A Low Complexity Multicarrier
Layered Space-Time Architecture for
Realizing High Data Rates

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by

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Abstract

The next generation wireless communication systems are expected to support a wide range of high-quality services which require high data rate transmission as well as high system capacities. One of the most recent and promising techniques for realizing extraordinary spectral efficiencies over wireless links is V-BLAST (Vertical-Bell Laboratories Layered Space-Time). This is a new system proposed by Bell Laboratories as an extremely efficient scheme for high capacity communications in wireless environments. The V-BLAST technique employs a multi-element antenna array technology at both the transmitter and receiver to increase system capacity, data rate as well as spectral efficiency.

In this thesis, we consider the performance of V-BLAST in a frequency selective fading channel. In particular, we investigate the effect of time delay spread on V-BLAST. In order to make V-BLAST more robust against the detrimental effects of frequency selective and time varying channels, we combine V-BLAST with Orthogonal Frequency Division Multiplexing (OFDM) transmission and Reed-Solomon coding. Owing to the required intensive computation involved, we propose a low complexity multicarrier V-BLAST based on the use of a sub-carrier grouping, an innovative sub-optimal decoding ordering procedure, and the Gram-Schmitt Orthogonalization (GSO) procedure instead of the pseudo-inverse operation used in V-BLAST. It is shown that the proposed system can greatly reduce the computational complexity required with a minimal penalty in performance compared with a standard multicarrier V-BLAST.
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Chapter 1

Introduction

During the past few years, there has been an explosion in wireless technology. This growth has opened a new dimension to future wireless systems whose ultimate goal is to provide universal personal and multimedia communications without regard to mobility or location[1, 2, 3]. To achieve such an objective, the next-generation personal communications networks (PCNs) will need to be able to support a wide range of services which will include high quality voice, data, facsimile, still pictures, and video. These future services that are likely to include applications which require high transmission rates of several Mbps. However, unlike wired channels, transmission over radio channels is extremely challenging.

The mechanisms behind electromagnetic wave propagation can generally be attributed to reflection, diffraction and scattering. Due to multiple interactions from various objects, the electromagnetic waves travel along different paths of varying lengths. The interaction between these waves causes multipath fading at a specific location, and the strengths of the waves decrease as the distance between the transmitter and receiver increases. These multiple propagation paths can in general have different time delays, attenuation, and phase shifts, which results in a spread in delay times imposing a limit on the maximum transmission rate. As a result, the channel induces multiple delayed version of the transmitted symbols which is called intersymbol interference (ISI) leading to the introduction of an irreducible error floor[4, 5]. Moreover, due to varying Doppler shifts on different multipath signals, spreading in the frequency domain
or random frequency modulation and time varying channels will also occur.

1.1 Motivation

Wireless communication systems have recently enjoyed great market success. So far, however, these developments have mostly been concentrated on voice and low data-rate paging applications. Third generation wireless access systems such as Wideband Code Division Multiple Access (WCDMA) and the evolution of second-generation systems such as TDMA IS-136+ and EDGE (Enhanced Data-rates for GSM Evolution) will provide bit rates of 50 to 384 kbps in macro-cellular systems [6]. In the near future, however, broadband data access services at peak rates beyond 2 Mbps will likely be needed to provide users with widespread support for universal personal communications coverage with few compromises when compared with fixed networks. Hence, one of the main objectives of the next generation wireless and mobile communications systems is to extend the services provided by the current second-generation systems with high-rate capabilities. Communicating at high transmission rates over the harsh and hostile wireless channels with a limited spectrum creates many difficult and challenging problems. This implies that countermeasures should be employed to mitigate the delay spread impairment and an increase in the capacity of the current wireless systems will need to be achieved [7, 8, 9]. To increase the system information capacity as well as the data rate, while using the spectral resource efficiently, provisions such as multi-element antenna array technology at both the transmitter and receiver have been proposed for the next generation wireless systems. This technique allow us to use of space diversity to compensate for channel impairment without increasing the transmitted power or bandwidth.

Recent information theory work has shown that the rich scattering wireless channel is capable of enormous theoretical capacity provided that the scattering is properly exploited through the use of an appropriate processing architecture [10, 11]. The Layered Space-Time architecture proposed by Foschini and Golden [10] is one such approach that utilizes a multi-element antenna array at both
the transmitter and receiver to provide high data rates. In an independent
Rayleigh scattering environment, this processing structure leads to theoretical
rates, which grow linearly with the number of antennas achieving very high
capacities [10, 12].

Frequency selective fading channels can severely degrade system performance
by causing intersymbol interference (ISI). This will then result in an irreducible
bit error rate (BER) which imposes an upper limit on the data rate [4, 5]. This
impairment is thus a crucial issue that becomes more important in the next gen-
eration wireless systems where broadband and high data rate communications
services are needed [9].

Another possible technique proposed for the next generation wireless sys-
tems, which has received a lot of attention recently, is multicarrier modulation
or Orthogonal Frequency Division Multiplexing (OFDM) in which the trans-
mitted data is divided into several interleaved bit streams. These streams are
then used to modulate several subcarriers. Such an approach has already been
proven to be one effective high speed transmission method for wireless personal
communications [13, 14, 15, 16].

The motivation for this thesis is to study the performance of Layered Space-
Time architecture in multipath fading channels. We aim to employ a delay
spread mitigation techniques such as OFDM with the Layered Space-Time ar-
chitecture in order to maintain system performance at acceptable levels in mul-
tipath fading channels. We also propose to investigate the complexity issue and
to seek for a low complexity algorithm.

1.2 Thesis Contributions

The next-generation of wireless personal communication systems is expected
to provide ubiquitous universal high speed multimedia transmission over the hos-
tile mobile environments without regard to mobility or location. To achieve this
objective various technical challenges must be overcome [6]. For example, the
deployment of broadband wireless access systems would require a transmission
technique which can mitigate the detrimental effects of frequency-selective fading [17]. In recent years, there has been much interest in applying multicarrier techniques in wireless systems because of its various advantages in mitigating the severe effects of frequency-selective fading [18, 13].

In the past decades, several antenna systems have previously been proposed, and demonstrated at the receivers of the wireless communication systems, and these have shown that significant increases in capacity and performance are possible [19, 20, 21, 22]. Further increases in capacity and performance may be possible by including multiple antennas at the transmitter in both the downlink and the uplink. Although only base station (BS) allows the use of smart antenna or adaptive array for current mobile communication systems because of the cost associated and the relatively small size of the mobile station (MS) (e.g. a handset) to incorporate more than one antenna in order to have multiple uncorrelated signals. However, recent developments in hardware miniaturization and advances in antenna design may make this feasible in the future [23]. In addition, the increasing use of notebook computers in various wireless computing applications also suggests the use of smart antennas at MS. Therefore, it is worthwhile to investigate the use of multiple antennas at the MS and/or BS and determine whether the increase of the system performance justifies the increase of the system complexity.

The BLAST (Bell Laboratories Layered Space-Time) architecture [16, 24], and more particularly Vertical BLAST or V-BLAST, is an approach that uses multi-element antenna arrays at both the transmitter and receiver in conjunction with an appropriate processing architecture. In a flat-fading scattering environment with independent channels, this processing structure leads to theoretical rates that grow approximately linearly with the number of antennas [10].

In previously published work, the performance of V-BLAST has been primarily investigated in flat-fading environments. However, frequency-selective fading channels can severely degrade system performance by causing intersymbol interference (ISI), and result in an irreducible bit error rate which imposes an upper limit on the data rate [4, 5]. This impairment is thus a crucial issue
that becomes more important in the next generation wireless systems where broadband and high data rate communications services are needed [7, 25, 8]. In this thesis, and in contrast to previous work, we present the performance of V-BLAST over frequency-selective fading channels for a variety of modulation schemes and different numbers of transmit and receive antennas with two types of delay spread distributions.

When the multipath is sufficiently rich, channel gains between any pairs of transmit and receive antennas are largely uncorrelated or even independent and identically distributed (i.i.d.). V-BLAST [16] is one technique that can exploit this potential and provide a significant increase in capacity and system performance. Space Division Multiplexing (SDM) is one technique that can exploit this potential and provide a significant increase in capacity and Bit Error Rate (BER) performance. Basically, different, parallel data streams are transmitted, simultaneously and on the same frequency, using an antenna array. When using multiple antennas at the receiver as well, these data streams, mixed-up by the wireless channel, can be recovered by SDM techniques such as V-BLAST [16]. In [25] and [26], a single-carrier modulation with equalization is adopted to overcome the delay spread limitations in V-BLAST. These equalizers require many taps and must be updated at the highest Doppler rate. In addition, the extensive period required to train the equalizer could be a major source for system inefficiency. An alternative approach, and the one taken here, is to use multicarrier transmission and in particular OFDM, to minimize the effects of the channel delay spread. Since no equalization is required, OFDM alleviates the need for a long training period. Currently, a lot of research is concentrated on applying transmitter or receiver diversity to multicarrier techniques. A number of transmitter and receiver diversity techniques for OFDM are proposed, respectively, in [27] and [28].

Closely related to the V-BLAST technique is Space-Time codes [29] which uses space (different antennas) and time dimensions to maximize the coding and diversity gains. Hence, one data-symbol is coded and all the antennas are used to transmit it. However, V-BLAST achieves higher bit rates by transmitting
simultaneously different data on the different transmit antennas.

In this thesis, we focus our attention on such a V-BLAST system in rich-scattering environments where it performs at its best. We propose to combine V-BLAST with coded OFDM to mitigate the effects of time delay spread and to make the system more robust against frequency selective fading. The proposed system is been investigated with different channel delay profiles and number of subcarriers in both stationary and time varying environments. and especially the proposed system delay spread tolerance.

Clearly, OFDM/V-BLAST system involves an intensive computation and hence it may be difficult to implement it for high data rate communications[30]. Thus, we propose a low complexity multicarrier V-BLAST based on subcarrier grouping, using sub-optimal ordering instead of the optimal one and applying GSO to the channel matrix $H$ to find the weight vectors.

1.3 Outline of the Thesis

The overall plan of the thesis is as follows. After some introductory remarks in this chapter, we begin in earnest in Chapter 2. Chapter 2 provides fundamentals about mobile communications. One major emphasis in the chapter in on the concepts of small-scale propagation effects such as fading, time delay spread and Doppler spread. In addition, we describe how to model the impact that signal bandwidth and motion have on the instantaneous received signal through the multipath channel. Chapter 2 also demonstrates how interference between the received signals affects the capacity of a mobile communication systems.

Next, Chapter 3 presents the concepts of multiple input multiple output (MIMO) channel. In particular, we focus on Layered Space Time systems, and as an example, we consider V-BLAST architecture and show how this system deals with the received signal in order to estimate correctly the transmitted data. We introduce the basic concepts of Layered Space Time processing.

At the beginning of Chapter 4, we introduce our research results on V-BLAST system performance in flat fading as well as in frequency selective for a
variety of modulation schemes, delay spread distribution, and different numbers of transmit and receive antennas.

Chapter 5 introduces traditional OFDM systems and some associated difficulties and requirements are also discussed. Furthermore, we propose a system which apply OFDM technique to V-BLAST architecture; namely, V-BLAST/OFDM, in order to mitigate the effects of time delay spread and to make the system more robust against frequency selective fading. The proposed system is investigated with different channel delay profiles and number of subcarriers in both stationary and time varying environments. In particular, we determine the key parameters for the system design and especially the proposed system delay spread tolerance. Analytical along with simulation results are provided to illustrate the feasibility of the system.

In Chapter 6, we examine the associated complexity of V-BLAST/OFDM system and present a possible way to reduce the system computation complexity while maintaining excellent performance.

We conclude the thesis in Chapter 6 by demonstrating the feasibility of these systems for wireless communication applications. In addition, some ideas worth future development to improve the current V-BLAST/OFDM system are presented. Unfortunately, the story is far from complete in this area, and research is continuing. The material in Chapter 6 will suggest additional ideas to the reader.

