Design of UWB Radar Sensors



Amnoiy Ruengwaree

kassel university



This work has been accepted by the faculty of electrical engineering / computer science of the University of Kassel as a thesis for acquiring the academic degree of Doktor der Ingenieurwissenschaften (Dr.-Ing.).

Supervisor: Prof. Dr.-Ing. G. Kompa Co-Supervisor: Prof. Dr. H. Hillmer

Commission members:

Prof. Dr.-Ing. J. Börcsök Prof. Dr. K. Geihs

Defense day:

15th November 2007

Bibliographic information published by Deutsche Nationalbibliothek The Deutsche Nationalbibliothek lists this publication in the Deutsche Nationalbibliografie; detailed bibliographic data is available in the Internet at http://dnb.d-nb.de.

Zugl.: Kassel, Univ., Diss. 2007 ISBN: 978-3-89958-358-8 URN: urn:nbn:de:0002-3586

© 2008, kassel university press GmbH, Kassel www.upress.uni-kassel.de

Printed by: Unidruckerei, University of Kassel Printed in Germany

Acknowledgements

First and foremost, I would like to express my deepest gratitude to my advisor, Professor Dr.-Ing. Günter Kompa, for his continual support and motivation. He has provided me with guidance and a sense of direction through all phases of this research. His level of understanding and expertise in work and in life has made my graduate experience enjoyable in and out of the department.

I would like to acknowledge the financial support of Rajamangala University of Technology Thunyaburi (RMUTT) in Thailand giving me the opportunity to do the research in Germany.

I would like to thank Dr.-Ing. A. Ghose, Dr.-Ing. B. Bunz and Dipl.-Ing. J. Weide, for their cooperation along with technical help. I am thankful to Mr. E. R. Srinidhi, Mr. A. Z. Markos, and Mr. E. Mengistu for doing a very good job in proof-reading the dissertation. Furthermore, I would like also to thank the secretaries of the HFT department, Frau P. Castillo (former) and Frau H. Nauditt for their help throughout the duration of the doctoral dissertation. Also, I am grateful to all my past and present colleagues of the HFT department for being the surrogate family and for their continued moral support over the past years. This work would not have been possible but for the help given by Mr. R. Yuwono and Mr. M. El-Hadidy who have worked on their respective Master's thesis and project works.

I would like to thank K. Marczykowski for the love and encouragement that she gave me at the time when everything looked impossible. Finally, my deepest gratitude goes to my father Prachoen, my mother Ampai and my sister Napa, who always encouraged and motivated me to continue my higher studies.

Table of Contents

1	Intro	oduction	1
	1.1	Objective of Work	1
	1.2	Organization of Work	4
		List of References	6
2	UW	B Radar Evolution and System Aspects	8
	2.1	Historical Review	8
	2.2	UWB Specification	11
	2.3	UWB Antennas	12
		2.3.1 Definition of UWB Antenna	12
		2.3.2 Determining Antenna Bandwidth	13
		2.3.3 Dispersion	14
	2.4	The UWB Radar Sensor Concepts	18
		2.4.1 Bi-Static Radar Sensor	19
		2.4.2 Mono-Static Radar Sensor	25
		List of Reference	28
3	UW	B Rugby-Ball Antenna	33
	3.1	Introduction	33
	3.2	Design Procedure of Antenna Structure	34
	3.3		36
		3.3.1 Simulations	36
		3.3.2 Fabrication and Measurements	48
		3.3.3 Results	53
	3.4	Reducing Weight of Antenna	64
		List of References	67
4	Desc	cription of the Developed Radar Modules	70
	4.1	Multi-Stage SRD-Pulse Sharpening Circuit	70
		4.1.1 Properties of the SRD	71
		4.1.2 Used ADS Model of SRD	76
		4.1.3 Circuit Design	79

		4.1.4 Experimental Results	81		
	4.2	Broadband Resistive Duplexer	85		
		4.2.1 Design	85		
		4.2.2 Fabrication and Results	87		
		List of References	90		
5	Appl	ication of New UWB Microwave Radar Sensor	92		
	5.1	Radar Sensor Measurement Setup	92		
	5.2	Distance Measurement to Metal Plates	97		
	5.3	Detection of Metal Plates Buried in Dry Sand	102		
	5.4	Water Level Control Measurement	103		
	5.5	Ranging Uncertainty	107		
		5.5.1 Measurement Accuracy of Bi-Static Radar Sensor	107		
		5.5.2 Measurement Accuracy of Mono-Static Radar Sensor	108		
		List of Reference	110		
6	Conc	lusion and Future Work	111		
Appe	ndix				
А	Reflection Properties of Water				
	A.1	Reflection Factor	114		
	A.2	Dielectric Constant of Water	115		
	List	of Reference	117		
Publi	cation	S	118		

Zusammenfassung

Gegenstand dieser Arbeit ist die Entwicklung von bi- und mono-statischen Mikrowellenradarsensoren zur Nahfelddetektion, Entfernungsmessung und Füllstandskontrolle. Wesentliche Komponenten des entwickelten Ultrabreitband-Radars (UWB), wie UWB-Antenne, mehrstufige Impulsversteilerungsschaltung and Breitband-Duplexer, werden detailliert beschrieben.

Es wurde eine neue Antenne mit der Bezeichnung "Rugby-Ball Antenna" entwickelt, welche als Sende- und Empfangsantenne eingesetzt wird. Die Einspeisung der Antenne erfolgt über eine Grundplatte in der Bildebene mit zusätzlichem Reflektor. Das Antennenelement wird durch die Schnittmenge zweier Kreisflächen mit unterschiedlichen Kurvenradien gebildet. Die Bandbreite der Antenne wird durch deren oberen und unteren Krümmungsradius bestimmt. Die entwickelte Antenne weist eine große Bandbreite von 187,4% auf, welche hervorragend zur Aussendung und Detektion von ultrakurzen elektrischen Impulsen geeignet ist. Das Ziel des Radarentwurfs war die Realisierung einer Antenne, welche kurze Impulse mit einer Anstiegszeit (t_r) von etwa 50 ps (10-90%) und einer Impulsbreite (FWHM) von 75 ps senden und empfangen kann. Zur Erreichung dieses Ziels wurden mit Hilfe des elektromagnetischen 3D Feldsimulators "High Frequency Structure Simulator" 10. (HFSS, Version Ansoft Corp.) Simulationen verschiedener Antennenkonfigationen durchgeführt. Bei der endgültigen Version erfolgt die maximale Amplitude der Impulsabstrahlung unter einem Erhebungswinkel von 60° in Quer- und Kantenrichtung. In azimuthaler Richtung wird die maximale Abstrahlung bei 0° und 90° erreicht, und zwar in Quer- bzw. Kantenrichtung.

Es wurde mit Hilfe der Software ADS (Advanced Design System) eine neue Konfiguration eines resistiven Duplexers entwickelt. Dieser 3-Kanal Duplexer wurde in Mikrostreifenleitungstechnik realisiert. Das im Impulsversteilerungsschaltkreis erzeugte Radar-Ausgangssignal wird im Duplexer in zwei Pfade aufgeteilt. Ein Teil speist die Antenne zur Abstrahlung des Radarsendesignals und der andere Teil dient als Radarreferenzpuls. Beide, Signalereferenzimpuls und der verzögerte Messimpuls (vom Ziel reflektierter Impuls), werden an die Samplingeinheit weitergeleitet, welche die Pulse zur präzisen Laufzeitmessung in zeitgedehnte Signale umwandelt. Der Duplexer deckt eine Bandbreite von 12 GHz ab. Der Eigen-Reflexionsfaktor ist kleiner als -10 dB. Der Übertragungsfaktor beträgt mehr als -3 dB.

Es wurden bi-statische und mono-statische Radarsensoren aufgebaut und Experimente innerhalb einer maximalen Messentfernung von 100 cm durchgeführt. Es wurden Metallplatten im freien Raum und vergraben in trockenem Sand gemessen. Zusätzlich wurden Füllstandsmessungen in einem Wassertank in einem Messbereich von 0 bis 80 cm ausgeführt.

Untersuchungen zur Messunsicherheit ergaben einen Wert von 4.5 mm für den bi-statischen bzw. 5.5 mm für den mono-statischen Radarsensor.

Abstract

The objective of this work is the development of concepts for the design of bi-static and mono-static microwave radar sensors for near-field detection, ranging and level control measurement. Design details are presented for relevant components of ultra-wideband (UWB) radar, which includes UWB antenna, multi-stage SRD-pulse sharpening circuit and broadband duplexer.

A new antenna named "rugby-ball antenna", has been developed as the radiating and receiving element. The antenna is fed through the image ground plane and backed by a reflector. Intersection of two circular areas with different radius of curvature shapes the antenna element. Antenna bandwidth is determined by upper and lower radius of curvature of the antenna element. The developed antenna exhibits a large bandwidth of 187.4%, which is well suited for the ultra-short electrical pulse detection. The goal of the radar design was to realize an antenna that can transmit and receive short pulses with a 10%-90% rise time (t_r) of approximately 50 ps and pulse width (FWHM) of 75 ps. To achieve this goal, simulation of diverse antenna configurations has been performed using the full-wave electromagnetic solver "High Frequency Structure Simulator" (HFSS, version 10, Ansoft Corp.). The pulse radiation with maximum amplitude of the antenna occurs at an elevation angle of 60° from broadside and edge-on directions. Regarding the azimuth angle, maximum radiation is observed at 0° and 90° from broadside and edge-on directions, respectively.

A new pulse sharpening circuit has been developed for the radar sensor. It includes a multi-stage design with beamlead packaged step recovery diode (SRD). The circuit produces electrical pulses with a pulse width (FWHM) of 75 ps, a rise time of 50 ps and a peak voltage of 1.25V. The pulses provide the output signal of the radar.

A new configuration of broadband resistive duplexer has been designed using ADS (Advanced Design System) software. This 3-port duplexer has been realized in microstrip technique. The radar output signal generated in the pulse sharpening circuit is split through the duplexer into two parts. One part is fed to the antenna for radiation and the other part serves as a reference pulse. Both reference pulse and the delayed measuring pulse (pulse reflected from the target) are transmitted to a sampling unit, which converts the pulses into time-extended ones for precise time-of-flight measurement. The duplexer covers a bandwidth of 12 GHz. The return losses are more than 10 dB. The insertion losses have turned out to be less than 3 dB.

Experiments were performed using a fabricated bi-static and a mono-static radar sensor with a maximum detection range of 100 cm. Targets such as metal plate in free space and metal plates buried in dry sand were measured. In addition, water level control measurement in rainwater tank was performed within a water level interval of 0 to 80 cm.

The long-term range measurement uncertainty of the radar sensor was found to be less than 4.5 mm for the bi-static system and 5.5 mm for the mono-static configuration.

Chapter 1

Introduction

1.1 Objective of Work

This thesis represents a continuation of previous work in the high frequency engineering (HFT) department, which has long experience both in near-field laser and microwave radar technology. Already in 1984 short-range pulsed laser radar was built, using the sampling principle for low-cost range detection [1]. Recent laser radar development [2] in the HFT department includes optical pulses of approximately 450W peak power, pulse width (FWHM) of 44 ps, and rise time of about 32 ps. The laser radar sensor has been used for 2D and 3D imaging of targets [3].

In 2001, the proven laser radar concept was used to realize the UWB microwave radar in bi-static configuration. The key point was to replace the optical head of the laser radar by an UWB antenna and to provide adequate fast electrical pulses. In his work, Duzdar [4] established a first version of an UWB antenna. The antenna, named "vertical inverted trapezoidal antenna", is fed through an image ground plane and backed by a reflector. The antenna is made of aluminium and has a relative bandwidth of 62.5%. This covers a frequency range of 1 GHz to 5 GHz. The respective voltage standing wave ratio (VSWR) is less than 2. Pulse radiation occurs both from the side edges and broadside of the antenna. Maximum pulse amplitude occurs at an elevation angle of 60° and

35° from edges and broadside directions, respectively. These excellent data were confirmed by researchers at Helsinki University of Technology [5].

For fast electrical pulse, the pulse sharpening circuit [4] was developed with physical dimension of 4 cm x 6 cm which was based on the well-known step recovery diode (SRD). This circuit was simulated using a table-based model of the SRD, which requires no DC bias and that can produce Gaussian pulses with a pulse width (FWHM) of 150 ps, rise time of 130 ps and peak amplitude of 5.5V. This electrical pulse was used as transmitting signal. The measurement uncertainty of the developed radar sensor was 6 mm.

In 2002, Abuasakar [6] has improved the sensitivity of the sampling unit by introducing a balanced sampling gate. For providing the balanced signal, an ultra-wideband balun was designed to operate in a frequency range from 600 MHz to 6 GHz. The circuits of the receiver were realized in microstrip technique.

Both laser and microwave radar sensors developed in the HFT department are based on the time-of-flight principle ([2]-[4], [7]-[8]). The target distance is obtained by the time difference between the time significant points of the reference pulse and the delayed pulse, which is reflected at the target (see Figure 1.1). The error (σ_R) in the range measurement of a radar sensor is dependent on a number of factors, i.e. the rise time (t_r) of the electrical pulse, the signal to noise ratio (SNR), and the *n* number of measurement which are averaged. It is shown in [9]-[10] that the error in the range measurement is proportional to the rise time of the electrical pulse and inversely proportional to the square root of the *n* number and the SNR. The resolution is then given by

$$\sigma_{R} = \frac{c}{2\sqrt{n}} \cdot \frac{t_{r}}{SNR}, \qquad (1.1)$$

where *c* is the speed of light in free space.

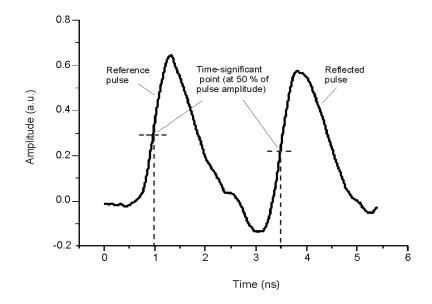


Figure 1.1 Measured time significant points in pulsed laser radar [7].

The objective of the current work can be classified into following 4-parts concerning the full development of the radar sensor envisaged in [4]. The first part focuses on improving its range accuracy. One technique for increasing range accuracy is to reduce the rise time of the radiated pulse as confirmed by equation (1.1). To achieve this aim, the pulse sharpening circuit should be designed to generate a fast electrical pulse with maximum rise time of about 50% of what was realized in [4]. Simultaneously, the rise time should not be too fast in order to comply with the rules of federal communication commission (FCC) [11]. The design and realization of the developed fast electrical pulse sharpening circuit is simulated using a table-based model of the SRD. This circuit does not need any DC power supply. The realization is in microstrip environment.

To satisfy the requirement of reduced pulse rise time for increased accuracy of radar sensor, the second part is needed to develop an UWB antenna which supports radiation and receive of those ultra-fast pulses. The new antenna should be designed to have relative bandwidth greater than 62.5% and operating frequency range covering 10.6 GHz, defined at VSWR < 2. Additionally, it is essential that the new antenna is compact and suited for portable application. To

reach this goal, the antenna structure should be reduced to at least 50% of previously designed size [4].

The third part is to reduce the size of the bi-static radar sensor for monostatic configuration. The mono-static radar sensor uses only one antenna for radiating and receiving fast electrical pulses. For signal separation a 3-port duplexer is required. The duplexer should exhibit low reflection loss, at least more than 10 dB and insertion loss less than 3 dB.

The final part of this current work is to assemble the developed system components to a functioning bi-static and mono-static radar sensor as well. System performance is confirmed by level control measurement and detection of metal objects buried in sand.

1.2 Organization of Work

In Chapter 2, the UWB radar history, concept, and antenna are reviewed. The antenna section includes the definition of antenna bandwidth and describes the dispersion of antenna. The functioning of bi-static and mono-static radar sensors is discussed.

In addition, a novel rugby-ball antenna is presented in Chapter 3. It has been derived from the monopole disc and inverted trapezoidal antennas. The rugby-ball antenna was developed using several hardware test models. The results show that the rugby-ball antenna has significantly increased radiation pattern bandwidth compared to the inverted trapezoidal antenna. Moreover, the rugby-ball shape has smaller size and is lighter than the inverted trapezoidal antenna. Finally, the rugby-ball antenna is optimized for reducing weight. The measurement results and their comparison with the simulation results are discussed.

In Chapter 4, the developed modules of the bi-static and mono-static radar sensor are presented. First, a table-based circuit model of the highly non-linear step recovery diode (SRD) element used in the Advanced Design System

4

(ADS[®]) software is introduced. After that, a circuit used for picosecond pulse sharpening comprising a multi-stage SRD circuit and a filter network is discussed. Finally, the design and realization of a 3-port broadband resistive duplexer is presented.

Chapter 5 deals with the measurement setups and measured results both for the bi-static and mono-static radar sensor. The experiments include detection and ranging of different targets and water level control measurements. In addition, test of range uncertainty measurement is presented for both radar sensors.

In Chapter 6, the results are summarized and conclusions are given for future work.

List of References

- G. Kompa, "Extended time sampling for accurate optical pulse reflection measurement in level control," *IEEE Transactions on Instrumentation & Measurement*, vol. IM-33, pp. 97-100, 1984.
- [2] G. Kompa, V. Gorfinkel, J. Sola, A Stolze, W. Vogt, and F. Volpe, "Power picosecond pulsed laser radar with micrometer ranging resolution," *Proceedings of the 26th European Microwave conference* (EuMC 96), pp. 147-152, 1996.
- [3] A. Biernat, *Erzeugung und Anwendung von ultrakurzen Laserradarimpulsen mit hoher Leistung*. Doctoral Thesis, HFT, University of Kassel, 1998.
- [4] A. Duzdar, *Design and Modeling of an UWB Antenna for a Pulsed Microwave Radar Sensor*. Doctoral Thesis, HFT, University of Kassel, July 2001.
- [5] P. Eskelinen, "Improvements of an inverted trapezoidal pulse antenna," *IEEE Transactions on Antennas and Propagation Magazine*, vol. 43, no. 3, pp. 82-85, 2001.
- [6] S. Abuasaker and G. Kompa, "A high sensitive receiver for base band pulse microwave radar sensor using hybrid technology," *IEEE Radar Conference*, Long Beach, California, USA, pp. 121-124, April 22-25, 2002.
- [7] A. Ghose, Pulsed Measurement Based Nonlinear Characterization of Avalanche Photodiode for the Time Error Correction of 3D Pulsed Laser Radar. Doctoral Thesis, HFT, University of Kassel, July 2005.
- [8] G. N. Kamucha, A Non-Invasive Registration Technique in Hip-Joint Replacement Surgery Using Laser Radar Imaging. Doctoral Thesis, HFT, University of Kassel, July 2003.
- [9] A. Wehr and U. Lohr, "Airborne laser scanning," *ISPRS Journal of Photogrammetry & Remote Sensing*, no. 54, pp. 68-82, 1999.
- [10] P. Webb and C. Wykes, "Analysis of fast accurate low ambiguity beam for non $\lambda/2$ ultrasonic arrays," *Ultrasonics*, no. 39, pp. 68-78, 2001.

[11] Federal Communications Commission, *Notice of inquiry in the matter of: Revision of part 15 of the commission's rules regarding ultra-wideband transmission systems.* Document # 02-48, April 2002.

Chapter 2

UWB Radar Evolution and System Aspects

2.1 Historical Review

The subject of this work is focused on microwave radar system using ultrawideband (UWB) signal waveforms. Application of UWB signal waveform implies advantages such as higher spatial resolution and easier target information recovery from reflected signals. Before starting the design of a radar sensor, the understanding of UWB concepts is essential. Therefore, a brief historical review of UWB technology, radar specification, antenna configuration, and radar sensor concepts is described in this chapter.

In this section, we first give a historical survey of UWB technology [1]. The term "ultra-wideband" has several similar meaning such as impulse, carrier-free, baseband, and large relative-bandwidth radio or radar signals. Contributions to the development of UWB RF signals and their application started in the late 1960's with the pioneering contributions of Harmuth, Ross and Robbins, and Etten. The Harmuth books and published papers [2]-[6], placed in the public domain, are based on the basic design for UWB transmitters and receivers. At approximately the same time, the Ross and Robbins patents [7]-[8] pioneered the use of UWB signals in several application areas, including communications, radar, and using coding schemes. In 1973, the work described in [9] is a landmark patent in UWB communications. Both Harmuth and Ross and Robbins applied the 50-year-old idea of matched filtering to UWB systems. Etten's [10] empirical test of UWB radar systems resulted in developing system design and

antenna concepts. In 1974, Morey [11] designed an UWB radar system for ground penetration, which afterwards had become a commercial success at Geophysical Survey Systems, Inc. (GSSI). Other subsurface UWB radar designs followed [12].

The development of sample and hold receivers at Tektronix Inc. (mainly for oscilloscopes), which have supported the evolution of the UWB field, were available commercially in the late 1960s [13]. Other advances in the development of the sampling oscilloscope were made at the Hewlett Packard Company. These activities were important to UWB system designs. Starting in 1964, both Hewlett Packard and Tektronix produced first time domain instruments for pulse diagnostics. In the 1960s both Lawrence Livermore National Laboratory and Los Alamos National Laboratory performed original research on pulse transmitters, receivers and antennas. Cook and Bernfeld [14] presented the developments in pulse compression, matched filtering and correlation techniques that began in 1952 at the Sperry Gyroscope Company.

In the 1970s Lawrence Livermore National Laboratory expanded its laserbased diagnostics research into pulse diagnostics. Thus, by the early 1970s basic designs of UWB signal systems were done, however, no strong efforts could be observed in perfecting such systems. In fact, by 1975 an UWB communication link or UWB radar could be constructed from components purchased from Tektronix. After the 1970s, the only innovation in the UWB field could come from improvements in particular of the electrical properties of subsystems, but not in the overall system concept itself, not even in the overall subsystem's concepts. The basic known components were pulse train generators, pulse train modulators, switching pulse train generators, detection receivers and wideband antennas. Moreover, particular properties of the subcomponents and methods were also known such as avalanche transistor switches, light responsive switches, integration and averaging matched filters, synchronous detectors and antennas driven by step pulse.

9

In 1978, Bennett and Ross [15] presented the known pulse generation methods. Since then, there have been many meetings at various conferences, at Society for Photo-optical Instrumentation Engineers (SPIE) meetings, meetings held by Los Alamos National Laboratory, and other national meetings, where many approaches to pulse generation techniques have been discussed till date.

In 1988, Barret was able to organize an UWB workshop for the United States Department of Defense's Director of Defense Research and Engineering attended by over 100 participants [16]. Then, there was already substantial progress in UWB in the Russian Federation and China, which paralleled the progress in United State. There were also active academic programs (e.g., at Lawrence Livermore National Laboratory, Los Alamos National Laboratory, and University of Michigan), which focused on the physics of short pulse transmissions that differed from the physics of continuous or long pulse signals.

