Estimation and Correction of RF Impairments in LTE Downlink

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Approval Sheet

This thesis entitled Estimation and Correction of RF Impairments in LTE Downlink by G.V.S.Santosh is approved for the degree of Master of Technology from IIT Hyderabad.

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Dedicated to

My Trio

Abstract

This Work focuses on refining the received signal at mobile receiver in 4G-LTE mobile communication. Impairments Such as Carrier Frequency Offset, Sampling Frequency offset , DC Offset and IQ Imbalance are studied and some possible ways to overcome those problems has discussed. Simulations are done in Matlab to verify algorithms to correct those impairments. They are also tested on received signals from some of the Test Equipments.

Soft Demapping of 16 and 64 QAM is discussed,a simplified approach to demodulate symbols has been presented. Also Channel Estimation and Interpolation techniques are discussed.

Contents

Introduction:

OFDM and LTE:

OFDM is simply defined as a form of multi-carrier modulation where the carrier spacing is carefully selected so that each sub carrier is orthogonal to the other sub carriers. Two signals are orthogonal if their dot product is zero. That is, if you take two signals multiply them together and if their integral over an interval is zero, then two signals are orthogonal in that interval. Orthogonality can be achieved by carefully selecting carrier spacing, such as letting the carrier spacing be equal to the reciprocal of the useful symbol period. As the sub carriers are orthogonal, the spectrum of each carrier has a null at the center frequency of each of the other carriers in the system.

Two periodic signals are orthogonal when the integral of their product over one period is equal to zero. For the case of continuous time:

$$\int_{0}^{T} \cos(2\Pi nft)\cos(2\Pi mf_1 t)\, dt = 0,$$

For the case of discrete time:

$$\sum_{k=0}^{N-1} \cos\left(\frac{2\pi kn}{N}\right)\cos\left(\frac{2\pi km}{N}\right) dt = 0,$$

Where m≠n in both cases.

OFDM transmits a large number of narrowband sub channels. The frequency range between carriers is carefully chosen in order to make them orthogonal one another. In fact, the carriers are separated by an interval of 1/T, where T represents the duration of an OFDM symbol

Each sub – carrier in an OFDM system is a sinusoid with a frequency that is an integer multiple of a fundamental frequency. Each sub – carrier is like a Fourier series component of the composite signal, an OFDM symbol.

The sub – carriers waveform can be expressed as

$$\begin{aligned} \mathrm{s(t)} &= \cos(2\pi fct + \theta k) \\ &= a_n \cos(2\pi n f_o t) + b_n \sin(2\pi n f_o t) \\ &= \sqrt{a_n^2 + b^2 n}\cos(2\pi n f_o t + \varphi), \end{aligned}$$

$$\text{Where } \varphi = \tan^{-1}(\frac{bn}{an})$$

The sum of sub carrier is the baseband OFDM signal

$$s(t) = \sum_{n=0}^{N-1} \{a_{ncos(2\pi n f_o t) - b_n \sin(2\pi n f_o t)\}}$$

It is well known, orthogonal signals can be separated at the receiver by correlation techniques. The receiver acts as a bank of demodulators, translating each carrier down to baseband, the resulting signal then being integrated over a symbol period to recover the data. If the other carriers all beat down to frequencies which, in the time domain means an integer number of cycles per symbol period (T), then the integration process results in a zero contribution from all these carriers.

Principle of OFDM:

In a conventional serial data system, the symbols are transmitted sequentially, one by one, with the frequency spectrum of each data symbol allowed to occupy the entire available bandwidth. A high rate data transmission supposes very short symbol duration, conducing at a large spectrum of the modulation symbol. There are good chances that the frequency selective channel response affects in a very distinctive manner the different spectral components of the data symbol, hence introducing the ISI. The same phenomenon, regarded in the time domain consists in smearing and spreading of information symbols such, the energy from one symbol interfering with the energy of the next ones, in such a way that the received signal has a high probability of being incorrectly interpreted. Intuitively, one can assume that the frequency selectivity of the channel can be mitigated if, instead of transmitting a single high rate data stream, we transmit the data.

Simultaneously, on several narrow-band subchannels (with a different carrier corresponding to each subchannel), on which the frequency response of the channel looks “flat”. Hence, for a given overall data rate, increasing the number of carriers reduces the data rate that each individual carrier must convey, therefore lengthening the symbol duration on each subcarrier. Slow data rate (and long symbol duration) on each subchannel merely means that the effects of ISI are severely reduced. This is in fact the basic idea that lies behind OFDM. Transmitting the data among a large number of closely spaced subcarriers accounts for the “frequency division multiplexing” part of the name. Unlike the classical frequency division multiplexing technique, OFDM will provide much higher bandwidth efficiency. This is due to the fact that in OFDM the spectra of individual subcarriers are allowed to overlap. In fact, the carriers are carefully chosen to be orthogonal one another. As it is well known, the orthogonal signals do not interfere, and they can be separated at the receiver by correlation techniques.

The input data sequence is baseband modulated, using a digital modulation scheme. Various modulation schemes could be employed such as BPSK, QPSK (also with their differential form) and QAM with several different signal constellations. There are also forms of OFDM where a distinct modulation on each subchannel is performed. The modulation is performed on each parallel substream that is on the symbols belonging to adjacent DFT frames. The data symbols are parallelized in N different substreams. Each substream will modulate a separate carrier through the IFFT modulation block.

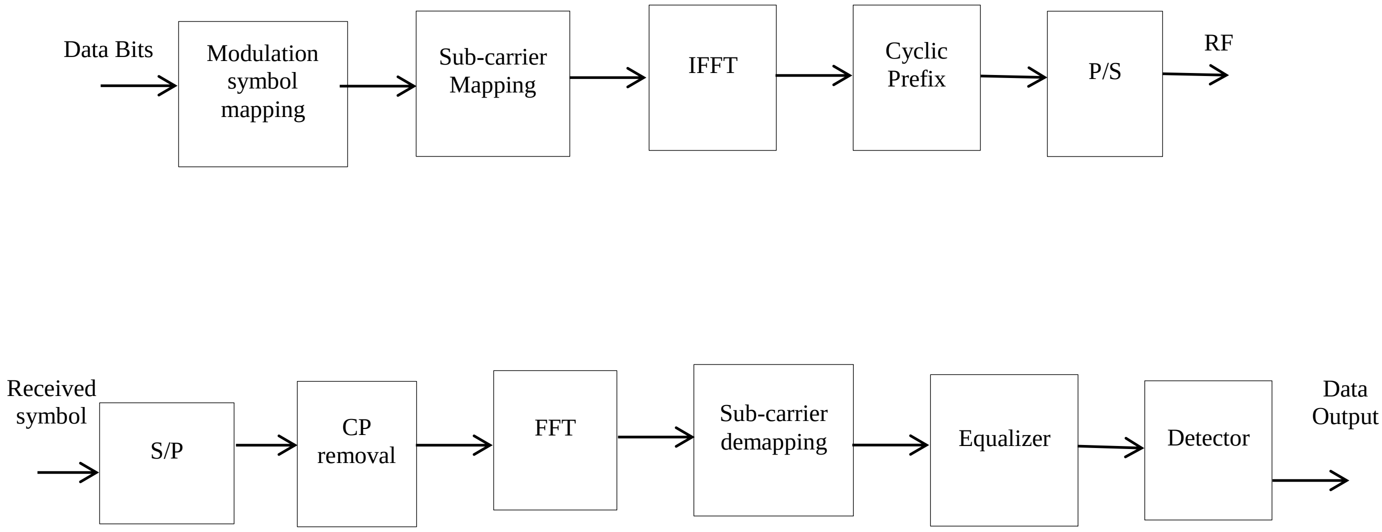


Figure: Modulation and Demodulation in OFDM

Cyclic Prefix :

The Cyclic Prefix or Guard Interval is a periodic extension of the last part of an OFDM symbol that is added to the front of the symbol in the transmitter, and is removed at the receiver before demodulation.

The cyclic prefix has to two important benefits –

- The cyclic prefix acts as a guard interval. It eliminates the inter – symbol interference from the previous symbol.
- It acts as a repetition of the end of the symbol thus allowing the linear convolution of a frequency – selective multipath channel to be modeled as circular convolution which in turn maybe transformed to the frequency domain using a discrete Fourier transform. This approach allows for simple frequency – domain processing such as channel estimation and equalization.

Disadvantages of OFDM:

1. High peak to average power ratio.
2. Susceptible to synchronization errors such as frequency offset.

LTE :

LTE stands for Long Term Evolution and it was started by the telecommunications body Third Generation Partnership Project (3GPP). The main goal of LTE is to provide a high data rate, low latency and packet optimized radio access technology supporting flexible bandwidth deployments. Same time its network architecture has been designed with the goal to support packet-switched traffic with seamless mobility and great quality of service. It provides an uplink speed of up to 50 megabits per second (Mbps) and a downlink speed of up to 100 Mbps. LTE will bring many technical benefits to cellular networks. Bandwidth will be scalable from 1.25 MHz to 20MHz. This will suit the needs of different network operators that have different bandwidth allocations, and also allow operators to provide different services based on spectrum.

The LTE PHY(Physical Layer) is a highly efficient means of conveying both data and control information between an enhanced base station (eNodeB) and mobile user equipment (UE). The LTE PHYemploys some advanced technologies that are new to cellular applications. These include Orthogonal Frequency Division Multiplexing (OFDM) and Multiple Input Multiple Output (MIMO) data transmission.

Features of LTE:

1. High throughput: High data rates can be achieved in both downlink as well as uplink. This causes high throughput.
2. Low latency: Time required to connect to the network is in range of a few hundred milliseconds and power saving states can now be entered and exited very quickly.
3. FDD and TDD: Frequency Division Duplex (FDD) and Time Division Duplex (FDD), both schemes can be used on same platform.
4. High spectral efficiency: The spectral efficiency associated with the high data rates is also high.

5. Simple architecture: Because of Simple architecture low operating expenditure (OPEX).
6. Plug and play: The user does not have to manually install drivers for the device. Instead system automatically recognizes the device, loads new drivers for the hardware if needed, and begins to work with the newly connected device.

Basic LTE parameters:

Time domain:

In LTE, the resources in the time domain are allocated in terms of subframes.

Frame:

A frame in LTE corresponds to time duration of 10ms.

1 Frame=10ms

Sub-Frame:

A frame is further divided into 10 sub-frames. So, the duration of 1 sub-frame is 1ms.

1 Sub-Frame=1ms

Slot:

Slot in LTE is defined as time resource of 0.5ms.Thus a sub-frame consists of 2 slots.

1Slot=0.5ms

Slot is composed of 7(or 6) OFDM symbols in case of normal CP (extendCP).

Chapter 1

Sampling offset correction:

OFDM systems are vulnerable to synchronization errors For example, carrier frequency offsets, which are caused by the inherent instabilities of the transmitter and receiver carrier frequency oscillators, can lead to severe system degradation due to inter carrier interference (ICI). Symbol timing synchronization must also be achieved in order to avoid inter symbol interference. In ofdm systems symbol synchronization is again divided into symbol synchronization and sampling clock synchronization.

