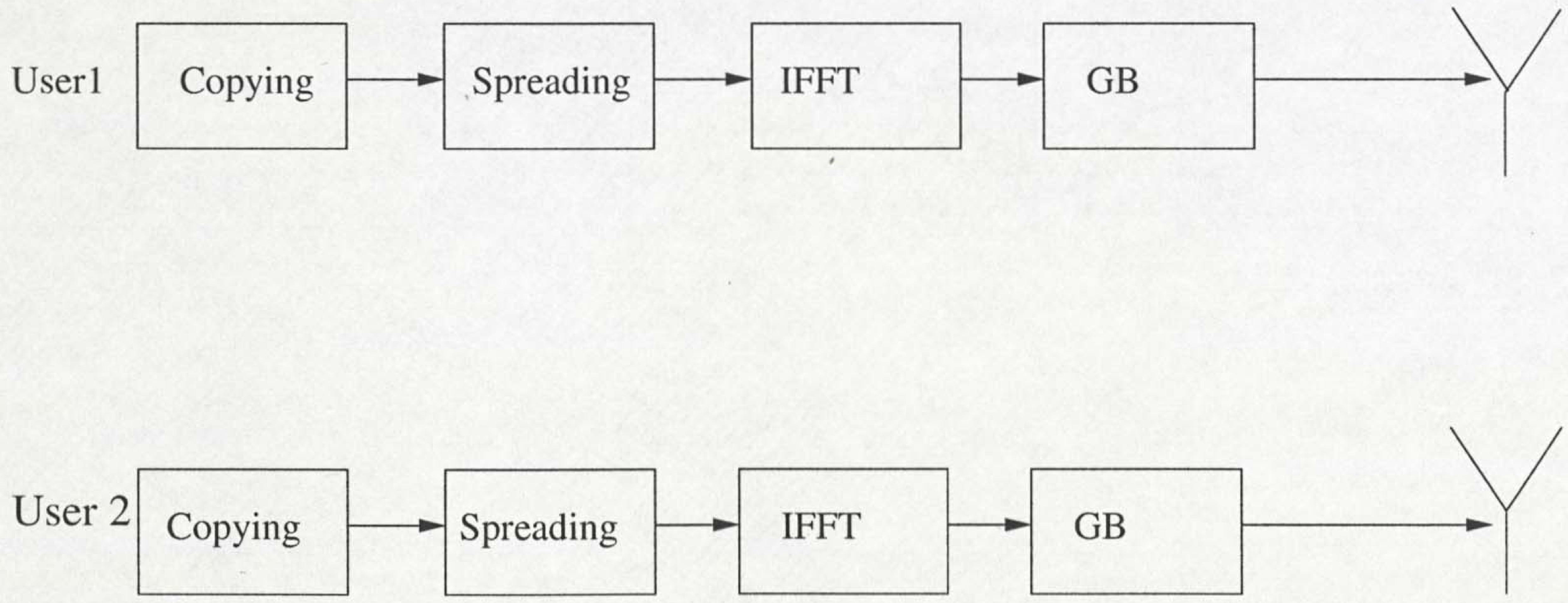

Figure 3.1: Uplink MC-CDMA Transmitter Verde [54]

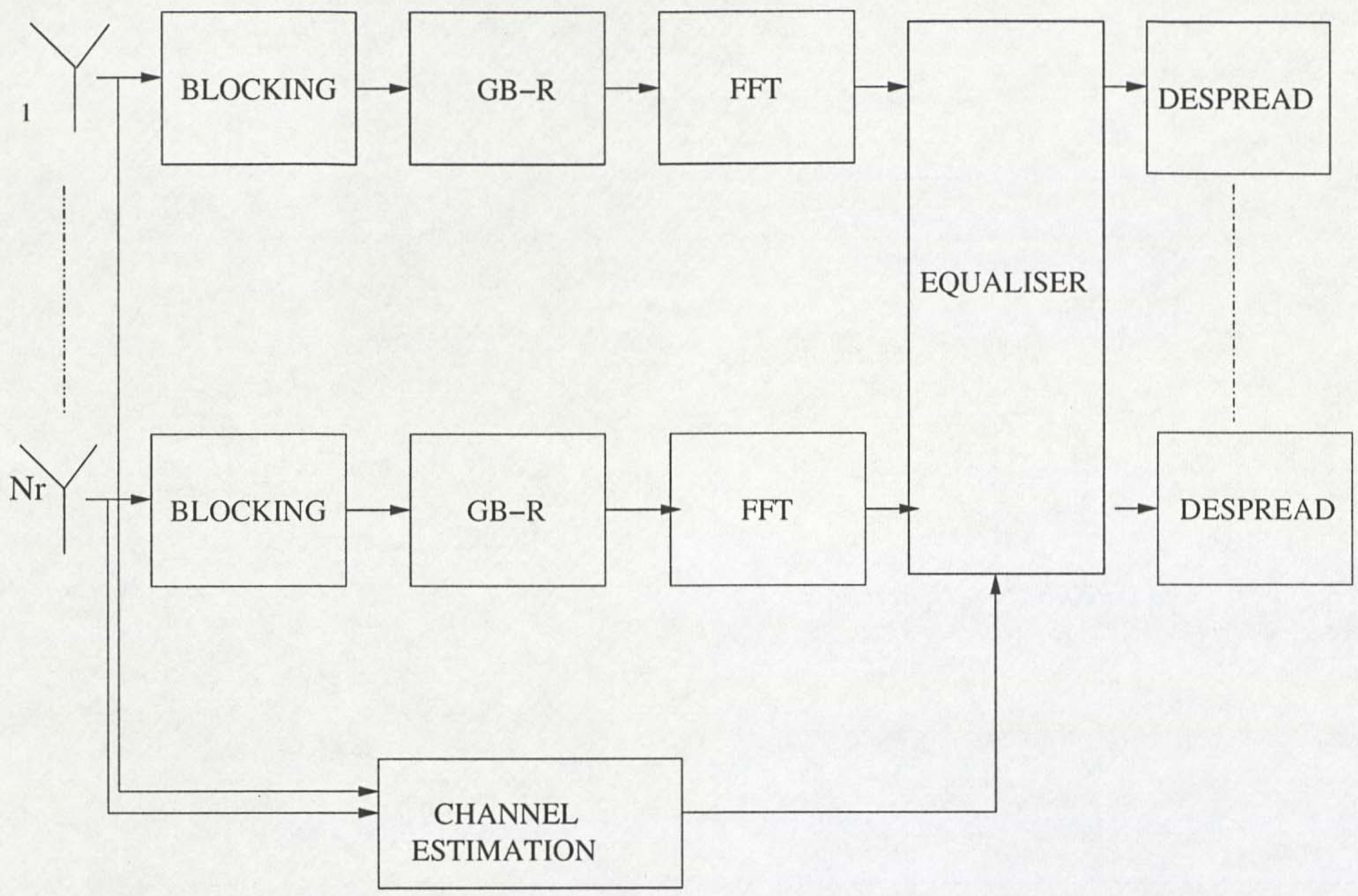

Figure 3.2: Uplink MC-CDMA Receiver Verde [54]

The data symbols are assumed to be i.i.d with unit variance. The following are the assumptions made regarding the system model considered here.

1. N_t and N_r are the number of transmit and receive antennas respectively and N_t is equal to number of users. Also a synchronous system model is assumed.

2. L is the channel order, which is assumed to be known at receiver. In practice the channel order is estimated along with the noise variance depending on the type of blind channel estimation algorithm used.

3. The signals are assumed to be white with zero-mean and unit variance.

4. The spreading codes used are Walsh-Hadamard codes of size N, where N is the number of subcarriers. Here the spreading gain is chosen to be equal to the number of subcarriers.

Contents

5. BPSK (Binary Phase Shift Keying) modulation is used for each user.

Figure 3.1 shows the uplink MIMO MC-CDMA transmitter. Each user's signal is first copied N times and then spread in the frequency domain using the corresponding code bit for each carrier. The spread signal is then transformed using the IFFT. A cyclic prefix is added at the beginning of each block to negate the effects of ISI caused by multi-path phenomenon. This is achieved by choosing the length of the cyclic prefix to be L. After serialisation, the different users spread signals are then transmitted from their respective antennas through a frequency selective Rayleigh fading channel.

At the j^{th} receive antenna, after removal of the guard band and demodulation using FFT, the received signal can be expressed as

$$\mathbf{X}_j = \sum_{i=1}^{N_t} \mathbf{H}_{ji}\mathbf{d}_i + \mathbf{V}_j \tag{3.1}$$

where $\mathbf{H}_{ji}$ is the channel frequency response matrix between the j^{th} receive antenna and the i^{th} transmit antenna. $\mathbf{V}_j$ is the noise at the j^{th} receive antenna while $\mathbf{d}_i$ is the i^{th} user's spread signal and $\mathbf{X}_j$ represents the received signal at antenna j.

The spreading is done in frequency domain using a diagonal spreading matrix $\mathbf{C}_i$ of size $N \times N$ as shown in equation below. It can also be expressed succintly as $\mathbf{diag}(c_i(0)c_i(1)\ldots c_i(N-1))$. Thus the ith user's signal can be represented as $\mathbf{d}_i = \mathbf{C}_i\mathbf{s}_i$, where $\mathbf{s}_i$ is the size $N \times 1$ vector of the i^{th} user's data block.

$$\mathbf{C}_i = \begin{bmatrix} c_i(0) & 0 & 0 & 0 \\ \vdots & c_i(1) & 0 & 0 \\ 0 & \vdots & . & \vdots \\ \vdots & \ldots & .. & 0 \\ 0 & 0. & . & c_i(N-1) \end{bmatrix} \tag{3.2}$$

MAI and ISI are the main obstacles facing MC-CDMA systems and there have been numerous approaches researched to aleviate this problem with Multi-User detection (MUD) [64] being one of the more prominent approaches investigated. In the model considered, the near-far problem is not taken into account assuming that appropriate power control is made use of to counter the effects of a strong user drowning out the weak signal of a far away user. Also the channel considered is a frequency selective fading one, which is assumed to be quasi-static i.e the channel varies slowly such that it is assumed to be constant for a block of symbols.

MC-CDMA, which does spreading in the frequency domain, can also be reformulated in the DS-CDMA mould. This is done as shown below in order to facilitate application of the ACM algorithm to the model later in this chapter.

The received signal at the j^{th} receive antenna in a N_t transmit, N_r receive antenna system can be written in matrix-vector form as

$$\mathbf{X}_j = \sum_{i=1}^{N_t} \mathbf{H}_{j,i}(\mathbf{C}_i \odot \mathbf{s}_i) + \mathbf{V}_j \tag{3.3}$$

where $\mathbf{H}_{j,i}$ is the diagonal channel matrix of size $N \times N$ between the jth receive antenna and the ith user. $\mathbf{C}_i$ represents the i^{th} user's diagonal spreading code matrix of size $N \times N$, while $\mathbf{s}_i$ denotes the information or data of user i. $\mathbf{V}_j$ represents the additive white Gaussian noise at the receiver with $\odot$ representing Hadamard product i.e. element by element multiplication.

3.3 Blind Channel Estimation

The blind channel estimation methods, as mentioned earlier, are all SOS based and utilise the knowledge of the spreading codes, which is assumed known at the base-station (receiver) end. SOS based methods provide fast convergence while yielding acceptable performance levels [65] - [66].

All blind channel estimation methods estimate the channel values upto a scalar ambiguity, which is removed by means of a single pilot symbol. The channel is assumed to be a quasi-static Rayleigh fading channel, which is assumed to be slowly-time varying and stays constant over a block. The next section focusses on the extension of three methods namely the SS method, the SA method and the ACM method to the MIMO MC-CDMA uplink case.