1.4 Publications

During these two years of master studies, achievements are evidenced by a publication record which includes


Chapter 2

Preliminaries

In contrast to wireline communications, mobile communications offers users mobility and allows users to communicate conveniently worldwide. However, transmission over the radio media is extremely random and does not offer easy analysis. The transmission may be reflected, diffracted and blocked by some obstacles inside the radio channel. In addition, the signal quality at the receiver may even be affected by the movement of the mobile subscriber. In terms of the stationarity and stability of the media, mobile communications is definitely not as good as wireline communications where the media is highly predictable. As a consequence, in mobile radio system design, it is critical to know the mechanisms behind electromagnetic wave propagation, so modeling of the radio channel is required. On the other hand, due to the limited spectrum and system channel characteristics, there are various interferences which greatly affect the signal quality. These impairments have a strong negative impact on mobile communications, and countermeasures like diversity must be used to improve the link performance in the hostile mobile environments. These characteristics are discussed in this chapter.

2.1 Channel

As radio waves propagate, the strengths of the waves decrease as the distance between the transmitter and receiver increases. This phenomenon is usually characterized by the Friis free space propagation model. In addition, due to the
fact that the surrounding environmental cluster may be vastly different at two different locations having the same distance from the transmitter, it has possibly a direct Line-of-sight (LOS) at one location, but there may be buildings blocking the LOS at another location. This is referred to as shadowing which describes the random shadowing effects caused by different levels of clutter on the propagation path [31]. Fortunately, these inherent impairments can be effectively overcome by some advanced power control techniques [21, 32, 33].

Apart from these effects, radio wave propagation can be attributed to reflection, diffraction and scattering. As radio wave propagates, multiple reflections, diffractions and scattering occur from various objects. The interaction between these waves causes multipath fading in which two or more versions of the transmitted signal arrive at different times and with different attenuation. These echoes cause various kind of interferences to the received signal depending on the speed of transmission and the channel bandwidth. With multipath fading, there will be rapid changes in signal strength over a small travel distance or time interval. Moreover, due to varying Doppler shifts on different multipath signals, spreading of the frequency spectrum or random frequency modulation may occur. In this section, the emphasis will be put on the modeling of the multipath fading channel.

2.1.1 N-Ray Multipath Channel Model

The small-scale variations due to multipath fading of a mobile radio signal can be modeled as the impulse response of a mobile radio channel. This impulse response is in general a time-varying continuous-time function and can reflect all the small-scale properties of the channel and simulate or analyze any type of radio transmission through the channel. The time-varying nature of the impulse response is due to the motion of the objects in the radio environment and the time variability mainly depends on the transmission rate and the speed of the receiver motion in space provided that the speeds of motion of surrounding objects at the transmitter and receiver are negligible. In this thesis, we par-
particularly consider high speed transmission where the channel can be thought of stationary for several packets, and usually a n-ray model is used to express the channel impulse response as [4, 35]

\[ c(t) = \sum_{n=0}^{N-1} \beta_n \delta(t - \tau_n) \]  

(2.1)

where \( \beta_n \) and \( \tau_n \) are the complex path gain, and time delay for the \( n \)th path of the mobile radio channel, respectively.

To determine \( \beta_n \), we can use either ray tracing (deterministic model) [31] or statistical approaches. Throughout the thesis, a statistical approach is used to allow easier control of the channel parameters. In the statistical approach, we assume that paths with different delays are uncorrelated (i.e., uncorrelated scattering), so that the path gains \( \beta_{n_1} \) and \( \beta_{n_2} \) are uncorrelated if \( n_1 \neq n_2 \). On the other hand, a commonly accepted model [36, 37] suggests that \(|\beta_n|\) be a random variable with Rayleigh distribution and \( \angle \beta_n \) be a random variable uniformly distributed from 0 to \( 2\pi \). Therefore, \( c(t) \) is a zero-mean complex Gaussian random variable. In addition, it is assumed that the paths with different delays within the excess delay are equally probable. Consequently, the path delay is assumed as constant increment delay, so \( \tau_n = \frac{\tau_{exc}}{N-1} n \) where \( \tau_{exc} \) is the excess delay of the multipath channel.

Most of the time, we will consider the composite channel impulse response, \( h(t) \), rather than the pure channel impulse response, \( c(t) \), as defined in (2.1). The composite channel impulse response, \( h(t) \), is equal to \((g \otimes c)(t)\) in which \( g(t) \) is the resultant pulse shaping filter at the transmitter and receiver, and \( \otimes \) denotes the convolution operator between two continuous time functions. The details of the pulse shaping filters at the transmitter and receiver will be discussed in the next section.

### 2.1.2 Channel Parameters

There are many quantitative parameters defined in the literature such as \( \text{rms} \) delay spread, Doppler spread, coherence time and coherence bandwidth, which
are commonly used to evaluate, quantify and compare different multipath channels, and these parameters usually provide useful design guidelines for engineers to develop wireless communication systems.

One of the most commonly used quantities to measure the time dispersive properties of wide band multipath channels is the \textit{rms} delay spread, $\tau$. It is defined as the square root of the second central moment of the power profile. That is,

$$
\tau = \left[ \frac{\int (t - d)^2 \mathbb{E}[|c(t)|^2] dt}{\int \mathbb{E}[|c(t)|^2] dt} \right]^{\frac{1}{2}}
$$

\hspace{1cm} (2.2)

where $d$ is the average delay or the centroid of $\mathbb{E}[|c(t)|^2]$ and is given by

$$
d = \frac{\int t \mathbb{E}[|c(t)|^2] dt}{\int \mathbb{E}[|c(t)|^2] dt}.
$$

\hspace{1cm} (2.3)

The delay spread, $\tau$, is a temporal or spatial average of consecutive impulse response measurements over a local area, and usually, $\tau$ is used to quantify how the multi-delay paths spread in time and measure the degree of seriousness of inter-symbol interference (will be defined in Section 2.2.1). Moreover, when the \textit{rms} delay spread is greater than the symbol period of transmission (i.e., $\tau > T$), the multipath channel is regarded as frequency selective fading channel. In contrast, the channel is termed frequency non-selective or flat fading channel if $\tau < T$.

In the past, many measurements were taken to determine a statistical range of $\tau$. Typically, in an urban environment, it is found that the \textit{rms} delay spread would be of the order of thousand nanoseconds while for indoor case, it would be a few hundred nanoseconds [38, 39].

Another channel parameter frequently used in this thesis is the coherence bandwidth, $B_c$. By definition, the coherence bandwidth is a frequency separation or bandwidth in frequency within which amplitudes of the frequency components are highly correlated. In other words, two signals in frequencies separated greater than $B_c$ passing through the same channel will be affected quite differently. Some analytical research [40] shows that an approximate coherence bandwidth can be
found by the knowledge of \( \text{rms} \) delay spread and is given by

\[
B_c \approx \begin{cases} 
\frac{1}{50\tau} & \text{for frequency correlation} > 0.9 \\
\frac{1}{5\tau} & \text{for frequency correlation} > 0.5.
\end{cases} \tag{2.4}
\]

In fact, exact close form formula of \( B_c \) does not exist, so simulation and spectral analysis should be done in order to know the exact impact of multipath fading on a particular mobile transmission. However, (2.4) gives a very good insight and feeling on the relationship between coherence bandwidth and \( \text{rms} \) delay spread, and this approximation will be used in our analysis to estimate the frequency correlation of the channel.

The time-varying nature of the channel is always characterized by two channel parameters namely, Doppler spread and coherence time. Because of the Doppler effects induced by the movement of objects in the channel, there are frequency spread or spectral broadening of the transmitted signal. The maximum shift in frequency is called the maximum Doppler shift, \( f_d \) which is given by \( f_d = v/\lambda \) where \( v \) is the speed of the mobile and \( \lambda \) is the wavelength of the radiation. In this equation, the motion of the mobile is assumed to be the dominant factor that makes the channel change, and this is almost the case in wireless communications.

The coherence time, \( T_c \), is defined similar to the coherence bandwidth, \( B_c \), discussed before, but in the time domain. The coherence time is the time duration where the channel does not vary much and is more or less the same. For modern digital communication,

\[
T_c \approx \sqrt{\frac{9}{16\pi f_d^2}} = \frac{0.423}{f_d} \tag{2.5}
\]

is often used to estimate the coherence time or stationarity of the channel [42]. Also, \( T_c \) is a measure to differentiate the fading rate of the channel. Normally, when coherence time is less than the symbol period of transmission (i.e. \( T_c < T \)), the multipath fading is regarded as fast fading. On the other hand, slow fading occurs when the coherence time is much greater than the transmitting symbol period (i.e. \( T_c \gg T \)).

Our analysis given in this thesis is based on a TDMA mobile transmission system. In particular, data are transmitted in packets over the radio channel.
and these packets may include a training sequence for equalization and/or synchronization purposes along with information data. As mentioned before, we assume the channel is quasi-stationary and can be considered as time invariant over a packet. In other words, the channel we have considered is a slow fading channel.

### 2.1.3 Exponential Power Profile

Instead of taking intensive measurement to get a realistic power delay profile of the channel, one of the idealized profiles which is commonly used to simulate or analyze the power delay profile of the channel is the exponential power delay profile. In terms of these channel parameters, the exponential power delay profile is expressed as

$$
E[|c(t)|^2] = \begin{cases} 
\frac{1}{D} \exp\left(-\frac{t}{D}\right) & \text{for } t \geq 0 \\
0 & \text{elsewhere.}
\end{cases}
$$

(2.6)

where $D$ is the normalized $rms$ delay spread which is defined as the $rms$ delay spread of the channel, $\tau$, divided by the symbol period of transmission, $T$, (i.e. $D \triangleq \tau/T$). Although other models of power profiles exist [5], the exponential power profile will be used throughout this thesis in order to have a more realistic channel model in analyzing the performance of our proposed systems. Moreover, to make the analysis more succinct and reduce the computing burden of simulations, we specifically consider only the paths with delay less than 5 normalized $rms$ delay spread or the excess delay spread, $T_{exc}$, is assumed to be $5D$. As a result, the path delay, $\tau_n$, becomes $\tau_n = \frac{5D}{N-1} n$. Moreover, the total power enclosed by the time period from 0 to $5D$ is above 99% of the total power of the channel and therefore, the truncation will only introduce negligible degradation of the analysis accuracy.

### 2.2 Interference

In mobile radio transmission, there are many types of interference jamming the desired radio signals, and all or at least most of these impairment should
be investigated and analyzed in the design of mobile transmission. Then, countermeasures should be designed aiming to mitigate the effects. In the following sections, we particularly consider the inter-symbol interference (ISI) and co-channel interference (CCI) which are usually the most dominant interferences in high speed transmissions.

2.2.1 Inter-Symbol Interference

The bandwidth of the channel is generally bandlimited and when a pulse with unlimited bandwidth (e.g., rectangular pulse) is transmitted through this channel, it will cause a truncation or distortion of the transmitting signal in the frequency domain. Equivalently, there is time dispersion of the pulse and the pulse for each symbol will smear into the time intervals of succeeding symbols. This type of interference is regarded as intersymbol interference (ISI) and this leads to an increased probability of making an error at the receiver in detecting a symbol [5].

Obviously, a bandlimited pulse can be chosen for transmission in order to avoid the distortion in the frequency domain due to the bandlimited channel. However, cutting the bandwidth of the transmitting pulse will in turns stretch or widen the width of the pulse in time. This causes overlapping of the symbols directly.

