High resolution indoor localization system using ultra wideband impulse radio

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**HIGH RESOLUTION INDOOR LOCALIZATION SYSTEM USING
ULTRA WIDEBAND IMPULSE RADIO**

ANKUSH VASHISTHA

SCHOOL OF ELECTRICAL & ELECTRONIC ENGINEERING

2019

HIGH RESOLUTION INDOOR LOCALIZATION SYSTEM USING ULTRA WIDEBAND IMPULSE RADIO

ANKUSH VASHISTHA

School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Doctor of Philosophy

2019

Statement of Originality

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Ankush Vashistha

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[Date Here]

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Date

[Supervisor's Signature Here]

A/Prof Choi Look Law

[Supervisor's Name Here]

Authorship Attribution Statement

This thesis contains material from the following peer-reviewed paper(s) where I was the first and/or corresponding author.

Chapter 3 is published as A. Vashistha and C. L. Law, “A novel E-DTDOA based one-way ranging using UWB-IR with unsynchronized anchors,” in IEEE Transactions on Industrial Informatics, 2019.

The contributions of the co-authors are as follows:

- A/Prof Choi Look Law provided the initial research direction.
- I did all the mathematical modeling along with the analysis.
- I performed all the simulations and experiments on the hardware.
- I prepared all the manuscript drafts. The manuscript was reviewed by A/Prof Choi Look Law.

Chapter 4 is published as A. Vashistha and C. L. Law, “E-DTDOA based Localization for Wireless Sensor Networks with Clock Drift Compensation,” in IEEE Sensors Journal, 2019.

The contributions of the co-authors are as follows:

- I did all the mathematical modeling along with the analysis.
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- A/Prof Choi Look Law provided the initial project direction.
- I prepared the manuscript drafts which was reviewed by A/Prof Choi Look Law.
- I performed the experiments along with Dr. Rejina and A. Gupta.

22/08/2019

Date

Ankush Vashistha

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Abstract

Internet of things (IoT) is an emerging technology which enables integration of billions of objects and complex systems over internet with the help of many technologies such as Wi-Fi, ZigBee, Bluetooth, UWB etc. Position estimation has become an essential and crucial factor in various monitoring applications of IoT. Precision localization and positioning has become an attractive area of interest for many new applications and business solutions. Although, the global navigation satellite systems (GNSS) can provide good performance and positioning in outdoor systems, they are not very accurate when it comes to indoor locations or in GNSS-denied environments. With the ease of availability of commercial transceivers, and the demand for accurate positioning systems by various industries, the research interest towards indoor positioning and navigation systems based on ultra-wideband (UWB) technology has been immense. A good localization accuracy can be achieved by using UWB pulses due to their high temporal resolution and multi-path immunity. The UWB systems can be operated in an unlicensed band in the frequency range of 3.1 - 10.6 GHz. The operation in the unlicensed band makes it even more exciting and accessible for different commercial applications. However, several GHz of Nyquist sampling rate is required to sample such a large bandwidth signal. A sampling rate of few tens of GHz is suggested for resolving large number of multipaths in UWB based systems. Sampling at such a rate is an expensive solution and thus practically limited by cost and complexity of the required hardware and thus, it is a bottleneck in designing low cost sensor nodes employing UWB for accurate positioning systems.

In this thesis, a novel analytical equation is proposed to determine the equivalent time of arrival (E-TOA) for achieving sub-ns resolution, with much reduced analog to digital converter (ADC) sampling frequency (in the order of 2-3 MHz). The timing information is extracted from high resolution channel impulse response (CIR) which is obtained using equivalent time sampling (ETS) technique. The proposed E-TOA equation is different from conventional real-time sampling equation due to

the presence of additional transmitter clock drift, and thus sensitive to both the transmitter and the receiver clock drift variations. The validation of the proposed E-TOA equation is carried out numerically using simulations along with experimental validation. The effect of timing uncertainties relating to the transmitter clock start time and the receiver clock offset is analyzed with variations in the transmitter and receiver clock drift. It is thus demonstrated, using in house designed sensor nodes, that high ranging accuracy, in the order of few cm, can be achieved by utilizing the proposed analytical E-TOA technique, even with low sampling rate.