3.3.1 Subspace algorithm

The subspace algorithm works by utilising the structure inherent in the signal. The EVD of the auto-correlation matrix $\mathbf{R}_j$ of the received signal at receive antenna j yields the signal and noise subspace denoted by $\mathbf{U}_{sj}$ and $\mathbf{U}_{nj}$ respectively [53], [66]. Subspace based channel estimation schemes have been applied to SIMO DS-CDMA multi-user systems [64]. In MIMO systems i.e. with more than one receive antenna, the subspace equations are applied separately for each receive antenna to estimate all the N_t channels.

$$\mathbf{R}_j = E[\mathbf{y}_j\mathbf{y}_j^H] = \begin{bmatrix}\mathbf{U}_{sj} & \mathbf{U}_{nj}\end{bmatrix} \begin{bmatrix} \lambda_s & 0 \\ 0 & \lambda_n \end{bmatrix} \begin{bmatrix} \mathbf{U}_{sj}^T \\ \mathbf{U}_{nj}^T \end{bmatrix} \tag{3.4}$$

$$\lambda_s = \mathbf{diag}(\lambda_1, \lambda_2, ..\lambda_{N_t}) \tag{3.5}$$

$$\lambda_n = \sigma_n^2 \mathbf{I}_{N-N_t} \tag{3.6}$$

From the EVD shown, $\lambda_\mathbf{s}$ denotes the diagonal matrix of signal eigenvalues of rank N_t, i.e the number of users, while $\lambda_\mathbf{n}$ denotes the corresponding noise subspace eigenvalues of size $N - N_t$. The noise subspace is obtained as the space spanned by all the singular values $\leq \sigma_n^2$, where σ_n^2 is the noise power. Assuming that the signal and noise are uncorrelated, the fundamental subspace relation can be expressed as

$$\mathbf{U}_{nj}\mathbf{h}_{j,i} = 0 \tag{3.7}$$

where $\mathbf{h}_{j,i}$ is the channel impulse response at receive antenna j from user i. The subspace equation involves the minimisation of the following quadratic cost function

$$\mathbf{h}_{j,i} = \arg \min_{||\mathbf{h}||=1} ||\mathbf{h}_{j,i}^H \mathbf{Q}_{j,i} \mathbf{h}_{j,i}||^2 \tag{3.8}$$

where the matrix $\mathbf{Q}_{j,i}$ for user i is given as follows [53].

$$\mathbf{Q}_{j,i} = \mathbf{P}^H \mathbf{C}_i^H \mathbf{U}_{nj} \mathbf{U}_{nj}^H \mathbf{C}_i \mathbf{P} \tag{3.9}$$

Here $\mathbf{C}_i$ denotes the diagonal spreading code matrix for user i and $\mathbf{P}$ is the phase rotation matrix due to the delay of the channel and it is represented as

$$\mathbf{P} = \begin{bmatrix} 1 & 1 & .... & 1 \\ 1 & \phi... & \phi^{L-1} & \\ : & :....: & & \\ : & :... & : & \\ 1 & \phi^{N-1} & \phi^{(L-1)(N-1)} & \end{bmatrix} \tag{3.10}$$

where $\phi = e^{-j\frac{2\pi}{N}}$. It is a submatrix of the FFT matrix, taking only the first L columns. Since each user transmits from a separate antenna, the code is used to help in identifying the particular channel. As is inherent in all blind estimation methods, there is a scaling factor, which can be resolved by using a training symbol.

3.3.2 Subspace Approximation

Subspace methods [65], [66] have been extensively studied. Assuming that certain conditions are satisfied, the principle idea as stated before involves obtaining the noise subspace and constructing a cost function to be minimised. This cost function utilises the orthogonality between the noise-subspace and the signal subspace to obtain the required channel estimates.

The initial step of the subspace approach requires the determination of the rank of the noise-subspace after an EVD. Rank detection methods however are dependent on SNR and sample size, which are two key factors that affect subspace based blind estimation methods [67],[68]. Blind channel estimation using the SA method was initially proposed for downlink DS-CDMA systems [67]. It utilises the Power of R method, initially proposed for multi-user detection in [69]. The key idea involves replacing the noise subspace component in the cost function minimisation problem with the inverse of autocorrelation matrix raised to a positive integer. As value of positive integer increases, the approximation error is reduced. It is also easily applied to adaptive tracking and estimation cases. The algorithm [67] is briefly explained in this section. The EVD of the autocorrelation matrix of the received signal can be expressed as

$$\mathbf{R}_j = \quad [\mathbf{U}_{sj} \quad \mathbf{U}_{nj}] \begin{bmatrix} \lambda_s & 0 \\ 0 & \lambda_n \end{bmatrix} \begin{bmatrix} \mathbf{U}_{sj}^T \\ \mathbf{U}_{nj}^T \end{bmatrix} \tag{3.11}$$

The cost function minimisation for the SS method in section 3.3.1 is obtained by substituting Eqn. (3.9) into Eqn. (3.8) yielding

$$\hat{\mathbf{h}_{j,i}} = \underset{||\mathbf{h}_{j,i}=1||}{\arg\min} \; \mathbf{h}^H \mathbf{P}^H \mathbf{C}_i^H \mathbf{U}_{nj} \mathbf{U}_{nj}^H \mathbf{C}_i \mathbf{P} \mathbf{h}_{j,i} \tag{3.12}$$

Rearranging Eqn. (3.11) leads to

$$\sigma^2 \mathbf{R}_j^{-1} = \mathbf{U}_{nj} \mathbf{U}_n^H + \mathbf{U}_{sj} \mathbf{diag} \left(\frac{\sigma^2}{\lambda_i^2 + \sigma^2} \right) \mathbf{U}_{sj}^H \tag{3.13}$$

As the power of R is raised to a positive integer i.e. $m = 1, 2, 3..$, the above equation becomes

$$\sigma^{2m}\mathbf{R}_j^{-m} = \mathbf{U}_{nj}\mathbf{U}_{nj}^H + \mathbf{U}_{sj}\mathbf{diag}\left(\frac{\sigma^2}{\lambda_i^2+\sigma^2}\right)^m \mathbf{U}_{sj}^H \tag{3.14}$$

Taking the limits of the previous equation to infinity, the following approximation is obtained [67].

$$\lim_{m\to\infty} \sigma^{2m}\mathbf{R}_j^{-m} = \mathbf{U}_{nj}\mathbf{U}_{nj}^H \tag{3.15}$$

The subspace approximation algorithm [67] is obtained, where the replacement by the positive power of the inverse of the auto-correlation matrix eliminates need for subspace rank determination as well as initial EVD stage of subspace method. It also alleviates the effects of noise and sample size. The scalar σ^{2m} can be neglected as it is not important in the channel estimation stage [67] and thus yields the final cost function equation for SA approach as

$$\hat{\mathbf{h}}_{j,i} = \arg \min_{||\mathbf{h}_{j,i}=1||} \mathbf{h}_{j,i}^H\mathbf{P}^H\mathbf{C}_i^H\mathbf{R}_j^{-m}\mathbf{C}_i\mathbf{P}\mathbf{h}_{j,i} \tag{3.16}$$

The value of m is usually chosen to be 2 or 3 as effects of right hand term in Eqn. (3.9) are already small with these values, thereby avoiding the need for further computational complexity.

3.3.3 ACM Method

The ACM method as proposed in [63] utilises the entire autocorrelation matrix to find the estimates, with its key advantages being that no rank or noise power estimation, is required as in subspace based approaches. The equivalence of MMSE algorithm with that of MSINR receiver is made use of in deriving the equations for ACM algorithm [63]. For simplicity, user i is assumed to be the user of interest and the channel at receive antenna j is estimated in the following description. Also the spreading code matrix $\mathbf{C}_{N_t}$ for each user is assumed to be known at the base-station. As in subspace approach for a MIMO system, the following is repeated for each receive antenna.

The general MMSE weight coefficients for user i at receive antenna j is given by

$$\mathbf{w}_{mmse,i} = \mathbf{R}_j^{-1}\mathbf{C}_i\mathbf{P}\mathbf{h}_{j,i} \tag{3.17}$$

The MSINR receiver for user i [63] is represented as

$$\mathbf{w}_{msinr,i} = \arg \max_{||\mathbf{v}||=1} \frac{\mathbf{v}^H\mathbf{h}_{j,i}\mathbf{h}_{j,i}^H\mathbf{v}}{\mathbf{v}^H\mathbf{R}_j\mathbf{v}} \tag{3.18}$$

The above equations are equivalent as a result of the MMSE receiver also maximising the SINR [63], thereby yielding for user i,

$$\hat{\mathbf{h}_{j,i}} = \arg \max_{||\mathbf{v}||=1} \frac{\mathbf{v}^H\mathbf{P}^{\mathrm{H}}\mathbf{C}_i^{\mathrm{H}}\mathbf{R}_j^{-1}\mathbf{R}_{j,b_i}\mathbf{R}_j^{-1}\mathbf{C}_i\mathbf{P}\mathbf{v}}{\mathbf{v}^H\mathbf{P}^{\mathrm{H}}\mathbf{C}_i^H\mathbf{R}_j^{-1}\mathbf{C}_i\mathbf{P}\mathbf{v}} \tag{3.19}$$

where $\mathbf{R}_{j,b_i} = \mathbf{h}_{j,i}\mathbf{h}_{j,i}^{\mathrm{H}}$ is the subspace matrix of desired user symbols $\mathbf{b}_i$ while $\hat{\mathbf{h}_{j,i}}$ denotes the estimate of channel between receive antenna j and user i. The subspace matrix $\mathbf{R}_{j,b_i}$ is replaced by the whole signal space $\mathbf{R}_{j,b}$, which can be expressed as $\mathbf{R}_{j,b} = \mathbf{h}_{j,i}\mathbf{h}_{j,i}^{\mathrm{H}} + \mathbf{H}_{\mathrm{MAI}}\mathbf{H}_{\mathrm{MAI}}^{\mathrm{H}}$. $\mathbf{H}_{\mathrm{MAI}}$ represents the contribution due to the other users (interference). The matrix $\mathbf{R}_{j,b}$ contains the subspace of the entire signal including that of the interfering users. The matrices $\mathbf{R}_{j,b}$ and $\mathbf{R}_j$ are related

$$\mathbf{R}_j = \mathbf{R}_{j,b} + \sigma^2\mathbf{I} \tag{3.20}$$

The ACM algorithm can also be represented equivalently as

$$\hat{\mathbf{h}_{j,i}} = \arg\max_{||\mathbf{v}||=1} \frac{\mathbf{v}^{\mathrm{H}}\mathbf{P}^{\mathrm{H}}\mathbf{C}_i^{\mathrm{H}}\mathbf{R}_j^{-1}(\mathbf{R}_j - \sigma^2 I)\mathbf{R}_j^{-1}\mathbf{C}_i\mathbf{v}}{\mathbf{v}^{\mathrm{H}}\mathbf{P}^{\mathrm{H}}\mathbf{C}_i^{\mathrm{H}}\mathbf{R}_j^{-1}\mathbf{P}\mathbf{C}_i\mathbf{v}} \tag{3.21}$$

Using the equation above, the channel impulse response for user i can be estimated as

$$\hat{\mathbf{h}_{j,i}} = \arg\min_{||\mathbf{v}||=1} \frac{\mathbf{v}^{\mathrm{H}}\mathbf{P}^{\mathrm{H}}\mathbf{C}_i^{\mathrm{H}}\mathbf{R}_j^{-2}\mathbf{C}_i\mathbf{P}\mathbf{v}}{\mathbf{v}^{\mathrm{H}}\mathbf{P}^{\mathrm{H}}\mathbf{C}_i^{\mathrm{H}}\mathbf{R}_j^{-1}\mathbf{C}_i\mathbf{P}\mathbf{v}} \tag{3.22}$$

It is evident from the above equations that there is no decomposition required and hence no rank estimation is required. Also no noise power estimation is required making it more robust to errors that may be caused as a result of rank estimation.

