The design of multi-band planar inverted-F antennas for mobile handsets with the aid of a novel genetic algorithm and their specific absorption rate

This item was submitted to Loughborough University's Institutional Repository by the/an author.

Additional Information:

• A Doctoral Thesis. Submitted in partial fulfilment of the requirements for the award of Doctor of Philosophy at Loughborough University.

Metadata Record: https://dspace.lboro.ac.uk/2134/34701

Publisher: © Omar Aqeel Sarereh

Rights: This work is made available according to the conditions of the Creative Commons Attribution-NonCommercial-NoDerivatives 4.0 International (CC BY-NC-ND 4.0) licence. Full details of this licence are available at: https://creativecommons.org/licenses/by-nc-nd/4.0/

Please cite the published version.
Please note that fines are charged on ALL overdue items.

FOR REFERENCE ONLY
The Design of Multi-Band Planar Inverted-F Antennas for Mobile Handsets with the Aid of a Novel Genetic Algorithm and their Specific Absorption Rate

by

Omar Aqeeq Saraereh, MSc

A Doctoral Thesis
Submitted in partial fulfilment of the requirements for the award of
Doctor of Philosophy of Loughborough University

May 2005

© by Omar Aqeeq Saraereh, 2005
To my family,
Abstract

Keywords: Planar Inverted-F Antenna (PIFA), Genetic Algorithm (GA) Optimisation Technique, Internal Antennas, Multi-band PIFA Handset Antenna, Specific Absorption Rate (SAR).

Wireless Communications have progressed very rapidly in recent years and mobile handsets are becoming smaller and smaller. Present-day mobile cellular communication systems include combinations of the AMPS, GSM-900, DCS-1800, PCS-1900, UMTS, and WLANs in the 2.4GHz and 5.2GHz bands. User requirements for access to the various aforementioned wireless telecommunication services have resulted in a rapid technological push to unify these different systems in a drastically decreased size single mobile handset. All this combined with strict limitations set for the energy absorbed by the users of mobile terminals has created a need for improved antenna solutions and better understanding of small antennas. The objective of this thesis is to develop novel multi-band handset antenna design solutions to satisfy the specific bandwidth requirements of mobile cellular communication systems. Devices having internal antennas have appeared to fill this need. In the past few years, new designs based on the planar inverted-F handset antenna (PIFA) have been used for handheld wireless devices because these types of antennas have low-profile geometry and can be embedded into mobile handsets. Therefore, the antenna topology proposed and researched to achieve the target of this thesis is the PIFA. The research involves the design of a genetic algorithm (GA) optimisation applicable to PIFAs. The technique is used to optimise the PIFA to produce a dual-band PIFA handset design suitable for personal communications at GSM-900/DCS-1800. Strategically extending and shaping the conductive plate of the optimised dual-band PIFA, a triple-band PIFA covering the DCS-1800/PCS-1900/UMTS bands is produced. The triple-band PIFA is extended to a quad-band design, by combining a meander-line planar monopole with the antenna structure to accommodate the GSM-900 band. The thesis also examines a penta-band PIFA handset design that is a potential candidate for small and low-profile structures to cover the AMPS/GSM-900/DCS-1800/PCS-1900/UMTS bands. Finally, a simulation and experimental study has been carried out to examine the relationship between the specific absorption rate (SAR) and the separation distance between a PIFA and a phantom head. The study also examines the relationship between the SAR and the antenna efficiency.
Acknowledgements

Many people supported me during the completion of this thesis with criticism, helpful assistance and invaluable ideas. This thesis would have never been possible without them. I would like to first thank Professor Yiannis Vardaxoglou for his guidance and encouragement. He was a wonderful supervisor whose assistance and motivation were greatly appreciated.

I would like to thank the Department of Electronic and Electrical Engineering at Loughborough University for the technical and financial support I have received from.

My greatest of thanks also goes to Mr. Patrick McEvoy, Dr. Mohan Jayawardene, Dr Chin Panagmuwa and Mr. James Kelly for their great help, invaluable contributions and discussions throughout my research. I wish to thank the antenna groups (CMCR and WiCR) at Loughborough University including Maria, Yiannis, Alex, Alford, George A, George G, George P, Nico, Umi, Guo and Shenhong. I owe a debt of gratitude for their advice, encouragement and friendship.

I feel a deep sense of gratitude for my father and mother who formed part of my vision and taught me the good things that really matter in life. I am grateful for my brothers and sisters Dhal and Najla for rendering me the sense and the value of brotherhood. Last but not least, my special gratitude is due to my nephews Hashem and Ahmed for the nice moments I spent with them.

Finally, I am very grateful for my wife Sahar Al-Hawawreh, for her love and patience during the PhD period. One of the best experiences that we lived through in this period was the birth of our daughter Noor, who provided an additional and joyful dimension to our life mission.
List of Publications from the Research


TABLE OF CONTENTS

ABSTRACT ......................................................................................................................................... I

ACKNOWLEDGEMENTS .................................................................................................................. II

LIST OF PUBLICATIONS FROM THE RESEARCH ......................................................................... III

TABLE OF CONTENTS ..................................................................................................................... IV

CHAPTER 1.0

1 INTRODUCTION .......................................................................................................................... 1

1.1 Common Mobile Handset Antennas ....................................................................................... 3

1.2 Objectives .................................................................................................................................. 4

1.3 Organisation of Thesis ............................................................................................................. 5

References ....................................................................................................................................... 8

CHAPTER 2.0

2 REAL-VALUED CODING GENETIC ALGORITHM (GA) OPTIMISATION TECHNIQUE ......................... 11

2.1 Introduction ............................................................................................................................. 11

2.2 Simulation Method (CST-MWS) ............................................................................................. 13

2.3 Genetic Algorithm Terminology and Concepts ....................................................................... 16

2.4 A Simple Genetic Algorithm .................................................................................................. 17

2.5 Proposed GA/CST-MWS Optimisation Technique ................................................................... 19

2.5.1 Dipole Antenna Front-End ................................................................................................. 21

2.5.2 GA/CST-MWS Structure Overview ................................................................................. 22

2.5.2.1 Initial Approximate Design and Chromosomal Representation .................................. 22

IV
CHAPTER 3.0

3 REAL-VALUED CODING GA OPTIMISATION TECHNIQUE APPLIED TO A PIFA HANDSET ANTENNA

3.1 Introduction

3.2 The Planar Inverted-F Antenna (PIFA)

3.2.1 PIFA Basis and Structure

3.2.2 PIFA Analysis

3.3 Dual-Band PIFA Handset Antenna Optimisation Example

3.3.1 Dual-band PIFA Design Procedure

3.3.2 Dual-band PIFA Input Parameters for Optimisation

3.3.3 Dual-band PIFA GA/CST-MWS Optimisation

3.3.4 Dual-band PIFA Optimisation Results and Discussion

3.3.5 Measured Results of Optimised Dual-Band PIFA

3.3.6 Effect of PIFA Parameters on Resonant Frequency and BW

3.4 Conclusion

References

CHAPTER 4.0

4 TRIPLE-BAND PIFA HANDSET DESIGN

4.1 Introduction

4.2 Triple-Band PIFA Design
CHAPTER 6.0

6 SAR AND EFFICIENCY STUDY FOR A DUAL-BAND PIFA HANDSET ANTENNA (GSM900/DCS1800) ................................................................. 149

6.1 Introduction ............................................................................................... 149

6.2 Specific Absorption Rate (SAR) Theory .................................................... 150

6.3 SAR and Efficiency Study for a Dual-Band PIFA ..................................... 151

6.3.1 Dual-Band PIFA Design Configuration .................................................. 152

6.3.2 Dual-band PIFA Simulation and Experimental Results ......................... 154

6.3.3 Dual-band PIFA SAR and Efficiency Simulation Results ...................... 156

6.3.4 Dual-band PIFA SAR Measurement Results .......................................... 165

6.3.5 Magnetic Field Behaviour in the Presence of Phantom Head............... 169

6.4 Conclusion .................................................................................................. 173

References ............................................................................................................. 174

CHAPTER 7.0

7 CONCLUSIONS AND FUTURE WORK ....................................................... 175

7.1 Conclusions ............................................................................................... 175

7.2 Recommendations and Future Work ........................................................ 179

7.2.1 Increasing GA/CST-MWS Efficiency and Effectiveness ....................... 179

7.2.2 Reducing PIFA Size and Improving Performance ..................................... 181

References ............................................................................................................. 183

VII
APPENDICES

APPENDIX I

GA/CST-MWS FRONT-END AND REPRODUCTION OPERATIONS............ 187

I.1 Introduction.................................................................................................. 187

I.2 Front-End Design........................................................................................ 187

I.3 Reproduction Operations.............................................................................. 189

I.3.1 Fitness Operation Using Linear Scaling Function................................. 189

I.3.2 Selection Operation Using Roulette Wheel Selection......................... 190

I.3.3 Crossover Operation Using Intermediate Recombination...................... 191

I.3.4 Real-Valued Mutation Operator of the Breeder GA.............................. 193

I.3.5 Reinsertion Operation Using Random Reinsertion.............................. 195

APPENDIX II

SIMULATED AND MEASURED AVERAGE SAR OVER 1G FOR ANTENNA
#1 AND ANTENNA #2..................................................................................... 196

References............................................................................................................. 200
Chapter 1

1 Introduction

As mobile communications progress rapidly, an increasing number of frequency bands are being used. Therefore, it is desirable for a single mobile handset to access the additional services such as voice, data and video at anytime and any place. The mobile handset size is one of the main concerns for the user. It is clear within the past few years that the size of mobile handset has shrunk rapidly, promoting internal antennas as the best candidates for a small size mobile handset [1]. As a result, the demand for a smaller mobile terminal antenna that has the ability to operate at the standard mobile communication cellular bands has increased. This makes it necessary to consider novel compact, low cost and complex small antennas as a part of the solution.

It is a commonly agreed fact that electrically small antennas have poorer performance than larger ones [2,3]. The complexity of small antennas is a result of the interrelationship between their size, efficiency and bandwidth [2]. Any of these antenna parameters can be improved only at the expense of the other. For instance, a small single-band antenna with a high efficiency will always have a narrow bandwidth. Increasing the bandwidth requires reducing the efficiency, increasing the size or both. Therefore, during the development stage of such antennas it is important that the designer keeps in mind many aspects like the physical size, bandwidth, efficiency, specific absorption rate (SAR) of the antenna and the overall cost. These specifications add more complexity to the design of such antennas.

To achieve increasingly complex small antenna designs, a systematic and accurate optimisation technique is required to shorten the design cycle that would otherwise be carried out by an engineer iterating manually. Existing designs of antennas have evolved empirically using prototypes that are
developed with intuitive trials which needs always the engineers concentration and intervention, prototype manufacture is slow compared to simulated iterations and design strategies are prone to error in human judgement. More recently the Genetic Algorithm (GA) has been successfully applied to electromagnetic problems, especially, to assist the antenna designer in the optimisation of various antenna types [4,5].

A GA is a stochastic search procedure modelled based on Darwin’s theory of natural selection and evolution. In a GA a set of suggested solutions or population of potential solutions is caused to evolve toward a global optimum solution. Evolution toward a global optimum occurs as a result of evaluating the suggested solution against a defined target through a fitness-weighted selection process. Exploration of the solution space is accomplished by recombination and mutation of existing characteristics present in the current population.

Most of the presented GAs in the literature have focused on optimising different types of antennas like wire, microstrip, patch and fractal antennas [6-9]. The representation method used to describe antenna parameters for these GAs was the binary format. Unfortunately, little research has been reported on optimising internal antennas, for mobile handsets like the planar inverted-F antenna (PIFA), using GAs. The few proposed ones have also focused on using the binary format for the antenna parameters representation [10,11]. This thesis has employed a real-valued coding GA optimisation technique coupled with computer simulation technology software CST MICROWAVE STUDIO® (CST-MWS). The whole programme is written in Visual Basic. The validity of the technique is demonstrated through some antenna designs. As a result, a promising optimisation technique suitable for mobile handset antenna designs is reported in this thesis.
1.1 Common Mobile Handset Antennas

Over the years, many types of antenna have been used in mobile handsets. They can be roughly placed in two categories. First, there are antennas that are similar to monopoles. Externals such as normal-mode helical, meander, and retractable antennas are of this type [12,13]. More recently, these antennas have been used internally, but they are not that common. These antennas do not work when they have a ground underneath them, so they cannot function with a populated section of the PCB below them.

The Second category is internal antennas that work on top of a ground plane, such as microstrip antennas (MSA) and PIFAs [12,13]. One of the current trends in mobile communications is the increasing popularity of internal antennas. Traditional external antennas like whip and helix have been replaced in many applications by internal antenna solutions. Owing to the protective casing of the terminal, internal antennas are mechanically more reliable than external whips and helices. A terminal with an internal antenna is also more convenient to handle. The low cost of such antennas makes them attractive also from the industrial point of view.

The internal antenna topology proposed and researched in this piece of work is the PIFA type antenna. The PIFA has desirable features like multi-frequency behaviour with high efficiency, low profile, low cost and lightweight [1]. These features make this antenna the best candidate to achieve the objective of this thesis in designing compact internal multi-band handset antenna. Recently PIFAs have attracted much attention, and a variety of dual-band or triple-band PIFAs, suitable for applications in mobile phones, have been demonstrated [14-23]. The PIFA usually occupy a compact volume and can be integrated within the mobile phone housing, leading to concealed or internal mobile phone antenna with a length that is approximately a quarter wavelength. In addition, in comparison to the conventional common whip antennas, which exhibit an omnidirectional radiation pattern, PIFAs have the advantage of relatively smaller backward radiation towards the mobile phone.
user. This suggests that the level of electromagnetic energy absorption by the user's head (SAR) can be reduced. These advantageous characteristics have led to the appearance of several multi-band PIFA designs, most of them capable of quad-band operation, in the literature [24-29] recently. Unfortunately, the reported designs have many serious drawbacks including the size and structure of the antenna. All antennas reported in [24-29] have a large size making them difficult to embed in the modern compact mobile handset. Also, some of these antenna structures were very complex, which makes them difficult to fabricate. For example, antenna designs reported in [24,25,28] have two layers of antenna elements and are loaded with many parasitic elements. The major limitation of many low-profile antennas, such as PIFAs, is narrow bandwidth. Bandwidth in these antennas is almost always limited by impedance matching. A variety of techniques for broadening bandwidth, tuning modes and improving match have been proposed in this thesis to overcome this problem.

1.2 Objectives

The main objective of this thesis is to satisfy the increasing demand for a single, compact, internal and small multi-band handset antenna covering the different mobile communication standard operations (AMPS/GSM-900/DSC-1800/PCS-1900/UMTS). Clearly, the mobile terminal should not only be capable of operating within the existing infrastructure, but also provide high performance operation for systems that may not have been standardised. With increasing importance of indoor systems such as DECT, Bluetooth/WLAN and HIPERLAN providing alternative schemes, research is now geared towards emerging markets of the 3G systems such as the UMTS, integrating several systems into common communication protocol.

The work in this thesis has been divided to three parts. The objective of the first part is to build a GA optimisation technique to shorten the design cycle of PIFA and increase the understanding of such complex antenna parameters. The GA has a faster processing time and does not need the user
intervention compared to traditional manual optimisation. The objective of the second part of the thesis is to design novel multi-band PIFAs. This was achieved by applying different parametric studies for sensitive parameters to integrate more bands, improve the bandwidth and tune the resonant frequencies of single or dual-band PIFAs. The objective of the third part of the thesis is to study the relationship between SAR parameter and antenna efficiency and separation distance for a PIFA placed in the vicinity of a phantom head.

1.3 Organisation of Thesis

Chapter 1 covers the introduction and the objectives of the thesis. It reviews some of the common handset antennas used for mobile communications in the literature. The PIFA type antenna is briefly introduced and the advantages of its low profile, multi-band behaviour over other common handset antennas are highlighted.

The main objective of Chapter 2 is to introduce the reader to a real coding genetic algorithm (GA) optimisation technique suitable for antennas design. The chapter shows the advantages of using the GA over other optimisation techniques. The GA optimisation technique is demonstrated using a dipole antenna example as a proof of concept.

Chapter 3 applies the presented real-valued coding GA optimisation technique to a PIFA mobile handset antenna. A single-band PIFA, after cutting slots into its top planar plate, is fed to the GA optimisation technique. The optimisation target is to achieve a dual-band PIFA design suitable for personal communication handsets at GSM-900MHz and DCS-1800MHz. Measured results are presented for comparison with simulation. Effect of PIFA parameters on the antenna's resonant frequency and bandwidth are also investigated.
In Chapter 4, the dual-band PIFA design proposed in Chapter 3 is extended to a low profile triple-band PIFA design after subtle modifications to the antenna parameters. The antenna is suitable for mobile communication handsets operating at DCS-1800MHz/PCS-1900MHz/UMTS bands. Details of the proposed antenna are described; experimental results of a constructed prototype are presented. The effects of radiating element slots and shorting pin (radius and position) on the resonant frequency and bandwidth are studied. The specific absorption rate (SAR) distributions in a head exposed to electromagnetic waves radiating from the proposed antenna prototype are predicted. A triple-band PIFA was extended to cover the GSM-900MHz band. Where a planar meander-line monopole is added to the triple-band PIFA structure to achieve a quad-band design. The quad-band antenna structure and the simulation and experimental results are presented and discussed.

Chapter 5 presents a compact penta-band PIFA design suitable for mobile handsets operating at AMPS (824-894) MHz, GSM-900 (890-960) MHz, DCS-1800 (1710-1880) MHz, PCS-1900 (1850-1990) MHz and UMTS (1900-2170) MHz bands. The chapter investigates the influence of the ground plane on the impedance bandwidth and resonant frequency of a PIFA. Different studies are applied to the ground plane in order to cover more frequency bands through variation in the geometry of the ground plane itself. These modifications in the ground plane involve slotting the ground plane, varying its length and considering a non-metallic ground plane underneath the antenna element. The penta-band PIFA is produced after studying and evaluating the performance of several designs. These designs are modelled in CST-MWS and their prototypes are fabricated to measure the input return loss, efficiency, gain and radiation patterns in an anechoic chamber. SAR measurements are also applied to these designs for further characterisation.

In Chapter 6 a SAR study is applied to a dual-band PIFA operating at GSM-900MHz and DCS-1800MHz bands. The effect of varying the distance between the PIFA and the human head on the SAR value is studied at the uplink frequencies of both bands. Furthermore, the relationship between the SAR values against the efficiency values is investigated using two identical
antennas, which differ only in the feeding mechanism. The impact of efficiency on the SAR values is also studied. The SAR distributions in a head exposed to electromagnetic waves radiated from the dual band PIFA are simulated and measured.

Conclusions and future work flowing from this thesis are presented in Chapter 7. The future work involves both improvements to the GA optimisation technique and the presented multi-band PIFAs designs.
References


Chapter 2

2 Real-Valued Coding Genetic Algorithm (GA) Optimisation Technique

2.1 Introduction

The existing designs of different types of antennas have evolved empirically using prototypes that are developed manually with intuitive trials. Unfortunately, the empirical method has many disadvantages including the frequent need for the engineer's intervention by intuitive and creative ideas, prototype manufacture is slow compared to simulated iterations and design strategies are prone to error in human judgement. A computer optimisation algorithm based on one of the available search algorithms should eliminate the intuitive trials of the inefficient empirical method and reduce the design cycle period. In the literature, there are various types of search algorithms that can be used for optimisation. These include traditional optimisation techniques, Tabu search, simulated annealing and genetic algorithm (GA).

The traditional optimisation techniques search for the best solutions, using gradients and/or random guesses [1]. Unfortunately, Gradient methods quickly converge to a minimum, once an algorithm is close to it. Furthermore, they could easily be trapped in local minima, requiring gradient calculations, working only on continuous parameters and being limited to optimise a few parameters [1]. Random-search methods do not require gradient calculations, but tend to be slow and can easily be trapped by a local minima [1]. On the other hand, the Tabu search and simulated annealing methods converge quickly to the required solution, given that the algorithm control parameters are optimal [2].
For antenna design, it was recently noted that the GA is the most applied optimisation technique [3-5]. Briefly, GAs are search algorithms based on the mechanics of natural selection and natural genetics. They operate on a population of randomly suggested solutions created within a defined range. By combining survival of the fittest of these solutions through information exchange they produce an approximate solution to a specific problem with some of the innovation ideas provided by the user [5]. In every generation, new sets of offspring (chromosomes) are created using random features of the preceding best-fit generations. They efficiently exploit historical information to speculate on new search points with expected improved performance.

Extensive research efforts have been reported on the applications of GAs to electromagnetic optimisation and problems in other fields, ranging from engineering to computer science to finance [3,4]. Goldberg [5] showed that the GA differs substantially from more traditional search optimisation methods. The GAs search a population of points in parallel, not a single point and use probabilistic transition rules, not deterministic ones. They do not require derivative information or other auxiliary knowledge; only the objective function and corresponding fitness levels influence the GAs performance and convergence. GAs also have the advantages of optimising continuous and discrete parameters, dealing with a large number of parameters, being well suited for parallel computing and providing multi close solutions.

In this chapter a real coding GA optimisation technique coupled with a computer simulation package (described in the next section) called CST MICROWAVE STUDIO® (CST-MWS) is introduced. Most of the literature reported on antenna optimisation has been focusing on using binary coding GAs for optimising antennas [6-9]. Not much research has been reported on optimising antennas using GAs with a real-valued representation for the parameters that describe the antenna configuration [10,11]. Recent research showed that the real-valued coding method has better performance and many advantages over the binary coding [12,16]. Here, the GA coupled with CST-MWS (GA/CST-MWS) has been applied to optimise a simple dipole antenna. The dipole antenna is chosen as a proof of concept because of its simplicity.
and the short period of time required for simulation. The average time required to simulate a dipole antenna using CST-MWS was 2 minutes using a Xeon-2GHz dual processor PC with 1GB of RAM.

2.2 Simulation Method (CST-MWS)

The recent progress in numerical mathematical methods and computational technologies has resulted in an increase in the number of simulation packages for solving large and complex electromagnetic problems. These numerical methods include Finite Difference Time Domain (FDTD) Method, Finite Integration Technique (FIT), Finite Element Method (FEM) and Method of Moments (MoM). They can be divided to two types based on the solver employed, a time domain or a frequency domain solver. FDTD and FIT employ the time domain solver, which can cover a wide frequency range in a single simulation run. In the time domain solver, Maxwell’s equations are modified to central-difference equations, then discretised and implemented in the modelling software. Equations are solved in an alternating manner, where the electric field E and the magnetic field H are alternately solved for given instants in time per unit cell. FEM and MoM employ the frequency domain solver, which deals with a single frequency at a time, although most popular software codes allow the solution to iterate over several frequencies. A simulation package CST-MWS employing the time domain solver has been chosen to run the simulations in this thesis.

CST-MWS is a general-purpose electromagnetic simulator based on the Finite Integration Technique (FIT), first proposed by Weiland in 1977 [17]. Unlike most numerical methods, the FIT discretises the integral form of Maxwell’s equations (2.0), rather than the differential form, on a pair of dual interlaced discretisation grids.
The FIT can be applied to different frequency ranges, from DC to THz. On Cartesian grids, the time-domain FIT can be shown to be equivalent to FDTD. However, whereas the classical FDTD has the disadvantage of the staircase approximation of complex boundaries, the Perfect Boundary Approximation (PBA) [18] technique applied in conjunction with FIT maintains all the advantages of the structured Cartesian grids, while allowing an accurate modelling of curved boundaries. The FIT technique combined with the PBA is the core of CST-MWS program. Due to the combination of FIT with the PBA technique, curved shapes can be modelled fast and accurately. Thus, even complex antenna shapes as present in many mobile phones can be simulated together with realistic models of the human head within a reasonable time compared to FDTD.

FDTD requires that the entire computational domain be meshed, and the grid spacing must be small compared to the smallest wavelength and smaller than the smallest feature in the model. Very large computational domains can be developed, which result in correspondingly long solution times. Models with long, thin features are difficult to model in FDTD because of the very small mesh size required, leading to an excessively large computational domain (long simulation time). For such structures, a "thin sheet" model is present in CST-MWS, which allows the use of a coarse mesh, while taking into account accurately the thin sheet's thickness. The mesh of the structure is generated automatically by an expert system, not only based on the geometry but also on the physical behaviour of the electromagnetic fields. For instance, frequency range, material parameters, as well as analytically well-known singularities directly influence the mesh construction. The advantage of this meshing technique compared to other refinement techniques such as adaptive
meshing becomes obvious if different similar structures have to be analysed, necessary during a parameter sweep or optimisation procedure. The knowledge and experience of former similar applications directly provides an optimised mesh at the correct places, without doing several simulation passes.

In CST-MWS, three different solvers can be used for solving high frequency electromagnetic problems: transient, frequency-domain and eigenmodes solver [19]. The transient (time-domain) solver is probably the most interesting for RF device modelling. It will be used to produce the results throughout this thesis because it allows time-efficient simulation of large structures, yielding broadband results with just one calculation. Due to the special, FDTD-like algorithm, there is no system of equations to be solved; therefore the necessary memory and computing time resources are much lower than for other numerical discretisation techniques (such as the Finite Element Method). The user can choose the time variation of the field source – sinusoid-modulated Gaussian pulse, rectangular digital pulses, or any other user-defined excitation. Thus, many parameters calculation including S-parameter are possible.
2.3 Genetic Algorithm Terminology and Concepts

There are some important terminologies and concepts that form the basic structure of GA optimisers [4]. Figure 2.1 shows these terminologies and the following summarise the most important concepts, many of which will be expanded in later sections.

