UNIVERSITY OF OULU
DEGREE PROGRAMME IN WIRELESS COMMUNICATIONS ENGINEERING

# RF TRANSCEIVER SYSTEM DESIGN FOR IOT IN WIDE AREA NETWORKS

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# ABSTRACT

**Wireless communication has grown rapidly in the last two decades. New applications and advancement in technology is boosting the demand. Internet of things (IoT) is nowadays topic of discussion for everyone related to the wireless communication industry. IoT is a system of interconnected devices which can be people, animals, things or machines each with a unique identifier and the ability to transfer data over a network without any interaction with humans or computers.**

**The aim of this thesis is system design of RF transceiver for IoT devices operating in wide area networks. Several service providers are struggling to capture the IoT market. In this thesis detailed system design of third generation partnership project (3GPP) newly specified user equipment category M1 also known as long term evolution machine (LTE-M) is presented. LTE-M can operate in both full duplex and half duplex and it uses the same signal structure as the current operational standard long term evolution (LTE). The designed transceiver is able to operate in half duplex and meet the performance requirement (95 % throughput) specified by 3GPP.**

**Radio frequency transceivers have various architectures and each architecture has its own pros and cons associated with it. This transceiver is designed to be integrated in a wearable device. Constraints like small size and low power restrictions led to the choice of direct conversion architecture for the design. Simulations were performed in ADS to verify the theoretical results.**

**Keywords: communication, interconnected, wide area networks, architecture, transceiver, direct conversion, radio frequency.**

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# FOREWORD

At first, I would like to thank ALLAH for providing me the strength and ability to compile and complete right on time. I would like to express my special gratitude to Prof. Aarno Pärssinen for assisting, directing and guiding me throughout my thesis. His benignant and friendly response let me finish my dissertation on time.

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Oulu, Finland March 21, 2017

Hamza Khan

# LIST OF ABBREVIATIONS AND SYMBOLS

| | |
|---|---|
| 3G | Third Generation |
| 3GPP | Third Generation Partnership Project |
| ACPR | Adjacent Channel Power Ratio |
| ADC | Analog to Digital Converter |
| ADS | Advanced Design System |
| BB | Baseband |
| BER | Bit Error Rate |
| BLE | Bluetooth Low Energy |
| CW | Continuous Wave |
| DAB | Digital Audio Broadcasting |
| DAC | Digital to Analog Converter |
| DCR | Direct Conversion |
| DL | Downlink |
| EVM | Error Vector Magnitude |
| FDD | Frequency Division Duplexing |
| FFT | Fast Fourier Transform |
| HD-FDD | Half Duplex Frequency Divison Duplexing |
| IF | Intermediate Frequency |
| IFFT | Inverse Fast Fourier Transform |
| IIP3 | Input Third Order Intercept Point |
| IMD | Intermodulation |
| IMRR | Image Rejection Ratio |
| IoT | Internet of Things |
| ISM | Industrial, Scientific and Medical |
| LNA | Low Noise Amplifier |
| LO | Local Oscillator |
| LPF | Low Pass Filter |
| LPWAN | Low Power Wireless Area Network |
| LTE | Long Term Evolution |
| LTE-M | Long Term Evolution Machine |
| M2M | Machine-to-Machine |
| MIT | Massachusetts Institute of Technology |
| NB-IoT | Narrowband Internet of Things |
| NF | Noise Figure |
| OFDM | Orthogonal Frequency Division Multiplexing |
| OFDMA | Orthogonal Frequency Division Multiple Access |
| OIP3 | Output Third Order Intercept Point |
| P1dB | 1 dB Compression Point |
| PA | Power Amplifier |
| PAPR | Peak to Average Power Ratio |
| QAM | Quadrature Amplitude Modulation |
| QPSK | Quadrature Phase Shift Keying |
| RB | Resource Block |
| RE | Resource Element |
| RF | Radio Frequency |

| | |
|---|---|
| RFIC | Radio Frequency Integrated Circuit |
| RFID | Radio Frequency Identification |
| SC-FDMA | Single Carrier Frequency Division Multiple Access |
| SEM | Spectrum Emission Mask |
| SHR | Superheterodyne |
| SIG | Special Interest Group |
| SNR | Signal-to-Noise Ratio |
| TDD | Time Divison Duplexing |
| TOI | Third Order Intercept |
| UE | User Equipment |
| UL | Uplink |
| WAN | Wireless Area Network |

| | |
|---|---|
| $\Delta P$ | Difference between Fundamental Signal & $n^{th}$ Order Output Signal |
| $\omega_n$ | Frequency of $n^{th}$ Sinusoid |
| $a_n$ | Amplitude of $n^{th}$ Sinusoid |
| $ACPR$ | Adjacent Channel Power Ratio |
| $BW$ | Bandwidth |
| $EVM_{CL}$ | Carrier Leakage Effect on Error Vector Magnitude |
| $EVM_{I-Q}$ | I-Q Imbalance Effect on Error Vector Magnitude |
| $EVM_{Noise}$ | Noise Effect on Error Vector Magnitude |
| $EVM_{PA}$ | Power Amplifier Effect on Error Vector Magnitude |
| $EVM_{PN}$ | Phase Noise Effect on Error Vector Magnitude |
| $EVM_{QN}$ | Quantization Noise Effect on Error Vector Magnitude |
| $EVM_{Total}$ | Total Non-idealities Effect on Error Vector Magnitude |
| $F_n$ | Noise Factor of Component n |
| $G$ | Gain of Receiver |
| $G_n$ | Gain of Component n |
| $IIP3$ | Input Referred Third Order Intercept Point |
| $IM3_{Output}$ | Intermodulation Products at Output |
| $IMR_2$ | Ratio of Wanted Signal to Intermodulation Product |
| $IP_n$ | $n^{th}$ Order Intercept Point |
| $k_n$ | $n^{th}$ Term Gain |
| $NF_{Total}$ | Total Noise Figure of Receiver |
| $OIP3_{(dBm)}$ | Third Order Output Intercept Point in dBm |
| $P$ | Power of Fundamental Signal |
| $P_{int}$ | Power of Interfering Signal |
| $P_{out(dBm)}$ | Output Power in dBm |
| $p_1$ | Output Power at Frequency $\omega_1$ |
| $p_{12}$ | Output Power at Frequency $\|2\omega_1 \pm \omega_2\|$ |
| $p_2$ | Output Power at Frequency $\omega_2$ |
| $P1dB$ | 1dB Compression Point |
| $Rx_{sensitivity}$ | Receiver Sensitivity |
| $SNR$ | Signal-to-Noise Ratio |
| $v_0$ | DC Component at Output |
| $v_{in}$ | Input Signal |
| $v_{out}$ | Output |

# 1. INTRODUCTION

Internet of things (IoT) is becoming an increasingly growing research topic in telecommunication industry and it is said to be a new revolution in the cellular ecosystem. The idea of IoT is not new it has been there for decades by the name of machine to machine communication (M2M). In the late 1990's, AutoID lab was built in Massachusetts Institute of Technology (MIT) with the help of Proctor & Gamble (P&G) and the goal was to link radio frequency identification system (RFID) with internet [1]. The idea was very simple but powerful to manage objects connected to the internet wirelessly. With the advancement of technology, the idea which was proposed in the late 1990's is on the verge of becoming a reality now.

IoT is a system of interconnected devices which can be people, animals, things or machines each with a unique identifier and the ability to transfer data over a network without any interaction with humans or computers. Moreover IoT can be explained as a category of devices which are low power consuming, operate in wide area networks and can perform several type of communication i.e. machine to machine, machine to network or network to machine.

It is expected that IoT will provide service in almost every sector i.e. medical, transportation, manufacturing, information technology, retail, electricity generation and distribution etc [2, 3]. In order to provide services to the above mentioned sectors there are some key requirements which battery operated IoT devices needs to fulfil [2]:

- Battery life of device should be in years.
- Deployment cost should be negligible.
- Cost of the device itself should be very low.
- Global coverage & network which can support massive number of devices.

RFID systems is an efficient tool for communication but the limiting factor of it is the range of communication. RFID system can work for distance up to few meters. An alternative for RFID is low energy bluetooth (BLE) which extends the range of communication up to 100 meters. These technologies can be used in local area network for communication between different nodes. Nodes might require to communicate with a central office located in another city or country for which requirement is to have a wide area network (WAN). WAN technology has grown and expanded over the years e.g internet. Driving factor for the WAN technology were mainly new services and applications. The primary advantage of WAN is that it allows companies to expand their networks through plug-in connections over locations and boost interconnectivity.

IoT is also known as low power wide area network (LPWAN). Prime concern with IoT devices is the battery life i.e. the devices should operate for years without recharging or changing the battery. To create low power consuming devices, standardization body has defined a new category long term evolution machine (LTE-M) which has narrow bandwidth. The second concern with IoT devices is that they should operate at much lower cost than the current deployed wide area network 'LTE'. Because of low cost factor there are some new companies entering the market such as Sigfox, LoRa.

Sigfox is a French company which deploys wireless networks to connect low-power devices. Their network uses very narrow bandwidth and they operate in unlicensed band using standard radio transmission protocols. Sigfox wants to deploy its own IoT network throughout the world and to become a global carrier of LPWAN.

LoRaWAN is a reconfigurable LPWAN network created by the LoRa Alliance. It allows companies to build their own IoT infrastructure which can operate in global or national network. The technology is designed to enable a gateway or base station to cover entire cities. Data from the end devices are transferred to the server via the gateway. Connection between the end devices and the gateway use single hop wireless communication while the connection between the gateway and the server is standard IP connection.

LTE-M is a new category of low powered devices introduced by the telecommunication standardization body third generation partnership project (3GPP) in 2015. LTE-M is also known as cellular IoT. LTE-M is evolved from the currently deployed LTE standard which uses bandwidth of up to 20 MHz per channel and has relatively high power consumption exceeding the IoT targets. Since many IoT devices will not require high data rates 3GPP has decreased the bandwidth of LTE-M to 1.4 MHz. The main difference between cellular IoT and non-cellular IoT service provider is that the former operates in licensed bands while the latter operates in unlicensed band which is susceptible to unwanted interference from other radio systems or new emerging service providers. 3GPP is updating the specifications for cellular IoT in every new release and it has already introduced a power saving mode (which will send data either periodically or on demand or when the device will be triggered). In addition to LTE-M 3GPP is working on a new standard named as narrow band IoT (NB-IoT) which uses the bandwidth of 200 kHz and it is designed for ultra-low powered devices. Whichever category of devices telecom operator will choose to work with, will greatly depend on the targeted market sector and the future technological aspects.

It is predicted that almost 15 Billion radio standard devices will be deployed by 2021 [3]. A comparison of different available LPWAN technologies is shown in Table 1. It's not sure at the moment that which standard will lead the IoT industry. Comparison of available solutions shows that less effort will be required to deploy cellular IoT as the network is already there (cellular coverage is global). A software upgrade in the current cellular network will provide connectivity to a diverse range IoT devices.

Table 1. Comparison of different LPWAN technologies

| | LTE-M | NB IoT | BLE | LoRa | Sigfox |
|---|---|---|---|---|---|
| Frequency | LTE-bands | LTE-bands | 2.4 GHz | < 1 GHz | < 1 GHz |
| Bandwidth | 1400 kHz | 200 kHz | 1000/2000 kHz | 125-500 kHz | 0.2 kHz |
| Data rate | 200 kbps | 20 kbps | 270 kbps | 0.29-50 kbps | 0.1 kbps |
| Modulation | OFDMA | OFDMA | DSSS | SS | BPSK |
| Range | 10 km | 10 km | 0.1 km | 21 km | 10-30 km |

In this thesis RF system design of a cellular IoT device is discussed in detail. System design of a trasmitter that can transmit a LTE-M signal according to the specifications of 3GPP was performed alongwith the design of a receiver that can receive a LTE-M signal and can forward it to baseband processing stage without inducing any errors. RF system design takes into account several consideration i.e. blocking, sensitivity, frequency control, transmitted power, received power, interference caused to other RF devices. This thesis is organized as follows: LTE-M signal structure is described in Chapter 2. Different RF architectures are outlined in Chapter 3. Detailed analysis and comparison of LTE-M specification with other technologies is perfomed in Chapter 4. Design of transceiver is outlined in Chapter 5. Discussion is listed in Chapter 6. Finally, Chapter 7 concludes the thesis.

# 2. LTE PHYSICAL LAYER

The demand for higher data rate is increasing and because of this increased demand telecommunication standardization body '3GPP' introduced LTE in Release 8. To support higher data rates a bandwidth efficient modulation scheme was required and as a result OFDM was chosen as a candidate for the radio access technology of LTE in 2005. OFDM has been known since mid-1960s [4, 5] but due to lack of technological advancements it was not a practical solution till 1995. In 1995 ETSI introduced digital audio broadcasting (DAB) standard and it was based on the OFDM technology. Since then OFDM has been used in many standards including digital video broadcasting (DVB), Wi-Fi (IEEE 802.11 a/g/n), WiMAX (IEEE 802.16), and LTE [6]. LTE-M was specified in Release 13 of 3GPP. Although the specification of LTE-M & LTE varies a lot, they both use the same signal structure. In the following sections physical layer of LTE uplink and downlink is discussed in detail.

## 2.1. LTE Downlink

LTE-M & LTE both use OFDMA as a channel access technology in the downlink. OFDM is a special case of frequency division multiplexing which uses a large number of closely spaced subcarriers that are modulated with low data rates. In OFDM these closely spaced subcarriers are orthogonal to each other hereby mitigating the interference between them as shown in Figure 1. Orthogonality between subcarriers is achieved by keeping the subcarrier spacing equal to the reciprocal of symbol period. Since the subcarriers are narrowband they don't experience frequency selective fading because subcarrier bandwidth is smaller than the channel coherence bandwidth. OFDM design also mitigates the need of complex equalization techniques that were used in 3G. [7]

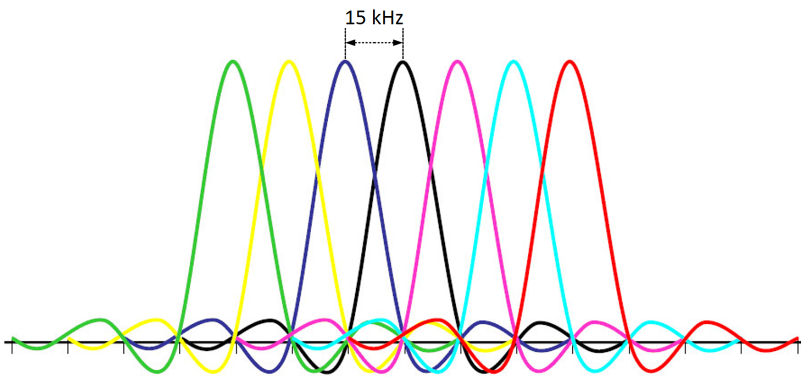


Figure 1. OFDM subcarriers.

Information that has to be sent is converted from serial to parallel and then spread across multiple subcarriers. When the signal comprising of these subcarriers is transmitted, multiple copies of the same signal are received due to reflection. Time dispersion is equivalent to frequency selective response of the channel. This results in loss of orthogonality and interference between subcarriers. To mitigate the effect caused by multipath, cyclic prefix is introduced at the start of OFDM symbol. Cyclic prefix is the copy of last part of the OFDM symbol as illustrated in Figure 2. To eliminate ISI, cyclic prefix should be longer than the delay spread of the channel. [7]

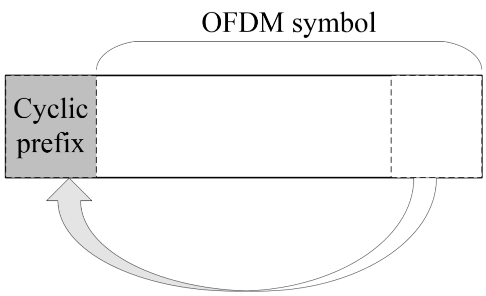


Figure 2. Cyclic prefix insertion in OFDM.

OFDMA is implemented using fast fourier transform (FFT) and inverse FFT (IFFT). At the transmitter IFFT modulates information across multiple subcarriers and creates a signal comprised of all these different subcarrier frequencies and at the receiving end FFT is applied to retrieve the data [6]. Figure 3 shows the block diagram of OFDM transceiver. In case of OFDM, IFFT does the same thing as multiple transmitter chains would do. After IFFT a parallel to serial converter is used because the samples generated by IFFT has to be transmitted one after the other. [7]

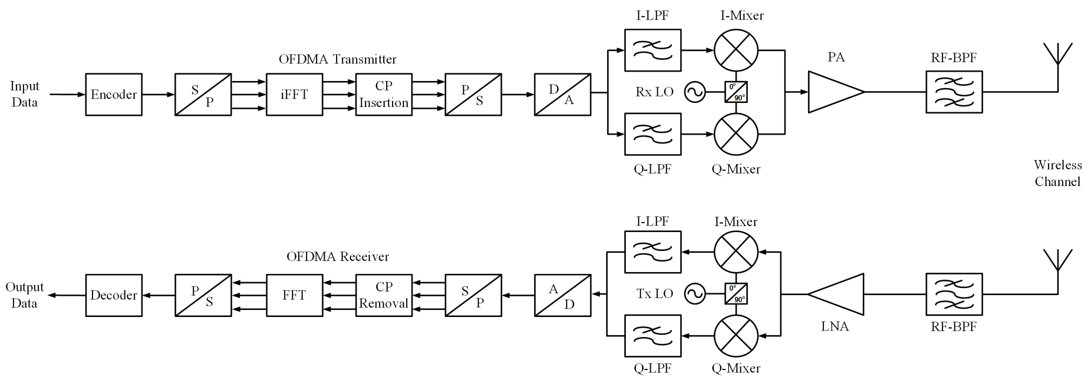


Figure 3. Block diagram of OFDM tranceiver.

