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Hardware Development of an Ultra-Wideband System for High Precision Localization Applications

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the Doctor of Philosophy Degree
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Cemin Zhang
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This dissertation is dedicated to my parents, Rong Zhang and
Qiling Zheng, and my loving wife Mingming Wang, for all your
support and aspirations,
and to my beloved daughter Vera Zimo Zhang

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ABSTRACT

A precise localization system in an indoor environment has been developed. The developed system is based on transmitting and receiving picosecond pulses and carrying out a complete narrow-pulse, signal detection and processing scheme in the time domain. The challenges in developing such a system include: generating ultra wideband (UWB) pulses, pulse dispersion due to antennas, modeling of complex propagation channels with severe multipath effects, need for extremely high sampling rates for digital processing, synchronization between the tag and receivers' clocks, clock jitter, local oscillator (LO) phase noise, frequency offset between tag and receivers' LOs, and antenna phase center variation. For such a high precision system with mm or even sub-mm accuracy, all these effects should be accounted for and minimized.

In this *work*, we have successfully addressed many of the above challenges and developed a stand-alone system for positioning both static and dynamic targets with approximately 2 mm and 6 mm of 3-D accuracy, respectively. The results have exceeded the state of the art for any commercially available UWB positioning system and are considered a great milestone in developing such technology. My contributions include the development of a picosecond pulse generator, an extremely wideband omni-directional antenna, a highly directive UWB receiving antenna with low phase center variation, an extremely high data rate sampler, and establishment of a non-synchronized UWB system architecture. The developed low cost sampler, for example, can be easily utilized to sample narrow pulses with up to 1000 GS/s while the developed antennas can cover over 6 GHz bandwidth with minimal pulse distortion.

The stand-alone prototype system is based on tracking a target using 4-6 base stations and utilizing a triangulation scheme to find its location in space. Advanced signal processing algorithms based on first peak and leading edge detection have been developed and extensively evaluated to achieve high accuracy 3-D localization. 1D, 2D and 3D experiments have been carried out and validated using an optical reference system which provides better than 0.3 mm 3-D accuracy. Such a high accuracy wireless localization system should have a great impact on the operating room of the future.

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Chapter 1

Introduction

1.1 Current Status of Wireless Local Positioning Systems

More than ever, business and organizations need reliable, real-time location information. In many cases, knowing the location of your resources/assets can be the difference between success and failure, and sometimes even life and death. Therefore, there is a great demand to develop a wireless local positioning technology as it has many diverse applications and has been extensively studied [1,2]. While Global Positioning Systems (GPS) use ultra high precision atomic clocks to measure the time-of-flight [3], a more standard method for indoor localization systems is use of Time Difference of Arrival (TDOA), where all of the base stations or receivers are synchronized, and the difference in time is measured between each pair of receivers to triangulate the position of an unsynchronized tag [1]. Two main technologies have recently emerged as possible solutions for the TDOA systems: frequency-modulated continuous-wave (FMCW) and ultra-wideband (UWB). FMCW systems can be found both in the literature [4-6] and have been commercially available [7] with various levels of accuracy. For example, Stelzer et al. achieved an accuracy of greater than 10 cm for an outdoor application when tracking a car around a 500 m² racecar track [4] while Wiebking et al. achieved an accuracy around 20 cm for an indoor application covering a 15x25 m² 2-D area [8]. Finally, Roehr et al. achieved an accuracy of 1 cm in a line-of-sight (LOS), multipath free environment using a novel chirp technique centered at 5.8 GHz [9].

Meanwhile, interest in UWB for radar applications in the last few years has increased greatly following the Federal Communications Commission's (FCC) decision to open up the bands from 3.1 - 10.6 GHz and 22 – 29 GHz for UWB use in 2002 [10]. UWB technology is well-known to have inherent advantages for indoor applications in terms of robustness to multipath interference and a great potential for high ranging accuracy [11]. Higher accuracy has been reported in the literature for indoor UWB positioning systems. For example, Low et al. achieved centimeter-range accuracy in a 1-D short range indoor LOS environment utilizing UWB pulse signals [12]. Zetik et al. reported sub-mm 1-D accuracy but with only extremely short displacements while accuracy decreased to 1.5 cm for 2-D localization over a 2x2 m² area [13]. Recently, Meier et al. designed a 24 GHz coherent system which uses a Kalman filter combined with correlation and phase information to reduce the uncertainty of a static point to 0.1 mm, although the uncertainty increases to 2 mm when the tag is in motion [14]. These experimental results have shown that UWB technology has the potential for high precision indoor localization even in harsh environments with significant multipath effects. However, commercial products of such technology are still limited even to this present day due to the complexities associated with the development of accurate wireless localization systems. Besides the research work mentioned above, there are a few companies that have addressed or produced such radars which may represent the state-of-the-art development using UWB or FMCW technology. These systems are (1) PulsON 350 from Time Domain Corporation [15]; (2) Sapphire DART from Multispectral Solutions [16]; (3) Ubisense system [17]; (4) LPR system from Symeo Corporation [18]. These systems have been demonstrated with varying levels of success in different

indoor/outdoor applications, such as personal tracking in a hospital/industrial facility, stock/asset tracking in a warehouse, crane/truck anti-collision detection, etc. A summary of these systems is given here for completeness where a comparison between these systems is given in Table 1.1.

Three different measurement principles are used in the commercial localization systems listed in Table 1.1: angle-of-arrival (AOA), roundtrip-time-of-flight (RTOF) and time-difference-of-arrival (TDOA). Figure 1.1 illustrates the foundation of each concept [1].

Table 1.1 – State-of-the-Art Commercial Wireless Localization Systems

	PulsON350	Sapphire DART	Ubisense	Symeo LPR
Technology	UWB	UWB	UWB	FMCW
Localization Method	TDOA	TDOA	TDOA/AOA	RTOF
Frequency Range	6.6 GHz (center freq.)	5.925 – 7.25 GHz	6-8 GHz	5.8 GHz
Typ. Accuracy	Sub-meter	< 30 cm	15 cm	+/- 5cm
Range	Building	200m (LOS)	> 160 m	400 m
Data Rate	1 Hz	200 Hz	20 Hz	25 Hz
Price	N/A	>\$20,000	>\$20,000	N/A

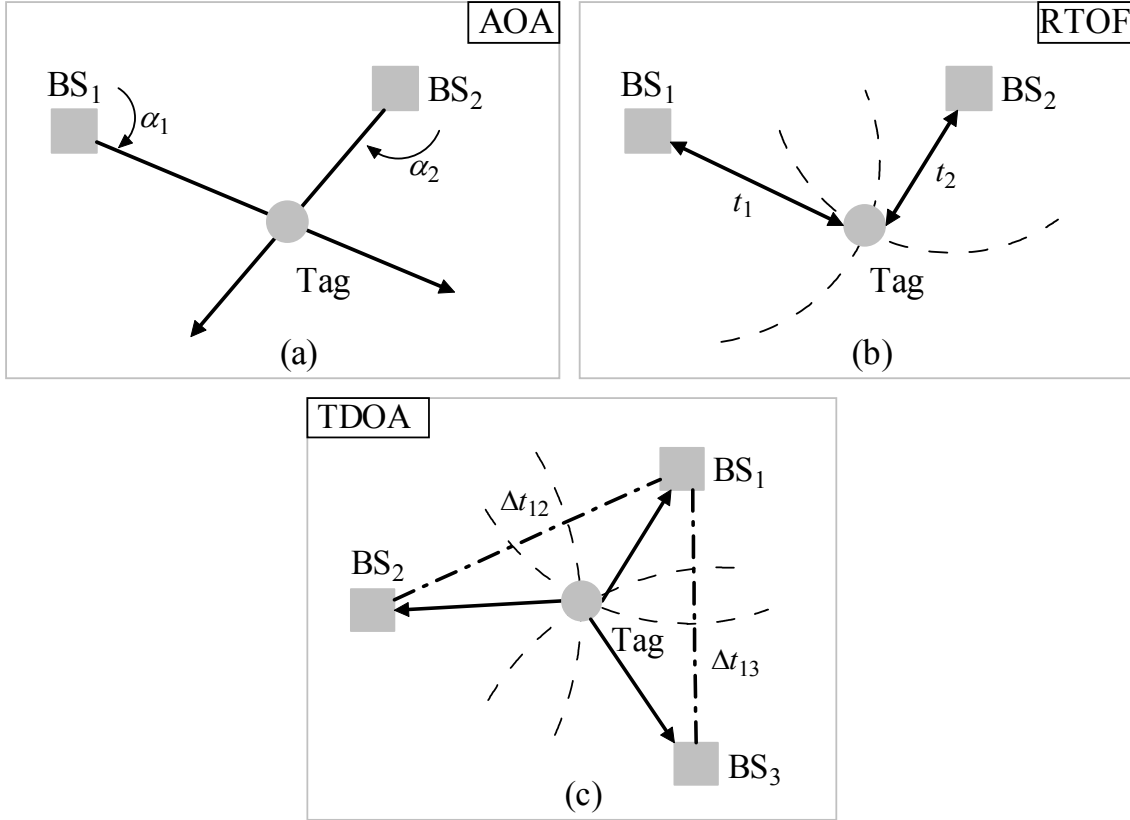


Fig.1.1 Measuring principles: (a) angle-of-arrival AOA, where α_1 and α_2 are the measured direction angles; (b) RTOF, where t_1 and t_2 denote the measured roundtrip signal propagation time, the spatial position is given by the intersection of circles centered at the BSs; (c) TDOA, where Δt_{12} and Δt_{13} denote the measured propagation time difference from a signal traveling from the tag to two different BSs and the position is given by the intersection of hyperbola with foci at the BSs.

As an illustration, Fig 1.2 shows Symeo's positioning system LPR-2D that is used for detection of cranes and vehicles for tracking of goods [18]. The transmitting tags are attached to vehicles and goods, with receivers fixed on the wall. This system uses FMCW techniques to provide 5 cm 2D accuracy up to 400 m distance.

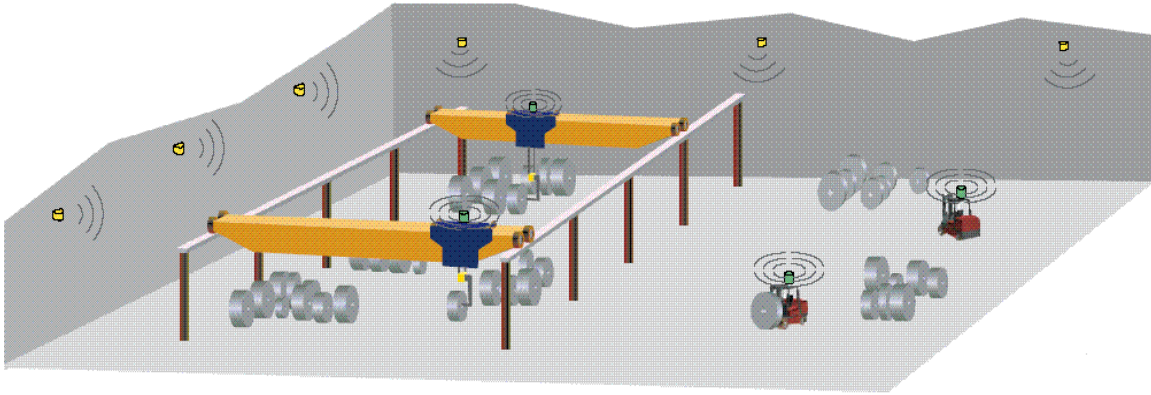


Fig.1.2 Symeo's positioning system LPR-2D

1.2 Challenges in Wireless Localization Systems with Millimeter Accuracy

According to Table 1.1, commercial wireless localization systems have limited ranging accuracy of 10 - 15 cm over a 50 m distance. Certain short-range industrial and medical applications such as dynamic part tracking, structural testing, and computer-assisted therapy require significantly higher accuracy, i.e. mm or even sub-mm range accuracy, than the above mentioned commercial systems. Current technologies used for these applications include infrared (IR), electromagnetic (EM), and ultrasound tracking, which have mm or even sub-mm accuracy. However, IR has a short transmission range and can be easily disturbed by a fluorescent lamp or other light sources in a room. Meanwhile, EM has reduced performance near metal and limited dynamic tracking ability. Finally, ultrasound has limitations due to multipath interference because of its limited bandwidth compared to UWB. Therefore, UWB tracking systems have inherent advantages over these existing technologies since UWB does not suffer from these mentioned drawbacks.

However, there are many challenges in developing such a real-time indoor UWB localization system which has accuracy orders of magnitude better than existing commercial systems. These challenges include multipath interference, sampling rate limitations, non-coherent system synchronization errors, and antenna phase center error. Errors associated with these design issues must be accounted for to develop a system for high accuracy applications. In this work, I will explore various system-level design issues in the context of prototyping a low cost high accuracy UWB localization system for a relatively short range. Through advanced sub-sampling techniques, antenna phase center calibration, and advanced non-coherent system architecture, mm-range accuracy for 3D localization in a real-time system will be demonstrated, however we believe further system optimization can lead to even sub-mm indoor localization accuracy. The developed system requires multiple team members with microwave circuit design and advanced signal processing expertise. Here I will present in detail my work on developing the UWB hardware system; the signal processing area of research has been handled by other members of our team.

1.3 Contributions

The established high accuracy UWB indoor positioning radar involves complex system requirements and hardware development. It follows a top-down approach: starting from the system level and coming up with the specifications for the RF front-end and the UWB antennas, respectively. My corresponding original contributions are listed below with details discussed further in the following chapters.

My major contributions are listed below:

- Development of a low cost reconfigurable pico-second pulse generator with a novel input matching network to suppress the pulse broadening, ringing and echoing;
- Development of a hybrid broadband high speed sub-sampler with reduced conversion loss and spurious level when compared to previously published sub-samplers; the developed low cost sampler can achieve higher than 100 GS/s;
- Design of various UWB transmitting and receiving antennas including the omnidirectional monopole antenna with improved radiation pattern and Vivaldi-rod receiving antenna for improved phase center variation;
- Investigation of carrier leakage suppression scheme using a band notch filter, and its effects on the pulse width and signal time delay spread of the transmitted pulse of the carrier based UWB system;
- Carried out extensive coherent 1D-3D experiments with different number/distribution of base stations, demonstrating mm-range accuracy using the developed UWB system;
- Development of a novel receiver side architecture: carrier based UWB receiver with a sub-sampling system -- combining the traditional energy detection and UWB techniques to address the synchronization problem;
- Development of the 1D and 3D real-time non-coherent systems to achieve mm-range accuracy in a dense multi-path indoor environment under both static and dynamic scenarios.

1.4 Dissertation Organization

Chapter 2 provides an overview of our approach in wireless localization systems using the UWB technology and TDOA approach. The tag and receiver block diagrams of our proposed UWB system are presented. Our basic mathematical equations are given to demonstrate the functionality of our system and the anticipated challenges and error sources.

Chapter 3 describes the development of a low cost tunable narrow pico-pulse generator using a step-recovery diode (SRD), which serves as the source for our UWB localization system. A simple but novel concept of an input-matching network was developed to significantly minimize pulse broadening and suppress pulse ringing and echoing. The pulse generator is adjustable from 300ps to 1ns and produces either monocycle or Gaussian pulses. In Chapter 3, the limitation of the SRD based pulse generator, and other technologies in developing the pulse generator such as GaAs FET, CMOS and non-linear transmission line (NLTL) will be discussed and compared.

Chapter 4 describes the development of a high speed sub-sampling mixer. The developed sampler is integrated with a step-recovery diode strobe-step generator to sub-sample UWB signals. The fabricated sub-sampler demonstrated a wide 3 dB bandwidth of up to 4 GHz and a reduced spurious level of better than -38 dBc. The time domain measurement is comparable with Tektronix TDS8200 digital sampling oscilloscope. Limitations of the designed sub-sampler and ways to further improve its performance will be addressed.

In Chapter 5, a novel elliptical disc monopole antenna with a modified ground plane has been developed. The antenna shows an excellent omni-directional radiation pattern, as well as a satisfactory input impedance match over an ultra-wide bandwidth. In addition, time domain impulse response experiments have demonstrated that the proposed UWB monopole introduces minimal pulse dispersion. Meanwhile, the effect of the receiving Vivaldi antenna phase center variation has been addressed and quantified which is an important factor for localization accuracy. A technique to improve the gain, narrow the H-plane beam-width, and minimize the phase center variations with frequency by utilizing a Vivaldi antenna with a protruded dielectric rod will be introduced.

Chapter 6 describes the system level analysis including link budget, time budget, and power budget.

Chapter 7 focuses on potential methods for local oscillator (LO) leakage rejection in a carrier based UWB system by using a notch filter located before the UWB transmitting antenna. Various filter parameters, such as the filter order and 3 dB rejection bandwidth have been studied to see their effects on providing sufficient band rejection level to reduce the unwanted LO leakage while minimizing the transmitted pulse dispersion.

Chapter 8 describes the experimental results of the high resolution coherent UWB positioning radar system based on time difference of arrival (TDOA). The results show millimeter accuracy in dense multipath indoor environments for 1D, 2D and 3D localization. The accuracy enhancement by increasing the number of base stations (BS) and optimizing the space distribution of these base stations have also been addressed. A brief error analysis has been conducted and will be presented.

In Chapter 9, to solve the synchronization problem for a noncoherent system, a novel receiver architecture is demonstrated by combining carrier based UWB system and the traditional energy detection techniques. Both simulation and measurement show a reduced system jitter error.

In Chapter 10, extensive noncoherent 1-D and 3-D localization experiments are performed where mm-range accuracy has been consistently achieved, in both dynamic and static modes, validating the theories of the novel UWB receiver architecture previously presented in Chapter 8. Sources of errors will be discussed in detail with suggestions to further improve the current system performance.

Chapter 11 finalizes this work and outlines the relevant contributions of this work.

Chapter 2

Localization System Overview

In recent years, an exponential growth of wireless localization systems has been observed using either FMCW or UWB techniques. Wireless technology has entered the realm of consumer applications as well as industrial and medical logistics applications, along with many other applications. Table 2.1 lists the recent localization systems including their performance and the type of technology used.

Table 2.1 – Recent Research Work in Wireless Localization Systems

	Tech.	Method	Freq. Range	Typ. Accuracy	Coherent/ Noncoherent	Environment /coverage
Schroeder et al.[19]	UWB	TDOA	N/A	58.1 cm (3D)	Noncoherent	Indoor
Krishnan et al. [20]		TDOA	3.1 – 5.1 GHz	15 cm (2D)	Noncoherent	Indoor (8 m)
Fujii et al. [21]		TDOA	3.1 – 5.1 GHz	10 cm (2D)	Noncoherent	Indoor (25 m)
Kitamura et al. [22]		TDOA	1- 2.7 GHz	20 – 70 cm (2D)	Noncoherent	Anechoic Chamber
Meier et al. [14]		N/A	22.58- 25.7 GHz	2 mm (1 D)	Coherent	Indoor (10 m)
Low et al. [12]		N/A	N/A	2 cm (1D)	Coherent	Indoor (8m)
Zetik et al. [13]		TDOA	DC-5 GHz	1.5 cm (2D)	Noncoherent	Indoor (4m)
Stelzer et al. [4]	FMCW	TDOA	5.8 GHz	10 cm (2D)	Noncoherent	Outdoor (500 m)
Wiebking et al. [8]		TDOA	5.8 GHz	20 cm (2D)	Noncoherent	Indoor (25 m)
Roehr et al. [9]		RTOF	5.8 GHz	4-5 cm (1D)	Noncoherent	Outdoor (200 m)

Based on Table 2.1, some useful results can be directly summarized:

- ◆ TDOA method was used in almost all 2D/3D research work;
- ◆ FMCW technique was applied in most of the outdoor applications whereas UWB technique was applied in most of the indoor applications;
- ◆ The reported accuracy listed in Table 2.1 is insensitive to the range coverage for both FMCW and UWB techniques.

The next question to be answered is which technique and method should we use in our precise indoor localization system and why. Table 2.2 compares FMCW and UWB techniques.

Table 2.2 – Comparison between FMCW and UWB Systems [23]

	FMCW	UWB
Measurement Feature	Frequency Domain	Time Domain
Indoor Application	Acceptable	Excellent
Outdoor Application	Excellent	Excellent
Achievable Accuracy	Outdoor: few cms Indoor: 20 cm	Sub-cm indoor and outdoor
Multipath Suppression	Poor	Excellent
Tracking Moving Target	Excellent	Acceptable
System Complexity	Complex	Simple
Power Consumption	High	Low

Based on the previous research works listed in Table 2.1 and the performance comparison in Table 2.2, UWB was chosen for our localization system because:

- ◆ UWB signals can provide a realistic way for sensing extremely small time differences, which in turn can be used for high-resolution ranging applications (mm-range or better);

- ◆ It offers time domain methods for gating out undesired multipath signals, ideal for indoor applications;
- ◆ UWB signals utilize narrow pulses with low duty cycles, reducing the need for high average power drives relative to FMCW radars [11].

However, with the current severe power restrictions of the FCC UWB regulation, mainly short range solutions are feasible. In this case, FMCW could be a better candidate for outdoor long range applications.

The measuring principles of AOA, RTOF and TDOA have been described in Chapter 1. Table 2.3 compares the performance of each method. As can be seen, AOA is not considered as a good localization scheme due to its unacceptable multipath rejection capability, complex tag design, and limited achievable accuracy. RTOF has been used in FMCW localization systems and provides a simple way of synchronization based on a reflected coherent approach [1]. However, when applied for an UWB localization system, it requires a very complex tag design, i.e. the tag has to be an UWB transceiver. Thus, it would be a great challenge and almost impossible to realize the tag with a low cost and compact size. Thus, TDOA method is the best candidate for the UWB localization approach and has been widely used in both commercial systems and research works as listed in Table 1.1 and Table 2.1. The biggest challenge in achieving high accuracy using the TDOA method is to solve the synchronization between the transmitter and receiver, which will be detailed in Chapter 9.

Table 2.3 – Comparison between Localization Methods

	AOA	RTOF	TDOA
Category	Energy Based	Time Based	Time Based
Multipath Rejection	Low	High	High
Tx-Rx Clock Synchronization	-	Not Required, reflected coherent	Required
Tag Complexity	Complex, require knowledge of direction	Complex, required to respond received UWB signal	Simple
Position Accuracy	Low	High	High

The goals and requirements in developing our wireless localization system are given in Table 2.4. The challenges in realizing such a system are:

- ◆ Narrow pulse signal generation;
- ◆ Receiver side UWB signal acquisition;
- ◆ Transmitter – receiver clock and carrier synchronization;
- ◆ Multi-tag communications;
- ◆ Millimeter range accuracy, or even sub-mm accuracy. This is orders of magnitude better than current commercially available UWB localization systems.

All of the above challenges have been addressed throughout this thesis. In our current localization system, many of the specs in Table 2.4 have been met except for tag integration using monolithic microwave integrated circuit (MMIC) and multi-tag communication.

Table 2.4 - Localization System Design Requirements

Environment	Indoor
Technology	UWB
Localization Method	TDOA
Range	5-10 m
Accuracy (Static and Dynamic)	mm-range*
Pulse Repetition Frequency (PRF)	10 MHz – 30 MHz
Frequency Range	6 – 10 GHz
Number of tags	10 – 100
Tag Power	Battery
Tag Integration	MMIC

* methods to achieve even sub-mm accuracy will be recommended

2.1 GPS-like System Scheme

The Global Positioning System (GPS) is a Global Navigation Satellite System (GNSS) developed by the United States Department of Defense. It uses a constellation of between 24 and 32 Medium Earth Orbit satellites that transmit precise microwave signals that enable GPS receivers to determine their current location, the time, and their velocity (including direction) [24]. In our approach, a GPS-like scheme is utilized along with time difference of arrival (TDOA) to locate 2-D and 3-D transmitting tag positions in an indoor environment, as shown in Fig. 2.1. In this section we will first describe the operation concept while the associated TDOA algorithm will be described at the end of this chapter in Section 2.4. Here we will use UWB signals to develop a scheme for precise localization, given that the average output power spectral density for indoor systems has an upper bound of -41.3 dBm/MHz [25] as specified by the FCC regulations. One typical UWB localization method is the use of Impulse Radio (IR) UWB, where a

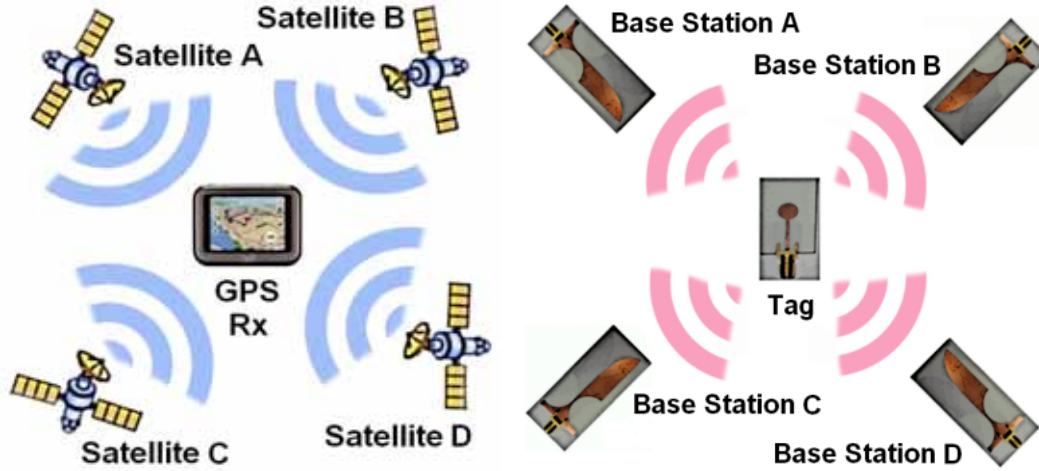


Fig.2.1 GPS system analogy: GPS system (left) and UWB positioning system (right).

baseband UWB pulse is transmitted by an UWB antenna while its level must conform to FCC regulations [12]. However, with IR-UWB systems, the carrier-less received signals are noisy due to the complex transmitted waveform and the added multipath signals, which make it difficult to accurately locate the position of the received line of sight (LOS) signals.

In our system, we modulate an UWB pulse with an 8 GHz carrier signal which resides at the upper end of the 3.1 – 10.6 GHz band. The use of this band reduces the size of the wideband RF components in the transmitter and receiver and also bypasses many of the interfering frequency bands that exist at the lower end of the 3.1 – 10.6 GHz band. A complete experimental setup of the developed system is shown in Fig. 2.2. In this developed system, we transmit a modulated narrow Gaussian pulse with a carrier frequency and demodulate it at the receiver side. The source of our UWB positioning system is a step-recovery diode (SRD) based pulse generator with a controlled pulse

width and a bandwidth greater than 1 GHz. We use a 300ps pulse that has produced greater than 3 GHz bandwidth signal in our implementation, as shown in Fig. 2.3. A detailed discussion about the development of this pulse generator can be found in Chapter 3, and the effect of the pulse width in our measurements will be discussed in Chapter 10. The modulated Gaussian pulse is then transmitted through an omni-directional UWB antenna. Multiple base stations are located at distinct positions in an indoor environment to receive the modulated pulse signal. The received double sideband (DSB) modulated Gaussian pulse at each base station first goes through a directional Vivaldi receiving antenna and then is amplified through a low noise amplifier (LNA). Next, through demodulation, we combine the upper and lower bands and get rid of the carrier to obtain I/Q signals. After going through a low pass filter (LPF) with a passband of DC-5 GHz to suppress the 8 GHz carrier leakage signal, the I/Q signals are sub-sampled using an UWB sub-sampling mixer (an equivalent time sampler), extending them to a larger time scale (i.e. μs range) while maintaining the same pulse shape. The sub-sampling mixer uses extended time techniques to achieve equivalent sampling rates in excess of 100 GS/s, which yields mm-range sample spacing and provides our peak detection algorithm with ample data. A detailed sub-sampler design can be found in Chapter 4. Finally, the extended I/Q signals are processed by a conventional low cost and low speed analog to digital converter (ADC) and a standard Field Programmable Gate Array (FPGA) unit.

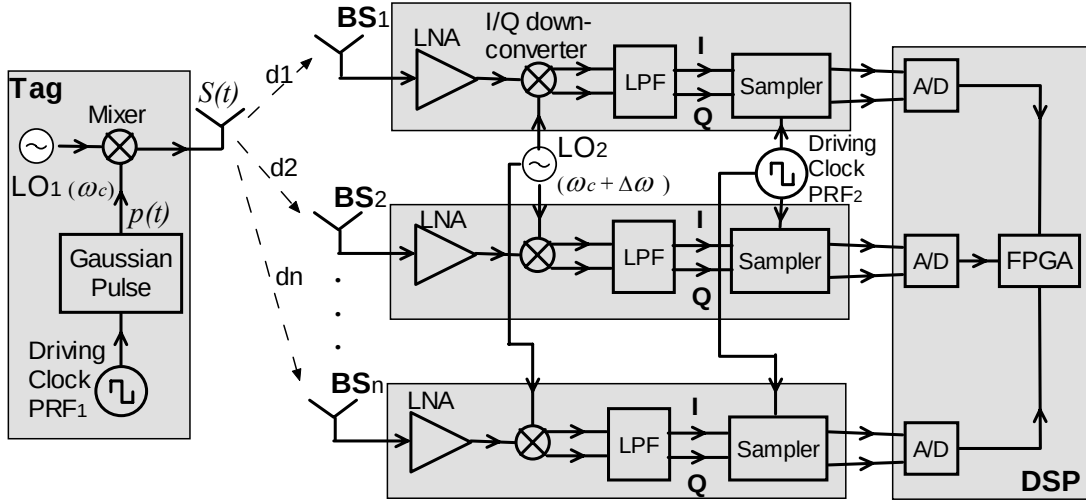


Fig.2.2 Block diagram of indoor localization system showing one tag and three base stations which feed into the main system controller.

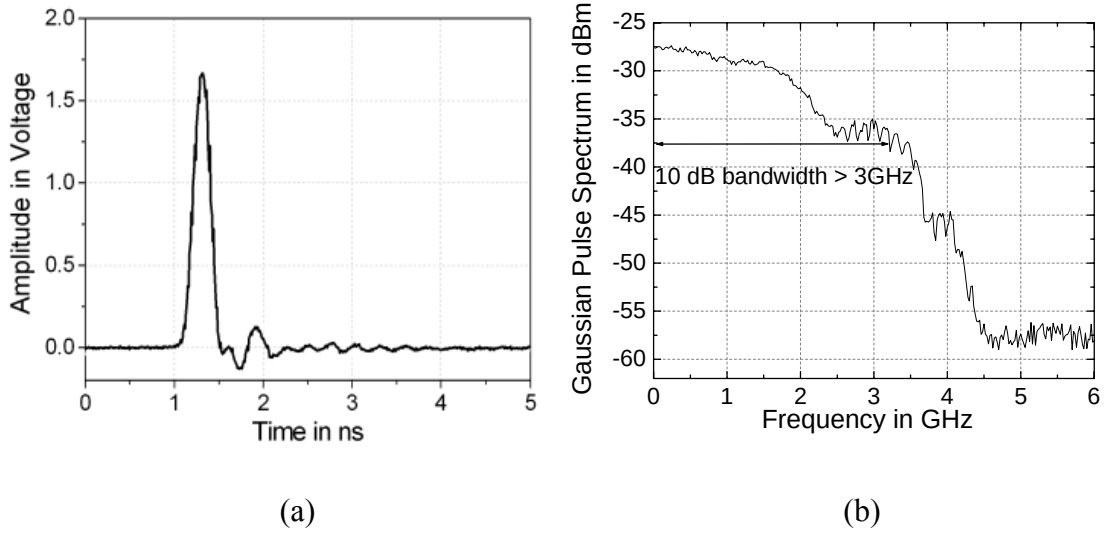


Fig.2.3 Gaussian pulse which serves as the system UWB source: (a) time domain exhibiting 300 ps pulse width (measured at 10% of the peak value), (b) frequency domain highlighting bandwidth in excess of 3 GHz.

The transmitting tag uses an elliptical monopole structure combined with a modified ground plane which provides an omni-directional radiation pattern [26], details of its design will be given in Chapter 5. The base stations, however, use a single element directional Vivaldi antenna with a flared-antipodal design which has demonstrated high gain and constant beamwidth over a wide band [27, 28], and again its design will be given in Chapter 5.

The output power spectral density of the transmitted signal in our system has been measured and plotted in Fig. 2.4. The modulated pulse signal has a 10 dB bandwidth of approximately 6 GHz, exceeding the 500 MHz minimum bandwidth required under the FCC rules governing UWB communication. Its average output power spectral density satisfies the FCC indoor limit by a margin of more than 3 dB for a majority of the useable bandwidth. The 8 GHz carrier signal leaks through the mixer and is shown as a peak at -16 dBm in Fig.2.4. This leakage could be suppressed by adding a band-notched filter or utilizing a band-notched monopole; however, this could distort the narrow pulse signal [29]. Therefore, the pulse distortions due to the notched filter will be separately addressed in detail in Chapter 7.

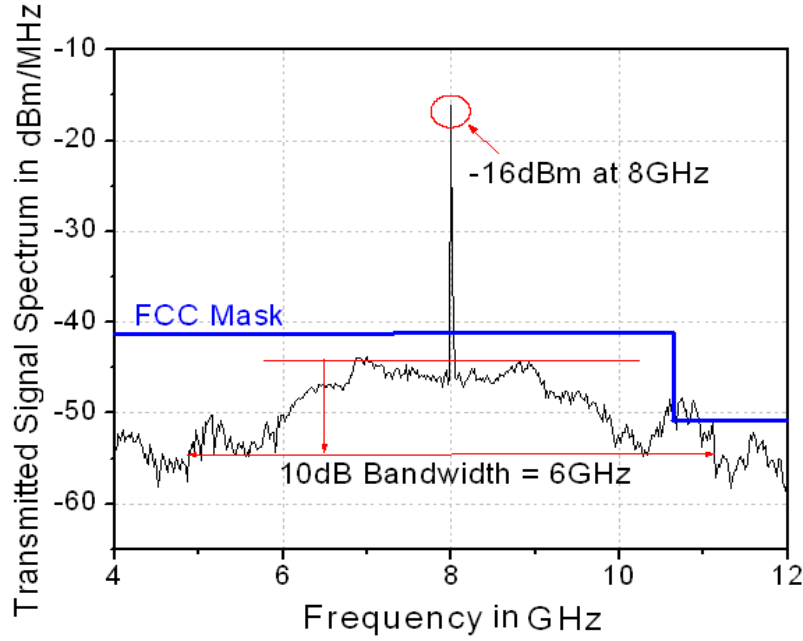


Fig.2.4 Power spectral density of modulated pulse signal showing double sideband modulated signal with a bandwidth of 6 GHz and carrier leakage at 8 GHz of -16 dBm.

2.2 Carrier Synchronization with I/Q Down Conversion

Carrier frequency synchronization between the transmitting tag and the receivers can *theoretically* be solved by direct I/Q down-conversion, however its practical implementation could be hindered by the effect of phase noise as will be explained in Chapter 9 in detail. The transmitted signal $s(t)$ from the tag is given by

$$s(t) = \sin(\omega_c t)(p(t) + K) \quad (2.1)$$

where $p(t)$ is the Gaussian pulse signal, K is the carrier signal leakage factor, and ω_c is

the carrier frequency generated by LO_1 . The received I and Q signals before passing through the low pass filter are given by

$$\begin{aligned} I &= [\sin(\omega_c t) \cdot (p(t) + K)] \cdot \sin(\omega_c t + \Delta\omega t) \\ &= \frac{1}{2} [p(t) + K] \{ \sin(2\omega_c t) \sin(\Delta\omega t) + [1 - \cos(2\omega_c t)] \cos(\Delta\omega t) \} \end{aligned} \quad (2.2)$$

$$\begin{aligned} Q &= [\sin(\omega_c t) (p(t) + K)] \cdot \cos(\omega_c t + \Delta\omega t) \\ &= \frac{1}{2} [p(t) + K] \{ \sin(2\omega_c t) \cos(\Delta\omega t) - [1 - \cos(2\omega_c t)] \sin(\Delta\omega t) \} \end{aligned} \quad (2.3)$$

where $\Delta\omega$ is a small offset frequency of LO_2 relative to the carrier ω_c generated by LO_1 from the tag. After passing through the LPF with a passband of DC-5 GHz, which suppresses the 8 GHz carrier leakage signal, the I and Q signals become

$$I = \frac{1}{2} [p(t) + K] \cos(\Delta\omega t) \quad (2.4)$$

$$Q = \frac{1}{2} [p(t) + K] \sin(\Delta\omega t) \quad (2.5)$$

Finally, the filtered I and Q data are sub-sampled and AC coupled, which are given by

$$I_{ex} = \frac{1}{2} P_{ex}(t) \cos(\Delta\omega_{eq} t) \quad (2.6)$$

$$Q_{ex} = \frac{1}{2} P_{ex}(t) \sin(\Delta\omega_{eq} t) \quad (2.7)$$

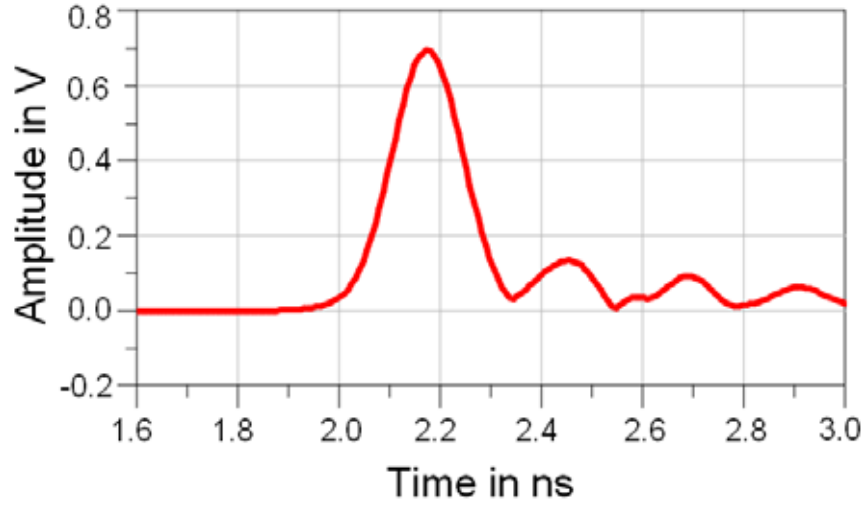
where $P_{ex}(t)$ is the pulse signal after time extension while maintaining the same pulse shape [30], and $\Delta\omega_{eq}$ is the equivalent offset frequency after sub-sampling, which can be expressed as

$$\Delta\omega_{eq} = |\Delta\omega - N \cdot PRF_2|, \quad \Delta\omega_{eq} \leq \frac{PRF_2}{2} \quad (2.8)$$

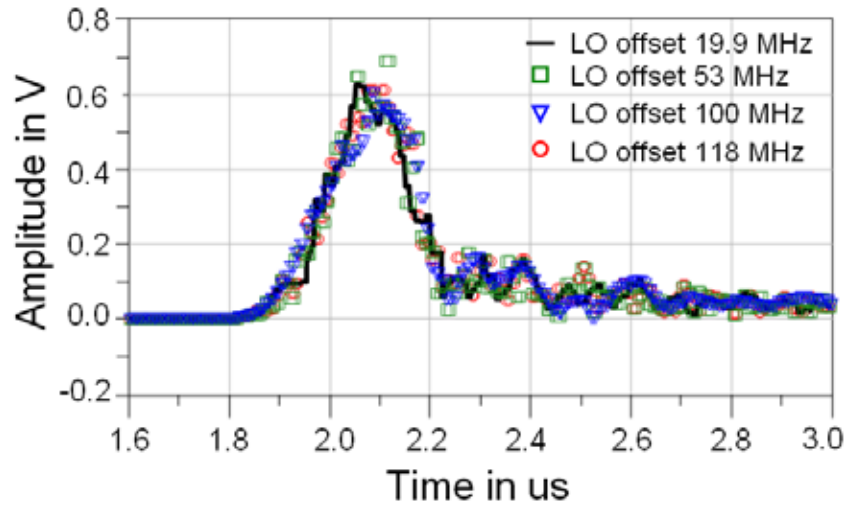
where N is an integral. The extended I_{ex} and Q_{ex} signals are then processed by FPGA circuitries, and the reconstructed received signal is given by

$$\bar{P}(t) = \sqrt{I_{ex}^2 + Q_{ex}^2} = \frac{1}{2} P_{ex}(t) \quad (2.9)$$

From (2.9), the recovered signal $\bar{P}(t)$ is not affected by the offset carrier frequency $\Delta\omega$ and contains the same information as the transmitted Gaussian pulse signal $p(t)$. To validate the above analysis and study how the offset frequency $\Delta\omega$ between the tag and base stations would affect the signal performance, a system level simulation using Agilent ADS2006A has been carried out. The simulation in Fig. 2.5 was conducted with arbitrarily picked four offset frequencies, 19.9 MHz, 53 MHz, 100 MHz and 118 MHz. As shown in Fig. 2.5, the original pulse has been successfully reconstructed with the four LO offset frequency conditions, which theoretically validates that the system performance is insensitive to the offset frequency $\Delta\omega$ between the tag and base stations with the help of I/Q demodulation.



(a)



(b)

Fig.2.5 ADS simulation showing the effect of Tx-Rx LO frequency offset on reconstructed sub-sampled pulse: a) original signal, b) reconstructed signal with LO offsets of 19.9 MHz, 53 MHz, 100 MHz and 118 MHz.

However, the above results shown in equation (2.9) and Fig. 2.5b are based on two assumptions: 1) there is no phase difference between the I and Q channels; 2) the phase

noise of both the tag and base station carrier are neglected, leading to a fixed offset carrier frequency $\Delta\omega$ without variation with time and temperature. However, practically those two effects will cause jitter and systematic error and thus must be taken into account. Error analysis and impact of the noise of such a scheme will be fully studied in detail in Chapter 9. An alternative receiver scheme using a single channel approach instead of I/Q down conversion will be fully explored in Chapter 9.

2.3 Multi-tag Control

Figure 2.6 shows the current layout of the tag. A microcontroller is currently used to implement a Time Division Multiple Access (TDMA) scheme for communicating with multiple tags. TDMA is the preferred multiple access scheme in current UWB commercial systems [31-32]. The modulation scheme used is On-Off Keying (OOK) with the 8 GHz carrier signal, although pulse modulation could be used to increase system dynamic range. A unique ID is stored on each tag. Each tag is in a low power or sleep state until activated by the main control station. The control station calls each tag in a round robin fashion. Figure 2.7 shows the prototype hardware. The measured results with OOK running at 10 kbps are shown in Fig. 2.8. At this stage of research, our system only implemented with one tag, but in principle it can be extended to a multi-tag system. Time and power budget for this scenario will be given in Chapter 6.

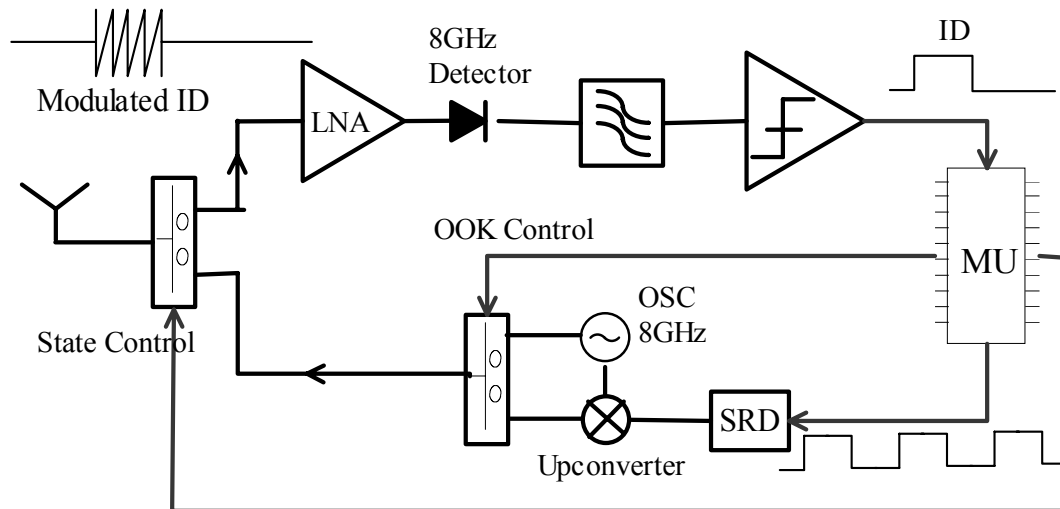


Fig.2.6 Block diagram of current tag layout showing OOK digital communication and UWB transmitting architecture.

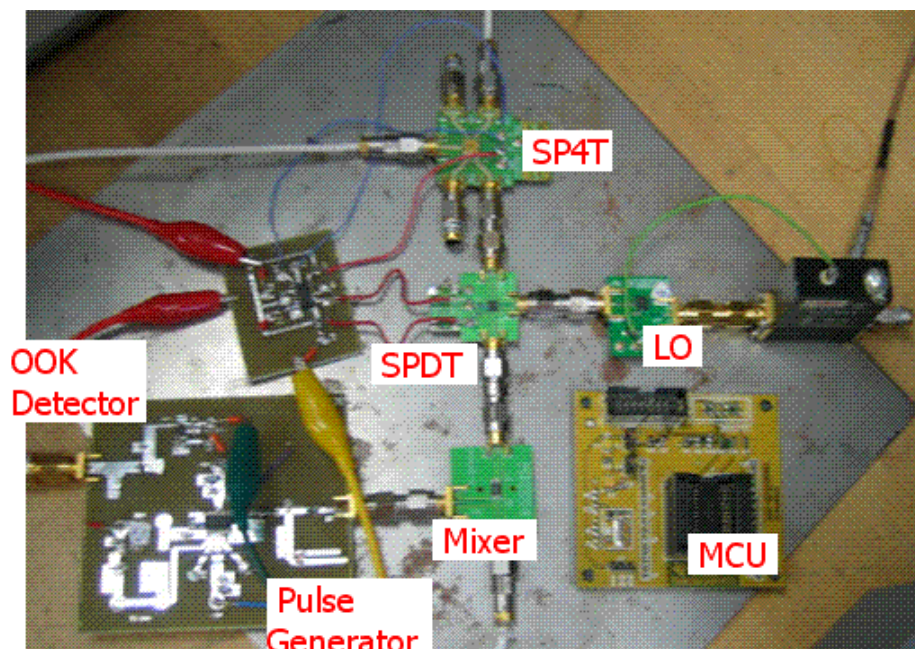


Fig.2.7 Prototype of tag with OOK digital communication and UWB transmitting architecture.

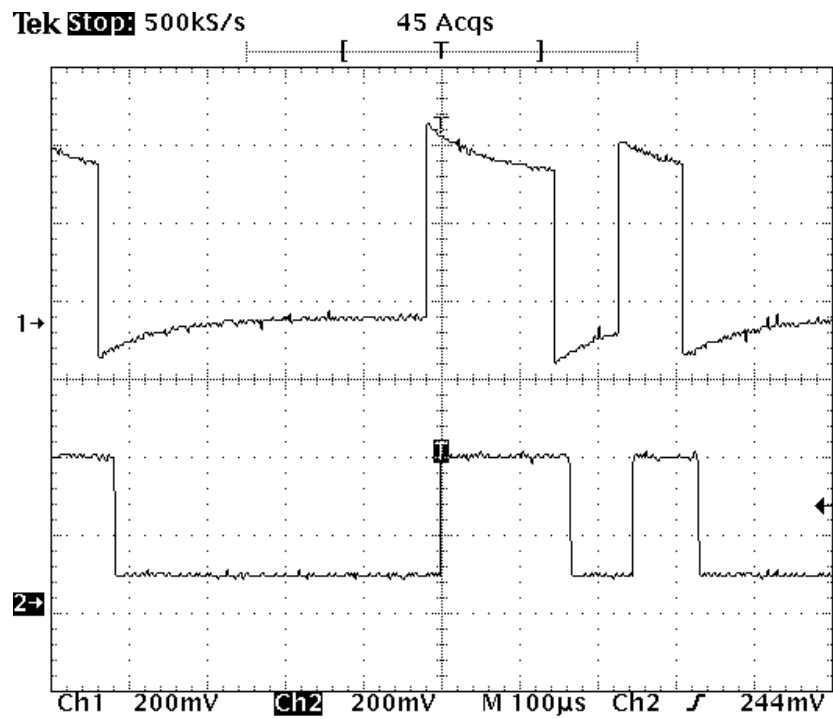


Fig.2.8 Measured OOK running at 10 kbps – top: recovered OOK signal; bottom: transmitted OOK signal.

Chapter 3

Tunable Picosecond Pulse Generator

A pulse generator is an essential component of UWB systems and generally has a fixed pulse width and shape. A fixed pulse width is adequate for localization and a narrower pulse width would be required to achieve even higher resolution. However, pulse generators capable of tuning the pulse duration and pulse shape electronically provide more flexibility and improved performance in other UWB systems. For example, in UWB see through wall radar an electronically tunable pulse generator allows the pulse width to be changed to achieve varying penetration abilities and resolutions. A wider pulse in the time domain has larger energy, which can be used to penetrate concrete walls that have higher attenuation. Shorter pulse features better ranging resolution, δR , and less energy is needed to penetrate dry wall, which has less attenuation than concrete due to its wider bandwidth (BW) as given by $\delta R = c / 2BW$ [33]. For localization applications, system performance can be significantly enhanced if extremely narrow pulses are utilized. An experiment demonstrating how pulse width affects the 3-D localization accuracy will be given in Chapter 10. In addition, Pulse Shape Modulation (PSM) can be achieved and utilized to encode data in logic states, which is widely used in communication systems.

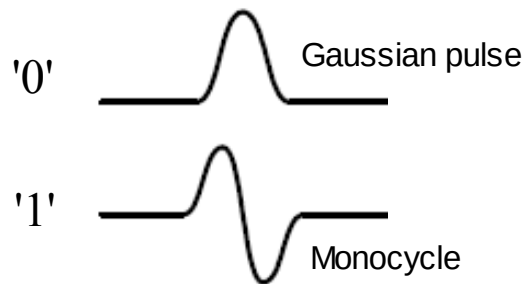


Fig.3. 1 Gaussian pulse represents data '0' and monocycle represents data '1'.

Table 3.1– Different Pulses and Their Applications

Pulses	Example of Applications
300 ps Gaussian	Localization
1 ns Gaussian	See Through Wall
Monocycle/Polycycle	UWB Communication, such as pulse shape modulation (PSM)

Table 3.1 summarizes some different pulses and their typical applications. In this chapter, a low cost pulse generator with tenability in both pulse width and shape will be described and the designed pulse generator has been successfully implemented in a high accuracy localization system as well as a see through wall radar system [34].

3.1 State of the Art Pulse Sources

There are various options to achieve the required pulse width. Table 3.2 compares current state of the art pulse sources. Generally step recover diodes (SRD) are suitable for a pulse with a duration of a few hundred picosecond, but for higher performance with much narrower pulse width, higher PRF and more compact size the GaAs Monolithic Microwave Integrated Circuits (MMIC) technology with short gate length and large current drive capability should be used. For example, Kawano et al. achieved less than 20 ps pulse width using 0.13 μm InP-HEMT technology with an f_T of 183 GHz [35]. Hafdallah et al. achieved around 80 ps pulse width using 0.3 μm GaAs MESFET and approximately 40 ps pulse width by using a 0.2 μm AlGaAs/InGaAs/GaAs HEMT process [36]. Seehausen described a sine wave triggered GaAs pulse generator based on delay line coupled NOR gates, demonstrating less than 100 ps duration and more than 5 GHz PRF [37]. Recently, there have been many researches using CMOS to generate

UWB pulses [38, 39]. However, the designs have provided wider pulse width and limited pulse amplitude.

In the proposed implementation for use in an UWB localization radar system SRD devices are utilized as they provide the following features:

- ◆ an affordable and quick turn around solution
- ◆ an adequate pico-second range pulse width
- ◆ relatively fast PRF
- ◆ sufficient output pulse voltage

In future studies, in order to achieve sub-mm accuracy one might need to investigate the pulse sources with more narrow durations using advanced GaAs technology.

Table 3.2 – State of the Art Pulse Sources

Technology	Pulse Width (10% peak)	PRF	Output Amplitude	Size
SRD	300 ps	Up to 40 MHz	> 1 V	Compact
Drift SRD [40]	~ 1 ns	KHz range	> 1 kV	Bulky
(Bi)CMOS [39]	~ 300 ps	MHz range	< 1 V	Ultra compact
GaAs FET [35]	20 ps	GHz range	> 1 V	Ultra compact
GaAs NLTL [41]	50 ps	GHz range	Depends on input signal	Bulky (Require extra circuits)

3.2 SRD Operation Concept

A comparison between an ideal rectifier and SRD device behavior is illustrated in Fig.3.2 [42]. The ideal PN diode rectifier does not store charge when driven into forward conduction. Consequently, when the terminal voltage reverses the diode voltage follows. Therefore, it is relatively straightforward to find the nature of the resulting diode voltage waveform. The ideal step recovery diode, on the other hand, stores charge which must be removed by negative current before the diode can follow the input voltage. By the time this charge has been removed, the generator voltage has already become quite negative. As a result, the terminal voltage on the SRD jumps to the negative generator voltage at a speed determined by two factors:

- 1) the RC time constant, where R is the parallel combination of the generator resistance and load resistance and C is the reverse bias capacitance of the diode
- 2) the transition speed of the diode

The physics of SRD devices are important since they serve as a guide in choosing the correct device that can yield the minimal pulse width. Selecting a SRD device with a small parasitic resistance and capacitance and short transition time is the first design step.

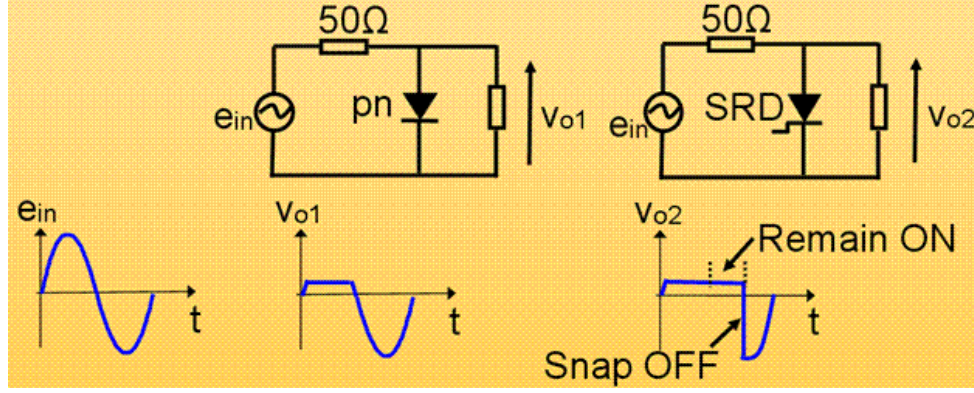


Fig.3. 2 Comparison of ideal PN diode rectifier and SRD behavior.