3.4 Equalisation

3.4.1 MMSE Equaliser

The received signal, after the blind channel estimation stage, is stripped of its guard band and then FFT processed. The frequency domain signal is then equalised using an MMSE equaliser [54], [70]. The soft estimate $\hat{\mathbf{d}}_i$ of $\mathbf{d}_i$ for user i is given as

$$\hat{\mathbf{d}}_i = \mathbf{w}_{mmse,i}^H \mathbf{X} \tag{3.23}$$

where $\mathbf{X}$ is the net received signal expressed as $\mathbf{X} = [\mathbf{X}_1 \mathbf{X}_2 \ldots \mathbf{X}_{N_r}]$ with $\mathbf{X}_j$ defined as in Eqn. (3.1). The MMSE equaliser for the i^{th} user, $\mathbf{w}_{mmse,i}$ is expressed as shown below with $\hat{\mathbf{H}}_i$ denoting the estimate of $\mathbf{H}_i$

$$\mathbf{w}_{mmse,i} = \left(\sum_{i=1}^{N_t} \hat{\mathbf{H}}_i \hat{\mathbf{H}}_i^{\mathrm{H}} + \sigma^2 \mathbf{I} \right)^{-1} \hat{\mathbf{H}}_i \tag{3.24}$$

The equalised signals are then despread using the respective user's codes before finally being fed into a hard decision device to obtain the final estimate of user's signal.

3.4.2 LSFE Equaliser

We extend the work of Zhu et al [45], [46] to the uplink MIMO MC-CDMA system. The block diagram of the LSFE is shown in Figure 3.3, which consists of a blind channel estimation stage followed by N_t LSFE stages performing interference cancellation.

The LSFE technique is modified such that the ordering of the streams is done based on the sum of all mean square error (MSE) values calculated for each user. After detection of each of the users, the effects of the detected user is cancelled from the remaining signal. After cancelling the interference caused by the detected signal, the corresponding channel values are also removed to leave the undetected stream with its corresponding channel coefficients.

Figure 3.4 shows a particular stage of the LSFE equaliser/detector [45],

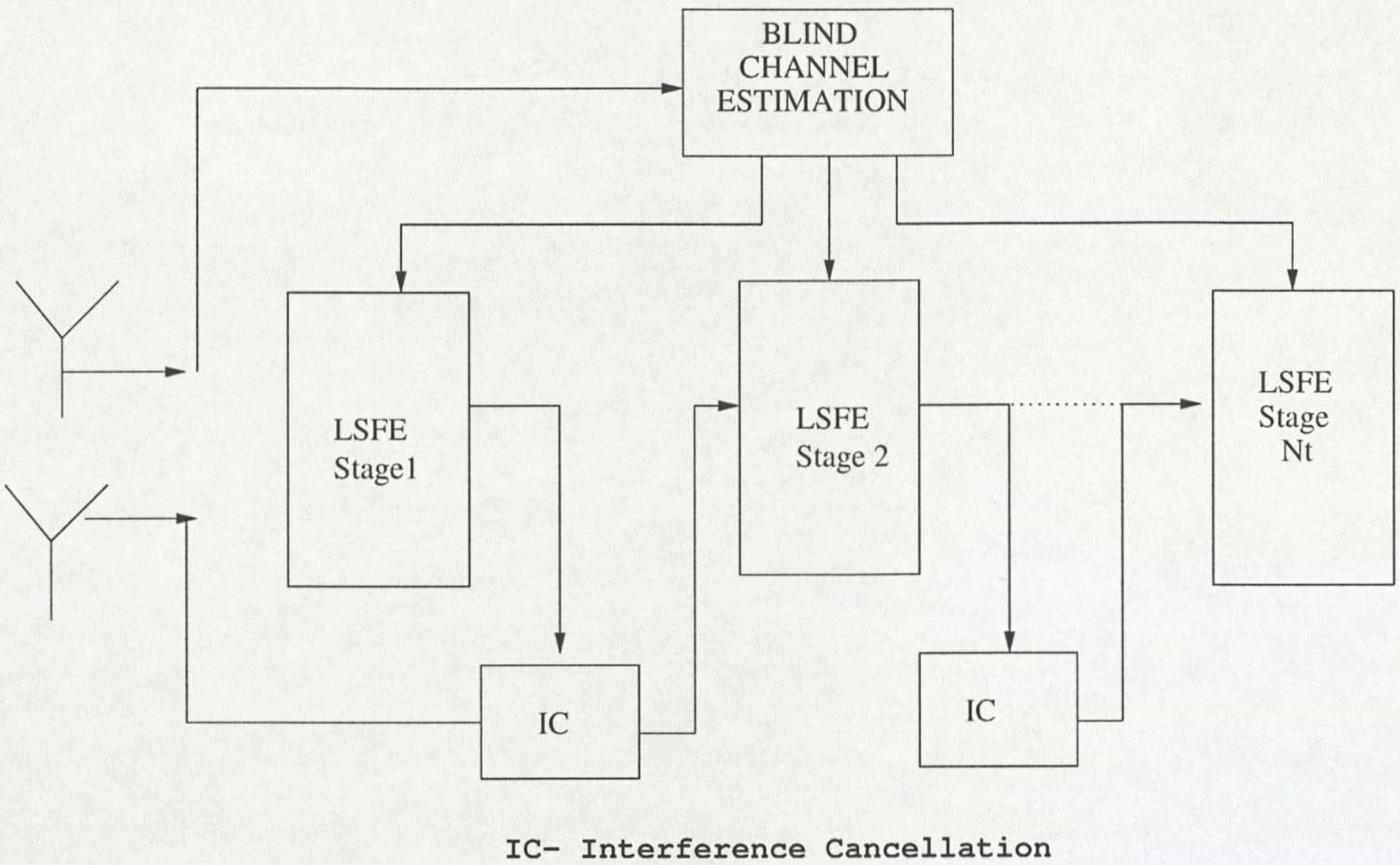

Figure 3.3: LSFE Receiver Zhu [45]

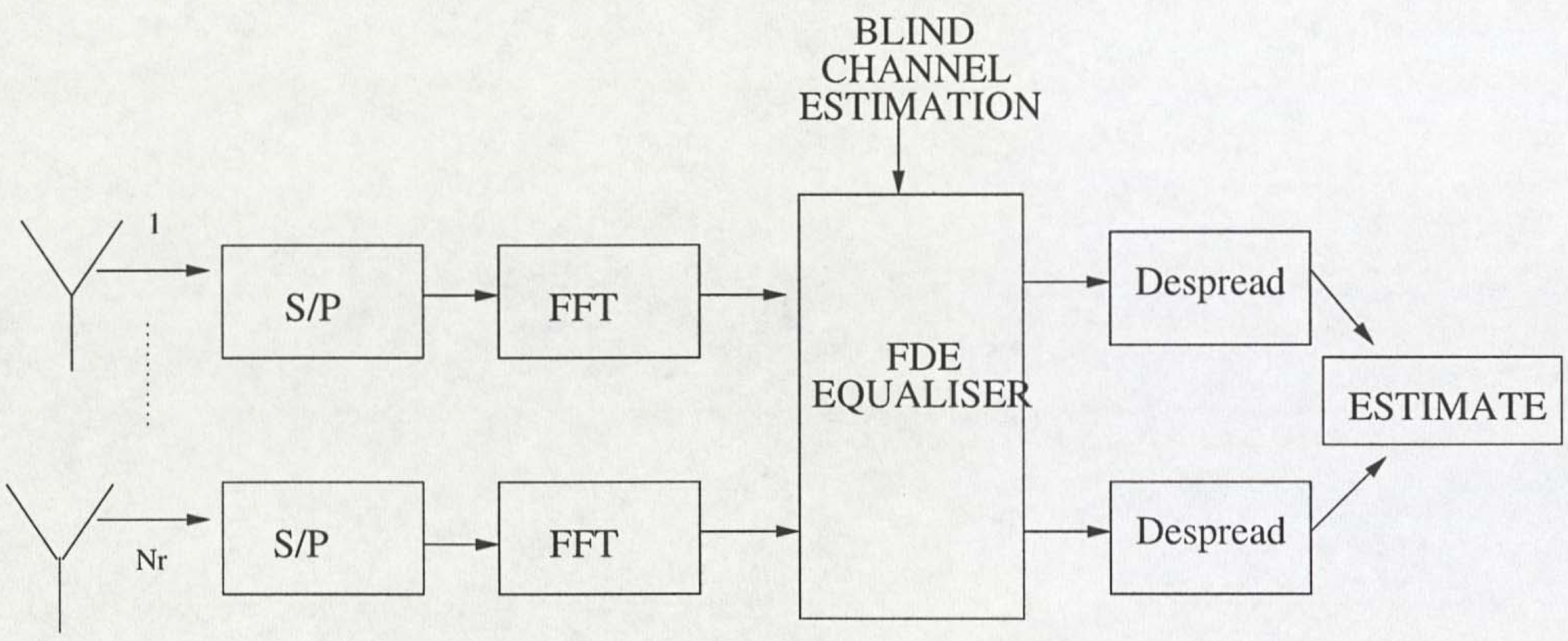

Figure 3.4: A Particular LSFE Stage Zhu [45]

where the FDE block performs equalisation as well as interference cancellation, before the hard decision is made by the decision device. The modified received signals with less interference are obtained after removing interference caused by previously detected signals.

Thus all users are detected and their corresponding interference are cancelled in a successive manner. The modified received signal at a particular stage on subcarrier n can be written as

$$\mathbf{X}[n] = \sum_{i} \hat{\mathbf{H}}_i[n]\mathbf{D}_i[n] + \mathbf{V}[n] \tag{3.25}$$

where the summation is over the undetected streams. $\hat{\mathbf{H}}_i[n]$ is the estimated channel frequency response matrix of size $N_r \times N_t$, with $\mathbf{D}_i[n]$ and $\mathbf{X}[n]$ representing the transmitted signal vector of size $N_t \times 1$ and the received signal vector of size $N_r \times 1$, while $\mathbf{V}[n]$ is the noise on subcarrier n. The FDE coefficients are calculated based on the MSE, which is defined as shown below with $\mathbf{s}_i$ representing a data block of user i.

$$MSE = E|\hat{\mathbf{s}}_i - \mathbf{s}_i|^2 \tag{3.26}$$

The weight vectors $\mathbf{w}_i$ can be expressed as

$$\mathbf{w}_i[n] = \mathbf{R}_i^{-1}[n]\mathbf{F}_i[n] \tag{3.27}$$

where $\mathbf{R}_i[n]$ is the autocorrelation matrix of the channel frequency response matrix estimate, denoted as $\hat{\mathbf{H}}_i[n]$ on subcarrier n for user i, expressed as

$$\mathbf{R}_i[n] = \sum_{i} \hat{\mathbf{H}}_i[n]\hat{\mathbf{H}}_i^H[n] + \sigma^2\mathbf{I} \tag{3.28}$$

where σ^2 is the noise power and $\mathbf{F}_i[n]$ is the channel sub-matrix corresponding to the user that is being detected expressed as

$$\mathbf{F}_i[n] = [\hat{\mathbf{H}}_{1,i}^T[n] \ldots \hat{\mathbf{H}}_{N_r,i}[n]^T]^T \tag{3.29}$$

The MSE on each subcarrier for each user is then calculated using the formula

$$MSE_{i,n} = 1 - \frac{1}{N}\sum_{n=0}^{N-1} \hat{\mathbf{H}}_i^{\mathrm{H}}[n]\mathbf{R}_i^{-1}[n]\hat{\mathbf{H}}_i[n] \tag{3.30}$$

The decision variable, which determines the ordering of the streams is given by the sum of all the MSE values as given by

$$MSE_i = \sum_n MSE_{i,n} \tag{3.31}$$

The ordering is done from the smallest MSE to the largest and detection of signals is then performed in this order.