With the conference held at W. J. Schafer Associates [16] and one at Los Alamos National Laboratory in 1991 [17], there have been many meetings held on impulse radar or radio - e.g., at the SPIE [18] and at the Polytechnic University [19]-[22], as well as numerous books on the subject e.g.: [23]-[25].

In 1994, McEwan, then at Lawrence Livermore National Laboratory, invented the micropower impulse radar (MIR) which provided for the first time an UWB operating at ultra-low power, besides being compact and inexpensive [26]. This was the first UWB radar that required only microwatts of power to operate.

In summary, the pioneering work of Harmuth, Ross, Robbins, Etten, and Morey defined UWB systems and did so in a very practical manner. There was never a time wherein a particular subcomponent invention was required for UWB systems to become possible, except, perhaps, the sample and hold oscilloscope [1]. In the commercial arena, UWB systems have been used and commercialized beginning in the early 1970s.

10

2.2 UWB Specification [24]

Terms such as narrowband and wideband can have several meanings depending on the subject, i.e., communications, radar, etc. This section will give details of what UWB is accepted to be. In the assessment, the Defense Advanced Research Project Agency (DARPA) [27] has proposed that "definitions need liberal interpretation and that mathematical definitions are difficult to achieve and not often useful in a practical sense". Therefore, the following definitions are given.

Bandwidth (*BW*) is simply the difference between the upper (f_H) and lower (f_L) operating frequency

$$BW = f_H - f_L \tag{2.1}$$

The bandwidth of a system is often described relative to the centre frequency (f_c) . Often, the centre frequency is defined as the arithmetic average of the upper and lower operation frequencies

$$f_{C} = \frac{1}{2} (f_{H} + f_{L})$$
(2.2)

An alternate definition of the centre frequency involves the geometric average

$$f_c = \sqrt{f_L f_H} \tag{2.3}$$

The fractional bandwidth (bw) of a system is the ratio of the bandwidth to the centre frequency

$$bw = \frac{BW}{f_c} \tag{2.4}$$

Using the arithmetic average definition of centre frequency, the fractional bandwidth is given by

$$bw = \frac{2(f_H - f_L)}{(f_H + f_L)}$$
(2.5)

Therefore, UWB refers to electromagnetic signal waveforms that have instantaneous *bw* greater than 0.25 with respect to the centre frequency [27]. There are two other radar classes identified by signal fractional bandwidth: narrowband, where the fractional bandwidth is less than 1%, and wideband, with a fractional bandwidth from 1% to 25%. Most narrowband systems carry information, also called the baseband signal, as a modulation of a much higher carrier frequency signal. The important difference between both of them is that the UWB waveform combines the carrier and baseband signal while narrowband waveform comprises only information. Baseband or impulse radar is other names for UWB radar and radio signals [28]. The UWB signal generally occurs as either a short duration impulse signal or as a nonsinusoidal (e.g. square, triangular, chirped) waveform.

2.3 UWB Antennas

In this work, the development of a microwave radar sensor is considered mainly for low-power and short-range applications due to the restriction of pulse-forming electronics, which is based on Gaussian-shaped pulse. In order to transmit and receive very short-time duration pulses of electromagnetic energy, an UWB antenna is needed. Therefore, in the following section some fundamental characteristic of UWB antennas are discussed.

2.3.1 Definition of UWB Antenna

UWB antennas exhibit very large bandwidth compared to antennas in general. There are two criteria available for identifying when an antenna may be considered UWB. A definition given by DARPA [29] says that an UWB antenna has a fractional bandwidth (*bw*) greater than 0.25. An alternate definition, fundamental by the United States federal communications

commission (FCC), places the limit bw at 0.2. Using the fraction bandwidth in equation (2.5) we can write

$$bw = \frac{2(f_H - f_L)}{(f_H + f_L)} \ge \begin{cases} 0.25 & DARPA\\ 0.20 & FCC \end{cases}$$
 (2.6)

Additionally, the FCC provides an alternate definition whereby an UWB antenna is any antenna with a bandwidth greater than 500 MHz.

2.3.2 Determining Antenna Bandwidth

The definition of the antenna bandwidth may comprise impedance pattern, gain, and radiated bandwidths [30]. However, the definition should include all properties which are important to a particular application. Thus, the IEEE (Institute of Electrical and Electronics Engineers) standard [31] defines the bandwidth of an antenna as "the range of frequencies within which the performance of the antenna, with respect to some characteristics, conforms to a specific standard." In this work, the impedance bandwidth is used and defined for a voltage standing wave ratio VSWR < 2. Using the relation $|\Gamma| = \frac{VSWR - 1}{VSWR + 1}$, the respective reflection coefficient becomes $|\Gamma| = 0.33$ corresponding to -10 dB. Then, the relative antenna bandwidth in percent (B_p) can be denoted as the difference of the upper and lower frequencies of operation (f_u , f_l) related to the centre frequency (f_c)

$$B_{p} = \frac{f_{u} - f_{l}}{f_{c}} \times 100\%$$
 (2.7)

where $f_c = \frac{f_l + f_u}{2}$. Antenna bandwidth may also be defined as a ratio (*B_r*) as [32]:

$$B_r = \frac{f_u}{f_l} \tag{2.8}$$

Also in this case f_u and f_l denote the frequency interval in which the impedance requirement is fulfilled. The upper operating frequency of many UWB antennas is limited only by the frequency dependence of the feed mechanism.

2.3.3 Dispersion

The properties of an UWB antenna depend strongly on the frequency. Therefore, the transmitted pulse waveform is filtered by the antenna structure. For free space propagation, impulse response depends only on the antenna's filtering behaviour and the distance [33]. The choice of an antenna concept needs consideration of size and cost as well as efficiency and low dispersion. Dispersion manifests itself as an extended pulse waveform. Alternatively, dispersion is a variation in pulse waveform as a function of phase angle of signal. The effective origin of signals from an antenna is referred to as the "phase centre [30]." In antenna design theory, the phase centre is the point from which the electromagnetic radiation spreads spherically outward, with the phase of the signal being equal at any point on the sphere. Apparent phase centre is used to describe the phase centre in a limited section of the radiation pattern (see Figure 2.1). If the phase centre moves as a function of frequency, the resulting radiated waveform will be dispersive [34].

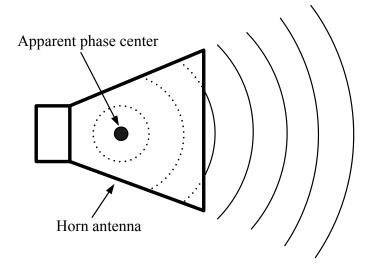


Figure 2.1 The apparent phase centre of horn antenna.

However, the dispersion of the pulse waveform of an antenna can be possibly compensated [35]. In a point-to-point link, the dispersion remains the same. Thus, it is possible to construct a filter to undo the dispersive effects of an antenna. Unfortunately, dispersion will tend to be different at different phase angles of signal. That is making the compensation procedures more complication. An overview on typical UWB antenna dispersion is given in Table 2.1.

Type of antenna	Name of antenna	Dispersion		
	- TEM horn [36]	No dispersion		
Impulse radiating	- Vivaldi or exponential tapered slot [37]	Very low		
antennas	- Monocone [38]	Extremely low		
	- Planar monopole [39]	Very low		
Frequency independent	- Spiral and sinuous [34]	High		
antennas	- Logarithmic periodic [37]	High		
Broadband antennas	- Elliptical dipole [40]	From 2.5 - 7 GHz		
Divadualid alitellias	- Multimode slot [38]	Medium		

 Table 2.1 Overview dispersion of typical UWB antenna.

Additionally, to study dispersion of several UWB antennas have been carried out [33]. In Andrews's work [33], he built up a short UWB radio link to demonstrate the principles of UWB transmission, reception and propagation. For his experiments he used monopole, conical, TEM horn and D-dot probe antennas. Both differentiation and integration effects in the time domain have been showed by these various antennas. The UWB test signal was a step pulse with amplitude of 4V, 10%-90% rise time of 9 ps as shown in Figure 2.2.

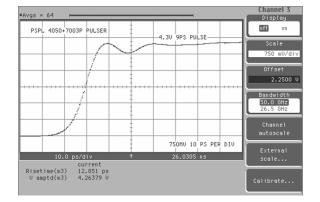
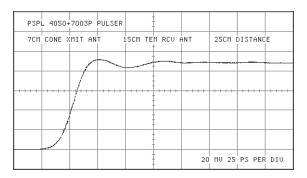


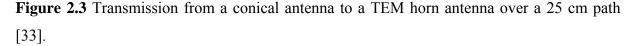
Figure 2.2 Step pulse with amplitude of 4V, rise time of 9 ps used for antenna testing [33].

In order to study the dispersion behaviour of various antenna structures, experiments were conducted using the HP-54752B oscilloscope as receiver to measure time-domain waveform. The used conical antenna was 7 cm high and had an impedance of 132 Ω . The TEM horn antenna was 15 cm long and had an impedance of 106 Ω . The separation between the antennas was 25 cm. The input step pulse was fed directly to the conical antenna. The output of the TEM horn antenna was connected to the oscilloscope. Figure 2.3 shows the output of the TEM horn antenna, which has rise time of 23 ps. The waveform pattern obtained from the TEM horn antenna (see Figure 2.3) is in good agreement with the input step pulse waveform (see Figure 2.2). The D-dot antenna is basically an extremely short monopole antenna. The equivalent antenna (V_{ant}) given by [33]

$$V_{ant} = h_{eff} \cdot E_{inc} \tag{2.9}$$

where h_{eff} is the effective height of antenna and E_{inc} is the incident electrical field.





For a very short monopole, the antenna capacitance is very small and the capacitor thus acts like a differentiator to transient electromagnetic fields. Therefore, the output from a D-dot probe antenna is the first derivative of the incident electrical field. Figure 2.4 shows the radiated field from a D-dot

transmit antenna. It was received by a TEM horn antenna. This shows that the transmitting transient response of D-dot antenna is the second derivative of the driving generator voltage.

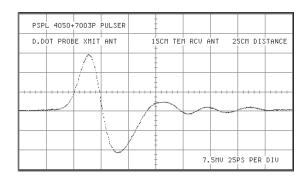


Figure 2.4 Transmission from a D-dot antenna to a TEM horn antenna over a 25 cm path [33].

PSPL 4050+7003P PU	_SER		
D.DOT PROBE XMIT A	NT D.DOT PROBE	RCUANT 25	5CM DISTANCE
		and the second second second	hiter and the state of the second
├── <u>├</u> ── \		2 MU 25P	

Figure 2.5 Transmission between a pair of identical D-dot antenna over a 25 cm path [33].

Figure 2.5 is the received output signal using a pair of D-dot antennas for both transmit and receive. The signal is the third derivative of the input step pulse.

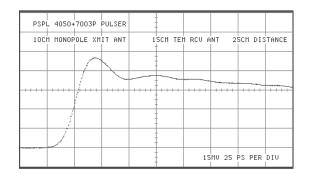


Figure 2.6 Transmission from a monopole antenna to a TEM horn antenna [33].

PSP	L 4050+	7003P	PULSER		-				
10CM	MONOPOL	E XMIT	ANT	10CM	MONOPOLI	E RCV A	NT 2	SCM DIS	TANCE
					-				
					-				
			and the second se		-				
					Ę				
					-	10	MU 25P	S PER D	DIU

Figure 2.7 Transmission between a pair of identical 10 cm monopole antennas [33].

A final experiment has been conducted on a monopole antenna to observe its dispersion behaviour. The monopole antenna is sometimes used as a simpler version of a conical antenna for transmitting UWB signals, which are similar in wave shape to the driving point voltage. However, its radiated fields are not as uniform as those for the conical antenna. Its driving point impedance is not constant, but rises with time. This leads to distortion of the radiated electromagnetic fields. Figure 2.6 shows the radiated step pulse of a 10 cm monopole. It resembles the step pulse from conical antenna (Figure 2.3). However, the top line is longer flat, but sags with increasing time. This is due to the non-uniform impedance of this antenna. Figure 2.7 shows the transmission response of a pair of 10 cm monopoles for both transmit and receive. This is essentially the integral of the generator's step pulse.

2.4 The UWB Radar Sensor Concepts

UWB microwave radar sensors have received significant interest recently [41]-[42]. Many of these systems operate in the time-domain and have an effective frequency spectrum of about 0.1 GHz to 10 GHz [42]. Very short pulses are used in UWB radar sensor and communications. Main advantages are [41]:

- Ability to mitigate multi-path as a result of its short duration.
- Ability to operate indoor as well as in cities and obstructed areas.
- Facilitate high-precision ranging and radar.

 Low-power wide-bandwidth characteristic enables low probability of interception by undesired receivers.

However, almost all UWB radar systems designed rely on sending a pulsed sinusoidal signal at specific frequency to obtain range information [43]-[49]. These systems are constructed for long-range application and have very high costs. Target range calculation of microwave radar sensor is a complicated process due to the system parameters, which are needed in the range equation [47]. An easier range computation method, which is known as the time-of-flight principle, can replace this complicated equation. This method is used extensively by pulsed laser radar sensors to calculate the range [50]-[54]. The range (R) can be calculated from the elapsed time between the transmitter pulse and the return target according to the equation:

$$t = \frac{2R}{c} \tag{2.10}$$

where *t* is the elapsed time between transmitted and reflected pulse from the target and *c* is the light velocity in the free space. The maximum measurable range R_{max} in such system is a function of the pulse repetition period *T* and is given by [55]

$$R_{\max} = \frac{cT}{2} \tag{2.11}$$

Within this range, the elapsed time between transmitted and reflected pulse from the target will include multiple transmitted pulses from unexpected targets that should be sometimes eliminated. But, beyond this range, the calculation of the distance from the target is ambiguous.

2.4.1 Bi-Static Radar Sensor

In a bi-static configuration (see Figure 2.8), the transmitting antenna (Tx) and the receiving antenna (Rx) are spatially separated while the target distance

(d) is calculated between the target point and the vertical centre to the antenna plane. In Figure 2.8, the bi-static system generates ultra-short pulses, which are fed to the transmitting antenna. Received pulses at the antenna (Tx) from the bi-static system are transmitted in two paths, i) path t_1 : target pulse and ii) path t_3 : reference pulse. Then, the target pulses are reflected from the target to the receiving antenna following path t_2 : reflected pulse. After that, reference and reflected pulses from the receiving antenna are then fed to the bi-static system. Finally, the target distance (d) is calculated using the time difference between reference and reflected pulses.

Several years ago, the basic idea of a bi-static radar sensor was developed in the HFT derivations from the pulsed laser radar sensor technology [55]. The respective microwave system comprises a clock synchronization module, a pulse generation module, a sampler module, an interface module and the UWB antennas. Figure 2.9 shows a detailed block diagram of the previous radar sensor version [55]. The description of the system components is given in the following sub-section.

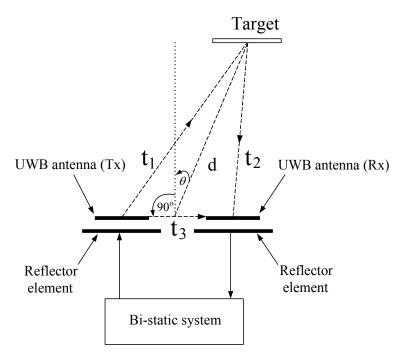


Figure 2.8 Bi-static configuration.

A. Clock Synchronization and Pulse Generation Module

The clock synchronization module controls the timing signals in the radar sensor and the sampling module to realize the extended time method. It comprises the quartz oscillator, frequency divider-I and frequency divider-II, a fast and slow RC-generator, and a comparator. It is based on a stabilized 20 MHz crystal oscillator to provide the basic reference clock signal. The frequency is converted by the frequency dividers into 100 kHz and 20 Hz, respectively. The frequency divider-I circuit, which is built using CMOS IC's, has a selectable output that ranges from 26 kHz to 6.66 MHz, which sets the pulse repetition frequency (PRF) of the radiated pulses and sampling rate of the sampler.

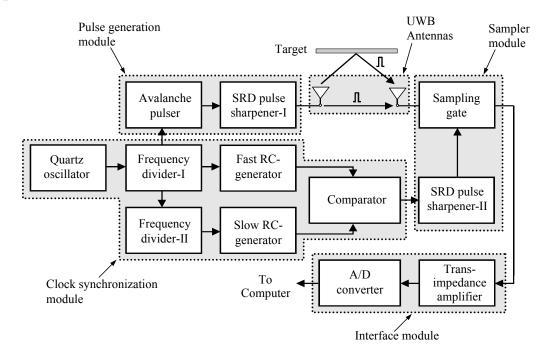


Figure 2.9 Block diagram of the developed bi-static microwave radar sensor [55].

The output pulse of the signal divider-I has logic 1 with a pulse width (FWHM) of 500 ns (see Figure 2.10 and Figure 2.11). This signal is fed into a high-pass filter when it is differentiated yielding a positive and a negative pulse. Then, the differentiated signal is fed into a 74HCT245 bi-directional transceiver digital logic IC. This logic device has TTL-compatible inputs having several

buffer amplifiers connected in parallel [56]. Due to the use of several of the buffer amplifiers the logic device delivers an output with short step pulse as shown in Figure 2.11. The block diagram and the output waveform of the signal obtained from a SPICE simulation at different nodes inside the frequency divider-I are shown in Figure 2.10 and Figure 2.11, respectively.

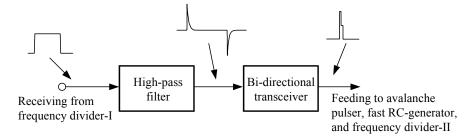


Figure 2.10 The block diagram of different nodes inside the frequency divider-I.

The sharpened step pulse from the bi-directional transceiver has a pulse width of 50 ns (FWHM), amplitude of 4.85V, and a 10%-90% rise time (t_r) of 1 ns (see Figure 2.11). This pulse triggers a pulse generation module (see Figure 2.9), which consists of an avalanche pulser and SRD pulse sharpener-I. In this module, the avalanche pulser circuit can be used as a first stage at the input of pulse sharpening circuit and is based on avalanche phenomena occurring in bipolar transistor operating in the breakdown region [57], [58].

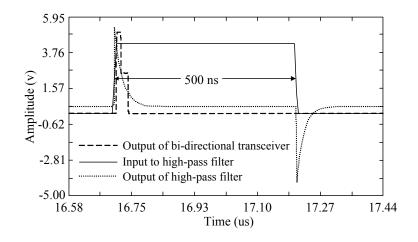


Figure 2.11 Output voltage waveforms of frequency divider-I [55].

To generate current pulses, the transistor is biased at an operating point that is near to the first breakdown point [57]. When the base of transistor is triggered (by the pulse from bi-directional transceiver), a breakdown in the transistor occurs. A collector current of about 1A to 3A and voltage drop over the collector-emitter path of about 10V to 20V [59] appear. The voltage pulse is fed into SRD pulse sharpener-I. The SRD pulse sharpener-I enhances the rise time and shapes the pulse. An improved 10%-90% rise time of 120 ps and a pulse width (FWHM) of 150 ps are achieved. The output of the pulse generation module excites directly the UWB transmitter antenna.

Another output from the frequency divider-I feeds both a fast RC-generator and a frequency divider-II. The fast RC-generator has a charging time constant τ_f of 100 ns and the output of the frequency divider-II, whose frequency is about 20 Hz feeds a slow RC-generator with the charging time constant τ_s of 375 ms. Output voltage from each RC-generator is used to trigger a comparator. The output of comparator is shifted in time at each clock signal input of the fast RC charger by a fixed time interval of around 6 ps. The output pulse from the comparator is shifted by 6 ps according to the equation [51]:

$$\Delta t \approx T_f \frac{\tau_f}{\tau_s} \quad \text{with} \quad \tau_s \gg \tau_f \,,$$
 (2.12)

where τ_f and τ_s are the charging time constants of the fast and slow RC circuit, respectively, which are monotonously increasing voltage functions. The Δt represents the system time resolution and T_f is the time period of the fast RCgenerator.

B. Sampling and Interface Module

The sampling module consists of a SRD pulse sharpener-II and a sampling gate (see Figure 2.9). After comparison of the output pulses from fast and slow

RC-generator, the comparator output pulse, which is used for sampling purpose, is delivered to a SRD pulse sharpener-II.

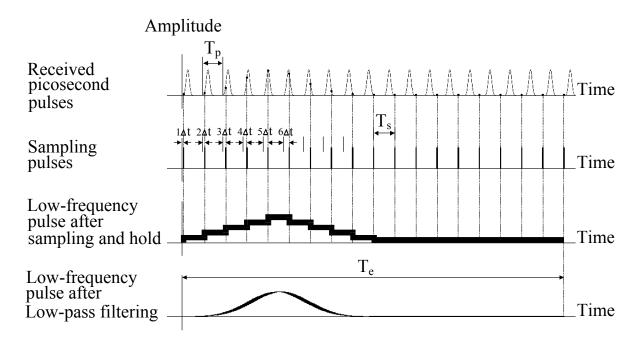


Figure 2.12 Graphical representation of the extended-time sampling technique [59].

The reshaped pulses from the SRD pulse sharpener-II are fed into the sampling gate through 50 Ω microstrip transmission lines. The second input pulses of the sampling gate are obtained from the receiver antenna. Extended-time sampling is used to downconvert the received picosecond pulses into the millisecond range by using the sampling pulse from the SRD pulse sharpener-II. A graphical representation of this sampling process can be seen in Figure 2.12, where T_p represents the received pulse repetition period. T_s and T_e are the sampling pulse repetition period and the time-extended period of the sampled pulses, respectively. In the final process, the sampled low-frequency (LF) output pulses from the sampling module are fed directly into the interface module, which consists of a transimpedance amplifier and analog-to-digital (A/D) converter. In this module, the transimpedance amplifier acts as an active low-pass filter with an upper 3 dB cut-off frequency of about 12 kHz. Further 8-bit analog-to-digital format. This converter is controlled via an I/O card on a

computer. In-house built software was used to read the data from the RAM and to save it on a disk. All data are used to display the target returned pulse and to extract target range information from the received signals.

2.4.2 Mono-Static Radar Sensor

Although in [55] the bi-static radar sensor demonstrated good performance for near-range target detection, the largeness of the system is a disadvantage for general use. To satisfy the requirement of reduced dimension and preserved lowcost aspect of the radar sensor, it is attractive to replace the bi-static configuration by a mono-static system. This mono-static radar sensor uses only one antenna for signal transmitting and receiving as shown in Figure 2.13.

