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MODELLING AND DESIGN OF COMPACT WIDEBAND AND ULTRA-WIDEBAND ANTENNAS FOR WIRELESS COMMUNICATIONS

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UNIVERSITY OF BRADFORD

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MODELLING AND DESIGN OF COMPACT WIDEBAND AND ULTRA-WIDEBAND ANTENNAS FOR WIRELESS COMMUNICATIONS

Simulation and Measurement of Planar Inverted F Antennas (PIFAs) for contemporary mobile terminal applications, and investigations of Frequency Range and Radiation Performance of UWB Antennas with Design Optimisation using Parametric Studies

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Abstract

MODELLING AND DESIGN OF COMPACT WIDEBAND AND ULTRA- WIDEBAND ANTENNAS FOR WIRELESS COMMUNICATIONS

Simulation and Measurement of Planar Inverted F Antennas (PIFAs) for contemporary mobile terminal applications, and investigations of Frequency Range and Radiation Performance of UWB Antennas with Design Optimisation using Parametric Studies

Hmeda Ibrahim Hraga

Keywords

Ultra-Wideband (UWB); Global System for Mobile Communications (GSM); Wireless Local Area Network (WLAN); Antennas; Planar Inverted F Antenna (PIFA); Multiple Inputs Multiple Outputs (MIMO); Radiation Pattern; Pattern Diversity; Envelope Correlation Coefficient; Capacity Loss

The rapidly growing demand for UWB as high data rates wireless communications technology, since the Federal Communications Commission (FCC) allocated the bandwidth of UWB from 3.1GHz to 10.6 GHz. Antenna also plays an essential role in UWB system. However, there are some difficulties in designing UWB antenna as compared to narrowband antenna. The primary requirement of UWB antennas is be able to operate over frequencies released by the FCC. Moreover, the satisfaction of radiation properties and good time domain performance over the entire frequency range are also necessary.

In this thesis, designing and analysing printed crescent shape monopole antenna, Planar Inverted F-L Antenna (PIFLA) and Planar Inverted FF Antenna (PIFFA) are focused. A Planar Inverted FF Antenna (PIFFA) can be created to reduce the potential for interference between a UWB system and other communications protocols by using spiral slot.

The antennas exhibits broadside directional pattern. The performances such as return loss, radiation pattern and current distribution of the UWB antennas are extensively investigated and carried out. All the results have been demonstrated using simulation and experimentally whereby all results satisfy the performance under -10dB point in the bandwidth of UWB.

In addition the miniaturization of MIMO/diversity Planar Inverted-F antenna (PIFA) which is suitable for pattern diversity in UWB applications is presented. This antenna assembly is formed by two identical PIFAs, a T-shaped decoupling structure which connects the two PIFAs and a finite ground plane with a total compact envelope dimension of 50 × 90 × 7.5mm³. The radiation performance of the proposed MIMO antenna was quite encouraging and provided an acceptable agreement between the computed and measured envelope correlation coefficient and channel capacity loss.
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Acronyms

AWGN   Additive White Gaussian Noise
CST    Computer Simulation Technology
DCS    Digital Cellular System
EM     Electro Magnetic
FCC    Federal Communications Commission
GSM    Global System Mobile
HFSS   High Frequency Structure Simulator
IEEE   Institute of Electrical And Electronics Engineers
LPD    Low Probability of Detection
LPI    Low Probability of Intercept
LTCC   Low temperature co-fired ceramic
PC     Personal Computers
PCB    Printed Circuit Board
PCS    Personal Communication System
PIFFA  Planar Inverted F-F Antenna
PIFLA  Planar Inverted F-L Antenna
RF     Radio Frequency
SNR    Signal-to-Noise Ratio
UMTS   Universal Mobile Telecommunication System
UWB    Ultra -Wideband
VSWR   Voltage Standing Wave Ratio
Wi-Fi  Wireless Fidelity
WiMax  Worldwide Interoperability for Microwave Access
WLAN  Wireless Local Area Network
WPAN  Wireless Personal Area Network
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CHAPTER 1

Introduction

1.1 Background History and Motivation

With the advent of the information era, numerous advanced communication technologies have arisen during the past two decades, which, have greatly influenced and benefited every field of human society. The first-generation (1G) mobile communication technology only enabled analogue voice communication while the second-generation (2G) technology achieved digital voice communication. Currently, the third-generation (3G) technology can offer a wide range of high speed mobile services, including video calling, internet access, messaging, e-mail and information services. In the near future, the fourth-generation (4G) technology will be able to provide a comprehensive IP (Internet Protocol) solution where voice, data and streamed multimedia can be offered to users anytime and anywhere, and at higher data rate than former generations.

In the past few years, wireless personal area network (WPAN) has been attracting considerable interest and undergoing rapid development worldwide. A WPAN is a network for interconnecting devices around an individual person’s workspace in which the connections are wireless. WPAN aims to achieve seamless operation among home
or business devices and systems. In addition, fast data storage and exchange among these devices will also be realised. This demands a data rate which is much higher than what has been achieved in currently existing wireless technologies.

The maximally available data rate, or capacity, for the ideal band-limited additive white Gaussian noise (AWGN) channel is linked with the bandwidth and signal-to-noise ratio (SNR) by Shannon-Nyquist criterion [1, 2], as given in equation 1.1.

\[
C = B \log_2 (1 + SNR)
\]  

(1.1)

Where \(C\) represents the maximum transmission data rate, \(B\) denotes the channel bandwidth.

Equation 1.1 illustrates that the transmission data rate can be raised by enlarging the bandwidth or amplifying the transmission power. Nevertheless, the signal power can’t be easily increased as many portable devices are powered by battery and the potential interference with other radio systems should be also suppressed. Therefore, a large frequency bandwidth will be the solution to realise a high data rate.

In 2002, the United States Federal Communications Commission (FCC) adopted the First Report and Order that validated the commercial operation of ultra wideband (UWB) technology [3]. Since then, UWB technology has been swiftly evolving as one of the most promising wireless technologies that provide the high bandwidth required
by the latest and future portable home and office devices for multiple digital video and audio streams [4, 5].

The UWB technology has undergone remarkable achievements during the past few years. In spite of all the promising prospects featured by UWB, there are still challenges in making this technology fulfil its full potential. One particular challenge is the UWB antenna. In recent years, many varieties of UWB antennas have been proposed and investigated. They present a simple structure and UWB characteristics with nearly Omni-directional radiation patterns. However, for some space-limited applications, UWB antennas need to feature a compact size while maintaining UWB characteristics.

UWB monopole antennas are planar structures [6-8] which require a ground plane, creating problems for integration with an integrated circuit. This drawback limits its practical application. However, one of the serious limitations of conventional microstrip antennas is their narrow bandwidth, which is normally only a few percent of the centre frequency. Therefore, printed planar monopole antennas may provide the candidate for its low cost, light weight and easy fabrication. Planar monopole antennas of various configurations have been proposed and offer different attractive feature [9, 10].

Planar inverted-F antennas (PIFAs) are widely used in mobile terminals, these antennas can be modified to operate in a wide single band and multiple bands, by using parasitic elements and slotting the antenna element to form several paths for surface currents.
PIFA structures capable of covering six telecommunication standards have been reported [11].

In this thesis, three types of new antennas were investigated.

A low profile multi-frequency band monopole antenna module has been presented. The radiator is a crescent shaped microstrip patch. The realised antenna structure shows a relative bandwidth of 51.8%, and a gain of 1.5dBi over the frequency range from 1.75 to 3.1GHz. The antenna size was optimised at 57mm×37.5mm×0.8mm, and is suitable for integration with a variety of mobile terminals operating over DCS, PCS, UMTS, Bluetooth or IEEE 802.11b/g wireless standards.

A wideband planar inverted F-L antenna (PIFLA) has been designed and studied. The design optimisation has found an objective compromise between the size constraints of the antenna assembly and the desired impedance bandwidth performance, with respect to the lower operating UWB frequency spectrum. The prototype package dimensions are 30×15×4 mm³.

A compact low profile antenna assembly with a finite ground plane is proposed and realised. This antenna constructively combines the resonant modes of the two PIFA elements. The prototype performance demonstrates sufficient impedance bandwidth, suitable radiation characteristics and gain stability for UWB radio applications.
1.2 Aims and Objectives of the Current Research Work

This thesis is concerned with the development of an UWB antenna for UWB communication systems and evaluating its performance. This work aims to perform basic research on improving the design and deployment of UWB antenna that are compliant with the published FCC regulations for UWB. The primary goal of the present work is to design such antennas that can cover the range of UWB with a reflection coefficient of $|S_{11}| < -10$dB.

The antenna performances have been evaluated to meet the specific wideband targets after attempting several optimising procedures of various antenna parameters. Two design approaches were considered in this present work. The first is a modified version of PIFAs and the second is a diffracted ground to alter the wideband performances. Both designs show wide bandwidth to cover the lower and upper UWB bands.

A new design concept was introduced for PIFA antennas to reject the WLAN within the wideband antenna operation by introducing spiral slot on feeding plate of the PIFAs elements. A parametric study was performed to check and validate the design concept in terms the gain and group delay.

A novel miniaturized UWB MIMO/diversity Planar Inverted-F antenna (PIFA) suitable for pattern diversity applications is designed and developed over a finite ground plane. The new MIMO antenna was able to operate on UWB from 3.1 to 10.6 GHz and
provide acceptable envelope correlation coefficient and channel capacity loss including other radiation performances.

1.3 Summary of Main Targets

This work aims to perform basic research on improving the design and deployment of UWB antennas that are compliant with the published FCC regulations for UWB.

- To design antennas that can cover the UWB range with a reflection coefficient of $|S_{11}| < -10\text{dB}$.
- To model and analyse a range of small, compact and low-profile printed and PIFA antennas suitable for integration with mobile handsets.
- To introduce a new design concept for PIFA antennas to reject the WLAN band within the wideband antenna bandwidth by introducing a spiral slot on feeding plate of the PIFAs elements.
- To implement and test several antenna prototypes and compare their result against simulations.
- To conduct parametric studies of several antenna designs including parasitic loading in order to optimise for best radiation performance.

In general, achieving a small antenna size is the main target and the real design challenge of this current research and to work to develop future UWB technologies which minimise interference.
The principle novel contribution of this work is thus to provide new designs for UWB antennas that are more compact and have better performance than current designs.

1.4 Scope of the Work

In brief, this thesis is divided into eleven chapters. Each chapter discuss the different issues which are related to this work. The outline of each chapter is stated in the paragraphs below.

Chapter one covers the introduction and overview of the thesis background, objectives, scope of work and methodology to carry out this thesis.

Chapter two presents a basic discussion of wireless antennas and its classification. Basic characteristics of UWB antenna along with parametric study and improved is studied. Various other design examples are also given pertaining to UWB antenna as to how ideas and concept are important.

Chapter three provides an explanation on the design procedure and antenna design of UWB printed antenna. The return loss and radiation pattern are compared between the simulation and measurement results. Apart from that, the current distribution is study by simulations. The figures of the prototype antenna are attached.
Chapter four presents the results and discussions of the Planar Inverted F-L Antenna (PIFLA) for Lower-band UWB Applications. The proposed antenna design, geometry structure and specification are presented. The simulation, fabrication process, and measurement stage are explained. In addition a miniature planar inverted F-L antenna (PIFLA) has been investigated experimentally and theoretically in this chapter. By balancing the size and bandwidth constraints, these prototype antennas have a compact envelope dimension of 30mm × 15mm × 8mm, and cover the required operating frequency band for WLAN and UWB applications.

Chapter five investigates a wideband plane inverted F-F antenna with minimum occupied dimension of 30 × 15 × 3 mm³. The simulated and measured results in terms of the return loss, radiation pattern, power gain and group delay, show that the proposed antenna is a good candidate for lower UWB spectrum and IEEE 802.11a WLAN applications.

Chapter six discusses the design of dual-planar inverted F-antenna for WLAN/WiFi, WiMAX and the lower band UWB wireless standard. The antenna has shown the return loss, consistent omni-directional radiation patterns and reasonable gain values across the operating bands. The proposed antenna is very well suited to multi-band wireless applications.

Chapter seven presents a planar inverted FF antenna (PIFFA) assembly. This antenna design electromagnetically (EM) couples two PIFA antennas over an air substrate. The simulation, fabrication process, and measurement stage are explained. Secondly, Ultra-
Wideband Planar Inverted FF Antenna with band-rejection using Spiral Slots is proposed. This antenna can be applied for UWB communication systems to avoid interference with other wireless communication systems.

Chapter eight investigates an UWB antenna assembly with an inbuilt HIPERLAN/2 band-rejection characteristic. The HIPERLAN/2 rejected band centred at 5.35 GHz was selected for this application. The band-stop response was achieved by embedding slot-spiral resonator on the broadband feed plate. This design allows the antenna to operate over the full UWB spectrum including HYPERLAN/2, without the need for additional specialized filtering. The antenna has demonstrated sufficient impedance bandwidth, suitable radiation characteristics, and adequate gain for UWB applications.

Chapter nine introduces a compact and low profile modified PIFA antenna. This antenna combines impedance matching techniques including top-loading, off-centre rectangular plate feeding and a shorting wall to achieve an ultra-wide impedance bandwidth from 3 GHz to 13 GHz. Despite exhibiting this superior impedance bandwidth, the antenna only occupies a relatively small dimensional envelope of 50 x 50 x 7.5 mm$^3$.

Chapter ten presents a miniaturized MIMO/diversity Planar Inverted-F antenna (PIFA) which is suitable for pattern diversity in UWB applications. This antenna assembly is formed by two identical PIFAs, a T-shaped decoupling structure which connects the two PIFAs and a finite ground plane. The compact envelope dimension of this antenna is 50 x 90 x 7.5mm$^3$. Theoretical and experimental S-parameters are illustrated for this
antenna that fully covers an operating frequency band of 3.1-10.6GHz for UWB applications, at a reflection and acceptable agreement is also obtained between computed and measured radiation patterns, gains, envelope correlation coefficient and channel capacity loss.

Chapter eleven presents the interim conclusions of this thesis, and considers recommendations for future work in order to enhance the performance of UWB antenna.
1.4 References


CHAPTER 2

Literature Review

2.1 Ultra-Wideband (UWB) Technology

UWB technology is a promising solution for short range wireless communications which provide high data rates, at the expense of low power consumption. Since FCC released the spectral emission masks for commercial ultra-wideband (UWB) applications in 2002, there has been a great deal of activity in this field, but there is little agreement in practice about the creation of a fully functional standard. In general UWB applications tend to be generic, or highly specific, in the latter case, the whole system design must be taken into account, even where the antenna is seen as the principal design aim. Ideally UWB antennas must be engineered for compact terminals, or unobtrusive location, frequently in the presence of one or more overlapping channels. The main body of this report considers UWB technology in the context of in-door wireless communications. Wider technological issues, within the scope of UWB, wall-imaging systems, through-wall imaging systems, medical systems, surveillance systems, vehicular radar systems and communications and measurements systems.

A typical UWB system will handle pulse data over a large spectrum at lower power, the main characteristic being the low power spectral density, which works by generating a large number of small pulses which are transmitted across a wide spectrum. The
receiver translates these pulses into useful data by scanning for a familiar pulse sequence. This is especially useful for short ranged, high definition content transfer in a WPAN [1]. Furthermore, the low power consumption makes it feasible for low operating costs; the UWB technology is an excellent choice to address the high-speed WPAN market.

The spectrum allocation for UWB is in the range from 1.99 GHz- 10.6 GHz, 3.1 GHz-10.6 GHz, or below 960 MHz depending on the particular application [2].

### 2.2 Various Technologies for UWB

Meaningful UWB technologies have been available for nearly 30 years; during the late 1980s, the technology originated from the baseband, carrier-free or impulse technology. It was given the term “ultra-wideband” in 1989 by the U.S Department of Defence, commercial uses have been more recent [3].

Commercial UWB operates under the legal frequency range from 3.1 GHz to 10.6 GHz and transmission power will be no more than -41 dBm/MHz, This FCC sponsored rule change allows the use and subsequent study of commercial ultra-wideband communications, and is intended to provide an efficient use of scarce radio bandwidth while enabling both high data rate personal-area network (PAN) wireless connectivity and longer-range, low data rate applications. In the meantime, Intel is working closely with the governments to promote and facilitate UWB regulations [3]. Other companies
have also participating in the development of UWB; companies include, U.S. Radar Inc., Time Domain Inc and Zircon Corporation among others. According to the FCC’s ruling, any signal that occupies at least 500MHz spectrum can be used in UWB systems. That means UWB is not restricted to impulse radio any more, it also applies to any technology that uses 500MHz spectrum and complies with all other requirements for UWB.

UWB signals occupy extremely large bandwidths where the RF (Radio Frequency) energy is spread over a comparatively large spectrum. It is wider than any incumbent narrowband wireless system by orders of magnitude and its emitted power seen by other narrowband systems is a fraction of their own power. If the whole 7.5 GHz band is optimally utilised, the maximum achievable power to UWB transmitters is approximately 0.556 mW or less. This is barely a fraction of available transmit power in the industrial, scientific and medical (ISM) bands such as the WLAN (Wireless Local Area Network) IEEE (Institute of Electrical and Electronics Engineers) 802.11a/b/g standards. This effectively confines the UWB scheme to indoor and short-range communications at high data rate or mid-range communications at low data rate. Applications such as wireless USB and WPANs have been proposed with hundreds of Mbps to several Gbps with distances ranging from 1 to 4 metres. For ranges beyond 20 metres, the achievable data rate by UWB is inferior to existing WLAN systems such as IEEE 802.11a/b/g [4].
UWB antenna designs present a higher challenge than most conventional antennas for wireless designers. UWB antenna behaviour and performance should be consistent across the bandwidth for efficient radiation of UWB signals. Successive UWB pulse transmission and reception entail minimization of spreading and distortion of the pulse which require the antenna characteristics of linear phase, low return loss, constant directivity and constant group delay. Wireless antenna can designed for high performance for extremely small bandwidth applications. These types of antennas fall under narrow band classification. These Wideband antennas possess tremendous bandwidth capability covering multiple octaves.

UWB antenna operation is further subdivided into two sub-bands depending on the bandwidth of operation; lower band and upper band, lower band (3.1 - 5.1) GHz, upper band (5.85 - 10.6) GHz for the indoor and handheld applications.

2.3 UWB Classification

UWB exhibits three major advantages which are High Data Rate, Low Power and Multipath Immunity. UWB antennas do not require high power; the power usage specification states that the power supply requirement is equivalent to that of a battery power appliance. It provides very good bandwidth for high throughput with low power and is more in demand over other technologies also because of its low production costs. UWB power transmission pulses are of very low density which makes it very difficult to
intercept as they are very much comparable to background noises. UWB uses -41dBm/MHz where other services are considerably high. UWB utilises wide bandwidth covering RF spectrum and SS consuming less energy as seen in below diagram. Very high data rates are achievable using UWB therefore making it a more viable solution in a multi user environment pairing with multiple devices using WPAN and 802.15.3 PAN standards. UWB has a very high range resolution making it possible to use over circumstances where it needs penetrating through obstacles [5].

Wireless antennas can be broadly classified depending on their radiation pattern and bandwidth. UWB antenna can be classified as omni-directional and directional [6]. Omni-directional antennas, as the name suggests has its radiation pattern distributed in all directions while directional antennas have concentrated radiation patterns thus having a high gain for longer distances compared to the forms.

This project work is focused on omni-directional UWB antennas, although some research interest is growing on directional ultra-wideband antennas. A directional antenna focuses on energy into a narrow solid angle compared with an omni-directional antenna. In general, the size of a directional antenna is large as compared to the omni-directional case.
2.4 UWB Characteristics

Initially UWB technology has been mainly used for radar-based applications [7] due to wideband nature of the signal resulting in very accurate timing information. UWB technology has also been of considerable interest with respect low probability of intercept (LPI) and detection (LPD), multipath immunity, high data throughput, precision ranging and localization. Multipath propagation is one of the most significant obstacles when radio frequency (RF) techniques are used indoors.

Since UWB waveforms are of such short time duration, they are relatively immune to multipath degradation effects as observed in mobile and in-building environments. Thus, UWB has gained recent attention and has been identified as a possible solution to a wide range of RF problems. In communication systems, UWB pulses can be used to provide extremely high data rate performance in multi-user network applications. Additionally, UWB applications can co-exist with narrowband services over the same [8].

2.5 Methodology and Technical Concepts

The types of antennas were taken into consideration include monopoles, planar and spiral antennas. It is known that when using time domain techniques, larger frequency range means better range resolution [9] and naturally, UWB requires a broadband antenna. However this would require the planar antenna to be infinite in order to fulfil
the self-scaling and self-complementary conditions. The rationale for pursuing planar designs include, the topology chosen to achieve broadband impedance matching has a compact size, low pulse and low profile distortion upon transmission and reception [10], the planar antenna is much easier to fabricate compared to other types of the antennas.

2.6 Previous Work Done on UWB Antennas

This section will highlight a few particularly noteworthy designs that have implemented the UWB antenna designs.

Bandwidth enhancement is a major issue in today’s fast evolving and demand communication system [11]. The design presented in [12] is based on a monopole printed on a ground plane of width of 30 x 15 mm². The substrate dimensions are 30 x 35 mm² and the circular patch radius R=7.5mm is printed on FR4 board with thickness 2mm and relative permittivity 4.4. Introducing a slot on the patch enhances the bandwidth. The optimal bandwidth performance indicates a rectangular slot of 0.466R.

Planar UWB antenna size constraints are investigated in [13]. This approach employs a high-dielectric-constant substrate material, which allows for a considerable antenna size reduction. Simulations were performed using HFSS for antennas assuming DuPont-951 (εᵣ = 7.8) and RT6010LM (εᵣ = 10.2) substrates. For the 1-mm-thick DuPont-951, the antenna dimensions were 22 × 28 mm, with a 10-dB return-loss bandwidth from 2.7 GHz to more than 15 GHz. For the 0.64-mm-thick RT6010LM prototype, the
dimensions were $20 \times 26$ mm, and the antenna exhibits a 10-dB return loss bandwidth from 3.1 to 15 GHz. Both antennas feature roughly omni-directional properties across the whole 10-dB return-loss bandwidth. Practical measurements taken on this antenna design reflects that of in simulation using RT6010LM substrate. The proposed radiating element in this design is formed by the intersection of two ellipses, while the feeding structure is a coplanar waveguide. The designed antennas using this method have an extremely wide 10-dB return loss bandwidth from 3.1-15 GHz and posses almost 90% efficiency. Rectangular patch designs with increased bandwidth are reported in [14], this was achieved by using a truncated rectangular patch as the feeding element; the substrate thickness and permittivity were 0.813mm and 3.38 respectively. A rectangular slot of width 40mm and length 27mm is introduced. The result demonstrate an omni-directional radiation pattern for more than of 146% (2 GHz-12.8GHz) covering the FCC band of ultra-wideband making it more suitable for UWB antenna system.

The design of compact components for UWB front ends frequently involves novel hybrid circuits and antennas for ultra wideband [15]. This work has reported on the design of front end of UWB communication system using compact planar components, such as couplers, dividers and antennas, Novel configurations of these devices accompanied by simple design formulas have been described. The proposed devices have been studied in terms of their return loss, coupling, isolation and radiation patterns, and support of distortion less pulse transmission. This project uses standard CAD tools and experimental measurements to achieve its requirements. All of the presented components have shown excellent UWB performances despite of its compact size. All
these attributes to attract more designers of UWB communication sub-systems, it is worthwhile to note that all of the presented designs use multi-layer dielectric substrates with electric properties similar to those of typical Low temperature co-fired ceramic (LTCC) materials. Therefore, it is apparent that the presented components can easily be developed in the LTCC technology, which is the present focus in the design of frontends of wireless transceivers.

A new UWB antenna operating over 3.2-12 GHz was proposed in [16]; the design employs rectangular patches, and a partial ground plane, and displays good impedance matching. The antenna has the following parameters: 35mm length, 30mm width printed on substrate FR4 of 1.6mm thickness and 4.4 permittivity, the results were obtained using Microwave Studio, and show reasonable agreement with the measured results. Return losses of -10 dB are achieved in the BW of 3.2-12 GHz. The group delay was tested and its variation is less than 0.5 ns. The measured radiation patterns at 3, 5, and 7 GHz and antenna gain in the 3–10-GHz frequency range were also presented.

[17] Presents a technique that can be used to optimise the performance of a diamond dipole antenna (DIDA) for ultra-wideband applications. The antenna has a bandwidth of 11.15 GHz range from 2.57 to 13.72 GHz, voltage standing wave ratio VSWR of less than 2. Omni-directional radiation pattern are produced.

A compact UWB monopole for multilayer applications has been presented in [18]. It was fabricated using the RT Duroid substrate with dielectric constant of 2.33. This
multilayer structure consists of two layers, of thickness 1.575 mm, with metal etched on the faces of each substrate. The rectangular shape of the ground plane was produced by etching the top and bottom metal clad of the top and bottom layer respectively. The antenna was fed by a 50 Ω shielded strip-line. The strip-line ground planes were terminated in the radiation region above and below the monopole disc which was located between the two substrate layers. The side walls of the shielded strip transmission line were fabricated using two arrays of metallised vias, diameter 0.5 mm, pitch 1 mm. The separation between the two via arrays was 7.5 mm. The total dimensions of the antenna were 20×35.5 mm. The ultra-wideband, monopole characteristics of the antenna are proved experimentally.

A design procedure using planar electric and magnetic monopole is presented [19]. Design equations and methods are explained for this type of radiator, a compared vs. analysis using HFSS. In this design, LTCC in the form of 1mm thick DuPont951 material is assumed as the antenna substrate. The results show that using this substrate, very compact UWB antennas can be developed and the simulations indicate high accuracy in the design method. The prototype antennas feature near omni-directional characteristics and good radiation efficiency.

A compact antenna with CPW-fed is described in [20]. It has dimensions of 15.5 × 17 mm², printed on one side of FR4 substrate with thickness 1.6 mm, and permittivity of 4.4. A sunken ground plane and narrow feeding-strip are introduced; a slot notched on the centre of radiator is used to fine-tune the impedance matching at higher frequency.
part. The optimal dimensions can be achieved through parametric analysis. The antenna operates over 3.08-10.6 GHz for VSWR better than 2. The antenna also displays an omni-directional radiation pattern and a good gain flatness that is suitable for UWB applications.

A clover-shaped antenna (3.1-10.6 GHz) was designed and fabricated in [21]. To obtain sufficient bandwidth with return loss better than -10 dB, a partial ground plane and coaxial probe source is used. The measured bandwidth of the fabricated antenna is 8.25 GHz and VSWR is less than 2. Referenced to the centre frequency, the gain is 3.20-4.00 dBi. The experimental 3-dB beam width (HPBW) in the azimuth and elevation are 55.35° and 62.27°, respectively.

A new planar circular disc-and-ring monopole antenna for dual-network applications was reported in [22]. The first band is specified for the universal mobile telecommunication system (UMTS, 1.92–2.17 GHz), and the second is for the FCC defined UWB system (3.1–10.6 GHz). Sufficient impedance bandwidth and omni-directional radiation performance in both bands have been observed and it shows interesting prospects for UMTS/UWB dual-network applications. A prototype for the UMTS/UWB dual-network applications has been built and demonstrated. As the measured and simulated results show, not only does it show omni-directional radiation characteristics but also the impedance bandwidth performance is suitable for the specified dual-network applications.
Another planar monopole design is presented in [23]. This antenna has excellent impedance bandwidth over the range from 1.926 to 4.9 GHz. The impedance bandwidth reaches 123.9% (10-dB return loss) for 2.4 GHz, which meets the required bandwidth specifications of IEEE 802.11 b/g. Adding a parasitic element and a ground plane that does not cover the whole of the other side of the patch is proposed to improve impedance bandwidth.

UWB planar microstrip-fed monopole antenna is studied in [24]. The design uses FR4 substrate, with a relative dielectric constant of 4.4, and thickness of 1.6 mm. The results show that the proposed antenna has dimensions of 43 mm × 50 mm, and a bandwidth from 2.85 GHz to more than 15 GHz. The bandwidth covers the whole of UWB commercial band from 3.1 to 10.6 GHz. By tuning the feed gap height between the ground plane and feed point, wideband performance was presented with the measured return-loss response of the designed antenna covering 2.85 GHz to more than 15 GHz. Moreover, the antenna features omni-directional characteristics and is suitable for UWB applications. The peak gain is 2.32–3.5 dBi in the operating range.

Dual-frequency planar inverted F-L-antenna (PIFLA) for WLAN and short range communication systems is studied in [25]. The design and analysis is presented of a low profile and dual frequency inverted L-F antenna for WLAN and short range wireless communications, providing a compromise between size reduction and attainable bandwidth. By balancing the size and bandwidth constraints, the proposed antenna has a compact envelope dimension of 30 mm × 30 mm × 8 mm and covers the required
operating frequency band for the IEEE 802.11a/b/g, Bluetooth and ZigBee standards. These features make the proposed antenna an attractive candidate for application in a range of mobile terminals.

2.7 Bandwidth Enhancement Techniques

It is desirable to find methods that can enhance the bandwidths of the PIFA antennas. Bandwidth is a crucial parameter in UWB system. Several methods exist to increase the antenna bandwidth. In order to improve the impedance bandwidth for UWB applications, several bandwidth enhancement techniques have been reported, such as the use of an asymmetrical feed arrangement [26], adjust the gap between radiating element and ground plane [27], a double feed [28], a bevelling radiating element [29], a bevelling ground pattern [30], parasitic metal plate places in parallel with the feeding plate [31], and so on. The bandwidth enhancement approaches mentioned above focus on making modifications on the radiating elements of PIFA antennas. There are also other bandwidth improvement methods concentrating on making modifications on the system ground plane of mobile terminals. For example, T-shaped ground plane [32], slotted ground plane for handset devices [33,34]. The bandwidth of the PIFA antenna design also enhances by using quarter-wavelength wave traps [35].
These antennas are either required large area for the feed circuitry or bevelling, and others are designed with radiating element erected vertically to the ground plane. The most common technique for planar antennas is to cut a slot on the patch, additionally, parasitic patches can enhance the impedance bandwidth, but this is not usually sufficient to cover the whole UWB band. Thicker substrates can also be used; this is expensive and awkward for most practical applications. Parameters in the design of a microstrip patch UWB antenna such as feed line position, feed line width dimensions ground of plane can all be optimised to increase the bandwidth [36]. Finally, multipath effects can create a major threat to any communication system. But as UWB system possess high data rate and short pulse range multipath effect is not predominant.
### 2.8 Reference


Correlation Detection” Microwave Symposium, IEEE/MTT-S International, 02 July 2007 page(s) 1471 – 1474.


CHAPTER 3

Small Wideband Antenna for GSM and WLAN

Application

3.1 Introduction

Multi-standard wireless communications systems combining high speed transmission rates for indoor and outdoor data communications, e.g. GSM, WLAN and UWB, offer the possibility of greater freedom and convenience in connecting various devices working over a number of narrow and broadband frequency bands, the UWB band being used for short range indoor services [1]. Eventually, a single broadband antenna design may be required for each device. Interesting design candidates include monopole antennas [2-8].

Monopole antennas offer low profile geometries, wide bandwidth, and ease of integration with planar RF circuitry [2]. Printed monopole antennas are usually fed by a microstrip network; various geometries such as circular, square and rectangular patches have been examined [3-7]. The ground plane is an important element in the monopole antenna design process, in which size vs. achievable bandwidth at the antenna input port is the most crucial part [8]. In this chapter we examine the possible bandwidth
expansion mechanism for a monopole antenna covering 3.1 GHz to 10.6 GHz with a return loss better than -10 dB across the band. A new multiband crescent-shaped microstrip fed antenna is simulated, built and tested. The antenna covers most of the existing wireless frequency band allocations from 1.7GHz to 3.1GHz.