3.5 Simulation Results

The simulation setup mainly involved a 2 transmit, 2 receive antenna uplink MIMO MC-CDMA system unless otherwise specified. The number of subcarriers was set to $N = 32$ with channel modelled as quasi-static block Rayleigh fading channel, which remains constant over the duration of a block. The channel impulse length was set to $L = 4$ with uniform power delay profile. A block size of $N_s = 100$ frames was used throughout. As blind channel estimation methods yield estimates upto a scalar factor, a pilot symbol was used to obtain the scaling factor and one pilot symbol was used for each transmit-receive antenna pair. Walsh-Hadamard codes, of size N i.e. equal to the number of subcarriers, were used throughout because of their orthogonal structure. The BER results are obtained as averaged over 100 Monte-Carlo runs for the setup described. BPSK modulation was used throughout. 'SS', 'SA' and 'ACM' represent the Subspace, Subspace Approximation and ACM method estimates respectively while 'Known' denotes the case with perfect channel knowledge at the receiver. In practical systems, a further error-correction step is also added at the receiver end in order to improve BER. The accepted limit for BER is usually in the range of $10^{-2} - 10^{-3}$.

Figure 3.5 compares the MMSE BER performances of a simple 2×2 MC-CDMA uplink system using known channel and blind channel estimates obtained via SS, SA and ACM based methods. As is clearly seen, the performance levels obtained using blind channel estimates are similar to each other and to the known channel case. The SS method has the least complexity with the ACM method requiring the most computations of the three but at the advantage of not needing an initial rank determination step.

Figure 3.6 shows the improvement in performance obtained as a result of using LSFE over the conventional MMSE equaliser. Here the interference cancellation of the detected users signal at each stage is responsible for the improvement in performance. The plot indicates the performance of the LSFE algorithm with blind channel estimates obtained using ACM, SS and SA based methods.

Figure 3.7 illustrates the improvements that are achieved through use of receive diversity at the receiver (base-station) end. The extra diversity afforded by multiple receive antennas provides significant gains. As is seen at SNR of 10 dB, an order of magnitude improvement is obtained with the SS,

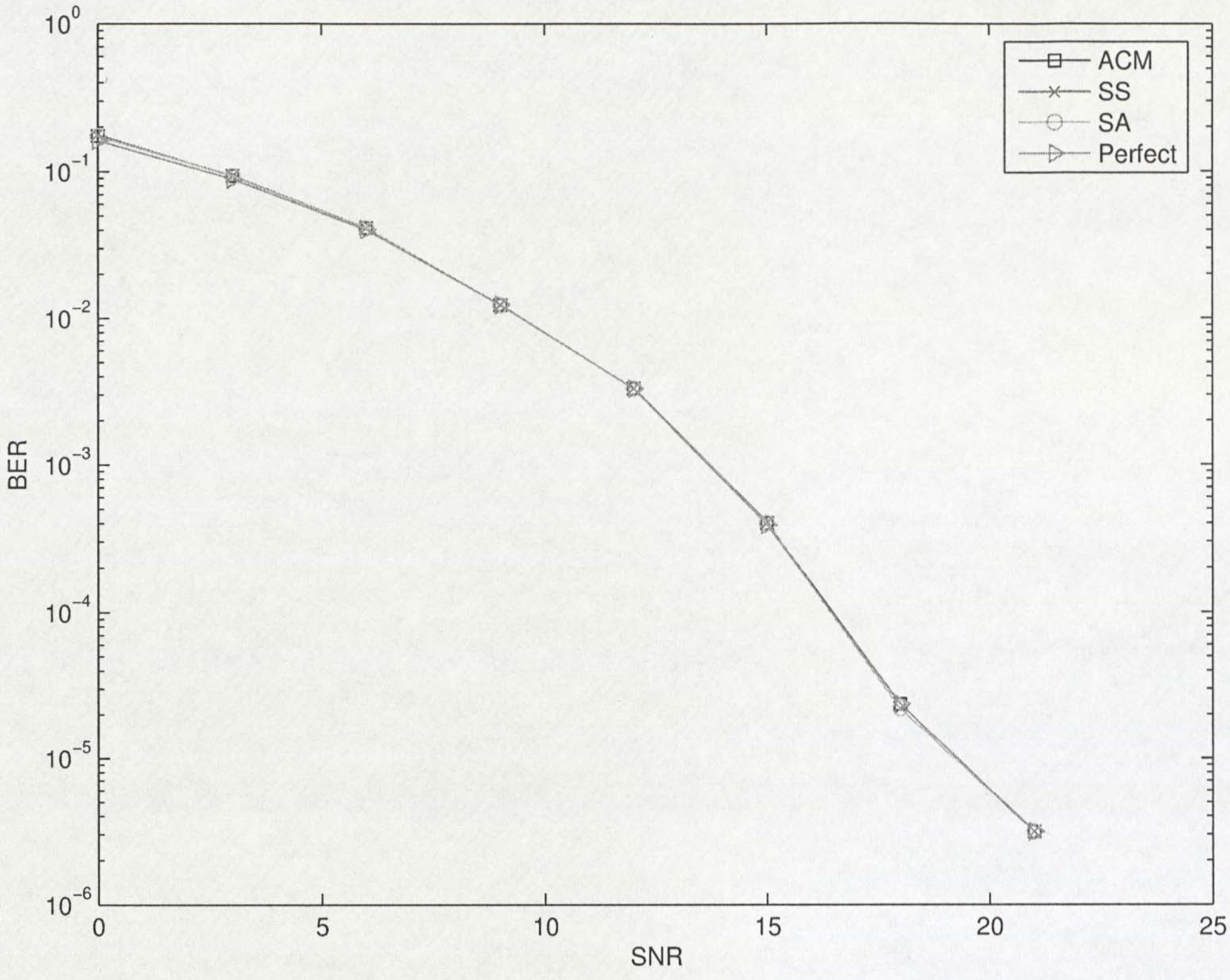

Figure 3.5: Performance of MMSE for SS, SA and ACM estimates

SA and ACM based MMSE equalisers achieving similar performance levels.

Figure 3.8 illustrates the effect of MAI on a 2 transmit, 2 receive antenna system at SNR of 20dB. As the number of users increase, the performance worsens but it is clear that LSFE based estimates perform better overall as compared to MMSE at lower loads but at higher loads i.e. large number of users, the overloaded system yields poor performance. Practical systems, as mentioned earlier do include error correction codes but they can only improve the performance to a limit. The degrading effect of MAI in the uplink scenario especially in an overloaded system provides an upper bound on achievable performance in this setup.

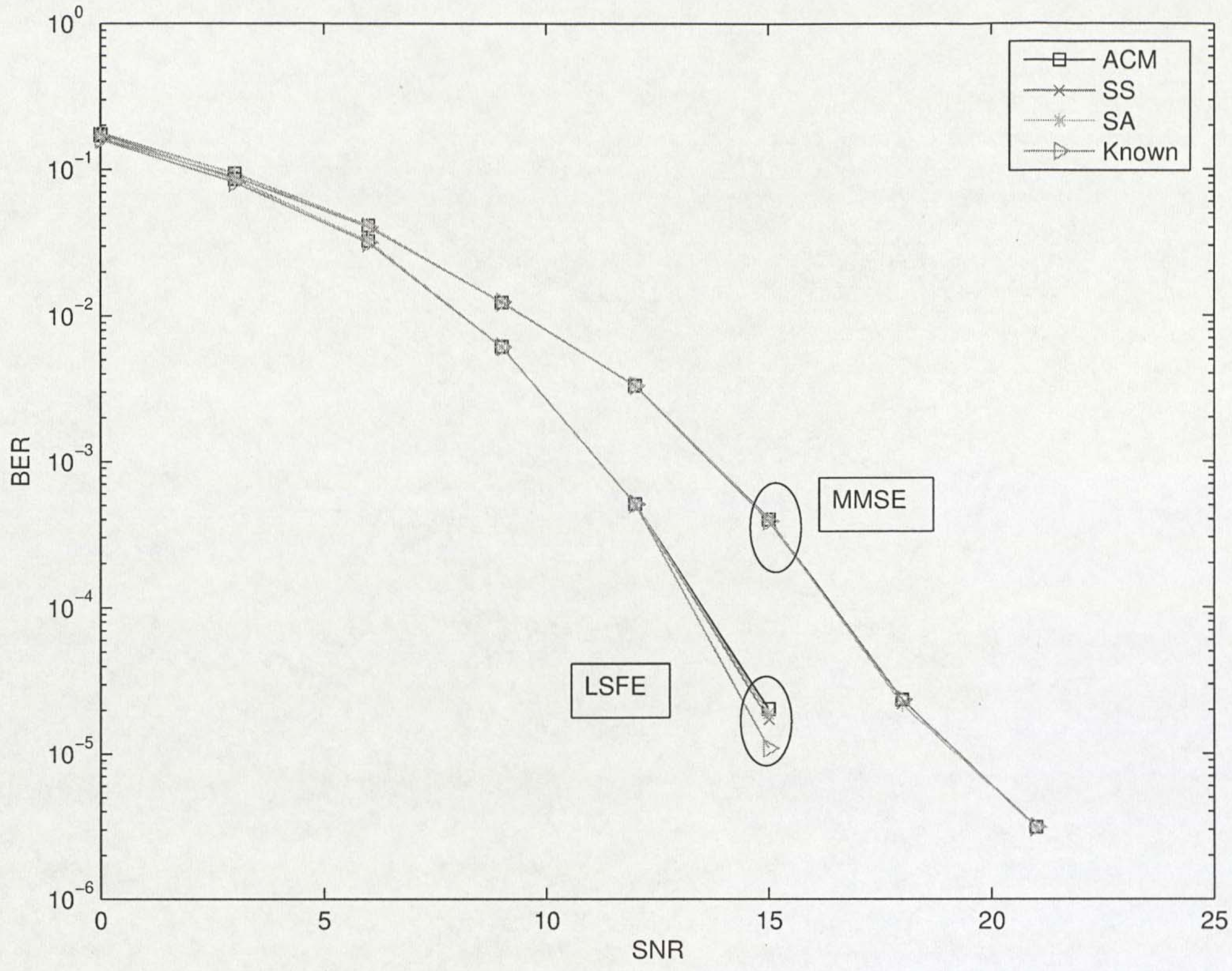

Figure 3.6: Performance of LSFE vs MMSE for 2x2 MIMO MC-CDMA Uplink with N=32,L=4

Figure 3.9 indicates the effect of varying channel orders on the 2 transmit, 2 receive antenna system at SNR of 15 dB. As the channel order increases, assuming that the guard band is of sufficient length (CP), the relative performance of the equalisers is not affected. The LSFE equaliser scheme yields improved performance compared to the MMSE scheme as a result of the SIC involved. As is seen, the performances of the 3 blind channel estimation methods is similar and quite stable with the LSFE scheme providing an order of magnitude improvement in performance when compared to the scenario utilising the conventional MMSE equaliser.