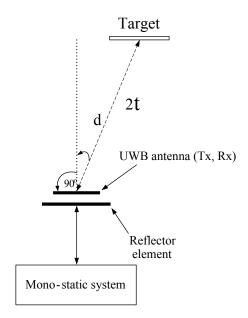


Figure 2.13 Mono-static configuration.

A block diagram of the mono-static radar sensor is shown in Figure 2.14. A pulse generator from Picosecond Pulse Labs (model 10000A) delivers rectangular pulses with amplitude of 8V, a pulse width (FWHM) of 20 ns and a 10%-90% rise time of 400 ps as shows in Figure 2.15. This pulse is formed to a Gaussian picosecond pulse with a pulse width (FWHM) of 75 ps and a 10%-90% rise time of 50 ps by using a SRD pulse sharpener, which consists of a 2-stage SRD circuit. Then, the picosecond pulse is split into two pulses by the

duplexer, which is designed in microstrip technique. A first output pulse is fed to the antenna, which acts both as radiating and receiving element covering an operation frequency range of 0.65 GHz to 20 GHz. A second output pulse (used as reference pulse) is fed to the sampling unit, which is a HP 54120B digital sampling scope with 50 GHz sampling head. Reflection pulses from the target are received via the antenna and fed to a sampling unit through the duplexer.

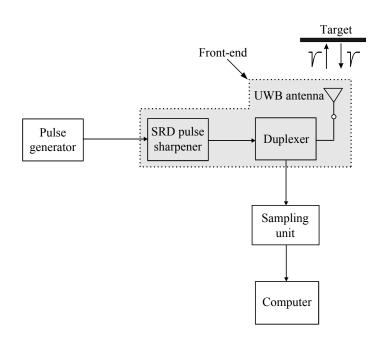


Figure 2.14 Block diagram of the mono-static microwave radar sensor.

picosecond After downconverting the pulses. the reference and measurement pulses are transferred from the digital sampling scope to a computer through an interface port using a General Purpose Interface Bus (GPIB) cable. Through digital signal processing, using in-house built software, noise reduction of the signal was achieved. Finally, the time difference between reference and reflected pulses is used to calculate the target information. The data of the target is displayed and saved in the computer. A description of the mono-static radar sensor including design details of the SRD pulse sharpener, duplexer, and antenna as well as measurement results are described in Chapter 3 and 4

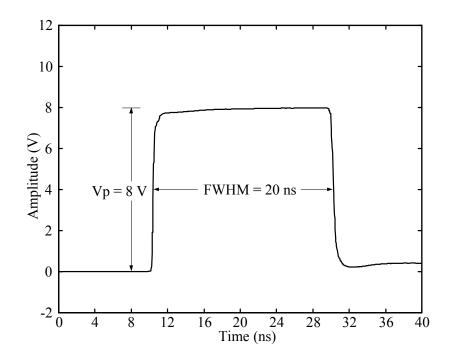


Figure 2.15 Measured output pulse of Picosecond Pulse Labs generator.

List of References

- [1] T. W. Barrett, "History of ultra-wideband (UWB) radar & communications: pioneers and innovators," *Progress in Electromagnetics Symposium 2000 (PIER200)*, pp. 1-29, 2000.
- [2] H. F. Harmuth, *Transmission of Information by Orthogonal Functions*. 1st *Edition*, New York: Springer, 1969.
- [3] H. F. Harmuth, "Range-doppler resolution of electromagnetic walsh waves in radar," *IEEE Transaction on Electromagnetic Compatibility*, EMC-17, pp. 106-111, 1975.
- [4] H. F. Harmuth, "Synthetic aperture radar based on nonsinusoidal functions pulse compression, contrast, resolution, and doppler shift," *IEEE Transaction on Electromagnetic Compatibility*, EMC-21, pp. 40-49, 1979.
- [5] H. F. Harmuth, *Nonsinusoidal Waves for Radar and Radio Communication*. New York: Academic, 1981.
- [6] H. F. Harmuth, *Antennas and Waveguides for Nonsinusoidal Waves*. New York: Academic, 1984.
- [7] G. F. Ross and K. W. Robbins, *Baseband Radiation and Reception System*. U.S. Patent 3,739,392, June 1973.
- [8] G. F. Ross and K. W. Robbins, *Narrow Range-Gate Baseband Receiver*. U.S. Patent 4,695,752, September 1987.
- [9] G. F. Ross, Transmission and Reception System for Generating and Receiving Baseband Duration Pulse Signals for Short Baseband Pulse Communication System. U.S. Patent 3,728,632, July 1973.
- [10] P. V. Etten, "The present technology of impulse radars," *Proceedings of the International Radar Conferences*, pp. 353-539, October 1977.
- [11] R. N. Morey, *Geophysical Survey System Employing Electromagnetic Impulses*. US Patent 3,806,795, April 1974.
- [12] D. L. Moffatt and R. J. Puskar, "A subsurface electromagnetic pulse radar," *Geophysics*, vol. 41, pp. 506-518, 1976.

- [13] S. W. Millikan Way, *Instruction Manual: Type S-2 Sampling Head*. Tektronix, Inc., P.O. Box 500, Beaverton, Oregon 97005, 1968.
- [14] C. E. Cook and M. Bernfeld, *Radar Signals: An Introduction to Theory and Application*. New York: Academic Press, 1967.
- [15] C. L. Bennett and G. F. Ross, "Time-domain electromagnetics and its application," *Proceedings of the IEEE*, vol. 66, pp. 299-318, 1978.
- [16] T. W. Barrett, "Impulse (time-domain) radar technology assessment colloquium," W. J., Arlington, VA, 16th-17th March, 1988.
- [17] B. Noel, Ultra-Wideband Radar: Proceedings of the First Los Alamos Symposium. Boca Raton, FL: CRC Press, 1991.
- [18] I. J. Lahaie, "Ultra-wideband radar," SPIE Proceedings Series, vol. 1631, 22nd - 23rd January, 1992.
- [19] H. L. Bertoni, L. Carin, and L. B. Felsen, *Ultra-Wideband Short-Pulse Electromagnetics*. New York: Plenum Press, 1993.
- [20] L. Carin, and L. B. Felsen, *Ultra-Wideband Short-Pulse Electromagnetics*. 2nd Edition, New York: Plenum Press, 1995.
- [21] C. E. Baum, L. Carin, and A. P. Stone, *Ultra-Wideband, Short-Pulse Electromagnetics*. 3rd Edition, New York: Plenum Press, 1997.
- [22] E. Heyman and B. Mandelbaum, *Ultra-Wideband Short-Pulse Electromagnetics*. 4th Edition, New York: Plenum Press, 1999.
- [23] H. F. Harmuth, *Radiation of Nonsinusoidal Electromagnetic Waves*. New York: Academic, 1990.
- [24] J. D. Taylor, *Introduction to Ultra-Wideband Radar Systems*. Boca Raton, FL: CRC Press, 1995.
- [25] L. Yu. Astanin and A. Kostylev, "Ultra-wideband radar measurements," *Analysis and Processing Radar Sonar, Navigation & Avionics Series, IEE*, London, UK, 1997.
- [26] T. E. McEwan, *Ultra-Wideband Radar Motion Sensor*. US Patent 5,361,070, November 1994.

- [27] OSD/DARPA, Assessment of Ultra Wideband (UWB) Technology. Ultra-Wideband Review Panel, Arlington, VA: DARPA, 1990.
- [28] R. Vicker, "Ultrahigh resolution radar," SPIE Proceedings, vol. 1875, SPIE, Bellingham, WA, 1993.
- [29] C. Foster, "Assessment of ultra-wideband (UWB) technology," *IEEE* Aerospace and Electronic System Magazine, pp. 45-49, November 1990.
- [30] H. Schantz, *The Art and Science of Ultra-Wideband Antennas*. Norwood: Artech House Inc., 2005.
- [31] Antenna Standards Committee of the IEEE Antennas and Propagation Society, IEEE Standard Definitions of Terms for Antennas, IEEE Std 145-1993. New York: the Institute of Electrical and Electronics Engineers Inc., 1993.
- [32] W. L. Stutzman and G. A. Thiele, *Antenna Theory and Design.* 2nd *Edition*, New York: John Wiley & Sons, 1998.
- [33] J. R. Andrews, *Antenna & Propagation UWB Signal Sources*. Application Note AN-14a, Boulder/USA: Picosecond Pulse Labs, August 2003.
- [34] H. Schantz, "Dispersion and UWB antennas," *International Workshop on Ultrawideband Systems and Technologies*, Kyoto, Japan, pp. 161-165, May 2004.
- [35] T. Hertel and G. Smith, "On the dispersive properties of the conical spiral antenna and its use for pulse radiation," *IEEE Transactions on Antennas and Propagation*, vol. 51, no. 7, pp. 1426-1433, July 2003.
- [36] R. T. Lee, G. S. Smith, "A design study for the basic TEM horn antenna," *IEEE Antennas and Propagation Magazine*, vol. 46, no. 1, pp. 86-92, February 2004.
- [37] W. Soergel, C. Waldschmidt, and W. Wiesbeck, "Transient responses of a vivaldi antenna and a logarithmic periodic dipole array for ultra wideband communication," *IEEE Antennas and Propagation Society International Symposium*, vol. 3, Columbus, Ohio, pp. 592-595, June 2003.
- [38] J. Kraus, R. Marhefka, *Antennas for All Applications*, 3rd Edition, Boston: McGraw-Hill, 2002.

- [39] M. J. Ammann, Z. N. Chen, "Wideband monopole antennas for multiband wireless systems," *IEEE Antennas and Propagation Magazine*, vol. 45, no. 2, pp. 146-150, April 2003.
- [40] H. Schantz, "Bottom fed planar elliptical UWB antennas," *IEEE Conference on Ultra Wideband Systems and Technologies*, pp. 219-223, November 2003.
- [41] H. L. Bertoni, L. Carin, and L. B. Felsen, *Ultra-Wideband Short Pulse Electromagnetics*. New York: Plenum, 1994.
- [42] S. Vitebskiy, L. Carin, M. A. Ressler, and F. H. Le "Ultra-wideband, short-pulse ground-penetration radar: Simulation and measurement," *IEEE Transactions on Geoscience and Remote Sensing*, vol. 35, No. 3, pp. 762-772, May 1997.
- [43] L. Carin and L. B. Felsen, *Ultra-Wideband, Short Pulse Electromagnetics*. 2nd Edition, New York: Plenum, 1995.
- [44] S. A. Hovanessian, *Introduction to Sensor Systems*. Norwood: Artech House, 1988.
- [45] M. Skolnik, *Introduction to Radar Systems*. New York: McGraw-Hill, 1962.
- [46] D. Barton and S. Leonov, *Radar Technology Encyclopedia*. Boston: Artech House, 1997.
- [47] D. Barton, *Modern Radar System Analysis*. Norwood: Artech House, 1988.
- [48] J. Minkoff, Signals, Noise, and Active Sensors. New York: Wiley, 1992.
- [49] J. Motzer, "A pulse radar gauge for level measurement and process control," *IEEE MTT-S International Microwave Symposium Digest*, Boston, pp. 1563-1566, 2000.
- [50] G. Kompa, "Pulsed laser radar for 3D-quality assurance of hot workpieces," OPTO 96 Kongressband, Vortrag 4.2, 2nd Congress and Exhibition for Optical Sensor Technology, Measuring Techniques, Electronics, Leipzig, Germany, pp. 93-98, 25-27 September 1996.

- [51] G. Kompa, "Extended time sampling for accurate optical pulse reflection measurement in level control," *IEEE Transactions on Instrumentation & Measurement*, vol. IM-33, pp. 97-100, 1984.
- [52] G. Kompa, V. Gorfinkel, J. Sola, A. Stolze, W. Vogt, and F. Volpe, "Powerful picosecond pulsed laser radar with micrometer ranging resolution," *Proceedings of the 26th European Microwave Conference* (*EuMC 96*), pp. 147-152, 1996.
- [53] A. Biernat and G. Kompa, "A laser radar for precise 2D-3D object imaging," *Proceedings of the 27th European Microwave Conference (EuMC 97)*, pp. 601-604, 1997.
- [54] R. Ahola, A Pulsed Time-of-Flight Laser Rangefinder for Fast, Short-Range, High Resolution Applications. Doctoral Thesis, Department of Electrical Engineering, University of Oulu, Oulu, Finland, February 1987.
- [55] A. Duzdar, *Design and Modeling of an UWB Antenna for a Pulsed Microwave Radar Sensor*. Doctoral Thesis, HFT, University of Kassel, July 2001.
- [56] M. Okhrawi, Entwurf und Aufbau einer Samplingbrücke zur Zeittransformation vom ps-Bereich in den μs-Bereich. Master-Thesis, HFT, University of Kassel, September 1991.
- [57] G. Kompa, *High Frequency Sensors*. Lecture Notes, HFT, University of Kassel, April 1999.
- [58] V. Fibich, "Avalanchetransistoren in Schaltungen zur Erzeugung kurzer und hoher Stromimpulse," *Frequenz*, pp. 2-10, January 1970.
- [59] W. B. Mitchell, "Avalanche transistors give fast pulses," *Electronic Design* 6, pp. 202-209, 1968.
- [60] F. Michel, *Entwurf und Aufbau einer schnellen AD-Wandlereinheit für ein 3D-Laserradar*. Master-Thesis, HFT, University of Kassel, October 1995.

Chapter 3

UWB Rugby-Ball Antenna

3.1 Introduction

Ultra-wideband (UWB) applications have stimulated a surge of interest in antenna design by providing new challenges and opportunities for antenna designers. The main challenge in UWB antenna design is achieving wide impedance bandwidth while still maintaining high radiation efficiency [1]. Recently, the need for UWB antenna with omni-directional coverage is increasing in both military and commercial applications. Metal-plate antennas are preferred in most situations. The classic solution is to obtain an omni-directional pattern using a thin wire dipole or its counterpart monopole version with a ground plane. However, the wire dipole and monopole suffer from narrow impedance bandwidth, but it can be widened by using flat metal rather than a thin wire structure [2].

In case of UWB or ultra-fast electrical pulse application, it is important to find solution to improve the operational bandwidth, to reduce unwanted reflection and dispersion in antennas. Previously, many types of UWB antenna were presented such as bow-tie antenna [3]-[4], trapezoidal flat monopole antenna [5], circular disc antenna [6]-[7] and vivaldi antenna [8]. As shown in [9], the bow-tie antenna configuration needs additionally capacitive and resistive loading to reduce unwanted reflections from the driving point and also from the top end of the antenna. Also, Duzdar [10] presented a planar trapezoidal antenna, which is useful only for a bandwidth of 1 GHz to 5 GHz. The limited

bandwidth restricts its application to electrical pulses with a minimum rise time of 70 ps. In order to overcome the frequency-band limitation, researchers [11]-[13] presented an antipodal vivaldi antenna operating in a frequency range of 3 GHz to 20 GHz applying the -10 dB bandwidth definition. However, Guangyou [13] found a large amount of unwanted reflection from a feeding transition configuration. He proposed a technique to reduce the reflection by realizing through microstrip technique with 3-layers of metallic and 2-layers of dielectric. However, this made the design and realization complicated. In another attempt, planar inverted cone antenna (PICA) was proposed in [14] for radiating ultra wideband signals having a bandwidth of 18.5 GHz (1.5 GHz to 20 GHz). The application of this antenna is restricted to frequencies below 1.5 GHz due to the condition of the PICA antenna structure. In order to overcome the discussed limitations, this work focuses on the design and realization of a new compact low-cost UWB antenna, which covers the specified frequency-band of an UWB ground penetrating radar (GPR), UWB communication and measurement systems. According to the federal communication commission (FCC) recommendation [15], the operational frequency ranges of GPR are below 960 MHz and/or from 3.1 GHz to 10.6 GHz, while communication and measurement systems operate from 3.1 GHz to 10.6 GHz.

3.2 Design Procedure of Antenna Structure

The main focus of this chapter is the design of an UWB antenna with improved bandwidth and reduced the physical dimensions. The upper frequency limit of 10.6 GHz would result in a 10%-90% rise time of fast electrical pulse of 33 ps. As will be shown the length of new antenna in bi-static radar senor could be reduced about 50% of the length of previous antenna design [16]. Thus, the new antenna is more compact and well-suited for mono-static radar sensor application. The excitation of the new antenna is done by a 50 Ω coaxial cable via SMA connector.

The proposed special shape of the UWB antenna may be regarded as a variation of the conventional circular disc antenna [6], [17] and [18], but has been introduced to overcome the previously mentioned limitations. The new antenna geometry is obtained from an intersection of two circular elements having different radii of curvature as shown in Figure 3.1. This antenna design is termed rugby-ball antenna because of the similarity of its shape with rugby ball in rugby sport. Lower cut-off frequency of the rugby-ball antenna [19] is determined by the height h of the antenna element (Figure 3.1).

$$h = \frac{\lambda_L}{4} \tag{3.1}$$

where λ_L is the wavelength at the lowest operating frequency of the antenna and

$$h = (R_1 + R_2) - D (3.2)$$

where R_1 and R_2 are the radii of upper and lower circles, respectively, and *D* is the distance between the centers of the two circles as shown in Figure 3.1.

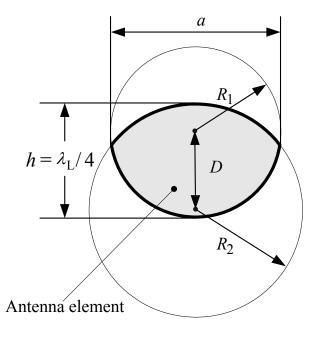


Figure 3.1 The geometry of the rugby-ball antenna.

Impedance matching of the antenna depends on the radii R_1 and R_2 . Length *a* determines the radiation property of the antenna. It is related to *D*, R_1 and R_2 by the following equation

$$a = \frac{1}{D}\sqrt{4D^2R_2^2 - (D^2 - R_1^2 + R_2^2)^2}$$
(3.3)

3.3 Simulation and Measurement Results

The prototype rugby-ball antenna was investigated using both simulations and measurements. Simulation software is based on numerical technique such as Method of Moment (MoM), Finite Element Method (FEM) and Finite Integration Technique (FIT). Each technique has its own benefits and disadvantages but none of them gives "exact" results. Validation of the design is needed and it is usually done by conducting test measurements for the prototype structures. In order to determine the input impedance, bandwidth, gain, polarization or other parameters of an antenna, usually S-parameter and radiation pattern measurements are conducted.

3.3.1 Simulations

The numerical simulation is an important stage in the modern antenna design. Although, the simulation itself is performed by a personal computer (PC), the preparation of the simulation setup requires understanding of the physical properties of the structure to be simulated. Also, the designer has to be aware of software-related issues in order to obtain proper results.

A. ADS (MoM)

The dimensions of the antenna element are obtained by using equation (3.1) - (3.3). Subsequently, simulations were first conducted using the Advanced Design System (ADS) software version 2004A and performed on a PC with single processor: Intel Pentium-III of 800 MHz and installed memory of 512 MB. The electromagnetic simulation could be performed by ADS in momentum environment. This software simulation package provides design and analysis of the antenna element by using Method of Moment (MoM). Therefore, the performance and behavior of the antenna are known before it is constructed. Electromagnetic fields are obtained by calculation of electric and magnetic surface currents on the conducting surfaces. Using this software, the simulated structure of the antenna includes only the antenna element itself but dose not take into account the reflector and ground plane. The antenna element was assumed to be constructed form aluminium, which is a lightweight metal. It is an excellent conductor of electricity. The aluminium plate has a thickness of 2 mm and a conductivity of 3.77×10^7 S/m [20]. Prior to the simulation of the antenna element, the following process should be fulfilled. First of all, the antenna element model is drawn in ADS as shown in Figure 3.2. Then, the substrate (metal) and surrounding environment (enclosure media for e.g. free space having permittivity and permeability equal to 1) are defined suitably. The conductivity and thickness of aluminium is set to 3.77×10^7 S/m and 2 mm, respectively. The current in each cell of antenna element is calculated by a precomputing step. The results of the pre-computing mesh are used to calculate Sparameter of the antenna structure. The final step, prior to running simulation, is to setup a frequency plan, which in this case was chosen to be from 45 MHz to 20 GHz.

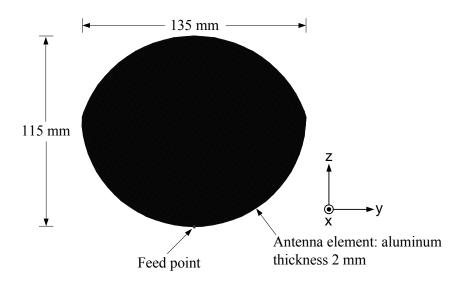


Figure 3.2 The antenna element layout implemented in ADS.

After several simulations and experimental trials, the optimal dimension of antenna element has been achieved as tabulated in Table 3.1.

Parameters	Dimension (mm)	
h	115	
R_1	70	
R_2	67	
D	22	
а	135	

 Table 3.1 Antenna element parameters.

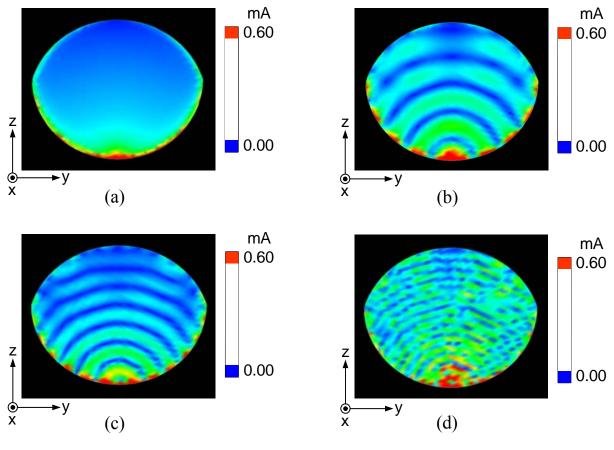


Figure 3.3 The current distributions on the surface of the rugby-ball antenna at(a) frequency of 1 GHz(b) frequency of 5 GHz(c) frequency of 10 GHz(d) frequency of 20 GHz.

Figure 3.3 shows the current distributions on the antenna surface, which are obtained from momentum visualization tool in ADS simulation. This surface current is helpful in identifying regions of the antenna geometry that needs to be optimized in order to minimize current reflections at discontinuities or bends.

The currents that flow on the antenna surface in the z-direction and y-direction were recorded from the simulation at different frequency. In several experiments, the behavior of the antenna surface current contour can be studied from three main positions namely positions 1, 2 and 3 (see Figure 3.4), which are the major points of the surface current reflection. As time progresses, the surface current propagates from the excitation point at position 1 to the top of antenna edge. The surface current reflects back from edge positions 2 and 3. Therefore, the current along these lines is attenuated. Additionally, it is important to notice the radial distribution of this surface current, which has higher amplitude at the side edges (see Figure 3.3).