Reconstructed signal obtained by ETS technique, that require periodic transmission of the same signal repetitively, is severely affected by the variation in transmitter and receiver clock drift as against the real time sampling where the variations are due to receiver node clock drift only. Thus, it requires a protocol to precisely estimate the transmitter and receiver clock parameters at the frame level to accommodate different (multiple) target node at each frame. A scheme, which uses a novel mathematical equivalent time of arrival (E-TOA) model for ETS based system, for clock drift estimation is presented. Based on clock drift estimation parameters, receiver nodes are tuned to the same frequency, while the estimated transmitter frequency is used to estimate the precise equivalent time resolution which is subsequently used to correct the E-TOA measurements. E-TOA measurements are further used to propose an equivalent differential time difference of arrival (E-DTDOA) based ranging algorithm, which relaxed the time synchronization requirement between the wireless nodes, and still achieving high time resolution. The E-DTDOA range measurements are subsequently used to obtain precise localization of the target node/s. The feasibility of the algorithm proposed is demonstrated experimentally using in house designed wireless sensor nodes.

Utilizing the E-DTDOA mechanism, this thesis further demonstrates experimentally a hybrid sub-1 GHz and UWB-IR based indoor positioning system. The system integrates sub-1 GHz technology for medium access control (MAC) and communication along with UWB-IR technology for the purpose of ranging. The presented positioning system employs energy efficient tags using UWB-IR in transmitter mode and the E-DTDOA as a measurement technique. A static experiment was performed for the system with 10 tags and 4 anchor nodes in an indoor environment. During the experimentation, the accuracy in tags positioning was found to be less than 30 cm in an area of 16m x 5.3m.

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List of Abbreviations

ADC	Analog to digital converter
AN	Anchor nodes
AOA	Angle of Arrival
A-TDOA	Asynchronous time difference of arrival
AWGN	Additive white Gaussian noise
BER	Bit error rate
BLE	Bluetooth Low Energy
CDF	Cumulative distribution function
CIR	Channel impulse response
CRLB	Cramer-Rao lower bound
DDS	Digital direct synthesizer
DSP	Digital signal processors
DTDOA	Differential time difference of arrival
EDGE	Enhanced Data GSM Environment
E-DTDOA	Equivalent differential time difference of arrival
EIRP	Equivalent isotropically-radiated power
ESPRIT	Estimation of signal parameters via rotational invariance technique
E-TOA	Equivalent Time of arrival
ETS	Equivalent time sampling
FCC	Federal Communications commission
FPGA	Field programmable gate array
FTSP	Flooding Time Synchronization Protocol
GCC	Generalized cross-correlation
GLONASS	GLOBAL NAVigation Satellite System

GPS	Global positioning system
GSM	Global system for mobile
HRTS	Hierarchy Referencing Time Synchronization
IMU	Inertial measurement unit
ISI	Inter symbol interference
ISM	Industrial, Scientific and Medical (radio spectrum)
IR	Infrared technology
LB	Locator box
LBS	Location based Services
LoRa	Long Range
LoRaWAN	Long Range Wide Area Network
LOS	Line-of-site
LPWAN	Low Power Wide area network
LTE	Long-Term Evolution
LTS	Lightweight Tree based Synchronization
MAC	Medium Access
MCU	Micro-controller unit
MN	Mobile node
MPC	Multi-path components
MUSIC	Multiple signal classification
MVDR	Minimum Variance Distortion less Response
NAVIC	NAVigation with Indian Constellation
NB-IoT	Narrow Band Internet of Things
NLOS	Non-Line of Sight
NTP	Network Time Protocol
OFDM	Orthogonal Frequency Division Multiplexing
PAM	Pulse amplitude modulation
PBS	Pairwise Broadcast Synchronization
PLL	Phased locked loop
POA	Phase of Arrival

