CHAPTER 1

Introduction

1.1 The Introduction to the Thesis

Software radio is a multi-mode/multi-band concept that has been developed as a solution for the growing number of incompatible radio air-interface standards. The radio handset should be able to use the same hardware for communications anywhere in the world. This means that the phone should accommodate the GSM (Global System of Mobile Communication) and W-CDMA (Wideband-Code Division Multiple Access) standards in Europe, the PDC (Pacific Digital Cellular) and PHS (Personal Handy Phone System) standards in Asia as well as the IS-54 (Interim Standard 54), IS-95 (Interim Standard 95) and CDMA2000 (Code Division Multiple Access 2000) standards in the United States. Further the phone should be capable of accommodating different functionalities such as cellular UMTS (Universal Mobile Telecommunications Systems), cordless (e.g. DECT (Digital Enhanced Cordless Telecommunications)), satellite UMTS, personal area networks (e.g. Bluetooth) and local area networks (e.g. IEEE 802.11). Over the
next years future fourth generation mobile systems will be added to this list. All schemes are capable of full duplex operation. No single standard covers all areas and all service requirements, so users will require multi-mode/multi-band handsets. To do this in a very small package is a difficult task.

The software radio approach is to replace many of the traditional analogue radio functions, such as channel selection, demodulation and synchronization, with digital signal processing (DSP). Configuring the radio to different standards becomes a software exercise. The remaining radio functions must be made generic (multi-band), the analogue to digital interface must have large dynamic range and the DSP sections must have low power consumption. It is interesting to note that over the past years tremendous progress has been made in reducing the part count of the baseband functions of a wireless mobile. However, little progress has happened in the RF functions. Today third generation mobile communication systems (3G) are at the beginning of their introduction. Through a number of technological advancements the evolution of practical software radio is accelerating. Some of these technologies making significant contributions include wideband RF devices, smart antennas, and integrated circuits.

An important issue in the design of a software radio is the replacement of fixed frequency components. One such component is the duplexing filter required for frequency division duplexing (FDD) systems. A new set of duplexing filters is required for each frequency band and each duplexing offset the radio must handle. In addition, a duplexer is required for each antenna in diversity systems. Normally two band-pass filters ($BPF_{Tx}$ and $BPF_{Rx}$) as shown in Figure 1-1, are combined to form a duplexer. The band-pass filter in the transmitter path ($BPF_{Tx}$) stops the transmitter noise artificially increasing the receiver noise figure while the band-pass filter in the receiver path ($BPF_{Rx}$) stops the transmitter signal overloading the receiver (desensitisation). These band-pass filters are designed to isolate the sensitive receiver circuits from the high power transmitter output.
A duplexer can be built in many ways using classical resonant circuits (e.g. ceramic or cavity resonators). Recently, duplexers are being built using multilayer technology with the use of slotline or microstrip line coupling structures to improve the level of integration. Other popular techniques include Surface Acoustic Wave (SAW) and Film Bulk Acoustic Wave Resonator (FBAR) devices.

Existing duplexers cannot be implemented in an integrated circuit. Here in this thesis an adaptive/active duplexer architecture that eliminates the need for multiple duplexers or reduces their requirements for software radio implementation is proposed. The adaptive duplexer architecture involves a reduced isolation circuit or device combined with a double loop cancellation process. Using this technique it is possible to obtain two adjustable nulls for cancelling the transmitter leakage signal (at frequency $f_{Tx}$) and its associated noise components (at frequency $f_{Rs}$) that fall in the receiver band (Figure 1-2).
CHAPTER 1 Introduction

Adaptive Duplexer for Software Radio

1.2 The Objective of the Thesis

The aim of this project is to study the feasibility of replacing or partially replacing the duplexing filter with an active cancellation unit. The unit will bleed a controlled amount of out-of-phase signal into the receiver front end to subtract out the jamming signal.

1.3 Contribution to Knowledge

The pace of change in telecommunications technology is increasing rapidly. In the area of mobile telephone systems, researchers are already exploring new technologies to realize a software radio system that supports multiple mobile telephony standards and multi-mode operation. The adaptive duplexer architecture in this work removes a significant barrier in the successful implementation of the software radio concept thus allowing the users of the mobile telephones to use a single handset to subscribe to all services. A duplexer that can handle multiple bands and which will enable a “future proof” solution by allowing the addition of new bands with different duplexing separations is at present unavailable. Hence,
this work on an adaptive duplexer with these capabilities will be a significant contribution to the realization of the software radio concept.

1.4 Summary of Findings

- A novel adaptive duplexer architecture that eliminates the need for multiple duplexers in a software radio front end is presented. Using this method the required transmitter leakage isolation and required transmitter noise cancellation levels that meet the 2G and 3G mobile wireless standards were achieved. A Tx leakage cancellation level of 66.8dB and a Tx noise cancellation of 58dB were achieved over a 5MHz bandwidth using a 190MHz duplexing frequency. This satisfies the typical requirement for a WCDMA duplexer of 48dB to 60dB.
- Cancellation performance is shown to be affected by delay differences between the main path and the cancellation paths. The tolerable delay variation depends on the dynamic range of the control elements in the cancellation paths. The achievable cancellation also depends on the duplexing frequency and systems with smaller duplexing frequencies achieve higher cancellation levels.
- The behaviour of the double loop cancellation technique is identified for the first time using geometrical analysis. This leads to an in depth knowledge of double loop cancellation theory that helps in determining important design parameters.
- It is shown that the delay difference close to zero in either of the cancellation paths results in a wide bandwidth. The coefficient values of the cancellation paths become very large when the difference of the path delays is close to an integral number of the reciprocal of the duplexing frequency and this condition must be avoided.
- The delay differences in the cancellation paths of the double loop cancellation should be chosen to straddle the expected range variation of the delay in the main path.
• An expression for the achievable cancellation level at a specified bandwidth is derived in terms of the delays and the duplexing frequency. It is shown that the relationship between the cancellation bandwidth and the achievable cancellation level is linear (6dB/octave). It is also shown that the residual signal power (|C|^2) is proportional to the bandwidth (B) squared, duplexing frequency (f_d) squared and the time delays (\tau_1, \tau_2) squared (|C|^2 = \pi^4 B^2 (f_d)^2 (\tau_1)^2 (\tau_2)^2).

• Cancellation in the Tx band and cancellation in the Rx band are shown to be interdependent. Therefore the control algorithm must use a cost function, which when minimised, provides cancellation in both bands. A new algorithm to achieve the required dual band cancellation is described.

1.5 Publication Summary

The following publications have resulted in from this research.

Conference Papers


CHAPTER 1 Introduction


Journal Articles


1.6 Organisation of the Thesis

This thesis is organized in eight chapters. Chapter 1, which corresponds to this introduction, states the objective, provides a contribution to knowledge, summary of findings, publications summary and a thesis overview.

Chapter 2 reviews the technology background. First it summarises the 1st and 2nd generations of mobile cellular systems. Next the 2.5 generation, that extends the 2nd generation platform to provide data rate enhancements and other premium services, is briefly looked at. Then the 3rd generation systems that bring about revolutionary changes are discussed. The convergence of different mobile cellular standards to form a unified 3G standard is also discussed along with a brief introduction to future 4th generation systems. The software radio as a concept to implement multi-band/multi-mode systems is also investigated. The duplexer front end component is recognised as one of the main challenges in realisation of a practical software radio. Finally some receiver architectures which are particularly suitable for integration and multi-band systems are examined.

Chapter 3 gives an introduction to the duplexer. Here the need for an adaptive duplexer for software radio is investigated. When a high transmitter leakage signal is present at the receiver input, a number of non-linear adverse effects can occur that degrade the receiver sensitivity. These include desensitisation, blocking, cross modulation and intermodulation and are discussed in this chapter. The main requirements associated with a duplexer are also specified.

A review of existing duplexing filter technologies along with a review of the literature associated with active duplexing structures are discussed in Chapter 4.

The new adaptive duplexer architecture which is based on a low isolation device and a cancellation unit is proposed and discussed in Chapter 5. Some design considerations for the low isolation device are also discussed.
Chapter 6 concentrates on the single loop cancellation method. This part of the research focussed only on the cancellation of the Tx leakage signal at the receiver. The work includes design and implementation details and also investigates phase and gain mismatches, delay effects and the control algorithm. Finally the experimental results are presented.

The new double loop cancellation system that suppresses both the Tx leakage and the Tx noise signals is presented in Chapter 7. The system is analysed using a geometric method. The influence of various parameters on the cancellation is examined. A new algorithm that controls the cancellation process is offered and the experimental results are presented.

Chapter 8 draws a number of conclusions based on the work completed and suggests some future work including a possible extension to four loop cancellation.
2.1 Introduction

Second generation mobile communication systems were created to address the need for increased capacity over first generation analogue systems. It is expected that at the beginning of 3G operation the second generation systems would coexist. Therefore transceivers should support both 2G and 3G functional requirements. Since the user preference is for a single mobile terminal, the transceiver requires multi-mode/multi-band capabilities. The software radio concept is being developed as a solution to this problem. There are many issues still to be addressed in the realisation of the software radio concept. The duplexer is one of the main problems of the RF front end since it is a fixed frequency component.

This chapter begins with a brief history and a summary of the first and second generation wireless system standards. Then it gives an overview to 3G systems and associated standards. The software radio and RF front end from the receiver
perspective are discussed in the next section. The adaptive duplexer in this research is based on a direct conversion receiver architecture (DCR) which is more suitable for software radio design. The direct conversion receiver is examined with some other receiver architectures in the final section.

2.2 1st Generation Systems

With the invention of microprocessors and the cellular communications concept in the 1970s and 1980s, the first generation (1G) mobile communication systems were born [1]. First generation systems use cellular coverage, where the coverage area is divided into small cell areas. The 1G systems were essentially analogue systems using Frequency Division Multiple Access (FDMA) to communicate and were designed for voice transmission only (no data). NMT (Nordic Mobile Telephone), AMPS (Advanced Mobile Phone Service), TACS (Total Access Communication System), ETACS (Extended Total Access Communication System), JDC (Japan Digital Cellular) etc., were among first generation systems. NMT was the first analogue cellular phone system that started operating in Scandinavia in 1979. In the beginning, it used the 450MHz band and therefore was named NMT 450. Later it used the 900MHz band because of the need for more capacity and was called NMT 900. AMPS was introduced in 1978 by the Bell telephone company in the USA and started operation in 1983 in Chicago. TACS was introduced in UK in 1982. ETACS was the extended version of TACS and was deployed in 1985. The cellular systems called C-450 (operated in the 450 MHz band) and Radicom 2000 (operated in the 200 MHz band) were also introduced in Germany and in France respectively in 1985.