- Genes and Chromosomes (Individuals):
As in natural evolution, the gene is the basic building block in the GA optimisation. Genes are a representation of the parameters used for optimisation. A chromosome is a one-dimensional vector of genes and can be represented in the GAs as binary strings or strings of real numbers.

![Figure 2.1 Concepts and terminology associated with genetic algorithms [4]](image)

- Population and Generations:
A population is a set of chromosomes used as a trial solution in the GA. This set of chromosomes is usually created randomly using a random number generator that uniformly distributes numbers in the desired range. The iterations in the GA are called generations, since there is a new generation with new features is produced by each cycle of the GA, using the reproduction operations.
Parents:
Once a population is created, pairs of chromosomes are selected from the population according to their relative fitness and designated as parents.

Children (offspring):
Children are the new chromosomes generated from the selected pair of parents by applying some reproduction operators.

Fitness:
Using an objective function to define the optimisation goal, fitness is a measure of how individuals have performed in the problem domain.

2.4 A Simple Genetic Algorithm

A flow chart of a simple GA is depicted in Figure 2.2. The first step in a simple GA process is to represent the parameters describing the problem in the form of a chromosome. This chromosome will then contain all the essential parameters about a particular suggested solution. There are two different methods of chromosomal representation (coding): binary format and real value representation [3,4]. When binary coding is used, \( N \) bits of zeros and ones are used to represent each parameter, where \( N \) can be different for each parameter. Placing all the encoded parameters one after another creates the chromosome in a form of a one-dimensional vector of zeros and ones. By decoding the chromosome, the original values of the parameters can be found.

When a real value representation is used, all the parameters are stored as real values in a one-dimensional vector that represents the chromosome. This means that encoding and decoding steps are not required when a real value representation is applied. Comparing both methods shows that a binary representation is more time consuming, since the encoded parameters need decoding each time. This is a clear disadvantage of the binary representation. Another disadvantage of binary representation is explained by Samii et al. [4]. It was found that binary coding does not always work well in the context of numerical function optimisation because similar numbers are not always coded similarly. For instance, using standard 4-bit binary coding, the number 7
is represented by 0111, whereas number 8 is represented by 1000. The number of positions in which two strings differ is called the Hamming distance between them, and the gaps between coding of adjacent integers separated by a Hamming distance greater than one are called Hamming cliffs [14,15]. The existence of Hamming cliffs implies that binary coding can convert an initially simple, unimodal function of a real number into a deceptive binary function. Recently, the binary coding has been replaced largely by the real value coding, which is a more efficient coding scheme and has many advantages over the binary one. At the very least, there is no penalty incurred for using real values. Further details of real value representation are described in [12,16].

The GA/CST-MWS used the real value representation. Therefore, the chromosome is constructed as a one-dimensional vector of real value parameters. Each chromosome has an associated objective value, assigning a relative merit to that chromosome. The algorithm begins with a large list of random chromosomes (suggested chromosomes). Objective values are
evaluated for each chromosome. After that the chromosomes are ranked, according to their respective objective values. Highly fit chromosomes, relative to the whole population (all the suggested chromosomes), have a high probability of being selected to become parents; whereas less fit chromosomes have a correspondingly lower probability of being selected. So genes that survive become parents, by crossover some of their genetic material to produce two new offspring (children). The parents reproduce enough children to keep the number of chromosomes constant after each generation. Mutations cause small random changes in a chromosome. Objective values are evaluated for the offspring and the mutated chromosome, and the process is repeated as required. The algorithm stops after a set number of generations, or when an acceptable solution is obtained. Briefly, the main duty for the GA is the reproduction process, which consists of three distinct steps: selection, crossover and mutation as shown in Figure 2.2.

2.5 Proposed GA/CST-MWS Optimisation Technique

The GA/CST-MWS code consists of two parts. The first part is the 'front-end' for inputting parameters of the antenna geometry and the second part contains the GA reproduction operations. The optimisation technique will be demonstrated by optimising the parameters of a dipole antenna shown in Figure 2.3. The dipole parameters are fed to the GA/CST-MWS as an initial chromosome through the front-end. The front-end then uses the initial chromosome to create a population of randomly suggested chromosomes. The GA/CST-MWS reproduction operations process the created population iteratively until convergence towards an optimum solution, is achieved. The GA/CST-MWS was written using Visual Basic (computer language) because of the ease of interfacing with CST-MWS.

The target of optimisation is to evaluate the input return loss for each generation of simulations (new population) to maximize the impedance bandwidth at the resonant frequency. Other fundamental parameters of antennas such as: gain, efficiency, directivity, beam width and radiation
pattern could be used as targets for optimisation instead of the input return loss. Unfortunately, all of these require simulating the antenna twice thus reducing the optimisation speed. It is also possible to use more than one target for optimisation. This leads to multi-objective GAs, which will be discussed in Chapter 7.

Figure 2.3 CST-MWS solid modelling front-end, dipole geometry
2.5.1 Dipole Antenna Front-End

Figure 2.4 shows the front-end of the Visual Basic programme designed to input the dipole parameters. Antenna parameters include Dipole Outer Radius, Dipole Length and Dipole Air Gap. GA parameters include Number of Individuals, Number of Selected Individuals and Maximum Number of Generations. Simulation parameters include Start Frequency, Stop Frequency and Monitor Frequency. The procedure of building a front-end is described in detail in Appendix I. The front-end’s task is to create a population of chromosomes within a defined range for each parameter using the first geometry parameters fed to the GA. The GA operates using a fixed population size specified by a keyboard input to the front-end shown in Figure 2.4. The GA processes the population of geometries simultaneously using a cluster of 15 computers.

![Figure 2.4 Dipole front-end input parameters as variables for a GA optimisation](image-url)

The population length (number of suggested chromosomes) is described by a parameter called Number of Individuals and the width, which is always equal to the number of parameters used to describe the problem. Typically, a population is composed of between 30 and 100 individuals. A
variant called the micro GA uses smaller populations, ~10 individuals, with a restrictive reproduction and replacement strategy in an attempt to reach real time execution [13]. The size of the population used should be chosen depending on the complexity of the problem under consideration and the computing resources available. Small population sizes are used during testing, and successively larger ones are used as more realistic problems are addressed.

Consequently, the output expected from the front-end is a population of geometries with randomly created parameter values. Once the front-end loaded with suitable parameter values a geometry files will be created automatically and read by CST-MWS to build different antenna models. Then the antenna models will be simulated and the $S_{11}$ values will be fed to the GA process for evaluation.

### 2.5.2 GA/CST-MWS Structure Overview

The basic blocks that construct the GA/CST-MWS are shown in Figure 2.5, where an overview of this optimisation technique is provided. The different GA/CST-MWS operations will be explained through the example of dipole optimisation. Some of these operations are common to all GAs and so are included in Appendix I. The optimisation target for the dipole is to resonate at 0.9GHz with a maximum impedance bandwidth at 10dB return loss. Therefore, an evaluation of the input return losses at the different generations created by the GA/CST-MWS is required. Individuals (suggested solutions) are evaluated by looking at the difference between the individual $S_{11}$ and the target $S_{11}$ and reducing the error distance between them. In the next few pages, each process in Figure 2.5 will be explained in detail, including description of its inputs and outputs.

#### 2.5.2.1 Initial Approximate Design and Chromosomal Representation

The first step (Block 1) in the optimisation process is to load the front-end shown in Figure 2.4 with the parameter values of an initial dipole design.
These parameter values will be called the seed input parameters because they form the first seed geometry of the dipole. This initial seed will be used later to create a population of random individuals (geometries or chromosomes) with different dimensions. The seed geometry parameter values of the initial dipole design are as follows:

\[
\begin{align*}
\text{Dipole Outer Radius} & = 3.0 \text{ mm} \\
\text{Dipole Length} & = 166.6 \text{ mm} \\
\text{Dipole Air Gap} & = 8.3 \text{ mm}
\end{align*}
\]

The input return loss for the seed geometry was simulated and a resonant frequency of 0.78GHz was obtained for the chosen parameter values. The reason for choosing these parameter values is that they make the dipole resonate at a resonant frequency relatively close to the target (0.9GHz). This should help the GA to converge toward a solution rapidly. The real value chromosomal representation for the dipole parameters will be as follows:

\[
\text{Chromosome} = [\text{Outer Radius}, \text{Length}, \text{Gap}] = [3.0, 166.6, 8.3] \text{ (mm)}
\]

The real numbers in the second vector are the parameter values of the seed chromosome that belong to the approximate initial dipole design (seed dipole).

The second step (Block 2) in the optimisation process is to create an initial population of suggested chromosomes by creating random values from each parameter in the seed chromosome. Generating a population of a number of potential solutions is achieved by using a random function generator \((Rnd)\) [20], which returns a single random number that is always in the range \(0 \leq Rnd < 1\). If \(Rnd\) truly produces random numbers, every number between 0 and 1 (but not including 1) has an equal chance (or probability) of being chosen each time \(Rnd\) is called. Multiple calls to \(Rnd\) will not generate any predictable pattern of numbers.
Genetic Algorithm
Optimisation Technique
Chapter 2

Figure 2.5 Genetic algorithm overview
It is very important to highlight that the GA is not a random search optimisation process [12]. It is an evolutionary simulation that relies mainly upon certain random processes in order to function properly, but the results should be distinctly non-random since they are obtained by applying the principle of survival of the fittest to produce improving approximations to a solution. It is also fair to say that the overall quality of the optimisation process and consequently the speed of convergence, can only be as good as the randomness of the random number generator used to simulate the random process upon which the GA depends.

For the dipole optimisation a population of 32 suggested random chromosomes has been chosen. Each chromosome contains 3 parameters. Therefore, the population is a two dimensional array of 32 rows x 3 columns as shown in Table 2-1. Usually, the user can initialise the GA with a seed chromosome, which has parameter values that are known to be in the vicinity of the target solution, as it has been done here. But this is only applicable if the nature of the problem is well understood beforehand. The random values suggested for each parameter in the chromosome are created within a defined range for each parameter to define the GA search space. The parameter ranges for the dipole antenna are defined within the following limits:

- **Dipole Outer Radius**: 1.0 – 12.0 mm
- **Dipole Length**: 20.8 – 166.6 mm
- **Dipole Air Gap**: 0.1 – 10.4 mm

The way that the *Rnd* function generator creates random numbers can be explained using the following equation:

\[
\text{New parameter Value} = (\text{Upper Limit} - \text{Lower Limit}) \times \text{Rnd} + \text{Lower Limit} \quad (2.1)
\]

Where **New parameter Value** is the new random value for any parameter in the chromosome, **Upper limit** is the maximum value of the parameter in the range, **Lower limit** is the minimum value of the parameter in the range and \(0 \leq \text{Rnd} < 1\).
After the random population of suggested chromosomes has been created, a geometry file for each chromosome will be produced as shown in Block 3 of the flow chart in Figure 2.5. These files will produce new dipole geometries, representing the new chromosomes parameter values. The next step (Block 4) is to simulate these geometries in CST-MWS to obtain the $S_{11}$ files, which will be used later in the evaluation operation to find the fittest solution. For the dipole example, 32 $S_{11}$ files will be produced after the simulation step. In Block 5 each $S_{11}$ file will be read as an array of frequency values.
points with its associated $S_{11}$ values and then passed to the evaluation operation in the next block.

### 2.5.2.2 Reproduction Operations

The first reproduction operation is the evaluation function (Block 6). This function will evaluate each suggested chromosome by giving it an objective value. The best-fit chromosome is given the highest numerical objective value in the case of maximisation problem and the lowest numerical objective value in the case of minimisation problem. Since the GA user is the only one who is aware of the way measuring the individual goodness compared to the target, then the evaluation function creation is a user task. For the dipole optimisation, the evaluation function evaluates each individual by calculating the difference between its $S_{11}$ and that of the target $S_{11}$. Consequently, the smaller the difference from the target, the higher the objective value of the associated individual. The objective value of the individual will be one divided by the average error of that individual. Below are the equations for calculating both the average error and the related objective value for an individual:

\[
\text{Point Error} = \frac{(Target S_{11} - Individual S_{11})^2}{(Target S_{11})^2} \tag{2.2}
\]

\[
\text{Average Error} = \frac{\text{Sum Point Errors}}{\text{Number of Frequency Points}} \tag{2.3}
\]

\[
\text{Objective Value} = \frac{1}{\text{Average Error}} \tag{2.4}
\]

Table 2-1 shows the calculated objective value of each chromosome in the population using the evaluation function. The best-fit chromosome has the highest numerical objective value, i.e. the least difference from the target. Most of the time, the target of optimisation is known or it can be an approximation of how the solution could be. So, another task for the GA user is to define the optimisation target or the target limits to help the GA to define a proper search space and converge rapidly toward an optimum. The target $S_{11}$ curve that has been used for evaluation in this example was the $S_{11}$ response of a bandpass filter. The target $S_{11}$ curve resonates at 0.9GHz with 13.4%
bandwidth (BW) at 10dB return loss as shown in Figure 2.6. The figure also shows the seed chromosome $S_{11}$ which resonates at 0.78GHz with 12.6% BW at 10dB return loss.

![Graph showing return loss vs frequency for Seed S11 and Target S11](image)

**Figure 2.6 Seed chromosome (initial dipole design) $S_{11}$ and the target $S_{11}$**

The next reproduction operation for the GA is the fitness scaling operation (Block 7). This operation helps the GA to avoid having the same number of copies of average individuals and best individuals in future generations and consequently avoids premature convergence occurrence. Usually this happens when the population’s average fitness becomes close to the population’s best fitness. Appendix I shows in detail the fitness linear scaling method used in the GA/CST-MWS. The fitness value of each chromosome in the initial population shown in Table 2-2 is based on the objective values showed previously (Table 2-1). It was noted that the average scaled fitness value is equal to the average objective value. It is desired that they are equal because subsequent use of the selection procedure (the next reproduction operation) will ensure that each average population member contributes one expected offspring to the next generation.
Table 2-2 Fitness values of each chromosome in the population

<table>
<thead>
<tr>
<th>Fitness Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.088</td>
</tr>
<tr>
<td>2.157</td>
</tr>
<tr>
<td>2.083</td>
</tr>
<tr>
<td>2.082</td>
</tr>
<tr>
<td>2.085</td>
</tr>
<tr>
<td>2.107</td>
</tr>
<tr>
<td>2.023</td>
</tr>
<tr>
<td>2.022</td>
</tr>
<tr>
<td>4.376</td>
</tr>
<tr>
<td>2.082</td>
</tr>
<tr>
<td>2.206</td>
</tr>
<tr>
<td>2.090</td>
</tr>
<tr>
<td>2.094</td>
</tr>
<tr>
<td>2.044</td>
</tr>
<tr>
<td>2.137</td>
</tr>
<tr>
<td>2.093</td>
</tr>
<tr>
<td>2.139</td>
</tr>
<tr>
<td>2.398</td>
</tr>
<tr>
<td>2.126</td>
</tr>
<tr>
<td>2.085</td>
</tr>
<tr>
<td>2.091</td>
</tr>
<tr>
<td>2.028</td>
</tr>
<tr>
<td>2.109</td>
</tr>
<tr>
<td>2.256</td>
</tr>
<tr>
<td>2.086</td>
</tr>
<tr>
<td>2.099</td>
</tr>
<tr>
<td>2.081</td>
</tr>
<tr>
<td>2.541</td>
</tr>
<tr>
<td>2.084</td>
</tr>
<tr>
<td>2.101</td>
</tr>
<tr>
<td>2.034</td>
</tr>
<tr>
<td>2.026</td>
</tr>
</tbody>
</table>

After fitness scaling assignment, a selection operation (Block 8) is applied to the initial population of suggested chromosomes. The selection operation determines the number of times that a particular suggested chromosome is chosen for reproduction and consequently the number of offsprings that this suggested chromosome will produce. The roulette wheel selection (RWS) described in Appendix I is the method that has been used to apply the selection operation for this GA. The number of chromosomes to be selected was defined in advance by a parameter called Number of Selected Individuals in the front-end in Figure 2.4. Using RWS method 14 chromosomes have been selected to be parents from the initial population of
suggested chromosomes (Table 2-1). The selected chromosomes are given below in Table 2-3.

<table>
<thead>
<tr>
<th>Selected chromosomes</th>
<th>Outer radius</th>
<th>Length</th>
<th>Air gap</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.6</td>
<td>94.9</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>5.2</td>
<td>28.9</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>7.3</td>
<td>159.7</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>2.8</td>
<td>147.2</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>1.8</td>
<td>45.1</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>1.3</td>
<td>134.8</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>1.4</td>
<td>86.1</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>4.0</td>
<td>165.9</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>10.1</td>
<td>97.4</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>5.7</td>
<td>85.0</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>8.7</td>
<td>36.3</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>11.2</td>
<td>116.1</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>2.7</td>
<td>157.1</td>
<td>8.3</td>
<td></td>
</tr>
<tr>
<td>4.9</td>
<td>132.2</td>
<td>8.3</td>
<td></td>
</tr>
</tbody>
</table>

Table 2-3 Selected chromosomes to be parents using RWS

After the selection operation, the selected chromosomes (parents) should be recombined to produce new chromosomes that have parts of both parent’s genetic material. This recombination has been applied using a reproduction operation called crossover (Block 9). There are different crossover techniques that can be used to apply the crossover operation; the crossover implementation heavily depends on the way the chromosomes were represented. Since real-valued coding was used in the GA/CST-MWS, an intermediate recombination function discussed in Appendix I, was applied to crossover the selected chromosomes.

The reproduction selection and crossover operators discussed above alone can generate a huge amount of dissimilar chromosomes. However, depending on the initial population which the GA initiated the run with, there may be insufficient variety of chromosomes to ensure that the GA sees the entire problem search space. The GA therefore may find itself converging on chromosomes that are not quite close to the optimum it seeks. These problems may be conquered by introducing another important reproduction operator.
called the mutation operation (Block 10). The mutation operation used in the GA/CST-MW is called the real-valued mutation operator of the breeder GA and is described in Appendix I. The selected chromosomes listed in Table 2-3 produced the offspring (new children) in Table 2-4 after applying to the crossover and mutation operations respectively. The mutation rate was a $\frac{1}{2}$ because only two parameters have been selected for optimisation. It can be noted that the mutation operation changes the parameter values significantly. The most important factors controlling the amount of change in the mutated chromosomes are the mutation rate and the parameter ranges.

<table>
<thead>
<tr>
<th>Mutated chromosomes</th>
<th>Objective Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Outer radius</td>
<td>Length</td>
</tr>
<tr>
<td>6.4</td>
<td>65.1</td>
</tr>
<tr>
<td>4.1</td>
<td>125.1</td>
</tr>
<tr>
<td>5.2</td>
<td>28.9</td>
</tr>
<tr>
<td>8.1</td>
<td>57.3</td>
</tr>
<tr>
<td>7.0</td>
<td>66.3</td>
</tr>
<tr>
<td>5.4</td>
<td>28.9</td>
</tr>
<tr>
<td>6.3</td>
<td>104.1</td>
</tr>
<tr>
<td>5.9</td>
<td>120.4</td>
</tr>
<tr>
<td>5.1</td>
<td>64.3</td>
</tr>
<tr>
<td>6.0</td>
<td>34.1</td>
</tr>
<tr>
<td>7.1</td>
<td>58.5</td>
</tr>
<tr>
<td>9.1</td>
<td>96.8</td>
</tr>
<tr>
<td>4.4</td>
<td>135.6</td>
</tr>
<tr>
<td>12.0</td>
<td>159.6</td>
</tr>
</tbody>
</table>

Table 2-4 New chromosomes (offspring) produced after crossover and mutation of the selected chromosomes

In the flowchart of Figure 2.5 the produced new chromosomes will go through blocks: 11, 12, 13 and 14, which are similar to blocks: 3, 4, 5 and 6 respectively. Block 14 contains the evaluation function that was used in block 6, but this time it evaluates the objective values for the new chromosomes after mutation. The objective value of each new mutated chromosome is shown in Table 2-4.

After the new set of offspring has been created, the next step is to reinsert the new chromosomes into the initial population in Table 2-1 to create
a new population (new generation) of chromosomes. Therefore, a reinsertion operation (Block 15) has been applied in order to fix the size of the population during the GA run. A popular reinsertion technique is ‘uniform reinsertion’. This technique was employed here and is discussed in Appendix I. 14 parents have been drawn from a population of 32 chromosomes to produce 14 new chromosomes. The reinsertion operation will insert the new 14 chromosomes into the initial population randomly. Therefore, the reinsertion process will keep the number of chromosomes in the population constant, i.e. 32 for each generation created by the GA. Table 2-5 shows the new population after reinserting the new chromosomes (shown in bold italic) into the initial population along with the associated objective value for each chromosome.

Finally, the new population of chromosomes (first generation created by the GA) will loop back to block 7, as shown in Figure 2.5, for a fitness assignment using the fitness scaling function discussed earlier. After that, the new population will go through the reproduction operations: selection, crossover, mutation and evaluation to produce the second generation of chromosomes. The GA process will continue running for a specified number of generations as defined (in the front-end in Figure 2.4) by the user through the Maximum Number of Generations parameter. Once the GA converges towards an optimum, it can be stopped depending on the user’s judgement of whether or not the obtained solution achieves its target.
## 2.5.3 Dipole Antenna Optimisation Results and Discussion

After the GA processed a sufficient number of generations a convergence was noted towards an optimum. The convergence was observed by monitoring the behaviour of the objective values versus created generations and the deviation between the target and achieved $S_{11}$. The higher the objective value, the less the difference from the target. Figure 2.7 shows the convergence history during the GA run after 110 generations ($110 \times 14 = 1540$...
simulations, time = 51 hours). The figure shows the average, best and worst objective values of each generation in addition to the elitist objective values, which have been obtained by selecting only the highest objective values in an ascending order. From Figure 2.7 one observes that the objective value increases rapidly, and then fluctuates in a noisy way due to the high population diversity. Finally the objective values settle down as the generations evolve in an upward trend trying to maximise toward an optimum. GAs always suffer from population diversity due to premature convergence caused by some extraordinary individuals at the beginning of the GA run. This is shown by the best objective values curve for the first few generations in Figure 2.7.

![Figure 2.7 GA run convergence history toward a solution showing average, best, worst and elitist objective value versus generations.](image)

In Table 2-1, it can be noted that the best chromosome is the one that has the highest objective value, which in this case is 26.584. Unfortunately, this chromosome has disappeared in the new produced population shown in Table 2-5 as a result of using the random reinsertion operation. It was found that a new chromosome of low quality with an objective value of 1.14 has
replaced the previous chromosome (objective value = 26.584). The best objective value in the new population in Table 2-5 was 11.318 and it belongs to a new chromosome never seen before. It was also noted that the average objective value of the initial population has decreased from 2.193 to 1.488 for the new population. This significant change of the best and average objective values, explains the diversity between the populations during the GA run. Particularly, for the first 20 generations shown in Figure 2.7. It can be concluded from the results that it is very likely for good chromosomes to be replaced without producing better ones. Thus, good information is lost and diversity increases. But this is not always the trend since the GA always tries to maximize the objective value towards a solution. However, some generations fluctuate because of diversity.

Figure 2.8 shows how the values of the parameters selected for optimisation (dipole outer radius and dipole length) and the best objective value of each generation vary against generation number. The graph reveals that the population diversity is very clear at the beginning of the run and it slowly vanishes at the end. The results show that the objective values and the dipole parameter values do not vary significantly for the last few generations of the run. This indicates that user should terminate the GA because any progress achieved after that will be insignificant. Figure 2.8 shows that the best chromosome is found at generation 107, where the highest objective value has been achieved to give in a solution that is very close to the target. The optimised parameter values of this chromosome were 3.8mm for the dipole outer radius and 142.4mm for the dipole length and they were within the defined ranges for these parameters. The values of these parameters before optimisation were 3.0mm for the outer radius and 166.6mm for the length. It can be noted that the dipole parameter values obtained are consistent with the dipole design in theory [21]. There is an inversely proportional relation between the length and the resonant frequency of the dipole and a proportional relation between the dipole’s outer radius and its bandwidth. Therefore, it can be concluded from Figure 2.8 (a) that the dipole outer radius was increased by 0.8mm to broaden the dipole bandwidth, while Figure 2.8 (b) showed that the
dipole length was decreased by 24.2mm to shift the resonant frequency up toward the target.

There is reasonably good agreement between the target $S_{11}$ and the achieved solution $S_{11}$ as depicted in Figure 2.9. It is clear how the GA forced the seed chromosome $S_{11}$ to converge towards that of the target, passing through least fit generations like 7, to fitter ones like 58 and converging towards the best fit at 107.