The next step is to perform a theoretical analysis of the operation of the SRD device in the time domain to improve the quality of the narrow pulse output signal. A step by step model of the output pulse formation is given in detail to demonstrate the source of distortion. The SRD device has low impedance in the ON state and much higher impedance in the OFF state [42]. After fast transition to the OFF state, a step signal is generated and would propagate towards both the positive x -axis (step 'A' in Fig. 3.3) and in the direction of the short-circuited stub. The step signal traveling along the short-circuited stub arrives at the end and is completely reflected back out of phase (i.e. inverted) and is shown as step 'B' in Fig. 3.3. Finally, step signal 'A' and step signal 'B' combine at $x=0$ to produce a Gaussian pulse with a width corresponding to the round trip delay along the short-circuited stub [43].

However, the previous description omits leakage caused by both the package parasitic capacitance, C_p , and the reverse depletion capacitance, C_r , of the SRD device during the OFF state as shown in Fig. 3.4. The reflected step from the short-circuited stub

will not only travel in the $+x$ -axis direction, but also travel in the $-x$ -axis direction across the SRD device as a leakage step. When this leakage step meets the triggering voltage source, V_s , it reflects back again due to the source mismatch as represented by step 'C' in Fig.3.3. The net result of the three combined step-waves is a distorted and broadened Gaussian pulse as demonstrated in Fig. 3.5. Hence, it is desirable to get rid of or reduce the effect of wave C.

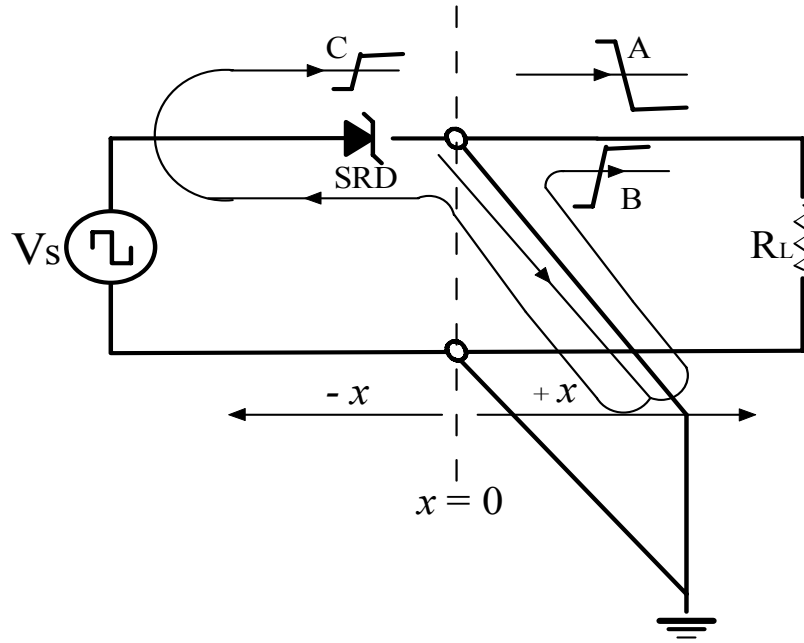


Fig.3. 3 SRD based Gaussian pulse shaping circuitry.

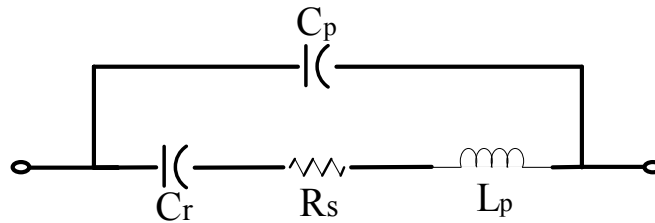


Fig.3. 4 Equivalent circuit of SRD under OFF state.

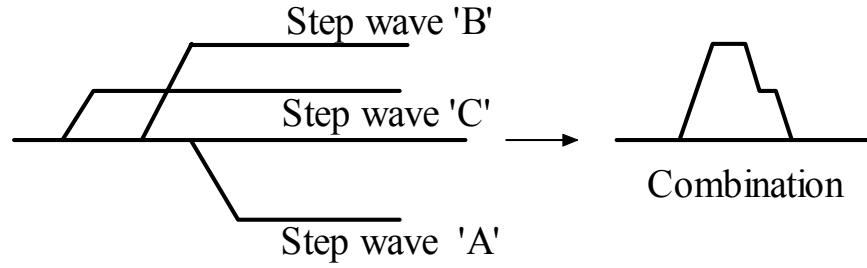


Fig.3. 5 tep-waves combination at the interface of $x = 0$.

3.3 Novel Input Matching Network

An input-matching network for the SRD pulse generator has been developed in order to bypass or suppress the leakage step wave. This eliminates Gaussian pulse distortion. The developed matching network is basically an RC low-pass filter which allows only the triggering signal (10MHz) to pass and bypasses the leaked fast step signal. Simulation was carried out using the Agilent ADS2003C transient simulator as shown in Fig. 3.6. The SRD spice model can be found from Aeroflex/Metelics Inc. application note [44], and is shown in Fig. 3.7 in ADS environment.

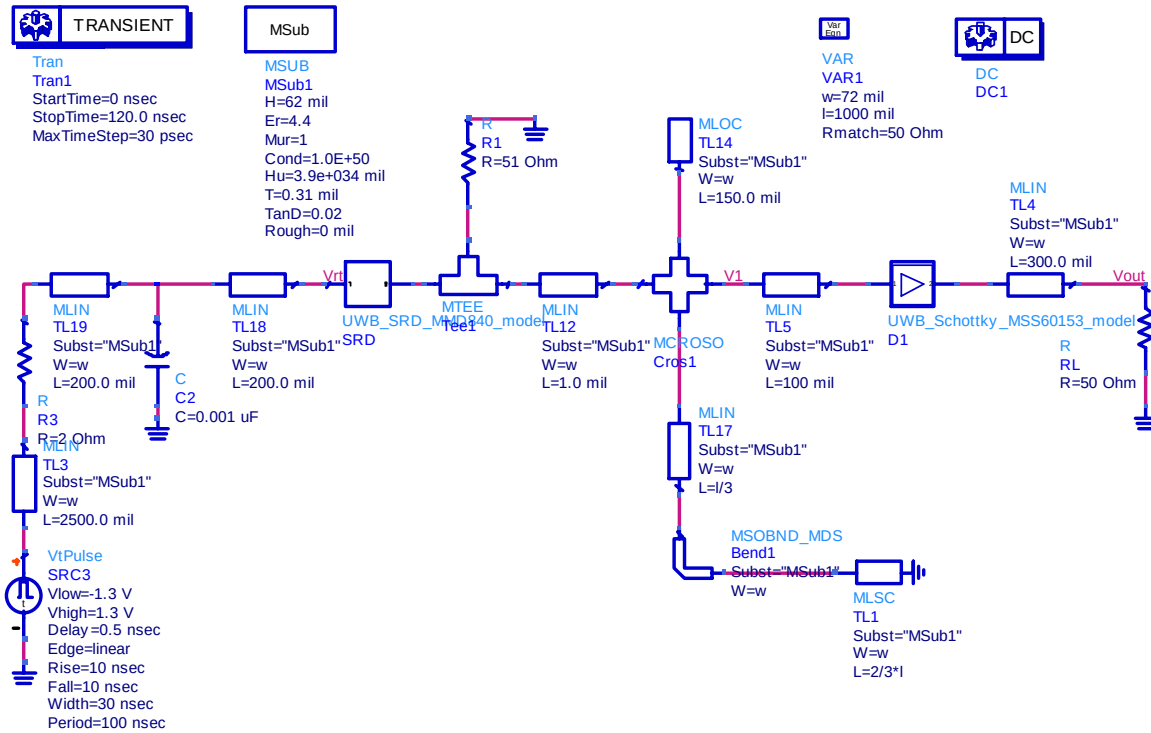


Fig.3. 6 Complete SRD transient simulation platform using ADS.

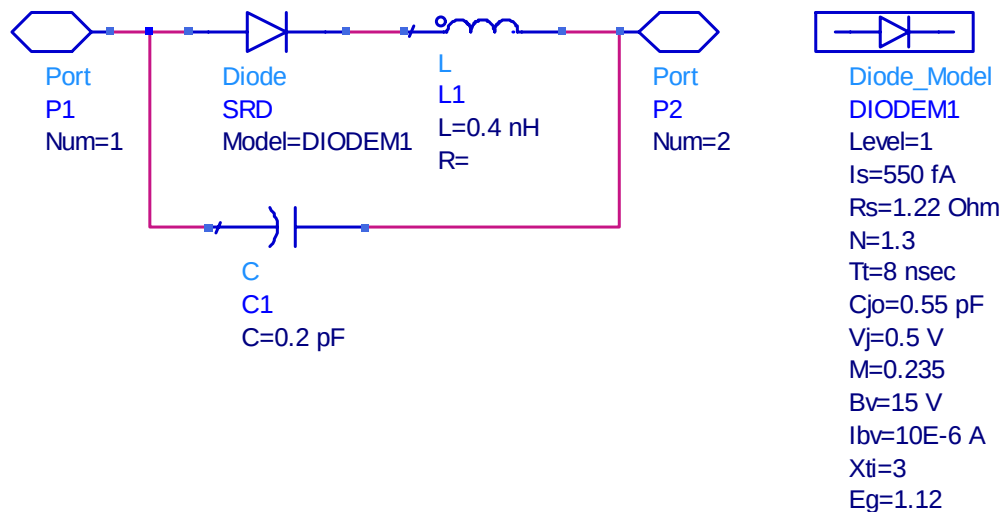


Fig.3.7 SRD spice model.

Other methods exist to model the SRD device such as the Hamilton's classical model (Fig. 3.8) [42]. Hamilton's model represents the nonlinearity as two abruptly switched capacitors with small reverse capacitance, C_r , in the pF range and relatively large forward polarization capacitance, C_f . Such representation of the switching process causes convergence problems in harmonic balance simulators. Zhang and Raisanen proposed a model [45] that includes a voltage ramp occurring during the transition process to overcome the convergence problems. On the basis of their I-V DC measurements of the diode, the capacitances can be determined and included in the modeling process. This new model introduces a parabolic function in place of the discontinuity represented by the switch. This parabolic function relaxes computational requirements while also making the whole model closer to a physical diode. A successful implementation of Zhang's SRD model can be found in [46]. In our analysis the previously described SRD Spice model has been chosen as part of the Agilent ADS simulation.

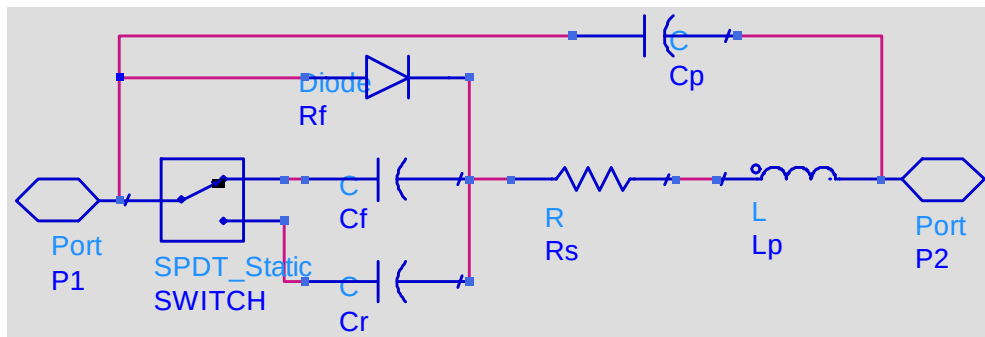


Fig.3.8 Classical SRD switching model.

To measure the generated pulses a Tektronix TDS8200 sampling oscilloscope has been used. Fig 3.9 shows both the ADS simulated and measured results before and after introducing the input-matching network. The pulse width is much wider and has severe distortion without the use of the newly developed input-matching network. After introducing the input-matching network the output pulse has much narrower duration and the pulse shape becomes more symmetric. As shown in Fig. 3.10, if the transmission line length between the trigger source and SRD is long the output signal may have severe ringing and echoing without the input-matching network. After introducing the input-matching network the ringing and echoing is suppressed.

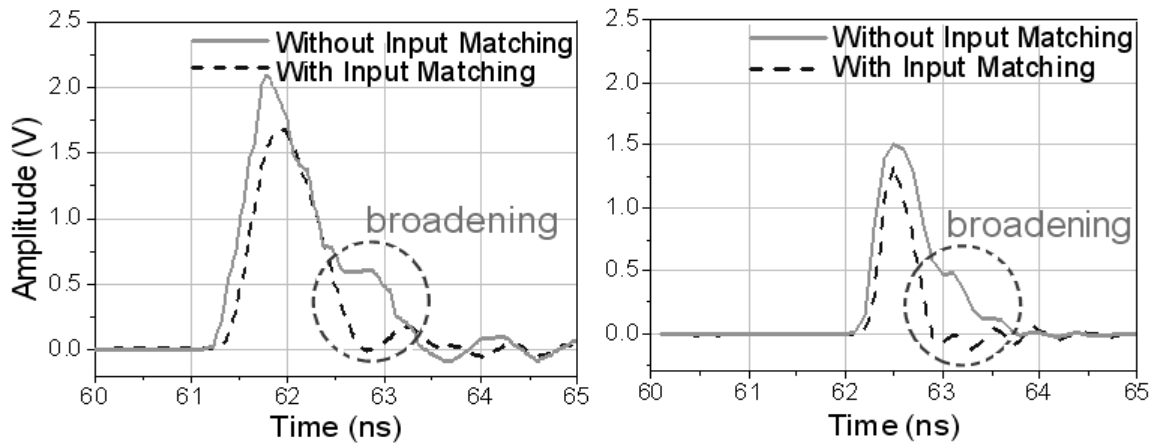


Fig.3.9 Suppressed pulse broadening: simulated (left) and measured (right) Gaussian pulse output with and without input matching network.

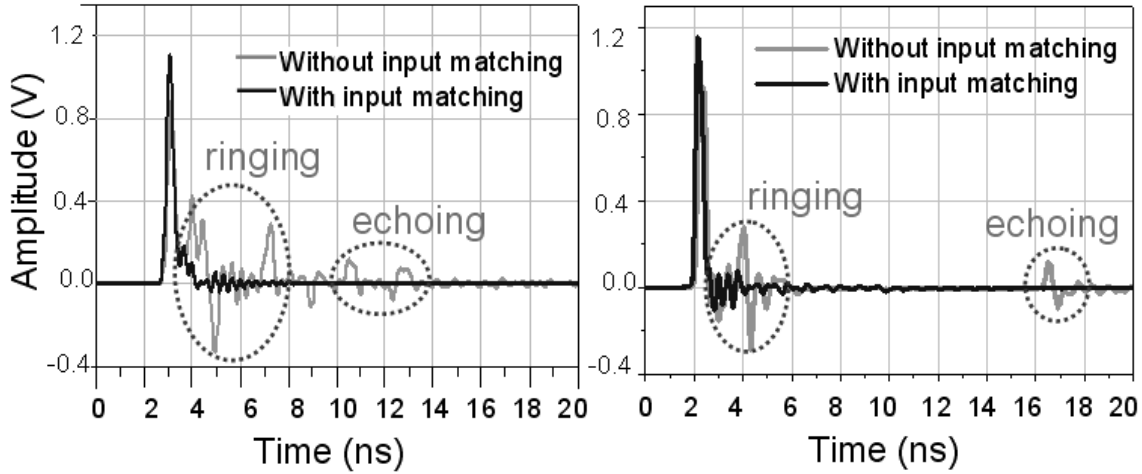


Fig.3.10 upressed ringing and echoing: simulated (left) and measured (right) Gaussian pulse output with and without input matching network.

3.4 Gaussian Pulse with a Tunable Duration

Recently J. Han et al [47] demonstrated a pulse generator with a tunable duration. However, they used a parallel-connected PIN diode structure, as shown in Fig. 3.11, which is only capable of producing a Gaussian pulse shape output signal.

Presented is an alternative circuit topology that provides more flexibility. The PIN diodes are connected in series to the different stubs, each with a distinct length, rather than in parallel, as shown in figure 3.12. Thus, only half the number of PIN diodes is required as compared with [47]. For a given pulse duration selection, only one stub is connected at a time as the other stubs are completely disconnected by turning their associated PIN diodes off.

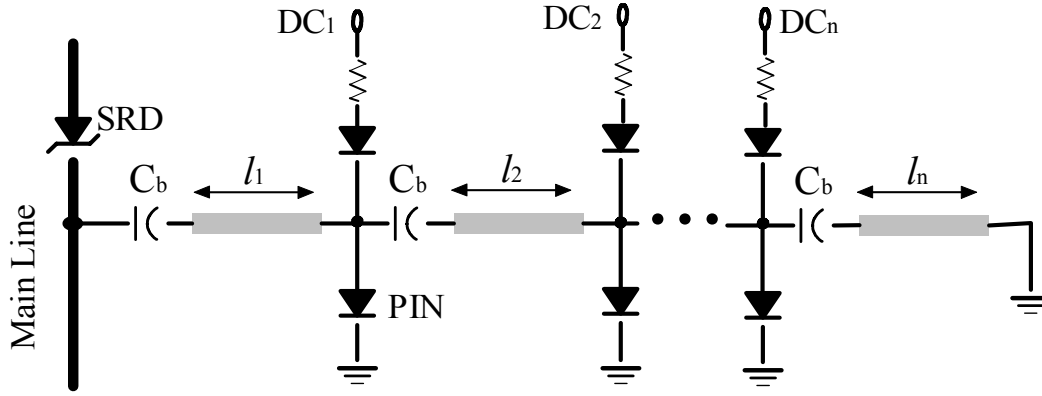


Fig.3. 11 Parallel PIN diode structure controlling the short transmission line stub.

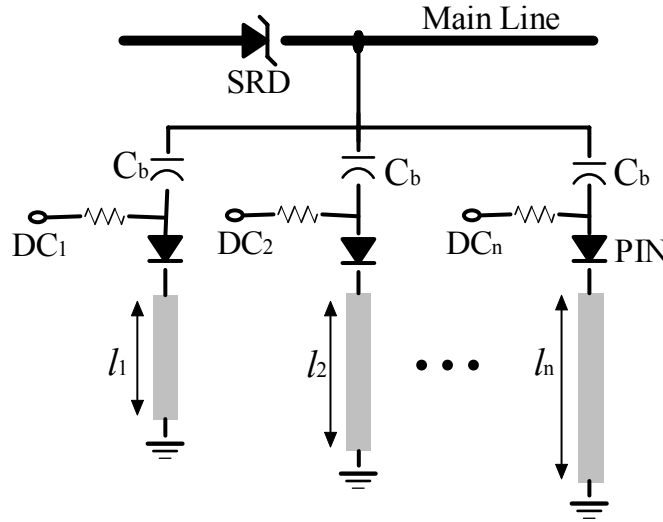


Fig.3. 12 Series PIN diode structure controlling short transmission line stubs.

Meanwhile, there is only one DC blocking capacitor, C_b , in series with each short circuited line stub. This capacitor mitigates the pulse distortion caused by having multiple DC blocking capacitors in a parallel PIN structure. Finally, the switching time between any two durations is doubled in parallel structures as compared to series structures. This occurs as the number of PIN diodes needed is doubled in order to complete the switching state. Table 3.3 summarizes the difference between parallel and series PIN diode structures.

Table 3.3 – A Comparison between Parallel and Series PIN Diode Structure

	Parallel structure	Series structure
Amount of PIN diodes	2n	n
Number of capacitors in series with short transmission line stub	n	1
Switching speed	20ns	10ns
Total cost	Higher	Lower

The design of the novel matching network is straightforward and is shown in Fig. 3.13. The R_f and C_f components comprise the novel input matching network used to significantly reduce the pulse duration and suppress the pulse distortion. Fig. 3.14 shows the measurement results of the fabricated tunable pulse generator. The output Gaussian pulse duration varies from 300ps to 1ns, measured at 10% of the pulse peak amplitude. All output pulses have minimal distortion and a very low ringing level.

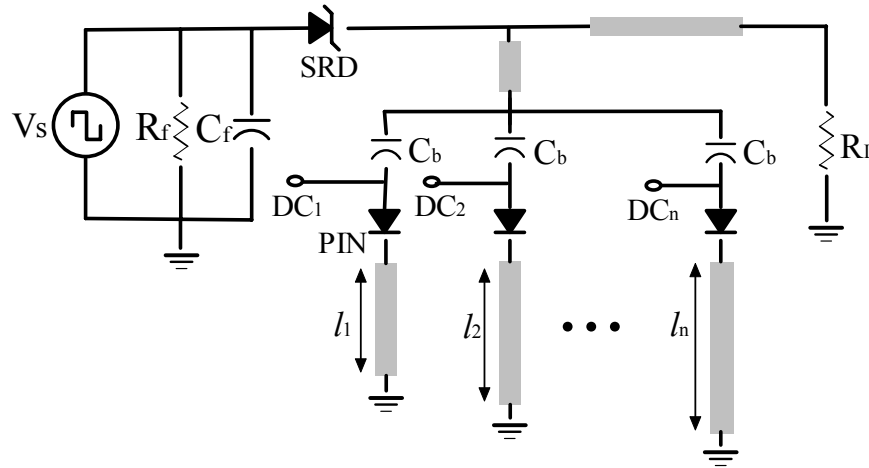


Fig.3.13 A schematic of pulse generator with tunable duration.

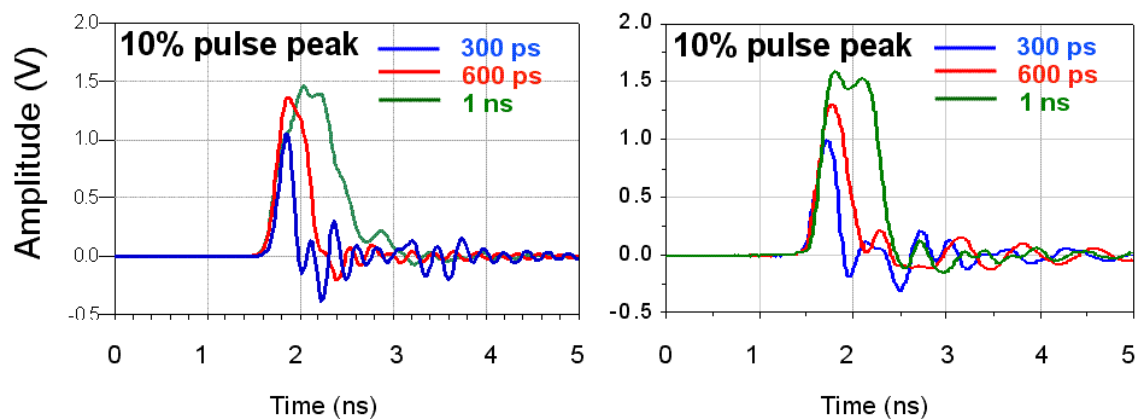


Fig.3.14 Simulated (left) and measured (right) Gaussian pulses with tunable pulse width from 300 ps to 1ns.

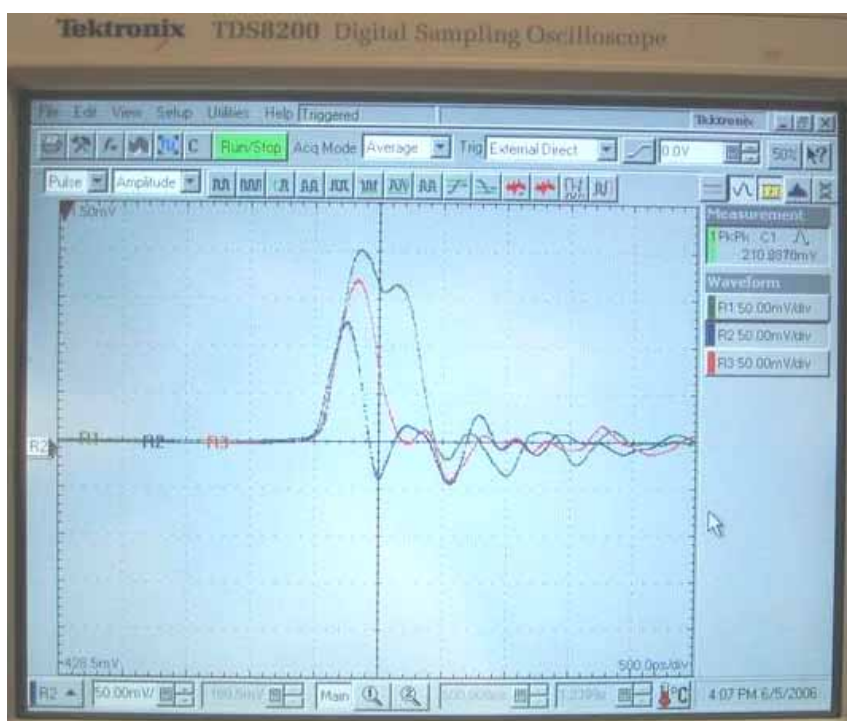


Fig.3.15 Photo of measured Gaussian pulses with tunable pulse width from 300 ps to 1ns using Tektronix TDS8200 digital sampling oscilloscope.

3.5 Tunable Pulse Shaping Circuitry

There are two types of pulse shaping circuits used to transform the Gaussian pulse to a monocycle pulse. A simple and straightforward way is to add an R-C differentiator circuit after the Gaussian pulse SRD circuit to produce a monocycle output [48]. This method lacks flexibility and does not offer tunable capability. In this chapter a new reconfigurable approach is proposed in order to deliver a clean monocycle output signal from an input Gaussian pulse. A second section of PIN diode controlled short-circuited transmission line stub is added along the main output line of the pulse circuit, as shown in Fig. 3.16. When the PIN diode is in the ON state, the Gaussian pulse propagating along the line is intentionally split between the main line and the shorted transmission line, l_s . The Gaussian pulse traveling along the short transmission line will reflect back as a negative pulse after a certain time and recombine with the former positive pulse to create a monocycle. When the PIN diode switches to the OFF state, only a Gaussian pulse will appear at the output load. Thus the ON/OFF state of the PIN diode will fully determine the pulse shape of the output signal. Furthermore, more complicated output pulse shapes such as polycycle can be generated by properly adding more sections of PIN diode controlled short-circuited transmission line stubs.

Fig. 3.17 shows the measured results of a tunable pulse shape output. The monocycle has very good symmetry and a low ringing level. High data rates can be achieved by employing the Pulse Shape Modulation scheme due to the fast switching speed of the PIN diode used to control the output shape of the monocycle and Gaussian pulses.

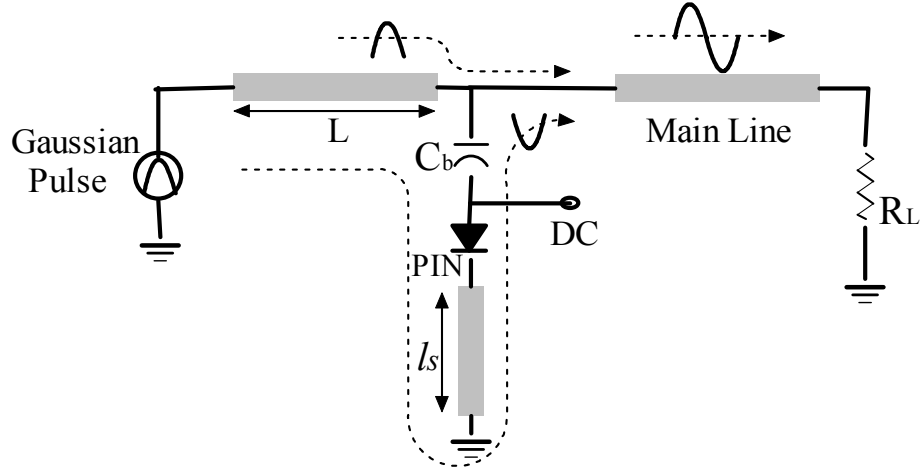


Fig.3. 16 Schematic of pulse generator with tunable pulse shaping output.

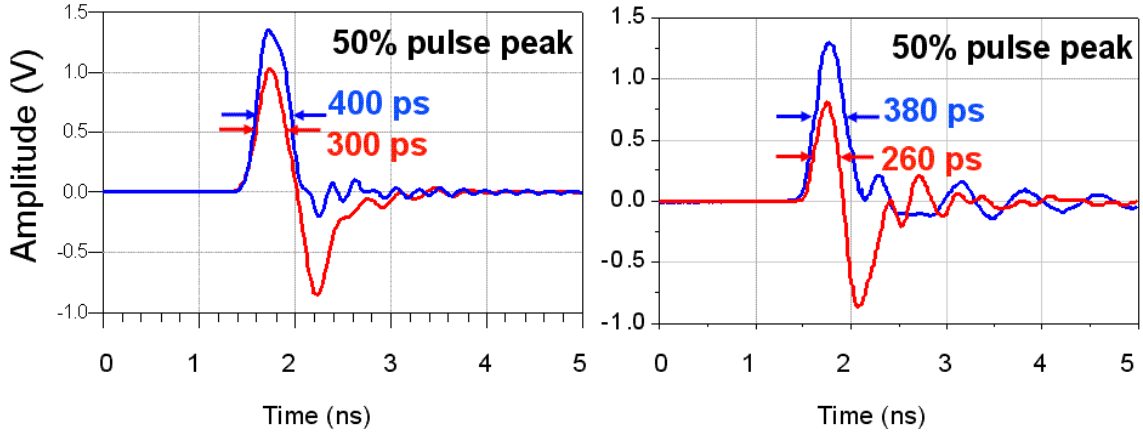


Fig.3.17 Measured pulse with both Gaussian and monocycle output pulses.

A summary of the circuit fabrication information regarding the reconfigurable pulse generator is listed in Table 3.4. It should be noted that an Agilent 33220A function generator first served as the driving clock to the SRD pulse generator which has now been successfully replaced by Texas Instrument's high speed buffer OPA2674 [49] and Vectron's high performance VTCO VTC4 [50].

Table 3.4– Circuit Fabrication Information

	Part Number	Key Feature
Substrate	FR-4	62 mil, $\epsilon_r=4.4$
SRD	MSD700 (Micrometrics Inc.)	Short transition time (60 ps)
PIN Diode	HSMS482x (Agilent Co.)	High freq. up to 3GHz, Low ON resistance
Trigger Clock	VTC4 (Vectron Inc.)	10 MHz, +/- 0.5 ppm
Trigger Buffer	OPA2674 (TI)	High gain, high drive capability

Fig. 3.18 shows the fabricated low cost SRD-based pulse generator that has an adjustable pulse duration in the range of 300 ps to 1ns with either monocyte or Gaussian pulse shape output. Pulse duration and shape can be adjusted based on the type of application. Switching time between the two output-states is within 10ns, corresponding to the time needed to switch between the ON and OFF states. PIN diodes are used for switching.

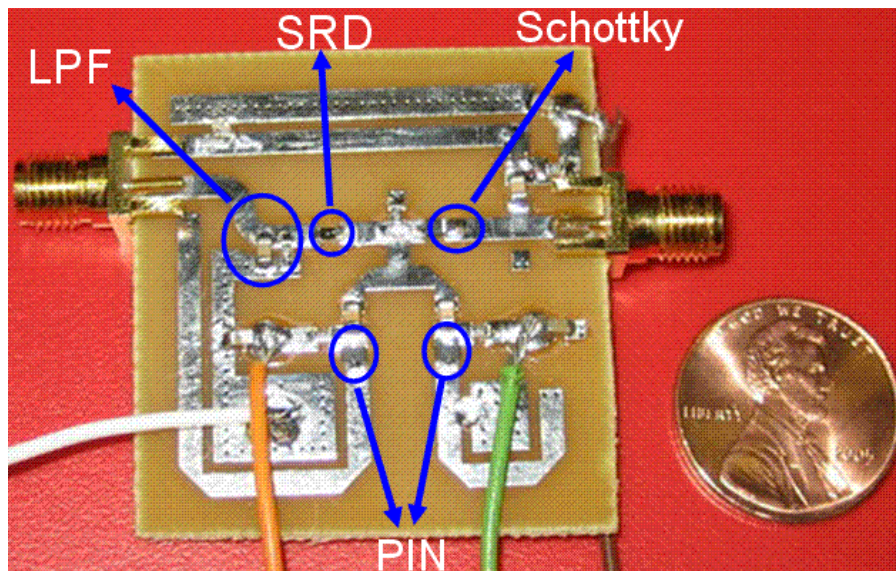


Fig.3.18 Photograph of the fabricated tunable pulse generator

3.6 Conclusion

A low cost, multi-functional pulse generator with electronic tuning of both pulse duration and shape has been developed. The developed tunable source is very useful in UWB applications such as see through walls radar with complex geometries. In the localization application, it is adequate to use a fixed pulse generator with a narrow pulse width. A novel input matching network has been introduced at the input of the tunable pulse generator to improve the quality of the produced pulse. Good agreement between the simulated and measured results has been achieved. Both simulated and measured results show that the input matching network can greatly prevent the pulse width broadening and suppress any significant pulse distortion. Novel series-connected PIN diode structures are utilized for the pulse duration and shape control. The developed pulse generator has a highly adjustable pulse duration ranging from 300ps to 1ns. The pulse generator is reconfigurable and could provide various output shapes such as monocycle and Gaussian pulses. These features will provide more flexibility in the design of adaptable UWB systems needed for such applications like see through walls radar.

The relationship between the pulse width and localization accuracy will be given in Chapter 10 which demonstrates that higher accuracy can be achieved using a narrower pulse width. In our implementation, a 300 ps pulse width is utilized as the source. The minimum achievable pulse width of 300 ps when using the SRD is limited by the diode transition time and device parasitic resistance and capacitance. In order to achieve a pulse width of less than 100 ps, a GaAs MMIC can be used.

Chapter 4

Sampling Mixer for UWB Signal Processing

4.1 Background

To detect the narrow pulses used in precise localization systems, usually in the range of a few hundred pico-seconds (i.e. about 5 GHz bandwidth), analog to digital converter (ADC) with at least 10 GSPS based on Nyquist criterion is required. Currently, such high performance ADC units are either not commercially available or too expensive for a majority of applications. As an alternative, a realistic approach is to sub-sample the UWB pulses upon extending their time scale while maintaining the pulse shape. Thus, the extended time scale of the UWB signals can then be handled by conventional ADC circuitries. This approach is not new and has been used since 1960s by many commercial sampling oscilloscope vendors. Here a low cost sub-sampling system will be presented.

4.1.1 The State of the Art Sampling Systems

There are many commercial off-the-shelf (COTS) high speed sampling modules exist such as the Tektronix 80E01 [51], Picosecond 7040 [52], AnaPico sampling system [53], and Maxtek high speed Data Converter [54]. A list of those commercial sampling modules can be found in Table 4.1. The Picosecond 7040 sampling module, for example, utilized the nonlinear transmission line (NLTL) technique to achieve a fast rise time of 14 ps, and an associated bandwidth of 25 GHz. However, it is not a standalone sampling system since it requires an external strobe triggering signal, where the strobe pulses work

Table 4.1– List of the State of the Art Commercial Sampling Modules

	Tektronix 80E01	Picosecond 7040	AnaPico	Maxtek Data Converter
Rise time	7 ps	14 ps	N/A	N/A
Bandwidth	50 GHz	25 GHz	35 GHz	6.5 GHz
Sampling rate	N/A	Up to 10 MHz	> 100 MHz	12.5 GSPS
Sampling Scheme	Equivalent time	Equivalent time	Equivalent time	Real time
Technology	N/A	GaAs MMIC & NLTL	GaAs MMIC	IBM SiGe BiCMOS
Module Size (cm ³)	13.5x7.9x2.5	5.1x3.8x1.3	40x30x15	N/A
Unit Cost	\$20,000	\$5,000	\$20,000	\$80,000

as the LO source to trigger the sampling mixer ON and OFF. The AnaPico sampling system recently announced a large bandwidth and fast sampling rate, but this comes with a bulky box and very high cost. The Maxtek Data Converter provides an extremely fast real time sampling solution but with an extremely high cost too.

For a lower cost approach, the sampling mixer (also called a sub-sampler or just a sampler in the rest of this chapter) circuits can be implemented using microwave integrated circuit (MIC) technology. Many MIC based sampling mixers have been proposed [55-57], but they suffer from a relatively large conversion loss. Recently, for example, J. Han et al. developed a coupled-slotline-hybrid (CSH) sampling mixer integrated with a strobe impulse generator for UWB applications [58]. The CSH sampler achieved a relatively low conversion loss of 4.5 to 7.5 dB up to 5.5 GHz, and a dynamic range of over 50 dB. However, its performance is sensitive to the location of the sampling diodes. This sensitivity would lead to an undesired side-lobe ringing with the

same polarity as the main peak of the strobe. This ringing may turn on the sampling diodes and might cause spurious effects leading to unacceptable signal distortions.

4.1.2 Hybrid Sampler Design Challenges

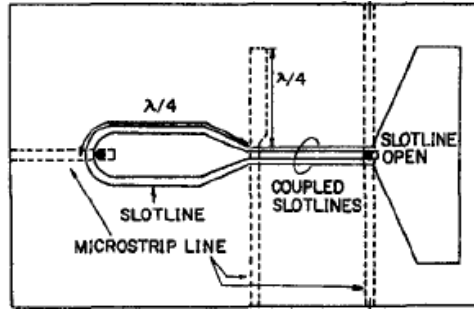
In the design of these hybrid sampler circuits, the most demanding task is to integrate an appropriate balun structure into the sampling circuit. The balun splits the strobe signal into two identical pulses with similar amplitudes but opposite polarities over a wide frequency range. Many authors have previously addressed this design challenge. For example, [59] realized a balun using a ferrite transformer but with a non-planar structure. Also, [58] utilized a balun based on a microstrip to a coupled-slot line transition where the RF and LO signals share the same traveling path. However, a strong coupling may exist between the coplanar-waveguide (CPW) and the coupled slot-line modes of [58], which would require an air-bridge (or a 0 Ohm resistor) to cancel these coupling effects. Recently, [60] gave a comparison between various approaches to design sub-sampling mixers including the utilization of a surface mount device (SMD) balun. Figure 4.1 shows various balun structures and Table 4.2 compares their performance.



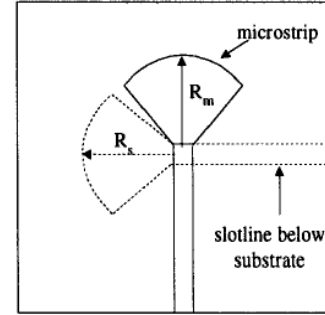
(a) Ferrite transformer



(b) SMD balun



(c) Coupled slot line balun



(d) Microstrip to slot line balun

Fig.4.1 Various Balun Structures: (a) M/A-COM's MABACT0039 Ferrite transformer [61]; (b) Tyco's CHM1608U SMD balun [62]; (c) Coupled slot line balun [63]; (d) Microstrip to slot line balun [64].

Table 4.2 - Comparison of Different Balun Structures

Balun Type	Bandwidth	Integration w/ Circuit	Phase Balance	Amplitude Balance
Ferrite Transformer	Up to 3 GHz	Easy	$\pm 13^\circ$	± 2.2 dB
SMD Balun	2.4 – 2.5 GHz	Easy	$\pm 10^\circ$	± 2 dB
Coupled Slot Line	Up to 5 GHz	Complex	$\pm 2^\circ$	± 0.3 dB
Microstrip to Slot Line	Up to 5 GHz	Easy	$\pm 2^\circ$	± 0.3 dB

According to Table 4.2, the ferrite transformer has a limited frequency range up to 3 GHz and suffers from a relatively large phase and amplitude imbalance, which would produce imbalanced strobe signals and cause significant sampling distortion. On the other hand the SMD balun has a very narrow bandwidth, prohibiting them to generate the required narrow strobe pulses. The coupled slot line balun features a wide band, and small phase/amplitude imbalance. However, circuit implementation based on this balun structure is complicated and suffers from a strong coupling between the CPW and the

coupled slot-line modes [58]. The microstrip to slot line balun has been chosen in our sampler design since it doesn't suffer from these aforementioned drawbacks.

It is still a challenge to design a highly efficient wideband sub-sampling mixer using MIC [55]. The challenge stems from:

- ◆ minimizing the RF to IF conversion loss;
- ◆ suppressing the strobe pulses ringing level;
- ◆ reducing the strobe waveform leakage to both the RF and IF ports;
- ◆ lowering the spurious levels of the down-converted signals.

In the following sections, the design efforts to develop a compact sampling mixer circuit based on a fully balanced structure will be explored. The circuit occupies only 32x20 mm², and has been fabricated on a double-sided substrate using a hybrid technology for lower production cost.

4.2 Sampling Concept

The fundamental principal of sampling is the repeated quasi-instantaneous capturing of a time-varying waveform by a sampling gate. The gate is opened and closed by narrow strobe pulses, which are triggered repeatedly by a precise time base [65]. During operation, a repetitive train of identical pulses are applied to the RF port with a pulse repetition frequency (PRF) denoted by f_0 . By firing the strobe with a slightly offset frequency given by $(f_0 - \Delta f)$ or $(f_0 + \Delta f)$, the strobe and RF signals are mixed in such a way that the strobe signal slowly scans across the RF signal being sampled. The complete

received signal can be reconstructed after a complete acquisition cycle with a total extended time equal to $(1/\Delta f)$. An extending factor “ α ” is defined as $(f_0/\Delta f)$, which corresponds to an extending ratio of the reconstructed signal over the un-sampled received signal.

The sampling circuit concept described above has been analyzed using Agilent ADS2005A transient simulation tool, as shown in Fig. 4.2, where a 300 ps Gaussian pulse (shown in Fig. 4.3a) was sampled. The PRF (f_0) of the pulse signal, and the strobe offset frequency Δf were set at 10 MHz and 1 KHz respectively. As seen in Fig. 4.3b, the simulated IF output shows an extended Gaussian pulse with a duration of 3 μ s, which confirms the calculated extending ratio of $\alpha = f_0/\Delta f = 10,000$. This corresponds to an equivalent time sampling rate of 100 GSPS. Higher equivalent sampling rates can be achieved upon reducing the frequency offset Δf . However, reducing the frequency offset will slow down the sampling speed and cause a systematic error in the noncoherent localization system which will be addressed in detail in Chapters 9 and 10.

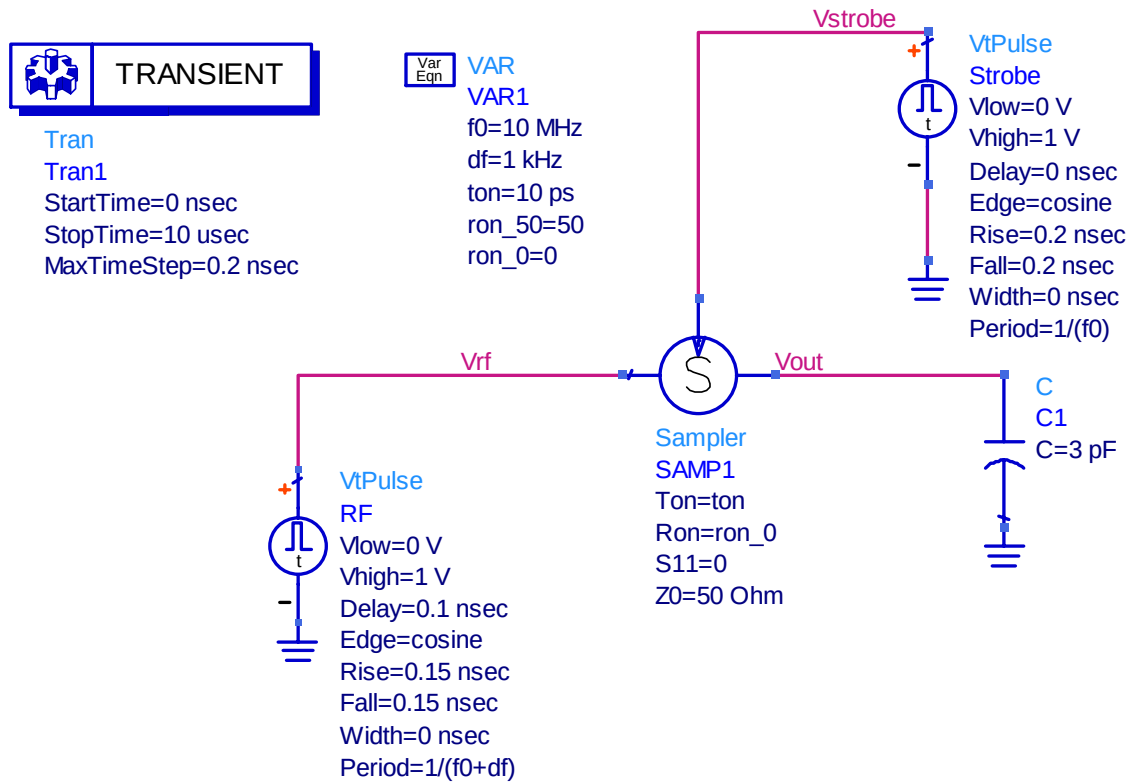


Fig.4.2 Demonstration of the sampling concept using ADS transient simulator.

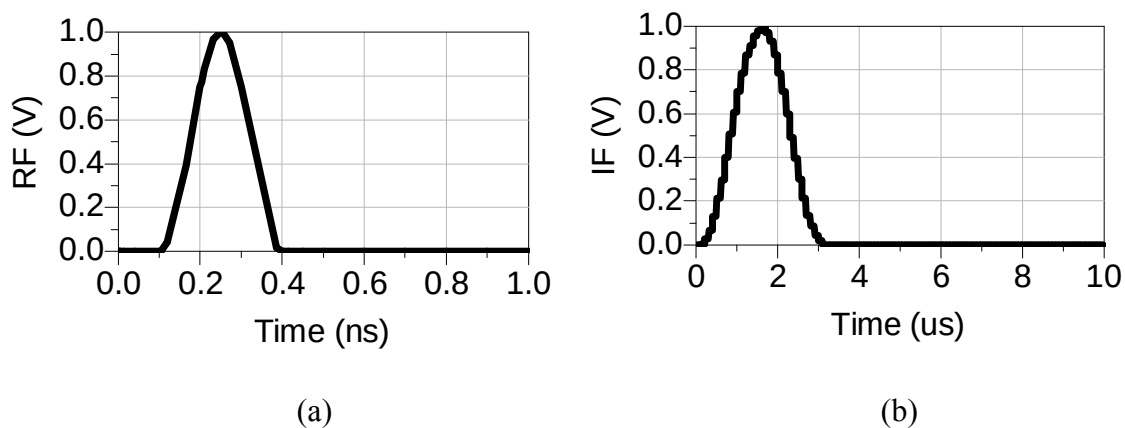


Fig.4.3 ADS simulation results: (a) 300 ps Gaussian pulse at RF input port; (b) 3 μ s Gaussian at IF output.

4.3 Strobe Generator Design

To enable a broadband performance of the UWB sampling mixer, a strobe signal with a fast rising (or falling) edge is required. Fig. 4.4 shows the schematic of the strobe-step generator based on a SRD connected in shunt. A resistor R_1 of $2\ \Omega$ was chosen to stabilize the driving clock source. A capacitor C_b , SRD and a resistor R_L comprise a clamping circuit, which generates an instantaneous self-biasing potential in order to generate a larger step. An optimized value of 22 nH for L_x was chosen to prevent the fast strobe from leaking back to the clock source and to allow the clock signal to pass without distortion.

Fig. 4.5 shows both the simulated and measured results of the output strobe signal when the strobe step generator is triggered by a 4 V square wave with a PRF of 10 MHz. The circuit simulation was carried out using ADS and the measured data was extracted from Tektronix TDS8200 digital sampling oscilloscope. Both the ADS simulation and measured results show a fast rising edge of approximately 100 ps. The strobe rising time is limited by a 70 ps transition time of the SMMD-840 Metelics SRD device and the utilized MIC fabrication, resulting in a limited 3 dB RF-IF conversion bandwidth.

Rapid advancement of the semiconductor technology has led to a much faster step rising edge, and thus, a significant sampler bandwidth improvement. For example, Rodwell's group achieved a fall time of 680 fs (725 GHz) by using GaAs nonlinear transmission lines, thus enabling millimeter-wave sampling [66]. Whiteley et al. [67] described the complete sampler hybrid circuit including the SRD pulser, balun, and GaAs

NLTL for strobe shortening. A 50 GHz bandwidth has been achieved. Double step sharpener using GaAs FET has been reported and achieved roughly a 25 ps rise time [68, 69].

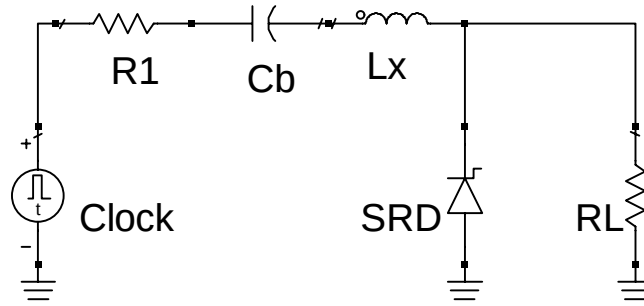


Fig.4.4 Schematic of the strobe-step generator.

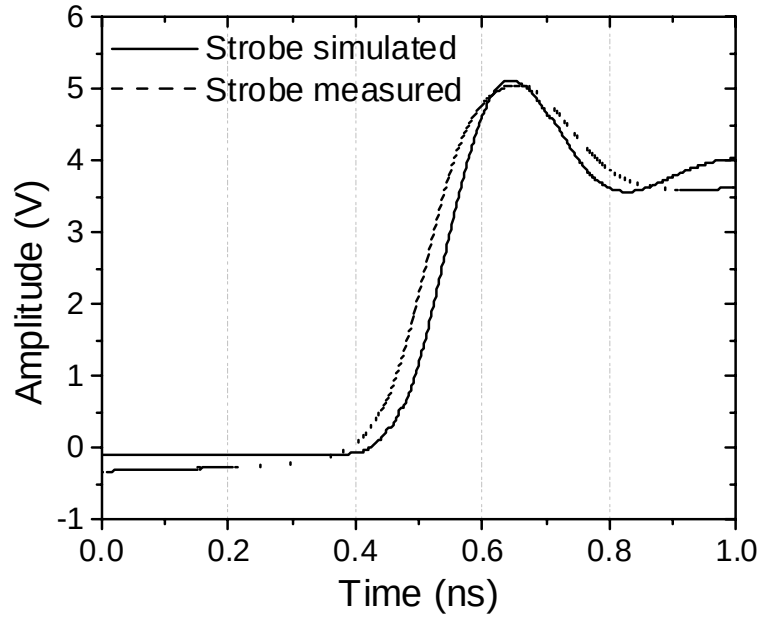


Fig.4.5 Simulated and measured output strobe-step signal.

4.4 Sampling Mixer Circuit Design

The developed sampling mixer is based on a traditional two-diode (D_1, D_2) bridge configuration. The schematic of the designed sampler is shown in Fig. 4.6. The diodes are high-barrier Schottky mixer diodes (MNH312 from MicroMetrics Inc), and have a small series resistance (R_s) of 10 Ohm, a high forward barrier voltage of 0.55V (at 1mA), and a tangential signal sensitivity of -52 dB. Meanwhile, an RC discharge path formed by the R_h and C_d network has been utilized to provide a proper time constant that is adjusted to be much slower than the RF signal charging time ($\tau \approx R_s C_h = 5$ ps) yet much faster than the driving clock period (100 ns when $f_0 = 10$ MHz). Additionally, the value of R_h is optimized using Agilent's ADS2006A to maintain a good conversion loss and low baseband noise. Hence, optimized values of 270 Ohm and 3 pF are selected for R_h and C_d respectively, corresponding to a discharging time of 810 ps. A good LO port matching can be realized by adding a 100 Ω resistance R_t , bridging the upper and lower strobe arms. A 50 Ohm terminating resistor R_f was placed at the end of the RF path and close to the sampling diodes to provide a proper matching for the RF port. A complete ADS simulation environment can be found in Fig. 4.7 and a Schottky diode model is given in Fig. 4.8.

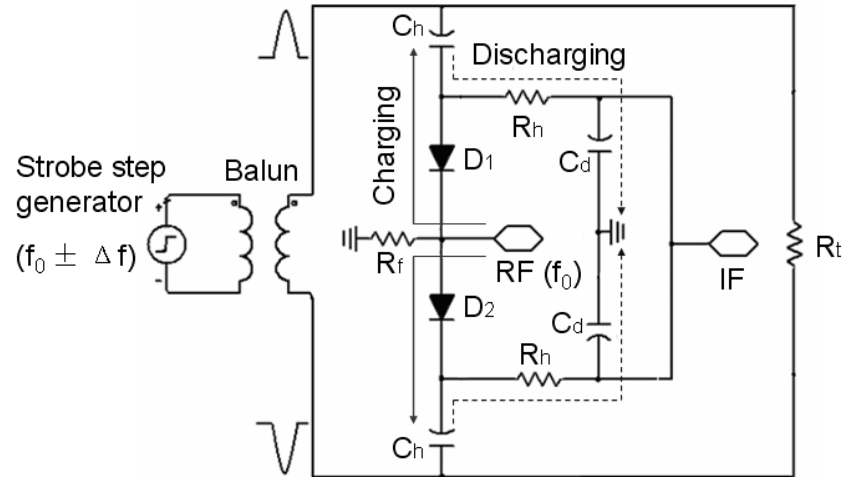


Fig.4.6 Schematic of the sampling mixer circuit topology.

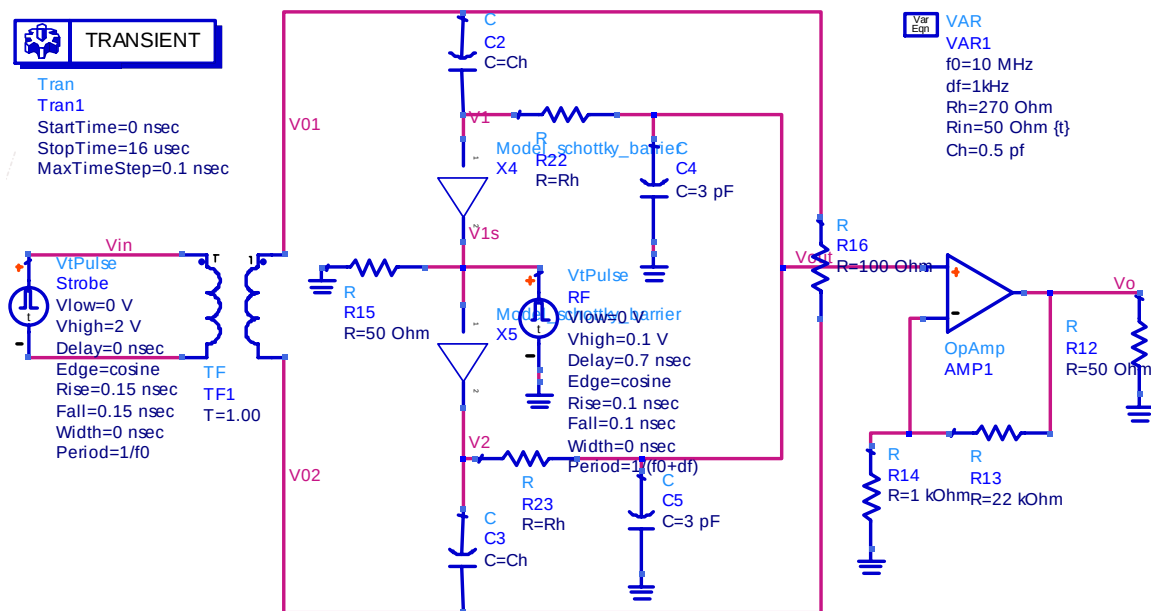


Fig.4. 7 ADS schematic of the sampling mixer circuit.