These systems had numerous problems such as capacity limitations of TDMA (Time Division Multiple Access), incompatibilities across geographies (USA, Japan and Europe), only nationwide coverage, no open interfaces except the radio interface, low speech quality and no security in speech transmission. The major first generation analogue cellular radio system standards [2] are compared in Table 2-1. (Please refer to the list of abbreviations for descriptions in the tables.)
2.3 2\textsuperscript{nd} Generation Systems

Second generation systems started to appear across the world in the early 1990s. Advances in integrated circuit technology brought digital transmission to mobile communications. Second generation systems are based on digital technology and offer data speed up to 9.6kb/s and use TDMA or CDMA access methods in combination with FDMA. Second generation systems are capable of providing voice, data, fax transfer as well as other services. 2G systems can be categorised as 2G cellular mobile systems and 2G Personal Communication Systems (PCSs). GSM, US-TDMA (IS-136), cdmaOne (IS-95) and PDC are included in second
CHAPTER 2 Technology Background

D-AMPS (Digital-Advanced Mobile Phone Services) is a digital version of AMPS. D-AMPS is also known as US-TDMA/IS-136. IS-54 (US Digital Cellular) service is an old version of the IS-136. GSM was originally designed to operate in the 900MHz band but was later adapted to operate in 1800MHz and 1900MHz bands. The GSM 450 (operate at 450MHz band) may start to operate in some

Table 2-2 Second generation digital cellular standards summary [3], [4]

<table>
<thead>
<tr>
<th>Specification</th>
<th>GSM</th>
<th>IS-54</th>
<th>PDC</th>
<th>IS-95</th>
</tr>
</thead>
<tbody>
<tr>
<td>Year of introduction</td>
<td>1990</td>
<td>1991</td>
<td>1993</td>
<td>1993</td>
</tr>
<tr>
<td>Frequencies</td>
<td>890-915 MHz(R) 935-960 MHz (F)</td>
<td>824-849 MHz (R) 869-894 MHz (F)</td>
<td>810-830 &amp; 1429-1453 MHz (R) 940-960 &amp; 1477-1501 MHz (F)</td>
<td>829-849 &amp; 1850-1910 MHz (R) 940-960 &amp; 1477-1501 MHz (F)</td>
</tr>
<tr>
<td>Multiple access</td>
<td>TDMA/FDMA/FDD</td>
<td>TDMA/FDMA/FDD</td>
<td>TDMA/FDMA/FDD</td>
<td>CDMA</td>
</tr>
<tr>
<td>Modulation</td>
<td>GMSK (BT=0.3)</td>
<td>π/4 DQPSK</td>
<td>π/4 DQPSK</td>
<td>QPSK QOQPSK</td>
</tr>
<tr>
<td>Carrier separation</td>
<td>200kHz</td>
<td>30kHz</td>
<td>25kHz</td>
<td>1.25MHz</td>
</tr>
<tr>
<td>Channel data rate</td>
<td>1 270.833kbps</td>
<td>48.6kbps</td>
<td>42kbps</td>
<td>19.2kbps</td>
</tr>
<tr>
<td>Number of voice channels</td>
<td>1000</td>
<td>2500</td>
<td>3000</td>
<td>4000</td>
</tr>
<tr>
<td>Spectrum efficiency</td>
<td>1.35bps/Hz</td>
<td>1.62bps/Hz</td>
<td>1.68bps/Hz</td>
<td>2.58bps/Hz</td>
</tr>
<tr>
<td>Speech coding</td>
<td>RELP-LTP @ 13kbps</td>
<td>VSELP @7.95kbps</td>
<td>VSELP @ 6.7kbps</td>
<td>QC Elep @ 9.6kbps / @ 14.4kbps</td>
</tr>
<tr>
<td>Channel coding</td>
<td>CRC with R=1/2; L=5 Conv.</td>
<td>7 bit CRC with r=112; =6 Conv.</td>
<td>CRC with Cony.</td>
<td>NA</td>
</tr>
<tr>
<td>Equalizers</td>
<td>Adaptive</td>
<td>Adaptive</td>
<td>Adaptive</td>
<td>Adaptive</td>
</tr>
<tr>
<td>Portable Tx. power max/avg.</td>
<td>1 W/125mW</td>
<td>600mW/200 mW</td>
<td>125mW</td>
<td>200mW</td>
</tr>
<tr>
<td>Duplexing method</td>
<td>TDD/FDD</td>
<td>TDD/FDD</td>
<td>TDD/FDD</td>
<td>FDD</td>
</tr>
</tbody>
</table>
countries to replace old analogue networks. Currently the maximum data rate for GSM is 14.4kbps. IS-95 is based on narrowband spread spectrum technology and uses 1.25MHz channel bandwidth. Therefore it offers increased capacity, wider bandwidth and is very flexible because it uses CDMA access method. IS-95 and IS-136 are capable of operating in the same band as AMPS and specified to be dual-mode systems. 2G systems compared with 1G systems allow more efficient use of the radio spectrum since they can handle more calls than analogue FDMA technology.

PDC is the Japanese 2G standard. It is somewhat similar to IS-54 standard, but uses 4-ary modulation for voice and control channels, making it more like IS-136 in North America [3].

DCS 1800 (Digital Communication System 1800) and PHS are also included in 2G PCSs [3] and are listed in Table A-1 in Appendix A. The major 2G digital cordless air interface standards [3] are also listed in Table A-2 in Appendix A.

2.4 2.5 Generation Systems

2.5 generation systems address the data capacity limitations associated with the 2nd generation systems. Even though the boundary between 2G systems and 2.5G systems is somewhat unclear, 2.5G systems provide clear upgrades to the 2G systems that almost make it possible to provide similar capabilities as 3G systems. A number of technologies are commonly used to provide these capabilities such as High Speed Circuit Switched Data (HSCSD), Enhanced Data rates for Global Evolution (EDGE) and General Purpose Radio Services (GPRS).

HSCSD is the easiest upgrade to achieve higher data rates. It improves the maximum user data rate of the air interface by using more than one time slot for data connections. Implementations that use up to four time slots for data connections are commercially available. This is an innovative and inexpensive
way to upgrade the current wireless platform as the changes involve only upgrading the software used in the network.

Using GPRS, data rates up to 115kbps with error correction are possible using approximately eight time slots. This technology is based on packet switching and thus makes efficient use of the available bandwidth using variable bit rates. It is also suitable for services that use bursty data due to its ability to dynamically allocate resources.

EDGE is an improvement over GSM which increases the traditional GSM data rates over 300%. It uses eight phase shift keying (8 PSK) method for modulation. This is an attractive solution for existing GSM networks as the change required is only a software upgrade. Due to its ability to co-exist with the Gaussian minimum shift keying modulation, it allows users to continue using their current handsets. IS-136 also can be upgraded using EDGE.

In addition to these methods, NTT DoCoMo (Nippon Telegraph and Telephone DoCoMo) from Japan has developed their own proprietary packet based technology called i-mode which provides users an efficient mechanism for wireless internet browsing and email access.

2.5 3rd Generation Systems

Third generation systems opened the way for a completely new era of wireless services that enabled access across multiple geographies. 3G systems provide a platform that is common for multiple wireless standards and technologies. They are aimed to carry data up to 2Mb/s, about 200 times faster than the 2G systems in indoor environment and a minimum of 144kbits/s in other environments [5]. Because of the high-speed data rate, 3G systems will be able to support services such as audio, video, multimedia, internet, data and speech.
The key objectives of 3G development are to provide [6]:

- multi-standard user terminals that operate effectively for all types of services, in all radio environments
- service quality that is comparable to the current fixed public network
- flexible new capabilities and services such as WWW, high bit-rate data and multimedia
- network to air interface flexibility
- compatibility with second generation e.g. GSM/DCS
- “Future proof” systems to easily accommodate added capabilities

The main requirements that apply to third generation systems are [7]:

- support for high data rates up to at least 144kbits/s (384kbits/s for full area coverage) in all radio environments and up to 2Mbits/s in low mobility and indoor environments
- support for symmetrical and asymmetrical data transmission
- support for packet-switched and circuit switched services, such as Internet (IP) traffic and real-time video
- support for good voice quality (compatible with wireline quality)
- support for greater capacity and improved spectrum efficiency compared with existing second-generation wireless systems
- support for several simultaneous services to end-users and terminals – that is, for multimedia service capabilities
- support for coexistence and interconnection with mobile satellite services
- support for roaming, including international roaming, between different IMT-2000 (International Mobile Telecommunications for the year 2000) operators
- support for scale-of economy and an open global standard to meet mass-market needs

The goal of 3G technologies is to create a single global standard that allows for global roaming. The International Telecommunication Union (ITU) and the United Nations organisation responsible for global telecommunications began its

The European Telecommunications Standards Institute (ETSI) regards 3G systems as UMTS. In 1998 the first decision in the standardisation process of UMTS was made by ETSI. ETSI chose the W-CDMA concept to be adopted in the spectrum (for uplink one band of spectrum and for down link another band of spectrum, - FDD duplex mode) of UMTS. It also chose the TD-CDMA (Time Division-CDMA) concept to be adopted in unpaired band (a single monolithic block of spectrum – TDD duplex mode) of UMTS. UMTS will consist of both satellite and terrestrial components as in IMT-2000 and will support both circuit-switched and packet-switched services. The Telecommunications Technology Association (TTA) in South Korea and Association of Radio Industries and Business (ARIB) in Japan have developed standards based on W-CDMA. The Telecommunications Industry Association (TIA) in United States proposed CDMA2000. The major difference between W-CDMA and CDMA2000 is that W-CDMA is backward compatible with GSM networks and CDMA2000 is backward compatible with IS-95 networks [9]. Due to the different technologies used in different regions in the world, a family of compatible standards were adopted under IMT-2000 umbrella.

