Time-Frequency Synchronization, Channel Estimation and Equalization of DL Channels in 3GPP Long Term Evolution
Introduction

3GPP LTE is a standard for wireless data communications technology and an evolution of the GSM/UMTS standards. The goal of LTE was to increase the capacity and speed of wireless data networks using new digital signal processing and modulation techniques, special attention has been given in selecting technologies for LTE DL. Technologies such as orthogonal frequency division multiplexing (OFDM) and multiple input, multiple output (MIMO), can enhance the performance of the current wireless communication systems. The high data rates and the high capacity can be attained by combining the advantages of the two technologies. These technologies have been selected for LTE DL and UL transmission. The received signal by user equipment (UE) undergoes many distortions and impairments due to the channel response, multipath delay, additive noise and RF front end mismatches. The receiver had to compensate the effects and retrieve the signal transmitted by the eNodeB (LTE Base Station). In order to aid the compensation of distortions the DL signals are transmitted with pilot tones termed as reference signals (RS). In addition to known RS, the inherent cyclic properties of OFDM signal and the CP (cyclic prefix) are used achieve the synchronization in time and frequency so that the loss of orthogonality among the subcarriers is restored. This project report aims at synchronization, channel estimation and equalization of PBCH (physical broadcast channel) signals of LTE. Synchronization techniques such as ML Estimation, CP correlated techniques are discussed. Channel estimation algorithms such as Least Squares (LS), Minimum Mean Square Error (MMSE) and Equalization algorithms ZF and MMSE, have been derived. Final equalized output was compared to with initial transmitted modulation. All the simulations are conducted using 4G LTE test-bed from Mymo Wireless where RF signals are captured over the wireless multipath propagation medium and suitable algorithms are used for the synchronization, channel estimation and equalization.
1. INTRODUCTION

Figure 1 Block diagram of a typical pilot-aided OFDM system

A brief block diagram of the OFDM system is shown in Figure 1. In an OFDM system, the available spectrum is divided into multiple subcarriers which are orthogonal to each other. Each of these subcarriers is independently modulated by a low rate QAM data stream. OFDM symbols are generated from a stream of serial QAM symbols. The N parallel streams are treated as samples in frequency domain, finally the N-point time domain blocks are obtained from the IFFT, which are subsequently serialised to create a time domain signal. OFDM has several benefits including its robustness against multipath fading and its efficient receiver architecture. Data symbols are independently modulated and transmitted over closely spaced orthogonal subcarriers. In LTE, modulation coding schemes (MCS) 4QAM, 16QAM, and 64QAM are used for DL and UL transmissions. In the time domain, a guard interval cyclic prefix (CP) is inserted prior to each OFDM symbol. As a fundamental principle of LTE, the data channels are shared channels, i.e. for each transmission time interval of 1 ms, a new scheduling decision is taken for the assignment of time-frequency resources during for DL and UL.

Transmitter:

- A serial complex QAM samples is converted to parallel data aligned as per the subcarrier index number for mapping onto the OFDM symbol subcarriers. See Figure 1.
- The IDFT block transforms the frequency domain data, $X_k$, on the $k^{th}$ subcarrier into time-domain samples $x_n$ as
  $$x_n = \text{IDFT} \{ X_k \} ; \quad n = 0,1,\ldots,N-1,$$
  where $N$ is total number of subcarriers
- Guard interval is added as below to remove the ISI
  $$x_n = x_{(n+g)} , \quad n = -N_g, -N_g + 1,\ldots, -1$$
  $$N_g = \text{Number of cyclic prefix samples prefixed to OFDM symbol for coping with time-domain dispersion of the channel}$$
- Parallel IFFT transformed samples are converted to serial data and fed to DAC for upsampling followed by RF upconversion.
Channel:

- Due to the channel impulse response $h$ which is a multipath fading channel and
- The additive noise $w$ which contaminates the transmitted signal

The received signal becomes $y_n = x_n \otimes h_n + w_n$.

Receiver:

- The receive data from the RF is fed to ADC. See Figure 2
- The synchronization is done by the receiver to make it free from the synchronization errors such as symbol time offset, CFO and sampling clock offset.
- The FFT is performed for extracting the data subcarriers onto which the QAM data was mapped followed by channel-estimation by using the reference symbols and then QAM data is extracted by channel equalization.

$$Y_k = FFT\{y_n\}, k = 0,1,...,N-1$$
$$Y_k = X_kH_k + I_k + W_k, k = 0,1,...,N-1,$$

where $H_k$ is the channel response, $I_k$ is the inter-carrier interference, $W_k$ is the additive Gaussian noise.

![Figure 2 OFDM receiver block diagram](image)

In this report we deal with the 3GPP LTE DL receiver processing which describes about

- LTE DL transmission scheme
- DL OFDMA (orthogonal frequency division multiple access) Frame Structure
- DL data transmission: Physical cell-search using primary and secondary synchronization signals combined with time and frequency synchronization
- Time-synchronization followed by carrier frequency offset (CFO) estimation and compensation of OFDM symbols
- Channel Estimation: DL reference signals used for channel estimation, channel interpolation and equalization
2. LTE DL Transmission Scheme

In the following sections the LTE DL receiver processing for 20MHz bandwidth is described. The TDD (Type-2 frame structure) mode of transmission, as shown in Figure 3, is considered.

2.1 OFDMA parameterization

DL and UL transmissions are organized into radio frames (also called as system frame number (SFN)) with \( T_f = 307200 \times T_s = 10 \text{ ms} \) frame duration, where \( T_s \) is the sampling time interval. Each frame (SFN) consists of 10 subframes, each subframe is of length \( T_{\text{slot}} = 30720 \times T_s = 1 \text{ ms} \), each subframe consists of two slots, each slot is of length \( T_{\text{slot}} = 15360 \times T_s = 0.5 \text{ ms} \). Each slot consists of 7 OFDM symbols where the 1st OFDM symbol has a length of 2208 samples (FFT length 2048 + CP length 160), 2nd to 7th OFDM symbols have each of length 2192 samples (FFT length 2048 + CP length 144).

The supported UL-DL configurations are listed in Table below where, for each subframe in a radio frame, “D” denotes the subframe is reserved for DL transmissions, “U” denotes the subframe is reserved for UL transmissions and “S” denotes a special subframe with the three fields DwPTS, GP and UpPTS.

The length of DwPTS and UpPTS is given by Table of special subframe below, subject to the total length of DwPTS, GP and UpPTS being equal to \( 10720 \times T_s = 1 \text{ ms} \). Each subframe \( i \) is defined as two slots, \( 2i \) and \( 2i+1 \) of length \( T_{\text{slot}} = 15360 \times T_s = 0.5 \text{ ms} \) in each subframe.