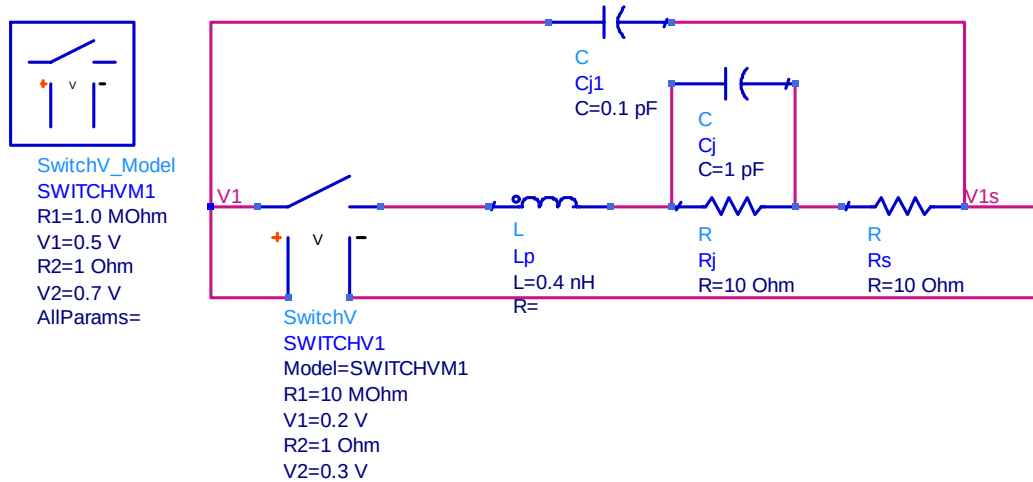
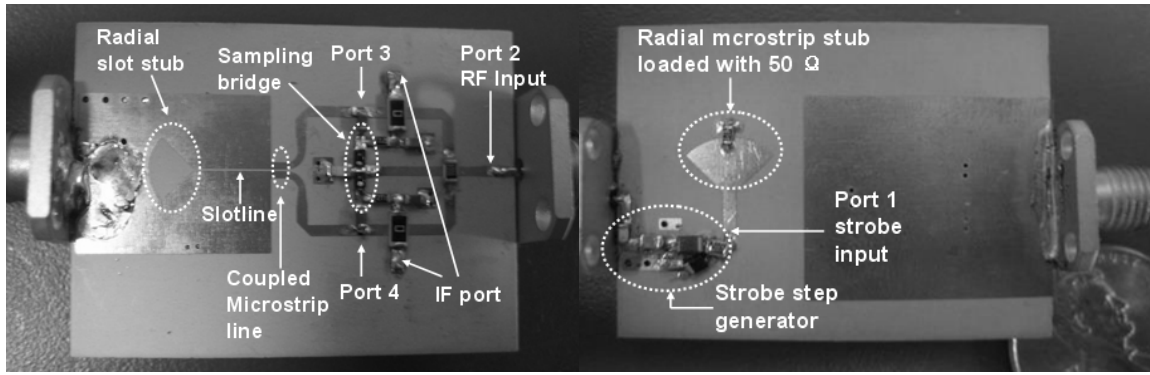


Fig.4.8 ADS Schottky mixer diode model.

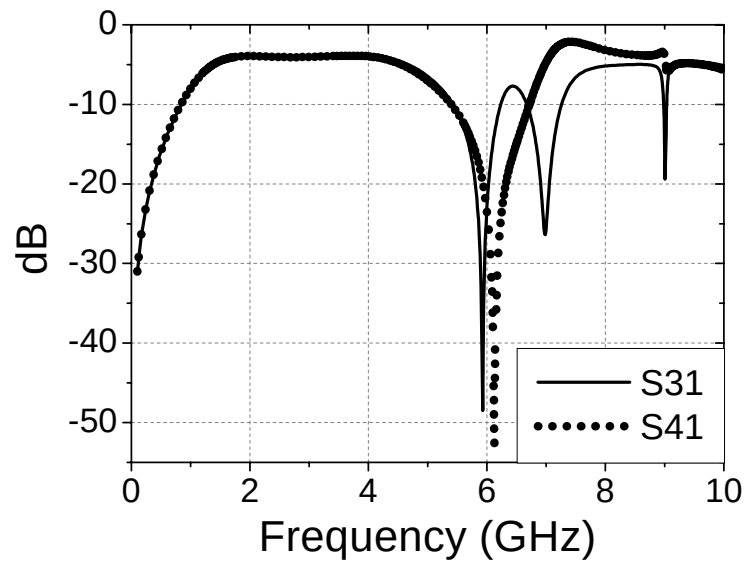
As mentioned earlier, a wideband balun structure that is comprised of a broadband radial microstrip to slot-line transition has been used [64], and followed by a slot-line to coupled microstrip line transition. Then the coupled microstrip line splits into two symmetrical microstrip arms (shown in Fig. 4.9), that are later connected to two sampling diodes (D_1 , D_2). Top and bottom views of the fabricated circuit are shown in Fig. 4.9. Figure 4.10 shows the simulated results of the amplitude and phase balance between the strobe-input port 1 and the two microstrip arms (i.e. port 3 and port 4 shown in Fig. 4.9) using ADS momentum, where a pass-band from 1 to 5.5 GHz has been predicted. This design provides two identical out-of-phase signals while suffering minimal loss.



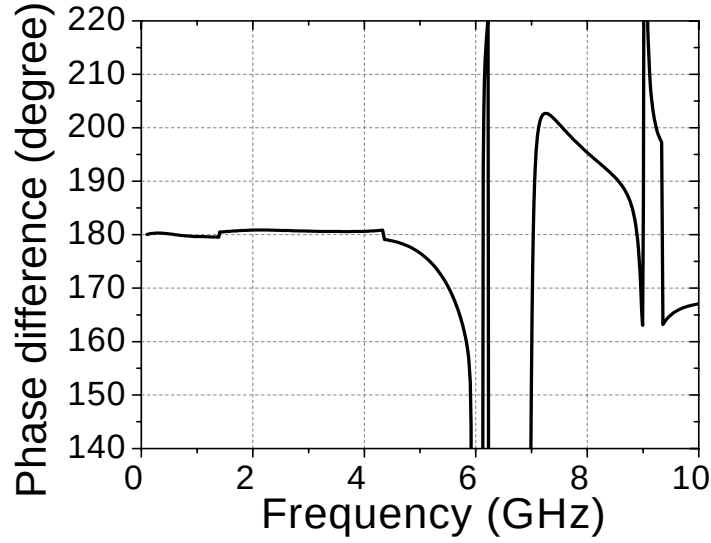
(a) Top layer

(b) Bottom layer

Fig.4.9 Top and bottom views of the fabricated sampling mixer.



(a) Balun amplitude response



(b) Balun phase response

Fig.4.10 Simulated results of: (a) S31 and S41, (b) phase difference between S31 and S41.

Moreover, the designed balun has an inherently high insertion loss at low frequencies (i.e. $< 1\text{GHz}$), which helps in blocking the 10MHz triggering clock signal. At the same time, it also differentiates the input strobe-step signal, which has a 100 ps rise time, into two strobe impulses with opposite polarities to trigger the sampling bridge. The measured two opposite strobe impulses are in excellent agreement with our predicted results, and as noted in Fig. 4.11b, the measured strobe impulses do not have unwanted side-lobe ringing. Meanwhile, the ringing with an opposite polarity (following the main peak) does not affect the sampler's performance as it enhances reverse biasing of the sampling diodes keeping them turned off. According to the approximate 3 dB bandwidth prediction formula, $\text{BW}(\text{GHz}) \approx 350 / \text{Gating Duration (ps)}$ [55], a 3.5GHz bandwidth is expected for a 100ps gating duration. Generally, a wider bandwidth could be potentially achieved by setting a proper gating duration through adjusting the strobe impulse amplitude.

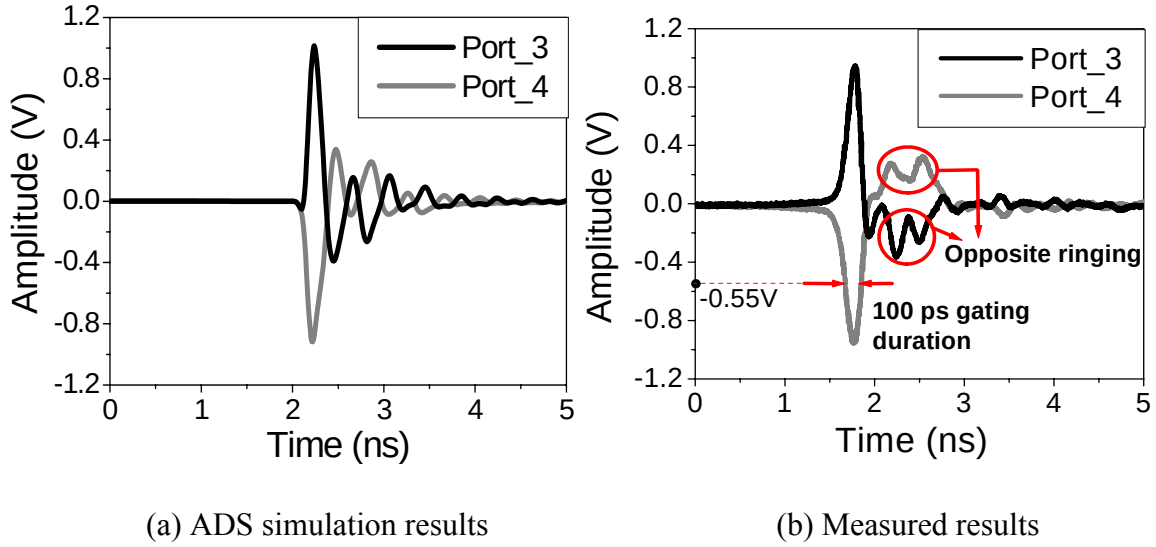
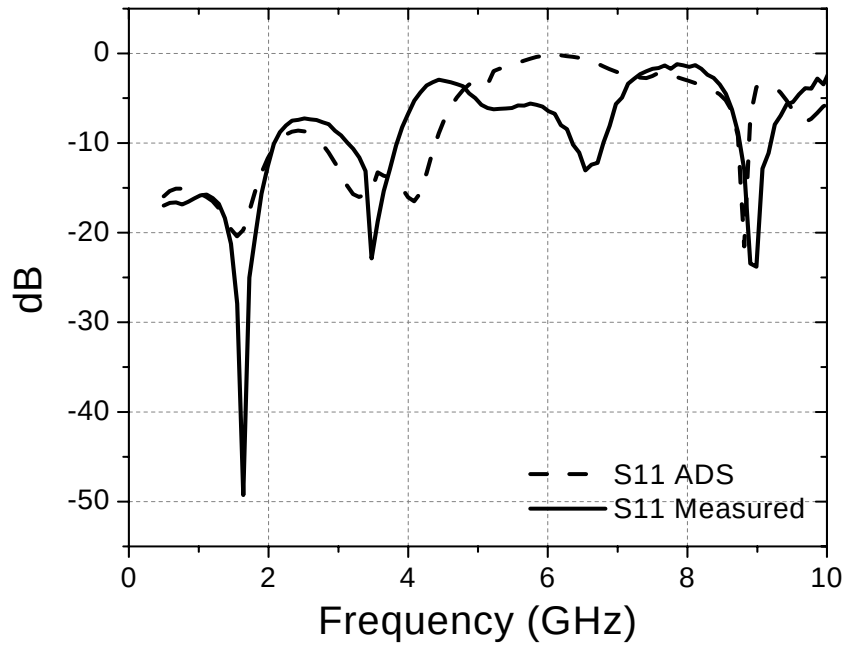
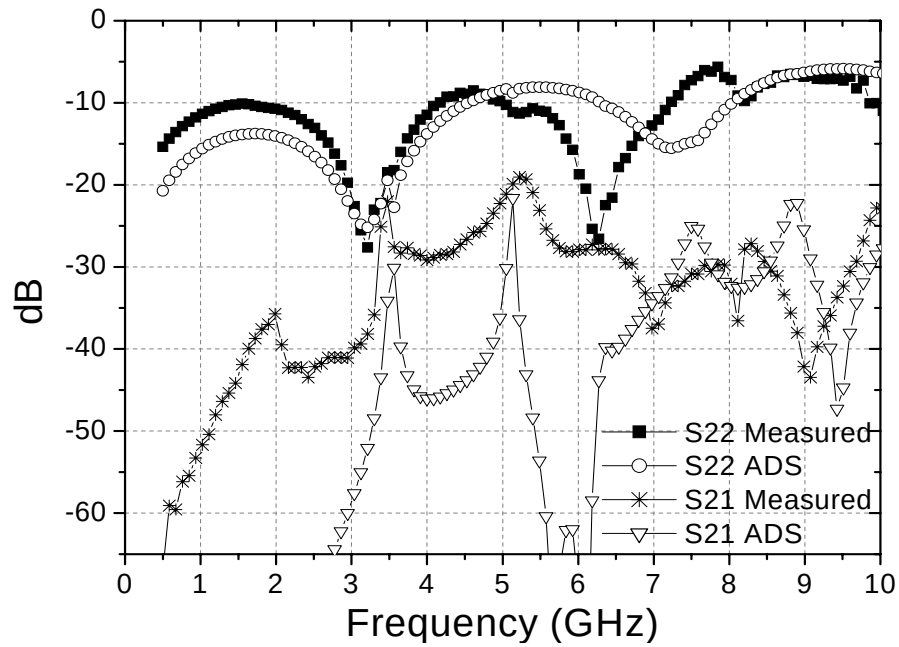


Fig.4.11 Strobe impulses at port 3 and port 4: (a) simulated and (b) measured results.

Good LO port and RF port matching and isolation has been achieved, as shown in Fig. 4.12. Meanwhile, the utilization of the balanced balun, and the physical displacement of the LO and RF signal paths (shown in Fig. 4.9) have improved the isolation performance. The measured RF return loss and the LO-RF isolation are better than 10-dB and 30 dB respectively up to 4 GHz, according to Fig. 4.12b.



(a)

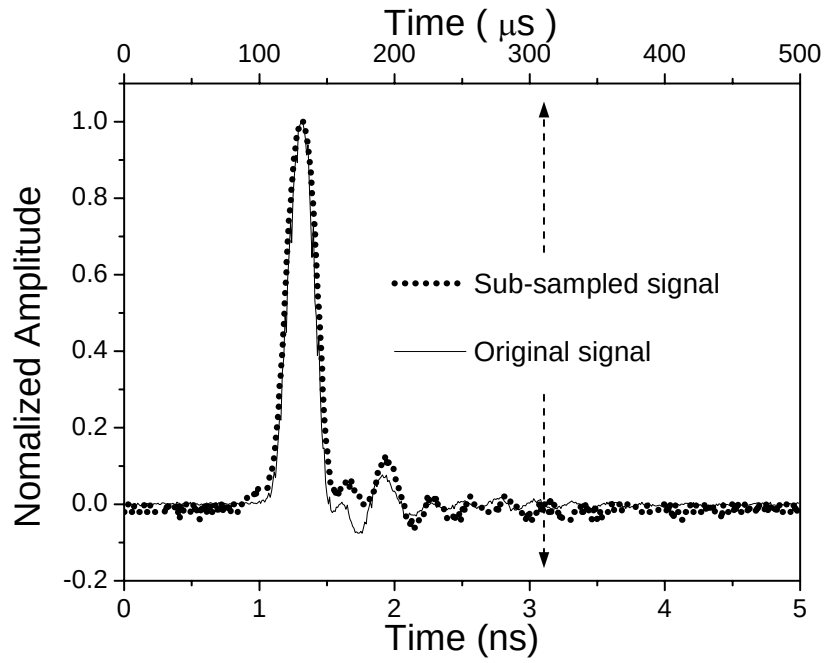


(b)

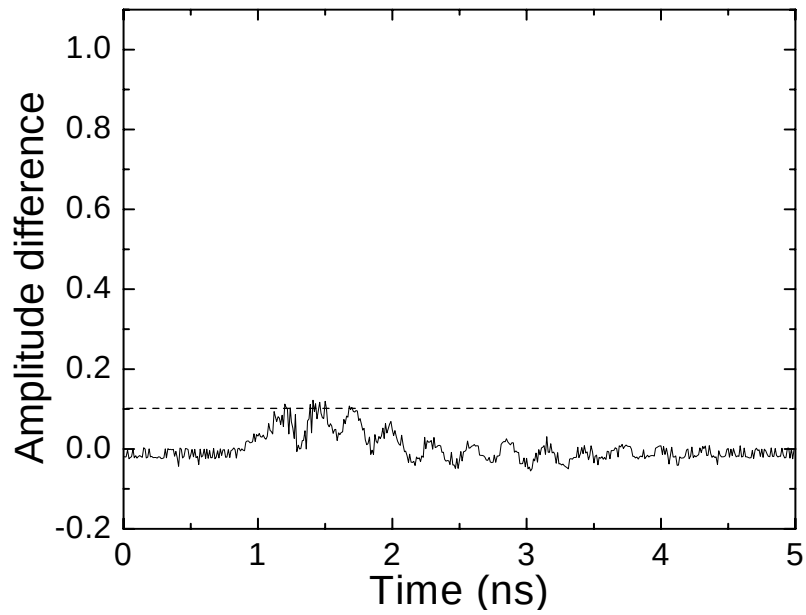
Fig.4.12 ADS simulation and measured: (a) LO port return loss. (b) RF port return loss and RF to LO port isolation.

4.5 Sampler Performances

The UWB sampling mixer (including the strobe-step generator) was fabricated on a double layer PCB board using Rogers RT/Duroid RO3010 materials with a relative dielectric constant of 10.2 and a thickness of 0.635 mm. For measurements, a dual-channel functional generator is used to trigger both the RF signal (i.e. 300 ps Gaussian pulse) with a pulse repetition frequency “PRF” (f_0) of 10 MHz, and the strobe-step generator with a frequency of $f_0 \pm \Delta f$. The added offset frequency (Δf) is set to be 100 Hz, corresponding to an extending ratio “ $\alpha = f_0 / \Delta f$ ” of 100,000. The down-converted signal from the sampling mixer is then amplified by an operational amplifier and measured by a Tektronix 340A oscilloscope with a 500 MSPS sampling rate. The output is compared to the original RF signal from the pulse generator measured by a Tektronix TDS8200 sampling oscilloscope. A comparison between our sub-sampled signal and the original signal is shown in Fig 4.13a. As demonstrated, the output pulse from the developed sampler extends the Gaussian pulse duration from 300 ps to 30 μ s, while maintaining almost the same pulse shape with minimal signal distortion. After subtracting the two normalized pulse signals, shown in Fig. 4.13b, a maximum amplitude difference of less than 10% was achieved.



(a)



(b)

Fig.4.13 (a) Measured 300 ps Gaussian pulse: comparison between our developed sub-sampler extended output and the original signal from Tektronix TDS8200; (b) Amplitude difference after subtraction, with a maximum difference of 10%.

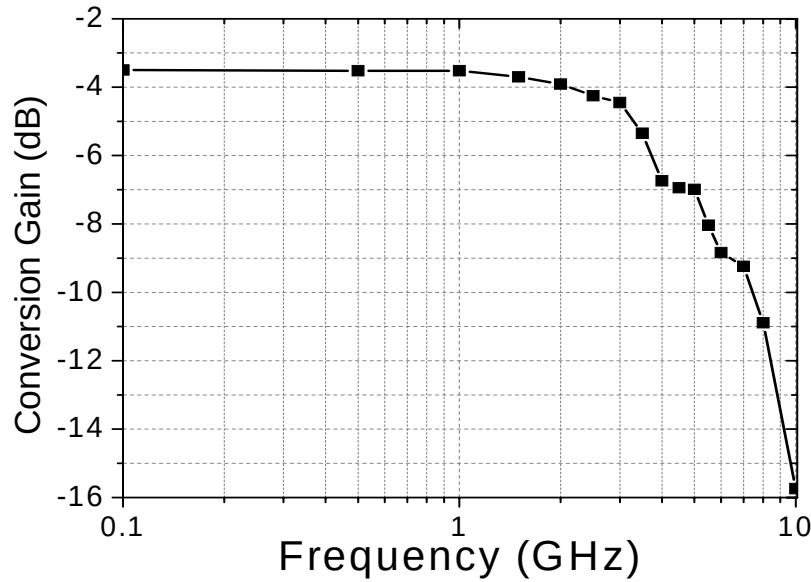


Fig.4.14 Measured conversion loss of the sampling mixer.

Additionally, the sub-sampler circuit has demonstrated an improved conversion efficiency performance. The measured conversion loss for a -10 dBm sinusoidal RF input signal over a wide frequency range is shown in Fig.4.14, which exhibits a 3.5 to 6.5 dB conversion loss up to 4 GHz. Meanwhile, the spurious level of the baseband signal has been determined by measuring the second harmonic of the down-converted signal, and it is better than -38 dBc over the entire RF operating band.

Three kinds of Schottky diodes (listed in Table 4.3) from MicroMetrics Company with different barrier levels have been utilized in the sampling mixer to study the relationship between the required strobe driving power and the circuit linearity for the Schottky mixer diodes with different barrier levels.

Table 4.3 – Comparison of Driving Power and Linearity with Different Barrier Levels

Schottky mixer diode	Forward Voltage @1mA	Strobe driving level (V-peak / power)	Measured input P1dB point
MNL112 (Low barrier)	0.25 V	2 V / 19 dBm	-5.5 dBm
MNM212 (Medium barrier)	0.45 V	3.5 V / 23.7 dBm	1.5 dBm
MNH312 (High barrier)	0.55 V	5 V / 26.8 dBm	5.2 dBm

As noticed from Table 4.3, there is a trade-off for the sampling mixer in choosing the right diode for a proper application. Low barrier Schottky diodes require less strobe driving power but are mostly nonlinear. High barrier diodes will have the best P1dB output level but the required strobe driving power is increased. Medium barrier diodes are somewhere in between. Fig. 4.15 shows the measured IF output power as a function of the RF input power with a fixed RF frequency of 3 GHz and a baseband amplifier of about 15.5 dB gain using the high barrier diode MNH312. The measured input 1-dB compression point is 5.2 dBm.

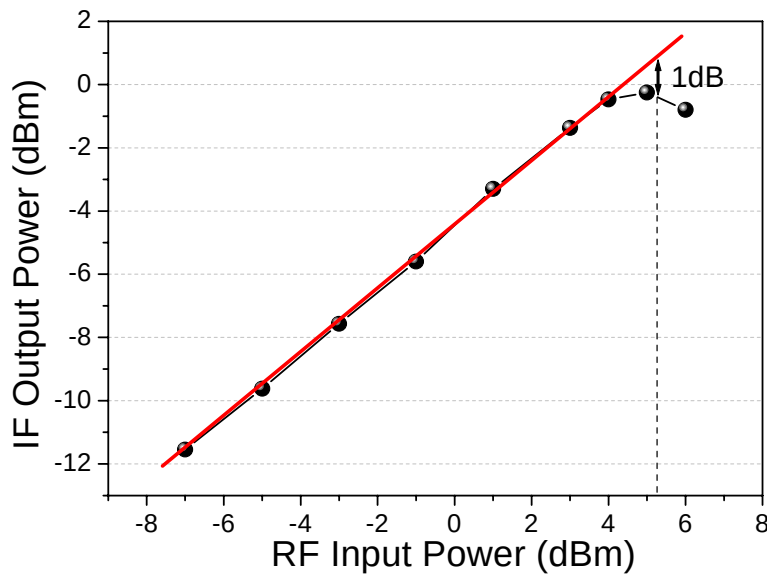


Fig.4.15 Measured 1-dB compression point.

Table 4.4– Comparison of Sampling Mixer Using Hybrid Technology

	Ref. [56]	Ref. [59]	Ref. [55]	Ref. [58]	Ref. [60]	This work
Year	1992	2002	2004	2005	2006	2007
Operating band	1-20 GHz	DC - 6GHz	DC - 9GHz	DC – 5.5 GHz	DC - 6.4GHz	DC - 4 GHz
Conversion loss without IF amp	36-42 dB	9 dB	16-19 dB	4.5-7.5 dB	N/A	3.5-6.5 dB
1-dB compression	6 dBm	N/A	8.5 dBm	2.5 dBm	5 dBm	5.2 dBm
Dynamic range	N/A	67 dB	N/A	> 50 dB	> 42 dB	> 50 dB
RF VSWR	$\leq 2:1$	N/A	$\leq 2.6:1$	1.3:1	N/A	$\leq 2:1$
Spurious level	N/A	N/A	N/A	-20 dBc	N/A	< -38 dBc
Dimensions (mm)	40 x 35	70 x 40	76 x 50	83 x 50	N/A	32 x 20

In addition, the measured 8 dB tangential sensitivity is better than -45 dBm, the dynamic range exceeds 50 dB, and the RF-IF isolation is over 42 dB. A detailed comparison between the design presented here and relevant published results is shown in Table 4.4. It can be seen that the proposed design is compact, has relatively low conversion loss, and insignificant spurious levels.

4.6 Conclusion

The performance of a recently developed sub-sampling circuit presented by J. Han et al. [58] was enhanced by utilizing a broadband balun. The developed balun is comprised by cascading two transitions: a radial microstrip to slot-line, and a slot-line to coupled microstrip line. The wide spacing between the RF and LO paths makes the various port-matching and isolation simpler to realize. The balun is optimized to differentiate the strobe step signal into two strobe impulses with opposite polarities and

without any unwanted side-lobe ringing, which leads to minimal pulse distortion and low spurious levels. The designed sampler has been successfully used for our UWB precise localization applications where high efficiency and minimal signal distortion are required [70].

Chapter 5

Ultra-wideband Antennas

One of the challenges for the implementation of an UWB localization system is the development of suitable or optimal transmitting and receiving antennas. From a system point of view, the response of the antenna should follow the specifications listed below:

- ◆ The frequency response needs to cover the entire operating bandwidth, i.e. 6 GHz to 10 GHz as shown in Fig. 2.4;
- ◆ The transmitting antenna, which is part of our tag, should be compact in size and have an omni-directional radiation pattern over a wideband, since the receivers are put around the transmitting tag in a 3D space;
- ◆ The UWB antennas should include minimal pulse signal distortion in time domain;
- ◆ The receiving antenna phase center should be as constant as possible with respect to both frequency and direction for a high precision localization system, which requires the antenna to maintain a constant radiation pattern over the frequency range of interest.

This chapter shows the developed transmitting and receiving antennas for the proposed high accuracy localization system, addressing all the above specifications. Firstly, an UWB monopole with a modified ground plane has been designed to improve the omni-directional radiation pattern over a wide band with minimal pulse distortion. Secondly, a novel time domain method to investigate the phase center variation of the UWB antenna versus direction is given. Finally, a Vivaldi Antenna with a protruded

dielectric rod has been designed to increase the antenna gain and minimize phase center variations.

5.1 Transmitting Monopole Antenna

A planar monopole antenna has many attractive features such as having a simple structure, exhibiting ultra-wideband characteristics, and producing a near omni-directional radiation pattern [71-73]. However, the ability to sustain an omni-directional pattern over an ultra-wide band is extremely important requirement for the transmitting antennas and should be thoroughly addressed. An omni-directional radiation pattern is a demanding feature in point-to-multipoint applications that require full horizontal signal coverage such as indoor localization, asset tracking, indoor GPS, mobile telemetry, base station antennas, etc. Such needs have been investigated and proposed herein is a novel elliptical monopole structure combined with a modified ground plane that achieves an excellent omni-directional radiation pattern as well as a satisfactory input impedance match over an ultra-wide bandwidth. The developed structure has been fabricated and tested, and good agreement between simulated and measured results has been achieved. Additionally, the developed antenna has been utilized as a part of the UWB localization radar system with millimeter accuracy as will be discussed in detail in Chapters 8-10. Presented here is the design methodology and perform time domain measurements that demonstrate the excellent signal integrity of the developed antenna.

5.1.1 Basic Omni-directional Antenna Design Concept

There are many options to design an antenna with omni-directional radiation pattern, as shown in Fig. 5.1.

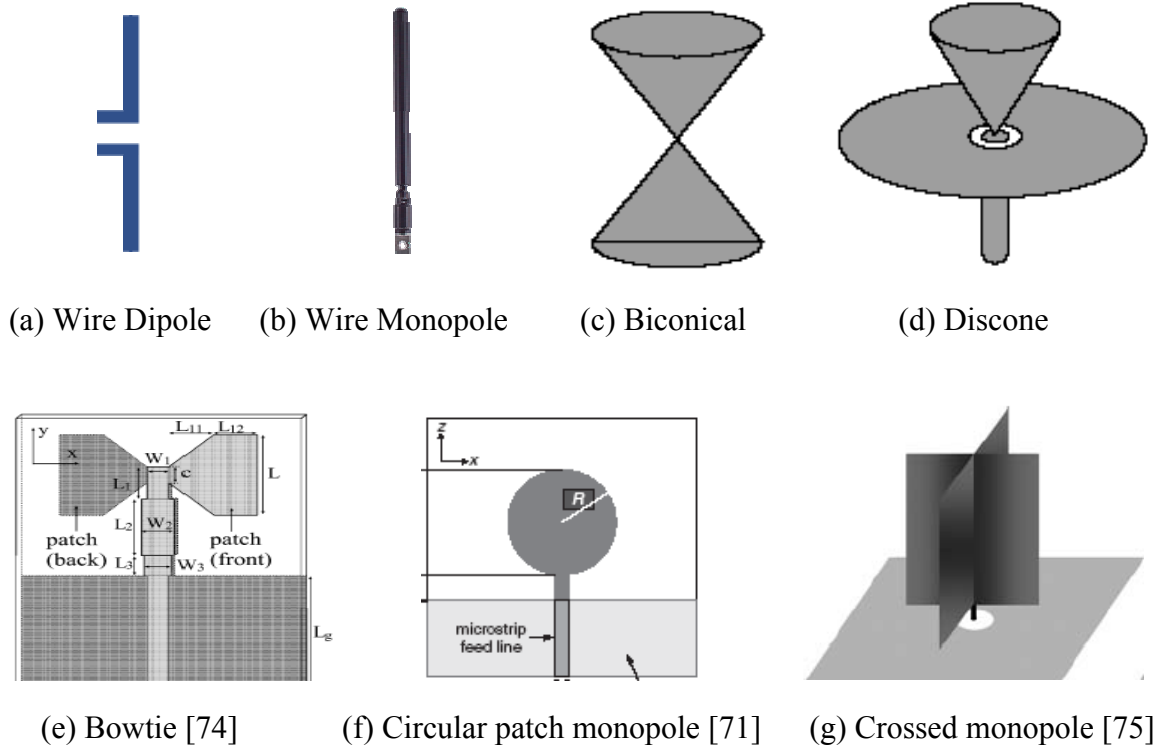


Fig.5.1 Antenna structures that feature omni-directional radiation pattern.

Wire dipoles and monopoles are the simplest structures that can produce omni-directional pattern, but limited to a very narrow bandwidth. Meanwhile, conical antennas, constructed by Sir Oliver Lodge in 1897, have relatively stable phase centers with broad impedance bandwidths due to the excitation of transverse electromagnetic (TEM) modes, as mentioned by John D. Kraus [76]. However, the conical antennas shown in Fig. 5.1(c, d) are seldom used in portable devices due to their size and cost constraints. Moreover, the bowtie antenna shown in Fig. 5.1 (e) features a planar structure and wideband

performance. However, the omni-direction radiation pattern is compromised in its H-plane [74]. Figure 5.1(f) shows a planar monopole fed by a microstrip line. The radiator is a circular patch and etched onto a dielectric substrate. The ground plane is etched onto the opposite side of the PCB. Such antenna can easily be integrated into the circuits with a compact design and are able to be fabricated at a low manufacturing cost. The drawback of this structure is that their gain in the H plane significantly degrades at higher operating frequencies, i.e. above 9 GHz [71]. In order to alleviate this problem, two planar radiators with a crossed configuration have been used to form a monopole as shown in Fig. 5.1(g) [75]. However, the crossed monopole has a larger volume than the planar monopole and is difficult to fabricate.

Table 5.1 summaries and compares the pros and cons of the above mentioned antenna structures. A circular patch monopole with a wide band performance, good omni-directional radiation pattern, compact size and low fabrication cost is the best candidate for the transmitting antenna in an UWB localization system.

Table 5.1– Comparison of Antennas with omni-directional radiation pattern

Antenna Type	Bandwidth	Omni-directional performance	Size	Cost
Wire Dipole/ Monopole	Very Narrow	Excellent	Compact	Low
Conical	Wide	Excellent	Bulky	High
Bowtie	Wide	Moderate	Compact	Low
Circular patch monopole	Wide	Good	Compact	Low
Crossed Monopole	Wide	Excellent	Bulky	Low

The radiator of the planar monopole antenna, which may be of any shape, is optimized to cover the UWB bandwidth and to miniaturize the antenna [77]. Experiments have been carried out on various planar monopole antennas such as circular, elliptical, rectangular, square, and hexagonal disc monopoles and show that an elliptical disc with certain ellipticity ratio can provide a maximum bandwidth ratio of more than 1:10.7 for $VSWR < 2$ [78]. Thus, elliptical patch is adapted here as the radiator in the proposed UWB planar monopole design.

The idea of using a modified ground plane to improve the omni-directional radiation pattern originated from the design of a modified reflector shown in Fig. 5.2 [79]. The so-called ultra-wideband reflectors ensure the frequency independence of the antenna radiation patterns. They can have a shape of a corner or a pyramid corresponding to the number of dipole branches as indicated in Fig. 5.2. The ultra-wideband reflectors create new radiation patterns (omni-directional, V-shaped, and conical). A good ultra-wideband impedance match can also be achieved with a properly designed reflector. The idea of modification of ground plane is then adapted for use in the proposed UWB planar monopole design to improve the omni-directional radiation pattern.

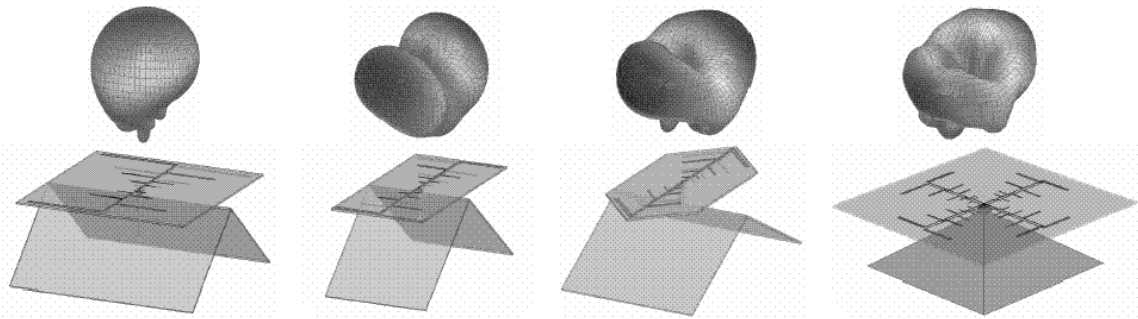


Fig.5.2 Different logarithmic-periodic dipole structures with different ultra-wideband metal reflectors. (after [79])

5.1.2 Proposed Monopole Antenna Structure and Performance

The proposed monopole antenna is shown in Fig.5.3. It is printed on Rogers 4003 substrate with a 20 mil thickness and a relative dielectric constant of 3.38, and the monopole antenna has very compact dimensions ($W \times L$). On the top surface of the substrate, an elliptical disc patch is printed with an ellipticity ratio (a/b) of 1.2. The disc is fed by a 50 Ohm microstrip feed line with a fixed width of Wl . On the bottom side of the substrate, the shape of the ground plane was modified to obtain a significantly improved omni-directional radiation pattern over a wide band. The optimized ground plane is comprised of a triangle shape combined with an ellipse that has an ellipticity ratio of 1.8 as shown in Fig.5.3. The elliptical patch on the top surface is spaced a distance h from the modified ground plane. The dimensions of the feed gap distance h and the width of the modified ground plane W are important parameters in determining the sensitivity of the wide band impedance matching of the monopole antenna. The optimal dimensions of the designed antenna are as follows: $L = 1000$ mil, $W = 450$ mil, $Wl = 47$ mil, $h = 40$ mil, $Ll = 200$ mil, $b = 130$ mil, and $\theta = 10^\circ$.

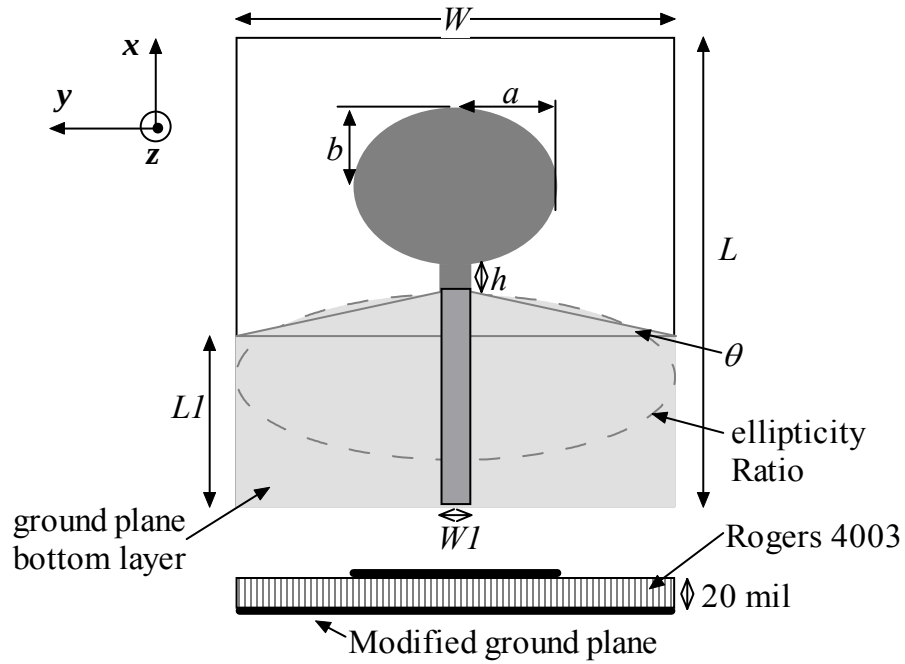


Fig.5.3 Configuration of an elliptical disc monopole with a modified ground plane.

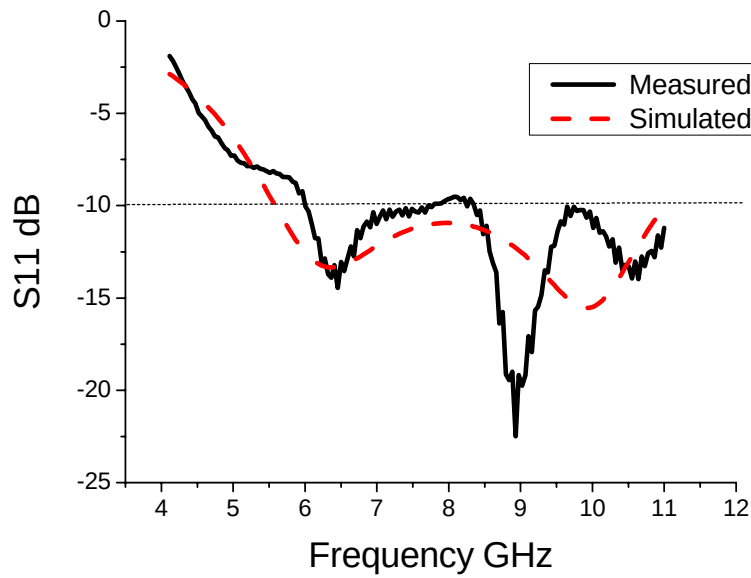


Fig.5.4 Measured and simulated return loss for the elliptical patch with a modified ground plane.

The structure was simulated using Ansoft high-frequency structure simulator (HFSS). Good agreement has been achieved between the measured and simulated return loss results of the elliptical disc monopole with a modified ground plane, as shown in Fig.5.4.

Figure 5.5 shows a comparison of the measured return loss performance between a circular and an elliptical disc patch with the conventional square ground plane. The ellipticity ratio of the elliptical patch is optimized to give the maximum bandwidth performance. The return loss at the lower frequency end has been further optimized by modifying the ground plane as shown in Fig. 5.6.

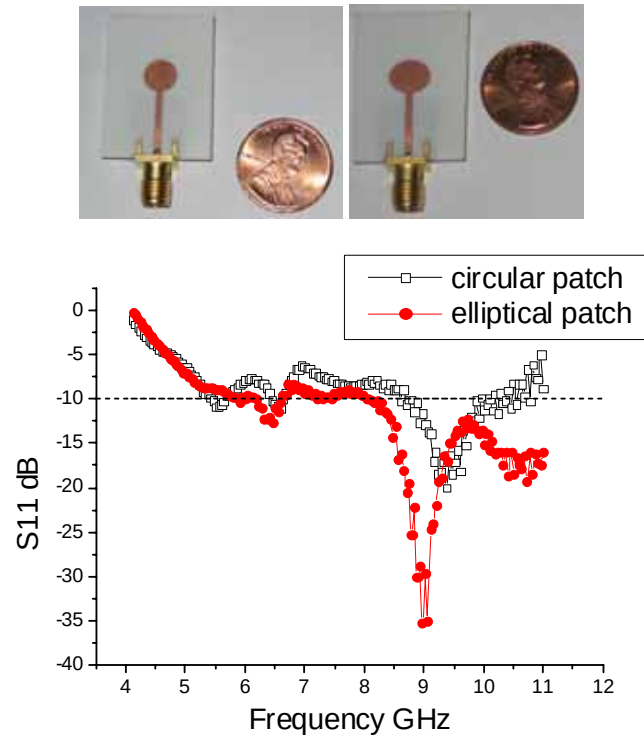


Fig.5.5 Measured return loss for the circular and elliptical patch.

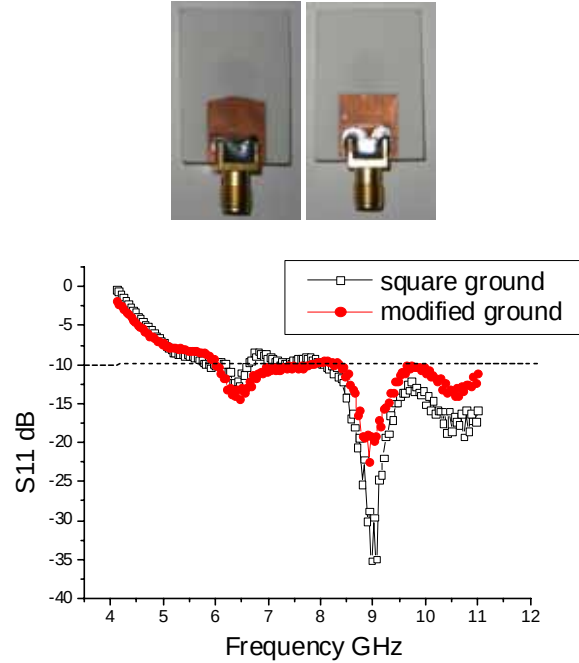


Fig.5.6 Measured return loss for an elliptical patch with a square and modified ground.

Improved omni-directional radiation patterns have been achieved by using a modified ground plane. Initially, a circular disc monopole with a conventional square ground plane was used, and its radiation pattern (shown in Fig. 5.7) in the y - z plane (H -plane) indicated a significant pattern distortion upon increasing the frequency. For example, at 10 GHz ripples greater than 12 dB can be seen in Fig. 5.7c. However, after modifying the ground plane, radiation pattern remains omni-directional from 6 to 10 GHz, and the distortion has been significantly reduced to less than 5 dB up to 10 GHz as seen in Fig. 5.8c. Further analysis of the structure using HFSS indicated that the omni-directional pattern can be extended to cover higher frequency range upon modifying the ground plane of the elliptical disc monopole as shown in Fig. 5.10. It indicates that this design can be scaled down to have even wider frequency response in our region of interest.

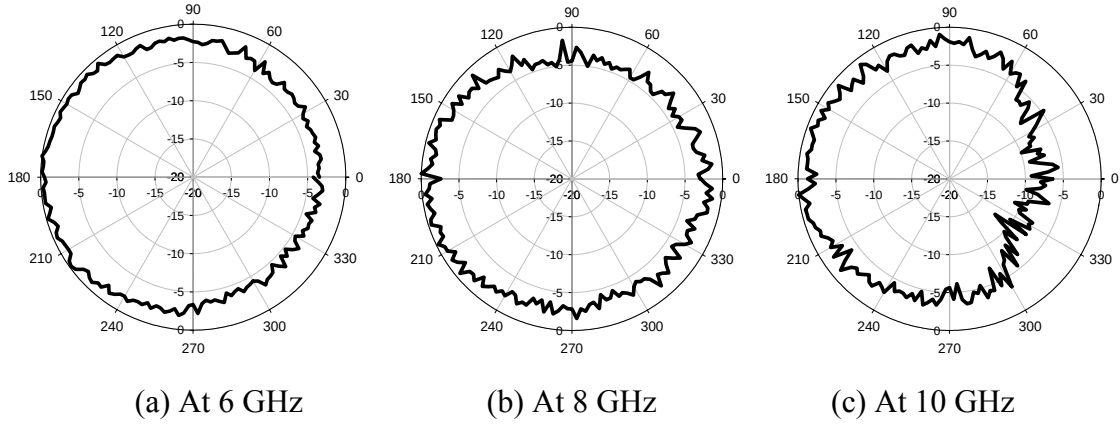


Fig.5.7 Measured y - z plane radiation patterns with a circular disc monopole and square ground.

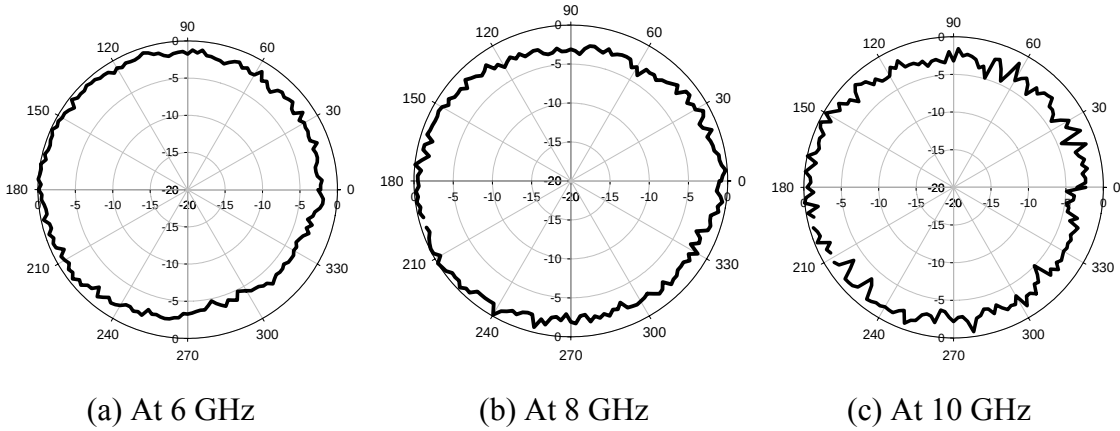


Fig.5.8 Measured y - z plane radiation patterns with an elliptical disc monopole and modified ground.

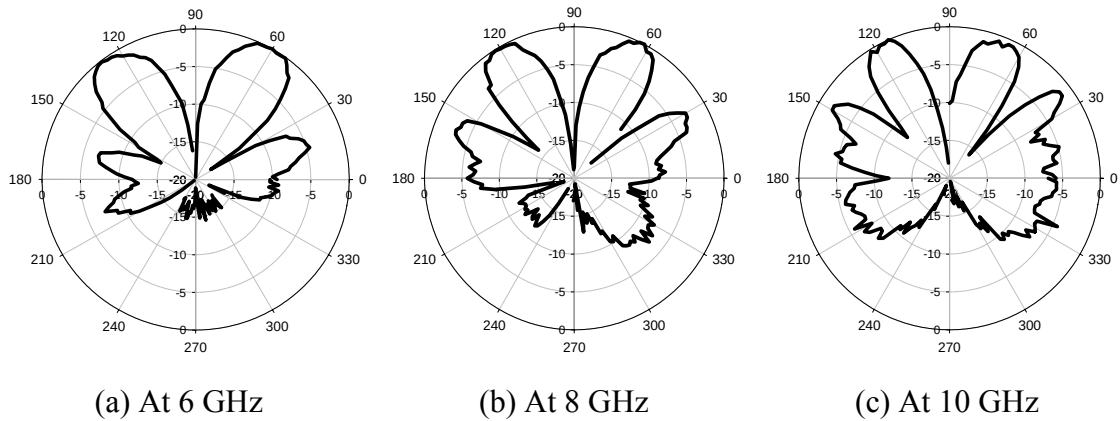


Fig.5.9 Measured x - z plane radiation patterns with an elliptical disc monopole and modified ground.

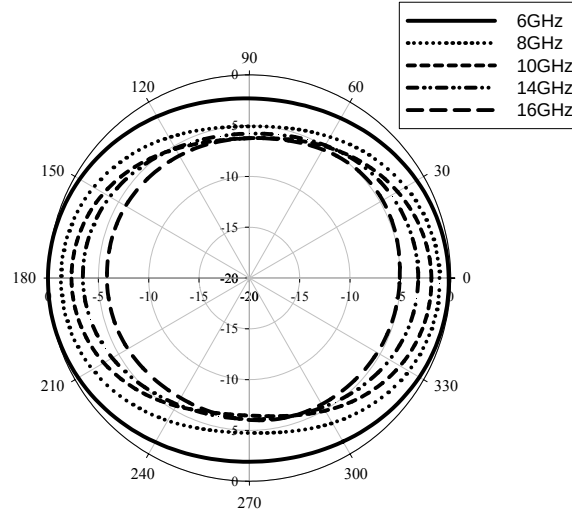


Fig.5.10 Simulated y - z plane radiation patterns with an elliptical disc monopole and modified ground.

5.1.3 Time Domain Measurements

Antennas can produce unpredictable distortion especially when transmitted narrow pulses. Hence, the developed UWB monopole with modified ground plane and elliptical patch was tested for signal integrity to assure minimal pulse distortion. In the signal integrity testing, near field impulse response measurement was carried out using the designed monopole at the transmitting end and an UWB TEM horn antenna at the receiving end (as shown in Fig. 5.11a). A 600 ps Gaussian pulse signal with a bandwidth of close to 2 GHz was up-converted to 8 GHz and then transmitted through our developed monopole. At the receiver side, the received signal using a TEM-UWB horn is then down-converted to recover the Gaussian pulse signal. Compared with a directly connection of the transmitter and receiver through a coaxial cable (Fig. 5.11b), the monopole-horn pair introduced insignificant distortion, as shown in Fig. 5.11c. Hence, our monopole design is adequate to handle such narrow pulses.

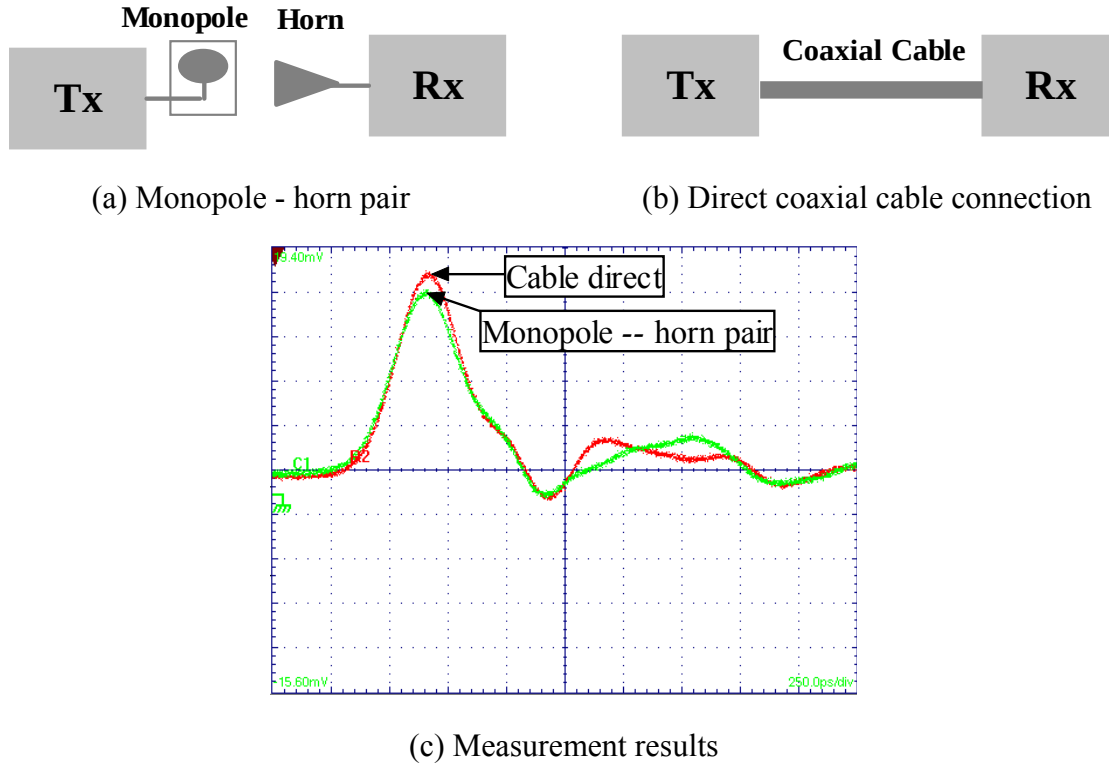


Fig.5.11 Time domain response measurement setup and results.