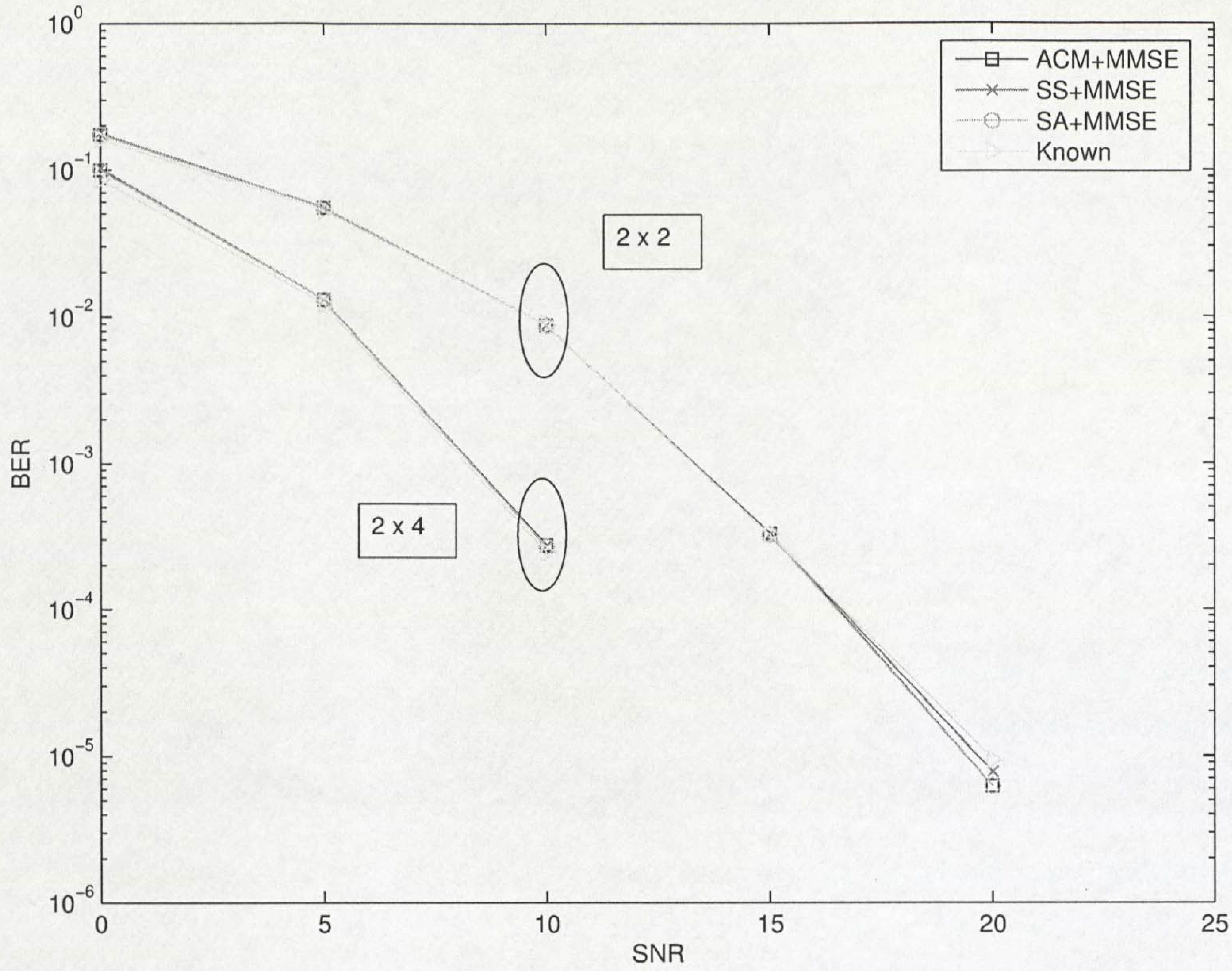

Figure 3.7: Performance of SS+LSFE, ACM+LSFE and SA+LSFE with different receive antennas for 2 transmit antenna MIMO MC-CDMA Uplink with N=32, L=4

3.6 Complexity Analysis

The numerical complexity is presented below for the two stages involved namely the Channel estimation and Detection. Some of the notations used are

- N_r - No. of receive antennas
- N_t - No. of transmit antennas i.e number of users
- N - No. of subcarriers

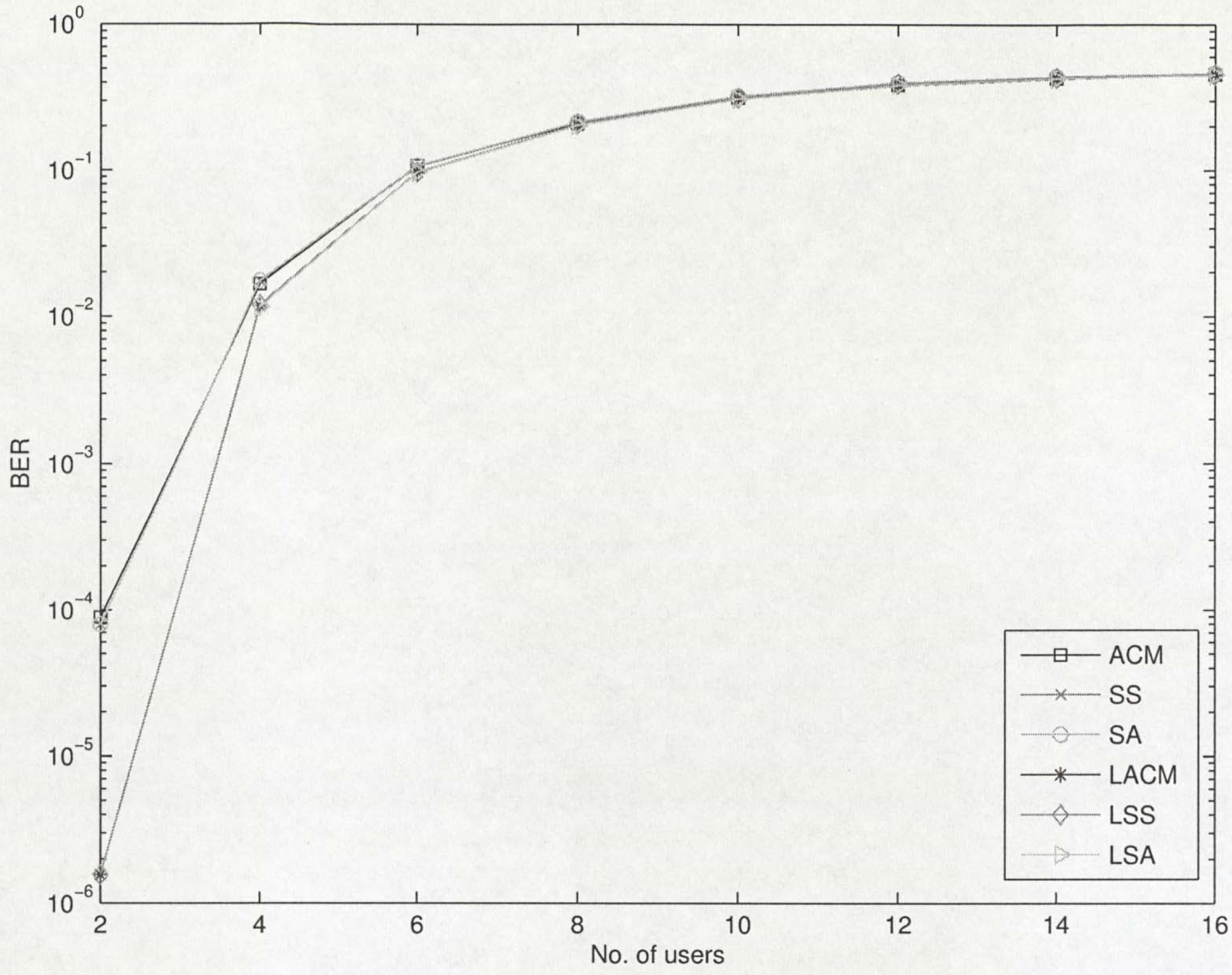

Figure 3.8: Performance of LSFE & MMSE for MIMO MC-CDMA uplink system with N=32,L=4 and varying no. of users(Nu)

- L - Channel order
- N_s - No. of symbols
- h_i - Channel Impulse response of user i

3.6.1 Subspace Algorithm

The subspace channel estimation method involves

- Correlation at each receive antenna

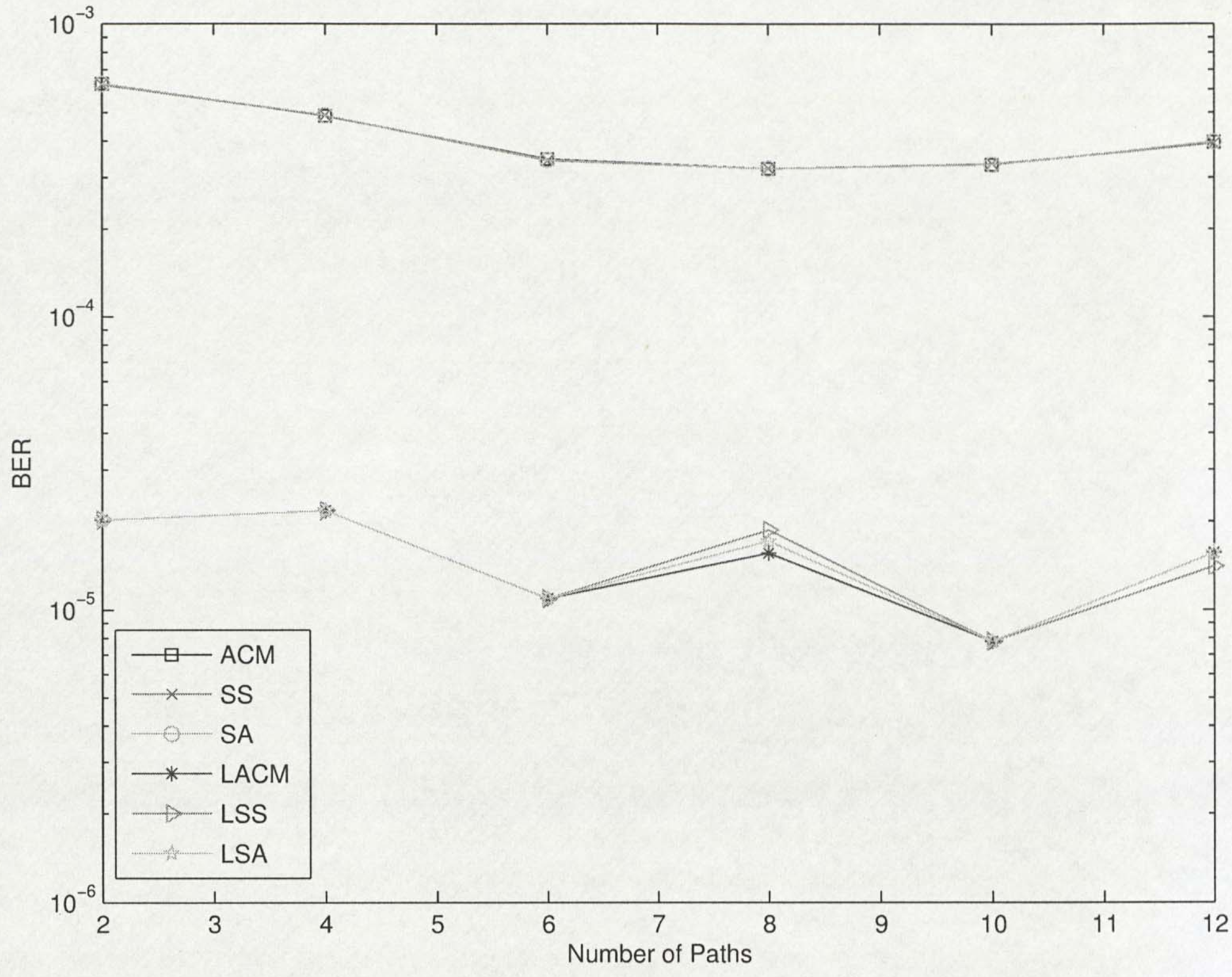

Figure 3.9: Performance of LSFE & MMSE for MIMO MC-CDMA uplink system with N=32,L=4 and varying channel order

