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COMMUNICATIONS AND ARRAY PROCESSING GROUP

Cyclic Prefix-Free MC-CDMA Arrayed MIMO Communication Systems

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Abstract

The objective of this thesis is to investigate MC-CDMA MIMO systems where the antenna array geometry is taken into consideration. In most MC-CDMA systems, cyclic prefixes, which reduce the spectral efficiency, are used. In order to improve the spectral efficiency, this research study is focused on cyclic prefix-free MC-CDMA MIMO architectures.

Initially, space-time wireless channel models are developed by considering the spatio-temporal mechanisms of the radio channel, such as multipath propagation. The spatio-temporal channel models are based on the concept of the array manifold vector, which enables the parametric modelling of the channel.

The array manifold vector is extended to the multi-carrier space-time array (MC-STAR) manifold matrix which enables the use of spatio-temporal signal processing techniques. Based on the modelling, a new cyclic prefix-free MC-CDMA arrayed MIMO communication system is proposed and its performance is compared with a representative existing system. Furthermore, a MUSIC-type algorithm is then developed for the estimation of the channel parameters of the received signal.

This proposed cyclic prefix-free MC-CDMA arrayed MIMO system is then extended to consider the effects of spatial diffusion in the wireless channel. Spatial diffusion is an important channel impairment which is often ignored and the failure to consider such effects leads to less than satisfactory performance. A subspace-based approach is proposed for the estimation of the channel parameters and spatial spread and reception of the desired signal.

Finally, the problem of joint optimization of the transmit and receive beamforming weights in the downlink of a cyclic prefix-free MC-CDMA arrayed MIMO communication system is investigated. A subcarrier-cooperative approach is used for the transmit beamforming so that there is greater flexibility in the allocation of channel symbols. The resulting optimization problem, with a per-antenna transmit power constraint, is solved by the Lagrange multiplier method and an iterative algorithm is proposed.

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List of Publications

- H. H. Peh, A. Manikas, T. T. Tjhung and W.-C. Wong, "An Investigative Study of Cyclic Prefix-free MC-CDMA Arrayed MIMO Communication Systems," *Proceedings of the International Symposium on Wireless Pervasive Computing 2007*, Feb. 2007.
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Abbreviations and Acronyms

3G	third generation
4G	fourth generation
AWGN	additive white Gaussian noise
BPSK	binary phase shift keying
BER	bit error rate
CCI	co-channel interference
CDMA	code division multiple access
(I)DFT	(inverse) discrete Fourier transform
DBPSK	differential binary phase-shift keying
DQPSK	differential quaternary phase-shift keying
DS-CDMA	direct sequence code division multiple access
(I)FFT	(inverse) fast Fourier transform
FIR	finite impulse response
ICI	intercarrier interference
ISI	intersymbol interference
MAI	multiple-access interference
MC-CDMA	multi-carrier code division multiple access
MC-DS-CDMA	multicarrier direct sequence code division multiple access
MIMO	multiple-input multiple-output
MISO	multiple-input single-output

MUSIC	multiple signal classification
MMSE	minimum mean squared error
OFCDM	orthogonal frequency code division multiplexing
OFDM	orthogonal frequency division multiplexing
PAPR	peak-to-average power ratio
PN	pseudo-noise
QoS	quality-of-service
QPSK	quaternary phase-shift keying
SIC	successive interference cancellation
SIMO	single-input multiple-output
SISO	single-input single-output
SIVO	scalar-input vector-output
SNR	signal-to-noise-ratio
SNIR	signal-to-noise-plus-interference-ratio
TDD	time-division duplex
UCA	uniform circular array
ULA	uniform linear array
V-BLAST	vertical-Bell Labs layered space-time
VIVO	vector-input vector-output
ZF	zero-forcing

List of Symbols

$\underline{\alpha}, \underline{\gamma}$	PN vector
α_k	k^{th} chip of the PN vector $\underline{\alpha}$
β	complex path fading coefficient
$\underline{\beta}$	column vector of complex path fading coefficients
$\mathbf{a}[n]$	channel symbol for the n^{th} symbol interval
$\underline{\mathbf{a}}[n]$	channel symbol vector for the n^{th} symbol interval
c	speed of light
$\mathbb{C}_i$	code matrix for the i^{th} transmitting element
$\mathbb{E}_n$	eigenvector basis for the noise subspace
$\mathbb{E}_s$	eigenvector basis for the signal subspace
F_c	RF carrier frequency
F_k	baseband frequency of the k^{th} subcarrier
$\underline{\mathbf{h}}$	MC-STAR manifold vector
$\underline{\mathbf{h}}$	diffused MC-STAR manifold vector
$\mathbb{H}$	MC-STAR manifold matrix
L	number of snapshots
M	number of users
N	number of antenna elements in the receiver antenna array
N_{sc}	number of subcarriers used
N_i^{sub}	number of channel symbol substreams for the i^{th} user

$\underline{n}(t)$	analogue AWGN vector
$\underline{n}[n]$	sampled AWGN for the n^{th} symbol interval
$\underline{r}_i$	3×1 vector of the Cartesian coordinates of the i^{th} element of an antenna array
$\mathbb{R}_{xx}$	covariance matrix of the data vector-signal $\underline{x}(t)$ (data covariance matrix)
$\underline{S}$	antenna array manifold vector
T_{cs}	channel symbol period
T_s	sampling period
μ	Lagrange multiplier
$\underline{w}$	weight vector
$\underline{x}(t)$	continuous time received vector-signal at baseband
$\underline{x}[n]$	discrete time received vector-signal at baseband
ϕ	elevation angle
τ	propagation delay
θ	azimuth angle
$\tilde{\theta}$	perturbation angle

Note: Parameters such as θ, ϕ and N are used for both the transmitter and receiver. In order to differentiate a transmitter parameter from a receiver parameter, $\overline{(\cdot)}$ will denote a transmitter parameter.

Notation

$\mathcal{Z}$	Set of integers
$\mathcal{R}$	Field of real numbers
$\mathcal{C}$	Field of complex numbers
$\mathcal{E}\{\cdot\}$	Expectation operator
a, A	Scalar
$\underline{a}, \underline{A}$	Column Vector
$\mathbb{A}, \mathbf{A}$	Matrix
$ A $	Magnitude of A
$\ \underline{A}\ $	Euclidian norm of the vector $\underline{A}$
$\ \mathbb{A}\ $	Frobenius norm of the matrix $\mathbb{A}$
$\underline{A}^b$	Element by element power
$\exp(\underline{A})$	Element by element exponential of vector $\underline{A}$
$[\underline{A}]_i$	i^{th} element of the vector $\underline{A}$
$(\cdot)^*$	Complex conjugate
$(\cdot)^T$	Transpose
$(\cdot)^H$	Hermitian transpose
$(\cdot)^\dagger$	pseudoinverse
$\otimes$	Kronecker product
$\odot$	Hadamard product
$\oslash$	Hadamard division

$\mathbb{I}_N$	Identity matrix of $N \times N$ dimension
$\mathbb{O}_{M \times N}$	Zero matrix of $M \times N$ dimension
$\underline{1}_N$	$N \times 1$ vector of all ones
$\underline{0}_N$	$N \times 1$ vector of all zeros
$\underline{1}_{N,k}$	$N \times 1$ vector with 1 at the k^{th} element and 0s elsewhere
$\mathbb{J}_N$	$N \times N$ downshifting matrix
$\text{col}_i(\mathbb{A})$	selection of the i^{th} column of $\mathbb{A}$
$\text{diag}(\underline{A})$	Diagonal matrix with the vector $\underline{A}$ as the leading diagonal
$\underline{\text{diag}}(\mathbb{A})$	Column vector with elements the diagonal elements of the matrix $(\mathbb{A})$
$\text{tr}(\mathbb{A})$	Trace of the matrix $\mathbb{A}$
$\mathbb{P}[\mathbb{A}]$	Projection operator onto the range space of $\mathbb{A}$
$\mathbb{P}^\perp[\mathbb{A}]$	Complementary projection operator of the matrix $\mathbb{A}$

Chapter 1

Introduction

With the introduction of 3G communications systems, there has been a dramatic increase in the variety of applications that are available to mobile users and these applications are often multi-media intensive, such as video phone calls. The prevalence of such data rate-hungry applications thus puts an upward pressure on the data rates in wireless communications systems. As such, in future 4G systems, even higher data rates, expected to be 10 to 100 times the target data rate in 3G systems, are to be anticipated in order to satisfy the data rate requirements to support such multi-media intensive applications. However, channel impairments, such as frequency-selectivity, present an obstacle to achieving such high data rates. In addition, the existence of high system overheads, such as pilot or training signals required for channel identification or synchronization, results in a reduction in the amount of resources that are available for the transfer of information through the wireless medium and contributes to a lower data rate that can be achieved.

However, with such high data rates, transmissions through the wireless medium become more susceptible to frequency selective fading as the signal bandwidth becomes greater than the channel coherence bandwidth. In addition, the wireless medium often subjects the transmitted signals to effects such as multipath and

multiple access and co-channel interference, making high data rates above 100 Mbps difficult to achieve. Mobility of the users also introduces time selective fading through Doppler spread which makes the accurate demodulation of the received signals more difficult.

With such high target data rates, multi-carrier (MC) modulation techniques are promising candidates to be implemented in 4G systems. MC techniques, such as orthogonal frequency division multiplexing (OFDM), can alleviate the frequency selective fading experienced by high data rate (wideband) signals. Such methods involve de-multiplexing the original signal stream onto a number of subcarriers, each with a lower data rate, so that each subcarrier would not be subject to frequency selective fading. Multiple-input multiple-output (MIMO) systems is another emerging communication technique which is capable of enhancing the achievable user data rate. Due to the possible gains in spectral efficiency that both MC techniques and MIMO systems promise, it is anticipated that these two techniques will be combined in 4G systems so as to bring out a synergistic effect on the spectral efficiency in 4G systems.

1.1 Channel Dispersion

In general, a wireless channel is multipath dispersive and time-varying due to the combined effects of reflection, diffraction and scattering. Thus, the propagation of radio signals through the wireless channel results in an alteration of the received signal. Measurements of the power of the received signal often exhibit two types of characteristics:

- Large-scale fading: This is due to path loss and shadowing effects and is observed through the variations in the average received power level. Path loss increases logarithmically with the distance between the transmitter

and receiver, while shadowing describes the attenuation effects due to the surrounding environmental clutter.

- Small-scale fading: Rapid fluctuations of the received signal over very short time durations or over very short distances. This is due to the superposition of multipath signals at the receiver which produces a resultant signal that varies widely in amplitude and phase over small time intervals or distances.

Small-scale fading can be divided into a number of different categories based on the delay spread and Doppler spread of the channel. Delay spread is due to the existence of multipath signals with different time delays. The transmitted signals typically arrive at the receiver with independent phase, time and amplitude which results in large variations in the envelope of the received signal. The delay spread is defined as the maximum excess delay between the first and last arrived path components, during which the signal power falls to some threshold level below that of the strongest received power. Coherence bandwidth, which is the inverse of the delay spread, refers to the frequency separation at which the attenuation of two frequency-domain samples of the channel becomes decorrelated. Thus, coherence bandwidth is a measure of the frequency-selectivity of the channel. If the coherence bandwidth is larger than the bandwidth of the received signal, a flat fading channel occurs. In a flat fading channel, the amplitude of the received signal varies over time but the spectrum of the transmitted signal is preserved. However, if the signal bandwidth is larger than the coherence bandwidth (i.e. the symbol period is lesser than the delay spread), intersymbol interference (ISI) occurs and the multipath fading is frequency-selective.

Doppler spread, on the other hand, is dependent upon the coherence time of the channel. Coherence time, which is the inverse of the Doppler spread, is

defined as the time separation at which the amplitudes of two samples of the channel becomes decorrelated. Thus, coherence time is a measure of the time duration for which the channel can be assumed to be approximately constant. If the coherence time is less than a symbol period, time-selective fading occurs and this results in a fast fading channel. This corresponds to a Doppler spread in the frequency domain and a distortion of the transmitted baseband pulse shape. On the other hand, in a slow fading channel, the coherence time is greater than a symbol period so that the channel is invariant for the duration of a transmitted signal.

1.2 Multi-Carrier Modulation

In the previous section, it is seen that the time and frequency dispersions of a signal are dependent on the coherence bandwidth and coherence time of the channel, respectively, as well as the bandwidth and duration of the signal. Hence, for a particular channel, the amount of distortion observed in the signal depends on the design of the signal. In order to overcome frequency-selective channels and the associated ISI of such channels, MC modulation schemes, such as OFDM, Multi-Carrier Direct Sequence Code Division Multiple Access (MC DS-CDMA) and Multi-Carrier Code Division Multiple Access (MC-CDMA) have been proposed. MC-CDMA is a type of multiple access scheme that is a combination of Direct Sequence Code Division Multiple Access (DS-CDMA) and OFDM [1, 2]. In contrast to DS-CDMA systems, each chip of the PN sequence modulates a different subcarrier for MC-CDMA systems and it has been shown that for the forward link, MC-CDMA has better multipath suppression capabilities than DS-CDMA [3]. Moreover, MC-CDMA was shown to be capable of supporting more users than DS-CDMA in [4].

A MC modulation scheme employs a set of subcarriers to transmit the information symbols in parallel over the channel. This is equivalent to the division of the available channel bandwidth into a number of subchannels and the bandwidth of each subcarrier is sufficiently narrow so that the frequency response characteristics of the subchannels are approximately flat. Figure 1.1(a) illustrates the comparison of the frequency spectrum of the transmitted signal in a MC modulation system with that of the transmitted signal in a single-carrier (SC) modulation system and it can be seen that the spectrum of the transmitted signal of a MC modulation system is composed of a number of subcarriers which have a narrowband spectrum. Figure 1.1 also illustrates the effect of a frequency-selective fading channel on the received signal in both SC and MC modulation systems. As shown in Figure 1.1(b), the spectrum of the received signal for a MC modulation system implies that a MC modulation system only requires a simple one-step equalization for each subcarrier whereas adaptive equalization is needed for a SC modulation system.

In addition, MC modulation schemes offer an advantage over single-carrier (SC) modulation schemes in terms of narrowband frequency interference since not all frequency subbands will be affected by the interference. Orthogonal subcarriers, which allows the subcarriers' spectra to overlap without causing interference, are often employed in an MC modulation scheme so that a higher level of spectral efficiency can be achieved. This is because the guardbands that are necessary to allow individual demodulation of non-orthogonal subcarriers will no longer be necessary. As long as the orthogonality of the subcarriers is maintained, it is still possible to recover the individual subcarriers' signals despite their overlapping spectrums.

In addition, since the system's data throughput is the sum of throughput of all the parallel channels, the data rate per subchannel is only a fraction of the data

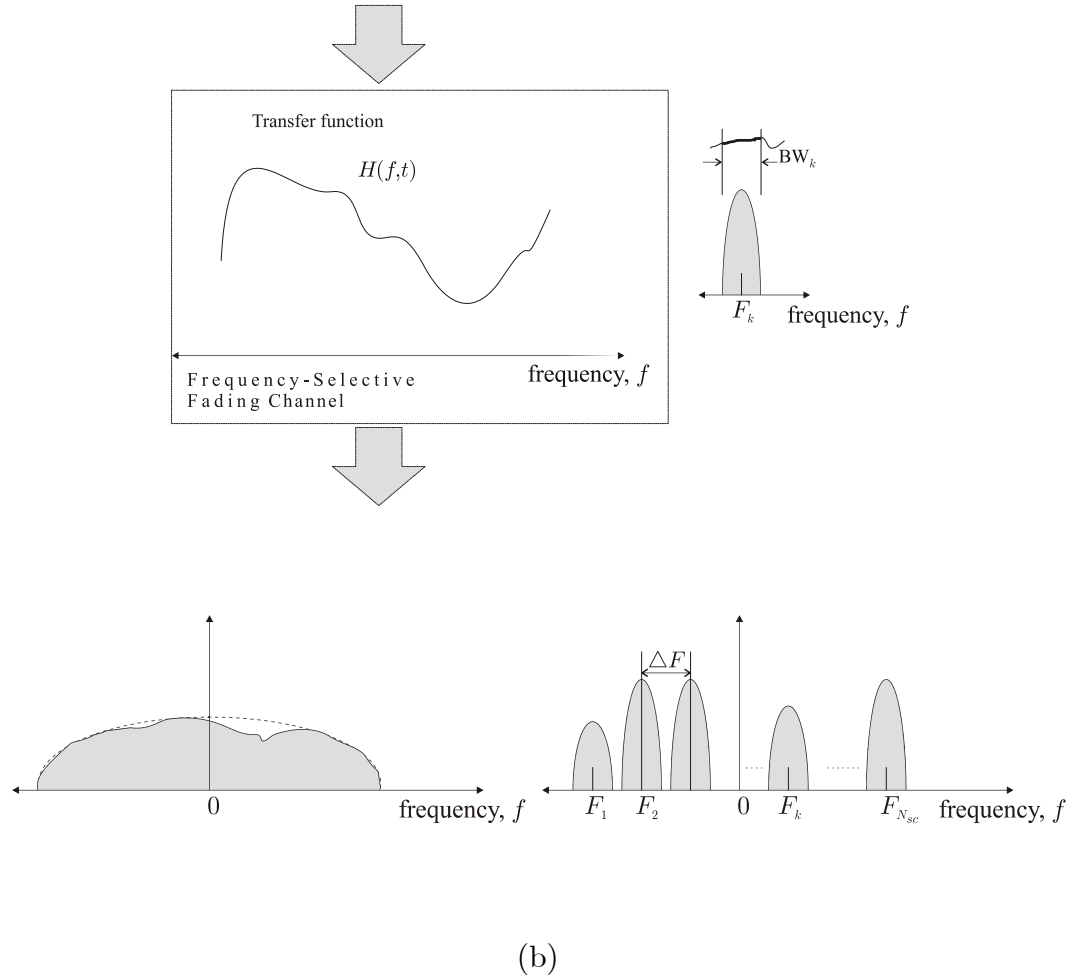
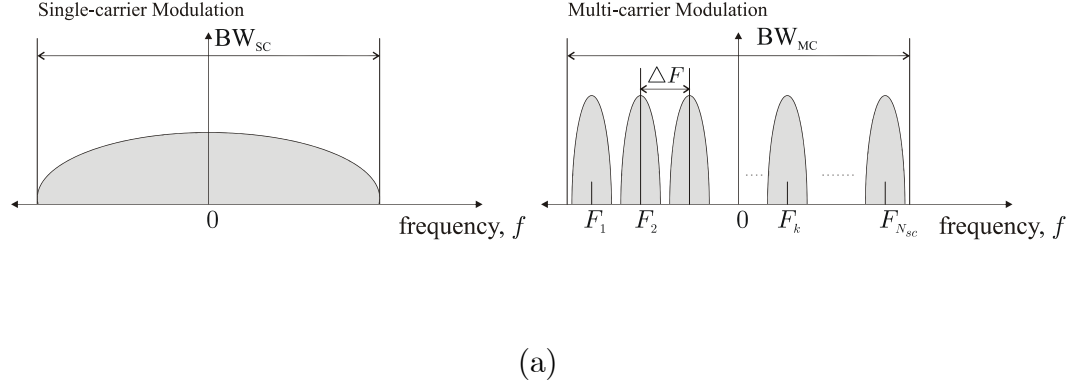


Figure 1.1: Comparison of single-carrier modulation and multi-carrier modulation systems: (a) frequency spectra of transmitted signals and (b) frequency spectra of received signals.

rate of a conventional SC system having the same throughput. This implies that MC systems which support high data rates can be designed while maintaining symbol durations much longer than the channel's memory. This is equivalent to the signal bandwidth being smaller than the coherence bandwidth of the channel. For instance, consider a SC modulation scheme with a symbol duration of T_{cs} which corresponds to a signal bandwidth of $\frac{1}{T_{cs}}$. However, in a OFDM system with N_{sc} subcarriers, the resultant OFDM symbol duration becomes $N_{sc}T_{cs}$ and this implies a signal bandwidth of $\frac{1}{N_{sc}T_{cs}}$ at each subcarrier. Thus, for a given channel, the bandwidth of the OFDM system has the possibility of being smaller than the coherence bandwidth of the channel. This implies that for the same channel, each subcarrier of the OFDM system experiences a frequency flat fading channel while in the SC system, a frequency-selective fading channel is observed. The details of the OFDM, MC DS-CDMA and MC-CDMA transmitters are now examined in detail.

1.2.1 OFDM

Figure 1.2 shows a baseband OFDM modulator. In this figure, the original channel symbol stream is de-multiplexed (serial-to-parallel converted) into N_{sc} substreams and each de-multiplexed symbol then modulates a subcarrier. The duration of each symbol on the subcarriers is N_{sc} times greater than the symbol period T_{cs} on the original symbol stream, i.e. the channel symbol duration on each subcarrier in an OFDM system is $N_{sc}T_{cs}$. In addition, each subcarrier is chosen to have a frequency separation of $\Delta F = 1/(N_{sc}T_{cs})$ from the next subcarrier so that they are orthogonal to each other over the duration $N_{sc}T_{cs}$. Through the careful selection of N_{sc} , the time interval $N_{sc}T_{cs}$ can be made much larger than the time duration of the channel-time dispersion and so, ISI can be made arbitrarily small and this results in each subcarrier undergoing frequency flat fading.

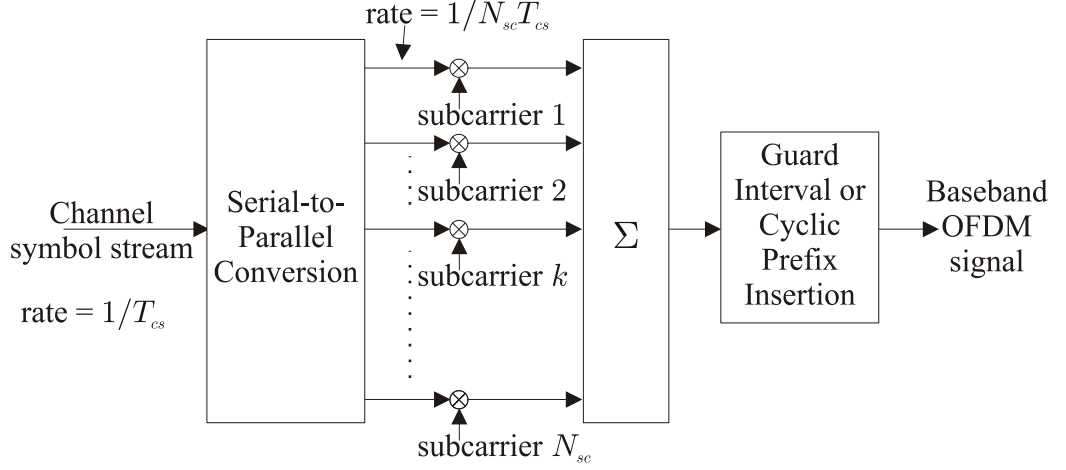


Figure 1.2: *Simplified baseband OFDM modulator.*

The idea behind the analog implementation of OFDM can be extended to the digital domain by using the discrete Fourier Transform (DFT) and its counterpart, the inverse discrete Fourier Transform (IDFT). These mathematical operations are widely used for transforming data between the time-domain and frequency-domain. These transforms are interesting from the OFDM perspective because they can be viewed as mapping data onto orthogonal subcarriers, see Figure 1.3. For example, the IDFT is used to take in frequency-domain data and convert it to time-domain data. Thus, the IDFT correlates the frequency-domain input data with its orthogonal basis functions, which are sinusoids at certain frequencies. This correlation is equivalent to mapping the input data onto the sinusoidal basis functions.

In practice, OFDM systems are implemented using a combination of fast Fourier Transform (FFT) and inverse fast Fourier Transform (IFFT) blocks that are mathematically equivalent but more efficient versions of the DFT and IDFT, respectively. An OFDM system treats the source symbols (e.g. the QPSK or QAM symbols that would be present in a single carrier system) at the transmitter as though they are in the frequency-domain. These symbols are used as the inputs to an IFFT block that brings the signal into the time-domain. This is

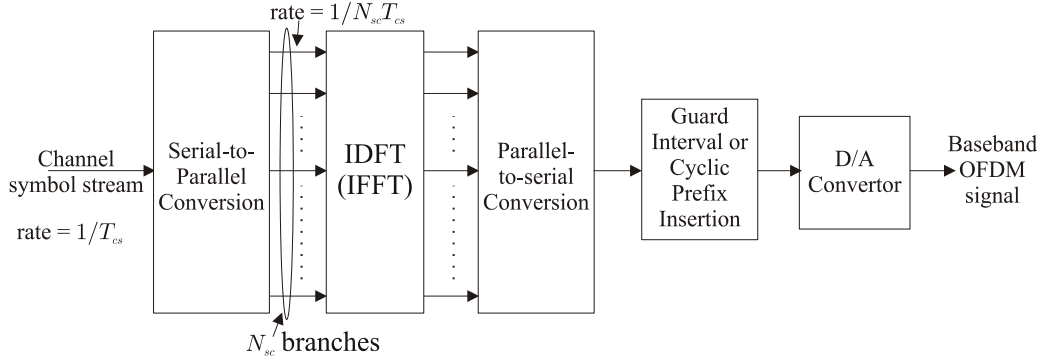


Figure 1.3: Block diagram of IDFT (IFFT) implementation of a baseband OFDM modulator.

shown in Figure 1.3 where the IFFT takes in N_{sc} symbols at a time. Each input symbol acts like a complex weight for the corresponding sinusoidal basis function. Since the input symbols are complex, the value of the symbol determines both the amplitude and phase of the sinusoid for that subcarrier. The IFFT output is the summation of all N_{sc} sinusoids. Thus, the IFFT provides a simple way to modulate data onto N_{sc} orthogonal subcarriers and the block of N_{sc} output samples from the IFFT make up a single OFDM symbol. The output of the IDFT operation is then parallel-to-serial converted. A guard interval or cyclic prefix is then inserted at the beginning of this OFDM symbol block. The resulting symbol block is then digital-analog converted for subsequent transmission.

1.2.2 MC DS-CDMA

An MC DS-CDMA system was first proposed in [5] which assigned a different channel symbol to each of the N_{sc} subcarriers. However, MC DS-CDMA has developed so that the most common definition considers a single channel symbol transmitted over every subcarrier, which provides frequency diversity at the receiver [6]. This is shown in Figure 1.4 where the symbol stream is spread by a PN-signal of length N_c before modulating N_{sc} subcarriers in parallel and the subcarrier separation is equal to the chip rate. Note that $N_c N_{sc} = N_c$, where N_c

is the length of the PN sequence used in a corresponding SC DS-CDMA system. Due to the MC modulation involved, a MC DS-CDMA modulator can also be implemented with the IFFT and this is shown in Figure 1.5.

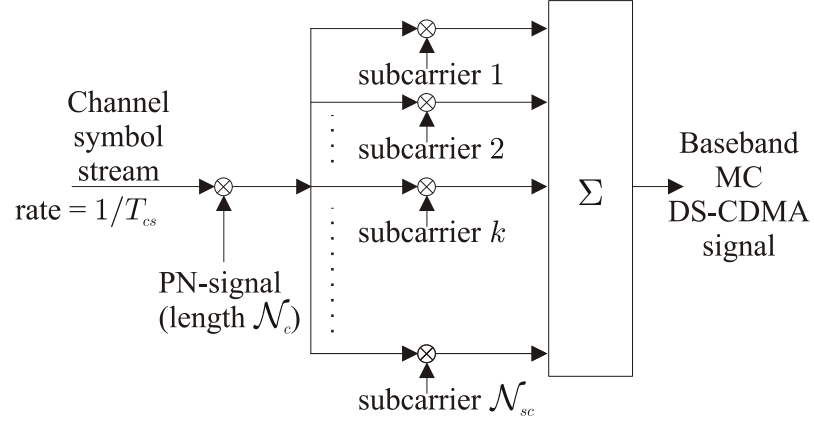


Figure 1.4: *Simplified baseband MC DS-CDMA modulator.*

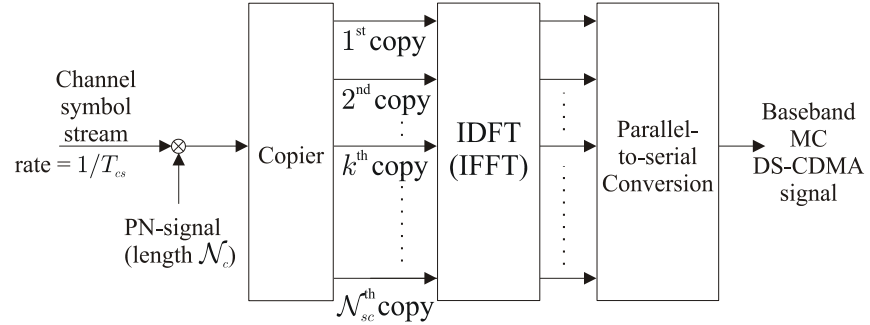


Figure 1.5: *Block diagram of IDFT (IFFT) implementation of a baseband MC DS-CDMA modulator.*

If each subchannel is assumed to be flat fading then multicarrier demodulation followed by despreading and diversity combining will yield the decision variables. In this case, the performance is very similar to that of SC DS-CDMA. Alternatively, if the system is designed so that the subchannels are frequency-selective and a RAKE receiver is used for each subcarrier, then performance gains are possible even though some intersubcarrier interference occurs [7].

1.2.3 MC-CDMA

Figure 1.6 illustrates a baseband MC-CDMA modulator with N_{sc} subcarriers. As seen in Figure 1.6, the channel symbol is repeated N_{sc} times and each copy is multiplied by a corresponding chip of a pseudo-noise (PN) code sequence [8]. The resultant N_{sc} signals then modulate the subcarriers in parallel and these modulated subcarriers are summed up to produce the baseband MC-CDMA signal.

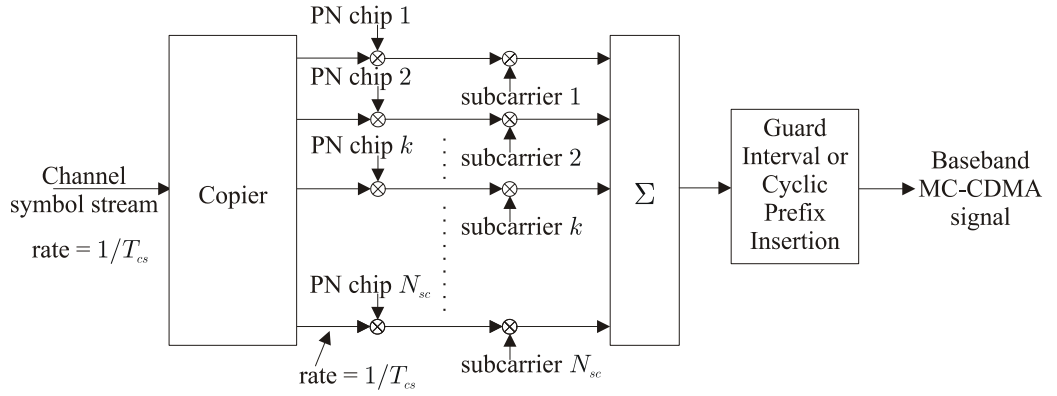


Figure 1.6: *Simplified baseband MC-CDMA modulator.*

The subcarriers are made to be orthogonal to each other over the channel symbol period by setting the frequency separation between two neighboring subcarriers to be equal to a multiple of the symbol rate, $\frac{1}{T_{cs}}$. Due to MC-CDMA's similarity with OFDM, MC-CDMA modulators can also be implemented with the IFFT. This is shown in Figure 1.7 where the N_{sc} copies of the channel symbol, after being multiplied by the PN sequence, are passed into the IDFT block. It is obvious that MC-CDMA is the frequency domain equivalence of DS-CDMA, where the spreading is carried out in the frequency domain, instead of the time domain. Due to the 'frequency-domain' spreading, each subcarrier carries a narrowband signal of duration T_{cs} . This is in contrast to DS-CDMA in which 'time-domain' spreading is carried out and the single carrier carries a wideband signal as the chip duration is $\frac{T_{cs}}{N_c}$ and $N_c = N_{sc}$ denotes the processing gain. As such,

MC-CDMA provides multiple access and also avoids frequency selective fading if the duration of the spread signal is large enough. Similar to the OFDM system, each symbol period is extended by a guard interval, shown in Figure 1.6 and Figure 1.7, which absorbs the delay spread of the channel.

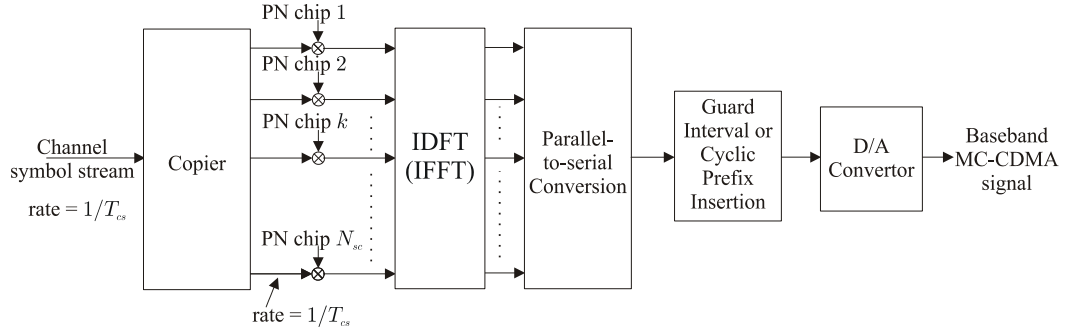


Figure 1.7: Block diagram of IDFT (IFFT) implementation of a baseband MC-CDMA modulator.

Table 1.1 shows a comparison of the bandwidth of each subcarrier in an OFDM, MC DS-CDMA and MC-CDMA with the bandwidth of the corresponding SC scheme. The narrow bandwidth at each subcarrier of MC signals, compared with their SC counterparts, provides MC signals with a higher immunity against delay spread, ISI and impulse noise. An appropriate choice of the number of subcarriers enables the length of the resultant MC symbol to become longer than the time span of the channel. As a result, the effect of ISI is confined to the first few samples, corresponding to the delay spread of the channel, of the received MC symbol. Hence, a guard interval (GI) can be employed to remove the effect of the ISI present.

1.2.4 Guard Intervals and Cyclic Prefixes

The GI can be a section of all zero samples transmitted in front of each MC symbol. Since the GI does not contain any useful information, the GI can be discarded at the receiver. A proper choice of the GI length is thus needed so

Table 1.1: Comparison of the bandwidth of each subcarrier in MC systems with their SC counterparts.

Modulation Scheme	Bandwidth at Carrier or Subcarrier
OFDM modulation (N_{sc} subcarriers)	$\frac{1}{N_{sc}T_{cs}}$
SC modulation	$\frac{1}{T_{cs}}$
MC DS-CDMA modulation (Processing Gain = $N_c = N_{sc}$)	$\frac{N_c}{T_{cs}} = \frac{N_c}{N_{sc}T_{cs}}$ or $\frac{N_{sc}}{N_{sc}T_{cs}}$
MC-CDMA modulation	$\frac{1}{T_{cs}}$
SC DS-CDMA modulation (Processing Gain = $N_c = N_{sc}$)	$\frac{N_{sc}}{T_{cs}}$ or $\frac{N_c}{T_{cs}}$

that it is longer than the time span of the channel so as not to distort the MC symbol itself. In the receiver, the GI is removed and in the process, the effects of ISI are removed as well. However, in practice, the guard interval is not used as it is unable to remove intrasymbol interference, also known as inter-carrier interference (ICI). The solution to the problem of ICI involves the discrete time property of signals. In continuous time, a convolution in time is equivalent to a multiplication in the frequency-domain. However, this property is true in discrete time only if the signals are of infinite length or if at least one of the signals is periodic over the range of the convolution. As it is impractical to have a MC symbol of infinite length, an alternative option is to make the MC symbol appear periodic. This periodic form is achieved by replacing the GI with a cyclic prefix which is a replica of the last few samples of the MC symbol. As in the case of the GI, the length of the cyclic prefix must be longer than the time span of the channel. Since it contains redundant information, the cyclic prefix is discarded at the receiver and in the process, the effects of ISI are removed. Because of the way in which the cyclic prefix was formed, the cyclically-extended MC symbol now appears periodic when convolved with the channel. Thus, the effect of the channel becomes multiplicative. In Figures 1.2, 1.3, 1.6 and 1.7, it is seen that GI or cyclic prefix is inserted into the MC symbol prior to transmission.

In a digital communications system, the symbols that arrive at the receiver

have been convolved with the time-domain channel impulse response. In order to undo the convolutional effects of the channel, another convolution must be performed at the receiver using a time-domain filter known as an equalizer. The equalizer processes symbols in order to adapt its response in an attempt to remove the effects of the channel and the length of the equalizer needs to be on the order of the time span of the channel. Such an equalizer can be expensive to implement in hardware and often requires a large number of symbols in order to adapt its response for a good performance.

In MC systems, the time-domain signal is still convolved with the channel response. However, the data will ultimately be transformed back into the frequency-domain by the DFT in the receiver. Because of the periodic nature of the cyclically-extended MC symbol, this time-domain convolution results in the multiplication of the spectrum of the MC signal with the frequency response of the channel. The result is that symbol at each subcarrier will be multiplied by a complex number equal to the channel's frequency response at that subcarrier's frequency. Thus, each received subcarrier experiences a complex gain due to the channel. An equalizer consisting of a single complex multiplication for each subcarrier can then be employed to undo these effects.

After the removal of the cyclic prefix, DFT is performed on the remaining received signal samples to demodulate the received signal. Due to the use of guard intervals and multiple subcarriers, MC systems are highly sensitive to time and frequency offsets. As such, time and frequency synchronization algorithms must be performed to ensure that OFDM has good performance. Time and frequency synchronization have often been performed with the use of pilot signals [9], leading to a loss of spectral efficiency as the pilot symbols use up valuable bandwidth. In [10], pilot symbols are used by the receiver for the acquisition and tracking of the carrier frequency. Pilot symbols have also been used for channel estimation

in MC-CDMA systems and much research has been done on the structure of pilot signals so that better performance can be achieved. For instance, in [11], the pilot signals are designed, using the weighted least squares (WLS) criterion, to have good signal-to-noise-plus-interference ratio (SNIR) and peak-to-average power ratio (PAPR) properties. The resulting pilot signals can be used for both synchronization and channel estimation purposes.

Besides using pilot symbols, the cyclic prefix can also be used to provide the synchronization. By removing the reliance on pilot signals, the system overhead is reduced and higher spectral efficiency is achieved. In [12], the cyclic prefix was used to perform joint maximum likelihood (ML) time and frequency offset estimation for non-dispersive channels. This removed the need for pilot symbols to be used for carrier frequency synchronization and delay estimation. The use of the cyclic prefix for estimating the delay spread and the power of the multipath signals was also proposed in [13]. However, the proposed method assumes a sparse multipath channel where the delayed signals have a large separation between them. Hence, the performance of the proposed method degrades when the multipaths are spaced closely together. Besides the use of cyclic prefix, it has been proposed that zero-padding can be used instead, where zero symbols are appended in place of the cyclic prefix [14, 15]. Zero-padding allows symbol recovery and finite impulse response (FIR) equalization of FIR channels regardless of the channel zero locations. However, such a method brings about an increase in receiver complexity as FIR filters will have to be used.

Although pilot symbols and cyclic prefix have been used for channel estimation, blind channel estimation techniques are much more attractive as they can completely remove the need for pilot signals or carriers and achieve higher spectral efficiency. In [16], a blind synchronization and carrier frequency offset estimator is proposed which introduces cyclostationarity into OFDM signals by

using time-frequency guard regions, pulse shaping or subcarrier weighting. Besides compensating for the timing and frequency offsets caused by the channel, estimation of the gain on each subcarrier also has to be performed. To remove the reliance on pilot symbols, a DBPSK-based MC-CDMA system was proposed in [17] for the downlink where the channel estimation is not carried out and the data symbols are recovered by making use of the property of the DBPSK modulation involved. However, the use of the proposed method is limited to the downlink where the transmission can be synchronous. Other than relying on the use of pilot symbols, subspace-based techniques are an attractive alternative. A subspace approach was proposed in [18] which makes use of virtual carriers in OFDM systems to carry out the estimation of the channel.

Although the presence of a cyclic prefix in a MC system enables the ISI effects of a channel to be removed by the receiver, its use lowers the spectral efficiency of the system. In order to improve the spectral efficiency of the system, cyclic prefix-free MC-CDMA systems are of interest and these have been investigated in [19, 20]. In addition, in [21], a cyclic prefix-free MC-CDMA system is proposed where the uplink finite impulse response (FIR) channel is estimated using subspace-based techniques. The method proposed is capable of estimating the channel up to a complex coefficient. However, in the method presented, there is an assumption that it is possible to obtain ISI-free received signal vectors from the received signals. Thus, there will have to be a certain degree of timing acquisition implemented in the receiver. However, the removal of the cyclic prefix has an adverse effect on the near-far resistance property of MC-CDMA systems. In [22], it was shown that a cyclic prefix-based MC-CDMA system, with a reduction in spectral efficiency, has much better near-far resistance capability than a cyclic prefix-free MC-CDMA system.

For multiuser applications of MC-CDMA, performance is severely reduced

even though the users are differentiated by their spreading codes. This is because the spreading codes only introduce a phase-shift of 0 or π in the subcarriers. As such, training-based or non-blind systems have to be used to attain a good performance. Moreover, synchronization of the users is required. Performance in asynchronous multiuser situations can be improved by the application of multi-user detection. In [23], a blind decorrelating detector based on subspace techniques is proposed for an asynchronous multi-user environment, while in [24], a blind subspace-based channel estimator and linear MMSE detector was proposed. However, the performance of the proposed receivers is limited by the assumption that the desired user is synchronized as this implies that synchronization to the desired user has to be performed.

In this thesis, the proposed communication systems are assumed to be operating in a channel with a delay spread equal to one channel symbol period. As such, due to the cyclic prefix-free nature of the MC modulation involved, the signal duration of the resultant MC-CDMA symbol in the proposed system is half of the signal duration in a conventional MC-CDMA system, which makes use of a cyclic prefix to overcome the frequency selectivity of the channel.

1.2.5 Crest Factor

All transmitters and receivers in communications systems contain devices such as amplifiers which have non-linear transfer functions. These non-linearities create an additional performance limitation. The receiver performance is typically limited by distortion generated in the input amplifier or mixer in the presence of strong undesired signals. On the other hand, the performance of the transmitter is limited by power amplifier linearity. A MC signal is made up of multiple simultaneous signals that, when combined together, have a higher peak signal level. Thus, for a given average power, MC signals result in an increase in the peak-to-

average power ratio (PAPR) of the signal, known also as crest factor (CF) of the signal, which are related as:

$$\text{PAPR} = (\text{CF})^2 = \left(\frac{\text{peak value}}{\text{rms value}} \right)^2$$

In multi-carrier systems, the PAPR value is often expressed in terms of statistics because of the low probability that all subcarriers will simultaneously reach peak amplitude, although the simultaneous peak amplitude value is large. These higher peak amplitude levels will create more severe distortion than a single carrier system even if the average power levels are the same. The higher distortion increases the SNR needed to maintain adequate performance. Linearity requirements in both the receiver and transmitter must be adjusted or ‘backed off’ to account for this increase in PAPR value. The PAPR value, and also the amount of linearity compensation, depends on a number of parameters such as the number of subcarriers and the level of SNR that must be maintained. Peak suppression techniques that have been proposed include coding, phase rotation and clipping.

Thus, MC modulation schemes, such as OFDM and MC-CDMA, are susceptible to the high PAPR problem. In [25], it is shown that a proper selection of the spreading codes used in a MC-CDMA system is capable of reducing the PAPR to a larger extent than the use of block coding to reduce the PAPR in an OFDM system.

1.2.6 General Objective

In this thesis, asynchronous cyclic prefix-free MC-CDMA systems are of interest due to the removal of reliance on the cyclic prefix as valuable signalling time can be used for the transmission of information instead. Moreover, MC-CDMA, in contrast with OFDM, has the potential to support more users on the same set of subcarriers due to code multiplexing in the form of CDMA. A discussion

on channel impairments that affect the performance of MC systems is presented in Appendix A. The cyclic prefix-free MC-CDMA system will be investigated in conjunction with MIMO systems, which will be described next.

