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5G PDSCH Simulator and DRL Approach For Designing The Optimal Precoder Vector and Power Allocation in MU-MIMO

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Chapter 1

Overview

1.1 The mobile industry in numbers

Mobile penetration is approaching saturation in most markets around the world, especially among adult and urban populations. In every region, the majority of new subscribers will be young consumers and rural dwellers. Despite increasing saturation in developed regions, there is still room for growth in many large, underpenetrated markets in developing regions. For example, India and Sub-Saharan Africa will account for around half of new mobile subscribers globally over the 2022–2030 period "Globally, there were 4.4 billion mobile internet users in 2022, equivalent to 55The mobile internet usage gap has narrowed markedly in the last five years – from 50% in 2017 to 41% in 2022 on average – as more people around the world rely on the internet for many daily activities, especially in the wake of the Covid-19 pandemic.

1.1.1 5G in the mobile market

"5G will overtake 4G in 2029 to become the dominant mobile technology by the end of this decade" 5G adoption continues to rise due to new network deployments and cheaper devices. As of January 2023, there were 229 commercial 5G networks around the world and over 700 5G smartphone models had been launched, including more than 200 in 2022. The number of connections on legacy networks (2G and 3G) will continue to decline in the coming years as users migrate to 4G and 5G, resulting in more network shutdowns. To date, operators have announced plans to shut down 96 2G networks and 107 3G networks around the world.

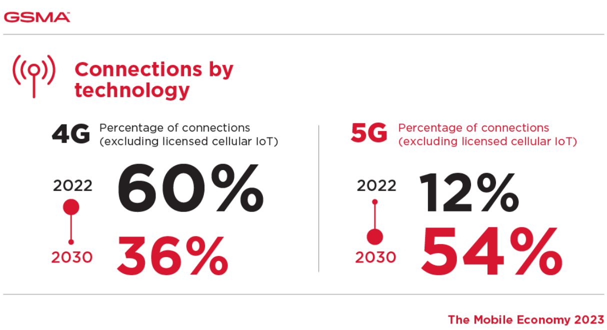


Figure 1.1: 4G vs 5G connections

1.2 The next generation – 5G NR

Despite LTE being a very capable technology, there are requirements not possible to meet with LTE or its evolution. Furthermore, technology development over the more than 10 years that have passed since the work on LTE was initiated allows for more advanced technical solutions. To meet these requirements and to exploit the potential of new technologies, 3GPP initiated the development of a new radio access technology known as NR (New Radio). A workshop setting the scope was held in the fall of 2015 and technical work began in the spring of 2016. The first version of the NR specifications was available by the end of 2017 to meet commercial requirements on early 5G deployments already in 2018. NR reuses many of the structures and features of LTE. However, being a new radio-access technology means that NR, unlike the LTE evolution, is not restricted by a need to retain backwards compatibility. The requirements on NR are also broader than what was the case for LTE, motivating a partly different set of technical solutions.

1.2.1 5G Use Cases

In the context of 5G, one is often talking about three distinctive classes of use cases: enhanced mobile broadband (eMBB), massive machine-type communication (mMTC), and ultra-reliable and low-latency communication (URLLC)

- eMBB corresponds to a more or less straight forward evolution of the mobile broadband services of today, enabling even larger data volumes and further enhanced user experience, for example, by supporting even higher end-user data rates.
- mMTC corresponds to services that are characterized by a massive number of devices, for example, remote sensors, actuators, and monitoring of various equipment. Key requirements for such services include very low device cost and very low device energy consumption, allowing for very long device battery life of up to at least several years. Typically, each device consumes and generates only a relatively small amount of data, that is, support for high data rates is of less importance.
- URLLC type-of-services are envisioned to require very low latency and extremely high reliability. Examples hereof are traffic safety, automatic control, and factory automation.

It is important to understand that the classification of 5G use cases into these three distinctive classes is somewhat artificial, primarily aiming to simplify the definition of requirements for the technology specification. There will be many use cases that do not fit exactly into one of these classes. Just as an example, there may be services that require very high reliability but for which the latency requirements are not that critical. Similarly, there may be use cases requiring devices of very low cost but where the possibility for very long device battery life may be less important.

1.2.2 Waveform, Numerology, and Frame Structure

The choice of radio waveform is the core physical layer decision for any wireless access technology. After assessments of all the waveform proposals, 3GPP agreed to adopt orthogonal frequency division multiplexing (OFDM) with a cyclic-prefix (CP) for both DL and UL

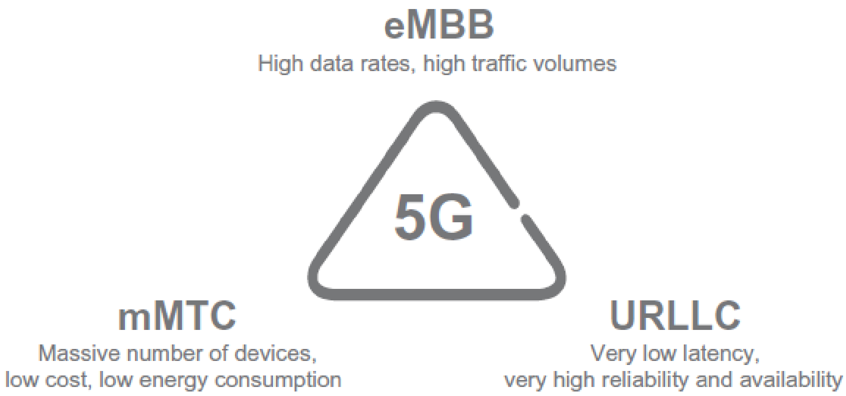


Figure 1.2: 5G Use cases classification

transmissions. CP-OFDM can enable low implementation complexity and low cost for wide bandwidth operations and multiple-input multiple-output (MIMO) technologies. NR supports operation in the spectrum ranging from sub-1 GHz to millimeter wave bands. Two frequency ranges (FR) are defined in Release 15:

- FR1: 450 MHz – 6 GHz, commonly referred to as sub-6 GHz.
- FR2: 24.25 GHz – 52.6 GHz, commonly referred to as millimeter wave.

Scalable numerologies are key to support NR deployment in such a wide range of spectrum. NR adopts flexible sub-carrier spacing of $2^\alpha \times 15$ kHz ($\alpha = 0, 1, ..., 4$) scaled from the basic 15 kHz sub-carrier spacing in LTE. Accordingly, the CP is scaled down by a factor of 2^α from the LTE CP length of 4.7 μs. This scalable design allows support for a wide range of deployment scenarios and carrier frequencies. At lower frequencies, below 6 GHz, cells can be larger and sub-carrier spacings of 15 kHz and 30 kHz are suitable. At higher frequencies, cells and delay spread are typically smaller and the CP lengths provided by the 60 and 120 kHz numerologies are sufficient.

A frame has a duration of 10ms and consists of 10 subframes. This is the same as in LTE, facilitating NR and LTE coexistence. Each subframe consists of 2^α slots of 14 OFDM symbols each. Although a slot is a typical unit for transmission upon which scheduling operates, NR enables transmission to start at any OFDM symbol and last only as many symbols as needed for the communication. This type of “mini-slot” transmission can thus facilitate very low latency for critical data as well as minimize interference to other links per the lean carrier design principle that aims at minimizing transmissions. Latency optimization has been an important consideration in NR. Many other tools besides “mini-slot” transmission have been introduced in NR to reduce latency.

1.3 5G Data Channels: Logical, Transport & Physical

In order to be able to carry the data across the 5G radio access network, the data and information is organized into a number of data channels. By organizing the data into various channels the 5G communications system is able to manage the data transfers in an orderly fashion and the system is able to understand what data is arriving and hence it is able to process it in the

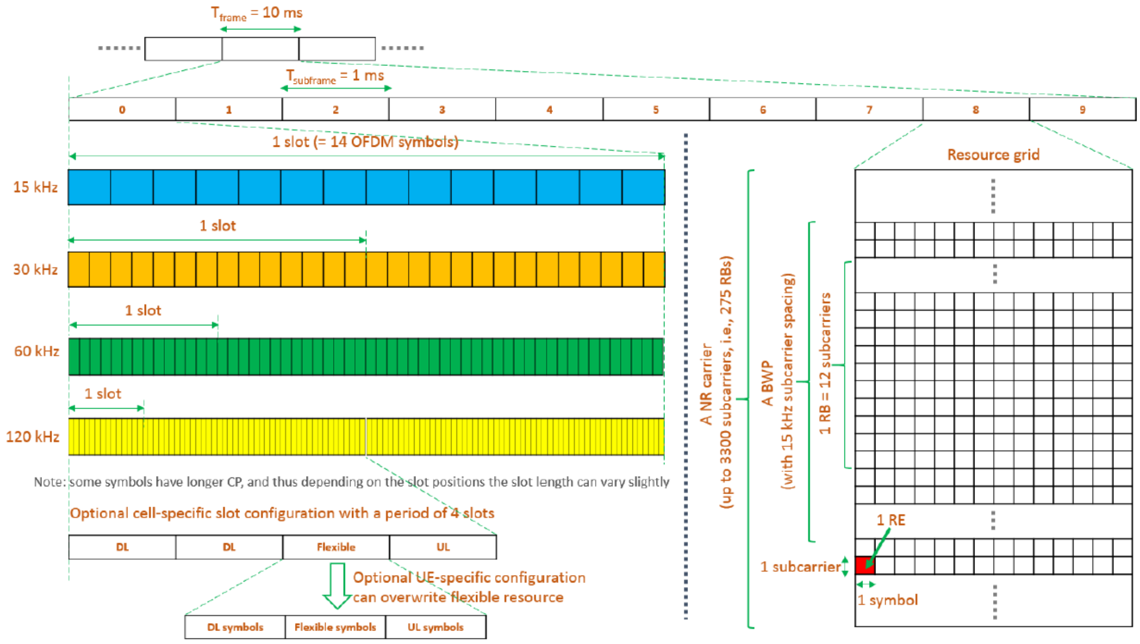


Figure 1.3: 5G Frame structure

required fashion. As there are many different types of data that need to be transferred - user data obviously needs to be transferred, but so does control information to manage the radio communications link, as well as data to provide synchronization, access and the like. All of these functions are essential and require the transfer of data over the radio access network. The 5G mobile or wireless communications system uses a similar access stratum to that used by 4G LTE. Although there are two protocol stacks: user plane and control plane, they still adopt the familiar OSI reference model. As a result there are various protocol layers and accordingly there are several data channel layers that are defined for the radio communications. In order to group the data to be sent over the 5G NR radio access network, the data is organised in a very logical way. As there are many different functions for the date being sent over the radio communications link, they need to be clearly marked and have defined positions and formats.To ensure this happens, there are several different forms of data "channel" that are used. The higher level ones are "mapped" or contained within others until finally at the physical level, the channel contains data from higher level channels. In this way there is a logical and manageable flow of data from the higher levels of the protocol stack down to the physical layer. There are three main types of data channels that are used within mobile communications systems. This is true for 5G systems, and accordingly the hierarchy is given below.