Lets assume that $L$ (no of subcarrier) dimensional data vector is to be transmitted $Z[i] = \begin{bmatrix} Z_0[i] & Z_1[i] & \dots & Z_{L-1}[i] \end{bmatrix}^T$ where $i$ is the index of OFDM symbol. This $L$-dimensional data vector is first modulated using conventional modulation scheme (QPSK, 16 QAM, 64QAM) onto different subcarriers which results in complex vector $Y[i] = \begin{bmatrix} Y_0[i] & Y_1[i] & \dots & Y_{L-1}[i] \end{bmatrix}^T$. Given this complex vector as input to $N$ point IFFT results in $N$ complex discrete samples $K[i] = \begin{bmatrix} K_0[i] & K_1[i] & \dots & K_{N-1}[i] \end{bmatrix}^T$. The size of IFFT should be greater or equal to the number of modulated subcarriers i.e. $L \leq N$. After IFFT cyclic prefix is inserted in the OFDM symbol, it is replica of the last $O$ samples of IFFT output. It is appended at the beginning of $K[i]$ resulting in a time domain signal $K[i] = \begin{bmatrix} K_{N-O}[i] & \dots & K_{N-1}[i] & K_0[i] & \dots & K_{N-1}[i] \end{bmatrix}^T$.

Purpose of inserting cyclic prefix is to avoid ISI. Cyclic prefix is chosen to be greater than the maximum delay spread of the channel. Since symbols of OFDM travel one by one and each symbol will encounter time delay caused by the channel. As a result symbol gets spread out and interfere with the next symbol creating ISI. ISI is avoided by inserting cyclic prefix at the start of OFDM symbol which is larger than the delay spread retaining the spread out symbol from interfering with the next symbol. [7]

Maximum number of subcarriers in an OFDM signal is 2048. Each subcarrier has a spacing of 15 kHz. Number of subcarriers depend upon the bandwidth with 2048 subcarriers used by 20 MHz. Subcarriers are grouped in the form of resource blocks (each resource block have 12 adjacent subcarriers) and these resource blocks are scheduled for transmission of different user, each resource block is given a time slot of 0.5 ms. Figure 4 shows one resource block which is divided into 12*7=84 resource elements (RE). RE is the smallest unit of LTE signal. Each resource element can carry one symbol which can be QPSK, 16 QAM or 64-QAM. [7]

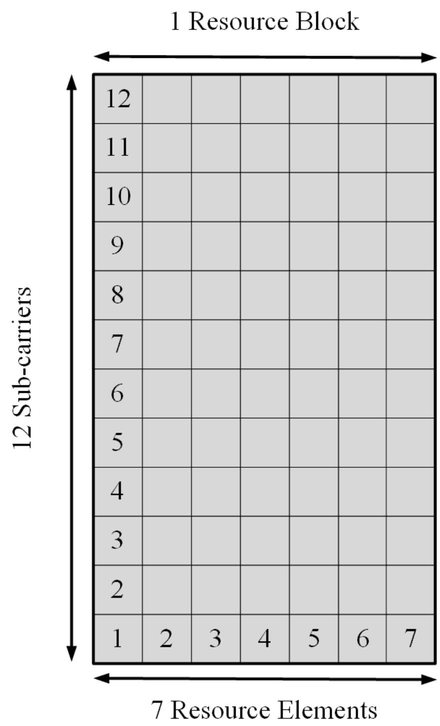


Figure 4. Structure of resource block.

## 2.2. LTE Uplink

Despite of many advantages of the OFDM technology it is not used in the uplink because of peak to average power drawback. As we have discussed earlier an IFFT operation modulates the information along with addition of subcarriers to make a combined signal. The addition of subcarriers creates high peaks in the output envelope. Variation of amplitude of the OFDM signal makes the linearity requirement for an RF amplifier stricter. If the amplifier is not linear enough to cover the entire range of signal's amplitude fluctuations the signal will be clipped from the peaks and as a result performance is degraded. To avoid degradation of performance, highly linear amplifiers are required which have poor efficeincy and consume a lot of power and in case of mobile station which has limited power these amplifiers should be avoided. [6]

Because of PAPR disadvantage in OFDM, another technique known as single carrier frequency division multiple access (SC-FDMA) is used for uplink in LTE. SC-FDMA combines low PAPR offered by single carrier system with multipath resilience offered by OFDM system. Unlike OFDMA where each subcarrier is individually modulated, SC-FDMA modulated signal is a combination of all the transmitted data symbols at a given time instant. In other words each transmitted subcarrier in the uplink contains information about all the transmitted symbols while in downlink each subcarriers contains information about one specific symbol. Figure 5 shows a comparison of OFDMA with SC-FDMA. [7]

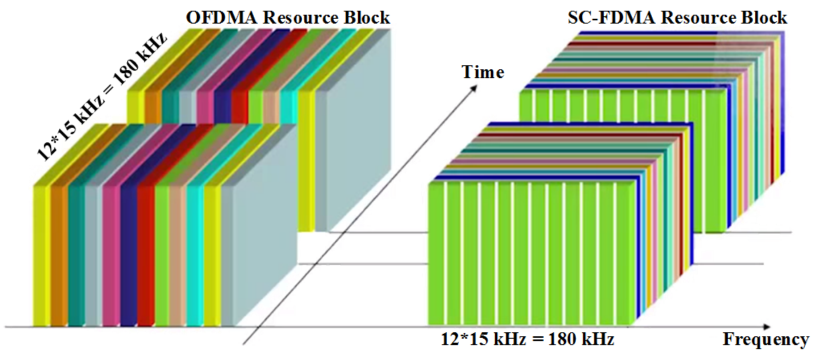


Figure 5. Comparison of OFDMA vs SC-FDMA.

In OFDM bits are grouped together and sent to IFFT for modulation of individual subcarrier, but in case of SC-FDMA a FFT operation is performed on the information bits first followed by an IFFT operation. Output of FFT is the basis for the creation of subcarriers for the following IFFT. Figure 6 shows the block diagram of SC-FDMA transceiver. [7]

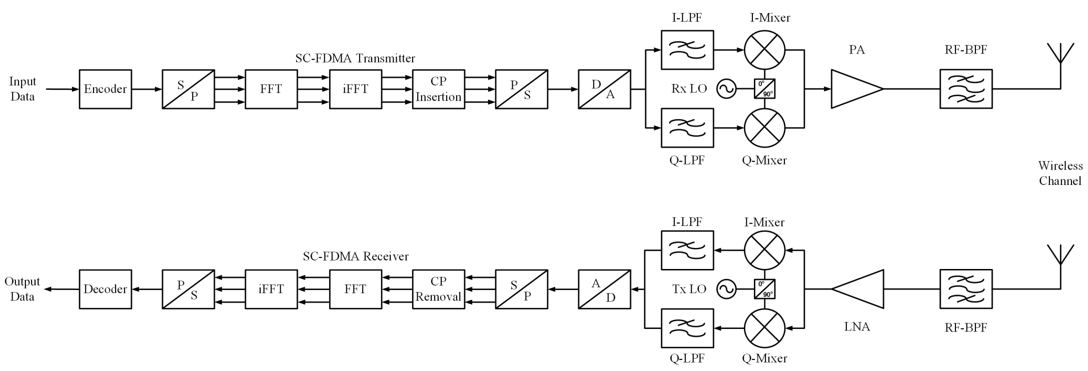


Figure 6. Block diagram of SC-FDMA transceiver.

Multiple access in case of SC-FDMA is made possible by assigning zeros to the subcarriers that are used by other users. Transmission structure for uplink and downlink is the same. Each resource element consist of one SC-FDMA data block on one subcarrier. Minimum amount of bandwidth utilized by a user equipment can be 180 kHz. In case of uplink adjacent resource blocks are assigned to users while in case of downlink non-adjacent RB's are allocated. [7]

As mentioned earlier OFDM can use several different modulation schemes, the choice of modulation scheme greatly depend upon the minimum signal-to-noise ratio (SNR) the system can achieve. The bigger the constellation size the more is the SNR required by it e.g. -1 dB SNR is required by QPSK (code rate 1/3) while 18.6 dB SNR is required by 64-QAM (code rate 4/5). The minimum SNR is chosen in order to meet the throughput requirement. Noise figure is a measure of SNR degradation caused by components in RF chain. Phase and gain imbalance occurs due to difference in phase and gain of I & Q branches. Quantization noise introduced by analog to digital converter (ADC) or digital to analog converter (DAC) also increases the noise floor hereby degrading the SNR. Received signal is also accompanied with blocking signals which desensitize the receiver and result in loss of performance. [7]

Aside from SNR other important criterion's are the error vector magnitude (EVM) and adjacent channel power ratio (ACPR) which are taken into account while designing the transmitter. EVM is a measure of error between the actual symbol and the reference symbol. ACPR is a measure of interference caused to adjacent radio channels. Choice of RF components depend upon the nonlinearities/losses which the system can afford while achieving minimum performance. [7]

# 3. TRANSCEIVER ARCHITECTURES

Transceivers incorporates both transmitter and receiver. RF transmitter is the electronics between the transmitting antenna and digital to analog converter (DAC). On the other hand RF receiver is the electronics between receiving antenna and analog to digital converter (ADC). Modern radio transceivers have to operate in a crowded radio environment and there are serious design challenges related to the architecture of transceivers. In the upcoming sections two of the most commonly used radio transceivers are discussed in detail.

## 3.1. Superheterodyne Transceiver

RF transceivers have different architectures ranging from complex architecture such as superheterodyne (SHR) shown in Figure 7 to simple architectures like direct conversion (DCR) shows in Figure 8. Choice of architecture depend upon the application and implementation requirement, as each architecture has its own pros and cons.

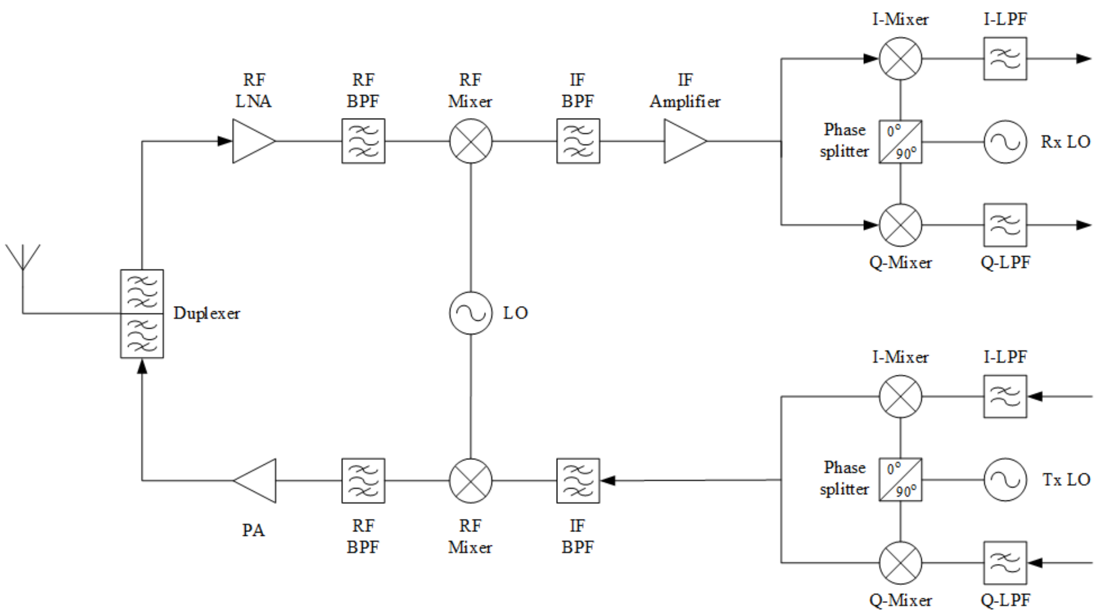


Figure 7. Simplified block diagram of superheterodyne transceiver.

Superheterodyne works well for both narrowband and wideband transceivers. Bandpass filters in superheterodyne provides protection against interfering signals. SHR has at least two frequency conversion stages the first one translates the signal into intermediate frequency (IF) while the second stage converts the signal into baseband (BB) in case of receiver or to radio frequency (RF) in case of a transmitter. IF signal split into I & Q branches which is necessary, since it is impossible to detect IF signal (utilizing amplitude and phase modulation) with single ended approach. Superheterodyne employs bandpass filters in RF/IF stages and quadrature up/down conversion which makes it possible to achieve good performance with moderate constraints. High performance levels can be achieved with the increase of complexity of transceiver along with increase of current consumption, component count and physical size. This architecture is not suitable for integration because of several connections with external lumped components, mainly RF and IF filters. Utilizing IF stage in the design relaxes filtering requirement but also causes image problems. [8]

### 3.2. Direct Conversion Transceiver

Direct conversion transceiver with IQ back-end is attractive especially for wideband systems. Direct conversion transceiver employs two mixers which down-converts the signal from RF to baseband in a receiver and vice versa in a transmitter. It is also known as homodyne of zero-IF architecture. From the last twenty years it has been used in several systems because of its compact design. DCR offer numerous advantages over SHR e.g. avoids image problem, number of components is reduced, integration is easier and the cost is lower. Many applications such as WiFi and Bluetooth utilize the direct conversion architecture. On the other hand there are some disadvantages associated with it as well e.g. dc offset, tough requirements for linearity and IQ balance along with susceptibility to number of interfering phenomenon's. IQ demodulation in case of DCR is necessary. [8, 9]

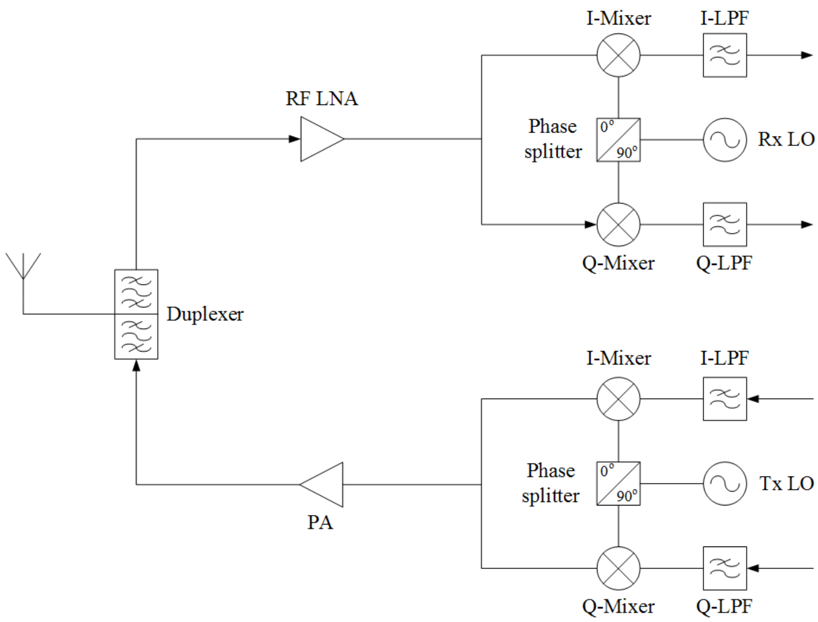


Figure 8. Direct conversion transceiver.

Signal entering the direct conversion receiver is filtered by the bandpass filter to remove far out interfering signals (avoiding desensitization). Low noise amplifier (LNA) amplifies the signal to ensure low system noise figure (NF) essential for the demodulation of a weak desired signal. Amplified signal is directly down converted using quadrature demodulation and the low pass filters provide protection from close in interfering signals (generated from within or outside the receiver). Linearity and quadrature accuracy of I & Q channels is basis for protection against parasitics as well as the intermodulation products. Consequently the constraints on back-end in case of DCR are more stricter than SHR. RF performance requirements for the receiver are discussed in Section 3.3. [8]

In case of transmitter the baseband I & Q signals are directly up-converted to the RF frequency. These up-converted signals are combined and fed to the power amplifier (PA). The amplified signal is passed through the bandpass filter which removes the PA harmonics as well as suppresses the power emitted in the adjacent channels. This transmitter architecture is subjected to 'injection pulling' interference, because PA and the voltage controlled oscillator (VCO) operates on the same frequency. Isolation of the synthesizer leads to degradation of this interference. Similar to the receiver, transmitter also put stricter requirement of linearity and IQ balance on the back-end. Compact design is achieved as only one LO is used and there is no IF chain. RF performance requirements for the transmitter is discussed in Section 3.4. [8]

The target is to design a wearable device RF transceiver which has a small size and low power consumption. Comparison of superheterodyne and direct conversion transceiver shows that, DCR architecture is well suited for this application.