1.3 Multiple-Input Multiple-Output

Multiple-input and multiple-output (MIMO) refers to the use of multiple antennas at both the transmitter and receiver to improve communication performance since it offers significant increases in data throughput and link range without additional bandwidth or transmit power. This is achieved by higher spectral efficiency and link reliability or diversity. However, MIMO is a general term for such multiple-antenna systems and other types of systems can be obtained by varying the number of antennas at transmitter and receiver. A multiple-input single-output (MISO) system results when the receiver has a single antenna while a single-input multiple-output (SIMO) system refers to the case where the transmitter has a single antenna. The simplest case, which occurs when neither the transmitter nor receiver have multiple antennas, is referred to as a single-input single-output (SISO) system. Figure 1.8 shows the four possible types of communications systems.

It has been shown in [26] that, contrary to common perception, the presence of multipath is an advantage in far-field MIMO systems and at high signal-to-noise ratios (SNRs), the capacity can be multiplied by adding antennas to both sides of the wireless link. As such, MIMO systems are an attractive option that can be used to realize the data rates that are required for future systems. MIMO systems are capable of bringing about such gains in capacity due to a combination of the following factors: array gain, diversity gain, spatial multiplexing gain, and interference reduction [27]:

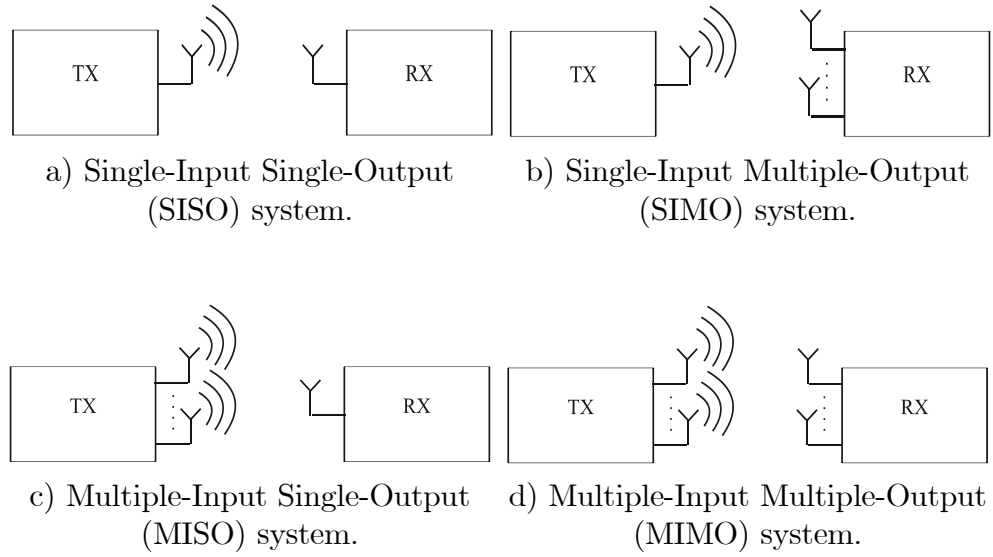


Figure 1.8: *Multi-antenna communications systems classification.*

- Array gain - refers to the increase in average receive signal-to-noise ratio (SNR) due to a coherent combining effect;
- Diversity gain - which is defined as the increase in signal-to-interference ratio due to some diversity scheme, or how much the transmission power can be reduced when a diversity scheme is introduced, without a degradation in performance.
- Spatial multiplexing gain - refers to a linear increase in capacity, without additional power or bandwidth expenditure, as a result of transmitting independent data signals from individual antennas; and
- Interference reduction - refers to the use of the difference in spatial signatures amongst users to reduce the interference to the desired user.

Transmit or receive array gain is dependent on the number of transmit and receive antennas as well as the availability of channel knowledge at the transmitter and receiver, respectively. For the receiver, channel knowledge can be obtained through blind channel estimation techniques or through the use of training sequences. On the other hand, channel knowledge can be made available at the

transmitter through feedback from the receiver. However, this assumes that the channel has the same characteristics in both the uplink and downlink, such as in time division duplex (TDD) systems. In MIMO systems, diversity gain is achieved in the form of spatial diversity and it is the least costly form of diversity as it does not involve the use of valuable bandwidth or signalling time, which occurs in the case of frequency and time diversity, respectively. At the receiver of a MIMO or SIMO system, the incoming signals can be combined after some processing so that the performance of the receiver can be improved. In the absence of channel knowledge, transmit diversity can be attained by using space-time coding techniques such as space-time block codes (STBCs) [28, 29]. However, in the Alamouti scheme [28], it is assumed that the transmitted signals undergo frequency flat fading and the channel is constant over at least two consecutive symbol periods. Moreover, channel knowledge must be made available to the receiver to carry out decoding of the received signal. Hence, there is a need to transmit orthogonal pilot symbols from each transmitting antenna so that the channel characteristics from each antenna to the receiver can be estimated. This results in a waste of valuable bandwidth. In [29], the Alamouti scheme is extended to MIMO systems with more than two transmitter elements through orthogonal space-time block codes.

Spatial multiplexing is another MIMO transmission technique method that can be used to overcome the lack of channel information at the transmitter. Spatial multiplexing is a transmission technique in MIMO wireless communication to transmit independent and separately encoded data signals from each of the multiple transmit antennas. Therefore, the space dimension is reused, or multiplexed, more than once, leading to spatial multiplexing gain. An example of such a scheme is vertical Bell Labs layered space-time (V-BLAST) architecture [30, 31, 32]. The V-BLAST architecture is a simple spatial multiplexing

scheme proposed for the flat-fading MIMO channel to support high rate data transmission. From the implementation perspective, V-BLAST is an attractive technique as each transmit antenna transmits an independently encoded data stream with equal transmission rate and power and the receiver detects each substream through successive cancellation. In simulations carried out, it was observed that spectral efficiencies of 20 - 40 bps/Hz can be achieved at SNRs from 24 to 34 dB. However, one drawback of the scheme is that accurate channel estimates must be obtained so that an accurate replica of the received substream prior to detection can be reproduced. This can then facilitate accurate cancellation of the replica, leading to better detection of the remaining substreams. On the other hand, in successive cancellation schemes, the receiver suffers from error propagation when errors are made in the detection of the substream. Another benefit of MIMO systems is that the multiple antennas at the receiver can exploit the spatial characteristics of the desired signal and co-channel signals and in the process, reduce the interference experienced by the desired signal.

1.4 Antenna Array Signal Processing

In mobile communication systems, the existence of multiple access interference (MAI) and co-channel interference (CCI) limit the ability to improve the performance of the system. The use of an antenna array introduces array gain into the receiver which allows the receiver to make use of diversity combining (or beam-forming) to improve its performance in overcoming the problem of fading in radio channels and makes use of the fact that the signals arriving at different locations fade at different rates. An antenna array is composed of a number of antenna elements that form an array system of a given geometry and measurements are taken with respect to the array reference point. Also, the beam pattern of the

antenna array depends on the geometry, amplitude and phase excitation of the elements.

Figure 1.9 shows a narrowband beamformer structure that exploits the space diversity at the receiver. In the narrowband assumption for array signal processing, the bandwidth of the signal is assumed to be narrow enough and that the array dimensions are small enough such that the baseband signal waveform remains almost constant at each array element. As shown in Figure 1.9, the resultant signal from each element, $[\underline{x}]_p(t)$, $p = 1, \dots, N$, is multiplied by a complex weight w_p^* , $p = 1, \dots, N$ and summed to form the array output. The effect of the complex weights w_p^* , $p = 1, \dots, N$ is to adjust the phase and amplitude of the signals induced on each element of the antenna array and this results in the formation of a beam in the direction where maximum gain is required. This process of combining the signals from different elements is known as beamforming and variations in the formation of the weights results in different types of beamformers.

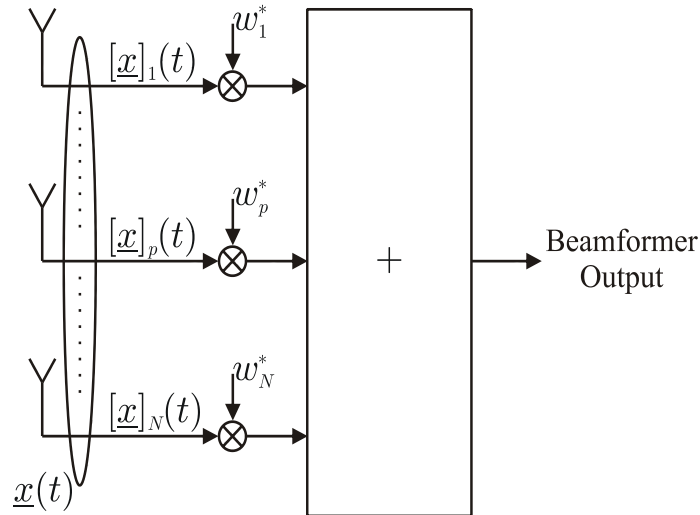


Figure 1.9: *Narrowband beamformer structure.*

There are many types of diversity combining schemes that are currently used in various wireless and cellular systems. For instance, in equal gain combining (EGC), the phases of the desired signals are adjusted and the signals are then

combined in-phase after equal weighting. In a selection diversity combiner, the signal from one of the antennas is selected for processing. The selection is performed based on a pre-defined criteria such as the power of the desired signal, the total power or the signal-to-interference ratio (SIR) at each antenna. This is in contrast to maximal ratio combining (MRC) where the weights are applied in proportion to the signal-to-noise ratio (SNR) and the weighted signals are combined in-phase. The MRC diversity combiner, which is equivalent to the Wiener-Hopf beamformer, suffers from a degradation in its resolution capabilities as the SNR is reduced. This implies that closely located interference may not be cancelled effectively. This has led to the development of super-resolution beamformers in which the resolution capabilities are independent of the SNR. However, super-resolution beamformers require the knowledge of the direction-of-arrivals (DOAs) of the incident signals and this information can be obtained from subspace-based techniques such as the Multiple Signal Classification (MUSIC) algorithm [33].

In the MUSIC algorithm, the array manifold is formulated as a set of all possible array responses and the array response is a function of the parameters to be estimated. The intersection of the array manifold with the signal subspace obtained from the array measurements then yield the estimates of the parameters of interest. Other subspace-based estimation algorithms include Estimation of Signal Parameters via Rotational Invariance Technique (ESPRIT) and the Weighted Subspace Fitting (WSF) approach.

1.5 Summary and Thesis Organization

This thesis investigates the use of an antenna array in an asynchronous MC-CDMA MIMO system where the cyclic prefix is omitted from the MC-CDMA modulation in order to conserve valuable bandwidth, which is a valuable resource

in future communications systems (4G and beyond). However, the lack of a cyclic prefix in MC-CDMA modulation implies that the ISI effects cannot be removed by the receiver and so, the resultant MC-CDMA signals would be susceptible to ISI in a frequency-selective fading channel. Such effects can be ameliorated through the use of antenna arrays as these provide additional space-time processing capabilities which are not available to systems that assume the use of independent antenna elements.

The main objective of the thesis is to design array receivers that are applicable for cyclic prefix-free MC-CDMA arrayed MIMO systems and are capable of overcoming the ISI arising from not using cyclic prefixes. The organization of the thesis is as follows:

- In Chapter 2, the concept of the array manifold vector is introduced. The use of the array manifold vector enables the spatial information to be incorporated into the channel model. This leads to the introduction of space-time channel models. A distinction is made between array-based and non-array-based multiple antenna systems. Both the diagrammatic and mathematical representations of these models are provided. This modelling forms the basic mathematical framework for later formulation in the subsequent chapters.
- In Chapter 3, a subspace-based receiver is proposed for a cyclic prefix-free MC-CDMA MIMO system. The MIMO system investigated here considers the case where antenna arrays are employed at both the transmitter and receiver. The proposed receiver achieves ISI cancellation for the cyclic prefix-free MC-CDMA system and its performance is shown to be superior than that of a representative existing system which relies on cyclic prefixes and ignores the array geometry. In order to reduce system overheads,

a blind subspace-based channel estimation technique is used to obtain the channel parameters required for the formation of the receiver weight matrix.

- In Chapter 4, the phenomenon of spatial scattering in a cyclic prefix-free MC-CDMA MIMO system with antenna arrays at both ends of the wireless link is addressed. Due to the spatial scattering, a diffused channel, where each path is made up of a cluster of inseparable paths, results. By taking the spatial scattering effects into consideration, it is seen that the proposed channel estimation technique outperforms a channel estimator that ignores the effects of spatial scattering. In addition, a decorrelating receiver is devised to cope with the effect of a diffused channel.
- In Chapter 5, the downlink of a multi-user cyclic prefix-free MC-CDMA MIMO system is considered and the problem of the joint optimization of the transmitter and receiver beamforming weights is addressed. Antenna arrays are assumed to be employed at the transmitter as well as at each user terminal. Lagrange multipliers are used in the optimization problem which seeks to minimize the mean square error (MSE) of all users subject to constraints in the PAPR of each transmitter antenna. An iterative algorithm is proposed for the determination of the transmitter and receiver beamforming weights.
- In Chapter 6, the thesis is concluded. The contributions of the work in this thesis are outlined and potential future research areas are identified.

Chapter 2

Space-Time Channel and System Architecture Modelling

In this chapter, the concept of the array manifold vector is first introduced which provides a basis for the subsequent derivation of the parametric channel models that are applicable for the array-based MIMO systems that are considered in this thesis. The parametric channel models that are developed form the basis for the subsequent channel estimation and reception algorithms proposed in Chapters 3 to 5 of the thesis. In addition, the general framework of cyclic prefix-free MC-CDMA arrayed MIMO communication systems is presented and its core components are outlined. A detailed modelling of each of these components will be provided in the subsequent corresponding chapters.

2.1 Antenna Array Manifold Vector

In order to exploit the spatial properties of the channel, an antenna array is implemented at the front-end of the receiver. This can be further extended to the case when the transmitter is similarly equipped with an antenna array. A mathematical model of the spatial information provided by the antenna array is obtained through the use of the array manifold vector, which is a function of a number of channel parameters, including the array geometry, the carrier frequency and the direction-of-arrival (DOA) of each multipath.

Figure 2.1 shows an illustration of the planewave propagation model of the planewave signals in real 3-D space. In this thesis, a narrowband array model is assumed to be valid as the array aperture is small enough such that the time required for the propagating wave to pass through the array is significantly lesser than the symbol duration [34]. Assume that the j^{th} path of the i^{th} user arrives at the reference point of the antenna array from the direction (θ_{ij}, ϕ_{ij}) , where θ and ϕ denotes the azimuth and elevation angles, respectively. Thus, the real 3×1 unit-vector $\underline{u}_{ij}$, pointing towards the direction (θ_{ij}, ϕ_{ij}) , can be written as

$$\underline{u}_{ij} = [\cos(\theta_{ij}) \cos(\phi_{ij}), \sin(\theta_{ij}) \cos(\phi_{ij}), \sin(\phi_{ij})]^T \quad (2.1)$$

Thus, taking the k^{th} subcarrier in a MC system into consideration, the relative phase variation at the p^{th} antenna element with respect to the reference point of the antenna array can be expressed as

$$\exp\left(-j \frac{2\pi}{c} (F_c + F_k) \underline{r}_p^T \underline{u}_{ij}\right) \quad (2.2)$$

where $\underline{r}_p \in \mathcal{R}^{3 \times 1}$ denotes the Cartesian coordinates, in metres, of the location of the p^{th} antenna element, c denotes the speed of light and F_c and F_k denote the carrier frequency and k^{th} subcarrier frequency, respectively, with $F_k = k\Delta F$ and ΔF denotes the subcarrier separation. The wavenumber vector $\underline{k}_{ijk}$ is now

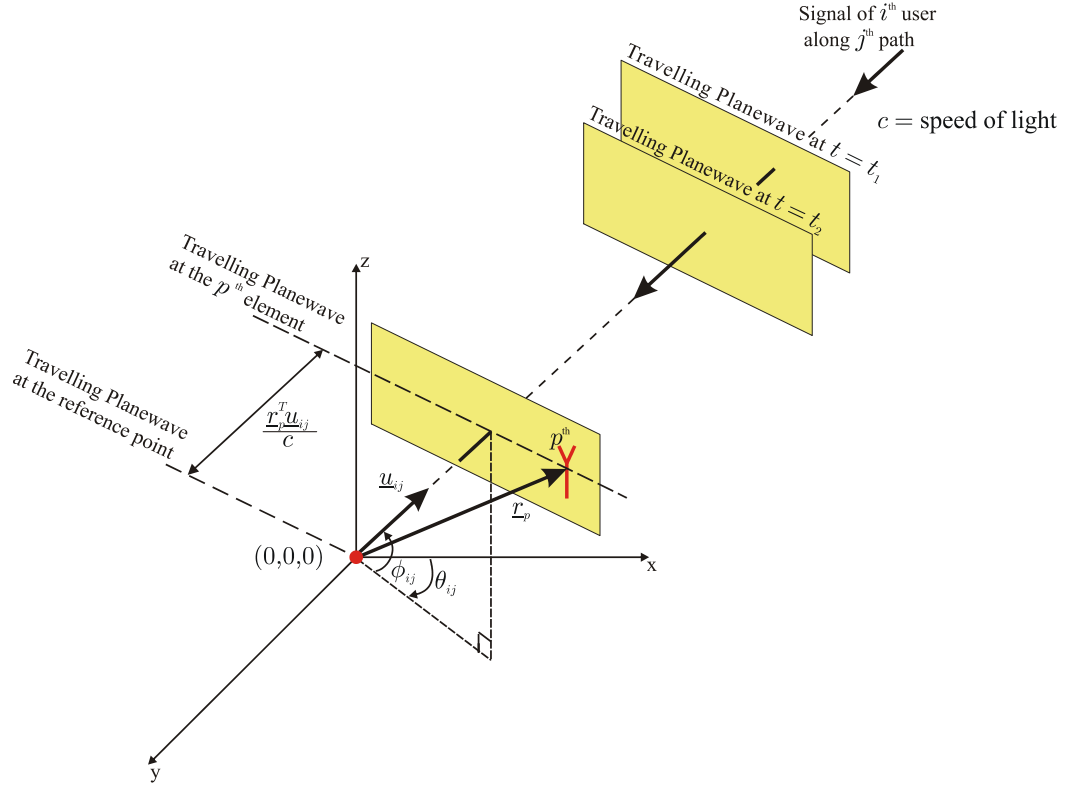


Figure 2.1: Illustration of the planewave propagation model based on the j^{th} path of the i^{th} user.

introduced as

$$\underline{k}_{ijk} = \frac{2\pi (F_c + F_k)}{c} \underline{u}_{ij} \quad (2.3)$$

which can be rewritten as follows:

$$\begin{aligned} \underline{k}_{ijk} &= \frac{2\pi (F_c + F_k)}{c} \underline{u}_{ij} \\ &= \frac{2\pi F_c}{c} \left(1 + \frac{F_k}{F_c} \right) \underline{u}_{ij} \\ &= \frac{2\pi F_c}{c} \left(1 + \frac{k\Delta F}{F_c} \right) \underline{u}_{ij} \end{aligned} \quad (2.4)$$

Thus, when the carrier frequency F_c is much greater than the maximum subcarrier frequency in a MC system, i.e. $\max_k \left(\frac{k\Delta F}{F_c} \right) \rightarrow 0$, the wavenumber vector $\underline{k}_{ijk}$ can be simplified to be independent of the subcarrier frequency F_k , $\forall k$. Thus, for all subcarriers, the wavenumber vector can be represented by

$$\underline{k}_{ij} \triangleq \frac{2\pi F_c}{c} \underline{u}_{ij} \simeq \underline{k}_{ijk} \forall k \quad (2.5)$$

Hence, for an antenna array with N omnidirectional elements, the array manifold vector associated with the j^{th} path of the i^{th} user's DOA (θ_{ij}, ϕ_{ij}) is given by

$$\begin{aligned}\underline{S}_{ij} &\triangleq \underline{S}(\theta_{ij}, \phi_{ij}) = \exp\left(-j \left[r_1^T \underline{k}_{ij}, r_2^T \underline{k}_{ij}, \dots, r_N^T \underline{k}_{ij}\right]^T\right) \\ &= \exp\left(-j \left[\underline{\mathbf{r}}^T \underline{k}_{ij}\right]\right) \in \mathcal{C}^{N \times 1}\end{aligned}\quad (2.6)$$

where $\underline{\mathbf{r}} = [r_1, r_2, \dots, r_N] = [r_x, r_y, r_z]^T$ is a $3 \times N$ matrix with its p^{th} column corresponding to the location r_p of its p^{th} antenna element. Note that Equation 2.6 is applicable for all subcarriers in a MC system due to the simplification that has been performed in Equation 2.5. The set of array response vectors (manifold vectors), $\{\underline{S}(\theta, \phi) \forall \theta, \phi \in [0, 2\pi]\}$ forms the array manifold. Typically, the signals are assumed to be on the (x,y)-plane, i.e. $\phi_{ij} = 0^\circ$. Thus, the array manifold vector, as shown in Equation 2.6 can be simplified to

$$\underline{S}_{ij} = \exp\left(-j\pi \left(r_x \cos \theta_{ij} + r_y \sin \theta_{ij}\right)\right) \in \mathcal{C}^{N \times 1}$$

where the sensor location is measured in units of half-wavelength, i.e. $\frac{\lambda_c}{2}$, where $\lambda_c = \frac{c}{F_c}$.

2.2 Space-Time Channel Modelling

In most MIMO systems that have been developed, one common approach has been that the channel matrix is a random matrix which results from a stochastic channel which conforms to certain distributions without characterizing individually the probability density function (pdf) of the random channel variables involved in the modelling. Such models have been widely used in capacity calculations for the development of MIMO systems [35, 36]. Thus, in such MIMO systems, the array geometry, due to the use of an antenna array, is often ignored. This implies array signal processing techniques are unavailable in such systems.

Antenna array signal processing techniques are an attractive option which can be used to further exploit the spatial properties of the channel and, in the process, achieve increases in capacity and spectral efficiency while minimizing transmission power. In addition, an extra layer of co-channel interference cancellation capability can be provided and furthermore, new methods of handling unwanted channel effects, such as Doppler spread and fading, can be proposed for even more efficient utilization of spectrum and space. An antenna array-based MIMO receiver is proposed in [37] for asynchronous multipath DS-CDMA systems. By using subspace-based techniques, the proposed receiver could estimate the spatial and temporal characteristics as well as the Doppler spread in the signals. Thus, antenna array-based MIMO can be implemented for improved performance over independent antenna elements in MIMO systems.

In the literature, there are many ways of referring to multi-antenna communications systems. One of the most common nomenclature takes the form of MIMO (and the related SISO, SIMO and MISO) systems (described in Section 1.3) which is based on the number of inputs and outputs of the system. Another type of classification takes the form of multiple-element multiple-element (MEME) systems where the ‘element’ refers to the number of antenna elements at the transmitter

or receiver. In this classification, single-element single-element (SESE), single-element multiple-element (SEME), multiple-element single-element (MESE) and MEME systems correspond to the SISO, SIMO, MISO and MIMO systems, respectively.

However, in this thesis, the classification of the multi-antenna systems based on the nature of the input signal is preferred. In particular, the multi-antenna systems are classified according to whether a single signal or a set of more than one signals enters or leaves the system and these are referred to as scalar-signal and vector-signal, respectively. Thus, the classification of the space-time wireless channels is based on the number of inputs and outputs of the channel, using the terms

- Scalar-input (output) to denote a single-input (output); and
- Vector-input (output) to denote a set of more than one inputs (outputs).

Such a classification enables a closer mathematical representation of the system and this facilitates the parametric modelling of such multi-antenna systems. The various types of multi-antenna systems classification is shown in Table 2.1 to illustrate the equivalence between the different types of nomenclature.

The various space-time channel models for both single user and multi-user systems, which are classified based on the structure of the signal, can be summarized as shown in Figure 2.2. Thus,

- *Scalar-Input Scalar-Output (SISO) channel*: e.g. In a single user system, single antenna elements are employed at both ends of the transmission link;
- *Scalar-Input Vector-Output (SIVO) channel*: e.g. In a single user system, a single antenna and an antenna array is employed at the transmitter and receiver, respectively;
- *Vector-Input Scalar-Output (VISO) channel*: e.g. Based on Figure 2.2, in the case of a single user system, a VISO channel is formed when an antenna

Table 2.1: Correspondence between the different types of nomenclature for multi-antenna systems.

Number of Inputs/Outputs	Number of Elements	Structure of Signal
Single-Input Single-Output (SISO)	Single-Element Single-Element (SESE)	Scalar-Input Scalar-Output (SISO)
Single-Input Multiple-Output (SIMO)	Single-Element Multiple-Element (SEME)	Scalar-Input Vector-Output (SIVO)
Multiple-Input Single-Output (MISO)	Multiple-Element Single-Element (MESE)	Vector-Input Scalar-Output (VISO)
Multiple-Input Multiple-Output (MIMO)	Multiple-Element Multiple-Element (MEME)	Vector-Input Vector-Output (VIVO)

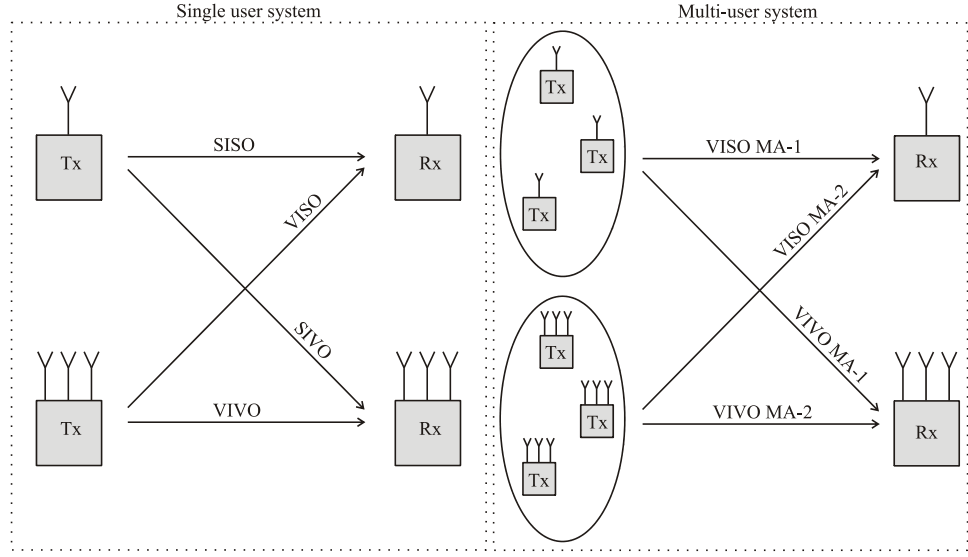


Figure 2.2: Space-time channel classification based on scalar input/output and vector input/output of the channel.

array is employed at the transmitter and a single antenna is used at the receiver. In a multi-user system, VISO channel can be further classified into: (a) VISO MA-1 which involves multiple users employing a single antenna and a single antenna is employed at the receiver; (b) VISO MA-2 which corresponds to the case where antenna arrays are used at the transmitters and a single antenna is employed at the receiver.

- *Vector-Input Vector-Output (VIVO) channel:* e.g. Similar to the VISO channel, there are different types of VIVO channels depending on how vector-signals are formed at the transmitter and receiver. In a single user system, a VIVO channel is formed when antenna arrays are used at both ends of the communication link. In the case of a multi-user system, the two types of VIVO channel are: (a) VIVO MA-1 where the system employs single antenna transmitters and an antenna array receiver; and (b) VIVO MA-2 where antenna arrays are used at both the transmitters and the receiver in the system.

2.2.1 Scalar-Input Scalar-Output (SISO) Channel

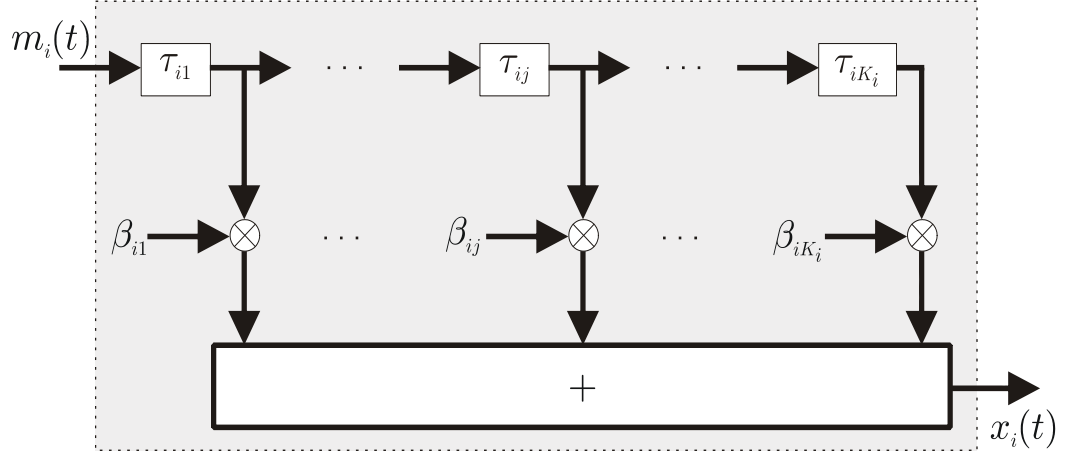


Figure 2.3: *Scalar-Input Scalar-Output (SISO) channel model.*

A basic wireless channel is formed when both the transmitter and receiver employ a single antenna. In such a system, the channel spatial information is not incorporated into the channel and this results in the SISO channel. The SISO channel model is illustrated in Figure 2.3 which considers the wireless link from a single antenna transmitter to a single antenna receiver. In the multipath slow fading channel considered here, the transmitted scalar-signal $m_i(t)$ arrives at the receiver via K_i multipaths. Hence, the channel impulse response between the transmitter and receiver can be represented by a summation of the multipath components, given by

$$\text{SISO channel impulse response} = \sum_{j=1}^{K_i} \beta_{ij} \delta(t - \tau_{ij}) \quad (2.7)$$

where each j^{th} multipath is characterized by its complex fading coefficient β_{ij} and path delay τ_{ij} . The fading coefficient β_{ij} can be described by a complex Gaussian distribution while the delay (TOA) τ_{ij} is uniformly distributed in the range $[0, T_{\text{spread}})$ where T_{spread} denotes the delay spread of the channel.

Based on Figure 2.3, the scalar-signal at the output of the channel $x_i(t)$ is a result of the convolution of the transmitted signal $m_i(t)$ with the channel impulse

response and is given by

$$x_i(t) = \sum_{j=1}^{K_i} \beta_{ij} m_i(t - \tau_{ij}) \quad (2.8)$$

2.2.2 Scalar-Input Vector-Output (SIVO) Channel

In contrast to the SISO channel, the implementation of a single antenna at the transmitter and an antenna array at the receiver results in a SIVO channel. Due to the presence of the antenna array at the receiver, the spatial properties of the channel can be incorporated into the model. This is accomplished through the use of the array manifold vector introduced in Section 2.1.

Similar to the SISO channel model, the multipath propagation in a SIVO channel model is modelled as a summation of a number of multipath components of the channel. This is shown in Figure 2.4 where each multipath is characterized by its own manifold vector (i.e. its own DOA (θ_{ij}) , TOA (τ_{ij}) and path fading coefficient (β_{ij})). As shown in the figure, the antenna array at the receiver is made up of N antenna elements.

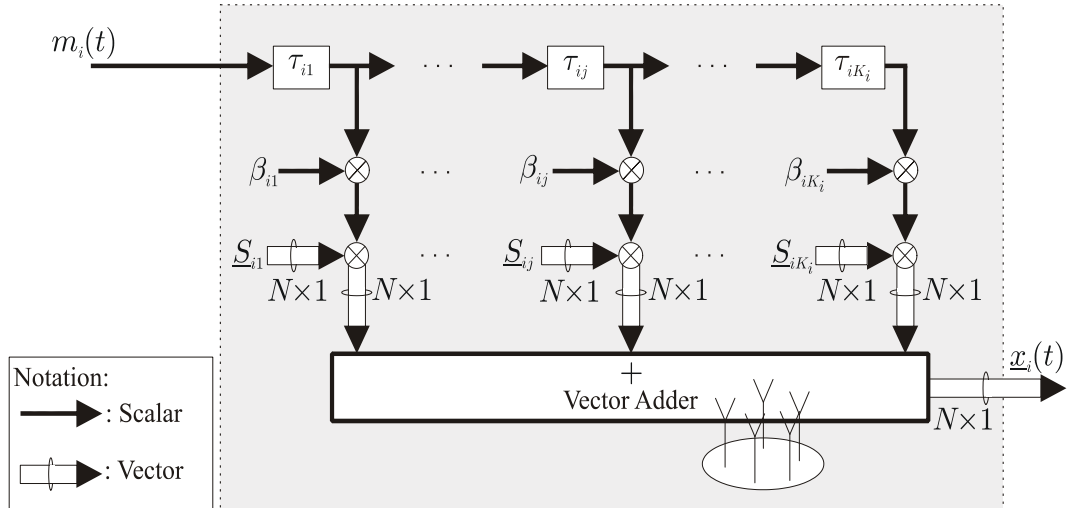


Figure 2.4: *Scalar-Input Vector-Output (SIVO) channel model.*

The resulting channel impulse response between the transmitter and the re-

ceiver is given by

$$\text{SIVO channel impulse response} = \sum_{j=1}^{K_i} \beta_{ij} \underline{S}_{ij} \delta(t - \tau_{ij}) \quad (2.9)$$

In this case, the vector-signal at the output of the channel $\underline{x}_i(t)$ can be expressed as

$$\underline{x}_i(t) = \sum_{j=1}^{K_i} \beta_{ij} \underline{S}_{ij} m_i(t - \tau_{ij}) \quad (2.10)$$

2.2.3 Vector-Input Scalar-Output (VISO) Channel

In the VISO channel model, the transmitter is equipped with an antenna array while the receiver employs a single antenna. An illustration of the VISO channel model is shown in Figure 2.5 for a point-to-point communication system with an antenna array of $\bar{N}$ elements at the transmitter and a single antenna array at the receiver. In this case, the transmit array manifold vector is denoted by $\bar{\underline{S}}_{ij}$, which is similarly defined as in Equation 2.6. The *transmit* array manifold vector, similar to its counterpart – the *receive* array manifold vector, is a function of the channel parameters associated with the transmitter antenna array. This includes the transmitter antenna array geometry and the direction-of-departures (DODs) of the channel.

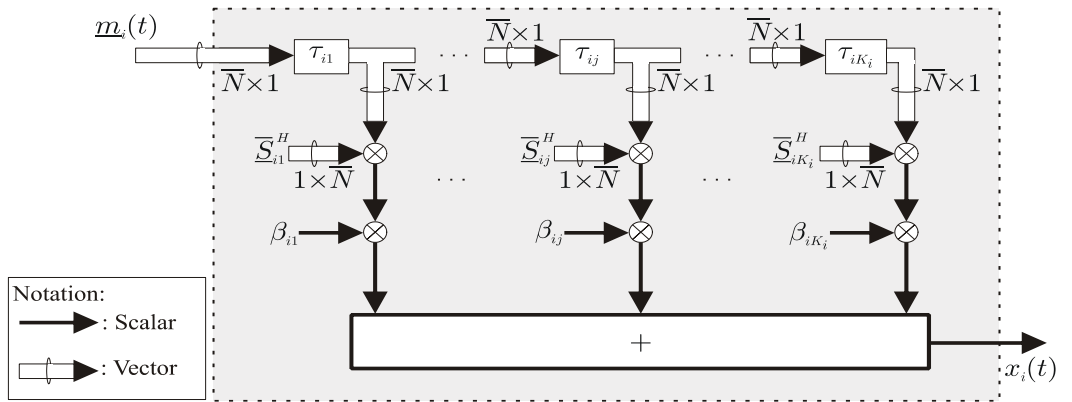


Figure 2.5: Vector-Input Scalar-Output (VISO) channel model for an array-based system.

Based on Figure 2.5, the VISO channel impulse response function can be

written as

$$\text{VISO channel impulse response} = \sum_{j=1}^{K_i} \beta_{ij} \underline{\bar{S}}_{ij}^H \underline{\delta}(t - \tau_{ij}) \quad (2.11)$$

and the output scalar-signal $x_i(t)$ is then obtained as

$$x_i(t) = \sum_{j=1}^{K_i} \beta_{ij} \underline{\bar{S}}_{ij}^H \underline{m}_i(t - \tau_{ij}) \quad (2.12)$$

2.2.4 Vector-Input Vector-Output (VIVO) Channel

Similar to the VISO channel, the VIVO channel model can be applied in point-to-point communication systems. The VIVO channel model in a point-to-point communication system, due to the use of antenna arrays at both the transmitter and receiver, is illustrated in Figure 2.6 with $\bar{N}$ and N elements in the antenna arrays at the transmitter and receiver, respectively.

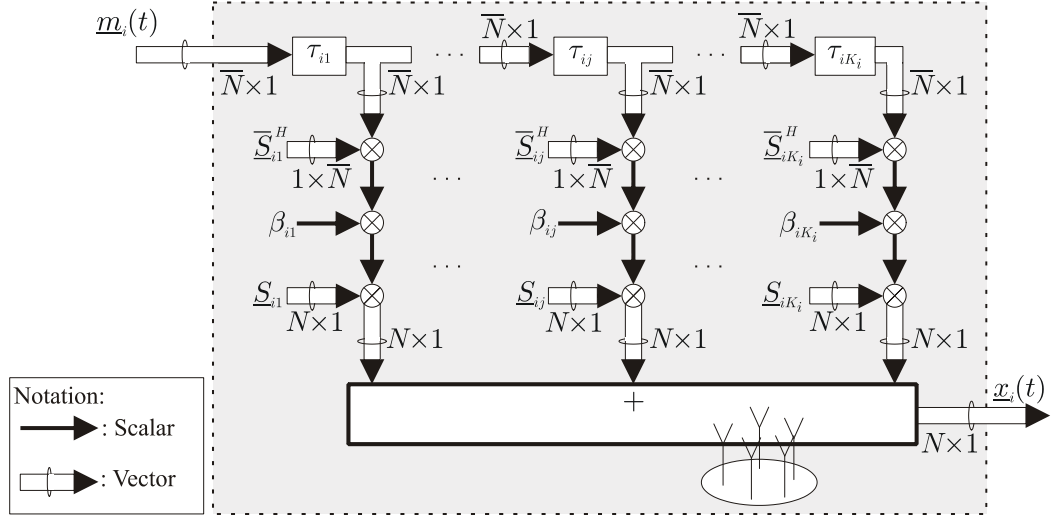


Figure 2.6: Vector-Input Vector-Output (VIVO) channel model.

Based on Figure 2.6, the VIVO channel model has a channel impulse response which can be represented by

$$\text{VIVO channel impulse response} = \sum_{j=1}^{K_i} \beta_{ij} \underline{S}_{ij} \underline{\bar{S}}_{ij}^H \underline{\delta}(t - \tau_{ij}) \quad (2.13)$$

Thus, the output vector-signal $\underline{x}_i(t)$ can be written as

$$\underline{x}_i(t) = \sum_{j=1}^{K_i} \beta_{ij} \underline{S}_{ij} \bar{S}_{ij}^H \underline{m}_i(t - \tau_{ij}) \quad (2.14)$$

2.2.5 Vector-Input Scalar-Output Multiple Access (VISO MA) Channels

The VISO channel model introduced in Section 2.2.3 is not limited to point-to-point communications systems and it is also applicable for the case of a multiple access communication system where multiple users, each equipped with a single antenna, are transmitting to a single-antenna receiver. Such a channel will be referred to as VISO MA-1 channel, as indicated in Figure 2.2. As seen in Figure 2.7, each SISO subchannel in the VISO MA-1 channel corresponds to each user for a system with M users. Thus, the received signal of the VISO MA-1 channel is given by

$$x(t) = \sum_{i=1}^M \sum_{j=1}^{K_i} \beta_{ij} m_i(t - \tau_{ij}) \quad (2.15)$$

Another VISO MA channel results when multiple users, each equipped with an antenna array, transmit to a receiver which employs a single antenna. The channel in such a system is referred to as a VISO MA-2 channel to differentiate it from the VISO MA-1 channel. The VISO MA-2 channel model is illustrated in Figure 2.8. In the case of the VISO MA-2 channel, the received scalar-signal is expressed as

$$x(t) = \sum_{i=1}^M \sum_{j=1}^{K_i} \beta_{ij} \bar{S}_{ij}^H \underline{m}_i(t - \tau_{ij}) \quad (2.16)$$

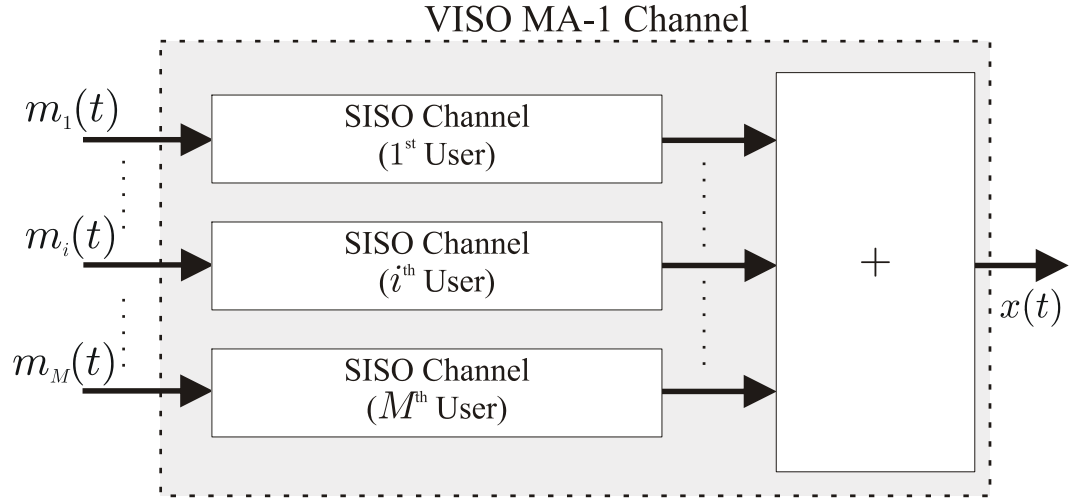


Figure 2.7: VISO MA-1 channel model where M users, each equipped with a single antenna, transmit to a single antenna receiver.

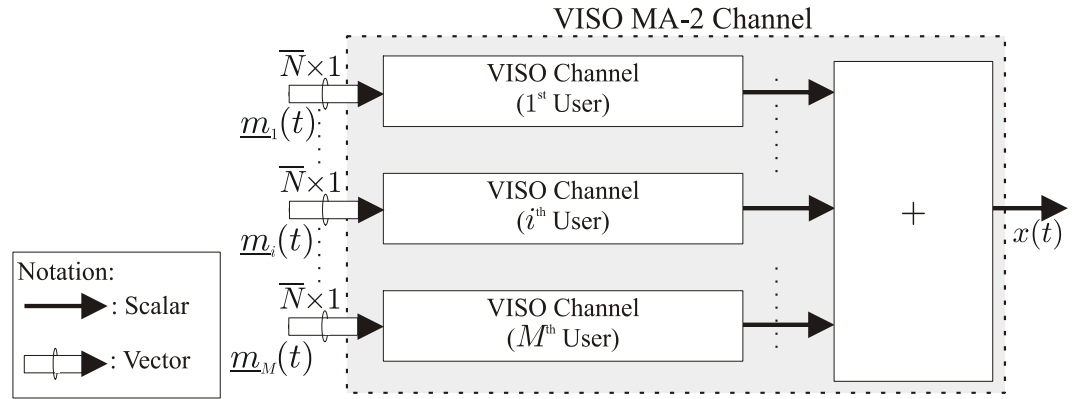


Figure 2.8: VISO MA-2 channel model where M users, each equipped with an antenna array, transmit to a single antenna receiver.