Fortunately, there are a number of well known pulse shaping techniques which can be used to simultaneously reduce the effects of ISI and the spectral width of a modulated digital signal. One of the most popular pulse shaping filters which is specifically considered in our analysis is the raised cosine rolloff filter. The raised cosine filter satisfies the Nyquist criterion [42] for ISI cancellation. Therefore, this pulse shaping filter can avoid ISI at every sampling time while keeping the signal bandwidth to be limited by a pre-defined finite boundary.
The transfer function of the filter is

$$
G(f) = \begin{cases} 
\frac{T}{2} \left[ 1 - \sin \left( \frac{\pi T}{\alpha} \left( |f| - \frac{1}{2T} \right) \right) \right] & 0 \leq |f| \leq \frac{1-\alpha}{2T} \\
\frac{1-\alpha}{2T} & \frac{1-\alpha}{2T} \leq |f| \leq \frac{1+\alpha}{2T} \\
0 & |f| \geq \frac{1+\alpha}{2T}
\end{cases}
$$

(2.7)

and the corresponding impulse response is

$$
g(t) = \frac{\cos \left( \frac{\alpha t}{T} \right) \text{sinc} \left( \frac{\pi t}{T} \right)}{1 - \left( \frac{2\alpha t}{T} \right)^2}
$$

(2.8)

where \( \alpha \) is the raised cosine rolloff factor which ranges between 0 and 1. For larger value of \( \alpha \), the impulse response of the filter decays faster, but with narrower signal bandwidth and vice versa. As a result, the rolloff factor, \( \alpha \), is in fact a design parameter for engineers and a spectral efficient mobile transmission without ISI can be possibly achieved if \( \alpha \) is properly adjusted.

Note that the raised cosine filter is theoretically time unlimited. In practice, a truncated version of the filter is used since it is impossible to transmit a signal forever. Normally, the filter is only generated for time period between \(-6T\) and \(6T\). However, this will only introduce very few performance degradation because of the fast impulse response decay at the zero-crossings (approximately as \(1/t^3\) for \( t \gg T \)).

Pulse shaping is done by incorporating a filter, \( \sqrt{G(f)} \), at both the transmitter and receiver in order to provide a matched filter [42] which is optimum in the sense that the received signal-to-noise ratio (SNR) is maximized in additive white Gaussian noise (AWGN) channel. By doing this, the performance of pulse shaping is not affected since the overall wireless communication system in our concern can be thought of as a linear time invariant (LTI) system, and hence the order of the subsystems is changeable or the splitting of the pulse shaping filter does not make any degradation in performance, but enhances noise immunity by matching the filters at the transmitter and receiver.

In fact, pulse shaping techniques can only avoid creation of ISI when a bandlimited transmission is considered in AWGN channel. As discussed in Section 2.1.1 and 2.1.2, radio channel induces multipath fading in mobile transmissions. As a consequence, even pulse shaping technique is employed, paths with delays greater the symbol period cause ISI by a multipath channel.
2.2.2 Co-Channel Interference

Apart from the interference caused by the channel, another interference which is also one of the most critical factors that limits the system performance and the capacity of the system is co-channel interference (CCI). CCI exists in any wireless multiple access system. In a frequency division multiple access (FDMA) or time division multiple access (TDMA) or space division multiple access (SDMA) system, frequency reuse is utilized to provide unlimited coverage of wireless system, so that there are users who share the same frequency band at the same time and hence, co-channel users surely create CCI one another. Consequently, there is always a trade-off between spectral efficiency and system performance.

Compared with FDMA and TDMA, SDMA reuses every frequency band more frequently to enhance the spectral efficiency and/or support more users by using spatial information of the co-channel users. In this system, multiple antennas are generally being used to form multi-directional beam pattern to capture and separate co-channel users transmitting at different locations [42]. The performance of this system highly depends on its rejection capability of CCI or the directivity of the antennas.

Another important wireless multiple access system, spread spectrum multiple access (SSMA) system also suffers the problem of CCI. Although the users, in theory, are using the same frequency band and can be separated by some orthogonal strategies such as direct sequence (DS), the near-far problem caused by imperfect correlation of the sequences and the impact of the relative distances of the co-channel users from the base is a major limitation to the performance and capacity of SSMA system [42]. As the above mentioning, some problems of CCI for various multiple access systems are addressed and thus, consideration of CCI must be included in investigating the performance of any wireless multiple access system.
2.3 System Model

A multipath radio propagation channel from the transmitter \( j \) \((j = 1, \ldots, M \text{ and } M \leq N)\) to receiver \( i \) \((i = 1, \ldots, N)\) can be described mathematically by its complex impulse response \( h_{ij}(t) \). In the following, we consider the baseband model for a multipath radio channel. Let \( u_j(t), j = 1, 2, \ldots, M \), be the baseband representation of the modulated waveform transmitted by the \( j-th \) antenna. The total received waveform by the \( i-th \), \((i = 1, 2, \ldots, N)\), antenna is then

\[
r_i(t) = z_i(t) + n_i(t)
\] (2.9)

where \( n_i(t) \) is the additive Gaussian noise at the \( i-th \) antenna and

\[
z_i(t) = u_j(t) * h_{ij}(t).
\] (2.10)

ISI caused by time delay spread results in frequency-selective fading. As we increase the SNR, an irreducible BER is approached because ISI increases in proportion to the signal level.

The impulse response of the channel from the transmitter \( j \) to receiver \( i \) can be expressed as [17]

\[
h_{ij}(t) = \sum_m A_m e^{i \Phi_m} \delta(t - \tau_m)
\] (2.11)

where \( A_m \) and \( \Phi_m \) are slowly varying random quantities that introduce a small Doppler shift for the wireless communication channel. A commonly used model [7], suggests that \( A_m \) be a random variable with Rayleigh distribution and \( \Phi_m \) be a random variable uniformly distributed on \([0, 2\pi]\). Therefore, \( h_{ij}(t) \) can be assumed to be a zero-mean complex Gaussian random variable. For \( t \neq t' \), it is reasonable to assume that \( h_{ij}(t) \) and \( h_{ij}(t') \) are uncorrelated since they are composed of signals from an independent set of paths with different scattering [5]. Consequently,

\[
<h_{ij}^*(t)h_{ij}(t)> = p(t) \delta(t - t')
\] (2.12)

where \(<.>\) denotes the ensemble average, and the function \( p(t) \) is the power delay profile, expressed as [5]

\[
p(t) = <|h_{ij}(t)|^2>
\] (2.13)
Chapter 3

Smart Antennas

3.1 Introduction

The technology of adaptive antennas for mobile communications has received enormous interest worldwide in recent years. The purpose of this chapter is to give a survey of the technology. A definition of adaptive antennas is used in this chapter, namely base station antennas with a pattern that is not fixed, but adapts to the current radio conditions.

The chapter starts with an explanation of the basic principles of MIMO systems. Different levels of intelligence are introduced, ranging from simple switching between predefined beams to optimum beamforming. Also, an evolutionary path for the development of smart antennas is suggested. A discussion of the consequences for mobile systems is given. The principle reason for introducing smart antennas is the possibility for a large increase in capacity: an increase of three times for TDMA systems and five times for CDMA systems has been reported. Other advantages include increased range and the potential to introduce new services. Major drawbacks and cost factors include increased transceiver complexity and more complex radio resource management. The basic principles for implementation of smart antennas at the base stations are explained. First, the general principles for beamforming using array antennas are explored, then the special cases of different smart antenna implementations are discussed.

It is foreseen that in the future an enormous increase in traffic will be experienced for mobile and personal communications systems. This is due both
to an increased number of users as well as new high bit rate data services being introduced. This trend is observed for second-generation systems and will most certainly continue for third-generation systems being introduced worldwide within a few years. The increase in traffic will put a demand on both manufacturers and operators to provide enough capacity in the networks. Presently, one of the most promising techniques for increasing the capacity in cellular systems by the amount necessary is smart or adaptive antennas.

This chapter will describe the basic principles behind the technology as well as provide a discussion about the advantages and technical challenges. Although it is feasible to also use these techniques in, for example, satellite systems and high frequency wireless local area networks, the discussion here is limited to cellular systems in the frequency range 1-2 GHz. At these frequencies it is difficult to introduce smart antennas into the mobile stations due to the limited size of the terminal in terms of wavelengths.

3.2 Smart Antennas

Traditionally, users communicating via the same base station have been separated by frequency, as in FDMA (frequency division multiple access); by time, as in TDMA (time division multiple access); or by code, as in CDMA (code division multiple access). Smart antennas add a new way of separating users, namely by space, through SDMA (space division multiple access). As will be explained later, SDMA, which means that users in the same cell can use the same physical communication channel, is the final step in an evolutionary path toward increasingly more advanced utilization. The physical communication channel is defined as a combination of carrier frequency, time slot and spreading code.

Base station antennas have up till now been omnidirectional or sectored. This can be regarded as a “waste” of power as most of it will be radiated in other directions than toward the user. In addition, the power radiated in other directions will be experienced as interference by other users. The idea of MIMO systems is to use base station antenna patterns that are not fixed, but adapt
to the current radio conditions. This can be visualized as the antenna directing a beam toward the communication partner only. Smart antennas will lead to a much more efficient use of the power and spectrum, reducing interference as well as increasing the useful received power and capacity.

In a cellular system the radio communication is between the user and a base station, which provides radio coverage within a certain area, called a cell. Capacity in such a system can be defined as the total bit rate per unit bandwidth per unit area, or bit/s/Hz/m². Because the available frequency band is limited, the capacity is given by the cell density, the frequency reuse distance and the number of users that can be served simultaneously by each base station. Techniques for increasing capacity in cellular systems have included using smaller cells, so-called microcells, and frequency hopping, a technique that disperses interference and averages the fading rate.

3.2.1 Basic Principles

The main philosophy is that interferers rarely have the same geographical location as the desired user. By maximizing the antenna gain in the desired direction and simultaneously placing minimal radiation pattern in the directions of the interferers, the quality of the communication link can be significantly improved. In personal and mobile communications, the interferers are other users than the user being addressed.

The difference between a smart/adaptive antenna and a “dumb”/fixed antenna is the property of having an adaptive and fixed lobe-pattern, respectively. Normally, the term “antenna” comprises only the mechanical construction transforming free electromagnetic (EM) waves into radio frequency (RF) signals traveling on a shielded cable and vice versa. We may call it the radiating element. In the context of smart antennas, the term “antenna” has an extended meaning. It consists of a number of radiating elements, a combining/dividing network and a control unit. The control unit can be called the smart antenna’s intelligence, normally realized using a digital signal processor (DSP). The processor
controls feeder parameters of the antenna, based on several inputs, in order to optimize the communications link. Different optimization criteria can be used. This shows that smart antennas are more than just the "antenna," but rather a complete transceiver concept.

We can define “levels of intelligence”. One set of definitions often used are described below.

- **Switched lobe (SL):** This is also called switched beam. It is the simplest technique, and comprises only a basic switching function between separate directive antennas or predefined beams of an array. The setting that gives the best performance, usually in terms of received power, is chosen. Because of the higher directivity compared to a conventional antenna, some gain is achieved. Such an antenna will be easier to implement in existing cell structures than the more sophisticated adaptive arrays, but it gives a limited improvement.

- **Dynamically phased array (PA):** By including a direction of arrival (DoA) algorithm for the signal received from the user, continuous tracking can be achieved and it can be viewed as a generalization of the switched lobe concept. In this case also, the received power is maximized.

- **Adaptive array (AA):** In this case, a DoA algorithm for determining the direction toward interference sources (e.g., other users) is added. The radiation pattern can then be adjusted to null out the interferers. In addition, by using special algorithms and space diversity techniques, the radiation pattern can be adapted to receive multipath signals which can be combined. These techniques will maximize the signal to interference ratio (SIR) (or signal to interference and noise ratio (SINR)).

Conventional mobile systems usually employ some sort of antenna diversity (e.g., space or polarization diversity). Adaptive antennas can be regarded as an extended diversity scheme, having more than two diversity branches. In this context, phased arrays will have a greater gain potential than switched lobe antennas because all elements can be used for diversity combining.
The introduction of smart antennas will have a large impact on the performance of cellular networks. It will also affect many aspects of both the planning and deployment of mobile systems. This chapter will discuss the potential benefits and cost factors, and will also briefly describe the implications on radio planning.