- Logical channel: Logical channels can be one of two groups: control channels and traffic channels:
  - Control channels: The control channels are used for the transfer of data from the control plane
  - Traffic channels: The traffic logical channels are used for the transfer of user plane data.
- Transport channel: Is the multiplexing of the logical data to be transported by the physical layer and its channels over the radio interface.
- Physical channel : The physical channels are those which are closest to the actual transmission of the data over the radio access network / 5G RF signal. They are used to

carry the data over the radio interface.

The physical channels often have higher level channels mapped onto them of provide a specific service. Additionally, the physical channels carry payload data or details of specific data transmission characteristics like modulation, reference signal multiplexing, transmit power, RF resources, etc.

1.3.1 5G NR logical channels

There are several different logical channels that are used within the 5G NR radio access network. Some of them will be familiar names from the 4G LTE system as the names have been carried over.

- **Broadcast Control Channel (BCCH)**: The BCCH is used within the downlink, and it is used for sending broadcast style information to the UE within that cell. The system information transmitted by the 5G NR BCCH is divided into different blocks:
  - Master Information Block (MIB): There is one MIB and this is mapped onto the BCH transport channel and then to the PBCH physical channel.
  - System Information Block (SIB): There are several system information blocks, SIBs. These are mapped onto the DL-SCH transport channel and then onto the PDSCH physical channel.
- **Paging Control Channel (PCCH)**: This is a Downlink channel. It is used to page the UEs whose location at cell level is not known to the network. As a result the paging message needs to be transmitted in multiple cells. The PCCH is mapped to the PCH transport channel and then to the PDSCH physical channel.
- **Common Control Channel (CCCH)**: This 5G channel is used on both the downlink and uplink for transmitting control information to and from the user equipments or mobiles. The channel is used for initial access, i.e. those mobiles that do not have a radio resource control, RRC connection.
- **Dedicated Control Channel (DCCH)**: The DCCH is used within the uplink and downlink to carry dedicated control information between the UE or mobile and the network. It is used by the UE and the network after a radio resource control, RRC connection has been established.
- **Dedicated Traffic Channel (DTCH)**: This 5G channel is present in both the uplink and downlink. It is dedicated to one UE and is used for carrying user information to and from a specific UE and the network.

1.3.2 5G NR transport channels

There are *five different transport channels*. Some are used on the uplink, others on the downlink, and some can be used on both.

- **Broadcast Channel (BCH)**: The BCH 5G channel is used in the downlink only for transmitting the BCCH system information and specifically the Master Information Block, MIB, information. In order that the data can be utilised, it has a specific format.

- **Paging Channel (PCH)**: The PCH is used for carrying paging information from the PCCH logical channel. The PCH supports discontinuous reception, DRX, to enable the UE to save battery power by waking up at a specific time to receive the PCH.
  In order that the PCH is received by all mobiles / UEs in the cell, the PCH must be broadcast over the entire cell as a single message, or where beam forming is used, this can be done using several different PCH instances.

- **Downlink Shared Channel (DL-SCH)**: As the name indicates, this is a downlink only channel. It is the main transport channel used for transmitting downlink data and it supports all the key 5G NR features. These include: dynamic rate adaptation; HARQ, channel aware scheduling, and spatial multiplexing.
  The DL-SCH is also used for transmitting some parts of the BCCH system information, specifically the SIB. Each UE has a DL-SCH for each cell it is connected to.

- **Uplink Shared Channel (UL-SCH)**: This is the uplink counterpart to the DLSCH that is, the uplink transport channel used for transmission of uplink data.

- **Random-Access Channel (RACH)**: The RACH is a transport channel, which carries the random access preamble which is used to overcome the message collisions that can occur when UEs access the system simultaneously.

1.3.3 5G NR physical channels

The 5G physical channels are used to transport information over the actual radio interface. They have the transport channels mapped into them, but they also include various physical layer data required for the maintenance and optimization of the radio communications link between the UE and the base station. The 5G mobile communications physical layer channels resemble those of 4G LTE, but PHICH and PCIICH have been removed. The HARQ operation has also been updated to be more flexible. Also the downlink control channel PDCCH is now administered by layer 3 procedures. There are *three physical channels* for each of the uplink and downlink

5G NR Downlink Physical Channels

Physical downlink shared channel (PDSCH): 5G NR physical downlink shared channel, PDSCH carries data sharing the capacity on a time and frequency basis. The PDSCH physical channel carries a variety of items of data: user data; UE-specific higher layer control messages mapped down from higher channels; system information blocks (SIBs) & paging.
The PDSCH uses an adaptive modulation format dependent upon the link conditions, i.e. signal to noise ratio. It also uses a flexible coding scheme. The combination of these means that there is a flexible coding and data rate.

Physical downlink control channel (PDCCH): As the name implies, 5G physical downlink control channel carries downlink control data. Its primary function is scheduling the downlink transmissions on the PDSCH and also the uplink data transmissions on the PUSCH.
The PDCCH uses QPSK as its modulation format and polar coding as the coding scheme, except for small packets of data.

Physical broadcast channel (PBCH): This 5G channel forms part of the synchronization signal block. Its function is to provide UEs with the Master Information Block, MIB. A further function of the PBCH in conjunction with the control channel is to support the synchronization of time and frequency. This aids with cell acquisition, selection and re-selection.
The PBCH uses a fixed data format and there is one block that extends over a TTI of 80ms, uses QPSK modulation and it transmits a cell specific demodulation reference signal, DMRS pattern that can be used aid with beam-forming.

5G NR Uplink Physical Channels

Physical random access channel (PRACH): This 5G channel is used for channel access. It transmits an initial random access pre-amble consisting of sequences which may be of two different lengths:

- A long sequence is 839 which is applied to the subcarrier spacings of 1.25kHz and 5 kHz
- Short sequence lengths of 139 are applied to subcarrier spacings of 15 kHz and 30 kHz (FR1 bands) and 60 kHz and 120 kHz (FR2 bands).

Physical uplink shared channel (PUSCH): The counterpart of the PDSCH. It is used to carry data from the UL-SCH and its higher mapped channels on a frequency and time-shared basis.
Like the PDSCH, The PUSCH also has a very flexible format. The allocation of frequency resources is undertaken using resource blocks along with a flexible modulation and coding scheme dependent upon the link signal to noise ratio.
To support the channel link estimation and demodulation, the PUSCH contains DMRS signals.

Physical uplink control channel (PUCCH): This carries the uplink control data. It is also possible that dependent upon the resource allocation the uplink control information or data may also be sent on the PUSCH, even though in the downlink direction, control information is always sent on the PDCCH.

The use of these 5G channels provide a method for organizing the flow of data over the radio interface of the 5G communications network. Using channels enables the communications system to recognize the type of data that is being sent, and to deal with it accordingly. The format used is very similar to that employed on 4G LTE and it built on the technology of previous mobile communications or mobile phone generations.

Figures 1.4 & 1.5 shows the mapping between logical, transport and physical channels in both DL and UL

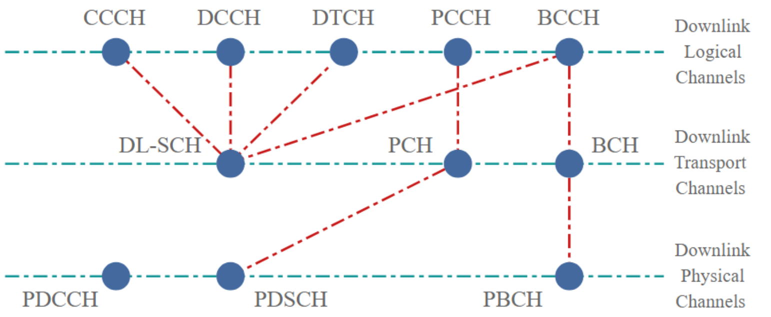


Figure 1.4: 5G Downlink logical, transport and physical channel mapping

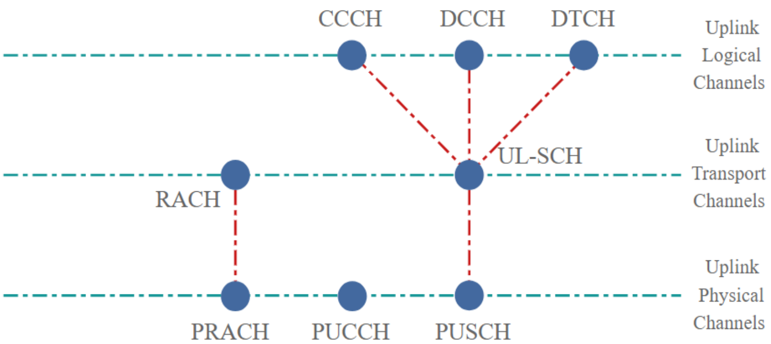


Figure 1.5: 5G Uplink logical, transport and physical channel mapping

Part I

5G PDSCH Simulator

PDSCH overview & Motivation

The Physical Downlink Shared Channel (PDSCH) is a crucial component of 5G wireless communication systems that enables high-speed data transmission with low latency. As a team, we were intrigued by the potential of 5G technology to revolutionize the way we communicate and connect with each other. We were particularly interested in the PDSCH channel, as it plays a critical role in achieving the high data rates and low latency that are essential for applications such as virtual reality, autonomous driving, and the Internet of Things (IoT).
We were also motivated by the technical challenges involved in simulating the PDSCH channel accurately. 5G is a complex and evolving technology that requires a deep understanding of wireless communication systems, signal processing, and coding theory. Our project aimed to develop a MATLAB based simulation model that could capture the unique characteristics of PDSCH and provide insights into its performance under different conditions.