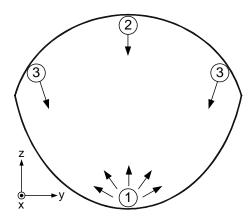
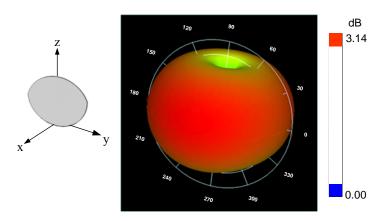
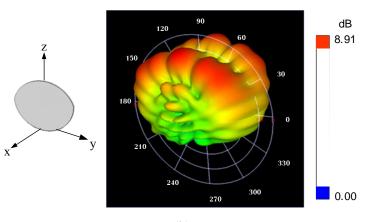


Figure 3.4 Critical positions of the surface current reflection of rugby-ball antenna.

Figure 3.5 shows a view of the 3D far-field pattern simulation results for the same rugby-ball antenna. These results depict symmetrical far-field pattern obtained in absence of reflector element. From these results, the major radiation lobes become visible in both negative and positive x-axis directions. However, in presence of a reflector element, it is possible to achieve the major radiation lobe of the antenna only in one direction. With increasing frequency, the major radiation lobes decrease while the minor radiation lobes increase due to the influence of skin effect [21] on the antenna surface current contour (see Figure 3.5).





(b)

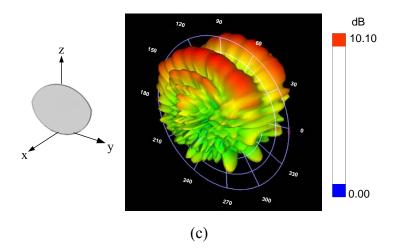


Figure 3.5 3D view of radiation pattern from ADS at(a) frequency of 1 GHz(b) frequency of 10 GHz(c) frequency of 20 GHz.

B. CST MWS (FIT)

The prototype rugby-ball antenna consists of 3-parts: antenna element, ground plane, and reflector. In ADS simulation only the properties of the antenna element could be analyzed. However, to obtain more realistic data, the antenna model must be completed by the ground plane and reflector. Ideally, the size of ground plane should be infinite which is impossible in practice to have size as large as possible. In practical vertical trapezoidal antenna [16], the size of the ground plane is chosen from the edge of the antenna element in all direction to be at least 1/15 of wavelength at the lowest operating frequency (λ_L) as, which is given by

$$L_g \ge \lambda_L / 15 \tag{3.4}$$

where L_g is the extension length of the ground plane from the edge of the antenna.

Thus, in the initial design, the size of the ground plane should be chosen on the periphery of antenna element to be at least 30.7 mm ($\lambda_L/15$ at frequency of 650 MHz.). This value is calculated from equation (3.4).

In order to control the directivity (edge and broad side) or major radiation lobe of the antenna, a reflector is needed as shown in Figure 3.6. The reflector dimension can be also calculated from equation (3.4). Thus, the reflector of the initial prototype rugby-ball antenna is designed having 30.7 mm extension from the edge of antenna element both in vertical and horizontal directions.

The next significant design procedure is the definition of the feed point of the antenna. A SMA connector through a ground plane is use for stimulation. The height between ground plane and antenna element is estimated to be 3.75 mm [16].

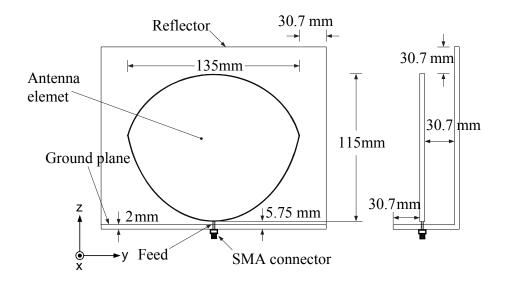


Figure 3.6 The initial prototype structure of the rugby-ball antenna.

After design of the initial prototype of the rugby-ball antenna shown in Figure 3.6 is completed, the structure model has been simulated using Computer Simulation Technology Microwave Studio (CST MWS) version 5. The CST simulations were performed on the PC with single processor Intel Pentium-III of 800 MHz and installed memory of 512 MB. This full three-dimensional (3D) electromagnetic (EM) simulation tool is based on the finite integration technique (FIT). The flexibility of FIT allows for problems to be formulated on Cartesian or general non-orthogonal grids both in the time domain as well as in the frequency domain. Initially, the frequency domain solver using the full Floquet modal expansion and periodic boundaries was used to calculate the resonant frequency of the unit cell.

The geometry of the antenna structure for simulation is drawn in a 3D form as shown in Figure 3.7. The dimensions are defined as shown in Figure 3.6. The aluminium plate was again 2 mm thin. The boundaries of the calculation domain were defined as Perfectly Matched Layer (PML) boundaries. The appropriate depth of air boundary surrounding the antenna is automatically determined by CST. An adapted waveguide port was used to model the coaxial line (50 Ω), which excites the antenna. Before starting simulation, the most important step is to set the frequency domain solver. In order to keep the number of calculated frequency points as small as possible, the automatic frequency sampling option is used to simulate the antenna model within a frequency range of 45 MHz to 20 GHz.

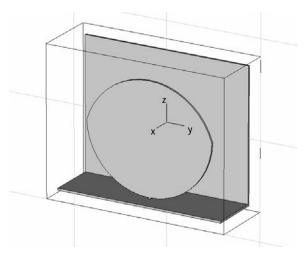


Figure 3.7 The geometry of the structure of proposed rugby-ball antenna in CST.

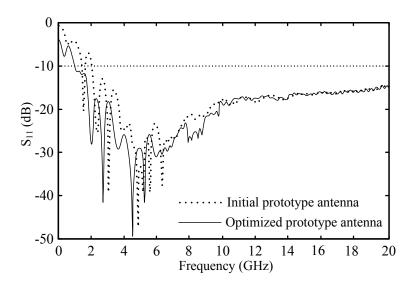


Figure 3.8 The magnitude of S_{11} of initial and optimized prototype antenna of frequency simulation from CST.

The magnitude of S_{11} of the initial prototype antenna obtained from CST simulations is shown in Figure 3.8 (dot line). The simulation result shows the resonance peaks of $|S_{11}|$ data occurring at 2.4, 3.0, 4.8, 5.4, 5.6, and 6.4 GHz with a peak dip amplitude of -26 dB, -38 dB, -47.5 dB, -39.75 dB, -39.75 dB, and -38 dB, respectively. It can be seen that the input frequency bandwidth of

initial prototype antenna, using the condition of $|S_{11}| < -10$ dB, is about 18.55 GHz (1.45-20 GHz) as shown in Figure 3.8.

After several CST simulations and experimental trials, the $|S_{11}|$ and frequency bandwidth of the initial prototype antenna could be improved by adjusting the height of the feed point of the antenna. For instance, when a height of the feed point of the antenna is reduced from the initial 3.7 mm value to 1.0 mm (see Figure 3.9), the simulation results of the optimized antenna show the dips occurring at 2, 2.6, 4.4, and 5.2 GHz with a peak dip of -28.2 dB, -40.8 dB, -49.8 dB, and -40.8 dB, respectively. Also, the frequency bandwidth has improved from 1.45-20 GHz to 0.9-20 GHz. This is due to the reduction of the height of the feed structure. Moreover, the ground plane and reflector size of the antenna have been optimised after several experimental trials. The optimized dimension decreased from 30.7 mm ($\lambda_L/15$ at frequency of 650 MHz.) to 20.0 mm ($\lambda_L/23$ at frequency of 650 MHz.). There was no significant difference observed in $|S_{11}|$ of the antenna in the whole frequency range. The final optimised structure is shown in Figures 3.9.

The 3D far-field pattern at frequencies of 1, 10 and 20 GHz are shown in Figure 3.10. It can be concluded that the radiation patterns have only one major lobe in the x-axis, which is different from ADS simulation. This is due to the inclusion of the reflector in the antenna structure model.

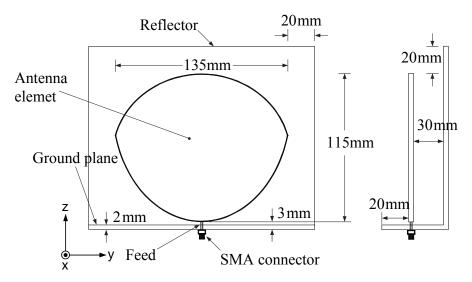
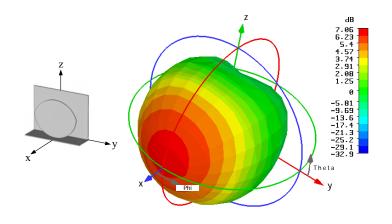
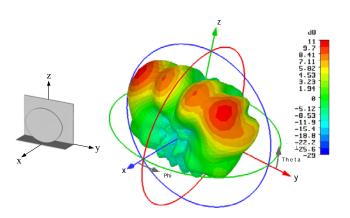


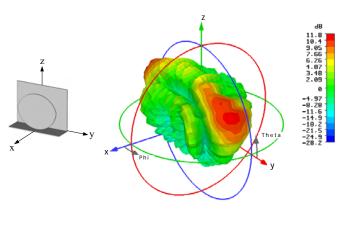
Figure 3.9 The optimized prototype structure of the rugby-ball antenna.



(a)







(c)

Figure 3.10 3D view of radiation pattern of optimized prototype antenna from CST at (a) frequency of 1 GHz (b) frequency of 10 GHz (c) frequency of 20 GHz.

C. HFSS (FEM)

The last simulation of the optimized prototype rugby-ball antenna (Figure 3.9) was then carried out using High-Frequency Structure (HFSS) version 10. The simulation is based on finite element method (FEM), which was performed on a PC with single processor: Intel Pentium-4 of 3.0 GHz and installed memory of 4.0 GB. In the HFSS, the geometry of the antenna model to be simulated is drawn in its 3D form as shown in Figure 3.11.

The antenna, ground plane, feed point and reflector element was simulated in HFSS using dimensions as depicted in Figure 3.9. All metal plate made of aluminium have thickness of 2 mm. In order to get proper results, the radiation boundary must be drawn around the structure. The radiation boundary is an approximation of free space. The air region surrounding the antenna was terminated by using a PML type of absorbing boundary. This type of boundary condition was chosen because of its excellent properties close to a radiating source. The size of the air box should be large enough to prevent distortion in the radiation and impedance characteristics. It is recommended that the distance from the radiating sources to the radiation boundary should be set to $\lambda_{\rm I}/10$ [22], where $\lambda_{\rm L}$ is the wavelength at the lowest operating frequency. In this simulation, the size of the air box is chosen to be 4.6 cm ($\lambda_{\rm L}/10$ at frequency of 650 MHz.).

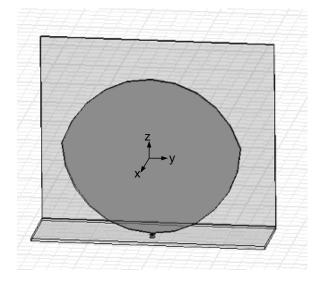
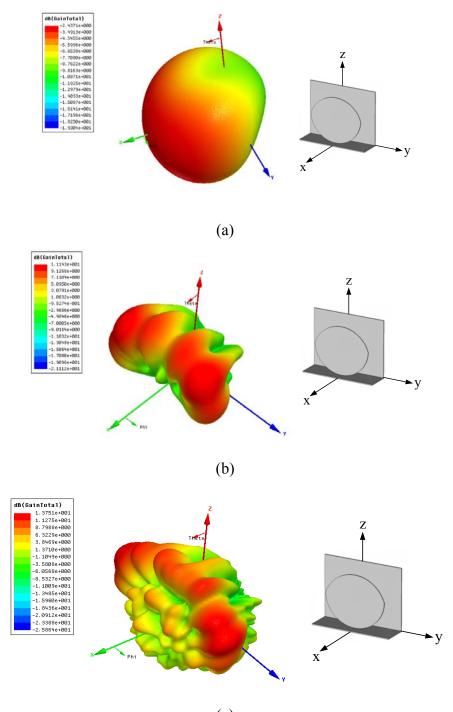


Figure 3.11 The geometry of the structures of rugby-ball antenna in HFSS.



(c)

Figure 3.12 3D view of radiation pattern of optimized prototype antenna from HFSS at (a) frequency of 1 GHz (b) frequency of 10 GHz (c) frequency of 20 GHz.

In S-parameter and input impedance sweep simulation, the frequency sweep data is generated using Adaptive Lanczos Pade Sweep (ALPS) from Ansoft, which provides broadband and reliable fast frequency sweep capabilities. The ALPS fast sweep capability coupled with a field calculator allows microwave cavity resonances and quality factors to be determined precisely.

The 3D far-field patterns at frequencies of 1, 10 and 20 GHz, which are recorded from HFSS simulation, are shown in Figure 3.12. It can be noticed that the radiation patterns have one major lobe in the x-axis. Also, with increasing frequency, the major radiation lobes decrease while the minor radiation lobes increase, which are related to CST results. The main influential factors to these phenomena are surface currents (propagate on the antenna element), feed point position and shape of antenna element.

In section 3.5, the continued simulation results, which are obtained from ADS, CST, and HFSS software, are presented and compared with measurement results.

3.3.2 Fabrication and Measurements

In this section, the procedure to build up the rugby-ball antenna and the measurement setups to study characteristics of antenna are presented. Firstly, the fabricated rugby-ball antenna will be presented. Then, calibration of the measurement system for S-parameter measurements is described. Measurements were conducted to determine the input matching of the antennas. Finally, radiation pattern measurements have been performed to test how the antennas radiate in reality.

A. Fabricated Antenna

The rugby-ball antenna, shown in Figure 3.13, was built using 2 mm thick aluminium for antenna element, reflector and ground plane. The fabricated antenna has same dimensions as shown in Figure 3.9. The input feed of antenna uses SMA connector. The inner conductor penetrates the ground plane and is connected to the antenna element. The outer conductor is connected with the flange to the image ground plane.

48

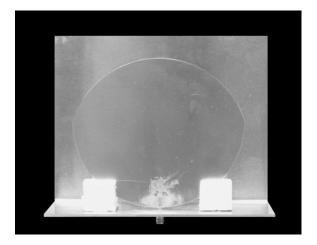


Figure 3.13 Photograph of the realized rugby-ball antenna.

The feed region of the radiating element is preferably arranged to have an impedance of 50 ohm for a well-matched coupling with common coaxial cable. The configuration of the feed region of the radiating elements can be approximated theoretically by regarding the spacing between the radiating element and the ground plane element as the slot portion of a slot radiator as described in [23]. A 50 ohm input impedance of the feed was estimated by using an iterative simulation when the feed was elevated 1 mm from the ground plane. A reflector was added to the antenna to control the major lobe and hinder any back lobe radiation. In several simulations and experimental trials, the reflector spacing from the antenna was experimentally determined to be about 30 mm so that the input impedance is minimally affected.

B. Measurement System Calibration (S-Parameter)

The instrument used for antenna measurements, is a Vector Network Analyzer (VNA) as shown in Figure 3.14. It consists of the HP8510B network analyzer, HP 8516A S-parameter test set, and HP 8360 sweep frequency synthesizer that acts as a microwave source. Before measurement, calibration of the VNA much be performed. The calibration method used in this work is a complete one-port network calibration called SOL (Short, Open, Load) [24], which requires a short circuit, open circuit, and matched load standard. The manufacturer, Hewlett Packard (now Agilent), supplies these standards for a coaxial cable calibration for the HP8510B. For the measurement of the reflection coefficient of the antenna, the SOL calibration standard was used for an APC-3.5 coaxial connector. The reference plane of the measurements was set at the connector just underneath the image ground plane of the antenna. The calibration was carried out over a frequency range from 0.1 GHz to 20.1 GHz in 0.1 GHz frequency steps which yields 201 measurement points. The antenna was connected to the VNA port, and the S₁₁ data was measured and recorded over the above specified frequency range in the laboratory.

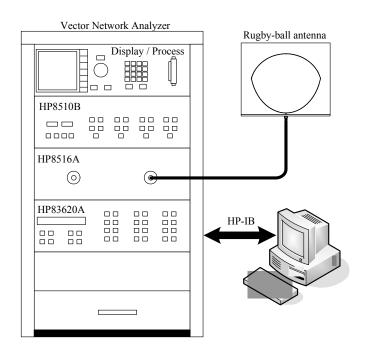


Figure 3.14 VNA Instrument setup for the characterization of the rugby-ball antenna.

C. Radiation Characteristics Measurements

In this section, the radiation measurement setups of the antenna are reported, which were used to measure the transient and gain radiation pattern. Both the radiation patterns represent radiation properties (such as radiation intensity, directivity, gain) of the test antenna as a function of space coordinates. In this case, the radiation patterns are determined for the far-field region, which is considered to exist at distances greater than $2D^2/\lambda$ [25] where *D* is the largest dimension and λ is the wavelength at operational frequency of antenna.

Transient Radiation Pattern

The measurement set up for transient radiation pattern is shown in Figure 3.15. Initially, the rectangular pulses from a picosecond pulse labs generator (model 10000A) with a 10%-90% rise time of 400 ps, pulse width (FWHM) of 20 ns and amplitude of 8V, are sent to an ultra-short pulse sharpener circuit using a step recovery diode (SRD). A full description of the SRD operation characteristics, modeling technique and circuit design will be discussed in the next chapter. Pulses with 10%-90% rise time of 100 ps, pulse width (FWHM) of 140 ps, and amplitude of 2.5V from the SRD pulse sharpening were fed directly to the antenna under test (AUT) in order to transmit ultra-short pulses. At the receiving side, output of the receiver rugby-ball antenna (test antenna) was measured using digitizing oscilloscope mainframe (HP54120B) via four-channel test set (HP54124A).

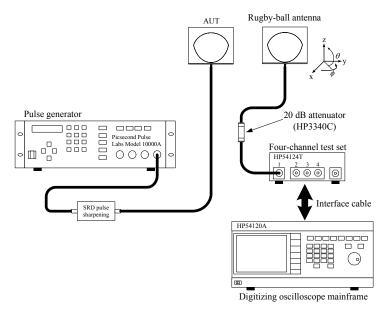


Figure 3.15 Transient radiation pattern measurement setup.

Measurements were carried out for maximum radiated power at different elevation angles θ . The maximum power radiated by the test antenna can be

obtained by measuring the amplitude of pulse, which is received form AUT as various elevation angles θ in broadside and edge-on direction according to [9]. The broadside and edge-on far-field transient radiation pattern results from measurement will be presented in next section. The setups of broadside and edge-on far-field measurement are shown in Figure 3.16 and Figure 3.17, respectively. They show electric field versus time for various elevation angle θ (0° to 90°) and fixed azimuth angle ϕ of 90° and 0° for both of broadside and edge-on orientation, respectively.

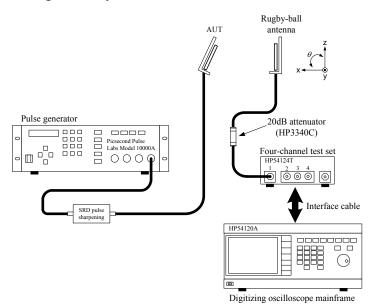


Figure 3.16 Transient radiation pattern measurement setup of broadside radiation with azimuth angle ϕ of 90° and elevation angle θ of 90°.

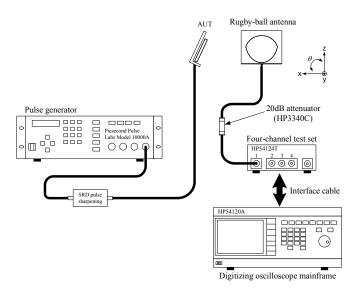


Figure 3.17 Transient radiation pattern measurement setup of edge-on radiation with azimuth angle ϕ of 0° and elevation angle θ of 90°.

Radiation (Gain) Pattern

The radiation (gain) pattern can be generally viewed in the two-dimensional or three-dimensional spatial distribution which is function of field-intensity over all angles of space. Therefore, it is necessary to specify various space angles with respect to the antenna under test and to take radiation pattern in the xyplane (elevation angle θ of 90° and azimuth angle ϕ of 0° to 360°) and xz-plane (elevation angle of 0° to 360° and azimuth angle θ of 90°). In this work, the radiation (gain) pattern measurements setup for characterization the rugby-ball antenna is shown in Figure 3.18. It consists of a signal generator (HP83650B) which delivers RF signals with an output power of 1 mW, the transmitting and receiving antenna (rugby-ball), a digital power meter (HP437B), and a power sensor (HP8487A).

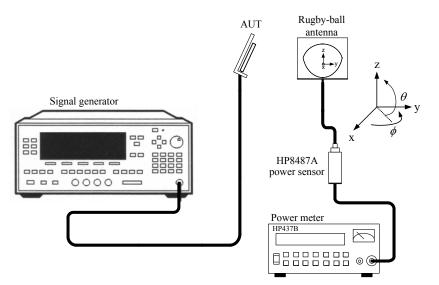


Figure 3.18 Radiation (gain) pattern measurement setup.

3.3.3 Results

In this section, the simulation and measurement results of rugby-ball antenna are presented and compared.

A. Antenna Matching

One important antenna parameter is impedance matching. A perfect impedance match maximizes the efficiency of the antenna. On the other hand,

the impedance mismatch yields undesirable reflection pulses and reverberation into the radar system. The quality of the antenna match is usually measured by voltage standing wave ratio (VSWR) [27] which is given in terms of reflection coefficient as,

$$VSWR = \frac{1+|\Gamma|}{1-|\Gamma|}$$
(3.5)

In this work, the VSWR has been calculated from simulated S-parameters using three commercial softwares: ADS[®], CST[®], and HFSS[®]. These simulation results are compared with the measurement results acquired with VNA (HP8510B). Figure 3.19 compares ADS simulated and measured results showing some discrepancy at frequencies below 1 GHz and above 13 GHz. This is expected as the 2D simulation setup in ADS® does not allow regarding the reflector and the ground plane. Specifically, it was discussed in section 3.3.1 that the tuning of the antenna element above the ground plane improves the impedance matching, and this feature is missing in the simulation setup and hence a disagreement between simulation and measurement occurs. However, considering the simulation results using HFSS and CST, there is very good agreement with measurement in the frequency range under consideration as shown in Figure 3.20 and 3.21. This indicates successful design and optimization of the antenna geometry as well as an optimal positioning of the antenna element with respect to the reflector and ground plane as has been described in section 3.3.1.