![Figure 2.8](image-url)
The percentage bandwidth of the achieved solution at -10dB return loss was:

\[ BW = \frac{f_2 - f_1}{f_c} \times 100 = 14.36\% \]

where:
\[ f_1 = 0.84 \text{ GHz}, \]
\[ f_2 = 0.97 \text{ GHz}, \]
\[ f_c = \frac{f_2 - f_1}{2} + f_1 = 0.905 \text{ GHz} \]

It should be noticed that the percentage BW for optimised solution is 2% more than that of the seed at resonant frequency ~0.9GHz and it reaches the 16dB return loss level. While the target S11 resonates at the same frequency exhibits a return loss of 30dB. This mismatch between the curves can be explained by two reasons. The first reason is that the target S11 curve has been obtained using a bandpass filter S11 response, which could be an impossible target for the GA to be achieved exactly. The second reason is that the GAs are not always expected to give the optimum solution, most of the time they give solutions that are close enough to the optimum.
Generally, the results expected from GAs should be clear. The optimisation theory includes the quantitative study of optima and methods for detecting them. Therefore, optimisation seeks to improve performance toward some optimum point or points. It should be noticed that this definition has two parts: (a) seeking improvement to approach some (b) optimum points. Goldberg [5] found that there is a clear distinction between the process of improvement and the destination or optimum itself. Thus, in judging optimisation results the user should not only focus on convergence (whether an optimum achieved or not) but must also look at the performance as a whole.

2.5.4 GA Disadvantage and Termination

The main disadvantage of GAs is that they sometimes converge to a local solution before discovering a globally optimum one due to premature convergence. When the objective value of a population remains static for a number of generations, it is a standard practice to periodically terminate the GA and compare the best individuals of the population against the target [3,22]. If no acceptable solutions are found, then the GA may be restarted with the same seed chromosome or a new run initiated using a new seed chromosome.

For example, before the successful run shown in Figure 2.7 achieved a desired solution, the GA was ran several times with the same seed chromosome but failed to optimise the dipole. The convergence histories for one of these unsuccessful runs are shown in Figure 2.10, where it is noted that the GA has not converged toward a solution. The objective value curves have an upward trend for the first 30 generations and they then remain static for the rest of the run. The results in Figure 2.10 show the best individual is found at generation 84 with an objective value of 2.473 and average objective value of 2.44. It is clear that these values are very small compared to the results achieved in the successful run of Figure 2.7. This best individual kept appearing for the rest of the run at generations 87, 96, 102,104 and 108 with
almost the same average objective value. In such a case, it can be judged from the GA’s behaviour that the best thing to do is to terminate the GA run and initiate a fresh one with either the same seed chromosome or a new seed chromosome. Another option is to select other parameters for optimisation and change the parameter ranges.

Figure 2.10 Unsuccessful GA run convergence history showing average, best, worst and elitist objective value versus generations.
2.6 Conclusion

A simple real-valued GA/CST-MWS optimisation technique was introduced to achieve increasingly complex antenna designs in the future instead of using traditional empirical method. It is known that the empirical method depends on innovative ideas and intuitive trials, which can lengthen the design cycle. The performance of the technique was tested through the optimisation of a dipole antenna. The GA/CST-MWS succeeded in optimising the dipole to resonate at the required target frequency of 900MHz with a 10dB bandwidth improvement by 2%.

Through monitoring the GA runs it was found that the main GA advantage is its ability to assess quickly whether the parameters selected for optimisation are appropriate for optimisation. Another advantage in using this optimisation technique is that it guarantees the simulation machines will be employed 24 hours a day, since the GA runs independently from the user.

On the other hand, it was noted that the optimisation technique sometimes remains static for a number of generations before a superior solution is achieved. In this case, the GA may be terminated and a fresh search initiated using a new seed chromosome. This termination action will increase the time needed to find an optimum solution since several runs will be needed to achieve the target.

Finally, it should be clear that the GA is not always expected to provide an identical solution that perfectly matches the target. Therefore, it can be concluded that the GA is a good tool for refinement purposes or for studying the behaviour of the parameters that describe the problem and their effect on the problem performance by monitoring the GA runs. In the next Chapter the proposed technique will be used to optimise a planar inverted-F antenna (PIFA) type antenna, which is more complex compared to a dipole antenna.
References


Chapter 3

3 Real-Valued Coding GA Optimisation Technique Applied to a PIFA Handset Antenna

3.1 Introduction

Most of the GAs presented in the literature have focused on optimising different types of antennas like wire, microstrip, patch and fractal antennas [1-4]. The representation method used to describe antenna parameters for these GAs was the binary format. Unfortunately, little research has been reported on optimising planar inverted-F antenna (PIFA) using GAs and the few proposed ones have also focused on using the binary format to represent the antenna parameters [5,6]. In this Chapter the real coding GA/CST-MWS optimisation technique, presented previously in Chapter 2 for a simple dipole, has been applied to a PIFA handset antenna.

The increased geometrical complexity of the PIFA structure brings an associated increase in design complexity. The size and aspect ratio of the top planar element, the height of the planar element above the ground plane, the width and position of the short-circuit plate and the feeding point location all have considerable influence on the electrical performance of the antenna [7]. With so many parameters that can be optimised, the potential time saved by using the GA/CST-MWS could be considerable compared to the traditional intuitive method. A single-band PIFA after cutting slots within its top planar plate has been fed into the GA/CST-MWS optimisation technique. The optimisation target was to achieve a dual-band PIFA design suitable for personal communication handsets at GSM-900MHz and DCS-1800MHz. A prototype of the obtained dual-band PIFA design was built and measured in the anechoic chamber to compare with the optimum GA/CST-MWS result.
3.2 The Planar Inverted-F Antenna (PIFA)

The PIFA consists of a rectangular planar element and a shorting plate. The width of the shorting plate is narrower than that of the shorted side of the rectangular planar element. The basis of the PIFA structure originated from two different types of antennas. The first type is the inverted-L antenna (ILA) and the inverted-F antenna (IFA) whose wire element was replaced by a planar plate [8,9]. The second type is the short-circuit microstrip antenna (short-circuit MSA) whose shorted element width set narrower than that of the shorted side of the antenna patch [7,10]. The next section will explain how the ILA/IFA and the short-circuited MSA can be converted into a PIFA.

3.2.1 PIFA Basis and Structure

The ILA is similar to a short vertical monopole element with a horizontal wire element connected to the end of the monopole [8] as depicted in Figure 3.1. The ILA is classified as a small antenna because of its low profile structure, which is gained by the restricted height of the vertical wire above the ground plane surface.

The ILA’s low profile structure was adopted in several other antenna designs. One of these designs is the IFA, which has an additional inverted-L wire attached to the ILA as shown in Figure 3.2. This addition to the ILA increases the radiation impedance of the IFA. Also the input impedance of the IFA can be matched to the load impedance, without using any additional circuit between the antenna and the load [8]. One major disadvantage of the modified IFA is its narrow bandwidth. To overcome this problem a planar plate has replaced the horizontal wire element. The resulting configuration, after this modification, shown in Figure 3.3 is called the planar inverted-F antenna (PIFA). The PIFA is a quarter wavelength resonant structure achieved by short-circuiting its radiating element (top planar element) to the ground plane using a short-circuit plate.
Figure 3.1 Configuration of an inverted-L antenna (ILA)

Figure 3.2 Configuration of an inverted-F antenna (IFA)

Figure 3.3 Configuration of a planar inverted-F antenna (PIFA) modified from IFA
However, the PIFA can also be derived from a rectangular short-circuit MSA. Placing a short circuit plate between the radiating element and the ground plane at the zero potential of the MSA, without affecting the field structure, can halve the length of the rectangular element of the MSA. By reducing the width of the shorting circuit plate to be narrower than that of the planar element, the current distribution is perturbed and hence the resonant frequency becomes lower than that of a conventional short-circuit MSA having a planar element of the same size [7]. After reducing the width of the shorting plate, the final structure is also called a PIFA.

3.2.2 PIFA Analysis

Hirasawa et al. [7] presented an analysis for the PIFA, which could be used as guidance for designing a single-band PIFA. The general structure of the PIFA is shown in Figure 3.3. It is formed by a planar rectangular element of length L and width W placed parallel to a ground plane at height H and a short-circuit plate of width S. The antenna is fed using a 50Ω transmission line. The resonant frequency of the PIFA can be calculated approximately using [7]:

\[ \lambda_r = 4 (L + W) \] (3.1)

Where \( \lambda_r \) is the guided wavelength at the resonant frequency.

However, equation (3.1) does not include the dimension variation of the shorting plate width S and the planar element height H. Therefore, Hirasawa et al. [7] reported a number of equations for different ratios of the shorting plate width to planar element dimensions:

when \( S/W = 1 \), the wavelength of the resonant frequency can be expressed by

\[ W + H = \lambda_r / 4 \] (3.2)

and when \( S = 0 \),

\[ W + L + H = \lambda_r / 4 \] (3.3)

finally, when \( 0 < (S/W) < 1 \), the resonant frequency can be expressed by

\[ W + L + H - S = \lambda_r / 4 \] (3.4)
Equation (3.4) shows that the size ratio of the planar element $L/W$, shorting plate width $S$ and antenna height $H$ are the major parameters, which control the characteristics of resonant frequency and bandwidth of the PIFA.

### 3.3 Dual-Band PIFA Handset Antenna Optimisation Example

The GA/CST-MWS optimisation technique has also been applied to a more complex antenna design with more parameters than the simple dipole. This complex candidate is the PIFA, which is one of the most common antennas used in mobile phones in the past few years. The fact that it must fit within a tiny volume allocated in mobile handsets, gives rise to geometrical complexity. A single-band PIFA was modified initially by inserting slots into its radiating element. Then it was optimised to yield a dual-band PIFA antenna suitable for personal communication handsets operating within the GSM-900MHz and DCS-1800MHz bands.

#### 3.3.1 Dual-band PIFA Design Procedure

The starting point for design of the dual-band PIFA was the single-band PIFA shown in Figure 3.4. The antenna consists of a top plate element, 38mm wide x 13mm long, which is supported 6.1mm above the surface of a ground plane (single sided PCB of FR4, $\varepsilon_r = 4$), 38mm wide x 100mm long x 1.6mm thick. The antenna element is matched to a $50 \Omega$ impedance coaxial cable and shorted to the ground plane by a vertical plate of 2.9mm wide.

The resonant frequency $f_r$, calculated using equation (3.1) was 1.47GHz. The equation neglects the antenna height and the shorting plate width. The calculated resonant frequency value was very close to the resonant frequency 1.455GHz obtained by simulating the antenna in CST-MWS, as shown in Figure 3.5. The antenna’s percentage BW achieved at 10dB return loss, which is the common criterion in to a $50 \Omega$ load, was 11%. But, because of the narrow bandwidth of low-profile antennas and the bandwidth dependency on the impedance matching, the above criterion reduced to 5dB
return loss into a 50Ω load resulting in a percentage BW of 22%. Equation (3.4) is used to calculate the resonant frequency, as $S/W = 0.08$ using the planar element height $H$ and the shorting plate width $S$; the calculated resonant frequency will be 1.3GHz. The reason for this difference from the simulated result is due to the size of the ground plane used and the position of the feed.

The next step in the design procedure is to integrate another band into the single-band PIFA. The single-band PIFA with its current configuration could not be optimised to yield a second resonance without a modification of the antenna configuration. The challenge was to introduce another band without increasing the volume allocated for the antenna in the handset. The requirements of the design were to produce a dual-band antenna with a width less than 40mm, length less than 15mm and a height less than 8mm.

The literature presented several ideas on integrating dual-band or multi-band performance into a single-band PIFA [10-13]. These ideas include multiple resonators, cutting slots in the radiating element and using parasitic resonators. Due to the limitation imposed on the volume allocated to antenna element in the required design, the best way to integrate another band is to cut slots in the antenna’s element top plate. Four different slots have been introduced in the top plate of the single-band PIFA to produce a dual-band PIFA as depicted in Figure 3.6. Figure 3.7 shows that the antenna operates at two different frequencies both of which deviate ~250MHz from the design objective. This difference leads to the last stage in the design procedure. The GA/CST-MWS optimisation technique was applied to the dual-band PIFA (seed geometry) to set the resonant frequencies at GSM-900MHz and DCS-1800MHz with the required bandwidth.
Figure 3.4 Single-band PIFA configuration modelled in CSTMWS

![Single-band PIFA Configuration](image)

**Figure 3.4 Single-band PIFA configuration modelled in CSTMWS**

Figure 3.5 Input return loss response of a single-band PIFA

![Input Return Loss Response](image)

**Figure 3.5 Input return loss response of a single-band PIFA**

Figure 3.6 A dual-band PIFA after cutting slots in the radiating element of the single-band PIFA

![Dual-band PIFA Configuration](image)

**Figure 3.6 A dual-band PIFA after cutting slots in the radiating element of the single-band PIFA**
3.3.2 Dual-band PIFA Input Parameters for Optimisation

The geometry of the seed dual-band PIFA is shown in Figure 3.6. Several parameters of the seed design have a strong influence on the antenna performance and can be used in the optimisation process. These parameters are the height of the radiating element (patch) above the ground plane, the position of the feed on the x-axis and y-axis, the shorting plate width and the dimensions of the four slots (Slot_1, Slot_2, Slot_3 and Slot_4). Each slot is characterised by four dimensions: Slot Width_Start, Slot Width_End, Slot Length_Start and Slot Length_End. The chromosome structure for the proposed dual-band PIFA will be a one-dimensional vector of 20 columns representing the variable parameters that describe the antenna geometry. These variables are listed in the front-end in Figure 3.8. Any variable can be selected for optimisation by ticking the check box adjacent to that parameter. Figure 3.8 also shows that there are some fixed parameters listed at the top part of the front-end such as the ground plane and the top plate dimensions. The handset dimensions will be predetermined along with the volume allocated for the antenna element and so will remain fixed during any run of the GA.
3.3.3 Dual-band PIFA GA/CST-MWS Optimisation

The variable parameters selected for optimising the seed dual-band PIFA are depicted in Figure 3.6 and are included by ticking the check boxes on the front-end (see Figure 3.8). The target for optimisation is to shift the resonance frequencies; 1.148GHz and 2.070GHz to resonate at GSM-900MHz and DCS-1800MHz respectively with the required bandwidth at 5dB return loss. The population size chosen is 40 individuals, where each individual contains 20 parameters. Consequently, the population is a two dimensional array of (40 Rows x 20 Columns). The minimum probability of mutation is 0.05 if 20 variables are selected for optimisation and the maximum probability is 1 if 1 variable is selected for optimisation. The number of selected individuals is limited to 14, as there are only 14 machines for simulations. Therefore, 14 offspring will be created in each generation of the GA run.

The evaluation function for this problem will be completely different from that used for the dipole example, because the current candidate is more complex and the input return loss response has two modes to match. In the literature, evaluation functions have been proposed for this kind of electromagnetic problem [5,14-17]. But, because the objective here is to minimise the antenna return loss values at two different bands, the best formula to use is (3.5). The formula averages the $S_{11}$ values that exceed $-5$dB within the desired frequency range for the different bands:

\[
\text{Objective Value} = \frac{(\text{Band}_{1} \text{Average} + \text{Band}_{2} \text{Average})}{2}
\]  

(3.5)

\[
\text{Band Average} = \left\{ \begin{array}{ll}
N_i \sum_{i=1}^{N_i} S_{11}(i) & \text{if } S_{11} \geq -5dB \\
-5 & \text{if } S_{11} < -5dB
\end{array} \right.
\]

Where $N_i$ is the number of frequency points in the desired frequency bands. After defining the evaluation function, the GA/CST-MWS was used to
optimise the selected variable parameters listed in Table 3-1 within a defined range for each parameter.

![Figure 3.8 Dual-band PIFA front-end input parameters as variables for optimisation](image-url)

<table>
<thead>
<tr>
<th>Parameter #</th>
<th>Selected parameters for optimisation</th>
<th>Value before optimisation (mm)</th>
<th>Value after optimisation (mm), Generation 74, Run D</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Slot 1 Width Start</td>
<td>8.00</td>
<td>10.00</td>
</tr>
<tr>
<td>2</td>
<td>Slot 1 Length Start</td>
<td>4.00</td>
<td>1.50</td>
</tr>
<tr>
<td>3</td>
<td>Slot 2 Width Start</td>
<td>32.5</td>
<td>26.5</td>
</tr>
<tr>
<td>4</td>
<td>Slot 3 Width Start</td>
<td>14.00</td>
<td>3.50</td>
</tr>
<tr>
<td>5</td>
<td>Slot 3 Width End</td>
<td>18.00</td>
<td>5.50</td>
</tr>
<tr>
<td>6</td>
<td>Slot 4 Width End</td>
<td>5.50</td>
<td>5.00</td>
</tr>
<tr>
<td>7</td>
<td>Slot 4 Length Start</td>
<td>10.00</td>
<td>9.20</td>
</tr>
<tr>
<td>8</td>
<td>Feed Position x-axis</td>
<td>3.50</td>
<td>2.40</td>
</tr>
<tr>
<td>9</td>
<td>Feed Position y-axis</td>
<td>6.00</td>
<td>4.90</td>
</tr>
<tr>
<td>10</td>
<td>Shorting plate Width</td>
<td>5.00</td>
<td>2.9</td>
</tr>
</tbody>
</table>

Table 3-1 Selected variable parameters for the seed dual-band PIFA optimisation
3.3.4 Dual-band PIFA Optimisation Results and Discussion

Four GA runs have been applied to the dual-band PIFA seed geometry shown in Figure 3.6 using different combinations of the variable parameters shown in Table 3-1. All the runs have the same population size and number of selected individuals. The convergence history of the four runs is shown in Figure 3.9. It is clear that runs A and B are unsuccessful, while run C showed some improvement. These runs were nonetheless very helpful as they were used to feed run D with individuals that were closer to optimum than the seed as they have a higher objective value.

![Convergence history for the best objective values of four different runs for the GA/CST-MWS optimisation technique.](image)

Run A used three parameters for optimisation, numbered: 8, 9 and 10 as indicated in Table 3-1. The best objective value 0.313 was achieved at generation 5. For subsequent generations the objective value remained constant due to premature convergence. Run B, included parameters: 4, 5, 8, 9 and 10 in the optimisation and the run was seeded with the best individual found in run A. This increment in the parameter number will expand the search space for the GA and should improve the results. The highest objective
value achieved for this run, was 0.317 at generation 8. It can be noted that the improvement obtained in run B is insignificant. Therefore, run C employed more parameters in the optimisation. Parameters numbered: 1, 3, 5, 7, 8, 9 and 10 were used and the run was seeded with the best-fit individual found in the previous run. Improvement was achieved in run C, where an objective value of 0.988 was achieved at generation 24. Unfortunately, the run also suffers from premature convergence and remained static for the rest of the generations.

In the last run (run D), all of the parameters listed in Table 3-1 were selected for optimisation. The run was seeded with the best individual obtained in run C. The best individual achieved in run D occurred at generation 74 with an objective value of 1.91. After this generation, the run remained static for a large number of generations indicating that a termination action should be taken. The best individual obtained at generation 74 should be analysed in order to decide whether it is fit enough to meet the target criteria or if another run is required.

Figure 3.10 compares the simulated input return losses of the achieved solution at generation 74 of run D and the seed geometry. It can be seen that the optimised results match the DCS-1800MHz band with 10% BW and almost match the GSM-900MHz band with a 6% BW at 5dB return loss. The results still need improving, especially for the lower band, where a 7.6% BW is required. This improvement can be achieved in one of the two following ways. The first method is to keep running the GA with different settings and varying parameters such as the population size, mutation rate and the number of parameters selected for optimisation. But unfortunately, this method will fail if the antenna’s physical size is not sufficient to achieve better results. The second method is to change the fixed parameters by altering the physical size of the ground plane and the antenna element.

The surface current distributions of the optimised antenna at the two operating frequencies are shown in Figure 3.11 (a), (b). The results indicate how the inserted slots in the antenna element force the currents to follow certain paths in order to achieve the targeted resonance. It can be noted that
there is a big difference in the surface current distributions for the different frequencies. It is also very important to notice the effect of the ground plane on the current distribution. The current distribution at 1.805GHz is more dominant under the antenna element compared with distribution at 0.915GHz. This is highlights the role of the PIFA’s ground plane in achieving the desired resonance. By observing the GA runs it was noted that one could vary the current flow and consequently shift the resonant modes by altering the location of the feeding point and the width of the shorting plate.

![Graph showing return loss vs frequency for two frequencies, 0.915GHz and 1.805GHz.](image)

**Figure 3.10** Best solution $S_{11}$ at generation 74 from run D and the seed geometry $S_{11}$

![Simulated surface current distribution for two frequencies, 0.915GHz and 1.805GHz.](image)

**Figure 3.11** Simulated surface current distribution of the optimised dual-band PIFA
Finally, the results obtained through successive GA runs showed that the principal factors controlling the GA convergence are the number of variable parameters per chromosome, the number of chromosomes per population (population size), the range defined for each parameter and the simulation time of an individual. In addition, there are also some GA operators that control the GA convergence. These are the performance of the random number generator, which generates the first population for the GA, and the mutation operation.

The effect of the random number generator can be analysed by comparing the different seeded GA runs in Figure 3.9. By inspecting the objective values for the first few generations of each run it is noted that the choice of the seed may have significant effects on the results obtained. In summary, the chosen seed determines the first population that will be created for the GA run and consequently, the initial genetic information available to the GA. The richer (high objective value) this information is the more efficiently the algorithm performs.

Other factors affecting the GA include the mutation operation. Different mutation rates have been used for the runs in Figure 3.9. A value of 33% for run A, 20% for run B, 14% for run C and 10% for D. The effect of different mutation probabilities can be observed by comparing the results obtained from each of the four runs. It was noted that the objective values were relatively low for the high mutation rate. Although the mutation operation produces new individuals that may not exist in the preceding generations, thus enlarging the search space, it can also destroy good individuals and lose important genetic information.

3.3.5 Measured Results of Optimised Dual-Band PIFA

A prototype for the optimised dual-band PIFA was constructed as shown in the photograph in Figure 3.12. The prototype was made out of 0.23mm copper sheet cut to the dimensions determined by the GA optimised
parameters listed in Table 3-1. The antenna’s top plate was shorted to the
ground plane by a shorting plate and fed by a 50Ω impedance coaxial cable.
The antenna prototype has been tested by measuring the antenna performance
(Gain, Efficiency and Radiation patterns) inside an anechoic chamber. The
antenna input return loss was measured using a HP-8753D (30kHz - 6GHz)
network analyser. Figure 3.13 shows the measured and simulated input return
loss results. There is good agreement between the measured and simulated
results. A slight shift is observed in the locations of the two operational bands
and the $S_{11}$ match of the prototype antenna due to fabrication errors, which
will be discussed in the next section, and the difference in the feeding cable
length between simulation and measurements. This difference in the feeding
cable length leads to a variation in the matching impedance value at the end of
the cable during the calibration process in the measurements. The reason for
using a short cable during simulation is to reduce the simulation time.
However, a longer cable length during the measurement is required to satisfy
physical constraints with mechanical positioning system. The GSM-900MHz
band resonates at 0.935GHz and deviates by 15MHz from the target with 6.2%
bandwidth (1.4% less than the target) at 5dB return loss. The DCS-1800MHz
band resonates at 1.855GHz and deviates by 55MHz from the target with
10.50% bandwidth at 5dB return loss, which matches the target bandwidth
precisely. Figure 3.14 provides the measured impedance characteristics for the
optimised antenna.

The measurement results concerning the antenna gain and efficiency
are depicted in Figure 3.15 for the GSM-900MHz and the DCS-1800MHz
bands. For the GSM-900MHz band, a maximum gain of 4.32dBi at 940MHz
with $-1.48$dBi average gain was obtained. For the DCS-1800MHz band, a
maximum gain of 3.88dBi at 1752MHz with $-0.92$dBi average gain was
obtained. The obtained values are compatible with usual specifications for this
type of mobile communications antenna. The antenna efficiency was also
measured, in order to further characterise the performance of the antenna. It
was noted that the efficiency results are consistent with the measured gain,
especially in the GSM-900MHz band. The radiated efficiency of the antenna
remains over 40% for most of the frequencies within the operating bands.
The measured cuts of the radiation patterns of the prototype antenna are shown in Figure 3.16 and Figure 3.17 for the GSM-900MHz and the DCS-1800MHz bands respectively. It can be seen that at the GSM-900MHz band frequencies the shape of the radiation patterns are almost omnidirectional for $E_\theta$ and resemble that of a half-wave dipole for $E_\phi$. For DCS-1800MHz band frequencies the patterns became more complex and somewhat more directive.

Figure 3.12 Photograph of the prototype dual-band PIFA GA optimised antenna

Figure 3.13 Simulated and measured input return loss response for the GA optimised dual-band PIFA
Figure 3.14 Measured impedance characteristics of the GA optimised dual-band PIFA

Figure 3.15 Measured gain and efficiency for the GA optimised dual-band PIFA prototype at: (a) GSM-900MHz (b) DCS-1800MHz
Figure 3.16 Measured radiation patterns (--- $E_{\phi}$, ----- $E_{\theta}$) for the GA optimised dual-band PIFA for GSM–900MHz at 930MHz and 950MHz.
Figure 3.17 Measured radiation patterns ($E_\Phi$, $E_\theta$) for the GA optimised dual-band PIFA for DCS–1800MHz at 1795MHz and 1922.5MHz
3.3.6 Effect of PIFA Parameters on Resonant Frequency and BW

As seen earlier (Figure 3.13) there was a difference between the simulated and measured results for the input return loss of the optimised dual-band PIFA, especially the upper band. This difference is mainly due to manufacturing errors during the antenna fabrication as noted earlier. The error can be quantified by studying some of the sensitive parameters of the optimised dual-band PIFA. Figure 3.18 shows results obtained through a simulation study of three parameters that have a significant effect on the antenna resonant frequency and the impedance bandwidth at the DCS-1800MHz band. These parameters are the height of the patch from the top surface of the ground plane and the position of the feeding cable along the x-axis and the y-axis.