In this Part, we will delve into the technical details of PDSCH and explain how our simulation model was designed and implemented. We will also discuss the results of our simulation experiments and highlight the insights we gained from them. Our hope is that this work will contribute to the ongoing efforts to improve the performance and reliability of 5G networks.
Designing and implementing a simulation model for PDSCH requires a detailed understanding of the 5G standard specifications, as well as knowledge of wireless communication systems, signal processing, and coding theory. Here is an overview of the steps we took to design and implement our simulation model:
Our simulation model for PDSCH in 5G wireless communication systems consists of several key blocks that are critical to achieving high-speed data transmission with low latency. These blocks include channel coding, MIMO setups, reference signals, and channel estimation.

The channel coding block is responsible for protecting the data from errors that may be introduced during wireless transmission. We simulated both linear block codes and LDPC coding to evaluate their performance under different conditions.
The MIMO setups block allows for multiple antennas to be used at both the transmitter and receiver, which can improve the quality and reliability of wireless communication. We simulated several MIMO setups, including SISO, SIMO, and MIMO, to gain insights into the performance of PDSCH under different conditions.
The reference signals block provides important information about the wireless channel, such as channel state information and data demodulation and decoding. We focused on two types of reference signals: CSI-RS and DMRS, which are used for channel estimation, synchronization, and data demodulation.

Finally, the channel estimation block is responsible for accurately estimating the wireless channel conditions and compensating for any distortion or interference that may be present. We spent a significant amount of time on channel estimation, which is essential for accurately simulating the performance of PDSCH under realistic conditions.
In the following sections, we will discuss each of these blocks in detail, including the technical details of how they were implemented and the results of our simulation experiments.

Chapter 2

Channel Coding

2.1 Overview

This part deals with linear block codes covering their fundamental concepts, generator and parity check matrices, error-correcting capabilities, encoding and decoding, and performance analysis. The linear block code discussed in this part is Hamming code.

2.1.1 Shannon's Noisy Channel Coding Theorem

Any channel affected by noise possesses a specific "channel capacity" C a rate of conveying information that can never be exceeded without error, but in principle, an error-correcting code always exists such that information can be transmitted at rates less than C with an arbitrarily low BER.

2.1.2 Channel Coding Principle

The channel coding principle is to add redundancy to minimize error rate as illustrated in Figure 2.1.

Figure 2.1: Illustration of the channel coding principle.

2.1.3 Channel Coding Gain

The bit error rate (BER) is the probability that a binary digit transmitted from the source received erroneously by the user. For required BER, the difference between the powers required for without and with coding is called the coding gain. A typical plot of BER versus Eb/No (bit energy to noise spectral density ratio) with and without channel coding is shown in Figure 2.2. It can be seen that coding can arrive at the same value of the BER at lower Eb/No than without coding. Thus, the channel coding yields coding gain which is usually measured in dB. Also, the coding gain usually increases with a decrease in BER.

Figure 2.2: Coding gain.

2.2 Block Codes

The data stream is broken into blocks of k bits and each k-bit block is encoded into a block of n bits with $n > k$ bits as illustrated in Figure 2.3. The n-bit block of the channel block encoder is called the code word. The code word is formed by adding $(n - k)$ parity check bits derived from the k message bits.

Figure 2.3: Coded data stream.

Properties of Block Codes

Block code rate

The block code rate (R) is defined as the ratio of k message bits and length of the code word n.

$$R = \frac{k}{n}$$

Code word weight

The weight of a code word or error pattern is the number of nonzero bits in the code word or error pattern.
For example, the weight of a code word $c = (1, 0, 0, 1, 1, 0, 1, 0)$ is 4.

Hamming distance

The Hamming distance between two blocks v and w is the number of coordinates in which the two blocks differ.

$$d_{hamming}(v, w) = d(v, w) = |\{i|v_i \neq w_i, i = 0, 1, \ldots, n-1\}|$$

Example: Consider the code words $v = (00100)$ and $w = (10010)$, then the Hamming distance $d_{hamming}(v, w) = 3$.
Hamming distance allows for a useful characterization of the error detection and error correction capabilities of a block code as a function of the code's minimum distance.

The Minimum Distance of a Block Code

The minimum distance of a block code C is the minimum Hamming distance between all distinct pairs of code words in C.

A code with minimum distance d_{min} can:

- *detect* all error patterns of weight less than or equal to $(d_{min} - 1)$.
- *correct* all error patterns of weight less than or equal to $(d_{min} - 1)/2$.

Example: Consider the binary code C composed of the following four code words.

$$C = \{(00100), (10010), (01001), (11111)\}$$

Hamming distance of (00100) and (10010) = 3
Hamming distance of (10010) and (01001) = 4
Hamming distance of (00100) and (01001) = 3
Hamming distance of (10010) and (11111) = 3
Hamming distance of (00100) and (11111) = 4
Hamming distance of (01001) and (11111) = 3
Therefore, the minimum distance $d_{min} = 3$.

2.2.1 Linear Block Codes

A block code C consisting of n-tuples $\{(c_0, c_1, ..., c_{n-1})\}$ of symbols from $GF(2)$ is said to be binary linear block code if and only if C forms a vector subspace over $GF(2)$.

Note: finite fields are also called Galois fields (GF).

The code word is said to be systematic linear code word if each of the 2^k code words is represented as linear combination of k linearly independent code words.

Linear Block Codes Properties

There are *two important properties* of linear block codes which are:

Property 1: The linear combination of any set of code words is a code word.

Property 2: The minimum distance of a linear block code is equal to the minimum weight of any nonzero word in the code.

Also, there are 2 well-known bounds on the minimum distance which are

Singleton Bound: The minimum distance of an (n, k) linear code is bounded by

$$d_{min} \leq n - k + 1 \tag{2.1}$$

Hamming Bound: An (n, k) block code can *detect* t_{ed} errors per code word and can correct up to t_{ec} errors per code word, provided that n and k satisfy the Hamming bound.

$$2^{n-k} \geq \sum_{i=0}^{t_{ec}} \binom{n}{i} \quad \text{where} \quad \binom{n}{i} = \frac{n!}{(n-1)!i!} \tag{2.2}$$

$$t_{ec} = \frac{(d_{min} - 1)}{2} \quad , \quad t_{ed} = d_{min} - 1$$

The relation is the upper bound on d_{min} and is known as *the Hamming bound.*

2.2.2 Generator & Parity Check Matrices

Let $\{g_0, g_1, \ldots, g_{k-1}\}$ be a basis of code words for the (n, k) linear block code C and $m = [m_0, m_1, \ldots, m_{k-1}]$ the message to be encoded. The Theorem that says the code word $c = (c_0, c_1, \ldots, c_{n-1})$ for the message is uniquely represented by the following linear combination of $g_0, g_1, \cdots, g_{k-1}$

$$c = m_0 g_0 + \ldots + m_{k-1} g_{k-1}$$

for every code word $c \in C$.
Since every linear combination of the basis elements must also be a code word, there is a one-to-one mapping between the set of k-bit blocks $(a_0, a_1, \ldots, a_{k-1})$ over $GF(2)$ and the code words

in C. A matrix G is constructed by taking the vectors in the basis as its rows.

$$G = \begin{bmatrix} g_0 \\ g_1 \\ \vdots \\ g_{k-1} \end{bmatrix} = \begin{bmatrix} g_{0,0} & g_{0,1} & \cdots & g_{0,n-1} \\ g_{1,0} & g_{1,1} & \cdots & g_{1,n-1} \\ \vdots & \vdots & \ddots & \vdots \\ g_{k-1,0} & g_{k-1,1} & \cdots & g_{k-1,n-1} \end{bmatrix}$$

This matrix G is a generator matrix for the code C. It can be used to directly encode k-bit blocks in the following manner:

$$mG = [m_0, m_1, \ldots, m_{k-1}] \cdot \begin{bmatrix} g_0 \\ g_1 \\ \vdots \\ g_{k-1} \end{bmatrix} = m_0 g_0 + m_1 g_1 + \ldots + m_{k-1} g_{k-1} = c$$

The dual space of a linear block code C is the dual code of C and a basis $\{h_0, h_1, \ldots, h_{n-k-1}\}$ can be found for dual code of C, and the following parity check matrix can be constructed:

$$H = \begin{bmatrix} h_0 \\ h_1 \\ \vdots \\ h_{n-k-1} \end{bmatrix} = \begin{bmatrix} h_{0,0} & h_{0,1} & \cdots & h_{0,n-1} \\ h_{1,0} & h_{1,1} & \cdots & h_{1,n-1} \\ \vdots & \vdots & \ddots & \vdots \\ h_{n-k-1,0} & g_{n-k-1,1} & \cdots & g_{n-k-1,n-1} \end{bmatrix}$$