Figure 3-1 Basic antenna structure; (a) top view, (b) bottom view
The antenna is mounted on top of a defected ground plane. A 50 Ω microstrip line is used for feeding purposes. The performance of the proposed antenna was analysed and optimised to cover the target bandwidth. The equivalent terminal size is 57×37.5×0.8 mm³. The performance of this antenna was fully characterised in terms of the reflection coefficient, radiation pattern and power gain. The calculated and measured results for the reflection coefficient show close agreement. The simulated gains and radiation patterns are given to fully describe the performance of this antenna.

3.2 Antenna Design Concept

The antenna geometry is given in Figure 3-1 above. The maximum dimension of this structure is approximately 0.68λ₀, computed at the lower frequency of the DCS band i.e. 1.7 GHz. The radiator is fed by an 18mm×1mm 50 Ω microstrip line. FR4 epoxy is used as the substrate throughout the module, with a thickness of 0.8mm, and the dielectric constant (εᵣ) is assumed to be uniformly 4.4, with a loss tangent of 0.017 over the target frequency range. A defected ground is located on one side of the substrate, and is truncated close to a point where the feed is coupled to the radiator.
Figure 3-2: Simulated input returns losses corresponding to the variation of parameters: (a) $r_1$; (b) $r_2$; (c) ground plane size

The underlying design principle is based on the manipulation of multiple resonances, tuned by the modification of the radiator’s geometry. It should be noted that the radiator is constructed from the sections of two circles; each having a different radius and centre, thus enabling the resultant patch, taken with the effect of the coupling to the defected ground plane, to radiate over two different frequency bands. The larger radius controls the fundamental frequency, whilst the shorter radius may be tuned to obtain the desired upper frequency. It should be noted that the variations of both resonances are subject to optimum positioning and size of the defected ground plane.
When two suitable modes are constructively merged, the desired wideband impedance bandwidth is obtained. The defected ground structure plays a significant role in improving the impedance matching across the target bands, but the actual ground plane dimensions are minimized as far as possible to fit the space available inside a transceiver casing.

![Current distribution images](image.png)

Figure: 3-3: The current distribution at (a) 1.8GHz, (b) 2.1GHz, (c) 2.4GHz and (d) 2.8GHz
Figure 3-4: Prototype antenna

Figure 3-5: Measured and simulated reflection coefficient $|S_{11}|$
Figure 3-6: Measured and Simulated gains of the proposed antenna

In Figure 3-2: the parameters $r_1$ and $r_2$ were varied as part of the design optimisation, using a commercial Conformal-FDTD program (SEMCAD) [9]. Initially a parametric study was made, varying each parameter by 1mm increments, whilst holding the remaining parameters at their initial values. These values were chosen arbitrarily to fit within the required envelope size; the values for $r_1$ and $r_2$ were found to be 12mm and 10mm, respectively.

It can be seen, in Figure 3-2 (a), that the fundamental resonant mode moves down in frequency as the length of $r_1$ is increased, but this does not impair the higher resonant mode. A similar pattern is observed for $r_2$, in that decreasing its length produces a slight upwards shift in frequency, but leaves the fundamental mode unaffected, as depicted in
The best performance fit over 1.7GHz to 2.5GHz is $r_1 = 11$ mm and $r_2 = 9$ mm. The effect of the ground plane was monitored by checking the variations in the input return loss against the size of the ground plane. Focusing on the fundamental mode at 1.7GHz, four different ground plane sizes were considered: 21×37.5mm$^2$ ($0.25\lambda_0 \times 0.385\lambda_0$), 42×42 mm$^2$ ($0.5\lambda_0 \times 0.5\lambda_0$), 57×37.5mm$^2$ ($0.68\lambda_0 \times 0.385\lambda_0$), and 84×84mm$^2$ ($\lambda_0 \times \lambda_0$). This is summarised in Figure 3-2 (c), where it can be seen that increasing the ground plane size from 21×37.5 mm$^2$ to 57×37.5 mm$^2$ significantly enhances the wideband performance of the impedance bandwidth, any further elongation diminishes the impedance bandwidth.

An investigation of the effect of the surface currents over the ground plane was also carried out. The computed surface currents for different operating frequencies were shown in Figure 3-3. These currents at 1.8, 2.1, 2.4 and 2.8 GHz are almost negligible over most of the ground plane except around the feeding antenna port.

### 3.3 Result and Discussion

The prototype test antenna is shown in Figure 3-4. Figure 3-5 illustrates the typical measured and computed antenna performance in terms of the impedance bandwidth. It may be clearly seen that the two adjacent resonant frequencies in the range are 1.7GHz and 3.1GHz (with $|S11| \leq -10$ dB). It is worth noting that this prototype’s impedance
bandwidth is 1.4GHz, or equivalently 58.3% with respect to the centre frequency at 2.4GHz. This provides adequate coverage for DCS, PCS, UMTS and IEEE 802.11b/g Standards.

Figure 3-7: Simulated and measured normalised radiation patterns of the proposed antenna for three planes ((a): x-z plane, (b): y-z plane, (c): x-y plane) at (i) 2100 MHz (ii) 2800 MHz xxxx’ simulated cross-polarization ‘oooo’ simulated co-polarization ‘----’ measured cross-polarization ‘——’ measured co-polarization
Figure 3-6 plots the simulated antenna gain in the broadside direction for several frequencies across the GSM1800, GSM1900, UMTS and 2.45 GHz bands. The overall simulated gain variations are less than 1.2 dBi. The antenna exhibits a good range of gain values in which the maximum and minimum measured gains were found between 2.85 dBi and 1.65 dBi over the aggregate bandwidth. The computed far field radiation patterns are depicted in Figure 3-7, three pattern cuts (i.e. the xz, yz and xy planes) were taken at two selected operating frequencies, i.e. 2100MHz and 2800MHz.

It is hard to state an objective criterion for optimising the pattern, since it depends on achieving a good compromise in the orientation within the assembly, and moreover, there is no completely satisfactory analysis of what constitutes a “good” antenna pattern under these conditions. Nevertheless it seems reasonably obvious that a good pattern should have its peak radiation in a near-horizontal direction, and not at a very high or very low elevation, when the set is operated in its most usual position with its axis at about 45° from horizontal.

It can also be said that a good compromise design target is to seek a pattern which has elevation directionality similar to that of a simple dipole; to ensure that a peak occurs at zero elevation; and to seek a design which radiates mainly omni directionally.

It can be seen that the azimuthally peak gain (the co-polar component shown in Figure 3-7(c)) varies over the range from 0 to 6 dB at 2100MHz, except at the null (approx. 10
which is observed at the azimuth angle of 45°. This because the pattern is substantially affected by the asymmetry of the ground plane, and its positioning within the module design.

The surface currents illustrated in Figure 3-3, indicate that strong excitation currents are present in the neighbourhood of the feeding port, producing an inherently asymmetric and weak pattern. It is very interesting as well to observe that the peak elevation gain values (the co-polar components of Figure 3-7(a) and 3-7(b) achieved were quite similar to that found at the azimuth direction. These pattern results will help to minimize the occurrence of poor reception.

3.4 Conclusion

A low profile multi-frequency band monopole antenna module has been presented. The radiator is a crescent shaped microstrip patch. The realised antenna structure shows a relative bandwidth of 51.8%, and a gain above 1.5dBi over the frequency interval (1.75GHz to 3.1GHz). Parametric analysis has been used to identify the principal structural parameters controlling the behaviour of the radiator geometry. Different parameter sets were optimised through HFSS. The best antenna performance, defined in terms of the reflection coefficient is obtained for the 57mm × 37.5mm × 0.8mm antenna volume, and is suitable for integration with a variety of mobile terminals covering the DCS, PCS, UMTS, Bluetooth or IEEE 802.11b/g wireless standards.
3.5 Reference


CHAPTER 4

Miniature Dual-Bands and Wideband Planar Inverted F-L-Antennas (PIFLAs) for WLAN and UWB Applications

4.1 Introduction

This chapter continues the theme of low cost multi-band antenna modules for user terminals, with greater focus on mobile handset applications. Mechanically, the module must be low profile, robust and cost effective through high volume production. From an electrical standpoint it must meet the requirement specification derived from the relevant communications standards, and have low coupling when placed or held in proximity to the human body. One such option is the planar inverted F antenna (PIFA) [1, 2].

However, it should be noted that the conventional PIFA structure has fundamental bandwidth constraints; typically 4-12% impedance bandwidth is achievable in an unmodified design. Modified PIFA designs include T-slot geometries [3] and the introduction of parasitic elements, such as the inverted L [4, 5], modified ground planes
Modifications to the feeding and shorting plates [8, 9] may also be considered. The design adopted here is a miniaturized ‘FL’ antenna, or PIFLA [10].

A novel feeding structure is introduced to achieve a good impedance bandwidth for the design frequency range, which is the lower sub-band (3100MHz-4800MHz) of the UWB spectrum [11]. The design optimisation is carried out using a frequency domain finite element analysis (Ansoft HFSS), the final model is also cross validated through the time domain using a Conformal-FDTD method (SEMCAD). A working prototype is constructed and tested on this basis.

Planar monopole antennas [12-17] have been extensive developed for WLAN and UWB applications, as outlined in the last chapter. This type of antenna suffers from low power gain, inconsistent radiation patterns and broadside gain direction, and high co-/cross-polarisation ratio over the operating frequency band (3100MHz to 10600MHz). Very limited published literature is available, e.g. [18] provides a unidirectional UWB antenna with stable radiation patterns and constant gain profile.

There were several antenna size reduction techniques have been proposed over last decades, that investigate the use of high permittivity substrate, shorting pins, shorting walls and modifying the geometry of the internal antenna [19-21]. Recently, another size reduction technique is proposed in [22-24] using the magnetic wall concept. It was found that the performances in terms of return losses, gains, radiation efficiencies,
radiation patterns of half size structures of the U-slot, E-shaped [22-23] and UWB microstrip patch antennas [24] are comparable to their full structures.

By further extending previous research work [25], this chapter presents a 50% size reduction for two dual-frequency/UWB planar inverted F-L antennas (PIFLAs) with overall size of 30mm × 15mm × 8mm, mounted on a 30mm × 15mm finite ground plane. The target frequencies were chosen to cover IEEE802.11x and UWB by using the same type of antenna [25] with different set of geometry parameters.

The first antenna design is constructed from a driven F-shaped element, and a parasitic L-shaped element. Both patches are inverted and aligned face-to-face over a finite ground plane as shown in Figure 4.1; the optimised dimensions of this antenna are 30×15×4 mm³. The realized antenna achieves a gain between 2.5 dB_i and 4.7 dB_i over a 66.6 % impedance bandwidth (on the interval 2.8 GHz to 5.6 GHz) for |S_{11}| ≤ -10 dB. The radiation patterns, antenna efficiency and antenna group delay are also measured and discussed.

The second antenna design with a compact volume size of 30mm × 15mm × 8mm is presented. By applying the magnetic wall concept a reduced size dual-band and a wideband half PIFLAs for WLAN (2.4GHz/5.2GHz) and UWB applications are achieved. The dual-band antenna shows a relative bandwidth of 12% and 10.2% at ISM2400 and IEEE802.11a frequency bands respectively for input return loss less than 10dB. By carefully tuning the geometry parameters of the dual-band proposed antenna,
the two resonant frequencies can be merged to form a wide bandwidth characteristic PIFLA, to cover 3000MHz to 5400 MHz bandwidth (57%) for a similar input return loss, that is fully covering the upper band UWB (3.1-4.8GHz) spectrum.
4.2 First Design: Planar Inverted F-L Antenna (PIFLA) for Lower-band UWB Applications

4.2.1 Antenna Design Concepts and Structure

The PIFLA schematic is shown in Figure. 4-2. It is constructed from a standard PIFA with a broadband feeding plate, and a parasitic planar inverted L-shaped element. For ease of integration with a practical enclosure, the assembly is mounted over a finite ground plane. The minimum dimensions for the PIFA are 17.5×13×3.5 mm³, where the 17.5mm is the length of F shape and 3.5mm is the antenna height. The minimum dimensions of PIFLA are: 9.5×14×3.5 mm³ for the parasitic element PIFLA and the ground plane dimensions for the final assembly are 30×15 mm².

Figure 4-1: Practical Prototype of proposed antenna
Optimal coupling is achieved with an element separation distance of 3mm. The lowest resonant mode of this structure is approximately 3000 MHz, so if the corresponding wavelength of this mode is $\lambda_0$, then the scaled optimal antenna dimensions are $0.30\lambda_0\times0.15\lambda_0\times0.04\lambda_0$.

The underlying design principle is straightforwardly based on the manipulation of multiple resonances, from the inverted F and L elements, both of which support strong current distributions. The driven PIFA element acts as the primary element, governing the lowest resonant frequency, whilst the higher resonant frequency is controlled by the parasitic element. Both the size of the ground plane, and the feed mechanism play a significant role in the determining the desired wideband characteristic for the impedance bandwidth.

**Figure 4-2:** Proposed antenna structure.
However, it should be noted that the lower frequency component is dominated by the large effect of the capacitive coupling between the feeding element and the finite ground. Thus, a broadband rectangular plate, with a 0.5 mm gap, is used to excite the PIFA. This technique provides an improved impedance matching over the conventional probe feed. The optimum gap width was achieved after several design iterations.

To further adapt this antenna for commercial wireless transceiver applications, understanding the effect of ground plane size is crucial. The influence of the ground plane size using the feeding plate mechanism is fully addressed in the following section.

### 4.2.2 Parametric Design Studies

This parametric study is useful because it provides a comprehensive picture of the antenna characteristic. The first cut design, analysed through the frequency domain, was made by varying each parameter by 1mm and 2mm increments, whilst holding the remaining parameters at their initial values. Figure 4-3(a) shows the simulated return loss for different antenna heights, a -10dB return loss bandwidth of the antenna is achieved at 3.5mm.

The optimisation results can be reviewed in Figure 4-3(b), which indicates that the best length for the PIFA is 17.5mm. Figure 4-3 (c) shows the variations of the reflection
coefficient $|S_{11}|$ with five selected ground plane sizes; once the ground plane size was reduced to $15 \times 30 \text{mm}^2$ the optimised result was obtained.

4.2.3 Result and Discussion

Figure 4-4 compares the simulated and measured reflection coefficient response; the impedance bandwidth of the prototype operates over the range 2800 MHz to 5600 MHz, with $|S_{11}| \leq 0$ corresponding to a 66.7% relative bandwidth with respect to a centre frequency of 4200 MHz. This operating range gives full coverage of the UWB uplink frequency spectrum (3100 MHz to 4800 MHz). There is small discrepancy between predicted and measured results, but this does not indicate the need for additional design optimisation, this may be a construction error.

The effect of the ground plane is a significant aspect in this design. Figure 4-3 (c) shows the variations of the reflection coefficient with five possible ground planes (scaled in units of $\lambda_o$): $\lambda_o \times \lambda_o$, $0.80 \lambda_o \times 0.80 \lambda_o$, $0.60 \lambda_o \times 0.60 \lambda_o$, $0.40 \lambda_o \times 0.40 \lambda_o$, and $0.30 \lambda_o \times 0.15 \lambda_o$. The performance is significantly degraded as the size of the ground plane is increased. This corresponds to the resonant mode of the ground plane being out of band, and therefore makes no constructive contribution to the impedance bandwidth.

Figure 4-5(a) gives the predicted and measured gain of the antenna in the broadside direction; over the interval 3000 MHz to 5750 MHz was interesting to find an average
measurement gain of 3.6\text{dB}_i with $\pm1.1\text{dB}_i$ fluctuation. Figure 4-5(b) exhibits the simulated radiation efficiencies of the antenna. The average measurement efficiency was observed to be 92.5\% with $\pm5.5\%$ efficiency fluctuation. This compares well with the simulations.
Figure 4-3: Simulated reflection coefficients $|S_{11}|$ with different dimensions of (a) antenna height $h$, (b) PIFA length $p_l$ and (c) ground plane.

Figure 4-4: Measured and simulated reflection coefficients $|S_{11}|$
The far field radiation patterns are presented in Figure 4-6. Two pattern cuts (the \(xz\) and \(yz\) planes) were taken at three selected operating frequencies which cover the aggregate bandwidth. The radiation patterns were found to stable and consistent at all the
designated frequencies, as shown in Figure 4-6. Significantly, it also indicates that the maximum co-polarized component appears at the direction of bore sight (+z) for both the E and H planes.

Figure 4-6: Simulated and measured normalised radiation patterns of the proposed antenna for two planes (left: x-z plane, right: y-z plane) at (a) 3000 MHz (b) 4000 MHz and (c) 5000 MHz. ‘xxxx’ simulated cross-polarization ‘oooo’ simulated co-polarization ‘------’ measured cross-polarization ‘———’ measured co-polarization
4.2.4 Group delay

An antenna in UWB system can be analysed as a filter by means of magnitude and phase responses. When a signal passes through a filter, it experiences both amplitude and phase distortion, depending on the characteristics of the filter. By representing the receiver/transmitter antenna as a filter, we can determine its phase linearity within the frequency band of interest by looking at its group delay.

Group delay is the measure of a signal transition time through a device. It is classically defined as the negative derivative of phase versus frequency given by;

\[ \text{Group delay} = - \frac{\partial \theta(\omega)}{\partial \omega} \]

The phase response and group delay are related to the antenna gain response. The group delay variation induced by the radiation pattern of the antenna is a significant parameter in the overall receiver performance; it can be used to track relatively large timing errors.
The group delay spread of the proposed design over the required bandwidth is approximately 0.5 ns. The group delay measurement is given in Figure 4-7.

Figure 4-7: Measured group delay of the proposed antenna
4.3 Second Design: Dual-Bands and Wideband Planar Inverted F-L-Antennas for WLAN and UWB Applications

4.3.1 Antenna Design Concepts and Structure

The PIFLA shown in Figure 4-8 is quite similar to the antenna design of previous work, reported in [25], but the size is reduced to half using the principle of the existing magnetic wall on the antenna surface [22-24]. The initial geometry parameters of the antenna are stated as follows: \( L_1 = 18.6 \) mm, \( L_2 = 10 \) mm, \( h_1 = 8 \) mm, \( h_2 = 4.5 \) mm, \( d = 3.5 \) mm and \( w = 0 \) mm. It should be noted that these dimensions are the same as in [25], except the width of F-shaped, L-shaped radiator and ground plane have been cropped to half, which are 7.5 mm, 8.5 mm and 15mm respectively. The copper metal plate thickness of the proposed antenna and the gap distance for the feed are 0.5mm.

Figure 4-8: Geometry of the proposed miniature PIFLA
The operation mechanism of this PIFLA is simple. As can be portrayed in Figure 4-8, this antenna is a combination of a planar F-shaped antenna with rectangular plate feed and a planar L-shaped antenna. The F-shaped antenna which has a longer electrical length is designed to control the lower resonant mode (2450MHz), whereas the L-shaped antenna is used to provide the higher resonant mode (5200MHz). Conventional wire feed PIFA have limited 4% to 12% bandwidth [23-24]. Changing different feed plate silhouette [23] e.g. triangular plate, bi-triangular plate, etc, the impedance bandwidth can be significantly increased from 18% to 25%. For this reason, we have used the rectangular plate feed for the PIFA radiator for bandwidth enhancement purpose.

The antenna miniaturisation is achieved by removing half of the patch antenna along the line symmetry, as in [22-24], hence, the volume of the patch antenna is reduced to half of the overall volume of the PIFLA antenna [25]. This has obvious advantages in the integration with module casings.

Figure 4-9: Current distribution of (a) full size and (b) half cut PIFLA at 2450 MHz
The current distribution of the full size PIFLA antenna [25] is investigated, as shown in Figure 4-9 (a). The maximum current mainly appears on the edges of the structure. Due to the minimum current appears on the centre of the structure that approximates the existing of magnetic wall. Thus, by cutting half of this full size structure, the half size structure will still hold the same properties of the same current distribution, as described in Figure 4-9 (b).

By keeping the same geometry parameters, the return losses of the half size PIFLA are studied. As can be seen in Figure 4-10, the bandwidth for lower resonant mode remains, but, the bandwidth for higher resonant mode degrades considerably in which 12.2% (5GHz to 5.65GHz) to work in [25] and 3.8% (5.15GHz to 5.35GHz) to the half structure.

![Simulated return Losses (Full Size PIFLA verse Half Size PIFLA (without optimised))](image)

Figure 4-10: Simulated return Losses (Full Size PIFLA verse Half Size PIFLA (without optimised))
4.3.2 Parametric Study Results

The primary goal of this parametric study is to understand the impedance bandwidth variation in the half-size PIFLA at the lower and higher resonant modes by tuning the geometry parameters. Therefore, throughout this study, there are two expected outcomes, i.e.

(i) to optimise bandwidth of the proposed half-size dual-band PIFLA for WLAN application, so that, the proposed half size PIFLA possesses the same bandwidth at the lower and higher resonant modes as full size PIFLA [25]

(ii) to establish wideband characteristic of this antenna by combining the two operating modes.

Each iteration runs for one parameter, with the others held constant. These fixed and variable parameters are shown in Figure 4-8. The variable parameters are considered to be critical in defining the limits of the operating bandwidth. The impedance bandwidth of the “half-structure” is heavily determined by the mutual coupling between the two radiating elements. This effect cannot be easily described analytically.
4.3.2.1 Effect of the length of F-shaped radiator (L1)

This parameter changes the electrical length of the F-shaped radiator, and it governs the lower resonant mode of this antenna. By extending the length of the F-shaped radiator (L1) from 16.5 mm to 30 mm, the lower resonant frequency varies from 2650 MHz to 1600 MHz, and the bandwidth is gradually diminished, as per Figure 4-11 (a). As for the higher resonant mode, it is clearly independent of the L1. However, as the L1 becomes shorter to 10.5 mm and 4.5 mm, the lower resonant mode tends to disappear and higher resonant mode tends to shift to the upper frequency band with wider bandwidth.

This is due to the both of the $F_L$ and $F_H$ are near to each other and good broadband impedance matching is attained.
Figure 4-11: Simulated return losses with variation of parameters: (a) L1; (b) L2; (c) h1; (d) h2; (e) d; (f) w.
4.3.2.2 Effect of the length of L-shaped radiator (L2)

The length of L-shaped radiator determines the upper resonant mode of the proposed PIFLA. The effect of variation of L2 is elaborated in Figure 4-11 (b). At lower resonant mode, as L2 is increased from 2 mm to 24 mm, the resonant frequency seems to move from higher to lower values, and the impedance bandwidth is shifts accordingly. This is due to the coupling effect as the L-shaped radiator is closer to the F-shaped radiator. Likewise, at the upper resonant mode, the resonant frequency is shown to shift from higher to lower resonant frequency (from 7500MHz to 4400MHz).

4.3.2.3 Effect of the height of F-shaped radiator (h1)

The height of F-shaped radiator serves the same role as L1 to control the fundamental resonant mode of the antenna, but, it also contributes to the capacitive coupling with the ground plane. As can be seen from Figure 4-11 (c), the smaller values of h1 are 1mm and 2.75mm. High mismatching of the impedance bandwidth at both higher and lower resonant modes were observed. When the h1 is varied from 4.5mm to 8mm, no significant variations on the impedance bandwidth at lower and upper resonant modes are found. Therefore, h1 can take any value between 4.5 mm to 8 mm, without deteriorating the performance of this antenna.
4.3.2.4 Effect of the height of L-shaped radiator (h2)

Influences of parameter h2 are shown in Figure 4-11 (d). By varying the height of the L-shaped radiator, both of the lower and upper resonant frequencies seem to move the lower resonant frequency simultaneously. However, at the lower resonant mode, the degree of the impedance bandwidth remains at the same level, whereas at higher resonant mode, impedance bandwidth seems to reduce drastically when h2 falls below 6.5mm.

4.3.2.5 Effect of the position of the rectangular plate feeding (d)

The position of the rectangular plate feeding (d) is the most sensitive parameter in this study. This is because its variation can introduce different excitation modes to the antenna, as indicated in Figure 4-11 (e). It is clear when d is 0.5 mm away from vertical metal plate (h1) of the F-radiator, the PIFLA results in an impedance mismatch at both modes. While varying d to 4.5 mm, this antenna shows an obvious dual-resonant operation, which completely covers the bands of IEEE 802.11b/g (2400MHz to 2485 MHz, BW=3.48%) and IEEE 802.11a (5.15 MHz to 5.35 MHz, BW=3.8%) for WLAN applications. As d is moved further away (d=8.5 mm), a broader impedance bandwidth can be obtained for two resonant operating modes, but, the two resonant operating band shift out of the WLAN 2400 MHz and 5200 MHz bands. However, it occupies the required operating frequency band for WLAN 5800MHz application (5725 MHz to 5825 MHz). By further increasing d to 12.5 mm and 16.5 mm, a wideband impedance
response can be achieved by merging the two resonant modes into one resonant mode. This wide impedance bandwidth is suitable for UWB uplink applications (3100 MHz to 5200 MHz).

4.3.2.6 Effect of the width of F-L shaped radiators (w)

The independent influence of the width of the F and L-shaped radiators has been discussed in [25]. It was shown that prolonging the width of the F- radiator will only affect the second resonant mode of this antenna. However, varying width of L-shaped radiator will not give any significant changing on its impedance bandwidth; so, the performance of this antenna is independent of width of L-shaped radiator. In this iteration, the width of F-L- radiator (w) will be altered at the same time to examine the antenna impedance bandwidth. It can be seen in Figure 4-11 (f) that widening the width (w) seems to offer a better impedance matching. Conversely, if the width (w) becomes shorter, the impedance mismatch will occur at upper frequency resonate mode.

4.3.3 Results and Discussion

The prototype antennas were characterised by RF and anechoic chamber measurements. The RF measurements were carried out using an HP-8510C VNA. The radiation patterns were measured as follows. The fixed reference antenna was a broadband horn (EMCO 3115), and the spacing with the antenna under test (AUT) was fixed at 4m.
Two pattern cuts (H-plane and E-plane) were taken for each design frequency covering the target bandwidth.

**A. Miniature Dual-band PIFA**

By scrutinizing the variations in the geometry against the impedance bandwidth, the optimal parameters can be recognised for dual-band operation. These parameters can be given as follows: \( L_1 = 18.6 \text{ mm}, \ L_2 = 10 \text{ mm}, \ h_1 = 8 \text{ mm}, \ h_2 = 5 \text{ mm}, \ d = 4.5 \text{ mm} \) and \( w = 1.5 \text{ mm} \).

Figure 4-12: Practical prototype of the proposed dual-bands PIFA
Figure 4-12 shows the first miniature dual-band PIFLA prototype. Figure 4-13 illustrates the typical measured and computed antenna performance in terms of impedance bandwidth for both full-size and half-size PIFLAs. Two adjacent resonant frequencies in the neighbourhood of $\leq -10$ dB are observed, i.e., 2450 and 5350 MHz. The lower and upper modes provide 12% and 10.2% relative bandwidth from 2350 MHz to 2650 MHz and 5100 MHz to 5650 MHz, respectively. The return loss is better than -10 dB, satisfying the desired IEEE802.11b/g frequency band (2400-2485 MHz) and IEEE 802.11a (5.15-5.35GHz) bands respectively. As can be observed, simulated and measured results for half-size PIFLA were found to be in excellent agreement. The half-size and full-size PIFLA seem to show identical measured impedance bandwidth at two resonant modes.

Figure 4-13: Measured and simulated return losses for the proposed dual-bands PIFLA
An investigation of the effect of the ground plane length on the antenna return loss was carried out. The size of ground plane was varied from 120 mm x 120 mm to 30 mm x 15 mm. The corresponding results are presented in Figure 4-14. If we define $\lambda_0$ as 2400MHz, then it can be seen that an insignificant variation in the simulated return loss was observed for the following cases

$\{\lambda_0 \times \lambda_0, 0.75\lambda_0 \times 0.75\lambda_0, 0.5\lambda_0 \times 0.5\lambda_0, 0.375\lambda_0 \times 0.375\lambda_0, 0.25\lambda_0 \times 0.125\lambda_0\}$. 

Figure 4-14: Simulated return losses for the proposed dual-bands PIFLA with various ground plane dimensions
However, the impedance matching for the antenna was notably improved while the resonant frequency of the antenna was unchanged. This indicates that the prototype also features a low degree of sensitivity to ground plane size, implying that the same antenna design can be potentially adopted for many other mobile devices with little modification. When the size becomes smaller than quarter wavelength, which is 30 mm × 15 mm (i.e. $0.24\lambda_o \times 0.12\lambda_o$), the return loss falls into the required dual-bands (2400MHz/5200MHz).

Figure 4-15 (a) and (b) illustrate the simulated and measured gain of the prototype in the broadside direction over the frequency ranges 2400 to 2480MHz and 5000MHz to 5600MHz respectively. At the lower band, a stable measured gain can be observed from 2.6 to 3.5 dBi. For the upper band, the range of the measured gain varies from about 4.5 to 5.2 dBi. It should be noted this measured gain for the half size PIFLA is in good agreement with the full size PIFL antenna. By analysing the gain variation over two specific operating bands, the peak antenna gain variations are found less than 1 dBi, as compared with the predictions.
Figure 4-15: Measured gains for proposed dual-bands PIFLA: (a) lower band; (b) upper band
The antenna efficiency of the prototype is determined from the EM simulation, and by the Wheeler cap method [26, 27]. The results are summarised in Figure 4-16. At the lower frequency band (see Figure 4-16 (a)), the variations of measured antenna efficiency for the Full and Half size PIFLA are 0.96 to 0.98 and 0.93 to 0.94 which corresponding to average of 97% and 93.5% over the operating frequency range. At the upper frequency band (see Figure 4-16 (b)), the average radiation efficiencies of 93.5% and 89% are achieved with ±9% and ±16% of radiation efficiencies fluctuation for the full size and half size proposed antenna respectively.
Figure 4-16: Measured radiation efficiencies for proposed dual-bands PIFLA: (a) lower band; (b) upper band

Figure 4-17 (a) and (b) show the simulated and measured co-polar and cross-polar radiation patterns in the x-z and y-z planes at 2400MHz and 5200MHz for the miniature PIFLA. The simulated and measured radiation patterns of the fabricated prototype are seen to be quite similar to each other at the two designated centre frequencies.

In addition, by comparing the radiation patterns for a full size PIFLA [25] and half size PIFLA, both of them are apparently indistinguishable. This leads to a conclusion that using half structure techniques, the sizes of this half dual-band antenna can be reduced without appreciably impaired its performance characteristic when compared to the full-size antenna.
Figure 4-17: Simulated and measured normalised radiation patterns of the proposed dual-bands PIFLA for two planes (left: x-z plane, right: y-z plane) at (a) 2400 MHz and (b) 5200 MHz. ‘xxxx’ simulated cross-polarization ‘oooo’ simulated co-polarization ‘---’ measured cross-polarization ‘———’ measured co-polarization
B. Wideband PIFLA

The optimised geometry parameters of the proposed PIFLA for UWB application are given as follows: \( L1 = 18.45 \text{ mm} \), \( L2 = 8.5 \text{ mm} \), \( h1 = 8 \text{ mm} \), \( h2 = 5.5 \text{ mm} \), \( d = 14.5 \text{ mm} \) and \( w = 4.5 \text{ mm} \). Figure 4-18 describes the experimental prototype of the PIFLA with finite ground plane of \( 30 \times 15 \text{ mm} \), while Figure 4-19 shows the typical measured and computed antenna performance in term of impedance bandwidth. As can be seen, the lowest and highest frequency edges i.e., 3000MHz and 5400MHz, of an input return loss \( \leq -10 \text{ dB} \) are observed. The impedance bandwidth of the proposed antenna, for -10dB return loss, is 2400 MHz or about 57% with respect to the centre frequency at 3950GHz (average of measured lower and higher frequencies with a -10dB return loss), which fully covers the frequency spectrum of uplink UWB (3100 MHz to 4800MHz). Basic agreement is achieved between the experimental and computed return loss over the desired operating frequency band.