2.2.6 Vector-Input Vector-Output Multiple Access (VIVO MA) Channels

The counterpart of the VISO MA-1 channel in the VIVO MA channel framework results when multiple users, each equipped with a single antenna, are transmitting to a receiver which employs an antenna array. The resultant channel model is thus referred to as a VIVO MA-1 channel model. However, instead of M SISO subchannels, the VIVO MA-1 channel is composed of M SIVO subchannels and the received signal at each antenna is a superposition of the transmitted signals

from each antenna via multipath fading. This is illustrated in Figure 2.9. Thus, the received signal of the VIVO MA-1 channel is given by

$$\underline{x}(t) = \sum_{i=1}^M \sum_{j=1}^{K_i} \beta_{ij} \underline{S}_{ij} m_i(t - \tau_{ij}) \quad (2.17)$$

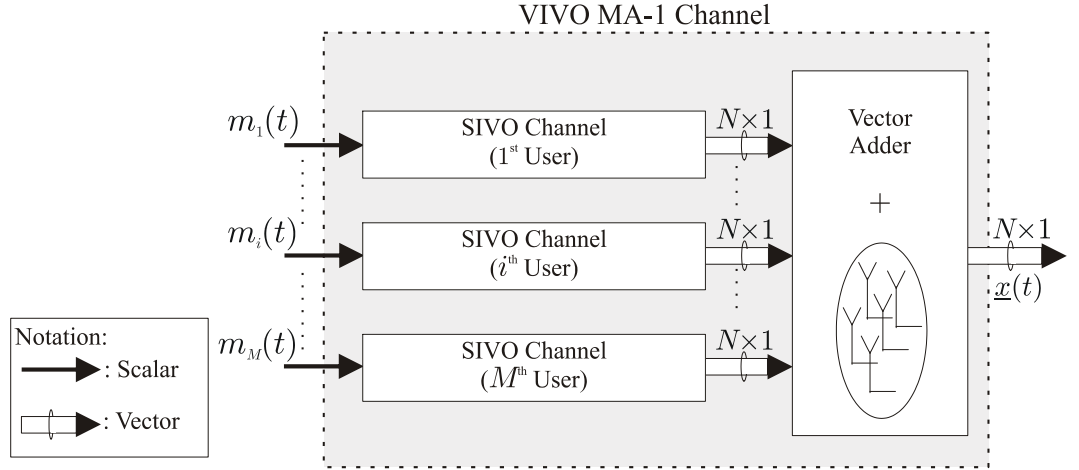


Figure 2.9: VIVO MA-1 channel model where M users, each equipped with a single antenna, transmit to a receiver equipped with an antenna array.

In addition, similar to the VISO MA-2 channel, a VIVO MA-2 channel, shown in Figure 2.10, arises when multiple users, each equipped with an antenna array, transmit to a receiver equipped with an antenna array. Thus, the VIVO MA-2 channel output is given by

$$\underline{x}(t) = \sum_{i=1}^M \sum_{j=1}^{K_i} \beta_{ij} \underline{S}_{ij} \overline{\underline{S}}_{ij}^H \underline{m}_i(t - \tau_{ij}) \quad (2.18)$$

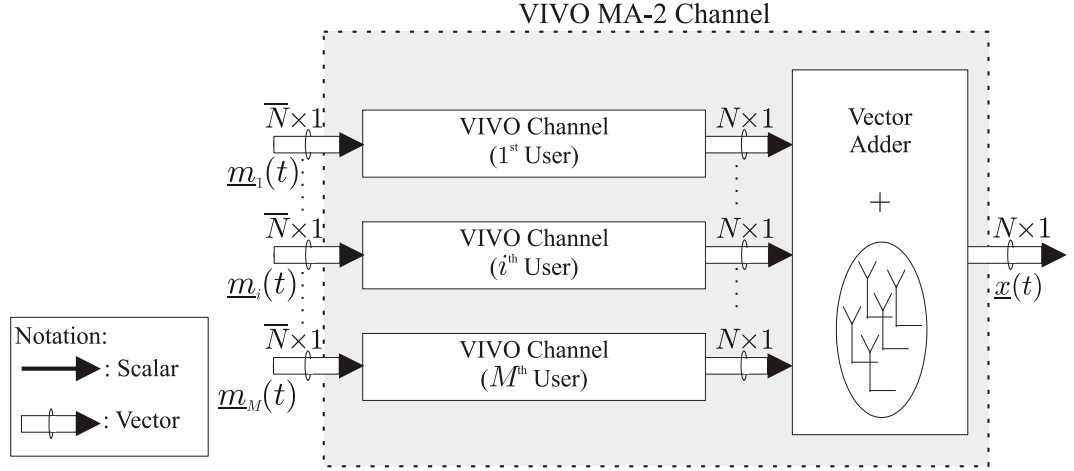


Figure 2.10: VIVO MA-2 channel model where M users, each using an antenna array, transmit to a receiver employing an antenna array.

2.3 Space-Time System Architecture

A wireless communication system is made up of three fundamental blocks:

- the transmitter;
- the radio channel; and
- the receiver.

In Section 2.2, various types of wireless channel models, based on the scalar/vector notation, have been presented. In the transmitter, the baseband signal is upconverted to the carrier frequency prior to transmission. The received signal is then downconverted to baseband in the receiver. However, without any loss of generality, baseband transmission will be assumed in this thesis. The space-time channel model represents the propagation of the baseband transmitted scalar-signal $m_i(t)$ or vector-signal $\underline{m}_i(t)$ from the transmitter to the receiver. The modelling of the transmitter and receiver blocks will now be presented so as to obtain a holistic view of the system architecture.

First, a single user MIMO system is shown in Figure 2.11 where antenna arrays of $\bar{N}$ and N elements are used at the transmitter and receiver, respectively. The use of an antenna array at the receiver is in contrast with conventional

MIMO systems which ignore the array geometry, i.e. the antenna elements are viewed as a collection of independent multiple elements instead of an antenna array. The MIMO system illustrated in Figure 2.11 is termed as an arrayed MIMO system in order to distinguish it from conventional MIMO systems which ignore the effects of the array geometry. At point-A of the transmitter, spatial multiplexing is performed and the data stream is first demultiplexed into $\bar{N}$ substreams (one substream per array element). The resultant vector-signal then undergoes cyclic prefix-free MC-CDMA modulation to form the baseband vector-signal $\underline{m}(t)$. The signal $\underline{m}(t)$ is then transmitted via the $\bar{N}$ element antenna array (point-C). The continuous transmitted signal propagates via a number of multipaths before reaching the receiver. As shown in Figure 2.11, the relevant channel model is the VIVO channel model, as described in Figure 2.6.

The task of the receiver is to capture the data and process it accordingly so that the information signals can be recovered. The use of an antenna array at the receiver can significantly increase the channel capacity by exploiting the spatial diversity [38] for example, to combat fading and to perform interference cancellation. The continuous signal $\underline{x}(t)$ is passed and stored in the bank of tapped delay lines (TDLs) to obtain the $2NL$ -dimensional discretized signal $\underline{x}[n]$ at point-E in Figure 2.11. In Chapter 3, the channel estimation and weight formation block at the receiver will be studied extensively. The details of the TDL block will be presented subsequently as it is a common feature among the three system architectures that are presented in this thesis for Chapters 3, 4 and 5.

In contrast with the MIMO system presented in Figure 2.11, the architecture shown in Figure 2.12 for Chapter 4 considers the extension of the VIVO channel to the diffused-VIVO channel, where each multipath is composed of a cluster of spatially inseparable paths instead of a single path (i.e. *point-source*). The details

of the diffused-VIVO channel model is examined closely in Chapter 4 and it will be shown that the diffused-VIVO channel can be simplified to a form similar to that of the VIVO channel. It can be seen that the transmitter structures in Figures 2.11 and 2.12 are similar. The receiver, based on the diffused-VIVO channel model, performs the estimation of the channel parameters, which includes the DOA, TOA and spatial spread, of the channel. In addition, a decorrelating receiver is introduced for the receiver weight matrix.

When channel state information (CSI) is available at the transmitter, advanced signal processing techniques can be applied at both the transmitter and receiver for jointly optimizing the system performance. The problem under consideration is joint transmitter-receiver (Tx-Rx) beamforming over the VIVO MA-2 (reverse link) channel which will be investigated in Chapter 5. The system architecture is illustrated in Figure 2.13. Figure 2.13 shows an M user cyclic prefix-free MC-CDMA arrayed MIMO downlink system with $\bar{N}$ transmit antennas at the base station and N receive antennas at each receiver terminal. Each user's symbol stream is first demultiplexed into N_i^{sub} substreams $\forall i = 1, \dots, M$. The resultant vector-signals $\underline{a}_i[n] \forall i = 1, \dots, M$, with dimensions $N_i^{\text{sub}} \times 1$ then undergo cyclic prefix-free MC-CDMA modulation before being transmitted together at point-E. The wireless channel is based on the VIVO MA-2 (reverse link) channel model and is composed of a set of VIVO channels, as described in Figure 2.10. At the i^{th} user, the interference is removed from the discretized signal $\underline{x}_i[n]$ (point-G) by applying the receive beamformer $\mathbb{W}_i \forall i = 1, \dots, M$.

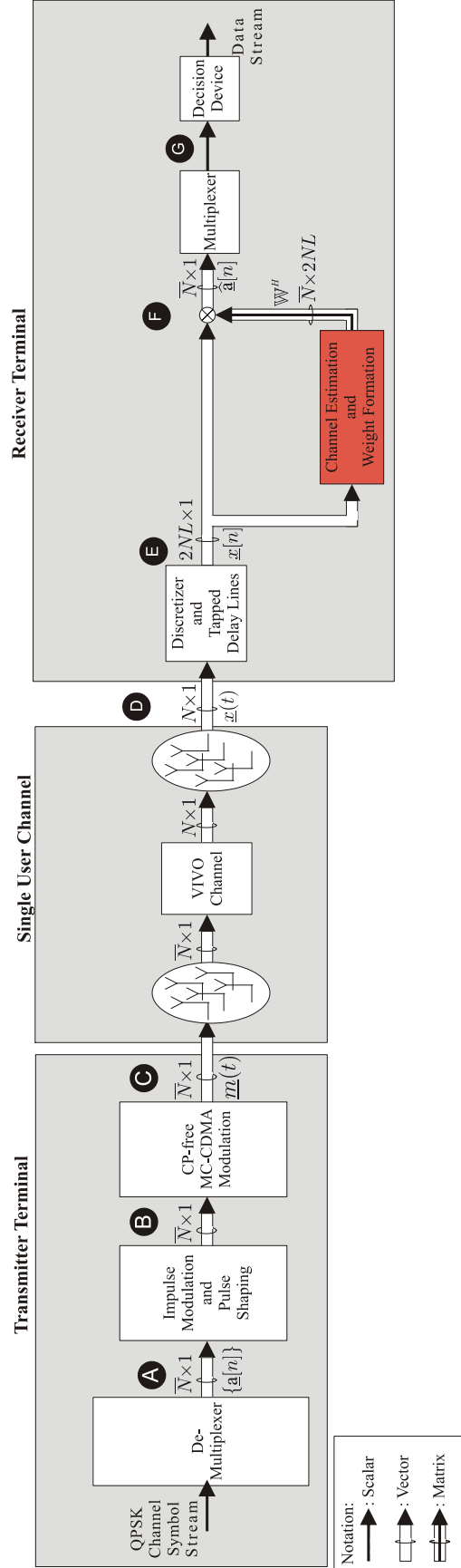


Figure 2.11: System architecture for Chapter 3.

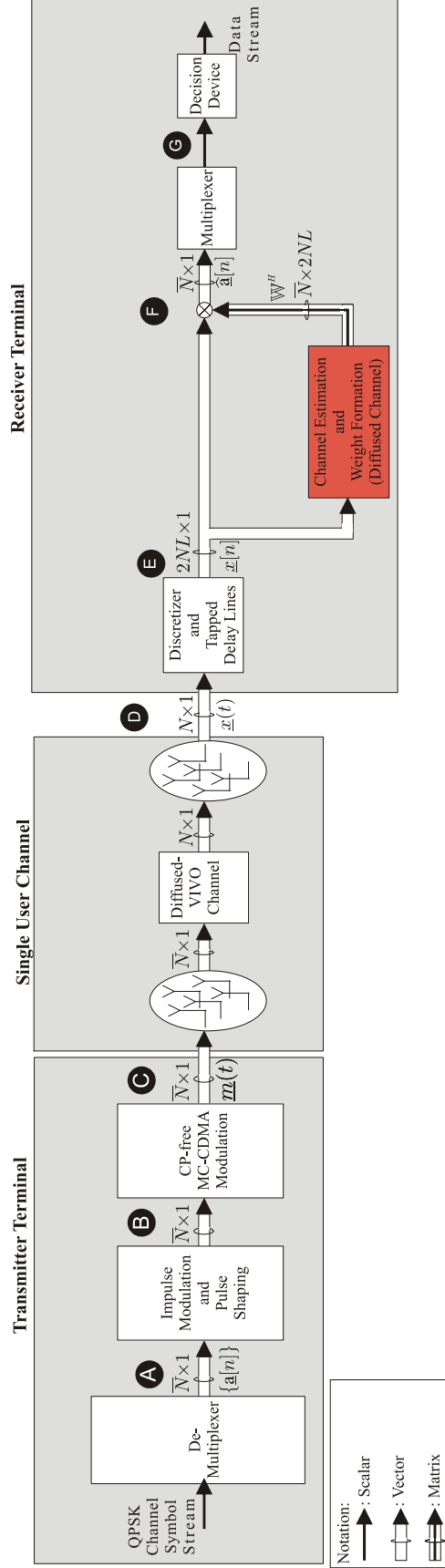


Figure 2.12: System architecture for Chapter 4.

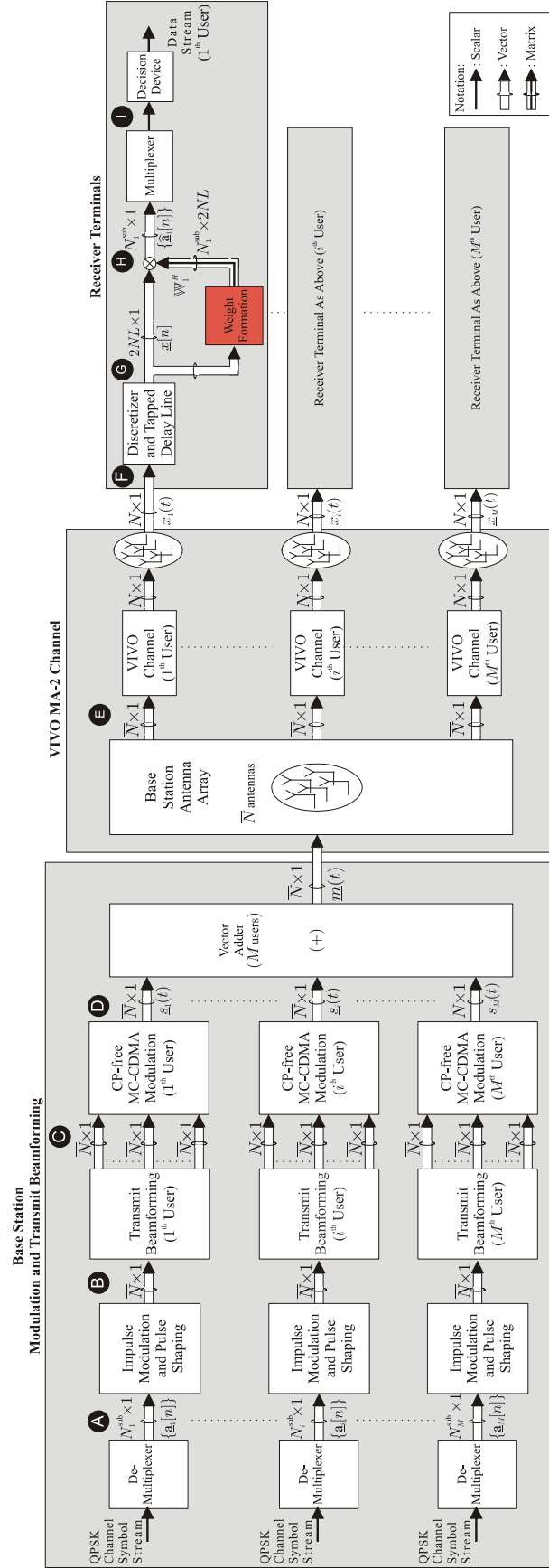


Figure 2.13: System architecture for Chapter 5.

2.3.1 Receiver Front-End Temporal Windowing

It is clear from Figures 2.11, 2.12 and 2.13 that the receivers employ an antenna array of N antenna elements. In order to extend the receiver to include space-time processing capabilities, TDLs are employed at each antenna of the antenna array and this is shown in Figure 2.14. The baseband received signal $\underline{x}(t)$ is observed at point-D (in Figures 2.11 and 2.12) and at point-F (in Figure 2.13) as shown. The asynchronous receiver makes use of a bank of N TDLs, each corresponding to each antenna element, to capture the full contribution of the channel symbol of interest.

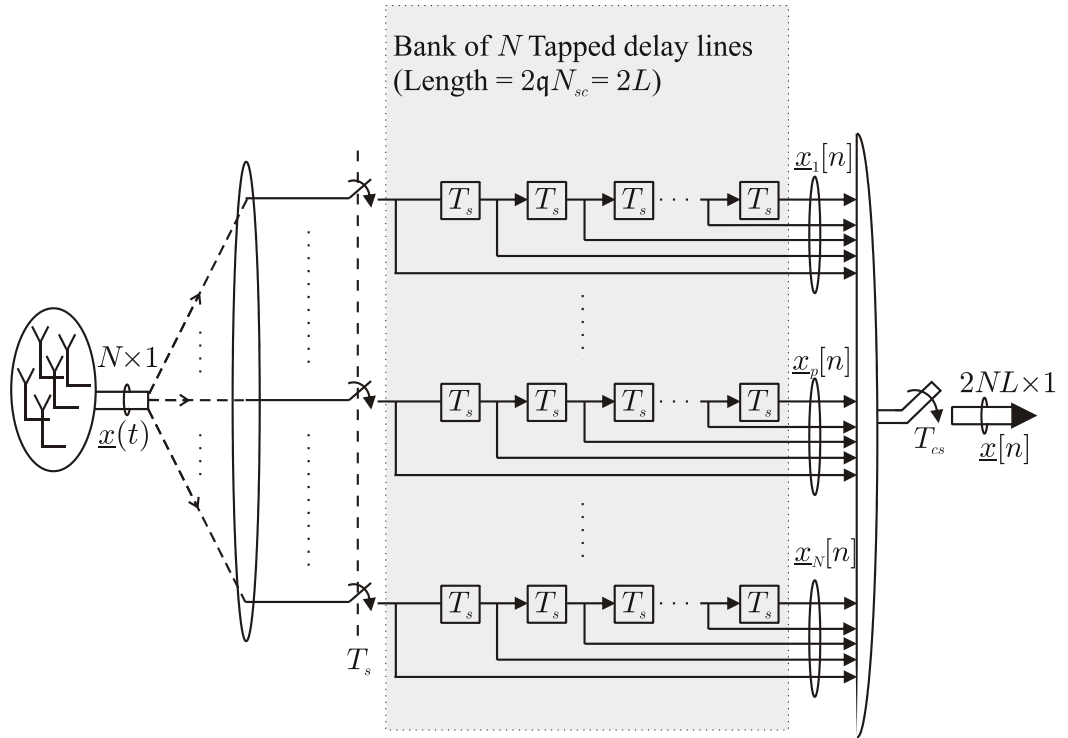


Figure 2.14: Discretizer and tapped delay line (TDL) structure in receiver front-end.

At the receiver front-end, the signal from each antenna is sampled with a rate of $\frac{1}{T_s}$, where $T_s = \frac{T_{cs}}{qN_{sc}}$ and q is the oversampling factor. As such, $L = qN_{sc}$ samples are obtained per channel symbol period T_{cs} . However, without any loss of generality, q is assumed to be 1, i.e. no oversampling is performed, in this

thesis. Hence, $L = N_{sc}$. As a result of sampling, the delay τ_j of the j^{th} path is quantized into integer and fractional components, i.e. $\tau_j = (\ell_j + \psi_j)T_s$, with $\ell_j \in 0, 1, \dots, (N_{sc}-1)$ and $\psi_j \in [0, 1)$. The phase shift due to the fractional part, i.e. $\exp(-j2\pi F_k \psi_j T_s)$, can be absorbed into the complex path fading coefficient β . Thus, it is only necessary to consider the integer component of the delay, i.e. $\ell_j T_s$.

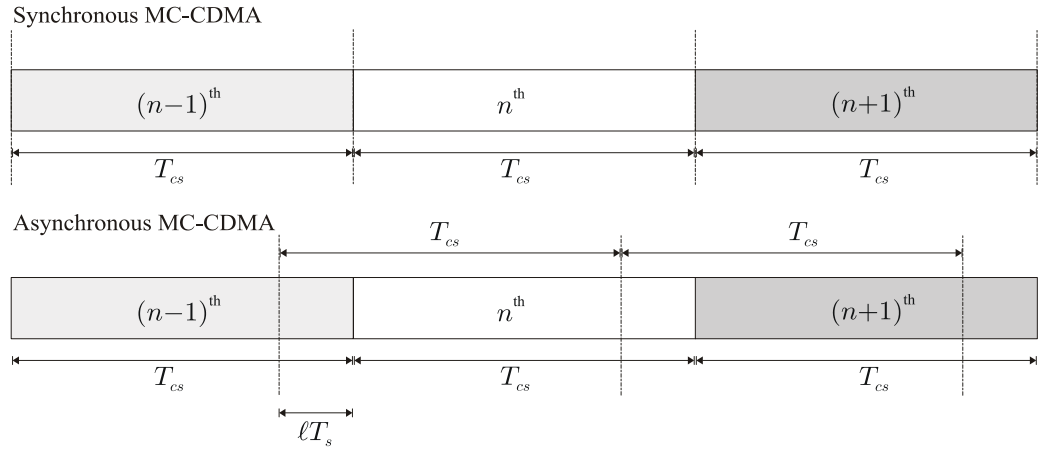


Figure 2.15: Illustration of the channel symbols captured by a tapped delay line (TDL) of length $2T_{cs}$.

A TDL of length $2N_{sc}$, which is equivalent to twice the symbol period T_{cs} , is employed at each antenna of the antenna array, as shown in Figure 2.14. This enables the asynchronous receiver to capture the contribution of a full symbol (current symbol). The choice of the length of the TDL is due to the assumption of delay spread of T_{cs} in the frequency-selective channel. The capturing of a full channel symbol, using a tapped delay line, is shown in Figure 2.15, in comparison with the synchronous receiver. As shown, for the asynchronous receiver, at the n^{th} symbol period, there is a delay of ℓT_s in the received signal with symbol period of T_{cs} . Hence, if the TDL only had a length of T_{cs} , it would not be able to capture the contributions of a full symbol. Thus, a TDL of length equivalent to $2T_{cs}$ is required. However, this also means that the receiver will also capture contributions from the preceding and succeeding channel symbols.

The details of the sampling carried out in the tapped delay lines are also shown in Figure 2.14. Through the use of the TDLs, $2N_{sc}$ samples are obtained from each TDL. The $2N_{sc}$ samples from each tapped delay line for the n^{th} symbol period are then concatenated to form a vector $\underline{x}[n]$ of length $2NN_{sc}$ in Figure 2.14 which is read at a rate of $\frac{1}{T_{cs}}$. Thus, the output of the TDLs can be expressed as

$$\underline{x}[n] = \begin{bmatrix} \underline{x}_1[n] \\ \underline{x}_2[n] \\ \vdots \\ \underline{x}_N[n] \end{bmatrix} \in \mathcal{C}^{2NN_{sc} \times 1} \quad (2.19)$$

where $\underline{x}_p[n]$ denotes the vector of length $2N_{sc}$ from the p^{th} tapped delay line.

2.4 Summary

In this chapter, the concept of the array manifold vector has been introduced. In addition, the array manifold vector is simplified to be independent of the subcarrier frequency for MC modulation schemes. This is based on the assumption that the maximum subcarrier frequency is much smaller than the carrier frequency. Hence, the effect of the subcarrier frequency on the array manifold vector can be ignored.

The mathematical modelling of the space-time fading channel models, which can be classified into four main categories based on the structure of the input or output signals, has also been presented with the channel impulse response derived for each channel type. These have been further extended to the case of multiple access systems. It is shown that the space-time channel plays an important role in transforming the signal from the transmitter-end to the receiver-end. The development of the communication system architecture involves the integration of three basic building blocks: the transmitter, the propagation channels and

the receiver and each of these fundamental units may vary in the designing of a space-time system architecture.

In the frequency-selective fading channel considered in Figures 2.11, 2.12 and 2.13, it has been assumed that the delay spread of the channel is equivalent to the symbol period. This implies that if cyclic prefixes are used in the MC-CDMA modulation, a cyclic prefix of length equal to one channel symbol period will be required. Hence, for the channel considered here, the use of a cyclic prefix will result in a huge loss in signalling time. Thus, cyclic prefix-free MC-CDMA modulation will be used in this thesis in conjunction with arrayed MIMO systems.

Chapter 3

Space-Time Cyclic Prefix-free MC-CDMA Array-Based MIMO System

This chapter addresses the problems of channel estimation and reception for the cyclic prefix-free MC-CDMA MIMO system presented in Figure 2.11, which is reproduced in Figure 3.1 for ease of reference. In order to reduce the system overheads, a blind channel estimation algorithm is of interest as resources that are used for the transmission of pilot signals can be used for the transmission of information instead. A parametric approach is taken in the channel estimation process where the space-time parameters of the channel are estimated. This is in contrast to non-parametric channel estimation techniques where the composite channel response, i.e. the channel matrix in MIMO systems, is estimated instead.

In addition to the joint space-time parameter estimator, linear receivers are also presented based on the model of the array-based MIMO system. The proposed system, as shown in Figure 3.1, makes use of a demultiplexer to distribute the channel symbols among the transmitter antenna elements. In contrast to V-BLAST systems, a linear receiver architecture is favoured over SIC techniques so

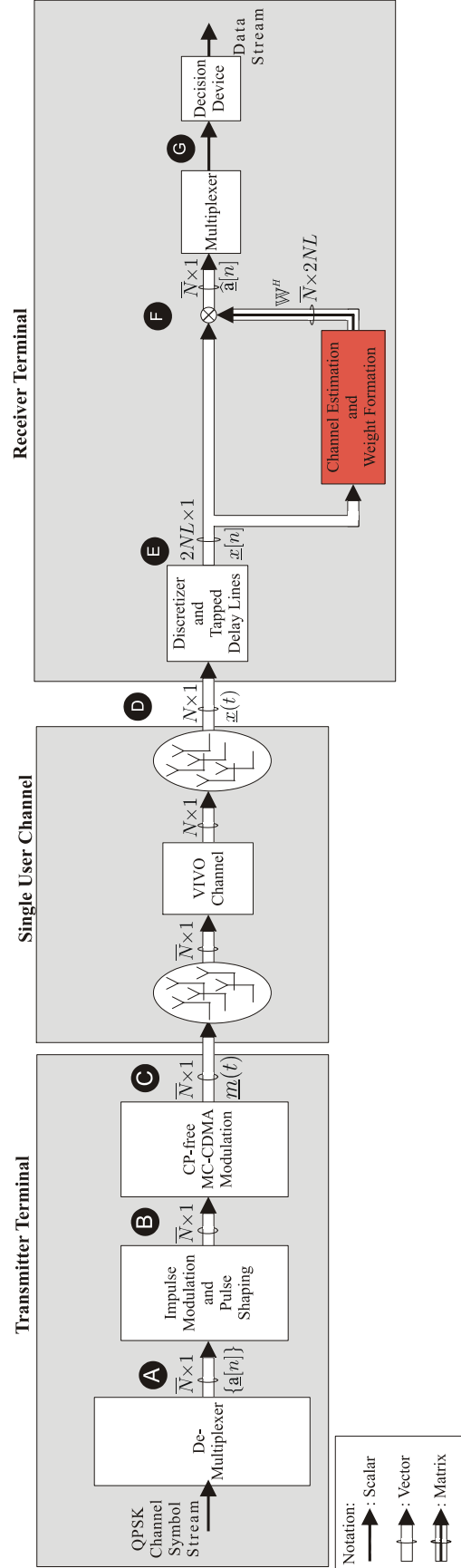


Figure 3.1: System Architecture of a cyclic prefix-free arrayed MIMO communication system (Figure 2.11 which is reproduced here for ease of reference).

that the problem of error propagation can be avoided. The linear weight matrix is applied to the output of the discretizer and tapped delay lines and the result is then multiplexed and passed into the decision device. A subspace-based receiver, which aims to cancel the ISI present in the system, is proposed. Furthermore, the RAKE and MMSE receivers are upgraded to spatio-temporal RAKE and spatio-temporal MMSE, respectively. The effectiveness of the joint estimation process is supported by simulation results and the performance of the proposed subspace-based receiver is compared with that of spatio-temporal RAKE and MMSE receivers.

3.1 Introduction

In MC-CDMA systems, the use of cyclic prefixes to overcome the effects of the frequency-selective fading channel reduces the spectral efficiency of the system. Thus, cyclic prefix-free MC-CDMA systems are of interest. Note that cyclic prefix-free MC-CDMA systems have been investigated also in [39] by Nabhane and Poor for, however, synchronous systems. Furthermore, in [40], a similar cyclic prefix-free MC-CDMA system is considered where, by performing time-domain equalization, it is shown to have a much better performance than the conventional frequency-domain equalization.

Due to the sensitivity of MC-CDMA systems to timing and frequency offsets, channel estimation is an integral part of a MC-CDMA system and this can be carried out with blind, or non-blind, techniques. Non-blind techniques, as the name implies, make use of pilot or training symbols to obtain the estimates of the channel. An example of such pilot-assisted channel estimation for a MC-CDMA MIMO system can be found in [41] where the channel estimation of the channel matrix is viewed as an extension of the SISO channel estimator for each

receive antenna.

However, as a result of the demand for high data rates in future communications for 4G and beyond, the reduction of system overheads is one of the objectives of this work. Thus, in the channel estimation for MC-CDMA MIMO systems, which is a likely candidate for 4G systems, semi-blind or blind channel techniques are an attractive alternative as such techniques reduce or remove the need for pilot or training symbols to be used for the estimation of the channel.

A number of blind channel estimation techniques have been proposed for MC-CDMA MIMO systems and subspace-based methods are a common feature of these blind channel estimation techniques. In [42], blind channel estimation techniques, based on eigendecomposition, have been proposed for the estimation of the channel impulse response in a MC-CDMA MIMO system which make use of STBCs. However, due to the subspace-based method used, a scalar ambiguity exists in the estimate of the CIR. In [43], a semi-blind subspace-based method was proposed to overcome the problem of scalar ambiguity. The proposed method requires some pilot symbols in order to resolve the scalar ambiguity inherent in subspace-based methods, but the number of pilot symbols used is much less than methods which are entirely pilot-based. It is shown that the proposed method is capable of resolving the scalar ambiguity through the use of only one training symbol. It must be noted that the systems proposed in [42, 43] are not array-based systems and thus, array signal processing techniques are not available in such systems. In contrast, in [44], a uniform linear array is implemented at the receiver. Thus, in addition to the CIR, the DOA of each user, as observed at the linear array at the receiver, is also estimated through the use of an ESPRIT-like algorithm. The obtained DOA information can then be used for beamforming to improve the performance of the system.

Due to the effectiveness of subspace-based channel estimation algorithms, the

proposed channel estimation technique is also subspace-based. However, instead of estimating the CIR of the MIMO channel, the channel parameters are estimated instead. The receiver terminal is thus designed to obtain estimates of the DOAs and TOAs of the multipaths impinging on its antenna array. The knowledge of these channel parameters can then be used in the design of the receiver for improved performance.

3.2 Cyclic Prefix-free MC-CDMA Arrayed MIMO Transmitter

The structure of the transmitter in a cyclic prefix-free MC-CDMA arrayed MIMO system is illustrated in Figure 3.2. As shown in this figure, the transmitter is equipped with $\overline{N}$ antenna elements. It is assumed that channel information is not available at the transmitter. Thus, the incoming channel symbols are demultiplexed among the transmitter antenna elements for transmission. The channel symbol stream is demultiplexed into $\overline{N}$ channel symbol substreams to obtain the n^{th} channel symbol vector length $\overline{N}$ given by $\{\underline{a}[n] \in \mathcal{C}^{N \times 1}, \forall n \in \mathcal{Z}\}$ and $\underline{a}[n] = [a_1[n], \dots, a_p[n], \dots, a_{\overline{N}}[n]]^T$ where each component $\{a_p[n], p \in 1, \dots, \overline{N}\}$, has a symbol period of T_{cs} . The vector sequence $\{\underline{a}[n], \forall n\}$ at point-A is then converted, through impulse modulation and pulse shaping, into the vector waveform, observed at point-B, which is defined as:

$$\underline{a}[n]c(t - nT_{cs}), \quad nT_{cs} \leq t < (n+1)T_{cs} \quad (3.1)$$

where $c(t)$ is the rectangular pulse of duration T_{cs} .

The signal vector, given by Equation 3.1, is then MC-CDMA modulated, as indicated in Figure 3.2. Due to the use of code diversity in the transmitter [45,46], each p^{th} element of Equation 3.1 is assigned an unique PN sequence $\underline{a}_p$ of length

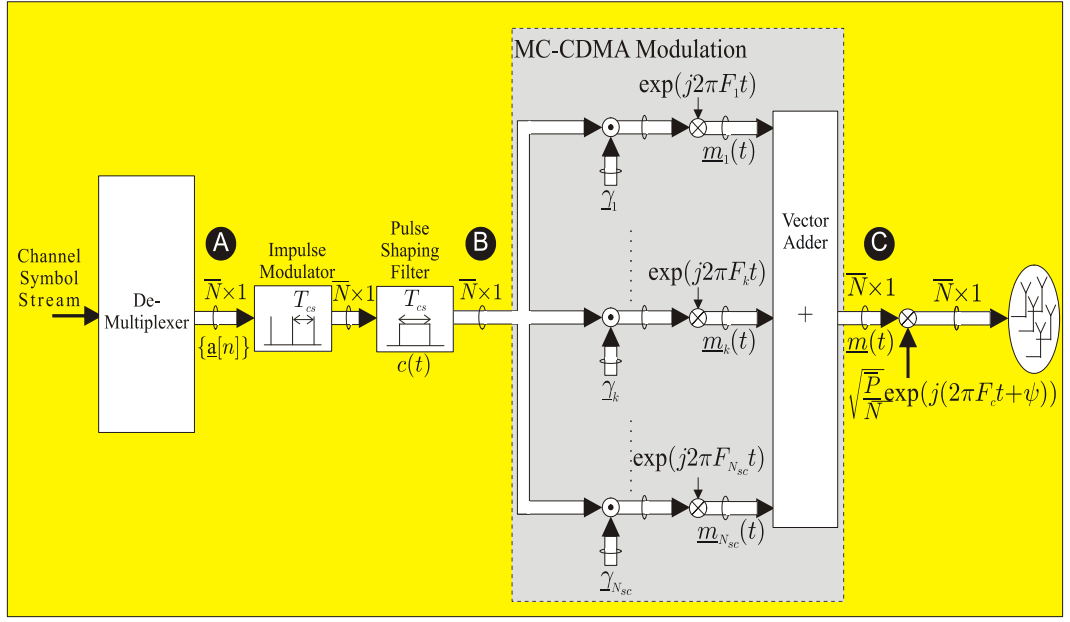


Figure 3.2: Structure of the transmitter in a cyclic prefix-free MC-CDMA arrayed MIMO system (transmitter terminal in Figure 3.1).

N_{sc} , i.e. $\underline{\alpha}_p = [\alpha_{p1}, \dots, \alpha_{pk}, \dots, \alpha_{pN_{sc}}]^T$ with $\alpha_{pk} = \pm 1$, $k \in 1, \dots, N_{sc}$. By employing code diversity, the $\bar{N}$ signals emitted from the transmitter terminal can be differentiated based on prior knowledge of the PN sequences used.

In the MC-CDMA modulation that follows, the vector-signal $\underline{a}[n]c(t - nT_{cs})$ is first copied into N_{sc} parallel streams. The p^{th} component of the k^{th} copy is then multiplied by a corresponding k^{th} chip of $\underline{\alpha}_p$. A subcarrier with frequency given by $F_k = k\Delta F$ is then used to modulate the resultant vectors, where $\Delta F = 1/T_{cs}$ is the subcarrier separation. The modulated subcarriers are then summed up to produce the MC-CDMA $(\bar{N} \times 1)$ vector signal $\underline{m}(t)$ at point-C, which is given by

$$\underline{m}(t) = \sum_{k=1}^{N_{sc}} \underline{m}_k(t) \quad (\bar{N} \times 1) \quad (3.2)$$

where

$$\underline{m}_k(t) = \underline{\gamma}_k \odot \underline{a}[n] \exp(j2\pi F_k t) \quad (3.3)$$

and

$$\underline{\gamma}_k = [\alpha_{1k}, \dots, \alpha_{pk}, \dots, \alpha_{\bar{N}k}]^T \quad (3.4)$$

with $\underline{\gamma}_k$ representing a $\bar{N} \times 1$ vector with elements of the k^{th} chips of the PN sequences $\underline{\alpha}_p \forall p$. The MC code matrix $\mathbb{C}_{\text{MC}} \in \mathcal{R}^{N_{sc} \times \bar{N}}$ for the user terminal is defined as

$$\mathbb{C}_{\text{MC}} \triangleq \begin{bmatrix} \alpha_{11}, & \alpha_{21}, & \cdots & \alpha_{p1}, & \cdots & \alpha_{\bar{N}1} \\ \alpha_{12}, & \alpha_{22}, & \cdots & \alpha_{p2}, & \cdots & \alpha_{\bar{N}2} \\ \vdots & \vdots & \ddots & \vdots & \ddots & \vdots \\ \alpha_{1k}, & \alpha_{2k}, & \cdots & \alpha_{pk}, & \cdots & \alpha_{\bar{N}k} \\ \vdots & \vdots & \ddots & \vdots & \ddots & \vdots \\ \alpha_{1N_{sc}}, & \alpha_{2N_{sc}}, & \cdots & \alpha_{pN_{sc}}, & \cdots & \alpha_{\bar{N}N_{sc}} \end{bmatrix} \triangleq \underline{\gamma}_k^T \quad (3.5)$$

$\triangleq \underline{\alpha}_p$

and the correspondence between the elements of $\mathbb{C}_{\text{MC}}$ and the PN sequences $\underline{\gamma}_k, \forall k$ and $\underline{\alpha}_p, \forall p$, is shown in Equation 3.5.

It can be seen from Equation 3.2 that neither cyclic prefixes nor guard intervals are used in the generation of the MC-CDMA symbol. The baseband vector-signal $\underline{m}(t)$ at point-C is then upconverted to a carrier frequency F_c prior to transmission, given by

$$\sqrt{\frac{\bar{P}}{N}} \underline{m}(t) \exp(j2\pi F_c t + \psi)$$

where ψ is a random phase offset relative to the receiver and $\bar{P}$ is the total transmitted power. Due to a lack of channel information, the transmitter power is distributed equally across the $\bar{N}$ transmitter antennas and so, the power from each transmitter element is given by $\frac{\bar{P}}{N}$. In this thesis, without any loss of generality, baseband transmission is assumed and thus, the carrier $\exp(j2\pi F_c t)$ at the transmitter and $\exp(-j2\pi F_c t)$ at the receiver are ignored.

3.3 Received Signal

In the VIVO space-time channel in Figure 3.1, it is assumed that there are K multipaths from the transmitter to the receiver. Thus, the received signal is given

by

$$\underline{x}(t) = \sum_{j=1}^K \beta_j \underline{S}_j \overline{S}_j^H \underline{m}(t - \tau_j) + \underline{n}(t) \quad (3.6)$$

where $\underline{n}(t)$ represents the Additive White Gaussian Noise (AWGN) of power σ_n^2 present in the channel that is observed by the receiver.

3.3.1 MC-STAR Manifold Matrix

Due to the sampling performed at a rate of $\frac{1}{T_s} = \frac{N_{sc}}{T_{cs}}$ between points D and E in Figure 3.1, the subcarrier $\exp(j2\pi F_k(t - \tau_j))$ can be represented by the subcarrier vector

$$\underline{f}_k[\ell] = \begin{bmatrix} \exp(j2\pi F_k(-\ell)T_s) \\ \exp(j2\pi F_k(1-\ell)T_s) \\ \vdots \\ \exp(j2\pi F_k(N_{sc}-1-\ell)T_s) \end{bmatrix}$$

which models the time variation of the k^{th} subcarrier after undergoing a path delay of ℓT_s . In addition, the following matrix

$$\mathbb{J} = \begin{bmatrix} \underline{0}_{2N_{sc}-1}^T & 0 \\ \mathbb{I}_{2N_{sc}-1} & \underline{0}_{2N_{sc}-1} \end{bmatrix} \quad (3.7)$$

is used to model the delay ℓ on the column vector $\begin{bmatrix} \underline{f}_k^T[\ell] & \underline{0}_{N_{sc}}^T \end{bmatrix}^T$, where $\underline{0}_{N_{sc}}$ is appended to the subcarrier vector $\underline{f}_k$ in order to extend the length of the subcarrier vector $\underline{f}_k$ to $2N_{sc}$ (which is the length of each TDL in the receiver). For instance, if $\mathbb{J}^\ell$ pre-multiplies $\begin{bmatrix} \underline{f}_k^T[\ell] & \underline{0}_{N_{sc}}^T \end{bmatrix}^T$, it downshifts the elements of $\begin{bmatrix} \underline{f}_k^T[\ell] & \underline{0}_{N_{sc}}^T \end{bmatrix}^T$ by ℓ elements. On the other hand, pre-multiplying $\begin{bmatrix} \underline{f}_k^T[\ell] & \underline{0}_{N_{sc}}^T \end{bmatrix}^T$ with $(\mathbb{J}^T)^\ell$ upshifts the column vector by ℓ elements.

At point-D in Figure 3.1, the received signal is given by

$$\underline{x}(t) = \sum_{j=1}^K \beta_j \underline{S}_j \overline{S}_j^H \sum_{k=1}^{N_{sc}} \left(\underline{\gamma}_k \odot \underline{a}[n] \exp(j2\pi F_k(t - \tau_j)) \right) \quad (3.8)$$

Thus, from Equation 3.8, the discretized and concatenated received vector-signal $\underline{x}[n]$, at point-E in Figure 3.1, can be written as

$$\begin{aligned}
\underline{x}[n] &= \sum_{j=1}^K \left(\beta_j \underline{S}_j \otimes \left(\sum_{k=1}^{N_{sc}} \left(\mathbb{J}^\ell \begin{bmatrix} \underline{f}_k[\ell_j] \\ \underline{0}_{N_{sc}} \end{bmatrix} \underline{\gamma}_k^T \right) \right) \text{diag} \left(\overline{\underline{S}}_j^H \right) \underline{a}[n] \right. \\
&\quad + \beta_j \underline{S}_j \otimes (\mathbb{J}^T)^{N_{sc}} \left(\sum_{k=1}^{N_{sc}} \left(\mathbb{J}^\ell \begin{bmatrix} \underline{f}_k[\ell_j] \\ \underline{0}_{N_{sc}} \end{bmatrix} \underline{\gamma}_k^T \right) \right) \text{diag} \left(\overline{\underline{S}}_j^H \right) \underline{a}[n-1] \\
&\quad \left. + \beta_j \underline{S}_j \otimes \mathbb{J}^{N_{sc}} \left(\sum_{k=1}^{N_{sc}} \left(\mathbb{J}^\ell \begin{bmatrix} \underline{f}_k[\ell_j] \\ \underline{0}_{N_{sc}} \end{bmatrix} \underline{\gamma}_k^T \right) \right) \text{diag} \left(\overline{\underline{S}}_j^H \right) \underline{a}[n+1] \right) + \underline{n}[n] \\
&= \sum_{j=1}^K \left(\beta_j \left(\underline{S}_j \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \mathbb{C}_{MC} \\ \mathbb{O}_{N_{sc} \times \overline{N}} \end{bmatrix} \text{diag} \left(\overline{\underline{S}}_j^H \right) \right) \underline{a}[n] \right. \\
&\quad + \beta_j \left(\mathbb{I}_N \otimes (\mathbb{J}^T)^{N_{sc}} \right) \left(\underline{S}_j \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \mathbb{C}_{MC} \\ \mathbb{O}_{N_{sc} \times \overline{N}} \end{bmatrix} \text{diag} \left(\overline{\underline{S}}_j^H \right) \right) \underline{a}[n-1] \\
&\quad \left. + \beta_j \left(\mathbb{I}_N \otimes \mathbb{J}^{N_{sc}} \right) \left(\underline{S}_j \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \mathbb{C}_{MC} \\ \mathbb{O}_{N_{sc} \times \overline{N}} \end{bmatrix} \text{diag} \left(\overline{\underline{S}}_j^H \right) \right) \underline{a}[n+1] \right) + \underline{n}[n] \\
&= \sum_{j=1}^K \left(\beta_j \mathbb{H}_j \underline{a}[n] \right. \\
&\quad + \left(\mathbb{I}_N \otimes (\mathbb{J}^T)^{N_{sc}} \right) \beta_j \mathbb{H}_j \underline{a}[n-1] \\
&\quad \left. + \left(\mathbb{I}_N \otimes \mathbb{J}^{N_{sc}} \right) \beta_j \mathbb{H}_j \underline{a}[n+1] \right) + \underline{n}[n] \tag{3.9}
\end{aligned}$$

The matrix $\mathbb{C}_{MC}$ is as defined in Equation 3.5 and

$$\mathbb{F}_{\ell_j} = [\underline{f}_1[\ell_j], \dots, \underline{f}_k[\ell_j], \dots, \underline{f}_{N_{sc}}[\ell_j]] \in \mathcal{C}^{N_{sc} \times N_{sc}}$$

Also, in Equation 3.9,

$$\mathbb{H}_j = \underline{S}_j \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \mathbb{C}_{MC} \\ \mathbb{O}_{N_{sc} \times \overline{N}} \end{bmatrix} \text{diag} \left(\overline{\underline{S}}_j^H \right) \tag{3.10}$$

is defined as the multi-carrier space-time array (MC-STAR) manifold matrix of the j^{th} path, which enables the modelling of both the spatial and the discretized

temporal characteristics of a particular signal path for MC-CDMA arrayed MIMO systems. Note that β_j represents the complex fading coefficient of the j^{th} path from the reference point of the transmitter to the reference point of the receiver while $\beta_j \bar{\underline{S}}_j$ represents the fading coefficient from the $\bar{N}$ transmitter elements to the reference point of the receiver antenna array. Furthermore, the matrix $\beta_j \underline{S}_j \bar{\underline{S}}_j^H$ represents the complex fading coefficients from all transmitter antennas to all receiver antennas. In Equation 3.10, it can be seen that the frequency-domain spreading in MC-CDMA system is transformed by the IDFT into a ‘time-domain’ spreading with a IDFT transformed path delay-dependent spreading matrix given by $\mathbb{F}_{\ell_j} \mathbb{C}_{\text{MC}}$. Thus, with the time-domain spreading representation [40,47], channel estimation and reception techniques that are used for DS-CDMA systems can also be applied to MC-CDMA systems. Indeed, it has been shown in [48] that the use of time-domain processing for MC-CDMA systems have produced better results than frequency-domain processing.