1.1 Symbol synchronization:

The purpose of symbol synchronization is to find the correct position of the fast Fourier transform (FFT) window. It can be done at receiver with the aid of the dedicated training symbols and also making use of cyclic prefix property of ofdm symbol to further increase the precision of finding the start of the window. To find the start of ofdm symbol in 4G LTE scenario, We have synchronization reference signals.

1. Primary synchronization signals
2. Secondary synchronization signals

Main objective of these signals is to know the Physical layer Cell Identification number of the Base station. Cell ID can be found upon on User Equipment understanding these signals which are being sent from base station periodically. There are 504 unique physical-layer cell identities. These identities are grouped into

168 physical layer cell identity groups, each group containing three unique identities(0,1,2). The grouping is such that each physical-layer cell identity is part of one and only one physical layer cell identity group. A physical-layer cell identity $N_{\mathrm{ID}}^{\mathrm{cell}} = 3N_{\mathrm{ID}}^{(1)} + N_{\mathrm{ID}}^{(2)}$ is thus uniquely defined by a number $N_{\mathrm{ID}}^{(1)}$ in the range of 0 to 167, representing the physical-layer cell-identity group, and a number $N_{\mathrm{ID}}^{(2)}$ in the range of 0 to 2, representing the physical-layer identity within the physical-layer cell-identity group.

1.2 Primary synchronization signal :

This signal provide us the Physical layer cell identity within the group i.e. $N_{\mathrm{ID}}^{(2)}$. Also these signals help UE to measure the impairments such as carrier frequency offset and sampling frequency offset. These signals are transmitted twice for a radio frame, in 10ms duration.

1.2.1 Sequence generation:

The sequence $d(n)$ used for the primary synchronization signal is generated from a frequency-domain Zadoff-Chu sequence according to

$$d_u(n) = \begin{cases} e^{-j\frac{\pi u n(n+1)}{63}} & n = 0,1,\ldots,30 \\ e^{-j\frac{\pi u (n+1)(n+2)}{63}} & n = 31,32,\ldots,61 \end{cases}$$

where the Zadoff-Chu root sequence index u is given by Table 1.

Table 1: Root indices for the primary synchronization signal

$N_{\mathrm{ID}}^{(2)}$	**Root index** u
0	25
1	29
2	34

1.2.2 Mapping to resource elements:

The mapping of the sequence to resource elements depends on the frame structure. For TDD frame structure PSS is located in the third symbol of the 3rd and 13th slots. For FDD frame structure it is the last ofdm symbol of 1st and 11th slots. In those symbols the sequence $d(n)$ shall be mapped to the resource elements according to

$$a_{k,l} = d(n), \qquad n = 0,\ldots,61$$
$$k = n - 31 + \frac{N_{\mathrm{RB}}^{\mathrm{DL}} N_{\mathrm{sc}}^{\mathrm{RB}}}{2}$$

In either frame type, PSS needs center 62 subcarriers in a OFDM symbol.

1.3 Secondary Synchronization signal:

The SSS consists of a frequency domain sequence, which is an interleaved concatenation of two length-31 m-sequences, named as even sequence and odd sequence here. Both even sequence and odd sequence are scrambled by an m-sequence whose cyclic shift value is dependent on physical layer cell identity within the group. The odd sequence is further scrambled by an m-sequence with cyclic shift value determined by even sequence. The combination of cyclic shifts of even sequence and odd sequence corresponds to the cell group identity. These two length-31 sequences are alternated every 5ms, which allows UE to detect the 10ms frame timing. Decoding this signal helps the UE in estimating the integer carrier frequency offset , which further aids in correcting the sampling timing error.

The even and odd sequence are generated as described below

$$d(2n) = \begin{cases} s_0^{(m_0)}(n) c_0(n) & \text{in subframe 0} \\ s_1^{(m_1)}(n) c_0(n) & \text{in subframe 5} \end{cases}$$
$$d(2n+1) = \begin{cases} s_1^{(m_1)}(n) c_1(n) z_1^{(m_0)}(n) & \text{in subframe 0} \\ s_0^{(m_0)}(n) c_1(n) z_1^{(m_1)}(n) & \text{in subframe 5} \end{cases}$$

where $0 \le n \le 30$. The indices m_0 and m_1 are derived from the physical-layer cell-identity group $N_{\text{ID}}^{(1)}$ according to

$$\begin{aligned} m_0 &= m' \bmod 31 \\ m_1 &= \left(m_0 + \lfloor m'/31 \rfloor + 1\right) \bmod 31 \\ m' &= N_{\text{ID}}^{(1)} + q(q+1)/2, \quad q = \left\lfloor \frac{N_{\text{ID}}^{(1)} + q'(q'+1)/2}{30} \right\rfloor, \quad q' = \left\lfloor N_{\text{ID}}^{(1)}/30 \right\rfloor \end{aligned}$$

where the output of the above expression is listed in Table 2

The two sequences $s_0^{(m_0)}(n)$ and $s_1^{(m_1)}(n)$ are defined as two different cyclic shifts of the m-sequence $\tilde{s}(n)$ according to

$$\begin{aligned} s_0^{(m_0)}(n) &= \tilde{s}\left((n+m_0) \bmod 31\right) \\ s_1^{(m_1)}(n) &= \tilde{s}\left((n+m_1) \bmod 31\right) \end{aligned}$$

Where $\tilde{s}(i) = 1 - 2x(i)$, $0 \le i \le 30$, is defined by

$$x(\bar{i}+5) = \left(x(\bar{i}+2) + x(\bar{i})\right) \bmod 2, \qquad 0 \le \bar{i} \le 25$$

with initial conditions $x(0)=0, \quad x(1)=0, \quad x(2)=0, \quad x(3)=0, \quad x(4)=1$.

The two scrambling sequences $c_0(n)$ and $c_1(n)$ depend on the primary synchronization signal and are defined by two different cyclic shifts of the m-sequence $\tilde{c}(n)$ according to

$$c_0(n) = \tilde{c}((n + N_{\mathrm{ID}}^{(2)}) \bmod 31)$$
$$c_1(n) = \tilde{c}((n + N_{\mathrm{ID}}^{(2)} + 3) \bmod 31)$$

where $N_{\mathrm{ID}}^{(2)} \in \{0,1,2\}$ is the physical-layer identity within the physical-layer cell identity group $N_{\mathrm{ID}}^{(1)}$ and $\tilde{c}(i) = 1 - 2x(i)$, $0 \le i \le 30$, is defined by

$$x(\bar{i} + 5) = \left(x(\bar{i} + 3) + x(\bar{i})\right) \bmod 2, \qquad 0 \le \bar{i} \le 25$$

with initial conditions $x(0) = 0, \quad x(1) = 0, \quad x(2) = 0, \quad x(3) = 0, \quad x(4) = 1$.

The scrambling sequences $z_1^{(m_0)}(n)$ and $z_1^{(m_1)}(n)$ are defined by a cyclic shift of the m-sequence $\tilde{z}(n)$ according to

$$z_1^{(m_0)}(n) = \tilde{z}((n + (m_0 \bmod 8)) \bmod 31)$$

$$z_1^{(m_1)}(n) = \tilde{z}((n + (m_1 \bmod 8)) \bmod 31)$$

where m_0 and m_1 are obtained from Table 2 and $\tilde{z}(i) = 1 - 2x(i)$, $0 \le i \le 30$, is defined by

$$x(\bar{i} + 5) = \left(x(\bar{i} + 4) + x(\bar{i} + 2) + x(\bar{i} + 1) + x(\bar{i})\right) \bmod 2, \qquad 0 \le \bar{i} \le 25$$

with initial conditions $x(0) = 0, \quad x(1) = 0, \quad x(2) = 0, \quad x(3) = 0, \quad x(4) = 1$.

Table:2 Mapping between physical-layer cell-identity group $N_{\mathrm{ID}}^{(1)}$ and the indices m_0 and m_1

$N_{\mathrm{ID}}^{(1)}$	m_0	m_1	$N_{\mathrm{ID}}^{(1)}$	m_0	m_1	$N_{\mathrm{ID}}^{(1)}$	m_0	m_1	$N_{\mathrm{ID}}^{(1)}$	m_0	m_1	$N_{\mathrm{ID}}^{(1)}$	m_0	m_1
0	0	1	34	4	6	68	9	12	102	15	19	136	22	27
1	1	2	35	5	7	69	10	13	103	16	20	137	23	28
2	2	3	36	6	8	70	11	14	104	17	21	138	24	29
3	3	4	37	7	9	71	12	15	105	18	22	139	25	30
4	4	5	38	8	10	72	13	16	106	19	23	140	0	6
5	5	6	39	9	11	73	14	17	107	20	24	141	1	7
6	6	7	40	10	12	74	15	18	108	21	25	142	2	8
7	7	8	41	11	13	75	16	19	109	22	26	143	3	9
8	8	9	42	12	14	76	17	20	110	23	27	144	4	10
9	9	10	43	13	15	77	18	21	111	24	28	145	5	11
10	10	11	44	14	16	78	19	22	112	25	29	146	6	12
11	11	12	45	15	17	79	20	23	113	26	30	147	7	13
12	12	13	46	16	18	80	21	24	114	0	5	148	8	14
13	13	14	47	17	19	81	22	25	115	1	6	149	9	15
14	14	15	48	18	20	82	23	26	116	2	7	150	10	16
15	15	16	49	19	21	83	24	27	117	3	8	151	11	17
16	16	17	50	20	22	84	25	28	118	4	9	152	12	18
17	17	18	51	21	23	85	26	29	119	5	10	153	13	19
18	18	19	52	22	24	86	27	30	120	6	11	154	14	20
19	19	20	53	23	25	87	0	4	121	7	12	155	15	21
20	20	21	54	24	26	88	1	5	122	8	13	156	16	22
21	21	22	55	25	27	89	2	6	123	9	14	157	17	23
22	22	23	56	26	28	90	3	7	124	10	15	158	18	24
23	23	24	57	27	29	91	4	8	125	11	16	159	19	25
24	24	25	58	28	30	92	5	9	126	12	17	160	20	26
25	25	26	59	0	3	93	6	10	127	13	18	161	21	27
26	26	27	60	1	4	94	7	11	128	14	19	162	22	28
27	27	28	61	2	5	95	8	12	129	15	20	163	23	29
28	28	29	62	3	6	96	9	13	130	16	21	164	24	30
29	29	30	63	4	7	97	10	14	131	17	22	165	0	7
30	0	2	64	5	8	98	11	15	132	18	23	166	1	8
31	1	3	65	6	9	99	12	16	133	19	24	167	2	9
32	2	4	66	7	10	100	13	17	134	20	25	-	-	-
33	3	5	67	8	11	101	14	18	135	21	26	-	-	-

1.3.1 Mapping to resource elements:

SSS symbol location is also dependent on the frame structure. For FDD type frame structure it is immediate preceeding symbol of PSS and for TDD type frame structure it is three ofdm symbols earlier.