In a systematic linear block code, the last k bits of the codeword are *the message bits*, that is

$$c_i = m_{i-(n-k)} \quad , \quad i = n-k, \ldots, n-1$$

While the first $n-k$ bits in the codeword are check bits generated from the k message bits according to

$$\begin{aligned} c_0 &= p_{0,0} m_0 + p_{1,0} m_1 + \ldots + p_{k-1,0} m_{k-1} \\ c_1 &= p_{0,1} m_0 + p_{1,1} m_1 + \ldots + p_{k-1,1} m_{k-1} \\ &\vdots \\ c_{n-k-1} &= p_{0,n-k-1} m_0 + p_{1,n-k-1} m_1 + \ldots + p_{k-1,n-k-1} m_{k-1} \end{aligned}$$

The above equation can be written in a matrix form as:

$$[c_0, c_1, \ldots, c_{n-1}] = [m_0, m_1, \ldots, m_{k-1}] \begin{bmatrix} p_{0,0} & p_{0,1} & \cdots & p_{0,n-k-1} & 1000 & \cdots & 0 \\ p_{1,0} & p_{1,1} & \cdots & p_{1,n-k-1} & 0100 & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \vdots \\ p_{k-1,0} & p_{k-1,1} & \cdots & p_{k-1,n-k-1} & 0000 & \cdots & 1 \end{bmatrix}_{k \times n} \tag{2.3}$$

or

$$c = mG$$

where G is the matrix on the right hand side of the equation 2.3.
The $k \times n$ matrix G is called a generator matrix of the code and it has the following form:

$$G = [P|I_k]_{k \times n} \tag{2.4}$$

The matrix I_k is the identity matrix of order k, and P is an arbitrary $k \times n-k$ matrix. When P is specified, it defines the (n,k) block code completely. The parity check matrix H corresponding to the above generator matrix G can be obtained as

$$H = \left[I_{n-k}|P^T\right]$$

$$H = \begin{bmatrix} 1000 & \cdots & 0 & p_{0,0} & p_{0,1} & \cdots & p_{0,n-k-1} \\ 0100 & \cdots & 0 & p_{1,0} & p_{1,1} & \cdots & p_{1,n-k-1} \\ \vdots & \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\ 0000 & \cdots & 1 & p_{0,n-k-1} & p_{1,n-k-1} & \cdots & p_{k-1,n-k-1} \end{bmatrix} \tag{2.5}$$

Theorem 2.2.1: Parity Check Theorem

For any (n,k) linear block code C with $(n-k)\times n$ parity check matrix H, a code word $c \in C$ is a valid code word ***if and only if*** $cH^T = 0$.

....

Example:

For the following generator matrix of (7,4) block code. Find the code vector for the message vector m = (1110) and check the validity of code vector generated.

$$G = \left[\begin{array}{ccc|cccc} 1 & 1 & 0 & 1 & 0 & 0 & 0 \\ 0 & 1 & 1 & 0 & 1 & 0 & 0 \\ 1 & 1 & 1 & 0 & 0 & 1 & 0 \\ 1 & 0 & 1 & 0 & 0 & 0 & 1 \end{array}\right]$$

Solution:

The code vector for the message block $m = (1110)$ is given by

$$c = mG = [1110] \cdot \left[\begin{array}{ccc|cccc} 1 & 1 & 0 & 1 & 0 & 0 & 0 \\ 0 & 1 & 1 & 0 & 1 & 0 & 0 \\ 1 & 1 & 1 & 0 & 0 & 1 & 0 \\ 1 & 0 & 1 & 0 & 0 & 0 & 1 \end{array}\right] = [0101110]$$

$$H = \left[\begin{array}{ccc|cccc} 1 & 0 & 0 & 1 & 0 & 1 & 1 \\ 0 & 1 & 0 & 1 & 1 & 1 & 0 \\ 0 & 0 & 1 & 0 & 1 & 1 & 1 \end{array}\right]$$

$$cH^T = [0101110] \cdot \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \\ 1 & 1 & 0 \\ 0 & 1 & 1 \\ 1 & 1 & 1 \\ 1 & 0 & 1 \end{bmatrix} = [000]$$

Hence, the generated code vector is valid.

2.3 Hamming Codes

Hamming code is a linear block code capable of correcting single errors having a minimum distance $d_{min} = 3$. It is very easy to construct Hamming codes. The parity check matrix H

(obtained in equation 2.5) must be chosen so that no row in H^T is zero and the first $(n-k)$ rows of H^T form an identity matrix and all the rows are distinct. We can select $2^{n-k}-1$ distinct rows of H^T.
Since the matrix H^T has n rows, for all of them to be distinct, the following inequality should be satisfied.

$$2^{n-k}-1 \geq n$$

Implying that

$$\begin{aligned}(n-k) &\geq \log_2(n+1)\\ n &\geq k+\log_2(n+1) \qquad (2.6)\end{aligned}$$

Hence, the minimum size n for the code words can be determined from equation 2.6.

Example: Design a Hamming code with message block size of eleven bits.
Solution: It follows from equation 2.6 that

$$n \geq 11+\log_2(n+1)$$

The smallest n that satisfies the above inequality is 15, and hence, we need a $(15,11)$ block code. Thus, the transpose of the parity check matrix H will be 4×15 matrix. The first four rows of H^T will be I_4 matrix. The last eleven rows are arbitrarily chosen, with the restrictions that no row is zero, and all the rows are distinct.

$$H^T = \left[\begin{array}{cccc} 1 & 0 & 0 & 0\\ 0 & 1 & 0 & 0\\ 0 & 0 & 1 & 0\\ 0 & 0 & 0 & 1\\ \hline 0 & 1 & 0 & 1\\ 0 & 1 & 1 & 0\\ 0 & 1 & 1 & 1\\ 0 & 0 & 1 & 1\\ 1 & 0 & 0 & 1\\ 1 & 0 & 1 & 0\\ 1 & 0 & 1 & 1\\ 1 & 1 & 0 & 0\\ 1 & 1 & 0 & 1\\ 1 & 1 & 1 & 0\\ 1 & 1 & 1 & 1 \end{array}\right] = \left[\frac{I_{n-k}}{P^T}\right] \qquad (2.7)$$

Then generator matrix G equals

$$G = \begin{bmatrix} 0&1&0&1&1&0&0&0&0&0&0&0&0&0&0\\ 0&1&1&0&0&1&0&0&0&0&0&0&0&0&0\\ 0&1&1&1&0&0&1&0&0&0&0&0&0&0&0\\ 0&0&1&1&0&0&0&1&0&0&0&0&0&0&0\\ 1&0&0&1&0&0&0&0&1&0&0&0&0&0&0\\ 1&0&1&0&0&0&0&0&0&1&0&0&0&0&0\\ 1&0&1&1&0&0&0&0&0&0&1&0&0&0&0\\ 1&1&0&0&0&0&0&0&0&0&0&1&0&0&0\\ 1&1&0&1&0&0&0&0&0&0&0&0&1&0&0\\ 1&1&1&0&0&0&0&0&0&0&0&0&0&1&0\\ 1&1&1&1&0&0&0&0&0&0&0&0&0&0&1 \end{bmatrix}$$

Example: Construct parity check and generator matrices for a $(7,4)$ Hamming code.
Solution: The parity check matrix H and generator matrix G for a $(7,4)$ Hamming code are

$$H = \begin{bmatrix} 1 & 0 & 0 & 1 & 0 & 1 & 1 \\ 0 & 1 & 0 & 1 & 1 & 1 & 0 \\ 0 & 0 & 1 & 0 & 1 & 1 & 1 \end{bmatrix}$$

$$G = \begin{bmatrix} 1 & 1 & 0 & 1 & 0 & 0 & 0 \\ 0 & 1 & 1 & 0 & 1 & 0 & 0 \\ 1 & 1 & 1 & 0 & 0 & 1 & 0 \\ 1 & 0 & 1 & 0 & 0 & 0 & 1 \end{bmatrix}$$

2.3.1 Syndrome Table Decoding

Consider a valid code word c for transmission and let e be an error pattern introduced by the channel during transmission. Then, the received vector r can be written as

$$r = c + e$$

Multiplying the r by the transpose of the parity check matrix gives the syndrome S which can be expressed as

$$\begin{aligned} S &= r \cdot H^T \\ &= (c+e) \cdot H^T \\ &= cH^T + eH^T \\ &= 0 + eH^T \\ &= eH^T \end{aligned} \tag{2.8}$$

Thus, the syndrome vector is independent of the transmitted code word c and is only a function of the error pattern e. Decoding is performed by computing the syndrome of a received vector, looking up the corresponding error pattern, and subtracting the error pattern from the received word.

Example: Construct a syndrome decoding table for a $(7,4)$ Hamming code
Solution: For a $(7,4)$ Hamming code, there are $2^{(7-4)}$ error patterns e as in Table 2.8

Table 2.1: Syndrome decoding table for a $(7,4)$ Hamming code

Error Pattern e	Syndrome
0000000	000
1000000	100
0100000	010
0010000	001
0001000	110
0000100	011
0000010	111
0000001	101

The syndrome for $(7,4)$ Hamming code is computed using the parity check matrix H (as given in the solution 2.7) as follows

$$S = e \cdot H^T$$

Thus, the syndrome decoding table for a $(7,4)$ Hamming code is as in Table 2.1

2.3.2 Hamming Codes Decoding

Syndrome table is used to decode the Hamming codes. The syndrome table gives the syndrome value based on the simple relationship with parity check matrix. The single-error-correcting codes (i.e., Hamming codes), are decoded by using syndrome value. Consider a code word c corrupted by e, an error pattern with a single one in the j^{th} coordinate position results a received vector r. Let $h_0, h_1, \ldots, h_{n-1}$ be the set of columns of the parity check matrix H. When the syndrome is computed, we obtain the transposition of the j^{th} column of H.

$$s = eH^T = [0, \ldots, 0, 1, 0, \ldots, 0] \cdot \begin{bmatrix} h_0^T \\ h_1^T \\ \vdots \\ h_{n-1}^T \end{bmatrix} = h_j^T \tag{2.9}$$

The above-mentioned process in equation 2.9 can be implemented using the following algorithm:

1. Compute the syndrome s for the received word. If $s = 0$, the received code word is the correct code word.
2. Find the position j of the column of H that is the transposition of the syndrome.
3. Complement the j^{th} bit in the received codeword to obtain the corrected code word.