The usable bandwidth of the antenna is defined as the frequency range for which the VSWR is less than 2. From the measured diagrams we get a bandwidth of 19.35 GHz (650 MHz to 20 GHz) or 187.4% (percent bandwidth) or 30:1 (ratio bandwidth).

54

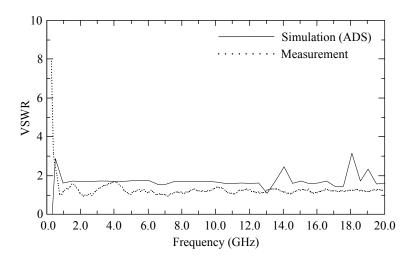


Figure 3.19 Simulated (ADS) and measured VSWR of antenna.

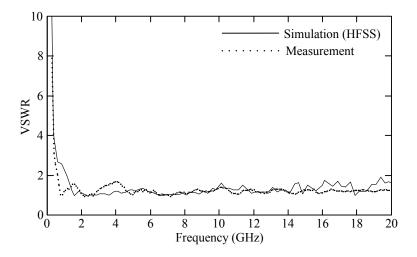


Figure 3.20 Simulated (HFSS) and measured VSWR of antenna.

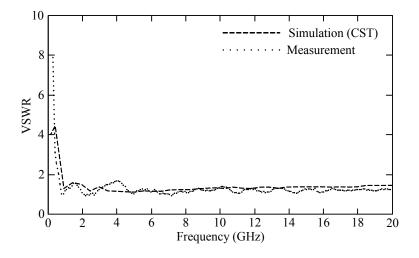


Figure 3.21 Simulated (CST) and measured VSWR of antenna.

B. Transient Radiation Patterns

The transient radiation measurements were carried out for maximum radiated power at different elevation angles θ . The broadside and edge-on far-field radiation from measurement results of rugby-ball antenna are shown in Figure 3.22 and Figure 3.24, respectively. Electric field versus time for various elevation angle θ for both the broadside (azimuth angle ϕ of 90°) and edge-on (azimuth angle ϕ of 0°) orientation are shown. These two planes of transient radiation were enough to characterize the far-field time-domain radiation of the antenna. Regarding the bi-static radar sensor, edge-on radiation is utilized to provide a reference pulse. Broadside radiation is used to detect targets and to receive their returns for both the bi-static and mono-static radar sensor.

Broadside Far-Field Radiation

Figure 3.22 shows the electric field radiated from broadside of antenna, i.e. at azimuth angle $\phi = 90^{\circ}$. The components were plotted for values of elevation angle θ ranging from 0° to 90°. From broadside radiation of antenna, it can be seen that at zero degree θ very little radiation is presented which is expected considering the geometry of the antenna. Maximum peak pulse amplitude is attained at an angle of θ equal to 60°. At broadside of 60°, the measurement pulse amplitude of 20 mV has been recorded by the sampling oscilloscope.

A 20 dB attenuator was used during the measurement with the sampling oscilloscope. Therefore, the scope display of 20 mV corresponds to the amplitude of 200 mV. The plot of the normalized peak amplitude of the pulse for co-polar and cross-polar radiation is shown in Figure 3.23. The co-polar component is at least 30 dB higher than the cross-polar radiation at all angles. There is no clear definition in the literature related to pulsed radiation as compared to the definition of the half power beamwidth (HPBW) with respect to frequency-dependant antennas, to quantify the opening angle of radiation. Nevertheless, in [16], a suitable definition has been proposed for the angle at

which the pulse amplitude drops to one-half of the maximum pulse amplitude for given a specific direction. In broadside case, shown in Figure 3.23, the opening radiation angle extends from about 30° to 90° giving a radiation aperture opening of 60° .

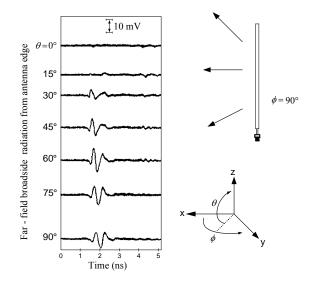


Figure 3.22 Measurement results for the broadside far-field radiation of antenna.

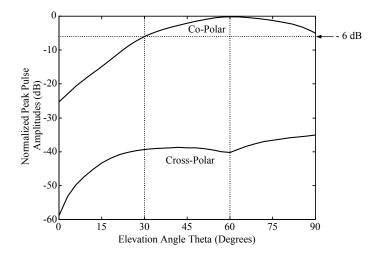


Figure 3.23 Normalized peak pulse amplitudes radiation from broadside of antenna as a function of elevation angle θ for both co-polar and cross-polar components.

Edge-on Far-Field Radiation

In edge-on face of rugby-ball antenna (see Figure 3.24), the maximum peak pulse amplitude of 50 mV occurs at an angle of θ equal to 60°. While the

minimum peak pulse amplitude of 10 mV occurs at an angle of θ equal to 0°. In general a symmetrical transient radiation pattern is observed on either side of z-axis. The normalized peak amplitudes in dB of the radiated pulses for both copolar and cross-polar in edge-on case are plotted as a function of the elevation angle θ in Figure 3.25.

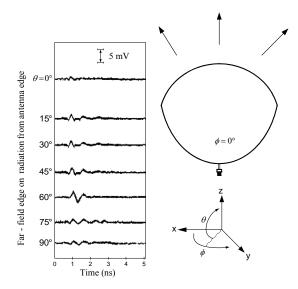


Figure 3.24 Measured results for the edge-on far-field radiation of antenna.

The co-polar component is higher than the cross-polar radiation by at least 17 dB for all angles. Maximum co-polar pulse radiation occurs at $\theta = 60^{\circ}$ and the radiation aperture opening is about 72° in the θ interval 12° to 84°.

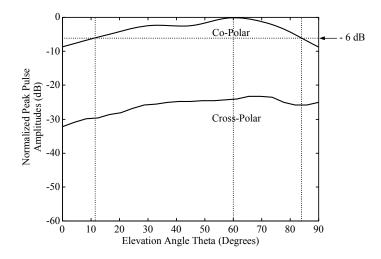


Figure 3.25 Normalized peak pulse amplitudes radiation from edge-on of antenna as a function of elevation angle θ for both co-polar and cross-polar components.

C. Gain Pattern

The measured and simulated (HFSS) radiation (gain) pattern in azimuthplane (xy-plane) and elevator-plane (xz-plane) of the proposed antenna are presented in Figure 3.26 to Figure 3.33 at 3.5 GHz, 7 GHz, 15 GHz, and 20 GHz, respectively. Regarding the azimuth-plane, the directional radiation beams for each frequency were separated in two directions symmetrical to x-axis, while in elevator-plane one directional radiation beam is observed. The antenna directional radiation, which is defined by HPBW, and the angle between the beams are tabulated in Table 3.2.

Frequency (GHz)	Azimuth-plane			Elevation-plane
	Directional radiation beam (degree)		Angle between beam (degree)	Directional radiation beam (degree)
3.5	50 (±10)	135 (±10)	85	55 (±15)
7.0	40 (±10)	140 (±10)	100	60 (±10)
15.0	30 (±10)	155 (±10)	125	75 (±8)
20.0	20 (±10)	170 (±10)	150	65 (±4)

Table 3.2 The directional radiation beam of the antenna.

It is evident from Table 3.2 that as the operation frequency of the antenna increases, the angle between the directional radiation beams in azimuth-plane diverges from 85° to 150°. This behavior of rugby-ball antenna can be attributed to the flat metal plate antenna. Furthermore, with respect to the elevation-plane, the directional radiation beam becomes narrower with the increase of frequency.

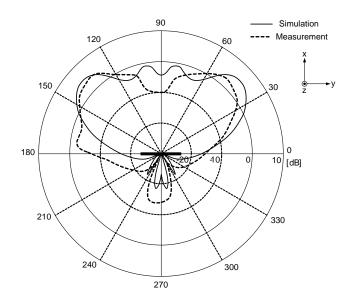


Figure 3.26 The azimuth gain pattern of the antenna for $\theta = 60^{\circ}$ at 3.5 GHz.

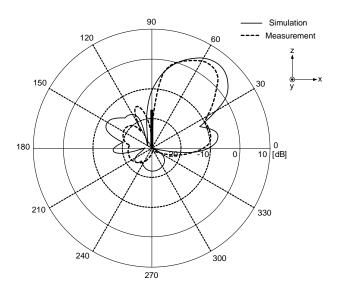


Figure 3.27 The elevation gain pattern of the antenna at 3.5 GHz.

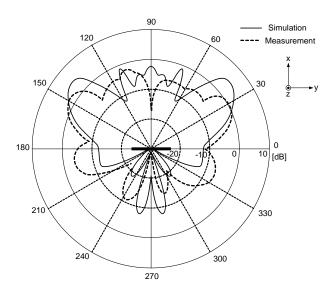


Figure 3.28 The azimuth gain pattern of the antenna for $\theta = 60^{\circ}$ at 7 GHz.

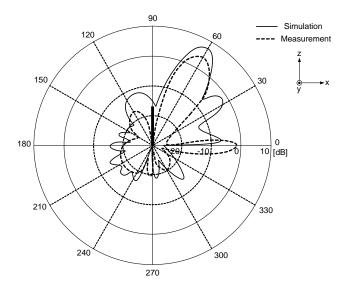


Figure 3.29 The elevation gain pattern of the antenna at 7 GHz.

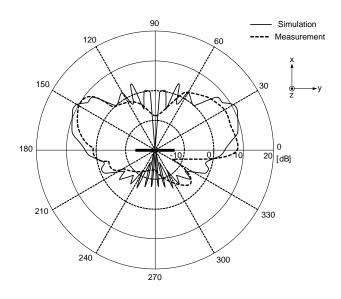


Figure 3.30 The azimuth gain pattern of the antenna for $\theta = 60^{\circ}$ at 15 GHz.

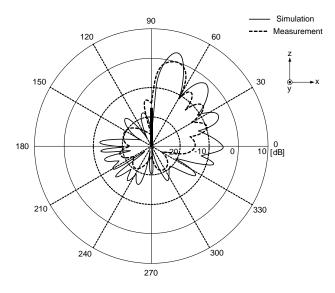


Figure 3.31 The elevation gain pattern of the antenna at 15 GHz.

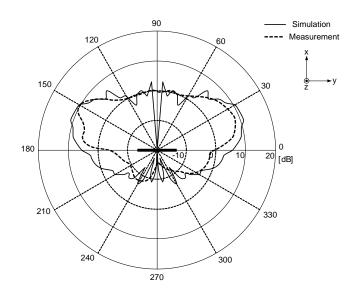


Figure 3.32 The azimuth gain pattern of the antenna for $\theta = 60^{\circ}$ at 20 GHz.

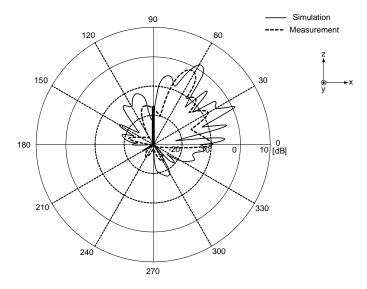


Figure 3.33 The elevation gain pattern of the antenna at 20 GHz.

3.4 Reducing Weight of Antenna

In previous sections, the design and analysis of the prototype rugby-ball antenna has been discussed and fabricated using aluminium. In order to use the antenna in a portable system, the antenna weight becomes importance. Weight reduction can be accomplished by drilling holes in the metal plate as reported in [26]. The size of a hole should be smaller than $0.1\lambda_h$ to $0.3\lambda_h$, where λ_h is the wavelength at the upper operating frequency of antenna. Therefore, 4 mm-diameter holes ($0.3\lambda_h$ is equal to 4.5 mm at 20 GHz.) were drilled into the reflector, the ground plane, and the radiation element as shown in Figure 3.34.

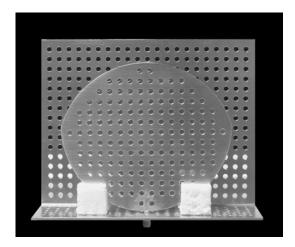


Figure 3.34 Photograph of the realized rugby-ball antenna with holes.

In Figure 3.35, VSWR was used to compare the impedance match of the original antenna to the light-weight design. The antenna with holes has become significant lighter. The VSWR characteristics remained very stable when compared with the performance of the original design. The original element had a total weight of 0.24 kg but the modification brought this down to 0.19 kg (20.83 % weight reduction). Both input resistance and reactance of the rugby-ball antenna (with and without holes) are calculated from S-parameter measurement results and are shown in Figures 3.36 and 3.37, respectively.

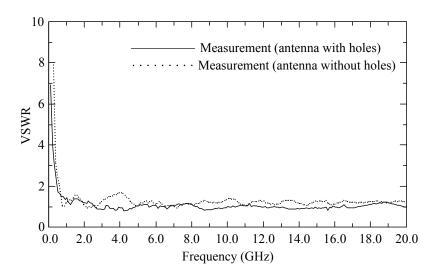


Figure 3.35 Comparison of measured VSWR of holes and without holes rugby-ball antenna.

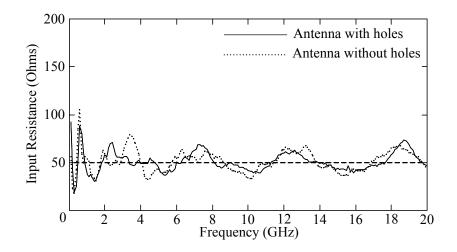


Figure 3.36 Measured antenna input resistance from rugby-ball antenna with holes (solid line) and without holes (dot line).

The input resistance measurement results of the antenna with and without holes are varying around 50 Ω for frequencies in the range of 0.6 GHz to 20 GHz (Figure 3.35). In this frequency range, the input resistance varies between a maximum of 80 Ω and a minimum of 35 Ω for antenna without holes. The antenna with holes has input resistance varying between a maximum of 67 Ω and a minimum of 35 Ω .

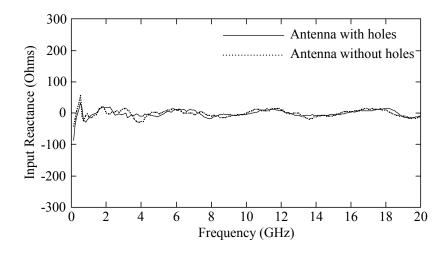


Figure 3.37 Measured antenna input reactance from rugby-ball antenna with holes (solid-line) and without holes (dot-line).

This oscillatory shape of the input resistance of both measurement results are attributed to the multiple resonance of the antenna due to the rounded shape of the lower circle. The input reactance measurement results of the both types of antennas are also varying about zero for the same frequency range (Figure 3.37).

Applications of the rugby-ball antennas with and without holes for bi-static and mono-static radar sensor are discussed in Chapter 5.

List of References

- [1] E. L. Mokole, M. Kragalott, and K. R. Gerlach, *Ultra-Wideband Shortpulse Electromagnetics.* 6th *Edition.* New York: Kluwer Academic/Plenum Publishers, 2003.
- [2] W. L. Stutzman and G. A. Thiele, *Antenna Theory and Design.* 2nd *Edition*, New York: John Wiley & Sons, 1981.
- [3] G. H. Brown and O. H. Woodward, "Experimentally determined radiation characteristics of conical and triangular antennas," *RCA Review*, vol. 13, pp. 425-452, December 1952.
- [4] M. C. Bailey, "Broad-band half-wave dipole," *IEEE Transactions on Antennas and Propagation*, vol. AP-32, no. 4, pp. 410-412, April 1984.
- [5] J. A. Evans and M. J. Amman, "Planar trapezoidal and pentagonal monopoles with impedance bandwidth in excess of 10:1," *IEEE International Symposium Digest (Orlando)*, vol. 3, pp. 1558-1559, 1999.
- [6] S. Honda, M. Ito, H. Seki and, Y. Jinbo, "A disc monopole antenna with 1:8 impedance bandwidth and omni-directional radiation pattern," *Proceedings of the ISAP '92 (Sapporo, Japan)*, pp. 1145-1148, September 1992.
- [7] P. P. Hammoud and F. Colomel, "Matching the input impedance of a broadband disc monopole," *Electronics Letters*, vol. 29, pp. 406-407, February 1993.
- [8] J. A. N. Noronha, T. Bielawa, D. G. Sweeney, S. Licul, and W. A. Davis, "Designing antennas for UWB systems," *Microwaves & RF Magazine*, pp. 53-61, June 2003.
- [9] K. L. Shlager, G. S. Smith, and J. G. Maloney, "Optimization of bow-tie antennas for pulse addiction," *IEEE Transactions on Antennas and Propagation*, vol. 42, no. 7, pp. 975-982, July 1994.
- [10] A. Duzdar and G. Kompa, "A novel inverted trapezoidal antenna fed by ground image plane and backed by a reflector," *Proceedings the 30th European Microwave Conference*, Paris, France, pp. 1-4, October 2000.

- [11] E. Gazit, "Improved design of a vivaldi antenna," *IEEE Proceedings*, vol. 135, no. 2, pp. 89-92, April 1988.
- [12] D. H. Schaubert and T.-H. Chio, "Wideband vivaldi arrays for large aperture antennas," *Perspectives on Radio Astronomy-Technologies for Large Antenna Arrays*, Netherlands Foundation for Research in Astronomy, pp. 49-57, 1999.
- [13] F. Guangyou, "New design of the antipodal vivaldi antenna for a GPR system," *Microwave and Optical Technology Letters*, vol. 44, no. 2, pp. 136-139, January 2005.
- [14] S.-Y. Suh, A Comprehensive Investigation of New Planar Wideband Antennas. Virginia Polytechnic Institute and State University, Virginia, U.S.A., 2002.
- [15] Federal Communications Commission, Notice of inquiry in the matter of: Revision of part 15 of the commission's rules regarding ultra-wideband transmission systems. Document # 02-48, April 2002.
- [16] A. Duzdar, *Design and Modeling of an UWB Antenna for a Pulsed Microwave Radar Sensor*. Doctoral Thesis, HFT, University of Kassel, July 2001.
- [17] P. P. Hammoud and F. Colomel, "Matching the input impedance of a broadband disc monopole," *Electronics Letters*, vol. 29, pp. 406-407, February 1993.
- [18] N. P. Agrawall, G. Kumar, and K. P. Ray, "Wide-band planar monopole antennas," *IEEE Transactions on Antennas and Propagation*, vol. 46, No. 2, pp. 294-295, February 1998.
- [19] A. Ruengwaree, R. Yuwono, and G. Kompa, "A novel rugby-ball antenna for pulse radiation," *Proceedings of the 35th European Microwave Conference (EuMC2005)*, Paris, France, pp. 1855-1858, October 2005.
- [20] R. Yuwono, *Ultra-Wideband (UWB) Antenna for Monostatic Microwave Radar*. Master-Thesis, HFT, University of Kassel, February 2005.
- [21] J. Rautio, "An investigation of microstrip conductor loss," *IEEE Microwave Magazine*, vol. 1, pp. 60-67, December 2000.

- [22] I. Bardi and Z. J. Cendes, *New Directions in HFSS for Designing Microwave Devices*, Microwave Journal, Horizon House Publications Inc., August 1998.
- [23] R. C. Johnson and H. Jasik, *Antenna Engineering Handbook.* 2nd Edition, New York: McGraw-Hill, 1984.
- [24] G. Kompa, *Microwave Integrated Circuits I.* Lecture Notes, HFT, University of Kassel, April 2004.
- [25] C. A. Balanis, *Antenna Theory Analysis and Design. 2nd Edition*, New York: John Wiley and Sons, 1997.
- [26] P. Eskelinen, "Improvements of an inverted trapezoidal pulse antenna," *IEEE Transactions on Antennas and Propagation Magazine*, vol. 43, no. 3, pp. 82-85, 2001.
- [27] D. M. Pozar, *Microwave Engineering*. 2nd Edition, New York: John Wiley and Sons, 1998.

Chapter 4

Description of the Developed Radar Modules

4.1 Multi-Stage SRD-Pulse Sharpening Circuit

UWB transmitters are attractive for pulsed radar and high data rate communication applications. This section describes design details for the formation of ultra-short electrical pulses. Measurement accuracy is increased if the rise time of the pulses is reduced in case of range measurement using time-of-flight principle [1]-[2].

Recently, many efforts have been given for generating and shaping of ultrashort monocycle Gaussian pulses [3]-[4]. There are several ways to reduce rise time of pulses. For example, with use of nonlinear transmission line (NLTL) [5], photoconductive switches [6], avalanche transistor switches together with step recovery diode (SRD) based pulse sharpening circuits. In case of NLTL and photoconductive switches, monolithic integrated circuits (MMIC) are preferred and useful only for low power applications. In case of UWB radar applications, in order to achieve short (sub-nanosecond) rise time of the transmitted pulses, SRD based circuits are extensively used as the pulse-forming network, where SRD is used as a charge-controlled switch [7]-[8]. SRD based circuits are easy to fabricate and cost-effective compared to NLTL or photoconductive switch methods. The rise time of the pulses is dependent on the diode biasing condition, and parasitics in the circuit [7]. Parasitics arise from the diode package and discontinuities in the biasing and matching circuitry. From the hardware designer's point of view, these circuits encounter number of challenges regarding broadband matching with the antenna. In the present approach, selfbiased mode of the SRD was used. Design of the circuit becomes much simpler as the parasitics are reduced in the circuit and broadband matching with the antenna becomes easier, keeping the switching performance unaltered.

The following section will describe the properties of the SRD and the SRD model used for the design of multi-stage SRD-pulse sharpening circuit. Finally, experimental results will be discussed for the single-stage and double-stage pulse sharpener.

4.1.1 Properties of the SRD

The SRD is a two-terminal P-I-N junction [7] whose static characteristics are similar to the usual p-n junction diode, but whose dynamic characteristics are quite different. The SRD dynamic characteristics are extremely important for switching circuit applications.

A. Ideal Dynamic Characteristics

The most distinguishing feature of the SRD is the very abrupt dependence of its junction impedance upon its internal charge storage [9]. This storage of charge occurs as a result of the non-zero recombination time of minority carriers that have been injected across the junction under forward bias conditions. In case of reverse biasing, the device continues as low impedance (generally less than 1 ohm) until all the charge is totally removed, at which point it rapidly switches from low impedance to high impedance, thereby stopping the flow of reverse current. The ability of the SRD to store charge and to rapidly change impedance levels can be exploited in the generation of extremely fast rise time pulses. The impedance transition time of the SRD diodes used in this work, which were supplied by Metelics Corporation [10], was about 30 - 38 ps for the beam lead-packaged SRD. The photograph and dimensions of SRD (MMDB-30-B11) diode is shown in Figures 4.1.