## 3.3. Receiver RF Performance Requirements

Receiver RF signal processing is the intermediate step between the reception of the signal (at the antenna) and the extraction of information from the signal at the baseband stage. Receiver should reliably demodulate the information along with rejection of interference. RF performance requirement ensures that the designed system can meet a specified criteria. RF components are generally nonideal and induce nonlinearities in the system, which impact the extraction of information from a modulated signal. Analyzing the performance requirements carefully is a critical step in the design of any RF system. Some of the performance criterion's are discussed in the following subsection which effects the system design. [7]

### *3.3.1. Sensitivity*

Sensitivity of a receiver is the most important system specification of any wireless communication standard. It defines the minimum level of input signal applied at the antenna port for which the receiver can produce sufficient SNR to demodulate the desired modulation scheme (including the coding gain). SNR deteriorates as the signal passes through the receiver, because a noisy receiver adds own noise to the signal. Deterioration of the SNR is described with noise figure (NF). NF limits the receiver performance. Large enough SNR is required at the input of an A/D converter to get the defined bit error rate. SNR requirement is affected by the chosen modulation scheme, algorithms etc. Receiver sensitivity can be calculated from [7].

$$Rx_{sensitivity} = -174 + 10 * log(BW) + NF + SNR, \tag{1}$$

where $NF$ is the noise figure of the receiver, $SNR$ is the required signal-to-noise ratio and $BW$ is the bandwidth of the receiver. Sensitivity calculated by Equation (1) can also be used to determine the maximum noise figure of the receiver. The desire of every system designer is to keep the $NF$ as low as possible as it decreases the sensitivity of receiver. [9]

### *3.3.2. 1dB Compression Point*

Linear amplifiers usually have a fixed gain for a certain frequency range. This means that if we plot the output power versus the input power we will get linear relationship, as shown in Figure 9. The gain of the amplifier is the slope of the line. The graph shows that as the input power increases, there comes a point after which the gain does not follow the linear relation any more. Saturation point of the amplifier is the point where the increase of input power doesn't cause any change in the output power. In this saturation region the response of the amplifier becomes strongly nonlinear and causes signal (amplitude) distortion, harmonics and intermodulation products. [7, 8]

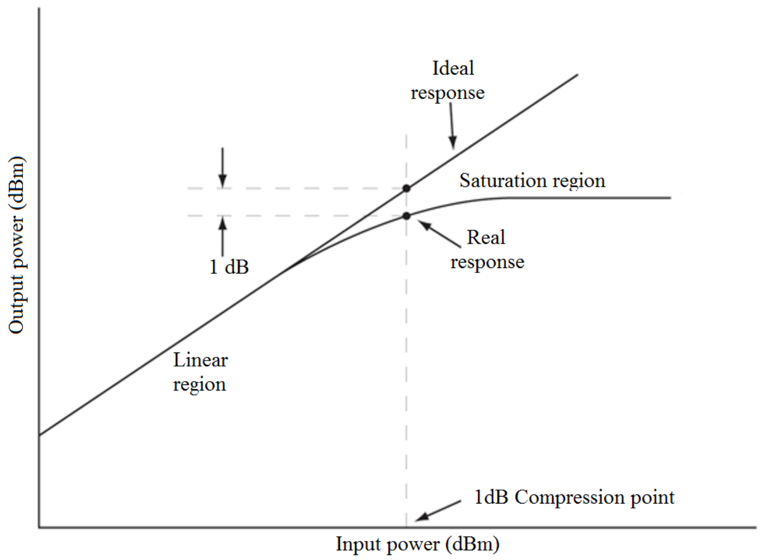


Figure 9. Amplifier response (input vs output power).

The 1dB compression point (P1dB) corresponds to the input power that causes the gain to decrease 1dB from the expected linear gain. We always want the amplifiers to operate in the linear region. Amplifier for any application is chosen in way that its operating point including signal swing headroom (amplitude fluctuations) should remain lower than the 1dB compression point. From the system design point of view P1dB is always chosen to be higher than the expected maximum signal level i.e. if the maximum signal level is -25 dBm then the P1dB point is chosen to be greater than -25 dBm. On the other hand in case of transmitter P1dB is always chosen to be higher than the desired output power, hereby avoiding saturation. [8]

### *3.3.3. Intermodulation Distortion*

When the amplifier operates in the nonlinear region, it starts producing frequency components that are multiple of the input signal frequency known as harmonics. The second, third, fourth and higher order of this harmonics are usually far outside the bandwidth of the amplifier, and therefore can be easily filtered. [8]

However, the nonlinearity will result into mixing of two or more signals which are called intermodulation (IMD) products. If these signals are close in frequency, then some of the difference or sum frequencies can result within the amplifier's bandwidth. If they are not removed, they can cause interference in the system. This happens because it is very difficult to filter these signals (since they are close to the desired frequency). Strict filtering requirements is imposed if these intermodulation products are to be filtered out. To loosen up this filtering requirements we try to control the biasing, signal levels and other factors so that the amplifier would have maximum possible linearity, resulting in low intermodulation distortion (IMD). [8]

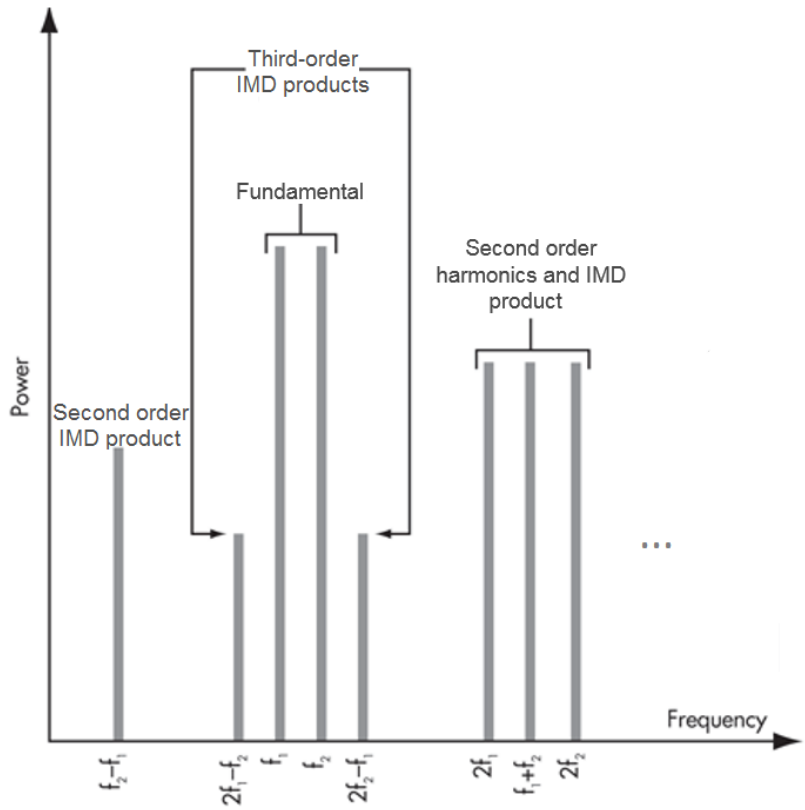


Figure 10. Harmonics and intermodulation products.

From Figure 10 we can see two signals with frequencies $f_1$ and $f_2$ which are within the amplifiers bandwidth. If the amplifier is operating in the nonlinear region, it will have many extra components, such as $f_1 - f_2$ and $f_1 + f_2$. These components are far from our desired frequencies, and therefore can be easily filtered in a superheterodyne receiver. In a direct conversion receiver $f_1 - f_2$ could be on the passband. However these signals will mix with the second, third, fourth and higher harmonics, to produce signals which is quite likely that will be within the amplifiers passband, thus interfering with the desired signals. The problem though is with the third-order products, which are $2 * f_1 \pm f_2$ and $2 * f_2 \pm f_1$. As can be seen from Figure 10 the IMD products which are more likely to be close to our desired frequencies are $2 * f_1 - f_2$ and $2 * f_2 - f_1$. Intermodulation distortion can be understood by considering the input output relation of a two port network which can be modelled with Taylor series as expressed in Equation (2) [10, 11]. [12]

$$v_{out}(v_{in}) = v_0 + k_1 v_{in} + k_2 v_{in}^2 + k_3 v_{in}^3 + ... + k_n v_{in}^n, \tag{2}$$

where $v_{out}$ is the output signal, $v_{in}$ is the input signal, $v_0$ is the dc component, $k_1$ is the small signal gain and $k_2$, $k_3$, ... , $k_n$ are the gains of n[th] order nonlinearities. If the input $v_{in}$ of the system is a sum of cosine signals i.e. $v_{in} = a_1 cos(\omega_1 t) + a_2 cos(\omega_2 t)$ [11], where $a_1$, $a_2$ correspond to the amplitude of sinusoids and $\omega_1$, $\omega_2$ corresponds to the frequency. By applying this signal at the input of two port network we get the output as mentioned in Equation (3) [10, 11].

$$\begin{aligned} v_{out}(v_{in}) = v_0 + k_1[a_1 cos(\omega_1 t) + a_2 cos(\omega_2 t)] + k_2[a_1 cos(\omega_1 t) + a_2 cos(\omega_2 t)]^2 + \\ k_3[a_1 cos(\omega_1 t) + a_2 cos(\omega_2 t)]^3 + ... \end{aligned} \tag{3}$$

When we expand the third term we get Equation (4) and from that we can see that the mixing of two input signals have caused an undesired signal at the wanted frequencies $\omega_1$ & $\omega_2$. Also other distortion products $2\omega_1 - \omega_2$ & $2\omega_2 - \omega_1$ are caused by this third order nonlinearity as shown in Figure 10 [10, 11].

$$\begin{aligned} k_3[a_1 cos(\omega_1 t) + a_2 cos(\omega_2 t)]^3 = \left[\frac{3k_3 a_1^3}{4} + \frac{3k_3 a_1 a_2^2}{2}\right] cos(\omega_1 t) + \\ \left[\frac{3k_3 a_2^3}{4} + \frac{3k_3 a_1^2 a_2}{2}\right] cos(\omega_2 t) + \left[\frac{k_3 a_1^3}{4}\right] cos(3\omega_1 t) + \left[\frac{k_3 a_2^3}{4}\right] cos(3\omega_2 t) + \\ \left[\frac{3k_3 a_1^2 a_2}{4}\right]\Big(cos((2\omega_1 + \omega_2)t) + cos((2\omega_1 - \omega_2)t)\Big) + \\ \left[\frac{3k_3 a_1 a_2^2}{4}\right]\Big(cos((2\omega_2 + \omega_1)t) + cos((2\omega_2 - \omega_1)t)\Big) \end{aligned} \tag{4}$$

Adding third order term with the first order and arranging them according to the frequency gives Equation (5) [10, 11].

$$\begin{aligned} k_1[v_{in}] + k_3[v_{in}]^3 = \left(k_1 a_1 + \frac{3k_3 a_1^3}{4} + \frac{3k_3 a_1 a_2^2}{2}\right) cos(\omega_1 t) + \frac{k_3 a_1^3}{4} cos(3\omega_1 t) \\ \left(k_1 a_2 + \frac{3k_3 a_1^2 a_2}{2} + \frac{3k_3 a_2^3}{4}\right) cos(\omega_2 t) + \frac{k_3 a_2^3}{4} cos(3\omega_2 t) \\ \left[\frac{3k_3 a_1^2 a_2}{4}\right]\Big(cos((2\omega_1 + \omega_2)t) + cos((2\omega_1 - \omega_2)t)\Big) + \\ \left[\frac{3k_3 a_1 a_2^2}{4}\right]\Big(cos((2\omega_2 + \omega_1)t) + cos((2\omega_2 - \omega_1)t)\Big) \end{aligned} \tag{5}$$

Intercept point can be calculated from $IP_n = P + \frac{\Delta P}{n-1}$ [13, 14], where $IP_n$ is the $n^{\text{th}}$ order intercept point, $P$ is the power of fundamental signal and $\Delta P$ is the difference between the fundamental signal and the $n^{\text{th}}$ order intermodulation product. Defining $p_1 = (k_1 a_1)^2$ as the output power at $\omega_1$, $p_2 = (k_1 a_2)^2$ as the output power at $\omega_2$ and $p_{12} = (3k_3 a_1^2 a_2/4)^2$ as the output power at $|2\omega_1 \pm \omega_2|$. From the relation of intercept point we can compute the OIP3 from Equation (6) [10].

$$OIP3 = p_1 + \frac{p_2 - p_{12}}{2} = (k_1 a_1)^2 + \frac{(k_1 a_2)^2 - (3k_3 a_1^2 a_2/4)^2}{2} \tag{6}$$

If we take the logarithm of Equation (6) we get the value of OIP3 and IIP3 (by replacing $p_1 = a_1^2$ in Equation (6)) in dBm as shown in Equation (7) [10].

$$OIP3_{(dBm)} = 20log\sqrt{\left(\left|\frac{4k_1^3}{3k_3}\right|\right)}, IIP3_{(dBm)} = 20log\sqrt{\left(\left|\frac{4k_1}{3k_3}\right|\right)} \tag{7}$$

Similarly, Equation (5) can by used to find the relationship between IIP3 and P1dB. The amplitude of second sinusoid is set to 0 i.e. $a_2 = 0$ and we get Equation (8) [10].

$$k_1[v_{in}] + k_3[v_{in}]^3 = \left(k_1 a_1 + \frac{3k_3 a_1^3}{4}\right) cos(\omega_1 t) + \frac{k_3 a_1^3}{4} cos(3\omega_1 t) \tag{8}$$

When the amplitude of fundamental signal $a_1$ is low the term $k_1 a_1$ dominates but as the amplitude increases the other term $3k_3 a_1^3/4$ starts dominating and drives the amplifier into compression. When the summation of these two terms is 1 dB below the first term alone that input power point ($a_1$) is known as 1dB compression point. This can be written mathematically as Equation (9) [10].

$$k_1 a_1 * 10^{-1/20} = \left(k_1 a_1 + \frac{3k_3 a_1^3}{4}\right) \tag{9}$$

Separating $a_1$ terms from Equation (9), we can re-write it as Equation (10) [10].

$$a_1 = \sqrt{\frac{-4k_1}{3k_3}} \sqrt{1 - 10^{-1/20}} \tag{10}$$

If we take the logarithm of Equation (10) we get 20log$\sqrt{\frac{-4k_1}{3k_3}}$ which is equal to the IIP3 as mentioned in Equation (7) so we replace it and and get the 1dB compression point as Equation (11) [10].

$$P1dB = IIP3 + 20log(\sqrt{1 - 10^{-1/20}}) = IIP3 - 9.65 \text{ dB} \tag{11}$$

If we plot the input vs output power, we get a curve which is similar to Figure 9 and can be seen in Figure 11. Figure 11 shows the fundamental and the third order intermodulation response. The point where the linear extension of these two curves intersect is called the Third Order Intercept (TOI) point. This point is never achieved in practice, but it is useful in case we want to determine the linearity condition of the amplifier. From this plot we can read the TOI with respect to both input and output. If the value is read from the output axis, it is called Output Third Order Intercept Point (OIP3), and if we read it from the input axis, then the value is called Input Third Order Intercept Point (IIP3). [12]

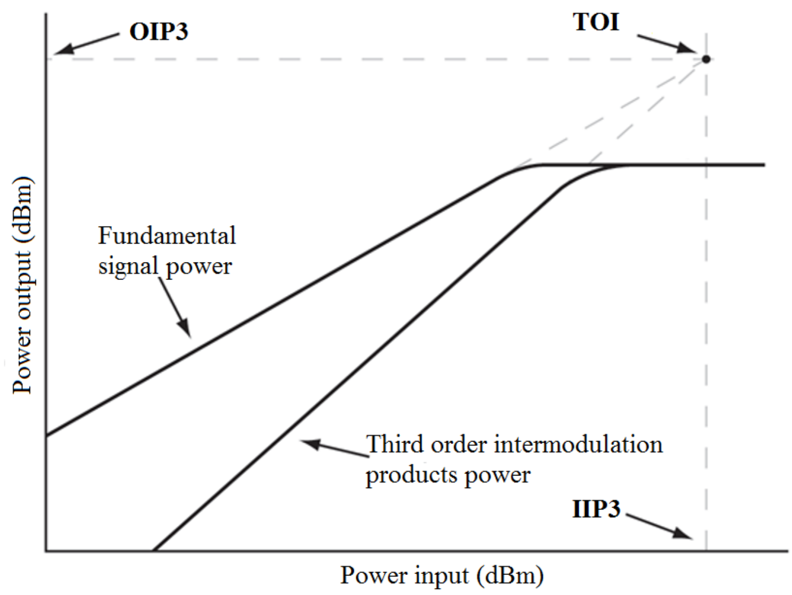


Figure 11. Third order intercept point.