The complexity of which is $N^2 N_s$, where N_s is the number of symbols and N is the number of carriers. Thus the net computations required will be

$$Cost_R = N_r N^2 N_s \tag{3.32}$$

- The Least squares approach to solving Eqn. (3.8) using Eqn. (3.9) leads to

$$Cost_{ls,i} = N^6 L^3 \tag{3.33}$$

Thus the total cost for all user is given as -

$$Cost_{ls} = N_t N^6 L^3 \tag{3.34}$$

Thus the total cost could be written as

$$Cost_{sub} = Cost_R + Cost_{ls} \tag{3.35}$$

which is simplified as

$$Cost_{sub} = N_r N^2 N_s + N_t N^6 L^3 \tag{3.36}$$

3.6.2 SA Algorithm

The SA blind channel estimation method involves

- Correlation at each receive antenna

  The complexity of which is $N^2 N_s$, where N_s is the number of symbols and N is the number of carriers. Thus the net computations required will be

$$Cost_R = N_r N^2 N_s \tag{3.37}$$

- The Least Squares (LS) approach to solving the equation Eqn. (3.16), where m is the order or positive integer power to which the correlation

matrix is inverted as described in section 3.3.2. For this analysis, m was chosen to be 2. Thus the complexity cost for LS method of a single user can be written as

$$Cost_{ls,i} = N^6 + N^7 L^2 \tag{3.38}$$

Thus the total cost for all users is given as -

$$Cost_{ls} = N_t N^6 + N_t N^7 L^2 \tag{3.39}$$

Thus the total cost could be written as

$$Cost_{sa} = Cost_R + Cost_{ls} \tag{3.40}$$

which is simplified as

$$Cost_{sa} = N_r N^2 N_s + N_t N^6 (1 + N L^2) \tag{3.41}$$

3.6.3 ACM algorithm

Similarly for ACM algorithm, the following steps are involved

- Auto-correlation for each receive antenna

  The same stage as in Subspace which yields a complexity cost of

$$Cost_R = N_r N^2 N_s \tag{3.42}$$

- Minimisation of the criterion for user 1 using Eqn. (3.22) which gives a complexity cost as

$$Cost_{ls,1} = N^6 + N^7L^2 + N^3 + N^7L^2 \tag{3.43}$$

Thus cost for all users will be

$$Cost_{ls} = N_tN^3(N^3 + 2N^4L^2 + 1) \tag{3.44}$$

Thus the total cost can be written as

$$Cost_{acm} = N_r + N_sN^2 + N_tN^3(N^3 + 2N^4L^2 + 1) \tag{3.45}$$

3.6.4 Detection

The complexity cose involved for detection is explained below.

MMSE Detector

The MMSE detector is expressed as in Eqn. (3.24) yields the following complexity

$$Cost_{wmmse,i} = N_r^2N^3 + N_r^3N^3 + N_r^2N_tN^3 \tag{3.46}$$

$$Cost_{wmmse} = Nt \times Cost_{wmmse,i} \tag{3.47}$$

yielding

$$Cost_{wmmse} = N_tN_r^2N^3(N_t + 1 + N_rN) \tag{3.48}$$

LSFE

The stages involved in LSFE are

- Ordering of users/streams based on mse. The main equation here is

$$Cost_{lsfe1} = N_r^3 + N_r^4 \tag{3.49}$$

- Detection of ordered stream using MMSE for each user /stream This would be the same value as above but multipleid by N_t times for each user

$$Cost_{lsfe2} = N_t^2 N_r^2 N^3 (N_t + 1 + N_r N) \tag{3.50}$$

- Cancellation of detected stream for all $N_t - 1$ stages

$$Cost_{lsfe3} = (N_t - 1) N_r N N_s \tag{3.51}$$

Therefore the total cost of LSFE can be written as

$$Cost_{lsfe} = Cost_{lsfe1} + Cost_{lsfe2} + Cost_{lsfe3} \tag{3.52}$$

Table 3.1: Channel Estimation Complexity for MIMO MC-CDMA Uplink

Cost	Subspace	ACM	SA
$Cost_R$	$N_r N^2 N_s$	$N_r N^2 N_s$	$N_r N^2 N_s$
$Cost_{ls}$	$N_t N^6 L^3$	$N_r + N_s N^2 + N_t N^3(N^3 + 2N^4 L^2 + 1)$	$N_t N^6 + N_t N^7 L^2$
$Total_{cost}$	$N_r N^2 N_s + N_t N^6 L^3$	$N_r + N_s N^2 + N_t N^3(N^3 + 2N^4 L^2 + 1)$	$N_r N^2 N_s + N_t N^6(1 + NL^2)$

Table 3.2: Detection Complexity for MIMO MC-CDMA Uplink

MMSE	LSFE
$N_t N_r^2 N^3 (N_t + 1 + N_r N)$	$N_r(N_r^2 + N_r^3 + N_t^2$ $N_r N^3 (N_t + N_r N + 1) + (N_t - 1) N N_s$

3.7 Summary

A review of the blind channel estimation methods available to MC-CDMA systems has been done followed by a detailed explanation of the working of three blind channel estimation algorithms namely the SS, the SA and the ACM methods.

It was shown that both ACM and SA provide similar performance to the SS method with the advantage that no noise-power estimation or rank estimation are needed. Integration of blind channel estimation with LSFE, yielding significant performance improvement due to the successive interference cancellation, over the MMSE scheme has been examined and illustrated. By increasing the number of receive antennas, LSFE utilising blind channel estimates, still achieves significant receive diversity.

A brief complexity analysis was also performed comparing the performances of SS, SA and ACM blind channel estimation methods when used with MMSE and LSFE equalisers. It was found that while the SA and ACM methods provide near comparable performance to the SS method, the SA method does so at lower complexity than the ACM if the order m for the SA method is chosen to be 2. The effect of order and the number of samples on SA was studied in [71], and usually the order is chosen to be 2 or 3. In the current analysis, m was set to 2.

Chapter 4

MIMO SC-CDMA with Block Transmission

4.1 Introduction

MC-CDMA [31], as described in Chapter 3, combines the advantages of CDMA and OFDM [19]. Linnartz et al [28] and Jourdan et al [30] investigated the effects of spreading coupled with MC schemes and advantages availed as a result of this combination. However MC-CDMA schemes have a few drawbacks, namely the PAPR [14] and the high sensitivity to carrier-frequency offset [32]. PAPR is affected by the choice of spreading codes as well as the modulation scheme used for CDMA systems [14].

This has resulted in a renewed interest in Single Carrier (SC) CDMA block transmissions with FDE [34], the key feature of which is that it utilises the block structure along with GB insertion in the form of CP or ZP as in MC-CDMA. Use of linear equalisers for MIMO systems is further complicated by the need for large number of taps [72]-[73]. To alleviate the high complexity needs, FDE based techniques have been proposed [73]-[74] for MIMO-CDMA schemes, which not only provide better performance but also lower complexity. The absence of large amplitude gains and the PAPR [14] as well as synchronisation problems [32] of MC-CDMA schemes, makes SC-CDMA with FDE, [34] an attractive alternative to MC-CDMA.

MIMO systems [39],[40],[75] have been shown to be effective in enhancing capacity without any additional power or bandwidth requirements. It requires a rich scattering environment as is the case in most wireless transmission channels. The combination of MIMO systems with SC-CDMA systems provides increased thoroughput and performance and the MIMO uplink scenario is investigated here.

Layered Space Frequency Equalisation (LSFE) [45], which provides performance improvement over conventional linear algorithms utilises FDE along with SIC for detecting the transmitted signals in stages. A low-complexity LSFE scheme was also proposed for downlink MIMO DS-CDMA systems [76], utilising an overlap-cut method to approximate guard-band insertion. However LSFE has not been used for the uplink SC-CDMA with block transmission, especially in a MIMO setup.

Conventional channel estimation schemes are based on training symbols [34]. This however leads to significant bandwith loss. Therefore blind channel estimation schemes are desirable. BSS [47] and other HOS [48] approaches are difficult to apply in CDMA based systems, due to the guassian nature of the signal. This is as a result of the addition of the users signals on all carriers as mentioned in Chapter 3. The SOS based methods [65] - [66] however are more suitable and the subspace approach is a key SOS based method that has been used for DS-CDMA based systems. Blind channel estimation techniques for DS-CDMA schemes based on the subspace idea [65]-[66] have been proposed in the downlink channel case [71], [77] as well as in the [67], which dealt with subspace approximation method. Liu [71] proposed a similar approach for a synchronous DS-CDMA setup, while Doukopoulos [77] utilised the same algorithm in an asynchronous DS-CDMA downlink system. However the subspace based method has not been used in MIMO SC-CDMA block transmission system with GB in the uplink case so far. Another SOS based approach is the ACM based blind channel estimation method [63]. It avoids the need for an initial EVD stage for determination of noise-subspace and also is insensitive to rank of the noise-subspace. It was used for a SIMO DS-CDMA system by Chen [63]. The key advantages of this approach are that it is more robust at lower SNR as a result of the use of the whole correlation matrix instead of the noise-subspace as in subspace based approach. However the ACM method has not been used in MIMO SC-CDMA uplink setup so far.

In this chapter, three blind channel estimation methods namely the SS method, the SA method and the ACM method are applied to the MIMO SC-CDMA uplink systems with block transmission. In addition, LSFE scheme was integrated along with blind channel estimation to yield performance improvements compared to the conventional MMSE scheme. The ACM method [63] is applied here to the uplink MIMO SC-CDMA system with block transmission, which eliminates the need for determination of rank criterion as in subspace based approach.

Simulation results show that all three methods provide close performance to the case with perfect channel knowledge at high SNR. There is the added advantage of eliminating the need for rank estimation when using the ACM method. The effect of receive diversity on performance attained are highlighted. Also plots are shown that illustrate the insensitive nature of SC-CDMA with FDE to various channel lengths. The effect of varying loads, i.e. different numbers of users on the system performance is also highlighted. A simple comparison plot is also shown that highlights the difference in performance of MIMO MC-CDMA and MIMO SC-CDMA with block transmission using FDE. The results obtained conform with the fact that performance of CDMA based systems depends on the spreading code used [78]. This chapter concludes with a brief complexity analysis of the three blind channel estimation methods and the equalisers utilised.

4.2 System Model

The system model for a MIMO SC-CDMA uplink system is shown in Figure 4.1. Here the system employs just a block transmission scheme with a GB insertion stage, which is the CP system. The SC-CDMA with CP scenario [34] is presented here while a brief description of the ZP system is provided later. The use of multiple antennas at the receiver (base station) is feasible and practical as against the mobile user, where power and complexity issues become paramount. At the base-station (receiver) end, receive diversity can be utilised by introducing multiple antennas.

Some assumptions made in the model described are as follows:-

1. N_t and N_r are the number of transmit and receive antennas respectively and N_t is equal to number of users. Also a synchronous system model is assumed.

2. L is the channel order which is assumed to be known at receiver. In practice the channel order is estimated along with the noise variance depending on the type of blind channel estimation algorithm used.

3. The signals are assumed to be white with zero-mean and unit variance

4. The spreading codes used are Walsh-Hadamard codes of size N where N is the number of subcarriers. Here the spreading gain is chosen to be equal to the number of subcarriers

5. BPSK modulation is used for each user.

Each user's signal is first spread using the corresponding spreading code. The spread signals after blocking are then transferred to GB insertion stage, where CP is appended to provide immunity against ISI.

As illustrated in Figure 4.1, the users spread signal is grouped into blocks of size N. These size N blocks are then passed through the CP insertion stage. The CP inserted signal of size P, where $P = N + L_{cp}$, is then transmitted through the multi-path fading MIMO channel under the assumption that $L_{cp} \geq L$ i.e. the CP is atleast as long as the channel order represented by L [79].