The five standards included in IMT-2000 terrestrial radio interfaces are shown below [10], [11].

1. IMT-DS (Direct Sequence CDMA - FDD duplex type) - This is known as UTRA FDD (UMTS Terrestrial Radio Access FDD) or W-CDMA and adopted in Europe and Japan. This has been specified by 3GPP (Third Generation Partnership Project) and operates in the IMT-2000 paired bands
at a chip rate of 3.84Mcps, spread over approximately 5MHz. This mode of operation will be used in the UMTS macro and micro-cellular environment.

2. IMT-MC (Multi-Carrier CDMA - FDD duplex type) - This is known as CDMA2000. This has been specified by 3GPP2 and operates on the downlink at a basic chip rate of 1.288Mcps, occupying 1.25MHz of bandwidth.

3. IMT-TC (Time Code CDMA - TDD duplex type) - This is known as TD-CDMA. It is a combination of the UTRA TD D and the TD-SCDMA. This operates in unpaired spectral bands at a chip rate of 3.84Mcps, spread over approximately 5MHz.

4. IMT-SC (Single Carrier - FDD duplex type) - This is known as UWC-136 and proposed by the Universal Wireless Communication (UWC) Consortium and Telecommunication Industry Association (TIA). It represents convergence between the TDMA-136, GSM and EDGE standards. UWC-136 will adopt the GPRS packet data network architecture, while enhancing the TDMA-136 radio interface to include GSM/EDGE compatibility and a high data rate indoor solution. The UWC-136 solution provides backward compatibility with the AMPS, IS-54, IS-136 and GSM networks.

5. IMT-FT (Frequency Time - TDD duplex type) - This is known as DECT.

In May 1999, the Operators Harmonization Group (OHG) concluded the harmonization discussion at a meeting in Toronto, Canada, which resulted in a single 3G CDMA standard with three modes; i.e. a direct sequence mode based on W-CDMA, a multi-carrier mode based on CDMA2000 and a TDD mode based on UTRA TDD [12]. Combined time division and code division multiple access scheme is used in UTRA TDD.

W-CDMA (UTRA FDD) technical summary [13] is shown in Table 2-3 and CDMA2000 technical summary [14] is shown in Table 2-4. TD-CDMA [15], TD-SCDMA [16] and W-CDMA (DoCoMo) [17] technical summaries are shown in Table A-3, Table A-4, and Table A-5 in Appendix A.
### Table 2-3  W-CDMA (UMTS) technical summary [13]

<table>
<thead>
<tr>
<th>Frequency band</th>
<th>1920MHz - 1980MHz and 2110MHz - 2170MHz (Frequency Division Duplex) UL and DL</th>
</tr>
</thead>
<tbody>
<tr>
<td>Minimum frequency band required</td>
<td>~ 2x5MHz</td>
</tr>
<tr>
<td>Frequency re-use</td>
<td>1</td>
</tr>
<tr>
<td>Carrier spacing</td>
<td>4.4MHz - 5.2MHz</td>
</tr>
<tr>
<td>Maximum number of (voice) channels on 2x5MHz</td>
<td>~196 (spreading factor 256 UL, AMR 7.95kbps) / ~98 (spreading factor 128 UL, AMR 12.2kbps)</td>
</tr>
<tr>
<td>Voice coding</td>
<td>AMR codecs (4.75kHz - 12.2kHz, GSM EFR=12.2kHz) and SID (1.8kHz)</td>
</tr>
<tr>
<td>Channel coding</td>
<td>Convolutional coding, Turbo code for high rate data, Duplexer needed (190kHz separation), Asymmetric connection supported</td>
</tr>
<tr>
<td>Tx/Rx isolation</td>
<td>MS: 55dB, BS: 80dB</td>
</tr>
<tr>
<td>Receiver</td>
<td>Rake</td>
</tr>
<tr>
<td>Receiver sensitivity</td>
<td>Node B: -121dBm, Mobile -117dBm at BER of 10⁻³</td>
</tr>
<tr>
<td>Data type</td>
<td>Packet and circuit switch</td>
</tr>
<tr>
<td>Modulation</td>
<td>QPSK</td>
</tr>
<tr>
<td>Pulse shaping</td>
<td>Root raised cosine, roll-off = 0.22</td>
</tr>
<tr>
<td>Chip rate</td>
<td>3.84Mcps</td>
</tr>
<tr>
<td>Channel raster</td>
<td>200kHz</td>
</tr>
<tr>
<td>Maximum user data rate (physical channel)</td>
<td>~ 2.3Mbps (spreading factor 4, parallel codes (3 DL / 6 UL), 1/2 rate coding), but interference limited.</td>
</tr>
<tr>
<td>Maximum user data rate (offered)</td>
<td>384 kbps (year 2002), higher rates (~ 2 Mbps) in the near future. HSPDA will offer data speeds up to 8-10 Mbps (and 20 Mbps for MIMO systems)</td>
</tr>
<tr>
<td>Channel bit rate</td>
<td>5.76Mbps</td>
</tr>
<tr>
<td>Frame length</td>
<td>10ms (38400 chips)</td>
</tr>
<tr>
<td>Number of slots / frame</td>
<td>15</td>
</tr>
<tr>
<td>Number of chips / slot</td>
<td>2560 chips</td>
</tr>
<tr>
<td>Power control period</td>
<td>Time slot = 1500Hz rate</td>
</tr>
<tr>
<td>Power control step size</td>
<td>0.5, 1, 1.5 and 2dB (Variable)</td>
</tr>
<tr>
<td>Power control range</td>
<td>UL 80dB, DL 30dB</td>
</tr>
<tr>
<td>Mobile peak power</td>
<td>Power class 1: +33 dBm (+1dB/-3dB) = 2W; class 2 +27 dBm, class 3 +24 dBm, class 4 +21 dBm</td>
</tr>
<tr>
<td>Number of unique base station identification codes</td>
<td>512 / frequency</td>
</tr>
</tbody>
</table>
The evolution of the third-generation mobile telecommunication system IMT-2000, in terms of data rate support [18] is shown in Figure 2-1. The capability targets for the 3G have been defined as 384 kbit/s for full area coverage and 2Mbit/s for local area coverage.

![Figure 2-1](image)

**Figure 2-1** Evolution towards third generation mobile systems in terms of data rate support [18].

W-CDMA is capable of providing more capacity as it uses a 4 times wider channel allocation compared to narrow band CDMA. It is based on the radio
access technique proposed by the ETSI Alpha group and the specifications were finalised in 1999 [13].

The AMPS, GSM, IS-95 (CDMA) and UMTS FDD (W-CDMA) standards are of particular interest to this thesis as they all specify FDD operation and therefore require a duplexing filter.

2.6 4th Generation Systems

As envisaged currently, the 4th generation systems will involve providing IP connectivity over the wireless interface. It will encompass all 2nd and 3rd generation wireless mobile technologies and will provide a standard that ensures the interoperability between satellite and terrestrial wireless mobile technologies and W-LANs (Wireless Local Area Networks). Interoperability of network technologies is going to be a major challenge in 4G systems. It is projected that the 4G systems will provide at least 100 Mbps peak data rates in full-mobility wide area coverage and 1 Gbps in low-mobility local area coverage [19]. OFDM (Orthogonal Frequency Division Multiplexing) modulation, multiple antenna technology and CDMA are being considered as elements for 4G systems.

2.7 The Software Radio

The software radio is a highly flexible radio base station or subscriber terminal platform, incorporating many advanced features and technologies, which enables it to provide flexibility and programmability well beyond conventional analogue or digital radios[20]. A software radio is a radio in which the channel modulation waveforms are defined in software [21]. It is a new radio architecture concept that is being researched to overcome a number of drawbacks associated with the traditional narrow band radio architecture. In a traditional narrowband receiver most of the functions such as filtering, amplification, down converting and demodulation are performed using analogue techniques before the signal is
digitised. In software radio the emphasis is given to replace most of the traditional RF functions by digital signal processing techniques. So that the user can have one single transceiver that can tune to different channels of different wireless systems. Software radio will allow the terminals to connect to multiple networks that use different air interface standards. Switching between different wireless standards becomes only a software upgrade without a need to change or upgrade hardware. An excellent overview and discussion on the software radio concept/architecture is presented in [22], [23].

A simple realization concept of a multi-mode receiver that consists of switched multiple traditional receiver chains is shown in Figure 2-2. The required channel is selected by using a fixed channel select filter at an intermediate frequency and channel bandwidth set by the parameters of the particular standard.

![Figure 2-2](image-url)  
*Figure 2-2  Traditional multi-mode/multi-band receiver.*

The ideal software radio architecture is shown in Figure 2-3. In this concept the analogue to digital conversion is carried out directly at the front end. Here, the entire band under consideration, which consists of number of channels, will be digitised so that all functions of the radio can be performed using programmable Digital Signal Processing (DSP) microprocessors or Field Programmable Gate Arrays (FPGAs) or some dedicated Application Specific Integrated Circuits.
(ASICs). Hence the ideal receiver architecture can be reconfigured to suit any RF band, modulation or data format. This makes it capable of operating within any communications network and would have significant benefits such as reduction of size, cost and power consumption of radio systems.

Figure 2-3  An ideal software radio architecture.

If the radio functions are performed using DSP, the new features and upgrades could be done using software upgrades, which become only a reprogramming task. The other significant advantage of using software radio is its ability to support multiple standards. With software radio the receiver could be programmed to receive incompatible technologies such as TDMA (Time Division Multiple Access), GSM and CDMA in the same receiver.