The value of IIP3 determines the magnitude of IMD products that appear at the output of receiver. Level of IMD products at the output can be calculated as [8].

$$IM3_{Input} = 3 * P_{int} - 2 * (IIP3), \tag{12}$$

where $IIP3$ is the input referred TOI point and $P_{int}$ is the power of interfering signal at the input. The value of TOI is almost 10 dB higher than the 1dB compression point as calculated from Equation (11). From Figure 11 and Equation (12) it can be clearly seen that higher third order intercept point (TOI) results into better linearity and lower level of IMD products. [8, 12]

Receiver's linearity is important since there are many interfering signals present in the medium. These signals when mixed together produce an interfering signal which is at the same frequency of desired signal. Filtering this signal is not possible so the only solution is to have high receiver linearity which in turn will produce low level IMD products. SNR determines the minimum level of intermodulation products at the output. [8]

When we design the transmitter concern changes from lowering the IMD products to having a desired output power and a defined adjacent channel power ratio (ACPR). Linearity of amplifier governs the maximum output power. In case of LTE the desired output for class 3 transmitter is +23 dBm so we want to have the transmitter 1dB compression point to be greater than this value (to avoid saturation) and in turn OIP3 should be almost 10 dB higher (to maintain ACPR and avoid performance degradation due to amplitude fluctuation) than this P1dB point. [7]

### *3.3.4. Desensitization*

In RF systems desensitization refers to the increase of system noise figure in the presence of high level unwanted signals. When a receiver is desensitized it is unable to receive a weak signal because the strong unwanted signal has increased the noise floor and decreased the sensitivity. So the receiver with sensitivity of -100 dBm if desensitized up to 5 dB will then have the decreased sensitivity of -95 dBm (signals below -95 dBm will not be detected). When RF transceivers are designed special attention is paid to high level blocking signals which can desensitize the receiver. Linearity and filtering approaches are often adopted to avoid desensitization. [15, 16]

Desensitization is also known as receiver blocking. Strong interfering signals affect the quiescent value and load lines of active devices and as a result receiver reaches saturation/blocking. Desensitization occur because of two phenomenons. One of the phenomenon is that large unwanted closely spaced signals appear at the input of receiver and creates third order intermodulation products at the desired signal frequency, hereby increasing the interference in receiver. Second phenomenon of desensitization occur because of the 2nd order nonlinearity in the receiver. Low frequency noise sources which are present in the amplifier are up-converted after being mixed with interfering signals resulting in an interfering signal at desired frequency. [17]

## 3.4. Transmitter RF Performance Requirements

Transmitter and receiver can operate simultaneously in a frequency division duplex system or both can communicate in different time slots in a time division duplex system. Architecture of transmitter is similar to that of receiver i.e. it can be superheterodyne, direct conversion. Important design parameters of the transmitter are error vector magnitude (EVM) which is a measure of modulation accuracy and adjacent channel power ratio (ACPR) which measures the level of unwanted emissions in adjacent channel. Nonlinear analysis in the transmitter is performed because the signal levels are high enough to drive the active devices into saturation. Nonlinearity in the transmitter mostly comes from the power amplifier (PA) which results in increased power being emitted in adjacent channels. [12]

### *3.4.1. Adjacent Channel Power Ratio*

Unwanted emissions from transmitters are restricted to avoid interference with other radio systems. Amongst all the unwanted emissions the most important ones are the emissions in adjacent channels. Adjacent channel power ratio (ACPR) is an important performance metric for power amplifier's in a transmitter design. It is defined as the $log_{10}$ ratio between the powers contained in a certain bandwidth $B$ at a certain offset frequency $f_0$ from the center frequency $f_c$ to the power contained in the bandwidth $B$ placed around $f_c$ as seen from Figure 12. ACPR is expressed in dB. [12]

ACPR measures how much a signal is spread into the adjacent channels. Signal spectrum is spread by the combined effect of nonidealities i.e. phase noise, spurious signals, nonlinear PA.

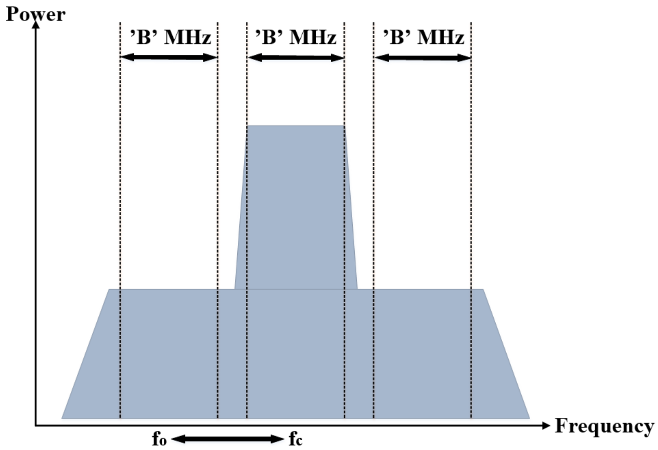


Figure 12. Adjacent channel power ratio measurement.

Careful system design makes sure that unwanted emissions are kept lower than the specified level. Highly linear PA's are required to meet the specifications of ACPR. When the output power is close to P1dB point contribution of fifth and higher order nonlinearities to the ACPR increases. Output power of the amplifier is usually backed off from the peak power (close to saturation) for operation in linear region which degrades the efficiency. When back-off is small efficiency is high and the resulting unwanted emissions are higher as well. When back-off is large the amplifier is deeply in the linear region which decreases the efficiency but we get the benefit of lowered unwanted emissions. [12]

### *3.4.2. Error Vector Magnitude*

In transmitters the signal quality is degraded by nonidealities of the components used such as modulators, in order to measure the total effect of this nonideality we use a quantity known as error vector magnitude (EVM). EVM is the ratio of the noise and distortion vector (with phase) power generated by all the nonidealities to the total power of ideal transmitted signal.

Error vector magnitude measures the difference between the ideal location of the symbol to the actual location of symbol in the constellation diagram as shown in Figure 13. [12]

Modulation accuracy of the signal degrades because of following reasons. [12]

- Non ideal filtering generates ISI (waveform is distorted due to filter group delay distortion and ripples).
- Phase noise of synthesizer.
- DC offset.
- Amplitude and phase imbalance of I & Q channel.
- Non-linearity of the PA.

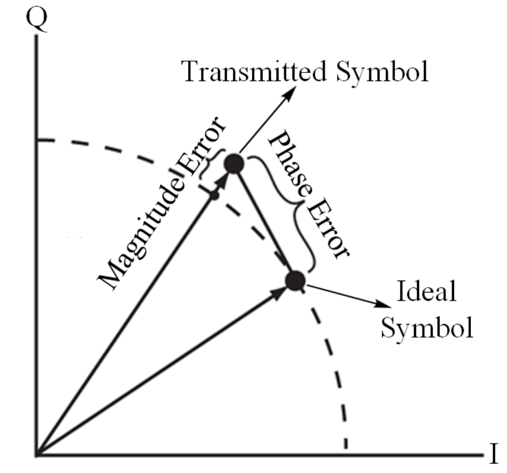


Figure 13. Error vector magnitude calculation.

EVM requirement is defined for each modulation scheme. High order modulation schemes has strict EVM requirement while low order modulation schemes has less stringent requirements. EVM values correspond to the loss of throughput. When designing a transmitter, effect of all the nonidealities on EVM is calculated and it is kept below the required level for the modulation scheme. EVM can be calculated from Equation (13) [12], where $Noise$ refers to the unwanted effect which is causing EVM degradation i.e. phase noise, carrier leakage, nonlinear PA etc. [7]

$$EVM_{Noise} = \sqrt{10^{\frac{Noise}{10}}} \tag{13}$$

If the nonidealities causing degradation of EVM are more than one and also they are statistically uncorrelated then the combined effect of EVM can be calculated from Equation (14) [12].

$$EVM_{Total} = \sqrt{EVM^2_{nonideality_1} + EVM^2_{nonideality_2} + ... + EVM^2_{nonideality_n}} \tag{14}$$

# 4. SYSTEM SPECIFICATIONS

Standardization body '3GPP' specifies the system requirements for transceivers operating in cellular network. These specification make sure that LTE-M operating devices achieve a minimum throughput of 95 % [18]. On the other hand Bluetooth special interest group (SIG) is responsible for the specification of Bluetooth devices and the criteria is to have 0.1 - 0.017 % bit error rate (BER) [19]. In the upcoming sections receiver and transmitter specification of Long Term Evolution Machine (LTE-M), Narrow Band Internet of Things (NB-IoT) and Bluetooth Low Energy (BLE) are discussed in detail. These specifications are regularly updated to meet evolving technology and market need. RF performance specification of LTE-M and NB-IoT are almost similar and in most of the bands devices on these two standards can co-exist.

## 4.1. Selected Bands

3GPP has specified 50 frequency bands for LTE. These bands support different bandwidths, some bands support frequency division duplex (FDD) and other support time division duplex (TDD). A typical user equipment will support subset of these frequency bands depending upon the targeted market. Choice of operating band defines the user equipment capability to roam between national and international operators. In line to roaming capabilities channel bandwidth also dictates the choice of bands since different bands have different channel bandwidths.

LTE-M is an upcoming standard and device manufacturer's want to target the global market, as a result band selection criteria is largely influenced by the geographic location. For the wearable device transceiver six bands are selected which are listed in Table 2 [20] based upon the geographic location. Another criteria for band selection of LTE-M is that each band should support 1.4 MHz channel bandwidth (occupied channel bandwidth is 1.14 MHz). Out of the six selected bands five can support LTE-M devices. LTE-M is an upcoming standard with the specification process on going so it is believed that band 17 will be made available to support LTE-M devices in the upcoming release.

Table 2. Bands selected for LTE-M

| E-UTRA bands | UL bands | | DL bands | | Support available | |
|---|---|---|---|---|---|---|
| | $F_{UL_{Low}}$ | $F_{UL_{High}}$ | $F_{DL_{Low}}$ | $F_{DL_{High}}$ | LTE-M | NB-IoT |
| 3 | 1710 MHz | 1785 MHz | 1805 MHz | 1880 MHz | Yes | Yes |
| 4 | 1710 MHz | 1755 MHz | 2110 MHz | 2155 MHz | Yes | No |
| 12 | 699 MHz | 716 MHz | 729 MHz | 746 MHz | Yes | Yes |
| 13 | 777 MHz | 787 MHz | 746 MHz | 756 MHz | Yes | Yes |
| 17 | 704 MHz | 716 MHz | 734 MHz | 746 MHz | No | Yes |
| 20 | 832 MHz | 862 MHz | 791 MHz | 821 MHz | Yes | Yes |

LTE-M is designed to operate in both full duplex FDD and half duplex FDD while NB-IoT can only operate in half duplex FDD mode. On the other hand BLE operates on full duplex in 2.4 GHz industrial, scientific and medical (ISM) radio band (2400 - 2483.5 MHz) [19].

## 4.2. Receiver Specifications

Radio receiver function is to receive the desired signal in the presence of unwanted signals and to keep the noise figure at a constant level. Different tests are recommended by 3GPP, SIG to make sure that the receiver is able to demodulate a desired signal and maintain at least a minimum throughput level. [7]

### *4.2.1. Sensitivity*

Sensitivity is the minimum received signal power at the antenna port which provides sufficient SNR for the specified modulation and coding scheme to achieve throughput of 95 % (for LTE-M and NB-IoT). [20]

When we are designing a receiver sensitivity value is mentioned and we have to keep our NF at a level which will achieve a certain SNR specified by the modulation scheme. Sensitivity values for LTE-M, NB-IoT and BLE are listed in Table 3 [18, 20, 19]. [7]

Table 3. Specification of reference sensitivity for LTE-M[1], NB-IoT[2]& BLE

<table>
<tr><th rowspan="2">E-UTRA bands</th><th>LTE-M</th><th>NB-IoT</th><th>BLE</th></tr>
<tr><th>(dBm)</th><th>(dBm)</th><th>(dBm)</th></tr>
<tr><td>3</td><td>- 100</td><td>- 108.2</td><td rowspan="6">- 70</td></tr>
<tr><td>4</td><td>- 103</td><td>N/A</td></tr>
<tr><td>12</td><td>- 100</td><td>- 108.2</td></tr>
<tr><td>13</td><td>- 100</td><td>- 108.2</td></tr>
<tr><td>17</td><td>$- 100^{*FDD}$</td><td>- 108.2</td></tr>
<tr><td>20</td><td>- 100.5</td><td>- 108.2</td></tr>
</table>

'*' shows that the value is scaled from some other band. Sensitivity values for band 17 is scaled from FDD band since the support for band 17 is not available currently. From Equation (1) we can see that increasing the bandwidth decreases the sensitivity which is the only reason why LTE-M have decreased sensitivity compared to NB-IoT as mentioned in Table 3. On the other hand bluetooth have a decreased sensitivity of -70 dBm due to decreased link range and high requirement of SNR [19].

### *4.2.2. Intermodulation*

Intermodulation (IMD) test is performed while designing a receiver to test the receiver's ability of receiving a desired signal in the presence of unwanted signal (at

[1] 3GPP, 36.101, section 7.3.1E
[2] 3GPP, 36.521, section 7.3F.1

the same frequency). IMD products arise because of the nonlinearity in receivers. The level of these intermodulation products is determined by the linearity of receivers and the power of interfering signals. If a receiver is operating in nonlinear region then the amplitude of intermodulation products will be large also if the power level of interfering signals are high then the IMD products will also have high power level. [7]

3GPP and SIG have specified the signal level and offset frequencies of interfering signals which will produce IMD products, along with signal level of desired signal. Intermodulation test specification for LTE-M & NB-IoT are listed in Table 4 [18, 20]. One interfering signal is continuous wave and the other interfering signal is modulated signal having a bandwidth of 1.4 MHz.

Table 4. Specification of intermodulation test for LTE-M[3]& NB-IoT[4]

| Receiver parameters | Units | LTE-M | NB-IoT |
|---|---|---|---|
| $P_{interferer-1}(CW)$ | dBm | -46 | -46 |
| $P_{interferer-2}(Modulated)$ | dBm | -46 | -46 |
| $BW_{interferer-2}$ | MHz | 1.4 | 1.4 |
| $F_{interferer-1}(Offset)$ | MHz | $\pm$ 2.8 | $\pm$ 2.2 |
| $F_{interferer-2}(Offset)$ | MHz | $\pm$ 5.6 | $\pm$ 4.4 |
| E-UTRA band 3 | dBm | - 88 | - 102.2 |
| E-UTRA band 4 | dBm | - 91 | N/A |
| E-UTRA band 12 | dBm | - 88 | - 102.2 |
| E-UTRA band 13 | dBm | - 88 | - 102.2 |
| E-UTRA band 17 | dBm | - 88$^{*FDD}$ | - 102.2 |
| E-UTRA band 20 | dBm | - 88.5 | - 102.2 |

Intermodulation test specification for BLE is specified by SIG and it is mentioned in Table 5 [19]. Wanted signal for BLE intermodulation test is 6 dB above sensitivity.

Table 5. Specification of intermodulation test for BLE

| Receiver parameters | Units | BLE | |
|---|---|---|---|
| | | 1 MHz Tx bandwidth | 2 MHz Tx bandwidth |
| $P_{interferer-1}(CW)$ | dBm | -50 | -50 |
| $P_{interferer-2}(Modulated)$ | dBm | -50 | -50 |
| $BW_{interferer-2}$ | MHz | 1 | 2 |
| $F_{interferer-1}(Offset)$ | MHz | $\pm f_1$ | $\pm f_1$ |
| $F_{interferer-2}(Offset)$ | MHz | $\pm f_2$ | $\pm f_1$ |
| $P_{wanted}$ | dBm | -64 | -64 |
| $F_{wanted}$ | MHz | $2 * f_1 - f_2$ | $2 * f_1 - f_2$ |

[3]3GPP, 36.101, section 7.8.1
[4]3GPP, 36.521, section 7.8.1F

When performing the test for 1 MHz transmission bandwidth frequencies $f_1$ and $f_2$ are chosen in way that $|f_1 - f_2| = n * 1$ MHz where $n$= 3, 4, 5. Similarly, when performing the test for 2 MHz transmission bandwidth frequencies $f_1$ and $f_2$ are chosen in way that $|f_1 - f_2| = n * 2$ MHz where $n$= 3, 4, 5.

Modulated interferer is chosen to see the impact of intermodulation distortion on all the sub-carriers If both the signal were continuous wave then only few sub-carriers would be effected and the performance impact can not be evaluated correctly. [7]

#### *4.2.3. Out-of-band blocking*

Receiver out-of-band band blocking characteristics are designed as a metric to evaluate receiver performance in the presence of higher level out-of-band signals. Out-of-band band blocking is defined for an unwanted CW interfering signal falling more than 15 MHz below or above the user equipment receive band. Specification of reference signal power for blocking test is mentioned in Table 6 [18, 20] along with blocking signal power levels and their corresponding offset frequencies. [7]

Table 6. Specification of out-of-band blocking signals for LTE-M[5] & NB-IoT[6]

| Receiver parameters | Units | LTE-M | NB-IoT |
|---|---|---|---|
| $P_{interferer(Range-1)}$ | dBm | -44 | -44 |
| $P_{interferer(Range-2)}$ | dBm | -30 | -30 |
| $P_{interferer(Range-3)}$ | dBm | -15 | -15 |
| $F_{Range-1}$ | MHz | $(F_{DL_{low}} - 15$ to $F_{DL_{low}} - 60)$ & $(F_{UL_{high}} + 15$ to $F_{UL_{high}} + 60)$ | Same as LTE-M |
| $F_{Range-2}$ | MHz | $(F_{DL_{low}} - 60$ to $F_{DL_{low}} - 85)$ & $(F_{UL_{high}} + 60$ to $F_{UL_{high}} + 85)$ | Same as LTE-M |
| $F_{Range-3}$ | MHz | $(F_{DL_{low}} - 85$ to 1 MHz) & $(F_{UL_{high}} + 85$ to 12750 MHz) | Same as LTE-M |
| E-UTRA band 3 | dBm | - 94 | - 102.2 |
| E-UTRA band 4 | dBm | - 97 | N/A |
| E-UTRA band 12 | dBm | - 94 | - 102.2 |
| E-UTRA band 13 | dBm | - 94 | - 102.2 |
| E-UTRA band 17 | dBm | - $94^{*FDD}$ | - 102.2 |
| E-UTRA band 20 | dBm | - 94.5 | - 102.2 |

Interfering signal power level varies from -44 dBm (at an offset of 15 to 60 MHz) to -15 dBm (at an offset of 85 to 12750 MHz) for both LTE-M & NB-IoT. These

[5] 3GPP, 36.101, section 7.6.2
[6] 3GPP, 36.521, section 7.6.2F

interfering signals when captured by the antenna causes desensitization of the receiver. The amount of desensitization caused depends upon the linearity of the receiver and the power level of blocking signal. As these signal fall outside the desired band of operation these signals can be filtered using band pass filtering and desensitization scenario can be avoided. When performing the out of band blocking test the reference signal level is increased 6 dB from the reference sensitivity value. [7]

Table 7. Specification of out of band blocking signals for BLE

| Interfering signal frequency range (MHz) | Interfering signal power (dBm) | Measurement bandwidth (MHz) |
|---|---|---|
| 30 - 2000 | - 30 | 10 |
| 2003 - 2399 | - 35 | 3 |
| 2484 - 2997 | - 35 | 3 |
| 3000 - 12750 | - 30 | 25 |

Out of band blocking specification for bluetooth are listed in Table 7 [19]. Interfering signal power level range from -30 to -35 dBm and measurement bandwidth is also different for different offset frequencies. Reference signal level in out of band blocking test is 3 dB above sensitivity level and the criteria which is to be met is BER $\leq 0.1$ %.