PPM	Pulse position modulation
PTP	Precision Time Protocol
PRF	Pulse repetition frequency
PRI	Pulse repetitive interval
QZSS	Quasi-Zenith Satellite System
RBS	Reference Broadcast Synchronization
RF	Radio Frequency
RFID	Radio Frequency Identification
RN	Reference Node
RSS	Received Signal Strength
RSSI	Received signal strength information
RTLS	Real-time location system
RTS	Real time sampling
RTT	Round trip time
RX	Receiver
SAN	Slave Anchor Node
SNR	Signal to noise ratio
TCXO	Temperature compensated oscillator
TDC	Time to digital converter
TDMA	Time division multiple access
TDOA	Time difference of arrival
TN	Target Node
TOA	Time of Arrival
TOF	Time of Flight
TPSN	Time synchronization Protocols for Sensor Networks
TWR	Two way ranging
TX	Transmitter
UHF	Ultra High Frequency
UWB	Ultra-wide Band
UWB-IR	Ultra-wideband Impulse Radio

VLC	Visible Light Communication
WLAN	Wireless Local Area Network
WSN	Wireless Sensor Network

1 Introduction

Localization and tracking are a key feature of emergent applications in our day to day life. Localization and tracking, in general, can be categorized as, 1) outdoor localization and tracking and, 2) Indoor localization and tracking.

Outdoor localization and tracking

We are well familiar with the outdoor positioning system, commonly known as global positioning system (GPS), and which is widely used in our day to day life. Broadly speaking, almost all the outdoor positioning system are based on satellites. Historically, positioning system based on satellites originated in early 1960's. The first satellite-based localization system developed was Global positioning system (GPS), which was originally developed for military operations. Later, it was opened for civilian operations. The details of the different satellite based localization system owned by different countries along with their frequency range of operation is summarized in Table. 1.1.

In general, the positioning and tracking done by the satellite based system, as shown in Fig.1.1, can be understood as follows: First, the receiver (target) finds the time of arrival (TOA) information of the signal received from the satellite. From the TOA information, it obtains the distance between the satellite and the GPS receiver. This distance information is used to find the position of the GPS receiver by using either three or four satellites using trilateration method. This position information is converted relative to the earth co-ordinate system. With the advancement of technology and the increasing demand of these localization systems in the civil community, there has been a rapid growth in localization market which further give rise to the new application areas. During the last two decades, there has been an

¹Global Positioning System

²GLOBAL NAVigation Satellite System

³NAVigation with Indian Constellation

⁴Quasi-Zenith Satellite System

Table 1.1: Different satellite based localization system

Year of Establishment	Country	Name of the Satellite system	Number of Satellites	Accuracy	Frequency of Operation
1960	USA	GPS ¹	30	5 m	1.563–1.587 GHz (L1) 1.215–1.239 GHz (L2) 1.164–1.189 GHz (L5)
1980	Russia	GLONASS ²	24	5m	1.593–1.61 GHz (G1) 1.237–1.2541GHz (G2) 1.189–1.214 GHz (G3)
2005	European Union	Galileo	22	1m	1.559–1.592 GHz (E1) 1.16–1.21 GHz (E5a/b) 1.260–1.300 GHz (E6)
2012	China	BeiDou	23	10m	1.5610 GHz(B1) 1.589GHz (B1-2) 1.20714 GHz (B2) 1.26852 GHz (B3)
2013	India	NAVIC ³	5	10m	1176.45 MHz (L5) 2492.0GHz (S)
2012	Japan	QZSS ⁴	4	1m	-

increasing effort to integrate these position and localization techniques within the system such as in mobile phones etc. This led to the emergence of location-based services (LBS). The main idea behind these LBS is that they gather the real time information from the participating devices, convert this information into the location or positioning and use this information for various applications. For instance, we can see the navigation of food delivery person in our phones or apps provided by the food delivery services. Another similar example could be when we book a taxi using online platforms, we can see the location of the taxi driver in our apps. Another possible example can be in power sector where the positioning solution can be very useful to track the workers working on a long power grid pylon, which is otherwise very difficult to track. It can be seen that there is high inter-dependency between these location-based services and different branches of engineering. For example, robotics can have some useful applications when combined with indoor and outdoor localization. Moreover, it can be useful in areas of wireless sensor networks, wireless

Figure 1.1: Satellite Positioning

communication, Mobile computing, Data mining etc.