### 3.2.2 Improvements and Benefits

- **Capacity Increase:** The principle reason for the growing interest in smart antennas is the capacity increase. In densely populated areas, mobile systems are normally interference-limited, meaning that interference from other users is the main source of noise in the system. This means that the signal to interference ratio, SIR, is much larger than the signal to thermal noise-ratio, SNR. Smart antennas will on average, by simultaneously increasing the useful received signal level and lowering the interference level, increase the SIR. Especially, the adaptive array will give a significant improvement. Experimental results report up to 10 dB increases in average SIR in urban areas [6]. In TDMA systems, the implication of the increased SIR is the possibility for reduced frequency reuse distance. Simulations performed on a FH-GSM network with 1/3 reuse distance utilizing SFIR reports that a capacity increase of 300 percent can be expected [5]. CDMA systems, such as IS-95 or UMTS, are more inherently interference-limited than TDMA systems. The main source of noise in the system is the interference from other users due to the spreading codes being non-ideally orthogonal. This means that the expected capacity gain is even larger for CDMA than for TDMA. A fivefold capacity gain has been reported for CDMA in [7].

- **Range Increase:** In rural and sparsely populated areas radio coverage rather than capacity will give the premises for base station deployment. Because smart antennas will be more directive than traditional sector or omnidirectional antennas, a range increase potential is available. This
means that base stations can be placed further apart, potentially leading to a more cost-efficient deployment. The antenna gain compared to a single element antenna can be increased by an amount equal to the number of array elements, e.g., an eight-element array can provide a gain of eight (9 dB).

- New Services: When using smart antennas the network will have access to spatial information about the users. This information can be used to estimate the positions of the users much more accurately than in existing networks. Positioning can be used in services such as emergency calls and location-specific billing. The FCC (Federal Communications Commission) in the United States has decided that by October 2001 user location information with an accuracy of 125 meters RMS error must be provided [8].

- Security: It is more difficult to tap a connection when smart antennas are used. To successfully tap a connection the intruder must be positioned in the same direction as the user as seen from the base station.

- Reduced Multipath Propagation: By using a narrow antenna beam at the base station the multipath propagation can be somewhat reduced. The actual reduction depends on the scenario, and is not always significant. Although channel equalizers and RAKE receivers most often will handle and even exploit the multipath components, on very high-speed connections this may not be the case. Potentially, the reduction of multipath propagation can be used to ease the requirement on future modem design.

Space Division Multiplexing (SDM) is one technique that can exploit the smart antennas potential and provide a significant increase in capacity and the Bit Error Rate (BER) performance, such as V-BLAST [16]. Using V-BLAST, bandwidth efficiencies of 20 - 40 bps/Hz have been demonstrated in an indoor environment at realistic Signal-to-Noise-Ratios (SNRs) and error rates.
3.3 V-BLAST System

A high-level block diagram of the V-BLAST system is shown in Figure 3.1. A single data stream is demultiplexed into $M$ substreams [47], and each substream is then encoded into symbols and fed to its respective transmitter. Transmitters $1 \cdots M$ operate at a symbol rate of $1/T$ symbols/sec, with synchronized symbol timing. Each transmitter is itself an ordinary QAM transmitter. This is done under the assumption that the same constellation is used for each substream, and that transmissions are organized into bursts of $L$ symbols.

 Receivers $1 \cdots N$ are, individually, conventional QAM receivers. Each one receive the signals radiated from all $M$ transmit antennas. At first, and for simplicity, flat fading is assumed, and the matrix channel transfer function is $H^{N \times M}$, where $h_{ij}$ is the (complex) transfer function from the transmitter $j$ to receiver $i$ with $M \leq N$.

 Let $a = (a_1, a_2 \ldots a_M)^T$ denote the vector of transmit symbols. Then, the corresponding received $N$-vector is

$$r_i = Ha + n$$

where $n$ is a noise vector with components drawn from a white Gaussian noise process with variance $\sigma^2$. 

25
3.3.1 V-BLAST Detection Algorithm

Let the ordered set

\[ S = \{k_1, k_2 \ldots k_M\} \]  \hspace{1cm} (3.2)

be a permutation of the integers \{1, 2 \ldots M\} specifying the order in which components of the transmitted symbol vector \( \mathbf{a} \) are extracted. Also, let \( \hat{\mathbf{a}} \) denote the estimation of \( \mathbf{a} \). Then, the detection process proceeds generally as follow [47]:

Step 1: Using a nulling vector, \( \mathbf{w}_{k_1} \), form a decision statistic:

\[ y_{k_1} = \mathbf{w}_{k_1}^T \mathbf{r}_1 \]  \hspace{1cm} (3.3)

Step 2: Slice \( y_{k_1} \) to obtain \( \hat{a}_{k_1} \):

\[ \hat{a}_{k_1} = \Psi(y_{k_1}) \]  \hspace{1cm} (3.4)

where \( \Psi(.) \) denotes the quantization (slicing) operation appropriate to the constellation in use [47].

Step 3: Let \( (\mathbf{H})_{k_1} \) denotes the \( k_1 - th \) column of \( \mathbf{H} \). Then, compute the modified received vector \( \mathbf{r}_2 \) obtained by canceling \( \hat{a}_{k_1} \) from the received vector \( \mathbf{r}_1 \), resulting in a modified received vector:

\[ \mathbf{r}_2 = \mathbf{r}_1 - \hat{a}_{k_1} (\mathbf{H})_{k_1}. \]  \hspace{1cm} (3.5)

Steps (3.3) to (3.5) are then performed for components \( \{k_2, k_3 \ldots k_m\} \) by operating in turn on the progression of the remaining modified received vectors \( \{r_2, r_3 \ldots r_m\} \). The specifics of the detection process depend on the criterion chosen to compute the nulling vectors \( \mathbf{w}_{k_i} \), the most common of these being MMSE or ZF [47]. The detection process is described here with respect to the latter. The \( k_i - th \) ZF-nulling vector is defined as the unique minimum norm vector satisfying

\[ \mathbf{w}_{k_i}^T \mathbf{H}_{k_j} = \begin{cases} 0 & \text{if } j > i \\ 1 & \text{if } j = i. \end{cases} \]  \hspace{1cm} (3.6)

Thus, \( \mathbf{w}_{k_i} \) is orthogonal to the subspace spanned by the contributions to \( \mathbf{r}_i \) due to those symbols not yet estimated and canceled. It is not difficult to show
that the unique vector satisfying (3.6) is just the $k_i$-th row of $H^+_{k_i-1}$, where the
notation $H^+_{k_i}$ denotes the matrix obtained by zeroing columns $k_2, k_3 \ldots k_i$ of $H$
and $\oplus$ denotes the Moore-Penrose pseudo-inverse. The post-detection signal-to-
one ratio (SNR), $\rho_{k_i}$, for the $k_i$-th detected component of $a$ is easily obtained
by substituting (3.1) and (3.6) into (3.5), and taking expected values. That is,

$$
\rho_{k_i} = \frac{\langle |a_{k_i}|^2 \rangle}{\sigma^2 ||w_{k_i}||}
$$

where the expectation in the numerator is taken over the constellation set.

### 3.3.2 Optimal Detection Ordering

The full ZF V-BLAST detection algorithm can now be described compactly
as a recursive procedure which includes the determination of the optimal ordering.
This is briefly described as follows [47]:

**Initialization:** Let $G_1 = H^+, i = 1$, and

$$
k_1 = \arg \min_j ||(G_1)_j||^2
$$

**Recursion:** Let

$$
w_{k_i} = (G_i)_{k_i}
$$

$$
y_{k_i} = w^T_{k_i} r_i
$$

$$
\hat{a}_{k_i} = \Psi(y_{k_i})
$$

$$
r_{i+1} = r_i - \hat{a}_{k_i}(H)_{k_i}
$$

$$
G_{i+1} = H^+_{k_i}
$$

$$
k_{i+1} = \arg \min_{j \in \{k_1 \ldots k_i\}} ||(G_1)_j||^2
$$

$$
i = i + 1
$$

in the above algorithm, $(G_i)_j$ denotes the $j$-th row of $G_i$. Thus, Eqns. (3.8)-(3.14)
determine the elements of the optimal ordering, $S_{opt}$. Eqns. (3.9)-(3.11)
compute, respectively, the ZF-nulling vector, the decision statistic, and the es-
timated component of $a$. Finally, Eqn. (3.12) performs the cancellation of
the detected component from the received vector, and (3.13) computes the new
pseudo-inverse for the next iteration. Note that this new pseudo-inverse is based on a "deflated" version of $H$, in which the columns $k_1, k_2 \ldots k_i$ have been zeroed. This is because these columns correspond to components of $a$ which have already been estimated and canceled. Hence, the system becomes equivalent to a "deflated" version of Figure 3.1 in which transmitters $k_1, k_2 \ldots k_i$ have been removed, or equivalently, a system in which $a_{k_1} = \ldots = a_{k_i} = 0$.

### 3.4 Conclusions

It is obvious that smart antennas at the base stations will be an important technology to provide the necessary capacity and coverage. It also helps to realize new services, e.g., based on user location. From a technology point of view, smart antennas can be seen as an extension of the "conventional" resource allocation schemes used in radio communications. In addition to dividing the space into cells, it will now also be possible to employ space division inside each cell. Different degrees of utilization of the spatial dimension are possible.

V-BLAST architecture is a particular concept of smart antennas systems. This technique can provide a significant increase in capacity and system performance. Basically, different parallel data streams are transmitted, simultaneously and on the same frequency, using an antenna array. When using multiple antennas at the receiver as well, these data streams, mixed-up by the wireless channel, can be recovered by a powerful Layered Space Time processing.
Chapter 4

Performance of V-BLAST Over Flat and Frequency-Selective Wireless Communication Channels

4.1 Introduction

In previous published work, the performance of V-BLAST has been investigated with 8 transmit and 12 receive antennas along with 16-QAM constellation in a flat fading environment. In this chapter, we investigate the performance of V-BLAST for a variety of configurations including different number of transmit and receive antennas. We are also interested in multipath fading channels. Frequency selective fading channels can severely degrade system performance by causing ISI. This will then result in an irreducible BER which imposes an upper limit on the data rate [4, 5]. This impairment is thus a crucial issue that becomes more important in the next generation wireless systems where broadband and high data rate communications services are needed [9]. Hence, in this chapter we will study, the performance of V-BLAST in multipath fading channels. In addition, we will investigate the effect of time delay spread tolerance on V-BLAST performance.

The organization of this chapter is as follows. Subsection 4.2 describe the key simulation steps. The results obtained and interpretations are given in Subsection 4.3. Finally, some conclusions are presented in Subsection 4.4.
TABLE I
TU CHANNEL: PATH DELAYS AND RELATIVE POWER LEVELS

<table>
<thead>
<tr>
<th>Path, k</th>
<th>Path delay, $t_k$ [μs]</th>
<th>$p(t_k)$ [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.0</td>
<td>-3.0</td>
</tr>
<tr>
<td>2</td>
<td>0.199</td>
<td>0.0</td>
</tr>
<tr>
<td>3</td>
<td>0.502</td>
<td>-2.0</td>
</tr>
<tr>
<td>4</td>
<td>1.606</td>
<td>-6.0</td>
</tr>
<tr>
<td>5</td>
<td>2.307</td>
<td>-8.0</td>
</tr>
<tr>
<td>6</td>
<td>5.017</td>
<td>-10.0</td>
</tr>
</tbody>
</table>

4.2 Key Simulation Steps

In our simulations, we consider coherent detection since it is more resistant to delay spread than differential detection [5].

4.2.1 Power Delay Profile

The power delay profile $p(t)$ is sampled at $t = t_n$ where $k = 1, 2, \ldots, K$ and where the delay spread is not too large.