UL-DL configurations with both 5 ms and 10 ms DL-to-UL switch-point periodicity are supported, see Table 1, Table 2 and Table 3. In case of 5 ms DL-to-UL switch-point periodicity, the special subframe exists in both half-frames. In case of 10 ms DL-to-UL switch-point periodicity, the special subframe exists in the first half-frame only.

Two cyclic prefix lengths are possible, depending on the delay dispersion characteristics of the cell. The longer cyclic prefix (16.67µs) should then target multi-cell broadcast MBMS (Multimedia Broadcast Multicast Services) and very-large-cell scenarios, for instance, for rural and low data rate applications at a price of bandwidth efficiency. The number of OFDM symbols per slot depends on the size of this cyclic prefix, which is configured by the upper layers.

- PSS is always transmitted in the 3rd OFDM symbol of DwPTS (subframes #1 and #6)
- SSS is always transmitted in the last OFDM symbol in slots 1 and 11 (subframe #0 and subframe #5)
- PDCCH (Physical Downlink Control Channel) in DwPTS (subframes #1 and #6) may span 1 or 2 OFDM symbols
- Data is transmitted after the control region as in other DL subframes
- In DwPTS the cell specific RS patterns are the same as in other DL subframes
- The reference signal REs in GP are muted
- SRS(Sounding Reference Signal) is transmitted on UpPTS
Seven UL-DL configurations with either 5 ms or 10 ms DL to UL switch-point periodicity are supported. In case of 5 ms switch-point periodicity, the special subframe exists in both half-frames. In case of 10 ms switch-point periodicity the special subframe exists in the first half frame only. Subframes 0 and 5 and DwPTS are always reserved for DL transmission. UpPTS and the subframe immediately following the special subframe are always reserved for UL transmission.

Table 1 Configuration of special subframe (lengths of DwPTS/GP/UpPTS in Samples)
Transmitted signal in each slot is described by a resource grid of subcarriers and available OFDM symbols, see Figure 4. Each element in the resource grid is called a resource element (RE) and each resource element corresponds to one complex-valued modulation symbol (subcarrier). The REs have a constant spacing of $f_{sc} = 15$ kHz. In both DL and UL, a basic

### 2.1.1 Resource Grid
scheduling unit is denoted by a resource block (RB). A RB is defined as 12 consecutive REs (180 kHz) in the frequency domain in one subframe. The number of resource blocks for different LTE bandwidths is given in Table 3. The number of OFDM symbols per subframe is 14 for normal cyclic prefix and 12 for extended cyclic prefix. Each OFDM symbol is appended with a cyclic prefix (CP), compare. Each subframe has 2 slots (even and odd), therefore, a slot consists of 7 OFDM symbols for normal CP and 6 OFDM symbols for extended CP. Each OFDM symbol consists of IFFT part and the CP part which is prefixed to IFFT part, see Table 4 for CP and number of samples for the 20MHz transmission bandwidth. Note that the subcarrier spacing of 15 Khz and the IFFT length (2048) remain same irrespective of the normal CP or extended CP, only the length of cyclic prefix changes, see Table 5. Figure 4 shows the placement of OFDM symbols, REs and the corresponding bandwidths.

Table 4
<table>
<thead>
<tr>
<th>Channel Bandwidth (MHz)</th>
<th>1.4</th>
<th>3</th>
<th>5</th>
<th>10</th>
<th>15</th>
<th>20</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transmission Bandwidth (N_{RB})</td>
<td>6</td>
<td>15</td>
<td>25</td>
<td>50</td>
<td>75</td>
<td>100</td>
</tr>
</tbody>
</table>

Table 5

<table>
<thead>
<tr>
<th>Configuration</th>
<th>OFDM Symbols</th>
<th>CP</th>
<th>CP duration</th>
</tr>
</thead>
<tbody>
<tr>
<td>Normal CP; Δf=15 kHz</td>
<td>7</td>
<td>160 for 1st symbol 144 for others</td>
<td>5.2 μs for first symbol 4.7 μs for other symbols</td>
</tr>
<tr>
<td>Extended CP; Δf=15 kHz</td>
<td>6</td>
<td>512</td>
<td>16.7 μs</td>
</tr>
</tbody>
</table>

2.2 DL data transmission

The DL scheduling or UL grants allocation is done by eNodeB in terms of RBs, i.e. one UE can be allocated integer multiple of RBs for uplink or downlink transmissions. The RB allocations can be contiguous or distributed. The DL scheduling or UL grants by eNodeB to a UE is done on a subframe basis, that is 1 ms duration. The allocation of RBs by the eNodeB to a UE for DL or UL is dependent on a number of factors such as the radio link quality measured in terms of channel quality indicator (CQI), RI (rank indicator) and PMI (precoding matrix indicator) and signal to interference and noise ratio (SINR). The DL and UL the transmissions are split into 3 categories, e.g. synchronization, control and data channels. The list of DL channels are given below.

— Synchronization Signals/Channels

  o Primary, Secondary Synchronization Signals (PSS, SSS) and Reference Signals (RS)

  o Physical Broadcast Channel, PBCH. The channels once detected can be used for time-frequency synchronization apart from the synchronization of system frame number being transmitted by eNodeB.

— Control Channels

  o Physical Control Format Indicator Channel, PCFICH.

  o Physical Hybrid ARQ Indicator Channel, PHICH

  o Physical DL Control Channel, PDCCH

— Data Channels

  o Physical DL Shared Channel, PDSCH

  o Physical Multicast Channel, PMCH

PSS and SSS are used for estimating the Sector ID (NSID) and Group ID (NGID) respectively, where \[\text{NSID} + 3 \times \text{NGID}\]. In the case of TDD PSS is transmitted on 62 subcarriers within 72 reserved subcarriers around DC subcarrier in the last OFDM symbol of subframe-0 and subframe-5 of every Frame (10ms). Similarly SSS is transmitted on 62.
subcarriers within 72 reserved subcarriers around DC subcarrier in the 3rd OFDM symbol of subframe-1 and subframe-6 of every Frame.

The RS signals are different for each cell-ID and their positioning in the frequency-domain is dependent on the cell-ID, their positioning in time-domain is dependent on the number of antennas. For example, as shown in Figure 5 and Figure 6 for 1x1, 2x1, 2x2 and 4x4 the RS occupy the selective REs only in the 1st OFDM symbol of each slot, for 4x4 MIMO the first 2 OFDM symbols are occupied. The RS denoted by R_0, R_1, R_2 and R_3 shown in Figure 5 and Figure 6. The UE estimates the Channel Impulse Response (CIR) from each transmitting antenna, therefore, when a RS is transmitted from one antenna port, the other antenna ports in the cell are idle (null), see the colour coding in the Figure 5 and Figure 6. The CIR estimates for REs that do not carry the RS are computed via time-frequency interpolation. The DL reference signal structure is important for channel estimation, it may be observed that there are 2 RSs per slot in the time domain 4 RSs per slot which results in a total of 8 per RB. The required spacing in the time domain between the reference symbols can be obtained by considering the maximum Doppler spread (highest speed) to be supported, for example maximum mobile speed to be supported is ~500Kmph. Consider the RF carrier frequency is 2.1 GHz.