Furthermore, we can say that when we are beneath a clear sky, GPS can provide a millimeter level accuracy. However, the disadvantage of the GPS is that it does not work well in GPS denied environments and in indoor environments. In the indoor environments, the GPS positioning error can be as high as 100 m, even with advanced GPS receivers having high sensitivity. This is because of the environment's characteristics such as walls, windows, furniture, ceiling etc. Due to environment characteristics, the transmitted signal gets reflected and reaches the receiver with multi-paths or gets blocked. The received signal, thus, gets affected in two ways: 1) Multi-path propagation, due to reflections from the environment. 2) Non-Line of Sight (NLOS) Propagation, due to the blockage of the direct path. Therefore, the complex environment in the indoors makes it difficult to decode the low power GPS signal with dense indoor multi-path and NLOS effects. With the inefficiency of the GPS in the indoors, it is important to have a separate indoor positioning and tracking systems and techniques through which high accuracy of positioning and tracking can be done. We discuss some of the techniques used for indoor localization and tracking systems.

Indoor localization and tracking

Indoor positioning services may be desirable at places such as large departmental stores, hospitals, railway stations, universities with possible applications such as

localizing blind persons in a mall, in warehouses for tracking goods, rescue operations etc. Construction sites can be another application which involves real time tracking of materials, equipment's, manpower as well as for safety managements. In medical domain, it is typically used for vital sign monitoring and motion tracking [1]. Another application of indoor positioning and navigation system that have attracted research community is in the wireless sensor networks which requires accurate localization of the sensor nodes. Moreover, it is also attracting big scientific and industrial community because of its big market potential [2]. In addition, availability of new integrated transceivers has attracted different projects to work in this field.

Generally, an indoor positioning system consists of few anchor nodes with known position, which are deployed in the tracking area, and few target nodes with unknown position. The main idea is to find the ranges between the anchor nodes and the target nodes. This range information is further used to calculate the position of the target node/s. Getting the range information is thus an important step for any positioning system. There are different ways to obtain the range information between the nodes. The classification of common range measurement techniques is given in Fig.1.2. The main parameters to consider while selecting the above techniques are:

Figure 1.2: Classification of range measurement techniques

1. **Accuracy:** Accuracy can be defined as the difference between the actual position and the estimated position. It is a statistical distribution which can be expressed in terms of the cumulative distribution function (CDF) or in terms of percentage. Accuracy is a very important parameter to decide which range measurement technique to use in the system. For instance, some of the applications like warehouse management requires accuracy of few meters to keep certain bulky goods and items,

however for a robot tracking in indoor areas require accuracy in the range of few centimeters.

2. Coverage: Next criteria to decide about the range measurement technique is the coverage area. It is defined as the range of an area in which a system can locate a person or an object. Albeit some technologies can provide wide range of coverage area outdoors, due to the complex environmental factors their coverage is restricted in indoor environments.

3. Cost: Cost refers to the amount spend over the whole positioning system, its hardware units, installation and maintenance. Cost of a system vary depending upon the requirement of the application. For example, systems using existing infrastructures such as Wi-Fi and mobile signal need less infrastructure and maintenance cost. Whereas, the systems which require dedicated hardware can have high installation and maintenance cost. However, there is always a trade-off between the cost and performance. This implies that to achieve high performance or accuracy, a system needs to have a dedicated hardware and such accuracy cannot be achieved with the existing infrastructures.