![Practical prototype of the proposed wideband PIFLA](image)

Figure 4-18: Practical prototype of the proposed wideband PIFLA
Figure 4-19: Measured and simulated return losses for the proposed wideband PIFLA

Figure 4-20: Simulated return losses for the proposed wideband PIFLA with various ground plane dimensions.
Moreover, in order to understand the effect of the ground plane of the proposed antenna, simulations were conducted to check the variations of the return loss against the size of the ground plane, as described in Figure 4-20. If we define $\lambda_0$ at 3000MHz as 100mm, from Figure 4-20, it follows that there are five different ground planes, which are $\lambda_0 \times \lambda_0, 0.8\lambda_0 \times 0.8\lambda_0, 0.6\lambda_0 \times 0.6\lambda_0, 0.4\lambda_0 \times 0.4\lambda_0, 0.3\lambda_0 \times 0.15\lambda_0$. It is noticeable that when the ground plane is greater or equal to $0.6\lambda_0$, the narrow impedance bandwidth characteristic was found and they do not contribute any significant bandwidth enhancement. As the ground plane becomes smaller than $0.4\lambda_0$, the impedance bandwidth becomes broader. According to [28], this is because the ground plane resonates at the operating frequency of the antenna element, so the bandwidth of the antenna-chassis combination will improve considerably. If the ground plane resonates far away from the operating frequency, the bandwidth will be decreased due to the insignificant contribution of the ground plane.
Figure 4-21: Measured and simulated gains for proposed wideband PIFLA.

The simulated and measured gains of the prototype in the broadside direction over the frequency range from 3000 MHz to 5200 MHz are shown in Figure 4-21. An average gain of 4.5 dBi is achieved with ±1.2dBi of gain fluctuation. The simulated results relate well with measured ones. Figure 4-22 shows the simulated and measured antenna efficiencies of the prototype. As can be seen, the maximum, minimum and average of measured radiation efficiencies are 87%, 98% and 92.5% respectively over the operating frequency range. Both simulated and measured radiation efficiencies are correlated well.
The far-field radiation characteristics of the prototype were also investigated at 3000MHz, 4000MHz and 5000MHz, as shown in Figure 4-23. It is observed that the radiation patterns at different frequencies are quite similar, which is expected in a wideband antenna. More importantly, it shows that the maximum co-polarised component appears at the direction of boresight (+z) for both E- and H-planes. At 3000 MHz and 4000 MHz, a co-pol/cross-pol ratio of better than 20dB is found at boresight in all of the measured radiation patterns. This indicates satisfactory polarization purity. While at 5000 MHz, due to the occurrences of high order mode resonances, this causes the increase in cross-polarisation levels and resulted a co-pol/cross-pol ratio of 10dB at boresight direction.
Figure 4-23: Simulated and measured normalised radiation patterns of the proposed wideband PIFLA for two planes (left: x-z plane, right: y-z plane) at (a) 3000MHz, (b) 4000MHz and (c) 5000MHz ‘xxxx’ simulated cross-polarization ‘oooo’ simulated co-polarization ‘------’ measured cross-polarization ‘———’ measured co-polarization
4.4 Conclusion

Both lower and dual-band planar inverted F-L (PIFLA) antennas for UWB application have been designed.

Firstly, various parameters of the lower band antenna are optimised to get the best performance over the required bandwidth and to reduce the overall size of the antenna. The optimised design is a compromise between antenna size and impedance bandwidth in the lower operating UWB frequency spectrum.

The prototype lower band antenna was fabricated from a copper sheet of thickness 0.5mm. The experimental result shows reasonable agreement with the simulated one. The antenna displays stable radiation patterns, gain and radiation efficiency for the entire operating bandwidth, thus making it suitable for UWB applications.

Secondly, a simple geometry of a miniature planar inverted F-L antenna (PIFLA) has been proposed and studied experimentally and theoretically. By carefully selecting different sets of optimal geometry parameters and applying the size reduction techniques to the proposed antenna, two 50% size reduction PIFLAs have been designed and tested. By applying the magnetic wall concept a reduced size dual-band and a wideband half PIFLAs for WLAN (2.4GHz/5.2GHz) and UWB applications are achieved. By balancing the size and bandwidth constraints, these proposed antennas
have a compact envelope dimension of $30\text{mm} \times 15\text{mm} \times 8\text{mm}$ and covers the required operating frequency band for WLAN and UWB applications. These features make the proposed antenna an attractive candidate for application in a range of future mobile terminals.
4.5 References


[26] HFSS ver.11, Ansoft Ltd.
[27] SEMCAD X ver.14, Schmid & Partner Engineering AG, Zeughausstrasse 43, 8004, Zurich, Switzerland.

CHAPTER 5

Planar Inverted F-F Antenna (PIFFA) for Wide-Band UWB Applications

5.1 Introduction

Planar inverted-FF antenna with suspended rectangular feeding mechanism is presented for lower band ultra-wideband and IEEE 802.11a wireless local area network (WLAN) applications. By manipulating the optimum strong electromagnetic coupling distance between driven and parasitic F-shaped radiators on a small ground plane, two closely separate resonances can be merged to form a wide impedance bandwidth of 73.7% (3.0 — 6.5 GHz) at $|S_{11}| < -10$dB. The overall occupied antenna volume is $30 \times 15 \times 3$ mm$^3$ to satisfy the size requirement of modern wireless devices. The impact of the different antenna geometry on the antenna performance is studied. The corresponding measured radiation patterns, gains and group delays are given to fully characterize the performance of this antenna.

Ultra-wide band antennas [1-7] have gained immeasurable attention for communication, geolocation, radar imaging and medical applications. Due to their compact size, low profile and ease of integration with other RF circuits, printed monopole antennas [1-4]
seem to be the most popular candidates to be designed to cater for this unabated demand. However, this type of antenna is ground dependent, where the electric currents are concentrated on both the radiator and the ground plane. This makes it inconsistent back radiation through its ground plane and bad impedance matching on its low resonant frequency band. Some techniques [3-4], such as modifying the shaped of the radiator [3] and introducing L-shaped slots on the edge of the ground plane [4] to suppress the electric current on the ground plane at lower edge operating frequencies, have been proposed to alleviate this problem.

To lower the cost of using dielectric filled printed antennas, some interesting UWB antennas with air dielectric have been designed and studied [5-7]. However these antennas have an average height of 10mm which is difficult to employ within the small volume of a portable electronic unit. In this chapter, a low profile and compact wideband UWB planar inverted FF antenna with a volume of 30 × 15 × 3 mm³ is presented to operate in the upper frequency spectrum of UWB and IEEE 802.11a WLAN.

5.2 Antenna Design Concepts and Structure

The geometry of the proposed planar inverted FF antenna (PIFFA) model is depicted in Figure 5-1. As can be seen, it is made of a driven and a parasitic planar inverted F (PIF) element which is very similar to our previous work [7-8]. The feeding position has been
moved to the centre of the structure and the L-shaped parasitic element has been replaced by an F-shaped element to increase the total volume of the antenna. By further optimising the proposed antenna geometry parameters, the size of this antenna has been further minimized and the impedance bandwidth has been widened in comparison with earlier work [7].

![Geometry of the proposed UWB antenna](image)

Figure 5-1: Geometry of the proposed UWB antenna

The optimised (minimized) dimensions of the PIFA and PIF element are $17.5 \times 15 \times 2.2$ mm$^3$ and $11.1 \times 15 \times 2.5$ mm$^3$ respectively. Both of them are separated by 1.5 mm for optimal coupling and are mounted on a 30 $\times$ 15 mm$^2$ ground plane to achieve a wide impedance bandwidth. The total volume of this proposed antenna is $30 \times 15 \times 3$ mm$^3$, taking account of the thickness of the copper which is 0.5 mm. As can be seen, it is about 25% smaller than the previous presented work [7].
The basic wide impedance bandwidth operating concept of this proposed antenna is simply controlling the multiple resonance characteristic of the antenna structure. Conventional probe feeding techniques invoke a large input inductance [9] and result in bad impedance matching over a wide frequency spectrum. To compensate for the inductance within a broadband frequency range, the suspended rectangular feed plate [7-8, 10-13] is adopted on the driven PIFA of this antenna structure. This feeding mechanism acting as a broadband impedance matching network allows two adjacent modes to be simultaneously excited. The designed geometry of the driven PIFA generates the fundamental resonant mode of this antenna, while the second resonant mode is excited by the contribution of both driven and parasitic PIFA elements, and the ground plane. The parasitic inverted F element is adopted due to introducing the additional shorting wall and creating capacitive coupling between the two shorting walls for better impedance matching on the higher frequency band. In general, the impedance matching can be improved by adjusting the electromagnetic coupling between the feeding rectangular sheet, the location of the feed point and the size of the ground plane. These effects will be addressed in the following section.

5.3 Results and discussion:

Two computational electromagnetic software packages, i.e. CST [14] and SEMCAD [15], were used to accurately predict the characteristics of the proposed antenna model
prior to constructing the prototype. Figure 5-2 shows the practical prototype of the proposed antenna model.

Figure 5-2: Practical Prototype of the proposed antenna

The comparison of the computed and simulated reflection coefficient $|S_{11}|$ is shown in Figure 5-3. As can be observed, the antenna covers the operating frequency spectrum from 3.0 GHz to 6.5 GHz which corresponds to 3.5 GHz or about 73.7% impedance bandwidth at $|S_{11}|$ less than -10 dB and two obvious resonance frequencies can be noticed, i.e. 3.5 GHz and 6.0 GHz. This indicates that the antenna is able to operate over the uplink UWB radio band (3.1 GHz to 4.8GHz) and IEEE 802.11a WLAN in the frequency band of 5.15 GHz to 5.825 GHz. A good agreement has been reached between theory and experiment.
Figure 5-3: Measured and simulated reflection coefficients $|S_{11}|$

However, some discrepancies between theoretical estimation and practical result can be most likely attributed to fabrication errors.

A parametric study was carried out to understand the impedance bandwidth variation of the proposed antenna by tuning the geometry parameters. Three parameters were selected for this study, i.e. the feeding position, the feed gap distance and the ground plane size. These parameters are considered as critical parameters in determining the lowest and highest frequency of the operating bandwidth. Each simulation was run with only one parameter varied, while other parameters stayed unchanged. The influences of feeding position are depicted in Figure 5-4(a).
In this study, the feeding position is initially set to the edge of the structure which corresponds to 0 mm, then, it is gradually moved with increments of 2.5 mm to the centre of the structure which is 7.5mm. As can be seen, for the lower resonant mode (3 GHz to 4.5 GHz), the impedance bandwidth remains essentially the same. At the higher resonant mode (4.5 GHz to 6.5 GHz), the impedance bandwidth decreases drastically when the feeding position is set to near the edge which is below 5mm. Therefore, this feeding position can take any value between 5 mm to 7.5 mm, without degrading the performance of the antenna.
The feed gap distance is one of the most sensitive parameter in this study. This is because its variation can significantly impair the impedance matching over the desired operating frequency band, as illustrated in Figure 5-4(b).

This parameter is investigated by starting from 0.25 mm and increasing to 1.25 mm with increments of 0.25 mm. It is clear that when it is 0.25 mm, the antenna exhibits an impedance mismatch across the frequency band of interest. As it is increased further (d = 0.5, 0.75 and 1 mm), a good impedance matching can be achieved over the band at $|S_{11}| < -10$ dB. By further increasing it to 1.25 mm, impedance mismatch will occur in the upper frequency resonance mode. This leads to the conclusion that the feed gap distance should keep between 0.5 to 1 mm for the best impedance matching.
An investigation of the effect of the ground plane length on the antenna reflection coefficient $|S_{11}|$ was performed, as shown in Figure 5-4(c).

![Graph showing simulated reflection coefficients $|S_{11}|$ with variation of different parameters: (a) feeding position, (b) feed gap distance and (c) ground plane size.](image)

(c)

Figure 5-4: Simulated reflection coefficients $|S_{11}|$ with variation of different parameters: (a) feeding position, (b) feed gap distance and (c) ground plane size

In this analysis, the proposed antenna is placed on one of the corners of the ground plane instead of in the centre of the ground plane in order to mimic the practical scenarios of implementing this antenna on a commercial electronic PCB. Defining $\lambda_o$ as the largest value (i.e. 100mm, at 3000MHz), Figure 5-4(c) shows the effect of five different ground planes: 30 mm $\times$ 15 mm ($\equiv 0.3\lambda_o \times 0.15\lambda_o$), 40mm $\times$ 40mm ($\equiv 0.4\lambda_o \times 0.4\lambda_o$), 60mm $\times$ 60mm ($\equiv 0.6\lambda_o \times 0.6\lambda_o$), 80mm $\times$ 80mm ($\equiv 0.8\lambda_o \times 0.8\lambda_o$), and 100mm $\times$ 100mm ($\equiv 1\lambda_o \times 1\lambda_o$).
0.4λ₀), 60 mm x 60mm (≈ 0.6λ₀ × 0.6λ₀), 80mm x 80mm (≈ 0.8λ₀ × 0.8λ₀) and 100mm x 100mm (≈ λ₀ × λ₀). It is noticeable that when the ground plane side length is greater than or equal to 0.4λ₀, a considerable impedance mismatch occurred within the lower frequency components from 3000 to 4000 MHz at |S₁₁| > -6 dB. Therefore, it is recommended to retune the antenna geometry parameters when it is mounted on different size ground plane for achieving |S₁₁| ≤ -6 over the frequency bands of interest. This is due to the strong capacitive coupling between the ground plane and the radiating elements, since the height of the antenna is 3mm which is 0.03 λ₀. However, when the ground plane size is reduced to 0.3λ₀ × 0.15λ₀, the impedance bandwidth at lower frequency is significantly improved, this might be due to antenna-ground plane combination has an important dependency on the ground plane resonant wave mode and on the coupling between the antenna and the ground plane [13].

Figure 5-5: Surface current distribution of the proposed antenna at (a) 3.5GHz and (b) 6.0GHz.
The surface current on the metal pattern of antenna is described in Figure 5-5 at 3.5 GHz and 6.0 GHz resonant frequencies. As can be seen clearly, at the low resonant mode of this antenna, intense current appears on the driven PIFA radiator and the total length of the current path is about 22.6 mm (10.2 + 2.2 + 10.2 mm) which is equivalent to $\lambda/4$ at 3.5 GHz. By examining the high resonant mode, strong current mainly concentrates on the surface of the parasitic PIF element and the total current path is approximately $\lambda/2$ at 6.0 GHz which is computed as 21.3 mm (9.4 + 2.5 + 9.4 mm). This verifies that the basic operating principle of this antenna where the driven PIFA radiator is excited by the low resonant mode and the high resonant mode is generated by parasitic PIF element.

The simulated and measured gains of the designed antenna in the broadside direction over the frequency range from 3000 MHz to 6500 MHz are shown in Figure 5-6. An average gain of 5.0 dBi is achieved with ±0.5dBi of gain fluctuation that might be due to alignment of the antenna elements. Nonetheless, the deviation from the simulated results is no more than 1.2 dBi and this can be said to be in reasonable agreement.
The far-field radiation characteristics of the proposed antenna were also investigated at 3000 MHz, 4000 MHz, 5200 MHz and 5800 MHz by both simulation and measurement. Two cuts of the radiation patterns, i.e. xz-plane and yz-plane are given to describe the far field performance of the antenna. As observed in the Figure 5-7, a satisfactory consistent omni-directional radiation behavior is obtained at all the presented frequencies, which is expected from a wideband antenna that is not substantially electrically large. Figure 5-8 illustrates the measured group delay of the proposed antenna. As can be found, the variation of the group delay is less than 0.5 ns over the lower band UWB frequency spectrum.
Figure 5-7: Simulated and measured normalised radiation patterns of the proposed antenna for two planes (left: x-z plane, right: y-z plane) at (a) 3000 MHz (b) 4000 MHz and (c) 5200 MHz (d) 5800 MHz ‘xxxx’ simulated cross-polarization ‘oooo’ simulated co-polarization ‘------’ measured cross-polarization ‘——’ measured co-polarization
5.4 Conclusions

This chapter proposed a wideband plane inverted F-F antenna with minimum occupied dimension of $30 \times 15 \times 3 \text{ mm}^3$ to be accommodated within a space limited modern wireless electronic terminal, has been presented. The simulated and measured results in terms of the return loss, radiation pattern, power gain and group delay, show that the proposed antenna is a good candidate for lower UWB spectrum and IEEE 802.11a WLAN applications.
5.5 References


[14] CST v.5.0, Microwave Studio [online], http://www.cst.com/ SEMCAD [online], www.semcad.com
CHAPTER 6

Dual Planar Inverted F-Antenna for WLAN/WIMAX and Lower-Band UWB Wireless Applications

6.1 Introduction

Wireless local area networks (WLANs) have gained popularity in indoor environments, owing to their licence free status and ease of installation: the WiFi WLAN card has thus become an essential element to be integrated into most current laptop computers and many mobile phones. IEEE 802.11 (WiFi) is a key industry standard for WLAN systems: this allocates operating frequencies in licence-free industrial scientific and medical (ISM) bands at 2.4, 5.2 and 5.8 GHz (specifically 2400–2485, 5015–5350 and 5725–5850 MHz) for portable electronic devices to access the internet wirelessly. Ignoring the newer and physically very different multiple-input multiple-output (MIMO)-based standard IEEE 802.11n, these WiFi systems are typically capable of handing a maximum channel rate of about 11 Mbps and a maximum user data rate of about 1.6 Mbps within the maximum transmission range of approximately 100 m [1].

In recent years, another two wireless standards, worldwide interoperability for microwave access (WiMAX) and ultra wideband (UWB), have been created,
respectively, defined by the IEEE 802.16 family of standards and the now-withdrawn IEEE 802.15.3a draft, with the aim of improving the data rate and/or communication distances, to cater to the increasing consumer demand. WiMAX technology operates in 2.5, 3.5 and 5.5 GHz bands (2500–2690, 3300–3700, 5250–5850 MHz), with a range that correlates it with the concept of a wireless metropolitan area network. It can theoretically reach 50 km radius coverage and achieve data rates up to 75 Mbps, for which the throughput is higher than the 1.5 Mbps performance of typical broadband services [2]. UWB is intended for short range (~10 m) and higher data rate communication (>500 Mbps) across a wide frequency spectrum (3.1–10.6 GHz). The lower and upper UWB spectra are 3.1 to 4.0 GHz and 6.0 to 10.6 GHz, respectively, [3].

Together with the MIMO-based IEEE 802.11n, these three wireless standards are major candidates for emerging commercial mobile wireless technologies. However, for terminals using these technologies to be truly mobile, multi-band miniaturised antenna designs are required and these pose challenges. Recently, several publications have discussed different types of antennas for WLAN/WiMAX applications [4–17]. These can be classified into three categories: dual band [4–7]; tri-band [8–14] and broadband [15–17], encompassing the desired operating frequency bands from 2.4 to 6.0 GHz. In the case of the dual-band and tri-band modalities, it was noted that to avoid frequency collision and to minimise interference from the unused licenced unlicenced frequency spectra, band-rejection functions were considered a desirable objective for all of the proposed antennas [4–13].
These band-notching methods include modifying the geometries of a planar inverted F-antenna (PIFA) [4, 5], adding a truncated corner on a printed inverted F-antenna [6], introducing a square conductor-backed plane on a printed monopole antenna [7], implementing a multi-arm inverted-F monopole [8], electromagnetic coupling of a stub-loaded openloop resonator on the underside of a CPW-fed mirrored-L monopole [9], applying ceramic chip radiating elements to the geometry of the antenna [10] and inserting band notching slits in the radiating structure of the antenna [11–14]. For the case of broadband operation, coaxial antennas [15], CPW-fed Koch fractal slot antennas [16] and printed E-shaped monopoles [17] were shown to provide sufficient impedance bandwidth to cover the required operating frequency for WiFi, WiMAX and lower-band UWB applications. However, it is significant that some of these antennas required a large ground plane [9, 14, 15, 17]; some produced inconsistent radiation patterns [9, 11, 17] and others showed substantial gain variations across the operating frequency bands [11, 13]. In addition, most of the works addressed only the needs of WiFi and WiMAX applications [4–16]: very few compact antenna designs have included discussion of application to the lower UWB band [17].

To overcome these limitations, a combination of two compact broadband planar inverted F-antennas was devised. This consists of a driven PIFA and a parasitic PIFA element and hence may be described as a dual PIFA (DPIFA). The driven and parasitic PIFAs are designed to control the lower and upper resonant modes of the antenna, respectively. By carefully adjusting the geometry parameters of the antenna, broadband impedance matching can be realised. The size of the antenna achieved was 30 mm × 15
mm × 8 mm, which is small enough to offer a high degree of freedom for designers to install into typical small electronic cases for wireless sensor network and short-range radio communication applications [18].

An extension of the planar inverted F-antenna (PIFA) concept is proposed, combining two such antennas (one of them fed parasitically) to extend the usable bandwidth to cover the WLAN, WiMAX and lower-band UWB spectra. The broad impedance bandwidth is achieved by combining a driven slotted PIFA, having bandwidth-enhanced rectangular strip feed, with a parasitic inverted F-shaped element, hence describable as a dual PIFA. The geometry parameters of these two radiators are selected to enhance the impedance bandwidth and hence to encompass the entire required set of operating frequency bands. The overall optimised dimensions of the antenna are 30 mm × 15 mm × 8 mm, thus making it compatible with installation on portable wireless electronic devices. The results show that the proposed antenna can achieve a gain between 2.0 and 4.0 dBi across the entire impedance bandwidth 2.4–6.2 GHz (i.e. 88.4% relative bandwidth) for a reflection coefficient of |$S_{11}$| < -10 dB. The radiation patterns of the antenna show low-gain broad-beam coverage, as is essential for a wireless electronic device. Computed and experimental results are compared and are shown to be in satisfactory agreement.
6.2 Antenna design concept and structure

The proposed antenna design is an evolution of earlier research [5, 19, 20]. In [5], it was found that variations of the widths of the driven and parasitic elements will not cause any significant change in the impedance bandwidth. In [20], the driven antenna element excites the higher frequency mode, while the parasitic antenna element controls the lower-frequency mode to provide dual-band operation for mobile telephone applications. In addition, using the concept of removing half of the antenna geometry along the line of symmetry of the structure [21, 22], the antennas in [5, 19, 20] might be reduced to half of its original size without deterioration of their overall characteristics. Based on the results of [5, 19], the aim of the present work was a compromise design process between size and impedance bandwidth constraints to realise a small wideband antenna on a small ground plane (ideally 30 mm × 15 mm), operating over the WLAN, WiMAX and lower-band UWB bands.

The design concept adopts the principle of multiple radiating elements, each supporting strong currents and radiation of one of the two resonant modes. It should be noted that the parasitic PIFA is adopted instead of the inverted-L element, as applied in the antenna structure reported in [5, 19]. This is because the parasitic PIFA can provide more degrees of freedom in its geometry parameters in the antenna design optimisation process, in comparison with the inverted-L element, and hence this provides more flexibility to achieve the optimum goals of the antenna performance. The driven PIFA is
the primary element that governs the lower resonant frequency, while the higher
resonant frequency is excited by the combination of the driven and parasitic PIFAs. This
is an inverse method of excitation in comparison with [20]. A corollary is that the
coupling gap between the two PIFAs is critical and a non-optimal choice of this can
result in narrow impedance bandwidth or bad impedance matching over the required
frequency band. It must be emphasised that the feeding method also plays a
considerable role in achieving wide impedance bandwidth: a broadband rectangular
strip with a 0.5 mm feed gap was used to excite the driven PIFA instead of using a
conventional probe (wire) feed. This improved the impedance matching because of its
reduced inductance compared with the probe feed [5, 19]. Further, a slot was introduced
in the vertical shorting wall of the driven PIFA in order to reduce the coupling between
the rectangular feed strip and the shorting wall, and hence to improve the impedance
matching. This became evident from the parametric study.

Initially, the geometry parameters of the proposed antenna and the resonant frequency
for the driven and parasitic PIFAs can be approximately predicted, using the following
formula [23]

\[ f \approx \frac{c}{4\pi(W+L)} \]  

(6.1)

Where \( c \) is the speed of light; \( W \) and \( L \) are the width and length of the radiating element,
respectively. To begin with, using the primary antenna dimensions shown in Figure 6-1, it is found that the driven PIFA is resonant at $f_L \approx 2.3$ GHz, whereas the parasitic PIFA is resonant at $f_H \approx 3.4$ GHz.

Hereafter, the driven PIFA is optimised with dimensions of $17.6$ mm $\times$ $15$ mm $\times$ $h_1$ mm, whereas the parasitic PIFA is $8$ mm $\times$ $14.1$ mm $\times$ $h_2$ mm. For ease and quick implementation in a typical commercial wireless casing or enclosure, both of them are mounted on a $30$ mm $\times$ $15$ mm finite ground plane. For optimal coupling, the separation between the two elements is $4.4$ mm. The thickness of the copper sheet and the feed gap distance are both $0.5$ mm. This configuration has overall dimensions of $30$ mm $\times$ $15$ mm $\times$ $h_2$ mm, which is equivalent to $0.24\lambda_0$ $\times$ $0.12\lambda_0$ $\times$ $0.064\lambda_0$, where $\lambda_0$ is the wavelength at lowest resonant mode ($\approx 2.4$ GHz), adopting the optimised value of $h_2$, $8$ mm.

### 6.3 Parametric study results

The impedance bandwidth was the main target to be optimised throughout the parametric study (defined at reflection coefficient $|S_{11}| < -10$ dB). Each simulation was run with only one parameter varied, while other parameters stayed unchanged and were held at previously determined optimum values. The fixed and variable parameters are shown in Figure 6-1, where the variable parameters are considered as critical in determining the lowest and highest frequencies of the operating bandwidth. Since the
impedance bandwidth of the proposed antenna is heavily determined by the strong mutual coupling between the two radiating elements, simple mathematical formulae cannot be found to calculate this effect analytically and hence the EM simulator HFSS [24] was employed to perform this study.

Figure 6-1: Geometry structure of the proposed antenna
Figure 6-2: Simulated reflection coefficients $|S_{11}|$ with variation of different parameters
(a) $d1$ (b) $d2$ (c) $h1$ (d) $h2$ (e) $ws$ (f) Ground plane size
6.3.1 Effect of the position of the rectangular strip feed point (d1)

The position of the rectangular strip feed point (d1) is the one of the most sensitive parameters because its variation can introduce different excitation modes to the antenna, as illustrated in Figure 6-2(a). When d1 is 0.5 mm, that is, 0.5 mm from the free end of the ‘tail’ of the ‘F’, the antenna exhibits a wide-band impedance response from 2.65 to 8.0 GHz, but only with unsatisfactory impedance matching of $|S_{11}| < -25$ dB. When d1 is increased to 5.5 mm, a broader impedance bandwidth (2.48–8.0 GHz) can be obtained, but with the same poor impedance matching condition. As d1 is increased to 10.5 mm, wide impedance bandwidth can be attained from 2.4 to 7.5 GHz with good impedance matching of $|S_{11}| < -10$ dB, which completely covers the WiFi 2.4, 5.2 and 5.8 GHz bands, the WiMAX 2.5, 3.5 and 5.5 GHz bands and lower band UWB (3.1– 4 GHz). However, by further increasing d1 to 10.5 and 15.5 mm, unacceptable impedance mismatch is presented over the desired operating frequency band.

6.3.2 Effect of the position of the variable shorting wall of parasitic PIFA (d2)

The position of the shorting wall of the parasitic PIFA controls the lower resonant mode of this antenna. As can be seen in Figure 6-2(b), by varying the position of the shorting wall (d2) from 1.25 to 5.75 mm from the tail of the F, with 1.5 mm increment, that is, gradually reducing the distance between the two shorting walls of the parasitic radiator,
the lower cut-off frequency could be varied from 2.35 to 2.5 GHz. Scrutinising the higher cut-off frequency within the frequency band of interest, it can be seen d2 changed from 1.25 to 4.25 mm.

However, further increasing d2 to 5.75 mm, the upper cut-off frequency shifts to a lower frequency (6.4 GHz) this is because of the coupling effect, as two shorting walls are then close to each other. It is worth noting that the required upper cut-off frequency for the present applications (around 6 GHz) is seen to be largely independent of d2.

6.3.3 Effect of the height of driven PIFA (h1)

The height of the F-shaped radiator has a similar effect to d2 in influencing the fundamental resonant mode of the proposed antenna, but it also contributes capacitive coupling with the ground plane. As can be noted from Figure 6-2(c,) with the smallest value of h1 (2 mm), substantial mismatching of impedance bandwidth at both the higher and lower a resonant mode was observed. When h1 is increased from 2 to 8 mm in steps of 2 mm, significant variations of the impedance bandwidth and impedance matching were noted over the whole desired frequency spectrum. It was observed that when h1 was 6 mm the antenna offered optimum impedance bandwidth as well as good impedance matching. Therefore it is expected that h1 can take values between 5 and 7 mm for further reengineering purposes for different commercial applications, without degrading the performance of the antenna.
6.3.4 Effect of height of parasitic PIFA (h2)

The influence of parameter $h2$ is very much like that of $h1$, as depicted in Figure 6-2(d). By varying the height of the parasitic PIFA, both the lower and upper cut-off frequencies move simultaneously. As can be noticed, when $h1$ is at its shortest dimension (2 mm), the antenna suffers from poor impedance matching in the low operating frequency band (2.4–3.2 GHz), but it shows reasonable matching from 3.2 to 5.8 GHz. Nevertheless, as $h1$ becomes longer (5, 8 and 11 mm), the required

Figure 6-3: Practical prototype of proposed antenna
Figure 6-4: Measured and simulated reflection coefficients $|S_{11}|$, for optimised parameters $d1 = 11.1$, $d2 = 5.5$, $h = 5.3$, $h2 = 8$ and $ws = 5$

Figure 6-5: Simulated and measured antenna boresight gains
Figure 6-6: Simulated and measured normalised radiation patterns of the proposed antenna for two planes (left: x-z plane, right: y-z plane) (a) 2.4 GHz  (b) 3.0 GHz  (c) 4.4 GHz  (d) 5.2 GHz  (e) 5.8 GHz ‘xxxx’ simulated cross-polarisation ‘oooo’ simulated co-polarisation ‘- - - -’ measured cross-polarisation ‘——’ measured co-polarisation

low resonant frequency mode is progressively being excited to cover the desired 2.4 GHz frequency with good impedance matching and bandwidth. Hence, by taking a compromise between antenna size and bandwidth, it is recommended that $h_2$ can take any value between 8 and 10 mm for ideal performance.
6.3.5 Effect of the width of the slot in the shorting wall

The width of the slot in the vertical shorting wall of the driven PIFA governs the impedance matching to cover the designated working frequency bands, particularly for the lower operating frequency mode. The width of the slot (ws) was chosen to be altered from 0 to 6 mm in increments of 2 mm. As illustrated in Figure 6-2(e), without the presence of the slot, unsatisfactory impedance matching over the frequency band is present. However, increasing the width (ws) improves impedance matching.

6.3.6 Effect of the ground plane size

The ground plane length determines the major dimension of the antenna. The size of the ground plane was varied from 30 mm × 15 mm to 120 mm × 120 mm, the antenna being located in the corner of the ground plane, as this is the practical reality for the majority of commercial designs. The results are presented in Figure 6-2(f): it is seen that the ground plane size has a major effect on the reflection coefficients |S_{11}|, but an adequate performance is achieved with the smallest ground plane 30 mm × 15 mm ≈ 0.24λ₀ × 0.12λ₀, where λ₀ is defined at 2.4 GHz). This makes the design particularly suited to miniature radio device applications. When the ground plane size is larger, a wide impedance bandwidth characteristic with an upper bound outside the measurement range is displayed and the ground plane thus appears to cause the moving of the entire impedance bandwidth to higher cut-off frequencies, but with no improvement in
reflection coefficient $|S_{11}|$ and with a lower cut-off frequency that is higher than the desired 2.4 GHz. As the ground plane becomes smaller than a quarter wavelength, for example, 30 mm × 15 mm ($\approx 0.24 \lambda \times 0.12 \lambda$), the resonances shift downwards such that they fall into the required bands (2.4–6.0 GHz) with good impedance matching. According to [25, 26], this counter-intuitive effect is because the ground plane resonates around the centre frequency of the antenna element, and so the bandwidth of the antenna–chassis combination will improve considerably. On the contrary, if the ground plane resonates far away from the operating frequency, the bandwidth will deteriorate owing to its contribution becoming insignificant.