### *4.2.4. In-band blocking*

In band blocking defines the receiver ability to extract the useful information out of the modulated signal in the presence of unwanted interfering signal falling into the UE receive band, or into the first 15 MHz below or above the UE receive band. [7]

Table 8. Specification of in-band blocking signals for LTE-M[7] & NB-IoT[8]

| Receiver parameters | Units | LTE-M | NB-IoT |
|---|---|---|---|
| Power in transmission bandwidth | dBm | Refsens + 6 | Refsens + 6 |
| $BW_{interferer}$ | MHz | 1.4 | 5 |
| $P_{interferer}$ | dBm | -56 | -56 |
| $F_{offset}$ | MHz | $\pm$ 2.8125 | $\pm$ 7.505 |
| E-UTRA band 3 | dBm | - 94 | - 102.2 |
| E-UTRA band 4 | dBm | - 97 | N/A |
| E-UTRA band 12 | dBm | - 94 | - 102.2 |
| E-UTRA band 13 | dBm | - 94 | - 102.2 |
| E-UTRA band 17 | dBm | - 94$^{*FDD}$ | - 102.2 |
| E-UTRA band 20 | dBm | - 94 | - 102.2 |

[7] 3GPP, 36.101, section 7.6.1
[8] 3GPP, 36.521, section 7.6.1F

Reference signal power level along with blocker signal power level, offset frequency and interferer bandwidth are mentioned in Table 8 [18, 20]. Unlike out of band blocking test where CW interferer's are present, in band blocking test use modulated signals which occupy the same bandwidth as that of desired signal in case of LTE-M and occupy 5 MHz bandwidth in case of NB-IoT. Reference signal in this case is 6 dB above the sensitivity level, same as out of band blocking test. While the interfering signal have a decreased power of -56 dBm when compared with out of band specifications. In band blocking increase the noise figure of receiver by reciprocal mixing with the local oscillator signal. [7]

#### *4.2.5. Narrow band blocking*

Narrow band blocking is a measure of the LTE-M receiver's ability achieve minimum throughput in the presence of an unwanted narrow band interferer at a frequency offset that is less than the channel spacing. [7]

Table 9. Specification of narrow band blocking signals for LTE-M[9]& NB-IoT

| Receiver parameters | Units | LTE-M | NB-IoT |
|---|---|---|---|
| Power in transmission bandwidth | dBm | Reference sensitivity + 22 | N/A |
| $P_{interferer}$ | dBm | -55 | |
| $F_{interferer}$ | MHz | 0.9075 | |
| E-UTRA band 3 | dBm | -78 | |
| E-UTRA band 4 | dBm | -81 | |
| E-UTRA band 12 | dBm | -78 | |
| E-UTRA band 13 | dBm | -78 | |
| E-UTRA band 17 | dBm | -78$^{*FDD}$ | |
| E-UTRA band 20 | dBm | -78 | |

3GPP specification of narrow band blocking test for LTE-M is listed in Table 9 [18, 20] and no specification of NB-IoT, BLE are available for this test. In this test reference signal level is 22 dB above the sensitivity level and the offset frequency of interfering signal is 0.9075 MHz. Interfering signal in this case is CW. Interfering signals in this case doesn't desensitize the receiver rather it increases the noise floor by reciprocal mixing with the local oscillator signal.

### 4.3. Transmitter Specifications

LTE-M transmitter specifications are designed to satisfy requirement concerning the output signal power, quality of desired signal and also the unwanted emissions which can interfere with other operational radio devices. The following subsections explain important transmitter specification for 1.4 MHz LTE-M channel.

[9]3GPP, 36.101, section 7.6.3

#### *4.3.1. Maximum Output Power*

Transmitter maximum output power is the most important criteria of transmitter design as it influences the interference experienced by the users using the same frequency channel in another cell. It also has impact on the magnitude of unwanted emissions. To mitigate the undesired phenomenon's it is desirable that the transmitters must set its maximum output power according to the specifications of 3GPP as mentioned in Table 10 [18, 20]. [7]

3GPP has defined the requirements for two power classes of user equipment (UE). For the selected bands the nominal output power is +23 dBm for class 3 and for class 5 the nominal output power is +20 dBm. Tolerance margin of 2 dB is also made available because of power fluctuation. When measuring the output power UE must utilize QPSK modulation scheme as specified by 3GPP. Also if we are using LTE-M with 1.14 MHz occupied channel bandwidth which consists of maximum of 6 resource blocks, the test should be conducted twice once with only one resource block and second time when all the resource blocks are utilized. Transmitter should be able to generate output power within the defined tolerance limit for both the tests. [7]

Table 10. Specification of maximum output power for LTE-M[10]& NB-IoT[11]

| E-UTRA band | LTE-M & NB-IoT | | | |
|---|---|---|---|---|
| | Class 3 (dBm) | Tolerance (dB) | Class 5 (dBm) | Tolerance (dB) |
| 3 | 23 | ± 2 | 20 | ± 2 |
| 4 | 23 | ± 2 | 20 | ± 2 |
| 12 | 23 | ± 2 | 20 | ± 2 |
| 13 | 23 | ± 2 | 20 | ± 2 |
| 17 | 23 | ± 2 | 20 | ± 2 |
| 20 | 23 | ± 2 | 20 | ± 2 |

SIG has specified four classes of BLE devices. Maximum output power for BLE devices are listed in Table 11 [19].

Table 11. Specification of maximum output power for BLE

| Power class | Maximum output power (dBm) |
|---|---|
| 1 | + 20 |
| 1.5 | + 10 |
| 2 | + 4 |
| 3 | 0 |

[10]3GPP, 36.101, section 6.2.2E
[11]3GPP, 36.521, section 6.2.2F

#### *4.3.2. Error Vector Magnitude*

Transmitted signal must fulfill the requirement of signal quality. Error vector magnitude measures the effect of all the nonidealites on the transmitted symbol with reference from the actual symbol. The requirement of LTE-M signal quality are listed in Table 12 as per 3GPP specifications for different modulation schemes. Whereas, NB-IoT can only utilize QPSK/BPSK so the requirement of EVM is 17.5 %. EVM is used to determine the maximum possible modulation order and code rate which can be opted by the transmitter. Table 12 [18] also mentions the required SNR to achieve the EVM requirement. Specification of EVM are designed to reflect only 5 % loss of throughput in typical RF scenarios. [7]

Table 12. Specification of error vector magnitude

| Modulation scheme | EVM requirement | Required SNR |
|---|---|---|
| QPSK | 17.5 % | 15.15 dB |
| 16-QAM | 12.5 % | 18 dB |
| 64-QAM | 8 % | 21.9 dB |

#### *4.3.3. Spectrum Emission Mask*

In ideal conditions radio transmitters should emit everything within the transmission band but this is not the case in practice. Out of band emissions are classified in many ways out of which one is the spectrum emission mask (SEM). SEM defines the limit of out of band emissions compared to in channel emissions. SEM is measured as a function of offset frequency from the corner of channel and the specification for LTE-M normalized to 1 Hz are shown in Figure 14 [18]. [7]

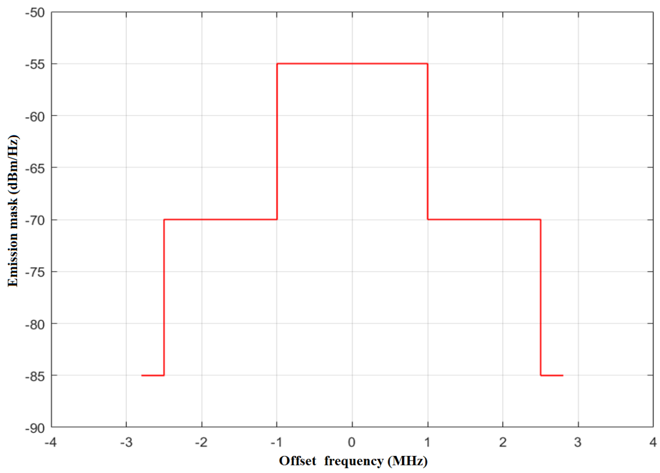


Figure 14. Specification of SEM for 1.4 MHz LTE-M signal[12].

[12] 3GPP, 36.101, section 6.6.2

Emission level at an offset frequency of ± 1 MHz is -55 dBm/Hz and similarly when the offset frequency is between ± 1-2.5 MHz the level of emissions is -70 dBm/Hz. Measurement bandwidth for offset frequency below 1 MHz is 30 kHz and for offset frequencies greater than 1 MHz measurement bandwidth is 1 MHz.

SEM specification for NB-IoT is listed in Table 13 [20]. Specification of LTE-M and NB-IoT are designed in a similar way.

Table 13. Specification of SEM for NB-IoT[13]

| Offset frequency (kHz) | Normalized emission limit (dBm/Hz) | Measurement bandwidth (kHz) |
|---|---|---|
| ± 0 | - 19 | 30 |
| ± 100 | - 50 | 30 |
| ± 150 | - 53 | 30 |
| ± 300 | - 74 | 30 |
| ± 500 - 1700 | - 80 | 30 |

Table 13 shows the offset frequency of unwanted emissions, normalized level of unwanted emissions and the measurement bandwidth. Out of band specification for BLE is not specified by SIG. It is the responsibility of device manufacturer to design the device in such a way that out of band emissions criteria for the intended region of sale is met [19].

#### *4.3.4. Adjacent Channel Power Ratio*

Another way of specifying out of band emissions in cellular communication is the adjacent channel power ratio (ACPR). ACPR measures the power leaked into adjacent channel and ACPR specifications are designed in a way that leaked power doesn't effect the performance of radio's operating in adjacent channel. It is defined as the ratio of power in the desired band to the power emitted in adjacent band. 3GPP specification of LTE-M ACPR are tabulated in Table 14 [18]. [7]

Table 14. Specification of ACPR for LTE-M[14]

| | Channel bandwidth (MHz) | ACPR (dB) | Offset frequency (MHz) | Measurement bandwidth (MHz) |
|---|---|---|---|---|
| E-UTRA$_{ACPR}$ | 1.4 | 30 | ± 1.4 | 1.08 |
| UTRA$_{ACPR}$ | 1.6 | 33 | ± 1.34 | 1.28 |

Table 14 shows that for LTE-M power emitted in adjacent E-UTRA channel at an offset frequency of ±1.4 MHz should be 30 dB below than the power emitted in desired channel. Similarly, the power emitted in the adjacent UTRA channel at an offset

[13] 3GPP, 36.521, section 6.6.2.1F
[14] 3GPP, 36.101, section 6.6.2.3

frequency of $\pm 1.34$ MHz should be 33 dB below the power emitted in wanted channel and the measurement bandwidth is 1.28 MHz.

Table 15 [20] specifies ACPR requirements for NB-IoT. Table 15 shows that the power in desired channel should be 20 dB above the power emitted in adjacent channel when measuring the interference with GSM at an offset frequency (measured from the center of channel) of $\pm$ 200 kHz while ACPR should be 37 dB when measuring interference to 3G system at an offset frequency of $\pm$ 2500 kHz. Measurement bandwidth also differs for both the GSM and 3G standard.

Table 15. Specification of ACPR for NB-IoT[15]

| | Channel bandwidth (kHz) | ACPR (dB) | Offset frequency (kHz) | Measurement bandwidth (kHz) |
|---|---|---|---|---|
| GSM $_{ACPR}$ | 200 | 20 | $\pm$ 200 | 180 |
| UTRA $_{ACPR}$ | 200 | 37 | $\pm$ 2500 | 3840 |

ACPR requirement in case of BLE is mentioned in Table 16 [19] and it is specified for two transmission bandwidths and the offset between channels should be at least 2 MHz in case of 1 MHz transmission bandwidth and it should be at least 4 MHz for transmission bandwidth of 2 MHz. If the transmission bandwidth is 1 MHz and the adjacent channel is at an offset of 2 MHz then the leaked power in that channel should be lower than 20 dB compared to desired channel.

Table 16. Specification of ACPR for BLE

| Transmission bandwidth | | | |
|---|---|---|---|
| 1 MHz | | 2 MHz | |
| Frequency offset (MHz) | ACPR (dB) | Frequency offset (MHz) | ACPR (dB) |
| 2 | 20 | 4 | 20 |
| $\geq 3$ | 30 | 5 | 20 |
| | | $\geq 6$ | 30 |

#### *4.3.5. Spurious Emission*

Spurious emission criteria is also used to specify emissions which are outside the desired band. 3GPP specifications for spurious emission are mentioned in Table 17 [18, 20].

Spurious emission arise because of harmonic emissions and intermodulation products. Spurious emission limit apply to frequency range offsets listed in Table 17. Spurious emission limit must be satisfied to protect receiver of the same UE or receiver of other devices operating nearby the transmitter.

[15] 3GPP, 36.521, section 6.6.2.3F

Table 17. Specification of spurious emission for LTE-M[16]& NB-IoT[17]

| Frequency range | Maximum level | Measurement bandwidth |
|---|---|---|
| 9 kHz $\leq f <$ 150 kHz | -36 dBm | 1 kHz |
| 150 kHz $\leq f <$ 30 MHz | -36 dBm | 10 kHz |
| 30 MHz $\leq f <$ 1000 MHz | -36 dBm | 100 kHz |
| 1 GHz $\leq f <$ 12.75 GHz | -30 dBm | 1 MHz |

Transmitting something on the same frequency of reception blocks the receiver in FDD system. In case of HD-FDD since the transmitter and receiver do not operate at the same time so blocking of own receiver doesn't occur, but these specifications must be met to protect nearby receivers. [7]

[16]3GPP, 36.101, section 6.6.3
[17]3GPP, 36.521, section 6.6.3F

# 5. TRANSCEIVER DESIGN

People have been using cellular communication for more than two decades and it has become a need. Wireless devices mostly rely on the batteries which are recharged frequently. With the advancement of technology a new generation of LPWAN devices are emerging, driven by the decrease in cost and power required for operation. Design of RF transceiver is the critical part in the designing of wirelessly operated devices.

Aim of this thesis is to develop RF transceiver for low powered category 'M1' devices (LTE-M) specified by 3GPP. Design of transmitter and receiver from specification is discussed in the following sections.

## 5.1. Receiver Design

Implementation of receiver is possible from different architectures as discussed in Chapter 3, the most important concern is that each architecture must be able to achieve a performance level defined by the standard. Figure 15 shows the block diagram of direct conversion receiver which is used in the analysis. Direct conversion receiver was adopted because of its advantages over superheterodyne receiver i.e. small size. For modelling purpose the whole receiver is divided into three parts SP6T switch, bandpass filtering, RFIC. Various design parameters have been discussed in Chapter 4, and in the following sub-sections design of receiver from the specification is explained.

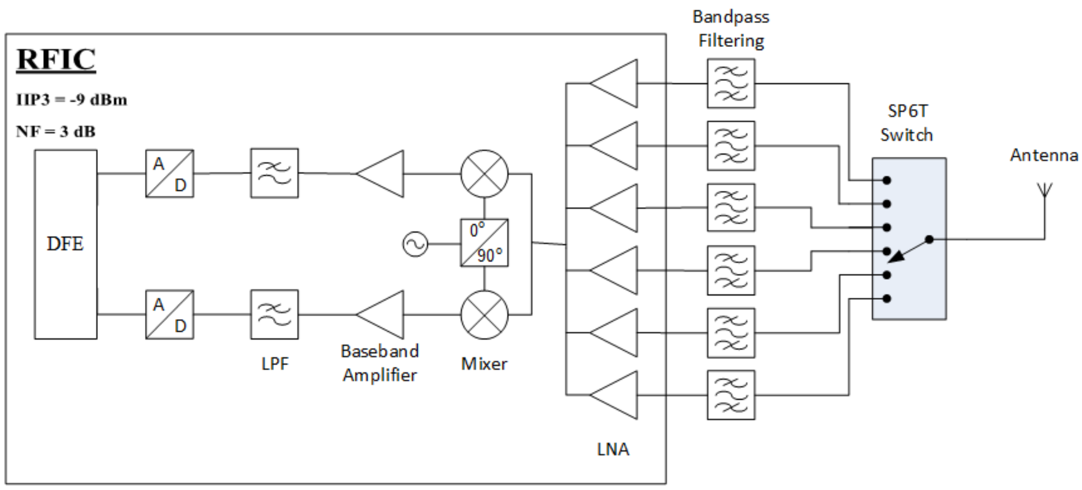


Figure 15. Block diagram of LTE-M receiver.