Technologies used for Indoor Positioning System

In this section, we discuss some of the technologies which are commonly used for indoor positioning and navigation system.

1. Optical Technologies: They use optical signals in the visible light spectrum range for obtaining the range information between the nodes. Following are the main technologies which come under optical based techniques:

a) Infrared technology (IR): They operate in the range with wavelength larger than visible light spectrum. Such system consists of infrared light emitting diode and a photo-diode receiver which can capture and detect the light pulses in order to receive the signal. The infrared signal is processed to obtain the range information. The accuracy of such technology can range from 1m - 2m in LOS scenarios. Moreover, a precision of 10 cm is claimed in [3]. The directivity of the optical signal is a major factor which make these systems unreliable. The coverage area can be a room with high installation cost. However, sunlight can be a cause of interference in such systems [3], [4], [5].

b) Visible Light Communication (VLC): This technique uses visible light for data transmission. LED lights are found to be the most appropriate. The basis of such

technique is the utilization of the ON-OFF property of light. This flickering property can be used with variety of modulation techniques. For instance, each of the LED lamps is provided with different encoding. The receiver or the sensor carried by the user receives the light and compares the encoding with the known encoding scheme. This way it determines the dominant LED and thus it can associate itself with the corresponding lamp in the nearby area. Authors in [6] propose an accuracy of 20 cm with the passive architecture. Moreover, these systems can have low installation costs as it can be used with the existing infrastructure. Some of the commercial products utilizing this technique are given in [7]. The commercial products claim to provide sub-meter accuracy with the provision of larger coverage area.

2. Sound-Based Technologies: Sound based location system are based on sound signals. These systems usually rely on the pressure wave which propagates in an environment. The benefit of using this technique is the slower propagation speed of sound waves than the radio signals. The slow speed provides the receiver more time to receive and detect the signal. These systems use the frequency in the range of 20 KHz. The sound signal with these frequency help determine the target node position by measuring the time of flight of the signal between the transmitter and receiver. Some of the related works are given in [8], [9], [10]. In addition to the ultrasonic signal, audible signal can also be used to find the range between the participating nodes [11]. Authors in [8] report a precision of 1 cm. Moreover, solutions using ultrasonic signal can have high implementation cost as compared to the audible sound signals.

3. Inertial based technologies: They are based on obtaining the parameters such as displacement, acceleration and rotational speed. The three are measured by an inertial measurement unit (IMU) which has odometer to measure physical displacement, accelerometer to measure acceleration and gyroscope to measure rotational speed. This information is used to obtain the relative position information. Since this information can only provide the relative information it is useful to combine them with other techniques to get the absolute position information. Authors in [12] combine this with the global positioning system to determine target position in GPS denied environment.

4. Radio Frequency Technologies: These techniques are commonly used for indoor localization and tracking. These techniques provide many features due to which they are most preferred for indoors. Some of them are known for their better penetrating power, large coverage area, good accuracy. Their ability to penetrate

through walls makes them a suitable candidate in non-line of sight (NLOS) scenarios. A detailed comparison of radio-based technologies is given in [2]. However, we discuss some of the commonly used radio-based technologies in the following sections.

a) *Wi-Fi*: This is the most common technology for positioning and localization. The main advantage of using this technology is its well-established infrastructure in buildings. Moreover, this can provide larger coverage area in indoors. The disadvantage of using such technology is poor localization accuracy. This technology is more prone to interference. The received signal strength (RSS) of the signal is used to find the ranges and due to which it usually provides accuracy in the order of few meters.