Finally, in the common representation for MIMO systems,

$$\underline{x}[n] = \mathbb{H}_{\text{desired}} \underline{a}[n] + \mathbb{H}_{\text{ISI}}^{\text{prev}} \underline{a}[n-1] + \mathbb{H}_{\text{ISI}}^{\text{next}} \underline{a}[n+1] + \underline{n}[n] \quad (3.11)$$

where

$$\mathbb{H}_{\text{desired}} = \sum_{j=1}^K \beta_j \mathbb{H}_j \in \mathcal{C}^{2NN_{sc} \times \bar{N}} \quad (3.12)$$

and $\mathbb{H}_{\text{ISI}}^{\text{prev}}$ and $\mathbb{H}_{\text{ISI}}^{\text{next}}$ are defined as

$$\mathbb{H}_{\text{ISI}}^{\text{prev}} = \left(\mathbb{I}_N \otimes (\mathbb{J}^T)^{N_{sc}} \right) \mathbb{H}_{\text{desired}} \quad (3.13a)$$

$$\mathbb{H}_{\text{ISI}}^{\text{next}} = \left(\mathbb{I}_N \otimes \mathbb{J}^{N_{sc}} \right) \mathbb{H}_{\text{desired}} \quad (3.13b)$$

Thus, it can be observed that the MIMO system represented by Equation 3.11 encompasses the ISI effects due to the previous and next symbols.

3.4 Receiver Weights

The problem of detecting the user's data symbol(s) from the received signal vector $\underline{x}[n]$ is now addressed. In the n^{th} received signal vector, $\underline{x}[n]$, there are 3 data symbol vectors, each corresponding to the $(n-1)^{\text{th}}$, n^{th} and $(n+1)^{\text{th}}$ symbol periods, respectively, and each data symbol vector contributes $\overline{N}$ channel symbols to $\underline{x}[n]$ due to the demultiplexing performed in the transmitter. This is shown in Equation 3.11.

In the following, the RAKE and MMSE receivers are described and explained and a subspace-based receiver is also proposed.

3.4.1 RAKE Receiver

The RAKE receiver, originally proposed by Price and Green [49], is a multipath diversity receiver based on correlation or matched filter receiver which aims to maximize the output signal-to-noise ratio by collecting delayed copies of the required signals. Based on the proposed arrayed MIMO system model, the RAKE receiver matrix can be expressed as

$$\mathbb{W}_{\text{RAKE}} = \hat{\mathbb{H}}_{\text{desired}} \in \mathcal{C}^{2NN_{sc} \times \overline{N}} \quad (3.14)$$

where $\hat{\mathbb{H}}_{\text{desired}}$ denotes the estimate of the channel matrix $\mathbb{H}_{\text{desired}}$. The RAKE receiver is a single user receiver and so, it is sensitive to co-channel interference (CCI), for instance, near-far problems such as that arising from imperfect power control. This can be overcome by the decorrelating (or zero-forcing) receiver. However, instead of the decorrelating receiver, the MMSE receiver, which is described in the next section, is implemented due to the ability of the MMSE receiver in compensating for the AWGN present.

3.4.2 MMSE Receiver

The MMSE linear receiver can be viewed as a solution which takes into account the relative importance of each interfering signal and the background noise present in the channel. This is in contrast to both the single-user matched filter receiver which is designed to cancel the background white noise as well as the decorrelating receiver which ignores the effect of the background noise while attempting to eliminate the interference present.

The MMSE receiver, as the name implies, minimizes the output mean square error (MSE) and it is obtained through the following optimization problem:

$$\mathbf{W} = \arg \min_{\mathbf{W}} \text{tr}(\mathbf{E}) \quad (3.15)$$

The matrix $\mathbf{E}$ denotes the MSE matrix, which is given by

$$\begin{aligned} \mathbf{E} &= \mathcal{E} \left\{ (\underline{\mathbf{a}}[n] - \mathbf{W}^H \underline{\mathbf{x}}[n]) (\underline{\mathbf{a}}[n] - \mathbf{W}^H \underline{\mathbf{x}}[n])^H \right\} \\ &= \mathbb{I}_N - \mathbf{W}^H \mathbf{H}_{\text{desired}} - \mathbf{H}_{\text{desired}}^H \mathbf{W} + \mathbf{W}^H \mathbf{R}_{xx} \mathbf{W} \end{aligned} \quad (3.16)$$

where it has been assumed that $\underline{\mathbf{a}}[n-1]$, $\underline{\mathbf{a}}[n]$ and $\underline{\mathbf{a}}[n+1]$ are uncorrelated and $\mathbf{R}_{xx}$ denotes the data covariance matrix, which is given by

$$\mathbf{R}_{xx} = \mathbf{H} \mathbf{H}^H + \sigma_n^2 \mathbb{I}_{2NN_{sc}}$$

The differentiation of $\text{tr}(\mathbf{E})$ with respect to $\mathbf{W}$ and equating the result to 0 results in

$$2\mathbf{R}_{xx} \mathbf{W} - 2\mathbf{H}_{\text{desired}} = 0$$

Thus, the closed form solution to the above optimization problem is then given by

$$\begin{aligned} \mathbf{W} &= \mathbf{R}_{xx}^{-1} \mathbf{H}_{\text{desired}} \\ &= (\mathbf{H} \mathbf{H}^H + \sigma_n^2 \mathbb{I}_{2NN_{sc}})^{-1} \mathbf{H}_{\text{desired}} \end{aligned} \quad (3.17)$$

The MMSE receiver is then described as follows

$$\mathbb{W}_{\text{MMSE}} = \left(\hat{\mathbb{H}}\hat{\mathbb{H}}^H + \sigma_n^2 \mathbb{I}_{2NN_{sc}} \right)^{-1} \hat{\mathbb{H}}_{\text{desired}} \in \mathcal{C}^{2NN_{sc} \times \overline{N}} \quad (3.18)$$

where

$$\hat{\mathbb{H}} = \begin{bmatrix} \hat{\mathbb{H}}_{\text{ISI}}^{\text{prev}}, & \hat{\mathbb{H}}_{\text{desired}}, & \hat{\mathbb{H}}_{\text{ISI}}^{\text{next}} \end{bmatrix} \quad (3.19)$$

Thus, the pre-multiplication of $\underline{x}[n]$ by $\mathbb{W}_{\text{MMSE}}^H$ will produce the decision variable vector, $\hat{\underline{a}}[n]$, for the current symbol vector, $\underline{a}[n]$.

3.4.3 Subspace-Based Receiver

In addition to the RAKE and MMSE receivers described earlier, a subspace-based receiver is now introduced. The proposed subspace-based receiver, which was first introduced in [50], is designed with the aim of suppressing the interference, i.e. ISI, present in the received signal, $\underline{x}[n]$, of the proposed MIMO system. Thus, the proposed method seeks to project the ISI components in $\underline{x}[n]$ onto their complementary subspace. This is achieved through the use of the ISI complement projection matrix, $\mathbb{P}_{\text{ISI}}^\perp$, which is defined as

$$\mathbb{P}_{\text{ISI}}^\perp = \mathbb{I}_{2NN_{sc}} - \hat{\mathbb{H}}_{\text{ISI}} \left(\hat{\mathbb{H}}_{\text{ISI}}^H \hat{\mathbb{H}}_{\text{ISI}} \right)^{-1} \hat{\mathbb{H}}_{\text{ISI}}^H \quad (3.20)$$

where $\hat{\mathbb{H}}_{\text{ISI}}$ denotes the estimate of the channel matrix of the ISI signals and it is given by

$$\hat{\mathbb{H}}_{\text{ISI}} = \begin{bmatrix} \hat{\mathbb{H}}_{\text{ISI}}^{\text{prev}}, & \hat{\mathbb{H}}_{\text{ISI}}^{\text{next}} \end{bmatrix} \quad (3.21)$$

$\hat{\mathbb{H}}_{\text{ISI}}$, or equivalently $\hat{\mathbb{H}}_{\text{ISI}}^{\text{prev}}$ and $\hat{\mathbb{H}}_{\text{ISI}}^{\text{next}}$, can be easily obtained from the estimate of the channel matrix ($\hat{\mathbb{H}}_{\text{desired}}$) of the current symbol period based on the relations given in Equations 3.13 for the MIMO system. Thus, $\mathbb{P}_{\text{ISI}}^\perp$, as defined by Equation 3.20, is a complement projection matrix of $\hat{\mathbb{H}}_{\text{ISI}}$. The subspace-based receiver weight matrix is then defined as

$$\mathbb{W}_{\text{subspace}} = \mathbb{P}_{\text{ISI}}^\perp \hat{\mathbb{H}}_{\text{desired}} \left(\hat{\mathbb{H}}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \hat{\mathbb{H}}_{\text{desired}} \right)^{-1} \in \mathcal{C}^{2NN_{sc} \times \overline{N}} \quad (3.22)$$

In Equation 3.22, $\mathbb{P}_{\text{ISI}}^\perp$ projects the estimated $\hat{\mathbb{H}}_{\text{desired}}$ to be orthogonal to the ISI subspace, thereby suppressing the ISI in the decision variable vector, $\hat{\mathbf{a}}[n]$.

Using the subspace-based weight matrix, the decision variable vector $\hat{\mathbf{a}}[n]$ is then given by

$$\begin{aligned}
\hat{\mathbf{a}}[n] &= \mathbb{W}_{\text{subspace}}^H \underline{\mathbf{x}}[n] \\
&= \overbrace{\left(\mathbb{H}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \mathbb{H}_{\text{desired}} \right)^{-1} \mathbb{H}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \mathbb{H}_{\text{desired}} \underline{\mathbf{a}}[n]}^{\mathbb{I}_{\bar{N}}} \\
&\quad + \overbrace{\left(\mathbb{H}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \mathbb{H}_{\text{desired}} \right)^{-1} \mathbb{H}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \mathbb{H}_{\text{ISI}}^{\text{prev}} \underline{\mathbf{a}}[n-1]}^{\mathbb{O}_{\bar{N} \times \bar{N}}} \\
&\quad + \overbrace{\left(\mathbb{H}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \mathbb{H}_{\text{desired}} \right)^{-1} \mathbb{H}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \mathbb{H}_{\text{ISI}}^{\text{next}} \underline{\mathbf{a}}[n+1]}^{\mathbb{O}_{\bar{N} \times \bar{N}}} \\
&\quad + \left(\mathbb{H}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \mathbb{H}_{\text{desired}} \right)^{-1} \mathbb{H}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \underline{\mathbf{n}}[n] \\
&= \underline{\mathbf{a}}[n] + \left(\mathbb{H}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \mathbb{H}_{\text{desired}} \right)^{-1} \mathbb{H}_{\text{desired}}^H \mathbb{P}_{\text{ISI}}^\perp \underline{\mathbf{n}}[n] \tag{3.23}
\end{aligned}$$

where it has been assumed that the perfect knowledge of the matrix $\mathbb{H}_{\text{desired}}$ is available. Moreover, it can be observed that the ISI complement projection matrix, $\mathbb{P}_{\text{ISI}}^\perp$, operates directly on the matrix $\mathbb{H}_{\text{desired}}$ and so, the multiplication of $\underline{\mathbf{x}}[n]$ by $\mathbb{W}_{\text{subspace}}$ has the effect of nulling the effects of the ISI matrices. It can be observed in Equation 3.23 that the subspace-based receiver enables the cancellation of the ISI effects from the received signal.

3.5 Space-Time Parameter Estimation for Asynchronous Multipath Propagation

It can be seen from the previous section in Equations 3.14, 3.18 and 3.22 that the knowledge of the matrices $\mathbb{H}_{\text{desired}}$, $\hat{\mathbb{H}}_{\text{ISI}}^{\text{prev}}$ and $\hat{\mathbb{H}}_{\text{ISI}}^{\text{next}}$ are needed for the formulation of the receiver weight matrices. This implies that the estimation of the channel parameters is required for the reconstruction of the matrices $\mathbb{H}_{\text{desired}}$, $\hat{\mathbb{H}}_{\text{ISI}}^{\text{prev}}$ and

$\widehat{\mathbb{H}}_{\text{ISI}}^{\text{next}}$. In this section, a channel estimation technique is proposed for the estimation of the channel parameters, in particular, the DOA and TOA. In order to obtain the estimates of the DOA and TOA, knowledge of the PN-sequence of one of the elements of the transmitter antenna array is required.

The received signal vector, from Equation 3.11, can be rewritten as

$$\begin{aligned} \underline{x}[n] = & \underbrace{\sum_{j=1}^K \beta_j [\underline{S}_j]_{\text{p}}^* \underline{h}_{pj} a_p[n]}_{\text{TOI}} + \underbrace{\sum_{p=1, p \neq \text{p}}^{\overline{N}} \sum_{j=1}^K \beta_j [\underline{S}_j]_p^* \underline{h}_{pj} a_p[n]}_{\text{CCI}} \\ & + \underbrace{\mathbb{H}_{\text{ISI}}^{\text{prev}} \underline{a}[n-1] + \mathbb{H}_{\text{ISI}}^{\text{next}} \underline{a}[n+1]}_{\text{ISI}} + \underline{n}[n] \end{aligned} \quad (3.24)$$

where it is assumed that the p^{th} transmitter antenna is the antenna whose PN sequence will be used for the estimation of the channel parameters. In Equation 3.24, $[\underline{S}_j]_{\text{p}}^* \underline{h}_{pj}$ corresponds to the p^{th} column of $\mathbb{H}_j$, where the vector $\underline{h}_{pj}$ is given by

$$\underline{h}_{pj} = \underline{S}_j \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \alpha_p \\ \underline{0}_{N_{sc}} \end{bmatrix} \in \mathcal{C}^{2NN_{sc} \times 1} \quad (3.25)$$

The vector $\underline{h}_{pj}$ is referred to as the MC-STAR manifold vector for the j^{th} path of the p^{th} antenna element.

It can be observed from Equation 3.24 that the first term is the term of interest (TOI), the second term represents the co-channel interference (CCI) due to the remaining $(\overline{N} - 1)$ transmitter antennas while the third and fourth terms represent the ISI effects and AWGN, respectively. In the desired term, the MC-STAR manifold vectors corresponding to the K paths of the p^{th} transmitter antenna are shown to be linearly combined by the fading coefficient vector β_j and the complex conjugate of the p^{th} element of the transmit array manifold vector for the j^{th} path, $[\underline{S}_j]_{\text{p}}^*$. Thus, the p^{th} transmitter contributes a desired signal subspace of only one dimension to the overall signal subspace. As a result, it is not possible to estimate the p^{th} transmitter antenna's space-time channel parameters (DOAs and TOAs) using signal subspace techniques such as MUSIC [33].

A solution, which makes use of the concept of the preprocessor scheme introduced in [51], is now proposed to overcome the ‘coherence’ problem caused by the linear combination of the MC-STAR manifold vectors. A matrix is first defined as:

$$\mathbb{C}_p = \left[\mathbb{J}^0 \begin{bmatrix} \mathbb{F}_0 \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix}, \mathbb{J}^1 \begin{bmatrix} \mathbb{F}_1 \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix}, \dots, \dots, \mathbb{J}^{N_{sc}-1} \begin{bmatrix} \mathbb{F}_{N_{sc}-1} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix} \right] \quad (3.26)$$

where $\mathbb{F}_{\ell_j} \equiv \mathbb{F}_\ell$ when $\ell_j = \ell$. Thus, $\mathbb{C}_p$ contains all the correspondingly down-shifted FFT transformed delay-dependent PN-sequences associated with the p^{th} transmit antenna. Based on $\mathbb{C}_p$ in Equation 3.26, a preprocessor matrix $\mathbb{P}_{p\ell}$ can be formed by

$$\mathbb{P}_{p\ell} = \mathbb{I}_N \otimes \mathbf{P}^\perp [\mathbb{C}_{p\ell}] \quad (3.27)$$

where $\mathbb{C}_{p\ell}$ is formed from the matrix $\mathbb{C}_p$ by removing its $(\ell + 1)^{\text{th}}$ column. In Equation 3.27, $\mathbf{P}^\perp [\mathbb{C}_{p\ell}]$ denotes the projection matrix onto the complementary subspace of $\mathbb{C}_{p\ell}$.

The preprocessor matrix $\mathbb{P}_{p\ell}$ is then applied to the signal $\underline{x}[n]$ given by Equation 3.24 to obtain

$$\underline{z}_{p\ell}[n] = \mathbb{P}_{p\ell} \underline{x}[n]$$

where the preprocessor matrix $\mathbb{P}_{p\ell}$ can be observed to be applied directly to the matrices $\mathbb{H}_{\text{desired}}$, $\mathbb{H}_{\text{ISI}}^{\text{prev}}$ and $\mathbb{H}_{\text{ISI}}^{\text{next}}$. Thus, for the p^{th} transmit antenna element, the preprocessor matrix $\mathbb{P}_{p\ell}$ will be able to null its MC-STAR manifold vectors contained in $\mathbb{H}_{\text{desired}}$ which do not have the same delay and at the same time, transform the MC-STAR manifold vectors which correspond to the delay ℓ .

The effect of $\mathbb{P}_{p\ell}$ on the MC-STAR manifold vectors of the p^{th} transmitter element is illustrated through the following example by considering paths with delay ℓT_s . Consider the TOI in Equation 3.24, which can be written as

$$[\underline{h}_{p1}, \dots, \underline{h}_{pj}, \dots, \underline{h}_{pK}] \left[\beta_1 [\underline{\bar{S}}_1]_p^*, \dots, \beta_j [\underline{\bar{S}}_j]_p^*, \dots, \beta_K [\underline{\bar{S}}_K]_p^* \right]^T \mathbf{a}_p[n]$$

instead.

In the simple case of a *single* path with delay corresponding to ℓT_s and paths with a delay of ℓT_s are to be detected, the preprocessor $\mathbb{P}_{p\ell}$ is used and so, the effect of $\mathbb{P}_{p\ell}$ on the K MC-STAR manifold vectors of the p^{th} transmitter is given by

$$\begin{aligned} & \mathbb{P}_{p\ell} [\underline{h}_{p1}, \dots, \underline{h}_{pj}, \dots, \underline{h}_{pK}] \\ = & \begin{cases} \underbrace{[\underline{0}_{2N_{sc}}, \underline{0}_{2N_{sc}}, \dots, \underline{0}_{2N_{sc}}]_{(j-1) \text{ terms}}}_{(j-1) \text{ terms}}, \mathbb{P}_{p\ell} \underline{h}_{pj}, \underbrace{[\underline{0}_{2N_{sc}}, \dots, \underline{0}_{2N_{sc}}]_{(K-j) \text{ terms}}}_{(K-j) \text{ terms}}, & \text{if } \ell_{pj} = \ell \text{ (transformation)} \\ \underbrace{[\underline{0}_{2N_{sc}}, \underline{0}_{2N_{sc}}, \dots, \underline{0}_{2N_{sc}}]_{(j-1) \text{ terms}}}_{(j-1) \text{ terms}}, \underline{0}_{2N_{sc}}, \underbrace{[\underline{0}_{2N_{sc}}, \dots, \underline{0}_{2N_{sc}}]_{(K-j) \text{ terms}}}_{(K-j) \text{ terms}}, & \text{if } \ell_{pj} \neq \ell, \forall j \text{ (simplification)} \end{cases} \end{aligned} \quad (3.28)$$

Thus, it is clear from Equation 3.28 that only paths with delay equal to ℓT_s (if any) will exist in the preprocessed signal vector $\underline{z}_{i\ell}[n]$. This is because all vectors $\underline{h}_{pj}$, $\forall j$, with $\ell_{pj} \neq \ell$ will be in the null space of the preprocessor matrix $\mathbb{P}_{p\ell}$. In addition to the simplification/transformation of $\underline{h}_{pj}$, $\forall j$, as shown in Equation 3.28, $\mathbb{P}_{p\ell}$ also has the effect of transforming the vectors $\underline{h}_{pj}$, $\forall p \neq p$ due to the code diversity employed in the transmitter. The matrices $\mathbb{H}_{\text{ISI}}^{\text{prev}}$ and $\mathbb{H}_{\text{ISI}}^{\text{next}}$ are also transformed. Thus,

$$\begin{aligned} \underline{z}_{p\ell}[n] = & \underbrace{\sum_{j=1}^K \beta_j [\underline{S}_j]_p^* \mathbb{P}_{p\ell} \underline{h}_{pj} a_p[n]}_{\text{transformed TOI}} + \underbrace{\sum_{p=1, p \neq p}^{\bar{N}} \sum_{j=1}^K \beta_j [\underline{S}_j]_p^* \mathbb{P}_{p\ell} \underline{h}_{pj} a_p[n]}_{\text{transformed CCI}} \\ & + \underbrace{\mathbb{P}_{p\ell} \mathbb{H}_{\text{ISI}}^{\text{prev}} \underline{a}[n-1] + \mathbb{P}_{p\ell} \mathbb{H}_{\text{ISI}}^{\text{next}} \underline{a}[n+1]}_{\text{transformed ISI}} + \mathbb{P}_{p\ell} \underline{n}[n] \end{aligned} \quad (3.29)$$

It is assumed, without any loss of generality, that the angle of elevation ϕ is equal to zero for all signals that impinge on the receive antenna array. This implies that all signals are located within the (x,y)-plane. As such, the MC-STAR manifold vector is only dependent on the azimuth θ and delay ℓ . The aim of the channel estimation algorithm is thus to obtain estimates of the parameters (θ, ℓ) .

The locus of all transformed MC-STAR manifold vectors, over the parameter space of θ , thus gives rise to a 1-dimensional continuum and the intersection of the

signal subspace with this manifold curve will provide the required spatial channel parameters for the p^{th} transmitter antenna. In order to obtain the signal subspace from the received signal, the covariance matrix of the preprocessed signal vector $\underline{z}_{\text{pl}}[n]$ is first obtained by

$$\mathbb{R}_{\underline{z}_{\text{pl}}} = \mathcal{E} \left\{ \underline{z}_{\text{pl}}[n] \underline{z}_{\text{pl}}^H[n] \right\} \quad (3.30)$$

The eigendecomposition of the preprocessed covariance matrix $\mathbb{R}_{\underline{z}_{\text{pl}}}$ is then carried out to partition the observation space of $\mathbb{R}_{\underline{z}_{\text{pl}}}$ into the signal and noise subspaces. The Akaike Information Criterion (AIC) or Minimum Description Length (MDL) criterion [52] can be used to separate the signal and noise subspaces.

If a bank of preprocessors $\mathbb{P}_{\text{pl}}$, $\ell = 0, 1, \dots, (N_{\text{sc}} - 1)$ is implemented at the receiver, a two-dimensional cost function over the parameters ℓ and θ can be obtained, which is given by

$$\zeta_{\text{p}}(\ell, \theta) = \frac{\left(\underline{S}_j(\theta) \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell} \underline{\alpha}_{\text{p}} \\ \underline{0}_{N_{\text{sc}}} \end{bmatrix} \right)^H \mathbb{P}_{\text{pl}} \mathbb{P}_{\text{pl}} \left(\underline{S}_j(\theta) \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell} \underline{\alpha}_{\text{p}} \\ \underline{0}_{N_{\text{sc}}} \end{bmatrix} \right)}{\left(\underline{S}_j(\theta) \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell} \underline{\alpha}_{\text{p}} \\ \underline{0}_{N_{\text{sc}}} \end{bmatrix} \right)^H \mathbb{P}_{\text{pl}} \hat{\mathbb{E}}_{\text{npl}} \hat{\mathbb{E}}_{\text{npl}}^H \mathbb{P}_{\text{pl}} \left(\underline{S}_j(\theta) \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell} \underline{\alpha}_{\text{p}} \\ \underline{0}_{N_{\text{sc}}} \end{bmatrix} \right)} \quad \forall \ell, \theta \quad (3.31)$$

for the p^{th} transmitter antenna. $\hat{\mathbb{E}}_{\text{npl}}$ denotes the estimated noise subspace of the corresponding preprocessed covariance matrix $\mathbb{R}_{\underline{z}_{\text{pl}}}$. Thus, a two-dimensional search of $\zeta_{\text{p}}(\ell, \theta)$ will produce the required estimates $\hat{\ell}$ and $\hat{\theta}$, which are given by the location of the peaks of $\zeta_{\text{p}}(\ell, \theta)$, for the p^{th} transmitter antenna. The proposed system is shown in Figure 3.3 for the p^{th} transmitter antenna in an arrayed MIMO system with $\overline{N}$ antennas at the transmitter.

It must be noted here that the proposed space-time multipath channel estimation algorithm described above considers the case where there are no co-delay

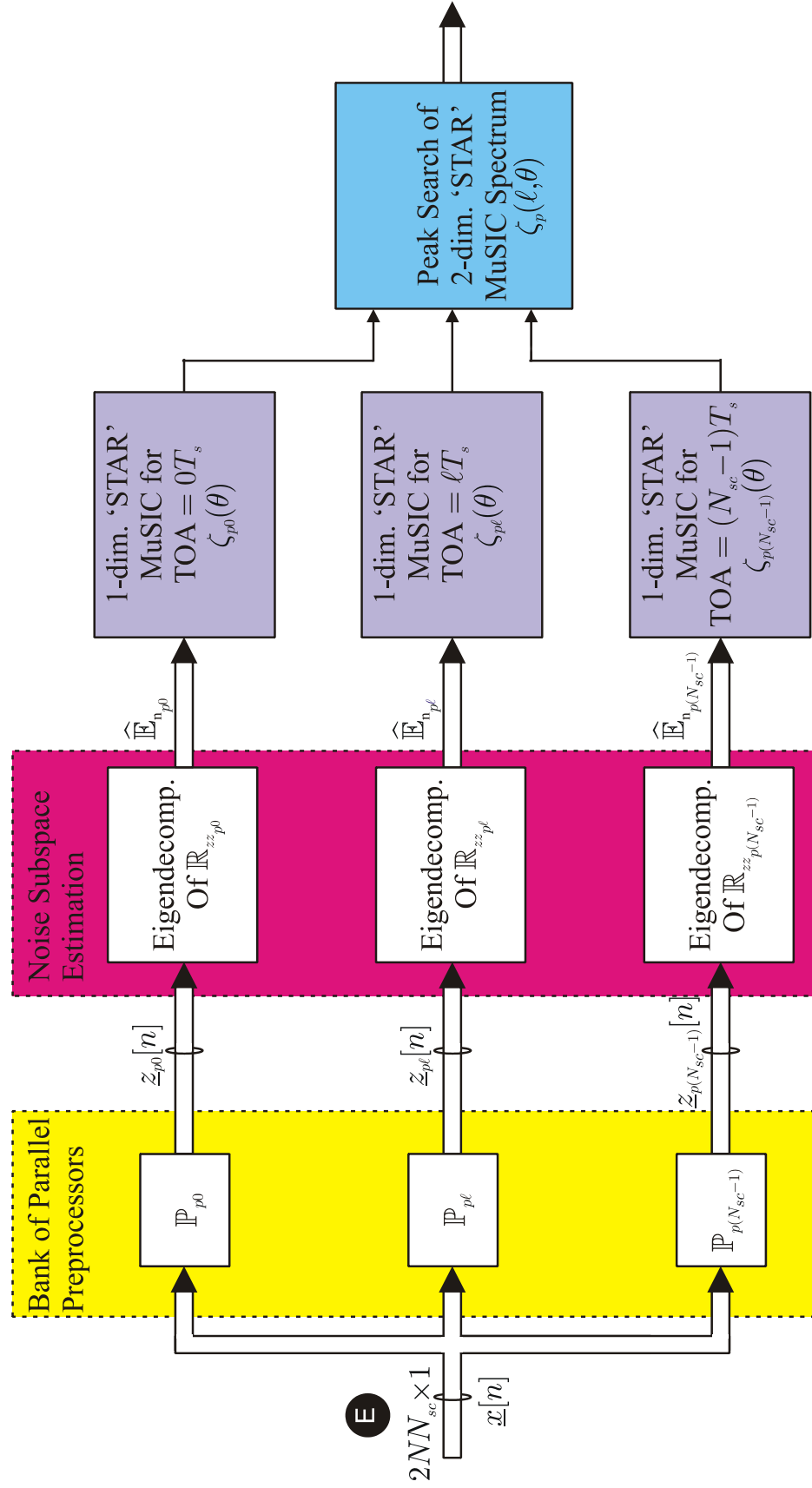


Figure 3.3: Bank of preprocessors required for the formation of the 2-dimensional STAR MUSIC spectrum to obtain estimates of the spatio-temporal channel parameters.

paths in the space-time multipath channel for each transmitter. Due to the nature of the preprocessor matrix used, the resultant transformed MC-STAR manifold vectors of co-delay paths will be linearly combined if co-delay paths are present in the space-time multipath channel for the transmitter antenna of interest. This is because the transformed MC-STAR manifold vectors with the same path delay will remain in the preprocessed received signal vector $\underline{z}_{p\ell}[n]$. In order to resolve this ‘coherence’ problem that arises from co-delay paths, spatial smoothing [53] has been performed on the preprocessed received signal vector $\underline{z}_{p\ell}[n]$ prior to the construction of the cost function $\zeta_p(\ell, \theta)$. However, this is only applicable if a ULA is applied at the receiver.

3.6 Simulation Studies

In this section, several examples are presented to demonstrate the feasibility of the proposed cyclic prefix-free MC-CDMA arrayed MIMO system. Consider a cyclic prefix-free MC-CDMA arrayed MIMO system with an uniform linear array at both the transmitter and receiver and in both antenna arrays, an antenna separation of half wavelength is assumed. Such a system shall be referred to as $(\bar{N}, N)$ cyclic prefix-free MC-CDMA arrayed MIMO system. Gold sequences of length N_{sc} are used and due to code diversity, each transmitter antenna is assigned a unique Gold sequence. The system is assumed to operate in a frequency-selective fading channel with channel length N_{sc} . Thus, the maximum path delay of the channel is $(N_{sc} - 1)T_s$. The DOAs, θ , and DODs, $\bar{\theta}$, are assumed to be uniformly distributed over $[0^\circ, 180^\circ)$ while the path delays, ℓ , are uniformly distributed over $[0T_s, (N_{sc} - 1)T_s]$.

3.6.1 Detection Performance

The detection capability of the proposed subspace-based receiver is first compared with the RAKE and MMSE receivers described in Section 3.4 for the proposed cyclic prefix-free MC-CDMA arrayed MIMO communications system. The number of antenna elements in the antenna array of the transmitter is fixed at $\overline{N} = 2$ while the number of antenna elements in the receiver antenna array are varied from 2 to 5. Gold sequences of length 7 are used for the PN sequences and the number of paths with distinct time delays is assumed to be 3, i.e. $K = 3$. In addition, the channel is assumed to be slowly time-varying so that the channel can be assumed to be quasi-stationary over each run.

In order to evaluate the performance of the proposed system, a reference MIMO system is also studied in order to compare their performance. The reference system is based on the model proposed in [54]. The reference model assumes the use of independent antenna elements for a MIMO-OFDM system. Due to the use of OFDM in the reference system instead of MC-CDMA, the symbols on each subcarrier in the reference OFDM system are re-interpreted as the PN sequence-modified symbols in a MC-CDMA system. Also, in order to mitigate the effects of the multipath channel, a cyclic prefix of length $N_{sc} - 1$ is used for the reference system. In the receivers of both the proposed and reference systems, perfect knowledge of the channel parameters are assumed to be available. A MMSE receiver is used for the receiver of the reference system to provide a basis for comparison. For the receivers of both the proposed and reference systems, perfect knowledge of the channel parameters are assumed to be available. For the reference system, a MMSE receiver is used to study the performance of the system. The simulation parameters are summarized in Table 3.1.

The constellation plots of the detected symbols for the receivers under consideration obtained for one simulation run at a SNR of 10dB are shown in Figure

Table 3.1: *Summary of Simulation Parameters for Space-Time Estimation*

Parameter	Value
No. of Transmitter Antennas, $\overline{N}$	2
No. of Receiver Antennas, N	2, 3, 4, 5
No. of paths, K	3
Processing Gain, N_{sc}	7
PN-codes	Gold sequences

3.4 for a (2, 2) cyclic prefix-free MC-CDMA arrayed MIMO system. From Figure 3.4, it can be observed that the proposed subspace-based receiver has comparable performance to that of the MMSE receiver and both the subspace-based and MMSE receivers have superior performance over the RAKE receiver. The reference system is observed to have similar performance as the RAKE receiver.

Figure 3.5 shows the BER performance of the proposed system with three different types of receivers, as described earlier, as well as the BER curve of the reference system. The BER performance curves are obtained by averaging over 20,000 simulation runs of 20,000 bits each. It can be seen that the performance of the subspace-based receiver is only slightly worse than that of the MMSE receiver. However, the MMSE receiver requires additional knowledge of the variance of the AWGN present in the channel. Among the three receivers, the RAKE receiver performs the worst, approaching an error floor after a signal-to-noise ratio (SNR) of 10dB. In comparison with the reference system, the proposed system performs much better, especially for the subspace-based and MMSE receivers. At a BER of 10^{-4} , the proposed system, using the subspace-based or MMSE receivers, has a gain of about 8dB over the reference system. However, when compared to the RAKE receiver, the reference system outperforms it at SNRs above 10dB, although the reference system has worse performance at SNRs below 10dB.

The BER performance curves for the (2,3), (2,4) and (2,5) cyclic prefix-free MC-CDMA arrayed MIMO system is next compared with the reference system, as shown in Figures 3.6, 3.7 and 3.8, respectively. It can be observed that increasing

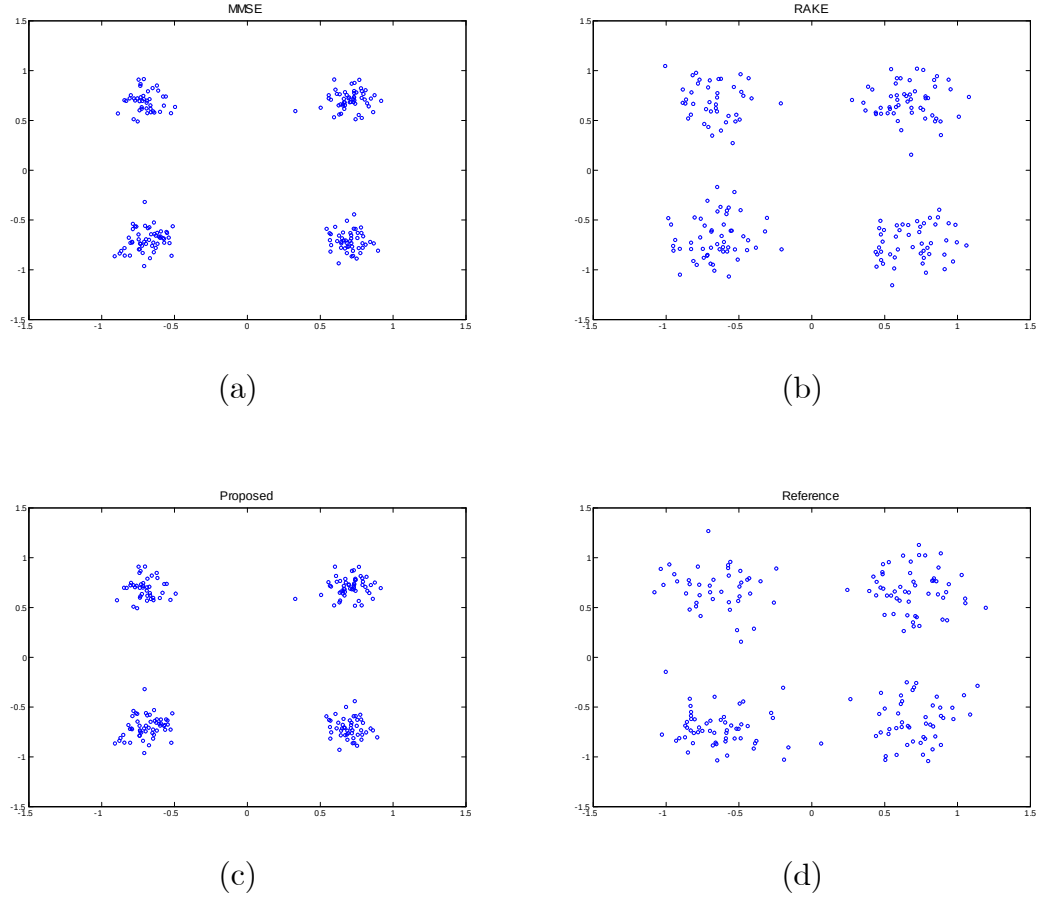


Figure 3.4: Comparison of the signal constellations produced by the receivers under consideration at $SNR = 10\text{dB}$ for a $(2,2)$ cyclic prefix-free MC-CDMA arrayed MIMO system. a) MMSE receiver for proposed system, b) RAKE receiver for proposed system, c) Subspace-based receiver for proposed system, d) MMSE receiver for reference system [54].

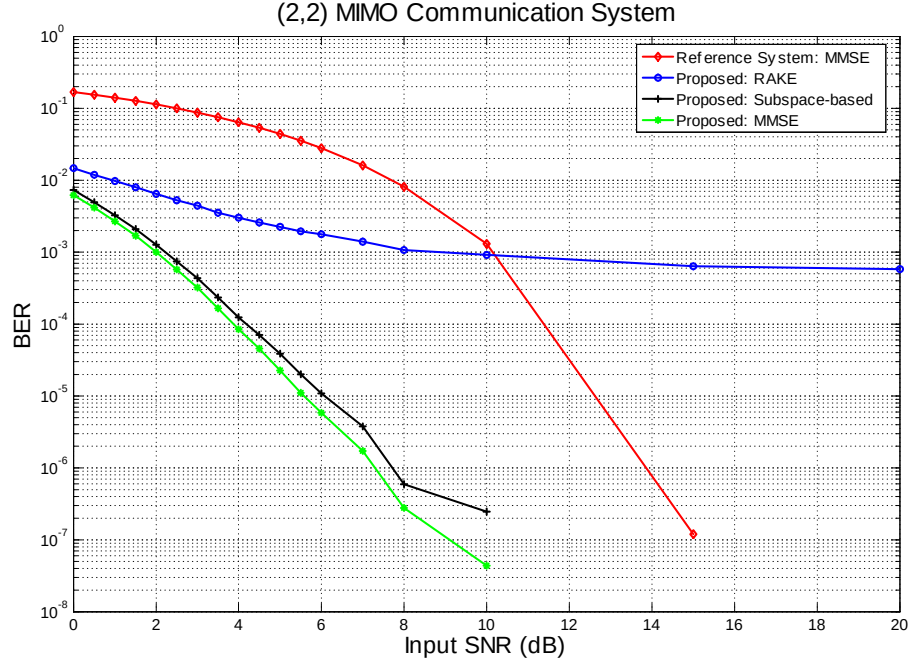


Figure 3.5: Comparison of the BER performance of the (2,2) cyclic prefix-free MC-CDMA arrayed MIMO communication system with the MIMO-OFDM based reference system [54].

the number of receive antenna elements improves the BER performance of both the proposed and reference systems.

In addition to comparing the BER performance of the proposed system with that of the reference system, the achievable system capacity of the two systems are also looked at. Figure 3.9. shows the comparison of the capacity of the proposed system and that of the reference system. As can be seen from the figure, the proposed cyclic prefix-free MC-CDMA arrayed MIMO system is able to achieve much higher capacity than the reference system.

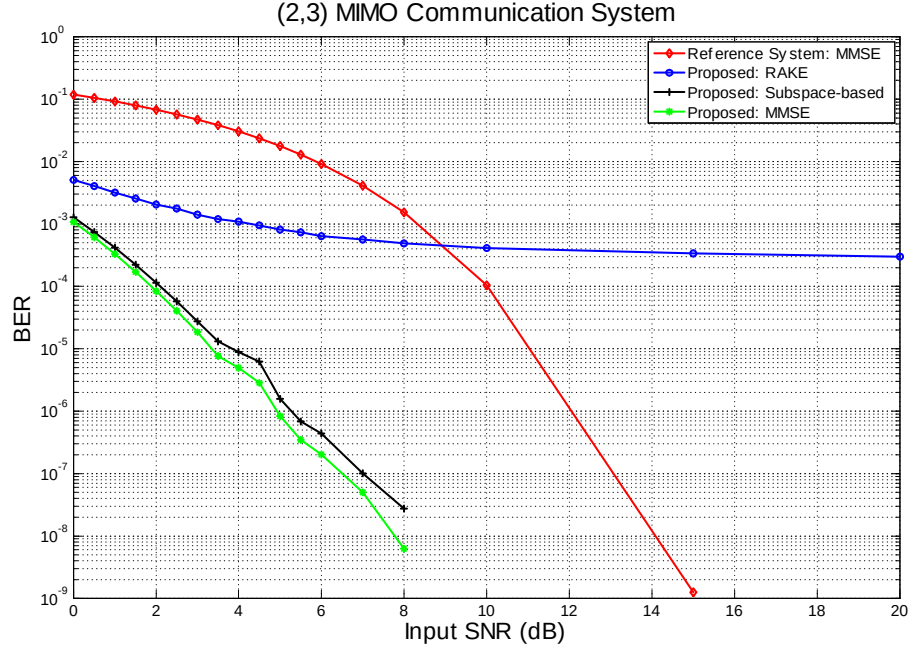


Figure 3.6: Comparison of the BER performance of the (2,3) cyclic prefix-free MC-CDMA arrayed MIMO communication system with the MIMO-OFDM based reference system [54].