Example: Decode the received vector $r = [001100011100000]$ using the $(15,11)$ parity check matrix.
Solution:

$$H = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\ 0 & 1 & 0 & 0 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 \\ 0 & 0 & 1 & 0 & 0 & 1 & 1 & 1 & 0 & 1 & 1 & 0 & 0 & 1 & 1 \\ 0 & 0 & 0 & 1 & 1 & 0 & 1 & 1 & 1 & 0 & 1 & 0 & 1 & 0 & 1 \end{bmatrix}$$

The received vector is $r = [001100011100000]$
The corresponding syndrome $s = r \cdot H^T$ is

$$s = [0011]$$

The syndrome is the transposition of 7^{th} column of H. Inverting the 7^{th} coordinate of r, the following code word is obtained

$$c = [001100001100000]$$

2.4 LDPC Coding

Chapter 3

Input/Output Setup

3.1 Overview

In this chapter, we will discuss various input-output setups (depending on *how many antennas are used in both the transmitter and the receiver*) that were implemented in the simulator. The availability of multiple antennas at the transmitter and/or the receiver can be utilized in different ways to achieve different aims:

- Multiple antennas at the transmitter and/or the receiver can be used to provide additional diversity against fading on the radio channel. In this case, the channels experienced by the different antennas should have low mutual correlation, implying the need for a sufficiently large inter-antenna distance (spatial diversity), alternatively the use of different antenna polarization directions (polarization diversity).
- Multiple antennas at the transmitter and/or the receiver can be used to "shape" the overall antenna beam (transmit beam and receive beam, respectively) in a certain way, for example, to maximize the overall antenna gain in the direction of the target receiver/-transmitter or to suppress specific dominant interfering signals.
- The simultaneous availability of multiple antennas at the transmitter and the receiver can be used to create what can be seen as multiple parallel communication "channels" over the radio interface. This provides the possibility for very high bandwidth utilization without a corresponding reduction in power efficiency or, in other words, the possibility for very high data rates within a limited bandwidth without an un-proportionally large degradation in terms of coverage. This feature is highly present in the MIMO setup.

The following table explains the different setups implemented in the simulator.

Table 3.1: Transmitter & Receiver for different setups

Setup	Transmitter	Receiver
SISO (Single-Input Single-Output)	One antenna	One antenna
SIMO (Single-Input Multi-Output)	One antenna	Many antennas
MISO (Multi-Input Single-Output)	Many antennas	One antenna
MIMO (Multi-Input Multi-Output)	Many antennas	Many antennas

Later in the chapter we will discuss how each setup works, its encoding and decoding techniques and benifits. We will also zoom in as if we were to send only one symbol. This will help us explain the techniques used in each setup better.

Before we discuss each setup in great detail, some information about the channel model must be cleared out:

- We are using an OFDM based system.
- Each sub-carrier carries only one modulated symbol.
- We let the channel model be a block fading channel where the channel stays the same for some time (Coherence time) and for some subcarriers (Coherence bandwidth).

The following assumptions were also made:

- We will ignore the modulation of the input signal for better understanding of how each setup works.
- We will also assume perfect knowledge of the channel at both the transmitter and the receiver.

3.1.1 Narrow-band wireless fading channel

One of the most common wireless channel model and the one that we will be using is the following

$$y[m] = h[m]x[m] + n[m]$$

Where the index m is a time index, $x[m]$ is the transmitted complex symbol at time m, $y[m]$ is the corresponding receives signal, and $n[m]$ is the AWGN at time m.
The different component compared to the AWGN channel is $h[m]$. This is referred to as the "fading" coefficient. It is a random value which captures the changes that happen in the transmitted electromagnetic waves due to the environment.
$h[m]$ is modelled as a complex number, where

$$h[m] = \mathrm{a} + j\mathrm{b} = |h[m]|e^{j\theta[m]} \tag{3.1}$$

where the magnitude of the channel $|h[m]|$ is a Rayleigh random variable

$$|h[m]| \; f_{\sigma^2}(x) = \frac{x}{\sigma^2} e^{\frac{-x^2}{2\sigma^2}}$$

The phase $\theta[m]$ is a uniform random variable. The figure shows a Rayleigh distribution for different values of σ. This shows the impact of fading on communications. Specifically, the value of $|h[m]|$ can be small with a considerable probability, and that effectively reduces the received power of the transmitted signal. When $|h[m]|$ is too small, the channel is said to be in deep fading.

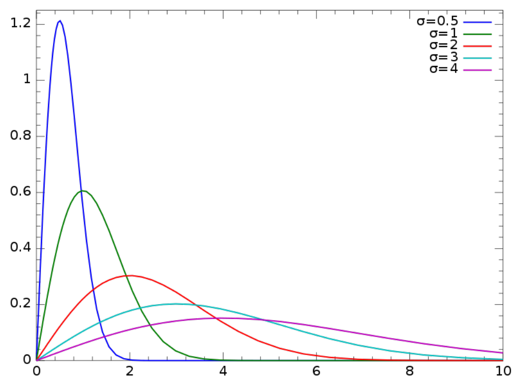


Figure 3.1: Rayleigh random variable

3.2 SISO

SISO stands for **"Single Input, Single Output"**. In communication engineering, SISO is the simplest way to describe a communication link between a transmitter and a receiver. It is used to describe the case where both the transmitter and receiver have single antennas. Pre-coding is not typically used in SISO communication systems.
For a SISO channel (we drop the time index for now)

$$y = hx + n$$

The noise power is N_o. The minimum distance between Constellation points without fading is $2a$, where a is the smallest transmitted amplitude per constellation points. Now, with fading coefficient h, minimum distance becomes $d_{min} = 2|h|a$. There, for this CSI value, we can upper bound the probability of error as

$$P_e\left(\bar{h}\right) \leq \ k\ Q\left(\frac{d_{min}}{\sqrt{2N_o}}\right) = \ k\ Q\left(\sqrt{\frac{4\left|h\right|^2 a^2}{N_o}}\right) = k\ Q\left(\sqrt{2\left|h\right|^2 \mathrm{SNR}}\right) \tag{3.2}$$

We assume that h is Rayleigh (as described in equation 3.1) with $\sigma^2 = 1$. Then, we want to compute the "average" P_e over CSI realization. This turns out to be

$$P_e \ = \mathbb{E}_{|h|^2}\{P_e\left(h\right)\} \leq k\left(\frac{1}{2} - \frac{1}{2}\sqrt{\frac{\mathrm{SNR}}{1+\mathrm{SNR}}}\right) = \frac{k}{2}\left(1-\mu\right) \tag{3.3}$$

How does this compare with AWGN?
Recall: P_e for AWGN is bounded by $kQ(\sqrt{2\mathrm{SNR}})$.
The following figure compares Pe for both AWGN and fading far BPSK.

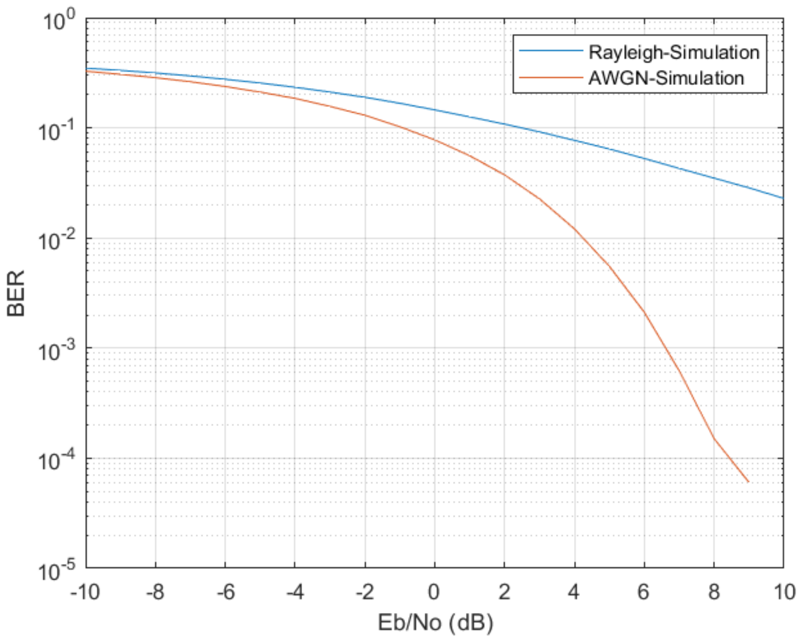


Figure 3.2: AWGN vs Rayleigh fading BER for SISO

At high SNR:
Probability of error in AWGN has a water fall behavior which indicates very good performance. However, it decays linearly in case of fading and that is too slow. This is because $|h|$ is many times too small and therefore the received power $|h|^2a^2$ is not enough to make decoding performance good. we say that a channel in *"deep fade"* if received signal power is less than or equal to the noise power.

$$|h|^2 a^2 \leq N_o$$
$$|h|^2 \leq \frac{N_o}{a^2} = \frac{1}{\text{SNR}}$$

What is the probability of this happening?
This probability is found to can be approximated to

$$10 \log \left(\mathbb{P}\{|h|^2 \leq \frac{1}{\text{SNR}}\} \right) = \text{const} - 10 \log (\text{SNR}) = \text{const} - \text{SNR}_{dB} \tag{3.4}$$

This looks like the linear decrease behavior we saw above in Figure 3.2.