### 5.1.1. *Sensitivity*

The minimum signal level specification by 3GPP are listed in Table 3. As mentioned in Equation (1) sensitivity is used to define the noise figure which the receiver can have to achieve a required SNR. Equation (1) can be re-written as Equation (15) to calculate the maximum noise figure of the system. [12]

$$NF = Rx_{sensitivity} + 174 - 10 * log(BW) - SNR \tag{15}$$

As described earlier occupied channel bandwidth of LTE-M signal is 1.14 MHz, required SNR is -1 dB (QPSK 1/3 modulation) [7], reference sensitivity is -100 dBm (band 3,12,13,17) and from Equation (15) the maximum allowed $NF$ is found to be 13.43 dB ($NF = -100 + 174 - 10 * log(1.14 * 10^6) - (-1) = 13.43$ dB). Noise figure of the cascaded system can be calculated by Equation (16) [8].

$$NF_{Total} = 10log(F_1 + \frac{F_2 - 1}{G_1} + ... + \frac{F_n - 1}{G_1 * ... * G_{n-1}}), \tag{16}$$

where $NF_{Total}$ is the total noise figure of the receiver, $F_1, F_2, ..., F_n$ is the noise factor of components in the receiver chain and their corresponding gain is $G_1, G_2, ..., G_{n-1}$. Direct conversion architecture is used for the receiver design and NF values of the available commercial components were used to calculate the cascaded noise figure.

- SP6T switch with 1 dB NF i.e. coming from the losses of passive components.
- Bandpass filter with 3 dB NF including ripple.
- RFIC (includes all the RF components i.e. LNA, Mixer, LPF, BB-Amplifier) with 3 dB NF.

When utilizing the commercial components the total noise figure of the system becomes 7 dB which is 6.5 dB lower than the allowed noise figure calculated from Equation (15). Ripple of the bandpass filter results in variable noise figure at different frequencies so in ADS a frequency sweep was done over band 3 (1805-1880 MHz) to calculate $NF$ at all the frequencies as shown in Figure 16. Output SNR for all the frequencies in band 3 is almost 7 dB above the -1 dB threshold which verifies the calculation of noise figure.

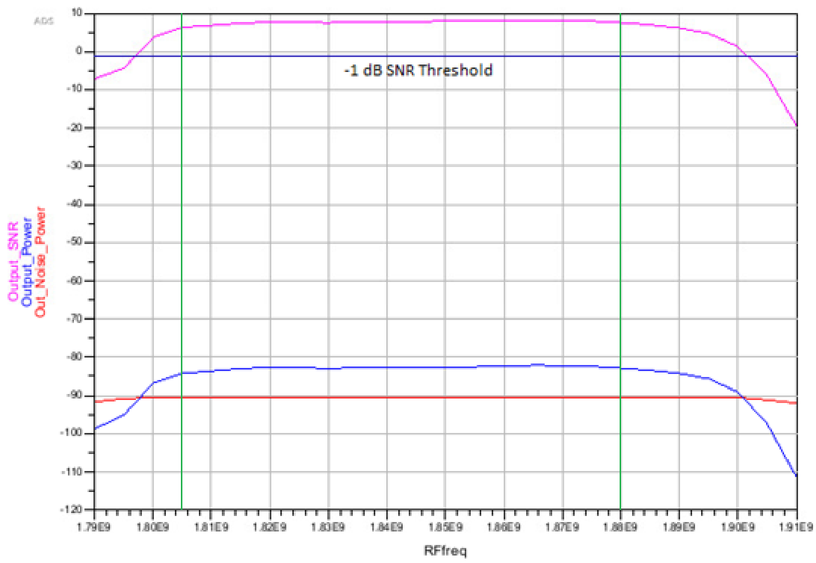


Figure 16. Frequency sweep of input signal at sensitivity level.

### *5.1.2. Third-Order Intercept Point*

Receiver third order intercept point is the metric that defines its linearity. Intermodulation test is carried out by applying interfering signals at the input of receiver as specified in Table 4. Interfering signal mix with each other and their third order intermodulation distortion product lies on the same frequency as the information signal and it increases the noise floor of the receiver. Receiver must achieve the same performance requirement as mentioned earlier in sensitivity test [7]. [12]

Third order intermodulation test bench in ADS was designed in such a way that the third order product rather than lying on the information signal lies at a slight offset from the information signal. As per 3GPP specification one interfering signal and the information signal should be modulated signal but these were taken as tone signals for ease in analysis. This test is performed to determine the linearity required by the nonlinear components.

After reading several research papers and data sheets of commercially available products a nominal value of TOI (-9 dBm) at the input of RFIC is chosen to be a realistic receiver performance level. Since the components preceding the RFIC in the receiver chain are lossy, the TOI is raised 4 dB resulting into a TOI value of -5 dBm at the input of receiver chain. Level of third order intermodulation product at the input can be found by using Equation (12) as $IM3_{Input} = 3 * (-46) - 2 * (-5) = -128$ dBm.

Calculation of intermodulation test are also verified by simulation in ADS software. Figure 17 shows the input to the receiver according to 3GPP specifications mentioned in Table 4. Interfering signals which are supposed to be at 2.8 and 5.6 MHz away from the desired signal are shifted further 0.3 MHz apart so that the third order intermodulation product lies at some offset as can be seen from output shown in Figure 18.

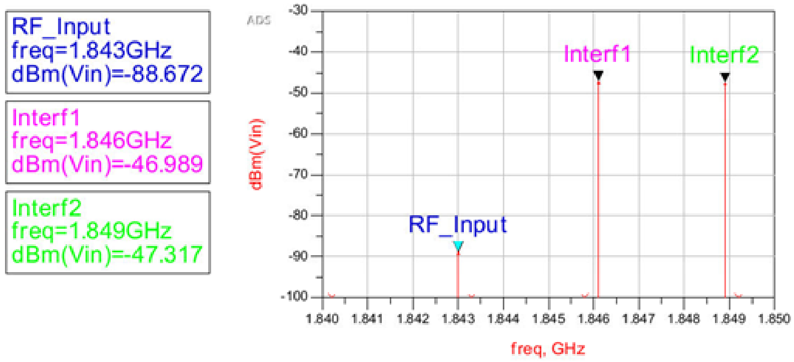


Figure 17. Receiver input for intermodulation test.

Intermodulation product at the output is found to be -108.32 dBm (when the gain is assumed 16 dB) from simulation which is approximately the same as calculated theoretically. RMS value of thermal noise and IMD are added to get the overall interference level and then it is subtracted from the desired signal to get the SNR. Output SNR in this case is 21.397 dB which is far above than the required SNR by QPSK 1/3 modulation scheme.

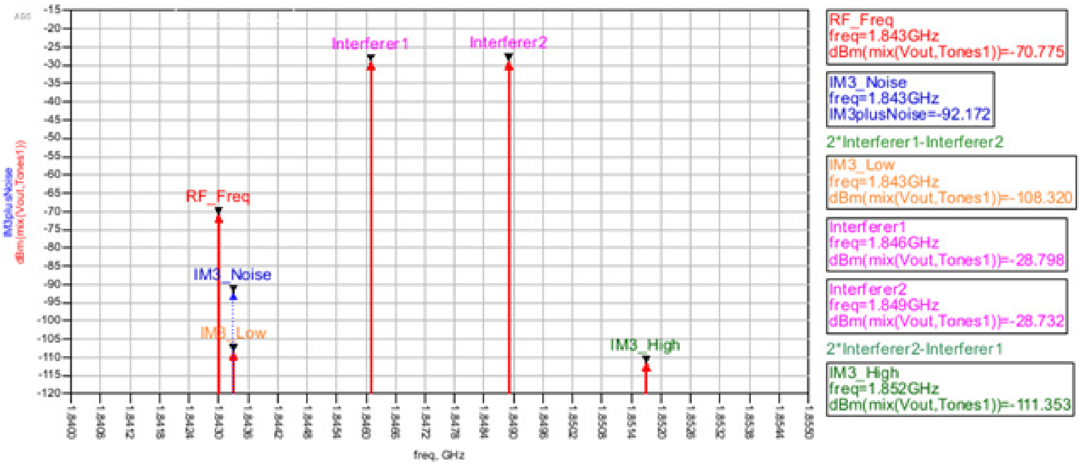


Figure 18. Receiver output for intermodulation test.

Theoretical and simulation results for intermodulation test shows that RFIC input TOI value of -9 dBm is sufficient for the receiver to achieve the performance requirement specified by 3GPP.

### *5.1.3. Desensitization*

Desensitization is the phenomenon where a strong blocking signal is able to increase the thermal noise floor of receiver due to additional noise paths established by the non-linearities. 3GPP has specified the offset frequencies and power levels of these out of band blocking signals as listed in Table 6. To prevent the receiver from desensitization these blocking signals should be attenuated or the receiver linearity should be increased. [17]

Increasing the linearity of the receiver increases the power consumption which can not be afforded in LPWAN devices so the other alternative of bandpass filtering was adopted to prevent the receiver from desensitization as shown in Figure 15.

To model desensitization TOI was chosen as a reference point. A general rule of thumb is that the 1 dB gain compression point (P1dB) is 10 dB lower than the TOI point. As can be seen from Figure 19 at -15 dBm input power the gain drops 1 dB from the nominal value. Receiver input TOI as theoretically calculated earlier was -5 dBm which gives the 1 dB gain compression point of -15 dBm. Theoretical and simulation results of 1 dB gain compression are similar. Similarly, the point where the noise figure of the receiver increases 1 dB was chosen to be 7 dB lower than the P1dB point which is a scaled version of [21]. The nominal value of noise figure is 7 dB but when the input power reaches -22 dBm the noise figure increases 1 dB as seen in Figure 19. This desensitization model is based on empirical data from [21]. However, desensitization is always dependant on the implementation type. Since, nonlinear mechanisms related to additive noise may not directly depend on the same components as the 1dB compression point. According to Table 6 out of band blocking signals are defined as interferers falling more than 15 MHz below or above the receive band. Signals that are between 15-85 MHz offset frequency have a maximum level of -30 dBm.

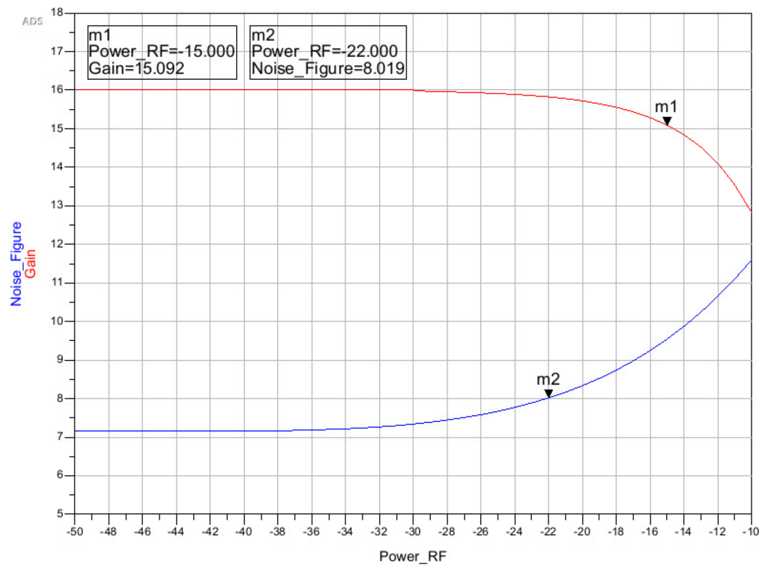


Figure 19. Noise figure and gain as a function of input power.

According to our desensitization model these signals can not cause any increase of noise floor, on the other hand signals lying at more than 85 MHz offset frequency have a power level of -15 dBm which causes an increase of 4 dB in the thermal noise floor. To prevent this increase of noise floor bandpass filtering providing attenuation of 8 dB for the blocking signals that are more than 85 MHz away from the receive band are used as shown in Figure 20. After attenuation the maximum level of these out of band blocking signal reduces to -23 dBm and desensitization is just 0.7 dB.

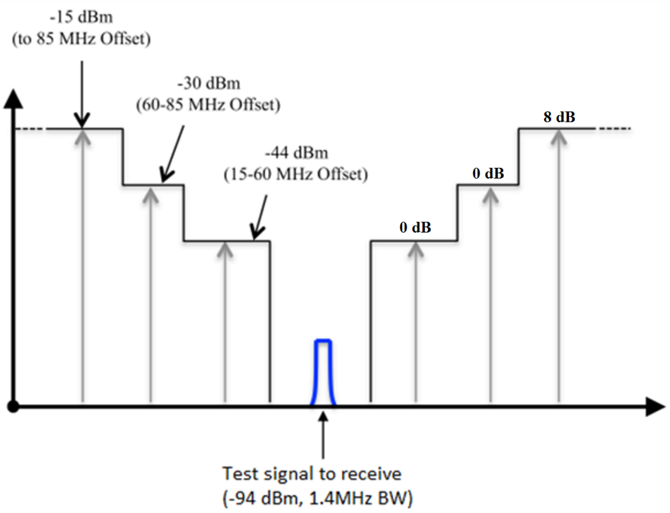


Figure 20. Filtering requirement for out of band blocking signals.

### *5.1.4. Reciprocal Mixing*

Reciprocal mixing (phase noise) is the process which increases the thermal noise floor of the receiver. Ideally the output of local oscillator signal should be an ideal sinusoidal waveform but this is far from the case in practice. The output of local oscillator with phase noise has side bands whose magnitude decreases as a function of offset frequency. When an interfering signal mixes with the LO signal, side bands of interfering signal extend into the receive band. Power of interfering signal determines the magnitude of these side bands i.e. higher level of blocking signal leads to more phase noise and vice versa. Figure 21 shows the phase noise in an oscillator output signal.

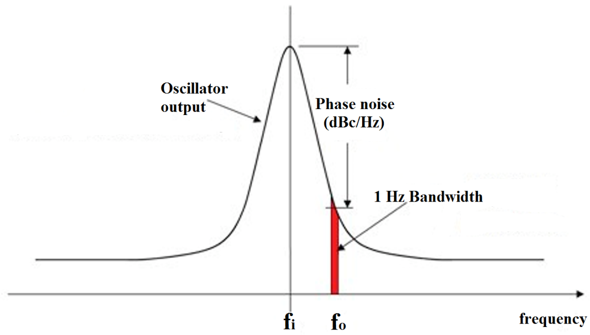


Figure 21. Phase noise in an oscillator output signal.

3GPP specification of out of band blocking translates into two phenomenon's first these blocking signals increase the noise floor of the receiver due to desensitization and secondly these blocking signal mix with the LO signal and further increases the noise floor by reciprocal mixing. Similarly, narrow band blocking signals and in band blocking signal also increase the noise floor by reciprocal mixing. Phase noise is produced when an interfering signal mixes with LO signal and creates high level side bands which increases the noise floor of desired signal frequency band. Phase noise is measured at frequency $f_o$ on 1 Hz bandwidth and the unit is dBc/Hz. Phase noise requirements for different 3GPP specification were determined and a synthesizer is chosen as synthesizer's normally have defined phase noise at different offset frequencies.

In the previous sub-section filtering was recommended to avoid the receiver from desensitization and the attenuated blocking signal along with others were used to find the requirements of phase noise. Phase noise requirements for synthesizer were found in a way that sum of desensitization, thermal noise and phase noise leaves 1 dB room for the modulation scheme to achieve its performance requirements. Phase noise of -134.2 dBc/Hz is required by the attenuated blocking signal which when integrated ($-134.2 + 10log(1.14 * 10^6)$ =-73.6) over the entire bandwidth (1.14 MHz) results into -73.6 dBc phase noise. To find the absolute level of phase noise blocking signal power (-23 dBm at 85 MHz offset) is added into -73.6 dBc which gives us -96.6 dBm noise level. The sum of all the noise's reaches a level of -94 dBm which is the same as reference signal so there is just 1 dB room for nonideality in case of out of band blocking and is shown in Figure 22.

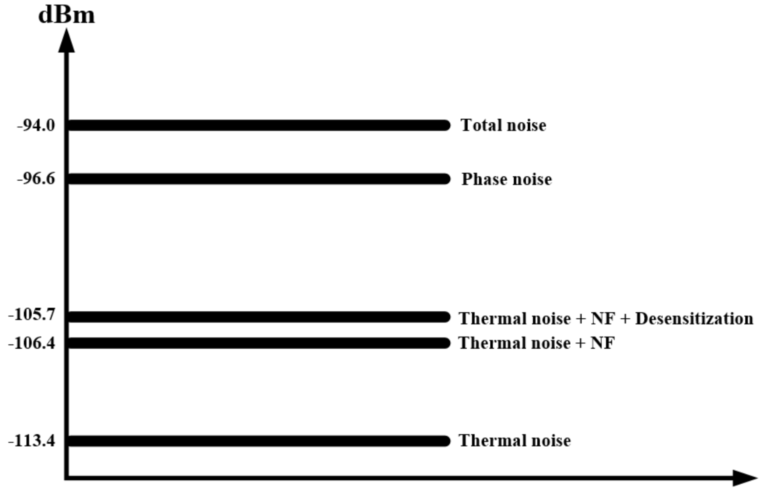


Figure 22. Sum of all the noise for blocking signal at frequency offset of 85 MHz.