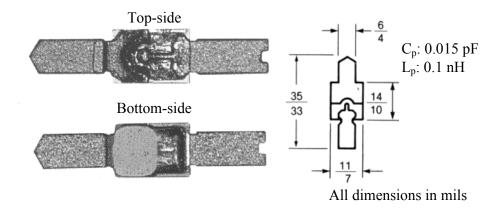


Figure 4.1 Photograph and dimensions in mils of the SRD (MMDB-30-B11).

The charge stored under forward bias can be obtained from the charge continuity equation [7]

$$i(t) = \frac{dQ}{dt} + \frac{Q}{\tau} \quad \text{for } Q > 0, \qquad (4.1)$$

where i(t) is the total instantaneous diode current, Q is the charge stored at junction, and τ is minority carrier lifetime of diode. In case of a constant charging current, the stored charge can be defined as

$$Q_F = I_F \tau (1 - e^{-t_F/\tau}), \qquad (4.2)$$

where Q_F is stored charge from forward current, I_F is the forward charging current and t_F is the length of time during which forward charging current is applied. When a constant reverse current is applied across the SRD, the time used to remove the stored charge is

$$\frac{t_s}{\tau} = \ln \left[1 + \frac{I_F (1 - e^{-t_F / \tau})}{I_R} \right],$$
(4.3)

where t_S is time required to remove the charge stored by I_F and I_R is the reverse current.

B. Actual Dynamic Characteristics

In a practical SRD, the presence of package parasitic effects should be taken into accounts, which in turn affect the ideal diode characteristics. Consequently, the differences between the output waveform of ideal and practical SRD become visible as shown in Figure 4.2. For detail understanding of actual dynamic characteristics of SRD, the equivalent circuit of a packaged SRD must be considered (see Figure 4.3).

The first parasitic effect in the waveform is the voltage drop across the diode under the forward bias condition and can be expressed as

$$V_F = \phi + I_F R_S, \qquad (4.4)$$

where V_F is the voltage drop, ϕ is the contact potential (0.7V to 0.8V) and R_s is dynamic parasitic resistance. This voltage drop becomes visible itself in the waveform before the sharpening occurs (see Figure 4.2).

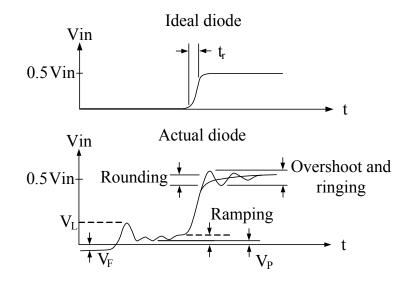


Figure 4.2 Ideal and actual dynamic characteristics of SRD diode [7].

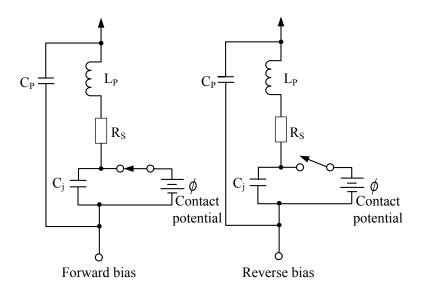


Figure 4.3 SRD equivalent circuits [7].

The second effect exhibits voltage spikes caused by the rapid change of reverse current through package inductance, which can be identified by the value of package inductance and reverse current as

$$V_{L}(\max) = L_{P} \left[\frac{di_{d}}{dt} \right]_{\max}$$
(4.5)

where V_L (max) is the maximum voltage spike and L_P is the package inductance. The value of voltage spike will be reduced when using the lower value of package inductance. In this work, the beam lead-packaged SRD (MMDB-30-B11) are used. For this model, the package inductance (L_P) is 0.1 nH. Other electrical parameters of SRD are tabulated in Table 4.1.

Model-outline	Breakdown voltage: VBR (V)		Capacitance: C _t (pF)		Lifetime: τ (ns)	Transition time: $t_t(ps)$	
	Min	Max	Typical	Max	Typical	Typical	Max
MMDB-30-B11	14	25	0.15	0.25	10	30	38

 Table 4.1 Electrical parameters of SRD [10].

The third effect is voltage plateau. This voltage appears because of the reverse current component flowing through the dynamic series resistance R_S of the SRD for the period of the storage phase. This voltage is defined as

$$V_{P} = (I_{F} + I_{R})R_{S}$$
 (4.6)

where V_P is the voltage plateau. In general, the value of voltage plateau is usually small (approximately: 0.16 V). This effect can be neglected when the sharpened pulse has large amplitude.

The last parasitic effect is the overshoot and ringing waveform (see Figure 4.2). This effect is due to a damped resonance of the SRD, package capacitance (C_P) with the package inductance (L_P) , and the stray inductance of the circuit. Exciting of the high frequency components of the output current occur this stray inductance. This effect can be reduced by minimizing stray circuit reactance and choosing a SRD with a minimized package capacitance and inductance.

C. Transition Rise Time

The transition rise time (t_r) of the SRD is defined as the time during which the SRD changes its impedance states within a circuit, which is dependent on the diode design, circuit constraints, and the level of stored charge. The transition rise time is composed of two components, as follows

$$t_r = \sqrt{t_t^2 + t_{RC}^2}$$
(4.7)

where t_t is the intrinsic diode transition time which depends on the diode itself and is usually specified by the manufacturer (in this work $t_t = 30$ ps), and t_{RC} is the circuit control rise time which is dependent on the diode reverse biased capacitance (C_{VR}) [junction capacitance (C_i) in parallel with package capacitance (C_P)] of SRD and the equivalent circuit resistance in parallel with SRD. In case of 10% to 90% rise time, the t_{RC} is given by

$$t_{RC} = 2.2R_{eq}C_{VR}$$
 (4.8)

and in case of 20% to 80% rise time, the t_{RC} is

$$t_{RC} = 1.4R_{eq}C_{VR}$$
 (4.9)

where R_{eq} is the equivalent resistance, consisting of the parallel combination of the source and load resistances.

4.1.2 Used ADS Model of SRD

A table-based SRD model used in this work was developed in [11]. The SRD model [11] was developed at the Department of High Frequency Engineering. This model is based on DC and the bias-dependent high-frequency (HF) measurements of the SRD. In [11], the circuit model used for each bias point was a modified version of the model proposed in [12] and [13], where the commonly used internal series resistance (R_n) are replaced by a more accurate resistance network (see Figure 4.4) where R_{max} is the maximum resistance of the diode at the reverse bias, R_v represents the forward resistance at high forward current (depends on voltage bias), and where R_s is the series resistance. Bias dependent junction capacitance and junction resistance were modelled using symbolically defined device (SDD) in Advanced Design System (ADS) software. The schematic of the SRD model is described in Figure 4.5. The two capacitances C_f and C_r represent the forward and reverse capacitance of the diode, respectively, C_i and R_i represent the p-n junction capacitance and resistance of the SRD diode, and Cp and Lp are package capacitance and inductance, respectively. SRD model was implemented in ADS. All the circuit parameters were determined from the DC and S-parameter measurements at difference bias point. First, the DC I-V characteristics of beam lead-package SRD were measured [11]. This was done to determine a starting value for the junction resistance (R_j) of diode before extraction was applied on the basis of the RF measurements. Measured DC properties of the diode are shown in Figure 4.6. Current limiter (maximum current of 20 mA) was used during DC forward measurement. Threshold voltage (V_T) is calculated to be 0.89V.

The circuit model parameters in Figure 4.5 were extracted according to [11]. The starting values for C_p , L_p , R_s , C_f and C_r were obtained from the manufacturer's data and from the DC measurements that were conducted.

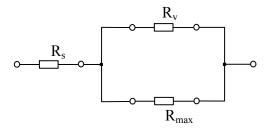


Figure 4.4 Internal series resistance (R_n) model [11].

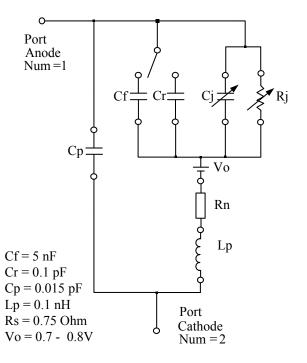


Figure 4.5 Table-based circuit model of the beam lead-package SRD [11].

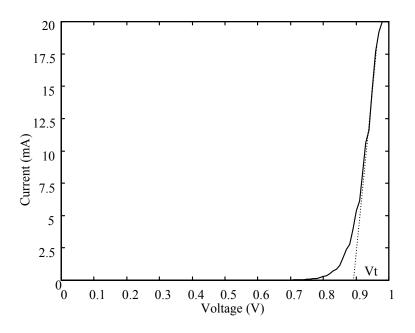


Figure 4.6 DC properties of the beam lead-package SRD [11].

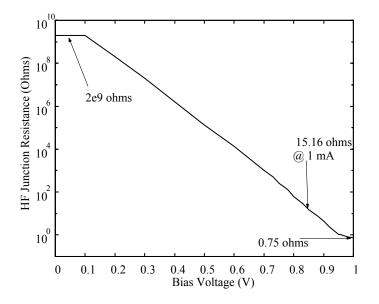


Figure 4.7 Junction resistance characteristic of beam lead-package SRD [11].

The results of the optimization process for both junction resistance and capacitance of the beam lead-packaged SRD are shown in Figure 4.7 and 4.8, respectively. The junction RF resistance, shown in Figure 4.7, is equal to 0.75Ω at a high-level forward bias current, which is the value of R_s. Moderate forward current level of 1 mA has been used to model R_v. At 0V bias, R_{max} is obtained as 2 GΩ. Figure 4.8 shows the C-V characteristics of the beam lead-packaged SRD

diode. At strong forward bias conditions, the junction capacitance is very large reaching a value of 5.28 nF at a forward voltage of 0.95V. Under reverse bias conditions, the capacitance becomes very small reaching a value of 0.12 pF at a reverse voltage of -5V. When lower reverse bias or higher forward bias voltages are needed, the value of the capacitances can be easily extrapolated from the graph.

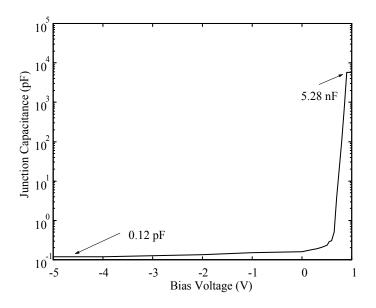


Figure 4.8 Junction capacitance characteristic of beam lead-package SRD [11].

4.1.3 Circuit Design

In general the cost of a pulse generator plays a major role in the conception and design process of a radar sensor [14]-[18]. A cost-efficient way of generating such pulses is through the use of a SRD. This section gives a detailed description of design and simulation of two-stage SRD pulse sharpening circuit, which is developed from single-stage SRD circuit. A block diagram of the 2stage SRD-pulse sharpening circuit, similar to the one discussed in [11], is shown in Figure 4.9. The circuit is composed of an input high-pass filter matched to a 50 Ω microstrip line, an input matching network to the SRD diodes, the 2-stage SRD for negative pulse rise time (in this case fall time) picosecond sharpening, a coupling network, an output matching network, and an output high-pass filter for Gaussian shaping purposes, which is also matched to a 50Ω microstrip line.

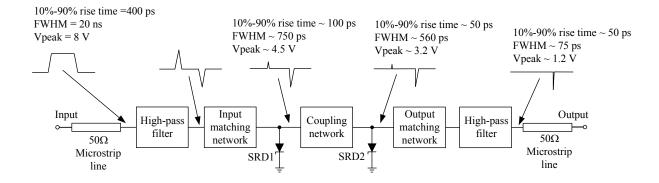


Figure 4.9 Block diagram of the SRD based 2-stage pulse sharpening circuit.

The high-pass filter differentiates the pulse obtained from the picosecond pulse labs generator (Model 10000A) with 10%-90% rise time of 400 ps, pulse width (FWHM) of 20 ns, and the peak amplitude of 8V (see Figure 2.15). The output of the filter is matched to the impedance of the SRD1 stage through an input matching network. The positive pulse supplies the forward bias current that charges the SRD1 diode, which causes the diode to act as a low impedance device. When the negative pulse arrives at the terminals of the SRD1, the stored energy is discharged, that was accumulated under forward bias. When the charge is totally removed, the SRD1 snaps to a high impedance state very rapidly. Using such a feeding configuration at the input of the SRD, helps to avoid DC biasing of the SRD in the forward direction. After sharpening the fall time of the negative-going pulses by the SRD1, a SRD2 is used to shape the pulse again to obtain faster fall time of negative pulse. Finally, a high-pass filter is used to shape the pulse to achieve a Gaussian waveform. Gaussian pulse shaping was necessary since the pulse is fed directly into the radiation antenna.

Designed pulse sharpening circuit using two-stage SRD diodes is shown in Figure 4.10. The SRD model was implemented using SDD elements available in ADS software. Model parameters were defined based on the extracted values from DC and S-parameter measurement. Junction capacitance of the diode was extracted to be 0.2 pF at the threshold bias point, and transition time was calculated to be 35 ps for the beam lead type SRD. The pulse sharpener was realized in microstrip technique using fiber-woven Teflon substrate with dielectric permittivity (ϵ_r) of 2.51, dielectric substrate height (h) of 0.381 mm, and conductivity thickness (t) of 18 µm. Lumped component models include parasitics, and the impedances are defined according to the datasheet supplied by manufacturer [10].

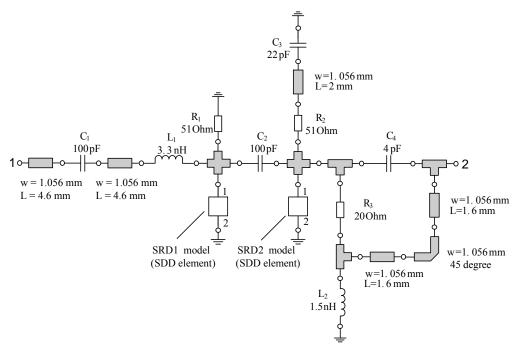


Figure 4.10 Schematic of SRD based 2-stage pulse sharpening circuit in ADS.

4.1.4 Experimental Results

To verify the 2-stage SRD pulse sharpener design, a sharpening circuit was fabricated and tested. Figure 4.11 shows the fabricated circuit of 2-stage SRD pulse sharpener in housing. The optimized physical size of 2-stage circuit is 3 cm x 3.5 cm, while the size of single-stage is 4 cm x 6 cm [11]. Thus, the dimension of new pulse sharpener is reduced to 50%. Here, the connecting input and output of pulse sharpener circuit is using 50 Ω coaxial cables. A 50 Ω

microstrip line was used to realize the circuit. Using HP-54120B Digital Oscilloscope, the input and output pulses of the circuit were measured.

Experimental and simulation results of single-stage [11] and 2-stage SRD pulse sharpener are shown in Figure 4.12 and Figure 4.13, respectively. Measured 10%-90% rise time (t_r) of the single-stage SRD circuit was found to be 100 ps (amplitude = 2.5V), whereas using the 2-stage SRD circuit, 10%-90% rise time of 50 ps and amplitude of 1.25V were obtained. In both cases, the circuits were excited by rectangular pulse with amplitude of +8V, 10%-90% rise time of 400 ps, 90%-10% fall time of 900 ps, and pulse width (FWHM) of 20 ns (see Figure 2.15) by picosecond pulse labs generator (model 10000A). Time domain measured pulse parameters for the single stage and 2-stage SRD circuits are summarized in Table 4.2.

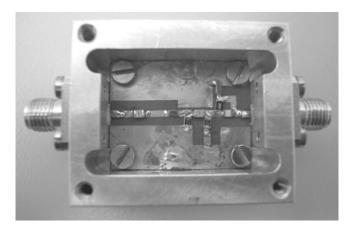


Figure 4.11 Fabricated 2-stage SRD pulse sharpener.

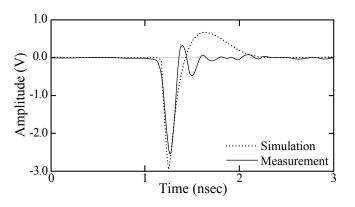


Figure 4.12 Transient simulation and measurement result of SRD single stage pulse sharpening circuit.

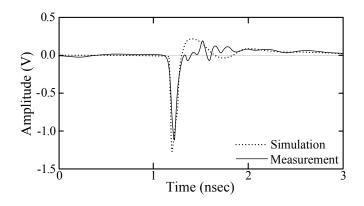


Figure 4.13 Transient simulation and measurement result of SRD 2-stage pulse sharpening circuit.

Parameter	Single-stage	Two-stage		
Rise time (10%-90%)	100 ps	50 ps		
Pulse width (FWHM)	140 ps	75 ps		
Pulse amplitude	2.5V	1.25V		

Table 4.2 Pulse parameters.

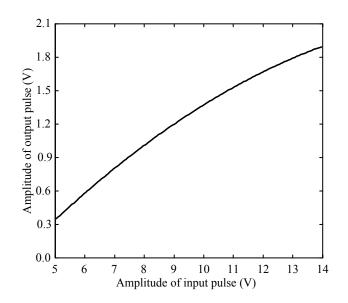


Figure 4.14 Measured output pulse amplitude as function of the input pulse amplitude using the beam lead-packaged diode.

Since the output pulse of the 2-stage SRD sharpening circuit depends only on the input pulse, several measurements were conducted using the digital sampling scope (HP 54120B) to measure the output pulse characteristics such as its amplitude, pulse width (FWHM) and 10%-90% rise time (t_r) as function of the input pulse amplitude. The input pulse rise time (10%-90%) was for all measurements 400 ps and its PRF was set to 100 kHz. The change in that parameter had no significant effect on the output pulse parameters. Each measurement was repeated 10 times. The average of these measurements was recorded. The results are shown in Figure 4.14 and Figure 4.15.

First, the input pulse width (FWHM) was set to 20 ns and its amplitude varied between 5 - 14V at 1V step intervals. Figure 4.14 shows the absolute output pulse amplitude as function of input pulse amplitude for the 2-stage SRD pulse sharpening circuits using the MMDB-30-B11 beam lead-package SRD. As can be observed in Figure 4.14, the relationship between the input and output pulse amplitude is nonlinear. This nonlinearity of the SRD can be attributed to the dynamic junction capacitance (C_j) and resistance (R_j) as shown in Figure 4.5, Figure 4.7, and Figure 4.8.

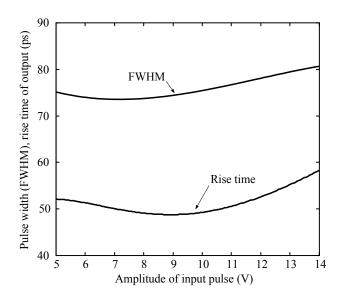


Figure 4.15 Measured output FWHM and t_r as a function of the input pulse amplitude.

Figure 4.15 shows the pulse width (FWHM) of the output pulse and its 10%-90% rise time (t_r) as a function of the input pulse amplitude. There are changes in both pulse width (FWHM) and rise time (t_r) as the input pulse amplitude increases. The 10%-90% rise time increases from 48 ps to 58 ps while the pulse width (FWHM) increases from 74 ps to 80 ps as the input pulse amplitude increased from 5V to 14V. The dependency of the output pulse rise

time and pulse width (FWHM) on the input pulse amplitude is due to the increase in the amount of injected charges in SRD under forward bias conditions as the input pulse peak amplitude is increased. This implies that the SRD needs longer time to remove these stored charges under reverse bias conditions before the change of state of impedance occurs.

4.2 Broadband Resistive Duplexer

A mono-static radar sensor has been developed using system components of the realized bi-static radar. Regarding a mono-static radar system, the transmitter and receiver antenna are contained within one element referred to as a transceiver antenna. A broadband resistive duplexer (BRD) is essential for the detection of the transmitted and received signal and couples these signals directly to the sample unit. The duplexer is principally a three port device and each port is matched to a 50 Ω . Port 1 is connected to the multi-stage SRD pulse sharpener. Port 2 is connected to the sampling unit and port 3 is connected to the rugby-ball antenna as shown in Figure 4.16.

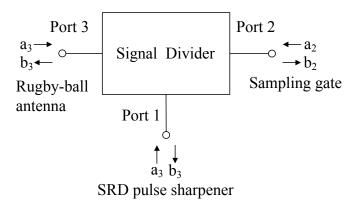


Figure 4.16 The three port broadband resistive duplexer.

4.2.1 Design

The design of a three-port duplexer depends on two important goals. The first one is to keep the eigen-reflection coefficients (S_{11} , S_{22} , and S_{33}) below -10 dB. The second goal is related to the transmission coefficients (S_{21} , S_{31} and S_{32}),

which should be more than -3 dB. In this work, the resistive bridge used in the duplexer is designed on the basis of reflectometer concept [19]. The three port divider contains lossy components to improve simultaneous matching of all ports. The circuit was realized with 50 Ω planar microstrip lines. Design and optimization of the duplexer circuit has been performed using the electromagnetic field simulator in ADS software. After having performed several simulation and optimization, the final network of this type is shown in Figure 4.17 (a).

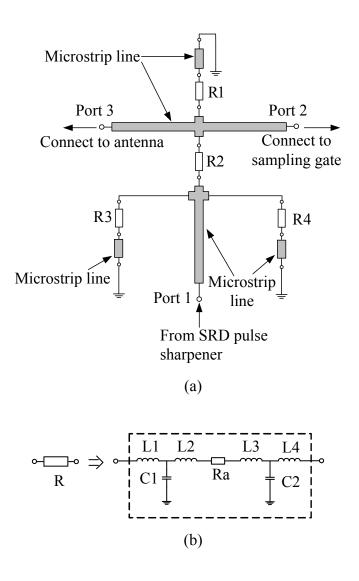


Figure 4.17 (a) The schematic of the potential divider. (b) Model of the resistance with parasitics.

The duplexer consists of symmetrical resistive films deposited on ceramic substrate. Models of resistor were extracted for thick film resistors considering its parasitic inductance and capacitance as in Figure 4.17 (b). Extracted values are given in Table 4.3. In Figure 4.17 (a), the duplexer splits the input pulse, which is received from the SRD pulse sharpener at port 1 into two pulses, a reference pulse at port 2, which is fed to the sampling unit, and the other as a transmitted pulse at port 3, which is fed to the antenna for radiation. Afterwards, the received return pulses are driven through the duplexer at port 3 and transmitted to port 2, which is then fed to the sampling unit. The measured and the simulated results of the duplexer are shown in the next section.

	Ra (Ω)	L1 (pH)	L2 (pH)	L3 (pH)	L4 (pH)	C1 (pF)	C2 (pF)
R1	1k	53.0666	112.367	112.367	53.0666	0.179331	0.179331
R2	5	53.0666	112.367	112.367	53.0666	0.179331	0.179331
R3	2k	53.0666	112.367	112.367	53.0666	0.179331	0.179331
R4	2k	53.0666	112.367	112.367	53.0666	0.179331	0.179331

 Table 4.3 Extracted values of resistor parasitics.