### 6.4 Results and discussion

In the parametric study process, a set of optimised geometry parameters has been identified as offering a broadband impedance matching response for the proposed antenna, where $d_1 = 11.1$ mm, $d_2 = 5.5$ mm, $h_1 = 5.3$ mm, $h_2 = 8$ mm, $w_s = 5$ mm and the ground plane size is minimised at 30 mm × 15 mm. An experimental prototype of the proposed antenna was fabricated, as depicted in Figure 6-3, to verify the simulated results. The measured and simulated (using HFSS [24]) reflection coefficient $|S_{11}|$ results are shown in Figure 6-4: these exhibit reasonable agreement although there is a frequency shift that can be attributed to fabrication errors in constructing this small antenna. As can be seen, the impedance bandwidth of the antenna encompasses the operating frequency spectra from 2.4 to 6.2 GHz for a reflection coefficient $|S_{11}| < -10$
dB, which corresponds to 3.8 GHz bandwidth or about 88.4% relative bandwidth with respect to the centre frequency 4.3 GHz. The bandwidth achieved fully covers the frequency spectrum of WLAN (IEEE 802.11a/b/g, 2.4/5.2/5.8 GHz), WiMAX (2.5/3.5/5.5 GHz) and the uplink UWB radio band (3.1–4.8 GHz). The simulated and measured gains of the designed antenna in the broadside direction over the frequency range from 2.4 to 6.0 GHz are shown in Figure 6-5. Again, there is some fluctuation because of fabrication errors but it can be observed that a practically useful average gain of 2.95 dBi was measured with ±1.0 dBi of gain fluctuation.

Measurements of the far field radiation patterns of the prototype antenna array were performed in a 100 m$^3$ anechoic chamber using an elevation-over-azimuth positioner, with the elevation axis coincident with the polar axis ($\theta=0^\circ$) of the antenna’s co-ordinate system. The azimuth drive thus generated cuts at constant $\phi$. The fixed antenna (reference antenna) was a broadband horn (EMCO type 3115) positioned at 4 m. The elevation positioner was rotated from $\theta = -180^\circ$ to $180^\circ$ in increments of $5^\circ$ for the selected measurement. Two pattern cuts (i.e. x-z and y-z planes) were recorded at five selected operating frequencies: 2.4, 3.0, 4.0, 5.2 and 5.8 GHz, covering the whole of the designated bandwidth in this study. The results are presented in Figure 6-6, which shows that the radiation patterns are stable and consistent at all of the designated frequencies. It can be noticed that the nulls exhibited in the patterns over the yz-planes, might be because of the consequence of the feed mechanism of the driven PIFA.
More importantly, it indicates that the maximum co polarised component appears in the boresight direction (+z) both for $E$- and $H$-planes and the simulated and measured co-polar radiation patterns are in good agreement with each other (the cross-polar components show more disagreements but they are weak and effectively noise like).

For implementation of the proposed antenna in a UWB system, that is, in particular, impulse-based systems, the shape of the transmitted electrical pulse should not be distorted by the antenna. Thus, a stable group delay response is desirable, which requires a highly linear phase response with respect to frequency. The measured group delay between two identical antennas of this type is portrayed in Figure 6-7. The variation is less than 0.5 ns over the frequency band from 3.1 to 4.8 GHz. The excellent pulse-handling abilities, with small pulse distortion, of the proposed antenna are thus demonstrated.

![Figure 6-7: Measured group delay](image_url)
6.5 Conclusions

A dual-planar inverted F-antenna of small size, covering the operating bands of WLAN/WiFi, WiMAX and the lower band UWB wireless standard has been presented. The antenna design concept was capable of covering 88.4\% relative impedance bandwidth with acceptable reflection coefficient $|S_{11}|$. The antenna has shown consistent omni-directional radiation patterns and reasonable gain values across the operating bands. The experimental results show satisfactory performance and good agreement with the computed results. With its broadband characteristic, the proposed antenna is very well suited to multi-band wireless applications.
6.6 References


CHAPTER 7

Planar Inverted FF Antenna for WLAN and UWB Applications

7.1 Introduction

The theme of the previous chapter is now considered with greater refinement. Once again we examine the properties of the PIFA type structure for UWB operations [2-6, 8]. Here we look at the need to avoid interference with other wireless channels within a general wideband, possibly multi-standard mode of operation. Again the emphasis is on mobile user terminals. For an illustrative example we look to the WLAN upper band frequencies, 5.2 GHz (operating over 5150 MHz to 5350 MHz) and 5.8 GHz (operating 5725 MHz to 5825 MHz). Some UWB designs have been reported for operation over this regime [9, 10]. However the UWB frequency range can cause interference with co-existing wireless channels, which in our case suggests the use of band notched PIFA structures. Some antennas have been proposed with a rejection band or band notch in the frequency range of 5 GHz to 6 GHz through the inclusion of a slot modification to the radiator [11-14]. Most of these designs do not attempt to address the bandwidth enhancement of the notch, or multi-band notches.
A miniaturised planar double inverted F antenna structure (PIFFA) is attempted, by controlling the geometry of the coupling distance between two PIFAs on an air substrate and the position of a broadband rectangular feeding plate. The final design is optimised with an antenna volume of $3 \times 15 \times 8 \text{mm}^3$. Details of this design and the results of the realised prototype assembly are discussed. The simulated and measured gain and radiation patterns are given also, to fully describe the antenna performance. In addition, a PIFFA with a spiral slot is attempted. The modelling approach is based on the previous chapter, using frequency domain analysis for generating systematic parameter tables (HFSS), and time domain analysis (SEMCAD) for verifying final performance of the final candidate structure.
7.2 First Design; PIFA Antenna for WLAN and UWB Applications

7.2.1 PIFA Antenna Design

The geometry of the proposed antenna (PIFFA) assembly may be visualised through Figure 7-1. The dimensions of the driven and parasitic PIFA elements are 18.5mm × 10mm × 4.7mm and 7.5mm × 14.5mm × 7.5mm, respectively. The antenna assembly is mounted on a finite ground plane whose dimensions are 30mm×15mm. The separation between the PIFA elements is set at 4mm, for optimal coupling. The conventional PIFA wire feed mechanism is replaced by a rectangular plate feed to the driven element, to improve the bandwidth performance. The resonant frequencies of both PIFA elements is approximated from [1]

\[ f \approx \frac{c}{4(W+L)} \]  \hspace{2cm} (7.1)

Where \( W \) and \( L \) are the width and length of the element respectively, and \( c \) is the speed of light. It follows that the driven PIFA is resonant at approximately 3.2GHz, and the parasitic PIFA is resonant at approximately 4.86GHz.
The lower resonance is dominated by the driven patch, whilst the higher resonance is excited by the combined area of the two PIFA elements. The coupling gap is critical for both impedance bandwidth and the quality of impedance matching over the required frequency band. The size of the ground plane is also a significant factor in realising a good UWB impedance match.

### 7.2.2 Results and Discussion for PIFA Antenna

Simulated and measured reflection coefficients $|S_{11}|$ are compared in Figure. 7-2. The individual structures were analysed to determine their impedance bandwidths. This is a necessary step in understanding the individual contribution of the driven and parasitic antennas. For the driven PIFA, the impedance bandwidth covers the range 3.6GHz to 12GHz, with a reflection coefficient of $|S_{11}| < -6\text{dB}$. The corresponding range for the
parasitic PIFA is 4.8GHz to 9.2GHz, for the same reflection coefficient. Constructive addition of these impedance bandwidths enables a good impedance match over 3.0GHz to 10.6GHz for $|S_{11}| < -10$dB.

Figure 7-2: Measured vs. simulated reflection coefficients $|S_{11}|$ of the antenna
The effects of ground plane size on antenna return loss were investigated parametrically. The physical dimensions of the ground plane were scaled in terms of $\lambda_0$, the wavelength corresponding to the lowest resonant frequency of the structure. The response was analysed for five different ground plane sizes 

$$\{0.3\lambda_0 \times 0.15\lambda_0, 0.4\lambda_0 \times 0.4\lambda_0, 0.6\lambda_0 \times 0.6\lambda_0, 0.8\lambda_0 \times 0.8\lambda_0, 1\lambda_0 \times 1\lambda_0\}.$$

When these particular dimensions are greater than or equal to $0.4\lambda_0$, there is a slight impedance mismatch within the lower frequency component in the range 3GHz to 3.5 GHz.

Figure 7-3: Simulated reflection coefficients $|S_{11}|$ corresponding to the variation of ground plane size.
GHz. The reflection coefficient in this range is -6dB, which is still acceptable for commercial applications. Reducing the ground plane dimensions to $0.3\lambda_0 \times 0.15\lambda_0$, improves the impedance bandwidth of the lower frequency components; this may be due to the combined effect of the antenna geometry and the ground plane, with the dual mode operation.

Simulated and measured antenna gains are compared in Figure 7-4. These values vary between 3.75dB and 5.5dB over the total UWB band; with maximum variations in the neighbourhood of $\pm 1.5dB$, and a mean of 4dB. Both simulated and measured results are in sufficient agreement. Figure 7-4 compares the simulated and measured antenna radiation patterns of the prototype assembly at 3GHz, 5GHz, 7GHz and 9GHz.

![Figure 7-4: Simulated vs. measured antenna gains.](image-url)
These patterns are approximately omnidirectional and consistent at the lower frequencies, except for the yz-plane in Figures 7-5(a) and (b). In these cases the nulls predicted in simulation are not fully realised; this is believed to be due to subtle differences in the effects of the simulated vs. physical feeding arrangements. As the frequency increases, this omnidirectional characteristic distorts, as a consequence of the feed mechanism of the driven PIFA.
Figure 7-5: simulated vs. measured normalised antenna radiation patterns for two planes (left: x-z plane, right: y-z plane); at (a) 3000 MHz, (b) 5000 MHz, (c) 7000 MHz, (d) 9000 MHz: ‘xxxx’ simulated cross polarisation ‘oooo’ simulated co-polarisation ‘----’ measured cross-polarisation ‘___’ measured co-polarisation
7.3 Second Design: PIFA Antenna for UWB Applications
with WLAN Band Rejection

7.3.1 PIFA Antenna Design with Spiral Slot

The geometry of the proposed UWB spiral slot antenna characteristic is illustrated in Figure 7-6. It is fabricated on 18.5mm × 10mm × 4.7mm and 7.5mm × 14.5mm × 7.5mm, respectively. The antenna assembly is mounted on a finite ground plane whose dimensions are 30mm×15mm. Once again, the separation between the PIFFA elements is set at 4mm, for optimal coupling. The conventional PIFA wire feed mechanism is replaced by a rectangular plate feed with spiral slot of 3mm×3mm to control the bandwidth performance of the notch.

Figure 7-6: PIFA antenna structure with spiral slot
7.3.2 Parametric Study

Figure 7-7(a) shows the variation of the reflection coefficient $|S_{11}|$ for three selected ground plane sizes, it can be seen that the 30x15mm$^2$ dimension should be chosen as the optimised size for the ground plane in order to obtain the desirable frequency range, as shown in Figure 7-7(a). The simulated the return loss for the three different antenna heights is indicated in Figure 7-7(b). The notched frequency requirement was achieved at 4.2mm. The feed point is the most sensitive parameter; three different positions were selected to get a good result. From Figure 7-7(c) can be noted that 7.5mm is the best position for the feed point.
Figure 7-7: Simulated reflection coefficients $|S_{11}|$ with different dimensions of (a) ground plane, (b) antenna height and (c) feed position
7.3.3 Results and Discussion for PIFA Antenna with Spiral Slot

Simulated and measured reflection coefficient $|S_{11}|$ is illustrated in Figure 7-8. The individual structures were analysed to determine their impedance bandwidths. For the driven PIFA, the impedance bandwidth of $|S_{11}| < -10\text{dB}$ covers the range 3GHz to 13GHz, with band rejection from 4.75GHz to 5.75GHz for WLAN communication system.

Figure 7-8: Simulated reflection coefficients $|S_{11}|$ of the proposed antenna
The far field radiation patterns are presented in Figure 7-9. Two pattern cuts (the $xz$ and $yz$ planes) were taken at four selected operating frequencies which cover the aggregate bandwidth. The radiation patterns were found to be stable and consistent at the designated frequencies, as shown in Figure 7-9. Significantly, it also indicates that the maximum co-polarized component appears at the direction of bore sight (+z) for both the E and H planes.
7.4 Conclusions

A compact low profile antenna with a finite ground plane is proposed and realised in two ways. Various parameters of the antenna are optimise the performance of the antenna has great effect with respect to the size of the ground in term of reflection coefficient. This antenna constructively combines the resonant modes of the two PIFA elements. The antenna was fabricated and tested and achieved good results over wide bandwidth. The simulation and the measurement show good agreement. The slight shift in the return loss curve can be attributed to the fabrication inaccuracies. The antenna also provides consistent radiation patterns with the reasonable gain over the operating frequency bandwidth for realistic UWB radio applications.

The band-notch ultra-wideband antenna using spiral slot has presented with the notch over the frequency 5.25 GHz (4.75-5.75 GHz) various parameters of the antenna are optimise and various technique have been adopted to create the notch which include, making a spiral slot in the feeding plate. The antenna shows narrow notched bandwidth suppression ability in WLAN band. The set of parameter studies of the antenna provides brief guidelines for the band-notched antenna design. Evaluations of return loss and radiation pattern confirm the antenna performance in term of the HFSS-simulation. These features demonstrate that the proposed antenna is suitable for UWB communication applications and prevents interference from the WLAN system.
7.5 References


CHAPTER 8

A Band-Suppression UWB Suspended Planar Antenna
Incorporating A Slotted Spiral Resonator

8.1 Introduction

The nascent ultra-wideband (UWB) wireless standard is designed for short range (~10m), low emitted power (<-41.3 dBm/MHz) and high data rate (>480Mbps) indoor communication [1]. It covers the frequency spectrum from 3.1 to 10.6 GHz. This technology is capable of delivering 1Gbit/s data rates which is nearly double the rate of most of the existing wireless standards. However, interference is a serious problem for UWB systems owing to the low emitted power of the UWB system and its necessary coexistence with other narrowband wireless services, the most significant of which are the wireless networking standards, i.e. IEEE 802.11a (5.15–5.35 GHz and 5.725–5.825 GHz) in US and HIPERLAN/2 (5.15–5.35 GHz and 5.47–5.725 GHz) in Europe [2].

Due to intensive research into the implementation of UWB systems, various printed band-suppression UWB monopole antennas have been proposed in the recent years [3-16]. To achieve a band-suppression function without additional cost incurred in the antenna design, several novel design techniques have been studied: these methods
include introducing different shapes of parasitic strip radiator [3-7], adding a truncated corner on a printed inverted F-antenna [8], supplementing a square conductor-backed plane on a printed monopole antenna [9], implementing a multi-arm inverted-F monopole [10], electromagnetic coupling of a stub-loaded open-loop resonator on the underside of a CPW-fed mirrored-L monopole [11], applying ceramic chip radiating elements to the geometry of the antenna [12], and inserting band notching slits in the radiating structure of the antenna [13-16].

These design innovations have gained advantages of broad impedance bandwidth and low profile but they suffer from low gain, severe back lobes in the radiation pattern, excessive currents in the ground plane and high cross-polarization radiation levels. Recently, several suspended plate antennas (SPAs) [17-20] have been proposed to overcome these drawbacks for UWB applications. However, since the envelope size of these antennas is physically large in comparison with printed antennas and limited space is available in the geometry of the antenna, the band-suppression feature is challenging to implement on this type of antenna. In this chapter, a novel UWB planar inverted F-L antenna (PIFLA) with a band-notched in a form of slotted spiral resonator is proposed and investigated. The spiral resonator is implanted on the feeding plate of the antenna to suppress the unwanted HYPERLAN/2 band (5.15-5.725 GHz) without compromising the UWB PIFLA antenna response. In this chapter will initially discuss the basic operation of the modified PIFLA antenna, then, it is followed by introducing the band-rejection technique, the results for both standard UWB and band-rejection version of
UWB antennas are demonstrated and the computed and experimental results of the antenna, a conclusion is drawn.

A novel miniaturized planar inverted F-L antenna assembly is considered for UWB radio operations. The antenna design utilizes the electromagnetic coupling between an air dielectric planar inverted-F antenna (PIFA) and a parasitic planar inverted-L (PIL) element, with broadband feeding from a rectangular plate. In order to improve the functionality of the channel, a simple notch filter has been introduced through a local modification to the broadband feed plate, this takes the form of a simple slotted rectangular spiral resonator which is etched directly onto the plate. This allows the proposed antenna to maintain its full band UWB coverage, with the HYPERLAN/2 band centred at 5.35 GHz to be effectively rejected over the sub-band 5.15 GHz – 5.725 GHz, without the need for substantial re-optimisation of its principal structure parameters. The impedance bandwidth operates over the full UWB band, with VSWR better than 2, this performance is not degraded by the presence of the band rejection. The observed gains, radiation patterns and group delay confirm that the antenna has appropriate characteristics for short range wireless applications.
8.2 Antenna Design Concept and Structure

The geometry of the proposed planar inverted F-L antenna (PIFLA) with single band-notch characteristic is shown in Figure 8-1. The antenna design concept is quite similar to the previous work reported [19-20] but the geometry parameters have been further optimised to cover the full UWB frequency band instead of the lower-band of UWB only as in [19-20].

Moreover, the suspended rectangular feeding plate was modified by engraving a slotted spiral resonator on it to provide the band-suppression characteristics. It should be noted that this proposed method offers a very simple band suppression design concept to the antenna designer without further optimising the antenna geometry parameters.

The dimensions of the driven PIFA are $18.3 \times 10 \times 5 \text{ mm}^3$, while those of the other parasitic PIL element are about $6 \times 14.5 \times 7.5 \text{ mm}^3$. Both of them are mounted on a $30 \times 15 \text{ mm}^2$ finite ground plane. For optimal coupling, the separation distance between the two radiating elements was found to be around 2.7 mm. A suspended broadband rectangular feeding plate with feeding gap of 0.5 mm is used to excite the driven PIFA, to achieve the lowest resonance frequency while the higher resonant frequency is excited by the combination of the driven and parasitic elements. Therefore, the coupling gap between the two suspended element planes is critical and has to be chosen with care, otherwise only narrow impedance bandwidth or bad impedance matching will be
observed over the required frequency band. In addition, the size of the ground plane also plays a vital role in this antenna in aiding achievement of good UWB impedance matching, especially at the lower band-edge frequency.

Figure 8-1: Geometry of the proposed antenna (a) Overall geometry, (b) Feeding plate
Figure 8-2 shows the computed and experimental return losses of the proposed PIFLA antenna without band-suppression. In this case, full rectangular feeding plate without presence of the slotted spiral is used to excite the proposed antenna model. As can be seen, the results cover the required UWB band from 3.1 to 10.6 GHz for return loss better than 10 dB. Due to fabrication error, a slight discrepancy can be noticed between the simulation and measurement results. By scrutinizing the simulation result, four adjacent resonant frequencies (showing a return loss ≥ 15 dB) can be observed over the 3.1 to 10.6 GHz frequency spectrum, i.e., 3.4 GHz, 6.0 GHz, 8.0 GHz and 10.4 GHz. In order to elaborate the operation of this UWB antenna in more detail, the vector plot of the surface current distributions at these resonant frequencies is investigated, as given in Figure 8-3. These plots give more insight into the contribution of the individual parts of this antenna at these frequencies, enabling adjustment to establish ultra-wide impedance bandwidth.

The first resonant frequency occurs at 3.4 GHz and the total length of the continuous current path is 44.2 mm which is ≈ 0.5λ (where λ is the free space wavelength). As can be seen, the current flows from the driven PIFA to the parasitic PIL element through the ground plane to excite this mode. Therefore, the major dimensions of these two elements and the separation distance between them can manipulate the first resonant frequency. This result also demonstrates that the ground plane has significant influence on the impedance matching at the lowest edge of operating frequency band, which
further confirms the finding in [19-20] that the $27 \times 15 \text{ mm}^2$ ground plane offers better impedance matching in comparison with other ground plane sizes.

![Graph showing return losses for UWB PIFLA antenna](image)

Figure 8-2: Simulated and Measured return losses for UWB PIFLA antenna

![Vector plot current distributions of the UWB antenna](image)

Figure 8-3: Vector plot current distributions of the UWB antenna, with major current flows indicated. (a) 3.4 GHz, (b) 6.0 GHz, (c) 8.0 GHz, (d) 10.4 GHz
At the second resonant frequency, 6.0 GHz, two discrete current paths are found on the driven PIFA and parasitic PIL element, both around 0.38\(\lambda\) which corresponds to \(\approx 19.5\) mm. It is not very obvious which element stimulates this resonant mode as both seem to contribute equally. However, it is believed that the mutual coupling distance between the two elements has considerable influence in determining this resonant frequency. By looking at the third resonant frequency, 8.0 GHz, a clear current path of length 19.5 mm is observed on the parasitic PIL element which is approximately 0.5\(\lambda\) at this frequency. In this case, it is apparent that total length of the parasitic PIF structure governs the third resonant frequency.

The last resonant frequency occurs at 10.4 GHz, as depicted in Figure 8-3 (d). The full length of the inner L-shaped structure of the driven PIFA is 14.7 mm, which is around 0.5\(\lambda\). This L-structure appears to control this resonant mode of the antenna. By scrutinizing all of these resonant modes, it can be seen that there is high current concentration on the shorting strip of the antenna at each of the resonant frequencies. Hence, it plays a pivotal role in enhancing the impedance matching of the proposed antenna for all desired frequency bands, particularly the 10.4 GHz resonant frequency band.

In order to introduce the band-stop characteristic into the proposed PIFLA antenna, a modified slotted rectangular spiral is etched on the suspended rectangular feeding plate as proof of concept. As can be seen in Figure 8-1, the slotted rectangular spiral is specified by the slot width, center-to-center spacing between the slots, and the number
of turns of the spiral, chosen to be 0.25 mm, 0.5 mm and 3 respectively. With these geometrical parameters, the total length of the slotted spiral resonator is found to be approximately $\lambda/4$ at the band-stop centre frequency 5.35 GHz. The resonant behavior of the slotted spiral is caused by the interaction between the spiral inductances, the distributed capacitance and the mutual inductance between the loops of the spiral. There is one paramount geometrical parameter, $L_1$, which is found to be critical for acquiring the desired band-suppression behaviour without deterioration of the impedance matching over the rest of the UWB frequency band, (where $L_1$ is the location of the slotted spiral on the feeding plate).

### 8.3 Results and Discussion

Figure 8-4 depicts the physical prototype of the proposed antenna which was fabricated by using 0.5 mm thickness of copper sheet, except the feeding plate. Due to the difficulty of etching 0.25 mm slots in copper sheet to form the spiral shape, the feeding plate was constructed using a low loss flexible substrate. In this case Arlon ‘FoamClad’ material [21] was used: the permittivity of this material is close to air, varying between 1.15 and 1.30. The thickness of the copper and the substrate are 36 $\mu$m and 1.6 mm respectively.
Figure 8-4: Practical prototype of the proposed antenna

Figure 8-5 illustrates the measured and computed antenna performance in term of voltage standing wave ratio (VSWR). The impedance bandwidth for VSWR < 2 is 114.3% extended from 3.0 GHz to 11 GHz, fully covering the frequency spectrum of the UWB standard. It also shows very effective suppression of the HYPERLAN/2 frequency band for both the simulation and the measurement results. The measured and simulated notched bands at VSWR > 2 are around 4.95 to 5.73 GHz and 4.90 to 5.75 GHz respectively. In addition, these bands at VWSR > 3, are around 5.0 to 5.64 GHz measured and 4.95 to 5.67 GHz simulated.
A parametric study was conducted to provide further useful information on the band-notch characteristics of the proposed antenna for the antenna designer. Hence, two geometric parameters were considered, these are the feeding plate length $L_1$ and the total length of the slotted spiral. The investigation was carried out in relation to the variation of VSWR as each parameter changes while others were kept constant.
Figure 8-6: Variations of geometry parameters against the VSWR over the UWB bands, (a) L1 and (b) Total length of spiral.
Figure 8-6 (a) shows results for the variation of the horizontal position of the slotted spiral loop on the feeding plate (L1) against VSWR. As can be seen, L1 is increased in 1 mm steps from 1 to 4 mm in this analysis. For VSWR = 2, as L1 increases from 1 to 4 mm, the centre frequency of the suppressed frequency band reduces slightly in frequency from 5.31 to 5.13 GHz, but the notch bandwidth becomes wider, from 4.96 – 5.65 GHz to 4.16 – 6.1 GHz, corresponding to 24.8% of bandwidth expansion without interfering with the impedance matching over the lower-edge and higher edge UWB operating frequency band. It can be noticed that the length of L1 should be between 1 to 2 mm to achieve an optimal band rejection between 5.15 and 5.725 GHz. For better understanding the variations of the band-notch, Table 8.1 summarizes the rejection bandwidth defined at VSWR=2 and VSWR=3 when L1 is varied from 1 mm to 4 mm.

Table 8-1: Computed rejected frequency band corresponding to the horizontal position of the slotted spiral loop on the feeding plate L1 when VSWRs are 2 and 3.

<table>
<thead>
<tr>
<th>Length</th>
<th>VSWR = 2</th>
<th>VSWR = 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>L1 = 1</td>
<td>4.96 ~5.65</td>
<td>5.07~5.59</td>
</tr>
<tr>
<td></td>
<td>(BW=13%, f_c=5.31)</td>
<td>(BW=9.8%, f_c=5.33)</td>
</tr>
<tr>
<td>L1 = 2</td>
<td>4.69 ~ 5.76</td>
<td>4.9~5.64</td>
</tr>
<tr>
<td></td>
<td>(BW=20.5%, f_c=5.23)</td>
<td>(BW=14%, f_c=5.271)</td>
</tr>
<tr>
<td>L1 = 3</td>
<td>4.4 ~ 5.9</td>
<td>4.7~5.75</td>
</tr>
<tr>
<td></td>
<td>(BW=29%, f_c=5.15)</td>
<td>(BW=20%, f_c=5.23)</td>
</tr>
<tr>
<td>L1 = 4</td>
<td>4.16 ~ 6.1</td>
<td>4.47~5.86</td>
</tr>
<tr>
<td></td>
<td>(BW=37.8%, f_c=5.13)</td>
<td>(BW=27%, f_c=5.17)</td>
</tr>
</tbody>
</table>
Table 8-2: Computed rejected frequency band corresponding to the total length of the slotted spiral when VSWRs are 2 and 3.

<table>
<thead>
<tr>
<th>Spiral geometry</th>
<th>VSWR = 2</th>
<th>VSWR = 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Proposed</td>
<td>4.95 ~ 5.73 (BW=14.6%, f_c=5.35)</td>
<td>5.03~5.64 (BW=11.4%, f_c=5.34)</td>
</tr>
<tr>
<td>Modified 1</td>
<td>5.0 ~ 6.0 (BW=18.2%, f_c=5.50)</td>
<td>5.24~5.9 (BW=11.85%, f_c=5.57)</td>
</tr>
<tr>
<td>Modified 2</td>
<td>5.14 ~ 6.35 (BW=21%, f_c=5.75)</td>
<td>5.5~6.25 (BW=12.7%, f_c=5.875)</td>
</tr>
</tbody>
</table>

The effect of the total length of the spiral resonator is investigated in Figure.8-6(b). This study was performed by progressively reducing the total length of the spiral. It is simply done by filling up the gap between the spiral loops from the most inner loop. Reducing the spiral length by 1 mm and 1.5 mm, two different spiral geometries might be considered, these are ‘Modified 1’ and ‘Modified 2’. It can be observed that the centre of the notch-band frequency for ‘Modified 1’ and ‘Modified 2’ can be shifted from 5.35 GHz to 5.75 GHz whereas the notch bandwidth is enhanced by 6.4%, corresponding to a frequency band of 4.95-5.73 GHz to 5.14-6.35 GHz respectively, defined at VSWR = 2. It also found that the upper and lower resonant frequencies of the UWB antenna remain undisturbed when this parameter is varied. In addition, it was interesting to notice that when the notched centre frequency is shifted to 5.5 GHz (referring to ‘Modified 1’ curve), the suppressed frequency band occupies the 5 to 6
GHz band which indicates the capability of rejecting the 802.11a band (5.15 to 5.825 GHz). Therefore, it is suggested that this new spiral geometry can be also applied to the proposed UWB antenna for rejecting the interferences generated by 802.11a wireless standard. Table 2 shows the notched band of the proposed antenna while this geometry parameter is changing for the case of VSWR = 2 and 3.

Figure 8-7 plots the measured gain of the standard PIFLA UWB antenna and the proposed band-suppression version in the broadside direction, over the interval 3.1 GHz to 10.6 GHz. As can be seen, the variation of the UWB antenna gains is about ±1.75dBi, in which the maximum and minimum gains are 3.75 and 5.5 dBi respectively, while an average measured gain of 3.6 dBi with ±0.5 dBi fluctuation is found from the proposed antenna. It was interesting to notice that the gain of the antenna decreases drastically at the suppression centre frequency of 5.35 GHz, falling as low as -5.8 dBi.

![Figure 8-7: Measured gain of the standard and band-suppression UWB antenna](image-url)
Figure 8-8 depicts the simulated and computed radiation patterns of the proposed antenna at 4, 6 and 8 GHz. The antenna is seen to be generally omnidirectional and consistent at lower frequencies. As the frequency increases, the omnidirectionality is distorted; believed to be due to subtle differences in the effects of the simulated versus the physical feeding arrangements.

Figure 8-8: Radiation patterns of the proposed antenna for x-z plane (left) and y-z plane (right) at 4 GHz, 6 GHz and 8 GHz (‘ooo’ represents simulated $E_{co}$, ‘——’ measured $E_{co}$, ‘xxx’ simulated $E_{cross}$ and ‘----’ measured $E_{cross}$)
In UWB application, a stable group delay over the frequency spectrum is important for achieving consistent and undistorted transmission of a pulse by the antenna. The measured group delay between two identical proposed band-suppression antennas is shown in Figure 8-9. In comparison with the standard UWB antenna (without band suppression), the group delay variation of the proposed antenna holds the same fluctuation, which is less than 0.5 ns over the UWB operating frequency band, except over the stop band, where the variation increases considerably: up to 4.5 ns at the centre frequency (5.35 GHz).
The current distributions on the proposed antenna are illustrated in Figure 8-10 at 3, 5.35, 7 and 9 GHz. As can be easily discerned, the majority of strong electric currents are concentrated only around the slotted spiral at 5.35 GHz, in comparison to 3, 7 and 9 GHz. This is due to the proposed antenna acting as a band-stop filter to suppress the unwanted 5.15- 5.725GHz HYPERLAN/2 band so; the proposed antenna does not radiate (or receive) in this band. This explanation can be validated by the measured gain of this antenna.