Figure 3.18 (a) shows that increasing the patch height in steps of 0.1mm has the effect of reducing the resonant frequency and increasing the bandwidth. This effect is explained by the proportional relationship between the bandwidth and the antenna volume and the inversely proportional relationship between the antenna volume and the radiation quality factor Q (see Chapter 5, section 5.2.3 for details).

Moreover, the position of the feeding cable has little impact on the radiation characteristics of the PIFA since it affects the impedance matching. For instance, varying the position the feed along the x-axis by 0.2mm steps away from the upper left corner of the patch increases the resonant frequency and bandwidth as depicted in Figure 3.18 (b). While varying the feed position along the y-axis by 0.2mm steps away from the upper left corner of the patch also increases the resonant frequency and bandwidth as shown in Figure 3.18 (c).

The input resistance and reactance, which together constitute the input impedance of the antenna, as a function of the feed position along the x-axis are shown in Figure 3.19 (a), (b). As expected, both resistance and reactance
tend to vary as the position of the feeding point altered. This feature of the antenna provides a convenient way for impedance matching.

Figure 3.18 Effect of varying the dual-band PIFA parameters on resonant frequency and impedance bandwidth (a) Patch height (b) Feed position dx (c) Feed position dy
So, it can be concluded that a small variation in the antenna’s parameters during manufacturing, can lead to a significant deviation in the results. For example, increasing the antenna patch height from 5.8mm to 6.6mm has shifted the resonant frequency by 33MHz and increased the bandwidth by 1%. Changing the feed position along the x-axis from 2.2mm to 3.4mm shifted the resonant frequency by 30MHz. While changing it along the y-axis from 4.3mm to 5.5mm shifted the resonant frequency by 46MHz and improved the bandwidth by 0.5%.
3.4 Conclusion

The GA/CST-MWS optimisation technique was used to optimise the performance of a PIFA mobile handset antenna. A single-band PIFA was modified to resonate at two frequency bands and then optimised, using the GA, to achieve dual-band performance suitable for personal communications at GSM-900MHz/DCS-1800MHz. After four different runs for the GA/CST-MWS technique an optimum was obtained and the antenna fabricated. The lower frequency band for the prototype PIFA resonates at 0.935GHz and deviates by 15MHz from the GSM-900MHz band. The design exhibits a bandwidth of 6.2% that is 1.4% less than the required GSM-900MHz standard bandwidth at 5dB return loss. The upper frequency band resonates at 1.855GHz and deviates by 55MHz from the required resonant, but covers perfectly the required DCS-1800MHz band with 10.50% bandwidth at 5dB return loss.

The deviation in the resonant frequencies was due to manufacturing tolerances and accuracy of the simulation package. Varying the antenna’s parameter values, such as the height of the planar radiating element and the position of the feeding cable during fabrication by just 1mm led to an average of 36MHz shift in resonant frequency and slight variation in the bandwidth (~0.5%). Therefore, the PIFA is particularly sensitive to parameter variations and so care must be taken during the manufacturing process. Errors associated with the simulation software include the size of mesh step, which decides the simulation period, the accuracy of simulating curved or circular structures like coaxial cables to match 50Ω impedance and built in errors in the simulation code.

The optimisation results showed that the speed of GA/CST-MWS convergence depends on the PIFA parameters selected for optimisation, their numbers, the range defined for each parameter, initial population size and suggested individuals quality. During the GA runs it was noted that the most important parameters for optimising the PIFA were the dimension of the slots in the planar radiating element, the feeding point position and shorting plate.
width. The time required for the GA/CST-MWS to find the optimum dual-band PIFA handset design was (80 generations × 14 simulations per generation × 18 minutes per simulation = 14 days). It should be highlighted that the speed of the technique convergence still depends, on the skill and the experience of the user. The GA user should show good judgment in monitoring the GA runs and selecting the appropriate parameters for optimisation. However, the time required for the GA to optimise the PIFA handset antenna with ten parameters selected for optimisation is still less than that required for the traditional intuitive trials which depend on slow manual parametric studies. At least the GA relieves the user from tedious and intuitive trials that require constant work and concentration by automating some of the design process.

It can be concluded that the GA is a promising optimisation technique that is best suited for complex antennas with many variable design parameters. In spite of the author's development work, the GA/CST-MWS method is still simple and considerable scope for improvement. Ideas for improving the GA performance and convergence speed will be presented in Chapter 7.
References


Chapter 4

4 Triple-Band PIFA Handset Design

4.1 Introduction

There is growing demand for small multi-band handset antennas, which fit easily in the small volume available within modern mobile handsets. The ability of PIFAs to integrate a multi-band performance, further enhances the merit of these antennas for future applications. Owing to PIFAs compact size, variety of dual-band or multi-band PIFAs suitable for applications in mobile phones, have recently been proposed [1-7]. Unfortunately, most of these designs occupy a larger volume than that allocated by mobile handsets manufacturer.

Therefore, based on the dual-band PIFA design proposed previously in Chapter 3 and presented by the author in [8,9], the antenna has been modified to yield a multi-band design after subtle modifications. In this chapter, a small dimension, relatively low profile triple-band PIFA compared with conventional PIFA designs has been introduced. The antenna is suitable for mobile handsets, which operate at the DCS-1800MHz, PCS-1900MHz and UMTS cellular frequency bands. Details of the proposed antenna are described; experimental results of a constructed prototype are presented and effects of shorting pin (radius and position) on resonant frequency and bandwidth have been studied.

The specific absorption rate (SAR) distributions in a human phantom head exposed to electromagnetic waves radiated from the proposed antenna prototype have been predicted, using the CST-MWS™ simulation package, and measured using a SPEAG Dosimetric Assessment System (DASY4™).
Finally, the proposed triple-band PIFA was modified to cover the GSM-900MHz band. This was achieved by incorporating a planar meander-line monopole added to the triple-band PIFA structure to achieve a quad-band design. Simulation and experimental results for the quad-band antenna are presented and discussed.

4.2 Triple-Band PIFA Design

As stated in section 4.1, the proposed triple-band PIFA is based on the dual-band PIFA design shown in Chapter 3. The dual-band PIFA was produced using the GA/CST-MWS optimisation technique. The multi-band PIFA antennas presented in the remainder of this dissertation were optimised using parametric studies with prior knowledge gained on the general behaviour of PIFA’s parameters concluded from the GA/CST-MWS runs in Chapter 3. The proposed GA is a high cost technique to run and still requires improvements and high-level code to cope with complex small multi-band antenna designs that have large number of parameters and optimisation targets, which are difficult to achieve. In the following sections a triple-band PIFA design will be presented based on parametric studies and empirical modifications applied to the dual-band PIFA as mentioned previously. The performance of the triple-band PIFA was studied and tested inside a plastic case and near a phantom head for further characterisation.

4.2.1 Configuration of the Triple-Band PIFA

The physical difference between the proposed triple-band PIFA and the dual-band PIFA antenna is the added s-shape strip, shown in the dashed area of Figure 4.1(a), and the shorting pin which replaced the shorting plate of the dual-band PIFA for ease of fabrication. Other modifications included changes to dimensions of the antenna element and the ground plane. The added s-shape to the antenna structure has been obtained after several trials and simulations. The idea was to disturb the surface currents flowing on the antenna element by changing its configuration. Therefore, the antenna’s
wavelength was increased by extending the antenna element from 13mm to 16mm long and slotted to the shown configuration (S-shape) in Figure 4.1(a) to achieve the desired bands. The antenna configuration consists of a rectangular patch located 5.9mm above the top of the ground plane. The patch is printed on a 0.765mm thick dielectric substrate ($\varepsilon_r = 2.2$) for support as shown in Figure 4.1.

![Figure 4.1 (a) Geometry of the triple-band PIFA (b) Photograph of a prototype](image-url)
The antenna is matched to a 50Ω impedance coaxial cable and shorted to the ground plane by a pin placed at the corner of the patch after some optimisation by changing the pin position along the x-axis and the y-axis and having at the same time the feed position and the antenna height from the ground fixed. The antenna element measures 36mm wide by 16 mm long and is supported 5.9mm above a single sided PCB (FR4, $\varepsilon_r = 4$), 78mm long x 36mm width x 1.6mm thick. The desired frequency bands have been obtained by cutting slots into the rectangular patch to force currents to follow certain paths, as depicted in Figure 4.1. The antenna’s resonant frequency and bandwidth are very sensitive to the size and location of these slots as indicated through the GA/CST-MWS results in Chapter 3.

4.2.2 Antenna Performance

The simulated and measured return losses for the triple-band PIFA are shown in Figure 4.2 and the impedance characteristics are provided in Figure 4.3. The minor differences between the simulation and measurement results may be attributed to the limited accuracy of fabrication and simulation as discussed in Chapter 3 (section 3.3.5). The antenna’s measured bandwidth is as large as 552MHz at 5dB return loss and covers the DCS1800, PCS1900 and 8.1% of the required standard bandwidth for the UMTS (12.2%). The band starts at $f_1 = 1.506$GHz and stops at $f_2 = 2.060$GHz having a percentage bandwidth of 31% at 5dB. The bandwidth should clearly be enhanced to cover the whole of the UMTS band. Several methods can be applied to expand (or tune) the bandwidth of the antenna under investigation to cover the UMTS band. The first method is to optimise the size of the ground plane with respect to the bandwidth, since the impedance bandwidth of the antenna depends on the ground plane dimensions [10]. This method will not be applied here since the current trend favours small mobile handsets, which attractive to users. Another method is to shift the whole band up by at least 100MHz. This can be done by tuning the resonant frequency of the antenna by repositioning the shorting pin, as will be discussed in section 4.2.4. Other parameters such as
the lengths and the widths of the different slots in the patch can also alter the resonant frequency and bandwidth.

Figure 4.2 Simulated and measured input return loss of the triple-band PIFA

Figure 4.3 Measured input impedance characteristics of the triple-band PIFA
Figure 4.4 shows the surface current distributions at different frequency points. At 1.665GHz it is clear that part of the outer L-shaped strip is the major radiating element at this frequency. For the DCS band (at 1.795GHz) the principle-radiating elements are the S-shaped strip and the outer L-shaped strip. However, for the PCS band (at 1.920GHz) it is clear that the S-shaped, outer L-shaped and the wide strip in the middle of the patch are the major radiating elements for this band. Finally, the major radiating elements for the UMTS band (at 2.035GHz) are the outer L-shaped strip and a part of the wide strip in the middle. It should be noted, from Figure 4.4, that the surface currents around the shorting pin on the top surface of the PIFA are always very small (almost zero) due to presence of the shorting pin. Most of the current flows to the ground plane at the point of insertion of the shorting pin into the PIFA's radiating element causes a low electric field as shown in Figure 4.5.

![Surface Current Distributions](image)

Figure 4.4 Simulated surface current distributions at various frequencies for the triple-band PIFA
The measured values of gain and efficiency, against frequency, for the proposed antenna in the DCS-1800MHz, PCS-1900MHz and UMTS bands are depicted in Figure 4.6. For the DCS-1800MHz band, it was noted that the gain drops as frequency increases. The lowest gain value (of 1.46dBi) occurs at 1880MHz. The efficiency measurements for this band showed good values at frequency points inside the bands of interest with a minimum efficiency of 42.37% at 1752.5MHz. For the PCS-1900MHz and UMTS bands, it can be seen that both gain and efficiency values decrease as the frequency increases. In the PCS-1900MHz band the antenna exhibits a minimum gain of 1.48dBi and a minimum efficiency of 51.77% at 1990MHz, while the minimum gain in the UMTS band is −1.4dBi and minimum efficiency is 15.01% at 2170MHz.

The measured radiation pattern in the azimuthal $\mathbf{E}_\phi$ and the elevation $\mathbf{E}_\theta$ planes for the proposed antenna at 1795MHz, 1920MHz and 2035MHz respectively are shown in Figure 4.7, in general a monopole-like radiation pattern was noticed.
Figure 4.6 Triple-band PIFA measured gain and efficiency for DCS1800, PCS1900 and UMTS bands.
Figure 4.7 Triple-band PIFA measured radiation patterns (--- $E_{\phi}$, --- $E_{\theta}$) at:
(a) 1795MHz  (b) 1920MHz (c) 2035MHz
4.2.3 Antenna Performance Inside ABS Plastic Case

This section of the thesis investigates the effect of the handset case on the performance of the antenna under investigation. Figure 4.8 shows the triple-band PIFA mounted inside a plastic case made of Acrylonitrile Butadine Styrene (ABS, $\varepsilon_r = 3.5$), 110mm long $\times$ 56mm wide $\times$ 18mm thick. Measurement results indicate that the plastic case did not significantly affect the antenna performance. Figure 4.9 shows the antenna’s measured input return loss response before and after mounting in an ABS plastic box. The ABS mounting results in a slight shift in resonant frequency as well as a 2.4% reduction in the bandwidth. The measured gain and efficiency for the ABS case mounted antenna are shown in Figure 4.10. The measurements generally show some degradation in the gain and efficiency as the frequency goes up compared to measurements without the plastic case (Figure 4.6). Generally the antenna’s performance inside the plastic case is still acceptable as it comparable to that of conventional PIFA. The gain values all lie above 0.5dBi with a maximum gain of 3.5dBi. The efficiency values range from 25% to 70%.

Figure 4.8 Measurement configuration of the presented triple-band PIFA mounted inside an ABS plastic case
Figure 4.9 Measured input return loss of the triple-band PIFA mounted in ABS plastic case, along with the measured input return loss in free space.

Figure 4.10 Triple-band PIFA mounted in ABS plastic case measured gain and efficiency for DCS1800, PCS1900 and UMTS bands.
4.2.4 Effect of Shorting Pin Radius and Position

Many methods for PIFA tuning have been reported in the literature [11-13]. Among these methods the repositioning of single and multiple shorting pins. The major disadvantage of these techniques is difficulty of fabrication. The effect of the shorting pin depends on several different parameters including the radius, position and length of the pin. The length of the pin is usually determined by the thickness of the antenna.

Shorting pin repositioning and radius variations will be adopted here due to its simplicity compared to other methods. A shorting pin may be analysed by modelling as a transmission line of length \( t \), where \( t \) is the distance between the ground plane and the radiating element [12]. The short piece of line of length \( \Delta z \) can be modelled as a lumped-element circuit, as shown in Figure 4.11, where \( R, L, G, C \) are per unit length quantities. Therefore, if the complete short circuit replaced by shorting pins, a series inductance \( L \) and a shunt capacitance \( C \) will be added to the antenna structure. The series resistance \( R \) represents a resistance due to finite conductivity of the shorting pins whilst the shunt conductance \( G \) accounts for dielectric loss between pairs of shorting pins.

![Figure 4.11 Lumped element equivalent circuit for an incremental length of transmission line [14]](image)

The values of both \( R \) and \( G \) are usually very small and can be neglected. The values of \( L \) and \( C \) depend on the number of the shorting pins employed, their radius \( a \), the separation between the centres \( d \), the permittivity
\( \varepsilon \) and permeability \( \mu \) of the antenna substrate [14]. For instance, if only two shorting pins are used, the values of \( L \) and \( C \) will be similar to those of a short piece of a two-wire transmission line and they are given by [14]:

\[
L = (t \mu / \pi) \cosh^{-1}(d / 2a) \\
C = (t \pi \varepsilon) / \cosh^{-1}(d / 2a)
\]

However, regardless of the number of shorting pins, the analogy between them and short pieces of transmission lines may be used to prove the following:

I. Increasing \((d / 2a)\) increases \( L \).

II. Increasing \((d / 2a)\) decreases \( C \).

III. Both \( L \) and \( C \) increase by increasing the thickness \( t \) of the antenna.

IV. Depending on the values of \( L \) and \( C \), the resultant reactance of the shorting pins will be either inductive or capacitive.

V. The resonant frequency of the short-circuited PIFA will be decreased if the reactance of the shorting posts is inductive, while capacitive reactance will increase the resonant frequency.

The initial parametric study, applied to the triple-band PIFA, investigated the effect of varying the radius of the shorting pin \( r_{sp} \), on the resonant frequency and the impedance bandwidth. For the purpose of this study the antenna parameters fixed and the shorting pin radius \( r_{sp} \) was varied from 0.3mm to 0.9mm. Table 4-1 shows the effect of the shorting pin radius on the antenna resonant frequency and bandwidth. It was noted that the resonant frequency has increased by 8MHz and there was a reduction in the bandwidth by 4.82% as the shorting pin radius increases. Figure 4.12 shows the simulated input return loss for the antenna with different shorting pin radii \( r_{sp} \). The impedance matching of the antenna has improved, especially in the UMTS band, where the upper frequency of this band has increased by 100MHz (from 2.06GHz to 2.16GHz). Increasing the radius of the shorting pin considerably increased its capacitive reactance whilst reducing its inductive reactance, as depicted in the smith chart of Figure 4.13. The
resonant frequency of the antenna is thus increased (according to equation (4.2)) as the shorting pin radius increased.

The antenna’s bandwidth is reduced as the shorting pin radius is increased. This can be explained by the following equations [16]:

\[
\text{Bandwidth} = \frac{(f_u - f_l)}{f_r} = \frac{1}{Q} \tag{4.3}
\]

\[
Q = \frac{1}{R \sqrt{LC}} \tag{4.4}
\]

Where \(f_u\) and \(f_l\) are upper and lower frequencies of bandwidth, \(f_r\) is resonant frequency and \(Q\) is quality factor. Equation (4.3) suggests an inversely proportional relation between \(Q\) and bandwidth. Equation (4.4) on the other hand shows if the antenna has a dominant reactive component, then the quality factor is large. Increasing the shorting pin radius increases the reactive component effect, which increases the quality factor, and consequently reduces the antenna’s bandwidth as shown in Table 4-1.

<table>
<thead>
<tr>
<th>Shorting pin radius (r_{sp}) (mm)</th>
<th>Resonant frequency (f_0) (GHz)</th>
<th>Bandwidth (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.3</td>
<td>1.8934</td>
<td>33.07</td>
</tr>
<tr>
<td>0.4</td>
<td>1.8970</td>
<td>30.93</td>
</tr>
<tr>
<td>0.5</td>
<td>1.8980</td>
<td>30.12</td>
</tr>
<tr>
<td>0.6</td>
<td>1.8991</td>
<td>29.52</td>
</tr>
<tr>
<td>0.7</td>
<td>1.9002</td>
<td>28.97</td>
</tr>
<tr>
<td>0.8</td>
<td>1.9005</td>
<td>28.50</td>
</tr>
<tr>
<td>0.9</td>
<td>1.9009</td>
<td>28.25</td>
</tr>
</tbody>
</table>

Table 4-1 Triple-band PIFA resonate frequency and bandwidth for different radii of the shorting pin \(r_{sp}\)
Figure 4.12 Simulated triple-band PIFA $S_{11}$ for different radii of the shorting pin $r_{sp}$

Figure 4.13 Simulated input impedance for the triple-band PIFA at different radii for the shorting pin $r_{sp}$
The function of next parametric study was to investigate the effect of the shorting pin’s position on the resonant frequency and impedance bandwidth of the triple-band PIFA. The position of the shorting pin is directly related to the current distribution in the antenna and as a direct consequence its input impedance. A simulation study was therefore, applied to the triple-band PIFA by varying the shorting pin’s position according to the reference point showed in Figure 4.14. The pin was moved either towards the long side of the ground plane along the x-axis or away from the feeding point along the y-axis.

For the first case (repositioning the shorting pin along the x-axis) it was noted that the resonant frequency increases as the distance from the reference point increases, as shown in Figure 4.15. This is due to the fact that, for higher frequencies, the impedance of the antenna approaches 50Ω and consequently the resonance frequency increases (as shown in Figure 4.16). For shorting pin positions between $d_x = 1\text{mm}$ and $d_x = 14.5\text{mm}$, the curves were similar to each other and fallen between the curves shown in Figure 4.16. The total shift in the resonant frequency for the total distance ($d_x=13.5\text{mm}$) travelled from the reference point was 43MHz. It was noted that the matching of the band has been improved, but at the same time the bandwidth decreased as depicted in Figure 4.15. There was ~3.33% reduction in the bandwidth compared to the original design before repositioning the shorting pin. To prove this improvement in matching the band, a prototype was build for the case when the shorting pin position is 9mm away from the reference point along the x-axis. The input return losses for this case compared to the original design before repositioning the shorting pin are shown in Figure 4.17. The measured results showed that the upper frequency of the band in the original design has been shifted by 62MHz after repositioning the shorting pin at 9mm from the reference point. This shifting in the frequency has improved the matching by including more frequency points to cover the UMTS band and simultaneously still covering the DCS-1800MHz and the PCS-1900MHz.
Figure 4.14 Triple-band PIFA antenna element with the directions for repositioning the shorting pin from the reference point.

Figure 4.15 Resonant frequency of the triple-band PIFA for various shorting pin positions along the x-axis from the reference point in addition to the (%) BW.

Figure 4.16 Simulated input impedance of the triple band PIFA at different positions for the shorting pin along the x-axis: $d_x = 1\text{mm}$, $d_y = 14.5\text{mm}$.
The second part of the study involved moving the shorting pin, along the y-axis, towards the feeding point. Figure 4.18 shows the simulated return loss results for the triple-band PIFA, with the shorting pin located at various distances from the reference point, towards the feeding point by steps of 1.5mm and up to maximum distance of 16.5mm. It was noted that the effect of this parameter on the matching was a negative effect, where the whole mode was shifted downwards leading to a poor match for the upper frequency points of the UMTS band. The resonant frequency of the lower mode, in Figure 4.18, was shifted from 1.897GHz for the original design to 1.571GHz when the shoring pin was positioned at 16.5mm from the upper left corner of the patch (15mm from the reference point). One may conclude that, reducing the distance between the shorting pin and the feeding point leads to degradation in matching for both bands. Figure 4.19 shows the input impedance variation with frequency with the shorting pin located in two different positions ($d_y = 1.5$mm, $d_y = 16.5$mm). For shorting pin positions between $d_y = 1.5$mm and $d_y = 14.5$mm, the curves were similar and lie between those shown in Figure 4.19. Results recently reported by Panayi [17] generally showed that “The
PIFA’s resonant frequency increases as the distance of the travelling shorting pin from the feeding point increases”.

Figure 4.18 Simulated $S_{11}$ of the triple-band PIFA with the shorting pin positioned at different distances $d_y$ from the reference point

Figure 4.19 Simulated input impedance of the triple band PIFA at different position for the shorting pin along the y-axis: $d_x = 1.5\,\text{mm}$, $d_y = 16.5\,\text{mm}$
4.2.5 Matching Improvement Using Slots

Another method to improve the antenna’s matching and impedance bandwidth at the frequency bands of interest is to add more slots in the available area of the top plate of the antenna’s active element. Such slots have been applied to the proposed antenna in order to force the surface currents to follow different paths, which will improve the antenna’s matching and bandwidth. The original triple-band PIFA design has been modified by removing the dielectric substrate supporting the patch and replacing the shorting pin by a shorting metal wall of 3mm wide located at the upper left corner of the patch. New slots were also cut into the patch. The design resulted from these modifications is depicted in Figure 4.20.

The design described above is a great improvement over the original design since all the bands of interest have been covered completely. Figure 4.21 shows the measured return loss together with the real and imaginary input impedance for the modified and original designs. The antenna’s percentage bandwidth was 34.50%, which represents an increase of about 3.5% over the original design. This increment in the antenna bandwidth was enough to cover the whole UMTS band, which was not covered completely by the original design, where the upper frequency of this band was 2.058GHz and the required value is at least 2.170MHz. In the new design the upper frequency of the UMTS band was 2.282GHz, which is 224MHz over the original design.