Doppler shift \( f_d = f_c \frac{v}{c} \)
\[ = 2.1 \text{GHz} \times \frac{500}{18} \times \frac{1}{3 \times 10^8} = 975 \text{ Hz} \]

According to Nyquist sampling theorem minimum sampling frequency needed to reconstruct signal is \( T_c = \frac{1}{2f_d} \approx 0.5 \text{ ms} \). Therefore, 2 reference symbols per slot are needed in the time domain in order to estimate the channel correctly.
Figure 5 Mapping of DL reference signals (normal cyclic prefix)
In the following sections we focus mainly on achieving the time-frequency synchronization, channel estimation and equalization for retrieving the Master Information Block (MIB) information from PBCH using the LTE Test-bed shown in Figure 12. The receiver signal processing chain described for PBCH in the following sections remains same for control and data channels except that the modulation coding scheme (MCS) for PBCH is QAM-4 whereas for data channels like PDSCH it can be QAM-4 or QAM-16 or QAM-64, the selection of MCS is decided by eNodeB based on the channel quality.

3. Cell Search and Synchronization

We consider the TDD UL-DL Configuration Index 1 for the discussion in the following sections. The position of the PSS, SSS and PBCH for TDD is shown in Figure 7. The Figure 8 shows the sequence of processing steps for PSS, SSS and then PBCH.

3.1 Primary Synchronization Signal (PSS)

Upon the UE power-on or while UE searching for a neighbouring cell information it first looks for the PSS and estimates the slot boundary (with ambiguity in subframe number) and the sector-ID (N_SID). Since PSS is transmitted twice in each Frame and both the transmissions are identical there is an ambiguity on the boundary whether start of Frame or middle of Frame.
3.2 Secondary Synchronization Signal (SSS)

The Frame boundary ambiguity is resolved by the detection of SSS as the transmission of SSS in first half of the Frame is different from the second half of the Frame. Detection of the SSS gives Frame boundary timing estimation and the group-ID ($N_{GID}$).

3.3 Cell-ID Estimation

The Cell ID is estimated by the formula ($N_{SID} + 3.N_{GID}$). LTE uses a hierarchical cell-search procedure in which an LTE radio cell is identified by a cell identity. There are 504 available physical layer cell identities which are divided into 3 groups (0, 1, 2) with each group having 168 layer identities (0, 1, 2,…, 167).

3.4 Physical Broadcasting Channel (PBCH)

As additional help during cell search, a PBCH is available which carries the MIB with basic physical layer information like system bandwidth, number of transmit antennas, and system frame number and PHICH (duration extended or normal). A 24-bit MIB bit-pattern is fed from higher layers to physical layer, the 24 bits plus the 16 bits CRC become 40 bits, the 40 bits are encoded by 1/3 rate Convolutional Encoder which gives 120 bits output. The 120 bits are repeated for 16 times resulting into a total number of bits as 1920 bits. The 1920 bits are QAM-4 modulated resulting into 960 QAM-4 samples. The 960 QAM-4 samples are divided into 4 parts each part is of 240 QAM-4 samples. Each part (240 QAM samples) is mapped to 4 OFDM symbols of each Frame.

The PBCH is spread across 4 Frames (40 ms), in each Frame a part of the PBCH is transmitted within 72 subcarriers of the 4 OFDM symbols centred around DC. Note that the decoding of MIB information from PBCH is achieved by processing any one of the Frames.

![Figure 7 PSS and SSS and PBCH signals positioning w.r.t. OFDM symbols, subframes and Frame for TDD mode.](image-url)
Successful execution of the cell search and selection procedure as well as acquiring initial system information is essential for a UE before taking further steps to communicate with the LTE network. Acquisition of IQ samples into DL receiver buffer and the sequence of steps during the cell-search procedure are shown in Figure. Once the cell-search is completed followed by the detection of cell system information the UE switches into ‘camp-on’ mode for acquiring the network identifier to associate with the eNodeB and join the network, Figure 9.

Figure 8 DL synchronization and PBCH processing flow

Figure 9 Synchronization and cell information acquisition until the camping state of UE.
4. Synchronization

OFDM offers many advantages in terms of robustness to ISI, multipath fading, simple method of channel estimation, equalization and spectrum efficiency. However the benefits come only if the orthogonality of the subcarriers in the OFDM symbol is preserved. The following time and frequency synchronization is critical for achieving a good OFDM receiver design.

**Synchronization error due to symbol time offset (SFO):** Timing offset problem is due to the unknown OFDM symbol arrival time. Symbol arrival timing may be different than that of the transmitted sequence due to multipath propagation and channel delay spread. It is necessary that the FFT windows are accurately picked and ISI is eliminated by discarding the CP.

**Synchronization error due to carrier frequency offset (CFO):** Frequency offset arises primarily due to carrier frequency mismatch between the transmitter and the receiver and also as a result of Doppler shift due to mobility of transmitter and/or receiver. The impact of the CFO is the loss of orthogonality between subcarriers that results into loss of data. It is necessary that CFO is sufficiently accurately estimated, compensated and tracked. The CFO may exceed the subcarrier spacing resulting into integer part of CFO (IFO) and fractional part of CFO (FFO). IFO and FFO are both to be estimated using different techniques and compensate the combined total CFO in the received OFDM symbol.

**Synchronization error due to sample clock frequency (SCO) mismatch:** It is also necessary to maintain OFDM synchronization in terms of the sampling clocks between DAC of transmitter and ADC of receiver. If the clock synchronisation is not accurate, relative sampling rate will be faster or slower which results into loss or addition of samples, the error can be compensated by interpolating additional sample or deleting the extra sample for maintaining the orthogonal property of the OFDM symbol.

In the report we look in to the estimation of STO, CFO followed by SCO. We use the ML method of joint time and CFO estimation and compensation.