Figure 8-10: Contour plot current distributions of the proposed antenna. (a) 3GHz, (b) 5.35 GHz, (c) 7 GHz, (d) 9 GHz
8.4 Conclusion

An electrically small, low profile, UWB antenna assembly, with an inbuilt HIPERLAN/2 band-rejection characteristic has been presented. The HIPERLAN/2 rejected band centered at 5.35 GHz was selected for this application. The design procedure considered the advantage of the coupling between an air-dielectric PIFA and PIL parasitic elements mounted over a finite ground plane. The band-stop response was achieved by embedding slot-spiral resonator on the broadband feed plate. This has the advantage of utilizing a well-defined antenna structure, and does not require substantial re-optimisation of the primary structural parameters defining the whole antenna geometry. This design allows the antenna to operate over the full UWB spectrum including HIPERLAN/2, without the need for additional specialized filtering. The antenna has demonstrated sufficient impedance bandwidth, suitable radiation characteristics, and adequate gain for UWB applications.
8.5 References


CHAPTER 9

A Low-Profile Ultra-Wideband Modified Planar Inverted-F Antenna

9.1 Introduction

Ultra-wideband (UWB) communication systems have several attractive features when compared with traditional narrow band communication systems. These key features include low maximum emitted power (<75nW/MHz), high data rate (>500Mbps) over short ranges (<10m), and large channel capacity which offers up to 7.5 GHz bandwidth over the 3.1 to 10.6 GHz frequency spectrum [1]. This has enabled reliable high-speed real-time high definition video streaming and multimedia file sharing in an indoor environment. In order to make UWB systems competitive with other narrow band communication systems, antennas with compact size, wide impedance bandwidth and consistent omni-directional radiation patterns are in unabated demand.

Due to the low profile, easy integration with circuitry and simple design, printed monopole antennas [2-4] have been widely proposed and studied over last decade. In general, these designs achieve broad bandwidth by electromagnetically coupling a radiator with optimised geometry to a defected ground plane with optimum dimensions...
on a dielectric substrate. However, these design innovations suffer from ground plane size sensitivity, severe backlobes in the radiation pattern and low, inconsistent gains. These drawbacks have limited these antennas to operation in unidirectional communication applications. To rectify this inadequacy, many planar or 3D metal plate monopole antennas [5-17] have been proposed, with associated new bandwidth enhancement techniques. These novel methods include modifying the feeding plate silhouette [5-6], inducing mode coupling between two closely placed elements [7-10], shorting and bevelling the planar structure [11], offsetting the feeding point from the centre of the structure [12], top-loading the structure with a circular disc patch [13], folding the planar structure into a 3D structure without dielectric [14] and with dielectric [15], gap loading [16] and dielectric loading [17]. Among these designs, it was found that some geometries of these antennas are complicated and difficult to fabricate [7-10,13-15,17], some antennas suffer from ground stability issues [7,9-10] and some antennas have a relatively large height above the ground (>0.1λ₀) [5-6,11,13,16-17]. These limitations have made such reported works [5-17] less favourable to be adopted as a commercial product.
In this chapter, a miniaturized low-profile modified planar inverted-F antenna (PIFA) is proposed and investigated. The multi-resonant characteristic of this antenna is realized by combining multiple bandwidth improvement methods including shorting wall, asymmetrical feeding point, broadband feeding plate and top loading. By implementing these techniques, this antenna exhibits a wide impedance bandwidth from 3 to 13 GHz with acceptable radiation characteristics. The detailed evolution procedures and operating principles will be addressed; the geometry parametric studies and measured results will be discussed and conclusions are drawn in this chapter.
9.2 Antenna design Concepts and Structure

The geometric configuration of the proposed antenna is shown in Figure 9-1. The antenna is constructed as a planar inverted F-shaped radiator with a planar rectangular parasitic strip which is parallel to the feeding plate of the radiator, electrically attached to the top plate, and not grounded. The structure and dimensions of the antenna were chosen on the basis of previous experience [7], combined with realistic constraints on acceptable size for incorporation in a mobile terminal: the dimensions were then optimised in parametric studies. A suspended broadband rectangular feeding plate [7] with optimised feeding gap of 0.75 mm is implemented to excite this antenna assembly and enable its multiple-resonance characteristic. As can be seen, the optimised
dimensions of this antenna’s radiator are $20 \times 10 \times 7.5 \text{ mm}^3$ which correspond to electrical dimensions of $0.2\lambda_o \times 0.1\lambda_o \times 0.075\lambda_o$, where $\lambda_o$ is defined at the lower edge operating frequency (LEOF) of the structure, taken as 3 GHz. For ease of integration with a practical transceiver enclosure, this antenna is mounted on a corner of a $50 \times 50 \text{ mm}^2$ finite ground plane. 0.5 mm thickness copper sheet was used for all conductors in this design.

Because of its aspect ratios, this antenna cannot be absolutely classified as a PIFA and it could be viewed as intermediate between this class and a folded dipole with top capacitive loading. It has some similarities to designs previously published [7,9-10,18]. In [18], the antenna uses a square patch on FR4 substrate as top loading for a rectangular monopole patch and two identical rectangular patches connecting the loading patch to a $50 \times 50 \text{ mm}^2$ ground plane. The total dimension of this previously reported antenna assembly is $50 \times 50 \times 10 \text{ mm}^3$ which is comparable to the presently proposed model, but only achieves a bandwidth of 8.7% (with a center frequency of 2.537 GHz) which is not sufficient to cover the entire UWB spectrum. In [7], the ultra-wide bandwidth is realized by controlling the multi-resonance characteristics of two closely coupled PIFAs. In other work [9], a similar working principle is applied but with a different positional arrangement of the two coupled radiators and an inverted L-parasitic element is used instead of a parasitic PIFA. Recently, work in [10] has further improved the impedance bandwidth of the structure in [9] by inserting an additional rectangular-shaped parasitic element. Comparing the size of the antennas in [7, 9-10], it
was found that the antenna in [7], including its ground plane, is smaller at $30 \times 15 \times 7.5$ mm$^3$ and it covers the operating frequency band from 3 to 10.6 GHz; in [9-10] both antenna dimensions including the ground are $28 \times 18.5 \times 4.5$ mm$^3$ and these can operate from 3 to 9.6 GHz [9] and 3.4 to 10.7 GHz [10] respectively. However, all of these antennas are very sensitive to variation of ground plane size in the lower part of the usable operating band, from 3-4 GHz for [7] and 3-5 GHz for [9-10], and this is due to significant contribution of the ground plane resonance over these bands: this is problematic as the ground plane behavior will be heavily influenced by a human hand holding a mobile terminal.

![Figure 9-3: Simulated reflection coefficients $|S_{11}|$ of proposed antenna in a design evolution process based on Figure: 9. 2(a)-(d).](image-url)
Figure 9-4: Simulated input impedance of proposed antenna in the design evolution process from Figure: 2(a)-(d).

The proposed antenna model combines different impedance bandwidth enhancing techniques together to enable coverage of the entire FCC UWB frequency band. Compared with the work in [7], the antenna structure (excluding the ground) has been
simplified and miniaturized, the impedance bandwidth has been widened and the ground size dependency of the previous designs has been minimized.

To understand the basic operational principles of this antenna, the structure has been split into the four differently sized rectangular metal plates which constitute the assembly. This will help to simplify the analysis of the contribution of each metal plate to the antenna in terms of impedance bandwidth and matching over the intended UWB operating frequency band (3.1 to 10.6 GHz). Figure 9-2 (a) to (d), illustrate the detailed design evolution of the proposed antenna from a planar monopole to a modified PIFA, while Figure 9-3 and Figure 9-4 show the corresponding performance of the antenna geometries, in terms of reflection coefficient and input impedance respectively.

This analysis starts by considering a rectangular metal plate with dimension of 6.75 × 14 mm² and the off-centre probe feed with a feeding gap of 0.75 mm which is used to excite this structure. This forms a conventional planar monopole antenna, as shown in Figure 9-2(a). To analytically estimate the LEOF of this antenna, the following simple formula based on the equivalent cylindrical monopole, as described in [19], is adopted.

\[
F(\text{GHz}) = \frac{75}{L + h + \frac{W}{2\pi}}
\]  

(9.1)

where L, W and h are the length, width and feeding gap distance of the planar monopole respectively.
Substituting the geometrical parameters \((L=6.75, \ W=14\) and \(h=0.75)\) in the above expression, it is found that the LEOF is predicted as 7.7 GHz. To validate this estimate, Figures 9-3 and 9:4 depict the numerical predictions of \(|S_{11}|\) and input impedance. As can be observed in Figure 9-3, the lower and higher edge operating frequencies (HEOF) are 8.2 GHz and 10 GHz respectively, with equivalent impedance bandwidth 19.8%, for a required \(|S_{11}| \leq -10\text{dB}\). Examining the response at 7.7 GHz, the corresponding \(|S_{11}|\) is around -6 dB which appears consistent with the estimated value using the formula (for which the LEOF return loss criterion was not stated explicitly). In the input impedance plot of Figure 9-4, the antenna exhibits a typical monopole antenna’s impedance curve response with optimum 50 ohm impedance matching at 8.7 GHz, and a parallel resonance occurs at 8.3 GHz.

To reduce the LEOF of the antenna without increasing the height of the monopole component, this is next top-loaded with a 20 x 10 mm\(^2\) rectangular capacitive plate to form a T-shaped antenna structure, as shown in Figure 9-2(b). This is also found to improve the bandwidth; as can be seen in Figure 9-3, the impedance bandwidth at the -10 dB points is now about 81% and encompasses the frequency band from 5.5 GHz to 13 GHz. This shows a 61.2% improvement of impedance bandwidth in comparison with the monopole. To further explain the physical contribution of the additional top-loading plate, Figure 9-4 exhibits the input impedance of this antenna. As can be observed, introducing the top-loading plate causes the single parallel resonance of the original monopole to split into two parallel resonances at 6 GHz and 11.6 GHz with associated resistances of 50 ohm and 83 ohm respectively, and with less variation of reactance (10
to 30 ohm) over the operating frequency band. Carefully tuning of the dimensions of this top-loading plate can bring these two parallel resonances close to each other and leads to good impedance matching over a wide bandwidth.

![Figure 9-5: Current distributions of the proposed antenna. (a) 3.4 GHz, (b) 6.6 GHz, (c) 9 GHz, (d) 12.5 GHz](image)

To achieve the LEOF required of a UWB antenna, i.e. 3.1 GHz, this antenna is further modified by shorting one edge of the top loading plate to the ground using a 5 × 7.5 mm² metal plate, as illustrated in Figure 9-2(c). With this modification, the antenna partially resembles a PIFA and the impedance bandwidth has further improved to cover 3.0 to 13 GHz for a \(|S11|\) better than -8 dB, as shown in Figure 9-3. Investigating the
input impedance of this antenna is shown in Figure 9-4. It is evident that this shorting wall increases the resistance and reduces the reactance of the antenna in the lowest usable frequency band of the previous antenna model.

Scrutinizing the $|S_{11}|$ plot in Figure 9-3, most of the operating frequency band now meet the criterion of $|S_{11}| < -10$ dB, except the band from 9.8 to 11.5 GHz which is $< -8$ dB. In order to enhance the impedance matching of this band, an $8 \times 5.5$ mm$^2$ parasitic metal plate is placed in parallel with the feeding plate. This result is final modified antenna structure as shown in Figure 9- 2(d). The $|S_{11}|$ plot in Figure 9- 3 proves that this parallel parasitic metal plate effectively improves the impedance matching of the required band without a deleterious effect on matching of the other frequency bands. Observing the input impedance response of this antenna further clarifies that this parallel plate increases the resistance of the antenna and acts as a capacitive reactance to eliminate part of the inductive reactance of the previous form.

By analyzing the $|S_{11}|$ plot of the proposed antenna in Figure 9-3 four adjacent resonant frequencies can be observed over the 3 to 13 GHz frequency spectrum, i.e., 3.4 GHz, 6.6 GHz, 9 GHz and 12.5 GHz. In order to have more indications on the contribution of the individual parts of this antenna at these frequencies, the vector plots of the surface current distributions at these resonant frequencies are investigated in Figure 9-5. The first resonant frequency occurs at 3.4 GHz and the total length of the continuous current path is 21.75 mm, which is $\approx 0.25\lambda$. As can be seen in Figure 9- 5(a), the current flows from the feeding plate to part of the top plate, then to the shorting plate and finally to
part of the ground plane to excite this mode. Therefore, the major dimensions of the geometry parameters, including the height of the antenna, coupling distance between the feeding plate and shorting plate and the width of the shorting plate, can be used to manipulate the first resonant frequency.

Figure 9-6: Practical Prototype of proposed antenna

Figure 9-7: Simulated and measured return losses for the UWB antenna
At the second resonant frequency, 6.6 GHz, one current path can be found from the feeding plate to the part of the top plate which is used to connect the shorting plate with the feeding plate. This current path forms an inverted L-structure, as illustrated in Figure 9-5 (b). This current path length is about 11.25 mm which corresponds to $\sim 0.25\lambda$ at this frequency. It seems that the geometric parameters that control the first resonant mode also determine the second resonant mode, although the shorting plate seems to have little influence on the second mode. By analysing the third resonant frequency, 9 GHz, it is seen that the stronger currents concentrate between the feeding plate and the parasitic plate. It is believed that the capacitive mutual coupling distance between the feeding plate and the parasitic plate has considerable influence on determination of this resonant frequency. The last resonant frequency occurs at 12.5 GHz, as depicted in Figure 9-5 (d). The full length of feeding plate structure is 6.75 mm, which is around $0.25\lambda$ at this frequency. This structure appears to control the resonant mode of this antenna.
Figure 9-8: Variations of geometric parameters against the reflection coefficients $|S_{11}|$. (a) Spacing of parasitic plate, $d$, (b) position of feed point, $fp$ and (c) dimensions of ground plane
9.3 Results and Discussion

To verify the simulated performance of the antenna, a physical prototype was fabricated by using 0.5 mm thickness copper plate, as illustrated in Figure 9-6. The reflection coefficient |$S_{11}$| of the antenna was measured by using a HP8510C vector network analyzer. Figure 9-7 shows the computed and experimental reflection coefficient |$S_{11}$| of the antenna. From these results, it can clearly be seen that four resonant frequencies occurred, at 3.4 GHz, 6.6 GHz, 9 GHz and 12.5 GHz in simulation and at 3.35 GHz, 7 GHz, 9.7 GHz and 12.5 GHz in measurement. Combining these resonant modes, the antenna achieves a wide impedance bandwidth of 125%, covering frequency spectrum from 3 GHz to 13 GHz for the criterion |$S_{11}$| better than -10 dB. Some disagreements can be found between the simulated and measured results and these can be attributed to the use of glue in the prototype and fabrication errors in constructing it.

![Figure 9-9: Simulated and measured gains of the proposed antenna](image-url)
A parametric study was conducted in order to understand the sensitivity of the geometric parameters in relation to the reflection coefficient. Three important geometry parameters, were chosen for this study, each being changed while the others were kept constant; they were: d, the coupling distance between the feeding plate and the parasitic plate; fp, the position of the feeding point along the edge of the feeding plate; and ground plane size (see Figure 9-1). From the previous section, it was evident that the presence of the suspended parasitic metal plate, which is placed in parallel with the feeding plate, plays a significant role in offering impedance matching over the 9.8 to 11.5 GHz band. However, the influence of the coupling distance (d) between the feeding plate and this parasitic plate had not been studied. To investigate this, Figure 9-8(a) shows variation of $|S_{11}|$ when changing the distance (d) from 0.5 mm to 3.5 mm in increments of 1 mm. When d is 0.5 mm, the contribution of the parasitic plate is negligible since it does not improve the impedance matching of the band of interest. However, by further increasing d to 1.5 mm, good impedance matching can be attained from 9.8 to 11.5 GHz without a deleterious effect on other frequency bands. This is due to an optimum capacitive coupling effect, as two metal plates are close to each other. By further moving d from 1.5 to 2.5 mm or 3.5 mm, the impedance matching over the required band does not show substantial improvement and impedance matching in other frequency bands will be impaired.

The position of the feeding point on the antenna governs the impedance matching and bandwidth of the antenna in the higher usable operating frequency band. As can be seen in Figure 9-8(b), moving the feed point along the edge of the feeding plate from 6.75
mm to 10.75 mm shows that optimum impedance bandwidth as well as good impedance matching can be found when \( f_p = 7.75 \) mm. However, when \( f_p \) is taken to values between 8.85 mm and 10.75 mm, impedance mismatch occurs only outside the UWB operating frequency band, i.e. 10.7 GHz-13 GHz. Therefore, it is expected that these values can be noted to be available for further reengineering purposes to satisfy requirements of different applications, without degrading the performance of the antenna.

(a)

(b)
Figure 9-10: Simulated and measured normalized radiation patterns of the proposed antenna for two planes (left: x-z plane, right: y-z plane) at (a) 3 GHz (b) 6 GHz (c) 9 GHz and (d) 12 GHz 'xxxx' simulated cross-polarization 'oooo' simulated co-polarization '------' measured cross-polarization '———' measured co-polarization

Figure 9-8(c) depicts the effect of the ground plane size of the antenna on the performance of $|S_{11}|$. In this study, the antenna is placed at one corner of the ground plane and the six ground plane sizes, i.e. $25 \times 25 \text{mm}^2 \ (\cong 0.25\lambda_0 \times 0.25\lambda_0)$, $50 \times 50 \text{mm}^2 \ (\cong 0.5\lambda_0 \times 0.5\lambda_0)$, $75 \times 75 \text{mm}^2 \ (\cong 0.75\lambda_0 \times 0.75\lambda_0)$, $40 \times 100 \text{mm}^2 \ (\cong 0.4\lambda_0 \times \lambda_0)$ and
100 x 100 mm$^2$ ($\equiv \lambda_o \times \lambda_o$) were used. Of these, 40 x 100 mm$^2$ is noted as a more practical ground plane size for a mobile terminal. As can be noticed, the entire UWB frequency band shows very little perturbation over the lower frequency band from 3 to 3.2 GHz for an improved $|S_{11}| < -9 \, \text{dB}$, which is only -1 dB drop, when the ground size is varied from 50 x 50 mm$^2$ to 100 x 100 m$^2$. This ground-size stability performance is comparable to the work in [14] and much better than the results in [7, 9-10]. When the ground plane size is reduced to 25 x 25 mm$^2$, due to the significant contribution of the ground plane resonant mode to the antenna mode [7], it distorts some parts of the lower and upper band impedance matching of the antenna.

Figure 9-11: The different orientations of the received antenna in the UWB pulse measurement.
Figure 9-12: Measured group delay for the proposed antenna. ($\theta = 0$)

Figure 9-13: Measured normalized excited and received pulses at different directions. ‘Dotted line’ — excited signal, ‘solid line’ — received signal

Figure 9-9 plots the simulated and measured peak gain over the operating frequency from 3 GHz to 13 GHz. As can be noticed, the measured average gain of $\approx 5.5$ dBi with
±1.5 dBi fluctuation is found from the proposed antennas and the worst variation between the simulated and measured gains is about ±1.3 dBi. The discrepancies between the simulated and measured gains are due to the fabrication error of the antenna prototype and insertion loss of the actual feed network is not taken into the account in the simulated model.

The radiation characteristics of the antenna were examined in an anechoic chamber. Figure 9-10 illustrate the simulated and measured radiation patterns of the proposed antenna at 3, 6, 9 and 12 GHz for two principal planes, i.e. xz-plane and yz-plane. Each pattern is presented in terms of co-polar and cross-polar components in a 5 dB scaled plot. As can be seen, both the computed and measured radiation patterns are in acceptable agreement. Some discrepancy can be found between the simulated and measured results, which can be attributed to the fabrication errors and the physical feeding arrangement. As a consequence of using the rectangular plate feeding mechanism, the current distributions on the antenna structure, as depicted in Figure 9-5, have strong frequency dependence, and this is evidenced in the patterns. The asymmetrical radiation patterns of the antenna are also caused by implementing an offset coaxial feed in the antenna’s structure. High co- and cross- polarizations of the radiation patterns, is also observed in Figure 9-10. However, this lack of polarization purity is probably insignificant when the antenna is employed in a rich scattering environment.
To confirm that the proposed UWB antenna has the ability of handling a pulsed transmission without significant distortion, the group delay and $|S_{21}|$ of two identical antennas which were placed in different orientations, i.e. $90^\circ$, $45^\circ$, $0^\circ$, $-45^\circ$, and $-90^\circ$ as shown in Figure 9-11 with a separation of 45 cm were measured. Figure 9-12 shows that the measured group delay of the proposed antenna is stable and holds the same fluctuation, which is less than 0.5 ns, over the 3-13 GHz operating band when two antennas were placed in face-to-face orientation. By substituting the measured $|S_{21}|$ values into the equations given in [20] and adopting the excitation signal from [21] as described in Equation (9.2), the normalized excited and received pulse signals are plotted and compared in Figure 9-13.
\[ S(t) = \sin(2\pi f_0 (t - t_0))e^{-i(t-t_0)/\tau^2} \]  \hspace{1cm} (9.2)

Where \(f_0 = 6.5\) GHz, \(\tau = 0.133\) ns, \(t_0 = 4\tau\). This pulse covers the UWB allocated band from 3.1 to 10.6 GHz. As can be seen, both of the pulse signals are almost indistinguishable at all the selected orientations. In addition, the fidelity factor of the antenna was also computed and found to be 80.2\%, 85.2\%, 91\%, 94\% and 82\% at 90°, 45°, 0°, -45° and -90° orientations respectively.

![Comparison of radiation efficiency and SAR between the antenna in free space and in proximity to the body model. (Normalized to 1 W)](image.png)

Figure 9-15: Comparison of radiation efficiency and SAR between the antenna in free space and in proximity to the body model. (Normalized to 1 W)

To further comprehend the performance of the antenna when it is in proximity with human body, a three layers body model as in [14, 23] is adopted in this study. This consists of skin, fat and muscle and their associated dimensions and electrical properties
are (120 x 110 x 1 mm$^3$, $\varepsilon_r = 38$, $\sigma = 2.7$ S/m, $\rho_{\text{skin}}=1200$kg/m$^3$), (120 x 110 x 3 mm$^3$, $\varepsilon_r = 5.1$, $\sigma = 0.18$ S/m, $\rho_{\text{fat}}= 1000$kg/m$^3$) and (120 x 110 x 40 mm$^3$, $\varepsilon_r = 50.8$, $\sigma = 3$ S/m, $\rho_{\text{muscle}}=1000$kg/m$^3$) respectively. In this analysis, the antenna is placed in the centre of the body model, with the ground plane facing the surface of the model and at two different proximity distances, i.e. 1 mm and 5 mm, from it. Figure 9-14 show the corresponding effects on $S_{11}$ and the peak gain. As can be clearly observed, the $S_{11}$ curves show that the low edge cut off frequency only shifts from 3 to 3.1 GHz when the coupling distance between the antenna and the body is 1 mm. As for the peak gain curves, they show that the gain drops significantly from approximately 3.8 dBi to 1 dBi at the lowest operating frequency (3 GHz) and there is a range of 0.5 to 2.8 dBi gain reduction over the 3-5 GHz band when the body is 1 mm away from the antenna. Interestingly, in the higher usable frequency band from 10-13 GHz, a slight gain improvement by 1 to 1.7 dBi can be noticed when the body is present.

Figure 9-15 shows the corresponding variations of the total radiation efficiency and peak Specific Absorption Rate (SAR) due to the body model effect. (SEMCAD X software [22] was used for this analysis.) As can be seen, efficiency is impaired most where SAR is highest, which is in the lower and upper usable frequency bands where the radiation patterns are more omni-directional, as depicted in Figure 9-10 (a), (b) and (d). The body-reduced efficiency reaches 90% around 9 GHz, as shown in Figure 9-10 (c), where there is least radiation in the lower hemisphere that is directed towards the body. It can be clearly pinpointed that the worst SAR value is predicted as 10.5 W/kg at
3 GHz when the body is 1 mm away. All other values are below 10 W/kg and are comparable to values given for UWB antennas in [14] and [23].

9.4 Conclusion

A compact and low profile modified PIFA antenna has been proposed and extensively investigated in this chapter. This antenna combines impedance matching techniques including top-loading, off-centre rectangular plate feeding and a shorting wall to achieve an ultra-wide impedance bandwidth from 3 GHz to 13 GHz. Despite exhibiting this superior impedance bandwidth, the antenna only occupies a relatively small dimensional envelope of 50 x 50 x 7.5 mm$^3$. The antenna prototype has demonstrated sufficient impedance bandwidth, suitable radiation characteristics, and adequate gains for UWB applications. These attractive characteristics have made it potentially suitable for practical wireless communication applications.
9.5 References


CHAPTER 10

Compact MIMO/Diversity Antenna for Portable and Mobile UWB Terminals

10.1 Introduction

This chapter presents a miniaturized MIMO/diversity antenna which is suitable for high data rate communication systems such as mobile UWB. This antenna assembly comprises two identical PIFAs, a T-shaped structure connecting them, and a finite ground plane. This T-shaped structure improves the impedance matching and suppresses the mutual coupling between the antenna elements over a wider bandwidth than previously reported. The compact envelope dimension of this antenna is $50 \times 90 \times 7.5$ mm$^3$. Theoretical and experimental S-parameters are illustrated for this antenna that fully cover the UWB operating frequency band of $3.1 - 10.6$ GHz, at a reflection coefficient and mutual coupling better than -10 dB and -20 dB respectively. Acceptable agreement is obtained between computed and measured radiation patterns, gains, envelope correlation coefficient and channel capacity loss. The proposed antenna is an attractive candidate to provide pattern diversity and enhance channel capacity in a rich scattering environment.
Conventional narrowband wireless technologies suffer from the limitations of signal fading, multi-path, low immunity to the possibility of arbitrary interference, and limited bandwidth when they operate in dense multipath environments such as occur in buildings and vehicles. Due to these deficiencies, they are less favorable for adoption in high-data-rate wireless applications such as high-speed internet access (>200Mbps) and high-definition TV video/audio streams. To tackle this problem, ultra-wideband (UWB) communication systems [1] have been introduced to cater for the ever increasing appetite for high data rate wireless applications. These systems feature low maximum emitted power (<75nW/MHz), high data rate (>500Mbps) over short ranges (<10m), and large channel capacity deriving from 7.5 GHz bandwidth over the 3.1 to 10.6 GHz frequency spectrum. However, because the UWB systems use low transmitted power this has restricted the application to short distance communication or moderate data rates.

Multiple Input and Multiple Output (MIMO) systems adopting more than one antenna on the transmitting and receiving end of the system have confirmed their capability to offer superior data rate, multipath fading resistance and co-channel interference reduction in the published literature [2, 3]. Therefore, the combination of both UWB and MIMO communication technologies is one of the more promising and cost effective solutions for maximizing the channel capacity and delivering a data rate exceeding 1 Gbps using a robust wireless link employed in an indoor environment [4]. To ensure the sustainability of this UWB-MIMO technology, it is imperative for antenna designers to continue seeking viable approaches for new compact size-reduced MIMO/diversity
antennas that exhibit low mutual coupling between elements while preserving good impedance matching and radiation performance over the UWB band. This must also keep abreast of the increasing demand for miniaturization in the latest portable mobile/handheld devices. Taking into consideration the requirement of small inter-element spacing while maintaining good impedance matching over ultra-wide bandwidth, significant efforts in finding solutions to reduce the wideband mutual coupling between two or more closely placed UWB antennas have been described in published literatures [5-19]. These coupling suppression methods can in general be classified as antenna element position adjustment [5-9], ground plane geometry modification [10-14], and combinations of both techniques [15-19]. By implementing one of these methods, the mutual coupling due to the near field induction, and the sharing of a common ground between two antenna elements [20], can be ameliorated.

In the case of optimising the element positions [5-9], this method reduces the mutual coupling through minimizing the near field coupling current between the antenna elements. Work in [5] demonstrates that it was possible to place two antenna elements side by side with an edge to edge spacing of $\approx 0.5\lambda_0$ for a low mutual coupling of better than -20 dB across the full UWB band. Other work in [6] proposes that one of the antenna elements can be rotated by $180^\circ$ and then both can be concatenated from feeding point to feeding point with a separate distance of $\approx 0.35\lambda_0$ while still achieving good isolation. With this arrangement, mutual isolation greater than 20 dB between two antenna elements can be realized from 1.7 to 11.4 GHz. To further reduce the spacing between two antennas, work in [7-9] show that, by using orthogonal orientation between
the elements, wideband inter-port isolation as good as 15 to 17 dB can be attained over the UWB spectrum for an edge-to-edge antenna element distance as low as 0.1\(\lambda_0\).

Figure 10-1: Geometry of the proposed antenna (a) Top view, (b) Perspective view
As for the case of ground plane geometry modification [10-14], this approach enhances the isolation by minimizing the induced current due to sharing a common ground. Several different geometries have been proposed to be included in a defected (partially cutaway and/or extended) ground such as a tree like structure [10], two T-shaped stubs [11], multiple stubs [12], inverted Y-slots [13] and a rectangular slot [14] to improve the inter-port isolation between the two antennas. The results in [10-14] reveal that this technique offers a wideband decoupling effect and gains a two port isolation level in the range of 15 to 20 dB for two planar radiators separated by a distance between 0.1λ₀ to 0.4λ₀.

Taking advantage of both of the above techniques, work in [15-19] attempt to further optimise the substantial decoupling effect for the two closely placed antenna elements. In [15-18], the mutual coupling was suppressed by tilting the two antenna radiator elements by 45° and -45° to form an orthogonal orientation and optimising the geometry of the ground plane, such as implementing a T-shaped protruded ground plane [15], introducing a slot in the center of the ground plane [16], and inserting a shorted diagonal element at 45° between feeding elements in an annular slot antenna [17]. In contrast to [15-17], inter-port isolation can be enhanced by embedding a 45° oblique slot and -45° oblique slot symmetrically in the antenna elements without physically rotating them, and having a modified ground plane with a long central strip [18]. This approach successfully improves the mutual coupling and coupling distance to levels as low as -20 dB and 0.03λ₀ respectively over the frequency band from 3.1 to 5 GHz. Moreover, the
work in [19] also demonstrated that the mutual coupling between two antennas can be alleviated by positioning the antenna elements in 90° relative orientation and including an inverted L-shaped vertical extension of the ground plane in the structure. With this method, edge-to-edge inter-element spacing becomes relatively small at less than 0.15λ₀ and inter-port isolation as good as 25 dB can be achieved over 2 to 6 GHz.

Figure 10-2: Simulated S-parameters of the proposed antenna with and without the T-shaped structure (a) |S₁₁|, (b) |S₂₁|
Examining all these methods as in [5-19], it is noticed that these antennas either do not fully cover the UWB band [7,13,15,18,19] with good inter-port isolation (≥20dB) [7-10,13,14,17] or have a large radiator edge-to-edge separation (>0.15λ₀) between the two antenna elements [5-6,11-12]. Interestingly, it was found that only work in [16] can satisfy both criteria and it uses a heavily slotted ground plane. In order to fill this deficiency, the present chapter proposes a MIMO/diversity planar inverted-F antenna (PIFA) which fully meets these criteria while retaining a complete ground plane. A T-shaped strip has been inserted between the antenna elements in order to improve the impedance matching and reduce the mutual coupling. To meet the standard size requirement of a commercial transceiver, the overall dimension of this antenna is optimised to an envelope dimension of 90×50×7.5 mm³, including the ground plane. Implementing the proposed coupling structure, it is found that impedance matching of better than 10 dB return loss and better than 20 dB isolation can be achieved over the whole UWB frequency band, with the antennas separated by 0.11λ₀.