![Figure 4.20 Slotted triple-band PIFA design with new slots after removing the dielectric board and replacing the shorting pin with a shorting wall 3mm wide.](image)
4.2.6 SAR Simulation and Experimental Results

It is well known that if the handset antenna is placed in the vicinity of a human biological tissue, some portion of the power delivered by the antenna will be absorbed in the tissue. This effect leads to the development of a parameter called specific absorption rate (SAR), which is now one of the critical antenna design parameters that should be considered in addition to the standard ones. The SAR is defined as the rate at which RF energy is absorbed per unit mass of biological tissue [15,18]. SAR will be discussed in detail in Chapter 6. This section discusses some simulation and experimental measurements that have been applied to the original triple-band PIFA design (presented in section 4.2.1) in order to determine its SAR value. The results were also compared to the basic SAR standard restrictions (ICNIRP guidelines define the basic limit for local exposure to be 2W/kg averaged over a volume of 10g, whereas FCC has slightly more stringent limits of 1.6W/kg over a volume of 1g than the ICNIRP [15]).
The SAR distributions in a phantom head exposed to electromagnetic energy, radiated from the triple-band PIFA presented, have been predicted using CST-MWS™. They were also measured using the in-house SPEAG Dosimetric Assessment System (DASY4)™. The simulation employed a four-tissue phantom head provided by CST-MWS. The model is shown in Figure 4.22 and comprises four concentric spheres (layers), each with different properties. These layers are skin, fat, brain and bone. Table 4-2 shows the electrical parameters of each layer in the phantom head. The simulated SAR results were obtained by positioning the triple-band PIFA antenna ground plane at a distance $d=7.5$ mm along the $z$-axis from the surface of the left flat part of the phantom head as depicted in Figure 4.23. During both simulation and measurements, it was noted that the antenna’s parameters change and the frequency detunes when it is placed close to the phantom head. This occurs because the head acts like a large dielectric load. This clearly evidenced from the input return loss curves of the antenna, before and after placing it in the vicinity of the phantom head, as viewed in Figure 4.24.

Figure 4.22 Four Tissue phantom head provided by CST-MWS
<table>
<thead>
<tr>
<th>Phantom Tissue</th>
<th>Electrical parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Dielectric constant $\epsilon_r$</td>
</tr>
<tr>
<td>Skin</td>
<td>41</td>
</tr>
<tr>
<td>Fat</td>
<td>5</td>
</tr>
<tr>
<td>Bone</td>
<td>12</td>
</tr>
<tr>
<td>Brain</td>
<td>45</td>
</tr>
</tbody>
</table>

Table 4-2 Electrical parameters of each layer in the phantom head

Figure 4.23 Triple-band PIFA in the vicinity of phantom head
Precise SAR measurements were performed using the DASY4 system showed in Figure 4.25. The system provides accurate measurements up to 6GHz in free space as well as inside tissue simulating liquids. The DASY4 system consists of a computer controlled, high precision robotics system which has a high accuracy 6-axis robot (with a working range greater than 1.1m) that positions the SAR measurements probes with a positional repeatability of better than ±0.02mm, extreme near-field probes, data acquisition electronics (DAE) which consists of a highly sensitive electrometer-grade preamplifier with auto-zeroing, a channel and gain switching multiplexer, a fast 16bit AD-converter and a command decoded and control logic unit [19]. The last component in the system is the Specific Anthropomorphic Mannequin (SAM). The SAM is a head shell phantom contains the brain equivalent material that allows a conservative measure of the exposure of all origins and all ages. The ear regions has been defined with reference points and planes to facilitate the reproducible positioning of DUT. Figure 4.25 shows that the SAM phantom is divided to three parts: left, right and flat sections.
The phantom sections were filled by a head simulation liquid of a dielectric constant $\varepsilon_r = 40.48$ and a conductivity $\sigma = 1.37$ Sm$^{-1}$ which is suitable for measuring frequencies in the DCS-1800MHz band. In the measurements, the prototype was attached under the left part of the phantom head in the standard cheek position. The standard position allows more realistic geometric relation between the head and the prototype during ordinary use by rotating the prototype by $59^\circ$ as shown in Figure 4.27 so that the prototype was positioned between the user's mouth and ear and tilted in way to touch the ear and the cheek at the same time during the measurements. When measuring the SAR for the DCS-1800MHz band the standard power transmitted is 125mW. Therefore, when calculating SAR the power has to be normalized to 125mW.

The simulated and measured SAR results are listed in Table 4-3. Unfortunately, it is not possible to compare the measured SAR results with the simulated ones. This is due to the fact that there is a difference in the human head equivalent materials used in the simulation method (using CST-MWS)
and the measurements method (using DASY4 system) as will be discussed in Chapter 6, section 6.3.4. Generally, the antenna’s measured and simulated SAR values met the 2W/kg for 10g volume-averaged SAR standard and only slightly exceeded the 1.6W/kg for 1g volume-averaged SAR standard for the simulated results. Table 4-3 shows also the measured SAR values after including the loss factor resulted from the radiation loss (dissipated power). It can be seen that including the loss factor makes the SAR values higher. The following equations in order to include the loss factor in the SAR were employed:

\[
SAR_{\text{with loss}} = \frac{SAR_{\text{without loss}}}{L_f}
\]  
\[\text{(4.5)}\]

where \(L_f\) is the loss factor.

If it is assumed that the antenna has a loss resistance \(R_L\) and a radiation resistance \(R_r\) with \(X = R_L/R_r\) then according to Balanis [20] the antenna efficiency is:

\[
effic = \frac{X}{(1 + X)} \quad \text{(dimensionless)} \]
\[\text{(4.6)}\]

The system loss factor is \(-L_f\) and can be calculated using the following equation:

\[
L_f = 10 \log_{10} \left( \frac{1}{\text{effic}} \right) \quad \text{dB}
\]
\[\text{(4.7)}\]

\[
L_f' = 10^{L_f/10} \quad \text{(dimensionless)}
\]

The high simulated and measured SAR values including the loss factor which exceeded the 1.6W/kg SAR standard limit can be referred to the fact that the antenna prototype is not embedded in a real mobile handset. Table 4-3 presents also the measured maximum SAR for the constructed prototype. The SAR maximums of the frequency points listed in Table 4-3 were located near the middle part of the ground plane of the antenna under investigation for both, simulation and measurements, as shown for example in Figure 4.26, Figure 4.27 at 1800MHz and Figure 4.28, Figure 4.29 at 1900 MHz.
<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Simulated SAR (W/kg)</th>
<th>Measured SAR (W/kg)</th>
<th>Measured Max SAR (W/kg)</th>
<th>Measured SAR (W/kg) Including loss factor</th>
</tr>
</thead>
<tbody>
<tr>
<td>1710.2</td>
<td>1.64086</td>
<td>1.97024</td>
<td>0.527</td>
<td>0.870</td>
</tr>
<tr>
<td>1747.4</td>
<td>1.63920</td>
<td>1.97565</td>
<td>0.329</td>
<td>0.545</td>
</tr>
<tr>
<td>1784.8</td>
<td>1.64061</td>
<td>1.98585</td>
<td>0.562</td>
<td>0.939</td>
</tr>
<tr>
<td>1800.0</td>
<td>1.63538</td>
<td>1.98192</td>
<td>0.668</td>
<td>1.120</td>
</tr>
<tr>
<td>1850.2</td>
<td>1.59381</td>
<td>2.00754</td>
<td>0.697</td>
<td>1.190</td>
</tr>
<tr>
<td>1880.0</td>
<td>1.57907</td>
<td>1.99761</td>
<td>0.686</td>
<td>1.190</td>
</tr>
<tr>
<td>1900.0</td>
<td>1.55457</td>
<td>1.97221</td>
<td>0.620</td>
<td>1.080</td>
</tr>
<tr>
<td>1909.8</td>
<td>1.56029</td>
<td>1.90980</td>
<td>0.586</td>
<td>1.020</td>
</tr>
</tbody>
</table>

Table 4-3 triple-band PIFA simulated and measured SAR results

The strong magnetic field H that is travelling horizontally between the head and the antenna ground plane causes the SAR maximum to be located at the middle of the ground plane, as depicted in Figure 4.30. Figure 4.30 shows the distributions of magnetic field predicted through computer simulations at 1800MHz and 1900MHz for a phase angle of zero degrees. It can be concluded that the ground plane radiation has a significant impact on the SAR values, in addition to the major radiations from the antenna element.
Figure 4.26 Power distribution of simulated 1g volume-averaged SAR at 1800MHz

Figure 4.27 Power distribution of measured 1g volume-averaged SAR at 1800MHz
Figure 4.28 Power distribution of simulated 1g volume-averaged SAR at 1900MHz

Figure 4.29 Power distribution of measured 1g volume-averaged SAR at 1900MHz
Figure 4.30 Simulated magnetic field H at: (a) 1800 MHz (b) 1900MHz
4.3 GSM-900MHz Addition to the Triple-Band PIFA

A design has been proposed in this section to extend the performance of the triple-band PIFA to operate at the GSM-900MHz band, in addition to the DCS-1800MHz, PCS-1900 and UMTS bands. The structure of the new antenna was inspired by designs for planar meander-line monopole antenna reported in the literature [21-29]. The designs reported in the literature are either a planar meander-line in the level of the top surface of the ground plane [21,26,27] or a folded meander-line with its lower part still in the level of the top surface of the ground plane [22-25,28,29]. The following section will present a quad-band handset antenna, based on these designs, that combines a meander-line monopole with the triple-band PIFA described in section 4.2.1. The monopole was connected to the radiating element of the triple-band PIFA at the same level at the feeding point. Details of the proposed design will be described and experimental results will be presented and discussed.

4.3.1 Meander-Line Monopole Connected to Triple-Band PIFA

The geometry of the proposed quad-band antenna is shown in Figure 4.31. The structure of the antenna is similar to that of the triple-band PIFA presented in Figure 4.1, except that the antenna’s radiating element has been extended to have a meander-line monopole in addition dielectric layer has been removed for ease of fabrication. Taking out the dielectric substrate did not affect the antenna’s performance significantly, since its permittivity is low ($\epsilon_r = 2.2$). The planar meander-line monopole consists of a single strip of copper (total length of 144mm), which is meandered to form the monopole shown within the dashed red rectangular of Figure 4.31. The meander-line monopole measures 36mm wide by 11mm long and connects to the top plate of the triple-band PIFA at 5.9mm above the ground plane at the feeding point. It is noted that the monopole has no ground plane and is separated from the active element of the triple-band PIFA by a distance of 1mm to prevent any coupling with the ground plane and the PIFA’s active element. The monopole is connected to the triple-band PIFA by 1mm wide metal strip. It is clear from Figure 4.31 that the planar meander-line monopole was connected to the
triple-band PIFA at the feeding point, in order to create an additional current loop which would ensure that the meander-line to operate within the GSM-900MHz band.

![Diagram of meander-line monopole connected to the original triple-band PIFA to operate at GSM-900MHz band](image)

**Figure 4.31 Planar Meander-line monopole connected to the original triple-band PIFA to operate at GSM-900MHz band**

### 4.3.2 Simulation and Experimental Results

The triple-band PIFA incorporating the new added planar meander-line monopole has been analysed using CST-MWS. The meander-line monopole geometry has been shaped as shown in Figure 4.31 for two reasons. The first reason is to shorten the length of the monopole making it about the same size as the width of the mobile handset ground plane (36mm). This will ensure that the antenna can be inserted into a mobile handset. The second reason is to achieve wideband performance at the low frequency band (GSM-900MHz) [26]. The total dimensions of the monopole strip are 1mm long by 144mm wide. By meandering, however the area was reduced to 11mm long by 36mm wide. As a result, the monopole can be mounted within a mobile handset.
Increasing the monopole length ($L = 11\text{mm}$) would shorten its width ($W = 36\text{mm}$) even further. The main advantages of the meander-line monopole are its compact structure and moderate bandwidth compared to PIFAs at low frequencies. For example, Figure 4.32 shows one of the intuitive trials used to extend operation of the triple-band PIFA into the GSM-900MHz band. The design shows a new strip of 1mm wide by 23mm long added to the original design, in addition to other modifications of the $\mathbf{S}$-shaped slots. Unfortunately, poor return loss at the lower band (see Figure 4.32) the modifications result in and have a detrimental effect on the bandwidth of the upper band.

Figure 4.32 Modified Triple-band PIFA to operate at GSM-900MHz and its $S_{11}$
A prototype for the planar mender-line monopole integrated to the triple-band PIFA is shown in Figure 4.33. The performance of the antenna was measured in an anechoic chamber. The simulated and measured return losses, depicted in Figure 4.34, are in a good agreement. The results indicate that the lowermost frequency band has a bandwidth of 116MHz (12%) at 5dB return loss with a resonant frequency 0.975MHz, covering 6% of the required GSM-900MHz band (bandwidth 8%) because of 36MHz shift in the resonant frequency. This slight difference between the measured and simulated results may be attributed to fabrication errors, in particular difficult aligning meandered parts of the monopole. The uppermost band has a bandwidth of 777MHz (40%) at 5dB return loss with a resonant frequency of 2.005GHz, covering the DCS-1800MHz, PCS-1900MHz and UMTS frequency bands. The bandwidth of the uppermost band was increased by 10% after adding the monopole. Finally, the narrow band in the middle can be shifted down in the future to couple with the lower band in order to increase the impedance bandwidth of the GSM-900MHz band. Figure 4.35 shows the impedance characteristics of the measured prototype.

Figure 4.33 A prototype of the triple-band PIFA with the integrated planar meander-line monopole
Figure 4.34 Input return loss $S_{11}$ of the presented triple-band PIFA with the integrated planar meander-line monopole

Figure 4.35 Measured input characteristics impedance of the presented triple-band PIFA with the integrated planar meander-line monopole
The simulation results for the presented design showed that the antenna resonates at 0.916GHz. A \( \lambda/4 \) resonant length of \(-82\text{mm}\) was used to achieve this. Figure 4.36 shows the surface current flow in the radiating meander-line monopole at 0.916GHz. The surface current distributions at other frequencies like 1.795GHz, 1.920GHz and 2.035GHz are approximately similar to those showed earlier in Figure 4.4 before integrating the monopole to the triple-band PIFA. The measured gain and efficiency for operating frequencies across the GSM-900MHz and the DCS-1800MHz/PCS-1900MHz/UMTS are depicted in Figure 4.37 and Figure 4.38, respectively. For the GSM-900MHz band the peak gain within the range of operating frequencies was 4.18dBi and the maximum efficiency was 75.07%. At the upper band, which covers the DCS-1800MHz/PCS-1900MHz/UMTS bands, the peak gain was 3.98dBi and the minimum was 1.74dBi, the maximum efficiency was 97%. Throughout the range of operating frequencies the efficiency values never fall below 36%.
The measured radiation patterns at 920MHz, 1795MHz, 1920MHz and 2035MHz are shown in Figure 4.39 and Figure 4.40 respectively. Radiation patterns at other operating frequencies in the bands were also measured and showed similar radiation patterns, as shown in the provided figures. The radiation patterns are generally stable across the respective operating bandwidth. It was also observed that the $E_0$ measured radiation patterns for the GSM-900MHz band, resemble those of a half wavelength dipole. The $E_\phi$ patterns are generally omnidirectional.

Figure 4.37 Measured gain and efficiency for the proposed triple-band PIFA with integrated planar meander-line monopole at GSM-900MHz

Figure 4.38 Measured gain and efficiency for the proposed triple-band PIFA with integrated planar meander-line monopole at DCS/PCS/UMTS bands
Figure 4.39 Measured radiation patterns (--- $E_\theta$, --- $E_\phi$) of the proposed triple-band PIFA with integrated planar meander-line monopole at 920MHz and 1795MHz
Figure 4.40 Measured radiation patterns (--- $E_\phi$, ----------- $E_\theta$) of the proposed triple-band PIFA with integrated planar meander-line monopole at 1922.5MHz and 2035MHz.
4.4 Conclusion

A triple-band PIFA that meets the bandwidth requirements of the DCS/PCS/UMTS cellular systems has been proposed and investigated. The antenna has small dimensions (antenna volumetric size=3.39cm$^3$), a rigid structure and relatively low profile (5.9mm height above the ground plane). Parametric studies of the antenna’s shorting pin radius, shoring pin position and the slots within the top patch of the antenna to tune the resonant frequency and broadening the antenna impedance bandwidth were undertaken. The techniques presented for tuning and broadening the bandwidth facilitate improvement of the antenna’s performance. The antenna performance was also tested inside an ABS plastic case for further characterisation. The results indicated that the antenna’s performance still acceptable inside a plastic case, with some degradation of the gain and efficiency compared with the antenna operating in free space.

The SAR distributions in a human head exposed to electromagnetic waves radiated from the triple-band PIFA have been determined. The results showed that the antenna’s measured and simulated SAR values met the 2W/kg SAR standard, even after including the loss factor. For the 1.6W/kg SAR standard, the results exceeded the limit, particularly after including the loss factor. In simulation and measurements the SAR maxima were located near the middle section of the antenna’s ground plane. This is explained by the presence of a strong magnetic field, which travels perpendicularly towards the head from the antenna’s ground plane.

Finally, based on the triple-band PIFA, a quad-band version was presented to operate at the GSM-900MHz in addition to the previously covered DCS/PCS/UMTS bands. The quad-band antenna was achieved by extending the triple-band PIFA by a planar meander-line monopole operating at the GSM-900MHz band. The quad-band antenna occupied a volume of 5.92cm$^3$. Simulation and measurement results were in a reasonable agreement and the antenna showed a good performance within the bands of interest.
References


Chapter 5

5 Penta-Band Planar Inverted-F Antenna (PIFA) Design With Modified Ground Plane

5.1 Introduction

In Chapter 4 a triple-band PIFA design was introduced and modified to incorporate a suspended planar meander-line monopole fed via a coaxial cable to operate in the low frequency band (GSM-900MHz) to achieve a quad-band design. The design process emphasised the challenges faced by researchers attempting to design multi-band handset antennas within the small space available in a mobile phone. Therefore, to enhance understanding and overcome this complexity in multi-band handset antenna design challenges, Chapter 5 investigates the influence of the ground plane on the impedance bandwidth and resonant frequency of a PIFA. Several studies concerning finite ground planes have been undertaken in the literature to cover more frequency bands through variations in the ground plane itself. These modifications involved etching slotting the ground plane [1-4], varying ground plane length [8-11] and considering a non-metallic ground plane underneath the antenna element to have the merit of the monopole antenna that has broad impedance bandwidth [12,13].

This Chapter describes a mechanism for improving the bandwidth of the lower frequency band of an initial dual band-PIFA design by slotting the ground plane. Then how the design was subsequently extended to a quad-band PIFA design by considering a non-metallic ground plane underneath the antenna element. Finally, the quad-band design was extended to cover a fifth band by optimising length of the ground plane to produce a penta-band PIFA design. The proposed penta-band PIFA was designed to cover the AMPS (824-894) MHz, GSM-900 (890-960) MHz, DCS-1800 (1710-1880) MHz,
PCS-1900 (1850-1990) MHz and UMTS (1900-2170) MHz bands. The penta-band PIFA was produced after studying and evaluating the performance of several designs. These designs were modelled in CST-MWS and their prototypes were fabricated to measure the return loss, efficiency, gain and radiation patterns in an anechoic chamber. SAR measurements were also applied to these designs for further characterisation.

5.2 Initial Design

5.2.1 Initial Design Configuration and Characteristics

The initial design that was modified to produce the subsequent penta-band PIFA is shown in Figure 5.1. It consists of a top plate measuring 42.2mm wide x 16mm long x 0.3mm thick. This component was supported 6mm above a finite conductive single sided FR4 substrate ($\varepsilon_r = 4$) serving as a ground plane, 100mm long x 42.2mm wide x 1.6mm thick. The antenna was fed at the left lower corner of the top plate by 1.4mm wide metal strip connected to a 50Ω coaxial cable and shorted to ground via 1.4mm wide metal strip and located 1.6mm from the feeding strip. The antenna’s feeding scheme was designed to allow the antenna to fit easily inside a real handset. The top plate was slotted as it illustrated in Figure 5.1 to allow the antenna to resonate at two frequency bands, one of them being the low frequency band at around 1GHz and the other one at the upper frequency band in the vicinity of 2GHz.
The simulated return loss of the initial dual-band PIFA design is shown in Figure 5.2. It can be observed that the antenna resonates at two different frequency bands. The first resonance occurs at 1.184 GHz and the second at 2.154 GHz. Figure 5.3 shows the surface current distributions at each of these frequencies. The difference in the current distributions is particularly noticeable around the feeding point and on the ground plane.

Figure 5.2 Simulated $S_{11}$ of the initial dual-band PIFA design

Figure 5.3 Initial dual-band PIFA design surface current distributions at:
(a) 1.184 GHz (b) 2.154 GHz
5.2.2 Initial Design Enhancement With Top Plate Slotting

In order to cover more frequency bands and shift the resonant modes shown in Figure 5.2 (for the initial design Figure 5.1) towards standard mobile communication bands, more slots have been cut within the top plate as depicted in Figure 5.4. The new configuration features two L-shape slots; one of them has been cut into the left part of the plate and the other one at the right part of the plate. The length of the notch between the feeding and the shorting strips together with the dimensions of the slots of the initial design were altered. It should be stated that all the slots within the top plate of the antenna have a very important impact on the return loss since they force the surface currents to follow in certain paths causing a shift in the resonant frequency, the reason for this will be explained in this section. The simulated return loss for the new configuration is shown in Figure 5.5. The new configuration showed a better return loss compared to the initial design, and the lower band has been shifted down by 190MHz to resonate at 0.995GHz with a bandwidth of 11%. On the other hand, the upper band has been split into two modes; one of them resonates at 1.965GHz and the other at 2.280GHz. This allows the antenna to cover a part of the DCS-1800MHz, PCS-1900MHz and UMTS bands at 5dB return loss. The surface current distributions for the new configuration are viewed in Figure 5.6. Figure 5.6 (a) shows the surface current distribution at 0.995GHz and highlights the importance of the new L-shape slot, introduced into the right hand side of the plate, in achieving this resonance. The mode supports a lower frequency resonance by providing an extended electrical length. Figure 5.6 (b) and (c) show the importance of both introduced L-shape slots and the notch in achieving other resonant frequencies.
Figure 5.5 Simulated $S_{11}$ of the improved initial dual-band PIFA design

Figure 5.6 Improved initial dual-band PIFA surface current distributions at:
(a) 0.995GHz  (b) 1.965GHz  (c) 2.280GHz
5.2.3 Expanding of Antenna Element Dimensions

Section 5.2.2 observed that the antenna presented there, still requires some improvement to cover the standard mobile communication bands. These improvements include shifting resonant modes and broadening their bandwidth to match those of the standard communication bands. In this section the length of the antenna's radiating element, shown previously in Figure 5.4 has been extended whilst its width was reduced slightly. In addition to this a greater number of slots were introduced into the top plate. These modifications have been applied to force the antenna to operate at the GSM-900MHz and the DCS-1800MHz in addition to the PCS-1900MHz and UMTS bands that already covered. The configuration of the new expanded design is shown in Figure 5.7. The antenna element is 25mm long by 38mm wide and placed 6mm above the ground plane. The ground plane measures 100mm long, 38mm wide and 1.6mm thick. The antenna volume is increased by 23% compared to the antenna design shown in Figure 5.4. Four fingers (slots) have been added to the new configuration in the right hand side of the top plate. These fingers replace the L-shape showed before in Figure 5.4. These fingers were very important in enabling coverage of the low frequency band, as will be explained later.

A prototype of the proposed design has been fabricated for measurements as depicted in Figure 5.7. The simulated and measured return losses are provided in Figure 5.8. Very good agreement between the simulated and measured return loss results was noted. The expanded PIFA design showed some performance improvement compared to the design showed in Figure 5.4, especially in the lower frequency band, which has been shifted downwards by 24MHz and the bandwidth increased by 5%. The upper band was also shifted slightly downwards to cover more frequency points within the DCS-1800MHz band whilst losing some frequency points belonging to the UMTS band. Figure 5.9 (a) shows the surface current distributions for the lower frequency band at 0.970GHz resonant frequency. This clearly illustrates the importance of the four fingers slotted in the right of the top plate. The fingers enable the antenna to operate at a resonant frequency lower than that
achieved for the antenna in Figure 5.4. At the 0.970GHz resonant frequency a quarter of a guided wavelength is 77.3mm. The resonant element was realised by slotting an antenna element measuring 25mm long by 38mm wide. Figure 5.9 (b) shows the surface current distribution for the upper frequency band at 1.970GHz resonant frequency. It was observed that the L-shape slot in the left hand side of the top plate is the main radiating element for this band, in addition to some coupling with other slots in the antenna element.

![Figure 5.7 Expanded PIFA design configuration and a prototype](image)

![Figure 5.8 Simulated and measured S11 of the expanded PIFA design](image)
The results for the expanded PIFA design show that the antenna’s performance requires further improvements. The lower band needs to be shifted and broadened to cover the GSM–900MHz and the AMPS bands. The upper band also needs broadening in bandwidth in order to cover the DCS-1800MHz, PCS-1900MHz and UMTS bands. Several ideas have been applied to the expanded PIFA design in an effort to realise these improvements. The first idea has already been discussed. The technique involves introducing more slots within the top plate of the antenna and varying the current slots dimensions to improve the antenna’s performance. Many intuitive trials have been applied to the antenna by varying the slots dimensions. But unfortunately the improvement in performance was insignificant, especially at the lower band which is the most difficult to achieve, because of the narrow bandwidth at low frequencies which is PIFAs suffer from. The narrow bandwidth occurs due to restrictions on the volume allocated to the antenna inside a mobile handset. Chapter 2 explained how is the PIFA’s bandwidth increases in relation to the height of the antenna’s element above the ground plane and size of the planar element. PIFAs are classified as small antenna because they are small compared to the wavelengths of the electromagnetic fields they radiate. More specifically, the term “electrically small antennas” has become understood to include any antenna which fits inside a sphere of radius $a = 1/k$. 