4.1 Time and Frequency Offset Estimation in OFDM Systems using ML method

A symbol clock and a frequency offset estimates may be generated at the receiver with the aid of pilot symbols known to the receiver or by maximizing the average log-likelihood function. ML estimation uses the cyclic prefix preceding the OFDM symbols, thus reducing the need for pilots.
Figure 10 OFDM system, transmitting subsequent blocks of N complex data

Let $s_n$ be QAM signal which is mapped onto the subcarriers, followed by IFFT and CP addition. The resulting OFDM signal $x_n$ is serially transmitted as shown in Figure 10. The transmitted signal undergoes the multipath propagation before it arrives at receiver along with addition of i.i.d. white Gaussian noise $w_n$. The first uncertainty is seen as a delay in the channel impulse response $\delta(n-\theta)$ where $\theta$ is the integer valued unknown arrival time of a symbol. The latter is the CFO which is modeled as a complex multiplicative distortion of the received data in the time domain $e^{j2\pi \varepsilon/N}$, $\varepsilon$ denotes the difference in the transmitter and receiver oscillators as a fraction of the intercarrier spacing $(1/N)$ in normalized frequency. Based on these two uncertainties and the additive Gaussian noise the received signal is written as,

$$r(n) = s(n-\theta)e^{j2\pi \varepsilon/N} + w(n)$$

OFDM symbol has $N + N_{CP}$ samples where the last $N_{CP}$ samples are placed in the starting of the symbol in the same order. So, when we take a window size of $N_{CP}$ in the starting and the ending of the symbol the two window sizes will be identical. The ML estimation uses the identical nature of the two windows and finds the timing and frequency offsets. Let us take the $2N + N_{CP}$ consecutive samples of $r(n)$, and that these samples contain one complete OFDM symbol of $N + N_{CP}$ samples. The position of this symbol within the observed block of samples is unknown because the channel delay is unknown to the receiver. Apply the auto-correlation of size $N_{CP}$ with a distance between the windows as $N$. Then the auto-correlation sum gets maximized only when the two windows match which indicates that $\theta$ is the starting sample of the OFDM symbol. Using this property $\theta$ STO can be determined.

$$E[r(n)r^*(n+m)] = \sigma_x^2 + \sigma_w^2 \text{ if } m = 0$$
$$= \sigma_i^2 e^{j2\pi m/N} \text{ if } m = N$$

Mymo Wireless Technology Pvt Ltd, www.mymowireless.com
The log-likelihood function for $\theta$ and $\epsilon$ is the logarithm $\Lambda(\theta, \epsilon)$ of the probability density function $f(r \mid \theta, \epsilon)$ of the $2N + N_{CP}$ received samples given the arrival time $\theta$ and $\epsilon$. Under the assumption that $r(n)$ is a jointly Gaussian vector then the log-likelihood function becomes

$$ \Lambda(\theta, \epsilon) = \left| \gamma(\theta) \right| \cos(2\pi \epsilon + \angle(\theta)) - \rho \phi(\theta). $$

Where

$$ \gamma(m) = \sum_{n=m}^{m+N_{CP}-1} r(n) r(n+N) $$

$$ \phi(m) = \frac{1}{2} \sum_{n=m}^{m+N_{CP}-1} |r(n)|^2 + |r(n+N)|^2 $$

$$ \rho = \left| \frac{E[r(n)r^*(n+N)]}{\sqrt{E[|r(n)|^2]E[|r(n+N)|^2]}} \right| = \frac{\sigma_s^2}{\sigma_s^2 + \sigma_n^2} = \frac{SNR}{SNR + 1} $$

From the joint time-frequency ML of $\Lambda(\theta, \epsilon)$, the estimates of $\theta$ and $\epsilon$ (refer[7]) are obtained as,

$$ \theta_{ML} = \arg \max_\theta \{ \left| \gamma(\theta) \right| - \rho \phi(\theta) \} $$

$$ \epsilon_{ML}(\theta) = -\frac{1}{2\pi} \angle(\theta_{ML}) $$

The ML estimation method above has the following issues, first issue is computationally very complex to do a 2-dimensional search (time and frequency) over the complete block of every OFDM symbol, 2nd issue is ambiguity in threshold setting for detecting the correlation peak. The recommended method is to first search in time domain for normalized correlation $\rho$ with a threshold setting $|\rho| \geq 0.9$, where $\max(|\rho|) = 1$, that indicates the presence of the OFDM symbol and the 2-dimensional search around the threshold can be initiated. Around the $|\rho| \geq 0.9$ use the ML method of search for detecting the $\hat{\theta}_{ML}$ and $\hat{\epsilon}_{ML}$. Note that the method gives the estimate of fractional CFO (FFO). After compensating the FFO the residual CFO can be estimated using the RS signals in frequency domain and will be compensated.

### 4.2 SCO Estimation

Sampling time difference $\zeta$ due to the mismatch between DAC and ADC clocks of transmitter and receiver respectively.

$$ \zeta = \frac{T_s(Tx) - T_s(Rx)}{T_s(Rx)}. $$

In the case of $\zeta = 0$, the separation between two consecutive $\hat{\theta}_{ML}$ estimations is $N + N_{CP}$ samples, that is, a $\hat{\theta}_{ML}$ estimation is obtained from every OFDM symbol. However, if $\zeta \neq 0$, the effect of the window drift is directly translated to the metrics, leading the separation
between correlation peaks to be \((N + N_{CP} + \zeta)\) samples. The calculation of the ML Estimation by J. J. van de Beek[7] metrics provides an estimation of the window drift, by comparing the \(\hat{\theta}_{ML}^{n}\) estimation for the nth symbol with the expected position of the estimation in the ideal case, that is \(\hat{\theta}_{ML}^{exp} = \hat{\theta}_{ML}^{n-1} + (N + N_{CP})\). In order to avoid deviations caused by instantaneous degradations on the calculated metrics, a filtering is required for tracking the variations due to \(\zeta\). For this reason, it is considered a maximum deviation of \(\pm 1\) sample between two consecutive symbols.

It is possible to have an estimation of \(\zeta\), by averaging the corrections performed over the symbol window. This estimation is obtained by averaging the value of \(\text{sign}(\hat{\theta}_{ML}^{n} - \hat{\theta}_{ML}^{exp})\), over the number of received symbols, the quotient between this average and the number of received samples gives an estimation of the stability of the sampling clocks.

5. CHANNEL ESTIMATION

Channel Estimation provides information about the channel delay spread in terms of the channel coefficients. The delay spread can be measured by taking the IFFT of the channel coefficients which are obtained on per-tone basis from the knowledge of received RS REs and known RS REs. This information is then used by equalizers so that the fading effect and co-channel interference can be removed and the original transmitted signal can be restored. Channel Estimation plays an important role in a communication receiver. In order to mitigate hostile channel effects on the received signal, precise channel estimation is required to provide information for further processing of the received signal.

**Channel Estimators** can be categorized into two types:

1) Non- Data-Aided (or Blind)
2) Data-Aided

**Non-Data-Aided:**

A Non-Data-Aided or blind channel estimator estimates the channel responses by statistics of the received signals. No specialized reference (training) signals are needed and the transmission efficiency is retained for systems using this type of estimators. Since the transmitted signals are not known to the receiver in this type, a large number of data must be collected in order to obtain reliable estimation.