However there are a number of challenges that lie ahead in achieving this concept, mainly in the speed of current Analogue to Digital Conversion (ADC) and DSP capabilities. The total digital processing requirements in such a radio may add up to more than 10GFLOPS (Giga Floating Point Operations per Second) [22]. By January 2005 Texas Instruments have accelerated their fastest DSP to 1500MFLOPS (Mega Floating Point Operations per Second) [24]. Still this is not sufficient for processing the entire radio functionality required by software radio. Since the band of interest could be many megahertz wide (3G mobile’s spectrum is 200MHz) and is in the microwave frequency range, it requires very high speed
ADCs to sample the signal. The ADCs must satisfy the Nyquist criterion for sampling so that the original signal can be digitised without the risk of losing information. Up to 500MSPS (Million Sampling per second) of sample rate will be required by 3G mobile with 200MHz bandwidth. Even if the ADCs capability of performing the required conversion rates were available, the power requirements associated with these devices would prevent these ADCs being used in mobile handsets. Kenington’s [25] work shows for 20bits (121.76 dB dynamic range) resolution, the theoretical minimum power consumption would be 600mW when operating at 48MSPS. This makes it highly unlikely that the ADC technology in current form would ever be appropriate for pure software radio implementation without a revolutionary change in the ADC architectures [26].

Therefore, radio frequency digitization as in ideal software radio architecture is currently not feasible even though it is an ultimate goal of technology developers. Therefore the receiver configuration should have a pre-selection filter and a low noise amplifier between the antenna and the ADC (and a reconstruction filter and a power amplifier between the DAC (Digital to Analogue Converter) and antenna in the Tx chain). Even then it is debatable whether an ADC exists which could perform well enough at reasonable cost to make a system economically viable [27].

A practical wideband software receiver architecture [20] is illustrated in Figure 2-4. Here, all the channel processing systems will share the wideband front end. The entire band or sub band is digitized by the ADC and the RF front end remains the same as in the traditional design.

A wideband software radio transmitter architecture [20], is shown in Figure 2-5. Here, channels are digitally combined and up-converted after baseband processing, using the counterparts of the wideband receive process.
Various software radio architectures and design considerations are described in [28], [29], [30], [31] and [32]. In the implementation of software radio architecture in a mobile handset, one of the main challenges lies in the design of a suitable RF front end. The latest sub micron CMOS (Complementary Metal-Oxide Semiconductor) VLSI (Very Large-Scale Integration) techniques available...
today have not been a great benefit to RF performance because the dynamic range is reduced by the lower supply voltage. Most of the miniaturisation of RF circuits has been accomplished through a reduction in the packaging size of passive components. Currently, Si bipolar and GaAs are used in most high performance analogue RF front-ends. CMOS is often used for low cost and lower performance front ends as used in W-LANs.

The other main hardware challenges in the RF front end are:

- The Antenna – consumer demand is for mobile communication devices which are lightweight and small in size. In such devices, transmission and receiving must be made by a common antenna [33]. Nearly all cellular phones use a single antenna. The antenna should be able to cover the whole radio spectrum (over 200MHz for 3G). Using a multi-band antenna, software radio can cover both the 2G and 3G spectrum. For example, the AN-40 super wideband antenna which is from US company “Adams-Russel” can cover over 300MHz of spectrum [34] in the 2GHz band.

- The Duplexer – to prevent the receiver from desensitising due to high interference signals, it requires a duplexer in the front end. Currently available duplexing filters are not capable of tuning to the entire required band. Therefore some sort of adaptive duplexing is required. This is discussed in chapter 3.

- The RF power Amplifier - wideband systems require amplifiers with very high dynamic range. Since a non-linear amplifier has dynamic range limitations, a highly linear wideband power amplifier is required.

A review of technical challenges of software radio can be found in [35]. Interoperability between incompatible wireless communication systems can be achieved using a fully programmable software radio. Clear benefits and flexibility lie when the reprogramming of the handset is done using the over-the-air interface. Further, reprogramming and software upgrades can be done using smartcards, over the internet and at kiosks etc. Software difficulties or requirements are not discussed in this section as it is out of the scope of this work.
2.8 Receiver Architectures

At present receiver architectures such as heterodyne, homodyne, image-reject digital intermediate frequency and sub-sampling exist. Super heterodyne receivers have been used widely for radio systems since this architecture has a number of significant benefits including image rejection and adjacent channel selectivity. There is a resurgence of interest in research into homodyne, low-IF and wideband-IF receiver architectures due to the high level of integration requirements for software radio. This section gives an overview of these receiver architectures. The research of the adaptive duplexer presented in this thesis is based on the homodyne receiver architecture. It can be adapted to other receiver architectures as well.

2.8.1 Heterodyne Receivers

![Simplified block diagram of a heterodyne receiver architecture.](image)

*Figure 2-6  Simplified block diagram of a heterodyne receiver architecture.*
A simplified block diagram of a heterodyne receiver is shown in Figure 2-6. The received RF signal from the antenna is filtered to remove out-of-band signals using a band select filter. It is then amplified by a Low Noise Amplifier (LNA) to amplify the signal without strengthening the noise component. The channel select filter down the receiver chain (which is used to filter the desired channel) requires high Q’s. In order to relax this requirement, the RF signal after the LNA is down converted to an Intermediate Frequency (IF), that is much lower than the received RF signal. The channel select filter then performs channel selection at IF. The selection of the IF is a major consideration in heterodyne receiver architectures. The frequency translation process generates an unwanted signal which may interfere with the desired signal. If the LO (Local Oscillator) frequency is \( \omega_{LO} = \omega_{RF} + \omega_{IF} \), then the desired receiver band centred at \( \omega_{RF} \) is translated to \( \omega_{IF} \). In addition to this, \( \omega_{RF} + 2\omega_{IF} \) is also translated to \( \omega_{IF} \). This unwanted signal is known as the image signal. The image problem is a critical one since image signal power can be much higher than that of the desired signal [36]. Therefore to remove the image signal, the output of the LNA signal is filtered by an image reject filter before the down conversion.

As shown in Figure 2-7 when the IF is high, the image rejection filter can have high attenuation in the image band [36]. If the IF is low, then the image band attenuation is low but the channel selection filter can produce higher attenuation to the adjacent channel interference signals. Therefore a significant trade-off between sensitivity and selectivity must be done in the design of heterodyne receivers. The channel selection filters in the first IF stage are normally made with SAW filters since highly selective transfer functions are impractical in current IC technology [37]. The centre frequencies and bandwidth of these filters are not flexible and not wide enough to support a multi-band receiver. Further, separate IF selection is required for each mode due to the fixed receive bandwidth of the heterodyne receiver architecture [38]. Therefore it is difficult to manufacture broadband terminals with heterodyne architecture. Another important drawback of this architecture is that the LNA must drive a 50\( \Omega \) load because the image reject filter is placed off-chip [39].
2.8.2 Homodyne Receivers

The homodyne receiver architecture eliminates many discrete components in the receiver chain. In a direct conversion receiver (homodyne, zero-IF), the incoming RF signal is amplified by an LNA and then the desired receiver band is directly translated to zero frequency using a single mixer stage (Figure 2-8). Here LO frequency is equal to the desired carrier frequency. Channel selection of the I and Q signals is done by low-pass filters. Quadrature I and Q channels are necessary in typical phase and frequency modulated signals because the two side bands of the RF spectrum contain different information and results in irreversible corruption if they overlap each other without being separated into the two phases [39]. The advantages over the heterodyne architecture are, it has a simple design,
no image reject filter is required and therefore the LNA need not drive a $50\Omega$ load simplifying the LNA design.

![Figure 2-8](image)  
*Figure 2-8  A simplified block diagram of a homodyne receiver architecture.*

### 2.8.2.1 Design Issues

There are well known design issues compared to the heterodyne architecture which are discussed in this section.

#### 2.8.2.1.1 DC Offset

Since the receive signal is down converted to zero frequency, DC offset voltages can corrupt the receive signal and can saturate the following stages. There are three types of DC offsets. One is caused when a large interferer leaks from the output of the LNA to the LO port and mixes with the signal itself (self-mixing). This interferer can be the Tx leakage signal in the receiver as discussed in Section

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3.6.4. The second is caused by the LO leakage signal mixing with the LO signal. The LO leakage signal can appear at the input of the LNA or the input of the mixer as shown in Figure 2-9. The third occurs when the LO leaks into the antenna and is radiated; it is subsequently reflected from moving objects back to the antenna to generate a time varying DC offset.

![Diagram of LO leakage signal](image)

Figure 2-9  Self-mixing in DCR due to LO leakage signal in the LNA/mixer input port.

2.8.2.1.2  I/Q Mismatch

The errors in the nominally 90° phase shifter and mismatches between the amplitudes of the I and Q paths, contribute to gain and phase errors [39] that cause unwanted side bands and distortions in the desired signal. Fortunately, pilot symbol assisted channel estimation is done in W-CDMA systems and this estimation leads to a correction of the I/Q phase and amplitude mismatches [38].

2.8.2.1.3  Even Order Distortion

If two strong interferers close to each other are present at the input to the LNA, a second order non-linearity produces a low frequency beat signal. This phenomenon is discussed in Section 3.6.3.2. In addition a second order distortion
creates a time variant DC offset when amplitude modulated signals are present at the input of the I/Q mixers [38].

2.8.2.1.4 Flicker Noise

Flicker noise is an intrinsic noise source found in semiconductor devices and has a spectral density that is inversely proportional to frequency (so named as $1/f$). Since in direct conversion receivers the down converted spectrum is located around zero frequency, the $1/f$ noise of devices can corrupt the signals. An analysis of $1/f$ noise in direct conversion receivers can be found in [40].

The major design issue out of all the above, is the DC offset component. It must be removed to prevent the large DC component from desensitising the baseband demodulator. For UMTS however, the problem is not so serious, because of the spread spectrum nature of the air-interface. With such a wide bandwidth, it is possible to use AC coupling in the receiver to filter out the DC offsets and some of the other $2^{nd}$ order products without significantly degrading the sensitivity [41]. These design issues put higher requirements on RF gain, LO-to-RF isolation, IIP3 (third order input intercept point), IIP2 (second order input intercept point) and noise performance. Despite these disadvantages, the direct conversion receiver architecture offers a higher degree of integration than the conventional heterodyne architecture and is more suitable for broadband terminals. The channel select filters (LPFs) and RF to baseband section, can be easily integrated into one chip using Bi-CMOS (especially with SiGe) technology. Recently published direct conversion receiver architectures for multi-band applications can be found in [42], [43], [44], [45], [46] and [47].