10.2 Antenna design concepts and structure

The geometric configuration of the proposed antenna is shown in Figure 10-1. The antenna assembly is constructed from two modified PIFAs with an individual profile 19×15.5×7.5 mm³. These are separated by a distance of 11 mm and a T-shaped structure is used to connect the feeding plates of the two PIFAs. This antenna assembly has a total dimension of 50×15.5×7.5 mm³ and it is placed on the end of a 50×90 mm²
electrically finite ground plane which is ideal for application to a PCMCIA network card. All elements including the ground plane are formed from 0.5 mm copper sheet. The structure and dimensions of a single antenna element were chosen on the basis of previous experience [21-22], combined with realistic constraints of acceptable size for incorporation in a mobile terminal. In the works [21-22], the ultra-wideband characteristic of the antenna is achieved by controlling the upper and lower resonant modes of a single-feed PIFA assembly, with either an F - shaped parasitic element as in [21] or an inverted - L parasitic element as in [22]. In these designs, a 30 × 15 mm² ground plane is used and it plays an important role in obtaining good impedance matching at the lower end of the frequency range. With these geometry combinations, the design in [21] exhibits an impedance bandwidth from 3.1 to 10.6 GHz with an optimised dimension of 30 × 15 × 7.5 mm³, while in [22] the antenna has a smaller dimension of 30 × 15 × 3.5 mm³ but only covers the band 2.8-5.6 GHz. In contrast to [21-22], the present chapter shows, in the proposed MIMO antenna which is augmented by a cross coupling, full UWB coverage can be achieved with only a PIFA (without parasitic) as the single element and hence that the element volume can be reduced by a further 34.5 %.

The design procedure for this proposed MIMO antenna started with optimising a PIFA on a corner of a 50 x 90 mm² finite ground plane. Manipulating the PIFA geometry parameters, i.e. feeding position, feeding gap, and the width and distance of the shorting wall and feeding plate, the multiple-resonance characteristic of this antenna could be
tuned to establish an ultra-wideband impedance bandwidth. In this initial design process, it was found that a response could be achieved with $|S_{11}|$ better than -8 dB, as shown in the circled curve of Figure 10-2(a). After optimizing the geometry for the single PIFA, the design process proceeded to find the optimum separation distance between two identical copies of the PIFA which were symmetrically placed on the ground plane. The aim of this process was to identify a best coupling distance between the two antenna elements for a good wideband inter-port isolation without degrading the good impedance matching bandwidth. Interestingly, it was found that the smallest separation distance was 11 mm for mutual coupling as low as -14 dB across the UWB band, while maintaining a return loss of at least 8 dB, as depicted by the solid lines in Figure 10-2. The final design stage was to introduce a T-shaped coupling structure between the two PIFAs and study its geometry parameters for achieving the best performance in term of port-to-port isolation and return loss. As can be seen in Figure 10-1(b), the horizontal section of the T-shaped structure is linked to the edges of the feeding plates of the two PIFAs and the vertical section of the T is connected to the ground. The optimal locations for integrating this structure are chosen on the basis that they are points of high current concentration when the antenna is operated without the cross coupling. Hence, the T-shape can pick up some current from the excited antenna and re-inject it in the coupled antenna with a suitable magnitude and phase to cancel out the existing coupling current between the two antenna elements. An interesting result is that the coupling structure achieves a very substantial improvement in impedance matching, as well as in isolation, over a very wide bandwidth.
Figure 10-3: Two possible impedance matching and decoupling current paths on the T-shaped structure Port 1 (right) is excited and port 2 is terminated in 50Ω

In principle, this approach is a modified version of [23-26]. In [23-24], a neutralization line is implemented between antenna elements to improve the port-to-port isolation over a narrow operating frequency band. To enable wider band operation, work in [25] proposes to implement decoupling lines which provide more current paths to suppress the mutual coupling between the antennas. Other work in [26] uses a 3-D meander line to improve the inter-port isolation. This arrangement avoids direct connection between the antennas and does not require retuning of the antennas, but works over a fairly narrow frequency band. These methods [23-26] are not capable of providing enough bandwidth for the UWB band. Methods [23-24] do disturb the initial operating frequency of the antenna elements and require some retuning of the geometry parameters after insertion of the line. The method described here enhances these techniques and improves the impedance matching over the designated UWB band as well as offering good ultra-wide band isolation.
In order to understand the basic operation of this impedance matching and decoupling structure, the T-shape has been divided into two sections, i.e. section AB and section CD, as illustrated in Figure 10-2. In this study, three scenarios, i.e. without T-shape, with only the horizontal section (AB), and with the complete structure (AB and CD) of the T-shape, are investigated. To evaluate the effectiveness of this structure, the simulated reflection coefficient $|S_{11}|$ and mutual coupling $|S_{21}|$ of the proposed antenna with and without the added element are compared in Figure 10-2. As can be observed, with inclusion of the complete T-structure, $|S_{11}|$ can be improved from a marginally acceptable case (-8 dB) to a commonly accepted level (-10 dB), a 2 dB average enhancement over the UWB band. Examining $|S_{21}|$, it has found that this is significantly reduced from a worst case (-14 dB) to a much improved level which is better than -23 dB across the UWB band 3.1 to 10.6 GHz. Moreover, some further insight into the operation of the T-shape without the grounded point is also studied in Figure 10-2. As can be observed, when only section AB is present in which the current is directly injected to the nearby antenna element, $S_{11}$ shows some improvement in the lower range of the useable frequency band, but $S_{21}$ exhibits rather poor performance (<-10 dB) compared to the case without the T-shape. Effective neutralization appears to require the presence of section CD, and more current appears to flow into path BCD or ACD than path AB.
Figure 10-4: Surface current distributions of the proposed antenna at (a) 3 GHz, (b) 6 GHz, (c) 9 GHz. Port 1 (right) is excited and port 2 (left) is terminated in 50 Ω. Left: proposed antenna, Right: reference antenna.

From the previous analysis, Figure 10-3 summarizes the possible current paths of the proposed antenna when port 1 is excited and port 2 is terminated with 50 Ω, for further understanding the decoupling and matching mechanisms. In the absence of the T-shaped structure, two coupling currents due to near field coupling and the sharing of a common ground can be represented by the black solid line and red solid line respectively. When the T-shaped structure is added between the antennas, two impedance matching and decoupling current paths, i.e. the dashed black line and dashed red line, will be
generated to suppress the unwanted near-field coupling current and the coupling current through the ground correspondingly. Compared to [23-26], this method can provide better isolation and impedance matching over bandwidth due to weakening two induced currents instead of only reducing the near field coupling current as in [23-26]. By calculating the total path length of these two continuous currents, impedance matching and decoupling currents, it was found that both current paths are 35.5 mm long which is approximately \(0.43\lambda_o\) \((\lambda_o = \text{free space wavelength})\) at 3.65 GHz. These two current paths have the vital role in improving the match and inter-port isolation at the lowest and highest edges of the operating band. This explanation can be validated by the \(S_{11}\) parameter curve Figure 10-2(a) of the antenna with the neutralization structure. As can be seen, at the lower operating band, a -25dB deep null occurs at 3.65 GHz when the T-structure is introduced between the antenna elements, whereas in the higher part of the operating band, a higher order resonance mode of the T-shape is excited and shows a -18 dB deep null at 9.8 GHz. It is also noticed that mutual coupling less than -20 dB can be found over these two operating regions. From this observation, the proposed impedance matching and decoupling network more resembles a branch line coupler than other microwave filters or resonators.
Figure 10.5: Simulated S-parameters with variation of different parameters of the T-shaped structure: (a) W1, (b) W2, (c) H. Left: reflection coefficients $|S_{11}|$, Right: mutual coupling $|S_{21}|$
Furthermore, the contribution of the T-shaped structure to the performance can also be elucidated by examining the surface current distributions with and without it, as shown in Figure 10-4. Three selected frequencies, i.e. 3 GHz, 6 GHz and 9 GHz, which represent the lower, middle and upper UWB frequency ranges, were used in this analysis. As can be clearly observed, when port 1 is excited and port 2 is terminated in 50 Ω, without the T-shaped structure, the surface currents flowing in the excited radiator induce current in the edge of the port 2 radiator and some part of the ground plane at all frequencies. Particular attention should be given to the 3 GHz case, where it can be noticed that higher current concentration appears on the ground plane between the antenna elements compared to the behaviour at 6 GHz and 9 GHz. This finding in a good agreement with the S_{11} and S_{21} results without the T-structure shown in Figure 10-2, where lower values of S_{11} and S_{21} can be found at 3 GHz and higher values of S_{11} and S_{21} can be pinpointed at 6 GHz and 9 GHz. These induced currents introduce a strong mutual coupling between the two antenna elements. By inserting the coupling structure, it is found that the surface currents on both radiators are transferred to the T-shaped structure and this weakens the induced current on the port 2 antenna element. This tends to decouple the currents on the port 2 radiator and hence it enhances the isolation between the two radiator elements effectively.
Figure 10-6: Practical Prototype of proposed antenna

Figure 10-7: Simulated and measured S-parameters of the proposed antenna
To check the sensitivity of the S-parameters ($S_{11}$ and $S_{21}$) to the geometry parameters of the T-shaped network, a parametric study was conducted as depicted in Figure 10-5. An ultra wide operating band was the main target for optimisation, with the requirements $|S_{11}| < -10$ dB and $|S_{21}| < -20$ dB. Three geometric parameters, namely the width ($W_1$) of the horizontal section of the T-structure and width ($W_2$) and height ($H$) of its vertical section, were considered for this investigation. These were optimised at 2.5 mm, 3 mm and 2 mm for $W_1$, $W_2$ and $H$ respectively. To show the dependencies, the simulation was run again varying only one parameter at a time from these starting values.

The width ($W_1$) of the horizontal section of the T-structure determines the amount of the current flowing from the excited antenna element’s feeding plate to the adjacent antenna element’s feed plate. Therefore, it is a paramount parameter for obtaining a good inter-port isolation and return loss, as illustrated in Figure 10-5. As can be observed in the $S_{11}$ plot of Figure 10-5(a), by varying $W_1$ from 0.5 mm to 3.5 mm with 1 mm increment, the lower cut off frequency could be varied from 2.9 GHz to 3.2 GHz, whereas the upper cut off frequency changes from 11 GHz to 10.6 GHz correspondingly. It was found that narrower $W_1$ gives a better impedance bandwidth, but when it is smaller than 0.5 mm, the $S_{21}$ in the lower and upper frequency ranges will be impaired. Further gradually increasing $W_1$ to 3.5mm shows that $S_{21}$ could be better than -20 dB across the desired UWB band. To compromise the performance between $S_{11}$ and $S_{21}$, a value of $W_1$ between 2 mm to 3 mm is suggested.
Figure 10-8: Simulated and measured MIMO characteristics of the antenna with and without T-structure, (a) correlation coefficient (b) capacity loss.
The width (W2) of the vertical section of the T-structure controls the upper cut off frequency of the UWB antenna. As can be seen in the $S_{11}$ plot of Figure 10-5(b), by changing W2 from 1 mm to 4 mm, the higher cut off frequency increases from 10.2 GHz to 11 GHz. Inversely, the cut off frequency point of the $S_{21}$ plot is decreased from 11 GHz to 10.5 GHz. From these observations, the value of W2 between 2.5 mm to 3.5 mm seems to offer a promising low mutual coupling and good impedance matching in the proposed UWB antenna. Figure 10-5 (c) plots the variation of the height (H) of the vertical section of the T-structure against the $S_{11}$ and $S_{21}$. This parameter varies both the characteristic impedance of the section AB, viewing it as a pair of transmission line sections, and the reactance of the vertical section CD, viewing it as approximating an inductance to ground. As can be illustrated, gradually increasing the H from 1 mm to 4 mm, the impedance bandwidth for $|S_{11}|$ less than -10 dB is progressively reduced from (3 to 11 GHz) to (3.2 to 10 GHz). However, in the case of $S_{21}$, modifying H from 1 mm to 3 mm does not impair it, but further increasing H to 4 mm will shift the upper cut off frequency to 9.5 GHz. Hence, it is recommended that the value of H should be selected between 1.5 mm to 2 mm.
10.3 Results and Discussion

From the parametric study results, the optimum values of the parameters $W_1$, $W_2$ and $H$ were verified as 2.5 mm, 3 mm and 2 mm respectively, and these values were used in fabricating a practical prototype of the antenna as in Figure 10-6 to validate the simulation results. The measured and simulated reflection coefficient $|S_{11}|$ and transmission coefficient $|S_{21}|$ results are shown in Figure 10-7: these exhibit reasonable agreement although there is a small frequency shift that can be attributed to reflections from the SMA connectors and feed cable effects, and small inaccuracies in assembling the real structure which is fabricated using conductive epoxy adhesive. In the $|S_{11}|$ plot, the simulated and measured results confirm that the impedance bandwidth of the
antenna encompasses the operating frequency spectrum from 3.0 to 11.0 GHz for a reflection coefficient $|S_{11}| < -10\text{dB}$; this corresponds to 8 GHz bandwidth or about 114.3\% relative bandwidth with respect to the centre frequency of 7 GHz. Likewise in the $|S_{21}|$ plot, the experimental result validates the computed result showing a low mutual coupling with $|S_{21}| < -21 \text{dB}$ over the UWB band from 3.1 to 10.6 GHz.
Figure 10-10: Simulated and measured normalised radiation patterns of the proposed antenna for two planes (left: x-z plane, right: y-z plane) at (a) 3 GHz (b) 6 GHz and (c) 9 GHz ‘xxxx’ simulated cross-polarization ‘oooo’ simulated co-polarization ‘------’ measured cross-polarization ‘———’ measured co-polarization

For diversity and MIMO application, the correlation coefficient between the two elements' patterns, and the channel capacity loss due to the presence of uncorrelated Rayleigh-fading MIMO channels, are another two important parameters that determine the performance of a MIMO system. These will therefore be discussed here.

The simulated and measured envelope correlation coefficient ($\rho_e$) of the MIMO antenna with and without the coupling element as presented in Figure 10-8(a), is calculated using the S-parameters method and the equation as given in [11, 27] is as follows:

$$\rho_e = \frac{|S_{11}'S_{12} + S_{21}'S_{22}'|^2}{\left(1 - |S_{11}'|^2\right)\left(1 - |S_{22}'|^2\right)}$$  \hspace{1cm} (10.1)
From the simulated and measured $\rho_e$ results, it was found that $\rho_e$ values for the antenna with and without the coupling element are less than 0.002 and 0.006 respectively. These results are comparable to the results given in [11, 23, 25]. Small discrepancies between the computed and experimental results can be mainly attributed to the fabrication errors and feed cable effects.

By taking the assumption that only the receiving antennas are correlated in a high SNR scenario, the capacity loss ($C_{\text{loss}}$) of a 2 x 2 MIMO system can be computed by using the equation given in [27]:

$$C_{\text{loss}} = -\log_2 \det(\psi^R)$$

(10.2)

Where $\psi^R$ is the receiving antenna correlation matrix that is given by: $\psi^R = \begin{bmatrix} \rho_{11} & \rho_{12} \\ \rho_{21} & \rho_{22} \end{bmatrix}$, $\rho_{ij} = 1 - \left| S_{ij} \right|^2 - \left| S_{ji} \right|^2$, and $\rho_{ij} = -\left( S_{ij}^* S_{ij} + S_{ji}^* S_{ji} \right)$, for $i, j = 1$ or 2.

Figure 10-8(b) shows the simulated and measured channel capacity losses of the proposed MIMO antenna with and without the T-structure. Considering the average capacity loss over 3.1 to 10.6 GHz for the antenna with this structure, it is found that neither curve exceeds 0.28 bps/Hz. The simulated and measured results show fair agreement and the variations in the curves are very similar to those of the correlation coefficient. Comparing with the result obtained from the antenna without the T-shape, it is notable that the channel capacity loss can be doubled (0.6 bps/Hz). This confirms the expectation that good return loss and low mutual coupling between two antenna elements lead to low envelope correlation coefficient and low capacity loss.
Figure 10-11: Simulated normalised radiation patterns of the single antenna, and proposed antenna without T-shaped structure, for three planes (left: x-z plane, right: y-z plane) at (a) 3.0 GHz (b) 6.0 GHz (c) 9.0 GHz. Port 1 (right) is excited and port 2 (left) is terminated in 50 Ω. ‘xxxx’ (single antenna alone) cross-polarization, ‘oooo’ (single antenna alone) co-polarization, ‘———’ (without T-shaped structure) cross-polarization, ‘———’ (without T-shaped structure) co-polarization.
Figure 10-9 shows the simulated and measured gains of the UWB MIMO antenna over the interval from 3 to 10.5 GHz. Agreement is reasonable and it can be observed that a practically-useful average gain of 5.3 dBi was measured with ±1.85 dBi of gain variation across the UWB frequency band, where the average simulated gain is found to be 5.5 dBi with ±1.5 dBi of gain variation.

The simulated and measured radiation patterns of the proposed antenna at 3, 6 and 9 GHz for two principal planes, i.e. xz-plane and yz-plane, are shown in Figure 10-10. Each pattern is presented in terms of co-polar and cross-polar components in a 5 dB scaled plot. As can be seen, both the computed and measured radiation patterns are in acceptable agreement. Additional small discrepancies may occur due to measuring chamber errors and shadowing. It should be noted that some disagreements between patterns at the lower and upper frequencies is due to the contribution of the ground plane resonant mode at the lower frequency [28]. The radiation patterns are inherently asymmetrical because an offset coaxial feed is used in the antenna’s structure, and because the ground plane is not symmetrical with respect to either element when a single port is excited.
Figure 10-12: The total radiation efficiency and peak gain of single antenna, proposed antenna without T-shaped structure and with T-shaped structure.

To check the direct contribution of the T-structure to the radiation field characteristics of the antenna, the simulated radiation patterns of the single element of the antenna and two element antennas without the T-shape are presented in Figure 10-11. As can be seen, these radiation patterns are very similar to those in Figure 10-10, with only insignificant distortions. To further comprehend the effect of introducing the coupling on the peak gains and total radiation efficiency, Figure 10-12 compares the performance of a single antenna, and the proposed antenna and without the coupling. T-shaped structure and with T-shaped structure. Observing the gain plots, compared with the single antenna, both antennas with and without the coupler have little gain variation in the lower usable operating frequency band; while in the higher usable frequency band,
the antenna with T-shaped and without T-shaped shows less than 1dBi and 2.2 dB gain variations respectively. Examining the radiation efficiency curves, the antenna with the T-shape exhibits greater than 90% efficiency over the UWB band, whereas the single element antenna and the antenna without the coupler have less than 90% efficiency for frequencies above 7GHz. These results have led to the conclusion that better impedance matching and higher isolation can improve the radiation performance of the MIMO antenna.

10.4 Conclusion

A compact and miniaturized MIMO UWB antenna has been studied in this chapter. By implementing a proposed T-shaped coupling structure, good impedance matching (<-10dB return loss) and low mutual coupling (<-20dB) between the antenna elements can be realized over very wide bandwidth covering the standard UWB frequency band from 3.1 to 10.6 GHz. Further investigation of the radiation and MIMO characteristics of the proposed antenna has confirmed that the antenna has improved capability in overcoming the multipath fading propagation problem through offering pattern diversity.


10.5 References


CHAPTER 11

Conclusions and Future Work

11.1 Conclusions

Wireless communications systems have been investigated with particular reference to modern ultra-wideband communications involving high speed transmission rates for indoor and possible outdoor, applications. Invention reports are presented for two new types of antenna, set against a review of the existing state of the technology.

A small and compact crescent shaped microstrip patch has been investigated first. The antenna was designed to bring a relative bandwidth of 51.8%. A parametric study was performed, and the optimised antenna performance indicates an antenna volume of $57\text{mm} \times 37.5\text{mm} \times 0.8\text{mm}$. Such a small antenna volume was found very attractive for use in mobile terminals operating over DCS, PCS, UMTS, Bluetooth or IEEE 802.11b/g wireless standards.

A new modified version of a wideband planar inverted F-L antenna or PIFLA has been presented. The design optimisation has found an objective compromise between the size constraints of the antenna assembly and the desired impedance bandwidth performance, with respect to the lower operating UWB frequency spectrum. The prototype and
simulated results were compared in terms of radiation patterns, gain performance, radiation efficiency, and group delay response. Consequently, such an antenna is suitable for UWB applications.

A dual-planar inverted F-antenna of small size, covering the operating bands of WLAN/WiFi, WiMAX and the lower band UWB wireless standard has been presented. The antenna design concept was capable of covering 88.4% relative impedance bandwidth with acceptable reflection coefficient $|S_{11}|$. The antenna has shown consistent omni-directional radiation patterns and reasonable gain values across the operating bands. The experimental results show satisfactory performance and good agreement with the computed results. With its broadband characteristic, the proposed antenna is very well suited to multi-band wireless applications.

A band-suppression UWB suspended planar antenna incorporating a slotted spiral resonator has been presented, from which narrow notched bandwidth suppression ability in WLAN band is obtained. The set of parameter studies of the proposed antenna provides brief guidelines for the band-notched antenna design. Evaluations of return loss and radiation pattern were confirmed the antenna performance. These features of the proposed antenna demonstrated that the proposed antenna is suitable for UWB communication applications and prevents interference from the WLAN system.
Finally, a compact and miniaturized MIMO UWB antenna has been designed and investigated. A new proposed T-shaped coupling structure was implemented to achieve good impedance matching (<-10dB return loss) and low mutual coupling (<-20dB) between the antenna elements over very wide bandwidth covering the standard UWB frequency band from 3.1 to 10.6 GHz. Further investigation of the radiation and MIMO characteristics of the proposed antenna has confirmed that the antenna has improved capability in overcoming the multipath fading propagation problem through offering pattern diversity.

### 11.2 Future Work

Future work plan will be constructed to extend the trends established in this report, resulting in detailed design strategies for the antennas and their response to the target operational environment. The following points have been selected as being the most likely threads for future development:

1. Further development of the design process in Chapter 3 is required. In particular the trade off between antenna volume and target bandwidth needs to be fully characterised. This could be achieved by reducing the size of both the radiating patch and ground plane to accommodate small volume antenna that more suitable for developing mobile handset applications.
2. Refinement of the PIFA design in order to reliably achieve both left and right handed circular polarizations. This would be as good task for future work considerations.

3. The use of genetic algorithms to explore the design space for the optimisation of the antenna in chapter 3

4. Tuning within the UWB frequency spectrum may be investigated, application to MIMO systems using tuned UWB antennas. It would be envisaged that suitable control mechanisms will be embedded into the antenna structure.

5. Frequency notching for the 802.11a interference band occurring at 5.15-5.35 GHz and 5.725-5.825 GHz will also be considered with regard to antenna designs in Chapter 5.
LIST OF PUBLICATIONS

JOURNAL PAPERS


INTERNATIONAL CONFERENCES


International Conference on Internet Technologies & Applications, Wrexham, North Wales, UK, 8-11 September 2011.


Ultra-wideband planar inverted FF antenna


A novel, miniaturised, planar inverted FF antenna (PIFA) assembly is presented. The antenna design electromagnetically couples two PIFA elements on an air substrate, and employs a broadband rectangular plate feeding mechanism to achieve an ultra-wideband characteristic. The dimensions of the complete prototype assembly are 30 × 15 × 8 mm. Good agreement is obtained between computed and measured impedance bandwidths over the range 3.1–10.6 GHz for \(|S_{11}| < -10\) dB. The simulated and measured gain and radiation patterns are given also, to fully describe the antenna performance.

Introduction: The Federal Communications Commission released the ultra-wideband frequency band (FCC Part 15, Subpart F, 3.1GHz–10.6GHz) as a free licence spectrum in 2002 for commercial use. The UWB radio concept uses narrow RF pulses between transmitter and receiver. Advocates of UWB technology propose such applications as real-time wireless video streaming within buildings, with data rates of 480 Mbit/s and beyond, combined with low power consumption, typically less than 2 mW/Mbit/s. The rationale for commercially available UWB antennas with large bandwidth (approximately 3:1), compact size, with consistent gain and radiation patterns is clear. One of the most promising candidates to meet these requirements is the planar inverted F antenna (PIFA); design commonly found in contemporary mobile terminal applications. PIFA design is attractive owing to its physical simplicity in generating multiple resonant frequencies [1–4], and has the potential for wideband operation [5–7]. In this Letter, a novel miniaturised UWB planar inverted FF antenna (PIFFA) is proposed. By controlling the geometry of the coupling distance between two PIFA elements and the position of a broadband rectangular feeding plate, the UWB frequency band can be achieved. The final design is optimised with a size of 30 × 15 × 8 mm for easy integration into a wireless radio unit. Details of this design, and the results of the realised prototype are discussed in the following Sections.

Antenna design: The geometry of the proposed antenna (PIFFA) assembly may be visualised through Fig. 1. The dimensions of the driven and parasitic PIFA elements are 18.5 × 10 × 4.7 and 7.5 × 14.5 × 7.5 mm, respectively. The antenna assembly is mounted on a finite ground plane, the dimensions of which are 30 × 15 mm. The separation between the PIFA elements is set at 4 mm, for optimal coupling. The conventional PIFA wire feed mechanism is replaced by a rectangular plate feed to the driven element, to improve bandwidth performance. The resonant frequencies of both PIFA elements is approximated from [1]

\[ f \approx \frac{c}{4(W + L)} \]

where \(W\) and \(L\) are, respectively, the width and length of the element, and \(c\) is the speed of light. It follows that the driven PIFA is resonant at approximately 3.2 GHz, and the parasitic PIFA is resonant at approximately 4.86 GHz.

Fig. 1 Proposed antenna structure

The lower resonance is dominated by the driven patch, while the higher resonance is excited by the combined area of the two PIFA elements. It follows that the coupling gap is critical for both impedance bandwidth and the quality of impedance matching over the required frequency band. The size of the ground plane is also a significant factor in realising a good UWB impedance match.

Results and discussion: The modelling of the individual antenna structure, and the optimisation of the final assembly were carried out, and cross-validated, using SEMCAD and Ansoft HFSS. Simulated and measured reflection coefficients \(|S_{11}|\) are compared in Fig. 2. The individual structures were analysed to determine their impedance bandwidths, as a necessary step in understanding the individual contribution of the driven and parasitic antennas. For the driven PIFA, the impedance bandwidth covers the range 3.6–12 GHz, with a reflection coefficient of \(|S_{11}| < -6\) dB. The corresponding range for the parasitic PIFA is 4.8 to 9.2 GHz, for the same reflection coefficient. Constructive addition of these impedance bandwidths enables a good impedance match over 3.0–10.6 GHz for \(|S_{11}| < -10\) dB.

Fig. 2 Measured against simulated reflection coefficients \(|S_{11}|\) of proposed antennas

The effects of ground plane size on antenna reflection coefficient \(|S_{11}|\) were investigated parametrically, as presented in Fig. 3. The physical dimensions of the ground plane were scaled in terms of \(a\), the wavelength corresponding to the lowest resonant frequency of the structure. The response was analysed for different ground plate sizes: 30 × 15 mm (≈ 0.33a × 0.15a), 40 × 40 mm (≈ 0.4a × 0.4a), 60 × 60 mm (≈ 0.6a × 0.6a), 80 × 80 mm (≈ 0.8a × 0.8a) and 100 × 100 mm (≈ a × a). When these particular dimensions are greater than or equal to 0.4a, there is a slight impedance mismatch within the lower frequency components in the range 3–3.5 GHz. The reflection coefficient in this range is –6 dB, which is still acceptable for commercial applications. Reducing the ground plane dimensions to 0.3a × 0.15a improves the impedance bandwidth of the lower frequency components; this may be due to the combined effect of the antenna geometry and the ground plane, with the dual-mode operation.

Fig. 3 Simulated reflection coefficients \(|S_{11}|\) corresponding to variation of ground plane size

Simulated and measured antenna gains are compared in Fig. 4. These values vary between 3.75 and 5.55 dB over the total UWB band; with maximum variations in the neighbourhood of ±1.5 dB, and a mean of 4 dB. Both simulated and measured results are in sufficient agreement. Fig. 5 compares the simulated and measured antenna radiation patterns of the prototype assembly at 3, 5, 7 and 9 GHz. These patterns are approximately omnidirectional and consistent at the lower frequencies, except for the y-plane in Figs 5a and 5b. In these cases the nulls predicted in simulation are not fully realised; this is believed to be due to subtle differences in the effects of the simulated against physical feeding arrangements. As the frequency increases, this omnidirectional characteristic distorts, as a consequence of the feed mechanisms of the driven PIFA.

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Fig. 4 Simulated against measured antenna gains

Fig. 5 Simulated against measured normalized antenna radiation patterns for two planes (left: x-z plane, right: y-z plane) at 3000, 5000, 7000 MHz.

a: 3000 MHz
b: 5000 MHz
c: 7000 MHz

d - - - - simulated cross-polarization
e-----e measured cross-polarization

Conclusions: The compact low profile antenna assembly with a finite ground plane is proposed and realised. This antenna constructively combines the resonant modes of the two PIFA elements. The prototype performance demonstrates sufficient impedance bandwidth, suitable radiation characteristics, and gain stability, for realistic UWB radio applications.

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References
Broadcast dual planar inverted F-antenna for wireless local area networks/worldwide interoperability for microwave access and lower-band ultra wideband wireless applications

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Abstract: An extension of the planar inverted F-antenna (PIFA) concept is proposed, combining two such antennas (one of them fed parasitically) to extend the usable bandwidth to cover the WiFi wireless local area network, worldwide interoperability for microwave access and lower-band ultra wideband spectra. The broad impedance bandwidth is achieved by combining a driven slotfed PIFA, having bandwidth-enhanced rectangular strip feed, with a parasitic inverted F-shaped element, hence describable as a dual PIFA. The geometry parameters of these two radiators are selected to enhance the impedance bandwidth and hence to encompass the entire required set of operating frequency bands. The overall optimised dimensions of the antenna are 30 mm x 15 mm x 8 mm, thus making it compatible with installation on portable wireless electronic devices. The results show that the proposed antenna can achieve a gain between 2.0 and 4.0 dBi across the entire impedance bandwidth 2.4–6.2 GHz (i.e., 88.4% relative bandwidth) for a reflection coefficient of |S11| < −10 dB. The radiation patterns of the antenna show low-gain broad-beam coverage, as is essential for a wireless electronic device. Computed and experimental results are compared and are shown to be in satisfactory agreement.

1 Introduction

Wireless local area networks (WLANs) have gained popularity in indoor environments, owing to their licence-free status and ease of installation: the WiFi WLAN card has thus become an essential element to be integrated into most current laptop computers and many mobile phones. IEEE 802.11 (WiFi) is a key industry standard for WLAN systems: this allocates operating frequencies in licence-free Industrial, Scientific and Medical (ISM) bands at 2.4, 5.2 and 5.8 GHz (specifically 2400–2485, 5015–5350 and 5725–5850 MHz) for portable electronic devices to access the internet wirelessly. Ignoring the newer and physically very different multiple-input multiple-output (MIMO)-based standard IEEE 802.11n, these WiFi systems are typically capable of handling a maximum channel rate of about 11 Mbps and a maximum user data rate of about 1.6 Mbps within the maximum transmission range of approximately 100 m [1]. In recent years, another two wireless standards, worldwide interoperability for microwave access (WiMAX) and ultra wideband (UWB), have been created, respectively, defined by the IEEE 802.16 family of standards and the now-withdrawn IEEE 802.15.3a draft, with the aim of improving the data rate and/or communication distances, to cater to the increasing consumer demand. WiMAX technology operates in 2.5, 3.5 and 5.5 GHz bands (2500–2690, 3300–3700, 5250–5550 MHz), with a range that correlates it with the concept of a wireless metropolitan area network. It can theoretically reach 50 km radius coverage and achieve data rates up to 75 Mbps, for which the throughput is higher than the 1.5 Mbps performance of typical broadband services [2]. UWB is intended for short range (~10 m) and higher data rate communication (>500 Mbps) across a wide frequency spectrum (3.1–10.6 GHz). The lower and upper UWB spectra are 3.1 to 4.0 GHz and 6.0 to 10.6 GHz, respectively, [3]. Together with the MIMO-based IEEE 802.11n, these three wireless standards are major candidates for emerging commercial mobile wireless technologies. However, for terminals using these technologies to be truly mobile, multi-band miniaturised antenna designs are required and these pose challenges.