5.2 Receiving Antenna Design

Vivaldi antennas belong to the class of tapered slot antennas (TSA) [80], and have been widely studied and applied due to its simple structure, light weight, wideband, high efficiency, and high gain. It has been utilized in many ultra wideband (UWB) applications such as, ground penetrating radar, UWB communication systems, UWB imaging system, etc. Theoretical and experimental analysis of Vivaldi antenna characteristics can be found in [81]–[84]. Variants of Vivaldi element have been

documented [85]–[87]. Recently, Vivaldi antennas became very popular in UWB pulse transmission since they cause slight distortion to the transmitted UWB pulses [88].

The antipodal Vivaldi element, one of those variants, shown in Fig. 5.12 is built on a Roger 4003C substrate with a relative dielectric constant of 3.38 and a thickness of 0.51 mm. The optimal dimensions are as follows: $w=24$ mm, $l=45.72$ mm, and $r=11.73$ mm. Exponential tapered profile, which is a common shape to obtain a wideband 10dB impedance match, is used for this element. This antenna is fed using a microstrip line through parallel-strips transition, and its exponential taper is determined by

$$y = c_1 e^{Rx} + c_2 \quad (5.1)$$

where R is the opening rate, c_1 and c_2 are determined by the coordinates of the first and last point of the exponential curve. The simulated and measured far field radiation patterns of a single element at 10 GHz are illustrated in Fig. 5.13. The measured maximum gain is approximately 5 dB across the operating bandwidth.

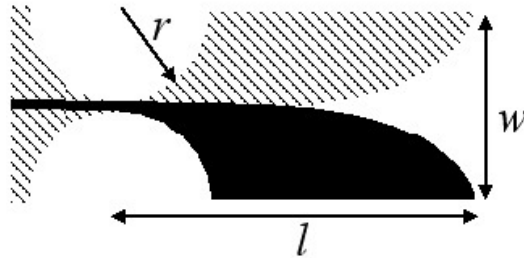


Fig.5.12 The antipodal Vivaldi element.

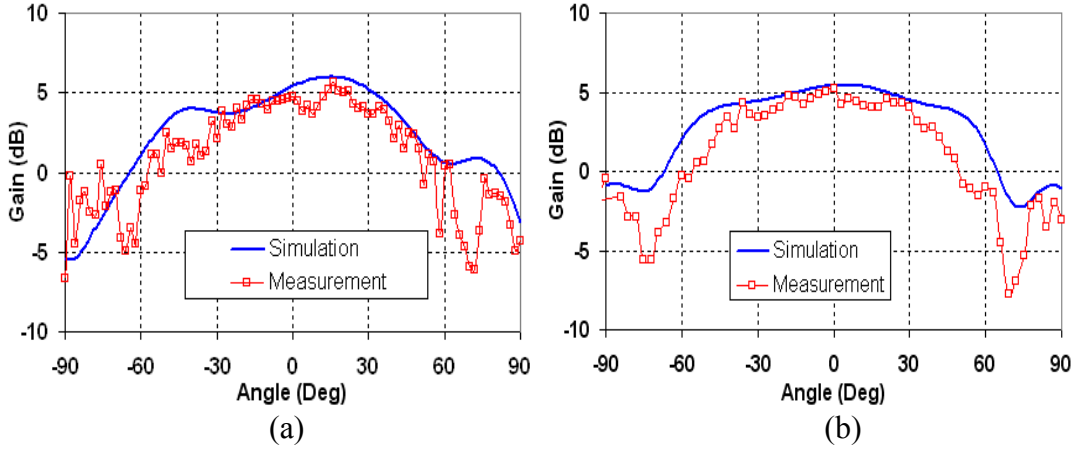


Fig.5.13 Pattern of Single Element at 10 GHz, (a) E-plane, (b) H-plane.

5.3 Antenna Phase Center Variation

The frequency domain phase center is defined as 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 center is used to describe the phase center in a limited section of the radiation pattern. However, for an UWB localization applications, although the variance of the frequency domain phase center position with the angle gives some indication of how well the antenna will perform in the time domain, a better and more intuitive definition would be to define a phase center for pulse radiation or a *time domain phase center*. The definition of the time domain phase center is very similar to the one given above for the frequency domain phase center except that instead of measuring the phase of the received signal, we measure the time of arrival of the peak of the transmitted pulse. Accounting for the antenna phase center variation at the transmitters and receivers is critical for high performance in precise accuracy localization systems. Ideally all frequencies contained in the pulse are radiated from the same point of

the UWB antenna and thus would have a fixed phase center [89]. In this case, all frequencies travel the same distance within the same time, and the pulse can be received undistorted.

In practice, however, the phase center varies with both frequency and direction. For localization systems that require a mm-range accuracy, this could result in an unacceptable localization errors. For example, to compensate for phase center variation in GPS antennas, automated high precision robots are used in a calibration procedure to move a GPS antenna into 6000 – 8000 distinct orientations [90]. In the case of our transmitting antenna, which is an UWB monopole, phase center variation is less than 1 mm and is considered negligible (both across the frequency band from 6 – 10 GHz and as the azimuth angle is varied). Phase center variation along the broadside direction was simulated to estimate the axial position of the Vivaldi phase center. Figure 5.14 shows the simulated phase center variation over the desired frequency band at broadside using CST software, with the original point set at the input of the Vivaldi antenna, as shown in Fig. 5.15. The average phase center position across the frequency band of 6 – 10 GHz is obtained at 39.5 mm from the feed point which is later used as the “apparent phase center” in directivity-dependent phase center measurements.

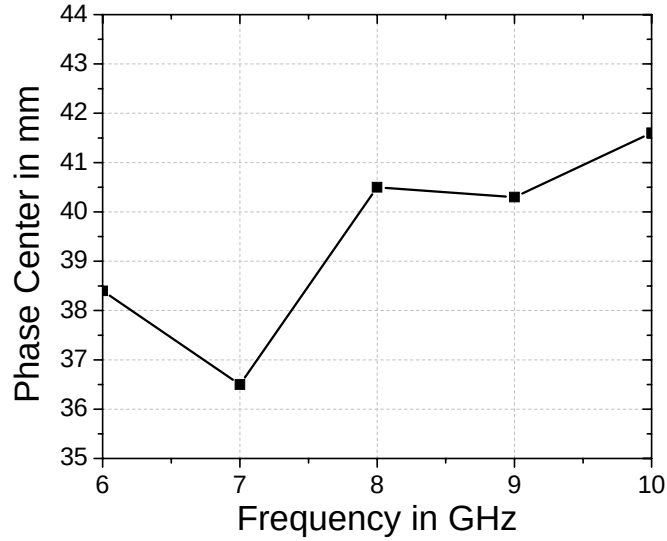


Fig.5. 14 Simulation of broadside Vivaldi phase center location versus frequency.

Since the UWB pulse contains broadband frequency information, a more accurate method for defining the phase center variation of the Vivaldi antenna is to employ time domain techniques. As shown in Fig. 5.15, an experiment was setup in an anechoic chamber to quantify how the phase center is affected by the directivity based on time domain measurements. Both the transmitting and receiving Vivaldi antennas were put face to face and separated by a distance of 1.5 m. The receiving antenna was rotated around the calculated “apparent phase center” (at 39.5 mm, shown in Fig. 5.15) from -45° to $+45^\circ$ at 5° per step. The apparent phase center was tracked on the receiving Vivaldi antenna as it was rotated from -45° to $+45^\circ$ with an optically tracked probe. These reference points from the optical system were used to calculate the actual center of rotation during the experiment. This allowed changes in the actual phase center as the receiving antenna was rotated to be separated from the physical movement of the apparent phase center, shown in Fig. 5.15.

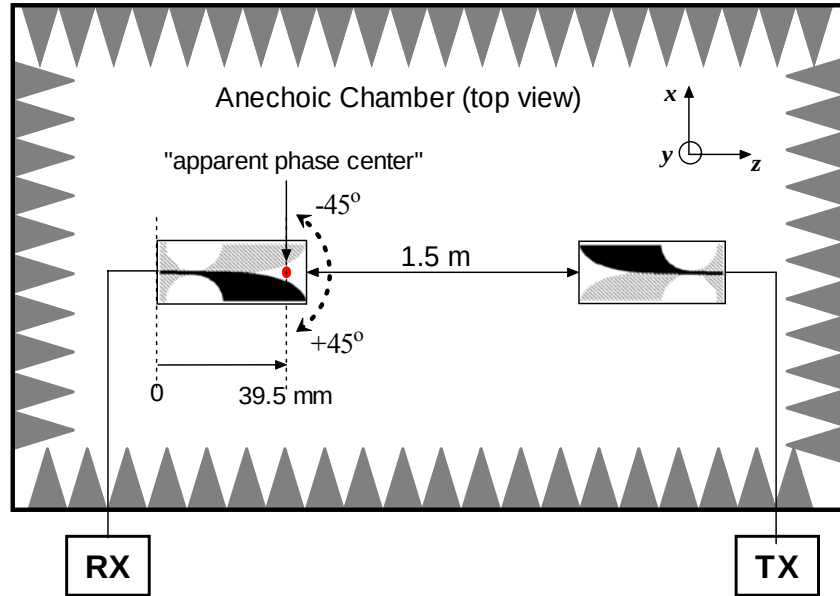


Fig.5.15 Experimental setup of Vivaldi antennas in an anechoic chamber used to measure the Vivaldi antenna directivity-dependent phase center variation.

Figure 5.16 shows the measured phase center displacement for both the E and H plane cuts. As shown in Fig. 5.16, the measured phase center variation versus the rotating angle indicates a small phase center variation of less than 2 mm within $\pm 20^\circ$ while the variation degrades dramatically with an angle greater than 30° . Calibration techniques exist to get rid of the phase center error in a practical system. These techniques are based on adding correction factors for the angular variation of the Vivaldi antenna phase center position into a set of nonlinear equations that are solved to determine the position of the transmitter. These correction factors are determined using a number of measurements and calibration against an optical localization system. The calibration procedure can be translated into TDOA based positioning systems where AOA effects result in degradation of the system accuracy. [91]

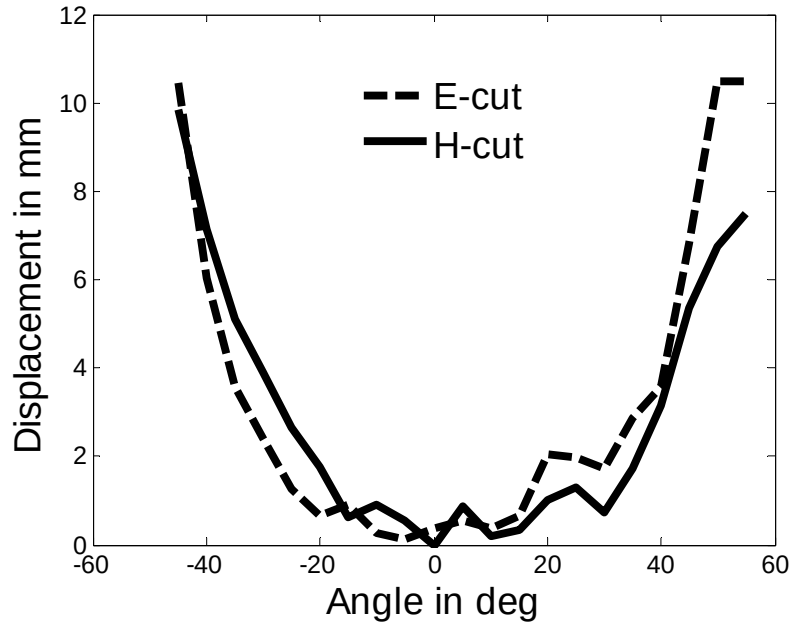


Fig.5.16 Measured Vivaldi phase center error versus angle for E-cut and H-cut.

5.4 Vivaldi Antenna Phase Center Improvement

To obtain symmetric patterns, it is required to significantly narrow down the H-plane beamwidth by using an H-plane array rather than a single element. Additionally, Vivaldi antennas have an unacceptable phase center variation in the H-plane [92], which may not have significant effects in pulse transmission, but could cause a noticeable error in the high precision localization applications since the phase center error will directly translate into ranging error in the TDOA algorithm. To accomplish this task, we have proposed a Vivaldi antenna with a protruded dielectric rod to improve the gain, narrow down the H-plane beamwidth, and minimize the phase center variations with frequency [93]. A sample antenna was fabricated and measured, and its preliminary measured results are very promising, and are in good agreement with our simulated results, design details will be given in the Appendix.

5.5 Conclusion

An elliptical disc monopole antenna with a modified ground plane has been developed. Good agreement between the simulated and measured results has been achieved. The antenna shows an excellent omni-directional radiation pattern, as well as satisfactory input impedance match over an ultra-wide bandwidth (almost 3:1 bandwidth). In addition, time domain impulse response experiments have demonstrated that the proposed UWB monopole introduces minimal pulse distortion. A novel time-domain phase center measurement has been established to test the phase center variation versus angle over a wide bandwidth. A new method to improve the performance of the Vivaldi antenna by inserting (protruding) a polystyrene rod along its flared slot has been presented. By optimizing the dimensions of the rod, both the radiation pattern and the gain of the antenna can be controlled. Use of the protruded dielectric rods has led to a phase center position “skew” reduction between the E- and H-planes, and minimal frequency dependence, which is very important for high precision UWB localization applications. In our current setup of 3D localization system, the volume of movement available to the tag is intentionally set within $\pm 20^\circ$ for each base station Vivaldi receiving antenna boresight, so the phase center related error is negligible according to our phase center measurement results shown in Fig. 5.16. When the tag movement exceeds $\pm 30^\circ$ of the Vivaldi antenna boresight, the phase center of the Vivaldi antenna will shift by more than 3 mm and such effect must be taken into account. In this scenario, either the phase center calibration technique or Vivaldi-rod antenna as indicated in the Appendix should be used to minimize the error from phase center variation.

Chapter 6

System Level Analysis

So far the hardware development comprising of the main components of the UWB localization system, such as the pulse generator, the sub-sampling mixer and the transmitting and receiving UWB antennas have been investigated in detail in the previous chapters. This chapter will focus on the full positioning system performance that has integrated the aforementioned hardware.

The overall system level analysis includes:

- ◆ A link budget – where the achievable distance between the tag and receiver in an indoor environment has been studied, both the simulation and measurement results show that the distance can exceed 5 m with a transmitted signal power level in compliance with the FCC regulations;
- ◆ A time budget – where the bottleneck of the time requirement in the system is analyzed, the total number of tags that the real time system can afford using time division multiple access (TDMA) algorithms are also given;
- ◆ A power budget – where the power consumption of the active components in the tag will be given, with an estimate of the lifetime of a tag with a coin battery, suggestions of saving power will also be discussed.

6.1 Link Budget

Path loss is considered to be a fundamental parameter in channel modeling as it plays a key role in the link budget analysis [94]. Also, path loss serves as an input for the

mean value of a large-scale fading, which in turn determines the small-scale fading characteristics. Measurements were made at distances between 1 and 3.8m in intervals of 0.2 m in a laboratory environment and a total of 15 locations were measured. The maximum measured distance is limited to 3.8 m due to the available cable length being less than 4 m. According to [95], the measured path loss at 1 m in a real environment can be modeled as a free-space path and its loss can be estimated using the Friis free-space equation given by

$$PL_{FS}(d) = \frac{G_{tx} G_{rx} \lambda^2}{(4\pi)^2 d^2} \quad (6.1)$$

where $PL_{FS}(d)$ denotes the free space path loss, λ is the wavelength of the carrier frequency in meters, d is the distance in meters, and G_{tx} and G_{rx} denote the transmitter and receiver antenna gains, which are about 2 dB and 6 dB respectively. The transmitting and receiving antennas were each set to a height of around 1.4 m for all measurements. The path loss in dB in the channel has a log-norm distribution with a mean that linearly changes with the distance and is modeled as:

$$PL_{dB}(d) = PL_{dB}(d_0) + 10\alpha \log\left(\frac{d}{d_0}\right) \quad (6.2)$$

where $PL_{dB}(d_0)$ denotes the mean path loss at an initial position d_0 which is 1 m for this case, and $10\alpha \log(d/d_0)$ denotes the mean path loss referenced to d_0 . The path loss exponent α is set to 1.65 for omni-directional to directional antenna cases [95]. As shown in Fig. 6.1, an UWB path loss model has been created using the ADS Ptolemy simulator. The multipath effects can also be included in the path loss simulation model. The ADS

simulation results are then compared to the calculated results based on both equation (6.2), and the measured results. Excellent agreement has been achieved as shown in Fig. 6.2, which plots the path loss versus distance between the transmitting and receiving antennas.

The signal to noise ratio (SNR) is defined as the ratio of the RMS signal voltage to the RMS noise voltage, as given by

$$SNR = 20 \log_{10} \frac{V_{rms,signal}}{V_{rms,noise}} \quad (6.3)$$

As noticed from Fig. 6.2, a Gaussian pulse signal with a SNR of about 20 dB was detected at a distance of 3.8 m between the transmitting and receiving antennas. Such SNR is required to achieve a mm-range accuracy.

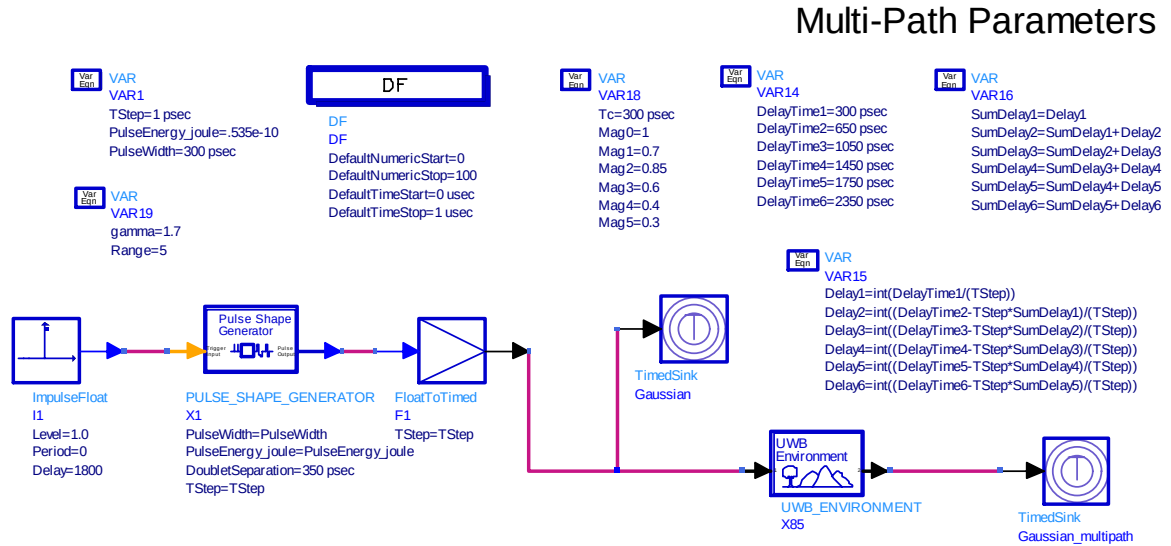


Fig.6.1 Path loss simulation using ADS Ptolemy simulator.

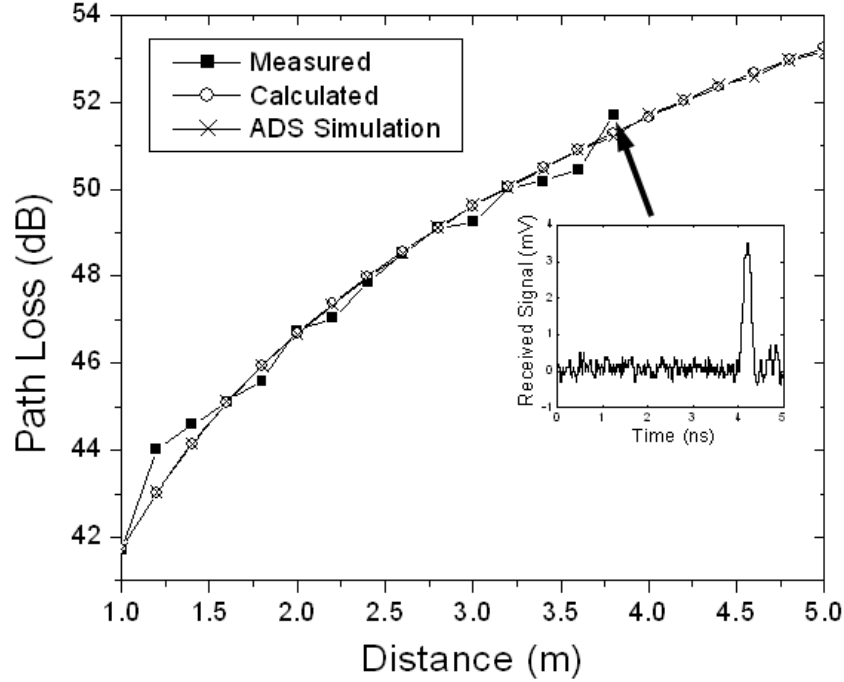


Fig.6.2 Path loss versus distance between the transmitting and receiving antennas.

A broadband LNA (HMC-C002) cascaded with a medium amplifier (HMC441) from Hittite has been used between the receiving antenna and the down-converting mixer. Both amplifiers are in their linear region during operation. The two stages of amplification increase the signal by a 27 dB while the down-converter Mixer, LPF and the sub-sampler has a total loss of 13 dB. Thus, the overall RF gain of the RF front end is 14 dB. As shown in Fig. 6.3, the RF front-end model is created in Agilent AppCAD software [96] to calculate both the overall noise figure and the input IP3, which are 2.34 dB and -10.48 dBm respectively.

		Stage 1	Stage 2	Stage 3	Stage 4	Stage 5
						
Stage Data	Units					
Stage Name:		LNA, HMC C002	MPA, HMC441	Mixer, HMC220	LPF	Sampler
Noise Figure	dB	2.0	4.5	7	2	4
Gain	dB	13	14	-7	-2	-4
Output IP3	dBm	26	32	10	100	15

Fig.6.3 System Receiver Noise Figure and IP3 Calculations using Agilent AppCAD.

The UWB localization system link budget is calculated and the key parameters are listed in Table 6.1. The equations (6.4 - 6.6) are used to provide the necessary values needed for the link budget calculations. These calculations followed similar steps to that in [97],

$$Tx_{average} = EIRP \cdot BW \text{ (in dBm)} = -41.3 + 10 \cdot \log 6000 = -3.52 \text{ dBm} \quad (6.4)$$

$$T_e = (F - 1) \cdot T_0 = (10^{2.34/10} - 1) \cdot 290 = 207 \text{ K} \quad (6.5)$$

$$N_{floor} = 10 \cdot \log(k \cdot T_e \cdot BW) = 10 \cdot \log \frac{1.38 \times 10^{-23} \times 207 \times 6 \times 10^9}{10^{-3}} = -77.66 \text{ dBm} \quad (6.6)$$

where EIRP is the Equivalent isotropically radiated power with a value of -41.3 dBm/MHz based on FCC regulations, BW is the system bandwidth (set to be 6 GHz with a double side band modulation scheme), T_e is the noise temperature, F is the overall receiver noise factor, N_{floor} is the total noise power at the receiver side, and k is Boltzmann constant.

Assuming SNR of 20 dB is required for the leading edge algorithm [98], the receiver sensitivity is calculated to be -57.66 dBm. Therefore, a link margin of 5.14 dB and 1.14 dB has been achieved for a 3m and 5m distance between the transmitter and

receiver respectively. Meanwhile, the dynamic range of the system is mainly limited by the sub-sampler, which has been previously described in Chapter 4, and it has a measured tangent sensitivity of -45 dBm and a P_{1dB} point of about 5 dBm [30]. Thus, the dynamic range of the system is over 50 dB. The proposed system can be extended to be more than 5 m considering the system link margin of 1.14 dB at 5 m. Further hardware refinements that can improve the link margin include: increasing the antenna gain, decreasing the LNA noise figure and reducing the system bandwidth. However, reducing the system bandwidth will give rise to a higher localization error, as will be demonstrated experimentally in Chapter 10.

Table 6.1– UWB Link Budget

Carrier Frequency		8 GHz	
Bandwidth		6 GHz	
Antenna Gain	Transmitter	2 dB	
	Receiver	6 dB	
Distance		3 m	5 m
Average Transmitted Power (P_t)		-3.52 dBm	-3.52 dBm
Total Path Loss (PL)		49 dB	53 dB
Average Received Power ($P_r = P_t - PL$)		-52.52 dBm	-56.52 dBm
Receiver Noise Figure (NF)		2.34 dB	2.34 dB
Total Noise Power (N_{floor})		-77.66 dBm	-77.66 dBm
System Required SNR		20 dB	20 dB
Receiver Sensitivity ($S = N_{floor} + \text{SNR}$)		-57.66 dBm	-57.66 dBm
Link Margin ($M = P_r - S$)		5.14 dB	1.14 dB

6.2 Time Budget

The bottleneck of the speed limitation in the system is from the equivalent time sampling procedure. In order to study the time budget of the system, the equivalent time sampling process must be fully understood.

6.2.1 Equivalent Time Sampling

Figure 6.4 depicts how equivalent-time sampling works: Pulses with a pulse repetition frequency PRF, at 10 MHz for example, are sent by the transmitter. These pulses have pulse duration of 300 ps and are generated by the developed pulse generator stated in Chapter 3. The sampler has to be turned on after a little longer time of Δt than the pulse repetition time (100 ns for PRF = 10 MHz). This can be accomplished by setting a slightly lower LO strobe frequency (e.g., $f_{LO} = 9.999$ MHz, $\Delta f = f_{PRF} - f_{LO} = 1$ kHz). The strobe clock can be easily frequency shifted with a direct digital synthesizer (DDS). At each period of the pulse, a sample is taken, which is marked by a dot in Fig. 6.4. The accumulated samples form the down converted pulses, with the same pulse shape as the input signal, but with an expanded time scale.

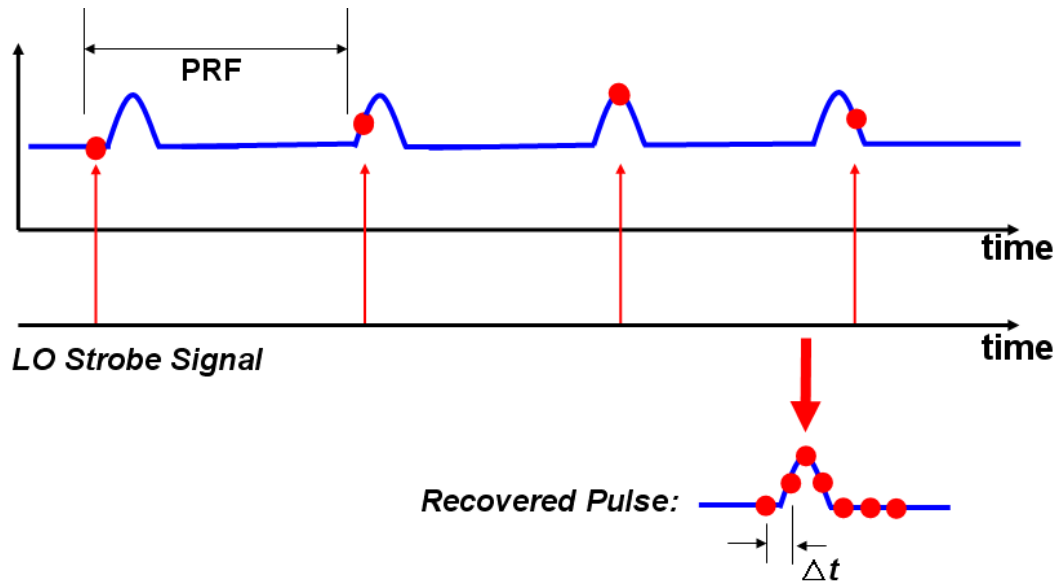


Fig.6.4 Principle of the equivalent time sampling.

Because the equivalent-time sampling requires a repetitive RF input signal, it has certain restrictions, for example, the equivalent-time cannot create a meaningful display from a single-shot event, hence the RF signal must repeat identically each time or the sub-sampled waveform will be distorted and meaningless. In our system, the transmitted 300 ps pulse train with a fixed PRF of a 10 MHz has satisfied the above mentioned restrictions.

6.2.2 Acquisition Time

For repetitive input signals, the equivalent-time sampling can provide similar accuracy to that of higher cost real-time oscilloscopes. This equivalent sampling rate can be adjusted to be very high by simply tuning the Δf , for example, $\Delta f = 1$ kHz with a signal PRF = 10 MHz corresponds to a Δt of 10 ps and an equivalent sampling rate of

100 GSPS. However, the major drawback of the equivalent time sampling scheme is that it is very time consuming. One efficient way to reduce the acquisition time is to increase the strobe frequency f_{LO} . For example, assuming the signal PRF and Δt is fixed at 10 MHz and 10 ps respectively, increasing the f_{LO} from 9.999 MHz to 10×9.999 MHz will improve the acquisition time from 1 ms to 0.1 ms. However, the SRD based strobe generator can only achieve a maximum f_{LO} of around 35 MHz due to the carrier life time. Other techniques such as utilizing GaAs MMICs could be used to generate much faster strobe signal which has been addressed in Chapter 3.

6.2.3 Maximum Number of Tags

Table 6.2 lists the time budget for our current system assuming the system is running at 10 MHz signal PRF. The number of tags in a real time system that can be sustained is calculated assuming a real time updating rate of 24 Hz under a TDMA scheme, which allows multiple tags to share the same UWB channel by dividing the signal into different time slots.

Table 6.2– Time Budget for a 10 MHz Signal PRF using a TDMA Scheme

	Throughout	Latency
Signal and Strobe PRF Clock Offset Δf	1 or 2 kHz	1000 or 500 μ s
FPGA Leading Edge Detection	~ 100 kHz	10 μ s
FPGA/PC RS232 Interface	2.05 kHz	488 μ s
TDOA Algorithm on PC	1.33 kHz	752 μ s
Tag ON/OFF	> 10 kHz	< 100 μ s

As seen from Table 6.2, the bottleneck of the time budget is mainly caused by 1) the signal and strobe PRF clock offset; 2) FPGA/PC RS232 interface; 3) TDOA algorithm running on a PC. These effects and methods to relax those limitations are discussed below:

- Signal and strobe PRF clock offset – 1 kHz offset corresponding to an equivalent sampling rate of 100 GSPS, and a 1 ms latency; this can be relaxed by increasing the offset frequency to 2 kHz corresponding to an equivalent sampling rate of 50 GSPS. Increasing the Δf to 2 kHz will not only double the speed, but also reduce the timing scaling effects due to the clock drift as will be mentioned in Chapters 9 and 10. With 2 kHz Δf , a maximum of 83 tags can be positioned at 24 Hz (i.e., $83 \times 24 = 1992$ Hz). Further increase of the Δf to 4 kHz will further increase the speed, however, it will degrade the localization accuracy due to the low equivalent sampling rate of 25 GSPS, as oversampling should improve the signal to noise ratio. Considering the signal pulse train has a very small duty cycle (300 ps / 100 ns = 0.3%). One efficient way to improve the sampling speed without compromising the sampling resolution is to smartly switch the offset frequency Δf between the strobe signal and the incoming RF PRF, by switching the Δf to a smaller value when there is a pulse signal coming, and then to a larger value when only noise is present.
- FPGA/PC RS232 interface – the series port running at 115,200 bps. One frame of raw data from the FPGA contains 7 bytes, i.e. 56 bits, 1 byte heading, and 6 bytes of time difference between every two base stations. The RS232 can transfer the

data at a speed of $115200 / 56 \approx 2.05 \text{ kHz}$. In such a scheme, a maximum of 85 tags can be positioned at 24 Hz (i.e., $85 \times 24 = 2040 \text{ Hz}$). To achieve faster interface transfer rate, a higher speed data link can easily be employed (i.e. USB) to avoid this constraint.

- TDOA algorithm running on a PC – For each TDOA algorithm iteration the PC will cost $\sim 100 \mu\text{s}$. Typically 4 to 8 iteration are required depending on the location of the tag. Integrating the TDOA algorithm into the FPGA will significantly reduce the time cost for this portion.

6.3 Power Budget

Table 6.3 lists the power budget for the active components comprising of the tag device. As can be seen, the major source of power consumption is from the LO. To reduce the average power consumption, the tag can work at a small duty cycle, e.g. 5%. Another challenge is that, to integrate a high performance LO source into a tag, such as a phase locked dielectric resonant oscillator (PLDRO), more power than a free running oscillator would usually be required.

Table 6.3 – Tag Power Budget

	Part Number	Current (mA)	Voltage (V)	Power (mW)
LO	HMC506	77	3	231
Buffer	OPA2674	Power mode: 18	± 5.5	198
		Shutdown: 1		11
MCU	MSP430	1.7	2.2	3.74
Clock	VTC4	1.5	5	7.5

The total power consumption thus adds up to about 451 mW. With a SONY CR2477 Lithium Coin Battery 1000 mAh @ 3V, the tag could last for about 6.6 hours. However, the challenges in integrating the battery are: 1) converting the 3 V to other required voltage value, which require additional power regulating circuits; 2) the buffer unit OPA2674 requires large instantaneous driving current (about 400 mA) to trigger the SRD device.

6.4 Conclusion

In this chapter, three system level issues: link budget, time budget and power budget have been investigated. Based on the link budget analysis and path loss experiment, the transmitter and receiver can work at a distance beyond 5 meters with a link margin of more than 1 dB in an indoor environment. Further hardware refinements that can improve the link margin are: increasing the antenna gain and decreasing the LNA noise figure.

From the time budget analysis, the equivalent time sampling process has been discussed to fully understand our system. Factors that limit the system speed have been investigated, those factors are: signal and strobe PRF clock offset frequency, FPGA leading edge detection algorithm, FPGA/PC interface, TDOA algorithm in PC and tag switching time. As calculation shows, a total number of 83 tags can be handled with the current setup of our localization system. Finally, the power consumption of the tag components is given, and potential challenges in battery integration have been addressed.

Chapter 7

Leakage Cancellation in the Carrier Based UWB System

LO leakage in the carrier based UWB system is a major design concern. In our transmitter, the mixer LO-RF isolation is not sufficient to comply with FCC strict signal level rules, as can be seen in the measured transmitted power spectrum shown in Fig. 2.4. However, this leakage can be substantially reduced by using a notch filter located before the UWB transmitting antenna as long as it will not lead to unacceptable signal distortion. Therefore, various filter parameters, such as the filter order and the 3 dB rejection bandwidth have been studied to see their effects on providing sufficient band rejection level to reduce the unwanted LO leakage while minimizing the transmitted pulse dispersion.

7.1 Carrier Leakage Cancellation Background

Carrier based UWB impulse radar systems have been widely used in many areas such as UWB localization and communication systems [99-100]. However, the carrier leakage, which is due to the limited LO-RF mixer isolation, will cause many problems such as: a) signal to noise ratio reduction and data loss [101]; b) signal demodulation degradation which would cause DC offset at the receiver side [102]; c) FCC transmitted signal limit violation; d) unacceptable interference with other existing services [103].

To address these problems, the most straightforward way is to increase the LO-RF isolation of the up-converter at the transmitter side. However currently available

commercial mixers have been reported to provide a maximum LO-RF isolation of around 45 dB with an LO driving power of 13dBm [104]. Thus the LO leakage at the RF port is around -32 dBm before amplification, which significantly exceeds the FCC regulation. Another technique is to divert a part of the LO signal, upon using a phase-shifter, and then add it back to the RF-output of the mixer to cancel the LO signal [105]. However, such a technique implies very complex mixer design. Recently, many researchers have investigated the use of a band-notched UWB monopole antenna to avoid the interference with the existing WLAN system. They integrated various kinds of slots to the monopole patch to create a notch filter effect [106, 107]. Similar ideas could be applied here to prevent the carrier leakage of the impulse UWB system. However, the use of a slot filter can cause deterioration of the antenna performance with respect to its gain and efficiency [108]. Even though the slot-notched monopoles only provide limited rejection level of less than 15 dB, a multi-order band-stop filter [109] could be utilized to provide an adequate rejection level to eliminate the carrier leakage. Nevertheless, the relations between the UWB signal dispersion and the filter parameters need to be studied to optimize the band-stop filter design and minimize subsequent signal distortion. Time domain UWB signal performance analysis such as the first pulse amplitude (FPA) and time delay spread (TDS) will be carried out as they are essential parameters in developing UWB positioning and communication systems.

7.2 Motivation and System Modeling

In our experimental localization system set-up shown in Fig. 2.2, a 300-ps Gaussian pulse modulates a carrier signal centered at 8 GHz. However, the up-converter

at the transmitter side requires a high LO driving power, and the up-converter cannot provide enough LO-RF isolation, which leads to an unacceptable carrier leakage at 8 GHz, as shown in Fig. 2.4. Such leakage needs to be filtered to comply with the FCC regulation using a notch filter.

To analyze the effect of using the notch filter on the transmitted signal, an Agilent-ADS2006A CAD model has been developed as shown in Fig. 7.1. The transmitter and receiver were directly connected through a simplified channel model, i.e. the antenna effects were not included in our model to focus on the signal dispersion due to the notch filter. In our analysis, the peak amplitude of the first pulse was calculated as a function of the filter order and its associated 3-dB rejection bandwidth. As noted from Fig. 7.2, in an ideal case, the higher amplitude is obtained when using a relatively narrow rejection band as most of the useful energy remains intact. In practice, however, there are a number of limitations that prevent the utilization of extremely narrow band notch filters such as: the limited realizable Q factor in a small volume, the fabrication tolerance, and the local oscillator's aging and temperature stability drifting effects. Meanwhile, the first pulse amplitude decreases as the filter order increases, so it is required to minimize the utilized filter order. For example, for a 500 MHz filter rejection bandwidth shown in Fig. 7.2, the relative first pulse amplitude decreases from 0.8 to 0.6 as the filter order increases from 1 to 4. However, notch filters with a single filter order generally can not provide adequate rejection level to eliminate the carrier leakage.

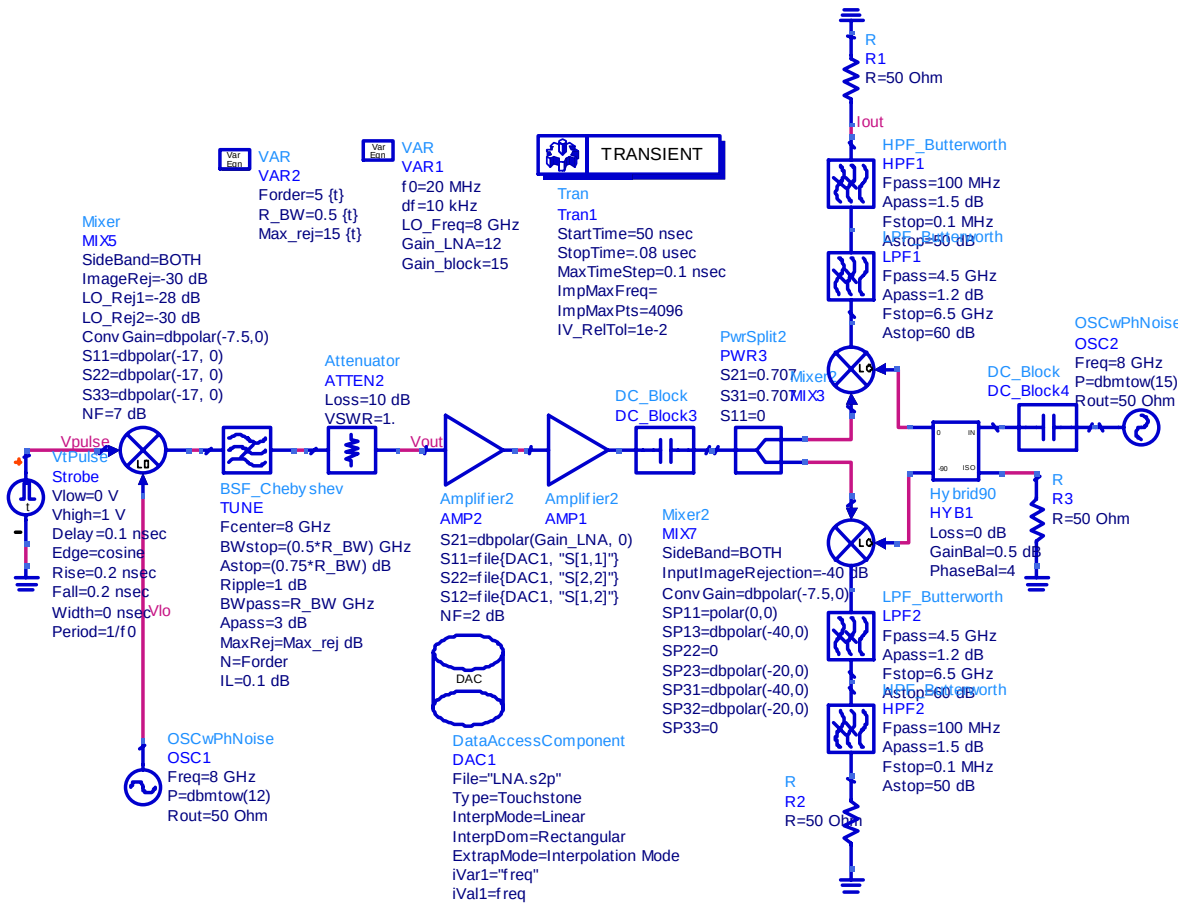


Fig.7.1 ADS Model of the simplified localization system where the signal dispersion due to antennas has not been considered.

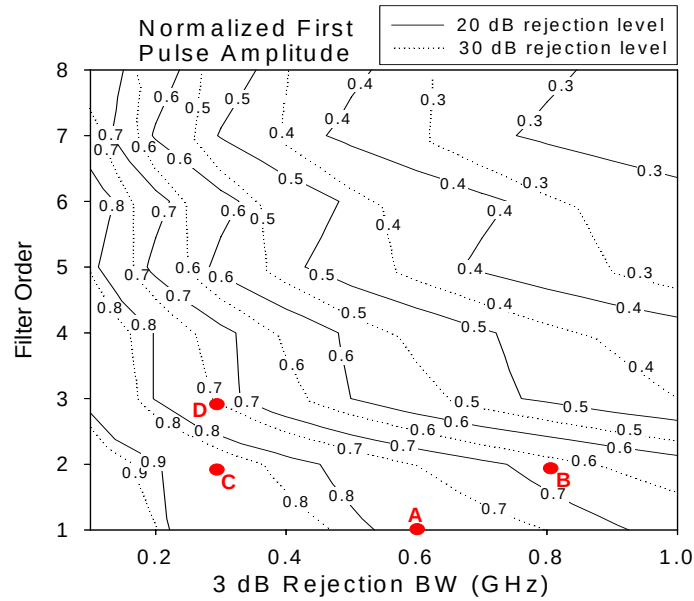


Fig.7. 2 The first pulse amplitude vs. the filter bandwidth and order: increasing the filter order for a given rejection bandwidth would lead to a less first pulse amplitude and using a wider rejection bandwidth for the same filter order would also lead to an amplitude reduction.

Another important consideration for the notch filter utilization in the carrier based UWB systems is the time delay spread. In this context, we will adapt a time delay spread definition as the time after which the pulse does not exceed -20 dBc of its first peak amplitude. As can be seen from Fig. 7.3, in general, increasing the filter order leads to an increase in the pulse time delay spread. Since a notch filter with a higher filter order features sharper notched band edges, the steeper transition in the frequency domain would result in a stronger and longer ringing effect in the time domain, giving rise to a longer pulse time delay spread. Hence, increasing the order to achieve a higher rejection level could lead to a higher dispersion, especially when the rejection bandwidth is relatively large according to Fig. 7.3. In summary, it is required to utilize a narrow band rejection filter with the possibly of a lower filter order for an adequate rejection level, i.e. about 30 dB for our carrier based UWB system.

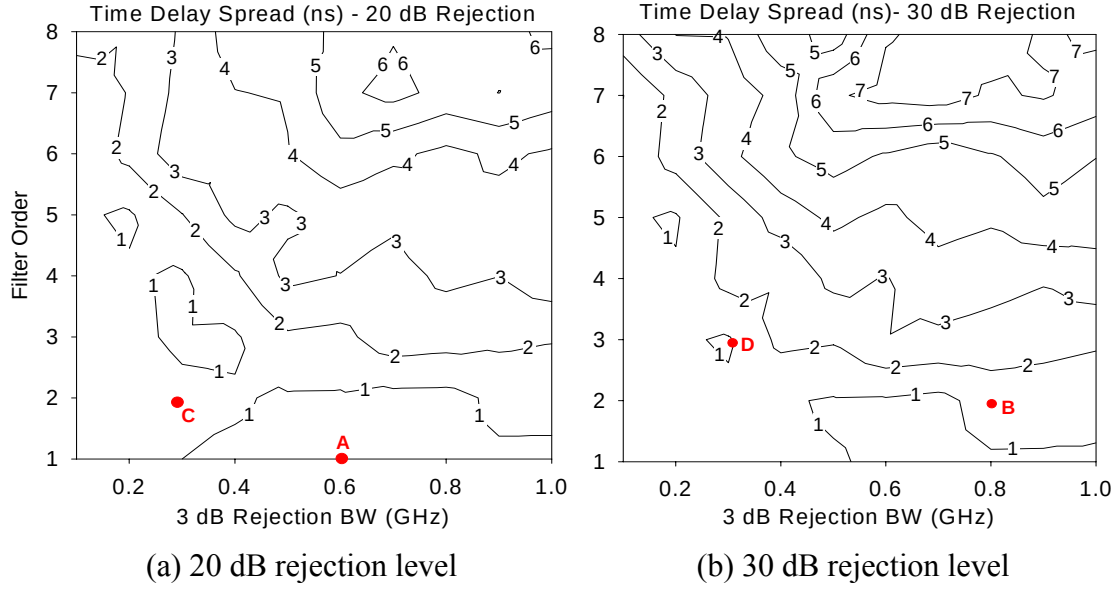


Fig.7.3 Time delay spread vs. filter bandwidth and order: for a given bandwidth the time delay spread increases upon increasing the filter order, especially for a relatively large rejection bandwidth.

7.3 Band Notched Filter Implementation

To validate the above analysis, various band-stop filters with different rejection bandwidths and filter orders were designed based on [109 - 111] and implemented here. The filters are fabricated on Rogers 4003 substrate with a 0.508 mm thickness and a relative dielectric constant of 3.38. One way to realize the band-stop filter is to integrate a compact coplanar waveguide (CPW) resonant cell (CRCC) to the CPW feed line [110], which is similar to the microstrip interdigital band-stop filter given in [111]. Fig. 7.4 shows 1st and 2nd order CRCC band-stop filters with their corresponding band-notch performances. As can be seen, 0.6 GHz and 0.8 GHz rejection bandwidths have been achieved for the 1st and 2nd order CRCC filters respectively. The 2nd order CRCC filter provides a rejection level close to a 30 dB, which is quite enough for eliminating the LO leakage indicated in Fig. 2.4.

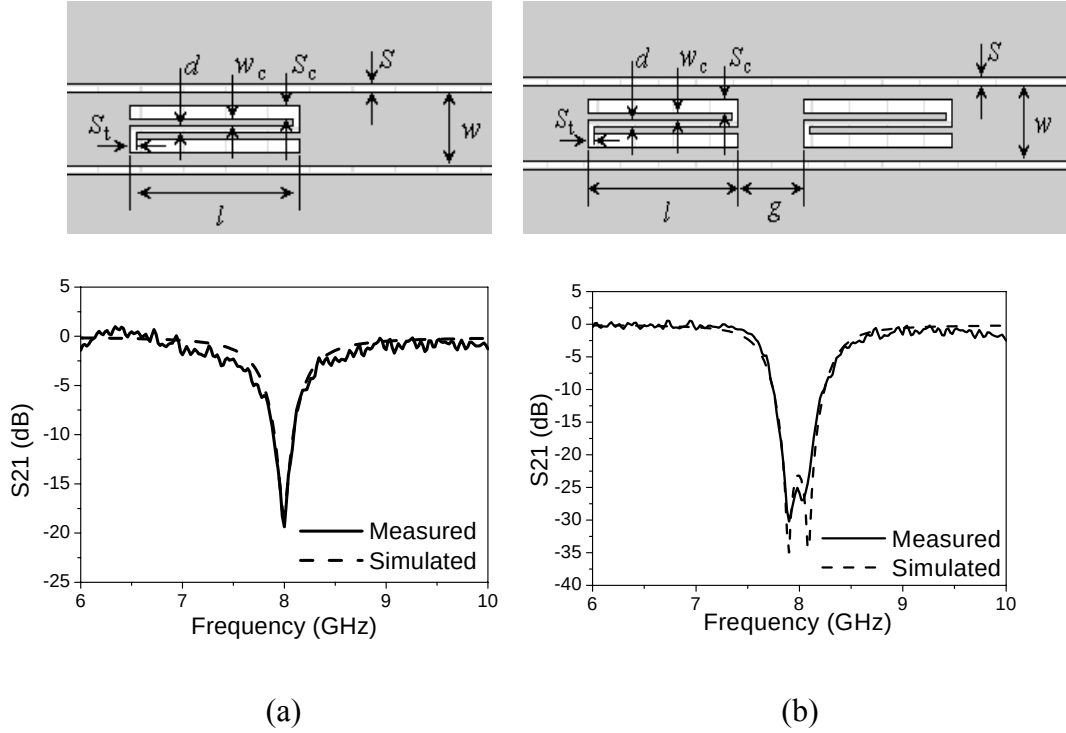


Fig.7. 4 (a) 1st order CRCC band-stop filter and its transfer characteristics; (b) 2nd order CRCC band-stop filter and its transfer characteristics. ($w = 1.7$ mm, $s = 0.2$ mm, $s_c = 0.325$ mm, $s_t = 0.15$ mm, $d = 0.15$ mm, $w_c = 0.15$ mm, $l = 4$ mm, $g = 2$ mm)

To achieve a narrower rejection bandwidth at 8 GHz, a U-slot shaped defected ground structure (DGS) of the microstrip line was utilized to construct a band-stop filter and obtain an improved Q factor [109]. Fig. 7.5 shows the 2nd and 3rd order DGS band-stop filters with their corresponding transfer characteristics. For a rejection bandwidth of approximately 0.3 GHz, the measured band rejection performance is in good agreement with the simulated results, but the center frequency is slightly shifted from 8 GHz to 8.06GHz. The sharp selectivity performance was observed in both the simulations and measurements. The slight bandwidth difference between the measured results and simulated results is due to the inaccuracy of fabricating the required 0.1 mm wide slots.

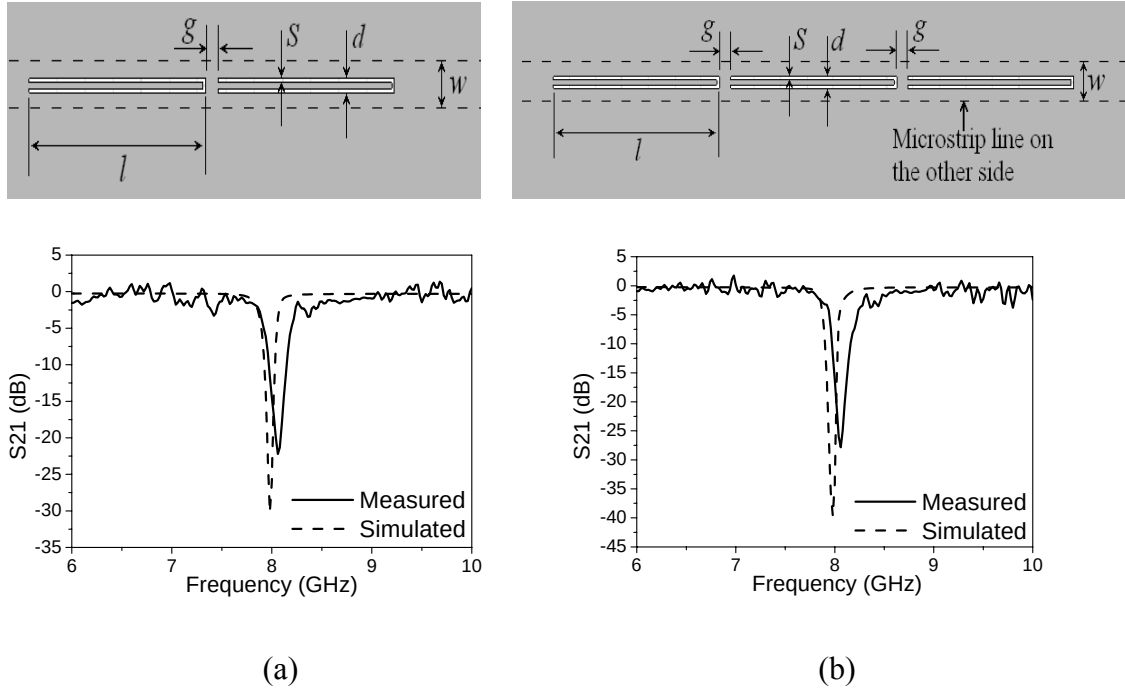


Fig.7.5 (a) 2nd order DGS band-stop filter and its transfer characteristics; (b) 3rd order CRCC band-stop filter and its transfer characteristics with a higher rejection. ($w = 1.17$ mm, $s = 0.1$ mm, $d = 0.35$ mm, $l = 6.3$ mm, $g = 0.4$ mm)

For higher rejection levels, it is necessary to increase the filter order. It is obvious that the rejection level increases as the filter order increases for both the CRCC and DGS band-stop filters as shown in Fig. 7.6.

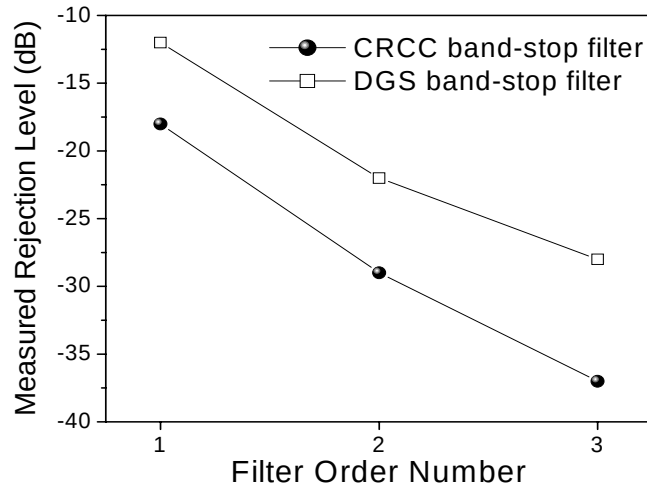


Fig.7.6 The measured rejection level vs. the filter order number.

7.4 Experimental Validation

The same experimental setup (as shown in Fig. 7.1) has been implemented in an anechoic chamber to measure the frequency and time domain responses of the various fabricated band-stop filters. To minimize the received UWB signal distortion due to antennas and clearly see the filter effects, low dispersion UWB monopole and Vivaldi antennas with a wide operating bandwidth from 5 to over 12 GHz were utilized. The UWB monopole transmitting antenna was placed in front of the UWB Vivaldi receiving antenna at a fixed distance of 0.5 m.