The channel impulse response at $t_n$, denoted by $h_{ij}(t_k) = h_{ik} + jh_{qk}$, is a complex zero-mean Gaussian random process with variance $p(t_k)$. Therefore, $h_{ik}$ and $h_{qk}$ are uncorrelated Gaussian random variables each with variance $p(t_k)/2$. A sample of the channel impulse response from the transmitter $m$ ($m = 1, \ldots, M$ and $M \leq N$) to the receiver $n$ ($n = 1, \ldots, N$) can then be constructed as:

$$h_{nm}(t) = \sum_{k=1}^{K} (h_{nm}^{ik} + jh_{nm}^{qk}) \delta(t - t_k).$$  \hspace{1cm} (4.1)

We evaluate the performance in frequency-selective fading channels using a stationary (no Doppler) 2-ray channel. Then, more investigations are performed using a Typical Urban (TU) channel with spatial rich scattering. The TU channel profile is detailed in Table I [57, 58].

30
4.2.2 Carrier Recovery

In a previous paper [5], it has been shown that the average of \( h_{nm}(t) \) over several symbols can be approximated by the average over the hole time axis. In this work, it is assumed that the bandwidth of the carrier recovery circuit used is much higher than the channel fading rate and the bandwidth of the carrier recovery loop is much lower than the symbol rate. Therefore, the carrier phase can be tracked as if \( h_{nm}(t) \) is time invariant. Hence, the recovered phase is

\[
\Psi_{nm}(t) \approx \text{Phase of } \left( \int h_{nm}(t) \, dt \right) = \text{Phase of } (H_{nm}(f)|_{f=0})
\]

(4.2)

where \( H_{nm}(f) \) is the Fourier transform of \( h_{nm}(t) \). In the simulations, the carrier phase is extracted from the average of \( h_{nm}(t) \) over several symbol periods. Thus, the carrier recovery is treated by replacing \( h_{nm}(t) \) by \( h_{nm}(t)e^{j\Psi_{nm}} \), where the recovered carrier is represented by

\[
e^{j\Psi_{nm}} = \frac{H_{nm}(0)}{|H_{nm}(0)|}.
\]

(4.3)

4.2.3 Timing Recovery and Sampling

For all modulations considered, the optimal receiver filter in the absence of delay spread is the matched filter with impulse response \( g^*(-t) \). This is the received filter assumed in the simulation. A squaring timing loop [4], is used in the simulation to recover timing clock. The multipath channel causes the timing delay, \( t_d \), expressed for any pulse signaling waveform, \( g(t) \), as

\[
t_d = \frac{T}{2\pi} \tan^{-1} \frac{F_{ic}}{F_{qc}}
\]

(4.4)

where

\[
F_{ic} = \int |h_{nm} * g(t)|^2 \sin(2\pi t/T)dt
\]

(4.5)

and

\[
F_{qc} = \int |h_{nm} * g(t)|^2 \cos(2\pi t/T)dt.
\]

(4.6)

In our simulation and for most cases, \( g(t) \) is real and even. Therefore, it does not introduce any time delay. Hence, the effects of the multipath channel on
Figure 4.1: V-BLAST performance with different QAM modulations

timing recovery can be studied by replacing $|h_{nm}(t) \ast g(t)|^2$ by $p(t)$, the power delay profile. Besides, for small values of $d$, the recovered timing tracks the centroid of $|h_{nm}(t)|^2$ [4].

Once $t_d$ is computed, the demodulation waveform $z_n(t)$ is sampled at $t = t_d + kT$ for the $k$-th bit; The resulting decision statistic is then used for symbol detection by comparing it with a threshold to determine the output bits [4].

4.3 Simulation Results in Flat Fading

4.3.1 Results for Different Modulation Schemes

In this section, we evaluate the block error probability (BLER) of V-BLAST as a function of the average SNR in a Rayleigh flat fading environment. Simulation results are presented to investigate the performance of V-BLAST with different modulation schemes, and different number of antennas at both transmitter and receiver.
At first, we compare our results with those of the Bell Laboratories prototype. To achieve this goal, we focus our attention on simulating V-BLAST under the same conditions as that of [16]. Hence, we use 8 antennas at the transmitter and 12 at the receiver \((M = 8, N = 12)\) with 16-QAM modulation over a Rayleigh flat fading channel. The simulation results, as shown in Figure 2, are consistent with the earlier available analytical and experimental results of V-BLAST (See Fig. 2 in [16]). It is noted that in Figure 4.1, the horizontal axis is the spatial average received SNR and the vertical axis is the block error rate \((BLER)\), where a block is defined as a single transmission burst. Throughout this chapter, the burst length is 80 symbols. We also assume ideal channel estimation.

In Figure 4.1, we also study V-BLAST performance using \(J\)-QAM modulation with \(J = 4, \ldots, 32\). A close observation of this figure indicates that to maintain the same level of performance at a \(BLER = 10^{-2}\), for example, a 3.5\(dB\) increase in average SNR is needed as we increase \(J\) from 4 to 8. However, note that the spectrum efficiency increases by 5.18 bps/Hz for each additional bit per symbol. In fact, in a 30 kHz bandwidth, with a symbol rate of 19.44 ksymbols/sec, the raw spectral efficiency for 4-QAM is

\[
E_s = \frac{(8TX) \times (2 b/sym/TX) \times (19.44 ksym/sec)}{30 kHz} = 10.36 \text{ bps/Hz.}
\]

The spectrum efficiency is about 15.5 bps/Hz for 8-QAM, 20.7 bps/Hz for 16-QAM, and 25.9 for 32-QAM. As a result, we can see that V-BLAST really achieves very high bandwidth efficiency over a Rayleigh flat fading channel. However, this is done at the expense of the block error rate.

### 4.3.2 Results for Different Number of Antennas

Here, we investigate the performance of the system architecture as we vary the number of transmit and receive antennas in a flat fading environment. It is noted that in Figure 4.2, the horizontal axis is the spatial average received SNR and the vertical axis is the \(BLER\), where a block is defined as a single
transmission burst. Throughout this chapter, the burst length is fixed to 80 symbols. We also assume ideal channel estimation. First, we fix the number of transmit antennas at 8 \((M = 8)\) and we vary the number of receive antennas form 8 to 30. As shown in Figure 4.2, the case of 8 antennas at both sides \((N = M = 8)\) corresponds to the worse case scenario. It requires an average SNR of 27dB to achieve a quality of service (QoS), or a BLER of \(10^{-1}\). The same QoS can be reached with 9RX \((N = 9)\) with 4.5dB reduction in SNR. A BLER of \(10^{-2}\) can be obtained with 15 antennas at the receiver at about 19dB SNR. As a result, we conclude that as the number of receive antennas increases, the performance of the system gets better but with a limited gain for more than 12RX \((N \geq 12)\).

Next, we fix the number of receive antennas and we vary the number of transmit antennas. Figure 4.3 presents the performance achieved as we vary \(M\) from 2 to 10. In this case, the trade-off between the QoS and the spectrum
Figure 4.3: *V-BLAST performance with a fix number of receive antenna (N = 12) and a variable number of transmit antennas (N).*

efficiency is studied. The more transmit antennas we use, the more the system capacity increases, and the larger the spectrum efficiency we obtain (2.59 bps/Hz per each transmit antenna). But, this is done at the expense of the BLER, which gets worse as $M$ increases. For example, for a BLER of $10^{-1}$, assume that we use $4TX$ ($M = 4$) instead of $2TX$ ($M = 2$), then we would require approximately an additional $4dB$ to achieve the same performance. However, this provides about 5.18 bps/Hz gain in the spectrum efficiency and a double system capacity.

We also investigate the system performance as we vary the number of transmit and receive antennas to achieve a certain *BLER*. We consider a flat fading environment and 16-QAM modulation scheme. Figure 4.4 shows the trade-off between the number of antennas required at both the transmitter and receiver for a fixed *BLER = 10^{-2}* at an average *SNR* of 17 dB, 12 dB and 7 dB. We can see that, at 17dB, for each additional transmit antenna, about two more antennas at the receiver are required for $M \leq 7$, and three for $M = 8 \cdots 10$. 

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Figure 4.4: Number of antennas required to attain $BLER = 10^{-2}$ with 16-QAM as a function of $SNR$.

However, when the average $SNR$ is $12dB$, the number of receive antennas $N$ required to achieve $BLER = 10^{-2}$ is approximately twice the number of transmit antennas.

4.4 Results in Frequency Selective Fading Channel using 2-ray power delay profile

In this section, we consider a frequency selective fading channel using a two ray equal amplitude power delay profile. We investigate the effect of time delay spread on V-BLAST performance as a function of the parameter $d$ using 16-QAM modulation. We also consider a rectangular signaling pulse with a matched receiver filter. The number of antennas is fixed at both the transmitter and the receiver ($M = 8, N = 12$).

Figure 4.5 shows a comparison between V-BLAST performance in flat fading
Figure 4.5: Performance of V-BLAST over a multipath fading channel.

and frequency selective fading environments with different values of $d$. The block error rate is represented as a function of the average SNR. A close observation of the results indicates that the BLER is severely degraded when the $rms$ time delay spread becomes relatively high.

To determine the influence of the normalized time delay spread on the irreducible BLER, more simulations are performed with different modulation schemes. Figure 4.6 shows the average irreducible BLER performance as a function of $d$. The parameter $d$ is varied from 0 to 1 in 0.05 steps. It is clear that for small values of $d$, ISI causes a very small perturbation. However, the performance degradation becomes relatively severe when $d$ increases. As a result, delay spread mitigation techniques such as adaptive multichannel equalization or orthogonal frequency division multiplexing (OFDM) should be considered. Nevertheless, we can see that V-BLAST architecture is more resistant to the effect of multipath delay spread than traditional systems.
Figure 4.6: The performance of V-BLAST as a function of the rms delay spread normalized by the symbol period.

4.5 Results in Frequency Selective Fading Channel using TU power delay profile

Using the configuration described in Section 4.2, we provide simulation results of V-BLAST for various J-QAM modulation (where $J=4, ..., 32$), different number of transmit and receive antennas ($M = 4-12$, $N = 8-18$) using both the TU and equal-power two-ray delay profiles. In the simulations, the substream symbol period is $T = 3.692\mu s$ (as for GSM) [57, 58] providing a normalized rms delay spread of 0.2886 for the TU model listed in Table 1. We assume a quasi-stationary channel, so that its time variation is negligible over the 80 symbol periods comprising a burst in our simulations. Perfect channel estimation is assumed and carrier and timing recovery are performed in the manner described in Section 4.2.

To verify the accuracy of our simulations, we first compare our results with
Figure 4.7: V-BLAST performance with various QAM modulations \((M=8, N=12)\) those of the Bell Laboratories prototype. To achieve this goal, we focus our attention on simulating V-BLAST under the same conditions as that of [16] with \(M = 8\) antennas at the transmitter, \(N = 12\) at the receiver and 16-QAM modulation scheme over a Rayleigh flat-fading channel. The simulation results, shown in Figure 1, are consistent with the earlier available analytical and experimental results of V-BLAST (See Fig. 2 in [16]). In Figure 4.7, the horizontal axis indicates the spatial average received \(SNR\) and the vertical axis the uncoded \(BLER\), where a block is defined as a single transmission burst.

### 4.5.1 Results for Different Modulation Schemes

In Figure 4.7, we also study the V-BLAST performance in a TU environment using \(J\)-QAM modulation with \(J = 4, \ldots, 32\). To maintain the same level of performance at \(BLER = 10^{-2}\), for example, a 5\(dB\) increase in average \(SNR\) is needed as we increase \(J\) from 4 to 8. However, note that the spectral efficiency increases by 5.18 bps/Hz for each additional bit per symbol. The spectral effi-
Figure 4.8: **V-BLAST performance with a fixed number of transmit antennas (M = 8) and a variable number of receive antennas (N) in a TU environment.**

Efficiency is about 15.5 bps/Hz for 8-QAM, 20.7 bps/Hz for 16-QAM, and 25.9 for 32-QAM. As a result, we can see that V-BLAST really achieves very high bandwidth efficiency over a wireless channel. However, this is done at the expense of the BLER.

Figure 4.7 also lists the V-BLAST performance over a flat fading channel using J-QAM modulation with $J = 4, \ldots, 32$. A close comparison shows that performance degrades most when we use a higher level modulation scheme.