**Data-Aided:**

Data-Aided channel estimators require known reference signals (RS) to be transmitted. Channel estimation can be achieved by comparing the received and transmitted reference or pilot signals. A sufficient number of such reference signals must be inserted according to the degree of channel variation, namely coherence time and coherence bandwidth of the channel under estimation. OFDM based communication standards, provide some forms of reference
signals, namely preamble or pilot signals. This paper will focus mainly on the data-aided channel estimation algorithms for OFDM communications.

5.1 PILOT BASED CHANNEL ESTIMATION

The least-square (LS) and minimum-mean-square-error (MMSE) techniques are widely used for channel estimation when training symbols are available. We assume that all subcarriers are orthogonal (i.e., ICI-free). Then, the training symbols for N subcarriers can be represented by the following diagonal matrix:

\[
X = \begin{bmatrix}
X[0] & 0 & \cdots & 0 \\
0 & X[1] & \vdots & \\
\vdots & \ddots & \ddots & 0 \\
0 & \cdots & 0 & X[N-1]
\end{bmatrix}
\]

Where \(X[k]\) denotes a pilot tone at the \(k^{th}\) subcarrier, with \(E\{X[k]\} = 0\) and \(\text{Var}\{X[k]\} = \sigma^2_x\) for \(k=0,1,2,\ldots,N-1\). Note that \(X\) is given by a diagonal matrix, since we assume that all subcarriers are orthogonal. Given that the channel gain is \(H[k]\) for each subcarrier \(k\), the received training signal \(Y[k]\) can be represented as

\[
Y = XH + Z
\]

Where \(H\) is a channel vector as \(H=[H[0],H[1],\ldots,H[N-1]]^T\) and \(Z\) is a noise vector given as \(Z=[Z[0],Z[1],\ldots,Z[N-1]]^T\) with \(E\{Z[k]\} = 0\) and \(\text{Var}\{Z[k]\} = \sigma^2_z\) for \(k=0,1,2,\ldots,N-1\). The \(\hat{H}\) denotes the estimate of channel \(H\).

5.1.1 Least Square Estimation

The least-square (LS) channel estimation method finds the channel estimate \(\hat{H}\) in such a way that the following cost function is minimized:

\[
F(\hat{H}) = \left\|Y - X\hat{H}\right\|^2 = (Y - X\hat{H})^H(Y - X\hat{H}) = (H^H\hat{H}^HX)(Y - X\hat{H}) = (H^HY - \hat{H}^HX\hat{H} + \hat{H}^HX\hat{H})
\]

By setting the derivative of the function with respect to \(\hat{H}\) to zero,
\[
\frac{\partial F}{\partial H} = -(Y^H X)^H X^H Y + (\hat{H}^H X^H X)^H + X^H X \hat{H} \\
= -X^H Y X^H Y + X^H X \hat{H} + X^H X \hat{H} \\
= -2X^H Y + 2X^H X \hat{H} \\
= 0
\]

Implies

\[
X^H X \hat{H} = X^H Y
\]

\[
\hat{H} = (X^H X)^{-1} X^H Y \\
= X^{-1} (X^H)^{-1} X^H Y \\
= X^{-1} Y
\]

Therefore

\[
\hat{H}_{LS} = X^{-1} Y
\]

The mean-square error (MSE) of this LS channel estimate is given as:

\[
\text{MSE}_{LS} = \mathbb{E}\{ \|H - \hat{H}\|^2 \} \\
= \mathbb{E}\{(H - \hat{H})^H (H - \hat{H})\} \\
= \mathbb{E}\{(H - X^{-1} Y)^H (H - X^{-1} Y)\} \\
= \mathbb{E}\{(X^{-1} Z)^H (X^{-1} Z)\} \\
= \mathbb{E}\{Z^H (X^{-1})^H Z\} \\
= \mathbb{E}\{Z^H (X^H X)^{-1} Z\} \\
= \frac{\sigma_e^2}{\sigma_X^2}
\]

The above mean-square equation is inversly proportional to the SNR, which implies that it may be subject to noise enhancement especially when the channel is in a deep null.

### 5.1.2 MMSE Channel Estimation

The MMSE channel estimation finds a better estimate in terms of a weighted matrix in such a way that the Mean Square Error is minimized when compared to LS estimation. The received signal was given by \(Y = XH + Z\). We have seen from the Least Square Estimation, the estimated channel was given as \(\hat{H}_{LS} = X^{-1} Y\).

Let \(\hat{H}_{LS} = \hat{H}\). Let us denote the MMSE Estimate of a channel as \(\hat{H}\) obtained by passing the \(\hat{H}\) through the weighted matrix \(W\), i.e. \(\hat{H} = W \hat{H}\) the actual channel being \(H\). The error is given as \(e = H - \hat{H}\). The MSE of the channel estimate \(\hat{H}\) is given as

\[
J(\hat{H}) = \mathbb{E}\{e^2\} = \mathbb{E}\{\|H - \hat{H}\|^2\}
\]
The Orthogonality principle states that the estimation vector $e = H - \hat{H}$ is orthogonal to $H$. Such that

$$E\{eH^H\} = E\{(H-\hat{H})H^H\} = E\{HH^H - \hat{H}\hat{H}^H\} = E\{HH^H\} - E\{\hat{H}\hat{H}^H\} = E\{HH^H\} - E\{WHH\} = E\{HH^H\} - WE\{HH^H\} = 0$$

$$R_{HH} - WR_{HH} = 0$$

$$W = R_{HH}^{-1}(R_{HH})^{-1}$$

$R_{HH}$ is the cross-correlation matrix between the true channel and temporary channel estimate vector in the frequency domain. Using the above equation the MMSE channel estimate is given as

$$\hat{H} = WH$$

$$= R_{HH}^{-1}(R_{HH})^{-1}\tilde{H}$$

$$= R_{HH}^{-1}(R_{HH} + \frac{\sigma^2}{\sigma^2_x})^{-1}\tilde{H}$$

The elements of the $R_{HH}$ and $R_{HH}$ in the above equation are given as

$$E\{h_k,\tilde{h}_{k,j}\} = E\{h_k,h_{k,j}^*\} = r_{f} [k - k] r_{f} [l - l]$$
Where \( k \) and \( l \) denote the subcarrier (frequency) index and OFDM symbol (time) index, respectively. In an exponentially-decreasing multipath PDP (Power Delay Profile), the Frequency-domain correlation \( r_f[k] \) is given as

\[
r_f[k] = \frac{1}{(1 + j2\pi r_{ms} k\Delta f)}
\]

Where \( \Delta f = 1/T_{sub} \) is the subcarrier spacing for the FFT interval length of \( T_{sub} \). Meanwhile, for a fading channel with the maximum Doppler frequency \( f_{max} \) and Jake’s spectrum, the time-domain correlation \( r_f[l] \) is given as

\[
r_f[l] = J_0(2\pi f_{max} l T_{sym})
\]

Where \( T_{sym} = T_{sub} + T_G \) for guard interval time of \( T_G \) and \( J_0(x) \) is the first kind of 0th-order Bessel function. Note that \( r_f[0] = J_0[0] = 1 \), implying that the time-domain correlation for the same OFDM symbol is unity.