2.8.3 Single Conversion Low-IF Receivers

In the single conversion low-IF receiver architecture [48], [49], [50] which is similar to zero-IF receiver architecture, all of the desired channels are down converted to a low intermediate frequency, as shown in Figure 2-10. This low-IF is of the order of one or two channel bandwidths. The main advantages over
zero-IF receiver are that there is no DC offset, flicker noise and LO self-mixing problems because the desired channel is not at DC; in fact it is offset from DC. Since this low-IF architecture avoids the use of discrete components such as image reject filters, it allows a higher level of integration. As the desired carrier is down converted to low-IF, image rejection must be performed by an on-chip image rejection mixer. The image-rejection is limited by matching considerations to about 40dB [50].

![Figure 2-10](image.png)

**Figure 2-10** A block diagram of a simplified single conversion low-IF receiver architecture.

### 2.8.4 Double Conversion Wideband-IF Receivers

The double conversion wideband-IF architecture [50], translates all of the receive signal channels to IF frequency using a mixer with a fixed frequency LO (LO1), thus maintaining a larger bandwidth. Then the up-converted frequency components are removed using a simple low-pass filter. All the received channels after the LPF (Low-Pass filter) are then down converted to baseband using a tuneable channel select low frequency synthesiser (LO2). Since the desired receive
channel is selected firstly in the first-IF and secondly in the lower second-IF, relaxed Q filters can be used. Here also digitally programmable baseband filters are used and multi-standard capability can be achieved as in the zero-IF receiver architecture. Although the LO\textsubscript{2} is at the same frequency as the IF desired carrier in the wideband-IF system, the offset at baseband which results from self-mixing is relatively constant and is easily cancelled [50].

![Block diagram of a simplified double conversion wideband-IF receiver architecture.](image)

The image rejection filter and channel select filter in heterodyne receiver are normally implemented off-chip. For different standards different image reject filters and different channel select filters have to be used, making it difficult to achieve a highly integrated low cost receiver for multi-band applications. In the DCR architecture no image rejection filter is needed and low-pass filters can be integrated with low power consumption. Although low-IF receiver architectures allow a similar level of integration as zero-IF architectures, they are less suitable
for multi-standard applications [50]. The wideband-IF architecture is also suitable for multi-band applications since the second LO (LO₂) is programmable.

The direct conversion receiver is chosen among the above receiver architectures to test the proposed duplexer architecture in this thesis, since it is more suitable for multi-band applications.

2.9 Conclusion

Mobile devices have evolved significantly and the need for unified standards has become even more important. It is rather unlikely that the whole world would adopt one wireless standard soon. Therefore for the end user the flexibility of having one handset that is tuneable to multiple modes and bands is quite important. Software radio attempts to address this need and one of the key requirements is to implement a low cost multi-band/multi-mode RF front end.

Since the significance of the duplexer in RF front end is primarily in FDD operation, this research focuses on implementing a solution for standards which use FDD operations. Some of the standards that require duplexing filters are AMPS, GSM (base stations), CDMA2000 and W-CDMA.

Near future wireless communication systems with multi-band capability require highly integrated programmable receiver architectures. Therefore the direct conversion receiver has become a more suitable choice due to significantly reduced component count and that has lead to significant size, cost and power reductions. The adaptive duplexer architecture in this thesis is based on a direct conversion receiver architecture.

The next chapter introduces the duplexer, TDD operation, FDD operation and discusses the need for an adaptive duplexer.
3.1 Introduction

Full duplex systems allow simultaneous radio transmission and reception between a subscriber and a base station, by providing two simultaneous but separate channels (frequency division duplex or FDD) or adjacent time slots on a single radio channel (time division duplex or TDD). These are the two duplex modes proposed in the 3rd Generation Partnership Project (3GPP). These two popular ways of achieving full duplex transmission and reception are explained at the beginning of this chapter. Then an introduction to the duplexing for FDD systems is presented. The duplexing requirement for multi-band systems is then looked at with an emphasis on its application to software radio. The effects of non-linearities in the receiver due to any Tx leakage signal are also examined.
3.2 TDD Systems

![Diagram of TDD Systems]

Figure 3-1  The principle of operation in TDD systems.

The principle of operation of a TDD system is shown in Figure 3-1. Here the radio communication system alternately transmits and receives a signal having the same frequency. There is no simultaneous transmission in both directions at a given instant of time. However, due to the fact that the data transmission rate is very much higher than the user’s data rate, it is possible to provide the appearance of a full duplex operation to the end user [51]. TDD is only possible with digital transmission formats and digital modulation, and is very sensitive to timing. It is for this reason that TDD has only recently been used, and only for indoor or small
area wireless applications where the physical coverage distances (and thus the radio propagation time delay) are much smaller than the many kilometres used in conventional cellular telephone systems [3]. Compared with FDD, TDD is advantageous in that there is no problem in allocating a frequency to each of the transmitting and receiving signals, and that the transmitting and receiving efficiencies (propagation losses) can be made substantially equal [51].

TDD systems use a switch inside the subscriber unit to switch between transmitter and receiver time slots, thus eliminating the need for a duplexer. This reduces the cost associated with duplexer component. Further this system reuses the filters, mixers, frequency sources and synthesizers, thereby eliminating cost and complexity.

3.3 FDD Systems

In frequency division duplexing (FDD), the subscriber and the base station use simultaneous radio transmission channels, so that they both may constantly transmit while simultaneously receiving signals from each other. The principle of operation in FDD systems is shown in Figure 3-2. Here, a pair of simplex channels \((f_1, f_2)\) with a fixed and known frequency separation (called as the duplexing frequency, duplexing offset or \(f_d\)) is used to define a specific radio channel in the system. The channel used to convey traffic to the mobile user from a base station is called the forward channel, while the channel used to carry traffic from the mobile user to a base station is called the reverse channel. Separate transmit and receive antennas can be used in the base station to accommodate the two separate channels. At the subscriber unit a single antenna is used for both transmission and reception from the base station, and a duplexer is used inside the subscriber unit to enable the same antenna to be used for simultaneous transmission and reception.
The basic problem in FDD radio design is transmitter to receiver isolation. Sufficient isolation is required to prevent the transmitter from desensitising and/or damaging the receiver. A frequency duplexing circuit is typically used to provide the required isolation. Alternatively two separate antennas could be used for sending and receiving signals in FDD systems.

To facilitate FDD, it is necessary to separate transmit and receive frequencies by about 5% of the nominal RF frequency, so that the duplexer can provide sufficient isolation while being inexpensively manufactured. In the United States’ (U.S.)
AMPS standard, the reverse channel has a frequency that is exactly 45MHz lower than that of the forward channel. In the UMTS W-CDMA standard, the duplexing frequency is 190MHz.

### 3.4 Introduction to the Duplexer

The duplexer is a device that isolates the receiver from the transmitter while permitting them to share a common antenna. The duplexer is often the key component that allows two way radios to operate in a full duplex manner. An ideal duplexer provides perfect isolation with no insertion loss, to and from the antenna.

A conventional duplexer is a three-port device and normally consists of two band-pass filters and an impedance transforming circuit to allow both filters to connect to a common antenna port (see Figure 1-1).

The band-pass filter in the receiving path (BPF$_{Rx}$ in Figure 1-1) stops the transmitter signal from jamming the receiver (receiver blocking or receiver desensitisation), by attenuating the transmit energy incident at the antenna port as shown in Figure 3-3. Receiver desensitisation, commonly called ‘receiver desense’, is caused when high RF signal levels enter a receiver’s antenna input. When desense occurs, the usual symptom is as though the desired signal was reduced; the signal becomes noisy or even fades out completely [52]. There could be a considerable difference between the frequency of the desensitising signal and the frequency of the desired signal. This interfering signal can be wideband noise and/or spurious emissions from the associated transmitter or other nearby transmitters [52].

The front-end of a radio receiver always consists of a low noise amplifier. The LNA’s main function is to amplify extremely low signals without adding noise and to amplify large signals without introducing any distortions [53]. The LNA provides sufficient gain to the smallest possible signal at its input. It is well
known that the gain of an amplifier starts to decrease, when the input power increases beyond a certain value (beyond its dynamic range). In FDD systems, a strong transmitter signal can leak into the receiver and bias the input of the receiver beyond this value. This desensitising phenomenon is mathematically analysed in Section 3.6.1.

Every transmitter emits signals other than those on the desired frequency. The function of the transmitter is to supply a modulated RF carrier at the required frequency and power level to the aerial system, with spurious components including harmonics below a required level [54]. The power amplifier is capable of increasing the power level of the Tx signal but at the same time it increases the noise floor across the whole amplifier bandwidth. The band-pass filter in the transmitting path (BPF$_{Tx}$ in Figure 3-4) stops this transmitter noise leaking into the

Figure 3-3  Receiver desensitisation.
receiver by attenuating the energy at the receive frequency present on the transmit carrier as shown in Figure 3-4.

![Diagram of transmitter noise](image)

**Figure 3-4** Transmitter noise.

The duplexing filters can also provide suppression of out-of-band spurious and image signals [55]. The duplexer can be the most expensive component in the mobile subscriber unit. Duplexers can be implemented in many ways. Some of the common implementations are based on hybrid ring, cavity notch and band-pass/band-reject designs.

A duplexer is normally characterised by Tx and Rx insertion loss and Tx-Rx isolation. As an example, if the Tx and Rx insertion loss is 3dB, then the Tx power should be increased by 3dB to compensate the loss in the Tx signal and Rx
noise figure is increased by 3dB. This means an increase in the insertion loss
degrades the receiver noise figure, and hence the receiver sensitivity, but increases
the Tx-Rx isolation. Therefore a good duplexer design should make accurate trade
offs among the Tx insertion loss, Rx insertion loss and Tx-Rx isolation.