3.3 SIMO

3.4 MISO

3.5 MIMO

3.5.1 Diversity

3.5.2 Multiplexing

Chapter 4

Reference Signals

4.1 Overview

Reference signals are predefined signals occupying specific resource elements within the down-link time-frequency grid. The NR specification includes several types of reference signals transmitted in different ways and intended to be used for different purposes by a receiving device. Unlike LTE, which relies heavily on always-on, cell-specific reference signals in the down-link for coherent demodulation, channel quality estimation for CSI reporting, and general time-frequency tracking, NR uses different down-link reference signals for different purposes. This allows for optimizing each of the reference signals for their specific purpose. It is also in line with the overall principle of ultra-lean transmission as the different reference signals can be transmitted only when needed. Later release of LTE took some steps in this direction, but NR can exploit this to a much larger degree as there are no legacy NR devices to cater for.
The NR reference signals include:

- Demodulation reference signals (DMRS) for PDSCH are intended for channel estimation at the device as part of coherent demodulation. They are present only in the resource blocks used for PDSCH transmission. Similarly, the DMRS for PUSCH allows the gNB to coherently demodulate the PUSCH.

- Phase-tracking reference signals (PTRS) can be seen as an extension to DMRS for PDSCH/PUSCH and are intended for phase-noise compensation. The PT-RS is denser in time but sparser in frequency than the DM-RS, and, if configured, occurs only in combination with DMRS.

- CSI reference signals (CSI-RS) are downlink reference signals intended to be used by devices to acquire down-link channel-state information (CSI). Specific instances of CSI reference signals can be configured for time/frequency tracking and mobility measurements.

- Tracking reference signals (TRS) are sparse reference signals intended to assist the device in time and frequency tracking

- Sounding reference signals (SRS) are uplink reference signals transmitted by the devices and used for uplink channel-state estimation at the base stations.

The previous section assumed full knowledge of the channel characteristics at both transmitter and receiver sides, this section discusses how such knowledge is gained by focusing on two important reference signals : CSI-RS and DMRS.
The propagation channel depends on the transmit frequency. Therefore, if up-link and down-link operate on two different frequencies as is the case in FDD, there is no choice but relying on the receiver to communicate information about the channel back to the transmitter. This is the case at the bottom right.
In the case of TDD, where the up-link and down-link share the same transmit frequency, it is possible, on the other hand, to estimate the down-link channel based on measurements on up-link transmission (or the opposite). In this section we focus on the FDD transmission case.

4.2 Demodulation Reference signal (DMRS)

The DM-RS in NR provides quite some flexibility to cater for different deployment scenarios and use cases: a front-loaded design to enable low latency, support for up to 12 orthogonal antenna ports for MIMO, transmissions duration from 2 to 14 symbols, and up to four reference-signal instances per slot to support very high-speed scenarios. To achieve low latency, it is beneficial to locate the demodulation reference signals early in the transmission, sometimes known as front-loaded reference signals. This allows the receiver to obtain a channel estimate early and, once the channel estimate is obtained, process the received symbols on the fly without having to buffer a complete slot prior to data processing. This is essentially the same motivation as for the frequency-first mapping of data to the resource elements. Two main time-domain structures are supported, differing in the location of the first DM-RS symbol:

- Mapping type A, where the first DMRS is located in symbol 2 or 3 of the slot and the DMRS is mapped relative to the start of the slot boundary, regardless of where in the slot the actual data transmission starts. This mapping type is primarily intended for the case where the data occupy (most of) a slot. The reason for symbol 2 or 3 in the down-link is to locate the first DMRS occasion after a CORESET located at the beginning of a slot.

- Mapping type B, where the first DMRS is located in the first symbol of the data allocation, that is, the DMRS location is not given relative to the slot boundary but rather relative to where the data are located. This mapping is originally motivated by transmissions over a small fraction of the slot to support very low latency and other transmissions that benefit from not waiting until a slot boundary starts but can be used regardless of the transmission duration.

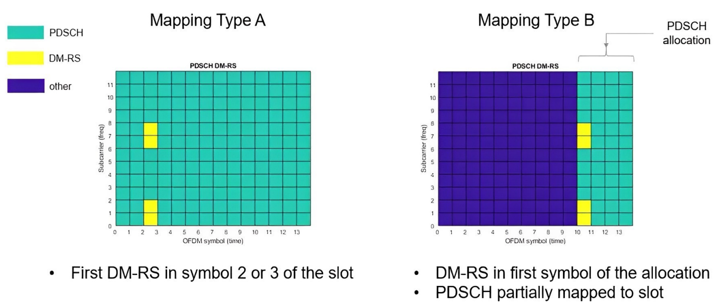


Figure 4.1: DMRS Mapping type A & type B

Although front-loaded reference signals are beneficial from a latency perspective, they may not be sufficiently dense in the time domain in the case of rapid channel variations. To support high-speed scenarios, it is possible to configure up to three additional DM-RS occasions in a slot. The channel estimator in the receiver can use these additional occasions for more accurate channel estimation, for example, to use interpolation between the occasions within a slot.

4.2.1 Pilot based DMRS channel estimation

Assume a single antenna port at both Tx and Rx.

$$y = h_{11}x$$

In order to know $h_1 1$, we need to send a symbol x which is known at the receiver, then $h_1 1$ can be calculated as:

$$h_{11} = \frac{y}{x}$$

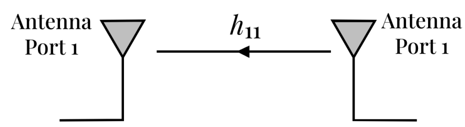


Figure 4.2: Single Tx Single Rx

This known symbol x is called a pilot symbol. The problem with this approach is that there exists a channel coefficient need to be known for each RE in the PDSCH which requires sending pilots in all REs, hence there will not be any place for data bits. The solution to this problem is to send pilot symbols in some REs as shown in the figure to determine the channel coefficients in those REs.

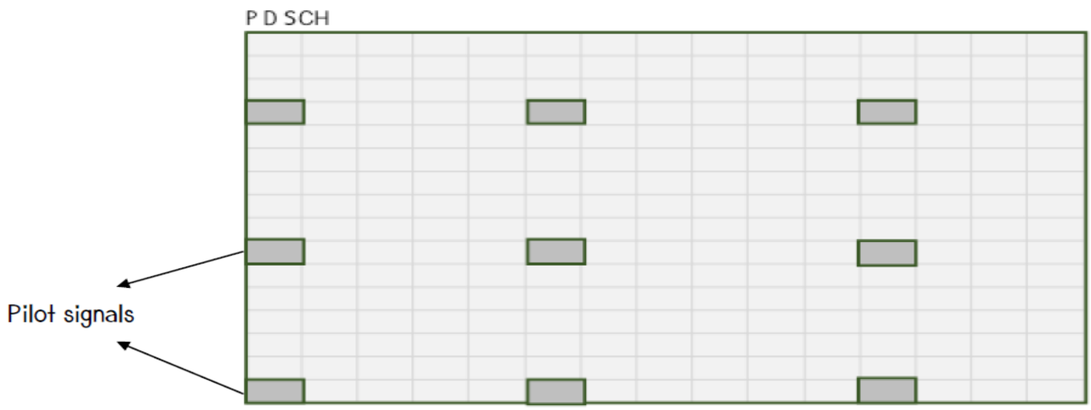


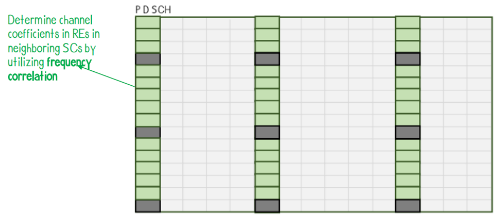
Figure 4.3: Pilot signals in some REs

Use time/frequency correlation to determine channel coefficients in other REs

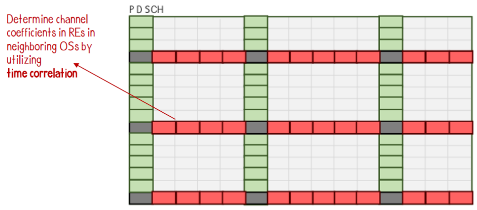
For the above property to be used time and frequency correlations must be determined, these are known using coherence time and coherence bandwidth.

Coherence bandwidth is a statistical measurement of the range of frequencies over which the channel can be considered "flat", or in other words the approximate maximum bandwidth or frequency interval over which two frequencies of a signal are likely to experience comparable or correlated amplitude fading. If the multi-path time delay spread equals D seconds, then the coherence bandwidth B_c is given approximately by the equation

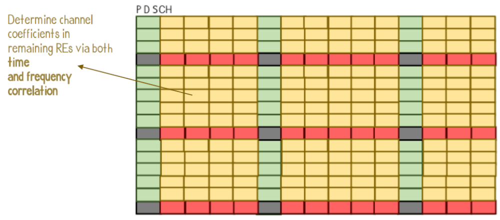
$$B_c \approx \frac{1}{D} \tag{4.1}$$



(a) Frequency correlation



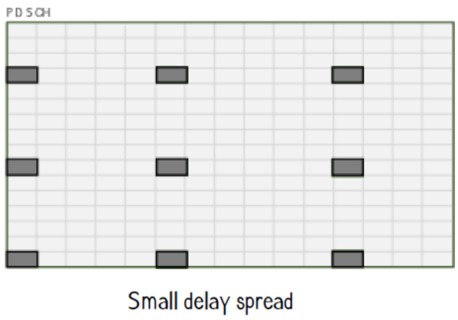
(b) Time correlation



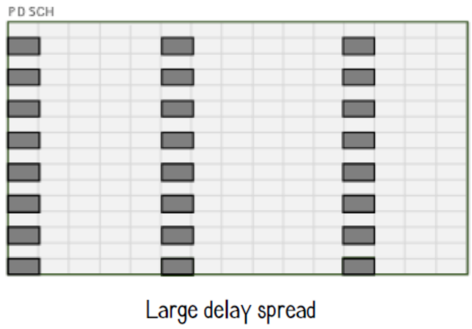
(c) Time / frequency correlation

Figure 4.4: Correlation in time & frequency

Hence if the delay spread is small, frequency correlation is high which means we need less pilots density in the frequency domain.