The sequence $d(n)$ shall be mapped to resource elements according to

$$a_{k,l} = d(n), \qquad n = 0,...,61$$

$$k = n - 31 + \frac{N_{\mathrm{RB}}^{\mathrm{DL}} N_{\mathrm{sc}}^{\mathrm{RB}}}{2}$$

$$l = \begin{cases} N_{\mathrm{symb}}^{\mathrm{DL}} - 2 & \text{in slots 0 and 10} \quad \text{for frame structure type FDD} \\ N_{\mathrm{symb}}^{\mathrm{DL}} - 1 & \text{in slots 1 and 11} \quad \text{for frame structure type TDD} \end{cases}$$

1.4 Carrier Frequency offset estimation:

There comes certain difference between carrier frequency transmitted from Base Station and User Equipment. Possible reasons for this offset is oscillator disturbance, might be either at transmitter or receiver or both, channel conditions. Receiver needs to mitigate the carrier frequency offset if any . Carrier frequency offset disturbs the orthogonality of sub carriers, and causes intercarrrier interference.

In context of 4G-LTE we estimate carrier frequency offset by splitting into integer part and fractional part.

Integer frequency offset is shift in terms of number of subcarriers ,where is subcarrier has a bandwidth of 15 KHz, it can be either positive or negative based on center frequency, for baseband center frequency is zero.

Carrier frequency offset can be estimated using Synchronization Signals transmitted in 4G-LTE communication system.

Among then we can estimate fractional frequency offset using Primary Synchronisation signal whereas Secondary Synchronisation Signal to get integer frequency offset.

1.4.1 System model :

OFDM system with carrier frequency and sampling frequency offset can be modeled as below

Δf = carrier frequency offset

$\varphi = (T'-T)/T$ where T= sampling time.

$n'=n+N_{cp}+lN_s$ N_{cp} length of cyclic prefix

N_s number of samples per symbol

Received samples can be written as

$$r_{l,n} = \left(\left(e^{j2\pi n'(1+\varphi)\Delta f T}\right)\right)\left(\sum_i h_i \sum_l \sum_k a_{i,k}\psi\left(n'T'-\tau_i\right)\right) + n_{l,n}$$

$$\psi\left(n'T'-\tau_i\right) = e^{j2\pi(k/N)\left(n'(1+\varphi)-N_g-lN_S-\tau_i/T\right)}$$

1.4.2 Fractional frequency offset estimate :

The carrier frequency offset (φ) can be split into fractional part and integer part

$$H(k) = b * IDFT\left(\tilde{G}(q)\right) \quad \Delta f = \epsilon + N_I$$

Where N_I is the integer part and ϵ is fractional part ,and it ranges -.5 < ϵ < .5

After detecting the start of the ofdm symbol and using cyclic prefix property of it we can get an estimate of ϵ

$$\gamma(n) = \sum_{n}^{n+L-1} r(k) r^*(k+N)$$

Max-loglikelihood of the above function will give us the estimate of ϵ . L denotes the CP length, measured in time samples.

$$\epsilon = -\frac{1}{2\pi} \arg\{\gamma(n)\}$$

The performance of this method can be improved by averaging over several OFDM symbols, In our context we perform the above operation on PSS signals that being transmitted periodically every 5ms by Base Station, it leads to assumption that carrier frequency offset is constant over that duration , this method also suffers from the channel conditions, which may not result in getting start of an OFDM symbol precisely. In our implementation we go for rigorous correlations spanning over large number of samples to find the exact start of PSS symbol.

1.4.3 Integer carrier frequency offset estimate:

In our 4G-LTE receiver design we estimate integer frequency offset in frequency domain making use of transmitted Secondary synchronization signals. To decode SSS signal we need to decode the physical layer group identity. Below are the steps to estimate integer frequency offset N_I

- Extract the SSS signal according to the estimated OFDM symbol timing and apply an N-point FFT.

- Separate 62 length sequence d(n) into sequence d(2n) and d(2n + 1), consisting of even and odd sub-carrier symbols

In the following we use the notation of sub-frame 0

- Divide d(2n)/$c_0(n)$ in order to obtain the sequence $s_0^{m_0}(n)$ .Sequence $c_0(n)$ is known at the receiver as it depends only on the already estimated sector-ID N_{ID}^2

- Build a reference sequence $S_{ref}(n)$, which is a duplicated version of $s_0^{m_0=0}(n)$ Or $s_0^{m_0=1}(n)$ with the length of 62

- Apply a cross-correlation between $s_0^{m_0=0}(n)$ and the reference sequence $S_{ref}(n)$ .The magnitude of the correlation term shows a significant maximum location which indicates the estimate of m_0
- After estimating the integer m_0 , we are able to compute $z_1^{m_0}(n)$ which is obtained by cyclically shifting already known reference sequence z1(n) by the estimated integer m_0 and afterwards divide d(2n + 1)/$c_1(n)$ $z_1^{m_0}(n)$ in order to obtain the sequence $s_1^{m_1}(n)$.
- Apply a cross-correlation between $s_1^{m_1}(n)$ and $S_{ref}(n)$. The magnitude of the correlation term shows a significant maximum at a position , which is estimated as integer m_1 .
- The pair of estimated m_0 and m_1 identities the group-ID N_{ID}^1
- Compute the overall cell-ID : N= $N_{ID}^2 + 3 * N_{ID}^1$

The described procedure is performed for several cyclic shifts of d(n) (where the cyclic shifts are varied for instance from -15 to +15 sub-carriers), in order to detect the integer CFO part N_I . Significant peaks will only be generated if the received sequence d(n) is compensated for the integer CFO N_I .The cyclic shift of d(n) for which significant peaks occurs gives the estimated integer CFO.

This way we will have an estimate for carrier frequency offset

1.5 Sampling frequency offset:

The minute difference in the sampling time results in drastic effects if unchecked and corrected. The sampling clock errors include the clock phase error and the clock frequency error. The

clock phase error effects are similar to the symbol timing errors and hence we treat the clock phase errors as a kind of the symbol timing errors. The sampling clock frequency errors can cause the Inter-Carrier Interference.

In frequency domain, affects of sampling clock offset can be Modelled as below

$$\varphi = (T' - T)/T \qquad \text{where } T = \text{sampling time.}$$

$$T_u = \text{ofdm symbol duration}$$

$$y_{(i,n)} = e^{(j2\pi i n\varphi(T/T_u))}.sinc(n\varphi).X_{(i,n)}.H_{(i,n)} + n_{(i,n)} + n_\varphi(i,n)$$

Where sinc(x) = sin*(πx)/ πx* and $n_\varphi(i,n)$ is the sampling clock frequency offset caused additional noise with variance

$$Var\left[n_\varphi\right] \approx \frac{\pi^2}{3}(n\varphi)^2$$

$n\varphi \ll 1,\ n_\varphi(i,n)$ can be usually be neglected. But sampling if goes unchecked can result in OFDM symbol timing drift. This drift need to be correct and symbol timing has to adjusted in continuous manner, otherwise we may completely move out of a symbol.

1.6 Sampling clock offset correction :

There are two cases here , under sampling and oversampling. sampling offset is mainly an hardware imperfection we can say, also temperature may also disturb the clock. Offset is usually measured in Parts Per million (PPM) . In Usual receivers same clock is used for down-conversion of received signal and sampling its Baseband Signal. So we can make use of carrier frequency offset estimate to calculate the sampling frequency offset. Sign of the carrier frequency offset estimate determines whether it is under sampling or oversampling.

Relation between carrier frequency offset and sampling clock offset is below

$$\frac{\Delta f_c}{f_c} = \frac{\Delta f_s}{f_s}$$

where f_c is carrier frequency and f_s is sampling rate

further

$$\Delta f_s = \frac{1}{T_s} - \frac{1}{T_s^{'}}$$

Where $T_s =$ sampling time

We have prior knowledge of standard sampling rate and carrier frequency.

$$T_s^{'} = \left(\frac{\Delta f_c}{f_c} * f_s + f_s \right)^{-1}$$

We can now have an estimate for

$$\Delta T_s = T_s - T_s^{'}$$

Addition or deletion of sample is done when

$$n * \Delta T_s = \pm T_s \qquad \text{where n is sample index}$$

For adding a sample at location ***n*** averaging of a sample preceeding and succeeding ***n*** are taken to get a new sample that is to be added in ofdm symbol. For estimating of carrier frequency offset we assumed that offset is remains unchanged in half frame. So that assumptions gets translated to here directly. We have to make a prior estimate of n and correct it before for entire half frame. Then proceed to decoding of ofdm symbols. This method has been tested rigorously for different channel conditions and noise levels. Though we have corrected the timing drift , affect of error on symbols in frequency domains still prevails and can corrected in channel equalization.

Chapter 2

Channel Estimation and Equalization:

2.1 Channel estimation techniques:

In order demodulate Physical downlink shared channel , we need to estimate channel before, to estimate the channel we make use of cell specific reference signals. Downlink shared channel is transmitted in resource blocks either of cell specific reference or UE specific reference signals to get estimate of channel. These reference signals are arranged in resource blocks based on following arguments. The required spacing in time between the reference symbols can be determined by considering the maximum Doppler spread (highest speed) to be supported, which for LTE corresponds to 500 km/h. The Doppler shift is $f_d = (f_c * v/c)$ where f_c is the carrier frequency, v is the UE speed in metres per second, and c is the speed of light($3 \cdot 10^8$ m/s). Considering f_c = 2 GHz and v = 500 km/h, then the Doppler shift is f_d =950 Hz. According to Nyquist's sampling theorem, the minimum sampling frequency needed in order to reconstruct the channel is therefore given by $T_c = 1/(2f_d) = 0.5$ms under the above assumptions. This implies that two reference symbols per slot are needed in the time domain in order to estimate the channel correctly.

In the frequency direction, there is one reference symbol every six subcarriers on each OFDM symbol that includes reference symbols, but the reference symbols are staggered so that within each Resource Block (RB) there is one reference symbol every three subcarriers This spacing is related to the expected coherence bandwidth of the channel, which is in turn related to the channel delay spread.

2.1 Cell-specific Reference Signal (CRS):

Cell-specific reference signals shall be transmitted in all downlink subframes in a cell supporting PDSCH transmission. These signals are transmitted on one or several of antenna ports 0 to 3.these reference signals are defined for $\Delta f = 15\ \text{kHz}$ only.