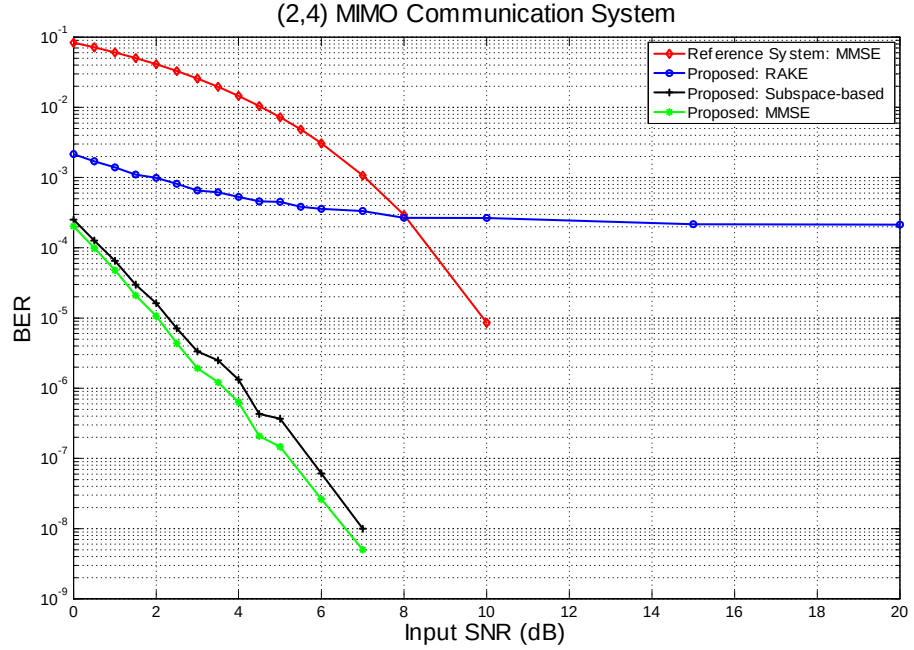


Figure 3.7: Comparison of the BER performance of the (2,4) cyclic prefix-free MC-CDMA arrayed MIMO communication system with the MIMO-OFDM based reference system [54].

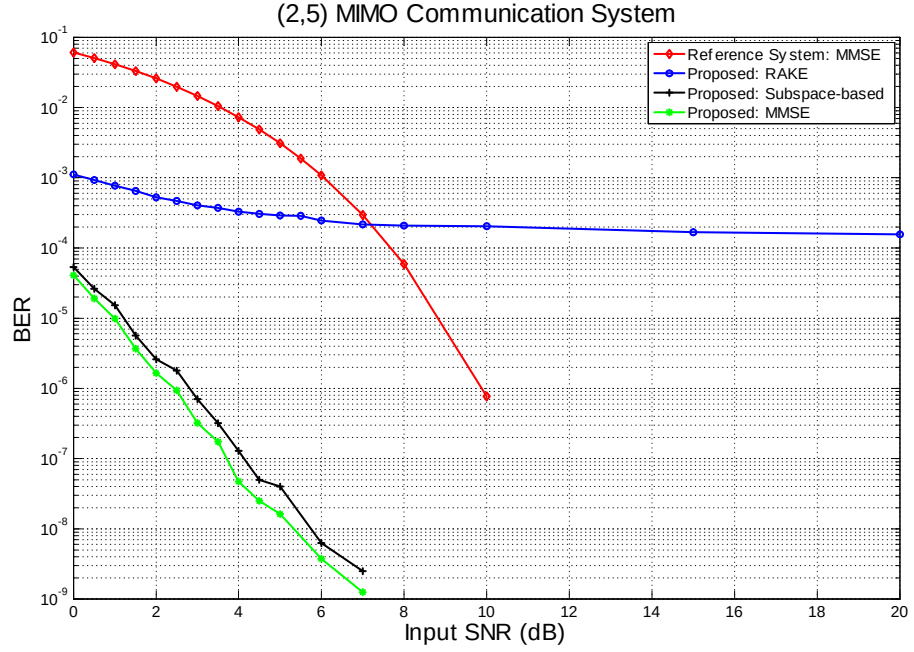


Figure 3.8: Comparison of the BER performance of the (2,5) cyclic prefix-free MC-CDMA arrayed MIMO communication system with the MIMO-OFDM based reference system [54].

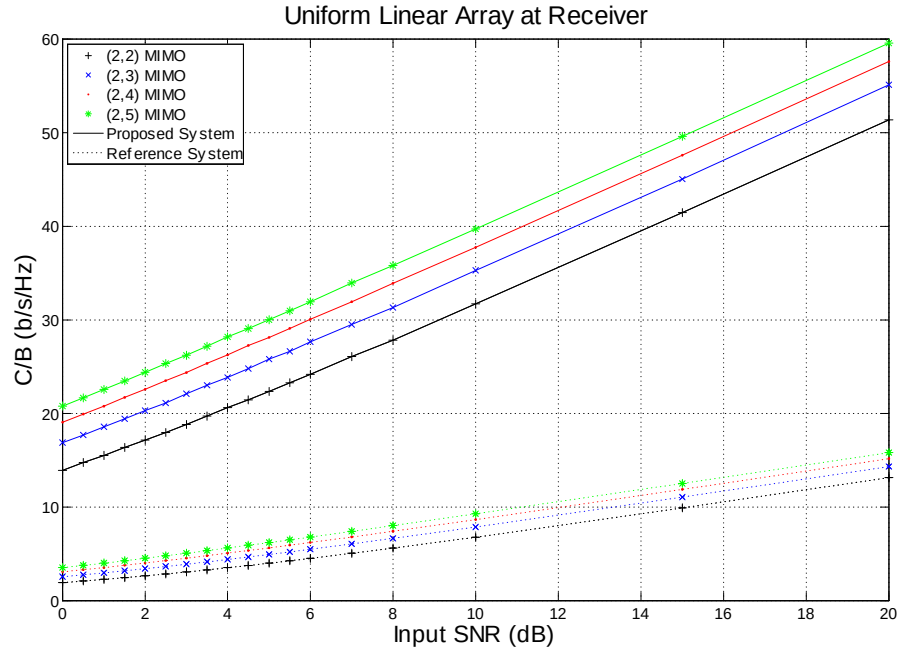


Figure 3.9: System capacity comparison of the arrayed MIMO system with the reference system [54] as the number of receive antennas is increased.

Table 3.2: *Spatial-temporal channel parameters with 2 co-delay and 2 co-directional paths for the (2,4) cyclic prefix-free MC-CDMA arrayed MIMO system under consideration.*

Path	TOA (T_s)	DOA ($^\circ$)	Path Fading Coefficient
j	ℓ_j	θ_j	$\beta_j, \beta_j = 1$
1	28	25	$0.1659 + j0.2733$
2	13	113	$0.1764 + j0.4070$
3	18	100	$0.3983 + j0.1630$
4	18	38	$0.2116 + j0.1803$
5	25	130	$0.1911 - j0.3984$
6	8	130	$0.4627 - j0.0948$
7	20	60	$0.0323 - j0.1385$

3.6.2 Space-Time Parameter Estimation

The performance of the space-time multipath channel estimation algorithm is now investigated for a (2,4) cyclic prefix-free MC-CDMA arrayed MIMO system. PN sequences of length $N_{sc} = 31$ are used. In order to estimate the channel parameters, the receiver is assumed to collect 500 QPSK channel symbols for processing. In addition, the channel is assumed to be slowly time-varying so that the channel can be assumed to be quasi-stationary over each observation interval and the signal-to-noise ratio (SNR) at the receiver is assumed to be 20dB.

The space-time channel parameters during a particular observation interval are listed in Table 3.2. In order to investigate the performance of the channel estimation in co-delay and co-directional situations, it can be seen from Table 3.2 that paths 3 and 4 in the channel have the same delay of $18T_s$. In addition, paths 5 and 6 demonstrate the co-directional situation where both paths have a DOA of 130° .

The 2-D MUSIC spectrum, obtained from the 2-D cost function given by Equation 3.31, is shown in Figure 3.10 for the case where spatial smoothing is not performed prior to the construction of the 2-D MUSIC cost function. It can be seen that only 5 peaks are present in the 2-D MUSIC spectrum obtained. Inspection of the 2-D MUSIC spectrum reveals that the peaks are non-existent

Table 3.3: *Spatial-temporal channel parameters with 2 closely-located paths for the (2,4) cyclic prefix-free MC-CDMA arrayed MIMO system under consideration.*

Path	TOA (T_s)	DOA ($^\circ$)	Path Fading Coefficient
j	ℓ_j	θ_j	$\beta_j, \beta_j = 1$
1	20	130	$0.8678 + j0.1723$
2	19	132	$0.2325 - j0.4039$

at the location for paths 3 and 4. Thus, the channel parameters for paths 3 and 4 cannot be estimated from the 2-D MUSIC spectrum that is obtained. In order to overcome this, spatial smoothing is first carried out and the resultant 2-D MUSIC spectrum is shown Figure 3.11. In the 2-D MUSIC spectrum that is obtained, it can be seen that 7 peaks are now present and each of these corresponds to the paths in the space-time multipath channel. Thus, a search of the 2-D MUSIC spectrum will provide the estimates of the DOAs and TOAs of the paths.

Thus, the proposed channel estimation algorithm has been shown to be effective in the estimation of channel parameters for a cyclic prefix-free MC-CDMA arrayed MIMO system. However, in the event of co-delay paths being present in the channel, spatial smoothing will have to be performed first in order to obtain estimates of the channel parameters of all the paths.

The ability of the channel estimation algorithm to resolve multipaths which are closely located in both time and space is investigated next. Table 3.3 shows the channel parameters for the (2, 4) cyclic prefix-free MC-CDMA arrayed MIMO communication system during a particular observation interval. The SNR at the receiver is assumed to be 20dB. The 2-D MUSIC spectrum that is obtained from the 2-D cost function, as given by Equation 3.31, is shown in Figure 3.12. It is clear that the two multipaths have been successfully resolved.

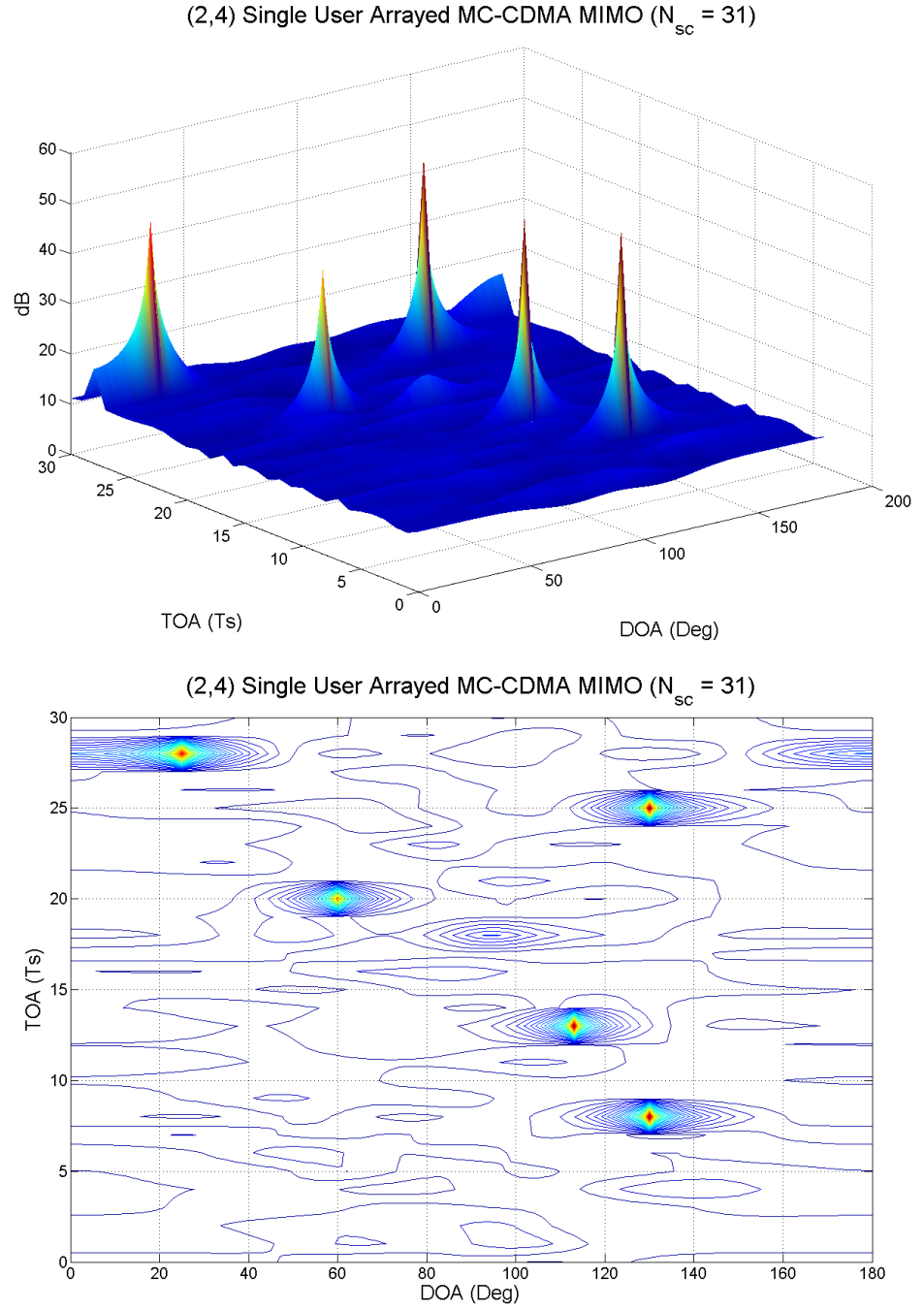


Figure 3.10: Spatial-temporal spectrum showing only 5 out of 7 multipaths of the frequency-selective fading channel in a (2,4) cyclic prefix-free MC-CDMA arrayed MIMO system. Only co-directional paths have been successfully resolved.

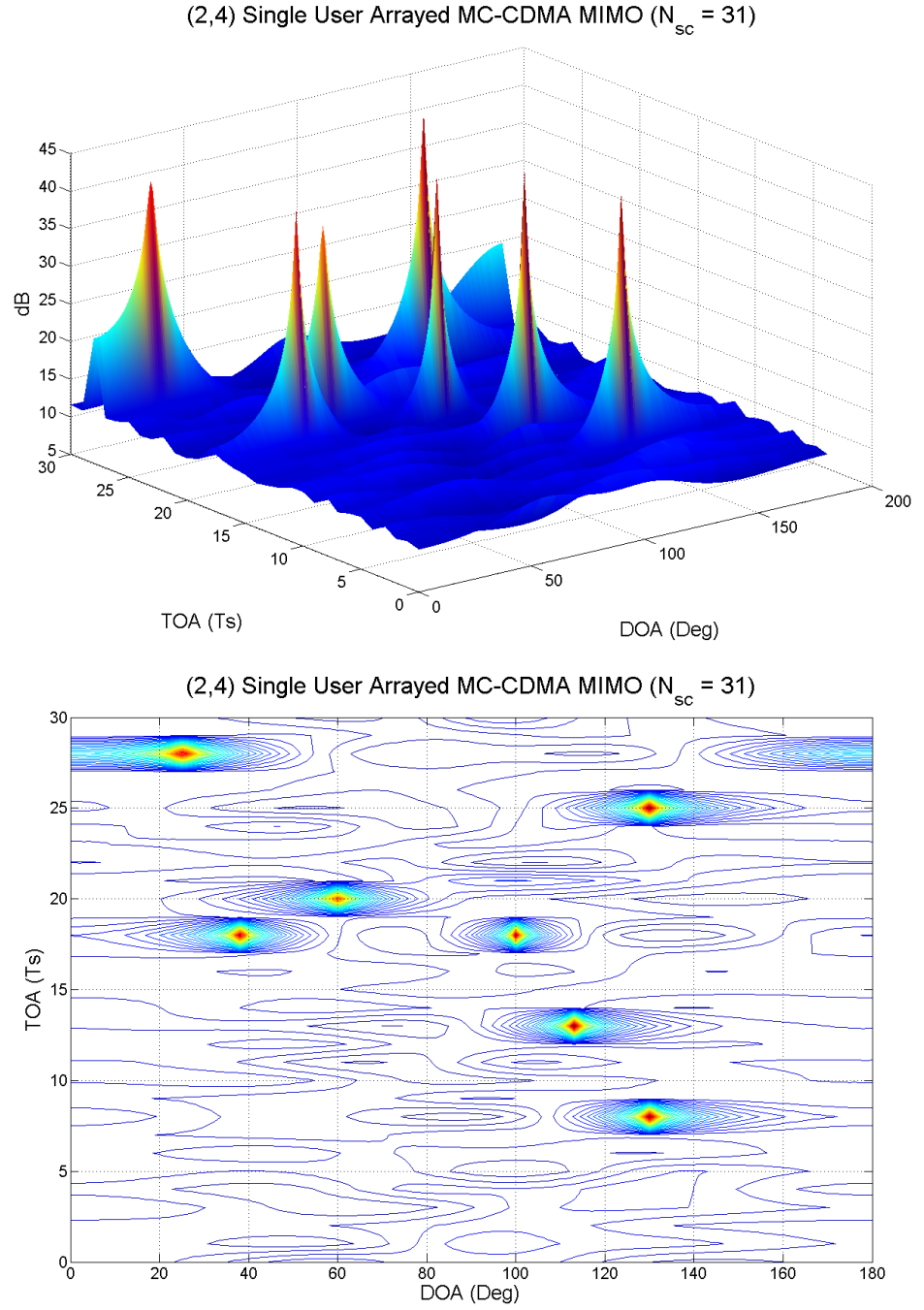


Figure 3.11: *Spatial-temporal spectrum showing all the 7 multipaths of the frequency-selective fading channel in a (2,4) cyclic prefix-free MC-CDMA arrayed MIMO system. 2 overlapping arrays of length 3 are used for the spatial smoothing process and the co-delay paths are successfully resolved.*

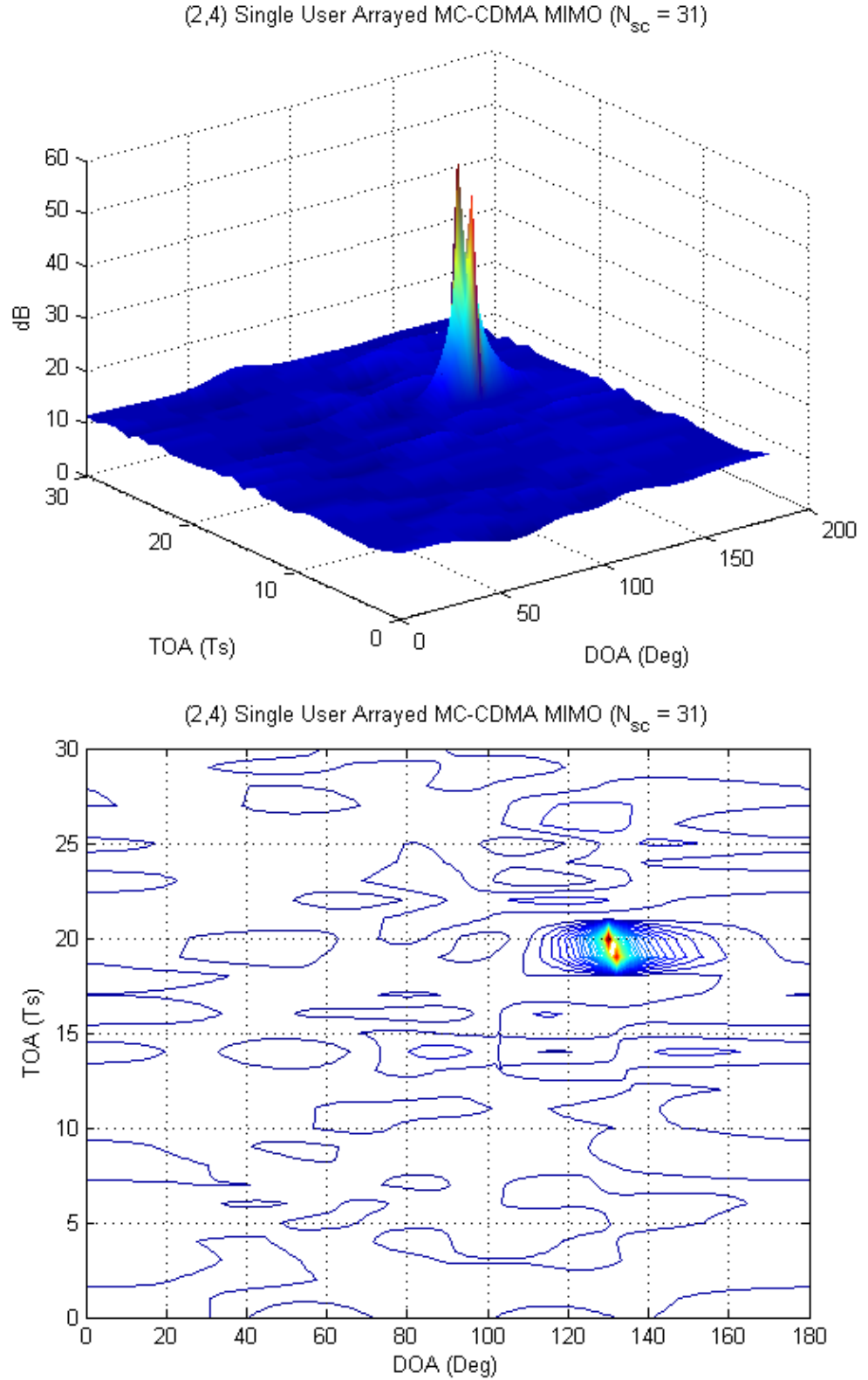


Figure 3.12: *Spatial-temporal spectrum showing both multipaths (which are closely located in both time and space) of the frequency-selective fading channel in a (2,4) cyclic prefix-free MC-CDMA arrayed MIMO system.*

3.7 Summary

The detection capabilities of the RAKE, MMSE and proposed subspace-based receivers in a cyclic prefix-free MC-CDMA arrayed MIMO system have been compared with that of a reference system and it has been observed that the array-based receivers have much better performance than the reference system. In addition, the BER performance has also been investigated and it has been shown that the inclusion of an antenna array at the receiver results in a far superior BER performance than the reference system. Moreover, it was seen that the subspace-based receiver has a similar BER performance to that of the MMSE receiver. The comparison of the proposed cyclic prefix-free MC-CDMA arrayed MIMO system with the reference system has also shown that the lack of cyclic prefix in the MC-CDMA signals used does not result in a worse performance than a system in which cyclic prefixes are used.

A space-time multipath channel estimation algorithm for asynchronous array-based MC-CDMA MIMO systems has also been presented in this chapter. The proposed algorithm only requires knowledge of the PN-sequence of the transmitter whose channel is to be estimated and it makes use of the concept of the MC-STAR manifold vector for the formulation of the preprocessor matrix so that the space-time channel estimation process can be carried out. The number of multipaths that can be resolved by the proposed channel estimation algorithm ability is not limited by the number of sensors available in the antenna array due to the inclusion of the temporal domain.

Chapter 4

Diffused Channel Estimation and Reception for Cyclic Prefix-free MC-CDMA Arrayed MIMO System

The problem of a spatially diffused vector channel for a cyclic prefix-free MC-CDMA arrayed MIMO system is addressed in this chapter. Based on the MC-STAR manifold vector framework, a model for the spatially-diffused vector channel is extended from the non-diffused vector channel model.

The diffusion process is an important channel impairment and failure to consider the effect of spatial diffusion will result in a deterioration in performance. Thus, the behavior of a spatially diffused channel is studied in consideration of parameter estimation and reception problems in a cyclic prefix-free MC-CDMA arrayed MIMO system. The transmitter structure of the system under consideration is identical to that introduced in Chapter 3. In this chapter, the MC-STAR manifold vector introduced in Chapter 3 is extended to the ‘spatially-diffused’

MC-STAR manifold vector appropriate for spatially-diffused channels. In the modelling of the spatially diffused channel, localized scattering is assumed to occur for each multipath. Hence, a diffused vector channel model of the wireless channel is presented. The derived channel model is then utilized together with the subspace-based channel estimation algorithm proposed in Chapter 3 such that the structure of the diffused channel is fully exploited to enhance the estimation performance of previously proposed joint channel parameter estimation (multipath DOA, TOA and Spread) for asynchronous MC-CDMA systems.

4.1 Introductory Background

Channel measurements have shown that each multipath in wideband systems undergoes localized scattering and arrives at the receiver array through a narrow angular region. Thus, in practical wireless channels, scattering of the multipaths invalidates the point-source assumption used by conventional algorithms. This is especially true in an urban or suburban setup where the transmitted signal often suffers multiple reflections, diffraction and scattering which will result in a diffusion of the signal. Several experiments have been carried out in [55,56,57,58] to underscore the importance of signal diffusion. Also, it is shown in [55] that the presence of diffused signals has a negative effect on the beamforming performance of the receiver. Thus, there is a degradation in performance when conventional array signal processing techniques are applied in diffused channels as these are based on a point source channel model.

In a MIMO communication system, the frequency selective channel is often modeled by assuming that each multipath has distinct spatial and temporal signatures. Thus, each multipath appears at the receiver's antenna array as a point-like source [27]. A similar assumption has also been made for the array-based MIMO

systems considered thus far. Moreover, the capacity gains that can be possibly attained in MIMO systems do not take into consideration the effect of signal scattering [27] and it has been shown in [59, 60] that the capacity gains promised in MIMO systems is adversely affected by the effect of scattering in MIMO channels. As such, it is important to address the scattering effect in MIMO channels so that a proper evaluation of MIMO systems in practical channels can be obtained. For instance, scattering models for MIMO systems have been proposed in [61, 62] so that the scattering effect can be studied.

As a result of the channel diffusion, conventional parametric channel estimation methods such as MUSIC do not provide accurate estimates of the channel parameters [63]. Thus, channel estimation algorithms appropriate for diffused channels will also have to be developed. Several MUSIC-based channel estimation algorithms have been proposed for the estimation of the nominal channel parameters for diffused sources. For instance, Valaee *et al* [64] proposed the DSPE (Distributed Signal Parameter Estimation) algorithm which involves the minimization of a norm of the transformed noise eigenvectors in the signal subspace and it is required that the spatial distribution of the distributed signal belongs to a parametric class. The DISPARE (Dispersed Signal Parametric Estimation) algorithm proposed by Meng *et al* [65] is based on the weighted projection of the eigenvectors of the signal covariance matrix onto the estimated quasi-noise subspace. The algorithm is shown to perform well with an assumption that the signal is uniformly distributed even when the actual distribution is otherwise. However, the algorithm is limited by the requirement of apriori knowledge of the number of sources. Another MUSIC-based algorithm is proposed in [66] where Wu *et al* introduces the vec-MUSIC algorithm which makes use of the properties of mathematical operators to obtain a geometric representation of the covariance matrix of the vectorized outer-product of the data. At the expense of higher com-

putational efficiency, the vec-MUSIC algorithm is able to identify both spread and point-like sources. Thus far, the DISPARE, DSPE and vec-MUSIC algorithms do not consider the angular spread of the diffused source.

The estimation of the nominal DOA as well as the angular spread has been considered in [67] where the estimation process is decoupled and the estimates of the nominal DOA and angular spread are obtained through two successive one-dimensional minimizations. The proposed method by Besson *et al* [67] combines a covariance matching estimation method with the extended invariance principle in the estimation process. However, it is shown in [68] that an ambiguity exists in the estimation of the nominal DOA in the method proposed by Besson *et al* [67] and a constraint is added by Zoubir *et al* in [68] to resolve the inherent ambiguity. With the estimates of the nominal DOA and angular spread, beamforming techniques such as the broad-null beamformer proposed in [69] can then be used for the reception process.

Existing literature on diffused multipath vector channels focuses mainly on DS-CDMA based wireless communication systems. In [70], a diffused channel framework for DS-CDMA systems is presented and a subspace approach is proposed to estimate vector channels having diffusion in the spatial domain. Two receiver techniques for DS-CDMA systems are proposed in [71] to effectively remove the perturbation due to such multipath spatial diffusion. However, no such work has been reported in the literature for multi-carrier MIMO systems operating in spatially diffused multipath channels. This study characterizes the spatially diffused multipath vector channel for a cyclic prefix-free MC-CDMA arrayed MIMO system and proposes a subspace-based channel estimation algorithm that is blind and robust. The estimated channel parameters are subsequently used to formulate the receiver weight vector.

4.2 Diffused-VIVO Channel Model

Due to the scattering of the multipaths, each multipath arrives at the receiver as a composition of multiple inseparable paths. Each of these paths has a distinct path coefficient, DOA and TOA and the inseparability of the paths is due to the space (DOA) and time (TOA) separation being so small such that the receiver system cannot identify them. Thus, these inseparable paths can be grouped together to form a cluster where each cluster appears to arrive at the receiver with a nominal DOA and TOA. The nominal DOAs and TOAs correspond to the directions and time delays from which the receiver observes the maximum signal strength. The channel parameters (e.g. DOA, TOA) of the rays within each cluster are statistically distributed around the nominal channel parameters of that cluster. The diffusion as seen by the receiver system is due to these spatio-temporally unresolvable point-like paths that arrive at the receiver after diffuse reflection through perturbations with respect to the nominal DOAs and TOAs respectively. The temporal dispersion of the paths within each cluster has been shown to follow an exponential distribution function and exponentially decay away from the nominal TOA [72]. On the other hand, the spatial dispersion observed within each cluster follows a Gaussian distribution. Thus, signals arrive at the receiver through multiple clusters and, in addition to the nominal DOA and TOA, each cluster is also characterized by its temporal and spatial spread. A scattering propagation channel is shown in Figure 4.1 for an arrayed MIMO system. It can be seen from the figure that as a result of scattering, the j^{th} cluster is made up of multiple paths, each with its own DOA and TOA.

A diffused channel model suitable for a MC-CDMA arrayed MIMO system is now introduced. However, with the presence of a rich scattering environment close to the receiver array, space-diffused propagation gives rise to spatially unresolvable point-like rays that arrive at the receiver after diffuse reflection through

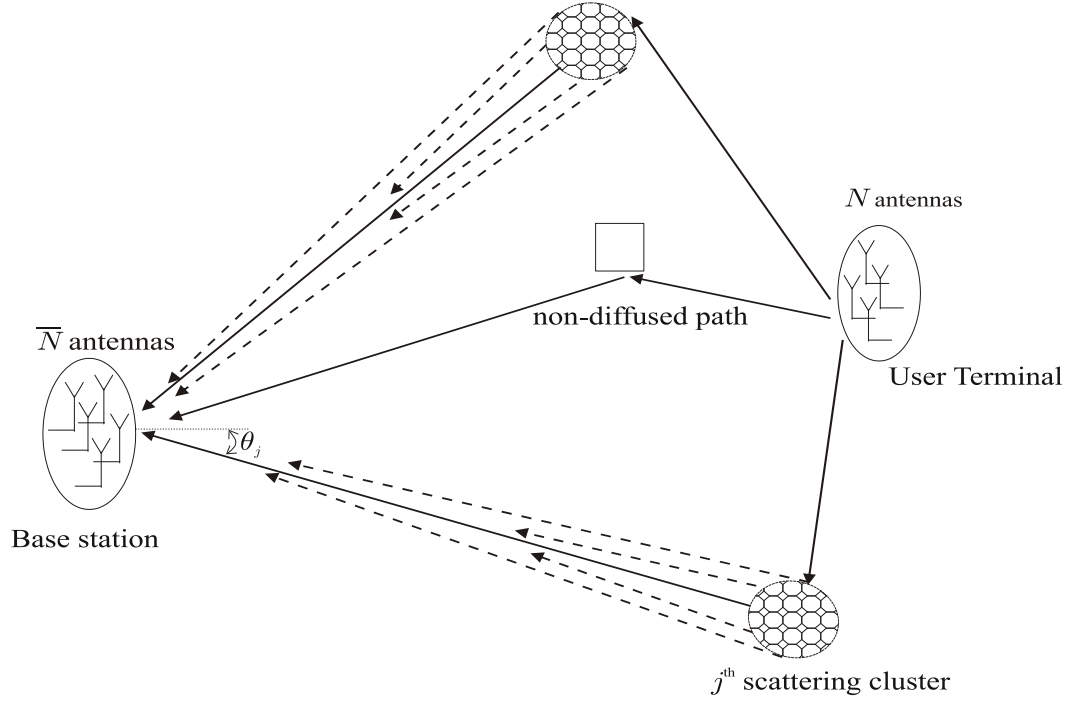


Figure 4.1: *Scattering propagation channel for a MIMO (VIVO) system.*

perturbations with respect to the nominal DOAs [73]. Thus, the vector channel arising from the spatial spread of multipaths is termed as a spatially diffused vector channel. In the diffused channel model considered here, it is assumed that each multipath cluster is subject to coherent local scattering such that the temporal diffusion within the cluster can be ignored.

4.2.1 Space-diffused Vector Channel Model

The model of the spatially diffused wireless channel for an arrayed MIMO system is presented in Figure 4.2, where the transmitted vector-signal $\underline{m}(t)$ arrives at the receiver via K diffused multipath clusters. The modelling of the j^{th} cluster is shown in detail in Figure 4.2. As shown, each multipath becomes a cluster of rays and each cluster is made up of W_j space-time inseparable point-like rays [72]. The figure shows that the l^{th} ray of the j^{th} cluster has a path coefficient β_{jl} (which is common for all subcarriers of the l^{th} ray), direction-of-arrival (DOA) θ_{jl} and

time-of-arrival (TOA) τ_j . Also, $\underline{S}_{jl} \triangleq \underline{S}(\theta_{jl})$ is the receiver array manifold vector for the j^{th} cluster's l^{th} ray arriving at an angle of θ_{jl} , and is a function of the receiver array geometry. Due to the spatial diffusion of the channel, the DOA of the ray θ_{jl} has a perturbation element $\tilde{\theta}_{jl}$ about nominal DOA θ_j such that $\theta_{jl} = \theta_j + \tilde{\theta}_{jl}$.

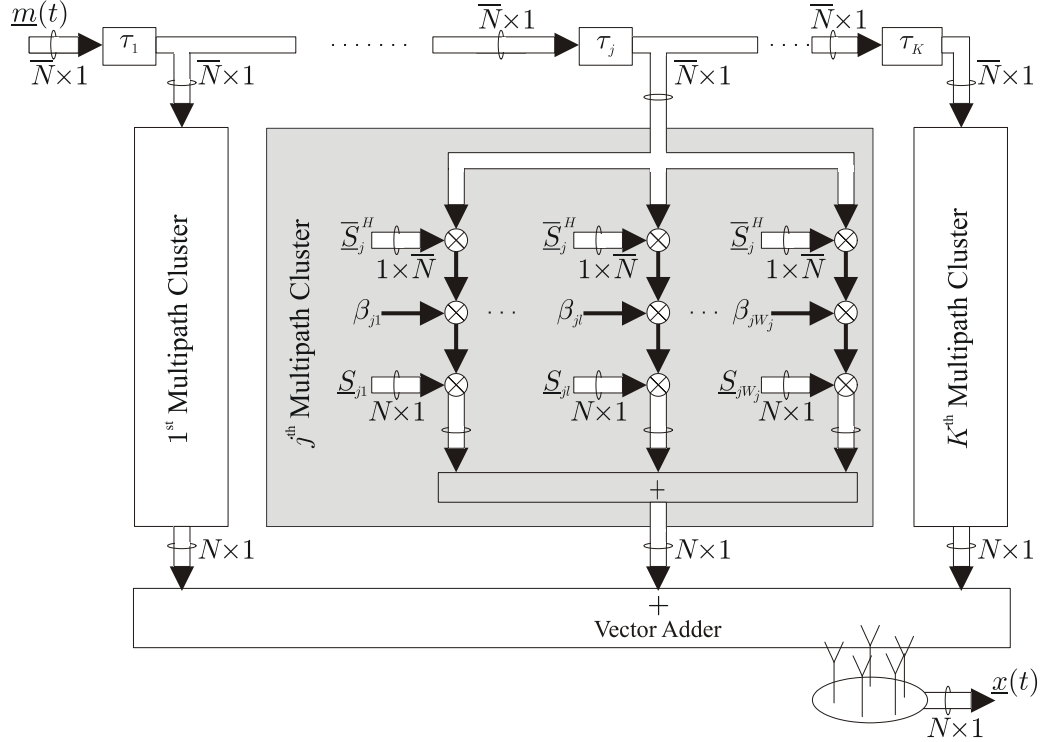


Figure 4.2: Diffused-VIVO channel for arrayed MIMO systems with transmit and receive antenna arrays of small aperture.

Based on Figure 4.2, the baseband signal-vector at the output of the antenna array receiver, through a diffused channel, can be expressed as

$$\text{Diffused-VIVO Channel Impulse Response} = \sum_{j=1}^K \sum_{l=1}^{W_j} \beta_{jl} \underline{S}_{jl} \underline{S}_j^H \delta(t - \tau_j) \quad (4.1)$$

In contrast, for a non-diffused (i.e. point source) channel, the channel impulse response for an arrayed MIMO system is given by:

$$\text{Point Source VIVO Channel Impulse Response} = \sum_{j=1}^K \beta_j \underline{S}_j \underline{S}_j^H \delta(t - \tau_j) \quad (4.2)$$

4.3 Received Signal

The baseband received vector-signal of a cyclic prefix-free MC-CDMA arrayed MIMO system, through a diffused channel, is thus given by

$$\begin{aligned}\underline{x}(t) &= \sum_{j=1}^K \sum_{l=1}^{W_j} \beta_{jl} \underline{S}_{jl} \overline{S}_j^H \underline{m}(t - \tau_j) + \underline{n}(t) \\ &= \sum_{j=1}^K \sum_{l=1}^{W_j} \beta_{jl} \underline{S}_{jl} \overline{S}_j^H \sum_{k=1}^{N_{sc}} \underline{m}_k(t - \tau_j) + \underline{n}(t)\end{aligned}\quad (4.3)$$

Based on the model of the discretized received signal vector for a point source channel in Chapter 3, the discretized received signal vector, observed at point-E in Figure 2.12, can be expressed as

$$\begin{aligned}\underline{x}[n] &= \sum_{p=1}^{\overline{N}} \sum_{j=1}^K \sum_{l=1}^{W_j} \left(\beta_{jl} [\underline{S}_j]_p^* \left(\underline{S}_{jl} \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \right) \underline{a}_p[n] \\ &\quad + \beta_{jl} [\underline{S}_j]_p^* \left(\underline{S}_{jl} \otimes (\mathbb{J}^T)^{N_{sc}} \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \underline{a}_p[n-1] \\ &\quad + \beta_{jl} [\underline{S}_j]_p^* \left(\underline{S}_{jl} \otimes \mathbb{J}^{N_{sc}} \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \underline{a}_p[n+1] \right) + \underline{n}[n].\end{aligned}\quad (4.4)$$

where $[\underline{S}_j]_p^*$ denotes the conjugated p^{th} element of the transmit array manifold vector due to the j^{th} path.

4.3.1 Diffused MC-STAR Manifold Vector

A first order Taylor series approximation of the array manifold vector about the nominal DOA can be written as:

$$\begin{aligned}\underline{S}(\theta_{jl}) &= \underline{S}(\theta_j + \tilde{\theta}_{jl}) \\ &= \underline{S}(\theta_j) + \tilde{\theta}_{jl} \dot{\underline{S}}(\theta_j)\end{aligned}\quad (4.5)$$

where $\dot{\underline{S}}(\theta)$ represents the first derivative of $\underline{S}(\theta)$ with respect to θ . Thus, the first term of Equation 4.4 can be re-written as

$$\begin{aligned} & \sum_{p=1}^{\bar{N}} \sum_{j=1}^K \sum_{l=1}^{W_j} [\underline{S}_j]_p^* \left(\beta_{jl} \underline{S}(\theta_j) \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix} + \beta_{jl} \tilde{\theta}_{jl} \dot{\underline{S}}(\theta_j) \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \mathbf{a}_p[n] \\ &= \sum_{p=1}^{\bar{N}} \sum_{j=1}^K \kappa_{pj} \left[\underline{S}(\theta_j) \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix}, \dot{\underline{S}}(\theta_j) \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix} \right] \begin{bmatrix} 1 \\ \varphi_j \end{bmatrix} \mathbf{a}_p[n]. \end{aligned} \quad (4.6)$$

and the parameters κ_{pj} and φ_j are defined as

$$\kappa_{pj} = [\underline{S}_j]_p^* \sum_{l=1}^{W_j} \beta_{jl} \quad (4.7)$$

$$\varphi_j = \frac{\sum_{l=1}^{W_j} \beta_{jl} \tilde{\theta}_{jl}}{\sum_{l=1}^{W_j} \beta_{jl}} \quad (4.8)$$

The factor κ_{pj} in Equation 4.7 is a random variable that represents the net channel fading coefficient whereas the random variable φ_j is the normalized spreading factor associated with each multipath cluster. It can be observed from Equation 4.8 that, for a large spatial perturbation $\tilde{\theta}_{jl}$, the spreading factor tends to be large when the corresponding channel coefficients are large. Moreover, due to the Laplacian distribution of the power levels of the spatially spread paths, the power levels associated with the paths having a higher spatial spread is much smaller than the paths corresponding to the nominal DOAs. Thus, the spreading factor φ_j can be used as a measure of the spatial diffusion inherent in each multipath cluster.

Thus, in the diffused channel model, the *diffused* MC-STAR manifold vector $\underline{\mathbf{h}}_{pj}$ is different than the standard *point-source* MC-STAR manifold vector $\underline{\mathbf{h}}_{pj}$ (in

Equation 3.25), and is represented by

$$\begin{aligned} \underline{\mathbf{h}}_{pj} &= \begin{bmatrix} \underline{S}(\theta_j) \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix}, \underline{\dot{S}}(\theta_j) \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix} \end{bmatrix} \begin{bmatrix} 1 \\ \varphi_j \end{bmatrix} \\ &= \left(\mathbb{M}_j \otimes \mathbb{J}^{\ell_j} \begin{bmatrix} \mathbb{F}_{\ell_j} \underline{\alpha}_p \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \underline{\mathbf{p}}_j \end{aligned} \quad (4.9)$$

where $\mathbb{M}_j = [\underline{S}(\theta_j), \underline{\dot{S}}(\theta_j)]$ and $\underline{\mathbf{p}}_j = [1 \ \varphi_j]^T$ is the spread factor of the j^{th} multipath cluster. It is important to point out here that for the point-source assumption $\varphi_j = 0$, and Equation 4.9 reduces to the *point-source* MC-STAR manifold vector given by Equation 3.25.

Thus, the received vector $\underline{x}[n]$ for the current symbol period can be written as

$$\underline{x}[n] = \sum_{p=1}^{\overline{N}} \sum_{j=1}^K \beta_j [\underline{S}_j]_p^* \underline{\mathbf{h}}_{pj} a_p[n] + \underline{\mathbf{n}}[n] \quad (4.10)$$

The n^{th} symbol vector, $\underline{x}[n]$, is thus composed of linear combination of $\underline{\mathbf{h}}_{pj}$'s and β_j 's weighted by the p^{th} element of $\underline{S}_j^H$ for the n^{th} input channel symbol.

The received signal vector, taking into account the contributions of the previous and next channel symbols, is then given by

$$\begin{aligned} \underline{x}[n] &= \sum_{j=1}^K \left(\mathbf{H}_j \beta_j \text{diag} \left(\underline{S}_j^H \right) \underline{\mathbf{a}}[n] \right. \\ &\quad + \left(\mathbb{I}_N \otimes (\mathbb{J}^T)^{N_{sc}} \right) \mathbf{H}_j \beta_j \text{diag} \left(\underline{S}_j^H \right) \underline{\mathbf{a}}[n-1] \\ &\quad \left. + \left(\mathbb{I}_N \otimes \mathbb{J}^{N_{sc}} \right) \mathbf{H}_j^{\text{next}} \beta_j \text{diag} \left(\underline{S}_j^H \right) \underline{\mathbf{a}}[n+1] \right) + \underline{\mathbf{n}}[n] \end{aligned} \quad (4.11)$$

where

$$\mathbf{H}_j = [\underline{\mathbf{h}}_{1j}, \dots, \underline{\mathbf{h}}_{pj}, \dots, \underline{\mathbf{h}}_{\overline{N}j}]$$

denotes the collection of the $\overline{N}$ diffused MC-STAR manifold vectors for the j^{th} path. In addition, $\underline{\mathbf{a}}[n] = [a_1[n], \dots, a_p[n], \dots, a_{\overline{N}}[n]]^T$ denotes the current channel symbol vector.