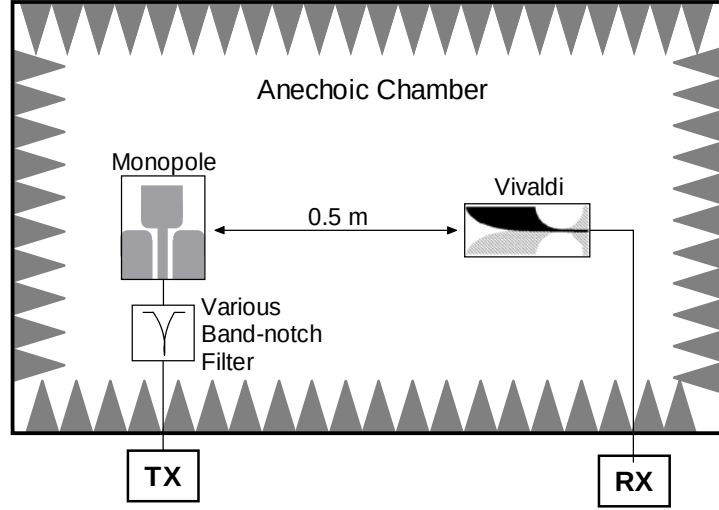


Fig.7.7 Experiment setup in an anechoic chamber.

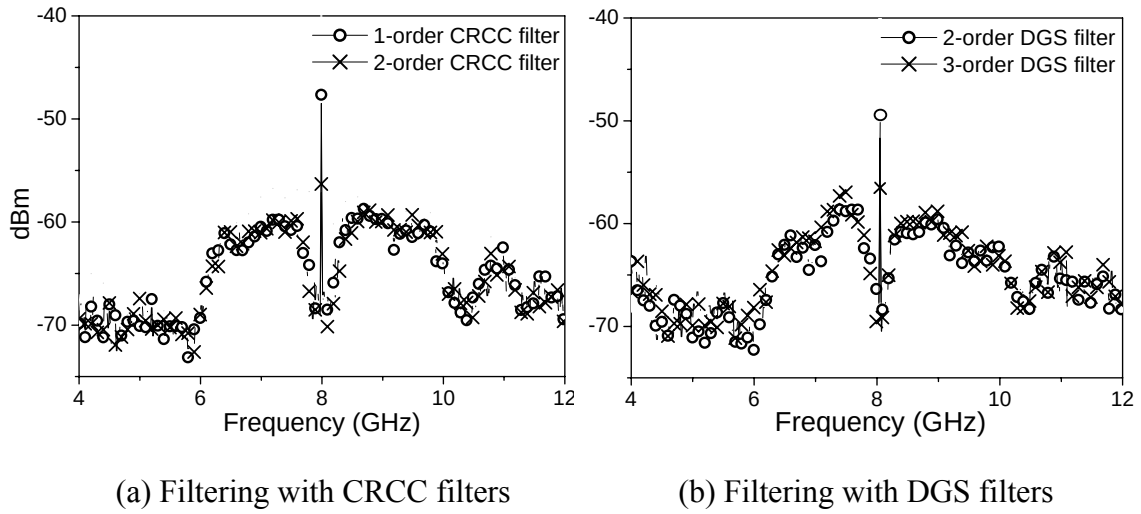


Fig.7.8 Received signal Power spectral densities without demodulation.

The spectrums of the received signals without demodulation are shown in Fig. 7.8. As can be seen, the 2nd order CRCC and 3rd order DGS notched filter provide sufficient rejections to eliminate the LO leakage as expected.

In the time domain measurements, we studied the impact of various band-stop filters on the first pulse amplitude and time delay spread. As shown in Fig. 7.9, the

measured pulse amplitudes are normalized with respect to the received signal without band-stop filter. As predicted, the 2nd order CRCC filter has the lowest pulse amplitude compared to the DGS designs due to its relatively wider rejection bandwidth. The performance of the 1st and 2nd order CRCC filters, as well as the 2nd and 3rd DGS filters are denoted by symbols A, B, C, D (in Figs. 7.2 and 7.3) respectively and has been summarized in Table 7.1. The simulated results show good agreement with the corresponding measured results. It should be noted that the measured time delay spread exhibits about a 0.2 ns longer in time than the simulated results. This extra delay is due to the extra received pulse ringing caused by other sources of hardware distortion that have not been considered in the simulation model, such as the antenna dispersion and the ringing from the 300 ps pulse generator itself. Based on Table 7.1, it is clear that the 3rd order DGS filter provides enough rejection level of (30 dB) while maintaining both a relatively large first pulse amplitude (0.8) and a small time delay spread (1.1 ns) and is suitable in our UWB system implementation.

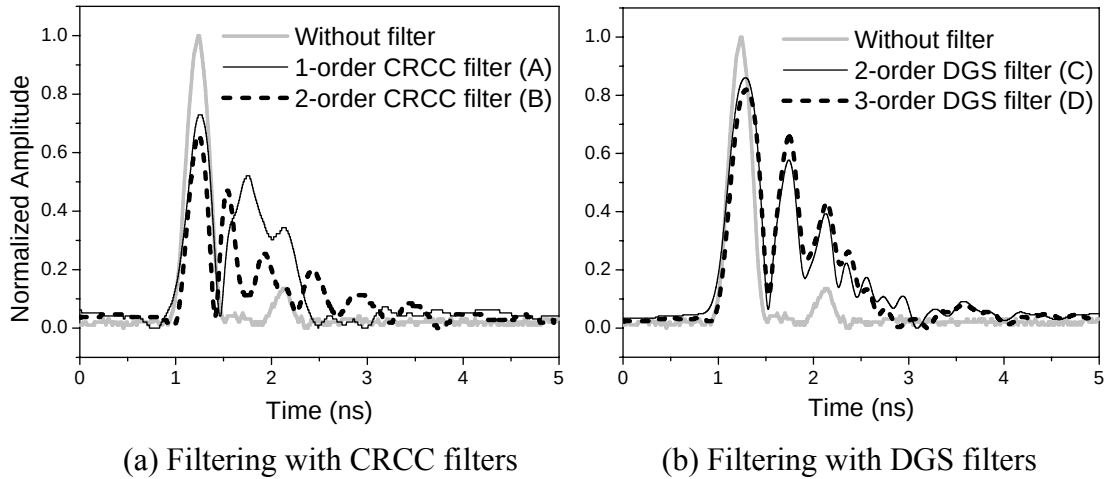


Fig.7.9 Measured time domain responses for (a) CRCC notched filters and (b) DGS notched filters.

Table 7.1– First Pulse Amplitude and Time Delay Spread for Various Notch Filters

Notch filter		Simulated	Measured	Simulated	Measured
Type	Rejection Level (dB)	first pulse amplitude	first pulse amplitude	time delay spread (ns)	time delay spread (ns)
1 st CRCC	20	0.78	0.74	0.89	1.0
2 nd CRCC	29	0.62	0.65	1.5	1.65
2 nd DGS	22	0.86	0.85	1.1	1.3
3 rd DGS	30	0.72	0.80	0.92	1.1

7.5 Conclusion

The LO leakage from the up-converter in the carrier based UWB localization system could be remedied by using a multi-stage band-stop filter. Based on our analyses and measurements, it has been found out that the filter order number needs to be optimized to satisfy specific pulse amplitude and delay spread requirements while providing sufficient band rejection level. Meanwhile, a narrower rejection bandwidth with the associated higher Q value would lead to larger first pulse amplitude and less time delay spread, subsequently benefiting the UWB system with a minimum pulse dispersion. For example, the fabricated 3rd order DGS filter has provided adequate rejection level while maintaining a relatively large first pulse amplitude and a small time delay spread, making it the most suitable candidate for filtering the LO leakage in our carrier based UWB system design.

Chapter 8

Coherent Localization Experiment

One of the major issues in UWB communication is the synchronization of the system's clock of both the transmitter and receiver. In our early steps, we evaluated the system's performance while bypassing this challenge to get the ideal set of results that can be used as a measurement reference, i.e. in a coherent system. Therefore in our early experimental trials, the clock of the transmitter and receiver has been wired to carry out the coherent 1-D, 2-D, and 3-D localization experiments. In order to test the feasibility of a high accuracy measurements in a short range system, the equivalent time Tektronix TDS8200 sampling oscilloscope was also used instead of the receiver architecture outlined in Fig. 2.2. In these experiments the tag clock was synchronized with the oscilloscope triggering clock to have a coherent system as previously mentioned. This isolates the errors due to the system architecture and TDOA algorithm by bypassing the clock synchronization, receiver-side sampling, etc. These experiments provided a benchmark on achievable system accuracy, i.e. the ideal scenario.

When testing a 3-D localization system for a mm-range accuracy, a highly accurate reference positioning system is required for calibration and validation. For this purpose the Optotrak 3020 [112] is used to obtain accurate reference data. The Optotrak is a 3-D IR tracking system that provides a 3-D positioning data with an accuracy of less than 0.3 mm, which is needed to validate the developed UWB positioning system. Figure 8.1 shows the Optotrak system and its optical probe. As shown in Fig. 8.2, the indoor lab

environment for the 1-D, 2-D, and 3-D experiments has dense multipath effects including reflections from the side walls, floor, furniture, ceiling, test equipment, and human bodies. Our system depends on calculating the differential time of flight between the transmitter and many receivers, however if an UWB signal passes through a human body (e.g. high permittivity materials) with huge attenuation, or if the LOS signal is completely blocked and the UWB signal would arrive at the receiver through multiple reflections, then it will exhibit a relatively large time delay [113]. In both cases, these signals would introduce significant errors into our TDOA system. Consequently, in these experiments only unblocked LOS cases will be studied. The monopole transmitting antenna was constantly within $\pm 20^\circ$ of the broadside direction from the Vivaldi antennas to minimize the phase center variation at the base stations.

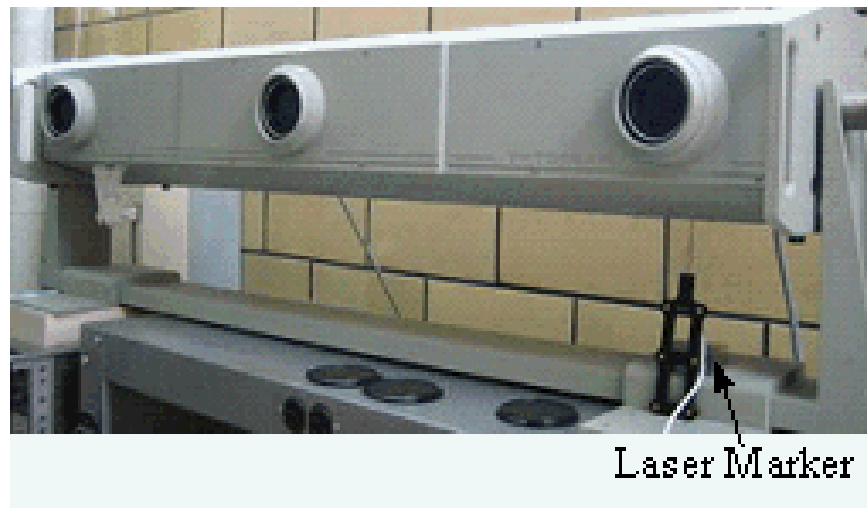


Fig.8.1 Optotrak system and optical probe.



Fig.8.2 Indoor experimental environment

8.1 Experiment Setup

The experimental setup of the coherent system utilizes a 300 ps Gaussian pulse to modulate a carrier signal centered at 8 GHz, which is then transmitted through an omnidirectional UWB antenna. The effect of the number of receiving antennas and their location will be discussed later in this chapter. Here for the coherent 3D experiments, up to six directional Vivaldi receiving antennas are located at distinct positions in an indoor environment to receive the modulated pulse signal. Meanwhile, all the receiving antennas are connected by a Single Pole Six Throw (SP6T) switch through coaxial cables having the same length which is extremely important to acquire highly accurate data sets. The received modulated Gaussian pulse is then amplified through a LNA, and demodulated to

the I/Q pulses. After going through a low pass filter, the I/Q pulses are fed into the multi-channel Tektronix TDS8200 sampling oscilloscope, and finally processed through the developed algorithm to accurately calculate the position of the transmitting tag. The complete experimental setup is shown in Fig. 8.3.

In this experiment we have used the Tektronix TDS8200 sampling oscilloscope for best scenario conditions as previously mentioned. However, for a practical portable system, the TDS8200 needs to be replaced by a combination of a sub-sampler [30] and a low cost data processing unit designed through FPGA implementation. A practical standalone system was successfully implemented and its results will be discussed later in Chapters 9 and 10.

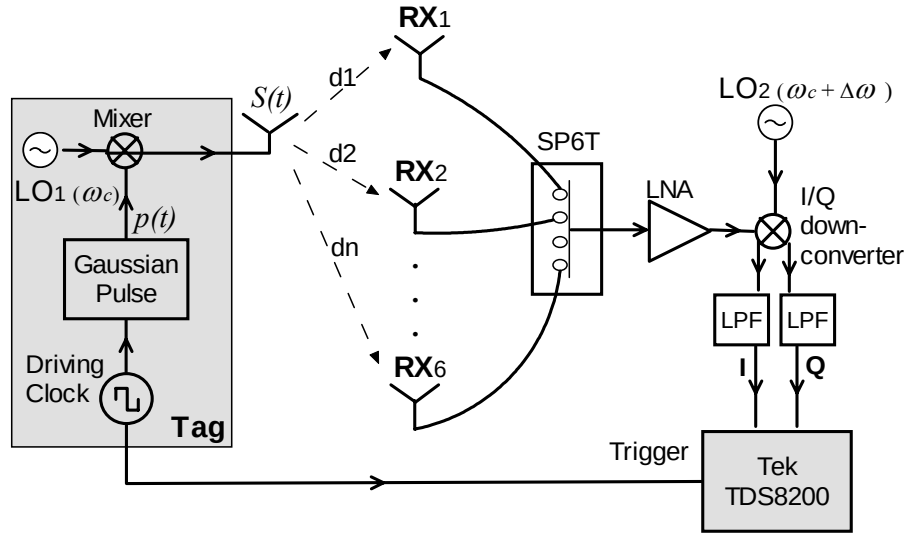


Fig.8.3 Block diagram of synchronized indoor localization system.

8.2 Simple Signal Processing Algorithm

In our early investigations simple peak detection algorithms have been utilized. To detect pulse peak in signals where pulses from multipath interference overlap significantly with the desired pulse, advanced algorithms need to be applied. Those other approaches can be found in [98]. Initially, a simple algorithm with a combination of signal strength (SS) and first peak-finding algorithms has been used for accurate detection of the pulse peak position. For example, a normalized received signal including a white noise and multipath pulses is shown in Fig. 8.4a. The white noise was removed after applying an averaging filter, as shown in Fig. 8.4b.

However, the detected signals could become more complex and can cause pulse overlapping in cases of severe multi-path conditions and could significantly impact its peak location estimate as shown in Fig. 8.5.

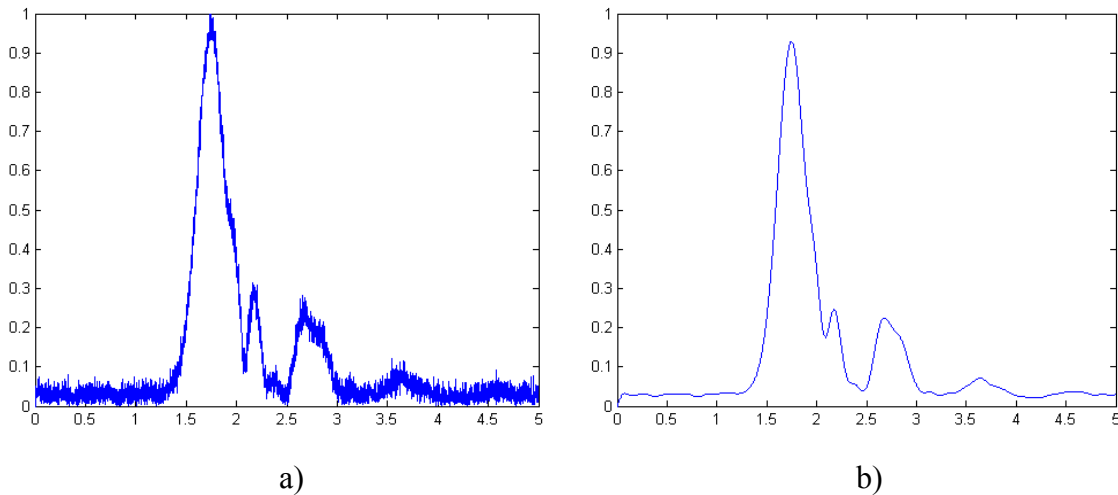


Fig.8.4 Pulse peak detection and filtering: a) Received pulse raw data; b) Received signal after averaging filter.

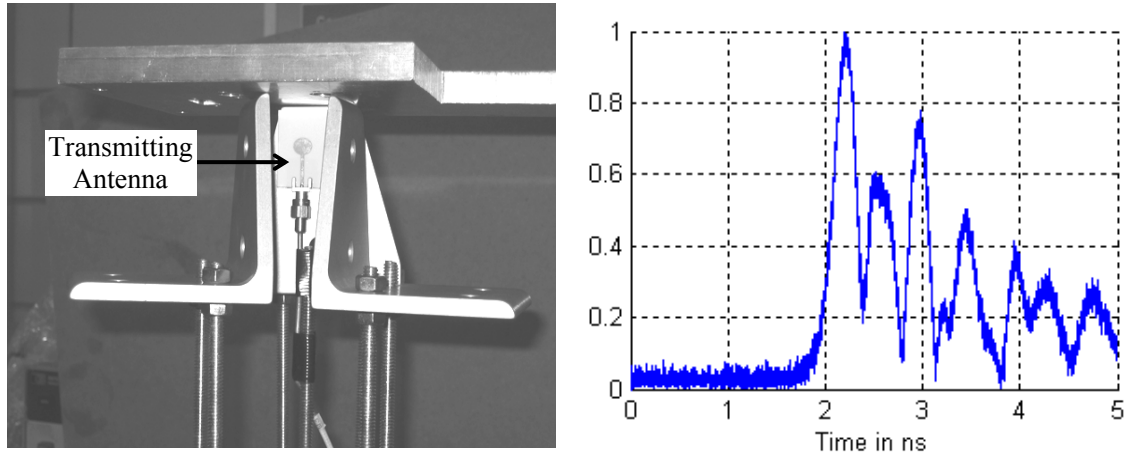


Fig.8.5 On the left: omni-directional transmitting antenna placed in a severe multipath environment surrounded by numerous closely spaced metal objects; on the right: the measured signal with pulse overlapping.

Relevant methods of advanced signal processing for UWB positioning in multipath environments generally can be decomposed into three major categories:

- 1) frequency-domain spectrum fitting [114,115];
- 2) covariance methods [116,117];
- 3) pulse detection and subtraction in the time domain [118-120] or in the frequency domain[121].

The computational complexity of the frequency-domain spectrum fitting involves *a priori* knowledge of the number of multipath components and a potentially expensive optimization step. Meanwhile, the covariance-based methods employ matrix inversion and eigen value decomposition which become costly when more points are taken to form the covariance matrix. Alternatively, simplified methods using only matched filters

employed by Low et al. [12] cannot offer high accuracy in cases where multipath components experience a delay of less than half of the pulse width. To address the strong multipath problem, a novel algorithm, called Iterative Peak Subtraction, follows the track of the “search and subtract” paradigm has been proposed by a member of our team, with some additional preprocessing and a peak selection evaluation step [70]. And further, a more advanced algorithm called leading edge detection has been successfully implemented in FPGA and utilized in our experiments [98].

8.3 Experiment and Results

8.3.1 1D Measurement

To carry out the 1D experiment, a Newport precision optical rail with a sub-mm accuracy was utilized for 1D reference positions. Both transmitting and receiving antennas were positioned on the optical rail. The transmitting antenna was initially placed a 10 cm away from the receiver and then precisely moved along the rail in 5cm increments for a total of nine measurements. The experimental setup is shown in Fig. 8.6, and the measured error distribution vs distance displacement is shown in Fig. 8.7, where an error of less than 3 mm was consistently achieved.

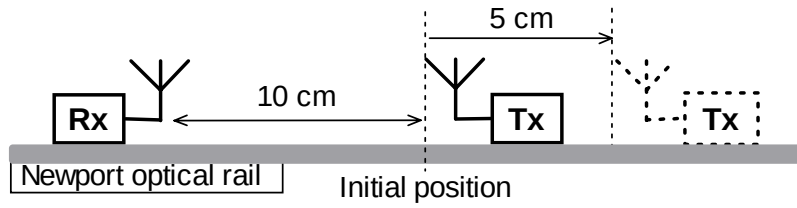
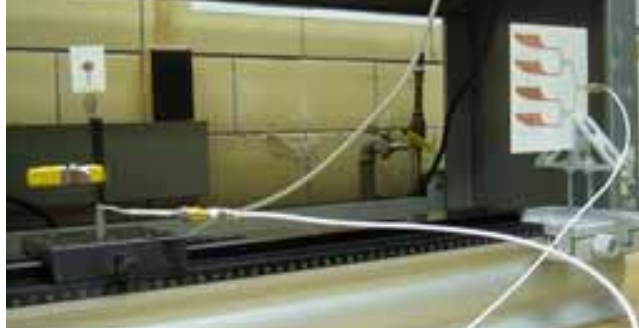


Fig.8.6 1D coherent measurement setup.

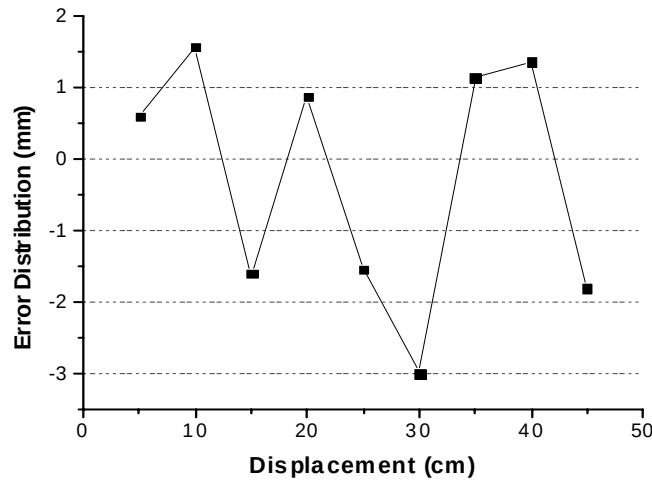


Fig.8.7 1D measurement error.

An experiment has been conducted to study how our positioning radar system would perform under partially blocked line of sight environment. In this case, an object with a high attenuation (e.g. human hand) was used to fully block the LOS path between the transmitting tag and the receiver. It is called partially LOS condition since the LOS signal still can reach the receiver but with substantial attenuation.

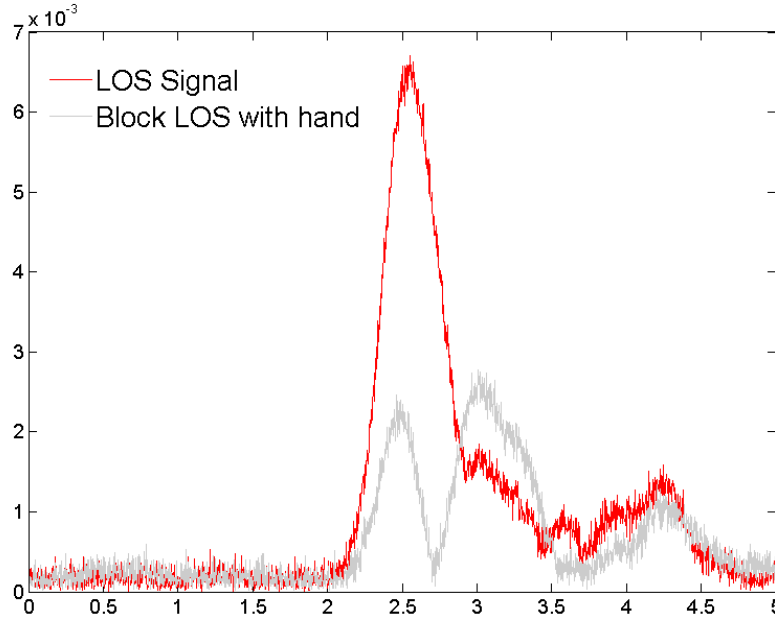


Fig.8.8 Received signals under LOS and partially blocked LOS conditions.

Figure 8.8 shows the received signals under LOS and partially blocked LOS conditions. While blocking the LOS path by hand, two clear pulses are detected indicated in a gray color. The first pulse is the LOS signal traveling through the human hand and the second pulse is the multipath/NLOS signal reflected from nearby instruments. The NLOS signal is stronger than the LOS signal due to the significant LOS signal attenuation of roughly 10 dB. First peak-finding algorithm is applied since Signal Strength yields incorrect results due to the stronger multipath/NLOS effect. An error of 2mm was achieved in this case at a 50 cm displacement even after partially blocking the LOS signal, such error could be from the slight signal delay while the LOS pulse penetrating or getting around the human hand. Further experiments should be carried out to thoroughly and quantitatively evaluate the effect of human body obstruction on the LOS signals.

8.3.2 2D Measurement

This experiment demonstrates the capability of the developed UWB system to accurately perform 2D local positioning within a dense multipath indoor environment. The TDOA technique is used for localization. Three base stations, BS1, BS2, and BS3, were placed in the 2D X-Y plane at the Optotrak system defined positions: (0 mm, 0 mm), (1546.63 mm, -122.87 mm) and (895.93 mm, 1600.51 mm). The transmitter was placed at an unknown position (X_u, Y_u) on the X-Y plane. A sketch of the experimental scenario is shown in Fig. 8.9, including the side-walls, tables, and measurement instruments. The X-Y plane is defined at 1.5 m above the ground and a 1.5 m below the ceiling.

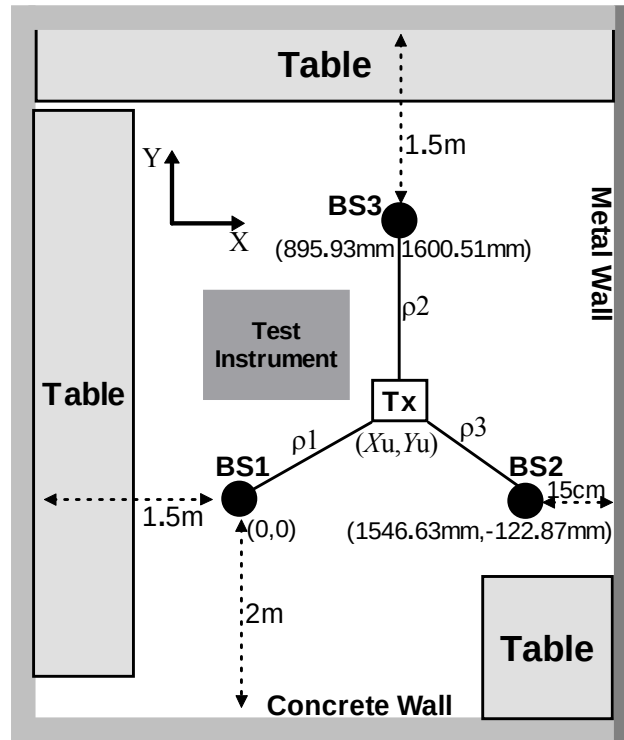


Fig.8.9 2D experimental scenario.

$\rho_i (i = 1, 2, 3)$ is the LOS distance between the transmitter and each receiver. The differential distances between the transmitter and the three receivers (i.e. $\delta\rho_{12} = \rho_1 - \rho_2, \delta\rho_{13} = \rho_1 - \rho_3$) are utilized in conjunction with the developed TDOA algorithm to triangulate the 2D transmitter location. The measurements were repeated, by moving the transmitter in discrete 5 cm steps for a total of nine movements. An optical probe was attached to the transmitting tag to provide a sub-mm accurate 2D reference data.

Figure 8.10 shows the 2D error distribution. A maximum error of approximately 3.5 mm was achieved for both (x,y) coordinates and the overall distance. A systematic error (underestimate of the distance) of about 1.75 mm for the 2D measurements has been noticed, as shown in Fig. 8.10a, this error was related to the difficulty to exactly mapping the Optotrak defined X-Y plane to the base stations defined X-Y plane.

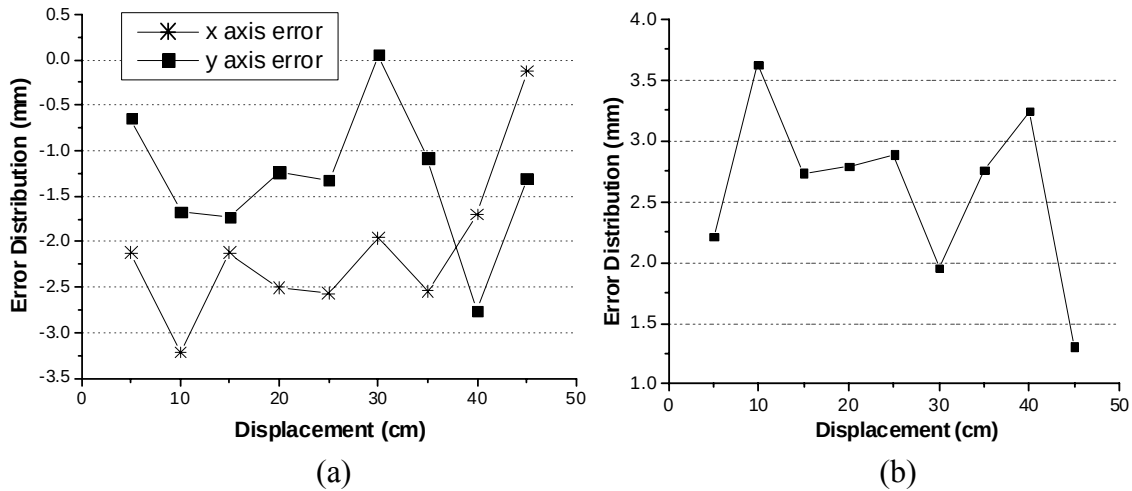


Fig.8.10 2D error distribution: (a) Errors in the x, y directions, (b) Overall distance error.

8.3.3 3D Measurement

Figure 8.11 shows the experimental setup with 6 base stations put at fixed locations around the transmitting tag under test. Both the tag and the base stations were put on a supporting rail. The tag was then moved along the rail with eight measurements applied. The 3D positions of the six base stations were measured by an optical tracking system and are listed in Table 8.1.

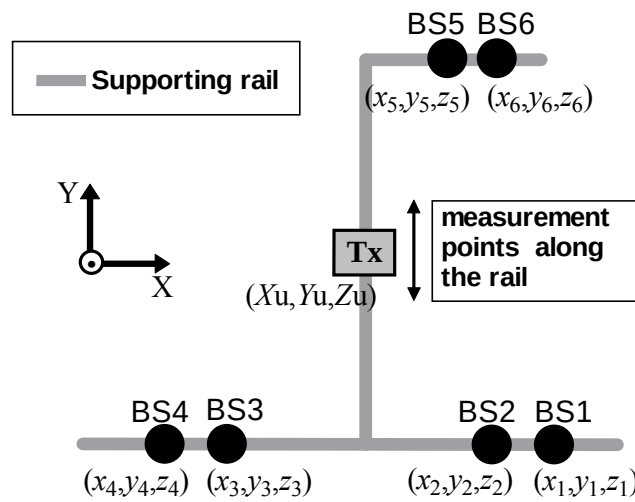


Fig.8. 11 Experiment setup using 6 base stations

Table 8.1– 3D Positions of 6 Base Stations Provided by the Optotrak 3020 system

	x (mm)	y (mm)	z (mm)
BS1	833.43	-4204.59	-411.15
BS2	736.72	-4183.98	157.68
BS3	-227.404	-4137.553	136.01
BS4	-331.388	-4153.172	-134.13
BS5	755.04	-2342.32	-377.49
BS6	788.99	-2361.42	-75.73

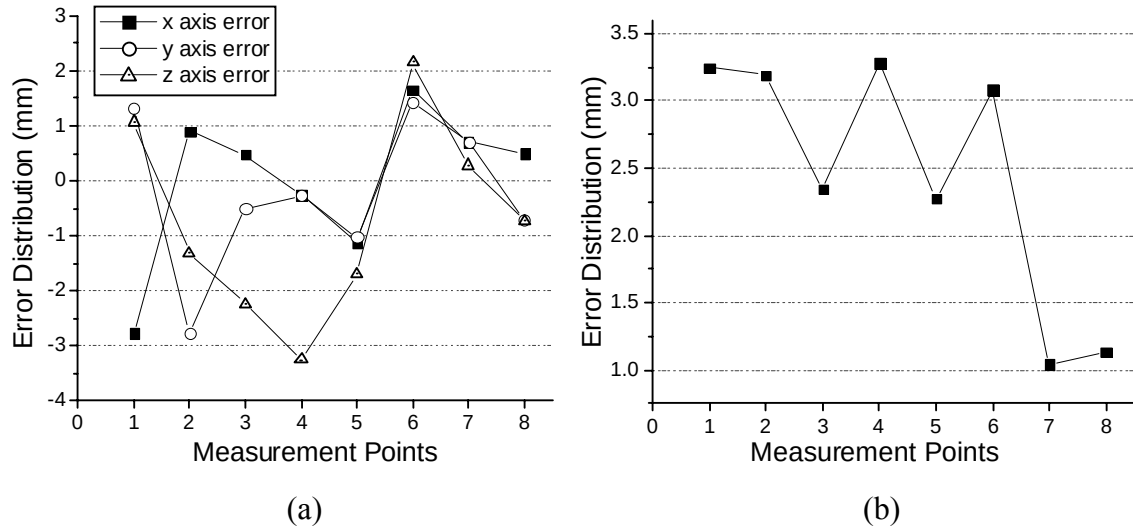


Fig.8.12 Error distribution with 6 base stations: (a) Error in x , y , and z , (b) Overall distance error.

Figure 8.12 shows the 3D error distribution with the 6 base stations. A maximum error of approximately 3.3 mm was achieved for both the (x,y,z) and the overall distance. The relatively the lager error in z -coordinate as compared to the x - and y -coordinates is caused by the six base stations have smaller fluctuations along the z -coordinate according to Table 8.1. Better overall accuracy can be achieved upon optimizing the location and number of these base stations.

The placement of the base stations is critical for the accuracy of a TDOA localization system. Ideally, the base stations would be spherically positioned around the volume of interest. If only four base stations are used, placing them at the nodes of a tetrahedron gives optimal results, although this may not be possible in actual indoor environments. Therefore, the number of base stations used for TDOA is varied, and the effect of this on the overall system accuracy has been investigated. Figures 8.13(b)-(d)

outline the other three configurations considered in our study where only 4 or 5 base stations were used in the TDOA localization. It should be noted that the data from the experiment outlined in Fig. 8.13(a) is used for all configurations by leaving out specific base stations to realize the other three configurations.

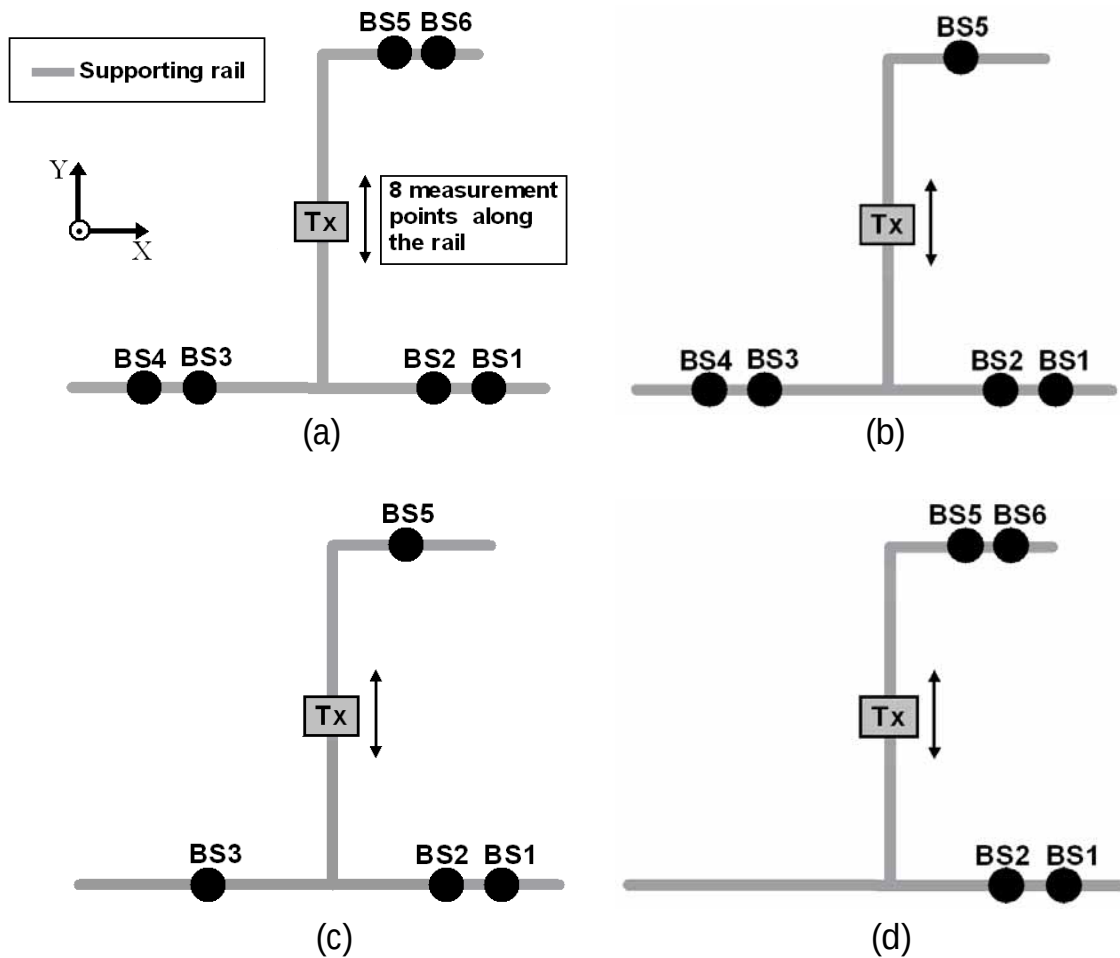


Fig.8.13 3-D synchronized localization experiments: (a) 6 base stations. (b) 5 base stations. (c) 4 base stations with low position dilution of precision (PDOP). (d) 4 base stations with poor coverage in the x -direction resulting in a high PDOP.

The 3-D experiment includes two 4 base station scenarios, one with 5 base stations, and one with 6 base stations configuration, as shown in Fig. 8.13. The system accuracy clearly gets progressively better as the number of base stations is increased from 4 to 6 and yields the highest accuracy of a 2.45 mm as indicated in Table IV. Obviously the largest error is seen using the 4-base station configuration as shown in Fig. 8.13(d). All of the base stations in this configuration reside on the $+x$ side of the tag. As shown in Fig. 8.14, this poor choice of placement has resulted in a large error in the x position, which can be attributed to the poor base station placement relative to the tag, also known as position dilution of precision (PDOP).

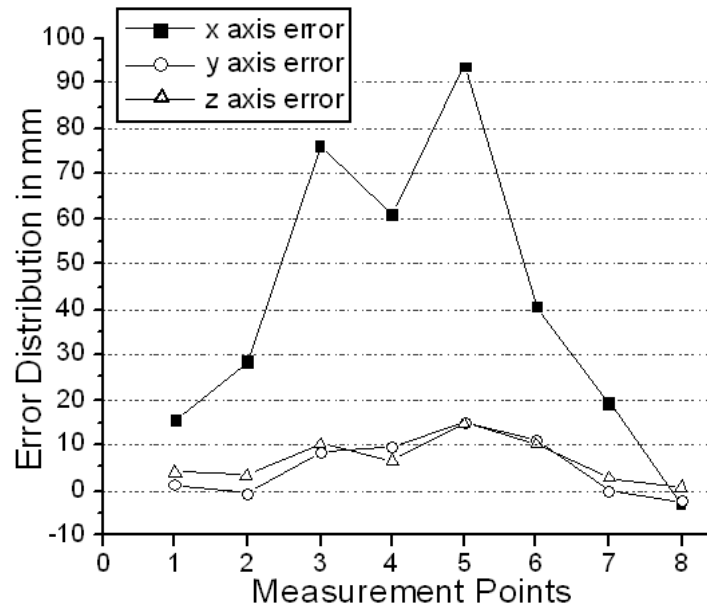


Fig.8.14 Error in the x , y , and z illustrating a high PDOP for the base station distribution shown in Fig. 8.13(d). This shows the reduction in accuracy that could result from the poor base station spatial arrangement.

PDOP can be computed upon convergence of the TDOA algorithm as in [122] and it is dependent solely on the base station geometry relative to the tag position. Our TDOA algorithm involves linearizing the relative range measurements of each of the base stations about a position estimate using a matrix H [123] defined by

$$H D_x = R \quad (8.1)$$

where R represents a matrix with the range difference elements, and D_x is the position update vector. The least squares solution of (8.1) is shown in (8.2)

$$D_x = (H^T H)^{-1} H^T R \quad (8.2)$$

where D_x can be used to update the current position estimate P_i

$$P_{i+1} = P_i + D_x \quad (8.3)$$

The dilution of precision parameters can be calculated using (8.4)

$$\begin{bmatrix} \sigma_x^2 & \sigma_{xy} & \sigma_{xz} \\ \sigma_{yx} & \sigma_y^2 & \sigma_{yz} \\ \sigma_{zx} & \sigma_{zy} & \sigma_z^2 \end{bmatrix} = (H^T H)^{-1} \quad (8.4)$$

Combining the diagonal elements, the overall PDOP is given by

$$PDOP = \sqrt{\sigma_x^2 + \sigma_y^2 + \sigma_z^2} \quad (8.5)$$

Alternatively, the PDOP can be explicitly defined for each coordinate direction as

$$PDOP_x = |\sigma_x| \quad (8.6)$$

Therefore the 3-D error can be estimated by combining the PDOP of the geometric configuration with the 1-D uncertainty of the system [3] as shown in (8.7)

$$\sigma_{3D} = PDOP \sigma_{1D} \quad (8.7)$$

For each of the base station configurations we report the mean PDOP across the eight measured tag positions for each of the coordinate axes as well as the overall PDOP. The results are reported in Table 8.2. It can be seen that the PDOP value is more strongly dependent on the base station distribution more than on the number of base stations and it is possible to have a small PDOP value with a minimum of 4 base stations located at optimized locations. Thus, in our noncoherent experiment, 4 base stations optimally located to have a small PDOP will be used. Table 8.3 summarizes the mean, the standard deviation and the worst case error of overall distance for the 1-D, 2-D, and 3-D experiments.

Table 8.2- PDOP Summary for the Synchronized Localization Experiments

	Mean PDOP _x	Mean PDOP _y	Mean PDOP _z	Mean Overall PDOP
6 BS	0.87	0.44	1.61	1.88
5 BS	0.90	0.59	1.70	2.02
4 BS Fig. 8.13(c)	1.40	0.60	1.89	2.43
4 BS Fig. 8.13(d)	20.2	1.71	3.33	20.6

Table 8.3 Error Summary – Coherent Localization Experiments

		Mean Error (mm)	Std. Dev. Error (mm)	Worst Case (mm)	Mean PDOP
1-D		1.49	0.69	3.0	-
2-D		2.61	0.69	3.6	-
3-D	6 BS	2.45	0.93	3.3	1.88
	5 BS	3.13	1.20	4.2	2.02
	4 BS Fig. 8.13(c)	3.62	1.53	5.6	2.43
	4 BS Fig. 8.13(d)	43.3	32.4	96.1	20.6

8.4 Error Discussion

The data in Tables IV and V can be analyzed by considering the following points.

1. The PDOP values for each coordinate direction can be understood by examining the 6 base station experiments as shown in Fig. 8.13(a). The spatial spread of the base stations along the y -axis is the largest (1.86 m), subsequently the $PDOP_y$ value listed in Table III is the lowest among the coordinate axes. This is a common trait to the experimental set up in Fig. 8.13(b-c), therefore the PDOP values are consistent across these experiments. In these experiments the $PDOP_z$ values are consistently the highest, due to the maximum range of base station positions that spread along the z -axis being relatively the smallest (i.e. only 0.57 m). This is consistent with our measurements of Fig. 8.13(d) set up, which represents a poor geometric configuration, and its $PDOP_x$ was the worst case due to the low spatial diversity of the base stations along the x -axis.
2. The phase center of the Vivaldi receiving antenna varies with the received signal direction, which can have large degradation effects on the system accuracy especially when the tag is in locations on the border or outside of a given target volume. We have insured in this experiment that all the tag positions remained within $\pm 20^\circ$ of the broadside direction of each single element Vivaldi in order to minimize the phase center effects. The error originating from this effect is estimated to be less than 1 mm, although if the angle from broadside increases beyond $\pm 20^\circ$, this phase center error increases significantly.

3. Multipath interference from extremely close metal (e.g. metal bar supporting the transmitting tag) could cause pulse peak shifting. These localization experiments were done with strong LOS signals and only minimal amounts of multipath interference. However, the developed algorithm can handle dense multipath situations but still has substantial uncertainty of around 3 mm under severe multipath conditions [70].
4. The overall system error can be significantly reduced by increasing the number of base stations. As shown in Table 8.2, the PDOP decreases with a higher number of base stations. Also, the system error due to multipath interference is reduced with an increased number of base stations since the received signals at each base station will traverse through a different UWB channel realization which is very similar to MIMO principles.

8.5 Conclusion

In this chapter, the experimental results of the high resolution coherent UWB positioning radar system based on TDOA have been described. 1D, 2D, and 3D experiments utilizing our UWB radar system have demonstrated a mm-range accuracy with reference to an Optotrak system. The measurement accuracy could be significantly enhanced by increasing the number of base stations. Another degree of freedom in designing these systems is the space distribution of these base stations. Spherical distribution proved to be the best but it could be impractical in some implementations.. The following useful practical conclusions can be derived:

- ◆ Four base stations can yield reasonably good PDOP with optimized base station locations;
- ◆ Experiments should be conducted in a LOS scenario, since that even a human hand will yield significant signal attenuation;
- ◆ Advanced algorithm such as the first peak finding or the leading edge detection algorithm should be utilized to obtain higher accuracy;
- ◆ Phase center induced error can be generally neglected when the tag movement is set to be within $\pm 20^\circ$ off the boresight of each Vivaldi receiving antenna.

Chapter 9

Noncoherent UWB Localization System Theory

In the previous chapter, the coherent experiment has been conducted, where the transmitter and the receiver PRF clock are synchronized with a Tektronix two channel functional generator AFG3102 or a DDS module. The LO carrier synchronization was discussed in Chapter 2 in detail.

This chapter will be focused on issues in realizing a fully noncoherent UWB localization system, i.e. a standalone tag without wire connecting to the receivers. The organization of this chapter is as follows: Firstly, various UWB receiver architectures for local positioning systems are compared and a novel UWB receiver scheme is proposed. Secondly, I/Q mismatch issues in our earlier UWB receiver scheme, stated in Chapter 2, are discussed. Then synchronization on PRF clock between tag and receiver are investigated. Finally, a novel single channel noncoherent system approach is given, with focus on jitter reduction, extensive simulation and experiment are conducted and the results agreed very well with each other. The noncoherent 1D and 3D localization experiment based on the novel receiver scheme will be presented in the next chapter.

9.1 UWB Localization Receiver Scheme

Carrier based impulse radio (IR) UWB systems have been widely used in many areas such as UWB localization, see through wall and communication systems, where the transmitted UWB signal is up-converted through a LO carrier, then down-converted at

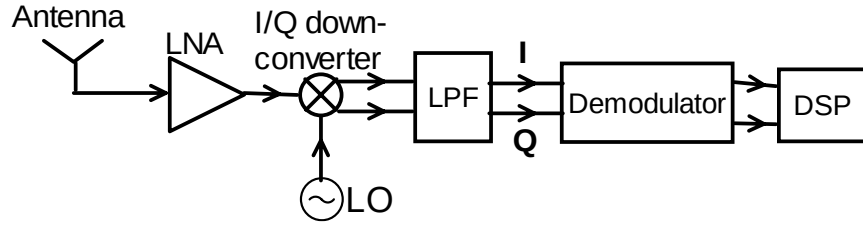


Fig.9.1 Carrier based non-coherent UWB receiver architecture.

the receiver side [70, 99, 124]. Figure 9.1 shows the carrier based low complexity UWB receiver architecture. In [99] the demodulator is implemented with a 2-FSK scheme, whereas in [70] the demodulator is realized through a low cost 2 channel sub-sampling mixer with an equivalent sampling rate over 100 GSPS [30], followed by a digital signal processing unit such as standard Field Programmable Gate Array (FPGA). However, the amplitude and phase differences between I and Q channels due to hardware variation always present, which would cause the distortion of the demodulated signal. Recently, Treyer et al. are able to correct the amplitude and phase error by using the Hartley phasing-type SSB modulator, but require a relatively complex circuitry and limited to narrow band applications [125].

Another widely used and low complex UWB architecture is the energy detection (ED), where the UWB signal is transmitted directly through the UWB antenna without up-conversion [21, 126-131]. At the receiver side, the energy detection of the signal is achieved by passing the signal through a square-law device, usually a Schottky or tunnel diode, followed by an integrator and sampler. Figure 9.2 shows the typical UWB receiver architecture using energy detection. However, due to the large bandwidth of the received UWB signal, it is difficult and costly for the ED based receiver to operate at above the

Nyquist rate [126]. Usually, a fast comparator is used as the sampling device. For example, T. Buchegger et al. realized an UWB communication link with a data rate of 1.2 Mbps using the tunnel diode detector [127]. J.P. Lie et al. realized the leading-edge pulse detection method using a tunnel diode combined with an envelop detector to maintain high accuracy for UWB ranging in multipath environment [128]. Recently, Fujii et al. achieved a 0.3 ns time resolution using the ED based impulse detector, corresponding to 0.1 m distance resolution [21].

This chapter presents analysis and development of a novel UWB receiver architecture for a low complexity, non-coherent real-time UWB localization system. As shown in Fig. 9.3, the proposed UWB receiver architecture combines the carrier based and the ED based UWB receiver scheme. The down-conversion requires only one channel instead of I/Q channel as compared to Fig. 9.1. The receiver has low complexity and the synchronization scheme does not increase drastically the complexity. The equivalent sampling rate is over 100 GSPS, i.e. well above the Nyquist rate to provide the required timing accuracy.

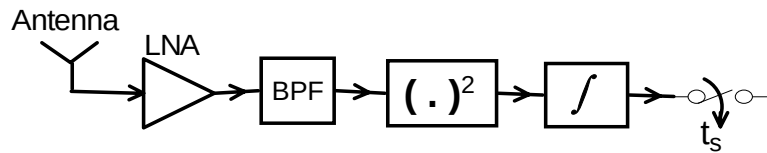


Fig.9.2 Energy detection based UWB receiver architecture.

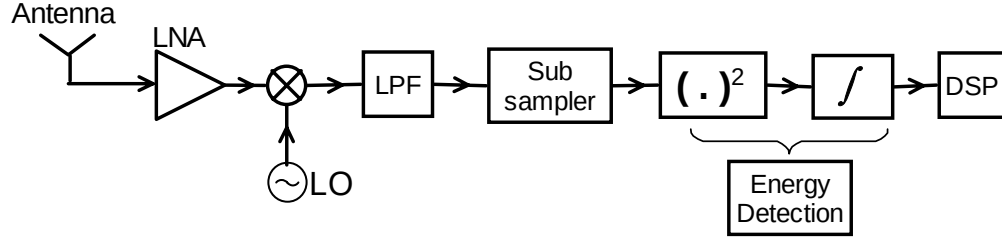


Fig.9.3 Proposed UWB receiver architecture combines the Carrier based and the ED based UWB receiver scheme.

9.2. I/Q Mismatch

The complete setup of our previous developed localization system was shown in Fig. 2.2. The 300 ps Gaussian pulse $p(t)$ was successfully recovered by I/Q down-conversion at the receiver side, given by equation (2.9)

$$\bar{P}(t) = \sqrt{I_{ex}^2 + Q_{ex}^2} = \frac{1}{2} P_{ex}(t)$$

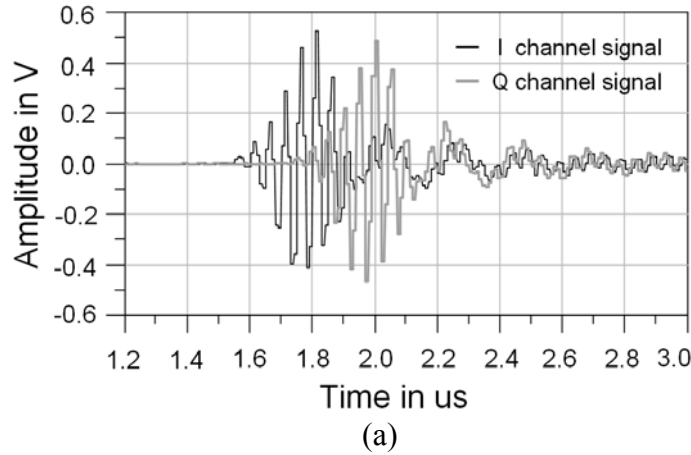
Where I_{ex} and Q_{ex} are extended I/Q signal from sub-sampler, and $P_{ex}(t)$ is the pulse signal after time extension while maintaining the same pulse shape as $p(t)$.

However, the above results are based on two assumptions: 1) there is no phase difference between I and Q channels; 2) the phase noise of both the tag and base station carrier are neglected, leads to a fixed offset carrier frequency $\Delta\omega$ without variation with time and temperature. However the I/Q mixer, HMC520 from Hittite, has a phase difference of up to 4 degree between I and Q channel, added up with unknown phase difference from the designed sub-samplers, and other phase unbalanced sources such as operational amplifier, LPF and cables in the I and Q channels, the phase difference between I/Q channel could render I/Q mismatch. Fig. 9.4a shows a simulated I_{ex} and Q_{ex}

signals at the samplers output, with a phase unbalance introduced time difference between I/Q channel of 0.2 ns. As can be seen in Fig. 9.4b, the reconstructed pulse from the phase unbalanced I/Q channels were distorted and failed to recover the original Gaussian pulse. As shown in Fig. 9.4c, the measured I_{ex} and Q_{ex} signals demonstrate a large mismatch. Unfortunately, such phase differences between I_{ex} and Q_{ex} channels are unknown and different for each base station due to hardware variations, and the extended I_{ex} and Q_{ex} signals have been modulated with the equivalent offset carrier frequency $\Delta\omega_{eq}$ after sub-sampling process, which makes it extremely difficult to calibrate the unknown phase difference error. Here we define a modulation factor β to be

$$\beta = \frac{\Delta\omega_{eq}}{\text{bandwidth of } P_{ex}(t)} = \frac{\alpha \cdot \Delta\omega_{eq}}{2\pi \cdot BW} \quad (9.1)$$

where $\alpha = \frac{PRF_1}{|PRF_1 - PRF_2|}$ is the time extension factor, BW is the -10 dB bandwidth of the transmitted UWB Gaussian pulse signal. For example, in Fig. 9.4a, the I_{ex} and Q_{ex} signal featured a modulation factor $\beta \approx 7$.