At the receiver end as illustrated in Figure 4.2, the received signal has its guard band removed before being passed onto FFT stage. After FFT, the FDE stage equalises the signal. The equalised signal is converted back to

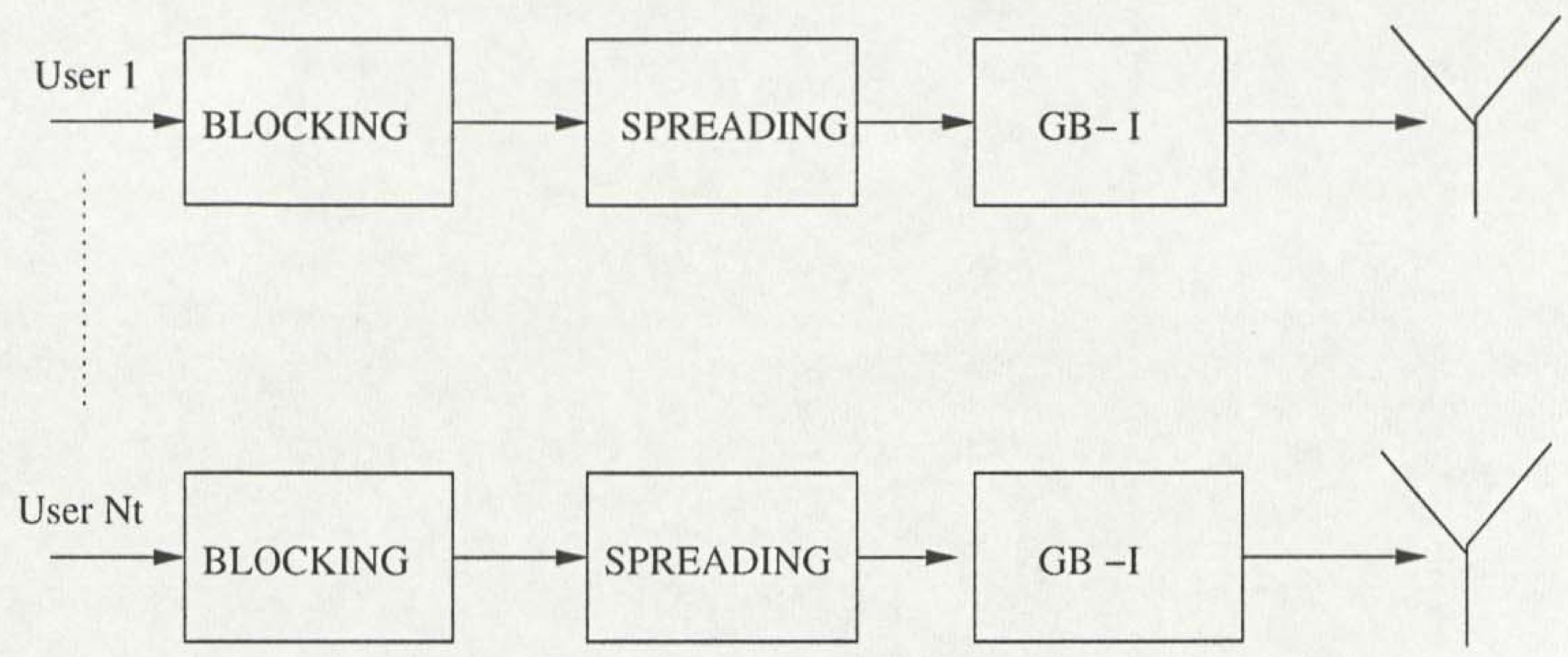

Figure 4.1: SC-CDMA Uplink Transmitter Madhukumar et al [34]

time domain via application of IFFT before hard decision device is used to obtain the estimates of the transmitted symbols.

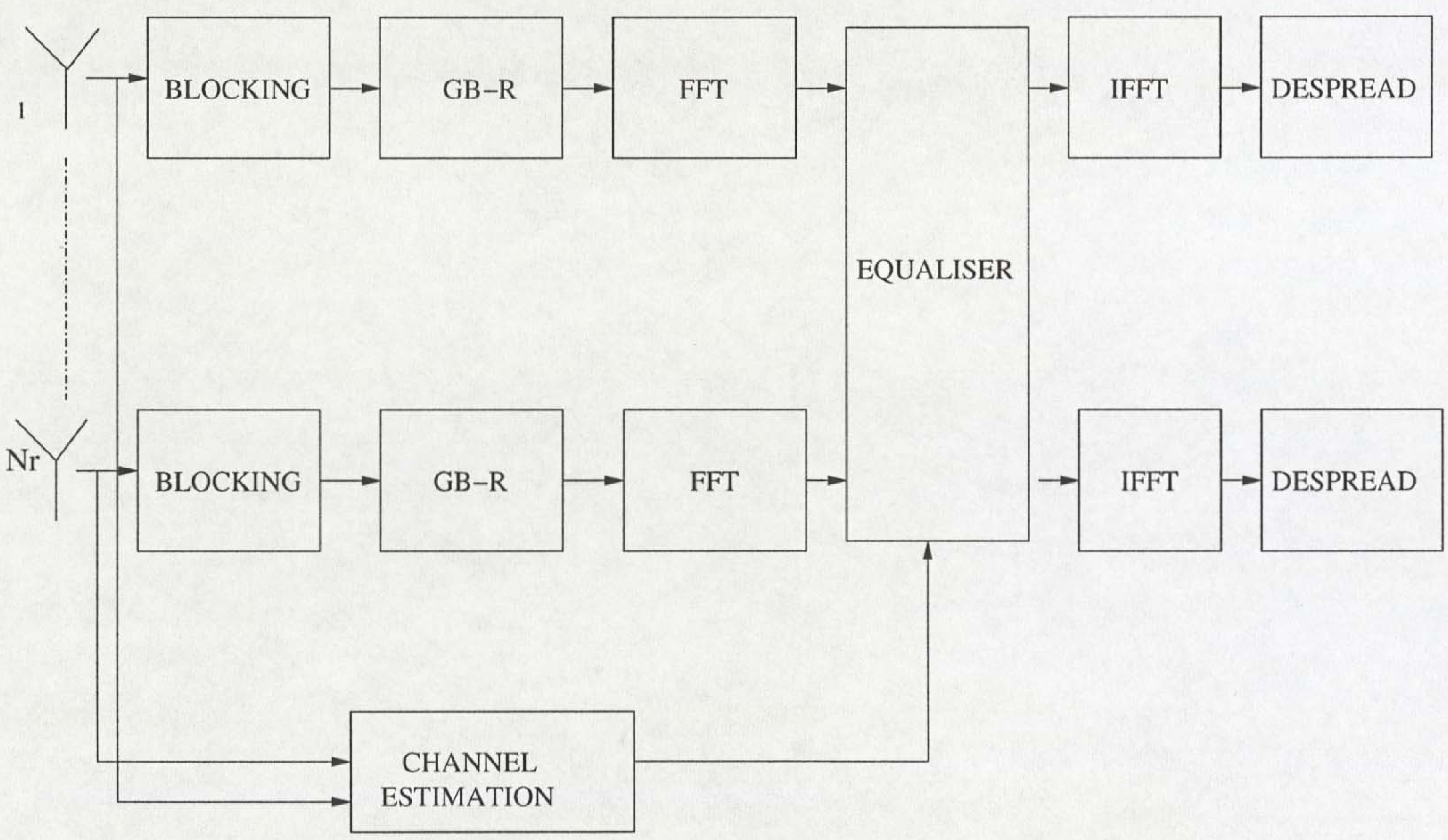

Figure 4.2: SC-CDMA Uplink Receiver Madhukumar et al [34]

After passing through the GB removal and FFT stages [34], the frequency domain model of the received signal at frequency bin n can be written as

$$\mathbf{X}_j[n] = \sum_{i=1}^{N_t} \mathbf{H}_{j,i}[n] c_i[n] s_i[n] + \mathbf{V}_j[n] \tag{4.1}$$

where $\mathbf{H}_{j,i}[n]$ is the fourier transform coefficient of the channel matrix between transmit antenna i and receive antenna j on sub-carrier n. $c_i[n]$ represents the n^{th} value of the spreading code for user i, while $s_i[n]$ is the copy of the data bit transmitted on subcarrier n and $\mathbf{V}_j[n]$ denotes the noise at the j^{th} receive antenna. The above equation over all subcarriers can be expressed as shown below with $\mathbf{H}_{j,i}$ denoting the N point FFT of the circulant channel matrix between j^{th} receive and i^{th} transmit antenna and is given as $\mathbf{H}_{j,i} = [\mathbf{H}_{j,i}^T[1] \dots \mathbf{H}_{j,i}^T[N]]^T$.

Let $\mathbf{X}_j = [\mathbf{X}_j^T[1] \dots \mathbf{X}_j^T[N]]^T$ represent the received signal at antenna j, while $\mathbf{c}_i = [c_i[1] \dots c_i[N]]^T$ denotes the spreading code vector of user i , a column vector of size $N \times 1$ and $\mathbf{s}_i = [s_i[1] s_i[2] \dots s_i[N]]$ represent the data transmitted by user i. $\mathbf{V}_j = [\mathbf{V}_j^T[1] \dots \mathbf{V}_j^T[N]]^T$ is the noise at the j^{th} receive antenna. Thus the following equation is obtained.

$$\mathbf{X}_j = \sum_{i=1}^{N_t} \mathbf{H}_{j,i} \mathbf{c}_i \mathbf{s}_i + \mathbf{V}_j \tag{4.2}$$

Letting $\mathbf{d}_i = \mathbf{c}_i \mathbf{s}_i$ denote the spread signal of user i, Eqn. (4.2) can be written as

$$\mathbf{X}_j = \sum_{i=1}^{N_t} \mathbf{H}_{j,i} \mathbf{d}_i + \mathbf{V}_j \tag{4.3}$$

The equivalent matrix vector form of the received signal model is of the form as expressed in Eqn. (4.4) with $\mathbf{X}$ denoting the net received signal over all received antenna given by $\mathbf{X} = [\mathbf{X}_1^T \mathbf{X}_2^T \ldots \mathbf{X}_{N_r}^T]^T$.

$$\mathbf{X} = \sum_{i=1}^{N_t} \mathbf{H}_i \mathbf{d}_i + \mathbf{V} \tag{4.4}$$

where $\mathbf{H}_i$ is defined as

$$\mathbf{H}_i = [\mathbf{H}_{1,i}^T \ldots \mathbf{H}_{N_r,i}^T]^T \tag{4.5}$$

and $\mathbf{V}$ represents the noise denoted as $\mathbf{V} = [\mathbf{V}_1^T \mathbf{V}_2^T \ldots \mathbf{V}_{N_r}^T]^T$.

4.3 Blind Channel Estimation

The use of training symbols and different pilot symbol structures for uplink and downlink scenarios have been proposed for the SC-CDMA with block transmission [34]. In this section, three blind channel estimation schemes for SC-CDMA with FDE, namely the SS, the SA and the ACM methods are proposed and their differences compared to the MIMO MC-CDMA setup in Chapter 3 are explained.

4.3.1 Subspace Method

The subspace based method utilizes the guard band to estimate the signature waveform, which is the convolution of the channel and user's spreading code [71], [77]. The system models used in [71],[77] and [80] were that of DS-CDMA system with no guard band structure and using the whole signal with ISI. The model used here is similar but with the significant difference of having a guard band to avoid IBI, while utilising the Toeplitz nature of the model to obtain the subspace cost function as will be explained below. The use of GB intact received signal without FFT processing for blind channel estimation is the key difference between the subspace method presented here and that applied to MIMO MC-CDMA scheme in the previous chapter.