In a more compact form,

$$\underline{x}[n] = \mathbb{H}_{\text{desired}} \underline{\mathbf{a}}[n] + \mathbb{H}_{\text{ISI}}^{\text{prev}} \underline{\mathbf{a}}[n-1] + \mathbb{H}_{\text{ISI}}^{\text{next}} \underline{\mathbf{a}}[n+1] + \underline{\mathbf{n}}[n] \quad (4.12)$$

where

$$\mathbb{H}_{\text{desired}} = \sum_{j=1}^K \mathbf{H}_j \beta_j \text{diag} \left(\overline{\underline{\mathbf{S}}}_j^H \right), \quad (4.13)$$

$$\mathbb{H}_{\text{ISI}}^{\text{prev}} = \sum_{j=1}^K \left(\mathbb{I}_N \otimes (\mathbb{J}^T)^{N_{sc}} \right) \mathbf{H}_j \beta_j \text{diag} \left(\overline{\underline{\mathbf{S}}}_j^H \right) \quad (4.14a)$$

$$\begin{aligned} &= \left(\mathbb{I}_N \otimes (\mathbb{J}^T)^{N_{sc}} \right) \mathbb{H}_{\text{desired}} \\ \mathbb{H}_{\text{ISI}}^{\text{next}} &= \sum_{j=1}^K \left(\mathbb{I}_N \otimes \mathbb{J}^{N_{sc}} \right) \mathbf{H}_j^{\text{next}} \beta_j \text{diag} \left(\overline{\underline{\mathbf{S}}}_j^H \right) \quad (4.14b) \\ &= \left(\mathbb{I}_N \otimes \mathbb{J}^{N_{sc}} \right) \mathbb{H}_{\text{desired}} \end{aligned}$$

In Equation 4.12, $\mathbb{H}_{\text{desired}}$ denotes the channel matrix of the current symbol period, while $\mathbb{H}_{\text{ISI}}^{\text{prev}}$ and $\mathbb{H}_{\text{ISI}}^{\text{next}}$, as defined in Equation 4.14, denote the channel matrices of the ISI present in a cyclic prefix-free MC-CDMA arrayed MIMO system.

4.4 Channel Estimation and Reception

The proposed blind channel estimation for the diffused vector channel makes use of the preprocessor matrix introduced in Section 3.5. As before, it is assumed that the p^{th} transmit antenna is the antenna of interest and the corresponding channel parameters are to be estimated. However, due to the structure of the arrayed MIMO system, the estimation of the channel parameters for the p^{th} transmit antenna is equivalent to the estimation of the channel parameters for all antennas. Thus, the preprocessor matrix $\mathbb{P}_{p\ell}$, defined by Equation 3.27, is then applied to the signal $\underline{x}[n]$ given by Equation 4.4 to obtain

$$\underline{z}_{p\ell}[n] = \mathbb{P}_{p\ell} \underline{x}[n] \quad (4.15)$$

The multiplication of the received signal $\underline{x}[n]$ by $\mathbb{P}_{p\ell}$ nulls the channel vectors, given by $\underline{\mathbf{h}}_{pj}$, which do not have the same delay and at the same time, transforms the channel vectors which correspond to the delay ℓ . In addition, the channel vectors of the remaining transmit antenna elements, i.e. $\underline{\mathbf{h}}_{pj} \forall p, p \neq p$, will be transformed.

As a result of the transformation performed by the $\mathbb{P}_{p\ell}$, the locus of all transformed diffused MC-STAR manifold vectors will result in a 4-dimensional $(\ell, \theta, \kappa, \varphi)$ continuum and the intersection of the signal subspace with this manifold will provide the required channel parameters. The projection of the vector $\left(\mathbb{M}_j \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell\alpha_p} \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \underline{\mathbf{p}}_j$, normalized to have unity norm, onto the estimated noise subspace of the preprocessed spatio-temporal covariance matrix $\mathbb{R}_{p,\ell}$ enables the following subspace cost function to be obtained:

$$\zeta_1(\ell, \theta, \kappa, \varphi) = \frac{\underline{\mathbf{p}}_j^H \left(\mathbb{M}_j \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell\alpha_p} \\ \underline{0}_{N_{sc}} \end{bmatrix} \right)^H \mathbb{P}_{p\ell} \hat{\mathbb{E}}_{n_{p\ell}} \hat{\mathbb{E}}_{n_{p\ell}}^H \mathbb{P}_{p\ell} \left(\mathbb{M}_j \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell\alpha_p} \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \underline{\mathbf{p}}_j}{\underline{\mathbf{p}}_j^H \left(\mathbb{M}_j \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell\alpha_p} \\ \underline{0}_{N_{sc}} \end{bmatrix} \right)^H \mathbb{P}_{p\ell} \mathbb{P}_{p\ell} \left(\mathbb{M}_j \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell\alpha_p} \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \underline{\mathbf{p}}_j} \quad (4.16)$$

where $\hat{\mathbb{E}}_{n_{p\ell}}$ is the estimated noise subspace obtained from the eigendecomposition of the $(\ell + 1)^{\text{th}}$ preprocessed spatio-temporal covariance matrix, given by $\mathbb{R}_{zz_{p\ell}} = \mathcal{E} \{ \underline{z}_{p\ell}[n] \underline{z}_{p\ell}^H[n] \}$. It can be observed from Equation 4.16 that an exhaustive search over all possible values of $(\ell, \theta, \kappa, \varphi)$ will produce all the required channel parameters ℓ, θ, κ and φ . The values of $(\ell, \theta, \kappa, \varphi)$ at which the minimum points occur thus provide the estimates for $(\ell, \theta, \kappa, \varphi)$. However, since $\mathbb{M}_j \underline{\mathbf{p}}_j \neq 0$, the search over all possible combinations of $(\ell, \theta, \kappa, \varphi)$ can be avoided by first minimizing ζ_1 with respect to ℓ and θ . This is achieved by searching for the

minimum generalized eigenvalue of $(\mathbb{D}, \mathbb{B})$ given by

$$\begin{aligned} \hat{\mathbb{D}} &= \left(\mathbb{P}_{\mathbf{p}\ell} \left(\mathbb{M}_j \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell\alpha_{\mathbf{p}}} \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \right)^H \hat{\mathbb{E}}_{\mathbf{n}_{\mathbf{p}\ell}} \hat{\mathbb{E}}_{\mathbf{n}_{\mathbf{p}\ell}}^H \left(\mathbb{P}_{\mathbf{p}\ell} \left(\mathbb{M}_j \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell\alpha_{\mathbf{p}}} \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \right) \\ &= \begin{bmatrix} \hat{d}_{11} & \hat{d}_{12} \\ \hat{d}_{21} & \hat{d}_{22} \end{bmatrix} \end{aligned} \quad (4.17)$$

$$\begin{aligned} \hat{\mathbb{B}} &= \left(\mathbb{P}_{\mathbf{p}\ell} \left(\mathbb{M}_j \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell\alpha_{\mathbf{p}}} \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \right)^H \left(\mathbb{P}_{\mathbf{p}\ell} \left(\mathbb{M}_j \otimes \mathbb{J}^\ell \begin{bmatrix} \mathbb{F}_{\ell\alpha_{\mathbf{p}}} \\ \underline{0}_{N_{sc}} \end{bmatrix} \right) \right) \\ &= \begin{bmatrix} \hat{b}_{11} & \hat{b}_{12} \\ \hat{b}_{21} & \hat{b}_{22} \end{bmatrix} \end{aligned} \quad (4.18)$$

It can be easily seen from Equations 4.17 and 4.18 that $\mathbb{D}$ and $\mathbb{B}$ are 2×2 matrices. Thus, an efficient alternative method of minimizing ζ_1 over (ℓ, θ) is given by

$$\zeta_2(\ell, \theta) = \frac{\mathcal{X} - \sqrt{\mathcal{X}^2 - 4 \det(\hat{\mathbb{B}}) \det(\hat{\mathbb{D}})}}{2 \det(\hat{\mathbb{B}})} \quad (4.19)$$

where

$$\mathcal{X} = (\hat{d}_{11}\hat{b}_{22} + \hat{d}_{22}\hat{b}_{11} - \hat{d}_{12}\hat{b}_{21} - \hat{d}_{21}\hat{b}_{12}) \quad (4.20)$$

Thus, the K minimum points of ζ_2 will produce the K (ℓ, θ) pairs of the multipaths. The spread factor $\underline{\mathbf{p}}$ for a particular cluster can then be obtained from the generalized eigenvector corresponding to the minimum eigenvalue $\zeta_{2,\min}$.

4.4.1 Estimation of Spatial Spread

In order to estimate the spatial spread of the multipath, the first order Taylor series approximation in Equation 4.6 is first rewritten as

$$\begin{aligned} \underline{S}(\theta) + \varphi \dot{\underline{S}}(\theta) &= \underline{S}(\theta) + \text{Re}(\varphi) \dot{\underline{S}}(\theta) + j \text{Im}(\varphi) \dot{\underline{S}}(\theta) \\ &\approx \underline{S}(\theta + \text{Re}(\varphi)) + j \text{Im}(\varphi) \dot{\underline{S}}(\theta) \end{aligned} \quad (4.21)$$

Thus, it can be seen from Equation 4.21 that the angular perturbation away from the nominal DOA can be approximated by $\text{Re}(\varphi)$. Hence, the mean of $|\text{Re}(\varphi)|$ can be approximated by the standard deviation σ_θ of the angular perturbation $\tilde{\theta}$ [74]:

$$\mathcal{E}\{|\text{Re}(\varphi)|\} = \sigma_\theta \quad (4.22)$$

In channels with small angular spreads, a uniform distribution can be assumed for the pdf of the angular spread of the cluster and so, the total angular spread Δ is obtained by

$$\Delta = \sqrt{12}\mathcal{E}\{|\text{Re}(\varphi)|\} \quad (4.23)$$

However, as φ is the normalized spreading factor, it can be replaced by the second component of the minimum eigenvector of the minimum eigenvalue of the generalized eigendecomposition of $(\mathbb{D}, \mathbb{B})$, expressed as

$$\underline{p}(2) = \frac{\zeta_{2,\min}\hat{b}_{11} - \hat{d}_{11}}{\hat{d}_{12} - \zeta_{2,\min}\hat{b}_{12}} \quad (4.24)$$

Therefore, the estimate of the cluster spatial spread is given by

$$\Delta = \sqrt{12}\mathcal{E}\left\{\left|\frac{\zeta_{2,\min}\hat{b}_{11} - \hat{d}_{11}}{\hat{d}_{12} - \zeta_{2,\min}\hat{b}_{12}}\right|\right\} \quad (4.25)$$

4.5 Receiver Weights

Once the space-diffused vector channel parameters, i.e. nominal DOAs θ_j 's, nominal TOAs ℓ_j 's and spatial spread factors φ_j 's, for all the clusters have been estimated using the above approach, the receiver weight vector can be formulated using the concept of diffused MC-STAR manifold vector given in Equation 4.9. By defining the estimated composite channel vector $\underline{\mathbf{h}}_{-p}$ as a linear combination of estimated diffused MC-STAR vector $\underline{\mathbf{h}}_{-pj}$'s

$$\underline{\mathbf{h}}_{-p} \triangleq \sum_{j=1}^K \kappa_{pj} \underline{\mathbf{h}}_{-pj} \quad (4.26)$$

the estimated composite channel matrix $\hat{\mathbb{H}}$ can be written as

$$\hat{\mathbb{H}} = [\mathbb{H}_1, \mathbb{H}_2, \dots, \mathbb{H}_N] \quad (4.27)$$

with $\mathbb{H}_p = [\underline{\mathbf{h}}_p^{\text{prev}}, \underline{\mathbf{h}}_p, \underline{\mathbf{h}}_p^{\text{next}}]$, $\underline{\mathbf{h}}_p^{\text{prev}} = (\mathbb{I}_N \otimes (\mathbb{J}^T)^{N_{sc}}) \underline{\mathbf{h}}_p$ and $\underline{\mathbf{h}}_p^{\text{next}} = (\mathbb{I}_N \otimes \mathbb{J}^{N_{sc}}) \underline{\mathbf{h}}_p$. A linear space-time decorrelating receiver is used to evaluate the performance of the proposed diffused channel estimation approach. The weight vector that decorrelates the data transmitted by the p^{th} transmit antenna is then given by

$$\underline{w}_p = \text{col}_{3p-1} \left((\hat{\mathbb{H}}^\dagger)^T \right) \quad (4.28)$$

where $\text{col}_p(\mathbb{A})$ selects the p^{th} column of $\mathbb{A}$, $(\cdot)^\dagger$ gives the pseudo-inverse and $\hat{\mathbb{H}}$ is the estimated composite channel matrix.

4.6 Simulation Studies

Simulations are carried out for a (3,5) cyclic prefix-free MC-CDMA arrayed MIMO system operating in a spatially diffused channel. Gold sequences of length 31 are used and the channel symbols are assumed to be QPSK modulated. The signals are assumed to be in the x-y plane and so, elevation $\phi_j = 0^\circ$ for all signals.

4.6.1 Channel Estimation using the Proposed Method

First, the estimation of the channel parameters using the proposed method is investigated. The (3,5) cyclic prefix-free MC-CDMA arrayed MIMO system is assumed to operate in an invariant AWGN channel with SNR = 0dB. The received signal is assumed to be composed of 7 multipaths and these parameters are shown in Table 4.1. Each multipath is also assumed to be a cluster of 100 inseparable multipaths which are spatially spread around the nominal DOA by 5° . Data is collected over an observation interval of 500 symbol periods and the parameters

are estimated from a 2-dimensional search of the 2-D MUSIC and the estimation results are presented in Table 4.1. As shown in Table 4.1, the estimated DOAs and TOAs agree well with the simulated values and they can be used to estimate the channel accurately.

The 2-D MUSIC spectrum results, obtained by ζ_2 in Equation 4.19, are shown in Figure 4.3. The MUSIC spectrum in Figure 4.3 shows 7 peaks which correspond to the user's nominal DOAs and TOAs. The estimates of the spatial spread of each cluster are also shown in Table 4.1 and these are estimated by evaluating Equation 4.25 over 50 observation intervals. However, some of the estimates of the spatial spread angles differ slightly from the actual spread of 5° . This is due to finite sample effects in the estimation of the signal covariance matrix and a small number of observation intervals used in the estimation.

Table 4.1: Actual and estimated multipath channel parameters at $SNR = 0dB$

	Path						
Parameters	1	2	3	4	5	6	7
Nom. DOD ($^\circ$)	31	50	62	140	74	111	35
Nom. DOA ($^\circ$)	65	130	40	100	120	40	80
Est. DOA ($^\circ$)	65.5	129.7	40.7	99.6	119.3	40.1	80.0
Nom. TOA (T_s)	3	4	10	29	16	20	24
Est. TOA (T_s)	3	4	10	29	16	20	24
Nom. Spread ($^\circ$)	5.0	5.0	5.0	5.0	5.0	5.0	5.0
Est. Spread ($^\circ$)	3.4	5.0	7.7	4.6	4.5	8.8	5.0

Next, the performance of the proposed diffused channel estimator is compared against an estimator which does not take into account the spatial spread of the channel, i.e. given by Equation 3.31 in Chapter 3 and the estimates are obtained from the location of the peaks in the two-dimensional MUSIC spectrum obtained. A cluster is assumed fixed at DOA of 90° and TOA of $7T_s$ and the standard deviation of the nominal DOA estimates, over 100 independent runs, is computed for different values of spatial spread at $SNR = 20dB$. Figure 4.4 and Figure 4.5 shows the standard deviation performance for a uniform linear array (ULA) and

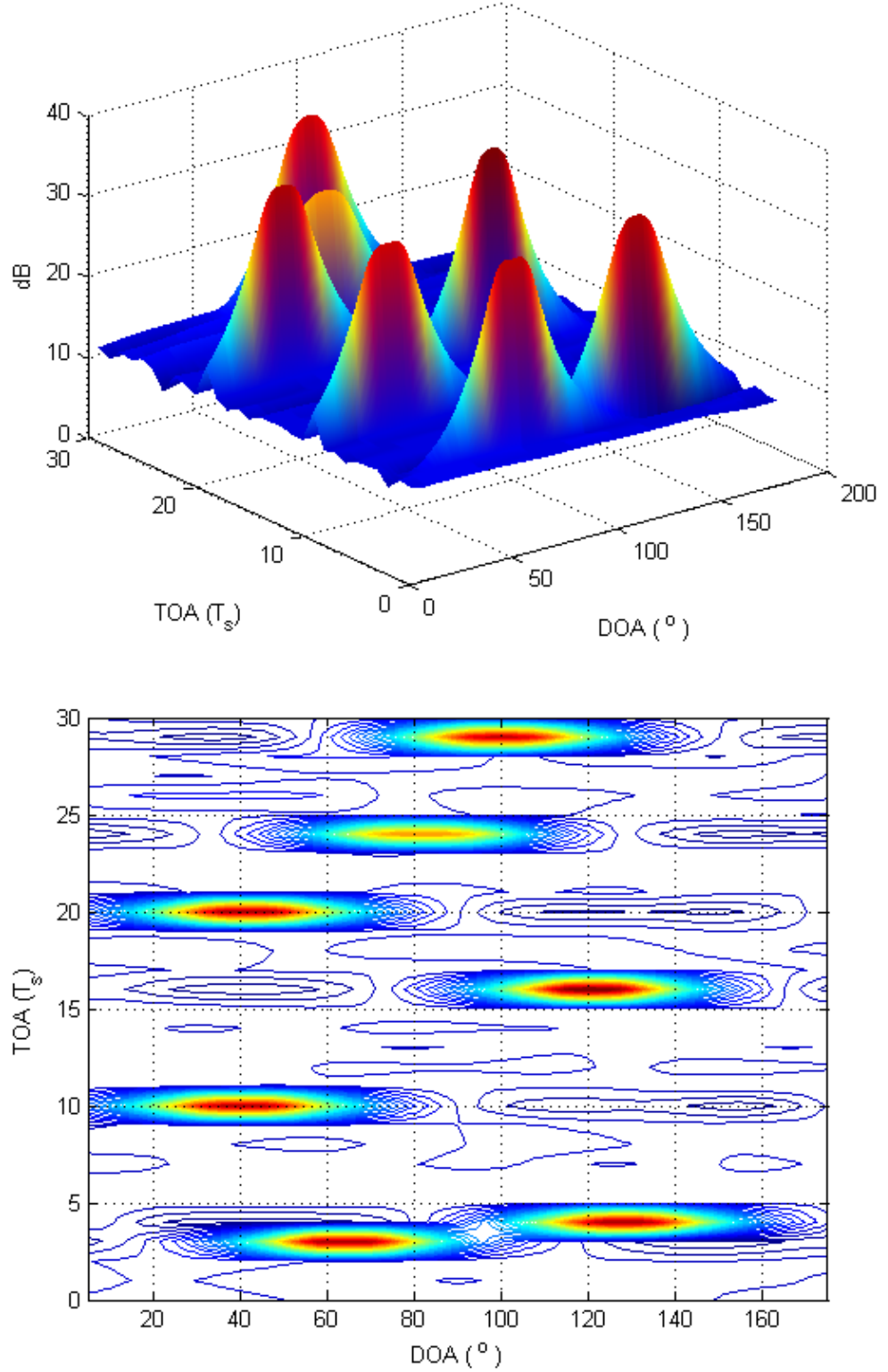


Figure 4.3: 2D MUSIC spectrum for a (3,5) cyclic prefix-free MC-CDMA arrayed MIMO system at SNR = 0dB.

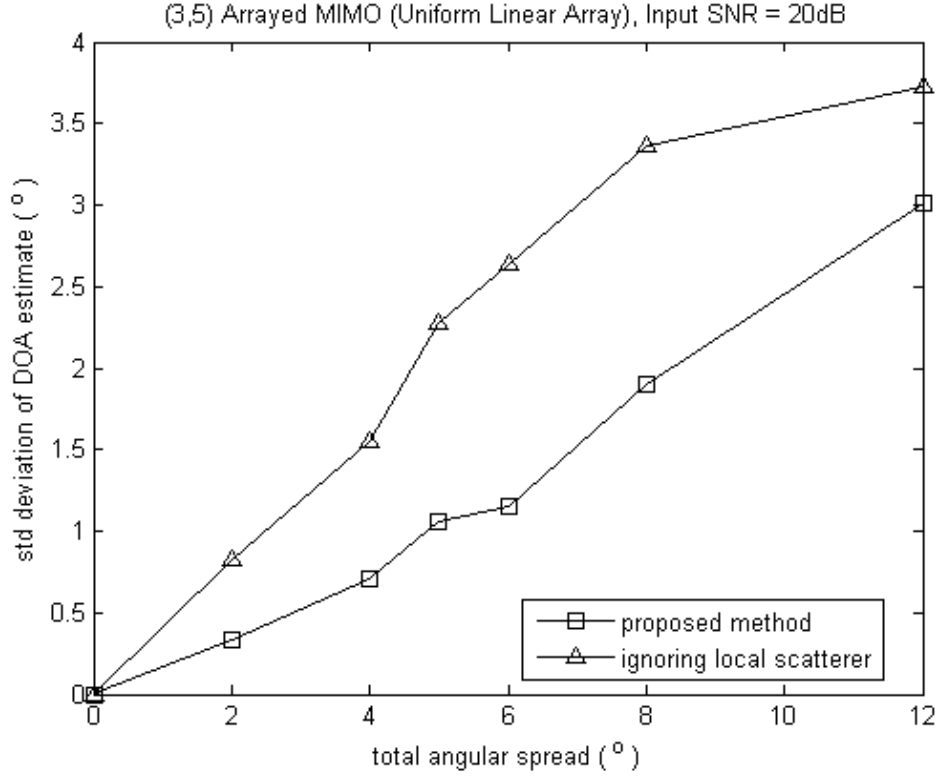


Figure 4.4: Standard deviation of DOA estimates versus spatial spread of multipath at input SNR of 20dB (ULA).

a uniform circular array (UCA), respectively. From the figures, it can be observed that the performance of the proposed estimation approach based on the diffused channel model is much better than methods based on the point-source channel model (which ignores the spatial spread of the multipaths).

The performance of the proposed diffused channel estimator is also compared with the performance of the point-source channel model estimator at spatial spreads of 5° and 10° for different values of SNR when a cluster is assumed fixed at DOA of 90° and TOA of $7T_s$. Figure 4.6 shows the standard deviation results, obtained over 100 independent runs, for the case of a uniform linear array at the receiver while Figure 4.7 shows the results for a uniform circular array. As shown in the figures, the proposed estimator for the diffused channel has a better performance than the estimator that ignores the spatial spread in the channel. However, it can be observed that the performance of the proposed diffused chan-

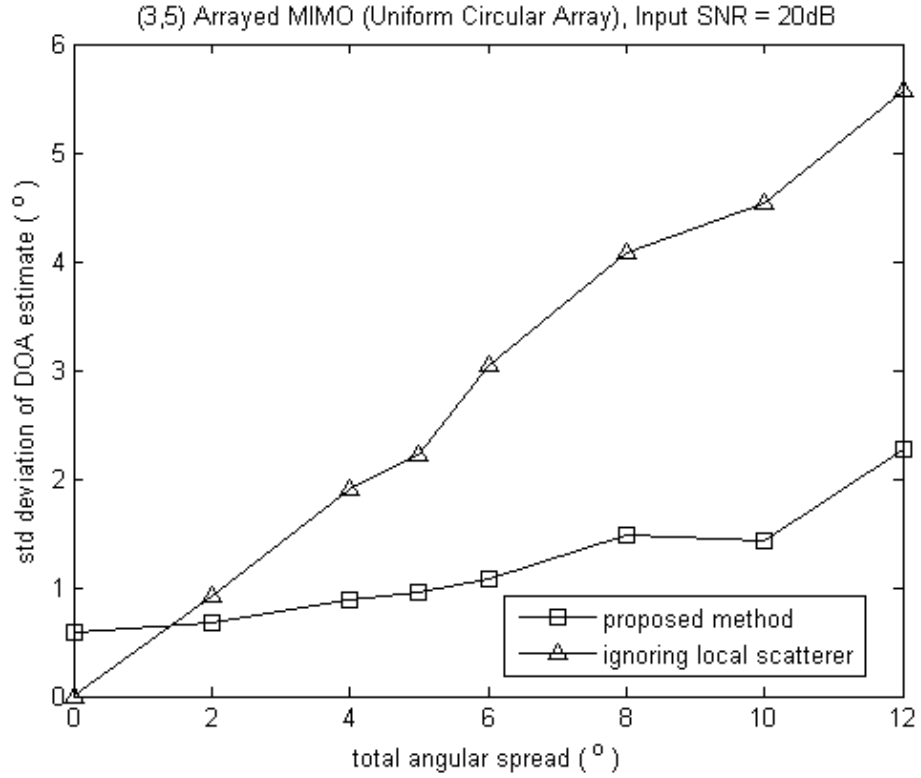


Figure 4.5: Standard deviation of DOA estimates versus spatial spread of multipath at input SNR of 20dB (UCA)¹.

nel estimator does not vary greatly as the SNR increases. This is due to the interference from the other antennas at the transmitter terminal which limits the performance of the estimator.

4.6.2 Reception using the Diffused Channel Framework

Numerical results are now presented to observe the effect of spatial diffusion on arrayed MC-CDMA reception and to highlight the key benefits of using the proposed diffused channel modelling and estimation approach. Once the space-diffused channel parameters have been estimated, the receiver weight vector can be formulated using the estimated parameters. Two scenarios are compared: in the first case the receiver ignores spatial diffusion by using the point-source channel model, whereas in the second case, the diffused channel model is used. The re-

¹The std deviation of the DOA estimate is different from 0 at an angular spread of 0 due to numerical errors.

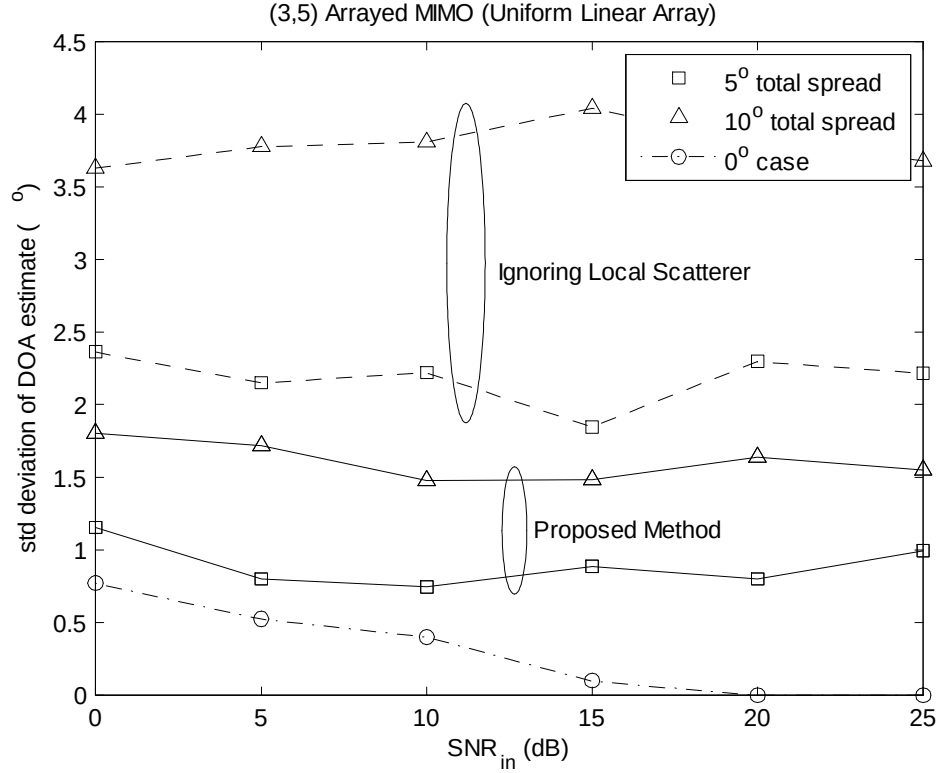


Figure 4.6: Standard deviation of DOA estimates versus input SNR for spatial spreads of 5° and 10° (ULA).

ceiver weight vector for the *point-source* case is based on Equation 3.25 in Chapter 3, while the proposed reception approach employs the *diffused* MC-STAR vector defined in Equation 4.9. The performance of the diffused channel framework will be evaluated using the metric SNIR_{out} (Signal-to-Noise-plus-Interference Ratio) at the receiver output.

Consider a single-user (3,5) cyclic prefix-free MC-CDMA arrayed MIMO system using ULAs with half-wavelength inter-element spacing at both the transmitter and receiver. The channel is diffused spatially with 6° spread, and assume that the nominal DOAs and TOAs of the multipath clusters have been estimated using the proposed approach described in Section 4.4. SNIR_{out} is calculated by averaging over 100 Monte Carlo runs. Figure 4.8 depicts the SNIR_{out} at the receiver output as a function of the input Signal-to-Noise Ratio (SNR). The SNIR_{out} of the linear space-time decorrelating receiver based on the proposed diffused chan-

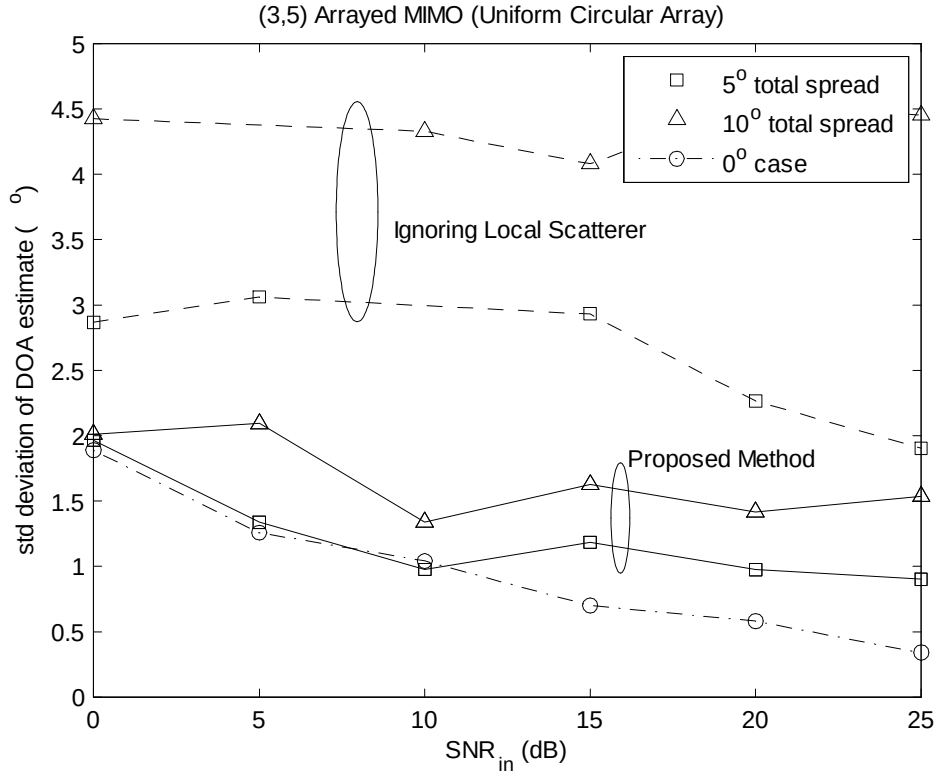


Figure 4.7: Standard deviation of DOA estimates versus input SNR for spatial spreads of 5° and 10° (UCA).

nel model increases linearly with input SNR. Although the channel is severely diffused spatially, the proposed approach models the diffused channel accurately to cancel the diffused channel interference and yield good performance. However, the decorrelating receiver based on point-source channel model starts to deviate from a linear trend at a relatively low input SNR of around 15 dB, due to uncanceled diffused channel interference.

The impact of varying degrees of spatial diffusion on the SNIR_{out} performance of MC-CDMA arrayed MIMO reception is now considered. Assuming an input SNR of 20dB, the spatial spread is varied from 0° to 10° and the SNIR_{out} is observed at the receiver output, as shown in Figure 4.9. It can be seen that reception based on the diffused STAR framework is robust to any changes in the spatial spread. Point-source channel model does not take into account the spatial diffusion of the multipath clusters, and hence fails to model the wireless

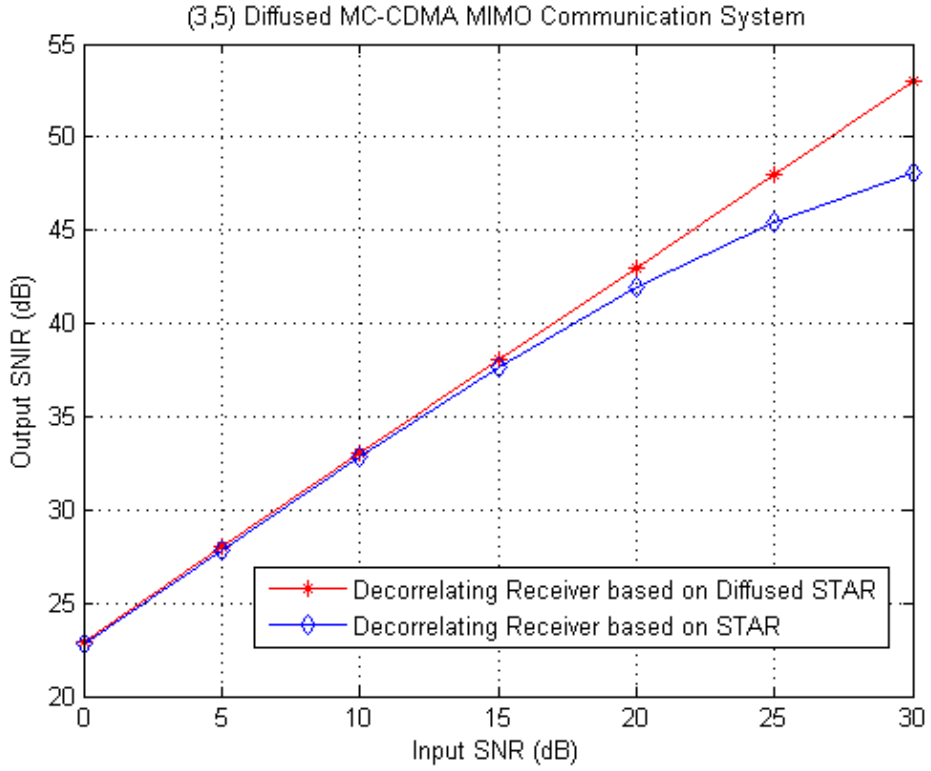


Figure 4.8: $SNIR_{out}$ versus input SNR plots for decorrelating reception based on diffused STAR and point-source STAR in a channel with 6° spatial diffusion.

channel accurately. Therefore, reception based on the point-source STAR shows degradation with increasing spatial spread. On the other hand, the diffused channel model considers the spatial diffusion of the multipath clusters and gives better $SNIR_{out}$ performance.

Finally, the near-far performance of the proposed diffused STAR based reception is studied by considering the uplink of a multiple access MC-CDMA arrayed MIMO system, with 3 co-channel arrayed users communicating with a receiver array. The transmitters are each equipped with 3-element ULAs, while the receiver ULA contains 5 antenna elements. An input SNR of 20dB and spatial diffusion with 6° spread is assumed. Figure 4.10 shows the variation of the $SNIR_{out}$ as the signal power of the interfering arrayed users increases exponentially. It can be observed the power imbalance causes the decorrelating receiver $SNIR_{out}$ to depreciate. However, diffused STAR-based reception comprehensively outper-

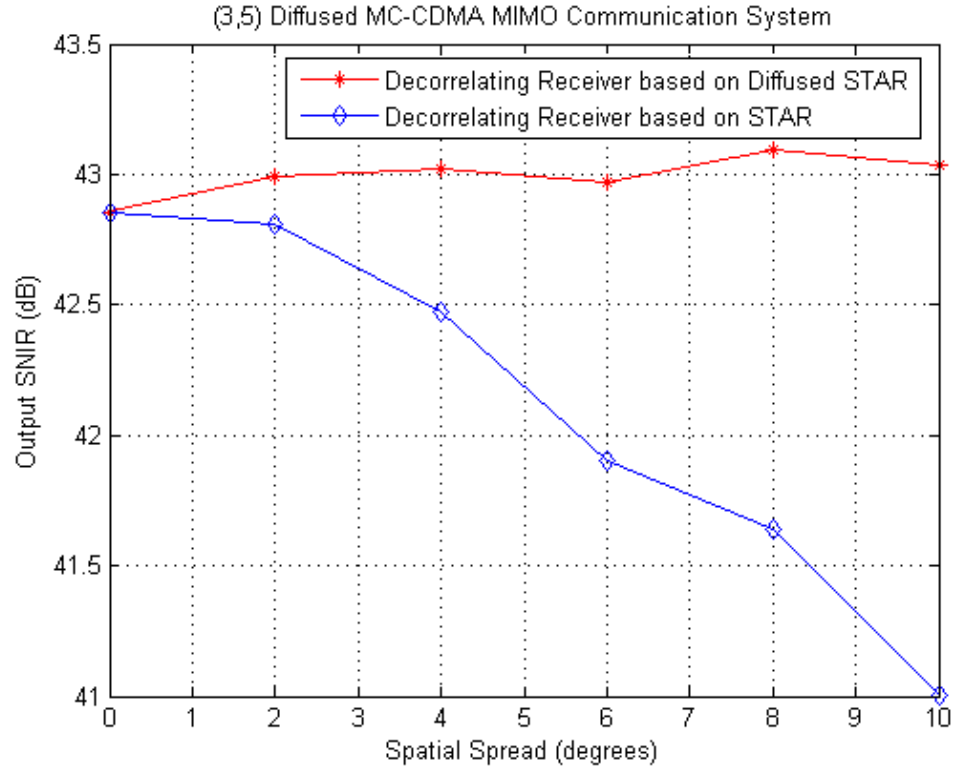


Figure 4.9: $SNIR_{out}$ versus spatial spread for decorrelating reception based on diffused STAR and point-source STAR at input SNR of 20dB (ULA).

forms point-source STAR-based reception, with a 40dB performance advantage at moderate to high power imbalances.

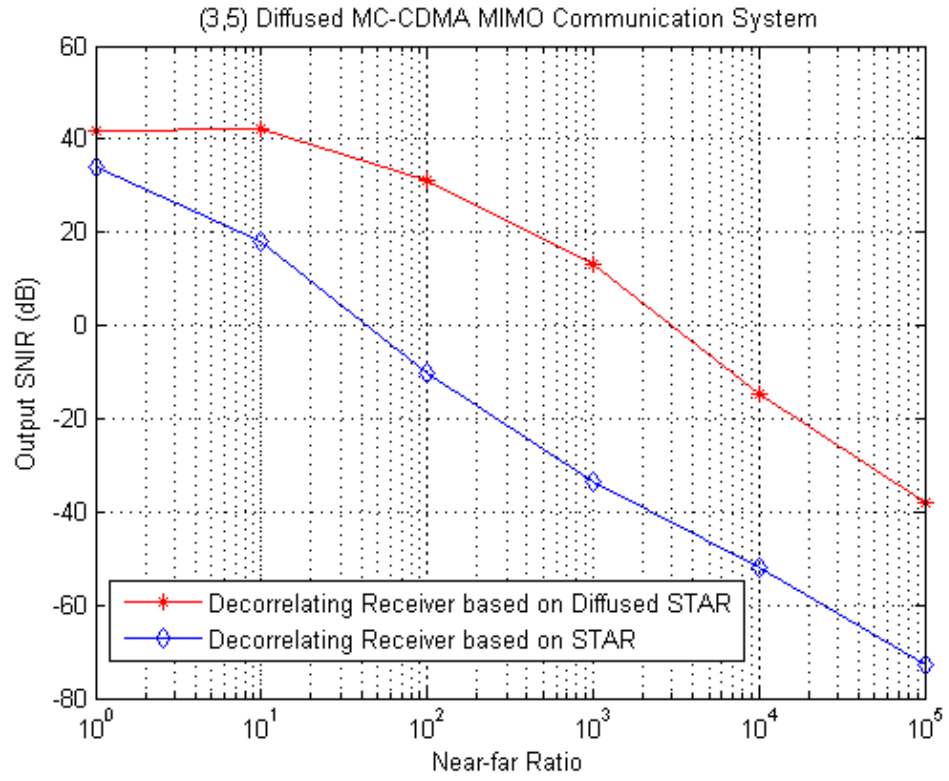


Figure 4.10: $SNIR_{out}$ versus near-far ratio for decorrelating reception based on diffused STAR and point-source STAR at input SNR of 20dB (ULA) and 6° spatial diffusion.

4.7 Summary

In this chapter, a diffused channel framework for a cyclic prefix-free MC-CDMA arrayed MIMO communication system is presented. Localized scattering is assumed to occur for each multipath; hence the wireless channel is modeled as a spatially diffused vector channel. Thus, the point source MC-STAR manifold vector is extended to the diffused MC-STAR manifold vector so that the effects of the spatially diffused vector channel can be accurately modeled. A robust blind estimation method is presented to estimate the parameters of the spatially diffused channel, followed by reception based on these parameters.

The proposed subspace-based channel estimation algorithm jointly estimates DOAs, TOAs and spread angles of spatially diffused multipaths of a user in a cyclic prefix-free MC-CDMA arrayed MIMO system. This is achieved by making use of the structure of the diffused channel. Moreover, it is shown that the proposed channel estimation method has better performance than an estimator which ignores the localized scattering in the channel. However, the performance of the proposed method is limited by the interference resulting from the other antennas at the transmitter terminal.

Point-source reception that ignores the presence of spatial diffusion is also compared to reception based on the diffused channel model. A linear space-time decorrelating receiver is proposed and reception based on the proposed diffused channel approach yields better SNIR_{out} performance than reception that ignores the presence of spatial diffusion.

Chapter 5

Joint Beamforming in Downlink Cyclic Prefix-free MC-CDMA Arrayed MIMO System

The work in this chapter extends the single user cyclic prefix-free MC-CDMA arrayed MIMO system (i.e. a point-to-point communication system) to a multi-user system and focuses on the downlink of a cyclic prefix-free MC-CDMA arrayed MIMO communications system where the base station transmits simultaneously to multiple users. Due to the interference present in such a system, joint optimization of the transmitter and receiver beamforming weights is used to improve the performance of the system. The joint optimization of the transmitter and receiver weights allows the transmitter pre-coding and receiver decoding weights to be optimized for the system conditions, thus enabling a higher data throughput for the system. Thus, it offers an advantage over transmitter-only, or receiver-only beamforming. Moreover, the computational burden can be distributed over both ends of the transmission link.

An iterative joint optimization algorithm, which seeks to minimize the overall MSE of the multi-user system, is thus developed in this chapter. In contrast

to previous joint optimization algorithms that have been developed for MIMO systems, the proposed algorithm takes into consideration the geometry of the antenna array. Thus, the inherent space-time structure of the multipath and multiple access channels can be exploited more effectively with the added information provided by the array manifold vector. Moreover, instead of a *subcarrier-noncooperative* approach, a *subcarrier-cooperative* approach is taken here. This implies that the subcarriers are not required to be orthogonal. Thus, there can be a flexibility in the allocation of channel symbols among the subcarriers and transmitter antennas to overcome channel fading effects.

Due to the multi-carrier nature of the system, the proposed algorithm also seeks to reduce the probability of clipping of the transmitted signals from each antenna. This is achieved by making use of the relation between the probability of clipping and the average transmit signal power of each antenna. The joint optimization of the precoding and equalization matrices is thus performed with the goal of minimizing the overall MSE of the system, subject to a power constraint of the transmitter power of each antenna element.

5.1 Introductory Background

In the downlink of a multi-user MIMO system, a base station communicates to multiple users simultaneously, i.e. a point-to-multi-point system. This results in interference among the users. A method to improve the performance of each receiver is through the implementation of multi-user detection (MUD) techniques [75] at each receiver to improve the performance of the detection process. However, such techniques are computationally intensive and they are thus not suited for implementation in the receivers where computational power is limited.

Thus, instead of the implementation of MUD techniques at each of the re-

ceivers, the interference between users can be reduced through the use of transmit precoding when channel information is available at the transmitter. The channel information can be obtained in a number of ways. For instance, the channel reciprocity of time division duplex (TDD) systems can be used to obtain the channel information from the channel information estimated during the uplink. Alternatively, the channel information can be fed back from the receivers to the transmitter through a feedback channel. The use of a feedback channel implies that system overheads are increased albeit with a potential improvement in system performance. Transmit precoding (or beamforming) has the advantage of transferring the computational load from the receivers to the base station which is better equipped to perform computationally intensive algorithms. Thus, the receivers can then be simplified and make use of simple detection techniques. This leads to a reduction in power consumption in the receivers.