(a) Small delay spread



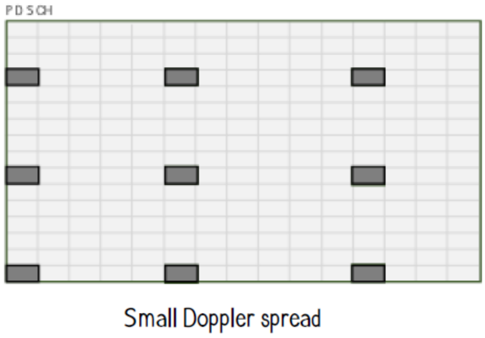
(b) Large delay spread

Figure 4.5: Delay spread effect

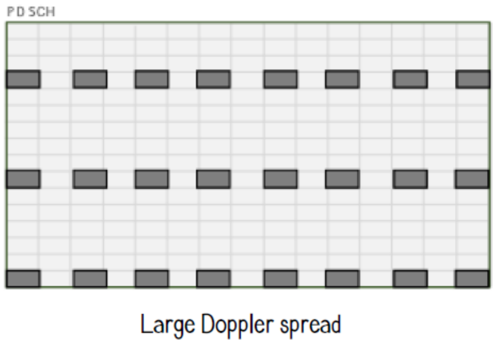
Coherence time is the time duration over which the channel impulse response is considered to be not varying. Such channel variation is much more significant in wireless communications systems, due to Doppler effects. If the maximum doppler spread equals f_m Hz, then the coherence time T_c is given approximately by the equation

$$T_c \approx \frac{1}{f_m} \tag{4.2}$$

Hence if the doppler spread is small, time correlation is high which means we need less pilots density in the time domain.



(a) Small Doppler spread



(b) Large Doppler spread

Figure 4.6: Doppler spread effect

4.3 Channel State Information Reference Signal (CSI-RS)

In the first release of LTE (release 8), channel knowledge for the downlink transmission direction was solely acquired by means of device measurements on the so-called cell-specific reference signals (CRS). The LTE CRS are transmitted over the entire carrier bandwidth within every LTE sub-frame of length 1 ms, and can be assumed to be transmitted over the entire cell area. Thus, a device accessing an LTE network can assume that CRS are always present and can be measured on.
In LTE release 10 the CRS were complemented by so-called CSI-RS. In contrast to CRS, the LTE CSI-RS are not necessarily transmitted continuously. Rather, an LTE device is explicitly configured to measure on a set of CSI-RS and does not make any assumptions regarding the presence of a CSI-RS unless it is explicitly configured for the device. The origin for the introduction of CSI-RS was the extension of LTE to support spatial multiplexing with more than four layers, something which was not possible with the release-8 CRS. However, the use of CSI-RS was soon found to be an, in general, more flexible and efficient tool for channel sounding, compared to CRS. In later releases of LTE, the CSI-RS concept was further extended to also support, for example, interference estimation and multi-point transmission.
As already described, a key design principle for the development of NR has been to as much as possible avoid "always on" signals. For this reason, there are no CRS-like signals in NR. Rather, the only "always-on" NR signal is the so called SS block which is transmitted over a limited bandwidth and with a much larger periodicity compared to the LTE CRS. The SS block can be used for power measurements to estimate, for example, path loss and average channel quality. However, due to the limited bandwidth and low duty cycle, the SS block is not suitable for more detailed channel sounding aimed at tracking channel properties that vary rapidly in time and/or frequency.
Instead the concept of CSI-RS is reused in NR and further extended to, for example, provide support for beam management and mobility as a complement to SS block. CSI-RS is a DL signal that can be used by the UE to measure some parameters and reports it back to the gNB to take actions based on that parameters. Some of this parameters can be :

- Channel quality indicator (CQI)
- Precoding matrix indicator (PMI)
- Rank indicator (RI)

4.3.1 Basic CSI-RS structure

A configured CSI-RS may correspond to up to 32 different antenna ports, each corresponding to a channel to be sounded. In NR, a CSI-RS is always configured on a per-device basis. It is important to understand though that configuration on a per-device basis does not necessarily mean that a transmitted CSI-RS can only be used by a single device. Nothing prevents identical CSI-RS using the same set of resource elements to be separately configured for multiple devices, in practice implying that a single CS-RS is shared between the devices.
As illustrated in Fig. 2.7, a single-port CSI-RS occupies a single resource element within a block corresponding to one resource block in the frequency domain and one slot in the time domain. In principle, the CSI-RS can be configured to occur anywhere within this block although in practice there are some restrictions to avoid collisions with other downlink physical channels and signals. Especially, a device can assume that transmission of a configured CSI-RS will not collide with:

- Any CORESET configured for the device.
- Demodulation reference signals associated with PDSCH transmissions scheduled for the device.
- Transmitted SS blocks.

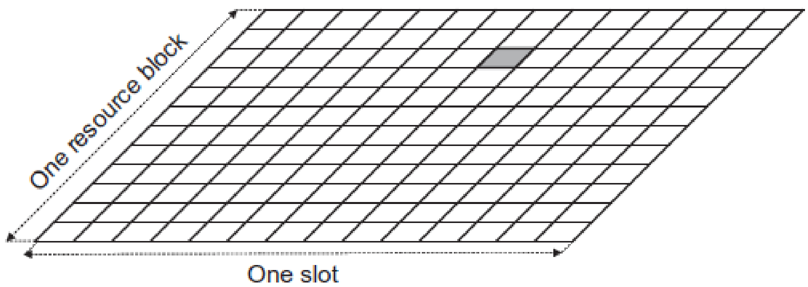


Figure 4.7: Single-port CSI-RS structure consisting of a single resource element within an RB/slot block.

4.3.2 Frequency-Domain structure of CSI-RS configurations

A CSI-RS is configured for a given downlink bandwidth part and is then assumed to be confined within that bandwidth part and use the numerology of the bandwidth part.
The CSI-RS can be configured to cover the full bandwidth of the bandwidth part or just a fraction of the bandwidth. In the latter case, the CSI-RS bandwidth and frequency-domain starting position are provided as part of the CSI-RS configuration.
Within the configured CSI-RS bandwidth, a CSI-RS may be configured for transmission in every resource block, referred to as CSI-RS density equal to one.
However, a CSI-RS may also be configured for transmission only in every second resource block, referred to as CSI-RS density equal to 1/2. In the latter case, the CSI-RS configuration includes information about the set of resource blocks (odd resource blocks or even resource blocks) within which the CSI-RS will be transmitted. CSI-RS density equal to 1/2 is not supported for CSI-RS with 4, 8, and 12 antenna ports.
There is also a possibility to configure a single-port CSI-RS with a density of 3 in which case

the CSI-RS occupies three subcarriers within each resource block. This CSI-RS structure is used as part of a so-called Tracking Reference signal (TRS).

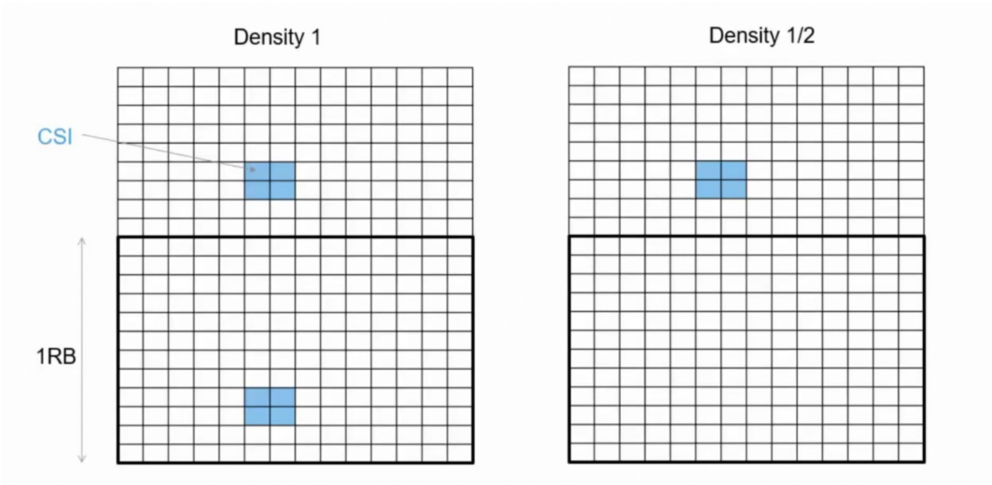


Figure 4.8: CSI-RS Frequency density

4.3.3 Time-Domain structure of CSI-RS configurations

The per-resource-block CSI-RS structure outlined above describes the structure of a CSI-RS transmission, assuming the CSI-RS is actually transmitted in a given slot. In general, a CSI-RS can be configured for periodic, semi-persistent, or aperiodic transmission.
In the case of periodic CSI-RS transmission, a device can assume that a configured CSI-RS transmission occurs every Nth slot, where N ranges from as low as four, that is, CSI-RS transmissions every fourth slot, to as high as 640, that is, CSI-RS transmission only every 640th slot. In addition to the periodicity, the device is also configured with a specific slot offset for the CSI-RS transmission
In the case of semi-persistent CSI-RS transmission, a certain CSI-RS periodicity and corresponding slot offset are configured in the same way as for periodic CSI-RS transmission. However, actual CSI-RS transmission can be activated/ deactivated based on MAC control elements (MAC CE). Once the CSI-RS transmission has been activated, the device can assume that the CSIRS transmission will continue according to the configured periodicity until it is explicitly deactivated. Similarly, once the CSI-RS transmission has been deactivated, the device can assume that there will be no CSI-RS transmissions according to the configuration until it is explicitly re-activated.
In the case of aperiodic CSI-RS, no periodicity is configured. Rather, a device is explicitly informed ("triggered") about each CSI-RS transmission instant by means of signaling in the DCI.