Figure 5.9 Surface current distributions of the expanded PIFA at: 
(a) 0.970GHz (b) 1.970GHz
where \( k \) is the wave number \((2\pi/\lambda)\) associated with the electromagnetic field [5,6]. Some investigations [6,7] have established a relationship between the volume of antenna such as PIFA (which can be enclosed in a sphere of radius \( a \)), its minimum radiation quality factor \( Q(a,k) \) and the maximum attainable bandwidth. These investigations suggested that there is an inversely proportional relationship between the antenna bandwidth and the radiation \( Q \). Consequently, a similar relationship between the radiation \( Q \) and the antenna’s volume was also reported.

The second idea is to insert slots within a region of the antenna’s ground plane directly underneath the antenna element and close to the feeding point. The aim of this method was to broaden the bandwidth of existing frequency bands or to add another band. The third idea is also based on ground plane modification, but this time the idea is to consider a non-metallic ground plane underneath the antenna element. These two ideas have been applied to the expanded PIFA design shown in Figure 5.7. Simulation and measurement results were used to examine the effectiveness of these techniques.

### 5.3 Ground Plane Modification for Expanded PIFA

#### 5.3.1 Inserting Slots in the Ground Plane

It is well known that the geometry of the PIFA’s ground plane has a significant effect on the antenna’s performance. The reason for this is that the ground plane radiates a considerable amount of energy delivered to the antenna. Figure 5.3 and Figure 5.9 showed the surface current distributions of two different PIFA designs. It is very clear from the figures that the surface current distributions on the ground plane are more dominant directly below the radiating element and near the feeding point, than in any other region. It is therefore expected that any modification of this region of the ground plane will affect the antenna performance.

Figure 5.10 shows three possible configurations for the slot cut into the ground plane. In all of these configurations the slots are located directly
underneath the antenna element because of the presence of the handset battery and other electronics, which should be considered. The simulation results in Figure 5.11 show that the configuration in Figure 5.10 (c) simultaneously broadens the bandwidth of the lower and upper frequency bands. Configurations (a) and (b) primarily broaden the bandwidth of the lower band and slightly the upper frequency band. The bandwidth of the antenna’s lower frequency band before slotting the ground plane was 12%. While after slotting, it was broadened to 19% using configuration (c) shown in Figure 5.10. The bandwidth of the upper frequency band was also broadened from 13% to 16% using configuration (c). It is very important to observe that all of the configurations shown in Figure 5.10 are always starting at a point that is very close to the feeding point and has an open end to the edge of the ground plane. This observation was reached after many simulations and investigations for different configurations. It has been noted that inserting a slot in the ground plane far away from the feeding point and/or not opened at one of the slot ends, does not improve the antenna bandwidth. This is due to the fact that the surface currents are concentrated around the feeding point area and flow at the edges of the ground plane.

![Figure 5.10 Different configurations of slot cut in the ground plane](image-url)
Figure 5.11 Simulated input return losses of the investigated slot cut configurations in the ground plane of the expanded PIFA compared to the case before slotting

The slot configuration shown in Figure 5.10 (c) occupies an area of 28mm long by 38mm wide within a ground plane measuring 100mm long by 38mm wide. This ground plane is situated below an antenna element measuring 25mm long by 38mm wide. Note that the dimensions of the area where the slot has been inserted are almost identical to those of the antenna element area. Recently, some results reported by [1,2], used the proposed technique either to add more bands or to broaden existing ones. The results in [1] were obtained by cutting a slot of 60mm long by 1mm wide in a ground plane measuring 114mm long x 50mm wide with an antenna element of 50mm wide x 24mm long x 7mm thick. This implies that the introduced slot is not accommodated underneath the antenna element. On the other hand, the results in [2] were obtained by meandering the overall ground plane geometry.

A prototype incorporating slot configuration (c) cut into the ground plane underneath the antenna element is depicted in Figure 5.12. The measured return losses before and after inserting the slot into the ground plane combined with the simulated return loss of the designed antenna after slotting are shown in Figure 5.13. The results showed that there is a reasonable agreement between simulation and measurements. A comparison between the measurements before and after slotting the ground plane revealed that the slot
significantly improves the impedance bandwidth, particularly within the lower frequency band, which is always forming a challenge to match the required bandwidth. The antenna achieved the required performance for the GSM-900MHz band with 25% bandwidth and 950MHz resonant frequency after cutting a slot cut into the ground plane. This means that the bandwidth has been improved by 11% compared to the design before slotting and all the frequency points in the GSM-900MHz band have been covered. The upper band was shifted, by the presence of the slot, by 50MHz to resonate at 1920MHz and the bandwidth remained constant (approximately 15% for both cases). It can be seen that the upper band covers the PCS-1900MHz band, but needs improvement to cover the DCS-1800MHz and UMTS bands. The measured input impedance of the expanded PIFA with the slotted ground plane is provided in Figure 5.14.

![Antenna element](image)

**Figure 5.12** Prototype of the expanded PIFA design after inserting slots in the ground plane based on configuration (c)

The simulated surface current distributions at 1.03GHz and 1.874GHz are shown in Figure 5.15. From the figure it is clearly evident how the slots in the ground plane influence the surface current distributions on the antenna element surface or on the ground plane. This is referred to the variation in the
phases of the flowing currents, which led to a substantial change of the antenna properties. The lower slot cut in the ground plane is located close to the longest side of the ground plane and controls performance of the lower frequency band, while the upper slot cut, located close to the short side of the ground plane controls performance at the upper frequency band, as shown by the surface current distributions in Figure 5.15.

![Graph 5.13](image1.png)

**Figure 5.13** Measured and simulated $S_{11}$ of the expended PIFA design before and after inserting slots in the ground plane

![Graph 5.14](image2.png)

**Figure 5.14** Measured input characteristic impedance of the expanded PIFA with slotted ground plane
Efficiency and gain measurements for the studied antenna under investigation are shown in Figure 5.16 for the lower and the upper frequency bands. The measurements for the lower frequency band (GSM-900MHz) indicate that the antenna has a good efficiency. The efficiency values lie between 50% and 85%, whilst the gain values range from 2dBi to 4.5dBi. This constitutes good performance in the GSM-900MHz band. The antenna also exhibits good performance in the PCS-1900MHz band and reasonable performance for the DCS-1800MHz and the UMTS bands. Generally, the efficiency values for the upper bands were between 20% and 75%, while the gain values range from 0.5dBi to 3.5dBi. The measured radiation patterns for the antenna after inclusion of ground plane slot configuration (c) are shown in Figure 5.17 for 920MHz and 1920MHz. At 920MHz the $E_\theta$ component has generally an omnidirectional behaviour, while the $E_\phi$ behaves similar to that of a half wavelength dipole. For 1920MHz, the patterns showed increased directivity in certain directions.
Figure 5.16 Expanded PIFA with slotted ground plane efficiency and gain measurements at: (a) Lower band (b) Upper band
Figure 5.17 Measured radiation patterns (\( \mathbf{E}_\phi \), \( \mathbf{E}_z \)) of the expanded PIFA after ground plane slotting at: 920MHz, 1920MHz
5.3.2 Non-Metallic Ground Plane Under Antenna Element

The method of slotting the ground plane proposed earlier was only successful in broadening the bandwidth of the lower band, to meet the specifications for the GSM-900MHz frequency band. The method was only partially successful in broadening the bandwidth of the upper frequency band. This section will therefore present an alternative technique for improving simultaneously the performance for both bands of the expanded PIFA design shown in Figure 5.7. This technique employs a non-metallic ground plane below the antenna element in order to improve its performance. The technique yields very good level of improvement in the antenna’s performance as confirmed by simulation and measurement results presented later in this section. Figure 5.18 shows the expanded PIFA design showed earlier in Figure 5.7 after the ground plane PCB has been divided to two parts.

![Diagram of expanded PIFA design](image)

**Figure 5.18** Configuration and a prototype of the expanded PIFA design after considering a non-metallic ground plane underneath the antenna element
The first part is a non-metallic area underneath the antenna element and the other part (rest of the ground plane) is a conductive part that serves as the ground plane for the antenna. The non-metallic part is 16mm long by 38mm wide and located underneath the antenna element, which is 25mm long by 38mm wide. It can be noted from the design that a part of the antenna element length (9mm) overlaps above a metallic area. The simulated and measured return losses for the antenna under investigation are shown in Figure 5.19. Reasonable agreement between the simulated and measured results was noted. Figure 5.20 shows the measured return loss of the expended PIFA design before and after considering a non-metallic ground plane, underneath the antenna element. The results indicate that this modification significantly improves the antenna’s performance in both, the lower and the upper frequency bands. The lower band covers the GSM-900MHz band with 21.6% bandwidth at 5dB return loss. The upper band covers the DCS-1800MHz, PCS-1900MHz and UMTS with a bandwidth of 35.34% at 5dB return loss. It is clear from these results that the technique is highly effective. The technique, not only, splits and broadens the upper band to cover three new communication bands but also broadens the lower band.

Simulated surface current distributions for the new design at resonant frequencies 1.01GHz, 1.795GHz and 2.162GHz are shown in Figure 5.21 (a), (b) and (c). The figures emphasise the important contribution of the ground plane to the radiated field at the lower frequency band at 1.01GHz. At this resonant frequency most of the ground plane area around the feeding system and the ground plane edges are radiating in addition to the antenna element. This occurs due to the long wavelength required at this low frequency band. For higher frequencies it was noted that the amount of radiation from the ground plane is less than that for low frequencies, which is expected. The antenna exhibits good radiation efficiency and gain values as depicted by the measurements in Figure 5.22 within the GSM-900MHz band (Figure 5.22 (a)). The peak gain varies from 1.5dBi to 4dBi and the antenna efficiency ranges from 30% to 90%. In the upper frequency band, which covers the DCS-1800MHz/PCS-1900MHz/UMTS, the peak gain varies from 0.15dBi to 4.5dBi and the efficiency ranges from 25% to 80%. Measured radiation
patterns for the proposed antenna at 920MHz, 1795MHz, 1920MHz and 2035MHz are presented in Figure 5.23 and Figure 5.24. The radiation patterns at 920MHz are approximately similar to those of a simple straight monopole. At 1795MHz, 1900MHz and 2035MHz no significant difference between the radiation patterns can be noted and is generally like omnidirectional radiation.

Figure 5.19 Simulate and measured $S_{11}$ of the expanded PIFA design after considering a non-metallic ground plane underneath the antenna element

Figure 5.20 Measured $S_{11}$ of the expanded PIFA before and after considering a non-metallic ground plane underneath the antenna element
Figure 5.21 Simulated surface current distributions of the expanded PIFA after considering a non-metallic ground plane underneath the antenna element at:
(a) 1.01GHz (b) 1.795GHz (c) 2.162GHz

Figure 5.22 Measured gain and efficiency of the expanded PIFA after considering a non-metallic ground plane underneath the antenna element for:
(a) GSM-900 (b) DCS/PCS/UMTS
Figure 5.23 Measured radiation patterns ($E_{\phi}$, $E_{\theta}$) of the expanded PIFA after considering a non-metallic ground plane underneath the antenna element at:

- 920MHz
- 1795MHz
Figure 5.24 Measured radiation patterns ($\cdots E_\phi$, $E_\theta$) of the expanded PIFA after considering a non-metallic ground plane underneath the antenna element at:
1920MHz and 2035MHz
The antenna’s SAR was measured using the Dosimetric Assessment System (DASY4) before and after considering a non-metallic ground plane underneath the antenna. The measured SAR values at 915.8MHz before considering non-metallic ground plane were 0.861W/kg for the 10g volume-averaged SAR and 1.26W/kg for the 1g volume-averaged SAR. After considering the non-metallic ground plane the SAR values rose considerably to 1.95W/kg for the 10g volume-averaged SAR and 2.98W/kg for the 1g volume-averaged SAR. At 1850.2MHz the measured SAR values before considering the non-metallic ground plane were 0.903W/kg for the 10g volume-averaged SAR and 1.73W/kg for the 1g volume-averaged SAR. After considering the non-metallic ground plane the SAR values rose slightly to 0.964W/kg for the 10g volume-averaged SAR and 1.96W/kg for the 1g volume-averaged SAR.

By comparing the measured values it can be concluded that the SAR is always higher after considering a non-metallic ground plane underneath the antenna. For both studied frequencies (915.8MHz and 1850.2MHz) the 1g volume-averaged values exceeded the standard limit (1.6W/kg for 1g volume-averaged). The antenna meets the 10g volume-averaged SAR standard limit (2W/kg for 10g volume-averaged SAR standard).

Figure 5.25 shows the power distribution for the measured SAR at:
(A) 915.8MHz before considering the non-metallic ground plane, (B) 915.8MHz after considering the non-metallic ground plane, (C) 1850.2MHz before considering the non-metallic ground plane and (D) 1850.2MHz after considering the non-metallic ground plane.

In cases (A) and (C) the SAR maximums (the cubes which represents the zoom scan around the SAR hot spot) are located towards the top part of the handset (within the area allocated to the antenna element). While in case (B) and (D), where a non-metallic ground plane was used underneath the antenna, the SAR maximums are located in the middle part of the ground plane and away from the antenna element. This explains the high SAR values observed in cases (B) and (D) compared to cases (A) and (C), since the ground plane is
closer to the user head than the antenna element, when the handset is in the talk position. The antenna’s ground plane in case (B) and (D) radiates strongly than it radiates in cases (A) and (C). Thus explaining the high SAR values for case (B) and (D). Measured SAR values at other frequency points for the expanded PIFA design after considering a non-metallic ground plane underneath the antenna element are listed in Table 5-1. The values generally met the 2W/kg for 10g volume-averaged SAR standard and exceeded the 1.6W/kg for 1g volume-averaged SAR standard.

![Graphs showing power distribution of measured SAR for different frequency points](image)

Figure 5.25 Power distribution of measured SAR for the proposed PIFA design at: (A) 915.8MHz before considering non-metallic ground plane, (B) 915.8MHz after considering non-metallic ground plane, (C) 1850.2MHz before considering non-metallic ground plane and (D) 1850.2MHz after considering non-metallic ground plane.
5.4 Penta-band PIFA Design

5.4.1 Effect of Ground Plane Length

After the quad-band PIFA design presented in section 5.3.2 there is a need to integrate a fifth band in the antenna to operate at the AMPS (824MHz-894MHz) band. In order to integrate an extra frequency band to the design the author began by considering the ground plane size should be taken, since the size of the ground plane plays an important role in defining the antennas performance (resonant frequency, bandwidth, gain and radiation patterns) [4], [8]. The influence of the ground plane size can be understood by recalling that the impedance of an antenna oscillates with ground plane size and converges toward a value as the ground plane size increases. In this section a parametric study has been applied to the quad-band PIFA design, shown in Figure 5.18, to investigate the effect of the ground plane length on the impedance bandwidth and resonant frequency.

The quad-band PIFA, shown in Figure 5.18 has a ground plane with dimensions 100mm long by 38mm wide. The parametric study applied to the antenna included the following ground plane lengths: 50, 60, 70, 80, 90, 100, 120, 130 and 140mm. The purpose of the study was to discover a means of

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>SAR (10g average) (W/kg)</th>
<th>SAR (1g average) (W/kg)</th>
</tr>
</thead>
<tbody>
<tr>
<td>890.2</td>
<td>1.930</td>
<td>2.930</td>
</tr>
<tr>
<td>915.8</td>
<td>1.950</td>
<td>2.980</td>
</tr>
<tr>
<td>950.0</td>
<td>2.250</td>
<td>3.49</td>
</tr>
<tr>
<td>1747.4</td>
<td>1.020</td>
<td>1.76</td>
</tr>
<tr>
<td>1784.8</td>
<td>1.040</td>
<td>1.82</td>
</tr>
<tr>
<td>1800.0</td>
<td>0.899</td>
<td>1.79</td>
</tr>
<tr>
<td>1850.2</td>
<td>0.964</td>
<td>1.96</td>
</tr>
<tr>
<td>1880.0</td>
<td>1.030</td>
<td>2.12</td>
</tr>
<tr>
<td>1900.0</td>
<td>0.965</td>
<td>1.98</td>
</tr>
<tr>
<td>1909.8</td>
<td>0.987</td>
<td>2.01</td>
</tr>
</tbody>
</table>

Table 5-1 Measured SAR values at different frequency points for the expended PIFA design after considering a non-metallic ground plane underneath the antenna element
broadening the lower GSM-900MHz band to cover the AMPS band without degrading the match at the upper band that covers the DCS-1800MHz/PCS-1900MHz/UMTS. Figure 5.26 shows simulated results of resonant frequencies and bandwidths of the quad-band PIFA (shown in Figure 5.18) with a range of different ground plane lengths. From the results it is clear that the maximum impedance bandwidth correspond to a ground plane length of 120mm. When the ground plane length increased beyond this value the impedance bandwidth started to decrease. The results indicate a significant variation in the impedance bandwidth values for the different ground plane lengths considered. The variation of resonant frequency with ground plane length is small. The upper most resonant frequency occurs for a ground plane length of 120mm. In brief, since the resonant frequency and the impedance bandwidth are both related to the impedance, they should exhibit the same behaviour as shown in Figure 5.26. Where it is clear from the curves that the resonant frequency values oscillate down and the bandwidth values oscillate up as the ground plane length increases. This oscillation continues until an optimum ground plane length is obtained.

Figure 5.27 shows the simulated and measured return losses for the proposed penta-band PIFA after increasing the length of the ground plane from 100mm to 120mm. Good agreement was noted between the simulated and measured results. Comparing the measured return loss for the penta-band PIFA design to the quad-band PIFA showed that there is a significant improvement in the lower frequency band, which has been broadened to cover most of the AMPS band in addition to the GSM-900MHz band as shown in Figure 5.28. The results showed that the impedance bandwidth of the lower frequency band has been increased from 21.6% to 31.37% at 5dB to cover the frequency points between 0.840GHz and 1.18GHz. The bandwidth completely covers the GSM-900MHz and 6% of the AMPS (required standard bandwidth 8%), where a 16MHz shift in the lower band is enough to cover both GSM-900MHz and AMPS bands completely. The impedance bandwidth of the upper frequency band was reduced by only 0.54%, which is not a significant variation.
Figure 5.26 Simulated impedance bandwidth and resonant frequency of the quad-band PIFA design for different ground plane lengths.

Figure 5.27 Simulated and measured $S_{11}$ of the proposed penta-band PIFA design with a ground plane of 120mm length.

Figure 5.28 Compared measured $S_{11}$ of the penta-band PIFA design to the quad-band PIFA design.
The measured gain and efficiency of the penta-band PIFA for the lower frequency bands (AMPS/GSM-900MHz) and the upper frequency bands (DCS-1800/PCS-1900MHz/UMTS) are depicted in Figure 5.29 and Figure 5.30 respectively. For the AMPS band, a peak antenna gain of about 3.63 dBi was observed, with an average gain of 2.34dBi. The efficiency values for this band were varied between 21% and 80%. The peak gain value for the GSM-900MHz band was 4.16dBi with an average gain of 3.66dBi and the efficiency values lay between 54% and 86%. For the DCS-1800MHz, PCS-1900MHz and UMTS bands, the measured peak antenna gain was 4.66, 4.11, 4.83dBi, respectively, and the gain averaged across all the bands was ~3.5dBi. The measured efficiency values for the DCS-1800MHz, PCS-1900MHz and UMTS bands were generally varied between 45% and 85%.

Figure 5.31 shows the simulated surface current distribution for the proposed penta-band PIFA at the lower resonant frequency 1.145GHz. The figure shows the role played by the ground plane in enabling the achievement of lower frequency bands such as the AMPS and GSM-900MHz bands. At these frequencies large surface currents flow along the edges and the top of the ground plane, since the resonant electrical length is large at these lower frequencies. Up to a certain limit, increasing the ground plane length was thus helpful in broadening the bandwidth to cover more frequency points. The measured radiation patterns of the antenna prototype at 859MHz and 1920MHz are shown in Figure 5.32. The patterns are monopole like radiation patterns for $E_\phi$ and showed good omnidirectional behaviour for $E_\theta$ at 859MHz. For the 1920MHz the radiation patterns were more directive in some directions and almost omnidirectional. Radiation patterns were also measured for other operating frequency points and the radiation patterns were generally stable.
Figure 5.29 Measured gain and efficiency of the penta-band PIFA design for the AMPS/GSM-900MHz bands

Figure 5.30 Measured gain and efficiency of the penta-band PIFA design for the DCS-1800MHz/PCS-1900MHz/UMTS bands
Figure 5.31 Simulated surface current distribution for the penta-band PIFA design at the lower resonant frequency (1.145GHz)
Figure 5.32 Measured radiation patterns \( \ldots E_{\text{eq}}, \ldots E_0 \) of the penta-band PIFA design at: 859MHz and 1920MHz
5.5 Conclusion

A penta-band PIFA design was presented in this chapter. The design almost meets the bandwidth requirements of the AMPS communication band (6% of the required 8% standard bandwidth). The design meets the GSM-900MHz, DCS-1800MHz, PCS-1900MHz and UMTS communication bands requirements precisely. The antenna had two resonant modes, one of them covers the lower bands AMPS/GSM-900MHz with 31.37% bandwidth and the other one covers the upper bands DCS-1800MHz/PCS-1900MHz/UMTS with 34.80% bandwidth.

The penta-band PIFA investigated in this chapter features a ground plane measuring 120mm long by 38mm wide and a radiating plate slotted and suspended 6mm above the ground plane. It occupies 20% of the total ground plane area. The antenna’s dimensions are practical for a real mobile handset operating over these five standard frequency bands. Simulation and measurement results are in good agreement for the multi-band behaviour of the proposed antenna. SAR measurements for the antenna before integrating the fifth band showed that the antenna meets the 2W/kg for 10g volume-averaged SAR standard and slightly exceeds the 1.6W/kg for 1g volume-averaged SAR standard.

The penta-band PIFA design evolved through several iterative stages before being finalised. Several ideas and techniques have been applied to an initial dual-band PIFA design to improve its performance to yield a penta-band PIFA design. These ideas were involving modifying two parts of the antenna. The first part was the antenna planar element (patch) and the second part was the antenna ground plane. The modifications made to the first part involved introducing additional slots within the antenna element to force the surface currents to follow certain paths, in order to vary the electric and magnetic fields distribution. Another modification was also employed; this involved increasing the antenna element’s physical size to vary the wavelength. The modifications to the ground plane involved inserting slots in the ground plane underneath the antenna element, considering a non-metallic
ground plane underneath the antenna element and finally optimising the antenna ground plane length. These modifications in the ground plane led to a substantial change in the antenna’s properties due to variations in the relating phases of the currents flowing in the ground plane. It was noted that most of these ideas were concentrating on making some modifications to the antenna ground plane. It can thus be concluded that the ground plane of a PIFA plays a vital role in defining its performance.
References


Chapter 6

6 SAR and Efficiency Study for a Dual-band PIFA Handset Antenna (GSM900/DCS1800)

6.1 Introduction

It is well known that using mobile phones lends to power absorption in human tissue (mainly the head and the hand). This effect is known as the Specific Absorption Rate (SAR). SAR is a function of the antenna radiation efficiency among other factors, since it consumes some of the power available for communication. All of the designs presented in the previous Chapters were PIFAs. For this reason, it is important to investigate specifically PIFAs SAR characteristics. In this chapter a SAR study has been applied to a dual-band Planar Inverted-F Antenna (PIFA) operating at the GSM-900MHz and DCS-1800MHz frequency bands. The effect of varying the distance between the dual band PIFA and the human head, on the SAR has been studied at the uplink frequencies of both bands. Furthermore, the relationship between the SAR and efficiency has also been investigated. This investigation was motivated by the fact that SAR and efficiency are both key parameters in developing high performance wireless devices. Calculating SAR independently presents an incomplete picture and leads to poor device performance. The Study is carried out using two identical antennas, which differ only in the feeding mechanism. Different feeding techniques were employed to obtain different performance characteristics from the antennas at particular frequency band. Consequently, the two antennas will have different efficiency values for the same frequency point. This facilitates the relationship between SAR and efficiency. SAR distributions in a human head exposed to electromagnetic radiation, from the dual-band PIFA, have been predicted using CST MICROWAVE STUDIO™ simulation package and measured using the in house SPEAG Dosimetric Assessment System (DASY4™).
6.2 Specific Absorption Rate (SAR) Theory

The SAR is a measure of the maximum energy absorbed by a unit of mass of tissue (Watts/kilogram) over a given time [1], or more simply the SAR can be defined as a unit of measurement of the amount of radio frequency energy absorbed by the body when using a radiation device at certain frequency. It can be calculated using the following equation provided by [2-4]:

\[
SAR = c \frac{\partial T}{\partial t} = \frac{P_{abs}}{\rho} = \frac{\sigma}{\rho} |E|^2
\]  

(6.1)

Where:
- \(c\) Specific heat of tissue phantom (Jkg\(^{-1}\) K\(^{-1}\))
- \(\frac{\partial T}{\partial t}\) Changes in heat over time (Ks\(^{-1}\))
- \(\sigma\) Conductivity of tissue (Sm\(^{-1}\))
- \(\rho\) Density of tissue (Kgm\(^{-3}\))
- \(E\) Electric field strength (Vm\(^{-1}\))
- \(P_{abs}\) Absorbed power within the 1g or 10g cube (W)

Equation (6.1) shows two possible methods to determine the SAR. The first method is to calculate the SAR by measuring the electric field strength in the human tissue. It is clear from the equation that this method requires the knowledge of conductivity and density values of tissue. The second method to determine the SAR is to measure the temperature rise in the tissue versus time. The specific heat of the tissue under investigation must be known for this purpose. Unfortunately temperature rise measurements in a tissue simulating liquids are extremely difficult to perform due to the extremely very small temperature increments to be measured. Due to the aforementioned difficulties it is more reliable to use the electric field to obtain the SAR. This technique was used to produce the results in this Chapter.