A nonlinear approach to transmit precoding can be found in the form of dirty paper coding (DPC) developed by Costa [76] which was shown to achieve capacity for the multi-user MIMO system in [77]. In such a method, the transmitter, being pre-equipped with the knowledge of the interference in a channel, designs a code to pre-compensate for the interference. However, DPC-based methods make use of unconventional coding techniques such as non-linear coding and high-dimensional lattices which leads to increased complexity at both the transmitter and receivers. As such, linear transmit beamforming is a more attractive option.

Transmit precoding (or beamforming) has been studied extensively for multi-user MIMO systems. One common approach is through the use of zero-forcing (ZF) criteria in the design of the precoding matrix. Such methods have been separately proposed in [78] by Spencer *et al* and in [79] by Choi and Murch. The objective of ZF precoding is to null the interference that is present at each receiver. This is thus equivalent to block diagonalization (BD) which decomposes

a multi-user MIMO downlink channel into multiple parallel independent single-user MIMO downlink channels. However, ZF-based precoding is limited by the restriction that the number of transmit antennas must be larger than the sum of receive antennas. This condition is required in order to provide the necessary degrees of freedom to ensure that the interference is nulled at each receiver and also results in a limit to the number of users that can be supported simultaneously. In order to overcome the restriction on the number of transmit and receive antennas in such ZF-based precoding schemes, Tarighat *et al* [80] propose to minimize the interference instead of nulling it and the proposed algorithm determines the transmit beamforming vectors by solving a generalized eigendecomposition problem. Alternatively, as proposed by Yoo and Goldsmith in [81], the users can be separated into different groups so that the restriction on the number of antennas is met by each group and these groups are further separated by assigned time slots. The application of transmit beamforming has also been explored in the downlink of MC-CDMA systems. For instance, in [82], Sälzer and Mottier show that the application of transmit beamforming to separate the users in space for the downlink of a multi-user MISO MC-CDMA system combined with simple detection in the receivers resulted in a superior performance when compared with conventional systems using MUD. A *subcarrier-noncooperative approach*, in which beamforming is performed on each subcarrier separately, is used in the method proposed by Sälzer and Mottier.

In order to further improve the system performance, joint optimization of the transmitter and receiver weights is a further development that is explored. The joint optimization of the transmitter and receiver weights allows the transmitter pre-coding and receiver decoding weights to be optimized for the system conditions, thus enabling a higher data throughput for the system. The optimization problem for single user MIMO systems has been investigated in [83,84,85]. In [83],

Yang and Roy propose a joint transmitter and receiver optimization for MIMO systems based on the MSE criterion and it is shown to have a much better performance than systems that employ receiver optimization only. In [84], Sampath *et al* consider a weighted MMSE criterion and the optimum transmit and receive matrices are determined through the eigendecomposition of the channel. A variety of design criteria are unified in a generalized framework in [85] by Palomar *et al* for the joint optimization process of a multi-carrier MIMO system. The optimization problem is formulated using the framework of convex optimization theory and it is further shown that a *subcarrier-cooperative approach*, instead of a *subcarrier-noncooperative approach*, can be used to obtain an improvement in performance.

The extension of joint transmit and receive beamformer design to a multi-user MIMO system is investigated in [86] by Tenenbaum and Adve and an iterative procedure is obtained based on the MMSE criterion. However, the proposed algorithm requires that the total number of data streams must be less than the number of transmit antennas. In [87], Zhao and Chan consider the joint optimization problem for the downlink of a multi-user MIMO MC-CDMA system with a *subcarrier-noncooperative approach* as the orthogonality of the subcarriers is maintained. In the optimization process, the transmit and receive beamforming weights are calculated alternately using second order cone programming (SOCP) algorithm and constrained optimization, respectively.

The downlink multi-user cyclic prefix-free MC-CDMA arrayed MIMO system is considered here and joint optimization of the transmitter precoding and receiver decoding weights is explored in order to improve the performance of the multi-user system. The use of antenna arrays enables the geometrical information of the antennas and channel to be leveraged upon as array signal processing techniques can be used. This is in contrast to the systems in [78, 88] which

ignore the antenna array geometry. The precoding and decoding matrices are designed to minimize the overall mean-squared-error (MSE) of the system subject to a transmit power constraint. In addition, for multi-carrier based systems, the peak-to-average power ratio (PAPR) is an important area of concern as the peak transmit power, often limited by regulatory or application restrictions [89], would have to be clipped when the threshold is exceeded. The PAPR, and hence average transmit power, of multi-carrier signals is closely related to the probability of clipping [90, 91]. Furthermore, it has been shown in [85] that the probability of clipping of an OFDM signal can be reduced through the imposition of a constraint on the transmitted power. Thus, the optimization process also seeks to reduce the probability of clipping of the transmitted signals from each antenna by making use of the relation between the probability of clipping and the average transmit signal power of each antenna.

5.2 Downlink Cyclic Prefix-free MC-CDMA Arrayed MIMO Transmitter

An overview of the downlink of a multi-user MIMO system is shown in Figure 5.1 where, for illustration, 3 users are depicted and the base station is simultaneously transmitting to the group of users. It can be observed that each user is affected by co-channel interference due to the transmitted signals for the other users. Joint optimization of the transmitter and receiver weights is thus required to minimize this inter-user interference.

The system architecture of a downlink cyclic prefix-free MC-CDMA arrayed MIMO system is shown in Figure 2.13. As shown in Figure 2.13, the base station is equipped with an antenna array of $\overline{N}$ elements, while each receiver terminal has an N element antenna array. In the base station, MC-CDMA modulation

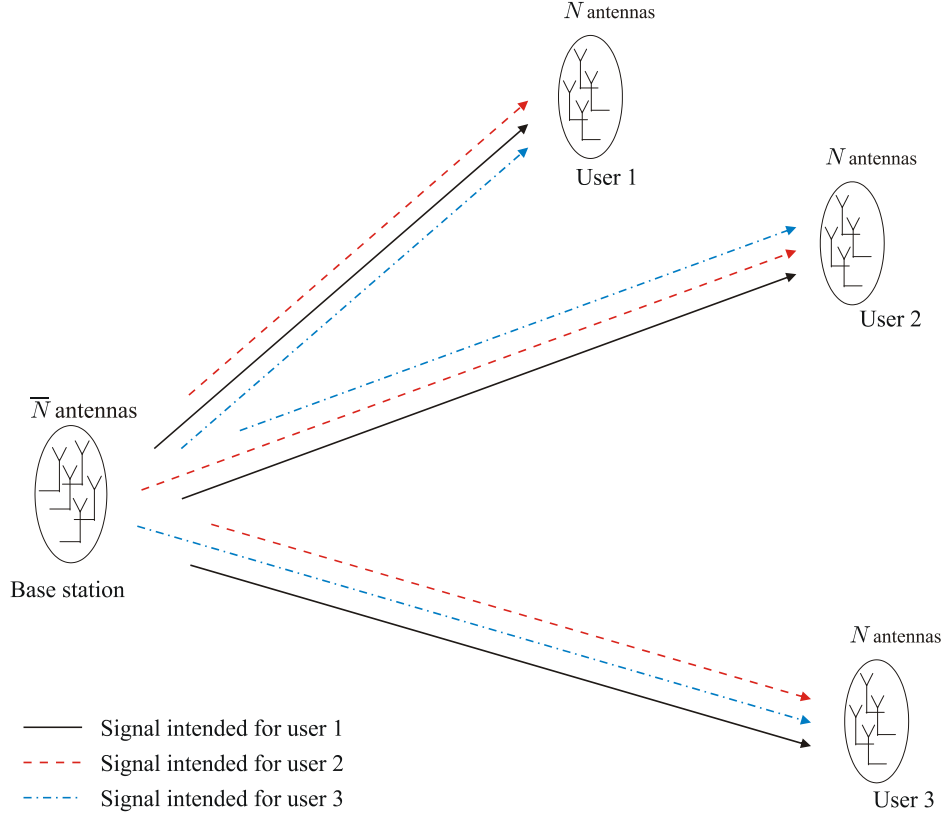


Figure 5.1: An illustration of a multi-user MIMO downlink with 3 users.

and transmit beamforming is performed on each user's channel symbols and the resulting vector signals are then summed up for transmission. Due to the implementation of transmit beamforming in the base station, it is now necessary to redevelop the cyclic prefix-free MC-CDMA modulation in the transmitter to include the transmit beamforming process.

The details of the MC-CDMA and transmit precoding that is carried out for each user is shown in Figure 5.2. Close inspection of Figure 5.2 and comparing it with Figure 3.2 in Section 3.2 reveals that there is an additional multiplication procedure in the MC-CDMA modulation and transmit beamforming process considered here due to the linear precoding that is used. In addition, code diversity is not implemented in the base station to differentiate the transmitted signals from each transmitter antenna.

As shown in Figure 5.2, we demultiplex the incoming channel symbols (e.g.

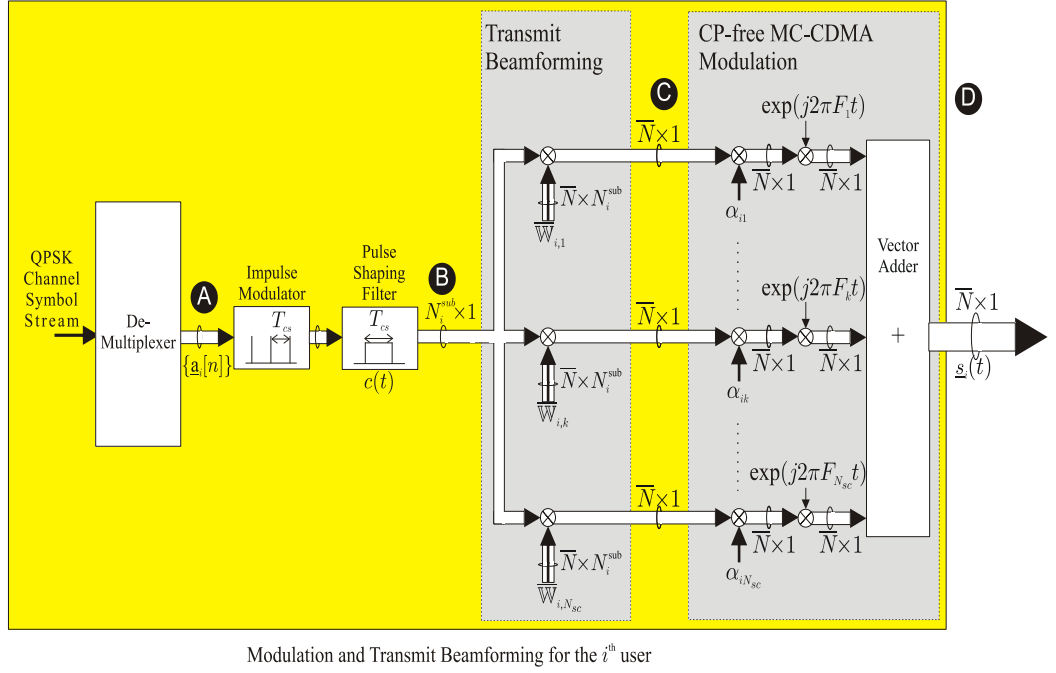


Figure 5.2: MC-CDMA modulation and transmit beamforming for the i^{th} user.

QPSK symbols) among the transmitter antenna elements. In particular, the QPSK channel symbol stream of the i^{th} user is demultiplexed into N_i^{sub} streams, where N_i^{sub} is the number of substreams of the i^{th} user. This results in the n^{th} channel symbol vector, at point-A, given by $\underline{a}_i[n] = [a_{i1}[n], a_{i2}[n], \dots, a_{iN_i^{\text{sub}}}[n]]^T$. Note that this is different from the MIMO systems that have been developed in Chapters 3 and 4 where $N_i^{\text{sub}} = \bar{N}$ as spatial multiplexing is assumed. The vector-symbol sequence $\{\underline{a}_i[n], \forall n\}$ is then converted into the vector-signal at point-B, defined as:

$$\sum_{n=-\infty}^{n=+\infty} \underline{a}_i[n] c(t - nT_{cs}), \quad nT_{cs} \leq t < (n+1)T_{cs} \quad (5.1)$$

where $c(t)$ is the rectangular pulse of duration T_{cs} . The vector-signal is first copied into N_{sc} parallel streams and the k^{th} copy of the signal given by Equation 5.1 then undergoes transmit beamforming through the application of the transmit beamforming matrix of the i^{th} user at the k^{th} subcarrier, $\bar{\mathbb{W}}_{i,k} \in \mathcal{C}^{\bar{N} \times N_i^{\text{sub}}}$, to obtain the signal at point-C of Figure 5.2. Cyclic prefix-free MC-CDMA modulation is then performed by multiplying the k^{th} branch with a corresponding

k^{th} chip $\alpha_{ik} = \pm 1$ of $\underline{\alpha}_i = [\alpha_{i1}, \dots, \alpha_{ik}, \dots, \alpha_{iN_{sc}}]^T$, which is the PN sequence of length N_{sc} corresponding to the i^{th} user. The resultant vector is then applied to a subcarrier with frequency given by $F_k = k\Delta F$, where $\Delta F = 1/T_{cs}$ is the subcarrier separation. The modulated subcarriers are then summed up to produce the baseband MC-CDMA vector-signal $\underline{s}_i(t) \in \mathcal{C}^{\bar{N} \times 1}$:

$$\underline{s}_i(t) = \sum_{n=-\infty}^{n=+\infty} \underline{a}_i[n] \sum_{k=1}^{N_{sc}} \alpha_{ik} \bar{\mathbb{W}}_{i,k} \exp(j2\pi F_k t), \quad nT_{cs} \leq t < (n+1)T_{cs} \quad (5.2)$$

In the case where there are M users, the baseband transmitted signal will then be given by

$$\underline{m}(t) = \sum_{i=1}^M \underline{s}_i(t) \quad (5.3)$$

The baseband signal $\underline{m}(t)$ is upconverted, using a carrier frequency F_c , to a bandpass signal given by

$$\underline{m}(t) \exp(j2\pi F_c t + \psi) \quad (5.4)$$

However, without any loss of generality, the carrier $\exp(j2\pi F_c t)$ at the transmitter and $\exp(-j2\pi F_c t)$ at the receiver is ignored and baseband transmission is assumed. Thus, based on Equation 5.3, the total transmitted power of M users can be expressed as

$$\bar{P} = \sum_{i=1}^M \text{tr} \left(\bar{\mathbb{W}}_i \bar{\mathbb{W}}_i^H \right) \quad (5.5)$$

where

$$\bar{\mathbb{W}}_i = \begin{bmatrix} \bar{\mathbb{W}}_{i,1} \\ \vdots \\ \bar{\mathbb{W}}_{i,k} \\ \vdots \\ \bar{\mathbb{W}}_{i,N_{sc}} \end{bmatrix} \quad (5.6)$$

Also, the total transmitted power from the p^{th} transmit antenna can be expressed as

$$\bar{P}_p = \sum_{i=1}^M \text{tr} \left(\bar{\mathbb{W}}_i \bar{\mathbb{W}}_i^H \left(\mathbb{I}_{N_{sc}} \otimes \underline{\mathcal{L}}_{\bar{N},p} \underline{\mathcal{L}}_{\bar{N},p}^T \right) \right) \quad (5.7)$$

where $\underline{\ell}_{\bar{N},p}$ denotes a $\bar{N} \times 1$ vector with 1 at the p^{th} element and 0s elsewhere.

5.2.1 PAPR and Transmit Power Requirement

In order to satisfy the requirement of reducing the probability of clipping of the transmitted signal to an adequate level, it is now necessary to obtain a relationship between the transmitted power $\bar{P}_p$ and the probability of clipping of the transmitted signal, which is defined as

$$\Pr \left(\text{PAPR} > \frac{A_{clip}^2}{\bar{P}_p} \right) = \mathcal{P} \quad (5.8)$$

where A_{clip} denotes the amplitude of the transmitted signal at which clipping of the transmitted signal will occur. In the downlink multi-user MIMO MC-CDMA system considered here, the PAPR of the transmitted signal from the p^{th} transmit antenna can be defined as

$$\text{PAPR} = \frac{\max_{0 \leq t < T_{cs}} |[\underline{m}]_P(t)|^2}{\frac{1}{T_{cs}} \int_0^{T_{cs}} |[\underline{m}]_P(t)|^2 dt} = \left(\frac{\text{peak value}}{\text{rms value}} \right)^2 \quad (5.9)$$

where $[\underline{m}]_P(t)$ denotes the p^{th} component of the baseband transmitted vector signal $\underline{m}(t)$. However, instead of performing the integration in Equation 5.9, sampling of the transmitted signal $[\underline{m}]_P(t)$ can be carried out to obtain an approximation to the PAPR of the continuous time signal. Hence, the transmitted signal is sampled at a rate of $\frac{1}{T_s} = \frac{\bar{q}N_{sc}}{T_{cs}}$ and $\bar{L} = \bar{q}N_{sc}$ samples are obtained per channel symbol period. As a result of the sampling, an approximate of the PAPR can now be expressed as:

$$\begin{aligned} \text{PAPR} &= \frac{\max_{0 \leq \bar{\ell} < \bar{L}-1} \left| [\underline{m}]_p [\bar{\ell} \bar{T}_s] \right|^2}{\mathcal{E} \left\{ \left| [\underline{m}]_p [\bar{\ell} \bar{T}_s] \right|^2 \right\}} \\ &= \frac{\max_{0 \leq \bar{\ell} < \bar{L}-1} \left| [\underline{m}]_p [\bar{\ell} \bar{T}_s] \right|^2}{\bar{P}_p} \end{aligned} \quad (5.10)$$

where $[\underline{m}]_p [\bar{\ell} \bar{T}_s]$ denotes the $\bar{\ell}^{\text{th}}$ sample of the sampled signal $\left\{ [\underline{m}]_p [\bar{\ell} \bar{T}_s], \bar{\ell} \in 0, 1, \dots, (\bar{L}-1) \right\}$.

The assumption of a Rayleigh distribution for $\left| [\underline{m}]_p [\bar{\ell} \bar{T}_s] \right|$ [89] results in

$$\Pr \left(\text{PAPR} > \frac{A_{clip}^2}{\bar{P}_i} \right) = 1 - \left(1 - \exp \left(-\frac{A_{clip}^2}{\bar{P}_i} \right) \right)^{\bar{L}}$$

Thus, the constraint for the total transmitted power from the p^{th} transmit antenna, $\bar{P}_p$, can be written as

$$\bar{P}_p = \frac{A_{clip}^2}{\ln \left(\frac{1}{1 - (1 - \mathcal{P})^{\frac{1}{\bar{L}}}} \right)} \quad (5.11)$$

Due to the assumption of a Rayleigh distribution for $\left| [\underline{m}]_p [\bar{\ell} \bar{T}_s] \right|$, the above expression in Equation 5.11 is obtained for the power constraint $\bar{P}_p$ instead of the expression in [85] which is based on a Gaussian assumption.

5.2.2 Received Signal

As shown in Figure 2.13, the wireless channel from the base station to each user corresponds to a VIVO channel which is assumed to be multipath dispersive with a maximum delay spread of T_{cs} . Thus, the baseband received signal $\underline{x}_{\mathbf{m}}(t)$ at the $\mathbf{m}^{\text{th}}$ user can be written as

$$\underline{x}_{\mathbf{m}}(t) = \sum_{j=1}^{K_{\mathbf{m}}} \beta_{\mathbf{m}j} \underline{S}_{\mathbf{m}j} \bar{\underline{S}}_{\mathbf{m}j}^H \underline{m}(t - \tau_{\mathbf{m}j}) + \underline{n}_{\mathbf{m}}(t) \quad (5.12)$$

where $\tau_{\mathbf{m}j}$ and $\beta_{\mathbf{m}j}$ represents the path delay and fading coefficient of the j^{th} path with $j \in 1, 2, \dots, K_{\mathbf{m}}$ and $K_{\mathbf{m}}$ is the number of resolvable multipaths from the base station to the $\mathbf{m}^{\text{th}}$ user. Also, $\underline{n}_{\mathbf{m}}(t)$ denotes the AWGN of power $\sigma_{\mathbf{m}}^2$ present in the channel.

In each receiver, a discretizer and tapped delay lines with length equivalent to $2T_{cs}$ are implemented and as a result, the discretized received signal $\underline{x}_{\mathbf{m}}[n]$ can

be expressed as

$$\begin{aligned} \underline{x}_m[n] = & \sum_{i=1}^M \sum_{j=1}^{K_m} \sum_{k=1}^{N_{sc}} \beta_{mj} \underline{S}_{mj} \otimes \\ & \left(\begin{aligned} & \alpha_{ik} (\mathbb{J}^T)^{N_{sc}} \mathbb{J}^{\ell_{mj}} \begin{bmatrix} \underline{f}_k[\ell_{mj}] \\ \underline{0}_{N_{sc}} \end{bmatrix} \bar{\underline{S}}_{mj}^H (\bar{\underline{W}}_{i,k} \underline{a}_i[n-1]) \\ & + \alpha_{ik} \mathbb{J}^{\ell_{mj}} \begin{bmatrix} \underline{f}_k[\ell_{mj}] \\ \underline{0}_{N_{sc}} \end{bmatrix} \bar{\underline{S}}_{mj}^H (\bar{\underline{W}}_{i,k} \underline{a}_i[n]) \\ & + \alpha_{ik} (\mathbb{J})^{N_{sc}} \mathbb{J}^{\ell_{mj}} \begin{bmatrix} \underline{f}_k[\ell_{mj}] \\ \underline{0}_{N_{sc}} \end{bmatrix} \bar{\underline{S}}_{mj}^H (\bar{\underline{W}}_{i,k} \underline{a}_i[n+1]) \end{aligned} \right) \\ & + \underline{n}_m[n] \end{aligned} \quad (5.13)$$

A single path of the received signal (j^{th} path, say) from the i^{th} user can thus be written as:

$$\begin{aligned} & j^{\text{th}} \text{ path of } i^{\text{th}} \text{ user's contribution to } \underline{x}_m[n] \\ = & \sum_{k=1}^{N_{sc}} \beta_{mj} \underline{S}_{mj} \otimes \left(\alpha_{ik} \mathbb{J}^{\ell_{mj}} \begin{bmatrix} \underline{f}_k[\ell_{mj}] \\ \underline{0}_{N_{sc}} \end{bmatrix} \bar{\underline{S}}_{mj}^H (\bar{\underline{W}}_{i,k} \underline{a}_i[n]) \right) \\ = & \beta_{mj} \underline{S}_{mj} \otimes \left(\mathbb{J}^{\ell_{mj}} \begin{bmatrix} \mathbb{F}_{\ell_{mj}} \text{diag}(\underline{\alpha}_i) \\ \mathbb{O}_{N_{sc} \times N_{sc}} \end{bmatrix} \right) (\mathbb{I}_{N_{sc}} \otimes \bar{\underline{S}}_{mj}^H) \bar{\underline{W}}_{i,k} \underline{a}_i[n] \\ = & \beta_{mj} \mathbb{H}_{m,ij} \bar{\underline{W}}_{i,k} \underline{a}_i[n] \end{aligned} \quad (5.14)$$

where for the sake of clarity, only the current channel symbol vector $\underline{a}_i[n]$ is considered and

$$\mathbb{H}_{m,ij} = \underline{S}_{mj} \otimes \left(\mathbb{J}^{\ell_{mj}} \begin{bmatrix} \mathbb{F}_{\ell_{mj}} \text{diag}(\underline{\alpha}_i) \\ \mathbb{O}_{N_{sc} \times N_{sc}} \end{bmatrix} \right) (\mathbb{I}_{N_{sc}} \otimes \bar{\underline{S}}_{mj}^H) \quad (5.15)$$

denotes the MC-STAR manifold matrix for the j^{th} path of i^{th} user, as observed at the m^{th} user, while $\mathbb{F}_{\ell_{mj}} = [\underline{f}_1[\ell_{mj}], \dots, \underline{f}_k[\ell_{mj}], \dots, \underline{f}_{N_{sc}}[\ell_{mj}]]$. $\mathbb{H}_{m,ij}$ is a function of the receive and transmit array manifold vectors, $\underline{S}_{mj}$ and $\bar{\underline{S}}_{mj}$, respectively, and the path delay ℓ_{mj} for the j^{th} path as observed at the m^{th} user as well as

the PN sequence of the i^{th} user. Thus, $\mathbb{H}_{\mathbf{m},ij}$ encompasses the spatial-temporal channel characteristics (DOA and TOA) for the received signal at the $\mathbf{m}^{\text{th}}$ and blind channel estimation algorithms such as that in Chapter 3 can be used to obtain estimates of the spatial-temporal channel parameters and $\bar{\mathbb{W}}_i$ is as defined in Equation 5.6. Thus, the received signal at the $\mathbf{m}^{\text{th}}$ user can be written as

$$\begin{aligned} \underline{x}_{\mathbf{m}}[n] &= \sum_{i=1}^M \sum_{j=1}^{K_{\mathbf{m}}} \beta_{\mathbf{m}j} \begin{pmatrix} (\mathbb{J}^T)^{N_{sc}} \mathbb{H}_{\mathbf{m},ij} \bar{\mathbb{W}}_i \underline{\mathbf{a}}_i[n-1] \\ + \mathbb{H}_{\mathbf{m},ij} \bar{\mathbb{W}}_i \underline{\mathbf{a}}_i[n] \\ + (\mathbb{J})^{N_{sc}} \mathbb{H}_{\mathbf{m},ij} \bar{\mathbb{W}}_i \underline{\mathbf{a}}_i[n+1] \end{pmatrix} + \underline{\mathbf{n}}_{\mathbf{m}}[n] \\ &= \sum_{i=1}^M \left(\mathbb{V}_{\mathbf{m},i} (\mathbb{I}_3 \otimes \bar{\mathbb{W}}_i) \begin{bmatrix} \underline{\mathbf{a}}_i[n-1] \\ \underline{\mathbf{a}}_i[n] \\ \underline{\mathbf{a}}_i[n+1] \end{bmatrix} \right) + \underline{\mathbf{n}}_{\mathbf{m}}[n] \end{aligned} \quad (5.16)$$

where $\mathbb{V}_{\mathbf{m},i} = [\mathbb{H}_{\mathbf{m},i}^{\text{prev}} \quad \mathbb{H}_{\mathbf{m},i} \quad \mathbb{H}_{\mathbf{m},i}^{\text{next}}]$ and $\mathbb{H}_{\mathbf{m},i}$, $\mathbb{H}_{\mathbf{m},i}^{\text{prev}}$ and $\mathbb{H}_{\mathbf{m},i}^{\text{next}}$ are defined as

$$\begin{aligned} \mathbb{H}_{\mathbf{m},i} &= \sum_{j=1}^{K_{\mathbf{m}}} \beta_{\mathbf{m}j} \mathbb{H}_{\mathbf{m},ij} \\ \mathbb{H}_{\mathbf{m},i}^{\text{prev}} &= (\mathbb{I}_N \otimes (\mathbb{J}^T)^{N_{sc}}) \mathbb{H}_{\mathbf{m},i} ; \quad \mathbb{H}_{\mathbf{m},i}^{\text{next}} = (\mathbb{I}_N \otimes \mathbb{J}^{N_{sc}}) \mathbb{H}_{\mathbf{m},i} \end{aligned}$$

In the $\mathbf{m}^{\text{th}}$ receiver, the receive beamforming weight, $\mathbb{W}_{\mathbf{m}}^H$, is used to obtain the soft decision variable vector, $\hat{\underline{\mathbf{a}}}_{\mathbf{m}}[n]$, which is given by

$$\hat{\underline{\mathbf{a}}}_{\mathbf{m}}[n] = \mathbb{W}_{\mathbf{m}}^H \underline{x}_{\mathbf{m}}[n] \quad (5.17)$$

5.3 Joint Transmit-Receive Beamformer Optimization

The joint transmit-receive optimization algorithm is now presented and in the optimization process, it is assumed that the base station and receivers have perfect channel knowledge. The objective of the optimization process is to determine the

precoder and equalization matrices, $\overline{\mathbb{W}}_{\mathbf{m}}$ and $\mathbb{W}_{\mathbf{m}}$, respectively, for each user with a predetermined number of substreams for each user. Thus, based on the soft decision variable vector, $\hat{\mathbf{a}}_{\mathbf{m}}[n]$, in Equation 5.17, the MSE matrix of the $\mathbf{m}^{\text{th}}$ receiver can be expressed as

$$\begin{aligned} \mathbf{E}_{\mathbf{m}} &= \mathcal{E} \left\{ (\hat{\mathbf{a}}_{\mathbf{m}}[n] - \mathbf{a}_{\mathbf{m}}[n]) (\hat{\mathbf{a}}_{\mathbf{m}}[n] - \mathbf{a}_{\mathbf{m}}[n])^H \right\} \\ &= \mathbb{W}_{\mathbf{m}}^H \mathbb{R}_{xx,\mathbf{m}} \mathbb{W}_{\mathbf{m}} + \mathbb{I}_{N_{\text{sub}}} - \mathbb{W}_{\mathbf{m}}^H \mathbb{H}_{\mathbf{m},\mathbf{m}} \overline{\mathbb{W}}_{\mathbf{m}} - \overline{\mathbb{W}}_{\mathbf{m}}^H \mathbb{H}_{\mathbf{m},\mathbf{m}}^H \mathbb{W}_{\mathbf{m}} \end{aligned} \quad (5.18)$$

where

$$\begin{aligned} \mathbb{R}_{xx,\mathbf{m}} &= \mathcal{E} \left\{ \underline{x}_{\mathbf{m}}[n] \underline{x}_{\mathbf{m}}^H[n] \right\} \\ &= \sum_{i=1}^M \mathbb{V}_{\mathbf{m},i} \left(\mathbb{I}_3 \otimes \overline{\mathbb{W}}_i \overline{\mathbb{W}}_i^H \right) \mathbb{V}_{\mathbf{m},i}^H + \sigma_{\mathbf{m}}^2 \mathbb{I}_{2N_{sc}} \end{aligned} \quad (5.19)$$

denotes the received signal covariance matrix of the $\mathbf{m}^{\text{th}}$ receiver. The optimization objective of minimizing the MSE of all users can thus be expressed as the following constraint optimization problem

$$\begin{aligned} &\min_{(\overline{\mathbb{W}}_1, \dots, \overline{\mathbb{W}}_M; \mathbb{W}_1, \dots, \mathbb{W}_M)} \sum_{\mathbf{m}=1}^M \text{tr}(\mathbf{E}_{\mathbf{m}}) \\ \text{s.t. } &\sum_{\mathbf{m}=1}^M \text{tr} \left(\overline{\mathbb{W}}_{\mathbf{m}} \overline{\mathbb{W}}_{\mathbf{m}}^H \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\overline{N},p} \underline{\ell}_{\overline{N},p}^T \right) \right) \leq \overline{P}_p \quad p = 1, \dots, \overline{N} \end{aligned} \quad (5.20)$$

In addition, instead of the *subcarrier-noncooperative* approach, the *subcarrier-cooperative* approach is used here which implies that the subcarriers are not required to be orthogonal. Thus, there is increased flexibility in the allocation of channel symbols among the subcarriers and transmitter antennas to overcome channel fading effects [85].

The optimization problem stated in Equation 5.20 shows that the constraint involves limiting the total transmitted power from each transmitter antenna and this is in contrast to [92] which imposes a constraint on the total transmit power for a DS-CDMA system. In the following, it is assumed that both the transmitter

and receivers have perfect channel knowledge and so, by using the Lagrange Multiplier method, the cost function for the optimization problem can be expressed as:

$$J = \sum_{m=1}^M \text{tr}(\mathbf{E}_m) + \sum_{p=1}^{\bar{N}} \mu_p \left(\sum_{m=1}^M \text{tr} \left(\bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \right) - \bar{P}_p \right) \quad (5.21)$$

where $\mu_p, p = 1, \dots, \bar{N}$, denotes the Lagrange multiplier that has been introduced for each transmit power constraint.

In order to obtain the solution for the precoding matrix $\bar{\mathbb{W}}_m$ of the m^{th} receiver, it is assumed that the receive decoding matrix $\mathbb{W}_i, \forall i$ are fixed and thus, by evaluating $\left(\frac{\partial J}{\partial \bar{\mathbb{W}}_m} \right)_{\bar{\mathbb{W}}_m = \bar{\mathbb{W}}_m^*} = 0$, the solution $\bar{\mathbb{W}}_m^*$ is then given by

$$\begin{aligned} \bar{\mathbb{W}}_m^* = & \left\{ \mathbb{L}^H \left(\sum_{i=1}^M (\mathbb{V}_{i,m}^H (\mathbb{W}_i \mathbb{W}_i^H) \mathbb{V}_{i,m}) \right) \mathbb{L} \right. \\ & \left. + \sum_{p=1}^{\bar{N}} \mu_p \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \right\}^{-1} \mathbb{H}_{m,m}^H \mathbb{W}_m \end{aligned} \quad (5.22)$$

where $\mathbb{L} = \mathbb{1}_3 \otimes \mathbb{I}_{N_{sc}}$. The derivation of the solution $\bar{\mathbb{W}}_m^*$ can be found in Appendix B.1. Also, the solution for the Lagrange Multiplier for the p^{th} antenna, μ_p^* , is given by

$$\begin{aligned} \mu_p^* = & \frac{1}{\bar{P}_p} \sum_{m=1}^M \text{tr} \left\{ \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) (\mathbb{H}_{m,m}^H \mathbb{W}_m \bar{\mathbb{W}}_m^H \right. \\ & \left. - \mathbb{L}^H \left(\sum_{i=1}^M \mathbb{V}_{i,m}^H (\mathbb{W}_i \mathbb{W}_i^H) \mathbb{V}_{i,m} \right) \mathbb{L} \bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \right\} \end{aligned} \quad (5.23)$$

Conversely, the precoding matrices are fixed in order to obtain the optimal equalization matrices, which can be expressed as

$$\mathbb{W}_m^* = \mathbb{R}_{xx,m}^{-1} \mathbb{H}_{m,m} \bar{\mathbb{W}}_m \quad (5.24)$$

corresponding to the MMSE receiver. The derivation of the solutions for μ_p^* and $\mathbb{W}_m^*$ can be found in Appendix B.2 and B.3, respectively.

Based on the derivations for the precoder and equalization matrices as well as the Lagrangian Multipliers, an iterative algorithm is an attractive option since

the precoder and equalization matrices as well as the Lagrangian Multipliers are dependent on each other. Thus, the joint optimization process can be performed iteratively.

1. Initialization: (iterative index $i = 0$) the precoder matrix for each user,

$\overline{\mathbb{W}}_{\mathbf{m}}$ is first initialized as

$$\overline{\mathbb{W}}_{\mathbf{m}}^{(i=0)} = \sqrt{\frac{\overline{P}}{MN_{\mathbf{m}}^{\text{sub}}}} \mathbb{U}_{\mathbf{m}}^{N_{\mathbf{m}}^{\text{sub}}}$$

where the columns of $\mathbb{U}_{\mathbf{m}}^{N_{\mathbf{m}}^{\text{sub}}}$ are the general eigenvectors corresponding to the $N_{\mathbf{m}}^{\text{sub}}$ largest general eigenvectors of the matrix $\mathbb{H}_{\mathbf{m},\mathbf{m}}^H \mathbb{H}_{\mathbf{m},\mathbf{m}}$. The coefficient $\sqrt{\frac{\overline{P}}{MN_{\mathbf{m}}^{\text{sub}}}}$ is required to ensure an equal distribution of power to each user and substream.

2. The equalization matrix $\mathbb{W}_{\mathbf{m}}^{(i)}$ is then calculated for each user $\mathbf{m} = 1, \dots, M$ based on Equation. 5.24
3. Using the i^{th} precoder and equalization matrices, the Lagrangian multipliers $\mu_p^{(i)}$, $p = 1, \dots, \overline{N}$ are then calculated using Equation. 5.23.
4. The equalization matrices and Lagrangian multipliers obtained in the previous 2 steps are used to update the precoder matrices $\overline{\mathbb{W}}_{\mathbf{m}}^{(i+1)}$ by using Equation. 5.22.
5. Replace i with $i + 1$ and repeat Steps 2 – 4 until a convergence criterion is satisfied.

5.4 Simulation Studies

The performance of the proposed iterative joint optimization algorithm is investigated in this section. The transmitter is assumed to be equipped with an uniform linear array (ULA) of $\overline{N} = 4$ elements while the receivers are each equipped with a ULA of $N = 2$ elements, all with half-wavelength spacing. Each user is assigned a Gold sequence and $N_i^{\text{sub}} = 2 \forall i$. It is also assumed that the transmitted signal

arrives at each receiver via $K_i = 4$ multipaths. The DOD and DOA associated with each path are assumed to be uniformly distributed over $[0, 180^\circ)$ while the path delay is assumed to be uniformly distributed over $[0, (N_{sc} - 1)T_s]$. A block of 250 channel symbols is collected for processing at each time, during which the channel is assumed to be stationary and both the transmitter and receivers have access to the channel state information.

The convergence behavior of the proposed iterative algorithm is first studied for a cyclic prefix-free MC-CDMA arrayed MIMO system with $M = 6$ users, hereafter known as $(6, 4, 2)$, i.e. $(M, \bar{N}, N)$ cyclic prefix-free MC-CDMA arrayed MIMO system. The convergence behavior is investigated for different transmit SNR values, defined as $\frac{\bar{P}}{\sigma^2}$. Figure 5.3 shows the average MSE per substream per user as a function of the iterative process. It can be seen that the convergence property of the proposed iterative algorithm improves with the transmit SNR.

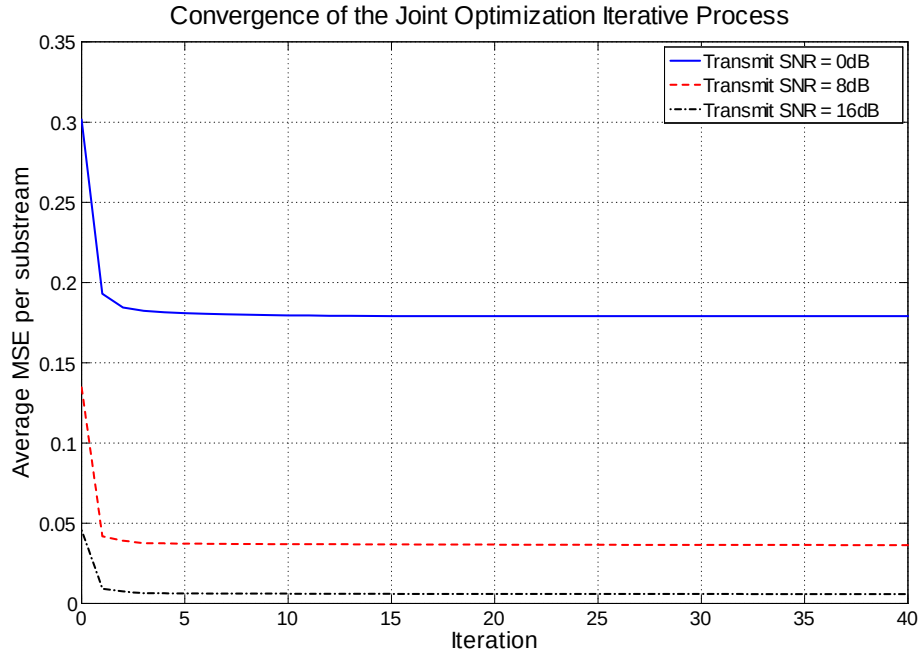


Figure 5.3: Study of convergence of the proposed algorithm with transmit SNR = 0, 8, 16dB in a $(6, 4, 2)$ cyclic prefix-free MC-CDMA arrayed MIMO system ($N_{sc} = 15$).

In Figure 5.4, the BER performance of the iterative joint optimization algo-

rithm is compared with the transmit precoding scheme described in Step 1 of the proposed algorithm coupled with equalization matrices based on RAKE and MMSE receivers for a $(6, 4, 2)$ MIMO MC-CDMA system while Figure 5.5 shows the BER performance for a $(10, 4, 2)$ MIMO MC-CDMA system. The precoder and decoder matrices are obtained from 30 iterations. In Figure 5.4, and 5.5, the effect of frequency diversity is also investigated by considering the use of Gold sequences of length 7 and 15 in the simulations. As shown in the figures, the iterative process results in a much better BER performance compared to the case where the precoder and decoder matrices are not obtained jointly. This is because of the adaptation of the precoder and decoder matrices, according to the system conditions, during the optimization process. In addition, it has been shown that with frequency diversity, the BER performance of the proposed iterative approach can be improved.

The probability of clipping of the transmitted MC-CDMA signal for various transmit SNR for a $(6, 4, 2)$ cyclic prefix-free MC-CDMA arrayed MIMO system with $N_{sc} = 15$ is next shown in Figure 5.6. It is assumed that the transmitted signal is oversampled by a factor of 4 so that the sampled transmitted signal can approximate the continuous time signal. It can be seen that the iterative process results in a lower probability of clipping of the transmitted signal compared to the non-iterative case.

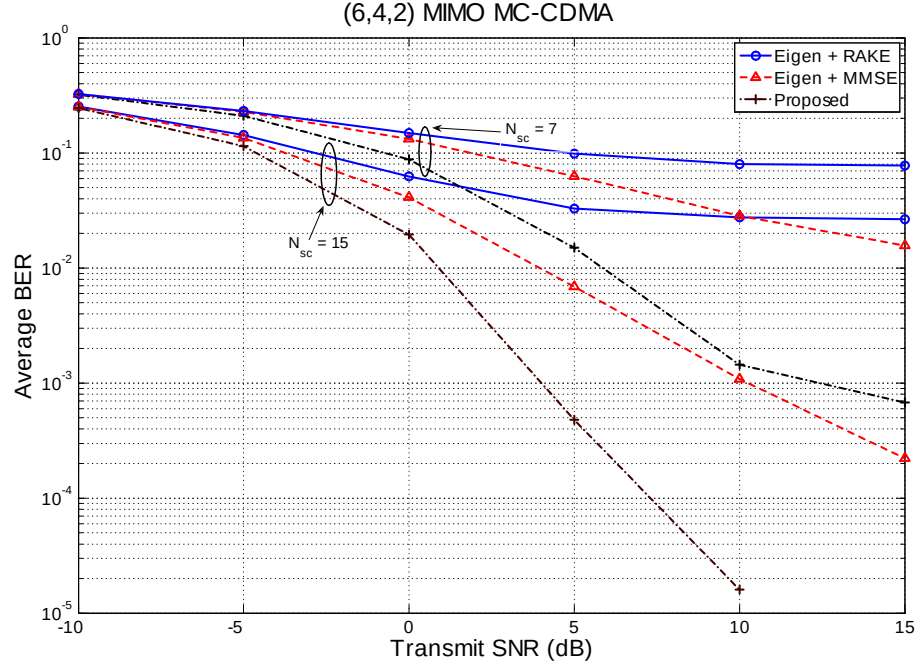


Figure 5.4: BER performance of the the proposed iterative algorithm and the eigendecomposition-based transmit precoding with RAKE or MMSE receiver equalization for a (6,4,2) cyclic prefix-free MC-CDMA arrayed MIMO system.

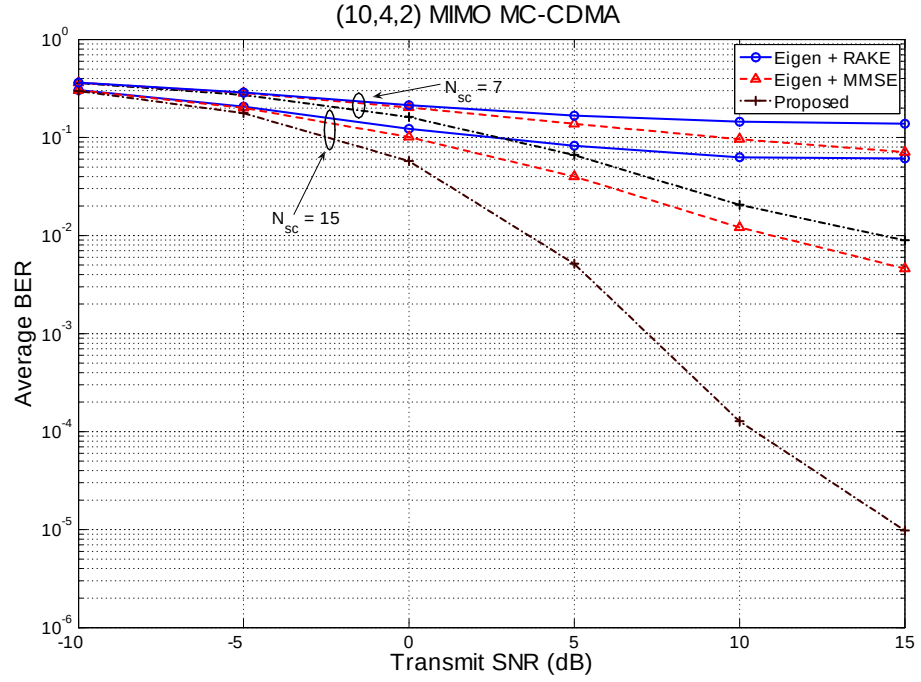


Figure 5.5: BER performance of the the proposed iterative algorithm and the eigendecomposition-based transmit precoding with RAKE or MMSE receiver equalization for a (10,4,2) cyclic prefix-free MC-CDMA arrayed MIMO system.