Making use of the Toeplitz matrix property [81], the Toeplitz channel matrix and the code vectors can be exchanged to obtain the equivalent user transmitted signal denoted by $\mathbf{g}_i$ as

$$\mathbf{g}_i = \mathbf{C}_{tj,i}\mathbf{h}_i\mathbf{s}_i \tag{4.6}$$

where $\mathbf{h}_{j,i}$ denotes the channel impulse response between receive antenna j and user i. $\mathbf{C}_{tj,i}$, the toeplitz spreading code matrix for user i, of size $P \times L$ is expressed as

$$\mathbf{C}_{tj,i} = \begin{bmatrix} b_{i1} & 0 & 0\ldots & 0 \\ b_{i2} & b_{i1} & 0 & 0 \\ : & : & b_{i1} & 0 \\ b_{iN} & \ldots & \ldots & b_{i1} \\ b_{i1} & 0 & \ldots & 0 \\ \vdots & b_{i1} & \vdots & 0 \\ b_{iL} & \ldots & \ldots & b_{i1} \end{bmatrix} \tag{4.7}$$

where $\mathbf{b}_{in}(n = 1, 2 \ldots N)$ is the n^{th} bit of the time domain spreading code sequence for user i. The autocorrelation matrix of the received signal is written as

$$\mathbf{R}_j = E[\mathbf{X}_j\mathbf{X}_j^H] \tag{4.8}$$

The EVD of $\mathbf{R}_j$ yields

$$\mathbf{R}_j = [\mathbf{U}_{sj} \quad \mathbf{U}_{nj}] \begin{bmatrix} \mathbf{\Lambda}_s & 0 \\ 0 & \mathbf{\Lambda}_n \end{bmatrix} \begin{bmatrix} \mathbf{U}_{sj}^T \\ \mathbf{U}_{nj}^T \end{bmatrix} \tag{4.9}$$

where $\mathbf{U}_{sj}$ and $\mathbf{U}_{nj}$ represent the orthogonal signal and noise subspaces respectively while $\mathbf{\Lambda}_s$ and $\mathbf{\Lambda}_n$ denote the diagonal matrices containing the corresponding eigen values of signal and noise subspaces. The orthogonality between the signal and noise subspaces results in Eqn. (4.6) being modified to yield the following equation [80]

$$\mathbf{U}_{nj}^H \mathbf{C}_{tj,i} \mathbf{h}_{j,i} = 0 \tag{4.10}$$

The key equation in the subspace method is the minimisation of the following quadratic cost function

$$\mathbf{h}_{j,i} = \arg \min_{||\mathbf{h}_{j,i}||=1} ||\mathbf{h}_{j,i}^H \mathbf{Q}_{j,i} \mathbf{h}_{j,i}||^2 \tag{4.11}$$

where the matrix $\mathbf{Q}_{j,i}$ for user i is given by

$$\mathbf{Q}_{j,i} = \mathbf{C}_{tj,i}^H \mathbf{U}_{nj} \mathbf{U}_{nj}^H \mathbf{C}_{tj,i} \tag{4.12}$$

The solution is obtained by taking the eigenvector corresponding to the lowest eigenvalue.

4.3.2 Subspace Approximation

The initial step of Subspace method requires the determination of the rank of the noise-subspace after an SVD or EVD. Rank detection methods however are dependent on SNR and sample size [67]. Blind channel estimation using Subspace Approximation method [67] was proposed for downlink DS-CDMA systems. It utilises the Power of R method as mentioned in Chapter 3, which involves replacing the noise subspace component in cost function with inverse of autocorrelation matrix raised to a positive integer. As value of positive integer increases, the approximation error is reduced. It is also easily applied to adaptive tracking and estimation cases [67],[69]. The extension to MIMO SC-CDMA case is presented below. The key point of difference from the MC-CDMA scheme in Chapter 3 is that the received signal with GB intact is utilised for blind channel estimation unlike the GB stripped and FFT processed signal used in previous chapter.

The EVD of the autocorrelation matrix of the received signal at the j^{th} receive antenna can be expressed as

$$\mathbf{R_j} = \begin{bmatrix}\mathbf{U}_{sj} & \mathbf{U}_{nj}\end{bmatrix} \begin{bmatrix} \lambda_{sj} & 0 \\ 0 & \lambda_{nj} \end{bmatrix} \begin{bmatrix} \mathbf{U}_{sj}^T \\ \mathbf{U}_{nj}^T \end{bmatrix} \tag{4.13}$$

Rearraning the above equation yields

$$\sigma^2 \mathbf{R_j}^{-1} = \mathbf{U}_{nj}\mathbf{U}_{nj}^H + \mathbf{U}_{sj}\mathbf{diag}\left(\frac{\sigma^2}{\lambda_{sj}^2 + \sigma^2}\right)\mathbf{U}_{sj}^H \tag{4.14}$$

As the power of R is raised to a positive integer i.e. $m = 1, 2, 3..$, the above equation becomes

$$\sigma^{2m}\mathbf{R_j}^{-m} = \mathbf{U}_{nj}\mathbf{U}_{nj}^H + \mathbf{U}_{sj}\mathbf{diag}\left(\frac{\sigma^2}{\lambda_{sj}^2 + \sigma^2}\right)^m \mathbf{U}_{sj}^H \tag{4.15}$$

Using the property that

$$\lim_{m\to\infty} \sigma^{2m}\mathbf{R_j}^{-m} = \mathbf{U}_{nj}\mathbf{U}_{nj}^H \tag{4.16}$$

The subspace approximation algorithm [67] is thus obtained, where the replacement by the power of inverse of the auto-correlation matrix eliminates need for subspace rank determination as well as initial EVD stage of subspace method. It also alleviates the effects of noise and sample size. The scalar σ^{2m} is negligible as regards the channel estimation stage. Thus the final cost function equation for SA approach for user i at the j^{th} receive antenna is given as

$$\mathbf{h}_{j,i} = \arg \min_{||\mathbf{h_{j,i}}||=1} ||\mathbf{h}_{ij}^H \mathbf{Q}_{ij} \mathbf{h}_{j,i}||^2 \tag{4.17}$$

where the matrix $\mathbf{Q}_{j,i}$ for user i at the j^{th} receive antenna is given by

$$\mathbf{Q}_{j,i} = \mathbf{C}_{ti}^H \mathbf{U}_{nj}\mathbf{U}_{nj}^H \mathbf{C}_{ti} \tag{4.18}$$

where C_{ti} is of size $P \times L$ expressed as

$$\mathbf{C}_{ti} = \begin{bmatrix} c_{i1} & 0 & 0\ldots & 0 \\ c_{i2} & c_{i1} & 0 & 0 \\ : & : & c_{i1} & 0 \\ c_{iN} & \ldots & \ldots & c_{i1} \\ c_{i1} & 0 & \ldots & 0 \\ \vdots & c_{i1} & \vdots & 0 \\ c_{iL} & \ldots & \ldots & c_{i1} \end{bmatrix} \quad (4.19)$$

$\mathbf{h}_{j,i}$ is the channel impulse response of user i at receive antenna j. The solution is obtained by taking the eigenvector corresponding to the lowest eigenvalue. Thus the cost function to be minimised is obtained as

$$\mathbf{Q}_{j,i} = \mathbf{C}_i^H \mathbf{R_j}^{-m} \mathbf{C}_i \quad (4.20)$$

which finally yields

$$\hat{\mathbf{h}}_{j,i} = \arg \min_{||\mathbf{h}_{j,i}||=1} \mathbf{h}_{j,i}^H \mathbf{C}_i^H \mathbf{R_j}^{-m} \mathbf{C}_i \mathbf{h}_{j,i} \quad (4.21)$$

where the value of m is usually chosen to be 2 or 3 as effects of right hand term in Eqn. (4.14) are already small with these values, thereby avoiding further computational complexity.

4.3.3 ACM method

The ACM method employs the whole received signal with the guard-band intact and utilises the structure afforded by the guard-band to blindly estimate the channel.

As in the MC-CDMA case, the equivalence of the SINR and MMSE receivers is used as the starting point for the derivation of the algorithm for the SC-CDMA case. There are significant differences between the application of ACM in MC-CDMA and SC-CDMA, the first and most important of which is that in SC-CDMA, the blind channel estimation is done before the guard removal and application of FFT in contrast to MC-CDMA, where the received signal is first stripped of its guard band and then transferred to frequency domain via FFT.

Thus the equations to be used will be different than in MC-CDMA case and these are derived as follows. The general MMSE weight vectors for the guard-band inserted received signals for user i at receive antenna j are given as

$$\mathbf{w}_{mmsej,i} = \mathbf{R}_j^{-1}\mathbf{C}_{tj,i}\mathbf{h}_{j,i} \tag{4.22}$$

where $\mathbf{R}_j$ is the autocorrelation matrix at receive antenna j and $\mathbf{h}_{j,i}$ is the channel impulse response from transmit antenna (i.e. user) i to receive antenna j with $\mathbf{C}_{tj,i}$ as defined in Eqn. (4.7). $\mathbf{R}_j$ can also be defined as

$$\mathbf{R}_j = \mathbf{h}_{j,i}\mathbf{h}_{j,i} + \sum_{\substack{k=1 \\ k\neq i}}^{N_t} \mathbf{h}_{j,k}\mathbf{h}_{j,k} + \sigma^2\mathbf{I} \tag{4.23}$$

where σ^2 is the noise power. The maximum signal to interference plus noise ratio (MSINR) weight vector for user i can be written as

$$\mathbf{w}_{msinrj,i} = \arg \max_{||\mathbf{v}||=1} \frac{\mathbf{v}^H\mathbf{C}_{tj,i}\mathbf{h}_{j,i}\mathbf{h}_{j,i}^H\mathbf{C}_{tj,i}\mathbf{v}}{\mathbf{v}^H\mathbf{R}_j\mathbf{v}} \tag{4.24}$$

Utilising the equivalence of the Eqns. (4.22) and (4.24), the following equation is obtained.

$$\hat{\mathbf{h}}_{j,i} = \arg \max_{||\mathbf{v}||=1} \frac{\mathbf{v}^H \mathbf{C}_{tj,i}^{\mathrm{H}} \mathbf{R}_j^{-1} \mathbf{R}_{j,b_i} \mathbf{R}_j^{-1} \mathbf{C}_{tj,1} \mathbf{v}}{\mathbf{v}^H \mathbf{C}_{tj,i}^{H} \mathbf{R}_j^{-1} \mathbf{C}_{tj,i} \mathbf{v}} \tag{4.25}$$

where $\mathbf{R}_{j,\mathrm{b}_i} = \mathbf{h}_{j,i}\mathbf{h}_{j,i}^{\mathrm{H}}$ is the subspace matrix of desired user symbols b_i while $\hat{\mathbf{h}}_{j,i}$ denotes the estimate of channel of user i. The matrix $\mathbf{R}_{j,b_i}$ contains the subspace of the entire signal including that of the interfering users. The matrices $\mathbf{R}_{j,b}$ and $\mathbf{R}_j$ are related as follows

$$\mathbf{R}_j = \mathbf{R}_{j,b_i} + \sigma^2 \mathbf{I} \tag{4.26}$$

The ACM algorithm for the SC-CDMA case can be expressed as

$$\hat{\mathbf{h}}_{j,i} = \arg \max_{||\mathbf{v}||=1} \frac{\mathbf{v}^{\mathrm{H}} \mathbf{C}_{tj,i}^{\mathrm{H}} \mathbf{R}_j^{-1} (\mathbf{R}_j - \sigma^2 \mathbf{I}) \mathbf{R}_j^{-1} \mathbf{C}_{tj,i} \mathbf{v}}{\mathbf{v}^{\mathrm{H}} \mathbf{C}_{tj,i}^{\mathrm{H}} \mathbf{R}_j^{-1} \mathbf{C}_{tj,i} \mathbf{v}} \tag{4.27}$$

Using the equation above, the channel impulse response between j^{th} receive antenna and user i can be estimated as

$$\hat{\mathbf{h}}_{j,i} = \arg \min_{||\mathbf{v}||=1} \frac{\mathbf{v}^{\mathrm{H}} \mathbf{C}_{tj,i}^{\mathrm{H}} \mathbf{R}_j^{-2} \mathbf{C}_{tj,i} \mathbf{v}}{\mathbf{v}^{\mathrm{H}} \mathbf{C}_{tj,i}^{\mathrm{H}} \mathbf{R}_j^{-1} \mathbf{C}_{tj,i} \mathbf{v}} \tag{4.28}$$

4.3.4 ZP-SC-CDMA

The above proposed blind channel estimation schemes can also be applied to MIMO SC-CDMA uplink schemes with ZP instead of CP guard band. Replacing the the last L rows of Toeplitz code matrix $\mathbf{C}_t$ with zeros, the same cost functions can then be used to obtain blind channel estimates for