### 4.5.2 Results for Different Number of Antennas

In this section, we study the performance of the system architecture as we vary the number of transmit and receive antennas in the TU fading environment for 16-QAM modulation. First, we fix the number of transmit antennas at $M = 8$ and we vary the number of receive antennas from $N = 8$ to $N = 18$. As shown in Figure 4.8, the case of $M = 8$ and $N = 8$ corresponds to the worse case scenario.
Figure 4.9: \textit{V-BLAST performance with a fixed number of receive antennas (N = 12) and a variable number of transmit antennas (M) in a TU fading environment.}

It requires an average \textit{SNR} of 28 dB to achieve a \textit{BLER} of $4 \times 10^{-2}$, for instance. The same \textit{BLER} can be reached with \textit{N} = 10 with 3 dB reduction in \textit{SNR}. A \textit{BLER} of $10^{-2}$ can be obtained with 16 antennas at the receiver at about 19 dB \textit{SNR}. This confirms that V-BLAST performance can be significantly improved by increasing the number of receive antennas [59].

Next, we fix the number of receive antennas at \textit{N} = 12 and we vary the number of transmit antennas. Figure 4.9 shows the performance as we vary \textit{M} from 4 to 12. Again, the trade-off between the \textit{BLER} and the spectral efficiency is studied. The spectral efficiency increases when we use more transmit antennas. However, this is at the expense of the \textit{BLER}. 

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Figure 4.10: The performance of V-BLAST as a function of the rms delay spread normalized by the symbol period.

4.5.3 V-BLAST performance with different power delay profiles

In this section, we investigate the effect of two different delay spread distributions (the TU channel described in Table 1 and an equal-power two-ray delay profile [5]) on V-BLAST. The number of antennas is fixed at both the transmitter and the receiver ($M = 8, N = 12$) and the system performance is investigated as a function of the parameter $d$, defined as the (rms) delay spread normalized by the substream symbol period $T$ [4].

Figure 4.10 shows the average irreducible BLER performance as a function of $d$. The $rms$ delay spread is varied from 0 to 1 in 0.05 steps. The system performance is severely degraded when the $rms$ time delay spread exceeds a certain limit that depends on the modulation scheme as shown in Figure 4.10. We notice that the irreducible BLER with 2-ray delay profile and TU profile are
very close. For small values of $d$, ISI causes a very small perturbation. However, the performance degradation becomes severe as $d$ increases. As a result, delay spread mitigation techniques should be considered for this range of delay spreads [25]. Since both profiles have the same $rms$ delay spread, it can be concluded that the irreducible $BLER$ is not too sensitive to the channel power delay profile for this range of delay spreads.

4.6 Conclusions

In this chapter, we first considered the performance of V-BLAST in a flat fading environment for several system parameters such as the number of receive and transmit antennas. In contrast to previous work, we have considered the performance of V-BLAST in a frequency selective fading channel. In particular, we investigated the effect of time delay spread on V-BLAST. Our results show that the presence of multipath can severely affect V-BLAST performance, especially when the delay spread becomes relatively high and this result in intolerable performance degradation. As a result, to maintain system performance at acceptable levels, delay spread mitigation techniques such as adaptive multi-channel equalization or OFDM should be employed with V-BLAST in multipath fading channels.
Chapter 5

A Layered Space-Time Coded Wideband OFDM Architecture

5.1 Introduction

The deployment of broadband wireless access systems would require a transmission technique which can mitigate the detrimental effects of frequency selective fading [17]. In recent years, there has been much interest in applying multicarrier techniques in wireless systems because of its various advantages in mitigating the severe effects of frequency-selective fading [60, 61, 18].

In this chapter, we focus our attention on such a system in rich-scattering environments where V-BLAST performs at its best. We propose to combine V-BLAST with coded OFDM to mitigate the effects of time delay spread and to make the system more robust against frequency selective fading. The proposed system is investigated with different channel delay profiles and number of subcarriers in both stationary and time varying environments. In particular, we determine the key parameters for the system design and especially the proposed system delay spread tolerance.

The remaining parts of this chapter are organized as follows. Section 5.2 describes the V-BLAST/OFDM system model. In Section 5.3, we present the different simulation parameters and results which demonstrate the potential of the proposed system to deal with the outdoor multipath environment. Finally, we conclude the chapter in Section 5.4.
5.2 Layered Space-Time OFDM System Model

The V-BLAST architecture is based on a single carrier signal processing algorithm. Therefore, to combine it with OFDM, the V-BLAST detection process has to be performed on every subcarrier at the receiver. The detailed system configuration of V-BLAST/OFDM is shown in Figures 5.1-5.2. For a \((M,N)\) antenna configuration system, a vector symbol of size \(M\) is transmitted by every subcarrier over a rich-scattering wireless channel, at each symbol time.

At the transmitter, each of the \(M\) data substreams is encoded, modulated and sent through a different transmit antenna. At each transmitter chain, we deploy an OFDM modem which consists of \(K\) subcarriers. Let \(\{X^i_{mp}\mid p = 0, \ldots, K-1\}\) denote the \(K\) subcarrier symbols where \(i, p\) and \(m\) represent, respectively, the indeces of the OFDM symbol and the corresponding subcarrier and transmit antenna.

A serial to parallel converter converts the transmitted information bit sequence into \(K\) subsequences. Each subsequence from the converter keys one of \(K\) orthogonal subcarriers and can be independently encoded and mapped to symbols selected from the signal space. These subsequences are passed through inverse fast Fourier transformation blocks (IFFTs). The keyed subcarriers are summed into an OFDM symbol. A cyclic prefix is also added in the guard interval to each of the resulting signals, for final transmission. This extension is set to the excess delay of the radio channel in order to reduce the effect of inter-symbol interference (ISI) and inter-subcarrier-interference. As a result, the output signal of each transmitter chain can be represented as

\[
S_m(t) = \sum_{i=-\infty}^{\infty} \sum_{p=0}^{K-1} X^i_{mp} \Psi_{i,p}(t). \tag{5.1}
\]

where \(\Psi_{i,p}(t)\) is the subcarrier pulse. That is,

\[
\Psi_{i,p}(t) = \begin{cases} 
  e^{j2\pi \frac{f_c}{T_s} (t-iT_s)} & iT_s \leq t < (i+1)T_s \\
  0 & \text{else} \end{cases} \tag{5.2}
\]
where $T_s = T_u + \Delta$ is the OFDM symbol duration, $1/T_u$ is the OFDM subcarrier spacing, $T = T_u/K$ is the sampling time interval, and $\Delta = k_1T$ is the guard interval length.

At the receiver, we assume perfect OFDM synchronization. Therefore, at the sampling time $t = iT_s + (k_1 + k)T$ corresponding to the $i^{th}$ OFDM symbol and the $k^{th}$ subcarrier, the output is

$$S_m^k = \sum_{i=-\infty}^{\infty} \sum_{p=0}^{K-1} X_{m,p}^{i,k} \Psi_{i,p}(iT_s + (k_1 + k)T).$$  \hspace{1cm} (5.3)
Therefore, the received signal can be written as

\[ R^k = H^k S^k + N^k \]

(5.4)

where

\[
H^k = \begin{bmatrix}
H_{1,1}^k & H_{1,2}^k & \cdots & H_{1,M}^k \\
H_{2,1}^k & H_{2,2}^k & \cdots & H_{2,M}^k \\
\vdots & \vdots & \ddots & \vdots \\
H_{N,1}^k & H_{N,2}^k & \cdots & H_{N,M}^k
\end{bmatrix}
\]

is the channel matrix corresponding to the \( k \)-th subchannel, \( S^k = (S_1^k, S_2^k, \ldots, S_m^k)^T \) is a \( M \times 1 \) transmitted vector, \( R^k = (R_1^k, R_2^k, \ldots, R_n^k)^T \) is a \( N \times 1 \) received vector, and \( N^k = (N_1^k, N_2^k, \ldots, N_n^k)^T \) is assumed to be an additive white Gaussian noise vector with zero mean and auto-correlation matrix

\[ E(N^kN^{kH}) = \sigma^2 I_n \]

(5.5)

where \( I_n \) denotes the identity matrix of dimension \( n \) and \((.)^T\), \((.)^H\) are, respectively, the transpose and the hermitien of \((.)\).

During the reception, each antenna receives the signal transmitted from all the \( M \) transmit antennas. First, the cyclic prefix of each received signal is removed. After passing through a serial-to-parallel converter and the fast Fourier transformation blocks (FFTs), the subcarriers are separated. Then, the \( N \) information symbols belonging to each subcarrier are routed to their corresponding V-BLAST multi-antenna processing unit. Hence, we estimate the \( M \) transmitted data signals per each carrier. Finally, the demapping and decoding are performed. The detected bits are converted back into serial form accordingly to recover the transmitted data bits.

### 5.3 Channel Model

We assume that the OFDM signal is transmitted over a wireless communication environment characterized by a multipath fading channel and a given coherence bandwidth. The channel response of each subcarrier is flat, in the
frequency domain, when the central frequency spacing between adjacent subcarriers is much less than the coherence bandwidth. For a practical channel, the multipath model is represented by its channel impulse responses using a L-ray model defined as [17],

$$h_{n,m}(\tau, t) = \sum_{l=0}^{L-1} h_{n,m}^l(t) \cdot \delta(\tau - \tau_{n,m}^l)$$  \hspace{1cm} (5.6)

where \(\{h_{n,m}^l(t)\}\) are the i.i.d. path complex Gaussian gains with zero mean and unit variance, \(\{\tau_{n,m}^l\}\) are the path time delays and \(L\) is the total number of the channel paths between the \(m^{th}\) transmit and the \(n^{th}\) receive antenna, respectively, at the mobile station (MS) and the base station (BS). The channel frequency response at the subcarrier frequency \(f_k = \frac{k}{T_s}\) can be written as

$$H_{n,m}^k = \sum_{l=0}^{L-1} h_{n,m}^l(kT) \cdot e^{-j2\pi \tau_{n,m}^l \frac{k}{T_s}}.$$  \hspace{1cm} (5.7)

The FFT output for the \(k^{th}\) subcarrier of the received signal at the \(n^{th}\) antenna can be represented by

$$R_n^k = \sum_{m=0}^{M} H_{n,m}^k \cdot S_m^k + N_n^k$$  \hspace{1cm} (5.8)

Here we assume that the channel does not change during one OFDM symbol and that the guard interval length is larger than the channel maximum delay spread.

### 5.4 Simulation Parameters and Results

The V-BLAST/OFDM system proposed in Figures 5.1-5.2 has been simulated in stationary and time varying environments. The antenna configuration consists of 4 transmit antennas, and 6 receive antennas. In order to increase the bandwidth efficiency, a 16-QAM multi-amplitude bit rate is also considered with coherent detection on each of the receiver branches. We apply forward error-correcting techniques to compensate for the problems produced by the environment and corrects for subcarriers in deep fades. By doing so, reliable performance for the system can be ensured.
For simplicity, we consider Reed-Solomon coding across the subchannels, with 1/2 coding rate. Reed-Solomon coding with 8-bit symbols (i.e., GF(256)), corresponding to grouping of two 16-QAM symbols, is used for error correction. We use a (24,12) code, which correct 6 erasures, based on the signal strength, and 3 randomly errored symbols. With the reduction in the delivered bit rate due to the 1/2-rate code, the peak rate for the sample 16-QAM system is divided by two.

5.4.1 Stationary Channel

In this section, we will evaluate the system performance for different number of subcarriers. Hence, we consider a frequency selective fading channel using a stationary (no Doppler), equal amplitude, two-ray delay profile with spatial rich scattering, expressed as

\[ p(t) = \frac{1}{2} [\delta(t - \tau) + \delta(t + \tau)] . \]  \hspace{1cm} (5.9)

A sufficient guard period (20 μsec) is considered to minimize the effect of delay spread as large as 20 μsec. In urban areas, the typical measured values of the Root Mean Square (RMS) delay spread range from 5μsec to 20μsec [31]. Therefore, the channel RMS delay spread considered in our simulations is \( \sigma_r = 5.0353 \mu\text{sec} \).