6. INTERPOLATION

At the receiver, the channel complex gain at the pilot symbol positions can be easily obtained from the received signal and the known pilot symbols. To estimate the channel for data symbols, the pilot subcarriers must be interpolated. Interpolation is then applied to derive the estimation of the channel knowledge at data symbol (non-pilot subcarriers) positions. There are many interpolators such as Wiener filter, minimum mean-square error (MMSE), Linear interpolation, second-order polynomial interpolation, spline, transform domain interpolation and low pass interpolation. Polynomial-based interpolators have been popular in the frequency-domain interpolation algorithms even though there are many better interpolators due to their low implementation complexity.

6.1 Linear Interpolation

Linear interpolation has been proposed to estimate the frequency-domain channel responses at data subcarriers. For the \( k \)-th subcarrier to be interpolated, let \( k/D = m + \mu \), where \( 0 \leq \mu < 1 \), and \( m = \lfloor k/D \rfloor \), the largest integer smaller than \( k/D \). Then, the linear interpolation method obtains the channel response at the \( k \)-th subcarrier as

\[
\hat{H}_k = \hat{H}_{D(m\nu)} = (1-\mu)\hat{H}_{mD} + \mu\hat{H}_{(m+1)D}
\]

The estimation quality can be improved by using higher-order polynomials. However, the implementation grows more complicated as the order increases. A piecewise second-order polynomial interpolation is as follows

\[
\hat{H}_k = \hat{H}_{D(m\nu)} = C_0\hat{H}_{mD} + C_{-1}\hat{H}_{(m+1)D} + C_{-2}\hat{H}_{(m+2)D}
\]

where
\[C_0 = \frac{(1-\mu)(2-\mu)}{2}\]
\[C_{-1} = \mu(2-\mu)\]
\[C_{-2} = \frac{-\mu(1-\mu)}{2}\]

Other high-order polynomial-based interpolators such as the piecewise parabolic interpolator and the cubic interpolator take in four base points for interpolation:
\[\hat{H}_k = \hat{H}_{(m+\mu)} = C_1\hat{H}_{(m+1)} + C_0\hat{H}_{(m)} + C_{-1}\hat{H}_{(m-1)} + C_{-2}\hat{H}_{(m-2)}\]

In the piecewise parabolic interpolator, the coefficients are given by
\[C_1 = -\alpha\mu + \alpha\mu^2\]
\[C_0 = 1 + (\alpha - 1)\mu - \alpha\mu^2\]
\[C_{-1} = (\alpha + 1)\mu - \alpha\mu^2\]
\[C_{-2} = -\alpha\mu + \alpha\mu^2\]

Usually, \(\alpha\) is set to provide better interpolation quality. On the other hand, the coefficients of the cubic interpolator are
\[C_1 = -\frac{1}{3}\mu + \frac{1}{2}\mu^2 - \frac{1}{6}\mu^3\]
\[C_0 = 1 - \frac{1}{2}\mu - \frac{1}{2}\mu^2 + \frac{1}{6}\mu^3\]
\[C_{-1} = \mu + \frac{1}{2}\mu^2 - \frac{1}{6}\mu^3\]
\[C_{-2} = -\frac{1}{6}\mu + \frac{1}{6}\mu^3\]

### 6.2 Shifted Raised-Cosine Interpolation

Receiver performance will be poor if there is uncertainty in the timing of the received signal. In real time situations, due to energy leakage, their exist a pre-cursor as well as a post-cursor in the reconstructed CIR.

Their should be good time domain window that can preserve the major portion of the reconstructed CIR and, reject the aliased and noisy components. Their may be need to shift the time-domain window right instead of centering at the origin due to multipath time delay. Shiting the window in time domain is equivalent to rotating the phase of the interpolation coefficients in the frequency domain.

Where the CIR is strong the window should be flat so that no distortion is introduced, the two ends of the window can weigh smaller in order to supress the unwanted components such as
noisy and aliasing effect. So, smoother weighting in the time domain entails faster fall-off in the frequency domain interpolation coefficients.

Spline interpolator in which raised cosine function is used as the frequency domain interpolation coefficients, satisfies the above requirements.

\[
W_{l,r,c} = \frac{\sin(\frac{\pi M_l}{N})}{\frac{\pi M_l}{N}} \ast \frac{\cos(\frac{\pi \beta M_l}{N})}{1 - 4\beta^2(\frac{M_l}{N})^2} * e^{\frac{-j2\pi dl}{N}}
\]

\(\beta\) is the roll-off factor, which decides the excess width of the window’s main lobe. \(d\) controls the position of the window shifted to the right.

### 6.3 Two-Dimensional MMSE Interpolation

The time-varying channel frequency response is a wide sense stationary 2D random process. To get a perfect reconstruction is hardly possible because of random noise and CIR. Among the linear interpolators, 2D MMSE interpolator is the optimum in terms of the minimum mean-squared error criterion. Since in the practical scenario the Referenc( or pilot ) signals are inserted in two dimensionally, so there is a need to do the interpolation in both dimensional(time and frequency. 2D MMSE interpolator can smooth the noise in the channel estimates for the pilot subcarrier; filter estimates on pilots; generate the the CIR for data subcarriers inside the grid; and predict the channel response for data subcarriers outside the pilot grid. Two 1D MMSE interpolations also achieve the performance nearly equal to 2D MMSE interpolator but with little bit reduction in the complexity than 2D MMSE. But high complexity in the usage of MMSE interpolator prevents their application in practical OFDM receivers.

### 7. EQUALIZATION

Channel Estimation provides information about distortion of the transmission signal when it propagates through the channel. This information is then used by equalizers so that the fading effect and/or co-channel interference can be removed and the original transmitted signal can be restored. Once channel estimates at data subcarriers are derived, the receiver performs equalization to compensate for signal distortion.

#### 7.1 One-Tap Equalizer

OFDM systems are favoured over single-carrier modulations in that the simple one-tap frequency-domain equalizer(FDE) can equalize OFDM signals that go through frequency-selective fading channels. In channels whose impulse responses remain constant within one OFDM symbol period, the received signal at each subcarrier takes the form of

\[
Z_{i,k} = H_{i,k}X_{i,k} + N_{i,k}
\]

One-Tap Equalizers restore the transmitted signal by \(\hat{X}_{i,k} = G_{i,k}Z_{i,k} \cdot G_{i,k}\), \(G_{i,k}\) the equalizer coefficient at the \(k^{th}\) subcarrier during the \(i^{th}\) symbol.