In a similar manner other out-of-band blocking signals were used to determine the phase noise requirement at respective offset frequencies. Figure 23 plots phase noise (dBc/Hz) which is required by the LO as a function of offset frequency and it can be seen that as the interfering signal power increases the requirement of phase noise also increases.

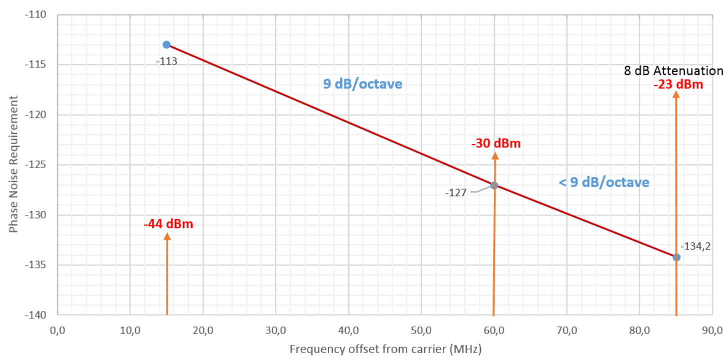


Figure 23. Phase noise requirement for out-of-band blocking signals.

Narrowband blocking and in band blocking specification of 3GPP does not cause any desensitization since the interfering signal level are much lower as mentioned in Table 8, 9 but these specification translates into phase noise. Similar to out of band blocking signal scenario phase noise of these two specifications are chosen in a way that 1 dB room is available for any other nonideality i.e. reference signal power equals the total noise power (phase noise, thermal noise).

Table 18 provides the requirement of phase noise as -102 dBc/Hz at an offset frequency of $\pm$ 2.8125 MHz for the in band blocking specification. Phase noise value of -102 dBc/Hz results into absolute noise of -96.4 dBm which when added with the thermal noise floor results into total noise floor of -94 dBm which is same as the reference signal power level.

Table 18. Phase noise calculation of in-band blocking

| Rx parameters | Units | Values |
|---|---|---|
| Blocker offset frequency | MHz | ± 2.8125 |
| Blocker power | dBm | - 56 |
| Synthesizer phase noise at offset frequency | dBc/Hz | -102 |
| Blocker phase noise power relative to carrier | dBc | -41.4 |
| Blocker phase noise absolute power | dBm | -96.4 |

Similarly, Table 19 provides the requirement of phase noise as -83.9 dBc/Hz at an offset frequency of ± 0.9075 MHz for the narrow band blocking specification. Reference signal level in case of narrow band blocking is 22 dB above the sensitivity level. Phase noise requirement in this case is relaxed because of increase in reference signal power. Phase noise value of -83.9 dBc/Hz results into absolute noise of -78.3 dBm which when added with the thermal noise floor results into total noise floor of -78 dBm. Total noise floor is equal to the reference signal power level, hereby meeting the -1 dB SNR criteria for chosen modulation scheme.

Table 19. Phase noise calculation of narrow band blocking

| Rx parameters | Units | Values |
|---|---|---|
| Blocker offset frequency | MHz | ± 0.9075 |
| Blocker power | dBm | - 55 |
| Synthesizer phase noise at offset frequency | dBc/Hz | -83.9 |
| Blocker phase noise power relative to carrier | dBc | -23.3 |
| Blocker phase noise absolute power | dBm | -78.3 |

Reciprocal mixing is the result of various blocking scenarios specified by 3GPP. The calculated requirements of phase noise at different offset frequencies are mentioned in Table 20 which are similar to the reference PLL of [22]. For the receiver to maintain the performance requirement defined by 3GPP, it is important that LO have the calculated level of phase noise at different offset frequencies.

Table 20. Phase noise requirements for LO

| Offset frequency (MHz) | Phase noise required (dBc/Hz) |
|---|---|
| ± 0.9075 | - 83.9 |
| ± 2.8125 | - 102 |
| ± 15 | - 113 |
| ± 60 | - 127 |
| ± 85 | - 134.2 |

PLL figure of merit is used as a metric of quantifying PLL quality, it plots the jitter variance of phase locked loop with respect to the power consumption. As the target in this thesis is to minimize the power consumption so a reference PLL which consumes around 2mW can be used. Figure 24 [22] shows the figure of merit of different available PLL's and we can see that the reference PLL has the lowest power consumption.

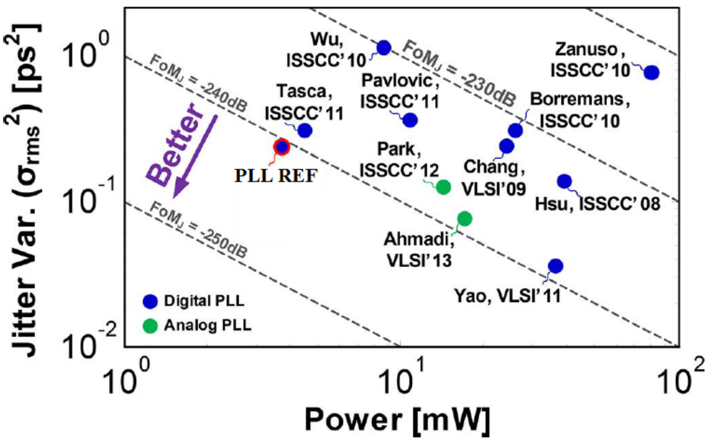


Figure 24. Figure of merit for PLL [22] © 2015 IEEE.

## 5.2. Transmitter Design

Transmitter design is dictated by two types of requirements one that concerns power level and quality of desired transmission and other that define unwanted transmissions. Transmitter can have the similar kind of architecture as the receiver i.e. superheterodyne or direct conversion. Selection of architecture depends upon the system requirements and available technologies. As mentioned earlier direct conversion architecture is used to keep the design of transceiver compact. Figure 25 shows the block diagram of direct conversion transmitter which is used in LTE-M transceiver design. [7]

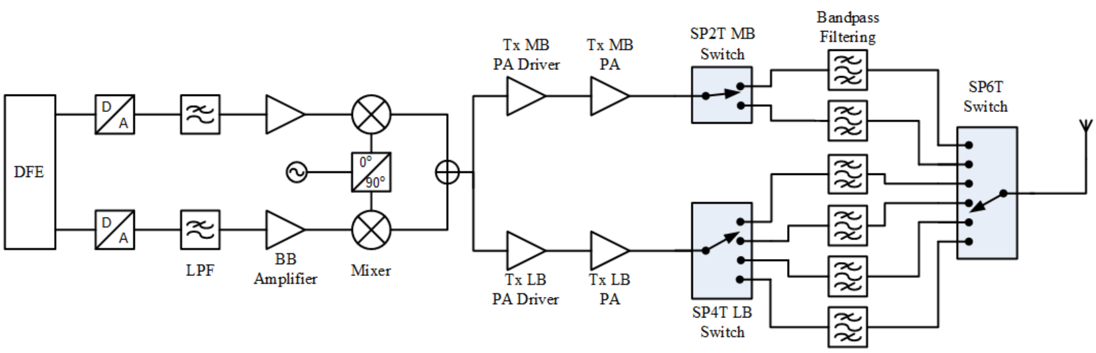


Figure 25. Block diagram of LTE-M transmitter.

Compared to the receiver which has dedicated LNA for each LTE-M band the transmitter doesn't have exclusive PA for every band rather it has two PA's one for low band (LTE band 3, 4) and the other for mid band (LTE band 12, 13, 17, 20). The size of

PA compared to LNA is big which is the reason why dedicated PA's were not used in the transmitter. Also two additional switches (SP2T, SP4T) were included in the design for switching with in the low and mid band. In the upcoming subsections design procedure of transmitter according to 3GPP specification is explained.

Components in the transmitter and receiver are alike but the scenarios associated with the components are different. In case of receiver demodulating a signal at sensitivity level is important while in case of transmitter transmitting a signal at maximum power is a critical factor. Similarly the quality of transmitted signal, emissions into adjacent band and spurious emission are important as well.

### *5.2.1. Error Vector Magnitude*

Error vector magnitude (EVM) also known as accuracy of modulated signal is a key parameter. EVM is the difference between the actual location of modulated signal compared to the theoretical location. The transmitter is designed to support QPSK modulation and the EVM requirement for it is 17.5 % [7]. EVM is degraded because of many factors which are discussed as follows. [12]

#### Phase Noise

Phase noise degrades the modulation accuracy of transmitted signal. Phase noise comes from the synthesizer which is up-converting the baseband signal to the RF frequency. Synthesizer used in the transmitter can be the same as in receiver in half duplex transceiver and it has -102 dBc/Hz in band phase noise at an offset of 500 kHz. When this phase noise is integrated over the entire signal bandwidth we get -41.43 dBc noise level compared to carrier. Influence of phase noise on EVM can be calculated from Equation (13) [12] and the degradation caused is 0.85 % i.e. $EVM_{PN} = \sqrt{10^{\frac{PhaseNoise}{10}}} = \sqrt{10^{\frac{-41.43}{10}}} = 0.0085.$

#### Quantization Noise

Quantization noise is produced when a digital to analog converter converts the discrete waveform into continuous signal. Resolution of DAC determines the amount of noise which will be associated with it. Equation (17) [23] determines the quantization noise of DAC and its unit is dBc.

$$QuantizationNoise = 6.02 * bits + 1.76 \tag{17}$$

Higher number of bits results in less error. 10 bit DAC is used in this transmitter and the quantization noise in this case equals -62 dBc as per Equation (17). Output power is backed off 10 dB from the maximum output power which increases the quantization noise to -52 dBc. Degradation of EVM due to quantization noise is 0.25 % as calculated from Equation (13) i.e. $EVM_{QN} = \sqrt{10^{\frac{QuantizationNoise}{10}}} = \sqrt{10^{\frac{-52}{10}}} = 0.0025.$

### I-Q Mismatch

IQ imbalances occur due to mismatches between I/Q branches of the transmitter chain. Theoretical phase difference between the I & Q branch is 90° and the gain of I & Q branches are the same. In the analog domain the difference of phase and gain between I & Q branch is never the same which results into unwanted image products. I-Q mismatch is defined in terms of image suppression/rejection which is the ratio of power in the image signal to that of desired signal. Image rejection ratio of -25 to -30 dBc can be achieved without any calibration techniques. To keep the design compact image rejection of -30 dBc is chosen which allows the system to have 0.27 dB gain imbalance and 1° phase imbalance and the corresponding EVM is calculated as 3.2 % from Equation (13) i.e. $EVM_{I-Q} = \sqrt{10^{\frac{ImageRejection}{10}}} = \sqrt{10^{\frac{-30}{10}}} = 0.032$. [12]

### Carrier Leakage

Carrier leakage in the transmitter will cause DC offset in the baseband I & Q channel [12]. Finite isolation between the LO and RF port causes the carrier signal to leak through. The leaked signal can not be filtered as it lies on the same frequency as desired signal and the effects are seen on EVM. Carrier leakage value of -35 dBc is achievable and the resulting EVM degradation is 1.77 % as calculated in Equation (13) i.e. $EVM_{CL} = \sqrt{10^{\frac{Carrierleakage}{10}}} = \sqrt{10^{\frac{-35}{10}}} = 0.0177$. [12]

### PA Non-linearity

If nonidealities are uncorrelated then the total EVM can be calculated from Equation (14) and can be expressed as Equation (18) [12]. Major contributing factor of EVM degradation is nonlinearity of the PA. In case of QPSK modulation scheme the total allowed EVM degradation is 17.5 %. Since we know the total allowed EVM degradation and we also know how much degradation is caused by other nonlinearities. So, almost 17 % of the total EVM budget can be allocated to the degradation caused by the PA.

$$EVM_{Total} = \sqrt{EVM_{PN}^2 + EVM_{QN}^2 + EVM_{I-Q}^2 + EVM_{CL}^2 + EVM_{PA}^2} \quad (18)$$

## *5.2.2. Maximum Output Power*

Transmitter maximum output power is an important criteria in the design of transmitter as it influences the interference experienced by other users using the same frequency channel in neighbouring cells. Along with the interference experienced by other user it also effects the unwanted emissions outside the defined frequency band. Maximum output power of mobile devices varies with the standard. Output power specification for LTE-M is listed in Table 10. Setting the output power accurately maximizes spectral efficiency. Requirements for maximum transmission power can be explained with the help of Figure 26. [7, 12]

Signal level at the output of digital to analog converter without backing off is 1 $V_{pp}$ (4 dBm). Output of DAC is backed off 10 dB which results into a signal of -6 dBm at the input of LPF as can be seen in Figure 26. Maximum output power at the antenna port for LTE-M is +23 dBm and to fulfil the criteria, input of low pass filter must be amplified. Signal is amplified with baseband amplifier, PA driver and PA. Along with amplifiers, output signal of DAC also pass through lossy components i.e. LPF, Mixer, Switch (SP2T/SP4T), Bandpass filter and SP6T switch which will attenuate the desired signal. To fulfil the criteria of maximum output power signal level at the following points were taken to determine the net gain.

- Input of LPF/Output of DAC.
- Input of bandpass filter.
- Input of antenna.

As we know that the signal level at the input of LPF is -6 dBm. If the amplifiers are able to provide 33 dB of net gain (sum of all losses and gains) between the DAC and bandpass filter, then the amplified signal will reach a level of +27 dBm at the input of bandpass filter which will be further attenuated 4 dB (nominal value of insertion loss for bandpass filter is 3 dB and SP6T switch is 1 dB). As a result the signal level at the antenna port will be +23 dBm.

Net gain of 33 dB can be achieved as nominal value of gain for PA driver is 10 dB and PA itself can have gain of 20 dB. Similarly, baseband amplifier can also provide 10 dB of gain. If the net gain provided by the transmitter is more than 33 dB then the level of unwanted emissions will increase.

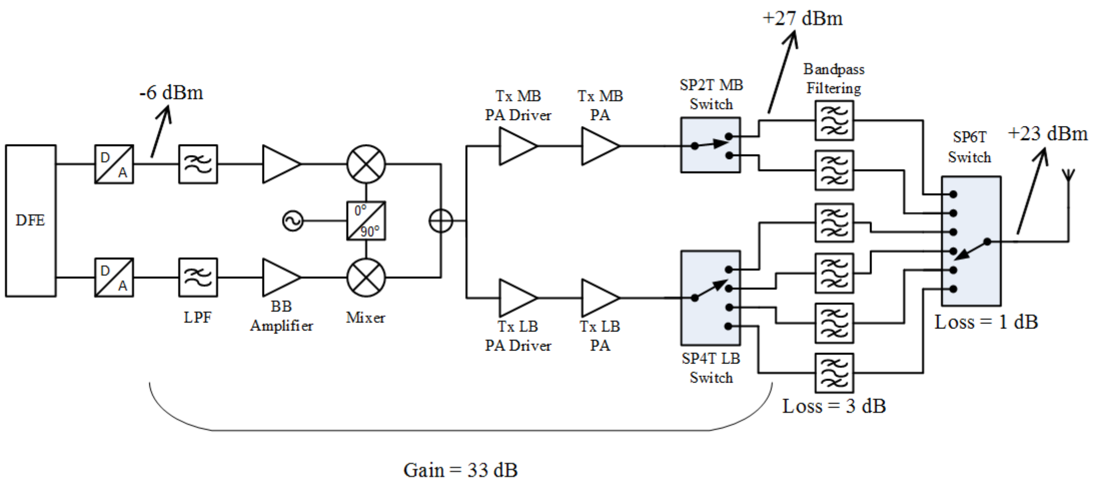


Figure 26. Block diagram of transmitter with power level.

### *5.2.3. Spurious Emissions*

Unwanted emissions outside the transmission band are tightly regulated to decrease the interference caused to nearby operating radio devices. In the transmitter these unwanted emissions mostly arise from aliasing, spurious tones of synthesizer, nonlinearity of PA and PA driver. [12]

### Aliasing

Digital to analog converter generates a staircase output of the input signal and it also produces aliases of the output signal at multiples of sampling frequency [24]. Aliasing can be visualized from Figure 27.

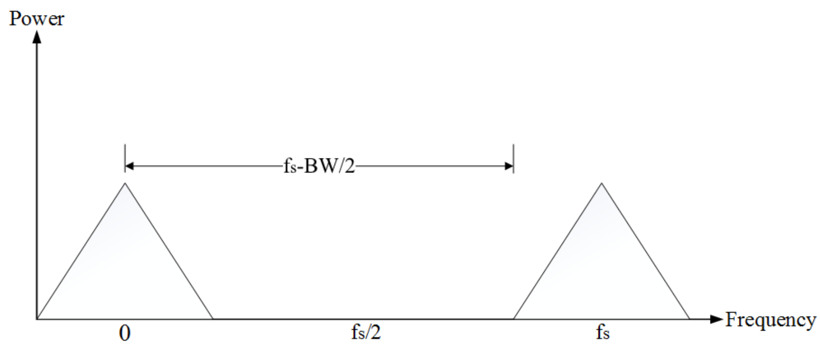


Figure 27. Aliasing in DAC.

Aliased signal is separated from the wanted signal by $f_s - BW/2$. Sampling frequency in the transmitter was chosen to be 76.8 MHz and the start frequency of aliased signal will become 76.23 MHz. Anti aliasing filtering is used to prevent the aliased signal from creating unwanted emissions. 3GPP specification of spurious emission is listed in Table 17 which must be satisfied by the transmitter.