An estimate of duplex circuit performance for W-CDMA [56] is listed in Table
3-1.

\begin{table}
\begin{center}
\begin{tabular}{|l|c|}
\hline
Duplexer Parameter & Anticipated Performance \\
\hline
Duplex filter Tx loss (dB) & <1.5 \\
\hline
Duplex filter Rx loss & <3.0 \\
\hline
Loss of system select switch and antenna feed (dB) & <1.0 \\
\hline
Combined Tx loss (dB) & <2.5 \\
\hline
Combined Rx loss (dB) & 2.0 to 4.0 \\
\hline
Duplex filter Tx-Rx isolation in Tx band (dB) & >60 \\
\hline
Transmitter power classes (hand-held and fixed mounted units) (dBm) & 33/27/24/21 \\
\hline
\end{tabular}
\end{center}
\caption{Anticipated performance for W-CDMA duplex arrangement [56]}
\end{table}

3.5 Need for an Adaptive Duplexer in Multi-Band Systems

The filters in FDD duplexers are generally not tuneable. It may not be possible to
implement duplexing filters with tuneable centre frequencies and tuneable
bandwidths for both 2G and 3G systems in the near future. Recent advances in
passive duplexer design have led to smaller, cheaper, and simpler designs [57],

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[58], [59], [60]. Tuneable passive duplexers that are tuneable over a limited frequency range are described in [61] and [62]. The first duplexer is designed for a dual-mode AMPS/CDMA cellular system and tunes the response of the Tx filter using voltage variable capacitors. The latter is designed for GSM1800/1900 and tunes the filters using voltage controlled varactor diodes. However these duplexer designs still do not cover multi-band requirements. Therefore the software radio that covers a number of bands will require more than one duplexer. This is shown in Figure 3-5 (a), where two duplexers are connected to the transmitter and receiver using a switch to enable operation on two bands. For multiple bands, the circuit becomes more of a problem. Having more than one duplexer is bulky and is not a viable solution for the mobile handset that has very strict space requirements. The same functionality can be implemented using an adaptive duplexer as shown in Figure 3-5 (b). The proposed adaptive duplexer architecture is explained in detail in Chapter 5.

Figure 3-5 The multi-band RF front end (a) using number of duplexers (b) using an adaptive duplexer.
3.6 The Effect of Receiver Non-linearity and the Tx Leakage Signal

Usually the Tx leakage power through the duplexer is significantly higher than any other RF input signal, including jamming signals and the desired signal. Tx leakage becomes the predominant factor that activates non-linearities in the receiver chain since this signal can interact with any other interference signal (in band, out of band or adjacent channel) or the desired signal itself to produce interference products in the receiver band. When the distortions are in band, they can no longer be filtered away. Therefore non-linearity effects need to be considered when determining the duplexer isolation.

An LNA is typically constructed from active devices that operate in the linear range, but the output is never quite linear. Non-linearities in the active devices (transistors) distort the amplified output signal. When there is a high transmitter leakage present at the input, a number of adverse effects including, desensitisation, blocking, cross modulation and intermodulation that degrade the receiver sensitivity can occur. In this section of the thesis, the effects of non-linearities are analysed.

3.6.1 Receiver Desensitisation

The desensitisation determines the receiver’s ability to operate successfully under strong interferers. As shown below, when the input power increases to a value that is high enough, the gain of the amplifier starts to decrease. This value is commonly characterised by a compression point measured at the input ($icp$) or at the output ($ocp$) of the amplifier. Hence when there is a high interfering signal, such as the Tx leakage, the gain of the desired signal is reduced causing the signal-to-noise ratio at the detector input to drop. This desensitisation effect can be mathematically analysed as follows.
According to Taylor’s series, the output of a nonlinear memory-less amplifier can be expressed as

$$y(t) = a_1x(t) + a_2x^2(t) + a_3x^3(t) + \ldots + a_i x^i(t) + \ldots$$ \hspace{1cm} Eq. 3-1

where $$x(t)$$ is the input to the system and $$a_i$$ is the $$i^{th}$$ order non-linearity constant.

In general, amplifiers have memory, i.e. the output signals are dependent on the previous instantaneous values of the input signals. If the reciprocal of the bandwidth of the input signal is much larger than the memory of the amplifier, the amplifier can be modelled as memory-less. For very wideband signals memory of the amplifier becomes a significant fraction of the reciprocal of bandwidth of the input signal. This results in a frequency dependent transfer function. The Volterra series approach should be used in such cases. Here, we use the power series approach because even W-CDMA, with 5MHz bandwidth, can be considered narrowband in terms of memory effects in the LNA.

When a desired signal ($$A_{Rx}, \omega_{Rx}, \phi_{Rx}$$) and a Tx interference signal ($$A_{Tx}, \omega_{Tx}, \phi_{Tx}$$) (this can be any interference signal) are present at the LNA input, i.e.

$$x(t) = A_{Rx} \cos(\omega_{Rx}t + \phi_{Rx}) + A_{Tx} \cos(\omega_{Rx}t + \phi_{Rx})$$ \hspace{1cm} Eq. 3-2

the output can be expressed as

$$y(t) = \left(\frac{a_1 A_{Rx}^2}{2} + \frac{a_2 A_{Rx}^2}{2}\right) + \left(a_1 A_{Rx} + \frac{3a_1 A_{Rx}^3}{4} + \frac{3a_1 A_{Rx} A_{Tx}^2}{2}\right) \cos(\omega_{Rx}t + \phi_{Rx}) +$$

$$\frac{a_1 A_{Rx}^2}{2} \cos 2(\omega_{Rx}t + \phi_{Rx}) + \frac{a_1 A_{Rx}^3}{4} \cos 3(\omega_{Rx}t + \phi_{Rx}) + \left(a_1 A_{Tx} + \frac{3a_1 A_{Tx}^3}{4} + \frac{3a_1 A_{Rx}^2 A_{Tx}}{2}\right) \cos(\omega_{Rx}t + \phi_{Rx}) +$$

$$+ \frac{a_1 A_{Rx}^2}{2} \cos 2(\omega_{Rx}t + \phi_{Rx}) + \frac{a_1 A_{Rx}^3}{4} \cos 3(\omega_{Rx}t + \phi_{Rx}) + \ldots$$ \hspace{1cm} Eq. 3-3
The second term represents the fundamental component of the desired signal. Some of its components are modified by the term $a_3$ and cause distortion. The receiver front end LNA causes compression in which case $a_3<0$. When the interference signal ($f_{Tx}$) is very large and the desired signal ($f_{Rx}$) is weak ($A_{Rx} << A_{Tx}$), the gain of the desired signal reduces to $[a_1A_{Rx} + (3a_3A_{Rx}A_{Tx}^2)/2]$, which is a function of $A_{Tx}$. This is called desensitisation. When the interference is large enough, so that $A_{Tx}^2 = (2|a_1|/3|a_3|)$, the gain of the desired signal drops to zero. Then the signal is blocked. The difference in dB between the blocking signal level and the sensitivity level is known as the receiver blocking ratio and is shown in Figure 3-6.

![Diagram](image)

**Figure 3-6** Receiver blocking.

The largest signal that can be received by a receiver establishes an upper power level limit of what can be handled by the system while preserving voice or data quality [63]. The dynamic range of the LNA (or the dynamic range of the receiver) which is the difference between the maximum receivable signal and the minimum receivable signal defines the quality of the receiver chain. According to the IS-98A/95B standard, the required minimum receiver sensitivity of a CDMA mobile station is $-104$dBm, while the maximum received signal power is $-25$dBm, giving a dynamic range of 79dB.
3.6.2 Cross Modulation

Cross modulation causes a transfer of the amplitude information of the Tx leakage signal to the desired receive signal or the adjacent channel. W-CDMA and CDMA systems exhibit an effective amplitude modulation (AM) of the Tx signal and one effect of interest is the Crest Factor [64]. There are two types of cross modulation.

**Type 1:** This type of cross modulation can occur when the modulation (or noise) on the amplitude of the Tx leakage signal transfers to the amplitude of the desired signal. This phenomenon is described in [36]. If the amplitude of the Tx leakage signal is modulated by \( (1 + m \cos \omega_m t) \), where \( m \) is the modulation index \((m<1)\), then the modulated leakage signal becomes \( A_{Tx}(1 + m \cos \omega_m t) \cos(\omega_r t + \phi_{Rx}) \). By substituting into the fundamental component of the Eq. 3-3 \((A_{Rx}<<A_{Tx} \text{ and } a_3<0)\) the output becomes

\[
y(t) = \left[ a_1 A_{Rx} + \frac{3}{2} a_3 A_{Rx} A_{Tx}^2 \left( \frac{m^2}{2} + \frac{m^2}{2} \cos 2\omega_m t + 2m \cos \omega_m t \right) \right] \cos(\omega_r t + \phi_{Rx}) + ...
\]

Eq. 3-4

Therefore the desired signal at the output of an LNA contains amplitude modulation at \( \omega_m \) and \( 2\omega_m \).

Cross modulation specifies the amount of AM which is transferred from an undesired signal to a desired signal. The percentage of modulation on the desired signal due to cross modulation can be expressed as [65]

\[
\%d = \frac{\%u(4P_u)}{(P_{ip} + 2P_u)} \quad \text{Eq. 3-5}
\]

where,

\( \%d \) = the percentage of modulation on the desired signal due to cross modulation

\( \%u \) = the percentage of modulation on the undesired signal

\( P_u \) = the power of the undesired signal

\( P_{ip} \) = the receiver third order input intercept point power
**Type 2:** This type of cross modulation can occur when there is an amplitude modulated Tx leakage signal \( A_{j_1}(1 + m \cos \omega_m t) \cos(\omega_{j_1} t + \varphi_{j_1}) \) and another interference signal (single tone jammer \( A_j, \omega_j, \varphi_j \)), which is close to the desired signal (adjacent channel). This can cause spectral contamination of the desired signal (Figure 3-7) by transferring the AM onto the adjacent channel, effectively increasing its bandwidth into the neighbouring channels. The \( \omega_j \) term of the output of the LNA becomes

\[
y(t) = \left[ a_1 A_j + \frac{3 a_3 A^3_j}{4} + \frac{3}{2} a_3 A_j A^2_{j_1} \left( 1 + \frac{m^2}{2} \cos 2 \omega_m t + 2 m \cos \omega_m t \right) \right] \cos(\omega_j t + \varphi_j) + \ldots \tag{Eq. 3-6}
\]

![Figure 3-7 Cross modulation due to the Tx leakage signal and a jammer signal.](image)

Aparin [66] analysed the jammer cross modulation from a Tx CDMA leakage signal in a common emitter circuit using the Volterra series and statistical theory. He showed that the cross modulation power depends on the circuit impedances not only at the jammer and Tx leakage centre frequencies, but also at the Tx baseband and at the sum and differences of the Tx leakage and jammer frequencies. In [67], Ko quantitatively dealt with cross modulation in CDMA
receivers and showed that duplexer isolation and LNA IIP3 are responsible for the cross modulation. He also showed that the required LNA IIP3 is about 4-5dBm with a duplexer isolation of 50dB.