4.2.2 Fabrication and Results

In this section, the realization, simulation and measurement results will be presented and discussed. The hybrid duplexer circuit has been fabricated in microstrip technique using fiber-woven Teflon substrate with dielectric permittivity (ε_r) = 2.51, dielectric substrate height (h) = 0.381 mm, and conductivity thickness (t) = 18 µm. In order to avoid mismatch at the ports of the circuit, uniform 50 Ω microstrip lines (width = 1.056 mm) have been realized. The SMD (surface mounted device) chip resistors were then soldered to the microstrip lines.

Measurement and simulation results of the reflection and transmission coefficients of the duplexer are shown in Figure 4.18 to Figure 4.20. The reflection and transmission coefficients were measured from DC to 12 GHz using a vector network analyzer of type HP8510. Figure 4.18 shows the transmission coefficient S_{31} from SRD pulse sharpener (port 1) to the antenna (port 3), and the eigen-reflection coefficient S_{33} of port 3 (antenna). In Figure 4.19, the transmission coefficient S_{23} from antenna (port 3) to sampling gate (port 2), and the eigen-reflection coefficient S_{22} of port 2 are shown. Figure 4.20 gives the transmission coefficient S_{21} , for the reference pulse from port 1 (SRD pulse sharpener) to port 2 (sampling gate), and the eigen-reflection coefficient S_{11} of port 1 (SRD pulse sharpener). In general, the simulated S-parameters agree very well with measurement results. For all ports of the duplexer, a return loss of more than 10 dB was measured for $S_{11},\,S_{22}$ and S_{33} up to 7 GHz. At higher frequencies, increase of the return loss was noticed due to parasitic effects in the microstrip-to-coaxial transition [20]. Also, the transmission coefficients (S₃₁, S₂₃, and S₂₁) are attenuated by 3 dB due to insertion of resistive components. The application of the broadband resistive duplexer in the monostatic radar system will be discussed in chapter 5.

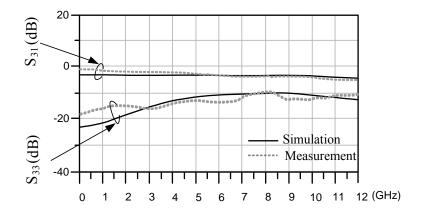


Figure 4.18 Simulated and measured eigen-reflection coefficients (S₃₃) and transmission coefficients (S₃₁) of the broadband resistive duplexer.

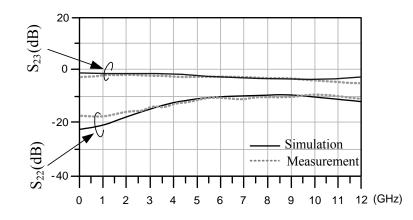


Figure 4.19 Simulated and measured eigen-reflection coefficients (S₂₂) and transmission coefficients (S₂₃) of the broadband resistive duplexer.

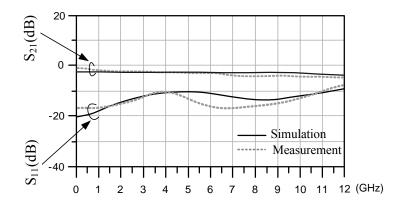


Figure 4.20 Simulated and measured eigen-reflection coefficients (S_{11}) and transmission coefficients (S_{21}) of the broadband resistive duplexer.

List of References

- [1] A. Ruengwaree, A. Ghose, and G. Kompa, "A novel UWB rugby-ball antenna for near-range microwave radar system," *IEEE Transactions on Microwave Theory and Techniques*, vol. 54, No. 6, pp. 2774-2779, June 2006.
- [2] A. Ghose, Pulsed Measurement Based Nonlinear Characterization of Avalanche Photodiode for the Time Error Correction of 3D Pulsed Laser Radar. Doctoral Thesis, University of Kassel, July 2005.
- [3] E. L. Mokole, M. Kragalott, and K. R. Gerlach, *Ultra-Wideband Shortpulse Electromagnetics.* 6th *Edition*, New York: Kluwer Academic/Plenum Publishers, 2003.
- [4] J. S. Lee and C. Nguyen, "Novel low-cost ultra-wideband, ultra-shortpulse transmitter with MESFET impulse-shaping circuitry for reduced distortion and improved pulse repetition rate," *IEEE Microwave and Wireless Components Letters*, vol. 11, no. 5, pp. 208-210, May 2001.
- [5] M. Case, M. Kamegawa, R. Yu, and M. J. W. Rodwell, "Impulse compression using solition effects in a monolithic GaAs circuit," *Applied Physics Letters*, vol. 58, no. 2, pp. 173-175, January 1991.
- [6] J. F. Holzman, F. E. Vermeulen, and A. Y. Elezzabi, "Recombinationindependent photogeneration of ultrashort electrical pulses," *Applied Physics Letters*, vol. 76, no. 2, pp. 134-136, January 2000.
- [7] Hewlett Packard, *Pulse and Waveform Generation with Step Recovery Diodes.* Application Note 918, California/USA, October 1986.
- [8] A. M. Nicolson, H. M. Cronson, and P. G. Mitchell, "Subnanosecond risetime pulse generators," *IEEE Transactions on Instrumentation and Measurement*, vol. 25, no. 2, pp. 104-107, June 1976.
- [9] J. L. Moll, S. Krakauer and R. Shen, "P-N junction charge-storage diode," *Proceedings of the Institute of Radio Engineers,* vol. 50, pp. 43-53, January 1962.

- [10] *State of Art Microwave Diodes.* Catalog of SRD's/Varactor Diodes, Metelics Corporation, 975 Stewart Drive, Sunnyvale, California 94086, USA (www.metelics.com).
- [11] X. Xu, Characterization and Modeling of SRD Diodes for the Computer Aided Design of a Generator of Ultrashort Pulses. Master-Thesis, HFT, University of Kassel, November 1999.
- [12] J. Zhang and A. Räisänen, "A new model of step recovery diode for CAD," *IEEE MTT-S Digest*, pp. 1459-1462, 1995.
- [13] J. Zhang and A Räisänen, "Computer-aided design of step recovery diode frequency multipliers," *IEEE Transactions on Microwave Theory and Techniques*, vol. 44, pp. 1503-1506, 1997.
- [14] J. Pulfer and B. Whitford, "A simple method of generating nano-second pulses at X band," *Proceedings of the IRE*, pp. 968, May 1961.
- [15] P. Paulus, W. Brinker, and D. Jäger, "Generation of microwave pulses by opto-electronically switched resonators," *IEEE Journal of Quantum Electronics*, vol. QE-22, pp. 108-111, January 1986.
- [16] D. Salameh and D. Linton, "Microstrip GaAs nonlinear transmission-line (NLTL) harmonic and pulse generators," *IEEE Transactions on Microwave Theory and Techniques*, pp. 1118-1122, July 1999.
- [17] Y. Qian and E. Yamashita, "Phase compensation and waveform reshaping of picosecond electrical pulses using dispersive microwave transmission lines," *IEEE Transactions on Microwave Theory and Techniques*, vol. 39, pp. 924-929, June 1991.
- [18] S. Salinas, *Compression Techniques of Asymmetric Ultra-Short Electrical Pulses*. Master-Thesis, HFT, University of Kassel, April 1999.
- [19] F. van Raay and G. Kompa, "A new active balun reflectometer concept for DC to microwave VNA application," *Proceedings of the 28th European Microwave Conference*, Amsterdam, The Netherlands, pp. 108-113, October 1998.
- [20] J. S. Wong, *Rectangular Coaxial Line to Microstrip Line Matching Transition and Antenna Subarray Including the SAME*. U.S. Patent 5,982,338, November 1999.

Chapter 5

Application of New UWB Microwave Radar Sensor

5.1 Radar Sensor Measurement Setup

The bi-static and mono-static radar sensor was tested in the laboratory for distance measurement to metallic targets, for the detection of buried metallic targets in dry sand and for water level control measurements. In case of mono-static radar measurements, the target is placed in line with the centre point at the top edge of the antenna (see Figure 5.1). In the bi-static setup the reference is taken as virtual centre point between the two central top edges of the antennas, spaced at a distance of 10 cm (see Figure 5.2). This was done to assure maximum illumination of targets and to obtain the shortest-range measurement possible.

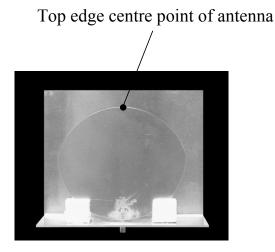


Figure 5.1 Top edge centre point of antenna used as a reference to measure a target range for mono-static configuration.

In both cases, the sensors are mounted on a single-legged stand, which determines the height of the antennas from the ground, for metallic target detection, hidden in dry sand, and for water level control measurements. In case of distance measurement to metallic targets, a three-legged stand was utilized. With the single-legged and three-legged stands angular orientation in both elevation and azimuth direction could be changed. To investigate the UWB microwave radar "front end", comprising UWB antennas, pulse sharpening circuit and broadband resistive duplexer circuit, experiments were conducted with absorber material surrounding the measurement setup to reduce unwanted reflection signals.

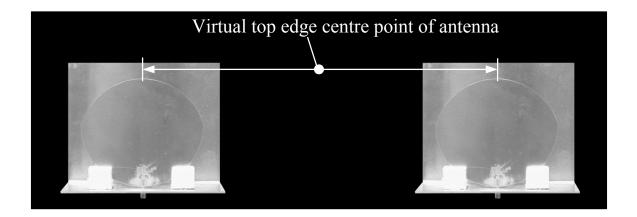


Figure 5.2 Virtual top edge centre point between the two antennas of the bi-static configuration used as a reference to measure a target range.

The measurement setup of bi-static and mono-static radar sensor is illustrated in Figure 5.3 and Figure 5.4, respectively. In all measurements, a rectangular pulse (see Figure 2.15) with amplitude of 8V and a pulse width (FWHM) of 20 ns was used to feed the SRD pulse sharpening circuit. The output of the sharpener was a Gaussian pulse with an amplitude of 1.25V, a 10%-90% rise time of 50 ps, and a pulse width (FWHM) of 75 ps (see Figure 4.13).

In the bi-static configuration, the output of the pulse sharpening circuit was fed directly to the transmitter of the rugby-ball antenna (TX). As a result the Rx

antenna receives, first the reference pulse (using the edge-on radiation characteristic of the antenna) and secondly the target return pulses (using the broadside radiation characteristic of the antenna). In case of the mono-static configuration, Gaussian pulses were fed to port 1 of the broadband resistive duplexer circuit. The output, at port 2 and port 3 of the duplexer, was fed to the sampler to act as a reference pulse and to the antenna for target detection.

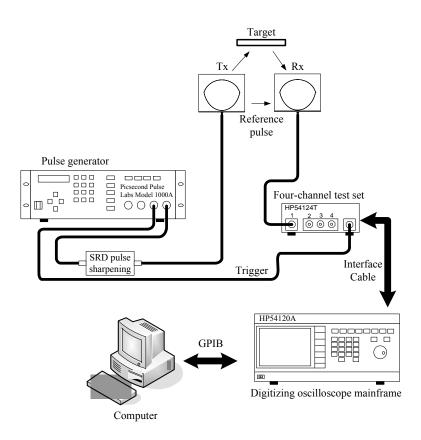


Figure 5.3 Measurement setup of the bi-static radar sensor.

In case of both bi-static and mono-static configuration, the Rx antenna and port 2 of duplexer, respectively, were connected to the digital sampling oscilloscope (HP 54120B having a 50 GHz sampling head). For data acquisition during measurements, general purpose interface bus (GPIB) was used to interface personal computer (PC) and oscilloscope.

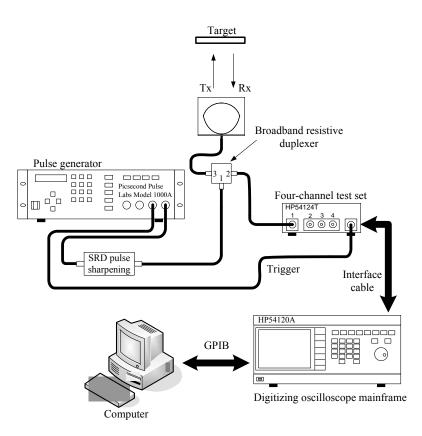


Figure 5.4 Measurement setup of the mono-static radar sensor.

The raw data obtained from measurement involved noise and unwanted peaks due to reflector edge diffraction, edge diffraction, and target ringing (see Figure 5.5 (a)). These unwanted pulses were reduced by using a companding technique [1]-[2] in which the pulse amplitude was raised to power 3. This resulted in a clear distinguishing characteristic of wanted and unwanted data. The processed data was again scaled back to the original pulse amplitude by which the noise and unwanted peaks disappear. Subsequently, digital signal processing (DSP) using one-dimensional multilevel stationary wavelet decomposition technique (wavelet toolbox of MATLAB® [3]) was used to reduce the noise of the target information pulse. This toolbox builds on the numeric and visualization capabilities of MATLAB software to provide point-and-click graphical tools and command line functions for analysis, synthesis, denoising, and compression of signals and images. In this case, the de-noising command line function was applied. The options of de-noise function have 5

level to remove noise of pulse signal, which can be varied from level 1-5. The level 1 implies minimum noise reduction while level 5 refers to maximum cancellation of noise from the measurement data. However, the higher level of de-noise function will considerably affect the amplitude and sharpness of the pulse signal. For this reason, level 3 was chosen in this work, which was also verified by conducting several experiment trails. The illustration of signal before and after using reduced unwanted pulse signal and noise process is shown in Figure 5.5. In this figure, the time-domain target response of an aluminium plate is depicted having the dimensions of 48 cm x 22 cm and thickness of 2 mm, with the target being placed at a distance of 75 cm from the virtual central antenna. The target's lower edge and the antenna ground image plane were both aligned. The virtual central antenna is labelled in the figure as the "reference pulse" where the first radiation pulse occurs, and the "reflector edge diffraction" presents the radiation of the edge reflector at the back of the antenna.

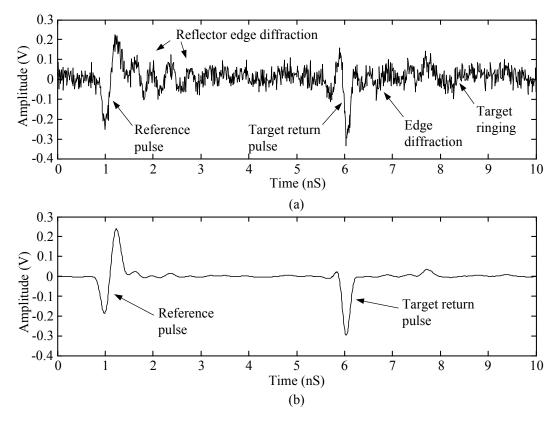


Figure 5.5 Time-domain response of the sensor in bi-static configuration (a) before and (b) after data processing using multilevel stationary wavelet decomposition.

After target return pulse, edge diffraction in time-domain is detected. This pulse is diffracted from the edge of the aluminium plate target. In addition, another small ringing pulse is detected after the edge diffraction has occurred. Creeping waves or surface waves, which move forward and backward on the target surface, causes this ringing. The target distance is estimated using the standard time-of-flight method. In this method, the 50% amplitude points at the falling front slope of both the reference pulse and the target return pulse should be estimated (see Chapter 1). The estimation is based on the detection of 10% and 90% amplitude points in each falling edge. The time difference between these two points is used in conjunction with equation (2.1) in Chapter 2 to calculate the distance.

5.2 Distance Measurement to Metal Plates

The first experimental test for bi-static and mono-static radar sensor is to measure the distance of metallic targets. In both configurations, the antennas and targets were supported by the stand, which are mounted 92 cm and 87 cm (see Figure 5.6) above the ground, respectively.

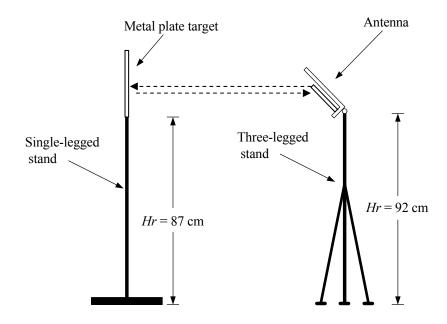


Figure 5.6 Position of the metal plate target and the antenna in distance measurement.

The antennas were tilted in such a way that the radar beam was directed to the target (elevator angle θ is approximately 60°). The transmitted Gaussian pulses had a 10%-90% rise time of 50 ps, an amplitude of 1.25V, and a pulse width (FWHM) of 75 ps. Figures 5.7 to 5.10 show the time-domain response of the sensor with a large aluminium plate with lateral dimensions of 48 cm x 22 cm and with a thickness of 2 mm. The target was placed in the horizontal position, at distances of 50 cm and 100 cm from the antenna. Figure 5.7 and Figure 5.8 show the results of the bi-static radar sensor while Figure 5.9 and Figure 5.10 show the results from the mono-static radar sensor.

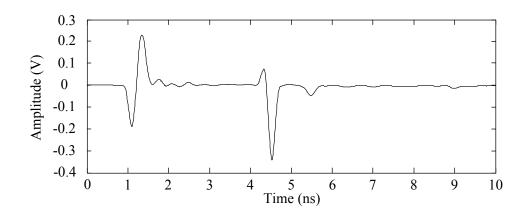


Figure 5.7 Time-domain response of the bi-static radar sensor [aluminium plate (48 cm x 22 cm x 0.2 cm) is located at 50 cm from the virtual centre of the antenna].

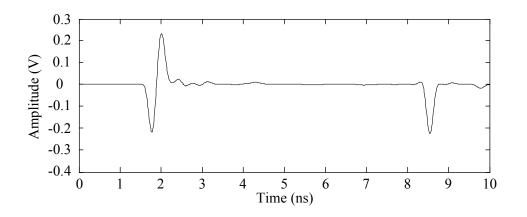


Figure 5.8 Time-domain response of the bi-static radar sensor [aluminium plate (48 cm x 22 cm x 0.2 cm) is located at 100 cm from the virtual centre of the antenna].

In the next step, bi-static and mono-static radar sensors were used to measure the distance of a smaller aluminium plate target. The plate had dimensions of 10 cm x 10 cm x 5 mm and was positioned at a distance of 50 cm and 100 cm away from the antenna. In these measurements, the placement of antennas and metallic targets were similar for both radar sensors as shown in Figure 5.6 but the height of the antennas and metallic targets were reduced to 85 cm and 82 cm, respectively, in order to maintain line of sight between the antenna and metallic target. The time-domain response of the radar sensor in bistatic configuration is shown Figure 5.11 and Figure 5.12. In both figures, the target lies in the horizontal position at a distance of 50 cm and 100 cm away from antenna. Figure 5.13 and Figure 5.14 show the respective response signals for the mono-static configuration.

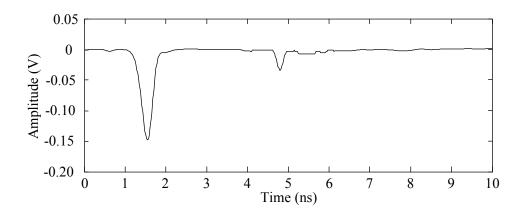


Figure 5.9 Time-domain response of the mono-static radar sensor [aluminium plate (48 cm x 22 cm x 0.2 cm) is located at 50 cm from the centre of the antenna].

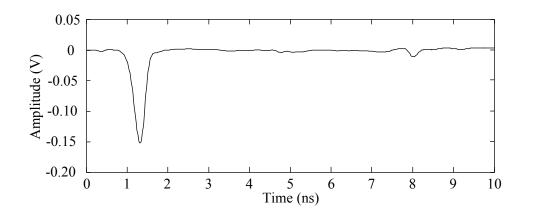


Figure 5.10 Time-domain response of the mono-static radar sensor [aluminium plate (48 cm x 22 cm x 0.2 cm) is located at 100 cm from the centre of the antenna].

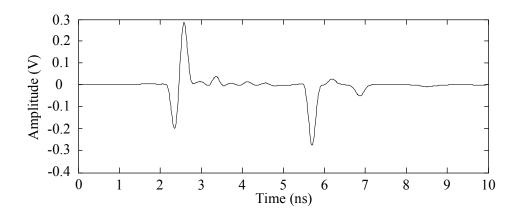


Figure 5.11 Time-domain response of the bi-static radar sensor [aluminium plate (10 cm x 10 cm x 0.5 cm) is located at 50 cm from the virtual centre of the antenna].

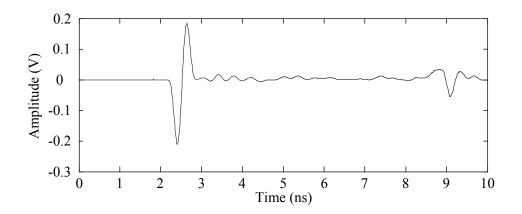


Figure 5.12 Time-domain response of the bi-static radar sensor [aluminium plate (10 cm x 10 cm x 0.5 cm) is located at 100 cm from the virtual centre of the antenna].

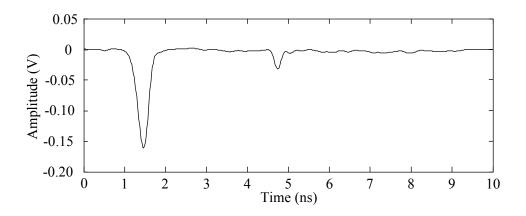


Figure 5.13 Time-domain response of the mono-static radar sensor [aluminium plate (10 cm x 10 cm x 0.5 cm) is located at 50 cm from the centre of the antenna].

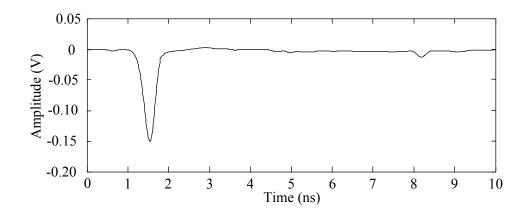


Figure 5.14 Time-domain response of the mono-static radar sensor [aluminium plate (10 cm x 10 cm x 0.5 cm) is located at 100 cm from the centre of the antenna].

In these measurements (involving small and large metallic targets), the targets were detected and the ranges were easily calculated. The amplitude of target return pulse has been found to be inversely proportional to the square of the distance between the antennas and metal target [4]. As shown in Figure 5.15, the normalized amplitude characteristics obtained from mono-static radar sensor shows good agreement with the theory ($\approx 1/R^2$) when compared to the bi-static case. In addition, the reflector edge diffractions (see Figure 5.5) occurred only in bi-static sensor since reference pulses in mono-static sensor were not sent through receiver antenna while in the case of bi-static sensor edge side radiation of the antennas was used as reference pulse. However, it has been observed that in the mono-static sensor, the amplitude of the target return pulses are lower than the pulses received from the bi-static sensor. This is due to the fact that, in mono-static sensor, the signal from the generator is attenuated by 3 dB due to the insertion loss of the broadband resistive duplexer. Moreover, an additional 3 dB attenuation is introduced when the signal returns from the target through the duplexer again. Consequently, the signal is attenuated by 6 dB, which results in small amplitude of the target return pulse.