As the aliased signal lies in the offset frequency range of 30 MHz < $f$ < 1000 MHz in which the specification of maximum emission level is -36 dBm on a measurement bandwidth of 100 kHz. If the maximum level of emissions is integrated over the entire bandwidth of LTE-M signal which is 1.14 MHz then the maximum level of emission becomes $-36 - 10log(100 * 10^3) + 10log(1.14 * 10^6)$ =-25.43 dBm. Maximum wanted signal level is 23 dBm, hence the maximum emission level relative to carrier is -48.43 dBc. So an anti aliasing low pass filter is required which can provide attenuation of 48.43 dB at frequency of 76.23 MHz to maintain the level of unwanted specification as per the standard.

### Adjacent Channel Power Ratio

Another way of quantifying out of band emission is the adjacent channel power ratio (ACPR) and spectrum emission mask (SEM). Measurement of SEM are carried out on a smaller bandwidth while measurement bandwidth of ACPR is much greater, so satisfying ACPR requirements means that SEM requirements are also met but this is not possible the other way i.e. satisfying SEM requirement doesn't mean ACPR requirement are also met [25]. SEM requirement are shown in Figure 14 and the normalized level of emission at an offset frequency of 1.34 MHz (adjacent UTRA channel center frequency) is -70 dBm/Hz which when integrated over the entire bandwidth gives the absolute value of -9.43 dBm ($-70 + 10log(1.14 * 10^6)$). Power level relative to carrier for spectrum emission mask is -32.43 dBc (as the maximum output power is +23 dBm and the absolute value of emission is -9.43 dBm). The required relative power level in case of UTRA$_{ACPR}$ at an offset frequency of 1.34 MHz is -33 dBc as mentioned earlier in Table 14. This shows that ACPR has more stricter requirement than SEM so achieving the ACPR requirement will satisfy the SEM criteria as well.

Main cause of ACPR is the third and fifth order nonlinearity of the transmitter chain whose main contributor is the PA. LTE-M signal is composed of sub-carriers and to find the ACPR caused by the transmitter these sub-carriers can be viewed as tone signals applied to the nonlinear PA. ACPR of -33 dBc is required by the transmitter to achieve the performance level specified by 3GPP. ACPR can be approximated from Equation (19) [26].

$$ACPR = IMR_2 + 10log(\frac{n^3}{16N + 4M}), \tag{19}$$

where $IMR_2$ is the ratio of wanted signal to intermodulation product and it is linked to OIP3 of the PA as Equation (20) [26], $n$ equals the number of sub-carriers, $N = \frac{2n^3-3n^2-2n}{24}$ and $M = \frac{n^2}{4}$.

$$IMR_2 = 6 + 2 * (OIP3_{(dBm)} - P_{out(dBm)})\text{dBc}, \tag{20}$$

where $P_{out(dBm)}$ is the output power in dBm and $OIP3$ is the third order output intercept point. LTE-M signal has 72 sub-carriers which gives the value of $N = 30450$ and $M = 1296$. From Equation (19) ACPR becomes equal to $IMR_2 + 1.2$. Equation (20) can be reordered to calculate the OIP3 also replacing the value of $IMR_2 = ACPR - 1.2$ (calculated for 72 sub-carriers) we get Equation (21).

$$OIP3 = \frac{ACPR - 1.2 + 2 * P_{out(dBm)} - 6}{2} \tag{21}$$

LTE-M specification of ACPR is 33 dBc and the maximum output power is +23 dBm so from Equation (21) we get the value of OIP3 as +35.9 dBm. OIP3 value of the transmitter chain should be greater than +35.9 dBm to fulfil the specifications of ACPR and SEM.

Table 21 lists the system level specification of LTE-M transmitter and receiver. BNF -1dB is also known as blocking noise figure -1dB and it is the point where the noise figure of the receiver increases 1 dB from the average value.

Table 21. Summary of LTE-M transceiver requirements in half duplex mode

| Receiver | | Transmitter | |
|---|---|---|---|
| Noise Figure (dB) | 5-7 | EVM | 17.5 % |
| IIP3 (dBm) | - 5 | Gain (dB) | 31-35 |
| P1dB (dBm) | - 15 | AAF (dBc) | -48.43 @ 76.23 MHz |
| BNF -1dB (dBm) | - 22 | OIP3 (dBm) | > 35.9 |
| LO (dBc/Hz) | - 83.9 @ ± 0.9075 MHz | D/A (bits) | 10 |
| | - 102 @ ± 2.8125 MHz | $F_{CLK}$ (MHz) | 76.8 |
| | - 113 @ ± 15 MHz | $P_{out}$ (dBm) | 23 |
| | - 127 @ ± 60 MHz | | |
| | - 134.2 @ ± 85 MHz | | |
| BPF Attenuation | 8 dB @ ± 85 MHz | | |

### 5.3. NB-IoT Design

Specification of NB-IoT and LTE-M are almost similar. NB-IoT always operates in half duplex and it is designed to enable a wide range of ultra low powered devices operating on standard telecommunication bands. In the upcoming subsections receiver and transmitter design of NB-IoT is discussed briefly.

#### *5.3.1. NB-IoT Receiver Design*

Narrow band IoT receiver can be designed in a similar as the LTE-M receiver. Sensitivity of NB-IoT signal is -108.2 as mentioned in Table 3. If the same QPSK (code rate 1/3) modulation scheme is used for NB-IoT then the required minimum SNR is -1 dB. Maximum allowed NF of the receiver can be calculated from Equation (1) as $NF = -108.2 + 174 - 10 * log(200000) - (-1) = 13.8$ dB.

Narrowband IoT receiver has increased sensitivity level by almost 8.2 dB compared to LTE-M which is because of the decreased bandwidth of NB-IoT signal. In this way the maximum allowed NF of narrowband IoT receiver is almost the same as LTE-M receiver.

Intermodulation specification for Narrowband IoT is listed in Table 4. Receiver input third order intercept point (IIP3) of -7 dBm can be used for the design of NB-IoT receiver. Level of intermodulation products at the input can be calculated using Equation (12) as $IM3_{Input} = 3 * (-46) - 2 * (-7)$ =-124 dBm. NB-IoT reference signal power level for intermodulation test is -102.2 dBm at the input. Achieved SNR in case of NB-IoT (IIP3 = -7 dBm) is 21.8 dB which is still 0.5 dB higher than in case of LTE-M (IIP3 = -5 dBm).

Out-of-band blocking specification of NB-IoT is listed in Table 6. Desensitization caused by high level blocking signals will be higher for NB-IoT receiver when compared with LTE-M. Which is due to the fact that receiver third order intercept point for NB-IoT receiver is 2 dB lower. Attenuation of 10 dB for signals at 85 MHz offset will limit the desensitization to less than 1 dB.

As mentioned earlier reciprocal mixing increases the noise floor of the receiver due to the presence of unwanted out-of-band blocking signals, narrowband blocking signals and in-band blocking signals. In case of NB-IoT receiver narrowband blocking is not specified so the other two causes of reciprocal mixing will be used to evaluate the performance requirement of synthesizer.

NB-IoT specification of out-of-band blocking are listed in Table 6 and the specification of in-band blocking are tabulated in Table 8. In case of out-of-band blocking phase noise of -133 dBc/Hz is required by the attenuated blocking signal which when integrated ($-133 + 10log(0.2 * 10^6)$ =-80) over the entire bandwidth (200 kHz) results into -80 dBc phase noise. To find the absolute level of phase noise blocking signal power (-25 dBm at 85 MHz offset) is added into -80 dBc which gives us -105 dBm noise level. The sum of all the noise's reaches a level of -102.2 dBm which is the same as reference signal so there is just 1 dB room for nonideality in case of out of band blocking. Similarly, other blocking signals are used to find the phase noise requirement of synthesizer at different offset frequencies as calculated earlier for LTE-M. The requirement of phase noise for NB-IoT is listed in Table 22 and these are almost similar to the requirement of LTE-M as mentioned earlier in Table 20.

Table 22. Phase noise requirement for NB-IoT receiver

| | Out of band blocking | | | In band blocking |
|---|---|---|---|---|
| Offset frequency (MHz) | $\pm$ 15-60 | $\pm$ 60-85 | $\pm$ 85-12750/1 | $\pm$ 7.505 |
| Phase noise (dBc/Hz) | -113.8 | -127.8 | -133.0 | -101.8 |

### *5.3.2. NB-IoT Transmitter Design*

Transmitter of NB-IoT can be designed in a similar way as LTE-M. Specifications of narrowband IoT transmitter are listed in Section 4.3. Maximum output power, EVM and spurious emission requirements are same for both the standards. ACPR requirement of NB-IoT is listed in Table 14. From Table 14 it can be seen that the specification of adjacent UTRA channel (offset frequency 2500 kHz) are more stringent than the adjacent GSM channel (offset frequency 200 kHz).

NB-IoT requires ACPR of 37 dB for the adjacent UTRA channel. The required OIP3 by the transmitter can be found from Equation (21) as $OIP3 = (37 - 1.2 + 2 * 23 - 6)/2 = 37.9$ dBm. Transmitter of narrowband IoT requires an overall OIP3 of 37.9 dBm to maintain the ACPR requirement. Bandwidth of narrowband IoT signal is less which makes it possible to use lower sampling frequency alongwith making the requirements of anti-alias filter reasonable. Spurious emission specification for NB-IoT is listed in Table 17. Spurious emission level of -23 dBm is specified for a bandwidth of 200 kHz at an offset frequency of 150 kHz < $f$ < 30MHz. Anti-aliasing filter must provide 46 dB of attenuation at offset frequency of 19.1 MHz to satisfy the requirement. Table 23 lists the system level requirement for the design of NB-IoT transceiver.

Table 23. Summary of NB-IoT transceiver requirements

| Receiver | | Transmitter | |
|---|---|---|---|
| Noise Figure (dB) | 5-7 | EVM | 17.5 % |
| IIP3 (dBm) | - 7 | Gain (dB) | 31-35 |
| P1dB (dBm) | - 17 | AAF (dBc) | -46 @ 19.1 MHz |
| BNF -1dB (dBm) | - 25 | OIP3 (dBm) | > 37.9 |
| LO (dBc/Hz) | - 101.8 @ $\pm$ 7.505 MHz | D/A (bits) | 10 |
| | - 113.8 @ $\pm$ 15 MHz | $F_{CLK}$ (MHz) | 19.2 |
| | - 127.8 @ $\pm$ 60 MHz | $P_{out}$ (dBm) | 23 |
| | - 133 @ $\pm$ 85 MHz | | |
| BPF Attenuation | 10 dB @ $\pm$ 85 MHz | | |

# 6. DISCUSSION

The thesis discussed system level design of RF transceiver operating on LTE-M standard. Transceiver is designed for a wearable device. As discussed earlier that RF transceiver can have various architectures and the choice of architecture is dictated by the application in which its used. The transceiver designed in this thesis is shown in Figure 28.

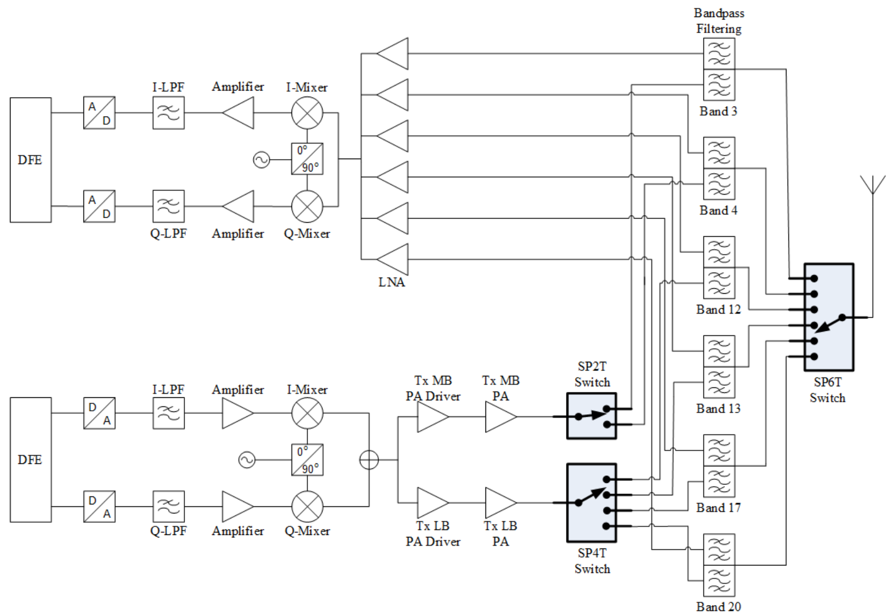


Figure 28. RF transceiver with receiver and transmitter bandpass filtering.

This architecture was designed to support six different bands so after the antenna a SP6T switch is used which can switch between the selected bands. After the switch bandpass filtering is used to remove far out interfering signals in case of receiver. Rest of the architecture of receiver is the same i.e. LNA's, demodulators, low pass filters etc. On the transmitter side there are two more switches SP2T and SP4T. The former switches between the two bands at mid range of the LTE bands at 2 GHz range and the latter switches between the four bands of the LTE at low range of under 1 GHz. The purpose of using switches on the transmitter side is to avoid the need for separate power amplifiers. Alternate approach to this architecture could be the use of six different PA's instead of the switching state. Since the size of PA is more than the size of switch so it was a feasible option to omit the use of more PA's in the architecture because of size restriction. An alternative architecture can be used which is shown in Figure 29.

On the receiver side band tunable filtering could be feasible due to relaxed requirements compared to FDD LTE instead of fixed bandpass filters. Tunable filter have in most scenarios lower Q-value and thus can't provide as sharp response as fixed (SAW) filters. Receiver can have two tuners one for the low band and one for the mid band which will have two LNA's and rest of the processing will be same. On the transmitter side we should try to avoid using any kind of filtering and this will make the performance requirement stricter for the PA. In the former architecture we used bandpass architecture to retain the transmission mask but in the latter architecture transmission mask must be retained with the PA. So the linearity of PA in the latter architecture plays the crucial role of maintaining the output of transmitter as per the defined standard. Also harmonic responses (Second order, Third order) of PA can be a major issue in the design hereby requiring internal LPF in the PA to meet spurious emission mask. Use of highly linear PA's will increase the power consumption which is not desired if the device is power limited. An advantage of this approach is that the size of components is reduced in the latter architecture as can be seen easily from the block diagram of both architectures.

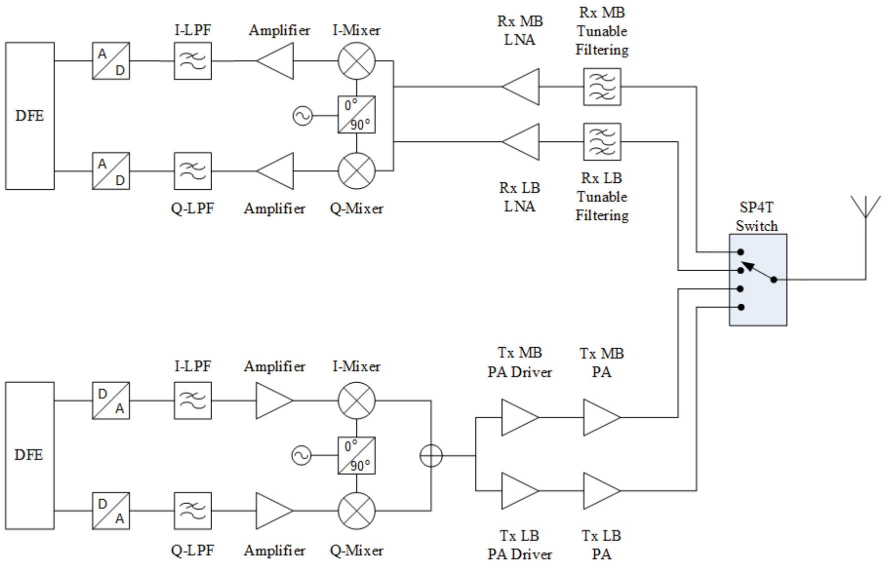


Figure 29. RF transceiver with receiver tunable filtering.

# 7. SUMMARY

This thesis presented a system level design of RF transceiver that can operate with 3GPP new standard variant LTE-M. Size of devices are getting smaller and their functionalities are increasing day by day. Wearable devices are nowadays equipped with Android and IOS operating systems and they can work in the similar way as a cell phone but integrating more components within the wearable device is a great challenge. The aim in this thesis is to design a small sized transceiver that can be integrated within the wearable device.

The design phase of transceiver started from the selection of LTE bands which the device will support. After selecting six different bands from different geographical locations RF specifications were studied and those specifications which influence LTE-M transceiver were used in the design. For the receiver design several test benches were created in ADS to evaluate the RF performance requirements. Third order intercept point of the receiver determines its linearity and additional requirements like filtering and more battery consumption are related to it. Desensitization of the receiver is caused by high level blocking signals for which filtering was used because increasing receiver linearity will in turn increase the power consumption of the receiver which was not feasible in this case. Also phase noise requirements of the synthesizer were found out by taking into account several blocking scenarios.

Unlike the receiver which has the main responsibility of receiving the desired signal out of the many unwanted signals present in the medium, the transmitter has to maintain its output mask in order to avoid interference with other users. Also EVM requirement of the transmitter must be met so that the receiver is able to extract the useful information out of the demodulated signal.

Transceiver design presented in this thesis can be integrated with any device. Great reduction in size can be achieved if the whole transceiver can be designed as an integrated circuit. Also the power consumption is crucial as IoT devices are expected to work for 10 years so low power consuming RFIC's must be built.

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