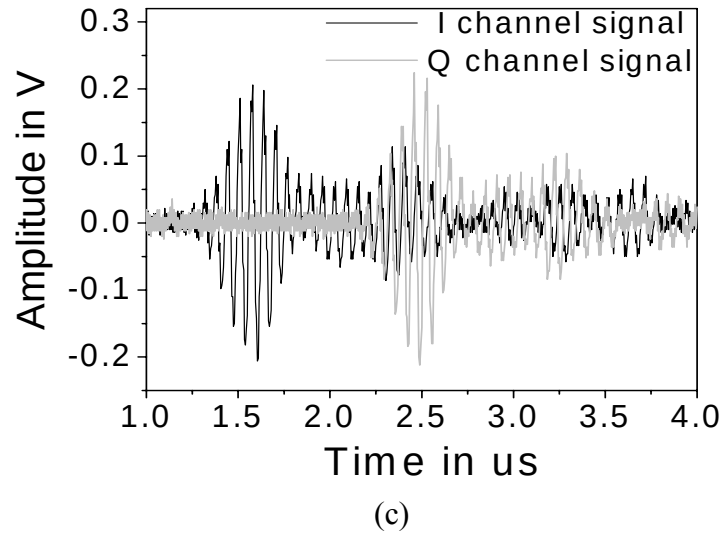
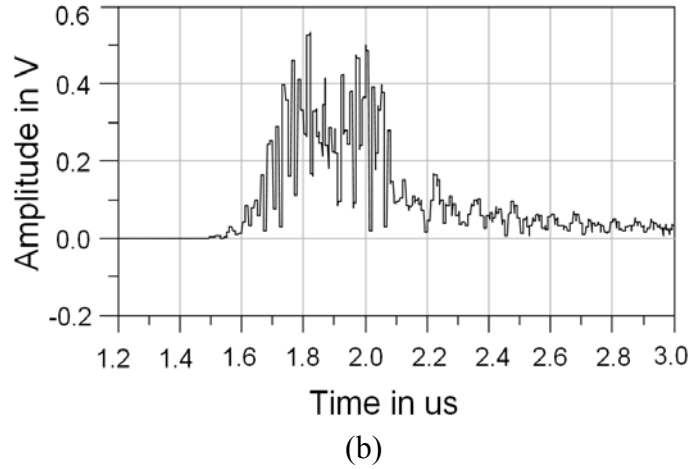


Fig.9.4 a) simulated results of I/Q mismatch; b) reconstructed pulse signal from mismatched I/Q signal, failed to recover the original Gaussian pulse; c) measured I/Q mismatch.

Considering the complexity and difficulties of applying I/Q scheme, and based on the results in equations (2.6) and (2.7), we could use single channel, i.e., either I_{ex} or Q_{ex} channel signal, for localization purpose as long as the equivalent offset carrier frequency $\Delta\omega_{eq}$ remain unchanged. However, the phase noise of carriers from transmitter and receiver are translated and included in the equivalent offset carrier frequency $\Delta\omega_{eq}$, leading to a unstable offset carrier frequency $\Delta\omega_{eq}$ that would vary with time and

temperature and cause jitter effect. To minimize such effect, the phase locked dielectric resonator oscillator (PLDRO) or any LO sources with extremely low phase noise and minimal temperature sensitivity could be used.

9.3 Tx-Rx PRF Synchronization

The signal clock frequency PRF_1 and strobe clock frequency PRF_2 between the tag and receiver in this system are not synchronized. This results in an interesting synchronization problem when incorporating the sub-sampling mixer (discussed in Chapter 4) since 10,000 pulses are needed to extend the pulse from 300 ps to 3 μ s, which corresponds to a 100 GS/s sampling rate. In our prototyped system, the extended time signal can then be adequately sampled with a 20 MS/s ADC, given the bandwidth of the sub-sampled signal is 0.3 MHz, although in the actual system the 20 MS/s ADC can be replaced by a higher performance ADC (e.g. 24 bit, 250 MS/s). This can potentially increase system performance by reducing quantization error through extending the dynamic range and increasing the sampling rate with little difference in chip and manufacturing costs.

Techniques exist to calculate clock jitter in the time domain given the phase noise of the crystal [132], and crystal manufacturers provide crystal stability specs in terms of parts-per-million (ppm). On the surface, the stability factor would appear to have less of an effect than actual clock jitter. For instance, the stability factor of ± 0.5 ppm for the 10.0 MHz Vectron VTC4-A0AA10M000 crystal yields clock jitter of 50 fs [50]. However, the effects of the stability factor are amplified during the sub-sampling process and cause time-scaling to occur.

Any frequency offset occurring between signal clock frequency PRF_1 and strobe clock frequency PRF_2 would cause apparent time-scaling in the sub-sampled signal. The ± 0.5 ppm stability of our crystal equates to 5 Hz of variation in the clock signal. This causes the nominal frequency of PRF_1 to be $10.000000 \text{ MHz} \pm 5 \text{ Hz}$ and the nominal frequency of PRF_2 to be $9.999000 \text{ MHz} \pm 5 \text{ Hz}$, with the worst case scenario being a difference of 10 Hz between the two signals. If the offset frequency (Δf) is set to be 1 kHz, the corresponding extending ratio is $\alpha = f_0 / \Delta f = 10,000$. However, with a potential offset of $\pm 10 \text{ Hz}$, Δf has the potential to be $1000 \pm 10 \text{ Hz}$, thus $\alpha = 10,000 \pm 101$. This time scaling may be time-varying depending on the thermal stability of the clocks, although this is a slow variation with a drift rate of less than 10 Hz, which allows the TDOA algorithm to calibrate such systematic error. Since all the base stations acquire synchronous samples clocked by PRF_2 , the time-scaling effect will be unknown, but consistent across receivers. Consequently, the 1-D ranging errors will be roughly $\pm 1\%$, and such estimation will be experimentally validated in the next chapter. The TDOA algorithm using time differences will likewise be affected. One effect way to reduce such error is to increase the strobe frequency of the sampling mixer, e.g. increase the strobe frequency from 9.999000 MHz to $10 \times 9.999000 \text{ MHz}$ will reduce the above 1D ranging scaling error from 1% to 0.1% without reducing the sampling resolution. However, the strobe frequency can not exceed 35 MHz due to SRD device limitation. Another straightforward way is to increase the offset frequency Δf , however, it will reduce the equivalent time sampling rate and increase the resolution error. Table 9.1 lists the scaling error as well as the resolution error with respect to different frequency offset Δf .

Table 9.1 – Scaling and Resolution Error vs. Offset Frequency (Assuming PRF=10 MHz,
Clock Stability = ± 0.5 ppm)

Freq. offset (kHz)		0.1	1	2	4	10
Sampling rate (GSPS)		1000	100	50	25	10
Sampling resolution (ps)		1	10	20	40	100
Worst case resolution error (mm)		0.15	1.5	3	6	15
Scaling error (%)		10	1	0.5	0.25	0.1
Scaling error related to tag coverage (mm)	0.5 m	50	5	2.5	1.25	0.5
	1 m	100	10	5	2.5	1.25
	5 m	500	50	25	12.5	5

The results in Table 9.1 has been plotted and indicated in Fig. 9.5. The resolution error increases linearly with the increase of the offset frequency. While the scaling error increases linearly with the range of the tag movement. The red dots denote the optimized point where the offset frequency would provide an overall minimum error.

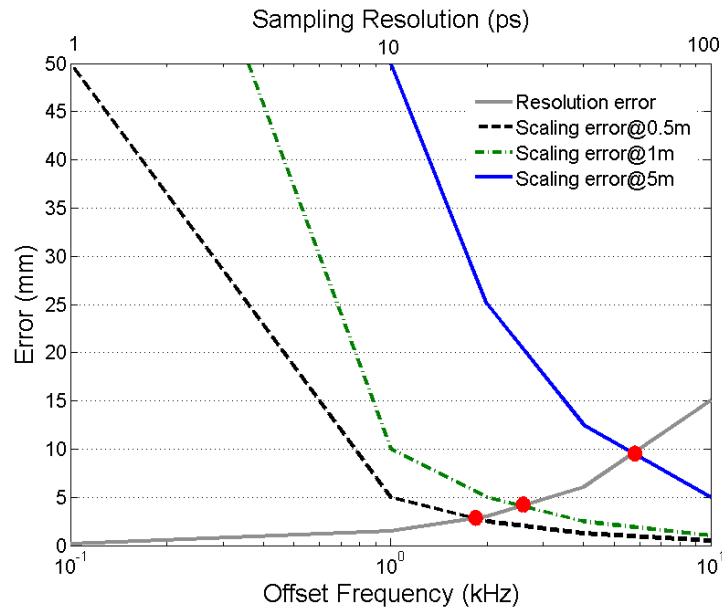


Fig.9.5 Scaling and Resolution Error vs. Offset Frequency.

Further validation through a static 3D experiment is shown in Fig. 9.6. After calibration, the tag is then put at a fixed random point inside a volume of interest (with about 0.5 m coverage), and the frequency is set to 1 kHz, 2 kHz and 4 kHz respectively. The minimum 3D static error occurred when the Δf is set at 2 kHz offset, as predicted in Fig. 9.5. Table 9.2 summaries the measured RMS error for these three cases. The results are obtained after averaging the raw TDOA data with 17 times for each data point.

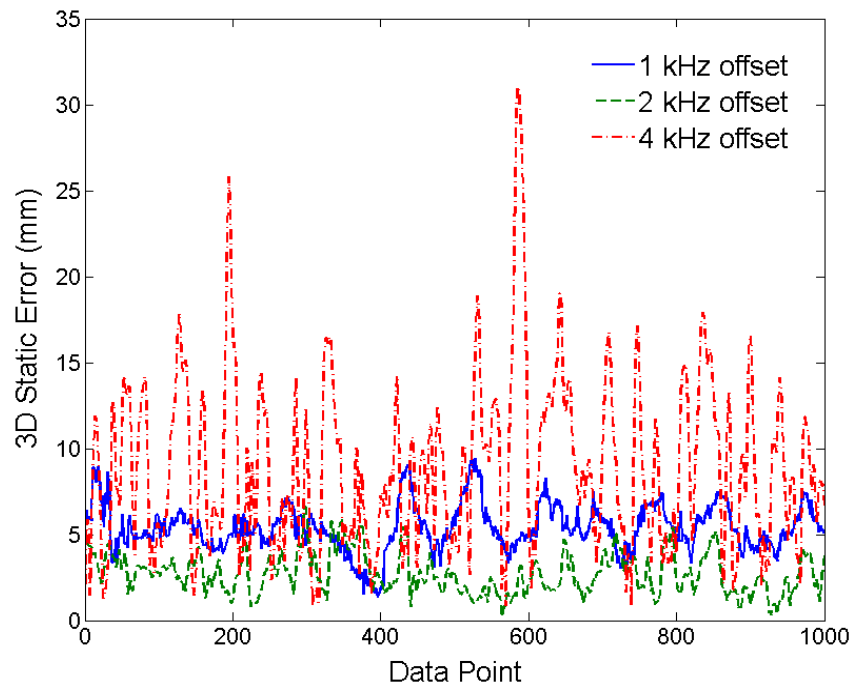


Fig.9.6 Experimental validation: offset frequency vs. 3D error

Table 9.2 – Measured Offset Effect

Offset (kHz)	3D RMSE (mm)
1	5.53
2	2.94
4	9.95

Besides stability, the jitter of these 10 MHz clocks must be examined. When the phase noise is integrated from 1 Hz – 5 MHz using an Agilent E5052A Signal Source Analyzer, RMS jitter for the crystal is found to be between 0.5 – 1.5 ps [133]. This technique is the most accurate way to measure clock jitter of highly stable crystals. If 1.5 ps of RMS jitter is assumed for both 10 MHz crystals, total system RMS jitter σ_{sys} due to the two unsynchronized clocks of 2.12 ps is obtained through (7)

$$\sigma_{sys} = \sqrt{\sigma_{clk1}^2 + \sigma_{clk2}^2} \quad (7)$$

where σ_{clk1} and σ_{clk2} are assumed to be uncorrelated normal random variables [132,134] of mean $\mu = 0$ and standard deviation $\sigma = 1.5$ ps. The jitter described in (7) will cause normally distributed noise of $\mu = 0$, $\sigma = 2.12$ ps (corresponds to 0.64 mm error) to be added to each sampled point. A simulation using Agilent ADS2006A has been carried out to study how such jitter affects the sampler performance during the sub-sampling process. Based on the ADS simulation results shown in Fig. 9.7, we notice that such jitter would cause tiny signal distortion, and could be significantly reduced after simple digital processing such as a low pass filter.

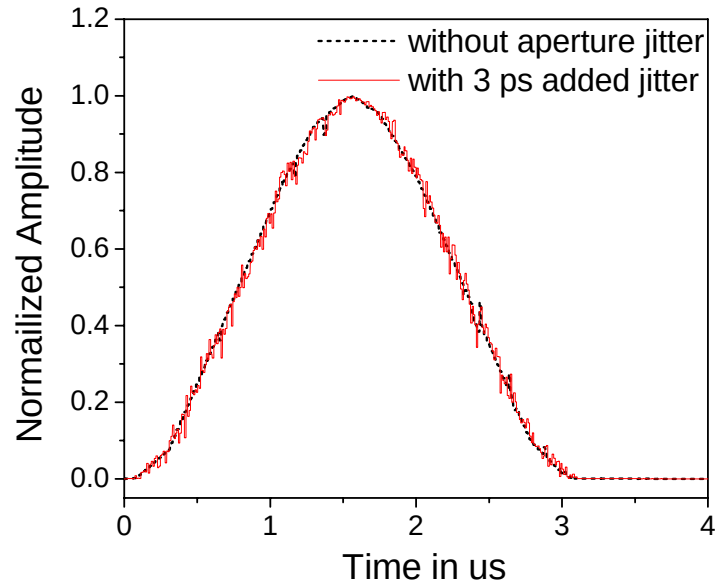


Fig.9.7 ADS simulation of the reconstructed sub-sampled pulse with and without 3 ps of PRF clock jitter.

Figure 9.8 outlines the experimental setup used to test the jitter effect caused by the unsynchronized PRF clock signals. A coaxial cable was utilized to connect the pulse generator and the sampler so that no channel noise is included. Figure 9.9 shows the measured time variation of the pulse peak position at consecutive measurement cycles. The peak to peak variation is below 10 ps while the RMS jitter is 3.48 ps. The measured RMS jitter is 1.36 ps larger than the theoretical result of 2.12 ps, which can be interpreted as the added jitter from the sub-sampling mixer, DDS, and ADC circuitry. The measured system clock jitter of 3.48 ps corresponds to 1.05 mm error.

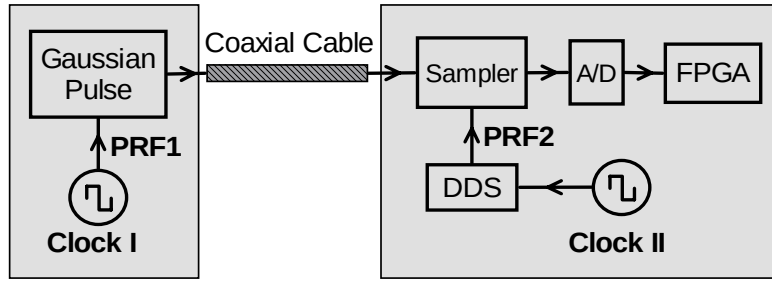


Fig.9.8 Experimental setup with the unsynchronized PRF clock sources to measure the effect of PRF clock jitter.

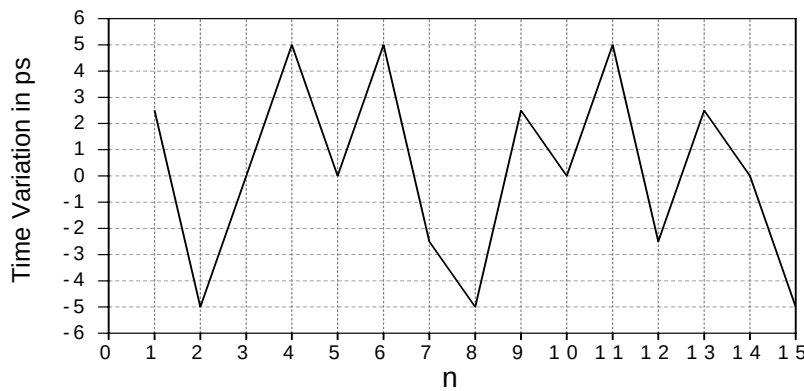


Fig.9.9 Time variation of the pulse peak position acquired over n sample points.

9.4 Noncoherent System – Single Channel Approach

9.4.1 Static Scenario

Carriers with high and low phase noise performance have been applied in the simulation model and the resulted jitter has been compared for each case. Table 9.3 lists the phase noise performance of Hittite LO HMC506, Agilent signal generator E8257D and 83622B, which will be used as the carrier sources. The Agilent signal generator features a much lower phase noise compared to the Hittite HMC506, especially when the offset from carrier is less than 10 kHz.

Table 9. 3 - Phase Noise Performance of Different Carrier Sources

Carrier source	Phase noise @1 kHz	Phase noise @10 kHz	Phase noise @100 kHz
HMC506	-46 dBc/Hz	-80 dBc/Hz	-103 dBc/Hz
E8257D	-106 dBc/Hz	-115 dBc/Hz	-115 dBc/Hz
83622B	-85 dBc/Hz	-90 dBc/Hz	-110 dBc/Hz

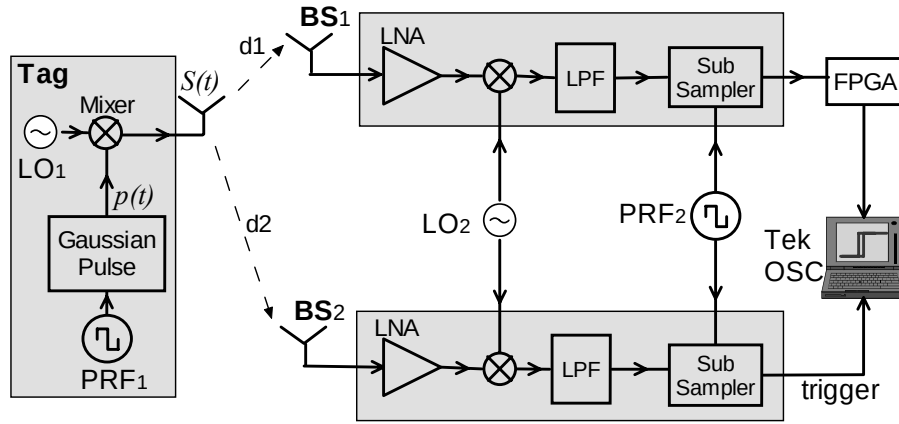


Fig.9.10 The jitter measurement setup

Both simulation and experiment have been carried out to study how the carrier phase noise translates to the system jitter. We first consider a static scenario, under this situation the tag is at a fixed position during simulation and measurement. The unsynchronized simulation/experiment setup has been carried out, as shown in Fig. 9.10, to study the relationship between the carrier phase noise and system jitter. The jitter was calculated by recording a continuous 40 time positions when the comparator is triggered at the rising-edge of a fixed voltage threshold setting at around 50% of the maximum pulse amplitude. The sub-sampled output signal from BS_2 served as the trigger signal for the Tektronix TDS340A oscilloscope. The simple received signal strength (RSS) method

with a fixed voltage threshold has been used. Three combinations of carrier sources have been utilized: 1) the HMC506 for carrier at both transmitter and receiver; 2) the HMC506 for the carrier at transmitter, the Agilent E8257D for the carrier at receiver; 3) the Agilent E8257D for the carrier at transmitter and the Agilent 83622B for the carrier at receiver.

Table 9.4 lists the results of simulated/measured RMS and Peak-to-Peak (PP) jitter under different combinations of carrier configurations at the transmitter and receiver sides. In case II of Table 9.3, by only replacing the receiver with a low phase noise carrier source did not improve the system jitter very much, since the carrier with high phase noise are included in the equivalent offset carrier frequency $\Delta\omega_{eq}$. Although the simulated results show smaller jitter error than the measured results, both of them show a same trend except in case III. In case III when low phase noise carriers were used at both transmitter and receiver sides, the measured jitter has not been reduced as expected in the simulation. This is due to the “shoulder” effect which will be explained in detail in the following section. The measured RMS and Peak-to-Peak jitter reduced to 7.2 ps and 30 ps respectively in case IV, where the modulation factor β and the threshold voltage have been tailored to minimize the “shoulder effect”.

Table 9.4 – Jitter Performance with Different Carrier Sources

	Tx Carrier	Rx Carrier	Measured RMS Jitter	Measured PP Jitter	Simulated RMS Jitter	Simulated PP Jitter
I	HMC506	HMC506	36.1 ps	130 ps	25.5 ps	100.1 ps
II	HMC506	E8257D	31.1 ps	115 ps	24.0 ps	70.1 ps
III	83622B	E8257D	31.9 ps	70 ps	15.9 ps	32 ps
IV	83622B	E8257D	7.2 ps	30 ps	1 ps	3 ps

9.4.2 Shoulder Effect and Dynamic Scenario

The experiment results from Table 9.4 reveal an interesting problem of the non-coherent carrier based UWB system using single channel approach. Figure 9.11 shows the single channel output after sub-sampling process. The “shoulders” are resulted from the equivalent offset carrier frequency $\Delta\omega_{eq}$, which modulates the time extended pulse signal $P_{ex}(t)$ as given by equation (6). The impact of phase noise from the carrier source will be translated not only as timing jitter, but also as “shoulder” amplitude variation of the modulated signal. When the RSS method with a fixed voltage triggering threshold was used, the random “shoulder” amplitude variation would produce another source of error. Such error created the large measured jitter in case III of Table 9.4 even when low phase noise carriers were used at both transmitter and receiver sides. By adjusting the equivalent offset carrier frequency $\Delta\omega_{eq}$ and the threshold voltage, the amplitude variation induced error can be eliminated and the measured jitter was reduced as in case IV of Table 9.4. However such optimization process is not practical since the threshold has to be set at very close to the peak amplitude to fully get rid of the “shoulder” amplitude variation. When the tag is moving, the peak amplitude may vary, and the threshold needs to be readjusted.

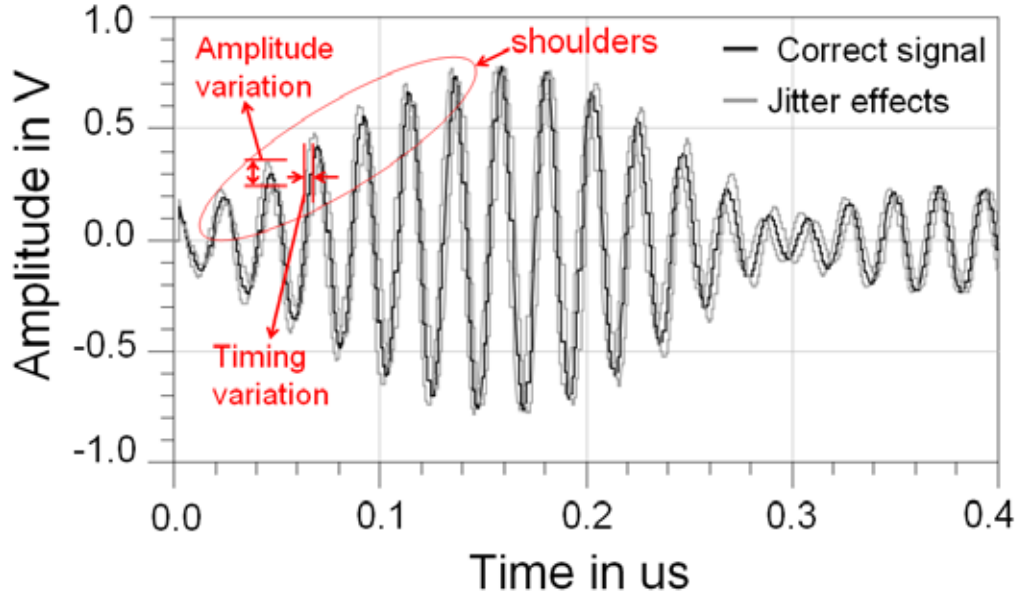


Fig.9.11 The “shoulder effect”: the received single channel sub-sampled signal modulated by the equivalent offset carrier frequency.

To further understand the “shoulder” effect problem, we need to consider the dynamic scenario. Under this situation the tag is considered moving continuously away from the base station up to one wavelength at the carrier frequency, the modulation factor β defined in equation (9.1) was set to be 5 and 15 respectively. In the simulation model, in order to study how the “shoulder” effect responds to the tag dynamic movement only, no phase noises are included in the carriers. Figure 9.12 shows the simulated results of how the “shoulder” amplitude is experienced a large variation while the tag was moving from 0° to 360° , 60° per step. By setting a fixed threshold, i.e. 0.4 V in both examples, the triggered time position doesn’t vary linearly with the tag position. As can be seen in Fig. 9.12, a large time position error would occur at certain tag position during the tag movement, which is caused by the “shoulder” amplitude variation during tag movement.

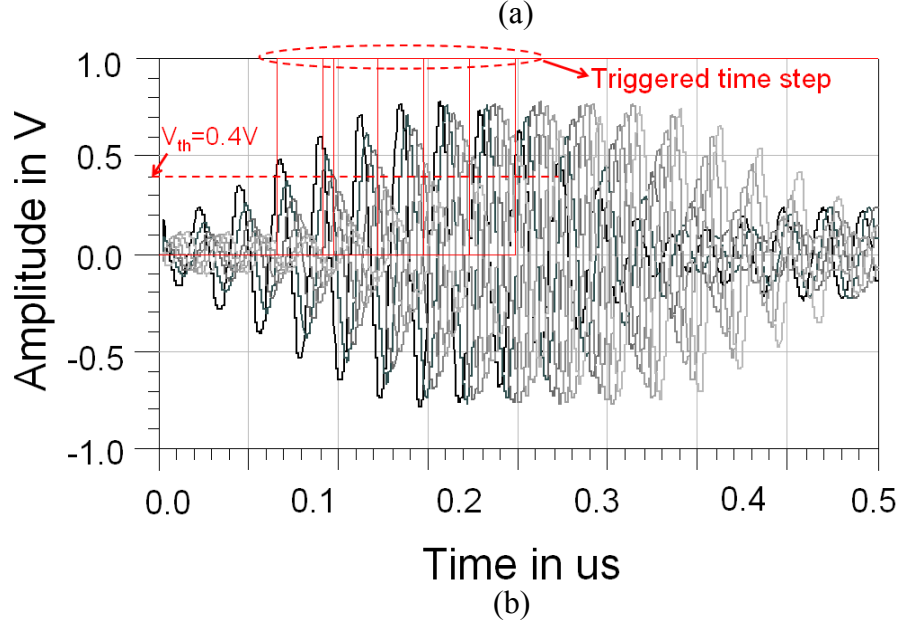
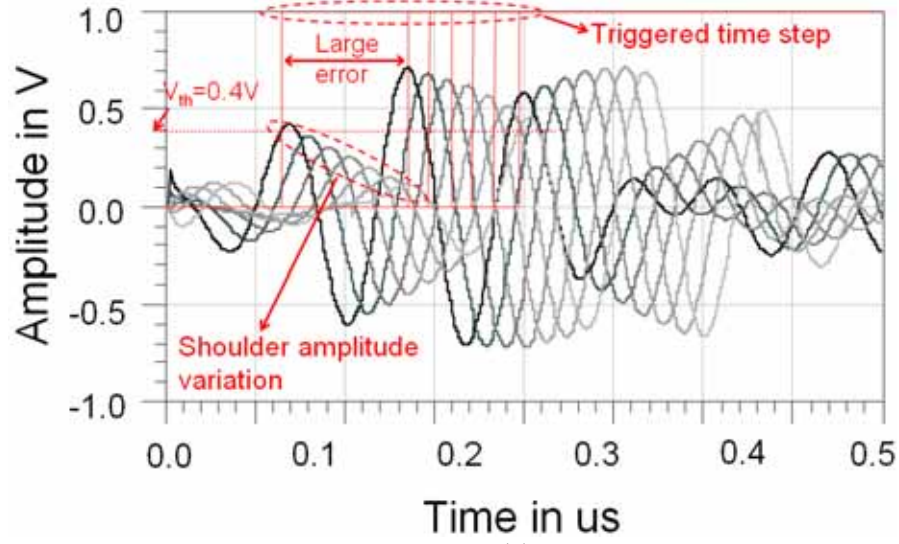


Fig.9.12 “Shoulder” effect when tag is moving one wavelength: a) $\beta=5$; b) $\beta=15$.

9.4.3 Envelope Detection

As discussed in the previous section, in the single channel approach, in order to reduce the system jitter and dynamic error which are directly translated to the localization error, the “shoulder” effect caused trigger error in static and dynamic situations needs to be minimized. According to equation (2.6), the UWB pulse signal information is

contained in $P_{ex}(t)$, the down-converted pulse signal after sub-sampling, which is modulated by the equivalent offset frequency $\Delta\omega_{eq}$. The useful information, i.e. the extended pulse signal $P_{ex}(t)$, has been demonstrated as the envelope of the received wave. Although the modulated single channel signal, i.e. I_{ex} or Q_{ex} , suffered from the “shoulder” amplitude variations, and the corresponding large trigger error, the envelope of the modulated signal remain relatively constant and less sensitive to the trigger error when fixed threshold is applied. A simple Schottky diode based envelope detector is used as an energy collector for the time extended single channel signal. The same dynamic simulation setup has been done with the tag moving continuously away from the base station up to one wavelength at the carrier frequency. Fig. 9.13 shows the simulated results of the same signal in Fig. 9.12, with the fixed threshold set at 0.4 V, but after the energy detection. As shown in Fig. 9.13, the triggered time position varies linearly with the tag movement. The error reduced dramatically as compared to Fig. 9.12. The “shoulder” effect has been minimized and the time position output is insensitive to trigger threshold voltage.

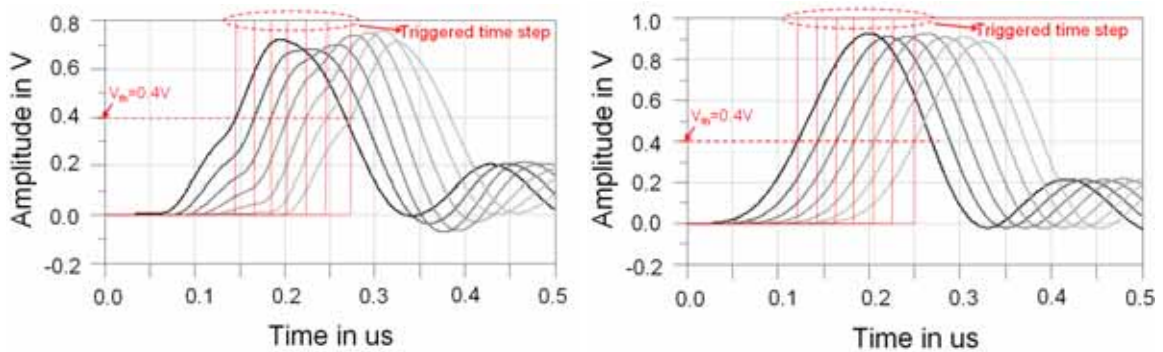


Fig.9.13 Minimize the “shoulder” effect using energy detection when tag is moving one wavelength: a) $\beta=5$; b) $\beta=15$.

Table 9.5 – Simulated Standard Deviation Error for Dynamic Scenario

β	Stand Deviation Error			
	No energy detection		w/ energy detection	
5	18.0 ps	5.40 mm	2.87 ps	0.86 mm
15	6.2 ps	1.86 mm	1.62 ps	0.49 mm

Table 9.5 compares and summaries the 4 cases from Figs. 9.12 and 9.13. The simulated time position errors have been recorded as the tag moving a distance of one wavelength, and the standard deviation errors have been calculated. It shows that the energy detection minimizes the “shoulder” effect and reduces the stand deviation error. All the simulated results here did not include the carrier phase noise effect.

It should be noted from Table 9.5 that for the no energy detection case, the error decreases as the modulation factor β increases. However in reality this is not true when there are phase noise presents, since if the modulation factor β is large, there will be an increased number of “shoulders” close to each other, thus an increased sensitivity to the “shoulder” amplitude variation when fixed threshold is applied. This also explains why in case IV of Table 9.4, both the modulation factor β and the threshold voltage need to be optimized to reduce the “shoulder” effect introduced jitter error.

Noncoherent experiments have been conducted to validate the above simulation results that the combination of low phase noise carrier and the energy detection would reduce the “shoulder” effect and minimized the jitter noise. The tag was put at a fixed position, with continuous 1000 data points recorded. The signal from sub-sampler went

through the A/D converter and fed into a FPGA, where an advanced leading edge detection algorithm was applied to calculate the incoming pulse position. Table 9.6 compares and summaries three measured jitter results: 1) Hittite oscillators without energy detection, 2) low phase noise Agilent carriers without the energy detection, and 3) low phase noise Agilent carriers with the energy detection. As can be seen from Table 9.6, the RMS jitter error reduced from 18.82 mm to 5.73 mm by applying low phase noise carrier sources and energy detection after the sub-sampler. Figure 9.14 plots the measured TDOA raw date error over a 1000 measurement points for these three cases. One interesting effect can be noticed from Fig. 9.15, which plots the case II of Table 9.6, using Agilent LOs without energy detection, where the error is oscillating between ± 10 mm. Such measurement results validate the theory of shoulder effect in the static mode, as shown in Fig. 9.11, where even carrier sources with low phase noise could produce a large error due to the shoulder effect. The error distributions for each of the three cases are plotted in Fig. 9. 16. The Agilent LOs without energy detection shows two distinct Gaussian distributions, corresponding to two different shoulders. After introducing the energy detection, the shoulder effect has been suppressed and the error distribution demonstrates a single Gaussian shape.

Table 9. 6 – Comparison and Summary of Measured Jitter Error

		RMS jitter	
I	Large phase noise carriers, no energy detection	62.73 ps	18.82 mm
II	Low phase noise carriers, no energy detection	41.3 ps	12.04 mm
III	Low phase noise carriers, w/ energy detection	19.1 ps	5.73 mm

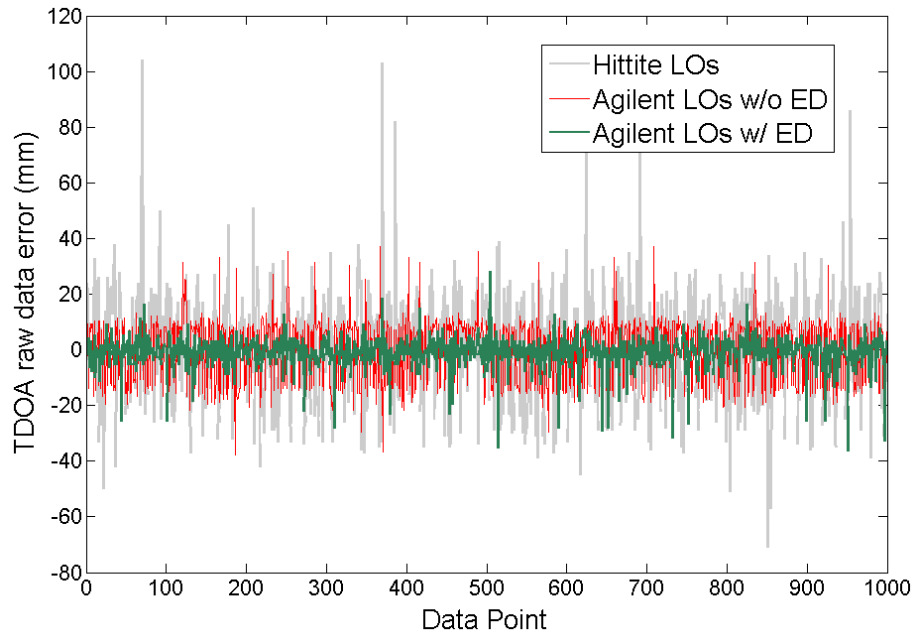


Fig.9.14 The measured TDOA raw data error under different carrier source configurations

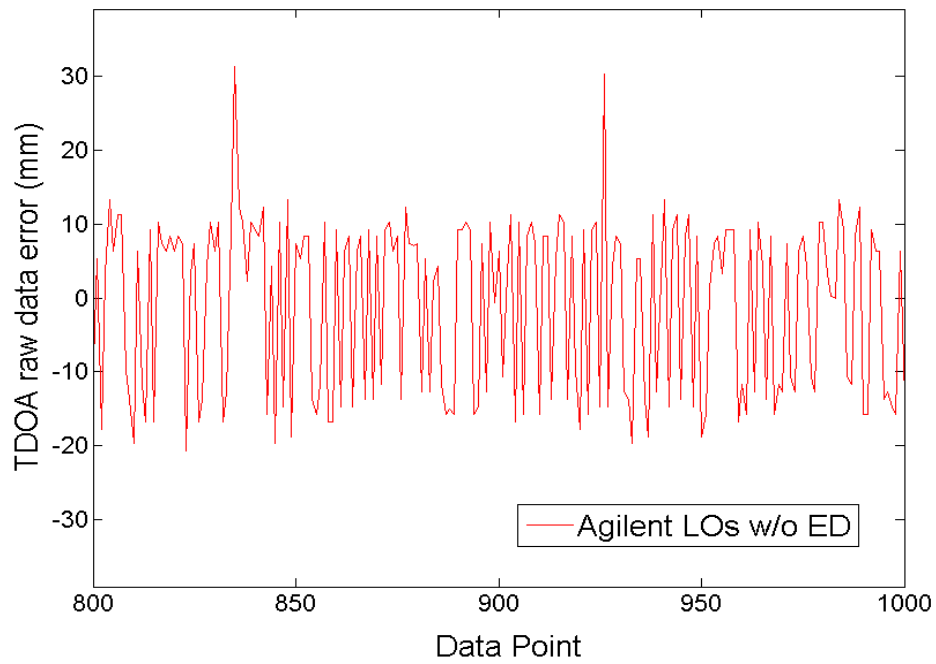


Fig.9.15 The measured TDOA raw data error using Agilent LOs without energy detection.

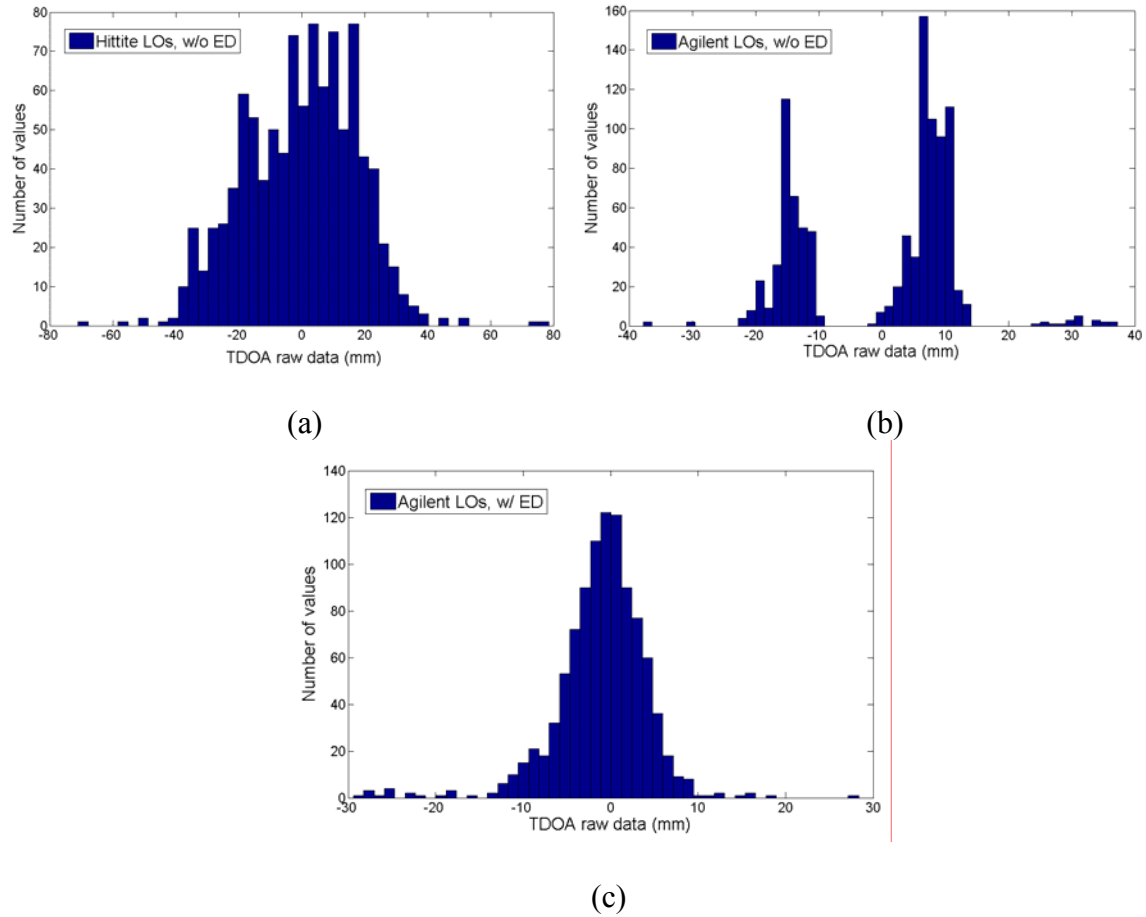


Fig.9.16 The measured TDOA raw data error distribution: (a) with Hittite LOs and no energy detection, (b) Agilent LOs without energy detection, (c) Agilent LOs with energy detection.

The resulted raw TDOA data error with Gaussian distribution could be easily minimized by increasing the number of average, which has been validated through a static 3D experiment, as shown in Table. 9.7. By increasing the number of average from 1 to 106, the 3D static RMS error reduced from 12.21 mm to 1.98mm. However, further increase the number of average could not effectively improve the RMS error, since the 1.98 mm 3D RMS error is mainly from other sources like PDOP and clock stability limitation.

Table 9.7 – Measured 3D RMS Error vs. the Number of Average

# of Average	3D static RMSE (mm)
1	12.21
5	5.87
14	3.66
22	2.84
30	2.67
106	1.98

Figure 9.17 shows the measured sub-sampled output waveforms before and after going through the energy detector. The high noise floor from sub-sampler is caused by the equivalent offset carrier frequency $\Delta\omega_{eq}$, which reduces the signal to noise ratio (SNR) and limits the system dynamic range. After passing through the energy detector, the noise floor has been blocked by the Schottky diode, leaving only the envelope signal. The system dynamic range has been relaxed substantially.

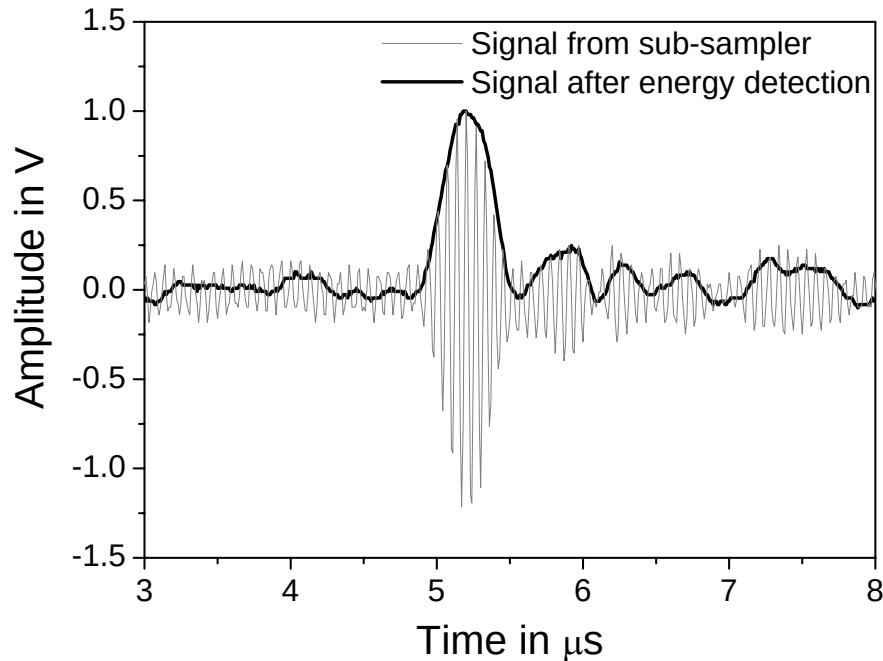


Fig.9.17 Measured sub-sampled output waveforms and the signal after energy detection.

9.5 Conclusion

A novel receiver architecture which combines the single channel carrier based UWB receiver and traditional energy detection based UWB receiver has been proposed. The UWB localization system is equipped with low noise carrier sources at both transmitter and receiver sides, and the advanced sub-sampling mixer for equivalent sampling the incoming pulse train. A proper equivalent offset carrier frequency $\Delta\omega_{eq}$, i.e. the modulation factor β , is intentionally chosen between the transmitter and receiver carrier, eliminating the requirement of carrier synchronization. We have addressed step-by-step the main challenges being faced that leads us to the finalized system architecture, including the I/Q mismatch, jitter errors due to phase noise in carrier offsets, “shoulder” effect in static and dynamic scenario, etc. Both simulation and measurement shows the robust of the proposed receiver architecture, with a reduced timing jitter error, improved received signal SNR, and insensitivity to the triggering threshold voltage. In the next chapter, 1-D and 3-D unsynchronized experiment results will be presented to further validate the theories discussed in this chapter, where constant mm-range accuracy in both static and dynamic scenario will be demonstrated by applying the novel receiver UWB scheme.

Chapter 10

Noncoherent Experiment

In the previous chapter, an approach of a novel noncoherent UWB receiver system has been investigated, i.e. a carrier based UWB system with an advanced sub-sampling procedure followed by an energy detection circuitry. In this chapter, extensive noncoherent 1-D and 3-D localization experiments were performed where a mm-range accuracy has been constantly achieved, validating the various hypothesis related to our newly developed system. One major design issue here is that the crystal clocks and the carriers between both the tag and the receivers are unsynchronized. The Optotrak 3020 has been used to obtain accurate reference data and has been extensively used to validate our results except for 1D experiment. In the 1D experiment, a Newport precision optical rail PRL-24 [135] is used for position reference with sub-mm accuracy.

The monopole transmitting antenna was within $\pm 20^\circ$ of the broadside direction of the Vivaldi antennas to minimize the phase center variation at the base stations. The achieved mm-range accuracy in the developed real time noncoherent system represents the world's best localization accuracy up to date based on UWB technique to the best of the author's knowledge.

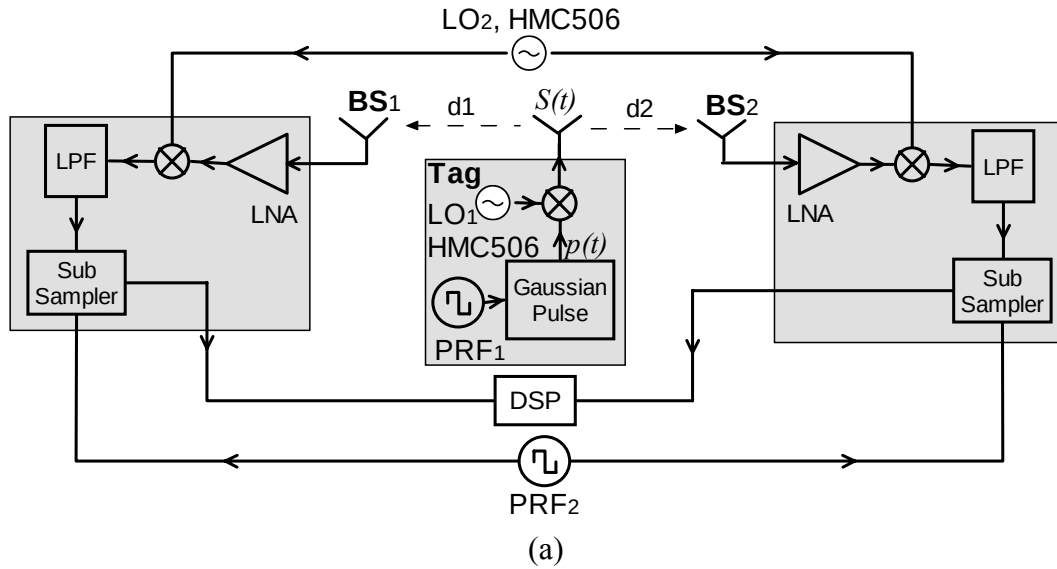
The organization of this chapter will be as follows. In the first section, two unsynchronized 1-D experiments are compared under different receiver schemes: the first

one with Hittite LOs and no energy detection has been applied; the second one with Agilent LOs (low phase noise compared to Hittite LOs) and with energy detection after sampling mixer. The 1D results show that the receiver architecture with energy detection and low phase noise LOs require less times of average and produce less ranging error. In the second section, how the energy detection has improved the 3-D accuracy under both the static and dynamic scenarios will be presented. Thirdly, a 3-D experiment with a tag moving back and forth along the rail has been performed to test the repeatability of the system. Besides we will investigate the effect of fictitious time scaling caused by the undesired clock drift and its impact on the system accuracy will be investigated. Next, a robot controlled 3-D experiment with metals (e.g., robot arm) in a close distance to the transmitting tag has been investigated to study the impact of multipath effects. Finally the error discussions are given followed by a conclusion.

10.1 Unsynchronized 1-D Experiment

Two 1-D experiments with unsynchronized LOs and PRF clock sources were carried out to validate the hypothesis and the simulations of the previous chapter and to test the robustness of our system. The two experimental setups are shown in Fig. 10.1, where only two base stations are needed for the 1-D measurements. The narrow bandwidth output of the sub-sampler (i.e. the original pulse bandwidth divided by the extending factor) is fed through an ADC circuitry. Next, the output signals are fed to the FPGA block which uses RSS algorithm to locate the pulse position, but later on in our 3D experiment we adapted a new leading edge detection algorithm. The differences between the two 1D experimental setups in Fig.10.1 are below:

- ◆ In Fig. 10.1(a) Hittite HMC506 LOs with a relatively high phase noise are used at both the transmitter and receiver. Whereas in Fig. 10.1(b) Agilent low phase noise LO sources E8257D and 83622B are used at both the tag and receiver respectively, and the frequency offset between these two LOs is set to give a modulation factor β approximately equals six (the definition of the modulation factor can be found in Chapter 9, equation (9.1));
- ◆ The envelope detectors are used following the sub-sampler at the receiver in Fig. 10.1(b), whereas in Fig. 10.1(a) no envelope detectors are used.



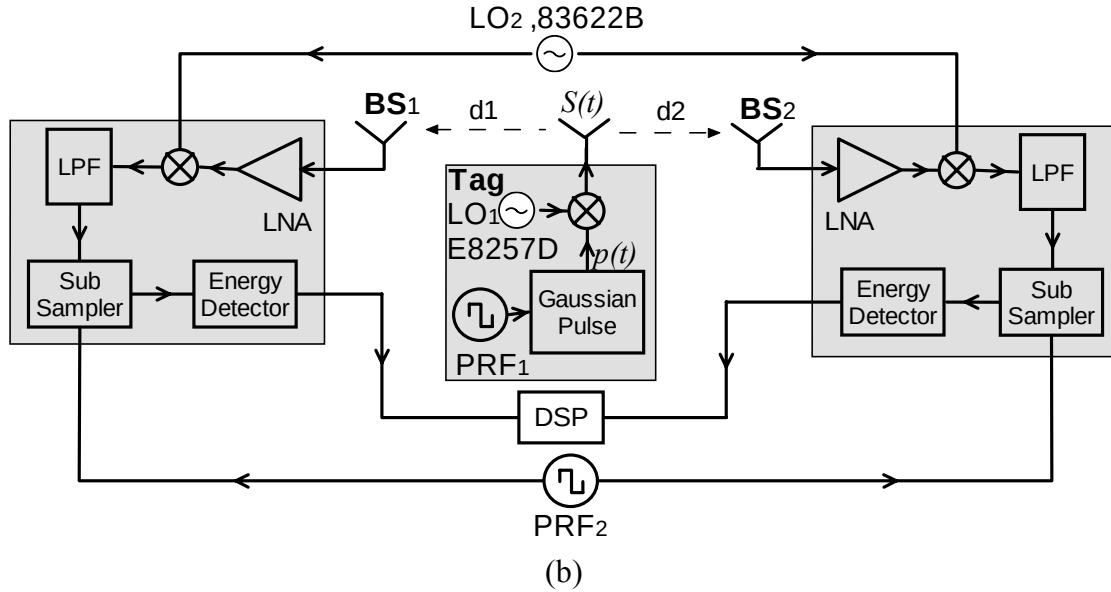
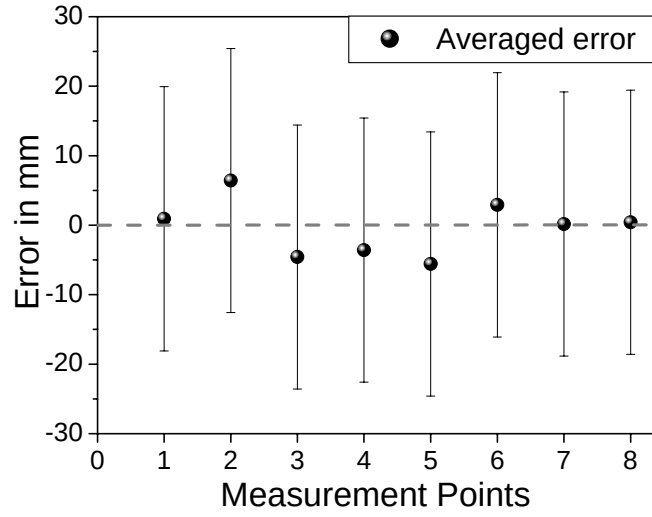
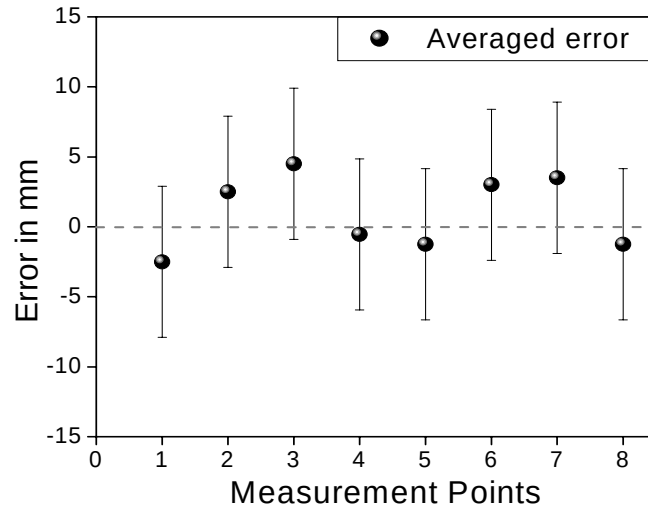


Fig.10.1 Experimental setup for 1-D unsynchronized positioning measurement: (a) Hittite LOs at the tag and receiver, no energy detector; (b) Agilent low phase noise LOs at the tag and receiver, with energy detection after the sub-sampler.