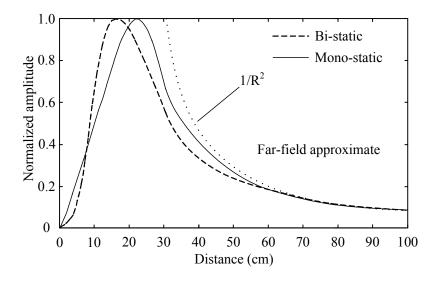


Figure 5.15 Received signal as function of distance (R).

5.3 Detection of Metal Plates Buried in Dry Sand

In this experiment, the metallic target is buried in dry sand. The bi-static radar sensor has been used in order to measure the target return pulse. This experiment could not be performed with the mono-static due to the weak amplitude of the return pulse. A test sand bucket was set up in the laboratory for testing the response of the sensor. In this case, an aluminium plate with dimensions of 26 cm x 9 cm x 0.2 cm was buried in the test sand bucket (46 cm x 60 cm x 28 cm). The sand bucket was placed under the antenna. The antenna was oriented to radiate in the direction perpendicular to sand bucket. The distance between sand bucket and antenna was 41 cm. The antenna setup was the same as used with the distance measurement to metal plates with same power, with a radar pulse rise time of 50 ps (10%-90%) and a pulse width (FWHM) of 75 ps. The amplitude of the transmitted pulse was 1.25V. In Figure 5.16, the target return pulse is clearly seen and easily detected, similar to the case of metallic targets in distance measurement.

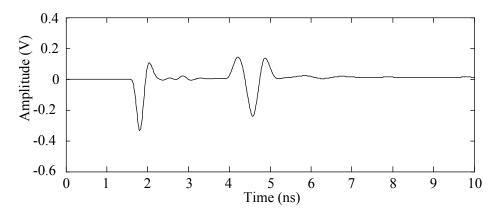


Figure 5.16 Measured time-domain response of the bi-static radar sensor [aluminium plate (26 cm x 9 cm x 0.2 cm) is buried 26 cm underneath of dry sand and located 41 cm from the virtual centre of the antenna].

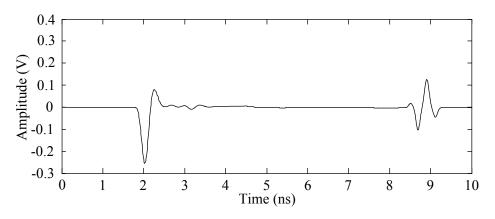


Figure 5.17 Measured time-domain response of the bi-static radar sensor [aluminium plate (26 cm x 9 cm x 0.2 cm) is buried 15 cm underneath of dry sand and located 100 cm from the virtual centre of the antenna].

The last measurement setup was used to increase the measurement distance to 100 cm away from the virtual centre of the antennas. In this case the metal plate was buried 15 cm in the sand. The measurement result is shown in Figure 5.17. In both experiments conducted, the increasing thickness of sand poses attenuation and dispersion on the measurement return pulses [5]-[6].

5.4 Water Level Control Measurement

In addition to metallic target detection, water level control was conducted using reflection and ranging. This was an interesting test to determine the water level inside the rainwater tank with mm accuracy. The measurement setup of bistatic and mono-static configuration for water level control measurement is shown in Figure 5.18 and Figure 5.19, respectively. In both configurations, a commercial 200 litres plastic rainwater tank was filled with predetermined increments of water level and having reached the specified water level, the water surface was determined. The water level was changed in steps of 5 cm till a level of 80 cm was reached. The antennas were hanged above the ground of 100 cm, which is supported by the single-legged stand (see Figure 5.18 and Figure 5.19). Figure 5.20 shows the time-domain response signal of the bi-static radar sensor for a water level of 40 cm, which shows the reference and return pulse with absolute peak-to-peak amplitude of 0.3V and 0.4V, respectively. Since, in Figure 5.20, the distance (D) between the antenna and water surface is calculated as 60 cm which implies that the water level is equal to 40 cm (100-D). Figure 5.21 shows a similar time-domain response for which the water was detected at 45 cm. The absolute peak-to-peak amplitude of reference and water surface return pulse are 0.15V and 0.025V, respectively.

With respect to water level control measurements, it has been shown that the amplitude of return pulse from surface of water are relatively large due to the high value of dielectric constant (ε_r) (see Appendix A). The amplitudes are comparable with the detected return pulses from the metallic plate.

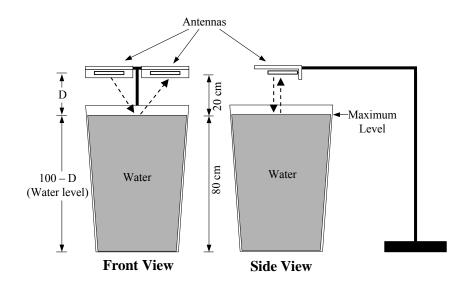


Figure 5.18 Water control measurement with bi-static microwave radar.

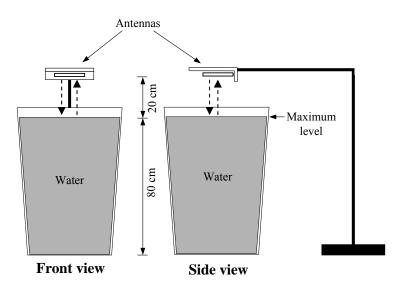


Figure 5.19 Water control measurement with mono-static microwave radar. (Note for simplification the broad-side of the antenna is draw parallel to the water surface and not under a rated angle of 60 degrees.)

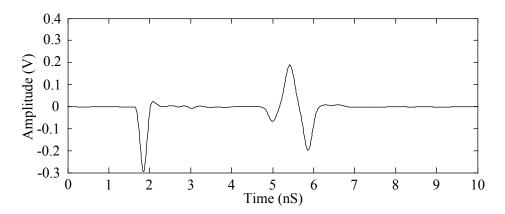


Figure 5.20 Measured time-domain response of the bi-static radar sensor at a water level of 40 cm.

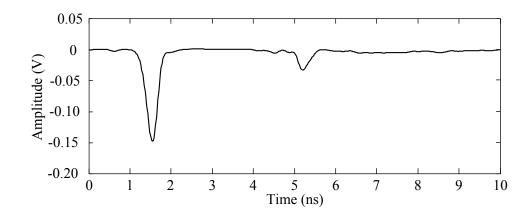


Figure 5.21 Measured time-domain response of the mono-static radar sensor at a water level of 45 cm.

In order to observe the variation of the amplitude of the return pulse for mono-static and bi-static radar sensor while increasing the water level a plot of detected and actual water level is shown in Figure 5.22 and Figure 5.23, respectively. For mono-static case, at higher water levels (upper 55 cm), the measurements become more uncertain when compared to bi-static case because of the sensitive of radar sensor (near-field region) as show in Figure 5.15.

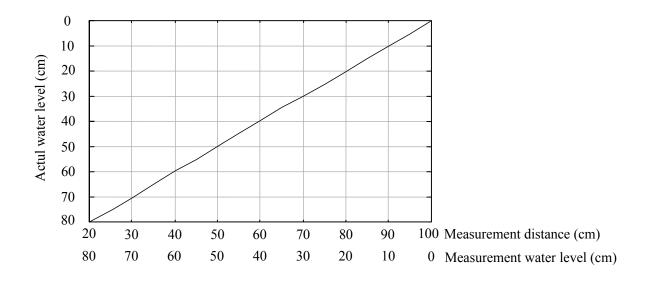


Figure 5.22 Detected water levels as function of actual water level (bi-static radar sensor).

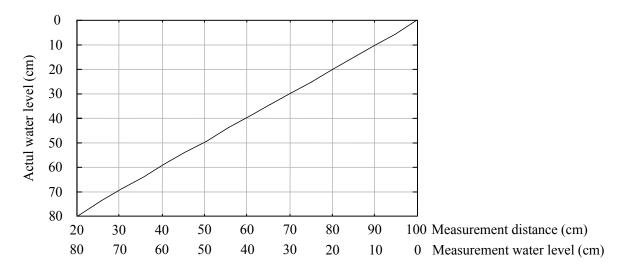


Figure 5.23 Detected water levels as function of actual water level (mono-static radar sensor).

5.5 Ranging Uncertainty

The ranging uncertainties of the mono-static and bi-static radar sensors were tested. The experiments should indicate the effect of long-term sensor temperature drift on the range measurement calculation. In both configurations, the measurement results were carried out by keeping a metal placed at 100 cm far from the antenna. Range measurement calculations were performed using 100 measurement results.

5.5.1 Measurement Accuracy of Bi-Static Radar Sensor

The first test of ranging uncertainty measurement is presented for the bistatic radar sensor. A statistical analysis of the measurement errors has been performed for the 100 measurements, which were computed for a metal plate (dimensions of 10 cm x 10 cm x 0.5 cm) at position 100 cm from the central virtual point of antennas.

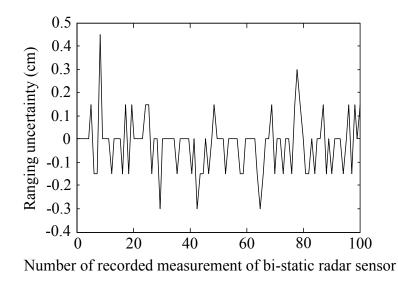


Figure 5.24 Measured ranging uncertainty of the bi-static radar sensor.

The ranging uncertainties are obtained by taking the difference of the actual and the measured distances. An uncertainty of 4.5 mm was observed as shown in Figure 5.24. In order to investigate the statistical occurrence of distance error, the distance errors are plotted as function of the probability as shows in Figure 5.25. The mean range uncertainty is found to be well within 0.075 mm.

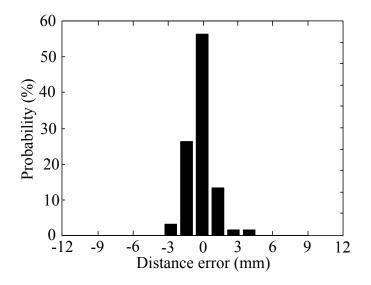


Figure 5.25 Plot of the statistical distribution of the distance error of range measurements with bi-static radar sensor.

5.5.2 Measurement Accuracy of Mono-Static Radar Sensor

In this section, the measured ranging uncertainty of the mono-static radar sensor is discussed. The setup and process of measurement accuracy is similar to the bi-static radar sensor (see section 5.5.1). Figure 5.26 shows the measured ranging uncertainty of 100 distance measurements. An uncertainty of 5.5 mm was achieved, which is more than the bi-static case. In addition, Figure 5.27 shows a statistical distribution of the distance error. The mean range uncertainty is found to be well within 3.18 mm. Consequently, it can be conducted that the mono-static sensor exhibits larger measurement uncertainty than the bi-static one due to the weaker power transmitted. Thus, the experiments have shown that both the bi-static and mono-static system are well suited for near range distance measurement.

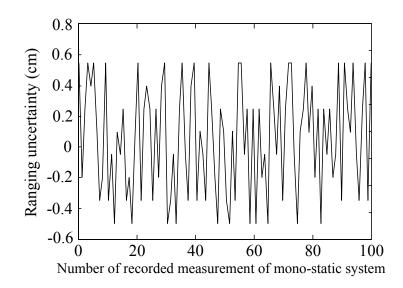


Figure 5.26 Measured ranging uncertainty of the mono-static radar sensor.

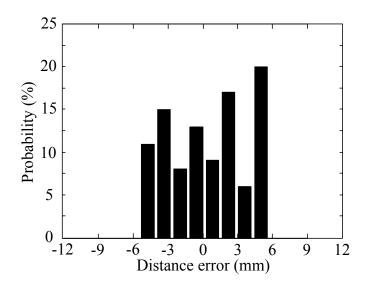


Figure 5.27 Plot of the statistical distribution of the distance error of range measurements with mono-static radar sensor.

List of References

- [1] R. Petelin and Y. Petelin, *Adobe Audition Soundtracks for Digital Video*. New Delhi: Laxmi Publications PUT. LTD., 2006.
- [2] R. L. Libbey, *Signal and Image Processing Sourcebook*. New York: Multiscience Press Inc., 1994.
- [3] MATLAB, Version 7, The Mathworks Inc., Natick, MA, USA, 2004.
- [4] G. Kompa, *High Frequency Sensors*. Lecture Notes, HFT, University of Kassel, April 2002.
- [5] E. G. Farr and C. A. Frost, *Ultra-Wideband Antennas and Propagation*. Farr Research Inc., Report WL-TR-1997-7051, vol. 4, July 1997.
- [6] Y.-H. Zhou, Q. Shu, and X. J. Zheng, "Attenuation of electromagnetic wave propagation in sandstorms incorporating charged sand particles," *European Physical Journal E*, pp. 181-187, May 2005.

Chapter 6

Conclusion and Future Work

This dissertation has been focused on near-field ranging using radar sensors which send ultra-short electrical pulses to the targets and measure the elapsed time between the sent and received pulses. Both bi-static and mono-static pulsed microwave radar sensors have been referenced. The proposed front-end of the mono-static radar sensor consists of an UWB rugby-ball antenna, a SRD (step recovery diode) based electrical pulse sharpening circuit, and a broadband resistive duplexer. In case of the bi-static radar sensor, two separate antennas are implemented to transmit and receive picosecond electrical pulses. On the other hand, in case of mono-static radar sensor, a single UWB antenna is used for both transmitting and receiving purpose.

Main focus of the research work was on the design of a UWB "rugby-ball" antenna. The three dimensional simulation and optimization were performed prior to practical antenna realization. The antenna was constructed using a metal plate with a thickness of 2 mm. The antenna is fed through an image ground plane and backed by a reflector. The percentage of bandwidth of the antenna is defined by the voltage standing wave ratio (VSWR) having a value less than 2 and has found to be 187.4%, extending from about 0.65 GHz to 20 GHz. The simulation results could be verified by measurements.

Based on measurements of far-field pulse radiation, the proposed antenna can transmit and receive ultra-short electrical pulses with 10%-90% rise time of 50 ps and pulse width (FWHM) of 75 ps. The pulse radiation with maximum amplitude of the antenna occurs at an elevation angle of 60° from broadside and

edge-on directions. Regarding the azimuth angle, maximum radiation is observed at 0° and 90° from broadside and edge-on directions, respectively. A radiation aperture (θ direction) at the broadside of the antenna was found to be 60° (between angle θ of 30° to 90°) while in edge-on direction of antenna it was 72° (between angle θ of 12° to 84°).

After fabrication and investigation of the optimized prototype ruby-ball antenna, holes with a diameter of 4 mm were drilled in the radiator element, ground plane, and reflector of the antenna which reduced the antenna weight considerably by 20%, without affecting the overall antenna characteristics. In this regard, the VSWR remained less than 2 retaining the impedance close to 50Ω with negligible reactance within a frequency range of 0.65 GHz to 20 GHz.

For feeding the antenna, a new electrical pulse sharpening circuit, based on the step recovery diode (SRD), was developed. The circuit requires no DC bias and can produce Gaussian pulses with a 10%-90% rise time of 50 ps, a pulse width (FWHM) of 75 ps, and a peak amplitude of 1.25V. By increasing the amplitude of the input pulse from 5V to 14V, the property of the output pulse changed. The pulse width (FWHM) increased from 75 to 80 ps and the rise time grew from 48 to 58 ps.

The bi-static radar sensor was reconfigured as a mono-static radar sensor wherein one antenna was used for both transmitting and receiving purpose. Consequently, the broadband resistive duplexer couples the transmitted and received pulses directly into the sample unit. This duplexer was realized in microstrip technique using lines with a characteristic impedance of 50 Ω and contains symmetrical resistive films deposited on ceramic substrate. The return loss of the duplexer was measured and simulated more than 10 dB up to 7 GHz. Also, the transmission coefficients (S₃₁, S₂₃, and S₂₁) are attenuated by 3 dB up to 10 GHz due to insertion of resistive components.

In the above described system, the developed mono-static and bi-static radar systems were used for water level control measurements and for the detection of metallic target, placed in air and buried in dry sand. Target ranging was performed using the standard time-of-flight concept wherein the range of obstacle was computed from the returned signal, which is distorted. The wanted information signal from the distorted signal was obtained after implementing companding and wavelet decomposition techniques. The range measurement uncertainty was achieved well below of 4.5 mm and 5.5 mm with the bi-static and mono-static configuration, respectively, by repeating the range measurements 100 times.

Thus, it has been shown that the developed sensors can be used for diverse non-contact measurement applications such as detection of buried metallic objects, fluid level control measurements, and nondestructive detection of materials used for construction (e.g. bridge, buildings, etc.).

For future work, both bi-static and mono-static low-cost pulsed microwave radar sensors should be enhanced for increasing range detection, which can be accomplished by developing a ultra-short electrical pulse generator and a sharpener to generate ultra-short electrical pulses with higher output power.

In multiple sensor application, such as three dimensional image surface measurement and long-range detection, both radar sensor configurations can be modified from single-sensor to multi-sensor. This modification could be carried out by enhancing the output power of ultra-short electrical pulse sharpener, multiple duplexer (applicable only to mono-static system) and antennas.

113

Appendix A

Reflection Factor and Dielectric Constant of Water

A.1 Reflection Factor [1]

Significant operating parameter of a radar level measuring device are dependent upon the reflected useful signal of the microwave, e.g. measurability, accuracy, repeatability, error probability and detectives in the case of non-ideal surfaces or interference reflectors. The power reflection factor (R_{pwr}) is defined here as being the ratio of reflected power density to the power density of the incident beam ($R_{pwr} = \rho_{ref} / \rho$).

Electromagnetic waves are reflected by electromagnetic interaction:

- From conductive surfaces (metals, and highly conductive liquids such as acids and saline solutions of sufficient concentration). In these cases, reflection is almost 100%.
- From dielectric liquids (described by the relative permittivity (ε_r), which describes the interaction with electric fields): the reflection factor (*R*) is a function of ε_r which is expressed in [1] as

$$R = \frac{(\sqrt{\varepsilon_r} - 1)^2}{(\sqrt{\varepsilon_r} + 1)^2}$$
(A1)

From Figure A.1, at $\varepsilon_r = 3.5$, the reflection factor is about 10% (-10 dB) whereas for $\varepsilon_r = 1.5$, about 1% (-20 dB) is observed (see Figure A.1). However,

from Figure A.2, according to the HFT optical laboratory environment, the ε_r of water is about 78 at room temperature of 20 C° and atmospheric pressure of 100 kPa. This implies that, form Figure A.1, the reflection factor is 63% (-2 dB), which is very high, aids in the determining the microwave level measuring system.

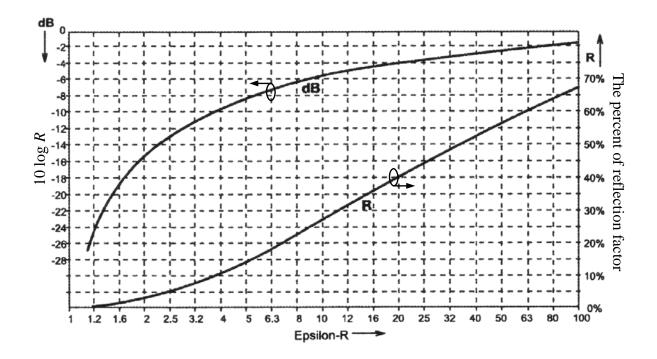


Figure A.1 Reflection factor (*R*) of dielectric media [1].

A.2 Dielectric Constant of Water

In [2], the dielectric constant of water as a function of pressure is described as

$$X = X(T, P_i) + a_0 \ln[(a_1 + P)/(a_1 + P_i)]$$
(A2)

where X is the dielectric constant of the water at temperature T. P_i (in Mpa) is the reference pressure and the coefficients a_i are adjusted to fit experimental values. The dielectric constants of water as a function of pressure with temperature as parameter are plotted in Figure A.2. The Figure shows the proposed equation adjusted to fit data from the international association for the properties of water and steam [3].

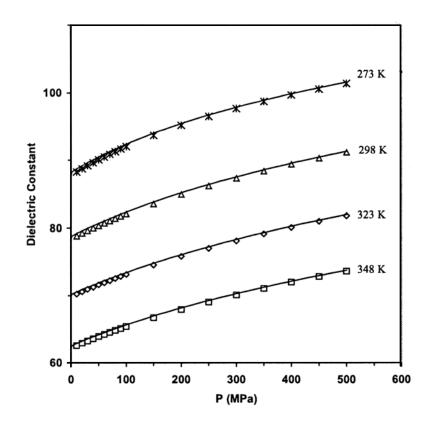


Figure A.2 Dielectric constant of water as a function of pressure at constant temperature [3].

List of References

- [1] D. Brumbi, *Fundamentals of Radar Technology for Level Gauging*. 4th *Edition*, Krohne, 2003.
- [2] W. B. Floriano and M. A. C. Nascimento, "Dielectric constant and density of water as a function of pressure at constant temperature," *Brazilian Journal of Physics*, vol. 34, no. 1, pp. 38-41, March 2004.
- [3] D. R., *CRC Handbook of Chemistry and Physics*. 74th Edition: CRC Press, 1994.

Publications

- 1. A. Ruengwaree, R. Yowuno, and G. Kompa "Novel rugby-ball UWB antenna for base band transmission," 7th International Conference On Telecommunications in Modern Satellite (TELSIKS05), vol. 1, pp. 16-19, Serbia, September 2005.
- 2. A. Ruengwaree, R. Yowuno, and G. Kompa, "A novel rugby-ball Antenna for pulse radiation", European Microwave Conference Proceedings, pp. 1855-1858, Paris, October 2005.
- 3. A. Ruengwaree, A. Ghose, and G. Kompa, "A novel rugby-ball UWB antenna for near-range microwave radar system," IEEE Transactions on Microwave Theory and Techniques, vol. 54, no. 6, pp. 2774-2779, June 2006.
- 4. A. Ruengwaree, R. Yowuno, and G. Kompa, "Design and performance of an UWB antenna for a mono-static microwave radar system", German Microwave Conference, Karlsruhe, pp. GM0084-F, March 2006.
- 5. A. Ruengwaree, R. Yowuno, and G. Kompa, "Ultra-fast pulse transmitter for UWB microwave radar", European Microwave Conference Proceedings, pp. 1833-1836, Paris, September 2006.