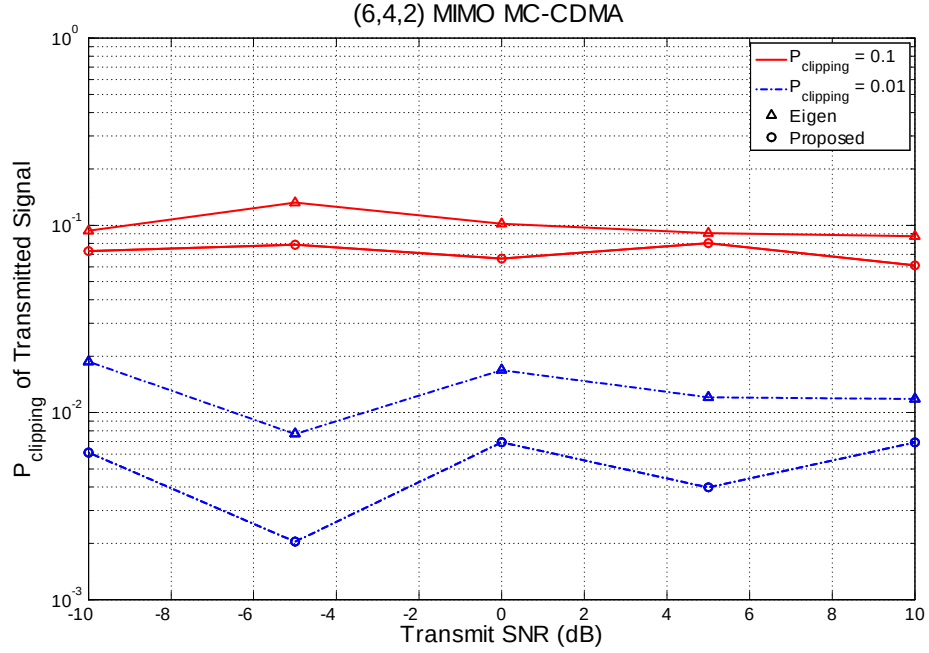


Figure 5.6: Average probability of clipping of transmitted MC-CDMA signal from each transmitter antenna with target $P_{\text{clipping}} = 0.1$ and 0.01 for various transmit SNR ($N_{sc} = 7$).

5.5 Summary

In this chapter, the problem of joint optimization of precoder and equalization matrices in a downlink cyclic prefix-free MC-CDMA arrayed MIMO system over multipath fading channels has been investigated. A *subcarrier-cooperative* approach has been taken in the optimization of the multi-carrier precoding matrices. This is in contrast with conventional multi-carrier systems where the orthogonality of the subcarriers is maintained and so, the processing is carried out independently for each subcarrier.

The proposed approach is based on minimizing the overall MSE of the system with a power constraint for each transmitter antenna, i.e. a per-antenna power constraint. Due to the link between the PAPR and the total transmit power for each transmitter antenna, a per-antenna power constraint enables the reduction of the probability of clipping of the transmitted MC-CDMA signals, which is an added advantage for multi-carrier systems.

In the proposed algorithm, the Lagrange Multiplier method is used to obtain solutions for the transmit and receive beamforming matrixes. The resulting solutions for the transmit and receive beamforming matrices as well as the Lagrange multipliers enables a iterative solution to be used as the equations obtained exhibit an interdependence.

Chapter 6

Conclusions and Future Work

In this thesis, the application of antenna arrays in MIMO systems is investigated with an emphasis on MC-CDMA-based systems. The combination of multi-carrier modulation with MIMO techniques is expected to be a key feature in future 4G systems due to the various benefits that are provided by each component and thus, MIMO MC-CDMA systems form the basis for the work in this thesis. However, in the MC-CDMA modulation scheme considered here, cyclic prefixes are not used so that the system overheads can be reduced. A gradual development of the array-based MIMO MC-CDMA systems is presented and the MIMO systems under consideration progress from a single user system, i.e. a point-to-point system, to a multi-user system. A summary of the work presented in the previous chapters is now provided together with a list of contributions. In addition, suggestions on the further development of the work that has been presented are given.

6.1 Thesis Summary

In Chapter 1, a discussion of the characteristics of MC modulation techniques is presented. In addition, a review of MIMO signaling techniques, such as spatial multiplexing, is provided. In order to harness the spatial information that is

present in MIMO systems, antenna arrays, instead of multiple antenna elements, are of interest and so, the concept of antenna array signal processing is introduced.

In Chapter 2, the concept of the array manifold vector is introduced. The array manifold vector enables the parametric modelling of the channel parameters. Based on the array manifold vector, space-time channel models are then introduced. The various space-time channel models are defined based on the structure of the input and output signals. The system architecture of the asynchronous array-based MC-CDMA MIMO systems that are considered in this thesis are then presented. In the receivers, a bank of TDLs is employed so as to equip the receiver with space-time processing capabilities. The system models presented herein provides a framework for the detailed investigations in the subsequent chapters.

The problem of space-time reception and channel estimation for a single user cyclic prefix-free MC-CDMA arrayed MIMO system is then investigated in Chapter 3. A subspace-based receiver is developed based on the proposed system model. The proposed receiver seeks to null the ISI effects in the received signal and it requires the same degree of information as the RAKE receiver. The performance of the proposed receiver is compared with the RAKE and MMSE and it is demonstrated that the proposed receiver has a similar performance as the MMSE receiver and is superior to the RAKE receiver. In addition, the performance of the cyclic prefix-free MC-CDMA arrayed MIMO system is compared with an existing MC-CDMA MIMO system (which is non-array based and makes use of cyclic prefixes) and the proposed cyclic prefix-free MC-CDMA arrayed MIMO system is shown to have a superior performance than the reference system. This is despite the presence of ISI in the proposed system whereas as the reference system is not susceptible to such effects. Thus, system overheads are reduced without a loss in performance.

In order to obtain the estimates of the channel parameters, the channel estimation algorithm is developed based on the concept of the MC-STAR manifold vector. For a specific transmitter antenna whose space-time channel parameters are to be estimated, a delay-dependent preprocessor matrix is defined for each delay and the preprocessor matrix that is obtained is then used to transform the MC-STAR manifold vectors with the same delay for the specific transmitter antenna. The preprocessor is designed to null the MC-STAR manifold vectors which do not have the same delay for the transmitter antenna under consideration and these MC-STAR manifold vectors are removed from the signal subspace. At the same time, the MC-STAR manifold vectors arising from the co-channel interference (CCI), resulting from the remaining transmitter antenna elements, as well as the ISI are transformed by the preprocessor matrix. However, the transformation of the desired MC-STAR manifold vectors is different from the transformation of the CCI and ISI MC-STAR manifold vectors. The proposed algorithm thus exploits this property and makes use of a MUSIC-type cost function to obtain the estimates of the space-time channel parameters for the desired transmitter antenna. The proposed algorithm is demonstrated to be capable of identifying greater number of coherent multipaths than the number of antenna elements.

In Chapter 4, the point-source channel model for a cyclic prefix-free MC-CDMA arrayed MIMO system is extended to a diffused-source channel model which considers the effects of localized scattering. As a result, the point-source MC-STAR manifold vector is extended to the diffused MC-STAR manifold vector. The diffused MC-STAR vector manifold is obtained from a first order Taylor series approximation of the array manifold vector about the nominal DOA. The diffused MC-STAR manifold vector that is developed enables the spatially-diffused multipath channel characteristics to be completely captured in the received signal model. A subspace-based channel estimation algorithm, which also makes use of

the preprocessor matrix developed in Chapter 3, is then developed to obtain the estimates of the DOA, TOA and spatial spread. The proposed algorithm is not limited by the number of antennas and it allows the estimates of the parameters of more multipaths than the number of antenna elements to be obtained. Moreover, it is shown that the proposed algorithm can has better performance than an algorithm which ignores the effects of local scattering in the channel.

A decorrelating receiver is then developed for the diffused channel and it is shown to have much better performance than a receiver which ignores the scattering present in the channel. The superior performance of the proposed receiver is also validated for the uplink of a multi-user cyclic prefix-free MC-CDMA arrayed MIMO system in which near-far effects are present.

The problem of joint transmitter-receiver beamforming in the downlink of a multi-user cyclic prefix-free MC-CDMA arrayed MIMO system is next investigated in Chapter 5. The work is based on the assumption that channel information is available at the base station and the optimization objective is to minimize the overall MSE of the system subject to a transmit power constraint. In addition, the link between the transmit power of each transmitter antenna and the PAPR (and consequently, the probability of clipping of the transmitted signal) is used to reduce the PAPR and hence, the probability of clipping, of the transmitted signals from each transmit antenna. The optimization problem is solved with the Lagrange Multiplier method and an iterative algorithm for the determination of the transmit and receive beamforming matrices is introduced. This is a result of the inter-dependence between the transmit and receive beamforming matrices. It is shown that the iterative process brings about a much performance than if the transmit and receive beamforming matrices were not optimized. In addition, it is demonstrated that the probability of clipping of the transmitted signal is indeed reduced by the per-antenna transmit power constraint.

6.2 List Of Contributions

The list below is a summary of the main results and contributions presented in this thesis:

- Development of a new space-time signal processing framework based on the STAR manifold vector concept, which is a function of various parameters such as DOA, DOD, array geometry and carrier frequency, for asynchronous cyclic prefix-free MC-CDMA arrayed MIMO systems that operate over frequency selective channels. The MC-STAR manifold matrix, thus developed, encompasses the spatial and temporal information of the space-time channel for the cyclic prefix-free MC-CDMA arrayed MIMO system. Due to the cyclic prefix-free nature of the considered system, the ISI effects are present in the received signal and so, such effects are modeled in the received signal so that the ISI effects can be removed by the receiver for an improved performance.
- A single-user space-time receiver is developed based on the model of the cyclic prefix-free MC-CDMA arrayed MIMO system that has been developed. The proposed subspace-based receiver aims to cancel the intersymbol interference in a MIMO system by projecting the ISI effects onto the complementary subspace and so, nulling the ISI contributions in the received signal. The proposed receiver is shown to have comparable performance as that of a MMSE receiver. In addition, the proposed cyclic prefix-free MC-CDMA arrayed MIMO system has been shown to have superior performance compared to a conventional MC-CDMA MIMO system, which employs cyclic prefixes and ignores the antenna array geometry.
- A blind channel estimation algorithm is developed which provides the spa-

tial and temporal parameters for each path in a multipath frequency-selective fading channel for the transmit antenna under consideration. The transformation is specified by the PN-sequence of the desired antenna. The signal processing is carried out in a transform domain which isolates and transforms the MC-STAR manifold vectors with a specified delay. The proposed algorithm is capable of identifying a greater number of multipaths, including co-directional paths, than the number of antenna elements. However, in the event of co-delay paths, spatial smoothing is required to resolve the coherence between the co-delay paths. Due to the assumption of code diversity, the channel estimation algorithm can be used for multi-user systems as well.

- Investigation of the effects of local scattering in a cyclic prefix-free MC-CDMA arrayed MIMO system. Localized scattering is an important channel impairment in practical systems which cannot be ignored. In order to investigate the effect of a diffused channel for a cyclic prefix-free MC-CDMA arrayed MIMO system, the ‘spatially-diffused’ STAR manifold vector, which is approximated as a function of the array manifold vector and its first derivative, is introduced to represent the spatial diffusion of the channel.
- The channel estimation algorithm that is previously proposed for a point-source cyclic prefix-free MC-CDMA arrayed MIMO system is adapted for the spatially-diffused cyclic prefix-free MC-CDMA arrayed MIMO system. The algorithm jointly estimates the spread, DOA and TOA of the multipaths in a spatially diffused channel.

- A linear space-time decorrelating receiver is developed based on the estimates of the channel parameters of the diffused channel and it is shown to have a better performance than a receiver which ignores the spatial diffusion of the channel. The effectiveness of the receiver is also verified for the uplink of a multi-user cyclic prefix-free MC-CDMA arrayed MIMO system which suffers from near-far effects.
- The joint optimization of the transmit and receive beamforming weights for the downlink of a multi-user cyclic prefix-free MC-CDMA arrayed MIMO system is investigated for the purpose of improving system performance. A Lagrange multiplier method is used to obtain a solution for the optimization problem and subsequently, an iterative algorithm is developed.
- A *subcarrier-cooperative* approach is taken in the modelling of the transmit beamforming weights. This is in contrast to the commonly used subcarrier-noncooperative approach in which the subcarrier weights are obtained separately.
- The relationship between the transmit power of an antenna, the probability of clipping and the peak-to-average power ratio (PAPR) is used to lower the probability of clipping of the transmitted multi-carrier signals from each transmit antenna. As such, per-antenna power constraint is imposed for the joint optimization problem.

6.3 Suggestions for Future Work

Time-Varying Channels

The array-based MIMO systems considered in this thesis are developed based on the assumption that the terminals are stationary. As a result, Doppler effects, and hence time-varying channels, which arise from the mobility of the terminals have been ignored. In future 4G systems, the mobility of the users is an important factor in determining the achievable data rates. A system architecture which takes the Doppler effect into consideration will enable the development of a receiver that can cope with the requirements of a time-varying channel.

Spatio-temporally Diffused Channel

The investigation of a temporally-diffused channel and subsequently, spatio-temporally diffused channels for MC-CDMA MIMO systems. In such channels, the multipaths appear diffused in both space and time at the receiver and are a more accurate reflection of the practical situation. The consideration of the spatial and temporal diffusion of the multipaths will lead to a more robust receiver which is capable of estimating the channel parameters associated with such channels.

Array Uncertainties

- Carrier Frequency Offset - The presence of carrier frequency offset in the system is another important phenomenon that can be investigated. In the presence of a carrier frequency offset, the received signal is inaccurately downconverted and so, the residual frequency component will translate into a time-varying received signal. Thus, accurate estimates of the carrier frequency offset will be required so that downconversion of the received signal can be performed accurately.

- **Calibration Uncertainties** - The proposed arrayed MIMO systems have assumed that the antenna arrays have been properly calibrated. However, array uncertainties, due to deviation of the true array characteristics or parameters from their nominal values, will introduce errors into the system. Such errors, if ignored, would degrade the performance of the system either slowly or abruptly. Thus, the design of robust receivers which are capable of overcoming such errors in the calibration are an attractive option.

Extension of MC-CDMA to OFCDM

The MC-CDMA modulation that has been considered in this thesis restricts the number of subcarriers to be equal to the length of the spreading sequence. Thus, the extension of the MC-CDMA MIMO systems to Orthogonal Frequency and Code Division Multiplexing (OFCDM) MIMO systems will result in more flexible system due to the high scalability of OFCDM systems in possible transmission rates. Thus, OFCDM systems can support low to high information transmission rates by employing code, time and frequency division multiplexing. However, the extension of the MC-CDMA system to an OFCDM system implies that the MC-STAR manifold vector that has been introduced in this thesis will have to be modified to be applicable for an OFCDM system. In addition, two-dimensional spreading has been proposed for OFCDM systems [93] and so, the investigation of the integration of antenna arrays in such OFCDM systems is an area that can be looked at.

Near-field Communications

Plane wave propagation, due to far-field communications, has been assumed for the array systems considered thus far. The consideration of near-field communications, which is characterized by spherical wave propagation, will introduce additional diversity to the system. This implies that the interference cancellation properties of the receiver can be improved by leveraging on the additional information available. However, this means that the channel estimation problem is made more complex due to the additional parameters that have to be estimated. Hence, a simplification of the channel estimation problem for such systems is necessary.

Combination of Various Channel Effects

The vector channel estimation and reception algorithms presented in this thesis are designed for separate channel effects. However, a combination of these separate scenarios have yet to be investigated. For instance, the Doppler effect on diffused sources, or the polarization effect on diffused sources, or the Doppler effect on spherical wave propagation, are some of the potential areas of research. Thus, a channel estimator can be devised to be integrated at the front-end of the receivers to cope with a multitude of such channel effects.

However, in these cases, it is expected that the cost function for each of its corresponding channel estimation algorithms will consequently become a multi-dimensional problem. Hence, further simplification to reduce the dimension of the search is required. This could be achieved by some novel means of merging either the polarization / diffused / Doppler / spherical wave propagation parameters together in the eigenvector component of the cost function. A unified architectural framework that is capable of considering such composite channels using a generalized cost function would also pose some of the challenging research

problems.

Multi-Rate Systems

In 4G systems, different users are expected to have different data rate requirements due to the diverse multimedia applications available such as high speed video transmissions and low rate voice communications. As such, the resultant 4G systems would have to be designed to cope with the different rates required for the wide spectrum of services available. The design of the receiver for such a multi-rate system gives rise to several challenges. For instance, it is difficult to exploit the interference structure for interference cancellation in such multi-rate systems. Hence, receivers that are suitable for such multi-rate systems are of interest. Moreover, complications arise in the channel estimation process with such multi-rate systems. Thus, the design of blind channel estimation algorithms may not be feasible and semi-blind channel estimation algorithms may have to be used instead.

Joint Optimization with QoS Requirements

In addition to multi-rate systems, it is also plausible that the quality-of-service (QoS) required by each user in a 4G system is different. Thus, the joint optimization of the transmit and beamforming weights, which in Chapter 5 focused on the minimization of the overall mean-squared-error (MSE) of the system, would have to focus on satisfying the different QoS requirements of each user. In multi-carrier modulation systems, the PAPR is also an important parameter in practical systems. As such, the investigation of the joint optimization with different QoS requirements, while ensuring that the PAPR is kept within an acceptable range, is a possible research problem. Moreover, when coupled with a multi-rate system as envisioned for 4G systems, a challenging problem arises with a multitude of

issues that have to be addressed.

Appendix A

Effect of Multipath Channels on MC Systems

One view of the inverse discrete Fourier transform (IDFT) is that the IDFT performs a correlation of its input signal with each of the sinusoidal basis functions. Hence, at a certain frequency where the input signal has some energy, the correlation of the input signal and the basis sinusoid of the corresponding frequency will produce a peak. Thus, the IDFT can be used in the transmitter employing MC technique to map an input signal onto a set of orthogonal subcarriers, i.e. the orthogonal basis functions of the IDFT. Conversely, the discrete Fourier transform (DFT) is employed again at the MC receiver to process the received subcarriers. The signals from the subcarriers are then combined to form an estimate of the source signal from the transmitter. The orthogonal nature of the subcarriers is thus exploited in MC schemes with powerful results. As the basis functions of the DFT are uncorrelated, the correlation performed in the DFT for a specific subcarrier only produces energy for that corresponding subcarrier while the energy from other subcarriers does not appear in the output of the correlation process. This separation of signal energy is the reason that the spectrums of the subcarriers spectrums in MC schemes can overlap without causing interference.

A discussion of the effect of multipath channels on MC systems and a subsequent discussion of the sources of non-ideal effects in such systems now follows.

A major problem in most wireless systems is the presence of a multipath channel which is due to the transmitted signal being reflected off of several objects. As a result, multiple delayed versions of the transmitted signal arrive at the receiver. The multiple versions of the signal result in a distortion of the received signal. This results in two problems for an MC system. The first problem is ISI. This problem occurs when the received MC signal is distorted by the previously transmitted MC symbol. The effect is similar to the intersymbol interference that occurs in a SCM system. However, in SCM systems, the interference is typically attributed to several other symbols instead of just the previous symbol as the symbol period in SCM systems is typically much shorter than the time span of the channel while the typical MC symbol period is much longer than the time span of the channel. The second problem is unique to MC systems and is known as intrasymbol interference, which is the result of interference amongst a given MC symbol's own subcarriers.

A.1 Intersymbol Interference

In a MC system, by choosing an appropriate value for the number of subcarriers, the length of the resultant MC symbol becomes longer than the time span of the channel. Thus, because of this configuration, the effect of ISI is the distortion of the first few samples, corresponding to the time span of the channel, of the received MC symbol. As only the first few samples of the symbol are distorted, a guard interval can be employed to remove the effect of the ISI present. For instance, the guard interval (GI) can be a section of all zero samples transmitted in front of each MC symbol. Since the GI does not contain any useful information,

the GI can be discarded at the receiver. Thus, a proper choice of the GI length such that it is longer than the time span of the channel ensured that the MC symbol itself will not be distorted. As a result, by discarding the guard interval, the effects of ISI are removed as well.

A.2 Intrasympol Interference (Intercarrier Interference)

The guard interval is not used in practical systems because it does not prevent an MC symbol from interfering with itself. This type of interference is called intrasympol interference, also known as intercarrier interference (ICI). The solution to the problem of ICI involves the discrete time property of signals. In continuous time, a convolution in time is equivalent to a multiplication in the frequency-domain. However, this property is true in discrete time only if the signals are of infinite length or if at least one of the signals is periodic over the range of the convolution. As it is impractical to have a MC symbol of infinite length, the alternative option is to make the MC symbol appear periodic. This periodic form is achieved by replacing the guard interval with something known as a cyclic prefix (CP) which is a replica of the last few samples of the MC symbol. As in the case of the GI, the length of the CP must be longer than the time span of the channel. Since it contains redundant information, the CP is discarded at the receiver and so, the effects of ISI are removed. Because of the way in which the CP was formed, the cyclically-extended MC symbol now appears periodic when convolved with the channel. Thus, the effect of the channel becomes multiplicative.

In a digital communications system, the symbols that arrive at the receiver have been convolved with the time-domain channel impulse response. In order

to undo the convolutional effects of the channel, another convolution must be performed at the receiver using a time-domain filter known as an equalizer. The equalizer processes symbols in order to adapt its response in an attempt to remove the effects of the channel and the length of the equalizer needs to be on the order of the time span of the channel. Such an equalizer can be expensive to implement in hardware and often requires a large number of symbols in order to adapt its response for a good performance.

In MC systems, the time-domain signal is still convolved with the channel response. However, the data will ultimately be transformed back into the frequency-domain by the DFT in the receiver. Because of the periodic nature of the cyclically-extended OFDM symbol, this time-domain convolution results in the multiplication of the spectrum of the MC signal with the frequency response of the channel. The result is that the symbol at each subcarrier will be multiplied by a complex number equal to the channel's frequency response at that subcarrier's frequency. Thus, each received subcarrier experiences a complex gain due to the channel. A frequency-domain equalizer is thus employed to undo these effects. The frequency-domain equalizer consists of a single complex multiplication for each subcarrier and is much simpler than a time-domain equalizer.

A.3 Non-ideal effects in MC systems

A brief discussion of the causes of non-ideal effects in an MC system is now presented and a detailed examination of the issues listed here can be obtained from [94, 95]. Due to the relation between multi-carrier modulation and Fourier transforms, the result of these effects are explained using Fourier transform theory.

A.3.1 Carrier Frequency and Phase Offsets

In practical systems, the local oscillator (LO) frequency at the receiver is typically different from the LO frequency at the transmitter. This results in a carrier frequency offset. A carrier tracking loop can be used to adjust the receiver's LO frequency in order to match the transmitter's LO frequency as closely as possible so that the carrier frequency offset (CFO) can be reduced as much as possible. An intuitive understanding of the effect of a CFO can be explained by Fourier Transform theory. The CFO can be expressed mathematically by multiplying the received time-domain signal by a complex exponential whose frequency is equal to the amount of the carrier frequency offset. Thus, the CFO is equivalent to a frequency shift in the received signal spectrum. This results in a loss of orthogonality of the subcarriers as the subcarriers are no longer orthogonal and the bins of the IDFT will no longer line up with the peaks of the received signal's pulses. The result is a distortion called inter-bin interference. Inter-bin interference occurs when energy from one bin spills over into adjacent bins and this energy distorts the affected subcarriers.

This effect is referred to as DFT leakage in Fourier Transform theory. This offset results in a loss of orthogonality, and the zero-crossings of the non-zero subcarrier's spectrum no longer line up with the FFT bins. In a practical system where almost all of the subcarriers would be actively used for transmitting data, a given subcarrier would experience inter-bin interference due to energy from all of the other active subcarriers in the system. Based on the central limit theorem, which states that the sum of a large number of random processes will result in a signal that has a Gaussian distribution, the inter-bin interference can be viewed as additive Gaussian noise which lowers the effective SNR of the system. The effect of a carrier frequency offset can be corrected by multiplying the signal by a correction factor which would be a sinusoid with a frequency that is ideally equal

to the amount of the carrier frequency offset. Various carrier tracking algorithms exist that can adaptively determine the frequency that will correct for the offset.

Besides a LO frequency offset, it is also possible to have an LO phase offset, separate from an LO frequency offset. The two offsets can occur together or either one can be present by itself. As the name implies, an LO phase offset occurs when there is a difference between the phase of the LO output and the phase of the received signal. This effect can be represented mathematically by multiplying the time-domain signal by a complex exponential with a constant phase. This results in a constant phase rotation for all of the subcarriers in the frequency domain. Thus, the constellation points for each subcarrier experience the same degree of rotation. This can be corrected by the frequency-domain equalizer if the phase rotation is small as long as the rotation does not lead to a rotation of the constellation points beyond the symbol decision regions. In the case of large phase rotations, a carrier tracking loop can be used to correct such effects.

A.3.2 FFT Window Location Offset

Another non-ideal effect that can occur in practical MC systems is an FFT window location offset. In practical MC systems, a correlation operation is often performed with a known preamble sequence located at the beginning of the transmission. This correlation operation enables the synchronization of the receiver with the received MC symbol boundaries. However, the synchronization process is usually not perfect and so, inaccuracies in the synchronization process still remain, and such errors appear as an offset in the FFT window location. The result is that the samples sent to the FFT will not line up exactly with the corresponding MC symbol.

The implication of a large offset is that part of the samples will be from one MC symbol, while the rest of the samples will be from another MC symbol. This

results in a severe distortion of the received subcarrier's constellations. However, a robust synchronization algorithm can be used to prevent such large offsets and an FFT window location offset of just a few samples will occur. The presence of the cyclic prefix in MC systems enables a small offset to be present without taking samples from more than one MC symbol. However, even an offset of just one sample will create some degree of distortion. From Fourier transform theory, the offset can be viewed as a shift in time. Thus, when the FFT window location offset does not go beyond a MC symbol boundary, this shift in time is equivalent to a linearly increasing phase rotation in the frequency-domain constellations. Thus, constellations on subcarriers corresponding to low frequencies will be rotated slightly, whereas constellations on higher-frequency subcarriers will experience a larger rotation. The correction of FFT window location offsets is often carried out by performing a time-domain correlation with a known training sequence embedded in the transmitted signal and the location of the peak of the correlation allows the receiver to synchronize itself with the incoming signal.

A.3.3 Phase noise

During the frequency conversion stage, noise can also be added to the signal as the local oscillator used in the converter will inherently have some phase noise which is due to an uncertainty in the actual frequency or phase of the signal. This phase noise will be transferred to the desired signal. Phase noise is usually shaped and is primarily concentrated near the carrier frequency of the signal. A MC signal set contains multiple subcarriers, each of which is a smaller percentage of the total frequency bandwidth than in a single carrier system. Hence, phase noise occupies a smaller percentage of the bandwidth in a SCM system. Thus, phase noise causes a larger degradation in the performance of a MC system more than in a single carrier system. Phase noise effects in a MC system can be separated into

two categories: phase noise maintained within one subcarrier spacing and phase noise that extends across subcarrier spacings. Phase noise within one subcarrier spacing essentially has a similar but scaled effect as for the single carrier system. On the other hand, phase noise that extends across subcarrier spacings results in demodulation errors. Pilot subcarriers are often used to overcome handle phase noise in MC systems and these pilot subcarriers are generated by the IFFT and can be used to provide a stable phase reference for the receiver circuitry. However, the use of pilot subcarriers lowers the achievable data rate of the system because these subcarriers are no longer available to transmit data.

A.3.4 Sampling frequency offset

A sampling frequency offset occurs when the analog-to-digital converter output is sampled either too fast or too slow. When sampling is performed at a frequency higher than the Nyquist frequency, a contracted spectrum results. Conversely, sampling at a frequency lower than the Nyquist frequency results in an expanded spectrum which can cause an aliasing of the spectrum to occur. The presence of sampling frequency offset results in inter-bin interference since the expansion or contraction of the spectrum prevents the received subcarriers from lining up with the FFT bin locations. A sampling frequency offset can be corrected by generating an error term that is used to drive a sampling rate converter.

A.3.5 Non-Linear Circuits in the Transmitter and Receiver

All transmitters and receivers in communications systems contain devices such as amplifiers which have non-linear transfer functions. These non-linearities create an additional performance limitation. The receiver performance is typically

limited by distortion generated in the input amplifier or mixer in the presence of strong undesired signals. On the other hand, the performance of the transmitter is limited by power amplifier linearity. A MC signal is made up of multiple simultaneous signals that, when combined together, have a higher peak signal level. Thus, for a given average power, MC signals result in an increase in the peak-to-average power ratio (PAPR) of the signal. In multi-carrier systems, the PAPR value is often expressed in terms of statistics because of the low probability that all subcarriers will simultaneously reach peak amplitude, although the simultaneous peak amplitude value is large. These higher peak amplitude levels will create more severe distortion than a single carrier system even if the average power levels are the same. The higher distortion increases the SNR needed to maintain adequate performance. Linearity requirements in both the receiver and transmitter must be adjusted or “backed off” to account for this increase in PAPR value. The PAPR value, and also the amount of linearity compensation, depends on a number of parameters such as the number of subcarriers and the level of SNR that must be maintained. Peak suppression techniques that have been proposed include coding, phase rotation and clipping.

Appendix B

Mathematical Derivations

B.1 Derivation of Transmit Beamforming Matrix

In the optimization problem stated in Equation 5.21, the constraint involves limiting the transmit power. Thus, assuming that the receive decoding matrix $\mathbb{W}_i, \forall i$ are fixed, and using Lagrange Multipliers, the optimization problem can be expressed as

$$\begin{aligned}
 J &= \min_{(\bar{\mathbb{W}}_1, \dots, \bar{\mathbb{W}}_M)} \sum_{m=1}^M \text{tr}(\mathbf{E}_m) + \sum_{p=1}^{\bar{N}} \mu_p \left(\sum_{m=1}^M \text{tr} \left(\bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \left(\mathbb{I}_{N_{sc}} \otimes \underline{L}_{\bar{N},p} \underline{L}_{\bar{N},p}^T \right) - \bar{P}_p \right) \right) \\
 &= \min_{(\bar{\mathbb{W}}_1, \dots, \bar{\mathbb{W}}_M)} \sum_{m=1}^M \text{tr} \left(\mathbb{W}_m^H \mathbb{R}_{xx,m} \mathbb{W}_m - \mathbb{W}_m^H \mathbb{H}_{m,m} \bar{\mathbb{W}}_m - \bar{\mathbb{W}}_m^H \mathbb{H}_{m,m}^H \mathbb{W}_m + \mathbb{I}_{N_m^{sub}} \right) \\
 &\quad + \sum_{p=1}^{\bar{N}} \mu_p \left(\sum_{m=1}^M \text{tr} \left(\bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \left(\mathbb{I}_{N_{sc}} \otimes \underline{L}_{\bar{N},p} \underline{L}_{\bar{N},p}^T \right) \right) - \bar{P}_p \right) \tag{B.1}
 \end{aligned}$$

Substituting $\mathbb{R}_{xx,m}$, as given by

$$\mathbb{R}_{xx,m} = \sum_{i=1}^M \mathbb{V}_{m,i} \left(\mathbb{I}_3 \otimes \bar{\mathbb{W}}_i \bar{\mathbb{W}}_i^H \right) \mathbb{V}_{m,i}^H + \sigma_m^2 \mathbb{I}_{2NN_{sc}}$$

into Equation B.1 yields

$$\begin{aligned}
J = & \min_{(\bar{\mathbb{W}}_1, \dots, \bar{\mathbb{W}}_M)} \sum_{m=1}^M \text{tr} \left(\mathbb{W}_m^H \left(\sum_{i=1}^M \mathbb{V}_{m,i} \left(\mathbb{I}_3 \otimes \bar{\mathbb{W}}_i \bar{\mathbb{W}}_i^H \right) \mathbb{V}_{m,i}^H + \sigma_m^2 \mathbb{I}_{2NN_{sc}} \right) \mathbb{W}_m \right. \\
& \left. - \mathbb{W}_m^H \mathbb{H}_{m,m} \bar{\mathbb{W}}_m - \bar{\mathbb{W}}_m^H \mathbb{H}_{m,m}^H \mathbb{W}_m + \mathbb{I}_{N_m^{sub}} \right) \\
& + \sum_{p=1}^{\bar{N}} \mu_p \left(\sum_{m=1}^M \text{tr} \left(\bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \right) - \bar{P}_p \right) \quad (\text{B.2})
\end{aligned}$$

$$\begin{aligned}
J = & \min_{(\bar{\mathbb{W}}_1, \dots, \bar{\mathbb{W}}_M)} \sum_{m=1}^M \text{tr} \left(\mathbb{W}_m^H \left(\sum_{i=1}^M \mathbb{V}_{m,i} \left(\mathbb{I}_3 \otimes \bar{\mathbb{W}}_i \bar{\mathbb{W}}_i^H \right) \mathbb{V}_{m,i}^H \right) \mathbb{W}_m + \sigma_m^2 \mathbb{W}_m^H \mathbb{W}_m \right. \\
& \left. - \mathbb{W}_m^H \mathbb{H}_{m,m} \bar{\mathbb{W}}_m - \bar{\mathbb{W}}_m^H \mathbb{H}_{m,m}^H \mathbb{W}_m + \mathbb{I}_{N_m^{sub}} \right) \\
& + \sum_{p=1}^{\bar{N}} \mu_p \left(\sum_{m=1}^M \text{tr} \left(\bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \right) - \bar{P}_p \right) \quad (\text{B.3})
\end{aligned}$$

In order to obtain the optimal solution $\bar{\mathbb{W}}_m, \forall m$, it is necessary to derive the first derivative of J with respect to $\bar{\mathbb{W}}_m, \forall m$, i.e. $\frac{\partial J}{\partial \bar{\mathbb{W}}_m}$. Based on Equation B.3, the $\frac{\partial J}{\partial \bar{\mathbb{W}}_m}$ thus be obtained as:

$$\begin{aligned}
\frac{\partial J}{\partial \bar{\mathbb{W}}_m} = & 2 \left(\sum_{i=1}^M \mathbb{L}^H \left(\mathbb{V}_{i,m}^H \left(\mathbb{W}_i \mathbb{W}_i^H \right) \mathbb{V}_{i,m} \right) \mathbb{L} \right) \bar{\mathbb{W}}_m \\
& - 2 \mathbb{H}_{m,m}^H \mathbb{W}_m + 2 \sum_{p=1}^{\bar{N}} \mu_p \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \bar{\mathbb{W}}_m \quad (\text{B.4})
\end{aligned}$$

where $\mathbb{L} = \mathbb{1}_3 \otimes \mathbb{I}_{N_{sc}}$. The optimal local solution $\bar{\mathbb{W}}_m^*$ can be obtained by evaluating $\left(\frac{\partial J}{\partial \bar{\mathbb{W}}_m} \right)_{\bar{\mathbb{W}}_m = \bar{\mathbb{W}}_m^*} = 0$ and so,

$$\begin{aligned}
& 2 \left(\sum_{i=1}^M \mathbb{L}^H \left(\mathbb{V}_{i,m}^H \left(\mathbb{W}_i \mathbb{W}_i^H \right) \mathbb{V}_{i,m} \right) \mathbb{L} \right) \bar{\mathbb{W}}_m^* \\
& - 2 \mathbb{H}_{m,m}^H \mathbb{W}_m + 2 \sum_{p=1}^{\bar{N}} \mu_p \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \bar{\mathbb{W}}_m^* = 0 \quad (\text{B.5})
\end{aligned}$$

Hence,

$$\left(\sum_{i=1}^M \mathbb{L}^H \left(\mathbb{V}_{i,m}^H \left(\mathbb{W}_i \mathbb{W}_i^H \right) \mathbb{V}_{i,m} \right) \mathbb{L} + \sum_{p=1}^{\bar{N}} \mu_p \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \right) \bar{\mathbb{W}}_m^* = \mathbb{H}_{m,m}^H \mathbb{W}_m \quad (\text{B.6})$$

and finally,

$$\overline{\mathbb{W}}_{\mathbf{m}}^{\star} = \left(\sum_{i=1}^M \mathbb{L}^H \left(\mathbb{V}_{i,\mathbf{m}}^H \left(\mathbb{W}_i \mathbb{W}_i^H \right) \mathbb{V}_{i,\mathbf{m}} \right) \mathbb{L} \right. \quad (\text{B.7})$$

$$\left. + \sum_{p=1}^{\overline{N}} \mu_p \left(\mathbb{I}_{N_{sc}} \otimes \underline{\mathbb{L}}_{\overline{N},p} \underline{\mathbb{L}}_{\overline{N},p}^T \right) \right)^{-1} \mathbb{H}_{\mathbf{m},\mathbf{m}}^H \mathbb{W}_{\mathbf{m}} \quad \forall \mathbf{m} \quad (\text{B.8})$$

B.2 Derivation of Lagrange Multiplier

The derivative of J with respect to μ_p can be expressed as

$$\frac{\partial J}{\partial \mu_p} = \sum_{m=1}^M \text{tr} \left(\bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \right) - \bar{P}_p$$

Letting

$$\frac{\partial J}{\partial \mu_p} = 0$$

results in

$$\begin{aligned} \sum_{m=1}^M \text{tr} \left(\bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \right) - \bar{P}_p &= 0 \\ \sum_{m=1}^M \text{tr} \left(\bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \right) &= \bar{P}_p \end{aligned}$$

which corresponds to the power constraint requirement. Thus, Equation B.5 can be modified to become

$$\begin{aligned} &\left(\sum_{i=1}^M \mathbb{L}^H \left(\mathbb{V}_{i,m}^H \left(\mathbb{W}_i \mathbb{W}_i^H \right) \mathbb{V}_{i,m} \right) \mathbb{L} \right) \bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \\ &- \mathbb{H}_{m,m}^H \mathbb{W}_m \bar{\mathbb{W}}_m^H + \sum_{p=1}^{\bar{N}} \mu_p \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H = 0 \end{aligned} \quad (\text{B.9})$$

To obtain the expression for μ_p , the factor $\left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right)$ multiplied to the above expression to yield

$$\begin{aligned} &\left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \left(\sum_{i=1}^M \mathbb{L}^H \left(\mathbb{V}_{i,m}^H \left(\mathbb{W}_i \mathbb{W}_i^H \right) \mathbb{V}_{i,m} \right) \mathbb{L} \right) \bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \\ &- \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \mathbb{H}_{m,m}^H \mathbb{W}_m \bar{\mathbb{W}}_m^H + \mu_p \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H = 0 \end{aligned} \quad (\text{B.10})$$

as

$$\left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) = \begin{cases} \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right), & \text{if } p = p \\ \mathbf{0}_{\bar{N}N_{sc} \times \bar{N}N_{sc}}, & \text{if } p \neq p, \forall p \end{cases}$$

Taking the trace and summation over M users results in

$$\begin{aligned}
& \sum_{m=1}^M \text{tr} \left(\left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \left(\sum_{i=1}^M \mathbb{L}^H \left(\mathbb{V}_{i,m}^H \left(\mathbb{W}_i \mathbb{W}_i^H \right) \mathbb{V}_{i,m} \right) \mathbb{L} \right) \bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \right) \\
& - \sum_{m=1}^M \text{tr} \left(\left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \mathbb{H}_{m,m}^H \mathbb{W}_m \bar{\mathbb{W}}_m^H \right) \\
& + \sum_{m=1}^M \text{tr} \left(\mu_p \left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \right) = 0 \tag{B.11}
\end{aligned}$$

$$\begin{aligned}
& \sum_{m=1}^M \text{tr} \left(\left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \left(\sum_{i=1}^M \mathbb{L}^H \left(\mathbb{V}_{i,m}^H \left(\mathbb{W}_i \mathbb{W}_i^H \right) \mathbb{V}_{i,m} \right) \mathbb{L} \right) \bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \right) \\
& - \sum_{m=1}^M \text{tr} \left(\left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \mathbb{H}_{m,m}^H \mathbb{W}_m \bar{\mathbb{W}}_m^H \right) + \mu_p \bar{P}_p = 0 \tag{B.12}
\end{aligned}$$

Thus, replacing p with p , the solution for the Lagrange multiplier μ_p^* , $\forall p$, can be expressed as

$$\begin{aligned}
\mu_p^* &= \frac{1}{\bar{P}_p} \sum_{m=1}^M \text{tr} \left(\left(\mathbb{I}_{N_{sc}} \otimes \underline{\ell}_{\bar{N},p} \underline{\ell}_{\bar{N},p}^T \right) \left(\mathbb{H}_{m,m}^H \mathbb{W}_m \bar{\mathbb{W}}_m^H \right. \right. \\
& \quad \left. \left. - \mathbb{L}^H \left(\sum_{i=1}^M \mathbb{V}_{i,m}^H \left(\mathbb{W}_i \mathbb{W}_i^H \right) \mathbb{V}_{i,m} \right) \mathbb{L} \bar{\mathbb{W}}_m \bar{\mathbb{W}}_m^H \right) \right) \forall p \tag{B.13}
\end{aligned}$$

B.3 Derivation of Receive Beamforming Matrix

The derivation of the receive beamforming matrix can be obtained in a similar way as the transmit beamforming matrix. The transmit beamforming matrix is assumed to be fixed and so, the derivative of J , as shown in Equation B.3, with respect to $\mathbb{W}_m$ can be expressed as:

$$\frac{\partial J}{\partial \mathbb{W}_m} = 2 \left(\sum_{i=1}^M \mathbb{V}_{m,i} \left(\mathbb{I}_3 \otimes \overline{\mathbb{W}}_i \overline{\mathbb{W}}_i^H \right) \mathbb{V}_{m,i}^H \right) \mathbb{W}_m - 2\mathbb{H}_{m,m} \overline{\mathbb{W}}_m + 2\sigma_m^2 \mathbb{I}_{2NN_{sc}} \mathbb{W}_m \quad (\text{B.14})$$

The solution for the receive beamforming matrix, $\mathbb{W}_m^*$, $\forall m$, can thus be obtained by defining

$$\left(\frac{\partial J}{\partial \mathbb{W}_m} \right)_{\mathbb{W}_m = \mathbb{W}_m^*} = 0 \quad (\text{B.15})$$

which results in

$$2 \left(\sum_{i=1}^M \mathbb{V}_{m,i} \left(\mathbb{I}_3 \otimes \overline{\mathbb{W}}_i \overline{\mathbb{W}}_i^H \right) \mathbb{V}_{m,i}^H \right) \mathbb{W}_m^* - 2\mathbb{H}_{m,m} \overline{\mathbb{W}}_m + 2\sigma_m^2 \mathbb{I}_{2NN_{sc}} \mathbb{W}_m^* = 0 \quad (\text{B.16})$$

Thus,

$$\begin{aligned} \mathbb{W}_m &= \left(\sum_{i=1}^M \mathbb{V}_{m,i} \left(\mathbb{I}_3 \otimes \overline{\mathbb{W}}_i \overline{\mathbb{W}}_i^H \right) \mathbb{V}_{m,i}^H + \sigma_m^2 \mathbb{I}_{2NN_{sc}} \right)^{-1} \mathbb{H}_{m,m} \overline{\mathbb{W}}_m \\ &= \mathbb{R}_{xx,m}^{-1} \mathbb{H}_{m,m} \overline{\mathbb{W}}_m \quad \forall m \end{aligned} \quad (\text{B.17})$$

It can be seen from Equation B.17 that the solution for the receive beamforming matrix corresponds to a MMSE receiver.

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