For both cases, a mm-range accuracy was consistently achieved for the 1-D unsynchronized measurements at 8 separate locations along the Newport rail with a 5 cm distance between any two successive measurements. As shown in Fig. 10.2(a), the system jitter can cause noticeable short term variation in the error at each static point of roughly ± 19 mm. This short term variation was mitigated by averaging 32 pulses at each static point. For the single channel scheme with a low phase noise carrier and energy detection, results shown in Fig. 10.2(b) demonstrate the system's jitter has a much smaller short term variation of roughly ± 6 mm at each static point, compared to the ± 19 mm shown in Fig. 10.2(a). Also such small short term variation is insensitive to the chosen triggering threshold voltage using RSS. This small short term variation was mitigated by averaging only 4 pulses at each static point.



(a)



(b)

Fig.10.2 Measured error of the 1-D unsynchronized experiment: (a) Hittite LOs in the tag and receiver, no energy detection; (b) Agilent low phase noise LOs at the tag and receiver, with energy detection after sub-sampling.

In Table 10.1 we compared the results between both cases, it is clear that the single channel scheme with a low phase noise carrier and energy detection requires less times of averaging and produces less 1D error, validating our assumptions in the previous section. Compared to the coherent experimental results shown in Table 8.3, the mean error in

Table 10.1 – Comparison of the Noncoherent 1D Experiment Results

	w/o energy detection	w/ energy detection
Static Variation	± 19 mm	± 6 mm
Mean Error (mm)	3.07	2.38
Std Dev. Error (mm)	2.39	2.66
Worst Case (mm)	6.4	4.50
Required times of averaging	32	4
LO Carrier Phase noise	High	Low
Energy Detection	No	yes

measuring the 1-D static data increases from 1.49 mm to 2.38 mm. The increase in error of 0.89 mm is comparable to the measured error of 1.05 mm due to the PRF clock jitter discussed in Section 9.3.

10.2 Unsynchronized 3D Experiment – Static and Dynamic

In our earlier stage of developing a coherent 3-D experiment as illustrated in Chapter 8, the tag clock PRF_1 was synchronized with the base station clock PRF_2 , and the received I/Q channels data were averaged with a large number, e.g. greater than 100 times, and then stored in the Tektronix TDS8200 digital sampling oscilloscope for post processing. Obviously, the previous coherent 3-D experiment was not a real-time system.

Figure 10.3 shows a 4 base stations distribution with pre-defined locations for each base station utilizing the Optotrak system. Notice that the spatial spread of the base stations along the z-axis is the largest (2498 mm), while along the x-axis is the smallest (1375 mm).

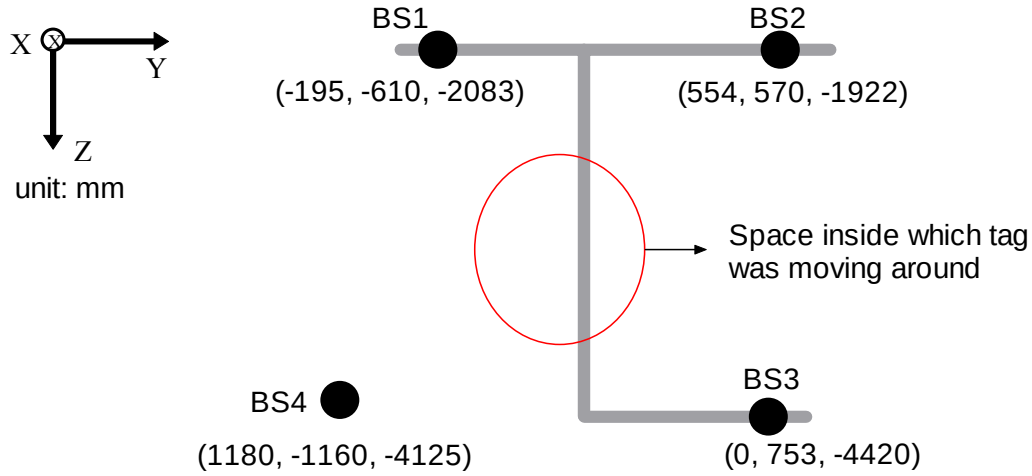


Fig.10.3 3-D unsynchronized localization experiments, 4 base stations distribution with locations for each base station.

To further validate the theory in Chapter 9 and evaluate how energy detection could improve the system performance, two 3-D experiments with unsynchronized LOs and PRF clock sources were carried out as shown in Fig. 10.4, where a minimum of four base stations are needed for the 3-D measurements. The signals from the sub-sampler are then fed to the FPGA which uses the newly developed leading-edge detection algorithm to locate the pulse position. This more powerful algorithm is written by a team member of our group and has been adapted in all our current systems [98]. The differences between the two experimental setups are highlighted in Fig. 10.4. The energy detectors are used following the sub-sampler at the receiver in Fig. 10.4(b), whereas in Fig. 10.4(a) no energy detectors are used. The raw time difference data between every two receivers went through 17 times of averaging before fed into the TDOA algorithm.

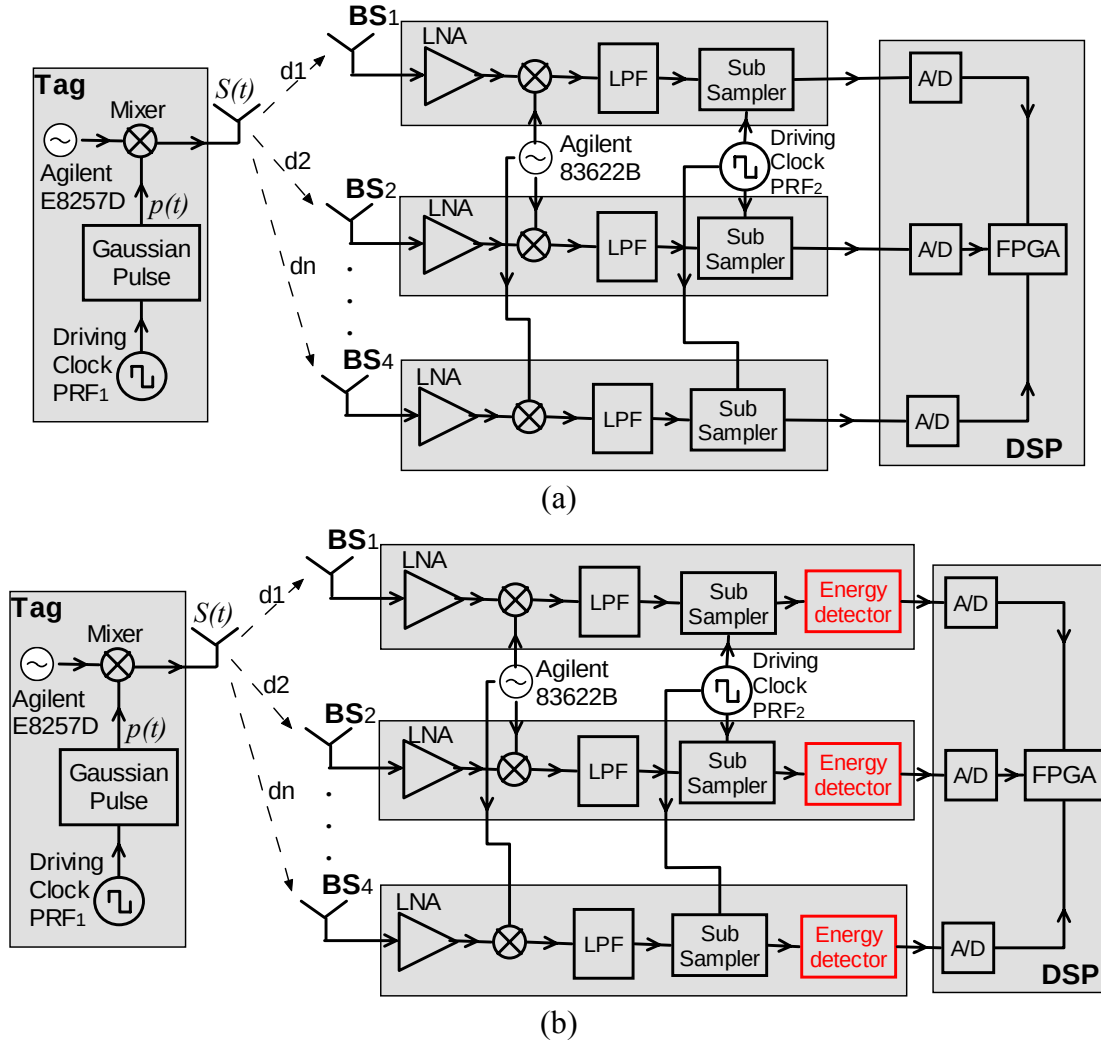


Fig.10.4 3-D experiment setup of unsynchronized localization system using a single channel demodulation: (a) without energy detection; (b) with energy detection.

Both static and dynamic measurements are conducted based on the setups shown in Fig. 10.4 in a similar experimental environment. In the static mode, the tag is located at a fixed position and a 1000 data points were taken and compared to the Optotrak measured data. Whereas in the dynamic mode, the tag is moving randomly inside the 3-D space indicated in Fig. 10.3. The 3-D motion traces of the tag are then plotted and UWB measured traces are compared with the Optotrak measured traces. Root mean square error

(RMSE) is used to report the error since it is a good measure of error resulting from both the accuracy and precision.

Table 10.2 lists a comparison of the noncoherent 3-D experiment results with and without energy detection under both static and dynamic mode. It shows that by adding energy detection, the static RMSE has been improved by 1.5 mm and the dynamic RMSE has been improved by 2.6 mm. Those results again confirm the assumptions in Chapter 9 and are related to a minimal “shoulder” effect when utilizing the energy detection method.

To visualize the above measured results we study the static and dynamic 3-D experimental results with energy detection more closely. Figure 10.5 shows the 3D static experiment with the energy detection results where the errors at x , y , and z axes over a 1000 measurement points are plotted. It should be noted that the error along the x -axis contributed most to the overall distance error, which can be explained by the largest PDOP value along the x -axis. Table 10.3 summaries the RMSE contributed in each coordinate direction and the corresponding PDOP value.

Table 10.2– Comparison of the Noncoherent 3D Experimental Results

	w/ out Energy Detection	w/ Energy Detection
Static RMSE	4.74 mm	3.26 mm
Dynamic RMSE	8.95 mm	6.37 mm

Table 10.3 – RMSE vs. PDOP at Each Axis in the Static Experiment with Energy Detection

	x	y	z
PDOP	1.315	0.891	0.678
RMSE	2.349 mm	1.777 mm	1.401 mm

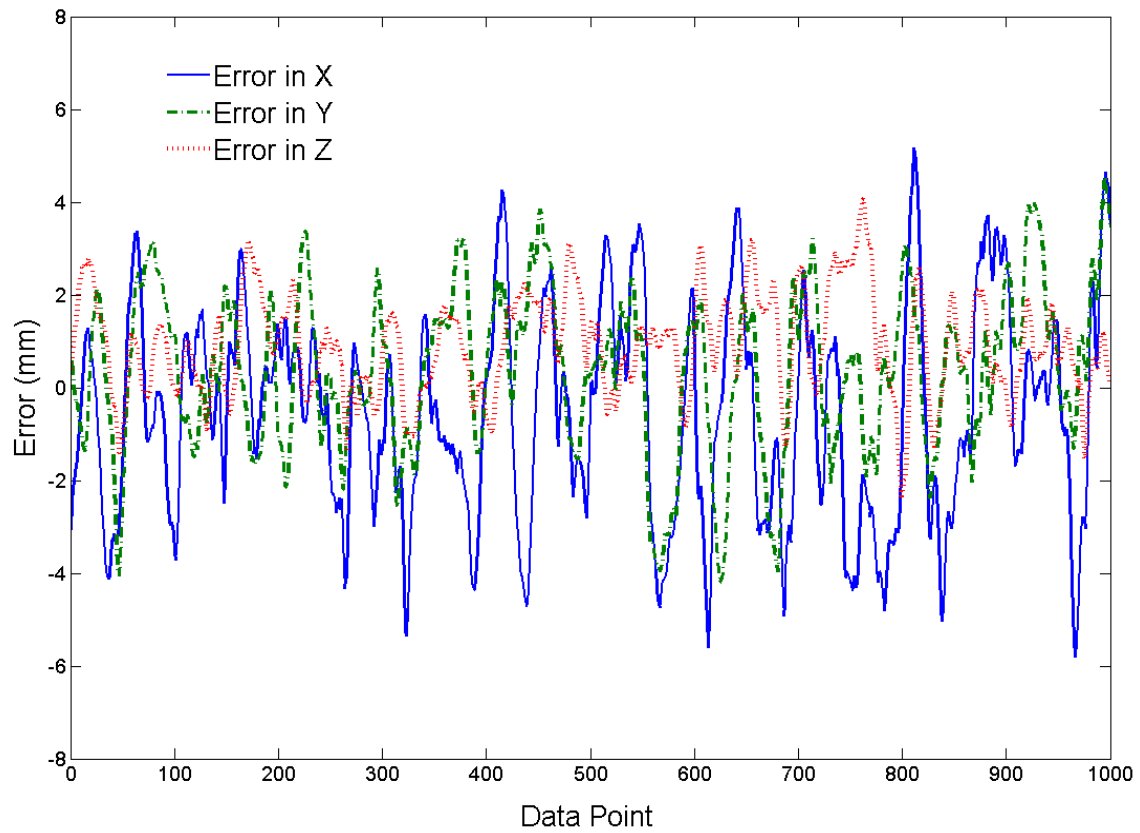
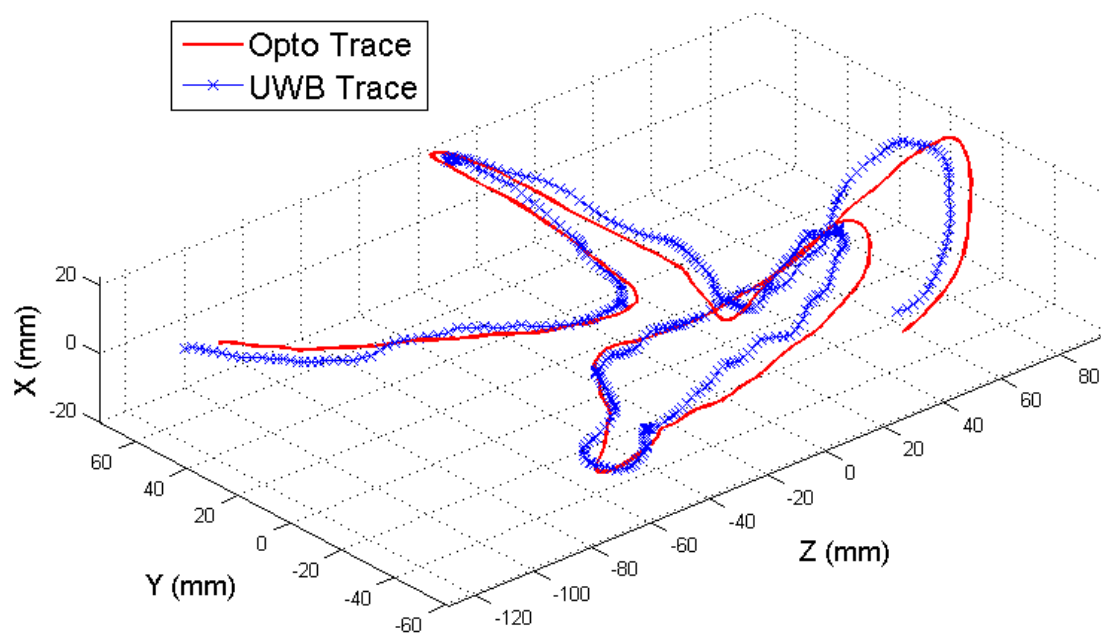
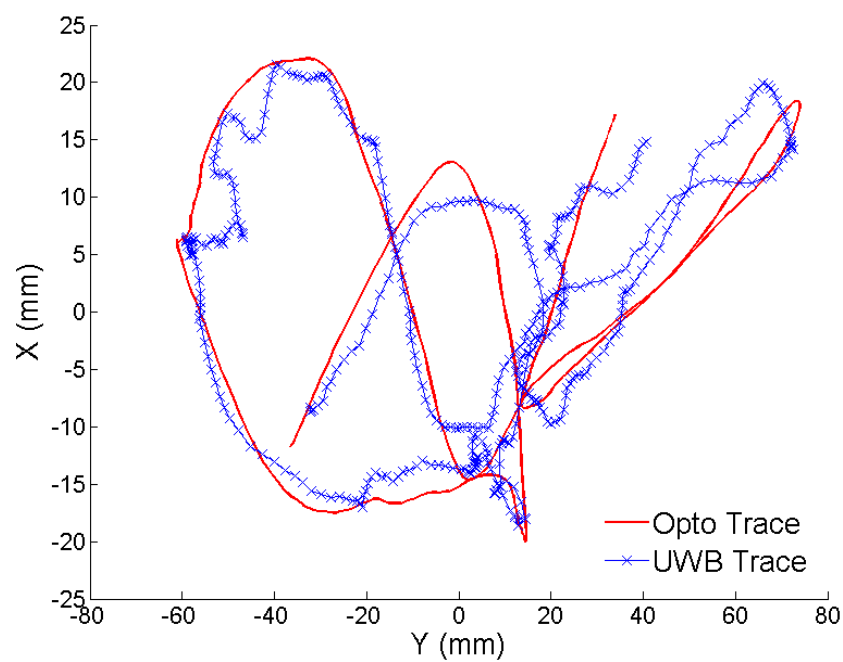


Fig.10.5 3-D static mode with energy detection, x , y and z axes error compared to the Optotrak System.

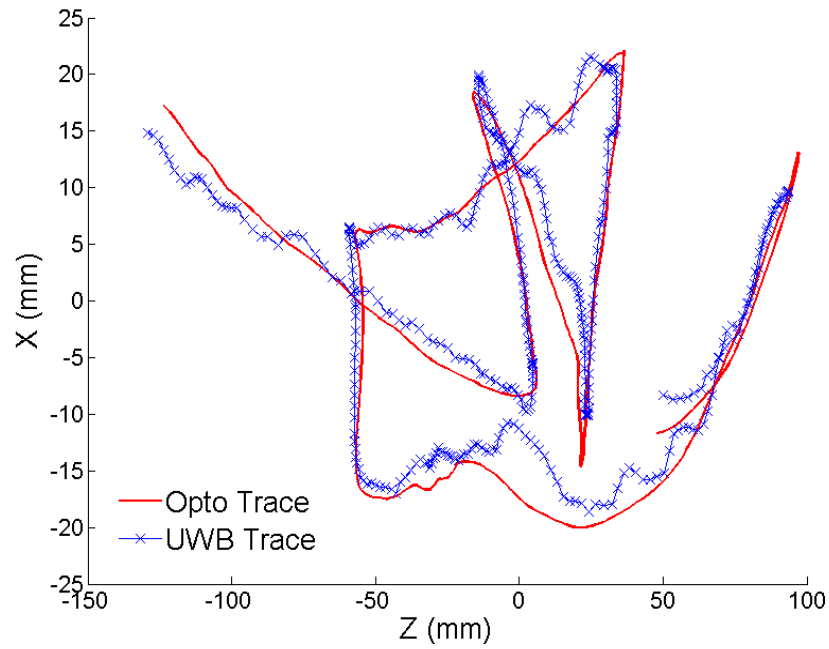
Figure 10.6 plots the UWB trace and Opto trace in the 3D dynamic mode with energy detection. Both the 3-D view and 2-D views from different planes are shown.



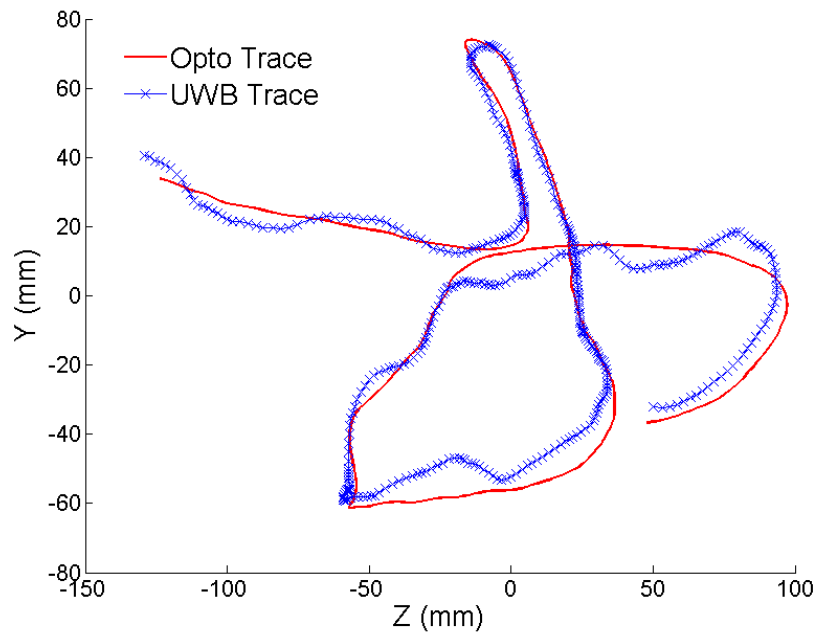
(a) 3-D view



(b) XY plane view



(c) XZ plane view



(d) YZ plane view

Fig.10.6 3-D dynamic mode with energy detection, UWB trace compared to Opto trace:
(a) 3-D view; (b) XY plane view; (c) XZ plane view; (d) YZ plane view.

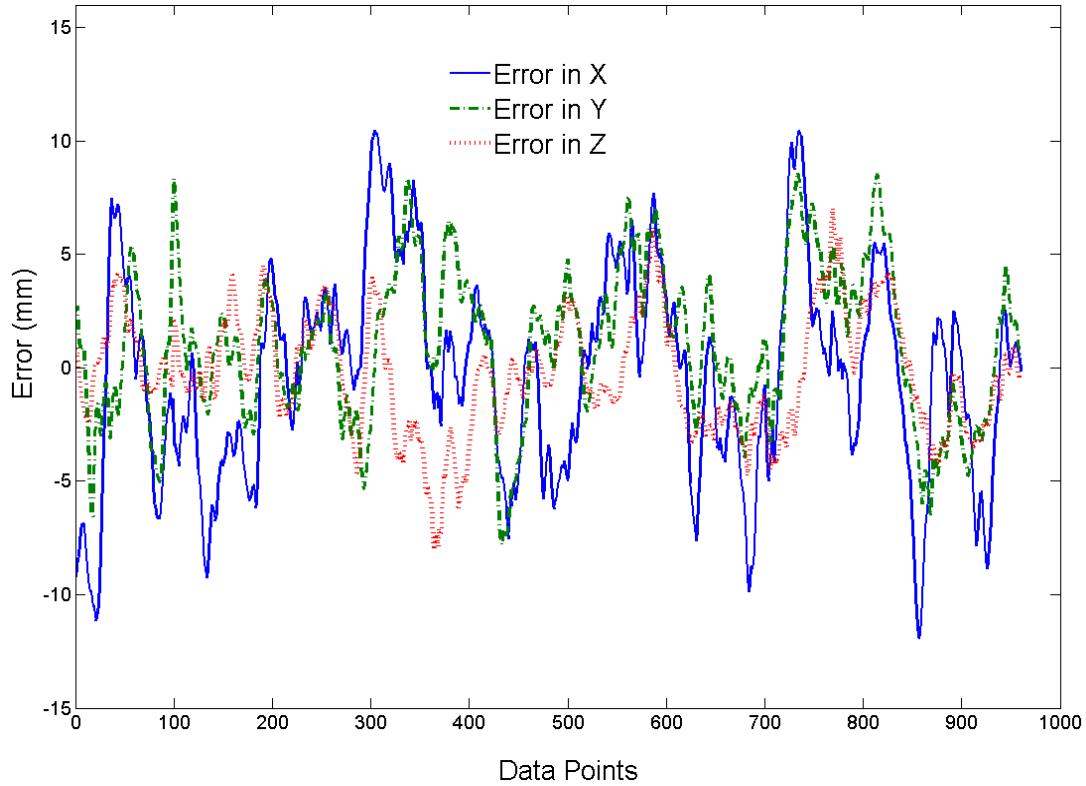


Fig.10.7 3-D dynamic mode with energy detection, x , y and z axes error compared to Optotrak System.

Figure 10.7 shows the 3D dynamic errors at x , y , and z axes over a 1000 measurement points. Once again, the error along the x -axis contributed most to the overall distance error, which can be explained by the largest PDOP value along the x -axis.

10.3 Unsynchronized 3D Experiment – Dynamic Rail Movement

To test the repeatability and robustness of our system, the tag is moving along the rail back and forth several times as shown in Fig. 10.8. Figure 10.9 plots the UWB trace together with the reference Opto trace and both the 3-D and 2-D views from different planes are shown.

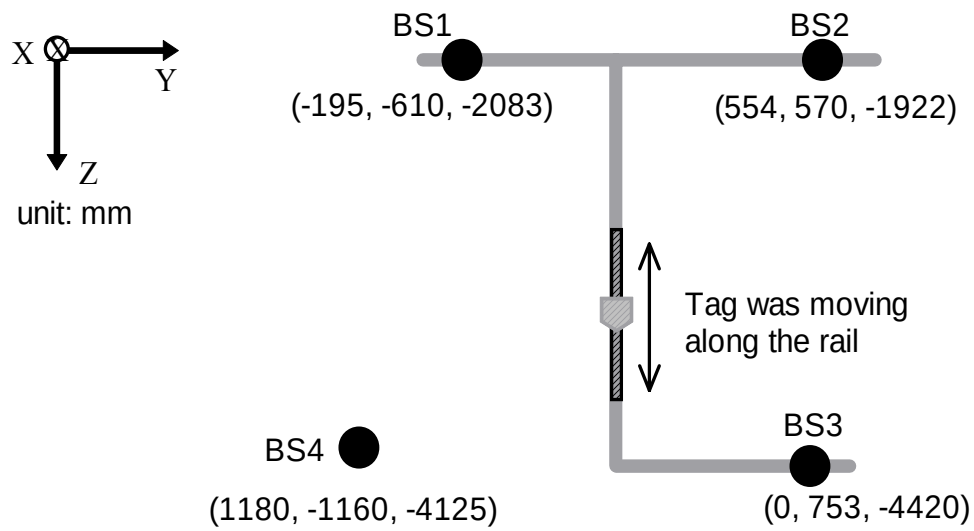
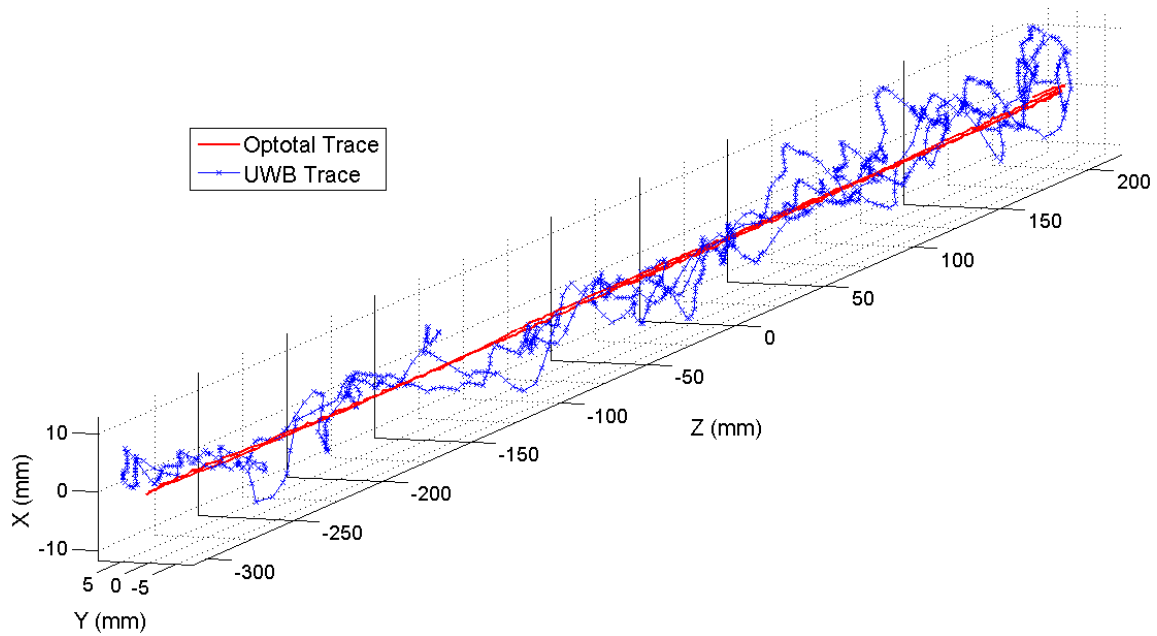
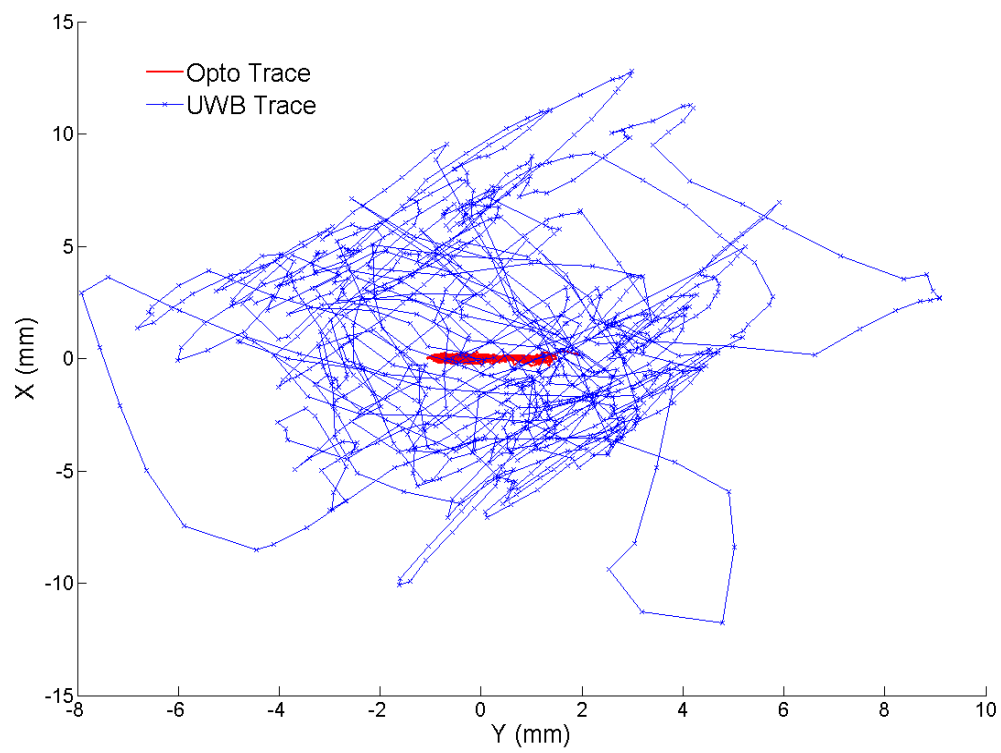


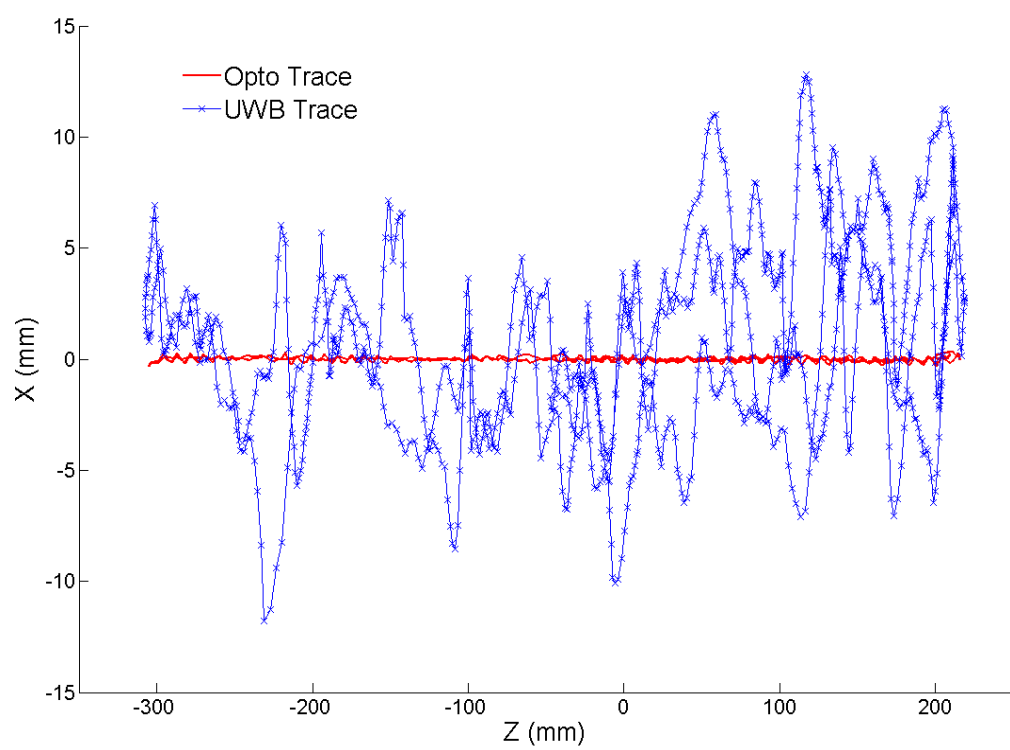
Fig.10.8 3-D unsynchronized localization experiments, with the tag moving back and forth along the rail.



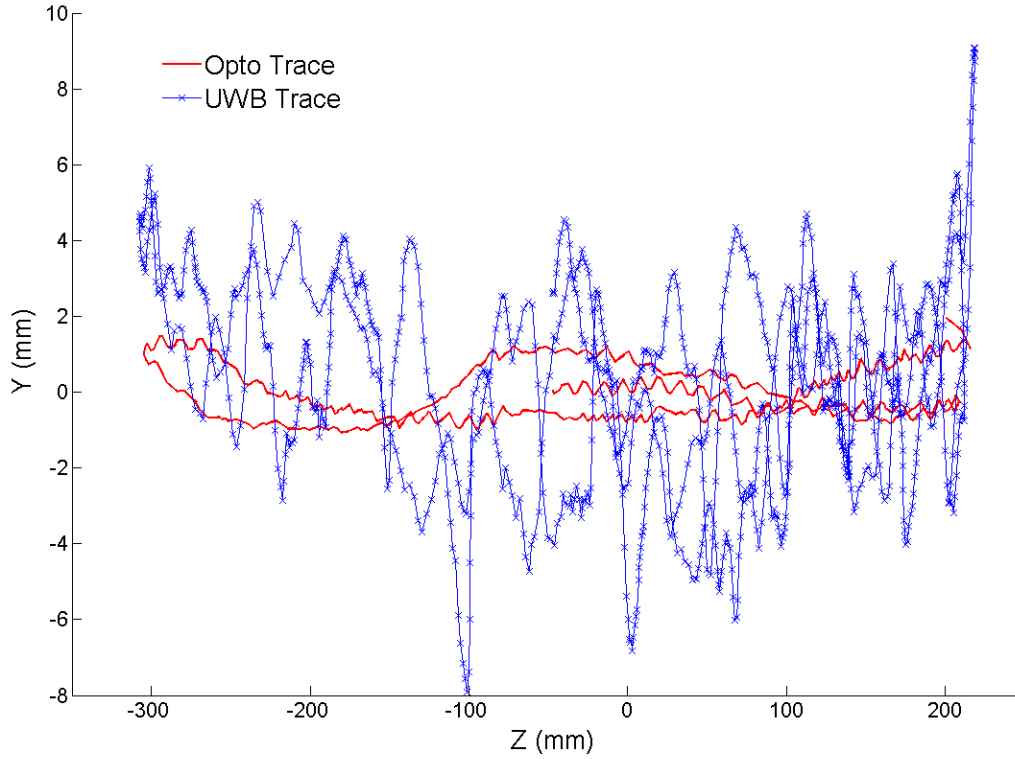
(a)



(b)



(c)



(d)

Fig.10.9 3-D dynamic measurement with the tag moving along the rail, UWB trace compared to Opto trace: (a) 3-D view; (b) XY plane view; (c) XZ plane view; (d) YZ plane view.

Figure 10.10 shows errors at x , y , and z axes over a 1500 measurement points. The error along the x -axis has contributed the most to the overall distance error, which can be explained by the largest PDOP value along the x -axis, as shown in Table 10.4. However, the PDOP in the z direction has a smaller value but contributes more to the RMSE error, which raises up an interesting question: why a larger PDOP value would yield a smaller error? does the PDOP theory fail? The answer is absolutely “NO”. To answer this question, the impact of the tag velocity on the system error needs to be addressed as well. As indicated in Fig. 10.8, the tag is moving back and forth along the z -axis thus the

velocity is mainly appeared in the z direction, as shown in Fig. 10.11. Figure 10.12 shows that the system error increases in a linear fashion with velocity. Thus, even though the PDOP value along the z -axis is smaller than that in the y -axis, the velocity effect renders a larger z -axis error than that for the y -axis. The velocity related error will be observed in a robot tracking experiment in the next section, where the faster robot motion produce the higher ranging error. Such effect can be explained to a certain extent by the time lag between the UWB system and the Opto reference system at higher speed.

Table 10.4– RMSE vs. PDOP at Each Axis

	x	y	z
PDOP	1.553	0.983	0.818
RMSE	3.979 mm	2.176 mm	2.776 mm

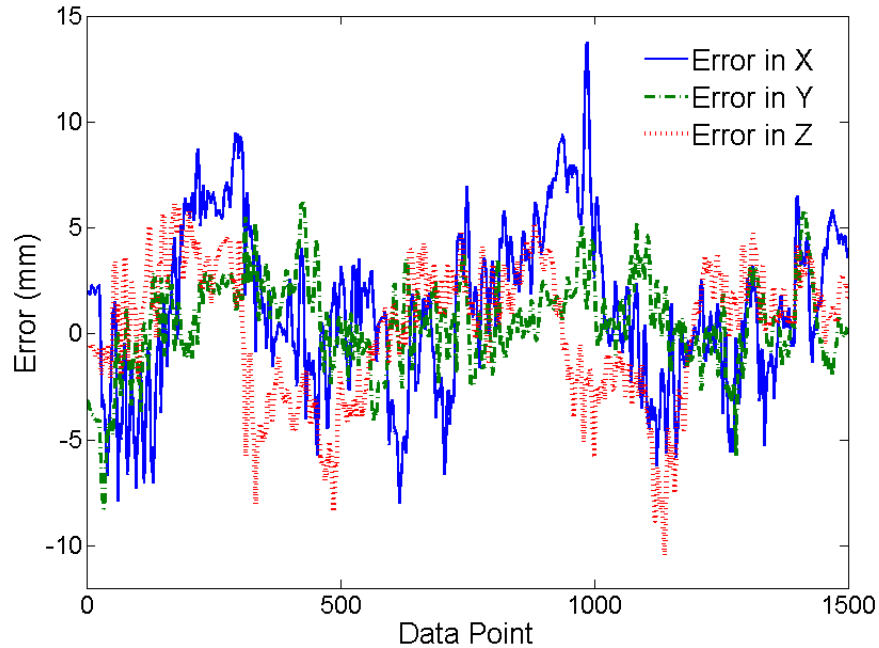


Fig.10.10 3-D noncoherent rail experiment, x , y and z axes error compared to Optotrak System.

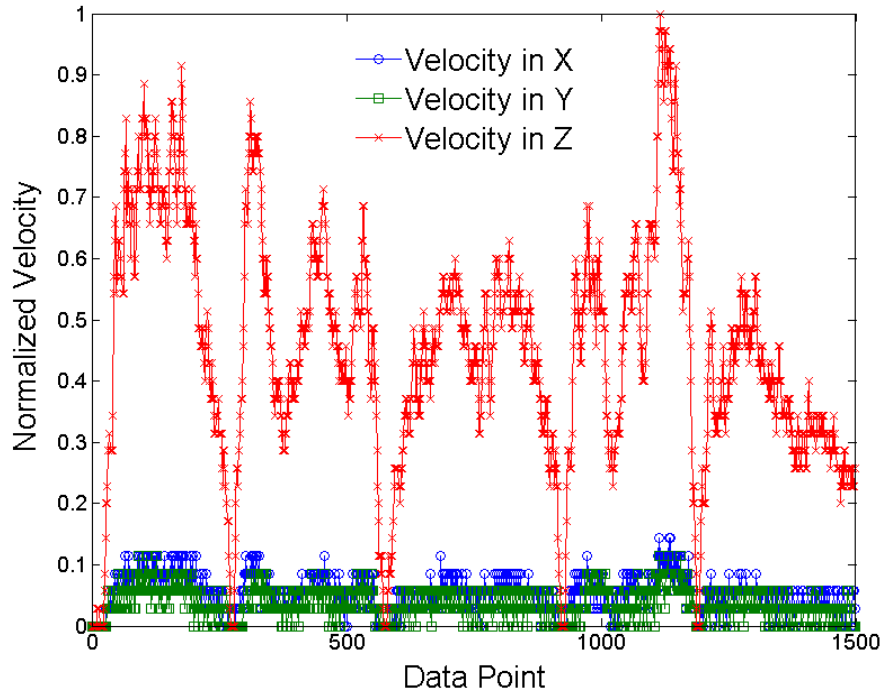


Fig.10.11 3-D noncoherent rail experiment: normalized velocity in x, y and z axes.

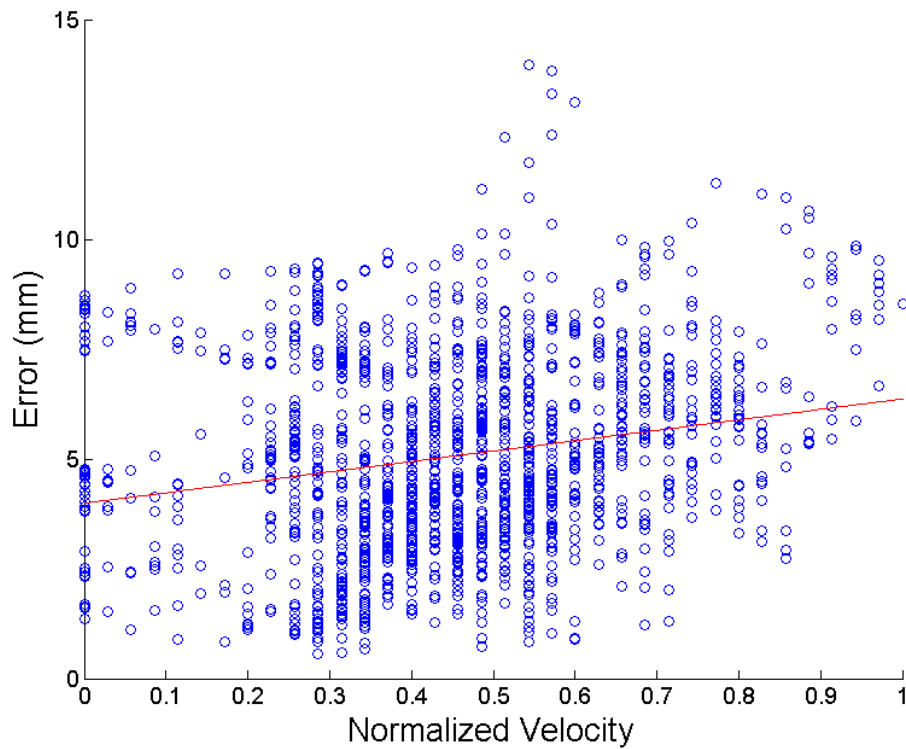
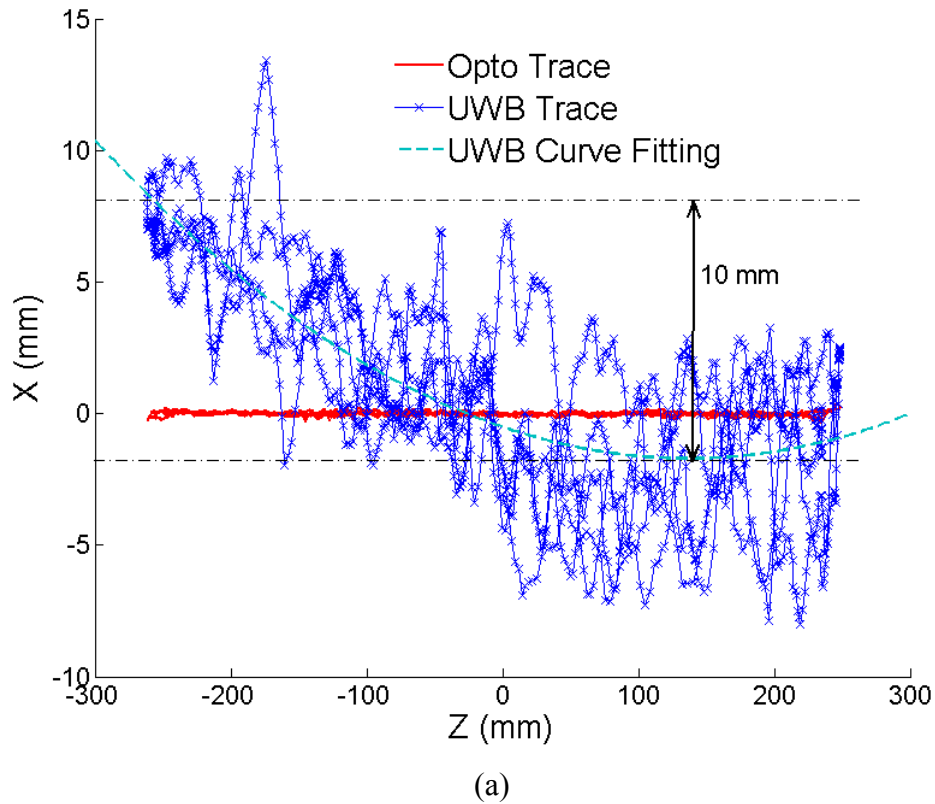


Fig.10.12 3-D noncoherent rail experiment: linear curve fitting shows that the RMSE increase with increasing the tag velocity.

The time scaling caused by the clock drift due to the thermal stability of the clocks has been discussed in Section 3 of Chapter 9, and a 1-D ranging error of roughly 1% was predicted for 1 kHz frequency offset. Such error was also observed in one of the dynamic rail experiments. As shown in Fig 10.13(a), in XZ plane, a 10 mm systematic error was observed with the tag movement exceeding a 50 cm along $-z$ direction, which is a 2% error. Whereas looking from the YZ plane, shown in Fig 10.13(b), the clock drift has demonstrated a 5.6 mm offset, which is roughly a 1% error. Such effect can be explained by the fact that the PDOP value of the x -axis is nearly as twice as the PDOP of the y -axis, as shown in Table 10.4. Such error can be effectively mitigated by increasing the strobe frequency or increase the offset frequency from 1 kHz to 2 kHz as discussed in Chapter 9.



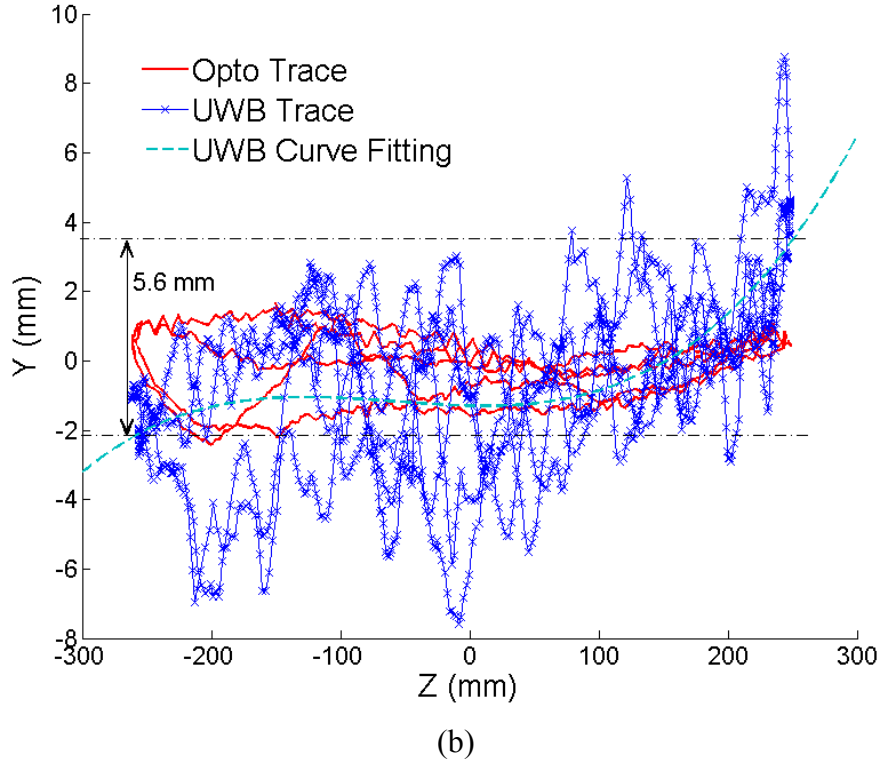


Fig.10.13 Systematic error caused by the clock drift in the 3-D dynamic measurement with tag moving along the rail: (a) XZ plane view; (b) YZ plane view.

10.4 Unsynchronized 3D Experiment – Robot Movement

The next noncoherent 3D experiment is to dynamically track the robot position. The experimental setup is shown in Fig. 10.14. The monopole antenna and the reference Optotrak probe are tied together and then fixed to the arm of the Robot CRS T475. The robot was pre-programmed to specifically cover 20 distinct static positions in a 3D volume, stopping for one second at one position than moving to the next position and so on. The measured traces by the UWB system are compared to the Optotrak reference system as shown in Fig. 10.15. Figure 10.16 shows the 20 distinct static positions taken by both the UWB and the Optotrak systems. The frequency offset is firstly set at 1 kHz.