SAR is the primary dosimetric parameter for the evaluation of the human exposure to electromagnetic energy in the frequency range between 100kHz-10GHz. It is very necessary to be judged by limits set down in
different standards, guidelines and regulations. These safety limits for electromagnetic exposure have been devised by national and international organisations, e.g., [5], [6]. In Europe, the exposure levels for mobile handsets are mainly evaluated by applying the International Commission on Non-Ionising Radiation Protection (ICNIRP) guidelines. The ICNIRP define the basic limit for local exposure to be $2\text{Wkg}^{-1}$ averaged over a volume of 10g over a period of 30 minutes [5]. In the United States, the Federal Communications Commission (FCC) has adopted the slightly less limits set by ANSI/IEEE standard [6]. The ANSI/IEEE standard limit is $1.6\text{Wkg}^{-1}$ averaged over a volume of 1g over a period of 6 minutes. Several standardisation bodies (i.e. IEEE SCC-34 SCC-2, CENELEC TC-211 WG-2, IEC TC106, ARIB, etc.) have drafted recommended practices for experimental verification of the compliance of mobile telecommunication devices with these basic limits. Scanning the electromagnetic fields within anthropomorphic phantom filled with homogeneous brain tissue simulation liquid usually does the compliance testing. For a particular frequency band, it is always assumed the worst-case absorption due to high losses.

6.3 SAR and Efficiency Study for a Dual-Band PIFA

Most of the SAR studies in the literature focused on the GSM-900MHz and the DCS-1800MHz cellular bands because guidelines and standards for these bands have been clearly established. It is expected that with the release of the third generation of mobile communications, a new guidelines and standards will shortly emerge. A dual-band PIFA handset antenna presented in the next section is designed to operate at the GSM-900MHz and the DCS-1800MHz. The handset antenna was placed at different distances from the head and at different frequencies to investigate the SAR and efficiency. The reason for not applying the study to the dual-band PIFA design presented previously at Chapter 3; is that the antenna’s lower frequency band did not cover completely all the uplink frequency points required for the GSM-900MHz band to carry out the SAR study.
6.3.1 Dual-Band PIFA Design Configuration

The subject of the SAR and efficiency study is the dual-band PIFA design depicted in Figure 6.1. The antenna consists of a rectangular patch, 40mm wide by 35mm long supported 7.9mm above a single sided PCB ($\varepsilon_r = 4$), 100mm long by 40mm wide by 1.6mm thick. The antenna is matched to a 50Ω coaxial cable connected to a vertical strip for feeding. A second strip shorts the antenna to the ground plane. The desired frequency bands have been obtained by cutting slots in the rectangular patch to force currents to follow certain paths as shown in Figure 6.1. Figure 6.2 (a) shows the surface current distribution at 0.923GHz. In the GSM-900MHz band there is a clear evidence of coupling between the area around the large slot (on the right hand side of the antenna) and the area around the feeding strip. However, for the DCS-1800MHz band at 1.787GHz, the principle radiating elements were the L-shape strip at the upper left corner and some areas around the large slot (at the right hand side of the antenna) as shown in Figure 6.2 (b). One also should note the difference between the surface current distributions on the top surface of the ground plane for the two frequencies under consideration. It is very clear from these differences how is the ground plane resonates for the GSM-900MHz band more than the DCS-1800MHz. The resonant frequencies of both bands can be further tuned by varying the position of the feeding strip and the width of the shorting strip. Other parameters such as the lengths and the widths of the slots in the patch also alter the resonant frequency, as explained previously (in Chapter 3 section 3.3.2).

![Figure 6.1 Dual-band PIFA design configuration (Antenna #1)](image-url)
As stated in section 6.1, one of the aims of the SAR study is to investigate the relationship between the SAR and the radiation efficiency at the same frequencies. In order to achieve this, the feeding mechanism for the dual-band PIFA design (labelled as Antenna #1) in Figure 6.1 was modified by changing the shorting strip width. The modification produced another antenna (labelled as Antenna #2) exhibiting different efficiency values to those of Antenna #1, at the same frequency points. Prototypes of Antenna #1 and Antenna #2 have been fabricated and are shown in Figure 6.3.

![Figure 6.2 Simulated surface current distributions for the studied dual-band PIFA at: (a) 0.923GHz (b) 1.787GHz](image)

![Figure 6.3 Prototypes of dual-band PIFA designs: Antenna #1 and Antenna #2](image)
6.3.2 Dual-band PIFA Simulation and Experimental Results

The simulated and measured return losses for Antenna #1 and Antenna #2 are depicted in Figure 6.4. For Antenna #1 there is reasonable agreement between the simulation and measurement. The first resonance occurs at \( f_1 = 910\text{MHz} \), \( \text{BW}_{(-5\text{dB})} = 87.2\text{MHz} \) (9.5%) which covers the GSM-900MHz band. The second resonance occurs at \( f_2 = 1830\text{MHz} \) to cover the DCS-1800MHz band, but it is shifted slightly up by 30MHz, \( \text{BW}_{(-5\text{dB})} = 246\text{MHz} \) (13.2%). The measured efficiency was reasonable in both bands, as depicted in Figure 6.5 and Figure 6.6. Measurements indicated a maximum efficiency of 37% at 897.5MHz with 0.02dBi measured peak gain for the GSM-900MHz band. While for the DCS-1800MHz band maximum efficiency of 65% at 1803.75MHz with a peak gain 4.14dBi was measured. For Antenna #2, Figure 6.4 shows the simulated and measured return losses, where a good agreement can be noted. The location of the first band for Antenna #2 has been shifted by 15MHz to resonate at \( f_1 = 925\text{MHz} \), \( \text{BW}_{(-5\text{dB})} = 122.9\text{MHz} \) (13.1%). The other band has been shifted by 50MHz to resonate at \( f_2 = 1880\text{MHz} \), \( \text{BW}_{(-5\text{dB})} = 256\text{MHz} \) (13.8%). From Figure 6.5 and Figure 6.6 one will note that the efficiency of Antenna #2 has decreased through the GSM-900MHz band and partially across the DCS-1800MHz band, in sympathy with the altered resonance.

![Figure 6.4 Antenna #1 simulated and measured S11 in free space and in the vicinity of SAM phantom head](image-url)
Figure 6.5 Antenna #1 and Antenna #2 measured efficiency at GSM-900MHz

Figure 6.6 Antenna #1 and Antenna #2 measured efficiency at DCS-1800MHz
6.3.3 Dual-band PIFA SAR and Efficiency Simulation Results

The simulation study carried out by simulating both antennas (Antenna #1 and Antenna #2) at different distances from a four tissue phantom head using CST-MWS™. Different frequency points within the transmitting sub-bands of the GSM-900MHz and DCS-1800MHz frequency bands were used to investigate the SAR values. The electrical parameters of the phantom head tissue (skin, fat, bone and brain) are listed in Chapter 4. Figure 6.7 shows one of the antennas investigated. This antenna was located at a distance $d$ mm from the phantom head. The distance $d$ was measured between the flat surface on the left part of the phantom head and the bottom surface of the antenna’s ground plane. The following values for $d = 1, 9, 17, 25, 33$ and $41$ mm employed to carry out the SAR study.

Figure 6.7 Studied antenna at a distance $d$ from CST-MWS phantom head
The measured return loses for Antenna #1 and Antenna #2 before and after placing them in the vicinity of the SAM phantom are depicted in Figure 6.8. Comparing the results with/without the influence of the head, it can be observed that the antennas are slightly detuned due to the dielectric properties of the phantom head. For instance, the lower band of Antenna #1 has been shifted by 12MHz and the match improved by 4dB. While at the upper band, for the same antenna, the shift was 10MHz and the match decreased by 5dB. On the other hand, for Antenna #2 the lower band was shifted by 17MHz and the match was improved by 6dB. The upper band for this antenna remained fixed and the match was decreased by 23dB. There was minor degradation of the antennas under investigation bandwidth at the upper and lower frequency bands when the antenna was placed close to the phantom head, this may be attributed to losses in the tissue. The simulated and measured S\textsubscript{11} results agree well. It may thus be concluded that the presence of the phantom head has significant influences on the performance of a PIFA.

![Graph](image-url)
Comparing the simulated 3-D radiation patterns of the antennas under investigation, with and without the presence of the phantom head, revealed that there are also significant differences in the radiation patterns. The reason for these differences is that the head has absorbed a part of the radiated power from the antenna and the radiation became directive towards the head. Figure 6.9 (A), (B) shows the 3-D radiation pattern for Antenna #1 at 0.915GHz without and with the presence of the phantom head. The antenna was placed 9mm from the head, which implies that the distance between the antenna element and the head is 18.5mm. It can be noted that the pattern became more directive in the presence of the head and that the maximum directivity increased from 2.461dBi to 4.656dBi. Figure 6.10 (C), (D) shows the 3-D radiation pattern for Antenna #1 at 1.785GHz without and with the presence of the phantom head respectively. It is very clear from the figures that the pattern is more directive in the presence of the head. The maximum directivity increased from 3.369dBi to 5.341dBi.

For both frequencies a distinct null, in the radiation patterns near the head, was also observed. For the 0.915GHz frequency the null is angle away from the z-axis rather than directed towards the head. While for the 1.785GHz frequency, the null is aligned with the z-axis and points directly towards the head. An explanation for these nulls can be obtained from the behaviour of the antenna’s magnetic field distribution in the absence of the head, as shown in Figure 6.11. From Figure 6.11 (a) one will note that the magnetic field distribution at 0.915GHz is concentrated in one spot on the z-axis and smaller elsewhere. This is leading to the expectation that there is no null aligned with the z-axis at this frequency. On the other hand, at 1.785GHz it was noted that the magnetic field distribution has been split into two spots along the z-axis, as depicted in Figure 6.11 (b), and was very low in the middle region. This means that the magnetic field is very low between these two spots due to the presence of a null along the z-axis. Generally, it was noted that the magnetic field intensity at 0.915GHz is greater and more concentrated than at 1.785GHz. For Antenna #2 the radiation patterns and the magnetic field distributions were similar to those for Antenna #1, with the only difference being in the directivity and magnetic field values.
Figure 6.9 Simulated 3-D radiation pattern of Antenna #1 at 0.915GHz for: (A) without the phantom head, maximum directivity = 2.164dBi (B) with the phantom head at 9mm distance, maximum directivity = 4.656dBi

Figure 6.10 Simulated 3-D radiation pattern of Antenna #1 at 1.785GHz for: (C) without the phantom, maximum directivity = 3.369dBi (D) with the phantom head at 9mm distance, maximum directivity = 5.341dBi

Figure 6.11 Simulated magnetic field distributions for Antenna #1 without the head at:
(a) 0.915GHz  (b) 1.785GHz
The simulated average SAR (Wkg$^{-1}$) over 10g for the frequency points in the transmitting side of the GSM-900MHz are depicted in Figure 6.12 and Figure 6.13 for both antennas at different distances from the phantom head. The figures also show the simulated installed efficiency. For frequency points in the transmitting side of the DCS-1800MHz the simulated results are shown in Figure 6.14 and Figure 6.15. The results for both antennas showed that the SAR values decrease as the distance between the antenna and the phantom head increases. This inversely proportional relationship between SAR and the distance was expected since the specific heat of the phantom’s tissue decreases when the distance increases [4]. On the other hand, the simulated installed efficiency results showed there is a proportional relationship between the installed efficiency and the distance.

![Figure 6.12](image)

Figure 6.12 The simulated 10g average SAR for Antenna #1 at frequency points in the transmitting side of the GSM900 band versus distance between the antenna and the phantom head. Also the simulated installed efficiency versus distance.
Figure 6.13 The simulated 10g average SAR for Antenna #2 at frequency points in the transmitting side of the GSM900 band versus distance between the antenna and the phantom head. Also the simulated installed efficiency versus distance.

Figure 6.14 The simulated 10g average SAR for Antenna #1 at frequency points in the transmitting side of the DCS1800 band versus distance between the antenna and the phantom head. Also the simulated installed efficiency versus distance.
Figure 6.15 The simulated 10g average SAR for Antenna #2 at frequency points in the transmitting side of the DCS1800 band versus distance between the antenna and the phantom head. Also the simulated installed efficiency versus distance.

The simulated 10g averaged-volume SAR distributions inside the phantom head at 0.915GHz and 1.785GHz with 9mm distance between the antenna and the head are depicted in Figure 6.16, for both antennas. It was noted that the power distribution towards the head at the studied frequencies is consistence to the magnetic field distribution of the antenna in the absence of the head, as shown previously in Figure 6.11. The power distributions profile shown in Figure 6.16 (A), (B) for Antenna #1 and Antenna #2 respectively, indicate that the maximum power is radiated from the antenna at 0.915GHz. The power distributions profiles at 1.785GHz are shown in Figure 6.16 (C), (D) for Antenna #1 and Antenna #2. At 0.915GHz the power is mainly concentrated within a specific area, whilst at 1.785GHz a lower intensity of power is distributed over a wider area. It was noted also that the magnetic field strength values follow the trend in the distributed power values, where at 0.915GHz was 0.509Am⁻¹ and at 1.785GHz was 0.305 Am⁻¹ as shown previously in Figure 6.11. It can thus be generally concluded that the GSM-900MHz frequency points have a higher SAR than the DCS-1800MHz frequency points.
The relationship between the SAR values and the installed efficiency can be observed from Table 6-1 and Table 6-2. Table 6-1 shows the simulated installed efficiency values and the SAR values for Antenna #1 and Antenna #2 for the transmitting side of the GSM-900MHz and the DCS-1800MHz bands. The distance between the antenna and the head was 9mm. Comparing the efficiency and SAR values at the same frequency point for both antennas showed that the SAR values decrease when the efficiency values decrease and vice versa. This suggests a proportional relationship between the SAR and efficiency. Table 6-2 showed also a similar relationship between the SAR and efficiency, but for a different distance between the antenna and the head (25mm). Furthermore one should note that the SAR values meet the standard limit outlined in the ICNIRP guidelines, when the antenna was between 2mm to 9mm from the head.
### Table 6-1
Comparison between the simulated installed efficiency and the SAR values for Antenna #1 and Antenna #2 at the frequency points for the transmitting side of GSM-900MHz and DCS-1800MHz. The distance between the head and the antenna is 9mm.

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Antenna #1</th>
<th></th>
<th>Antenna #2</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Efficiency (%)</td>
<td>SAR [10g] (Wkg⁻¹)</td>
<td>Efficiency (%)</td>
<td>SAR [10g] (Wkg⁻¹)</td>
</tr>
<tr>
<td>880</td>
<td>27.87</td>
<td>1.39</td>
<td>17.66</td>
<td>0.86</td>
</tr>
<tr>
<td>897.5</td>
<td>32.45</td>
<td>1.64</td>
<td>24.8</td>
<td>1.23</td>
</tr>
<tr>
<td>915</td>
<td>33.85</td>
<td>1.74</td>
<td>31.99</td>
<td>1.61</td>
</tr>
<tr>
<td>1710</td>
<td>51.74</td>
<td>1.29</td>
<td>43.92</td>
<td>1.06</td>
</tr>
<tr>
<td>1747.5</td>
<td>49.26</td>
<td>1.19</td>
<td>53.38</td>
<td>1.24</td>
</tr>
<tr>
<td>1785</td>
<td>43.25</td>
<td>1.02</td>
<td>54.32</td>
<td>1.21</td>
</tr>
</tbody>
</table>

### Table 6-2
Comparison between the simulated installed efficiency and the SAR values for Antenna #1 and Antenna #2 at the frequency points for the transmitting side of GSM-900MHz and DCS-1800MHz. The distance between the head and the antenna is 25mm.

<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Antenna #1</th>
<th></th>
<th>Antenna #2</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Efficiency (%)</td>
<td>SAR [10g] (Wkg⁻¹)</td>
<td>Efficiency (%)</td>
<td>SAR [10g] (Wkg⁻¹)</td>
</tr>
<tr>
<td>880</td>
<td>48.01</td>
<td>0.59</td>
<td>30.66</td>
<td>0.37</td>
</tr>
<tr>
<td>897.5</td>
<td>55.60</td>
<td>0.69</td>
<td>43.73</td>
<td>0.53</td>
</tr>
<tr>
<td>915</td>
<td>57.18</td>
<td>0.72</td>
<td>55.94</td>
<td>0.70</td>
</tr>
<tr>
<td>1710</td>
<td>37.21</td>
<td>0.33</td>
<td>62.16</td>
<td>0.41</td>
</tr>
<tr>
<td>1747.5</td>
<td>69.36</td>
<td>0.37</td>
<td>74.37</td>
<td>0.40</td>
</tr>
<tr>
<td>1785</td>
<td>60.47</td>
<td>0.31</td>
<td>75.10</td>
<td>0.38</td>
</tr>
</tbody>
</table>

Simulated values of average SAR (Wkg⁻¹) over 1g for frequency points in the transmitting side of the GSM-900MHz/DCS-1800MHz together with simulated installed efficiency values provided in Appendix II. A similar pattern is observed to those for the average SAR (Wkg⁻¹) over 10g and
supported the same relationship between the SAR and both the distance and the efficiency. It was also discovered that the average SAR (Wkg$^{-1}$) over 1g results meet the ANSI/IEEE standard when the distance between the antenna and the head is greater than 9mm.

### 6.3.4 Dual-band PIFA SAR Measurement Results

SAR measurements have been applied to Antenna #1 and Antenna #2 using the DASY4 system, described in Chapter 4. Briefly, the DASY4 system measures the induced electric field using an electric field probe. The probe scans tissue equivalent materials with appropriate permittivity representing the human head (SAM phantom head) by moving to selected locations within the phantom. The head simulation liquid parameters for the GSM-900MHz frequency points are: dielectric constant $\varepsilon_r = 41.28$ and conductivity $\sigma = 0.96$Sm$^{-1}$. For the DCS-1800 MHz frequency points the head simulation liquid parameters are: dielectric constant $\varepsilon_r = 40.48$ and conductivity $\sigma = 1.37$Sm$^{-1}$.

The handset antenna was located on the left hand side of the SAM phantom in the standard talk position during ordinary use. When measuring the SAR for GSM-900MHz the standard transmitted power is 250mW. Therefore, when calculating the SAR the power had to be normalized to 250mW. For DCS-1800MHz the standard transmitted power is 125mW.

For frequency points in the transmitting side of the GSM-900MHz/DCS-1800MHz the measured average SAR (Wkg$^{-1}$) over 10g for both antennas at different distances from the phantom head are depicted in Figure 6.17 to Figure 6.20. The measured results showed similar behaviour to the simulated results. Again an inversely proportional relationship has been noted between the SAR values and the distance between the antenna and the head. For each of the different distances and frequencies investigated the SAR values met the standard limit outlined in the ICNIRP guidelines. Unfortunately the facility to measure installed efficiency was unavailable hence its absence from these results.
Figure 6.17 The measured 10g average SAR for Antenna #1 at frequency points in the transmitting side of the GSM900 band versus distance between the antenna and the phantom head.

Figure 6.18 The measured 10g average SAR for Antenna #2 at frequency points in the transmitting side of the GSM900 band versus distance between the antenna and the phantom head.
Figure 6.19 The measured 10g average SAR for Antenna #1 at frequency points in the transmitting side of the DCS1800 band versus distance between the antenna and the phantom head.

Figure 6.20 The measured 10g average SAR for Antenna #2 at frequency points in the transmitting side of the DCS1800 band versus distance between the antenna and the phantom head.
Figure 6.21 shows the power distributions of the measured 10g volume averaged SAR for Antenna #1 and Antenna #2 (at 0.915GHz and 1.785GHz) when the antenna located 9mm from the phantom head. The figure shows that there is only one SAR maximum at 0.915GHz for both of the antennas under investigation, while at 1.785GHz there is an additional SAR maximum. The reason for this will be explained in section 6.3.5.

Measured SAR (Wkg\(^{-1}\)) values averaged over 1g for frequency points in the transmitting side of the GSM-900MHz/DCS-1800MHz are provided in Appendix II. The results showed similar behaviour to those shown for the SAR (Wkg\(^{-1}\)) averaged over 10g and indicate the same relationship between the SAR and the distance. It was noted that the SAR (Wkg\(^{-1}\)) averaged over 1g for the DCS-1800MHz frequency points meet the ANSI/IEEE standard limit for each of the different distances investigated. While for the GSM-900MHz the distance between the antenna element and the head should be greater than 10mm in order to meet the standard limit.

Unfortunately, it is not possible to compare the measured SAR results with the simulated ones. This is due to a difference between the method of simulation (using CST-MWS) and measurement (using DASY4 system). The first difference lies in the electrical properties of the phantom head equivalent materials, used to simulate a human head. The phantom head used for simulations was a simple four-tissue phantom that does not need a long period of time to simulate and huge computing resources in terms of memory issue. The other reason for the difference in the results is that the induced electric field is a complex function of several physical and biological variables, which includes the microwave frequency, the source size and polarisation, the tissue type, composition, and geometry, as well as its orientation [4]. Generally, the accuracy and reliability of simulation and measurements results are sensitive to the models used to represent the user handset combination and to the parameters assumed. The measured SAR results were reasonable compared to SAR measurements applied to PIFAs reported in the literature [7-9].
6.3.5 Magnetic Field Behaviour in the Presences of Phantom Head

During the SAR simulation and measurement for both of the antennas under investigation, the following have been observed. Firstly, the SAR values for the GSM-900MHz frequency points are always higher than those for the DCS-1800MHz frequency points. Secondly, an additional SAR maximum was noted during measurement for the DCS-1800MHz frequency points compared to the GSM-900MHz frequency points, as shown in Figure 6.21. Section 6.3.3 explained these observations with reference to the magnetic field distribution of the antenna in the absence of the head (Figure 6.11). Figure 6.16 shows the power distribution towards the head. In this section the observed points will be explained again by analysing the magnetic field H distributions, but this time inside the phantom head.
Figure 6.22 and Figure 6.23 show the simulated magnetic field distributions towards the head for Antenna #1 and Antenna #2 respectively, at 0.915GHz with a 9mm separation between the head and the antenna. While Figure 6.24 and Figure 6.25 show the simulated magnetic field distributions towards the head for Antenna #1 and Antenna #2 respectively, at 1.785GHz with a 9mm separation between the head and the antenna. The figures showed that the magnetic field of the low frequency points is stronger than the one for the high ones. For example, the maximum magnetic field values for Antenna #1 and Antenna #2 are 28.5Am⁻¹, 27.5Am⁻¹ at 0.915GHz. While at 1.785GHz the values drop to 17.4 Am⁻¹, 21.6 Am⁻¹. This increased strength of magnetic field at the low frequency points explains the higher level of the associated SAR values compared to the high frequency points SAR values.

Since the SAR power distribution is related to the magnetic field, which is travelling perpendicularly inside the phantom head. Then the fact of the presence of more than one SAR maximum can be explained with reference to the magnetic field distributions showed in Figure 6.22 to Figure 6.25. The magnetic field distributions showed that the distribution of the magnetic field for the lower frequencies of both antennas travels in one concentrated spot inside the phantom head. While the magnetic field for the high frequency points travels inside the head as two separate spots approximately. This behaviour of the magnetic field for the high frequency points explains the existence of the additional SAR maximum observed in the SAR power distribution measurements in Figure 6.21.

Finally, for both studied antennas it should be clear that the antenna element (top plate) and the ground plane drive power towards the phantom head. This is very clear in the simulated magnetic field distributions graphs.
Figure 6.22 Simulated magnetic field H distribution for Antenna #1 at 0.915GHz with 9mm distance from the head

Figure 6.23 Simulated magnetic field H distribution for Antenna #2 at 0.915GHz with 9mm distance from the phantom head
Figure 6.24 Simulated magnetic field H distribution for Antenna #1 at 1.785GHz with 9mm distance from the phantom head

Figure 6.25 Simulated magnetic field H distribution for Antenna #2 at 1.785GHz with 9mm distance from the phantom head
6.4 Conclusion

The characteristics of SAR distribution in a human head exposed to electromagnetic radiation from a dual-band PIFA operating at GSM-900MHz/DCS-1800MHz communication frequency bands were investigated. The SAR was investigated with the aid of two PIFAs, which differ only in their feeding mechanism (the shorting plate width). These PIFAs were located at different distances from a phantom head for the uplink band frequencies.

The results indicate that the introduction of a phantom head deteriorates the performance of the PIFA. The resonant frequencies were detuned by about 10MHz. This detuning led to a slight variation in the bandwidth of the frequency bands considered, together with impedance match of 5dB improvement within the lower band and more than 6dB of degradation in the upper band.