2.1.2 Sequence generation:

The reference-signal sequence $r_{l,n_s}(m)$ is defined by

$$r_{l,n_s}(m) = \frac{1}{\sqrt{2}}\left(1 - 2\cdot c(2m)\right) + j\frac{1}{\sqrt{2}}\left(1 - 2\cdot c(2m+1)\right), \qquad m = 0,1,\ldots,2N_{\text{RB}}^{\text{max,DL}} - 1$$

where n_s is the slot number within a radio frame and l is the OFDM symbol number within the slot. The pseudo-random sequence $c(i)$ is defined in clause 7.2. The pseudo-random sequence generator shall be initialised with $c_{\text{init}} = 2^{10}\cdot\left(7\cdot\left(n_s + 1\right) + l + 1\right)\cdot\left(2\cdot N_{\text{ID}}^{\text{cell}} + 1\right) + 2\cdot N_{\text{ID}}^{\text{cell}} + N_{\text{CP}}$ at the start of each OFDM symbol where

$$N_{\text{CP}} = \begin{cases} 1 & \text{for normal CP} \\ 0 & \text{for extended CP} \end{cases}$$

2.1.3 Mapping to resource elements:

The reference signal sequence $r_{l,n_s}(m)$ shall be mapped to complex-valued modulation symbols $a_{k,l}^{(p)}$ used as reference symbols for antenna port p in slot n_s according to

$$a_{k,l}^{(p)} = r_{l,n_s}(m')$$

where

$$k = 6m + \left(v + v_{\text{shift}}\right) \bmod 6$$
$$l = \begin{cases} 0, N_{\text{symb}}^{\text{DL}} - 3 & \text{if } p \in \{0,1\} \\ 1 & \text{if } p \in \{2,3\} \end{cases}$$
$$m = 0,1,\ldots,2 \cdot N_{\text{RB}}^{\text{DL}} - 1$$
$$m' = m + N_{\text{RB}}^{\text{max,DL}} - N_{\text{RB}}^{\text{DL}}$$

The variables v and v_{shift} define the position in the frequency domain for the different reference signals where v is given by

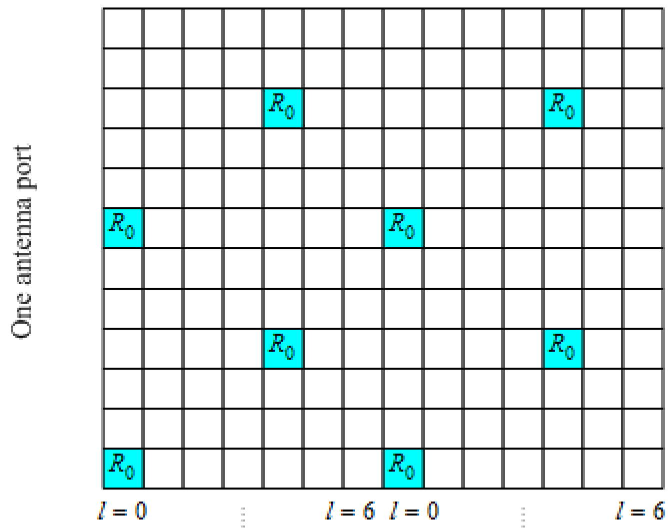
$$v = \begin{cases} 0 & \text{if } p = 0 \text{ and } l = 0 \\ 3 & \text{if } p = 0 \text{ and } l \neq 0 \\ 3 & \text{if } p = 1 \text{ and } l = 0 \\ 0 & \text{if } p = 1 \text{ and } l \neq 0 \\ 3(n_{\text{s}} \bmod 2) & \text{if } p = 2 \\ 3 + 3(n_{\text{s}} \bmod 2) & \text{if } p = 3 \end{cases}$$

The cell-specific frequency shift is given by $v_{\text{shift}} = N_{\text{ID}}^{\text{cell}} \bmod 6$.

Resource elements $\left(k,l\right)$ used for transmission of cell-specific reference signals on any of the antenna ports in a slot shall not be used for any transmission on any other antenna port in the same slot and set to zero.

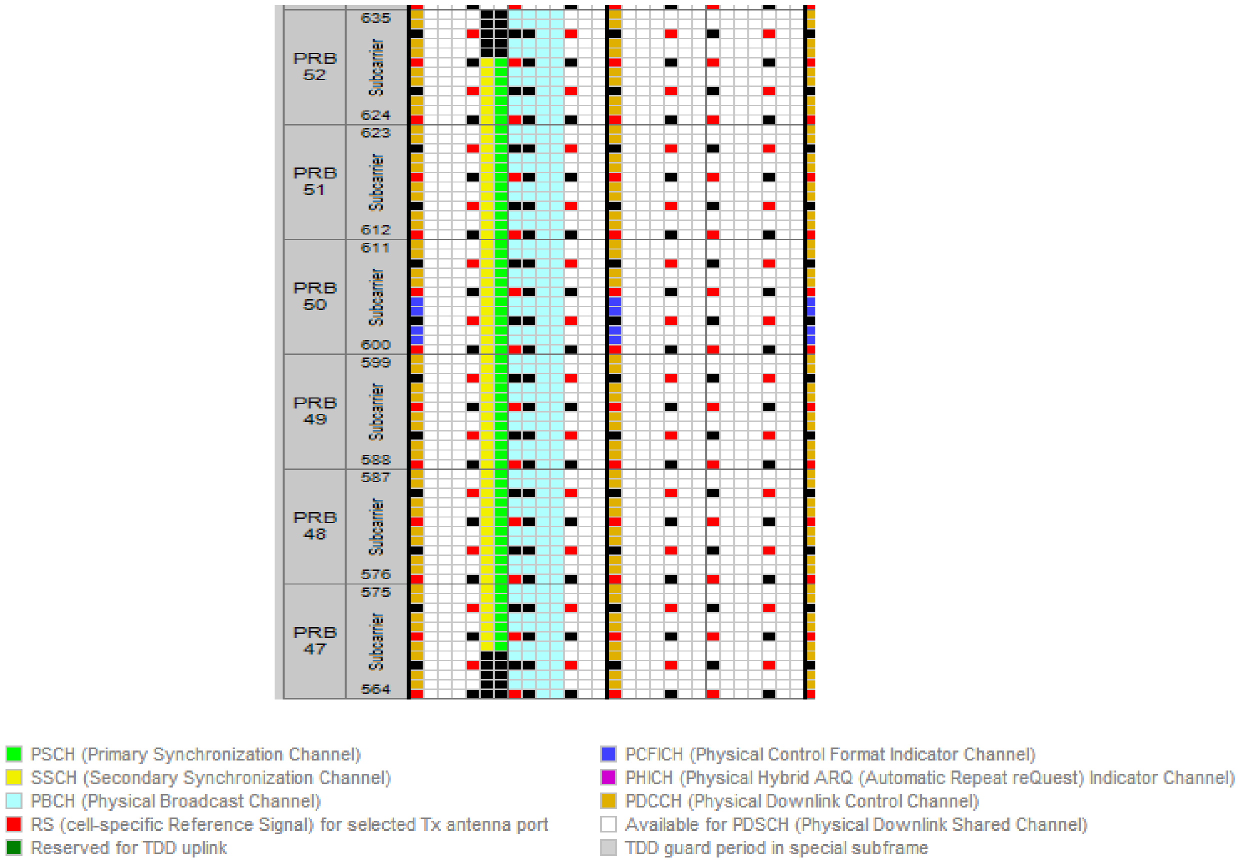
In an MBSFN subframe, cell-specific reference signals shall only be transmitted in the non-MBSFN region of the MBSFN subframe.

Figures below illustrate the resource elements used for reference signal transmission according to the above definition. The notation R_p is used to denote a resource element used for reference signal transmission on antenna port p.



Mapping of downlink reference signals (normal cyclic prefix)

PDSCH can be sent on resource blocks available, which ever blocks not scheduled for other reference signals or any other channels



In order to decode modulation symbol sent over PDSCH , channel need to be estimated and has to be corrected, here In our we will estimate channel making use of Cell Specific Reference Signals, how well we estimate and interpolate directly affect the Bit Error rate of receiver. We studied the existing Channel interpolation Techniques and compared them based on the Bit Error Rates.

2.2 Signal Model:

The received signal in frequency domain can written as below

$$y(k) = H(k)X(k) + N$$

k is sub-carrier index, H(k) channel frequency response and N in additive White Gaussian Noise.

2.3 Least Squares Channel Estimation:

The least square (LS) channel estimation is a simple estimation technique with very low complexity. It does not require any prior knowledge of the channel statistics. It is widely used because of its simplicity. However, it suffers from a high mean square error. The LS estimation channel frequency response H(k) is obtained by minimizing

$$\left\| Y(k) - X(k)\tilde{H}(k) \right\|^2$$

Least squares estimate is given by

$$H_{LS}(k) = Y(k) / X(k)$$

Once the channel is estimated we need to interpolate the response to all the other sub-carriers.

2.4 Interpolation methods:

2.4.1 Linear Interpolation (LI)

In Linear Interpolation algorithm, two successive or adjacent pilot subcarriers are used to determine the channel response for data subcarriers that are located in between the pilot signals. Using Linear Interpolation method, the estimated channel response for the data subcarrier d, $ks_f \le d < (k+1)s_f$ is given by the following equation .

$$\bar{H}_{LI}(d) = \bar{H}_{LI}(ks_f + 1)$$

$$= \left(1 - \frac{l}{s_f}\right)\bar{H}_p(k) + \frac{l}{s_f}\bar{H}_p(d)$$

$$= \bar{H}_p(k) + \frac{l}{s_f}\left(\bar{H}_p(k+1) - \bar{H}_p(k)\right)$$

$$\text{for } \; 0 \le l < s_f$$

This interpolation can be implemented with least complexity can also give better performance when subcarriers are close.

2.4.2 Second order linear interpolation:

Second order interpolation gives better channel response than linear interpolation
The channel estimated by second-order interpolation is given by:

$$\bar{H}_{LI}(d) = c_1\bar{H}_p(m-1) + c_0\bar{H}_p(m) + c_{-1}\bar{H}_p(m+1)$$

Where $c_1 = \gamma(\gamma - 1)$

$$c_0 = -(\gamma-1)(\gamma+1), \quad \gamma = \frac{l}{N}$$

$$c_{-1} = \frac{\gamma(\gamma+1)}{2}$$

N is the number of sub-carriers between two pilots. Simulations results show this methods performs better than the first order linear interpolation.

2.4.3 Spline cubic interpolation:

In this algorithm a smooth and continuous polynomial is fitted to the given data points is produced. The transfer function of each subcarrier is approximated by a third order polynomial with respect to l/S_f . this method uses higher order interpolation, which gives better performance among the all other methods going to discussed in this work. Using Spline Cubic Interpolation method, the channel response estimated for the data subcarrier d, $ks_f \le d < (k+1)s_f$ is given by following equation

$$\bar{H}_{sci}(d) = \bar{H}_{sci}(kS_f + l); 0 \le l < S_f$$

$$= \alpha_1 \bar{H}_p(k+1) + \alpha_0 \bar{H}_p(k) + \alpha_1 S_f \bar{H}_p^{'}(k+1) - \alpha_0 S_f \bar{H}_p^{'}(k);$$

$$k = 0, 1, \ldots N_p - 1$$

Where $\bar{H}_p^{'}(k)$ is first order derivative of $\bar{H}_p(k)$

And

$$\alpha_1 = 3\frac{(S_f - l)^2}{s_f^2} - 2\frac{(S_f - l)^3}{s_f^3}$$

$$\alpha_0 = 3\frac{(l)^2}{s_f^2} - 2\frac{(l)^3}{s_f^3}$$

2.4.4 Low Pass Filter interpolation:

After getting estimates of channel through Least Squares estimation , pad zeros at other sub carrier locations and this is passed through a low pass Finite Impulse Response filter. Interpolated response obtained after low pass filtering has less mean squared error with Ideal channel response. Performance of this method depends mostly on filter design, In matlab for simulation, use Interp function, to understand this interpolation method.