In Figure 5.4, the system Frame Error Rate (FER) is shown for different number of subcarriers versus the average SNR per receive antenna for 128, 256, 512 and 1024 subcarriers. The frame here refers to a block of 72 bits for the uncoded system, or one Reed-Solomon code word for the coded system. We can see that as we increase the number of subcarriers from 128 to 256, there is a significant improvement in the system performance. However, the improvement in the FER, obtained when we double the number of subcarriers from 512 to 1024, is not as great. This is can be explained by examining the channel response of each subchannel, which changes from being frequency selective to flat fading in the frequency domain when the frequency spacing between adjacent subcarriers become much less than the coherence bandwidth.
Figure 5.3: Performance of V-BLAST/OFDM system in a 2-ray equal gain fading channel.

Assuming that the entire channel bandwidth of 4.096 MHz, is divided into K subchannels, which results in subchannel spacing of $\Delta f = B/K = 8$ kHz, for $K = 512$ Sub-carriers for example. To make the tones orthogonal to each other, we must have a symbol duration, $T_s = 1/\Delta f = 125$ $\mu$sec (for $K = 512$). On the other hand, if we define the coherence bandwidth as the bandwidth over which the frequency correlation function is above 0.9 [31], then $B_c = 1/(50 \sigma_f) = 3.9720$ kHz. Therefore, it is clear that for less than 512 subcarriers, some ISI will occur and they will degrade the system performance. However, for 1024 subcarriers, each subchannel sees almost a flat fading.

Figure 5.4 shows that there is a significant improvement in the system performance when we use Reed-Solomon coding. In fact, with only 256 subcarriers, we can achieve better performance than uncoded flat fading. Hence, coding for this system is essential for achieving high transmitting rates with a relatively small number of subcarriers.
TABLE 5.1
HT CHANNEL: PATH DELAYS AND RELATIVE POWER LEVELS

<table>
<thead>
<tr>
<th>Path number</th>
<th>Path delay (μsec)</th>
<th>Average power</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.0</td>
<td>0.3933</td>
</tr>
<tr>
<td>2</td>
<td>0.2</td>
<td>0.2481</td>
</tr>
<tr>
<td>3</td>
<td>0.4</td>
<td>0.1566</td>
</tr>
<tr>
<td>4</td>
<td>0.6</td>
<td>0.0785</td>
</tr>
<tr>
<td>5</td>
<td>15.0</td>
<td>0.0988</td>
</tr>
<tr>
<td>6</td>
<td>17.2</td>
<td>0.0248</td>
</tr>
</tbody>
</table>

Figure 5.4: Performance of V-BLAST/OFDM system in HT channel.

In order to investigate the performance of V-BLAST/OFDM system for different power delay channel profiles, more simulations are performed in a hilly terrain with the same parameters. The HT channel profile is detailed in Table 5.1.

Figure 5.4 shows the uncoded and coded FER versus the average SNR for different number of subcarriers using the HT channel profile. As can be seen,
the performance of the system is improved by using a larger number of subchannels. For example, for a target uncode FER of $10^{-2}$, less than 24 dB SNR is required when we use 1024 subcarriers, rather than a higher SNR (more than 35 dB) when using only 128 subcarriers. This represents a large reduction in the link budget. However, for a target code FER of $10^{-2}$, we have only 2.5 dB reduction in the link budget. This is achieved at the expense of system complexity. Therefore, there is a tradeoff between the number of subcarriers and the system performance. This tradeoff becomes more significant when we consider the outdoor (macrocell) wireless propagation characterized by a relatively high time delay spread in comparison with the indoor (microcell) propagation. Thus the V-BLAST/OFDM system in outdoor environment is more sensitive to delay spread than in indoor environment. Hence, again we show the importance of coding in our V-BLAST/OFDM system for mitigating the effect of ISI caused by time delay spread. Moreover, when we use 128 subcarriers, we can achieve an FER of $10^{-2}$ with only 22 dB SNR when we use R-S coding. However, 24.5 dB are required when coding is not used.

The use of 16-QAM modulation, improves the system spectral efficiency at the expense of the SNR required. This is because, as we increase the modulation order, the distance between the constellation points is reduced. Hence, clearly, the coded V-BLAST/OFDM system mitigates the effect of time delay spread and ensures a high spectral efficiency.

A close comparison of our system performance using the two-ray and HT channel delay profiles for the same RMS delay spread, show that the system is slightly more robust to ISI when we use the latter model. This can be explained by the fact that the number of paths in the HT profile is greater, so we have more multipath components. This may give somewhat an optimistic diversity effect because diversity is proportional to the number of paths. However, these models do not affect V-BLAST/OFDM system delay tolerance, since it does not depend on the number of paths but rather on the number of paths exceeding a certain excess delay. Thus, the HT model seems a good basis to determine the V-BLAST/OFDM delay spread tolerance which exceeds 5 \microsec. This delay
Figure 5.5: *FER versus the average SNR for the HT delay profile RMS delay spread is a) 5μsec, b) 2.5μsec, c) 1.5μsec, d) 0.5μsec.*

spread tolerance allow the use of V-BLAST/OFDM architecture to deal with multipath and time delay spread in outdoor (macrocell) environment.

5.4.2 Time varying channel

In this section, the OFDM signal is composed of 120 subchannels so that data modulates each tone with a 160 μsec symbol period. An additional 40 μsec guard interval is used to eliminate any ISI due to the channel delay spread. This results in subchannels which are spaced by 6.25 kHz, block rates of 5 kbaud, and a total rate of 600 kbaud, or, equivalently, channel rates of 2.4 Mbps for 16-QAM before coding. In these simulations, Doppler rates as high as 100 Hz are considered. For these parameters, we can assume that the OFDM subchannels are orthogonal since the blocks are short enough. We also consider an HT environment and the same Reed-Solomon code as before.

From Figure 5.5, it can be concluded that the FER performance decreases
when the delay spread increases. This performance deterioration can be explained as follows. If the delay spread increases, the fading changes from flat to frequency selective especially when the delay exceeds the coherence time. However, coding helps the system to maintain an acceptable performance level. In fact, there are several fades within the OFDM signal bandwidth, with relatively strong subcarriers in between. The coding benefits from these stronger subcarriers to compensate for the attenuated subcarriers. However, if the delay spread becomes larger than 1.5 μsec, the performance deteriorates, because in this case the paths with a large delay can’t be resolved and will appear as ISI. Hence, for relatively high delay, 120 subcarriers is not enough to achieve an acceptable performance level.

5.5 Conclusions

In this chapter, we proposed a combination of V-BLAST algorithm with OFDM modulation to achieve high speed services for mobile users in macrocells over frequency selective fading and time varying channels. The proposed technique is a promising solution for achieving spectral efficient transmission in a scattered environment and overcoming the dispersive fading limitations of the radio channels. To demonstrate the potential of this proposed technique, we investigated the V-BLAST/OFDM system performance in multipath time varying outdoor environments. We also studied the effect of error forward coding on the system robustness using different channel power delay profiles. From the simulation results, it is concluded that the proposed system can provide a robust solution against frequency selective fading in high capacity wideband wireless and mobile communication systems.
Chapter 6

A Low Complexity Multi-carrier BLAST Architecture

6.1 Introduction

Clearly, the OFDM/V-BLAST system involves intensive computation and hence it may be difficult to implement it for high data rate communications\[30\]. Thus, we propose a low complexity multicarrier V-BLAST based on subcarrier grouping, using sub-optimal ordering instead of the optimal one and applying GSO to the channel matrix $H$ to find the weight vectors.

The remaining parts of the chapter are organized as follows. Section 6.2 describes the low complexity layered space time OFDM system. In Section 6.3, we present the different simulation parameters and results which demonstrate the potential of the proposed system. Finally, we conclude the chapter in Section 6.4.

6.2 The Low Complexity Receiver Structure

At the multi-antenna processing unit and for a given subcarrier, the strongest signal is picked out first and this involves the calculation of the Moore-Penrose pseudo-inverse of the channel matrix $H$. Second, the signal is decoded and after signal re-generation, it is subtracted from the $N$ received signals. Third, the corresponding column in $H$ is nulled. Then, this process is repeated and in each
step, the strongest signal, among the remaining undecoded signals, is always
decoded first. Hence, we estimate the $M$ transmitted data signals per each
carrier. It has been proved [16] that this decoding ordering leads to optimum
performance. In the above process, a series of calculations of pseudo-inverse is
required which is computationally intensive.

In this chapter, we consider a method to reduce the total number of arith-
metic operations for calculating the weight vectors $w'_{k_i}$. This sub-optimal or-
dering is obtained only from the first pseudo-inverse by sorting $\| (G_1)_j \| ^2$ in
an ascending order, for $j = 1, \ldots, M$, where $G_1 = H^H$. Then, the series of
calculations of pseudo-inverse can be replaced by applying Gram-Schmidt Or-
thogonalization (GSO) once to $H$, provided that the decoding ordering is known
in advance. This is a modification of the Least Square Method, which is a sim-
ple method to calculate the pseudo-inverse. This method consists of two parts.
The first part is to find a basis of the columns of $H$ and the second part is an
operation like backward substitution.

6.2.1 Finding the weight vector by applying GSO to the
channel matrix

Given the sub-optimal ordering $S' = [k'_1, k'_2, \ldots, k'_M]$, we can obtain the
weight vectors $w'_{k_i}$ simply by applying GSO to $H$. This is a modification of the
Least Square Method [62], which is a simple method to calculate the pseudo-
inverse. This method consists of two parts. The first part is to find a basis of the
columns of $H$ and the second part is an operation like backward substitution.
Denote $h_n$ as the $n^{th}$ column of $H$, $b^*$ as the complex conjugate of $b$ and $< u, v >$
as the dot product of the vectors $u$ and $v$.

Given the ordering $S' = [k'_1, k'_2, \ldots, k'_M]$

for $i = 1$ to $M$

\[ l = M - i + 1 \]
\[ u_i = h_{k_i} - \sum_{j=1}^{i-1} h_{k_{j'}} b_j > b_j \]

\[ b_i = \frac{u_i}{\| u_i \|} \]

\[ w_{k_i}' = \frac{b_i^*}{\| u_i \|} \]

end

By taking advantage of the Least Square Method, the series of calculations of pseudo-inverse can be replaced by applying GSO once as shown above.

Table 6.1 is the comparison of the number of arithmetic operations between V-BLAST and the proposed method. It is assumed, in Table 6.1, that the method for obtaining the detection ordering in both V-BLAST and the proposed system are the same and that method is the same as the one described in Section 3.3.2. The number of arithmetic operations for the calculation of the weight vectors in the V-BLAST algorithm is an order of magnitude larger than that of the proposed method. Therefore, the reduction increases with the number of transmitter antennas and the number of receiving antennas.

### 6.2.2 Finding Norms

The norm of each row of \( H^l \) are required for finding both the optimal ordering and the sub-optimal ordering. Normally, one needs to first calculate the pseudo-inverse of \( H \) and then compute the norm of each row of \( H^l \). However, through the modification of the Least Square Method [62], one can find out the norms by applying GSO to \( H \) with some more calculations.