Mymo Wireless Technology Pvt Ltd, www.mymowireless.com
7.1.1 Zero-Forcing Equalizer

Regardless of noise, the Zero-Forcing Equalizer simply uses the inverse of the channel response \( G_{ik} = H_{ik}^{-1} \) and forces the frequency-selective-faded signals back to flat faded ones. However it may result in noise enhancement in the subcarriers that suffer deep fading.

Consider the noise-free MIMO System that can be interpreted as a linear system of \( M_R \) equations (\( M_R \) received signals) with \( M_T \) unknowns (\( M_T \) i/p symbols).

\[
\tilde{y} = Hx
\]

Where dimensions are as follows:
- \( H = M_R * M_T \);
- \( x = M_T * 1 \);

In the ZF approach, this system is inverted to find the unknown i/p symbols

The ZF receiver is a \( M_T * M_R \) matrix denoted as \( F \). such that: \( F\tilde{y} = x \) (or) \( FH = I \)

The ZF receiver inverse the MIMO channel matrix.

The existence of the channel inversion is subjected to some conditions on the channel. 3 cases can be distinguished.
- \( H \) doesn’t have full column rank (e.g. when \( M_T > M_R \)). There are no unique solutions for the i/p symbols. A ZF receiver cannot be defined.
- \( H \) is a square and invertible. There exists a single ZF receiver \( F_{ZF} = H^{-1} \).
- \( H \) is tall (\( M_T <or= M_R \)) and has full column rank. The system is over determined. That mean several receivers exit satisfying \( FH = I \).

Those receivers can be written as \( (C^H H)^{-1} C^H \), where \( C \) is a \( M_T * M_R \) matrix with full column rank. When there is no additive noise, any of those receivers can be applied to recover the i/p symbols. When there is noise, an error will be made on the estimation of the symbols, the error depends on the coefficients of the receiver. So appropriate ZF receiver to be selected.

**MMSE-ZF Receiver:** ZF receiver that minimizes the mean squared estimation error.

Assume the channel has full column rank, their may exist many receivers that can eliminate the ISI.

Let the ZF receiver be \( F; FH = I \)

O/P of \( F \) is: \( \hat{x} = Fy = x + Fn \)

Estimation Error: \( \hat{x} - x = Fn \)

Applying MMSE to the estimated error.
\[
E_a \|\hat{x}-x\|^2 = E\|F_n\|^2 \\
= E\{(F_n)^H(F_n)\} \\
= E\{n^H F^H F_n\} \\
= E\{\text{tr}F_n n^H F^H\} \\
= \text{tr}E\{F_n n^H F^H\} \\
= \sigma_n^2 \text{tr}F^H F
\]

The ZF receiver minimizing the MSE is unique.

MMSE-ZF receiver or ZF equalizer is equal to \( F_{ZF} = (H^H H)^{-1} H^H \)

### 7.1.2 MMSE-Receiver

The purpose of the MMSE receiver is to minimize the average estimation error on the transmitted symbols. The average is taken over the transmitted symbols and the noise: the MSE is \( E_{x,n} \|\hat{x}-x\|^2 \). Even though the ZF receiver also minimizes the output MSE but the constraint of complete ISI elimination.

Let the inputs, outputs, noise and channel be defined as:

\[
y = [y_0 \ y_1 \ y_2 \ldots y_{(R-1)}]^T; \\
x = [x_0 \ x_1 \ x_2 \ldots x_{(T-1)}]^T; \\
n = [n_0 \ n_1 \ n_2 \ldots n_{(R-1)}]^T; \\
\text{and } H \text{ is } R \times T \text{ matrix}
\]

The output of a receiver \( F \) is of the form \( F y \) and the vector of errors is

\[
\hat{x}-x = FHx + F_n-x \\
= (FH-I)x + F_n
\]

Applying the MMSE condition:

\[
E\{\|\hat{x}-x\|^2\} = E\{\|(FH-I)x + F_n\|^2\} \\
= E\{\|(FH-I)x + F_n\|^H(FH-I)x + F_n\}\} \\
= E\{(X^H(FH-I) + n^H F^H)((FH-I)x + F_n)\} \\
= E\{(X^H(FH-I) + n^H F^H)(FH-I)x + E\{n^H F^H F_n\} \\
= P_x \text{tr}((FH-I)^H(FH-I)) + \sigma_n^2 \text{tr}(F^H F) \\
= P_x \text{tr}((H^H F^H - I)(FH-I)) + \sigma_n^2 \text{tr}(F^H F) \\
= P_x \text{tr}((H^H F^H FH - H^H F^H F H + I) + \sigma_n^2 \text{tr}(F^H F)
\]

Differentiating the above equation w.r.t \( F \) and equating it zero gives the MMSE receiver.
\[
\frac{\partial E}{\partial F} = P_x (F^H H^H + HH^H F^H + H - H) + \sigma_n^2 (F^H + F^H) \\
= P_x (F H^H H^H + HH^H F^H + H - H) + \sigma_n^2 (F^H + F^H) \\
= 2P_x (F H^H H^H) + 2 \sigma_n^2 F^H \\
= 0
\]

\[
P_x (F H^H H^H) + \sigma_n^2 F^H = 0
\]

\[
P_x H^H + \sigma_n^2 F = H^H
\]

\[
F (H^H + \sigma_n^2 I) = H^H
\]

\[
F = H^H (H^H + \sigma_n^2 I)^{-1}
\]

**Derivation for \( n H^H F^H n \) by using a 2*2 matrix for \( F \) and 2*1 for \( n \):**

\[
n H^H F^H n = \begin{bmatrix} n_1 & n_2 \end{bmatrix} \begin{bmatrix} F_{11} & F_{12} \\ F_{12} & F_{22} \end{bmatrix} \begin{bmatrix} n_1 \\ n_2 \end{bmatrix}
\]

\[
= \begin{bmatrix} n_1 & n_2 \end{bmatrix} \begin{bmatrix} |F_{11}|^2 + |F_{21}|^2 & F_{11}^* F_{12} + F_{21}^* F_{22} \\ F_{12}^* F_{11} + F_{22}^* F_{21} & |F_{12}|^2 + |F_{22}|^2 \end{bmatrix} \begin{bmatrix} n_1 \\ n_2 \end{bmatrix}
\]

\[
= \begin{bmatrix} n_1 & n_2 \end{bmatrix} \begin{bmatrix} n_1 (|F_{11}|^2 + |F_{21}|^2) + n_2 (F_{11}^* F_{12} + F_{21}^* F_{22}) \\ F_{12}^* F_{11} + F_{22}^* F_{21} + n_2 (|F_{12}|^2 + |F_{22}|^2) \end{bmatrix} \begin{bmatrix} n_1 \\ n_2 \end{bmatrix}
\]