2.4.5 FFT Based interpolation:

We can see the channel estimates as the down sampled version of actual channel. This interpolation is pretty convincing to understand and also reduces affect noise in channel estimates in interpolation. The Discrete fourier transform of channel estimates can be

$$G(n) = \sum_{m=0}^{M-1} \bar{H}_{LS}(m) \exp(-j2\pi mn / M)$$

If we observe the transform domain plot signal component is spread over low frequencies , and noise is spread over all frequency region. We can realize low pass filter by making samples in high frequency region zero

$$\dot{G}(k)=\begin{cases}G(k) & 0\le p<p_c\ \ M-p_c\le p<M-1\\ 0, & \end{cases}$$

Where p_c is cutoff frequency of filter . after filtering noise component is reduced to $2p_c$/M

Interpolation approach is as below , this signal is extended to N-Samples by padding N-M zeros at high frequency region around p=M/2.

.

This $\tilde{G}(q)$ is fourier transform of desired estimates of channel transfer function, by performing N-point IDFT, estimated transfer function is obtained.

$$H(k)=b*IDFT\left(\tilde{G}(q)\right)$$

a constant b is needed for calibration. Also the cutoff frequency p_c is chosen where 95% of energy is concentrated .

Chapter 3

Soft Demapping of QAM constellation:

We studied a simplified soft-output demapper for 16-QAM and 64-QAM constellations for 4G-LTE. The main objective is to de-map the received signals into soft bits which would have same sign as of the transmitted symbol. Let us denote bits of I-component as $b_{I,k,}$ whereas $b_{q,k}$ for Q-component

Decision on bits can be taken by

$$\hat{b}_{I,k} = \propto if \ \ P\Big[\big[\ b_{I,k} = \propto \big| r[i] \big\rangle P\Big[b_{I,k} = 1 - \propto | \, r[i] \ \ \Big], \propto = 1, 0$$

For $\propto = 1,$ then

$$\hat{b}_{I,k} = 1 \text{ if } \quad log \frac{P[b_{I,k} = 1 \,|\, r[i]],}{P[b_{I,k} = 0 \,|\, r[i]],} > 0$$

Log-Likelihood Ratio(LLR) of decision $\hat{b}_{I,k}$ is defined as

$$LLR\left(\hat{b}_{I,k}\right) = \log \frac{P[b_{I,k} = 1 \,|\, r[i]]}{P[b_{I,k} = 0 \,|\, r[i]]}$$

$$= \log \frac{\sum_{\beta \, \epsilon \, S_{I,k}^{(1)}} P[a[i] = \beta \,|\, r[i]]}{\sum_{\beta \, \epsilon \, S_{I,k}^{(0)}} P[a[i] = \beta \,|\, r[i]]} \ ,$$

If $\propto = 0$ then the LLR function have opposite sign.

This is the soft bit information of bit $b_{I,k}$ by applying bayes rule assuming that transmitted symbols are equally distributed then

$$LLR\left(b_{I,k}\right) = \log \frac{\sum_{\beta \in S_{I,k}^{(1)}} P[r[i] \mid a[i] = \beta]}{\sum_{\beta \in S_{I,k}^{(0)}} P[r[i] \mid a[i] = \beta]}$$

Where a sub optimal simplified LLR can be

$$LLR\left(b_{I,k}\right) = \log \frac{max_{\beta \in S_{I,k}^{(1)}} P[r[i] \mid a[i] = \beta]}{max_{\beta \in S_{I,k}^{(0)}} P[r[i] \mid a[i] = \beta]}$$

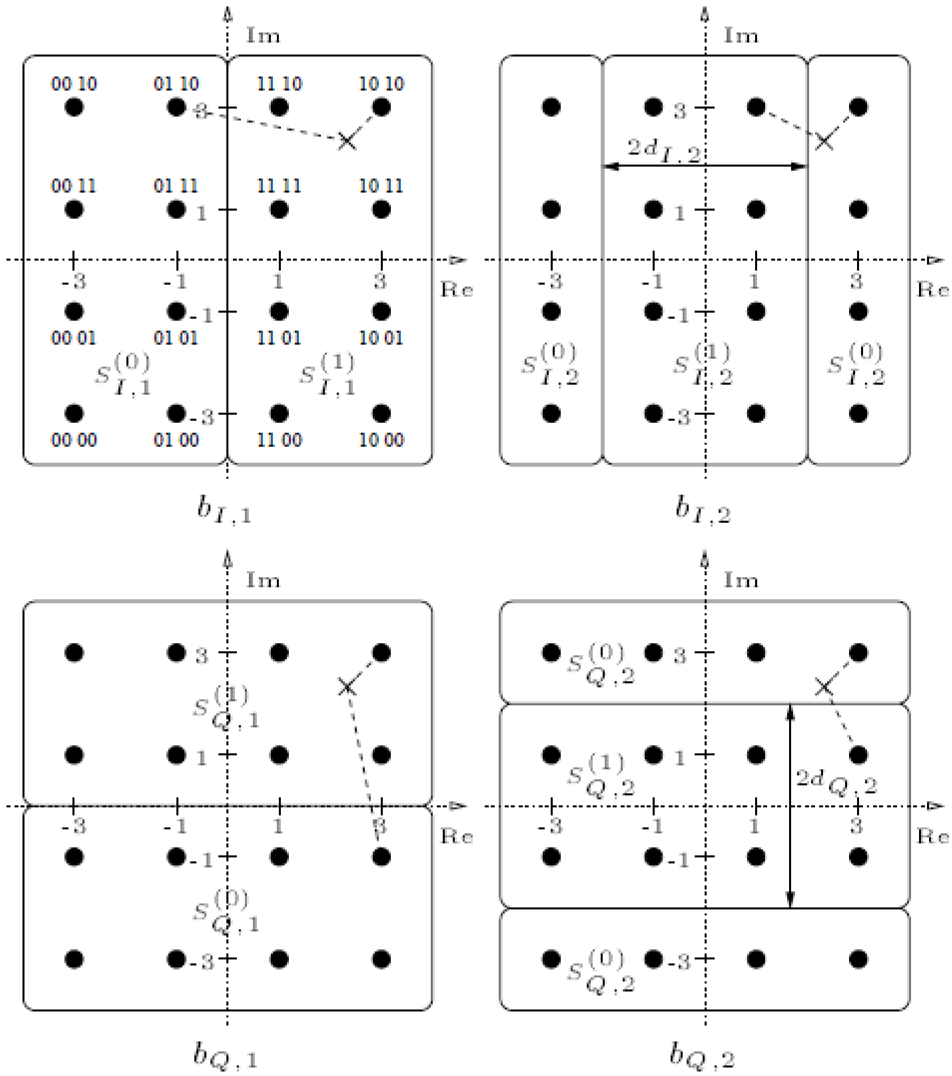
Final soft bit values can be calculated as:

$$LLR\left(b_{I,k}\right) = \frac{\left|H(i)\right|^2}{4} \left\{ min_{\beta \in S_{I,k}^{(0)}} \left|y[i] - \beta\right|^2 - min_{\beta \in S_{I,k}^{(1)}} \left|y[i] - \beta\right|^2 \right\}$$

3.2 Simplified LLR computation:

Below figure shows the partitions $S_{I,k}^{(1)}, S_{I,k}^{(0)}$ for generic case of bit $b_{I,k}$

And $S_{Q,k}^{(1)} S_{Q,k}^{(0)}$ of bit $b_{Q,k}$



the QAM constellation is split into two partitions of complex symbols, namely $S_{I,k}^{(0)}$ comprising the symbols with a `0' in position (I, k) and similar is case with $S_{I,k}^{(1)}$.

$$LLR\left(b_{I,k}\right)=\frac{\left|H\left(i\right)\right|^2}{4}\left\{min_{\beta \in S_{I,k}^{(0)}}\left|y\left[i\right]-\beta\right|^2-min_{\beta \in S_{I,k}^{(1)}}\left|y\left[i\right]-\beta\right|^2\right\}$$

$$\triangleq\left|H\left(i\right)\right|^2.D_{I,k}$$

Where $D_{I,k}$. for 16 QAM is

$$. \quad D_{\mathrm{I},1}=\begin{cases} y_I\left[i\right] & \left|y_I\left[i\right]\right|\leq 2 \\ 2(y_I\left[i\right]-1) & y_I\left[i\right]>2 \\ 2(y_I\left[i\right]+1) & y_I\left[i\right]<-2 \end{cases}$$

$$D_{I,2}=-\left|y_I\left[i\right]\right|+2.$$

For

Where $D_{Q,k}$ for 16 QAM is

$$D_{Q,1}=\begin{cases} y_Q\left[i\right] & \left|y_Q\left[i\right]\right|\leq 2 \\ 2(y_Q\left[i\right]-1) & y_Q\left[i\right]>2 \\ 2(y_Q\left[i\right]+1) & y_Q\left[i\right]<-2 \end{cases}$$

$$D_{Q,2}=-\left|y_Q\left[i\right]\right|+2.$$

And formulas derived for 64-QAM constellation, similar are equations for Q component.

$$D_{I,1} = \begin{cases} Y_I[i] & , \quad |Y_I[i]| <= 2 \\ 2(Y_I[i] - 1) & , \quad 2 <= Y_I[i] <= 4 \\ 4(Y_I[i] - 3) & , \quad Y_I[i] > 6 \\ 3(Y_I[i] - 2) & , \quad 4 <= Y_I[i] <= 6 \qquad 4(Y_I[i] - 3) \quad , \\ 2(Y_I[i] + 1) & , \quad -4 <= Y_I[i] <= -2 \\ 3(Y_I[i] + 2) & , \quad -6 <= Y_I[i] <= -4 \\ 4(Y_I[i] + 3) & , \quad Y_I[i] < -6 \end{cases}$$

$$D_{I,3} = \begin{cases} |Y_I[i]| - 2 & , \quad |Y_I[i]| <= 4 \\ -|Y_I[i]| + 6 & , \quad |Y_I[i]| > 4 \end{cases}$$

$$D_{I,2} = \begin{cases} 2(-|Y_I[i]| + 3) & , \quad |Y_I[i]| <= 2 \\ 4 - |Y_I[i]| & , \quad 2 < |Y_I[i]| <= 6 \\ 2(-|Y_I[i]| + 5) & , \quad |Y_I[i]| > 6 \end{cases}$$

$$D_{I,3} = \begin{cases} |Y_I[i]| - 2 & , \quad |Y_I[i]| <= 4 \\ -|Y_I[i]| + 6 & , \quad |Y_I[i]| > 4 \end{cases}$$

Chapter 4

DC-offset correction:

3.1 DC-offset

Along with carrier frequency offset, OFDM may also suffer from other disturbances such as direct current offset (DCO). For the sake of cost and power efficiency, mobile receiver architecture is under evolution from super heterodyne to direct conversion . DCO is one of the most common disturbances of direct conversion receiver . It comes from self-mixing of local oscillator (LO) signal or radio frequency (RF) signal due to the finite isolation between input ports of mixer. In OFDM systems, DCO not only degrades demodulation performance but also violates CFO estimation, Impacts of static DCO on CFO estimation can either be eliminated by analog high pass filter (HPF) or be compensated in digital domain with data-aided approaches.