The generated cross modulation power in CDMA can be expressed as [67]

\[ P_{\text{cm}} = 2P'_{\text{Tx}} + P_{\text{jam}} - 2\text{IIP}3_{\text{LNA}} + 20\log m(dB) + 12(dB) + \alpha(dB) \]  
\[ \text{Eq. 3-7} \]

Where

- \( P_{\text{cm}} \) = cross modulation power referred to input of the receiver expressed in dBm/1.23MHz
- \( P'_{\text{Tx}} \) = the Tx CDMA signal leakage power at the input of the receiver expressed in dBm/1.23MHz
- \( P_{\text{jam}} \) = the single tone jammer signal at the input of the receiver
- \( m \) = the modulation index
- \( \alpha \) = the amount of cross modulation power that penetrates into the in-band signal (the soaking factor)

The authors in [67] have also noted that \( P_{\text{cm}} \) increases by 2dB as the Tx leakage power increases by 1dB, while \( P_{\text{cm}} \) increases by 1dB as the single tone jammer power increases by 1dB.

The generated cross modulation power in W-CDMA can be expressed as [64]

\[ P_{\text{cm}}(\text{dBm}) = (2P'_{\text{Tx}} + P_{\text{jam}}) - 2\text{IIP}3 + c \]  
\[ \text{Eq. 3-8} \]

where \( c \) = the correction factor

It is often the cross modulation that determines the required IIP3 of the LNA in mobile W-CDMA receiver [68]. Here the single tone jammer is replaced by an adjacent W-CDMA channel, although not narrow band, the effect of the spectral spreading due to the cross modulation is still the same. The adjacent channel signal can be as high as –52dBm [64] and the transmit leakage signal can go as high as –20dBm, depending on the transmit power. Desensitisation occurs when
the resulting cross modulated interfering spectrum approaches the receiver sensitivity limit of –92.7dBm.

3.6.3 Intermodulation Distortion

When two sufficiently strong interfering signals are present at the input of the non-linear system, they will mix and create spurious signals at the output known as intermodulation products. One of these interfering signals can be the Tx leakage signal. These intermodulation products can be obtained as shown below.

When the Tx leakage signal \((A_{Tx}, \omega_{Tx}, \varphi_{Tx})\) and jamming signal \((A_j, \omega_j, \varphi_j)\) are present at the input, i.e.

\[
x(t) = A_{Tx} \cos(\omega_{Tx} t + \varphi_{Tx}) + A_j \cos(\omega_j t + \varphi_j)
\]

Eq. 3-9

Substituting into Eq 3-1 gives the following expression for the output

\[
y(t) = a_1(A_{Tx} \cos(\omega_{Tx} t + \varphi_{Tx}) + A_j \cos(\omega_j t + \varphi_j)) + a_2(A_{Tx} \cos(\omega_{Tx} t + \varphi_{Tx}) + a_3 \cos(\omega_j t + \varphi_j))^2
\]

\[
+ a_3(A_{Tx} \cos(\omega_{Tx} t + \varphi_{Tx}) + A_j \cos(\omega_j t + \varphi_j))^3 + ........
\]

Eq. 3-10

The frequencies, at which the intermodulation distortion (IMD) products are generated, is illustrated in Figure 3-8, and can be expressed as

\[
IMD = \pm p\omega_{Tx} \mp q\omega_j
\]

Eq. 3-11

\(p\) and \(q\) are positive integers

The second and third order intermodulation products are very common and are difficult to control. We limit our discussion in this section to these distortions only.
3.6.3.1 Third Order Intermodulation

Third order intermodulation products occur at $2\omega_{Tx} \pm \omega_j$ and $2\omega_j \pm \omega_{Tx}$ by mixing the second harmonic of $\omega_{Tx}$ with the fundamental $\omega_j$, and by mixing the second harmonic of $\omega_j$ with the fundamental $\omega_{Tx}$. Since these distortion products come from the cubic term in the non-linearity, they cause a broadening of the spectrum to approximately three times its original bandwidth, thus leaking into both lower and upper adjacent channels [68]. When the weak desired Rx signal is accompanied by a strong Tx leakage signal ($f_{Tx}$) and a jamming signal ($f_j$), third order IMD is generated. One of the products can fall in the receiver band corrupting the desired signal. This can occur particularly, when the difference between $f_{Tx}$ and $f_j$ is small. This effect is shown in Figure 3-9.

![Figure 3-8](image1.png)  
Intermodulation due to LNA non-linearity (two tone test signal). $2\omega_{Tx} - \omega_j$ and $2\omega_j - \omega_{Tx}$ are the third order intermodulation products.

![Figure 3-9](image2.png)  
The effect of third order intermodulation between the Tx leakage signal and a narrowband jammer.
The third order intercept point (IP3) is used to compare the linearity of different circuits. The IP3 is the theoretical point at which the desired signal and the third order distortion products are equal in amplitudes. This parameter is measured by a two tone test \((A_{Tx} = A_j = A)\) in which \(A\) is chosen to be sufficiently small so that higher order nonlinear terms are negligible and the gain is relatively constant and equal to \(a_1\) [36]. Then the third order inter modulation products are

\[
2\omega_{rx} \pm \omega_j : \frac{3a_3A^3}{4} \cos((2\omega_{rx} + \omega_j)t + (2\phi_{rx} + \phi_j)) + \frac{3a_3A^3}{4} \cos((2\omega_{rx} - \omega_j)t + (2\phi_{rx} - \phi_j))
\]

Eq. 3-12

\[
2\omega_j \pm \omega_{rx} : \frac{3a_3A^3}{4} \cos((2\omega_j + \omega_{rx})t + (2\phi_j + \phi_{rx})) + \frac{3a_3A^3}{4} \cos((2\omega_j - \omega_{rx})t + (2\phi_j - \phi_{rx}))
\]

Eq. 3-13

From the above two equations we can see that the third order intermodulation products increase in proportion to \(A^3\) as is shown in Figure 3-10. IIP3 is called the third order input intercept point and OIP3 is called the third order output intercept point. When there is a higher intercept point, the amplifier can handle higher level signals before it starts generating intermodulation products.

![Figure 3-10](image-url)  
*Linear and third order intermodulation products vs amplitude. \(a_3\) is normally negative, causing amplifier compression.*
3.6.3.2 Second Order Intermodulation

The even terms of the non-linear system characteristics cause second order intermodulation. When a strong modulated signal with time-varying envelope, experiences a second order intermodulation distortion, it produces a spurious baseband signal. This strong modulated signal can be the Tx leakage signal. The spurious baseband signal is proportional to the squared envelope and has twice the bandwidth of the interference signal. It can corrupt the desired baseband signal, and cannot be removed by filtering. This can be mathematically analysed as follows:

If the modulated Tx leakage signal is \( A_{\text{tx}}(1 + m \cos \omega_m t) \cos(\omega_t t + \phi_{\text{tx}}) \), the second order term in the output becomes

\[
a_x^2(t) = \frac{a_x^2 A_{\text{tx}}^2 (1 + m \cos \omega_m t)^2}{2} + \frac{a_x^2 A_{\text{tx}}^2 (1 + m \cos \omega_m t)^2}{2} \cos(2(\omega_t t + \phi_{\text{tx}})) \quad \text{Eq. 3-14}
\]

This baseband output component (the first term) can be expanded to show a DC component and other components including a high frequency component with twice the bandwidth \(2\omega_m\) of the Tx leakage signal. This DC component can be large compared to the desired receive signal. Therefore direct conversion receivers are very vulnerable to second order distortions. Figure 3-11 shows the second order distortion expected from a W-CDMA signal.

![Figure 3-11](Image)

*Figure 3-11 Spectrum of the baseband component due to second order distortion caused by the Tx leakage signal in a W-CDMA terminal.*
Another example of even order distortion can be presented as follows. Consider Eq. 3-11, with $p = q = 1$, the IMD products become

$$\omega_{tx} \pm \omega_j : a_2 A_{tx} A_j \cos((\omega_{tx} + \omega_j) t + \phi_{tx} + \phi_j) + a_2 A_{tx} A_j \cos((\omega_{tx} - \omega_j) t + \phi_{tx} - \phi_j)$$

Eq. 3-15

Therefore when a Tx leakage signal and another jamming signal are present at the input of the LNA and especially when the difference between these two interferers are very small, the output generates a low frequency beat. This becomes awkward in direct conversion receivers because this low frequency beat can leak through the mixer and corrupt the down converted desired signal. This effect is shown in Figure 3-12.

Figure 3-12  Effect of second order distortion from the LNA feeding through the mixer.

Second order distortion can be characterised using a second order intercept point, IP2 (Figure 3-13). This can be measured using a two-tone test, similar to IP3.

In [56], the required receiver IIP2, was shown to be about 47dBm for W-CDMA terminal with transmitter leakage of -50dB. This was based on the assumption that the second order product should be suppressed at least 10dB below the noise level. A multi-band receiver may require the use of a wide-band LNA to cater for the different RF bands. Wide-band input matching is necessary and this increases the frequency range over which interference signals can be present. A higher
second order input intercept point is required to lower the effect of interference originating from these other bands [69].

![Diagram of IIP2](image)

*Figure 3-13  Illustration of IIP2. The second order response has a slope of 2.*

### 3.6.4 DC Offset in Direct Conversion Receivers Due to the Tx Leakage Signal

In most receiver architectures, such as the superheterodyne receivers, an additional filter is introduced after the LNA to suppress the residual transmitter leakage signal. This protects other components down the receiver chain from being influenced by the transmitter leakage. For the direct conversion receiver architecture this is not necessarily effective. The Tx leakage can still be high at the input to the mixer, even if it is not high enough to activate the mixer nonlinearities. When some Tx leakage signal leaks from the LNA circuit (avoiding the filter after the LNA) or from the mixer input to the local oscillator port, it produces (by self-mixing) another source of second order intermodulation distortion. This results in a DC component in the baseband. This DC component varies with the square of the Tx leakage signal’s amplitude and has twice the bandwidth of that signal since the squaring (self-mixing) action makes the IM2 (second order intermodulation) signal bandwidth dependent on the amplitude modulation of the Tx leakage signal [70]. The effect of self-mixing in DCR due to Tx leakage signal is shown in Figure 3-14. The cure for this type of IM2 signal is to reduce leakage into the LO port, not to reduce distortion in the mixer itself.
A summary of receiver requirements for W-CDMA for the entire receiver and the part of the receiver [56] that follows the duplexer circuits is listed in Table 3-2.