\[
= \begin{bmatrix} n_1^2(|F_{11}|^2 + |F_{21}|^2) + n_1 n_2 (F_{11}^* F_{12} + F_{21}^* F_{22}) + n_2 (F_{11}^* F_{12} + F_{21}^* F_{22}) + n_2^2 (|F_{12}|^2 + |F_{22}|^2) \end{bmatrix}
\]

\[
E \{ n H^H F^H n \} = E \left\{ \begin{bmatrix} n_1^2(|F_{11}|^2 + |F_{21}|^2) + n_1 n_2 (F_{11}^* F_{12} + F_{21}^* F_{22}) + n_2 (F_{11}^* F_{12} + F_{21}^* F_{22}) + n_2^2 (|F_{12}|^2 + |F_{22}|^2) \end{bmatrix} \right\}
\]

\[
= \sigma_n^2 \left( |F_{11}|^2 + |F_{21}|^2 + |F_{12}|^2 + |F_{22}|^2 \right)
\]

\[
= \sigma_n^2 \text{tr}(F^H F)
\]

**8. Project summary**

Processing of received data received through 4G LTE Test-Bed (named as MW1000), and retrieving back the transmitted signal at PBCH slots. The processing of the LTE data was done which was transmitted and received in MW1000 test bed through (i) wired channel and (ii) wireless channel. The data transmission and receiving was done in real time and the test-bed which consists the eNodeB and UE. And their specifications meet the LTE standards.

The TDD frame structure of Normal cyclic prefix type is taken for the project and an Uplink-DL configuration of Type 1. In the test-bed simulation cases the channel is assumed to
remain constant for 1 ms, and it is enough to interpolate the channel in frequency domain and taking it as the constant for the symbols near to the actual estimated symbols.

**LTE Test-Bed Specifications:**

- **MW1000** - 3GPP LTE eNodeB and UE Test-bed with reprogrammable capability in C and Linux OS with RF front end integration
- **3GPP LTE** - Release 9
- **RF carrier frequency** - 2.42 GHz
- **The sampling frequency** - 30.72 MHz,
- **Transmission Bandwidth** - 20MHz
- **Number of Resource Blocks** - 100
- **FFT/IFFT size** - 2048
- **IQ Samples** - 16 bit I, 16 bit Q
- **Cyclic Prefix** - Normal
- **Duplexing Mode** - TDD (UL-DL Config Index-1)
- **Transmission Mode** - 2 (Transmit Diversity)
- **Transmission Medium** - RF, Wireless

**The MW1000 Test Bed**

The entire setup of the test bed is as shown below.

Figure 12 Mymo’s 3GPP LTE 2x2 MIMO Test-bed (MW1000)
Description of the above set up:

It is first of its kind for building and validating the LTE baseband and protocol designs rapidly in a real-time environment. The uniqueness of the test-bed is that the designers can design and implement and verify the algorithms, system design and Phy and protocol stack layers. The test-bench enables the design team to quickly build the complete system models in ANSI C and Linux OS platform and for validating the algorithms performance in real RF environment.

The 3GPP LTE UE and eNodeB layers are seamlessly integrated, and the baseband at IQ sampling rate 30.72MHz is interfaced with RF-Mixed-Signal card for operation at desired ISM RF band or LTE Band-37, 38 or 7. The design test-bench is built with a flexibility to access and modify any part of the signal processing or bit rate processing functional blocks of LTE UE or eNodeB written in ANSI C source code. The test-bed can be configured to radiate and capture the RF signals in a free-space multipath propagation environment or directly connecting eNodeB and UE RF ports through SMA cables and RF attenuators to avoid the free-space radiation. The TDD transmitted signal of UL-DL Configuration Index 1 for 1 SFN is shown in Figure 13.

![Figure 13 LTE TDD One Frame (10ms) Time Domain Signal Captured from Test-Bed](image)

Processing of the received Data:
The DL received IQ signal is captured for multiple frames (each frame 10ms) at 30.72 MHz sampling rate. The cell search by PSS and SSS followed by time-frequency synchronization is performed. The approximate signal window of PBCH in slot-1 of subframe 0 is processed for demonstrating the time-frequency synchronization, channel estimation and equalization results.

Retrieving the PBCH data

- The received data by UE has undergone many distortions due to the channel response, multipath delay, additive noise etc…
- The receiver had to compensate all the effects and retrieve the original signal back i.e. transmitted by the eNodeB.
- The effects that the signal and their affect was discussed above.
- Processing of the data involves
  - Synchronization
  - Extracting the PBCH slots
  - Channel Estimation
  - Interpolation
  - Equalization

These processes are described above and the brief notes about the steps followed in the project are:

- For synchronization the window start position was taken few samples before to avoid encroachment into the ISI part of the next symbol. The few samples taken before add upto symbol offset which will be corrected during the equalization.
- Frequency synchronization was done at multiple stages (in time domain and frequency domain) to get an effective compensation.
- For channel estimation the LS (least squares) estimation algorithm was used even though the MMSE estimator was the optimum because of the complexity involved in the usage of MMSE in real time.
- The assumption is that the channel is quasi-stationary for 1ms (one subframe). It is enough to estimate the channel in OFDM symbol-1 and symbol-5 (where RS signals are present) and interpolate in frequency domain.
- The equalization was done using the zero forcing equalizer.

The project results are as shown below:

The received OFDM IQ frequency domain signals with CFO uncorrected and unequalized is shown in Figure 14.
The above shown is the received signal IQ data of the symbol which contains PBCH, the actual transmitted data is 4QAM data. We can see the distortion caused by the channel dispersion and the CFO.

Figure 15 below shows the time and frequency plots during the ML Estimation of Time and Frequency offsets.
The above plot shows the ML Estimation of the Time and Frequency Offsets. The peak position in the time offset indicates the start position of the OFDM symbol. And the corresponding position in below subplot indicates Frequency offset for that entire symbol.

Unequalized PBCH IQ data after Time and frequency synchronization compensation is plotted in Figure 16.
Figure 16 Extracted compensated but Unequalized PBCH IQ data

This plot in Figure 17 indicates the PBCH data after synchronization in time and frequency and post equalization. The coarse and fine timing synchronization, the integer CFO, coarse and fine synchronization are estimated and compensated before equalization. The restored PBCH constellation is a 4QAM and as shown in Figure 17.
9. CONCLUSION:

The retrieving of the PBCH data from a received distorted signal was done. And the required results were obtained.

10. REFERENCES

During the project execution the following References are

[1] 3GPP, TS 36.211 V9.0 and TS 36.212 V9.0