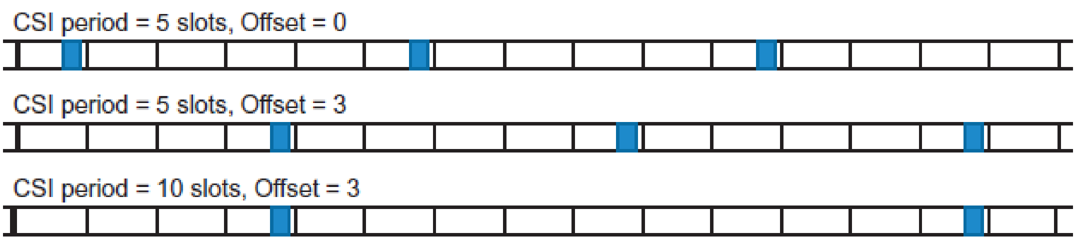


Figure 4.9: CSI-RS Periodicity and slot offset

4.3.4 CSI-RS and DMRS for full channel estimation

DMRS is used to help the receiver determine the effective channel, i.e., after taking into account the effect of the precoding matrix.
The question is how to determine the best precoding matrix ? CSI-RS is the answer. Channel estimation process steps are as follows:

- CSI-RS is transmitted to the UE to make initial channel estimation.
- A CSI feedback is sent to the gNB to be used to choose the necessary precoding, keep in mind that it is not necessary that the gNB uses the precoding suggested by the UE.
- PDSCH and its DMRS are transmitted after applying the chosen precoding.
- DMRS is used by the UE to estimate the effective channel after taking into account the effect of the precoding.

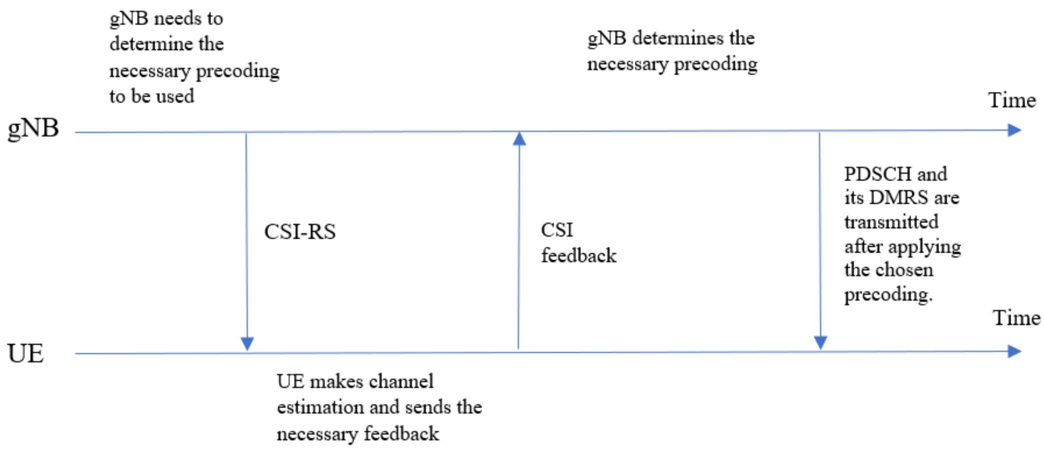


Figure 4.10: Channel estimation process

4.4 Channel Estimation

In general, there are two types of MIMO channel estimation methods:

Training based approach: which uses known training symbols.

Blind-based approach: which perform CE without the benefit of known training symbols.

In training-based CE, known training symbols are transmitted at certain prescribed times and frequencies that are known by the receiver. Since the receiver knows the training symbols, as well as when and where (i.e., at which frequencies) they are transmitted, it uses that information to estimate the gain and phase rotation imparted by the channel at each point in time and frequency based on the characteristics of the received training symbols. Although blind-based methods have higher bandwidth efficiencies because they do not use any resources for transmitting training symbols, they tend to have lower speed and poorer performance than training-based methods. For this reason, training-based CE is used more than blind-estimation, and it is the method we focus on in our project.

The placement of training symbols in time, frequency, and space (i.e., the transmit antenna's) dimensions is a key part of the design of a MIMO communication system. In general, training symbols should be spaced as far apart as possible to reduce training overhead, while still maintaining a required performance level. For example, in a high Doppler, fast fading environment, training symbols need to be placed relatively often in time. Similarly, in a highly frequency-selective channel, training symbols need to be placed close together in the frequency dimension. In this section we discuss two channel estimation methods :

- Simple pilot based estimation
- MMSE estimation

4.4.1 Simple pilot based estimation

Assume a single antenna port at both Tx and Rx

$$y = h_{11}x$$

In order to know h_{11}, we need to send a symbol x Which is known at the Rx, then h_{11} can be known as

$$h_{11} = \frac{y}{x}$$

This known symbol x is called a *pilot symbol.*

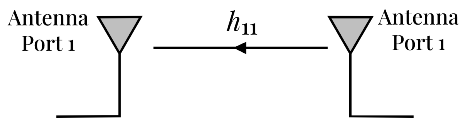


Figure 4.11: Single Tx Single Rx

What about a MIMO channel ?
Assume a 2x2 MIMO setup as shown in Figure 4.12
The received signal y at the receiver side is

$$y = \begin{bmatrix} y_1 \\ y_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{21} \\ h_{12} & h_{22} \end{bmatrix} \times \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}$$

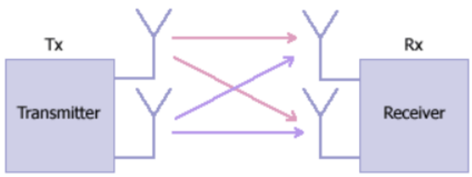


Figure 4.12: 2x2 MIMO setup

Estimating the 4 channel coefficients is not as simple as the SISO case, to do so one pilot symbol won't be enough, a number of pilot symbols equal the number of transmit antennas will be needed.

On the grey REs, pilots from the first layer only are sent

$$y = \begin{bmatrix} y_1 \\ y_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{21} \\ h_{12} & h_{22} \end{bmatrix} \times \begin{bmatrix} p_1 \\ 0 \end{bmatrix} = \begin{bmatrix} h_{11} \\ h_{12} \end{bmatrix} p_1$$

Knowing the value of the pilot symbol at the receiver side the channel coefficients corresponding to the first Tx layer can be estimated.
The same is done on the red REs with pilots from the second layer only are send, hence, the channel coefficients corresponding to the second Tx layer can be estimated.
This means that a number of pilot symbols equals the number of Tx layers is needed to fully

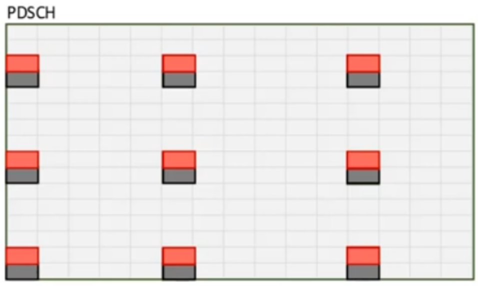


Figure 4.13: pilot symbols for 2x2 MIMO channel

estimate a MIMO channel.

The problem with this estimation method is that in the presence of noise the estimated channel coefficients are not exactly the real coefficients which results in errors in both precoding and decoding operations at the Tx and Rx. This problem appears clearly specially when estimation is done for two reference signals (i.e., CSI-RS and DMRS) where CSI-RS is used to choose the necessary precoding then DMRS is used to estimate the effective channel at the receiver side for decoding operation, the accumulated error in both estimations becomes big which results in a poor BER performance.

4.4.2 Minimum mean square error (MMSE) estimation

In MMSE the same approach as simple pilot based estimation (as mentioned in subsection 4.4.1) is done but the difference is that MMSE estimation takes into account the noise effect to get a better estimate of the channel coefficients.
For the channel model $y = Hx + n$ the channel estimate Ĥ is given as

$$\hat{H} = R_{hy} R_{yy} (\bar{y} - \bar{\mu_y}) + \mu_n \tag{4.3}$$

$$\begin{aligned} \text{where} \quad & R_{hy} = \mathbb{E}\{(h - \mu_h)(\bar{y} - \bar{\mu_y})^T\} \quad \text{is the cross covariance of } H, y \\ \text{and} \quad & R_{yy} = \mathbb{E}\{(\bar{y} - \bar{\mu_y})(\bar{y} - \bar{\mu_y})^T\} \quad \text{is the covariance matrix of } y \\ & \bar{\mu_y} \quad , \quad \mu_n \quad \text{are the expected value of } y \text{ and } n \text{ respectively.} \end{aligned}$$

Using some mathematical manipulation we get that the estimate value of the channel can be given by this simplified expression

$$\hat{H} = \frac{\frac{\bar{x}^H \bar{y}}{N_o} + \frac{\mu_h}{\sigma_h^2}}{\frac{\|\bar{x}\|^2}{N_o} + \frac{1}{\sigma_h^2}} \tag{4.4}$$

Although MMSE estimation is more computationally complex than simple pilot estimations it gives a better estimate of the channel coefficients and hence a better BER performance.

Chapter 5

Simulations

Part II

Deep Reinforcement Learning Model