Recently, several publications have discussed different types of antennas for WLAN/WiMAX applications [4–17]. These can be classified into three categories: dual band [4–7]; tri-band [8–14] and broadband [15–17], encompassing
the desired operating frequency bands from 2.4 to 6.0 GHz. In the case of the dual-band and tri-band modalities, it was noted that to avoid frequency collision and to minimise interference from the unused licensed/licensed frequency spectra, band-rejection functions were considered a desirable objective for all of the proposed antennas [4–13]. These band-notching methods include modifying the geometries of a planar inverted F-antenna (PIFA) [4, 5], adding a truncated corner on a printed inverted F-antenna [6], introducing a square conductor-backed plane on a printed monopole antenna [7], implementing a multi-armed inverted-F monopole [8], electromagnetic coupling of a stub-loaded open-loop resonator on the underside of a CPW-fed mirrored-L monopole [9], applying ceramic chip radiating elements to the geometry of the antenna [10] and inserting band-notching slots in the radiating structure of the antenna [11–14].

For the case of broadband operation, coxial antennas [15], CPW-fed Koch fractal slot antennas [16] and printed E-shaped monopoles [17] were shown to provide sufficient impedance bandwidth to cover the required operating frequency for WiFi, WiMAX and lower-band UWB applications. However, it is significant that some of these antennas required a large ground plane [9, 14, 15, 17]; some produced inconsistent radiation patterns [9, 11, 17] and others showed substantial gain variations across the operating frequency bands [11, 13]. In addition, most of the papers addressed only the needs of WiFi and WiMAX applications [4–16]; very few compact antenna designs have included discussion of applications to the lower UWB band [17].

To overcome these limitations, a combination of two compact broadband planar inverted F-antennas was devised. This consists of a driven PIFA and a parasitic PIFA element and hence may be described as a dual PIFA (D-PIFA). The driven and parasitic PIFAs are designed to control the lower and upper resonant modes of the antenna, respectively. By carefully adjusting the geometry parameters of the antenna, broadband impedance matching can be realised. The size of the antenna achieved was 30 mm × 15 mm × 8 mm, which is small enough to offer a high degree of freedom for designers to install into typical small electronic cases for wireless sensor network and short-range radio communication applications [18].

2 Antenna design concept and structure

The proposed antenna design is an evolution of earlier research [5, 19, 20]. In [5], it was found that variations of the widths of the driven and parasitic elements will not cause any significant change in the impedance bandwidth. In [20], the driven antenna element excites the higher-frequency mode, while the parasitic antenna element controls the lower-frequency mode to provide dual-band operation for mobile telephone applications. In addition, using the concept of removing half of the antenna geometry along the line of symmetry of the structure [21, 22], the antennas in [5, 19, 20] might be reduced to half of its original size without deterioration of their overall characteristics. Based on the results of [5, 19], the aim of the present work was a compromise design process between size and impedance bandwidth constraints to realise a small wideband antenna on a small ground plane (ideally 30 mm × 15 mm), operating over the WLAN, WiMAX and lower-band UWB bands.

The design concept adopts the principle of multiple radiating elements, each supporting strong currents and radiation of one of the two resonant modes. It should be noted that the parasitic PIFA is adopted instead of the inverted-L element, as applied in the antenna structure reported in [5, 19]. This is because the parasitic PIFA can provide more degrees of freedom in its geometry parameters in the antenna design optimisation process, in comparison with the inverted-L element, and hence this provides more flexibility to achieve the optimum goals of the antenna performance.

The driven PIFA is the primary element that governs the lower resonant frequency, while the higher resonant frequency is excited by the combination of the driven and parasitic PIFAs. This is an inverse method of excitation in comparison with [20].

A corollary is that the coupling gap between the two PIFAs is critical and a non-optimised choice of this can result in narrow impedance bandwidth or bad impedance matching over the required frequency band. It must be emphasised that the feeding method also plays a considerable role in achieving wide impedance bandwidth a broadband rectangular strip with a 0.5 mm feed gap was used to excite the driven PIFA instead of using a conventional probe (wire) feed. This improved the impedance matching because of its reduced inductance compared with the probe feed [5, 19]. Further, a slot was introduced in the vertical shorting wall of the driven PIFA in order to reduce the coupling between the rectangular feed strip and the shorting wall, and hence to improve the impedance matching. This became evident from the parametric study.

Initially, the geometry parameters of the proposed antenna and the resonant frequency for the driven and parasitic PIFAs can be approximately predicted, using the following formula [23]

\[ f \approx \frac{c}{4 \times (W + L)} \]  

where \( c \) is the speed of light; \( W \) and \( L \) are the width and length of the radiating element, respectively. To begin with, using the primary antenna dimensions shown in Fig. 1, it is found that the driven PIFA is resonant at \( f_1 \approx 2.3 \) GHz, whereas the parasitic PIFA is resonant at \( f_2 \approx 3.4 \) GHz.

Hence, the driven PIFA is optimised with dimensions of 17.6 mm × 15 mm × 1 mm, whereas the parasitic PIFA is 8 mm × 14.1 mm × 2 mm. For ease of implementation in a typical commercial wireless casing or enclosure, both of them are mounted on a 30 mm × 15 mm finite ground plane. For optimal coupling, the separation between the two elements is 4.4 mm. The thickness of the copper sheet and the feed gap distance are both 0.5 mm. This configuration has overall dimensions of 30 mm × 15 mm × 2 mm, which is equivalent to 0.24λₘ × 0.12λₘ × 0.064λₘ, where \( λₘ \) is the wavelength at lowest resonant mode (≥2.4 GHz), adopting the optimised value of \( k2, 8 \) mm.

3 Parametric study results

The impedance bandwidth was the main target to be optimised throughout the parametric study (defined at reflection coefficient \( |S_11| \) < −10 dB). Each simulation was run with only one parameter varied, while other parameters stayed unchanged and were held at previously determined optimum values. The fixed and variable parameters are shown in Fig. 1, where the variable parameters are considered as critical in determining the lowest and highest frequencies of the operating bandwidth. Since the
Fig. 1 Geometry structure of the proposed antenna

Fig. 2 Simulated reflection coefficients $|S_{11}|$ with variation of different parameters

- $a$ $d_1$
- $b$ $d_2$
- $c$ $h_1$
- $d$ $h_2$
- $e$ $w$
- $f$ Ground plane size

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3.4 Effect of height of parasitic PIFA (h2)

The influence of parameter h2 is very much like that of h1, as depicted in Fig. 2d. By varying the height of the parasitic PIFA, both the lower and upper cut-off frequencies move simultaneously. As can be noticed, when h1 is at its shortest dimension (2 mm), the antenna suffers from poor impedance matching in the low operating frequency band (2.4–3.2 GHz), but it shows reasonable matching from 3.2 to 5.8 GHz. Nevertheless, as h1 becomes longer (5, 8 and 11 mm), the required
Fig. 6 Simulated and measured normalised radiation patterns of the proposed antenna for two planes (left: x-z plane, right: y-z plane).

- 2.4 GHz
- 3.0 GHz
- 4.0 GHz
- 5.2 GHz
- 5.8 GHz

'xxxx' simulated cross-polarisation
'xxoo' simulated co-polarisation
'xxxx' measured cross-polarisation
'xxoo' measured co-polarisation

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low resonant frequency mode is progressively being excited to cover the desired 2.4 GHz frequency with good impedance matching and bandwidth. Hence, by taking a compromise between antenna size and bandwidth, it is recommended that $h_2$ can take any value between 8 and 10 mm for ideal performance.

3.5 Effect of the width of the slot in the shorting wall

The width of the slot in the vertical shorting wall of the driven PIFA governs the impedance matching to cover the designated working frequency bands, particularly for the lower operating frequency mode. The width of the slot (ws) was varied from 0 to 6 mm in increments of 2 mm. As illustrated in Fig. 2c, without the presence of the slot, unsatisfactory impedance matching over the frequency band is present. However, increasing the width (ws) improves impedance matching.

3.6 Effect of the ground plane size

The ground plane length determines the major dimension of the antenna. The size of the ground plane varied from 30 mm × 15 mm to 120 mm × 120 mm, the antenna being located in the corner of the ground plane, as this is the practical reality for the majority of commercial designs. The results are presented in Fig. 2f; it is seen that the ground plane size has a major effect on the reflection coefficient $S_{11}$, but an adequate performance is achieved with the smallest ground plane (30 mm × 15 mm ≈ 0.24λe × 0.12λe, where λe is defined at 2.4 GHz). This makes the design particularly suited to miniature radio device applications. When the ground plane size is larger, a wide impedance bandwidth characteristic with an upper bound outside the measurement range is displayed and the ground plane thus appears to cause the moving of the entire impedance bandwidth to higher cut-off frequencies, but with no improvement in reflection coefficient $S_{11}$ and with a lower cut-off frequency that is higher than the desired 2.4 GHz. As the ground plane becomes smaller than a quarter wavelength, for example, 30 mm × 15 mm (≈0.24λe × 0.12λe), the resonances shift downwards such that they fall into the required bands (2.4–6.0 GHz) with good impedance matching. According to [25, 26], this counter-intuitive effect is because the ground plane resonates around the entire frequency of the antenna element, and so the bandwidth of the antenna—chassis combination will improve considerably. On the contrary, if the ground plane is far away from the operating frequency, the bandwidth will deteriorate owing to its contribution becoming insignificant.

4 Results and discussion

In the parametric study process, a set of optimised geometry parameters has been identified as offering a broadband impedance matching response for the proposed antenna, where $d_1 = 11.1$ mm, $d_2 = 5.5$ mm, $h_1 = 5.3$ mm, $h_2 = 8$ mm, ws = 5 mm and the ground plane size is minimised at 30 mm × 15 mm. An experimental prototype of the proposed antenna was fabricated, as depicted in Fig. 3, to verify the simulated results. The measured and simulated (using HFSS [24]) reflection coefficient $S_{11}$ results are shown in Fig. 4; these exhibit reasonable agreement although there is a frequency shift that can be attributed to fabrication errors in constructing this small antenna. As can be seen, the impedance bandwidth of the antenna encompasses the operating frequency spectrum from 2.4 to 6.2 GHz for a reflection coefficient $S_{11} < -10$ dB, which corresponds to 3.8 GHz bandwidth or about 88.4% relative bandwidth with respect to the centre frequency 4.3 GHz. The bandwidth achieved fully covers the frequency spectrum of WLAN (IEEE 802.11a/b/g, 2.4/5.2/5.8 GHz), WiMAX (2.5/3.5/5.5/5 GHz) and the uplink UWB radio band (3.1–4.8 GHz). The simulated and measured gains of the designed antenna in the broadside direction over the frequency range from 2.4 to 6.0 GHz are shown in Fig. 5. Again, there is some fluctuation because of fabrication errors but it can be observed that a practical useful average gain of 2.95 dBi was measured with ± 1.0 dB of gain fluctuation.

Measurements of the far field radiation patterns of the prototype antenna array were performed in a 1 m³ anechoic chamber using an elevation-over-azimuth positioner, with the elevation axis coincident with the polar axis $(\theta = 0^\circ)$ of the antenna’s co-ordinate system. The azimuth drive thus generated cuts at constant $\phi$. The fixed antenna (reference antenna) was a broadband horn (EMCO type 3115) positioned at 4 m. The elevation positioner was rotated from $\theta = -180$ to $180^\circ$ in increments of $5^\circ$ for the selected measurement. The gain pattern cuts (i.e. $x-z$ and $y-z$ planes) were recorded at five selected operating frequencies: 2.4, 3.0, 4.0, 5.2 and 5.8 GHz, covering the whole of the designated bandwidth in this study. The results are presented in Fig. 6, which shows that the radiation patterns are stable and consistent at all of the designated frequencies. It can be noticed that the nulls exhibited in the patterns over the $yz$-planes, might be because of the consequence of the feed mechanism of the driven PIFA. More importantly, it indicates that the maximum co-polarised component appears in the boresight direction $(+z)$ both for $E$- and $H$-planes and the simulated and measured co-polar radiation patterns are in good agreement with each other (the cross-polar components show more disagreement but they are weak and effectively noise like).

For implementation of the proposed antenna in a UWB system, that is, in particular, impulse-based systems, the shape of the transmitted electrical pulse should not be distorted by the antenna. Thus, a stable group delay response is desirable, which requires a highly linear phase response with respect to frequency. The measured group delay between two identical antennas of this type is portrayed in Fig. 7. The variation is less than 0.5 ns over the frequency band from 3.1 to 4.8 GHz. The excellent
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pulse-handling abilities, with small pulse distortion, of the proposed antenna are thus demonstrated.

5 Conclusion

A dual-planar inverted F-antenna of small size, covering the operating bands of WLAN/Wi-Fi, Wimax and the lower band UWB wireless standard has been presented. The antenna design concept was capable of covering 88.4\% relative impedance bandwidth with acceptable reflection coefficients \( |S_11| \). The antenna has shown consistent omni-directional radiation patterns and reasonable gain values across the operating bands. The experimental results show satisfactory performance and good agreement with the computed results. With its broadband characteristic, the proposed antenna is very well suited to multi-band wireless applications.

6 References

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A Low-Profile Ultra-Wideband Modified Planar Inverted-F Antenna
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Abstract—A miniaturized modified planar inverted-F antenna (PIFA) is presented and experimentally studied. This antenna consists of a planar rectangular monopole top-loaded with a rectangular patch attached to two rectangular plates, one shorted to the ground and the other suspended, both placed at the optimum distance on each side of the planar monopole. The fabricated antenna prototype had a measured impedance bandwidth of 125%, covering 3 to 13 GHz for reflection coefficient better than $-10$ dB. The radiator size was $20 \times 10 \times 7.5$ mm$^3$, making it electrically small over most of the band and suitable for incorporation in mobile devices. The radiation patterns and gains of this antenna have been cross-validated numerically and experimentally and confirm that this antenna has adequate characteristics for short range ultra-wideband wireless applications.

Index Terms—Impedance bandwidth, monopole antenna, PIFA, ultra-wideband.

I. INTRODUCTION

ULTRA-WIDEBAND (UWB) communication systems have several attractive features when compared with traditional narrow band communication systems. These key features include low maximum emitted power (<75 nW/MHz), high data rate (>500 Mbps) over short ranges (<10 m), and large channel capacity which offers up to 7.5 GHz bandwidth over the 3.1 to 10.6 GHz frequency spectrum [1]. This has enabled reliable high-speed real-time high definition video streaming and multimedia file sharing in an indoor environment. In order to make UWB systems competitive with other narrow band communication systems, antennas with compact size, wide impedance bandwidth and consistent omni-directional radiation patterns are in unabated demand.

Due to the low profile, easy integration with circuitry and simple design, printed monopole antennas [2]–[4] have been widely proposed and studied over last decade. In general, these designs achieve broad bandwidth by electromagnetically coupling a radiator with optimized geometry to a defected ground plane with optimum dimensions on a dielectric substrate. However, these design innovations suffer from ground plane size sensitivity, severe backlobes in the radiation pattern and low, inconsistent gains. These drawbacks have limited these antennas to operation in unidirectional communication applications. To rectify this inadequacy, many planar or 3D metal plate monopole antennas [5]–[17] have been proposed, with associated new bandwidth enhancement techniques. These novel methods include modifying the feeding plate silhouette [5], [6], inducing mode coupling between two closely placed elements [7]–[10], shorting and beveling the planar structure [11], offsetting the feeding point from the centre of the structure [12], top-loading the structure with a circular disc patch [13], folding the planar structure into a 3D structure without dielectric [14] and with dielectric [15], gap loading [16] and dielectric loading [17]. Among these designs, it was found that some geometries of these antennas are complicated and difficult to fabricate [7]–[10], [12],–[15], [17]. Some antennas suffer from ground stability issues [7], [9], [10] and some antennas have a relatively large height above the ground ($>0.1\lambda_c$) [5], [6], [11], [13], [16], [17]. These limitations have made such reported works [5]–[17] less favorable to be adopted as a commercial product.

In this paper, a miniaturized low-profile modified planar inverted-F antenna (PIFA) is proposed and investigated. The multi-resonant characteristic of this antenna is realized by combining multiple bandwidth improvement methods including shorting wall, asymmetrical feeding point, broadband feeding plance and top loading. By implementing these techniques, this antenna exhibits a wide impedance bandwidth from 3 to 13 GHz with acceptable radiation characteristics. In section II, the detailed evolution procedures and operating principles will be addressed; the geometry parametric studies and measured results will be discussed in section III. Conclusions are drawn in section IV.

II. ANTENNA DESIGN CONCEPTS AND STRUCTURE

The geometric configuration of the proposed antenna is shown in Fig. 1. The antenna is constructed as a planar inverted-F-shaped radiator with a planar rectangular parasitic strip which is parallel to the feeding plate of the radiator, electrically attached to the top plate, and not grounded. The structure and dimensions of the antenna were chosen on the basis of previous experience [7], combined with realistic constraints on acceptable size for incorporation in a mobile terminal: the dimensions were then optimized in parametric studies. A suspended broadband rectangular feeding plate [7] with optimized feeding gap of 0.75 mm is implemented to excite this antenna assembly and enable its multiple-resonance characteristic. As can be seen, the optimized dimensions of this antenna’s radiator are $20 \times 10 \times 7.5$ mm$^3$ which correspond
to electrical dimensions of $0.2\lambda_s \times 0.1\lambda_s \times 0.075\lambda_s$, where $\lambda_s$ is defined at the lower edge operating frequency (LEOF) of the structure, taken as 3 GHz. For ease of integration with a practical transceiver enclosure, this antenna is mounted on a corner of a $50 \times 50$ mm$^2$ finite ground plane. 0.5 mm thickness copper sheet was used for all conductors in this design.

Because of its aspect ratios, this antenna cannot be absolutely classified as a PIFA and it could be viewed as intermediate between this class and a folded dipole with top capacitive loading. It has some similarities to designs previously published [7], [9], [10], [18]. In [18], the antenna uses a square patch on FR4 substrate as top loading for a rectangular monopole patch and two identical rectangular patches connecting the loading patch to a $50 \times 50$ mm$^2$ ground plane. The total dimension of this previously reported antenna assembly is $50 \times 50 \times 10$ mm$^3$ which is comparable to the presently proposed model, but only achieves a bandwidth of 8.7% (with a center frequency of 2.537 GHz) which is not sufficient to cover the entire UWB spectrum. In [7], the ultra-wide bandwidth was realized by controlling the multi-resonance characteristics of two closely coupled PIFAs. In other work [9], a similar working principle was applied but with a different positional arrangement of the two coupled radiators and an inverted L-parasitic element was used instead of a parasitic PIFA. Recently, authors in [10] have further improved the impedance bandwidth of the structure in [9] by inserting an additional rectangular-shaped parasitic element. Comparing the size of the antennas in [7], [9], [10], it was found that the antenna in [7], including its ground plane, is smaller at $30 \times 15 \times 7.5$ mm$^3$ and it covers the operating frequency band from 3 to 10.6 GHz; in [9], [10] both antenna dimensions including the ground are $28 \times 15.5 \times 4.5$ mm$^3$ and these can operate from 3 to 9.5 GHz [9] and 3.4 to 10.7 GHz [10] respectively. However, all of these antennas are very sensitive to variation of ground plane size in the lower part of the usable operating band, from 3–4 GHz for [7] and 3–5 GHz for [9], [10], and this is due to significant contribution of the ground plane resonance over these bands; this is problematic as the ground plane behavior will be heavily influenced by a human hand holding a mobile terminal.

The proposed antenna model combines different impedance bandwidth enhancing techniques together to enable coverage of the entire FCC UWB frequency band. Compared with the work in [7], the antenna structure (excluding the ground) has been simplified and miniaturized, the impedance bandwidth has been widened and the ground size dependency of the previous designs has been minimized. To understand the basic operational principles of this antenna, the structure has been split into the four differently sized rectangular metal plates which constitute the assembly. This will help to simplify the analysis of the contribution of each metal plate to the antenna in terms of impedance bandwidth and matching over the intended UWB operating frequency band (3.1 to 10.6 GHz). Figs. 2(a) to (d) illustrate the detailed design evolution of the proposed antenna from a planar monopole to a modified PIFA, while Fig. 3 and Fig. 4 show the corresponding performance of the antenna geometries, in terms of reflection coefficient and input impedance respectively.

This analysis starts by considering a rectangular metal plate with dimension of $6.75 \times 14$ mm$^2$ and the off-centre probe feed.
with a feeding gap of 0.75 mm which is used to excite this structure. This forms a conventional planar monopole antenna, as shown in Fig. 2(a). To analytically estimate the LEOF of this antenna, the following simple formula based on the equivalent cylindrical monopole, as described in [19], is adopted

$$F(\text{GHz}) = \frac{75}{L + h + \frac{W}{2\pi}}$$  \hspace{1cm} (1)

where L, W and h are the length, width and feeding gap distance of the planar monopole respectively. Substituting the geometrical parameters ($L = 0.75$, $W = 14$ and $h = 0.75$) in the above expression, it is found that the LEOF is predicted as 7.7 GHz. To validate this estimate, Figs. 3 and 4 depict the numerical predictions of $|S_{11}|$ and input impedance. As can be observed in Fig. 3, the lower and higher edge operating frequencies (HEOF) are 8.2 GHz and 10 GHz respectively, with equivalent impedance bandwidth 19.8%, for a required $|S_{11}| < -10$ dB. Examining the response at 7.7 GHz, the corresponding $|S_{11}|$ is around -6 dB which appears consistent with the estimated value using the formula (for which the LEOF return loss criterion was not stated explicitly). In the input impedance plot of Fig. 4, the antenna exhibits a typical monopole antenna’s impedance curve response with optimum 50 ohm impedance matching at 8.7 GHz, and a parallel resonance occurs at 8.3 GHz.

To reduce the LEOF of the antenna without increasing the height of the monopole component, this is next top-loaded with a $20 \times 10$ mm$^2$ rectangular capacitive plate to form a T-shaped antenna structure, as shown in Fig. 2(b). This is also found to improve the bandwidth; as can be seen in Fig. 3, the impedance bandwidth at the $-10$ dB points is now about 81% and encompasses the frequency band from 5.5 GHz to 13 GHz. This shows a 61.2% improvement of impedance bandwidth in comparison with the monopole. To further explain the physical contribution of the additional top-loading plate, Fig. 4 exhibits the input impedance of this antenna. As can be observed, introducing the top-loading plate causes the single parallel resonance of the original monopole to split into two parallel resonances at 6 GHz and 11.6 GHz with associated resistances of 50 ohm and 83 ohm respectively, and with less variation of reactance (10 to 30 ohm) over the operating frequency band. Carefully tuning of the dimensions of this top-loading plate can bring these two parallel resonances close to each other and leads to good impedance matching over a wide bandwidth.

To achieve the LEOF required of a UWB antenna, i.e. 3.1 GHz, this antenna is further modified by shorting one edge of the top loading plate to the ground using a $5 \times 7.5$ mm$^2$ metal plate, as illustrated in Fig. 2(c). With this modification, the antenna partially resembles a PIFA and the impedance bandwidth has further improved to cover 3.0 to 13 GHz for $|S_{11}| < -8$ dB, as shown in Fig. 3. Investigating the input impedance of this antenna in Fig. 4, it is evident that a shorting wall increases the resistance and reduces the reactance of the antenna in the lowest usable frequency band of the previous antenna model.

Scrutinizing the $|S_{11}|$ plot in Fig. 3, most of the operating frequency band now meet the criterion of $|S_{11}| < -10$ dB, except the band from 9.8 to 11.5 GHz which is $< -8$ dB. In order to enhance the impedance matching of this band, a $8 \times 5.5$ mm$^2$ parasitic metal plate is placed in parallel with the feeding plate. This results in a novel final modified antenna structure as shown in Fig. 2(d). The $|S_{11}|$ plot in Fig. 3 proves that this parallel parasitic metal plate effectively improves the impedance matching of the required band without a deleterious effect on matching of the other frequency bands. Observing the input impedance response of this antenna further clarifies that this parallel plate increases the resistance of the antenna and acts as a capacitive reactance to eliminate part of the inductive reactance of the previous form.

By analyzing the $|S_{11}|$ plot of the proposed antenna in Fig. 3, four adjacent resonant frequencies can be observed over the 3 to 13 GHz frequency spectrum, i.e., 3.4 GHz, 6.6 GHz, 9 GHz and 12.5 GHz. In order to have more indications on the contribution of the individual parts of this antenna at these frequencies, the vector plots of the surface current distributions at these resonant frequencies are investigated in Fig. 5. The first resonant frequency occurs at 3.4 GHz and the total length of the continuous current path is 21.75 mm, which is $\approx 0.25\lambda$. As can be seen in Fig. 5(a), the current flows from the feeding plate to part of the top plate, then to the shorting plate and finally to part of the ground plate to excite this mode. Therefore, the major dimensions of the geometry parameters, including the height of the antenna, coupling distance between the feeding plate and shorting plate and the width of the shorting plate, can be used to manipulate the first resonant frequency.

At the second resonant frequency, 6.6 GHz, one current path can be found from the feeding plate to part of the top plate which is used to connect the shorting plate with the feeding
plate. This current path forms an inverted L-structure, as illustrated in Fig. 3(b). This current path length is about 11.25 mm which corresponds to \( \sim 0.25\lambda \) at this frequency. It seems that the geometric parameters that control the first resonant mode also determine the second resonant mode, although the shorter plate seems to have little influence on the second mode. By analyzing the third resonant frequency, 9 GHz, it is seen that the stronger currents concentrate between the feeding plate and the parasitic plate. It is believed that the capacitive mutual coupling distance between the feeding plate and the parasitic plate has considerable influence on determination of this resonant frequency. The last resonant frequency occurs at 12.5 GHz, as depicted in Fig. 3(d). The full length of the feeding plate structure is 6.75 mm, which is around 0.25\( \lambda \) at this frequency. This structure appears to control the resonant mode of this antenna.

III. RESULTS AND DISCUSSION

To verify the simulated performance of the antenna, a physical prototype was fabricated using 0.5 mm thickness copper plate, as illustrated in Fig. 6. The reflection coefficient \( |S_{11}| \) of the antenna was measured by using a HP8510C vector network analyzer. Fig. 7 shows the computed and experimental reflection coefficient \( |S_{11}| \) of the antenna. From these results, it can clearly be seen that four resonant frequencies occurred, at 3.4 GHz, 6.6 GHz, 9 GHz and 12.5 GHz in simulation and at 3.35 GHz, 7 GHz, 9.7 GHz and 12.5 GHz in measurement. Combining these resonant modes, the antenna achieves a wide impedance bandwidth of 125\%, covering frequency spectrum from 3 GHz to 13 GHz for the criterion \( |S_{11}| \) better than \(-10\) dB. Some disagreements can be found between the simulated and measured results and these can be attributed to the use of glue in the prototype and fabrication errors in constructing it.

A parametric study was conducted in order to understand the sensitivity of the geometric parameters in relation to the reflection coefficient. Three important geometry parameters, were chosen for this study, each being changed while the others were kept constant; they were: d, the coupling distance between the feeding plate and the parasitic plate; \( z_p \), the position of the feeding point along the edge of the feeding plate; and ground plane size (see Fig. 1). From the previous section, it was evident that the presence of the suspended parasitic metal plate, which is placed in parallel with the feeding plate, plays a significant role in controlling wave matching over the 9.8 to 11.5 GHz band. However, the influence of the coupling distance (d) between the feeding plate and the parasitic plate had not been studied. To investigate this, Fig. 8(a) shows variation of \( |S_{11}| \) when changing the distance (d) from 0.5 mm to 3.5 mm in increments of 1 mm. When d is 0.5 mm, the contribution of the parasitic plate is negligible since it does not improve the impedance matching of the band of interest. However, by further increasing d to 1.5 mm, good impedance matching can be attained from 9.8 to 11.5 GHz without a deleterious effect on other frequency bands. This is due to an optimum capacitive coupling effect, as two metal plates are close to each other. By further moving d from 1.5 to 2.5 mm or 3.5 mm, the impedance matching over the required band does not show substantial improvement and impedance matching in other frequency bands will be impaired.

The position of the feeding point on the antenna governs the impedance matching and bandwidth of the antenna in the higher usable operating frequency band. As can be seen in Fig. 8(b), moving the feed point along the edge of the feeding plate from 6.75 mm to 10.75 mm shows that optimum impedance bandwidth as well as good impedance matching can be found when
Fig. 8. Variations of geometric parameters against the reflection coefficients $S_{11}$: (a) spacing of parasitic plate, $d$, (b) position of feed point, $fp$ and (c) dimensions of ground plane.

$fp = 7.75$ mm. However, when $fp$ is taken to values between 8.85 mm and 10.75 mm, impedance mismatch occurs only outside the UWB operating frequency band, i.e., 10.7 GHz–13 GHz. Therefore, it is expected that these values can be noted to be available for further re-engineering purposes to satisfy requirements of different applications, without degrading the performance of the antenna.

Fig. 8(c) depicts the effect of the ground plane size of the antenna on the performance of $S_{11}$. In this study, the antenna is placed at one corner of the ground plane and the five ground plane sizes, i.e., $25 \times 25$ mm$^2 (\equiv 0.25\lambda_0 \times 0.25\lambda_0)$, $50 \times 50$ mm$^2 (\equiv 0.5\lambda_0 \times 0.5\lambda_0)$, $75 \times 75$ mm$^2 (\equiv 0.75\lambda_0 \times 0.75\lambda_0)$, $100 \times 100$ mm$^2$ were used. Of these, $40 \times 100$ mm$^2$ is noted as a more practical ground plane size for a mobile terminal. As can be noticed, the entire UWB frequency band shows very little perturbation over the lower frequency band from 3 to 3.2 GHz for an improved $S_{11} < -9$ dB, which is only $-1$ dB drop, when the ground size is varied from $50 \times 50$ mm$^2$ to $100 \times 100$ mm$^2$. This ground-size stability performance is comparable to the published work in [14] and much better than the results in [7], [9], [10]. When the ground plane size is reduced to $25 \times 25$ mm$^2$, due to the significant contribution of the ground plane resonant mode to the antenna mode [7], it distorts some parts of the lower and upper band impedance matching of the antenna.

Fig. 9 plots the simulated and measured peak gain over the operating frequency from 3 GHz to 13 GHz. As can be noticed, the measured average gain of $\approx 5.5$ dBi with $\pm 1.5$ dBi fluctuation is found from the proposed antennas and the worst variation between the simulated and measured gains is about $\pm 1.3$ dBi. The discrepancies between simulated and measured gains are due to the fabrication error of the antenna prototype and insertion loss of the actual feed network is not taken into the account in the simulated model.

The radiation characteristics of the antenna were examined in an anechoic chamber. Fig. 10 illustrates the simulated and measured radiation patterns of the proposed antenna at 3, 6, 9 and 12 GHz for two principal planes, i.e., $x$-$z$ and $y$-$z$ plane. Each pattern is presented in terms of co-polar and cross-polar components in a $5 \, \text{dB}$ scaled plot. As can be seen, both the computed and measured radiation patterns are in acceptable agreement. Some discrepancy can be found between the simulated and measured results, which can be attributed to the fabrication errors and the physical feeding arrangement. As a consequence, the asymmetrical radiation patterns of the antenna are also caused by implementing an offset coaxial feed in the antenna’s structure. High co- and cross-polarizations of the radiation patterns is also observed in Fig. 10. However, this lack of polarization purity is probably insignificant when the antenna is employed in a rich scattering environment.
Fig. 10. Simulated and measured normalized radiation patterns of the proposed antenna for two planes: (left: y-z plane; right: x-z plane) at (a) 1 Ghz; (b) 6 Ghz; (c) 9 Ghz; and (d) 12 Ghz.