3.2 DC offset correction in the context of 4G-LTE

We introduced DC offset as given by the below equation and estimated DC offset from Frequency domain of received Primary syncronisation signal and also estimated value is refined from estimate made based on Secondary sychronisation signal. Thus can be done in two ways , either suppressing it to zero in frequency domain or by subtracting the value from the time domain received signal. Estimating DC offset is not that trivial, if the received signal has integer carrier frequency , you have check at corresponding shifted DC carrier for estimating it.

Chapter 5

IQ Imbalance correction:

5.1 Introduction

In future generation's wireless communication markets such as LTE, it is very advantageous to develop low cost transceivers. Zero IF modulators and demodulators are used in the recent designs to reduce the cost. However, this includes IQ demodulation at RF, which therefore cannot be done digitally and thus introduces IQ mismatch. Unfortunately, OFDM is very sensitive to receiver IQ imbalance. . IQ imbalance is not apparent in lower-order QAM modulation. But, in higher-order QAM modulation, it will become serious interference.

5.2 I-Q imbalance

. A low cost implementation of OFDM is a challenging task because of the RF impairment due to the RF analog components. There are two receiver architectures differentiated on the basis of how the RF is down converted. One is the direct conversion RF receiver known for its low cost and low power implementation on silicon and the other one is superheterodyne receiver.

. IQ imbalance arises when a front-end component doesn't respect the power balance or the orthogonality between the I and Q branch. We can therefore characterize this effect by 2 parameters: the amplitude imbalance ε, and the phase mismatch $\Delta\varphi$. The complex baseband equation for the IQ imbalance effect on the ideal signal x is given by

$$
\begin{aligned}
y &= (1+\varepsilon)\cos\Delta\varphi\Re(x) - j(1-\varepsilon)\sin\Delta\varphi\Re(x) \\
&\quad + j(1-\varepsilon)\cos\Delta\varphi\Im(x) - (1+\varepsilon)\sin\Delta\varphi\Im(x) \\
&= (\cos\Delta\varphi \text{ - } j\varepsilon\sin\Delta\varphi)x - (\varepsilon\cos\Delta\varphi \text{ - } j\sin\Delta\varphi)x^* \\
&= \alpha.x + \beta.x^*
\end{aligned}
$$

Where y is the signal with I-Q imbalance. $\Re()$ denotes the real part, $\Im()$ the imaginary and $()*$ the complex conjugate.

$$
\begin{aligned}
\alpha &= \cos\Delta\varphi \text{ - } j\varepsilon\sin\Delta\varphi \\
\beta &= \varepsilon\cos\Delta\varphi \text{ - } j\sin\Delta\varphi
\end{aligned}
$$

In the absence of IQ imbalance α = 1 and β = 0.

5.3 I-Q imbalance in OFDM

To analyse the effect of IQ imbalance in OFDM let us consider a data vector $\mathbf{x_t}$ in frequency domain which needs to be transmitted and $\mathbf{x_r}$ is received data. IFFT($\mathbf{x_t}$) is the time domain signal that is being transmitted from the transmitter. Applying the IQ imbalance (3) and taking the fft leads to

$$
\begin{aligned}
\boldsymbol{x_r} &= FFT\{\alpha.IFFT(\boldsymbol{x_t}) + \beta.[IFFT(\boldsymbol{x_t})]^*\} \\
&= \alpha.\boldsymbol{x}_t + \beta.(\boldsymbol{x_{t,m}}^*)
\end{aligned}
$$

Where bold variable stands for a vector.

5.4 I-Q Imbalance Estimation

The estimator for the imbalance can be written as

$$
d_{cor} = \frac{d_r - (d_t^*)\beta}{\alpha}
$$

If we can estimate α and β then we can correct the distortion that has been introduced by I-Q imbalance. One of the approaches to estimate is to α and β is to consider the above equation as a complex equation in complex variables α and β. If we consider 2 non-zero carriers together, we get a set of 2 linear complex equations, which can be solved easily. Solving a pair of equations we get

$$\alpha = \frac{(d_r)_i(d_t^*) + (d_r)_j(d_t^*)}{(d_t)_i(d_t) - (d_t)_j(d_t^*)}$$

$$\beta = \frac{(d_r)_i - \alpha(d_t)_i}{(d_t^*)}$$

the above equations are valid if we don't consider channel affects on received symbols, which may not be the case always. re writing above equations , suffix indicates the rotation of symbols in frequency domain around the center carrier.

$$y_{\boldsymbol{r}} = FFT\{\alpha.IFFT(\mathrm{h}\,\boldsymbol{x_t}) + \beta.[IFFT(\mathrm{h}\,\boldsymbol{x_t})]^*\}$$
$$= \alpha.\mathrm{h}\,\boldsymbol{x}_t + \beta.(\mathrm{h}_m\,\boldsymbol{x_{t,m}})^*$$

Where h is the frequency domain response of channel, accordingly the estimates of alpha and beta are derived as below

$$\alpha = \frac{(d_r)_i(\mathrm{h}_m\,d_{t,m})^*{}_j + (d_r)_j(\mathrm{h}_m\,d_t)^*{}_i}{(\mathrm{h}\,d_t)_i(\mathrm{h}_m\,d_t)^*{}_j - (\mathrm{h}\,d_t)_j(\mathrm{h}_m\,d_t)^*{}_i}$$

$$\beta = \frac{(d_r)_i - \alpha(\mathrm{h}\,d_t)_i}{(\mathrm{h}_m\,d_t)_i^*}$$

5.5 Correction:

The received signal can be corrected as shown below , better estimates of alpha, beta directly implies the correction accuracy. In 4G-LTE Scenario, to estimate alpha and beta, we used Primary Synchronization Signal and Secondary Synchronization signal . channel knowledge needed for estimation of parameters is obtained from Secondary Synchronization Signal and estimates are refined using Primary Synchronisation sequence

As there are 62 symbols are available.

$$d_{corrected} = \frac{(d_r)_i * \alpha - \beta(d_{r,m})^*{}_i}{|\alpha|^2 - |\beta|^2}$$

Chapter 6

Results

All impairments discussed in the above chapters are simulated in our LTE link simulator, We tested our algorithms of correction of these impairments using Primary synchronization signal and Secondary synchronization Signals. This chapter presents results of work.

Channel Estimation and Equalization:

We simulated the Physical Downlink Shared Channel with Vehicular A channel and Doppler spread with speeds of user ranging from 3kmph to 300kmph. Below are results for 16 QAM and 64 QAM constellation transmitted in the downlink shared channel , channel is estimated using Cell Specific Reference Signals and least squares channel estimation and Below figures show the various interpolated tested and 30dB SNR

1. **Second order channel interpolation**

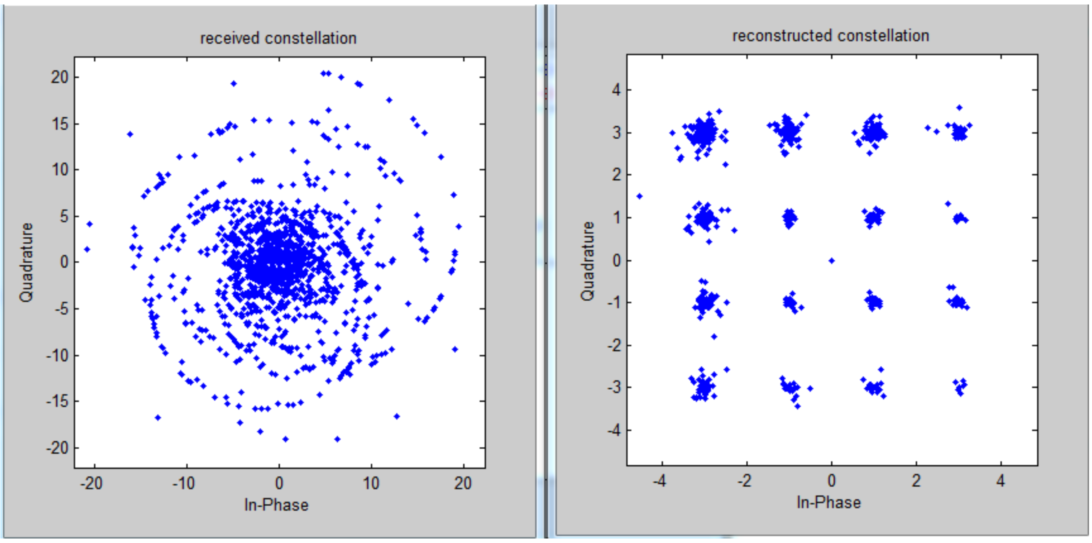


Figure 6.1 Simulation result for 16 QAM and Second order channel interpolation

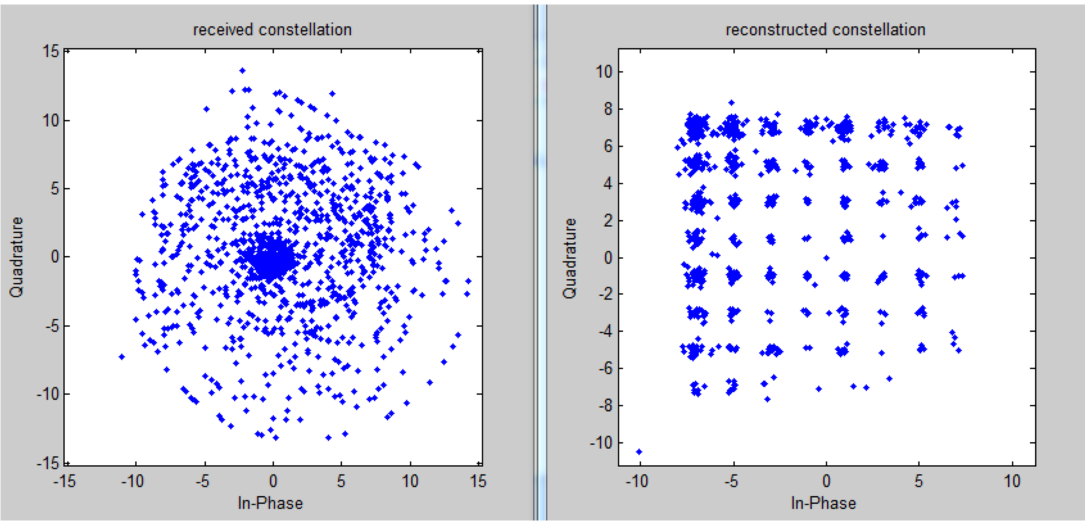


Figure 6.2 Simulation result for 64 QAM and Second order channel interpolation

Spline cubic interpolation:

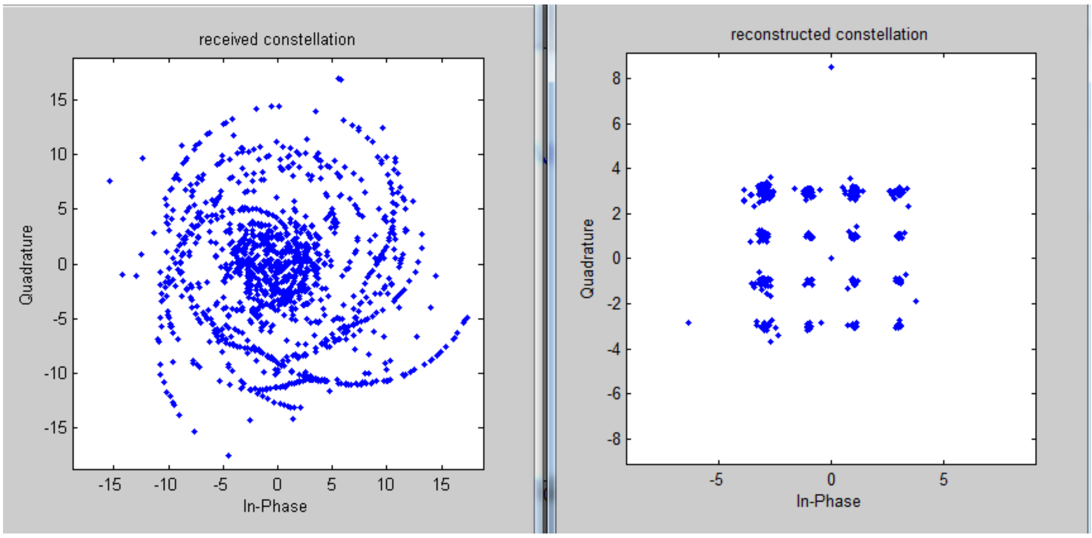


Figure 6.3 Simulation result for 16 QAM and Spline cubic interpolation

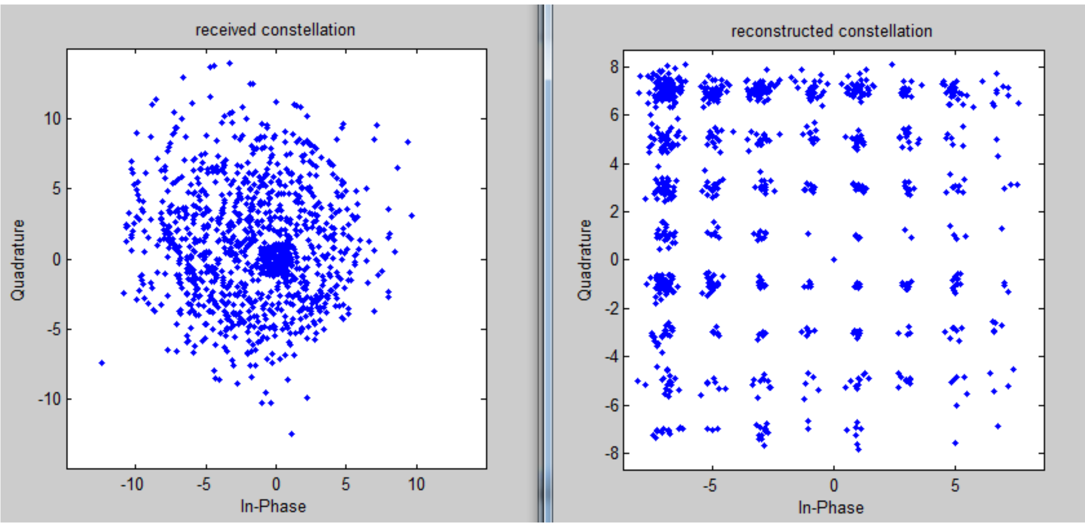


Figure 6.4 Simulation result for 64 QAM and Spline cubic interpolation

Low Pass filter interpolation

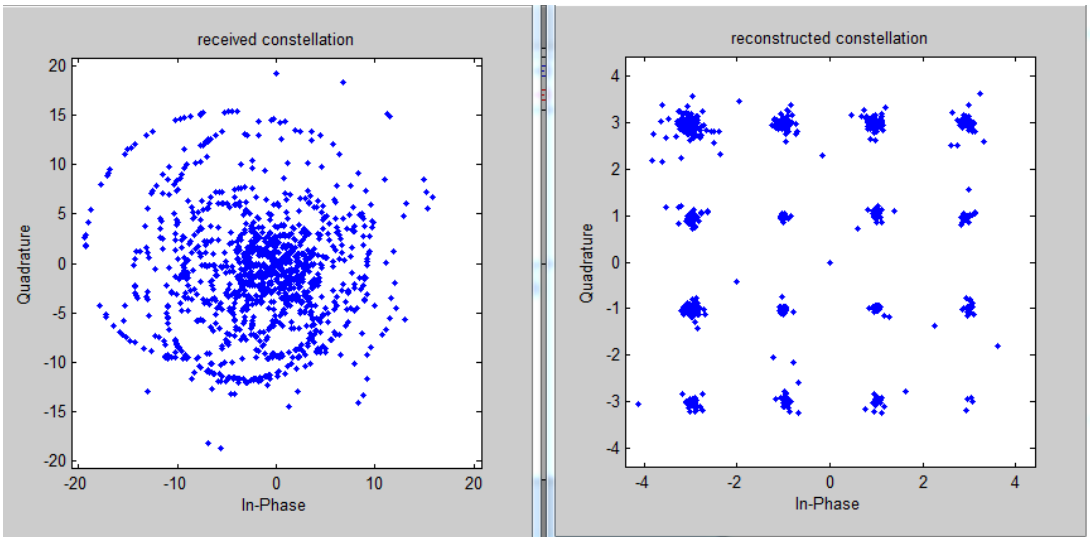


Figure 6.5 Simulation result for 16 QAM and Low Pass Filter interpolation

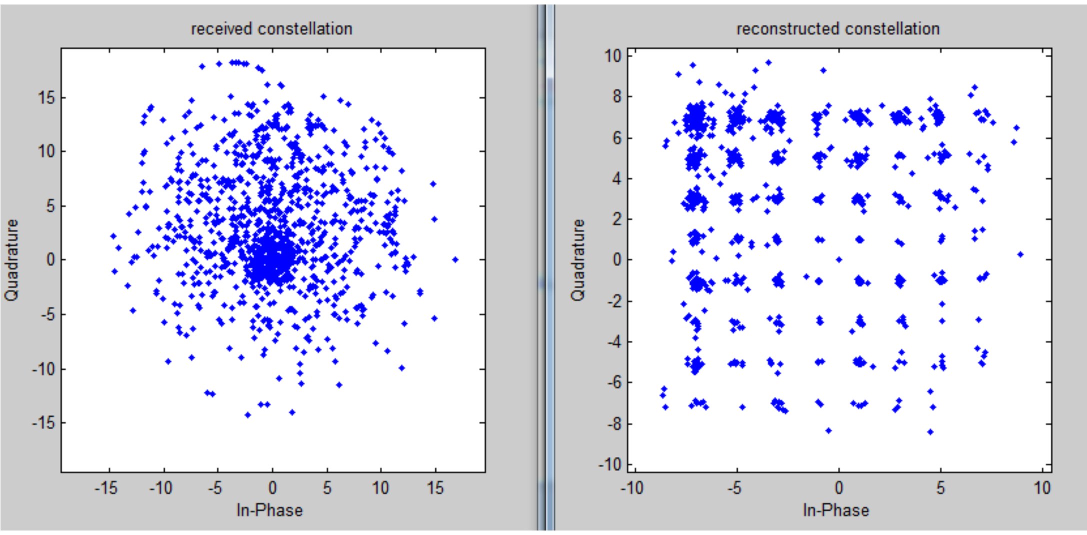


Figure 6.6 Simulation result for 64 QAM and Low Pass Filter interpolation

FFT based interpolation:

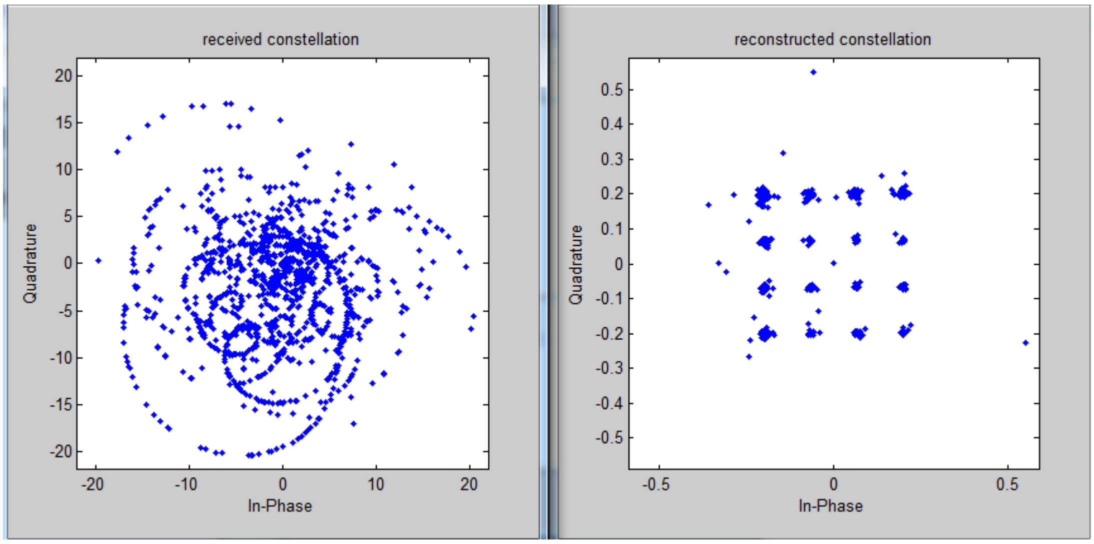


Figure 6.7 Simulation result for 16 QAM and FFT based interpolation

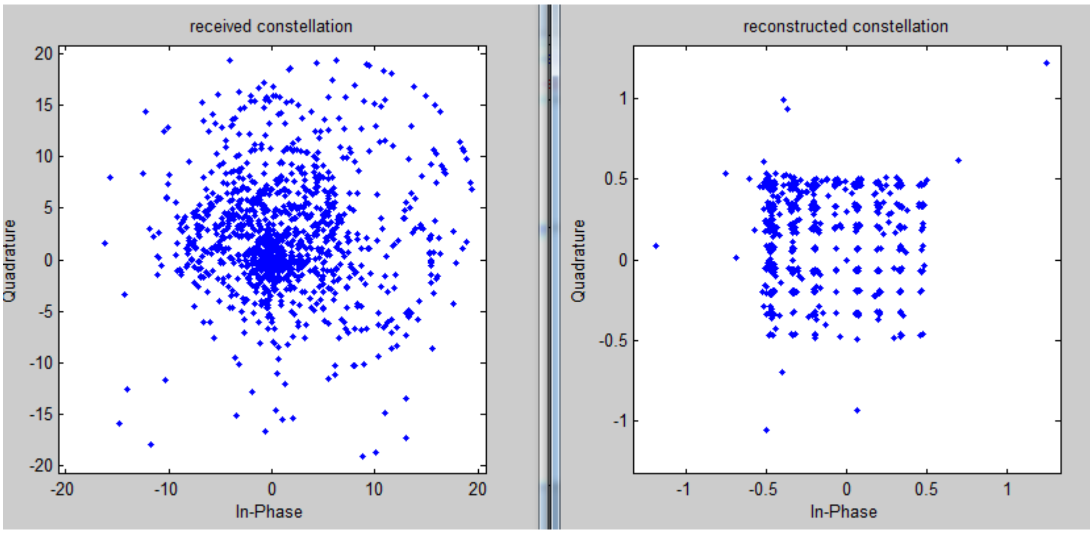


Figure 6.8 Simulation result for 64 QAM and FFT based interpolation

Second order interpolation method 2:

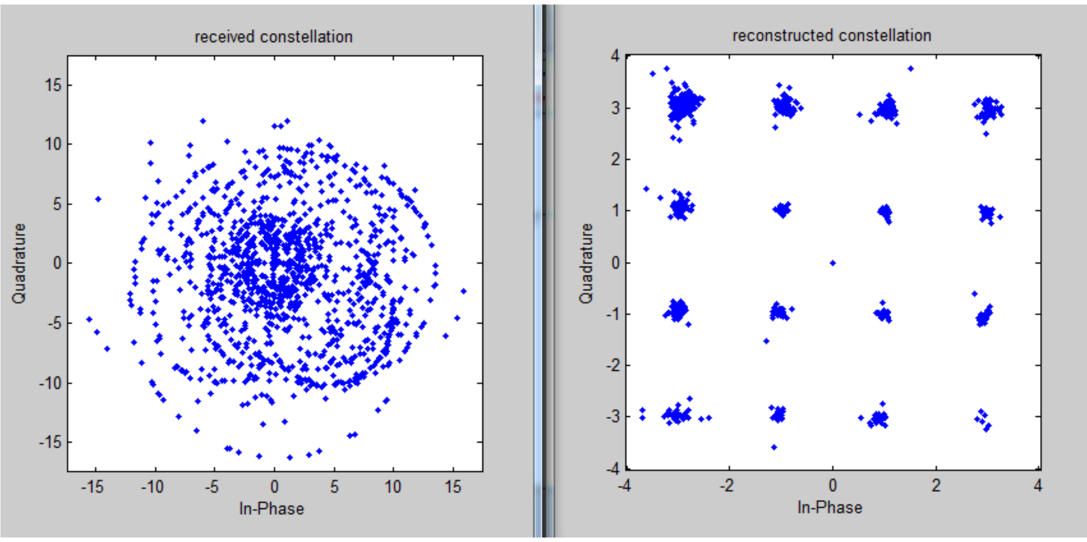


Figure 6.9 Simulation result for 16 QAM

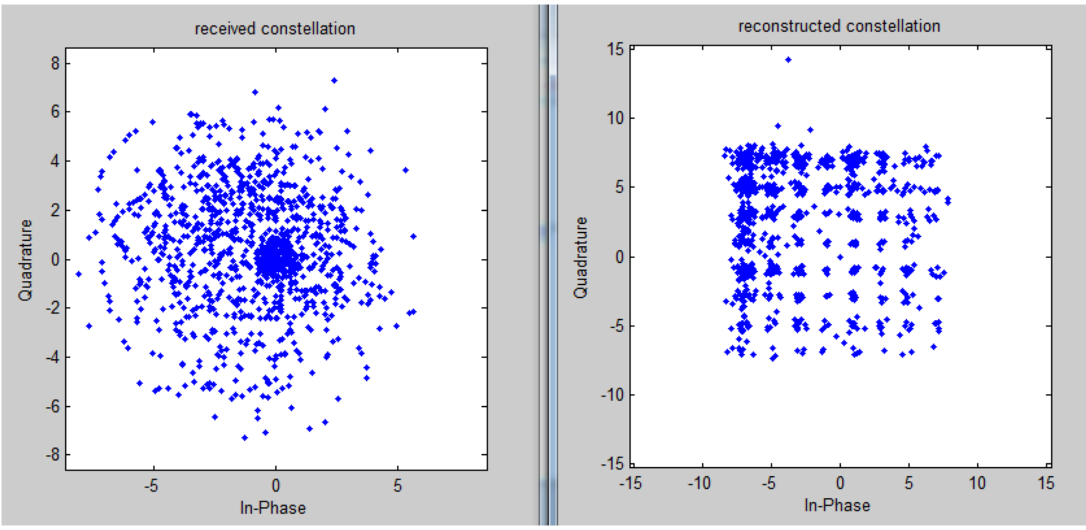


Figure 6.10 Simulation result for 64 QAM

Time domain channel taps:

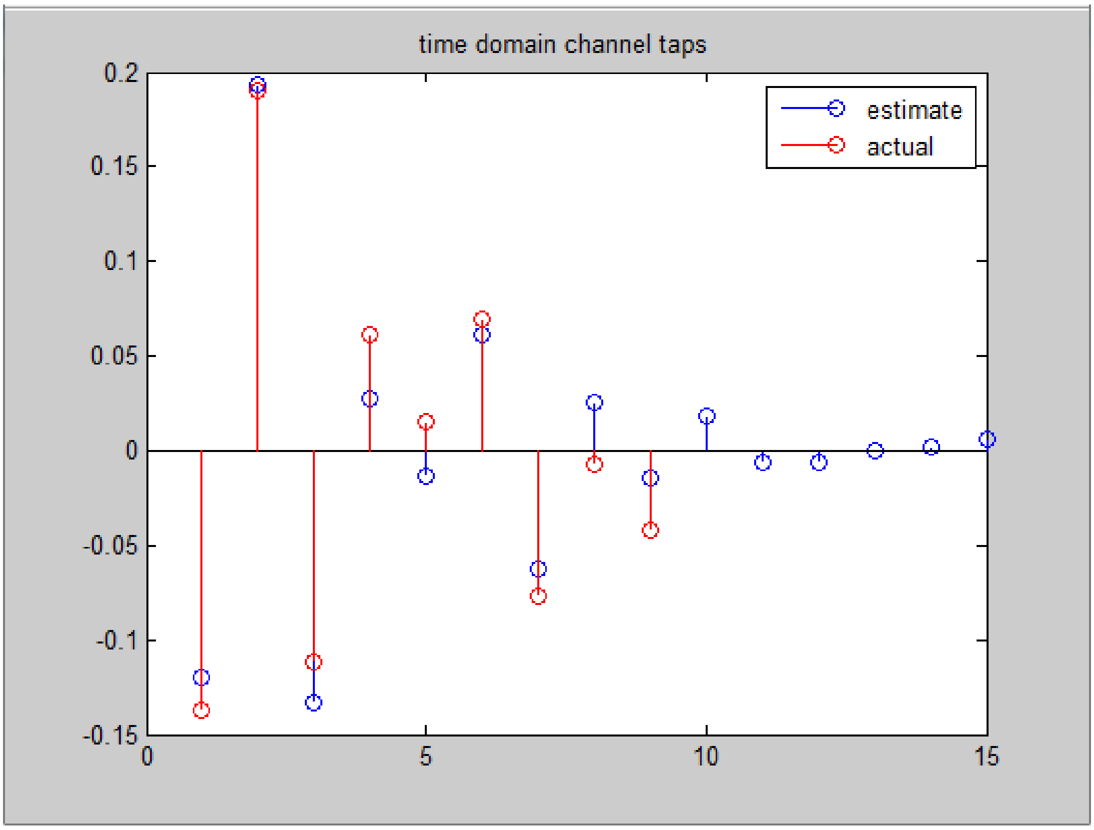


Figure 6.11 time domain channel taps

I Q Imbalance

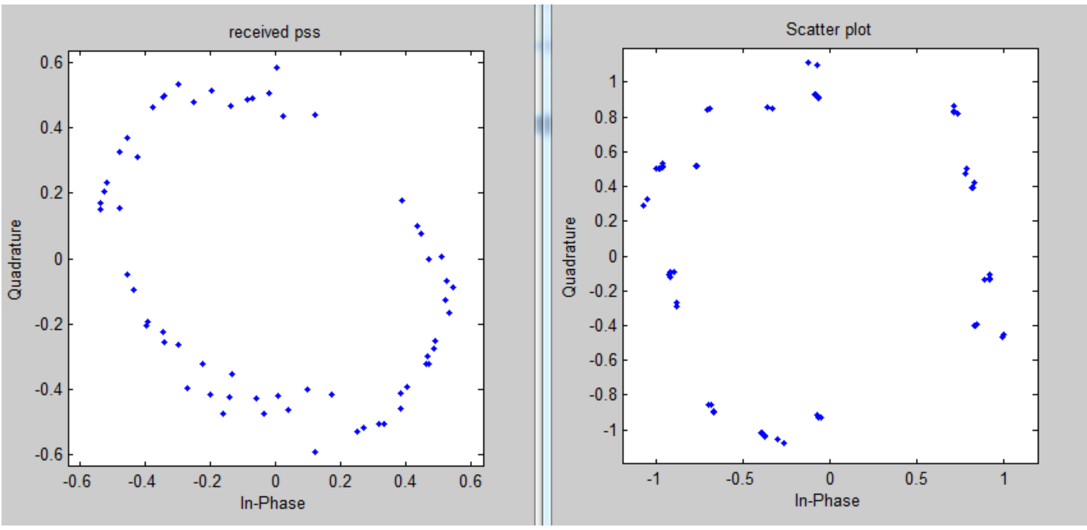


Figure 6.12 IQ imbalance correction in the presence of channel

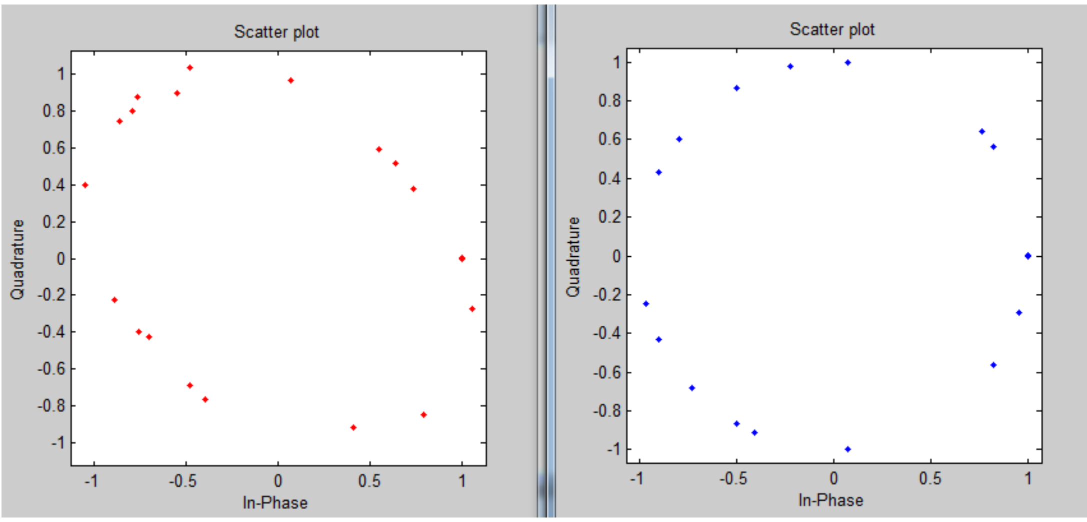


Figure 6.13 IQ imbalance correction

The above are constellation plots for Primary Synchronisation Signal before and After correcting

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