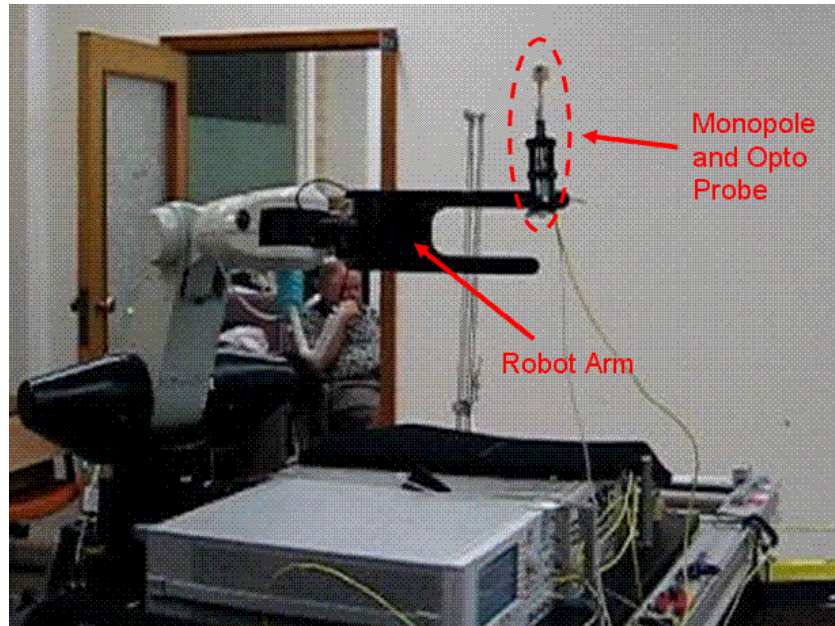
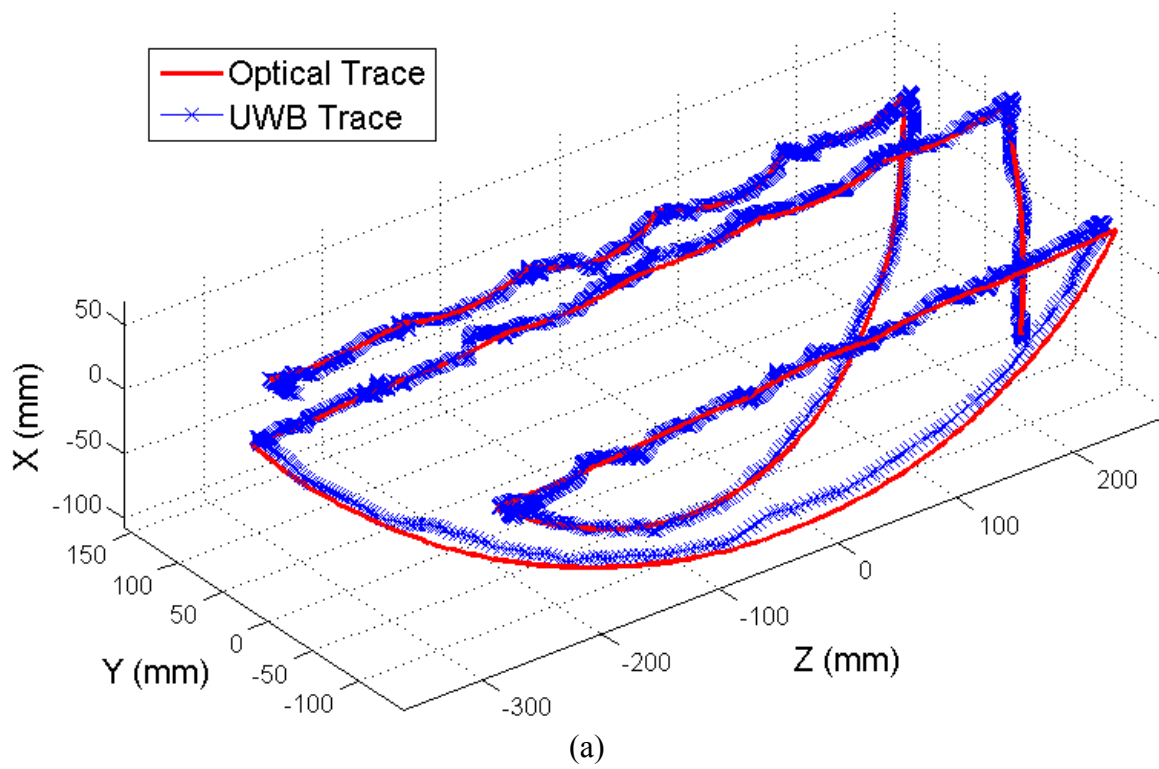
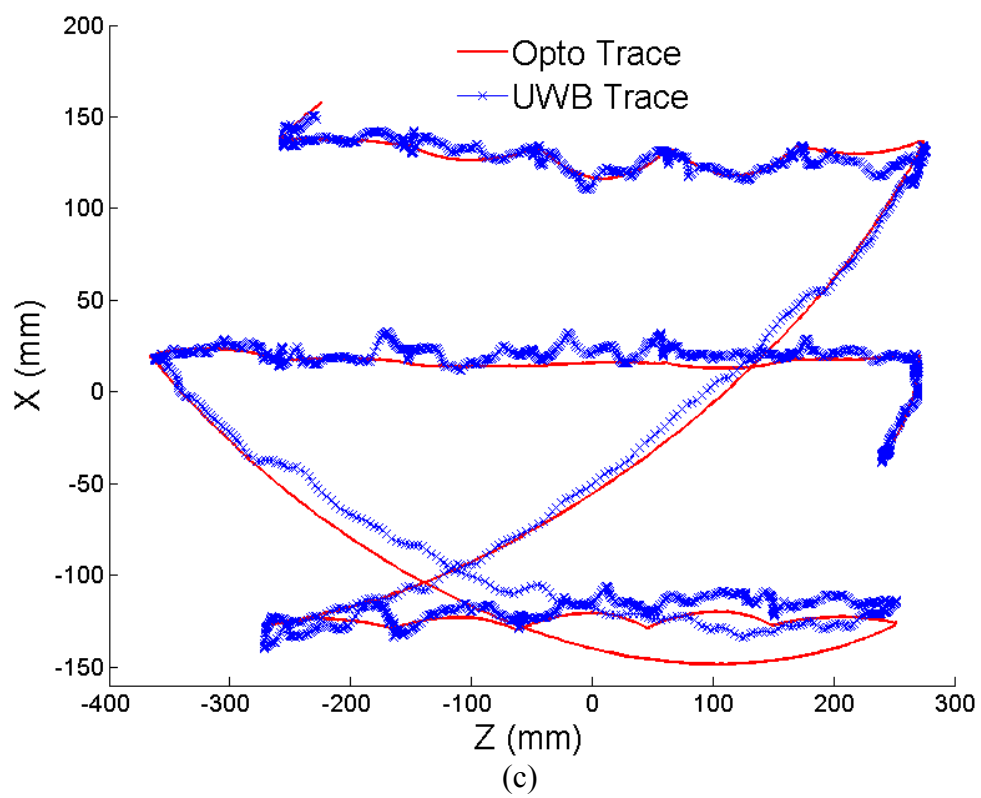
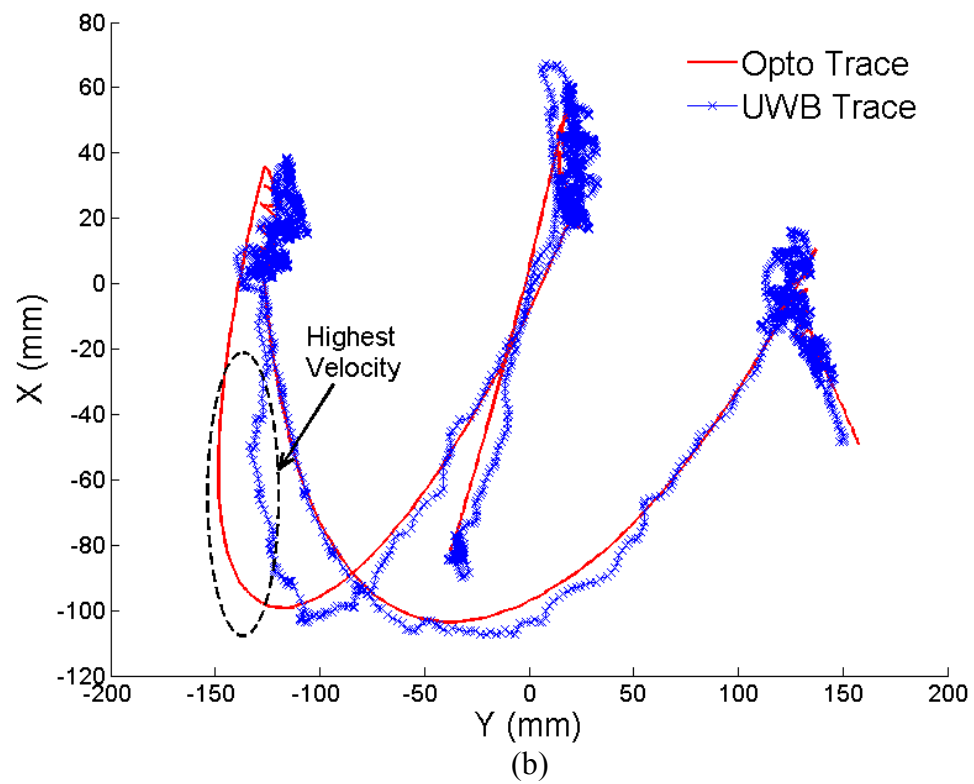


Fig.10.14 Experimental setup of robot tracking using the developed noncoherent UWB system.





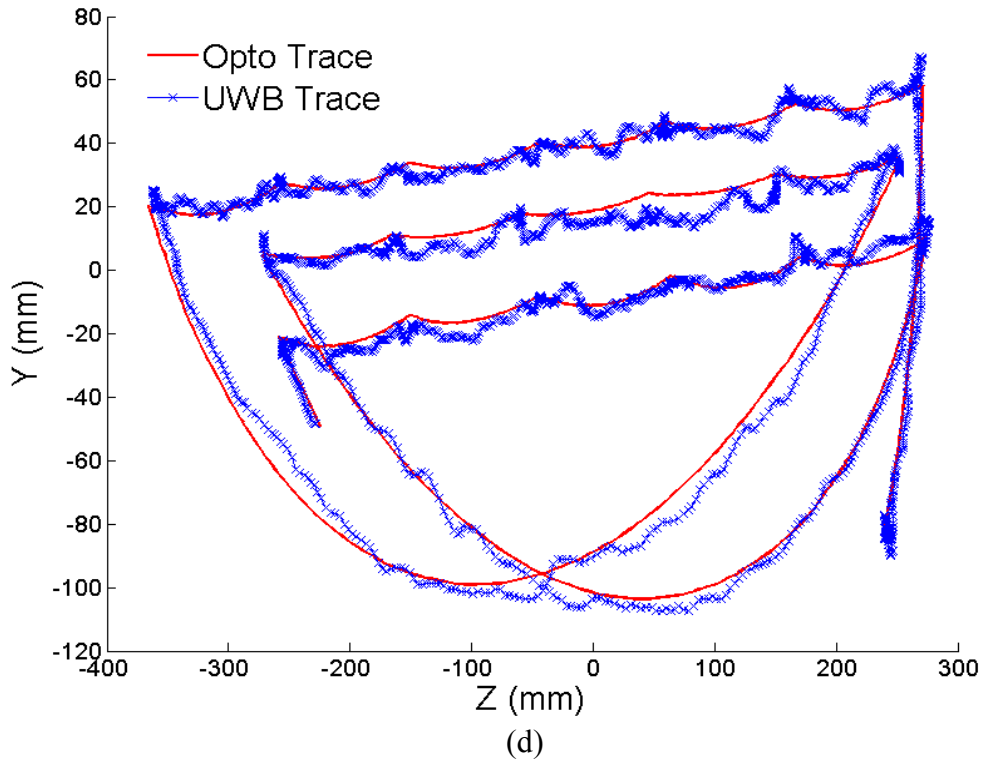


Fig.10.15 3-D dynamic robot tracking with 1 kHz offset, UWB trace compared to Opto trace: (a) 3-D view; (b) XY plane view; (c) XZ plane view; (d) YZ plane view.

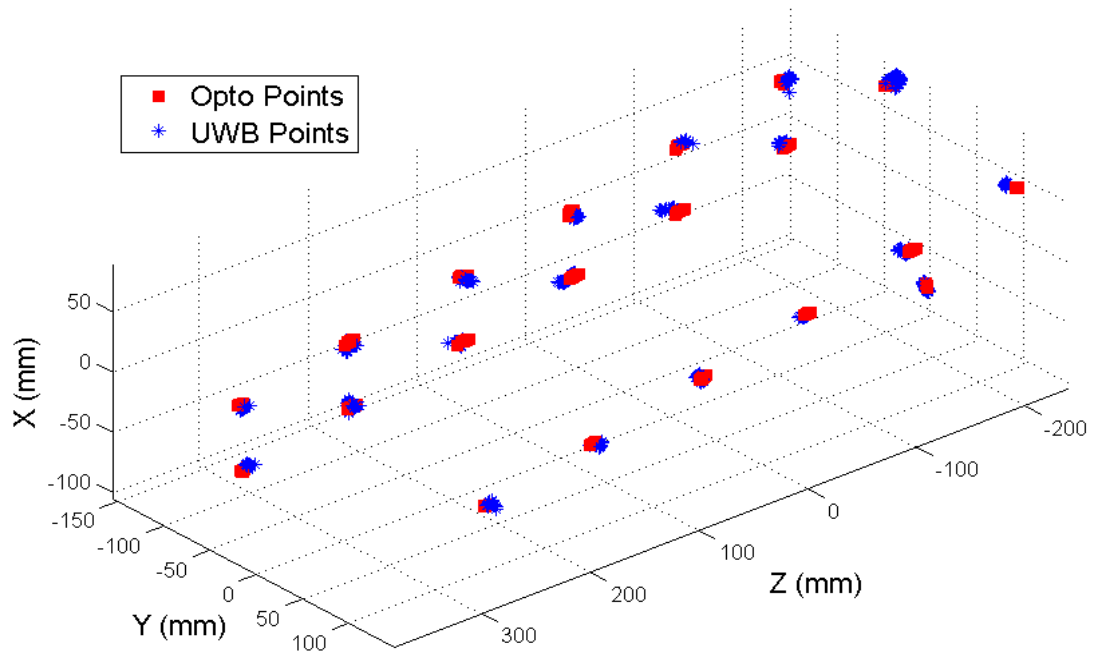


Fig.10.16 3-D robot tracking at static positions, UWB points compared to Opto points.

Figure 10.17 shows the errors at the x , y , and z axes over 2500 measurement points. The error along the x -axis contributed most to the overall distance error, which can be explained by the largest PDOP value in the x direction. As shown in Fig. 10.17, the highlighted region with the maximum error in all directions is when the robot is running at its maximum speed. Such highlighted region corresponds to the circled trace marked in Fig. 10.15 (b). Figure 10.18 shows that the system error increases in a linear fashion with the velocity and all the major errors (i.e. RMSE > 20 mm) happens when the robot running at a velocity of more than 50 mm/sec.

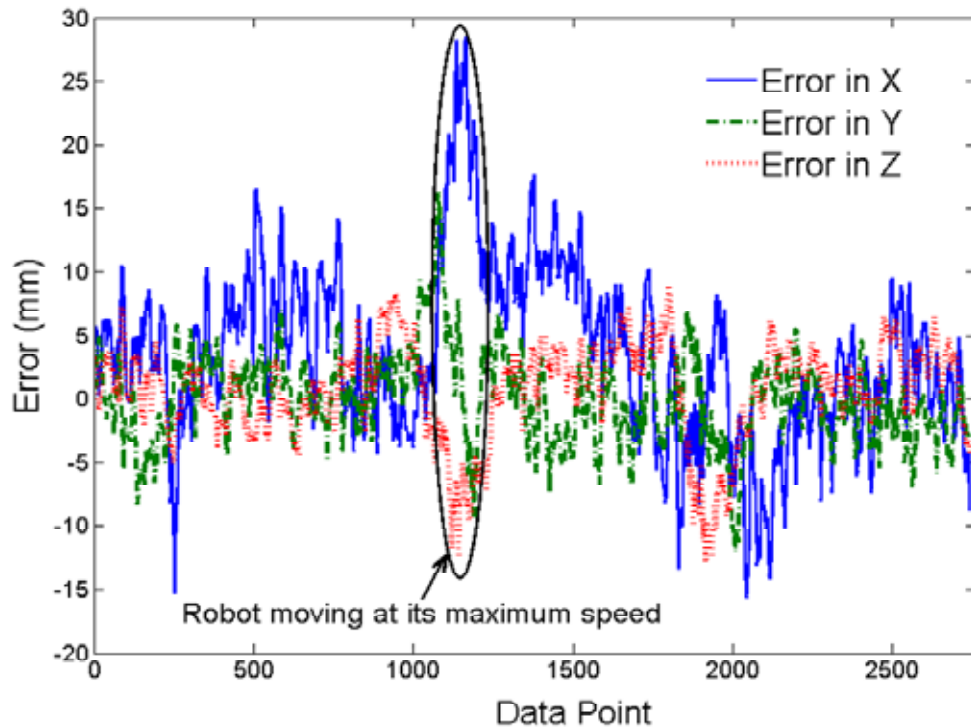


Fig.10.17 3-D robot tracking experiment, x , y and z axes error compared to the Optotrak System.

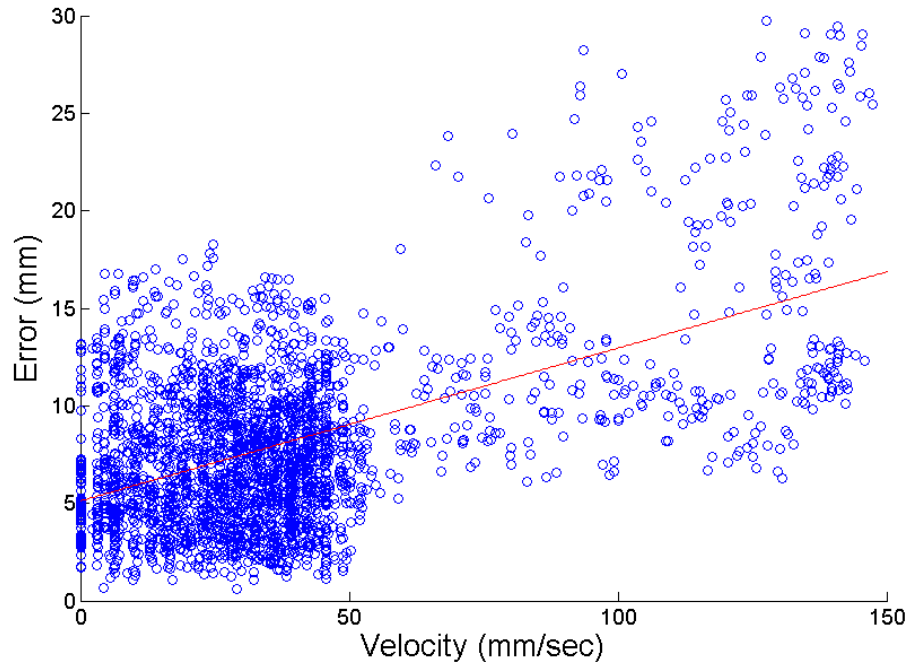
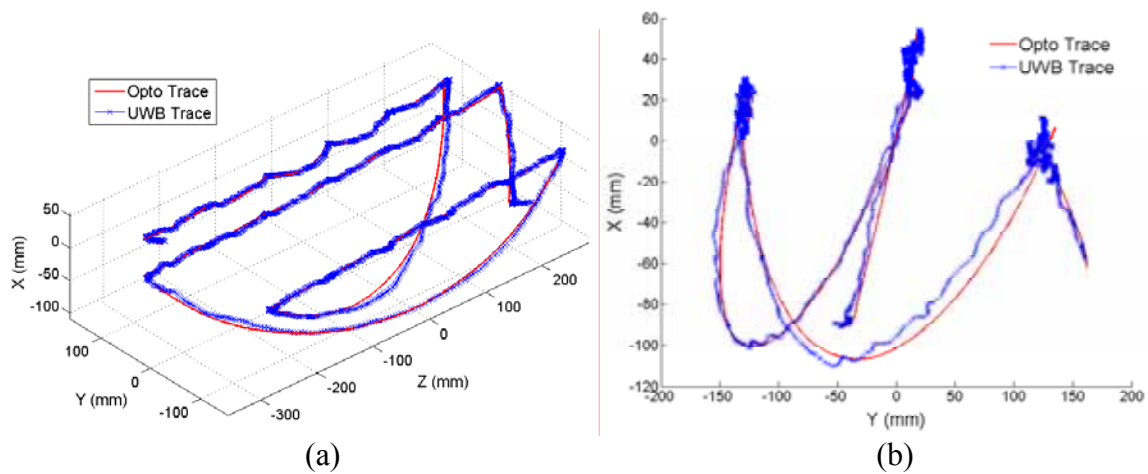


Fig.10.18 3-D robot tracking experiment: linear curve fitting shows that the RMSE increases with the increase in tag velocity.

The same robot experiment has been rerun with a 2 kHz frequency offset. The measured traces by the UWB system are compared to the Optotrak reference system as shown in Fig. 10.19. Figure 10.20 shows the 20 distinct static positions taken by both the UWB system and the Opto system.



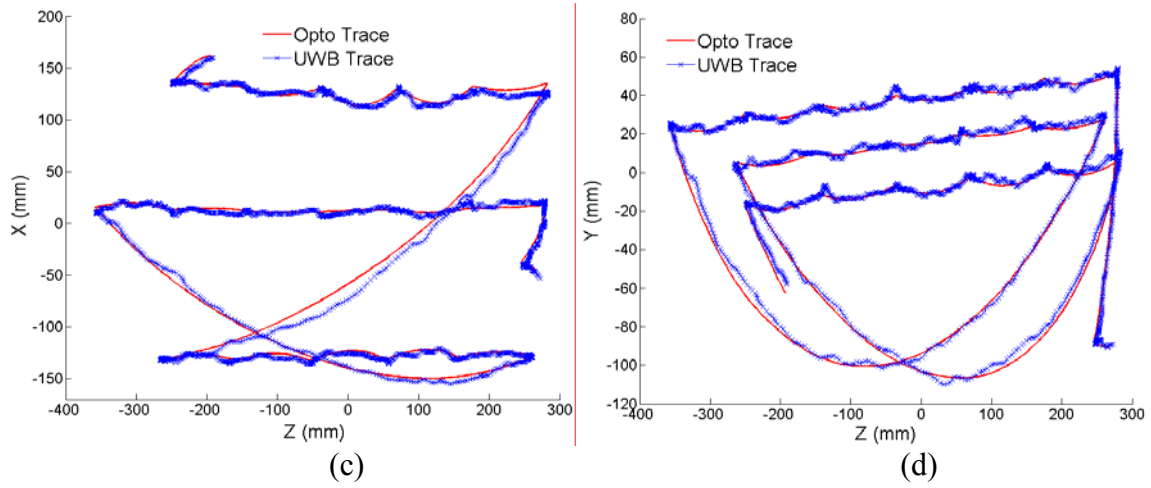


Fig.10.19 3-D dynamic robot tracking with 2 kHz offset, UWB trace compared to Opto trace: (a) 3-D view; (b) XY plane view; (c) XZ plane view; (d) YZ plane view.

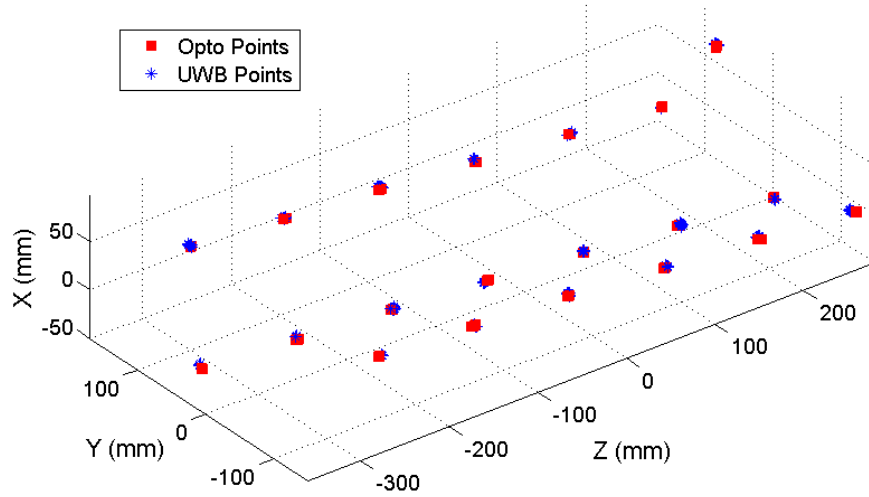


Fig.10.20 3-D robot tracking at static positions, UWB points compared to Opto points.

Table 10.5 lists the RMSE error for the overall trace as well as the 20 distinct static positions for 1 kHz and 2 kHz frequency offset respectively. It is clear that at 2 kHz offset the RMS error is 3-4 mm less than 1 kHz offset case which validates the results discussed in Chapter 9.

Table 10.5 – RMSE in Robot Tracking

Offset	Overall Trace	20 Static Positions
1 kHz	9.256 mm	7.480 mm
2 kHz	5.24 mm	4.67 mm

10.5 Unsynchronized 3D Experiment – Pulse Width Effect

To study how would pulse width, i.e. the signal bandwidth, affects the 3D localization accuracy, the pulse width is varied from 300 to 600 ps in a 3D static measurement. Figure 10.21 plots the 3D static error over a 1000 measurement points and Table 10.6 lists the 3D RMS error for 300 ps, 450 ps and 600 ps respectively. The localization error increases with the increase of the pulse width. Thus, further reducing the pulse width has a potential to further improve the accuracy.

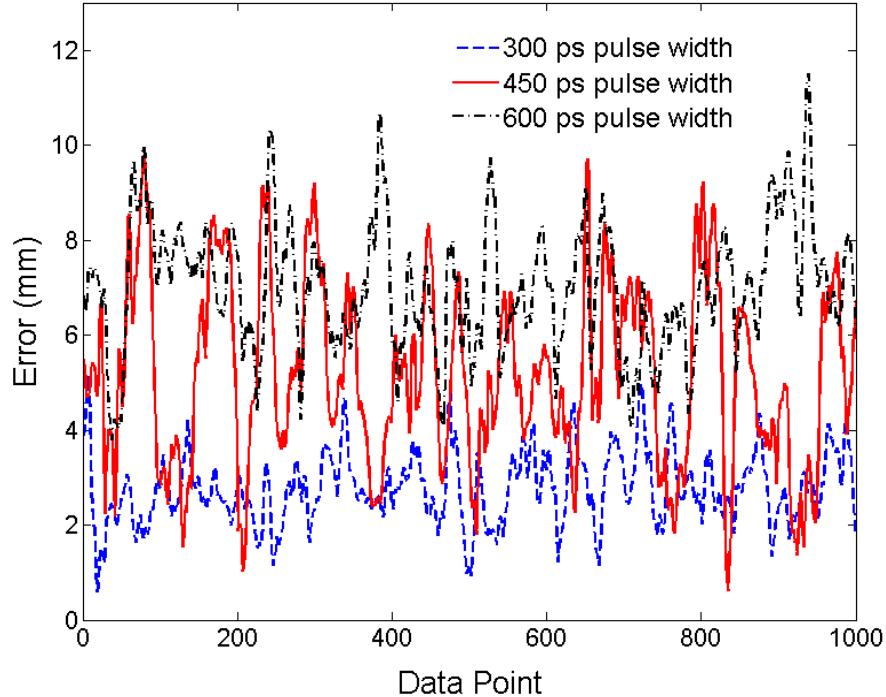


Fig.10.21 Measured 3D static error vs. pulse width

Table 10.6 – Pulse Width vs. 3D Static Error

Pulse Width (ps)	3D RMSE (mm)
300	2.84
450	5.45
600	7.09

10.6 Error Discussion

Below is a list of the error sources in our current noncoherent UWB localization system:

1. As the spatial spread of the base stations along the z -axis is the largest (~ 2.5 m), subsequently the $PDOP_z$ value listed in Table 10.3 is the lowest among the three coordinate axes. This is a common trait to the experimental set up throughout this chapter. Therefore the PDOP values are consistent across these experiments. In these experiments the $PDOP_x$ values are consistently the highest, due to the fact that the maximum range of base station positions along the x -axis is relatively the smallest (~ 1.37 m) when compared to all other setups. This is a major error source in our current localization setup, and such effect causes the error in the x -axis to be nearly as twice as the error in the z -axis. Better arrangement of the base station distribution and addition of redundant base station would greatly reduce this error.
2. The high fidelity clock at both the transmitter and receiver side still have an overall RMS jitter of 2.12 ps, corresponding to 0.64 mm distance error. Also, the best commercially available clock source features a stability of ± 0.5 ppm, which can produce a systematic drift. The clock drift caused by the time scaling effect has been addressed theoretically in Chapter 9 and experimentally in this chapter with a 1% systematic error. Such error can be remedied either by increasing the Tx/Rx PRF clock offset at the price of sampling resolution degradation or by increasing the

strobe frequency, i.e. increasing the real time sampling rate without degrading the sampling resolution.

3. The phase center of the Vivaldi receiving antenna varies with direction, which can have large effects on the system accuracy especially when the tag is in locations on the border or outside of the target volume. We have insured in all our experiments in this chapter that all the tag positions remained within $\pm 20^\circ$ of the broadside direction of each single element Vivaldi in order to minimize phase center effects. The error originating from this effect is estimated to be less than 1 mm, although if the angle from broadside increases beyond $\pm 20^\circ$, the phase center error increases significantly.
4. Multipath interference from extremely close metal (e.g. metal bar supporting transmitting tag) causes pulse overlapping and pulse position shifting. The developed leading edge detection algorithm by a member of our team can handle dense multipath situations but still has substantial uncertainty of around 3 mm under severe multipath conditions [70]. Except for the robot tracking experiment, the other localization experiments in this chapter were done with a strong LOS signal and only minimal amounts of multipath interference.
5. The velocity caused-error has also been observed in dynamic experiments. The system error increases in a linear fashion with the increase of tag speed. Such effect can be explained to a certain extent by the time lag between the UWB system and the Opto reference system at higher speed. The limitations of the time requirement in the UWB system are from the equivalent sampling, averaged filter, FPGA signal processing and TDOA program running in PC. The detailed system time budget has been given in Chapter 6.

Table 10.7 – Error Source Summary

Error Sources	Worst Case Error
Clock Stability	2.5 mm
Sampling Resolution	3 mm
PDOP	3 mm
Pulse Width (BW)	1 mm
Phase Center	1 mm
Multipath	7.65 mm
Velocity	0.08/(mm/s) mm

10.7 Summary and Conclusion

In this chapter, extensive noncoherent 1-D and 3-D localization experiments have been performed where a mm-range accuracy has been constantly achieved, validating the hypothesis of the novel UWB receiver architecture presented in the previous chapter. The Optotrak 3020 is used to obtain accurate reference data throughout the experiments. By comparing with two 1-D experiments, the ranging error has been improved significantly with the reduced timing jitter and “shoulder” effect through applying a low phase noise carrier based IR-UWB architecture together with the advanced sub-sampling and energy detection. Extensive 3-D static and dynamic experiments have been performed, including a tag random movement in the 3D space, a tag movement on a rail, a tag robot tracking etc. The RMSE errors in all these experiments are in the mm-range. The error sources from PDOP, clock drift, tag velocity etc. have been investigated and various recommendations were given to further improve the system performance. The achieved mm-range accuracy in the developed real time noncoherent system represents the world’s best localization accuracy based on UWB technique up to date to the best of the author’s knowledge.

Chapter 11

Conclusions

The established noncoherent mm-range accuracy UWB indoor positioning radar involves very complex system requirements and hardware development. It follows a top-down approach: starting from the system level and coming up with the specifications for the RF front-end and the UWB antennas, respectively. Several contributions have been presented in this dissertation and they will be summarized in this chapter.

First, a low cost reconfigurable pico-second pulse generator has been developed with a simple but novel input-matching network designed to significantly minimize pulse broadening and suppress pulse ringing and echoing. The pulse generator is adjustable from 300ps to 1ns and produces either monocycle or Gaussian pulses. These features will provide more flexibility in the design of adaptable UWB systems needed for such applications like see through wall radar and UWB communications.

Second, a hybrid broadband high speed sub-sampler, with reduced conversion loss and spurious level when compared to previously published sub-samplers has been developed. The sampler is integrated with a step-recovery diode strobe-step generator to sub-sample UWB signals. The fabricated sub-sampler demonstrated a wide 3 dB bandwidth of up to 4 GHz, a reduced spurious level of better than -38 dBc and an equivalent sampling rate higher than 1000 GS/s. The time domain measurement is comparable with Tektronix TDS8200 digital sampling oscilloscope.

Third, various UWB transmitting and receiving antennas have been designed, including the omni-directional monopole antenna with improved radiation pattern and Vivaldi-rod receiving antennas for improved phase center variation. A novel elliptical disc monopole antenna with a modified ground plane has been developed. The antenna shows an excellent omni-directional radiation pattern, as well as a satisfactory input impedance match over an ultra-wide bandwidth. In addition, time domain impulse response experiments have demonstrated that the proposed UWB monopole introduces minimal pulse dispersion. Meanwhile, the effect of the receiving Vivaldi antenna phase center variation has been addressed and quantified which is an important factor for localization accuracy. A technique to improve the gain, narrow the H-plane beam-width, and minimize the phase center variations with frequency by utilizing a Vivaldi antenna with a protruded dielectric rod will be introduced.

Fourth, for local oscillator leakage rejection in our carrier based Ultra-wideband system, a notch filter is proposed to locate before the UWB transmitting antenna. Various filter parameters, such as the filter order and 3 dB rejection bandwidth have been studied to see their effects on providing sufficient band rejection level to reduce the unwanted LO leakage while minimizing the transmitted pulse dispersion. The fabricated 3rd order DGS filter has provided to have adequate rejection level while maintaining a relatively large first pulse amplitude and a small time delay spread, making it the most suitable candidate for filtering the LO leakage in our carrier based UWB system design.

Fifth, to solve the synchronization problem for a noncoherent system, a novel receiver architecture has been demonstrated by combining carrier based UWB system and the traditional energy detection techniques. Both simulation and measurement shows the robust of the proposed receiver architecture, with a reduced timing jitter error, improved received signal SNR, and insensitivity to the triggering threshold voltage.

Last but not least, a stand alone real time noncoherent system has been realized by integrating all the aforementioned hardware and techniques. Extensive 1D and 3D static and dynamic experiments have been performed to test the system robustness and repeatability, including a tag random movement in the 3D space, a tag movement on a rail, a tag robot tracking etc. Where tracking both the static and dynamic targets, an RMS error of approximately 2 mm and 6 mm has been achieved respectively. Our results have exceeded the state of the art of any commercially available UWB positioning systems and are considered a great mile stone in developing such technology.

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Appendix A

TDOA Algorithm

There are four receivers at known positions Rx1 or (x_1, y_1, z_1) , Rx2 or (x_2, y_2, z_2) , Rx3 or (x_3, y_3, z_3) , and Rx4 or (x_4, y_4, z_4) , and a tag at unknown position (x_u, y_u, z_u) . The measured distance between the four known receivers to the unknown tag can be represented as ρ_1 , ρ_2 , ρ_3 , and ρ_4 , which is given by

$$\rho_i = \sqrt{(x_i - x_u)^2 + (y_i - y_u)^2 + (z_i - z_u)^2} + c\tau_u \quad (\text{A.1})$$

where $i = 1, 2, 3$, and 4 , c is speed of light, and τ_u is the unknown time delay in hardware.

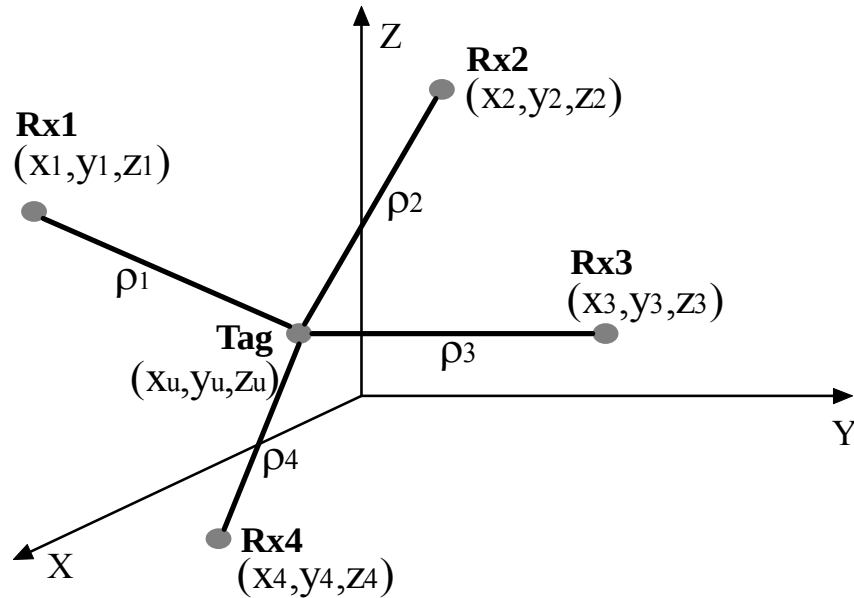


Fig. A.1. Calculation of the tag position based on TDOA approach with 4 base stations.

The differential distances between the four receivers and the tag can be written as

$$\begin{aligned}
\Delta\rho_{1k} &= \rho_1 - \rho_k \\
&= \sqrt{(x_1 - x_u)^2 + (y_1 - y_u)^2 + (z_1 - z_u)^2} \\
&\quad + \sqrt{(x_k - x_u)^2 + (y_k - y_u)^2 + (z_k - z_u)^2}
\end{aligned} \tag{A.2}$$

where $k = 2, 3$, and 4 , and the time delay τ_u in hardware has been cancelled.

Differentiating this equation (A.2) will give

$$\begin{aligned}
d\Delta\rho_{1k} &= \frac{(x_1 - x_u)dx_u + (y_1 - y_u)dy_u + (z_1 - z_u)dz_u}{\sqrt{(x_1 - x_u)^2 + (y_1 - y_u)^2 + (z_1 - z_u)^2}} \\
&\quad + \frac{(x_k - x_u)dx_u + (y_k - y_u)dy_u + (z_k - z_u)dz_u}{\sqrt{(x_k - x_u)^2 + (y_k - y_u)^2 + (z_k - z_u)^2}} \\
&= \left(\frac{x_1 - x_u}{\rho_1 - c\tau_u} + \frac{x_k - x_u}{\rho_k - c\tau_u} \right) dx_u \\
&\quad + \left(\frac{y_1 - y_u}{\rho_1 - c\tau_u} + \frac{y_k - y_u}{\rho_k - c\tau_u} \right) dy_u \\
&\quad + \left(\frac{z_1 - z_u}{\rho_1 - c\tau_u} + \frac{z_k - z_u}{\rho_k - c\tau_u} \right) dz_u
\end{aligned} \tag{A.3}$$

In equation (A.3), x_u , y_u , and z_u are treated as known values by assuming some initial values for the tag position. dx_u , dy_u , and dz_u are considered as the only unknowns. From the initial tag position the first set of dx_u , dy_u , and dz_u can be calculated. These values are used to modify the tag position at x_u , y_u , and z_u . The updated tag position at x_u , y_u , and z_u can be considered again as known quantities. The iterative process continues until the absolute values of dx_u , dy_u , and dz_u are below a certain predetermined threshold given by

$$\varepsilon = \sqrt{dx_u^2 + dy_u^2 + dz_u^2} \quad (\text{A.4})$$

The final values of x_u , y_u , and z_u are the desired tag position. The matrix form expression of equation (A.3) is

$$\begin{bmatrix} d\Delta\rho_{12} \\ d\Delta\rho_{13} \\ d\Delta\rho_{14} \end{bmatrix} = \begin{bmatrix} \alpha_{11} & \alpha_{12} & \alpha_{13} \\ \alpha_{21} & \alpha_{22} & \alpha_{23} \\ \alpha_{31} & \alpha_{32} & \alpha_{33} \end{bmatrix} \begin{bmatrix} dx_u \\ dy_u \\ dz_u \end{bmatrix} \quad (\text{A.5})$$

where

$$\alpha_{k-1,1} = \frac{x_1 - x_u}{\rho_1 - c\tau_u} + \frac{x_k - x_u}{\rho_k - c\tau_u}$$

$$\alpha_{k-1,2} = \frac{y_1 - y_u}{\rho_1 - c\tau_u} + \frac{y_k - y_u}{\rho_k - c\tau_u}$$

$$\alpha_{k-1,3} = \frac{z_1 - z_u}{\rho_1 - c\tau_u} + \frac{z_k - z_u}{\rho_k - c\tau_u}$$

The solution of equation (A.5) is given by

$$\begin{bmatrix} dx_u \\ dy_u \\ dz_u \end{bmatrix} = \begin{bmatrix} \alpha_{11} & \alpha_{12} & \alpha_{13} \\ \alpha_{21} & \alpha_{22} & \alpha_{23} \\ \alpha_{31} & \alpha_{32} & \alpha_{33} \end{bmatrix}^{-1} \begin{bmatrix} d\Delta\rho_{12} \\ d\Delta\rho_{13} \\ d\Delta\rho_{14} \end{bmatrix} \quad (\text{A.6})$$

where $[]^{-1}$ represents the inverse of the α matrix. If there are more than four receivers, a least-square approach can be applied to find the tag position with a better accuracy.

Appendix B

Vivaldi Antenna with a Protruded Dielectric Rod

To obtain symmetric patterns, it is required to significantly narrow down the H-plane beamwidth by using an H-plane array rather than a single element. To accomplish this task, we are proposing a Vivaldi antenna with a protruded dielectric rod to improve the gain, narrow down the H-plane beamwidth, and minimize the phase center variations with frequency. A sample antenna was fabricated and measured, and its preliminary measured results are very promising, and are in good agreement with our simulated results, design details will be given in the following sections.

B.1 The Proposed Structure

The proposed structure is shown in Fig. B.1. It consists of an antipodal Vivaldi antenna inserted into a polystyrene rod. The rod consists of three parts: (1) a launcher section with length l_{inset} , (2) a propagation section with length l_{rod} and (3) a radiating section with length l_{tip} . The Vivaldi antenna is made on a 1.5mm thick FR-4 substrate with a relative permittivity of $\epsilon_r = 4.4$. Meanwhile, the Vivaldi antenna is fed by a microstrip line whose width is chosen to be a 3mm to obtain a 50Ω characteristic impedance. The dimensions of the Vivaldi antenna are shown in Fig. B.1.

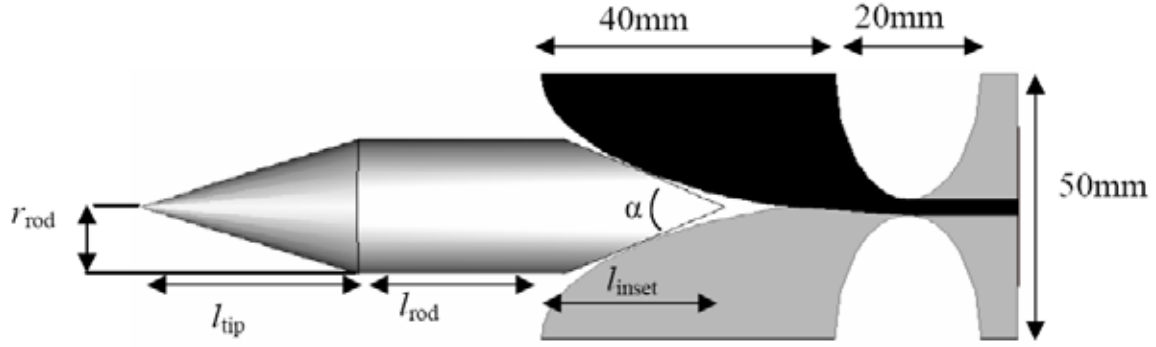


Fig. B.1. The Vivaldi-Rod Structure

The E-plane of the Vivaldi antenna is the plane parallel to the substrate, and the H-plane is the plane perpendicular to the substrate. The antenna structure is designed to operate over a 4 GHz to 8 GHz, and its feeding structure and the radius of the polystyrene rod are selected to make the HE_{11} mode of the cylindrical rod to be the dominant mode up to about 11.5GHz. The selected dimensions for the rod are $l_{inset} = 25\text{mm}$, $l_{rod} = 20\text{mm}$, $l_{tip} = 30\text{mm}$ and $r_{rod} = 12.5\text{mm}$.

B.2 The Gain and the H-Plane Beamwidth Improvement

The dielectric rod can be viewed as a traveling wave antenna with radiating polarization currents. The rod is fed by the Vivaldi antenna for a wide band operation. Since the aperture size of the rod is larger than that of the Vivaldi antenna in the H-plane (perpendicular to the substrate) and it is axially symmetric, we are expecting narrower H-plane beamwidth of the overall radiation pattern and an improvement on the symmetry of the pattern. This has been validated by the measured and simulated radiation patterns shown in Fig. B.2. A gain increase of about 3dB over most of the band as compared to the normal Vivaldi antenna is shown in Fig. B.3(a). Meanwhile, the input-match

performance was not considerably affected when adding the rod as shown in Figs. B.3 (b) and (c).

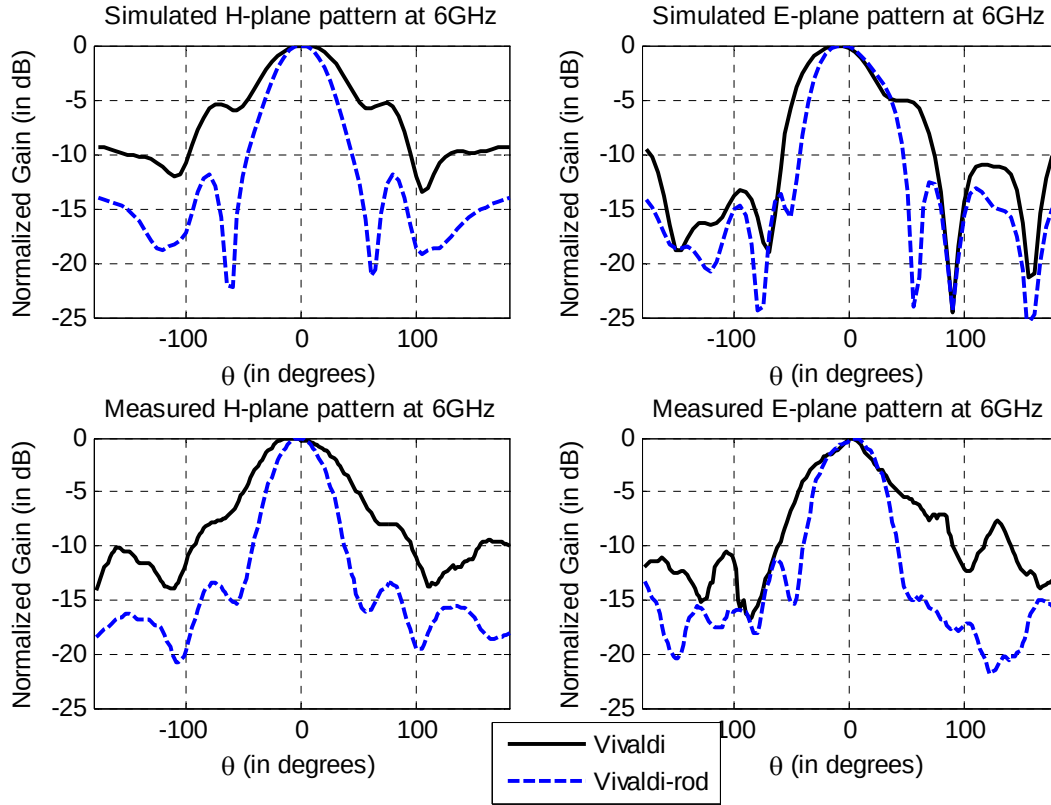


Fig. B.2. The simulated and measured radiation patterns in the E- and H-planes of the two antennas (θ is the angle from the bore-sight direction)

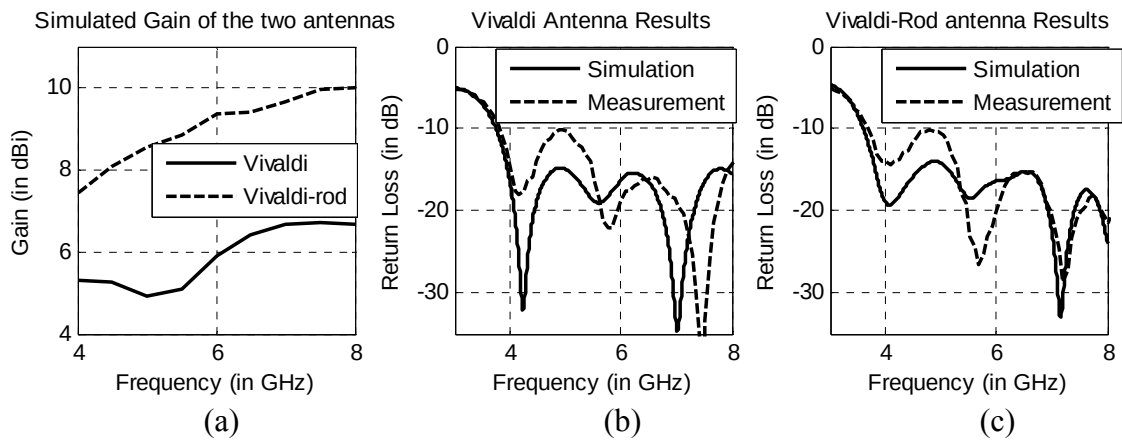


Fig. B.3. (a) the gain vs. frequency, (b) and (c) the simulated and measured return loss for both antennas

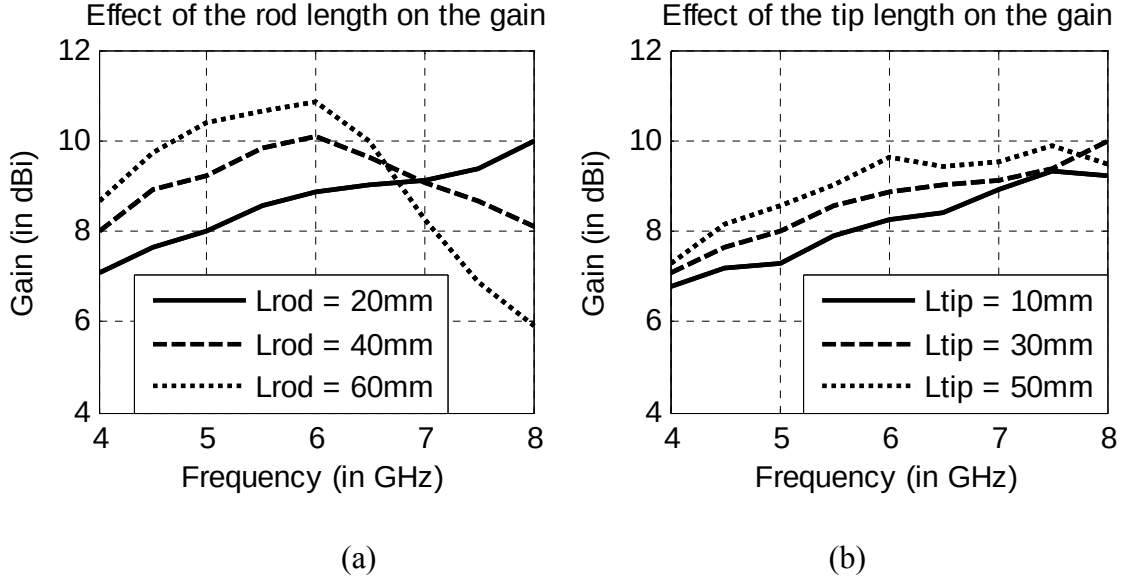


Fig. B.4 (a) The effect of the rod length on the gain and (b) the effect of the tip length on the gain

However, the pattern bandwidth of the Vivaldi-Rod is narrower than that of the Vivaldi alone, but we can optimize both the pattern bandwidth and its gain by adjusting the length of the rod and/or the tip-length (as shown in Fig. B.4). It is clear that the gain reaches a maximum, and then significantly drops due to the destructive interference between the feed and the tip radiation at higher frequencies. At these frequencies, the main beam even splits and the antenna becomes no longer end-fire.

B.3 Phase Center Compensation

Our UWB localization system, in its simplest form, has four receivers and one transmitter. By measuring the time difference of arrival for the pulse to reach the antennas, and by knowing the phase center position of the transmitting and receiving antennas and their spatial positions, we can locate the transmitter. However, the phase center variations have a major impact on the system performance. One problem with the

conventional Vivaldi antennas is that they have some “skew” between their phase centers when calculated in the H- and E-Planes individually as shown in Fig. B.5(a) which causes inevitable localization errors if we defined a unique phase center for the Vivaldi antenna. But, we found that by adding the rod to the antenna, this skew is significantly reduced over most of the band as shown in Fig. B.5(b).

Figure B.6 shows the simulated phase center displacement of the Vivaldi/Vivaldi-rod for both the E and H cuts using CST software. As shown in Fig. B.6(b), by using the Vivaldi-rod antenna, the phase center variation versus rotating angle indicates a small and stable phase center variation within the $\pm 40^\circ$ while the variation degrades dramatically with an angle greater than 30° for the Vivaldi antenna.

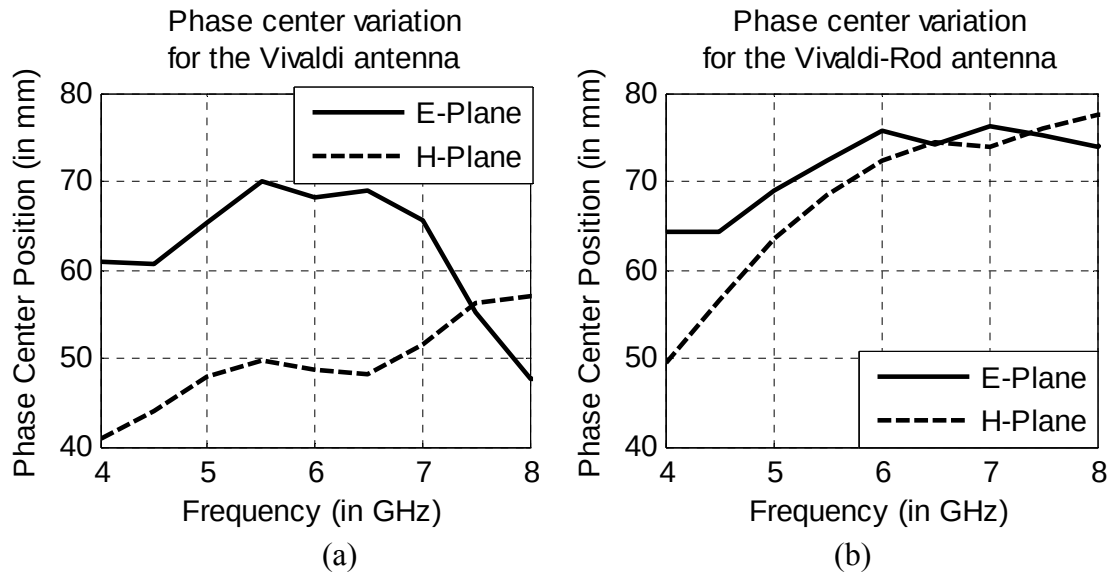
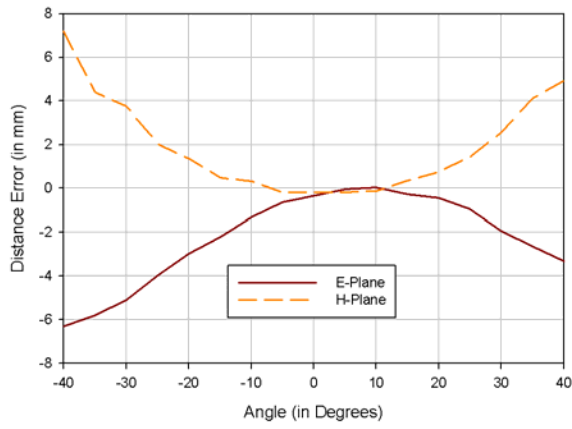
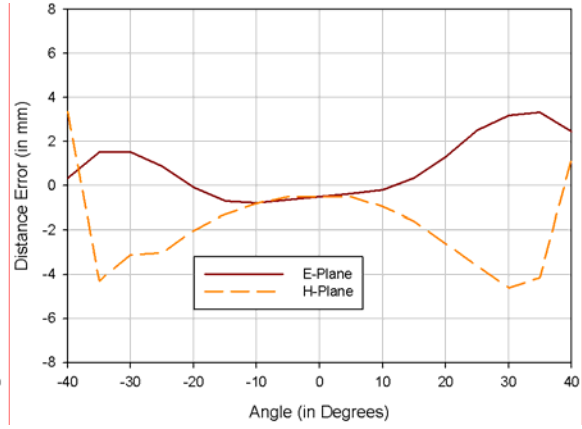


Fig. B.5. Improvement of the Vivaldi phase center variation using the dielectric rod.



(a) Vivaldi Antenna



(b) Vivaldi-rod Antenna

Fig. B.6. Simulated phase center error versus angle for the E- and H-cuts

VITA



Cemin Zhang (S'06–GS'07) was born in Chengdu, China, in 1978. He received the B.S. and M.S. degrees in electronic engineering from Zhejiang University, Hangzhou, China, in 2001 and 2004, respectively, and received the Ph.D. degree in electrical engineering at the University of Tennessee, Knoxville, TN, in 2008. In 2003, he worked as a RF engineer with UTStarcom Co. Ltd, Hangzhou, China, where he was involved with the development of antenna switch and mobile base station hardware. In early 2004, he worked as a product engineer at Intel Co., Shanghai, China, where he was involved with developing novel flash memories. Since 2008, he joined Hittite Microwave Corporation as an MMIC design engineer. He has developed various microwave components for a high precision UWB localization system including a tunable pico-second pulse generator, high speed sampler, UWB antennas etc., and established a novel unsynchronized UWB system architecture to achieve the mm-range 3D localization accuracy. His current research interests are to integrate the microwave system into a MMIC chip. He has authored/co-authored more than 25 journal/conference papers and presented at numerous international conferences. Mr. Zhang is a member of Sigma Xi, Phi Kappa Phi and National Scholars Honor Society. He was the recipient of the 2007 URSI Student Fellowship and 2008 UT Chancellor's Citation for Extraordinary Professional Promise. He has served as a reviewer for IET Signal Processing and many international conferences including the IEEE International Conference on Ultra-Wideband.