<table>
<thead>
<tr>
<th>Requirement</th>
<th>Entire Receiver</th>
<th>After Duplex</th>
</tr>
</thead>
<tbody>
<tr>
<td>Noise figure</td>
<td>( \leq 9 )</td>
<td>( \leq 5 )</td>
</tr>
<tr>
<td>In-band selectivity (dB)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>First adjacent channel (5MHz)</td>
<td>( \geq 33 )</td>
<td>( \geq 33 )</td>
</tr>
<tr>
<td>CW interferer (10MHz)</td>
<td>( \geq 58 )</td>
<td>( \geq 58 )</td>
</tr>
<tr>
<td>Third adjacent channel (15MHz)</td>
<td>( \geq 58 )</td>
<td>( \geq 58 )</td>
</tr>
<tr>
<td>Modulation blocker (&gt;15MHz)</td>
<td>( \geq 58 )</td>
<td>( \geq 58 )</td>
</tr>
<tr>
<td>Intercept points (dBm)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>IIP2 (10MHz)</td>
<td>( \geq -16 )</td>
<td>( \geq -18 )</td>
</tr>
<tr>
<td>IIP2 (15MHz)</td>
<td>( \geq 8 )</td>
<td>( \geq 6 )</td>
</tr>
<tr>
<td>IIP2 (Tx)</td>
<td>( \geq 47 )</td>
<td></td>
</tr>
<tr>
<td>IIP3 (10/20MHz)</td>
<td>( \geq -17 )</td>
<td>( \geq -19 )</td>
</tr>
<tr>
<td>IIP3 (67.4/134.8MHz)</td>
<td>( \geq -8 )</td>
<td></td>
</tr>
<tr>
<td>Image rejection (&gt;85MHz) (dB)</td>
<td>( \geq 84 )</td>
<td>n/a</td>
</tr>
<tr>
<td>Oscillator noise sidebands at &gt;8MHz offset (dBc/Hz)</td>
<td>( \leq -129 )</td>
<td></td>
</tr>
</tbody>
</table>
3.7 Duplexer Requirements

There are five main requirements a duplexer should satisfy when considering multi-band applications:

1) The duplexer should be able to handle the frequency specifications of the different standards. Since the software radio has to work with the standards of second and third generation, the frequency region should be between 800-2200 MHz. The signal frequency bandwidths are from 25KHz (PDC) to 5MHz (UMTS) [66]. Table 3-3 lists the duplexing frequency and frequency bandwidth for the most common air interface standards. The last three entries have no TDD component and so require duplexing filters.

2) The duplexer must be able to handle the maximum output power of the transmitter. The transmitter power levels depend on the standards and the terminal classes or levels. FDD maximum output powers are shown in Table 3-4.

<table>
<thead>
<tr>
<th>Air Interface Standard</th>
<th>[Downlink] / [Uplink] (MHz)</th>
<th>Duplexing Frequency (MHz)</th>
<th>Channel Spacing</th>
</tr>
</thead>
<tbody>
<tr>
<td>PDC</td>
<td>[940-956] / [810-826]</td>
<td>130</td>
<td>25kHz</td>
</tr>
<tr>
<td>GSM 900</td>
<td>[890-915] / [935-960]</td>
<td>45</td>
<td>200kHz</td>
</tr>
<tr>
<td>DCS 1800</td>
<td>[1710-1785] / [1805-1880]</td>
<td>95</td>
<td>200kHz</td>
</tr>
<tr>
<td>PCS 1900</td>
<td>[1850-1910] / [1930-1990]</td>
<td>80</td>
<td>200kHz</td>
</tr>
<tr>
<td>IS-95</td>
<td>[824-849] / [869-894]</td>
<td>45</td>
<td>1.25MHz</td>
</tr>
<tr>
<td>UMTS FDD (Europe)</td>
<td>[1920-1980] / [2110-2170]</td>
<td>190</td>
<td>5 MHz</td>
</tr>
</tbody>
</table>
Table 3-4  The maximum output power

<table>
<thead>
<tr>
<th>Air-Interface Standard</th>
<th>Maximum Tx Output Power(dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>GSM 900</td>
<td>Terminal class</td>
</tr>
<tr>
<td></td>
<td>2: 39</td>
</tr>
<tr>
<td></td>
<td>3: 37</td>
</tr>
<tr>
<td></td>
<td>4: 33</td>
</tr>
<tr>
<td></td>
<td>5: 29</td>
</tr>
<tr>
<td>DCS 1800</td>
<td>Terminal class</td>
</tr>
<tr>
<td></td>
<td>1: 30</td>
</tr>
<tr>
<td></td>
<td>2: 24</td>
</tr>
<tr>
<td></td>
<td>3: 36</td>
</tr>
<tr>
<td>PCS 1900</td>
<td>Terminal level</td>
</tr>
<tr>
<td></td>
<td>1: 30</td>
</tr>
<tr>
<td></td>
<td>2: 24</td>
</tr>
<tr>
<td></td>
<td>3: 33</td>
</tr>
<tr>
<td>IS-95</td>
<td></td>
</tr>
<tr>
<td></td>
<td>3: 30</td>
</tr>
<tr>
<td>UMTS FDD (Europe)</td>
<td>Terminal level</td>
</tr>
<tr>
<td></td>
<td>1: 33</td>
</tr>
<tr>
<td></td>
<td>2: 27</td>
</tr>
<tr>
<td></td>
<td>3: 24</td>
</tr>
<tr>
<td></td>
<td>4: 21</td>
</tr>
</tbody>
</table>

3) The duplexer must be able to provide sufficient isolation to prevent receiver desensitisation. As previously described, when there is a high interfering signal, the gain of the desired signal is reduced, becomes noisy or even fades out completely (see Figure 3-3). In a conventional duplexer the duplexing filter in the receiver path, prevents the receiver desensitisation. A typical value for the acceptable maximum spurious Tx signal at the receiver input is –27dBm. The duplexer isolation requirement for W-CDMA would therefore be between 48dB to 60dB depending on the power class (see Table 3-4).

4) The duplexer must be able to provide adequate rejection of the transmitter noise at the desired receive frequency. The noise level of a system sets the minimum signal strength that can be detected. Receiver sensitivity levels for most common FDD air interface standards are shown in Table 3-5. If the requirement for the Tx noise is not to decrease the receiver sensitivity, then the Tx noise at the receiver input should be below the Rx noise floor (\(-174+N_{\text{Rx}} \text{ dBm/Hz}\)). According to the noise figure calculations in [56],
the noise figure requirement for W-CDMA entire receiver is $N_{RX} \leq 9$. This noise figure must be met in the presence of the Tx leakage signal. Even if the Tx chain has a relatively low noise, the duplexer requires an attenuation of about 40dB [4].

Table 3-5  The receiver sensitivity level

<table>
<thead>
<tr>
<th>Air Interface Standard</th>
<th>Reference Sensitivity Level (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>GSM 900</td>
<td>Small MS</td>
</tr>
<tr>
<td></td>
<td>Other MS</td>
</tr>
<tr>
<td>DCS 1800</td>
<td>Class 1 or Class 2</td>
</tr>
<tr>
<td></td>
<td>Class 3</td>
</tr>
<tr>
<td>PCS 1900</td>
<td>Normal</td>
</tr>
<tr>
<td></td>
<td>Other</td>
</tr>
<tr>
<td>UMTS (FDD)</td>
<td>12.2kbps</td>
</tr>
<tr>
<td></td>
<td>64kbps</td>
</tr>
<tr>
<td></td>
<td>144kbps</td>
</tr>
<tr>
<td></td>
<td>384kbps</td>
</tr>
</tbody>
</table>

5) The duplexer should add a minimum of additional insertion loss to the Tx and Rx paths. Insertion loss occurs in both the Tx and Rx paths of a duplexer. The greater the insertion loss, the lower the output level, the higher the noise figure and the greater the power dissipation and temperature rise in the filters. In a conventional duplexer, a high Tx insertion loss may reduce its maximum power handling capability. As shown in Table 3-1, for W-CDMA a 3dB duplex filter Rx loss is common and this will increase the Rx noise figure by a similar amount. Due to the small physical size requirements of the duplexer, this increase is unavoidable. However in the adaptive duplexer architecture (Section 5.2), careful consideration has been given to select components that minimize Rx insertion loss.
3.8 Conclusion

The duplexer in FDD system is used to provide simultaneous transmission and reception. The continuous presence of the Tx leakage signal in FDD systems create problems such as receiver desensitisation, cross modulation, intermodulation distortion and DC offset in direct conversion receivers due to non-linearities and leakage in the RF front end. These problems highlight the need for high performance duplexers.

Multi-band radios require multiple switched high isolation duplexers. This is difficult due to the stringent space constraints of multi-band software radio handsets because existing duplexers cannot be implemented in an integrated circuit. Further, multiple duplexers consume significant cost. Therefore an integratable adaptive/active duplexer with the same functionality is needed. An adaptive duplexer with a low isolation device and an active cancellation unit is proposed in this thesis. Adaptive duplexer eliminates many external components in a multi-band transceiver front end. The adaptive duplexer architecture is described and analysed in the following chapters.

There are five main requirements for an active duplexer in multi-band applications. The duplexer should be able to handle multi-bands (800-2200MHz), handle the maximum output power of the Tx (33dBm for UMTS FDD), provide sufficient Tx-Rx isolation (60dB), provide adequate rejection of Tx noise at the desired Rx frequency (40dB) and give minimum insertion loss to the Tx path and Rx path (1.5dB-Tx, 3dB-Rx). A literature survey of duplexing methods is presented in the next chapter.