As expected, the simulated and measured results showed that there is an inversely proportional relationship between the SAR and the separation distance of the dual-band PIFA handset antenna from the head. In addition, the results show also that there is a proportional relationship between the antenna efficiency and the separation distance between the antenna and the head. On the other hand, there was a proportional relationship between SAR and the antenna efficiency. The results also show that the SAR for the GSM-900MHz frequencies is higher than the SAR for DCS-1800MHz frequencies. This is due to the increased strength of the magnetic field, which travels toward the head, at the GSM-900MHz compared to the DCS-1800MHz.

Finally, it has been noted that the PIFAs under investigation met the safety limits stipulated by the ICNIRP guidelines. While exceeding the ANSI/IEEE safety guidelines and the optimum distance between the antenna and the head to meet this standard is between 10mm and 15mm. It is also expected that the values will drop to meet the ANSI/IEEE standard once the antenna embedded in a real mobile handset.
References


Chapter 7

7 Conclusions and Future Work

7.1 Conclusions

The research described in this thesis was motivated by the increasing need for compact, low cost, wideband and multi-band antennas having acceptable SAR to sustain the rapid growth in the mobile communications market. The design and analysis of multi-band PIFA structures for use in mobile communications was the main theme of this thesis.

The thesis began by introducing the reader to a GA/CST-MWS optimisation technique employing a real-valued representation. This GA code was developed by the author together with a visual basic interface with CST-MWS. The utility of the technique was first tested by optimising a dipole antenna as a proof of concept. The technique was subsequently applied to a PIFA handset antenna. The presented optimisation technique proved to be an excellent tool for reducing the time duration of design cycle of PIFAs. By providing an assessment for the performance effect due to combinations of parameters, the optimisation technique convergence histories led to better understanding of the effect of the PIFA’s geometric parameters. Consequently, different multi-band PIFAs were constructed by applying parametric studies to single or dual-band PIFAs. Four different multi-band PIFA geometries were investigated in this thesis. The SAR parameter for PIFAs together with their relationship to the antenna’s efficiency and distance from the user’s head were also investigated.

The author’s original work achieved in the duration of the research includes; a real-valued GA/CST-MWS optimisation technique suitable for optimising handset antennas modelled in CST-MWS simulation package. The
technique showed the ability to achieve increasingly complex antenna designs by optimising a dipole antenna and a more complex structure such as the PIFA handset antenna. The GA/CST-MWS technique succeeded in optimising the dipole to resonate at the required target frequency of 900MHz and the bandwidth at 10dB return loss was improved by 2%. The technique was also fed by a single-band PIFA after slotting its radiating planar element to resonate at two frequency bands and then optimised to achieve a dual-band PIFA handset design suitable for personal communications at GSM-900MHz/DCS-1800MHz. Different runs of the GA/CST-MWS showed that the antenna's feeding probe position, radiating element height, the shorting plate width and dimensions of slots in the radiating element play a key role in defining the PIFA's performance. The advantages of the GA/CST-MWS optimiser were clear in evaluating quickly whether the selected parameters are appropriate for optimisation and what is possible with antenna structures. The optimiser relieves the researchers from tedious and speculative trials of optimisation by hand (manually), allowing being faster, more creative and more effective in achieving antenna designs for the future mobile communications.

The optimised dual-band PIFA using the GA/CST-MWS technique was modified and extended to produce a triple-band PIFA, which meets the bandwidth requirements of the DCS-1800MHz/PCS-1900MHz/UMTS cellular systems. The antenna has small dimensions (36mm wide by 16mm long by 5.9mm thick, occupied volume = 3.39cm³) and can easy fit inside a modern mobile handset. The antenna also has a rigid structure and relatively low profile compared to triple-band PIFAs reported recently in the literature [1-6]. Parametric studies have been applied to the triple-band PIFA to investigate the effect of the antenna’s shorting pin radius, position and the slots within the radiating planar element on resonant frequency and bandwidth. The results of these parametric studies were useful in improving the antenna performance and can be used as guidance for designing PIFAs in the future. The antenna under investigation was tested inside an ABS plastic case, and a slight degradation was noted in the gain and efficiency values compared to those observed for the antenna operating in the free space. For further
characterisation, the SAR distributions in a head exposed to electromagnetic radiation from the studied triple-band PIFA under investigation have been determined through simulation and measurement. The results showed that the antenna’s measured and simulated SAR values meet the 2W/kg for 10g volume-averaged SAR standard.

The triple-band PIFA was extended to produce a quad-band PIFA handset design. The quad-band PIFA operates within the GSM-900MHz/DCS-1800MHz/PCS-1900MHz/UMTS communication bands. The new antenna design was produced by extending the triple-band PIFA radiating element using a planar meander-line monopole. This modification enabled operation within the GSM-900MHz frequency band. The antenna occupies a volume of 5.92cm³. The simulation and measurements were in a good agreement and the antenna showed a good performance within the bands of interest. The antenna’s volume and structure still requires optimisation, however to fit inside modern small mobile handsets. Future work should include a study of techniques for reducing size of the entire antenna. This may perhaps be achieved by folding the mender-line monopole into a compact structure [7-9] to end up with a small, low profile quad-band mobile phone antenna. The resulting quad-band design would be a combination of PIFA type antenna and a folded compact monopole antenna.

A novel penta-band PIFA antenna was also presented in this thesis. The antenna was placed 6mm above a ground plane measuring 120mm long by 38mm wide. The antenna meets the bandwidth requirements of the AMPS, GSM-900MHz, DCS-1800MHz, PCS-1900MHz and UMTS cellular bands. The antenna’s radiating element was a slotted plate, which occupies 20% (38mm wide by 25mm long) of the total ground plane area and placed 6mm above the top surface of the ground plane to occupy a volumetric size of 5.4cm³. Generally, the antenna’s dimensions are acceptable for integration within a real mobile handset operating at five standard communication frequency bands. The dimensions of the penta-band PIFA are small compared with those of PIFAs reported recently in the literature [10-15]. In [11-13] the widths of the reported antennas exceeded 40mm and in [10,11,14,15] the
antennas height exceeded 8mm. Some of these antenna structures were also very complex, which makes them very difficult to fabricate. For example, the reported antenna designs in [10,11,14] comprise two layers of antenna elements and loaded with many parasites. Simulation and measurement results for the studied penta-band PIFA were in good agreement for the multi-band behaviour of the antenna. SAR measurements showed that the antenna meets the 2W/kg for 10g volume-averaged SAR standard. During the antenna design stages, many modifications and parametric studies were applied to the antenna’s ground plane. It was concluded that the ground plane of a PIFA plays a vital role controlling in the performance of the antenna because slight variations in the phases of the flowing currents on the ground plane lead to a substantial change in the antenna performance.

The last study undertaken in this thesis concerned the relationship between the SAR parameter and the PIFA’s installed efficiency and distance from the user’s head. The characteristics of the SAR distributions in a human head exposed to electromagnetic radiation from a dual-band PIFA operating within the GSM-900MHz and DCS-1800MHz communication bands were investigated. The simulation and measurement results showed that there is an inversely proportional relationship between the SAR and the separation distance of the dual-band PIFA handset antenna from the user’s head. The results also showed that there is a proportional relationship between the antenna efficiency and the separation distance between the antenna and the head. There was a proportional relationship between SAR and the antenna efficiency. Generally, the SAR values for the PIFAs investigated during this thesis showed three important facts. The first fact is that the presented PIFAs meet the safety limits stipulated by the ICNIRP guidelines (2W/kg for 10g volume-averaged SAR). The second fact is that the presented PIFAs should be positioned between 10mm and 15mm from the user’s head to meet the more stringent ANSI/IEEE safety guidelines (1.6W/kg for 1g volume-averaged SAR). The last fact is that the SAR values for the studied PIFAs at GSM-900MHz band frequencies are higher than at the DCS-1800MHz band frequencies.
Finally, the PIFAs investigated during this research showed that the PIFA is the most appropriate type of antenna that meets the current design constraints (limitation of design space) on antenna designs for small handheld devices such as mobile phones and personal digital assistants (PDAs). Generally, it was noted that the PIFA has the features of compactness, moderate bandwidth, high gain for both states of polarisation, and low SAR.

7.2 Recommendations and Future Work

The work and results that have been achieved in this thesis can be directed in two phases for future work. Phase one, some work remains to be done in increasing the GA/CST-MWS optimisation efficiency and effectiveness. Phase two, several possible avenues for future research in reducing the PIFA size and improving its performance can be applied to the PIFAs investigated during this thesis or future designs.

7.2.1 Increasing GA/CST-MWS Efficiency and Effectiveness

In spite of the fact that the presented GA/CST-MWS optimisation technique is not sophisticated enough to cope with the increasing complexity of small multi-band antennas, it is still a promising tool that may be enhances developing the technique to meet more demanding applications in the future. Many extensions and improvements reported in the literature can be applied to this optimisation technique to increase its robustness and ability to optimise more complex antenna structures, such as multi-band and ultra-wideband (UWB) antennas. Future work on the presented GA/CST-MWS can thus be summarised by applying the following ideas and concepts:

- Multiple Populations (Parallel GAs): In the literature, the use of multiple populations has been demonstrated, in most cases, to improve the quality of the results obtained using GAs compared to those derived a single population GA [16]. In a multiple population model, the population is divided into several subpopulations that are assigned
Conclusions and Future Work

Chapter 7

to different processors. Each processor applies a traditional simple GA independently to its own subpopulation. Occasionally, individuals are exchanged between subpopulations, in a process called migration [17].

- Multi-objective GA: Multi-objective or multi-criteria optimisation could be used instead of single objective optimisation when several objectives are simultaneously present and it is not possible to combine them into a single number [18-20]. In the case of antenna optimisation, many objectives can be added to the evaluation function, which is the key function responsible for procreation in a typical GA. Among these objectives are the gain, bandwidth, radiation efficiency, antenna size, input impedance and specific absorption rate (SAR). Multi-objective optimisation implements a high fidelity fitness evaluation function and improves the GA’s performance, but increase the simulation and the run time.

- Queen-bee Evolution for GA: A new evolution scheme for enhancing the performance of GAs recently has been introduced. This scheme, termed queen-bee evolution [21], is similar in its behaviour to nature in that the queen-bee (the fittest individual in a generation) crossbreeds with the other bees selected by a selection algorithm as parents. This increases the exploitation of GAs. However, it also increases the probability that the GA will converge prematurely, resulting in degradation in the GA’s performance. In order to reduce the probability of premature convergence, some individuals in queen-bee evolution are strongly mutated. This reinforces the exploration of GAs [21].

- Micro genetic algorithm (MGA): It was noted that the general choice of population size for the presented conventional GA/CST-MWS optimisation technique could range from 30 to 100. Such a large population size is the reason for the long computational run time of the technique. Unfortunately the use of a smaller population size for
conventional GAs leads to poor performance due to insufficient information for processing and reduces convergence speed [18,22]. Recently, a numerous number of researchers have investigated the MGA extensively with the aim of avoiding aforementioned problems associated with using small population size. The key strategy of MGAs for achieving small population size is that in a consequence way replaces the whole population (except the best individual) with a new population, once the old population converges. The MGA normally operates with a very small population size (usually 6 to 20) for each generation and reaches near-optimal regions faster than the conventional GAs, that deal with a large population size.

7.2.2 Reducing PIFA Size and Improving Performance

As shown through this thesis, the requirement for mobile handset antenna designs for mobile communications is to develop small-size, lightweight, wide bandwidth and highly integrated antennas in the handset. The PIFA type antenna, with its finite metallic ground plane has been used to produce the antennas investigated throughout this research. The PIFA facilitates simple impedance matching in low-profile designs. Although the PIFA has been widely used as a mobile communication antenna, it is still requires reduction in size and broadening in bandwidth. As a result of making the antenna smaller the bandwidth becomes lower, resulting in a higher overall loss in the antenna performance. To reduce the size of the antenna without incurring penalty, two concepts can be considered as a future work that can be applied to the PIFAs produced out of this thesis.

The first concept is to modify the PIFA’s ground plane, which affects the characteristics of the antenna significantly, using the idea of high-impedance surface (photonic band gap type (PBG-Type) or artificial magnetic conductor (AMC-Type)) ground plane [23-26]. PBG materials, also referred as electromagnetic bandgap (EBG) structures, are periodic structures capable of prohibiting the propagation of electromagnetic waves within certain directions and frequency bands [23]. With the development of EBGs, high-
impedance electromagnetic surfaces have been proposed [23,24]. The high-impedance surface helps to suppress the surface waves on the ground plane of the antennas, resulting in high radiation efficiency with little backward radiation. The high-impedance surface is thus a strong candidate for designing low profile, high efficiency antennas. These valuable features of high impedance surfaces suggest designing a PIFA with a PBG-type ground plane to reduce the size of the PIFA. Recently, it has been shown in the literature [25,26] that employing high-impedance surface in PIFA's ground plane reduces antenna size and surface waves, thus leading to an increase in directivity, forward and backward radiation ratios and efficiency. The results reported in [25,26] motivate on doing further research and investigations on the proposed concept. The employment of high-impedance surface as a means of suppressing the surface waves in mobile handsets can be also investigated to find the possibility of improving the antenna performance by reducing the coupling effect between the human hand/head and the antenna. Another important parameter, which could be investigated using the high-impedance surface, is the SAR since this structure allows control over the backward radiation pattern toward the mobile handset user's head.

The second concept is to design PIFAs with switchable slots [27,28]. The switches could be implemented using pin diodes, varactor diodes, microelectromechanical systems (MEMS) switches or small mechanical switches. This would enable one to change the antenna’s resonance frequency and bandwidth needed for the system used at any given time by a much smaller antenna. Switches have been applied in the literature to dipole antennas, patch antennas and microstrip antennas. Compared to normal designs, they exhibit attractive features of compact structure, similar radiation pattern and little co-site interference [28]. Therefore, the idea of using switches in the PIFA’s structure should be investigated during a future phase of research, with the aim of achieving further reduction in the PIFA’s size, especially, for multi-band behaviour.
References


Appendices
Appendix I

GA/CST-MWS Front-End and Reproduction Operations

I.1 Introduction

The procedures for designing the software of the GA/CSTM-WS front-ends, shown in Chapters 2, 3, for the dipole and single-band PIFA optimisation examples are described in this appendix. The appendix also describes the reproduction operations used to build the GA/CST-MWS structure together with their mathematical foundation.

I.2 Front-End Design

The procedure described below can be applied to any model built in CST-MWS to create a front-end suitable for the GA/CST-MWS optimisation engine. Once an initial design of a dipole modelled in CST-MWS using constant values for the parameters describe the geometry dimension obtained. A dipole front-end can be programmed in Visual Basic language using a macro extracted from the CST-MWS history list for that initial design. The macro language is used for the automation of common tasks in CST-MWS. To build the dipole front-end, the initial design with constant values of the parameters is required only. The structure modelling operations provided by CST-MWS are used to modify the structure of the model and need to be stored in the history list showed in Figure 1.

Usually these kinds of macros are defined to extend the built in primitives by some often used structure elements. The most convenient way to produce such a macro is to open the History List (shown in Figure 1) for the initial dipole model and selecting the corresponding operations from the list and pressing the Macro button. In the next dialog box, a name should be assigned e.g. 'Dipole Geometry Macro's' to the macro and decides whether the
macro shall be locally or globally available, as depicted in Figure 1. After this, the macro can be edited in the integrated development environment by usage of the Edit macros dialog box. Finally, the macro can be executed from the Macros menu and the current contents of the macro will then be stored in the History List.

![Dipole Geometry Macro](image)

Figure 1 History list and macro for a dipole antenna modelled in CST-MWS

After the initial geometry modelled in CST-MWS and the macro obtained, geometries with different parameters values can be produced by changing their constant values in the history list. This is however an impractical technique to forming a population of different individuals from the initial geometry design for the GA to process them for optimising. This means to go through the history list each time a new model with different parameters values required. To overcome this complexity, the initial geometry’s macro has been investigated, rewritten to interface with the GA code and the constant
parameters values, that describe the geometry, have been replaced by variable parameters.

The second step in building the dipole's front-end was to identify the dipole's physical parameters that control its electric and magnetic behaviour. It was noted that the dipole antenna geometry could be described using the following parameters: Dipole Outer Radius, Dipole Length and Dipole Air Gap. In addition to these parameters, there are other parameters that are also required for simulation purposes, these include: Start Frequency, Stop Frequency and Monitor Frequency. The next step was to rewrite the dipole geometry macro edited from the history list using Visual Basic and replaces the parameter constant values with variable ones that accept any value within a range defined by the user. The final layout of the dipole antenna front-end shown in Chapter 2.

1.3 Reproduction Operations

1.3.1 Fitness Operation Using Linear Scaling Function

The linear scaling method described by Goldberg [1], Samii et al. [2] and Chipperfield et al. [3] requires a linear relationship between the Fitness Value and the Objective Value of the suggested chromosome as described in equation (1):

\[
\text{Fitness Value} = (A \times \text{Objective Value}) + B
\]  

(1)

Where A is a positive scaling factor if the optimisation is maximising and negative if the optimisation is minimising. B is a factor used to ensure that the resulting fitness values are non-negative. Both coefficients (A and B) in equation (1) can be determined using the following method:
Oavg = Average Objective Value = \frac{\text{sum(ObjectiveValues)}}{\text{Number of Individuals}}

Omin = \text{minimum(Objective Value)}

Omax = \text{maximum(Objective Value)}

Sp = \text{Optional Scaling Parameter, usually by default 2}

If(Omin > ((Sp \times Oavg) - Omax) / (Sp - 1))Then

\quad \text{Delta} = Omax - Oavg

\quad A = ((Sp - 1) \times Oavg) / \text{Delta}

\quad B = (Oavg \times (Omax - (Sp \times Oavg))) / \text{Delta}

Else

\quad \text{Delta} = Oavg - Omin

\quad A = Oavg / \text{Delta}

\quad B = - (Omin \times Oavg) / \text{Delta}

End

1.3.2 Selection Operation Using Roulette Wheel Selection

The roulette wheel selection (RWS) has been described by [3,4] in an algorithmic style. RWS mechanism probabilistically selects chromosomes based on some measure of their performance. In brief, the mechanism starts with summing the fitness values of all population chromosomes produced by the preceding reproduction operation. If this sum is called Fitness\text{sum} then a real-valued interval could be determined as [0, Fitness\text{sum}]. Within this interval a random number is generated to return the first population chromosome whose fitness, added to the fitness of the preceding population individuals, is greater than or equal to the randomly generated number.

Figure 2 illustrates RWS graphically, where the size of each chromosome interval corresponds to the fitness value of the associated chromosome. The circumference of the Roulette Wheel is the sum of the chromosomes fitness values (Fitness\text{sum}). Chromosome 5 is the fittest as it occupies the largest interval on the circumference, whilst chromosomes 4 and 6 are the least fit and so have the smallest intervals. To select a chromosome a random number is generated and the chromosome whose segment spans the
random number will be selected. This procedure will be repeated until a sufficient number of chromosomes are selected.

![Figure 2 Roulette Wheel Selection method [3]](image)

**I.3.3 Crossover Operation Using Intermediate Recombination**

Intermediate recombination is a method only applicable to real-valued variables. Using this function the variable values of the offspring are chosen to lie between the variable values of the parents [3-6]. Offspring are produced according to the following equation:

\[
\text{Offspring} = \text{Parent}_1 + \text{Alpha}(\text{Parent}_2 - \text{Parent}_1)
\]  

(2)

Where *Alpha* is a scaling factor chosen uniformly at random over an interval \([-d,1+d]\). The value of the parameter \(d\) defines the size of the area for possible offspring. A value of \(d = 0\) defines the area for offspring the same size as the area spanned by the parents. This method is called (standard) intermediate recombination. Because most variables of the offspring are not generated on the border of the possible area, the area for the variables shrinks over the generations. This shrinkage occurs just by using (standard) intermediate
recombination. Using a larger value for $d$ can prevent this effect. A value of $d = 0.25$ ensures (statistically), that the variable area of the offspring is the same as the variable area spanned by the variables of the parents [5]. Each variable in the offspring is the result of combining the variables according to the equation (2) with a new Alpha chosen for each variable. Figure 3 shows the area of the variable range of the offspring defined by the variables of the parents.

For example, if two selected chromosomes named Parent 1 and Parent 2 produced by the selection operation RWS (each chromosome consists of 3 parameters, the first and the second parameters made variables while the third one left constant):

- Parent 1: [7.3, 159.7, 8.3]
- Parent 2: [2.8, 147.2, 8.3]

And the randomly chosen Alphas for this example are:

- Factor 1: [0.986, -0.213, NotNeeded]
- Factor 2: [1.235, 0.711, NotNeeded]

Then the new individuals are calculated as:
Intermediate recombination is capable of producing any point within a hypercube slightly larger than that defined by the parents. Figure 4 shows the possible area of offspring after intermediate recombination.

![Figure 4 Possible areas of the offspring after Intermediate recombination][6]

### 1.3.4 Real-Valued Mutation Operator of the Breeder GA

Mutation operation is a background operator that provides a guarantee that the probability of searching any given chromosome will never be zero and acts as a safety net to recover good genetic material that may be lost through the action of selection and crossover operations. Mutation operation specifies how the GA makes small random changes in the individuals in the population to create mutation children. Mutation provides genetic diversity and enables the genetic algorithm to search a broader space by the addition of small random values (size of the mutation step), with low probability. The probability of mutating a variable is set to be inversely proportional to the number of variables (dimensions) per chromosome ($n$):

\[
\text{Mutation Probability} = \frac{1}{n}
\]  

(3)
The more dimensions one individual has the smaller the mutation probability [1,6]. Mühlenbein [7] found that a mutation rate of $1/n$ produce good results for a broad class of test functions. With real-valued representation, mutation is achieved by a random selection of new values within an allowed range. Michalewicz [8] explained how real-valued GA might take advantage of higher mutation rates than a binary format GA, increasing the level of possible exploration of the search space without adversely affecting the convergence characteristics. Many articles have reported results for the optimal mutation rate [7, 9]. The mathematics behind the mutation operator of the breeder GA is defined by [3,6,7] using the following equation:

$$\text{Mutated Variable} = \text{Variable} \pm (\text{Range} \times \text{Delta})$$

$$\text{Range} = 0.5 \times \text{Variable Domain}, \quad \text{Search Interval}$$

$$\text{Delta} = \text{sum}(a(i) \times 2^i)$$

where $a(i) = \begin{cases} 1 & \text{for mutation, with probability } \frac{1}{\text{Accur}}, \quad \text{Accur} = 20 \\ 0 & \text{no mutation} \end{cases}$

This mutation algorithm is able to generate most points in the hypercube defined by the variables of the individual and range of the mutation. Figure 5 shows possible mutations for a real valued chromosome in two dimensions. However, it tests more often near the variable, that is, the probability of small step sizes is greater than that of bigger steps. The parameter Accur (mutation accuracy) indirectly defines the minimal step-size possible and the distribution of mutation steps inside the mutation range. The smallest relative mutation step-size is $2^{-\text{Accur}}$ and the largest is $2^0 = 1$. With Accur equal to 20, the mutation algorithm is able to locate the optimum up to a precision of $(\text{Range} \cdot 2^{-19})$. Thus, the mutation steps are created inside the area $[\text{Range}, \text{Range} \cdot 2^{-\text{Accur}}]$. Typical values for the parameter Accur are: {4,5...20} (see [6] for more details).
1.3.5 Reinsertion Operation Using Random Reinsertion

Different reinsertion schemes have been reported in the literature by [3,6,10]. One of these schemes is called the uniform reinsertion. This technique produces less offspring than parents and replace parents uniformly at random. Thus, for a chromosome to survive successive generations, it must be sufficiently fit to ensure propagation into future generations. Another popular technique is the fitness-based reinsertion, which replaces the least fit parents relying on their fitness values. The fitness based reinsertion scheme implements a truncation selection between offspring before inserting them into the population (i.e. before they can participate in the reproduction process).
Appendix II

Simulated and Measured Average SAR Over 1g for Antenna #1 and Antenna#2

Figure 6 The simulated 1g average SAR for Antenna #1 at frequency points in the transmitting side of the GSM900 band versus distance between the antenna and the phantom head. Also the simulated installed efficiency versus distance.

Figure 7 The simulated 1g average SAR for Antenna #2 at frequency points in the transmitting side of the GSM900 band versus distance between the antenna and the phantom head. Also the simulated installed efficiency versus distance.
Figure 8 The simulated 1g average SAR for Antenna #1 at frequency points in the transmitting side of the DCS1800 band versus distance between the antenna and the phantom head. Also the simulated installed efficiency versus distance.

Figure 9 The simulated 1g average SAR for Antenna #2 at frequency points in the transmitting side of the DCS1800 band versus distance between the antenna and the phantom head. Also the simulated installed efficiency versus distance.
Figure 10 The measured 1g average SAR for Antenna #1 at frequency points in the transmitting side of the GSM900 band versus distance between the antenna and the phantom head.

Figure 11 The measured 1g average SAR for Antenna #2 at frequency points in the transmitting side of the GSM900 band versus distance between the antenna and the phantom head.
Figure 12 The measured 1g average SAR for Antenna #1 at frequency points in the transmitting side of the DCS1800 band versus distance between the antenna and the phantom head.

Figure 13 The measured 1g average SAR for Antenna #2 at frequency points in the transmitting side of the DCS1800 band versus distance between the antenna and the phantom head.
References