- "xxx" simulated co-polarization
- "xxx" simulated cross-polarization
- "xxx" measured co-polarization
- "xxx" measured cross-polarization

To confirm that the proposed UWB antenna has the ability of handling a pulsed transmission without significant distortion, the group delay and $|S_21|$ of two identical antennas which were placed in different orientations, i.e. 90°, 45°, 0°, −45° and −90° as shown in Fig. 11 with a separation of 45 cm were measured. Fig. 12 shows that the measured group delay of the proposed antenna is stable and holds the same fluctuation, which is less than 0.5 ns, over the 3–13 GHz operating band when two antennas were placed in face-to-face orientation. By substituting the measured $|S_21|$ values into the equations given in [20] and adopting the excitation signal from [21] as described in (2), the normalized excited and received pulse signals are plotted and compared in Fig. 13

$$S(t) = \sin (2\pi f_0(t - t_0)) \exp \left( \frac{(t - t_0)^2}{\tau} \right)$$  (2)

where $f_0 = 6.5$ GHz, $\tau = 0.133$ ns, $t_0 = 4\tau$. This pulse covers the UWB allocated band from 3.1 to 10.6 GHz. As can be seen, both of the pulse signals are almost indistinguishable at all the selected orientations. In addition, the fidelity factor of the antenna was also computed and found to be 89.2%, 85.2%,
81%, 90% and 82% at 90°, 45°, 0°, 45° and 90° orientations respectively.

To further comprehend the performance of the antenna when it is in proximity with human body, a three layers body model as in [14], [23] is adopted in this study. This consists of skin, fat and muscle and their associated dimensions and electrical properties are (120 × 110 × 1 mm³, ϵ_r = 38, σ = 2.7 S/m, ρ[skin] = 1200 kg/m³), (120 × 110 × 3 mm³, ϵ_r = 51, σ = 0.18 S/m, ρ[fat] = 1000 kg/m³) and (120 × 110 × 40 mm³, ϵ_r = 50.8, σ = 3 S/m, ρ[muscle] = 1000 kg/m³) respectively. In this analysis, the antenna is placed in the centre of the body model, with the ground plane facing the surface of the model, at two different proximity distances, i.e. 1 mm and 5 mm, from it. Fig. 14 shows the corresponding effects on S11 and the peak gain. As can be clearly observed, the S11 curves show that the low edge cut off frequency only shifts from 3 to 3.1 GHz when the coupling distance between the antenna and the body is 1 mm. As for the peak gain curves, they show that the gain drops significantly from approximately 3.8 dB to 1 dBi at the lowest operating frequency (3 GHz) and there is a range of 0.5 to 2.8 dB gain reduction over the 3–5 GHz band when the body is 1 mm away from the antenna. Interestingly, in the higher usable frequency band from 10–13 GHz, a slight gain improvement by 1 to 1.7 dBi can be noticed when the body is present.

Fig. 15 shows the corresponding variations of the total radiation efficiency and peak Specific Absorption Rate (SAR) due to the body model effect. SEMCAD X software [22] was used for this analysis. As can be seen, efficiency is impaired most where SAR is highest, which is in the lower and upper usable frequency bands where the radiation patterns are more omni-directional, as depicted in Fig. 10(a), (b) and (d). The body-reduced efficiency reaches 90% around 9 GHz, as shown in Fig. 10(c), where there is least radiation in the lower hemisphere that is directed towards the body. It can be clearly pinpointed that the worst SAR value is predicted at 10.5 W/kg at 3 GHz when the body is 1 mm away. All other values are below 10 W/kg and are comparable to values given for UWB antennas in [14] and [23].

IV. CONCLUSION

A compact and low profile modified PIFA antenna has been proposed and extensively investigated in this paper. This antenna combines impedance matching techniques including top-loading, off-centre rectangular plate feeding and a shorting wall to achieve an ultra-wide impedance bandwidth from 3 GHz to 13 GHz. Despite exhibiting this superior impedance bandwidth, the antenna only occupies a relatively small dimensional envelope of 50 × 50 × 7.5 mm³. The antenna prototype has demonstrated sufficient impedance bandwidth, suitable radiation characteristics, and adequate gains for UWB applications. These attractive characteristics have made it potentially suitable for practical wireless communication applications.

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SMALL WIDEBAND ANTENNA FOR GSM AND WLAN APPLICATIONS

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Abstract—In this paper, a small printed antenna for mobile handset and WLAN applications is presented and discussed. The proposed antenna is a printed crescent shape monopole which is mounted on top of a defected ground plane. A 50 ohm microstrip line is used for feeding purposes. The performance of the proposed antennas was analysed and optimized to cover a wide bandwidth from 1.7 to 3.1 GHz. For validation, the antenna prototypes were fabricated and tested. The performance of this antenna was characterised in terms of the reflection coefficient, radiation pattern and power gain. The calculated and measured results in terms of reflection coefficient show good agreement. The simulated gain and radiation patterns are given to fully describe the performance of this antenna.

Keywords— Monopole antenna, GSM, WLAN, defected ground plane.

1. INTRODUCTION

Wireless communications systems have been investigated and developed over the last few decades, in which there were quite number of standards that deliver various range of high-speed transmission rate for indoor and outdoor data communications. GSM and WLAN including the release of ultra wideband (UWB) wireless communication standard for short range indoor services are good examples of these systems [1]. These varieties of wireless standards offer a greater freedom and convenience in connecting various types of devices working over a number of narrow and broadband frequency bands. A single broadband antenna design is crucial for each device. Thus, many antennas have been designed and studied to cater this unmet demand for these new wireless standards like proposed monopole antennas [2-5].

Monopole antennas basically are becoming very popular for this design requirement, especially in terms of their low profile, wide bandwidth, ease to integrate with PCB, and compact size on a printed circuit board (PCB) [7].

In this paper, a new multiband crescent-shaped microstrip fed antenna, covering most of the existing wireless frequency band allocations from 1.7 GHz to 3.1 GHz, has been designed and analysed. The overall dimensions of this antenna, including the ground size (i.e., equivalent terminal size is 57×37.5×9.8 mm).

II. ANTENNA DESIGN CONCEPT

The antenna geometry is given in Fig. 1. The maximum dimension of this structure is approximately 0.68λ0, computed at 1.7 GHz (i.e., the lower frequency of the DCS band). The radiator is fed by an 18:1 microstrip line. FR4 epoxy is used as the substrate material, with a thickness of 0.8mm, and the dielectric constant (εr) is assumed to be uniformly 4.4, with a loss tangent of 0.017 over the target frequency range. A defected ground plane is located on one side of the substrate, and is truncated close to a point where the feed is coupled to the radiator.

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The underlying design principle is based on the manipulation of multiple resonances, tuned by the modification of the radiator's geometry. It should be noted that the radiator is constructed from the sections of two circles; each having a different radius and centre, thus enabling the resultant patch, taken with the effect of the coupling to the defected ground plane, to radiate over two different frequency bands. The larger radius controls the fundamental frequency, whilst the shorter radius may be tuned to obtain the desired upper frequency. It should be noted that the variations of both resonances are subject to optimum positioning and size of the defected ground plane. When two suitable modes are constructively merged, the desired wideband impedance ground structure plays a significant role in improving the impedance matching across the desired operating bands, but the actual ground plane dimensions are minimized as far as possible to fit the space available inside a transceiver case.

In Fig. 2, the parameters $r_1$, and $r_2$, were varied as part of the design optimisation, using a commercial FDTD analysis program (SEMCD) [9]. Initially a parametric study was made, varying each parameter by 1 mm increments, whilst holding the remaining parameters at their initial values. These values were chosen arbitrarily to fit within the required envelope size; the values for $r_1$ and $r_2$ were found to be 12 mm envelope size; the values for $r_1$ and $r_2$ were found to be 12 mm and 10 mm, respectively. It can be seen, in Fig. 2(a), that the fundamental resonant mode moves down in frequency as the length of $r_1$ is increased, but this does not impair the higher resonant mode. A similar pattern is observed for $r_2$, in that decreasing its length produces a slight upwards shift in frequency, but leaves the fundamental mode unaffected, as depicted in Fig. 2(b). The best performance fit over 1.7 GHz to

Fig. 2: Simulated input return loss corresponding to the variation of parameters (a) $r_1$; (b) $r_2$; (c) ground plane size

Fig. 3: The current distribution at (a) 1.8 GHz; (b) 2.1 GHz; (c) 2.4 GHz and (d) 3 GHz

Fig. 4: Prototype antenna
The effect of the ground plane was monitored by checking the variations in the input return loss against the size of the ground plane. Focusing on the fundamental mode at 1.7GHz, four different ground plane sizes were considered: 21x37.5mm\(^2\) (0.25λ_0x0.385λ_0), 42x42 mm\(^2\) (0.5λ_0x0.5λ_0), 57x37.5mm\(^2\) (0.68λ_0x0.385λ_0), and 84x84mm\(^2\) (λ_0xλ_0). This is summarised in Fig. 2(c), where it can be seen that increasing the ground plane size from 21x37.5 mm\(^2\) to 57x37.5 mm\(^2\) significantly enhances the wideband performance of the impedance bandwidth. Further elongation diminishes the impedance bandwidth. An investigation of the effect of the surface currents over the ground plane was carried out. The computed surface currents for different operating frequencies were shown in Fig. 3, in which almost these currents at 1.8, 2.1, 2.4 and 2.8 GHz are negligible over almost the ground plane except around the feeding antenna port.

Fig 5: Measured and simulated reflection coefficient S\(_{11}\)

Figure 6: Simulated gain of the proposed antenna

III. RESULTS AND DISCUSSION

A prototype of the proposed antenna is portrayed in Fig. 4 for test purposes. Fig. 5 illustrates the typical measured and computed antenna performance in terms of the impedance bandwidth. It is clearly seen that the two adjacent resonant frequencies in the range |S\(_{11}\)| ≤ -10 dB, are 1.7GHz and 3.1GHz. Thus, it is worth noting that the prototype's impedance bandwidth is 1.4GHz, or equivalently 58.3% with respect to the centre frequency at 2.4GHz. This provides adequate coverage for DCS, PCS, UMTS and IEEE 802.11b/g.

Fig 7: Simulated normalised radiation pattern of the proposed antenna for three planes: (a): x-z plane (b): y-z plane (c): x-y plane) at (i) 2100 MHz (ii) 2800 MHz

'xxxx' simulated cross-polarization
'oooo' simulated co-polarization

Fig. 6 plots the simulated antenna gain in the broadside direction for several frequencies across the GSM1800, GSM1900, UMTS and 2.45 GHz bands. The overall simulated gain variations are less than 1.2 dB. The antenna exhibits a good range of gain values in which the maximum and minimum measured gains were found between 2.85 dB and 1.65 dB, over the aggregate bandwidth. The computed far field radiation patterns are depicted in Fig. 7. Three pattern cuts (i.e. the xz, yz and xy planes) were taken at two selected operating frequencies, i.e. 2100MHz and 2800MHz. It is hard to state an objective criterion for optimising the pattern, since it depends on how much effort the proposed antenna mounted on the wireless device is prepared to obtain a good orientation, and there is no completely satisfactory analysis in the literature as to what constitutes a "good" antenna pattern. Nevertheless it seems reasonably obvious that a good pattern should have its peak radiation in a near-horizontal direction, and not at a very high or very low elevation, when the set is operated in its usual position with its axis at about 45° from horizontal. It can also be said that a good compromise design target is to seek a pattern which has elevation directivity similar to that of a simple dipole, to ensure that a peak occurs at zero elevation, and to seek a design which radiates mainly omnidirectional. It can be seen, the peak azimuthal gain variations (i.e., the co-polar component shown in Fig. 7c) was varied between 0 – 6 dB except at the
the lower selected frequency 2100 MHz. This is due to that
the pattern is quite substantially affected by the asymmetry of
the ground plane and its position in the proposed design. In
addition, the surface current presented in Fig. 3, have shown
that strong excitation current at and near the feeding port that
can produce an inherently symmetric and weak pattern. It is
very interesting as well to observe that the peak elevation gain
values (the co-polar components of Figs. 7a and 7b) achieved
were quite similar to that found at the azimuth direction.
These pattern results will confirm to minimize the occurrence
of periods of very bad reception.

IV. CONCLUSION

In this paper a low profile multi-frequency band monopole
antenna module has been presented. The radiator is a crescent
shaped microstrip patch. The realised antenna structure shows
a relative bandwidth of 51.8%, and a gain above 1.5dBi over
the frequency interval 1.75GHz to 3.1GHz. The antenna size
was optimised at 57mm×37.5mm×0.8mm, and is suitable for
integration with a variety of mobile terminals operating over
DCS, PCS, UMTS, Bluetooth or IEEE 802.11b/g wireless
standards.

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Design of a Planar Inverted F-L Antenna (PIFLA) for Lower-band UWB Applications


Abstract — This paper examines the case for an ultrawideband planar inverted F-L antenna design intended for use in the lower sub-band. The antenna is based on the conventional inverted F, and inverted L as its feed element, and parasitic element, respectively. The optimized antenna size is 30×15×4 mm$. The prototype antenna has a good return loss of -10 dB, and a 66.6% impedance bandwidth (2.8 GHz - 5.6 GHz), the gain varies between 3.1 dBi and 4.5 dBi.

Keywords — Planar inverted-F-L antenna, broadband, ultrawideband, impedance bandwidth.

I. INTRODUCTION

A reliable low cost antenna module is a necessity for a multi-band mobile user terminal. Mechanically the module must be low profile, robust and cost effective through high volume production. From an electrical standpoint it must meet the requirement specification derived from the relevant communications standards, and have low coupling when placed or held in proximity to the human body. One such option is the planar inverted F antenna (PIFA) [1, 2]. However, the conventional PIFA has bandwidth constraints; typically 4-12% impedance bandwidth is achievable in an unmodified design. Modified PIFA designs include T-slot geometries [3], the introduction of parasitic elements, such as the inverted L [4, 5], modified ground planes [6, 7] and modifications to the feeding and shorting plates [8, 9]. The design adopted here is a miniaturized "FL" antenna, or PIFLA [10]. A novel feeding structure is introduced to achieve a good impedance bandwidth for the design frequency range, which is the lower sub-band (3.100 MHz - 4.800 MHz) of the UWB spectrum [11]. The design optimization is carried out using a frequency domain finite element analysis (Ansoft HFSS), the final model is also cross validated through the time domain using a conformal FDTD method (SEM-CAD). A working prototype is constructed and tested on this basis.

II. ANTENNA DESIGN

The PIFLA schematic is shown in Fig. 1. It is constructed from a standard PIFA with a broadband feeding plate, and a parasitic planar inverted L-shaped element. For ease of integration with a practical enclosure, the assembly is mounted over a finite ground plane. The minimum dimensions for the PIFA are 17.5×13×3.5 mm$^3$, where the 17.5 mm is the length of F shape and 3.5 mm is the antenna height. The minimum dimensions of PIFLA are 9.5×14×3.5 mm$^3$ for the parasitic element PIFLA and the ground plane dimensions for the final assembly are 30×15 mm$^2$. Optimal coupling is achieved with an element separation distance of 3 mm. The lowest resonant mode of this structure is approximately 3000 MHz, so if the corresponding wavelength of this mode is $\lambda_0$, then the scaled optimal antenna dimensions are $0.30\lambda_0\times0.15\lambda_0\times0.04\lambda_0$.

Fig. 1: Geometry of the proposed antenna.

The underlying design principle is straightforwardly based on the manipulation of multiple resonances, from the inverted F and L elements, both of which support strong current distributions. The driven PIFA element acts as the primary element, governing the lowest resonant frequency, whilst the higher resonant frequency is controlled by the parasitic element. Both the size of the ground plane, and the feed mechanism play a significant role in determining the desired wideband characteristic for the impedance bandwidth.
However, it should be noted that the lower frequency component is dominated by the large effect of the capacitive coupling between the feeding element and the finite ground. Thus, a broadband rectangular plate, with a 0.5 mm gap, is used to excite the PIFA. This technique provides an improved impedance matching over the conventional probe feed. The optimum gap width was achieved after several model attempts. To further adapt this antenna for commercial wireless transceiver applications, ground plane size effect of the proposed antenna is crucial. The influence of the ground plane size using the proposed feeding plate mechanism is fully addressed in the following section.

![Prototype of proposed antenna](image)

Figure 2: Practical Prototype of proposed antenna

III. PARAMETRIC DESIGN STUDIES

The parametric study is useful because it provides a comprehensive picture of the antenna characteristic. The first cut design (in the frequency domain) was made by varying each parameter by 1mm and 2mm increments, whilst holding the remaining parameters at their initial values. Fig. 3(a) exhibits the simulated return loss for different height of the antenna as can be seen that the -10dB return loss bandwidth of the antenna is achieved at 3.5mm. It can be seen very clearly in Fig. 3(b) that the lengths of PIFA has come at the optimised value of 17.5mm is the best size of the PIFA length. Fig.3 (c) shows the variations of the reflection coefficient $|S_{11}|$ with five selected ground plane sizes; once the ground plane size was reduced to 15x30mm$^2$ the optimised result was obtained Fig.3 (c).

![Reflection coefficients](image)

Figure 3: Simulated reflection coefficients $|S_{11}|$ with different dimensions of (a) antenna height h, (b) PIFA length $p_1$ and (c) ground plane.
IV. RESULTS AND DISCUSSION

Fig. 4 compares the predicted (frequency domain, HFSS) and measured reflection coefficient response, the impedance bandwidth of the prototype operates over the range 2800 MHz to 5600 MHz, with $|S_{11}| \leq -10$ corresponding to a 66.7% relative bandwidth with respect to a centre frequency of 4200 MHz. This operating range gives full coverage of the UWB uplink frequency spectrum (3100 MHz to 4800 MHz). There is small discrepancy between predicted and measured results, but this does not indicate the need for additional design optimization, this may be a construction error.

![Figure 4: Measured and simulated reflection coefficients $|S_{11}|$](image)

The effect of the ground plane is a significant aspect in this design. Fig. 3 (c) shows the variations of the reflection coefficient with five possible ground planes (scaled in units of $\lambda_o$): $0.80 \lambda_o$, $0.80 \lambda_o$, $0.60 \lambda_o$, $0.60 \lambda_o$, $0.40 \lambda_o$, and $0.30 \lambda_o$. The performance is significantly degraded as the size of the ground plane is increased, as the resonant mode of the ground plane is essentially out of band, and therefore makes no constructive contribution to the impedance bandwidth.

Fig. 5(a) gives the predicted and measured gain of the antenna in the broadside direction, over the interval 3000 MHz to 5750 MHz. It was interesting to find an average measurement gain of 3.6 dBi, with ±1.1 dB fluctuation. Fig. 5(b) exhibits the simulated radiation efficiencies of the antenna. The average measurement efficiency was observed to be 92.5% with ±5.5% efficiency fluctuation. This compares well with the simulations.

![Figure 5: Measured gains and radiation efficiencies for proposed antenna; (a) antenna gain, (b) radiation efficiency.](image)

The far field radiation patterns are presented in Fig. 6. Two pattern cuts (the $xy$ and $yz$ planes) were taken at three selected operating frequencies which cover the aggregate bandwidth. The radiation patterns were found to be stable and consistent at all the designated frequencies, as shown in Fig. 6. Significantly, it also indicates that the maximum co-polarized component appears at the direction of bore sight ($yz$) for both the E and H planes.
Figure 6: Simulated and measured normalized radiation patterns of the proposed antenna for two planes (left: x-z plane, right: y-z plane) at (a) 3000 MHz (b) 4000 MHz, and (c) 5000 MHz. 'xxxx' simulated cross-polarization 'oooo' simulated co-polarization '------' measured cross-polarization '-----' measured co-polarization

The group delay spread in the required bandwidth is approximately 0.5 ns. The group delay measurement is given in Fig. 7.

V. CONCLUSION

A wideband planar inverted F-L antenna (PIFLA) has been designed and studied. The optimized design is a compromise between antenna size and impedance bandwidth in the lower operating UWB frequency spectrum. The prototype package dimensions are 30×15×6 mm$^3$. This antenna displays stable radiation patterns, gain performance, and radiation efficiency for the entire bandwidth of operation, thus making it suitable for UWB applications.

REFERENCES

PIFA Antenna for UWB Applications with WLAN Band Rejection using Spiral Slots


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Abstract—This paper presents a band-suppressed PIFA antenna for UWB application using spiral slot. The modeling of the antenna geometry structures and the optimization of the full antenna assembly was carried out in the frequency and time domains using Ansoft HFSS and CST respectively for comparison. The impedance bandwidth for reflection coefficient \(S_11\), less than 10 dB covers the range from 3 GHz to 12 GHz, with band rejection from 4.75 GHz to 5.75 GHz. The simulated antenna gains and radiation patterns are given also, to fully describe the antenna performance.

I. INTRODUCTION

Ultra-wideband (UWB) wireless technology in most commercial applications follows the IEEE802.15a standard as a guideline, alongside other regulatory requirements [1]. In practice these are not very well defined, and in addition to the standard design requirements for impedance matching and radiation stability, the antenna designer must also be aware of the potential interference from other mobile service bands lying within the 3.1 GHz - 10.6 GHz range. Further system level design constraints follow from the precise nature of the application. Multiple-band operation with several overlapping standards is becoming a familiar requirement, however, in addition, the wider electromagnetic environment arising from ubiquitous networking must also be considered. Here we are concerned with creating a viable mobile UWB antenna which must co-exist with the WLAN service bands centered at 5.2 GHz and 5.8 GHz.

In this example we concentrate on band rejection of UWB 5.2 GHz band which covers the frequency spectrum from 5.15 to 5.35 GHz. Planar inverted-F antennas (PIFAs) are among the most widely used antennas for mobile terminals, and have the potential for wide-band operation [2-4]. PIFAs may be modified to operate over multiple frequency bands by using, e.g., parasitic antenna elements, and or cuts to the antenna element to form several paths for surface current. Some UWB variants have been reported [5, 6], however antennas with band notched characteristics are preferred. Some antennas have been proposed with a rejection band or band

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II. ANTENNA DESIGN CONCEPT

In this work, the proposed antenna is very similar to author's previous work, as in [11], except the feeding plate has been modified to enable the band filtering function. The geometry of the proposed UWB spiral slot antenna is illustrated in Fig 1(a), while Fig 1(b) shows the dimension of the feeding plate with spiral slot. This antenna is constructed by using two PIFA with dimension of 18.5mm x 10mm x 4.7mm and 7.5mm x 14.5mm x 7.5mm, respectively. The antenna assembly is mounted on a finite ground plane whose dimensions are 30 x 15 mm$^2$. Once again, the separation between the two PIFA elements is set at 4mm, for optimal coupling. The conventional PIFA wire feed mechanism is replaced by a rectangular plate feed with spiral slot of 3 x 3 mm$^2$ to control the bandwidth performance of the notch.

III. RESULT AND DISCUSSION

Fig 2(a) shows the variation of the reflection coefficient $S_{11}$ for three selected ground plane sizes, it can be seen that the 30 x 15 mm$^2$ dimension should be chosen as the optimized size for the ground plane in order to obtain the desirable frequency range, as shown in Fig 2(a). The simulated the reflection coefficient $S_{11}$ for the three different antenna heights is plotted in Fig 2(b). The notched frequency requirement was achieved at 4.2mm. The feed point is the most sensitive parameter, three different positions were selected to get a good result. From Fig 2(c) can be noted that 7.5mm is the best position for the feed point. Reflection coefficient $S_{11}$ was simulated by using two commercial packages (HFSS and CST) [12, 13] for cross-validation purpose. With the optimum geometry parameters, the simulated reflection coefficient $S_{11}$ of the proposed antenna is illustrated in Fig 4. By examining the reflection coefficient $S_{11} < -10$dB, the impedance bandwidth covers the range from 3 GHz to 13 GHz, with band rejection from 4.75 GHz to 5.75 GHz for WLAN communication system.

Measurements of the antenna gain in the entire operating band have been also performed and results are presented in Fig 5. As observed, a sharp gain decrease is obtained in the notched frequency.

The far field radiation patterns are presented in Fig 3. Two pattern cuts (the $x_2$ and $y_2$ planes) were taken at four selected operating frequencies, i.e. 3 GHz, 5 GHz, 7 GHz and 9 GHz which cover the aggregate bandwidth.

The radiation patterns were found to be stable and consistent at the designated frequencies, as shown in Fig 3. Significantly, it also indicates that the maximum co-polarized component appears at the direction of bore sight ($-z$) for both the $E$ and $H$ planes.

IV. CONCLUSION

The band-notch ultra-wideband antenna using spiral slot has presented with the notch over the frequency 5.25 GHz (4.75-5.75 GHz). Various parameters of the antenna are optimised and various techniques have been adopted to create
Figure 3: Simulated antenna radiation patterns for two planes (left: x-z plane, right: y-z plane): (a) 3 GHz, (b) 5 GHz, (c) 7 GHz and (d) 9 GHz. Where 'coo' is the co-polarized component and 'xpol' is the cross-polarized component.


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Miniaturised UWB Antenna for Wireless Body Area Network


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Abstract—In this paper, a miniaturised modified planar inverted-F UWB antenna with a broadband rectangular feeding structure for Wireless Body Area Network (WBAN) is proposed. The proposed antenna is designed to cover a wide frequency band from 3 GHz to 13 GHz. Details of the proposed antenna design and measured results are presented and discussed. To validate the results, a prototype design was fabricated and tested. The measured and computed results are in good agreement. Overall size of the antenna including ground plane is 50 mm × 50 mm × 8 mm making it suitable to fit into a standard enclosure of a UWB wireless transceiver.

Keywords—UWB antenna, Wireless Body Area Network.

I. INTRODUCTION

Since the Federal Communication Commission (FCC) released the UWB frequency spectrum from 3.1 to 10.6 GHz in 2002, it has opened up tremendous business opportunities to the mobile wireless transceiver manufacturers to update their products by incorporating ultra-wideband (UWB) technology. UWB technology holds great promise for a vast array of new applications that have the potential to provide significant benefits for public safety, businesses and consumers in a variety of applications such as radar imaging of objects buried under ground or behind walls and short-range, high-speed data transmissions.

The ultra wideband (UWB) systems require antennas having a very broad bandwidth [1]. In order to make UWB systems competitive with other narrow-band communication systems, UWB antennas with compact size, wide impedance bandwidth and consistent omni-directional radiation patterns are in unmet demand.

It is well known that conventional planar inverted-F Antenna (PIA) is a good candidate for mobile handheld devices. However, this antenna has bandwidth constraints; typically 4-12% impedance bandwidth is achievable in an unmodified design. Many bandwidth enhancement methods have been proposed to overcome this deficiency. These methods include adding a T-slot geometry [2], the introduction of parasitic elements, such as the inverted-L [3, 4], modified ground planes [5, 6] and modifications to the feeding and shorting plates [7, 8].

In this paper, a new small antenna design of UWB system for wireless body area network is studied and investigated. The antenna is formed by a rectangular top plate, shorting wall, rectangular feeding plate, a parasitic plate and a finite ground plane. To evaluate the performance of the antenna when it is placed next to the human body, a simulated model was constructed to analyze this scenario.

Figure 1: Proposed antenna structure. (a) top view, (b) 3D
II. ANTENNA DESIGN CONCEPT AND STRUCTURE

Fig. 1 shows the geometry of the antenna. The antenna is constructed by four metal plates, which including a top plate (20 x 10 mm²), shorting wall plate (5 x 7.5 mm²), feeding plate (14 x 6.75 mm²) and parasitic plate (8 x 3.5 mm²), as shown in Fig. 1. A optimum feeding gap of 0.75 mm is used to excite the antenna for the best impedance matching of the antenna. Off centre feeding method is also introduced in order to widen the impedance bandwidth of the antenna. By carefully optimising the geometry parameters, i.e. the dimension of all the metal plates, feeding gap, feeding position and the position of the parasitic plate, of the antenna, the wide band impedance bandwidth can be achieved.

Parametric study has been carried out to optimize the impedance matching bandwidth for the proposed antenna in order to achieve the required impedance matching covering the frequency band of interest from 3 GHz to 13 GHz. The feed pin position (fp), the distance between feed plate and parasitic plate (d) and the ground plane size were considered to be the most sensitive parameters to control the impedance bandwidth of the proposed antenna for meeting the design goals. Therefore, these parameters were considered for this analysis.

The position of the feed point (fp) is one of the most sensitive parameters in this study. By varying the position of the feed point from 4.75 to 7.75 mm with 1 mm each step as illustrated in Fig. 2(a). As can be seen from this figure, when the fp is set at 4.75 mm, 7.75 mm and 6.75 mm the target frequency range would not be obtained. However, the antenna will exhibit the target frequency range when fp set at 7.75 mm (i.e. close to the middle of the plate).

Effect of the position of the additional rectangular plate (d) influences the impedance matching of the FCC required UWB band. This is because its variation can introduce different excitation modes to the antenna, as indicated in Fig. 2(b). It is clearly seen, when d is 7 mm away from the shorting pin metal, antenna shows an obvious wide bandwidth with good impedance matching, which completely covers the band of UWB frequency range from (3.1 to 10.6) GHz.

The effects of ground plane size on antenna reflection coefficient |S11| were investigated parametrically, as presented in Fig. 2(c). The physical dimensions of the ground plane were scaled in terms of λg, the wavelength corresponding to the lowest resonant frequency of the structure. The response was analyzed for four different ground plane sizes: 25 mm x 25 mm (0.25λg x 0.25λg), 50 mm x 50 mm (0.5λg x 0.5λg), 75 mm x 75 mm (0.75λg x 0.75λg), and 100 mm x 100 mm (1λg x 1λg), where λg is set at 3 GHz. As can be clearly noticed that, the antenna with ground plane dimensions larger than 0.5λg will have a better impedance matching over the required UWB band, while further reducing this ground plane size to 0.25λg will deteriorate the impedance bandwidth at lower usable frequency band.

III. RESULTS AND DISCUSSION

To validate the simulated results, a antenna prototype is fabricated, as show in Fig. 3. The simulated reflection coefficient of the reference model and the measured reflection coefficient of the physical prototype are compared in Fig. 4. As can be observed, the simulated and measured |S11| are in satisfactory agreement, in which both of them encompassing the frequency band from 3 GHz to 13 GHz defining at |S11| = 10 dB. There are some discrepancies between the simulated and measured results can be attributed to the fabrication errors.
Fig. 5. Shows the simulated gain of the proposed antenna across the frequency range between 5GHz and 10.6GHz. It can be seen that the maximum simulated gain over the frequency range is 7.26dBi and the minimum is 2.65dBi.

To comprehend the antenna performance when it is placed in proximity to a human body model, a simulated model has been set up to investigate this problem. Three layers body model as in [9] is adopted in this study. This consists of skin, fat and muscle, the antenna is placed in the centre of the body model with the ground plane facing the surface of the model and at two different proximity distances, 1 mm and 5 mm as shown in Fig. 6. The radiation efficiency and peak Specific Absorption Rate (SAR) are presented in Table 1 at three frequencies 5GHz, 6GHz and 9GHz. (SEM-CAD X software [10] was used in this analysis.

| Three Layers Body model | | |
|---|---|---|---|
| Body feed distance | 1mm | 5mm |
| 5 | 40% | 10.5 | 48% | 6.5 |
| 6 | 70% | 5.3 | 76% | 2 |
| 9 | 90% | 1 | 90% | 0.7 |

Fig. 7. Depicts simulated radiation patterns of the proposed antenna. Two pattern cuts (i.e., y-z plane), each pattern presented in terms of co-polar and cross-polar, the patterns of the proposed antenna were taken at three selected operating frequencies 5GHz, 6GHz and 9GHz as shown below that cover the whole of the designated bandwidth in this study.
Figure 7: simulated radiation patterns for two planes (left: x-z plane, right: y-z plane); at (a) 2000 MHz, (b) 4000 MHz and (c) 6000 MHz; —— from co-polarization and —— co-polarization

IV. CONCLUSION

An UWB planar antenna with optimized size of 50 x 50 x 3 mm for wireless body area network is studied and investigated in this paper. The proposed design offers a wide frequency spectrum from 3 to 13 GHz. To validate the computed results, a prototype design has been manufactured and tested. Both measured and computed results have shown a good agreement in terms of reflection coefficient. The simulated far field radiation patterns and gains also were plotted to further confirm its radiation type. These attractive features make the antenna very promising candidate for UWB application.