To find the norms by the proposed method, we first apply GSO to the columns of \( H \). The ordering of the columns, i.e. \( S = [k_1, k_2, \ldots, k_M] \) is arbitrary. In fact, by choosing an appropriate ordering, further reduction can be achieved as described in section VI. After the calculation of GSO, a matrix \( D \) is obtained and the \((i, j)^{th}\) element of \( D \) is

\[ \text{for } i, j = 1, \ldots, M \]
\[ D_{i,j} = \begin{cases} <h_{k_{M-i+1}}^{*}, b_j> & \text{if } i > j \\ \| u_i \| & \text{if } i = j \\ 0 & \text{otherwise} \end{cases} \]

Denote \( Y \in \mathbb{C}^{M \times M} \) be a square matrix, which is used to store the norms (the results) and some intermediate values. The following equations are the remaining part for calculating the norms:

for \( i = M \) to 1

for \( j = M \) to \( i + 1 \)

\[
Y_{j,i} = \sum_{k=i+1}^{M} \frac{D_{k,i}Y_{j,k}}{-D_{i,i}}
\]

\[
Y_{i,j} = Y_{j,i}^{*}
\]

end

\[
Y_{i,i} = \frac{1 + \sum_{j=i+1}^{M} \| D_{j,i} \|^2 Y_{j,j}}{D_{i,i}^2} + \frac{2 \sum_{j=i+1}^{M} \sum_{k=j+1}^{M} Re[D_{j,i}D_{k,i}^{*}Y_{k,j}]}{D_{i,i}^2}
\]

end

\( Y_{i,i} \), where \( i = 1, \ldots, M \), is the norm of the \((k_{M-i+1})^{th}\) row of \( H^\dagger \).

Further reduction can be gained since the weight vectors are unchanged or changes a little between successive subcarriers. Thus, for each group of \( g \) subcarriers, we can use the same weight vectors corresponding to those of the subcarrier in the middle of the group. This subcarrier grouping leads to a huge complexity reduction with insignificant degradation in performance as long as the the number of subcarriers per group is small enough.

Table 6.1 shows the comparison of the required number of arithmetic operations per subcarrier for obtaining both the decoding order and the weight vectors between V-BLAST/OFDM and the proposed system. This complexity reduction in the calculation of the weight vectors increases with the number of transmitting and receiving antennas. Hence, by using the proposed receiver,
every type of arithmetic operation can be reduced by over 50% in the (4,6) OFDM/V-BLAST system, for instance. Furthermore, totally the proposed system leads to a 63% reduction.

<table>
<thead>
<tr>
<th></th>
<th>Total</th>
</tr>
</thead>
<tbody>
<tr>
<td>V-BLAST/OFDM</td>
<td>$4NM^3 + 6NM^2 + 2NM$</td>
</tr>
<tr>
<td></td>
<td>$-\left(\frac{1}{3}M^3 + 0.5M^2 + \frac{1}{3}M\right)$</td>
</tr>
<tr>
<td>Proposed System</td>
<td>$[16NM^2 + \frac{13}{3}M^3 + \frac{14}{3}M + 1$</td>
</tr>
<tr>
<td></td>
<td>$-\left(4NM + 8M^2 + 6N\right)]$</td>
</tr>
<tr>
<td>Reduction (M=4, N=6, g=8)</td>
<td>63%</td>
</tr>
</tbody>
</table>

### 6.3 Simulation Parameters and Results

The proposed system has been investigated with different channel delay profiles, different number of subcarriers and different subcarrier group sizes in both stationary and slow time varying environments.

#### 6.3.1 Stationary Channel

In the simulation, the antenna configuration consists of 4 transmit antennas, and 6 receive antennas. In order to increase the bandwidth efficiency, the 16-QAM multi-amplitude bit rate. We apply forward error-correcting techniques to compensate for the problems produced by the environment and correct for subcarriers in deep fades. For simplicity, we consider Reed-Solomon coding across the subchannels, with 1/2 coding rate. Reed-Solomon coding with 8-bit symbols (i.e., GF(256)), corresponding to grouping of two 16-QAM symbols, is used for error correction. We use a (24,12) code, which corrects 6 erasures, based on the signal strength, and 3 randomly symbols received incorrectly.
Figure 6.1: The proposed system performance in a 2-ray equal gain fading channel with different subcarrier group sizes.

First, we consider a frequency selective fading channel using a stationary (no Doppler), equal amplitude, two-ray delay profile. The channel RMS delay spread considered in our simulations is $\sigma_r=5.0353\mu$sec.

In Figure 6.1, the system coded FER is shown for different number of subcarrier sizes versus the average SNR per receive antenna for 8, 16 and 32 subcarriers per group. The frame here refers to a block of 72 bits for the uncoded system, or one Reed-Solomon code word for the coded system. It is clear that the proposed system performance degradation is very small in comparison with the exhaustive V-BLAST/OFDM. We can also see that as we decrease the group size from 32 to 16 and then to 8, there is a significant improvement in the system performance. This can be explained by effect of the weight error on the data estimation which increases when the group size increases.

In order to investigate the performance of V-BLAST/OFDM system for different power delay channel profiles, more simulations are performed in a hilly
Figure 6.2: Performance of the proposed system in HT channel with different number of subcarriers.

terrain (HT) with 8 subcarriers per group and different numbers of subcarrier. Figure 6.2 shows the coded FER versus the average SNR for different number of subcarriers using the HT channel profile.

As can be seen, the performance of the system is improved by using a larger number of subchannels. This can be explained by examining the channel response of each subchannel, which changes from being frequency selective to flat fading in the frequency domain when the frequency spacing is between adjacent subcarriers become much less than the coherence bandwidth. Hence, clearly, the coded V-BLAST/OFDM system mitigates the effect of time delay spread and ensures a high spectral efficiency.

A close observation of our system performance using the two-ray and HT channel delay profiles for the same RMS delay spread, show that the system is slightly more robust to ISI when we use the latter model. This can be explained by the fact that the number of paths in the HT profile is greater, so we have more
Figure 6.3: FER of the proposed system versus the average SNR for the HT delay profile.

multipath components. This may give an optimistic diversity effect because diversity is proportional to the number of paths.

6.3.2 Time varying Channel

In Figure 6.3, the OFDM signal is composed of 120 subchannels so that data modulates each tone with a 160 μsec symbol period. An additional 40 μsec guard interval is used to eliminate any ISI due to the channel delay spread. In these simulations Doppler rates as slow as 10 Hz are considered and each group has 10 subcarriers. It can be concluded that the FER performance decreases when the delay spread increases. This performance deterioration can be explained as follows. If the delay spread increases, the fading changes from flat to frequency selective especially when the delay exceeds the coherence time. However, coding helps the system to maintain an acceptable performance level. In fact, there are several fades within the OFDM signal bandwidth, with relatively strong
subcarriers in between. The coding benefits from these stronger subcarriers to compensate for the attenuated subcarriers. However, if the delay spread becomes larger than 1.5 $\mu$sec, the performance deteriorates, because in this case the paths with a large delay can not be resolved and will appear as ISI. Hence, for relatively high delay, 120 subcarriers is not enough to achieve an acceptable performance level. By the fact that the decoding ordering is fairly stationary in a slow fading channel, we can obtain a further reduction of 14.69%. Totally, our proposed method can lead to a reduction of 77.84% with a very small performance degradation.

6.4 Conclusions

In this chapter, we proposed a combination of V-BLAST algorithm with OFDM modulation to achieve high speed services for mobile users over frequency selective fading and time varying channels. The proposed technique, is a promising solution for achieving spectral efficient transmission in a scattered environment and overcoming the dispersive fading limitations of the radio channels. To demonstrate the potential of this proposed technique, we have investigated the V-BLAST/OFDM system performance in multipath time varying environments. We have also studied the effect of error forward coding on the system robustness using different channel power delay profiles. In addition, we have proposed a low complexity system based on subcarrier grouping with negligible performance degradation and significant computation reduction. Simulation results demonstrate that the proposed system can provide a robust solution against frequency selective fading in high capacity wideband wireless and mobile communication systems.
Chapter 7

Conclusions

Smart antennas have received a lot of attention in research and development of wireless personal communication systems. V-BLAST is such an example that uses multi-element antenna arrays at both the transmitter and receiver and a powerful Layered Space Time processing in order to increase the system capacity. In this thesis, we investigated three aspects including the V-BLAST performance in flat and frequency selective fading channels, the role of coded multi-carrier modulation to improve system performance in wideband channels, and the huge complexity reduction achieved by the proposed algorithm.

7.1 V-BLAST in Flat and Multipath Channels

In this thesis, we considered the performance of V-BLAST in a flat fading environment for several system parameters such as modulation schemes as well as the the number of receive and transmit antennas. We also investigated the performance of V-BLAST in frequency-selective fading channels. We have looked at the system performance with different power delay profiles and different antennas configurations.

Our results show that the presence of multipath can severely affect V-BLAST performance, especially when the delay spread exceeds a certain limit that depends on the modulation scheme and this results in intolerable performance degradation. As a result, we have employed OFDM, as a delay spread mitigation technique, to maintain V-BLAST system performance at acceptable levels.
in multipath fading channels.

7.2 Multi-carrier V-BLAST Architecture

In recent years, there has been much interest in applying multicarrier techniques in wireless systems because of its various advantages in mitigating the severe effects of frequency-selective fading. We proposed a combination of V-BLAST algorithm with Orthogonal Frequency Division Multiplexing (OFDM) modulation to achieve high speed services for mobile users over frequency selective fading and time varying channels.

The proposed technique, is a promising solution for achieving spectral efficient transmission in a scattered environment and overcoming the dispersive fading limitations of the radio channels. To demonstrate the potential of this proposed technique, we have investigated the V-BLAST/OFDM system performance in multipath time varying environments. We have also studied the effect of error forward coding on the system robustness using different channel power delay profiles. From the simulation results, it is concluded that the proposed system can provide a robust solution against frequency selective fading in high capacity wideband wireless and mobile communication systems.

7.3 A Low Complexity Multi-tone V-BLAST

We have proposed a low complexity system based on subcarrier grouping with negligible performance degradation and significant computation reduction. This method reduces the total number of arithmetic operations for calculating the weight vectors $w'_k$. This sub-optimal ordering is obtained only from the first pseudo-inverse by sorting $\| (G_1)_j \|^2$ in an ascending order, for $j = 1, \ldots, M$, where $G_1 = H^H$. Then, the series of calculations of pseudo-inverse can be replaced by applying Gram-Schmitt Orthogonalization (GSO) once to $H$, provided that the decoding ordering is known in advance. This is a modification of the Least Square Method, which is a simple method to calculate the pseudo-inverse.
This method consists of two parts. The first part is to find a basis of the columns of $H$ and the second part is an operation like backward substitution.

Further reduction can be gained since the weight vectors are unchanged or changes a little between successive subcarriers. Thus, for each group of subcarriers, we can use the same weight vectors corresponding to those of the subcarrier in the middle of the group. This subcarrier grouping leads to a huge complexity reduction with insignificant degradation in performance as long as the number of subcarriers per group is small enough.

Simulation results demonstrate that the proposed system can provide a robust solution against frequency selective fading in high capacity wideband wireless and mobile communication systems with a 50% computation reduction using $(4,6)$ antennas configuration and 16QAM modulation scheme.

### 7.4 Suggestions For Further Research

In most communication systems multiple users must be supported. Multi-carrier modulation can be applied in a multiuser application producing a highly flexible, efficient communications system. Little work has been previously done on multiuser OFDM. The system design of a multiuser OFDM system is dependent on the intended application and the system complexity. One possible approach is to look at multiuser OFDM/V-BLAST as a potential technique that could be used to make it a highly efficient and reliable communication system. Additionally, we can look to a reduced complexity Digital Signal Processors (DSP) algorithm that simplifies our multiuser OFDM/V-BLAST system. These techniques can potentially enhance the system performance and allow a high spectral efficiency and reliability.

Adaptive modulation has recently received a lot of interest since it can significantly increase the spectral efficiency of the overall system. Our multi-carrier V-BLAST system uses a fixed modulation scheme over all carriers for simplicity. The carrier modulation is designed to provide an acceptable BER under the worst channel conditions. However, each carrier in our system can poten-
tially have a different modulation scheme depending on the channel conditions. Any coherent or differential, phase or amplitude modulation scheme can be used including BPSK, QPSK, 8PSK, 16QAM, 64QAM. Each modulation scheme provides a trade off between spectral efficiency and the BER. The spectral efficiency can be maximized by choosing the highest modulation scheme that will give an acceptable BER.
REFERENCES