b) *Bluetooth*: Bluetooth operates at the same frequency as Wi-Fi. Bluetooth Low Energy (BLE) is easily available in most of the current smart phones and thus, makes it a feasible candidate for indoor positioning. Apart from the ease of availability, other advantages such as low-cost usage and low-power consumption makes BLE a suitable candidate. Due to low power consumption, it can last for several months on the batteries. The accuracy reported using this technology is somewhere between 1-3 m. Authors in [13] report an accuracy of 2 m. The disadvantages of such techniques are signal attenuation and reflection due to obstacles. A comparison between Wi-Fi and Bluetooth is presented in [14]. A hybrid system which combines both Wi-Fi and BLE is presented in [15]. Many interesting commercial projects such as Apple's inc. iBeacons which use BLE for location purposes are in place [16].

c) *ZigBee*: ZigBee is another interesting technology used for indoor localization. RSSI is obtained to find the position of the nodes. Authors in [17] demonstrate the working of the position system using several Zig Bee nodes using methods like fingerprinting and propagation.

d) *RFID*: This technology uses radio waves to transfer the digital data encoded in RFID tags to the reader. The advantages of using such a system for positioning is its functionality in NLOS. However, it provides low accuracy along with high power consumption when operating in active mode [2], [18].

e) *UWB*: Ultra-wide band signal provides a high localization accuracy due to its fine time resolution signal. It has the ability to resolve multi-paths. The accuracy achieved using this technology can be in the order of few centimeter. The other

advantages of using this technology are low interference with other radios and high penetration capability. However, the disadvantage of using UWB signal is the precise recovery of such a high band signal at the receiver makes it costly.

f) *Cellular Networks*: Many cellular networks which use GSM, EDGE and LTE networks are exploited for positioning of cell phones. The advantage of using such a technology is ease of infrastructure. Additionally, they can cover very large areas with very low interference due to the dedicated licensed frequency. However, the major disadvantage being the lower accuracy (around 100m). Coverage of the signal inside buildings is weak and moreover, it does not perform well in crowded areas [19]. The standardization of positioning in 5G is still under discussion within dedicated task in Release 16 [20]. Localization will be based on the characteristics of the uplink and down-link signals of new radio (NR) (3GPP-technologies) but also on new technologies and network configurations, for example, GNSS (e.g. BeiDou, Galileo, GLONASS, and GPS), Terrestrial Beacon Systems (TBS), Bluetooth, WLAN, RFID, and sensors [21]. The main breakthrough in 5G is due to the employment of massive multiple input–multiple-output (MIMO) beamforming and of millimeter wave (mmWave) signals. The use of mmWave brings a two-fold advantage: large available bandwidth and the possibility to pack a large number of antenna elements even in small spaces (e.g., in a smartphone). Wideband signals offer better time resolution and robustness to multi-path thus improving the performance of OTDOA/UTDOA schemes, as well as paving the way to new positioning methods such as multipath-assisted localization exploiting specular multipath components to obtain additional position information from radio signals [22]. A large number of antenna elements enables massive MIMO and very accurate beamforming. This will make possible the introduction of single-anchor approaches providing cm-level and degree-level accuracy in 6D positioning (3D position and 3D orientation) [23], thus overcoming the problem of deploying a redundant ad-hoc infrastructure which is, nowadays, a major bottleneck for the widespread adoption of indoor localization systems.

g) *LoRa*: It operates in an unlicensed band of 868 MHz. Long Range (LoRa) alliance released this technology with two versions based on the connectivity between the nodes. 1) Point to Point LoRa or star topology, 2) Cell structure-based topology. This is known as LoRaWAN and which follows Low Power Wide area network (LPWAN) group of communication standards [24]. LoRa based indoor localization

system is detailed in [24]. The advantages of such a technology is large coverage range of 30 Km in LOS along with the operation in unlicensed band. However, the drawbacks of such a technique are high cost of modules, lower accuracy and high signal interference and dedicated infrastructures.

h) *Sigfox*: The operation frequency range of this technology is in the ISM band i.e. between 902-928 MHz. This is also based on LPWAN network standards. The bandwidth is 100 Hz [25], [26]. The localization method used with these technologies are RSS along with machine learning techniques to map the measurements [25].

i) *NB-IoT*: 3GPP recently defined this emerging technology to meet the demands of internet of things. This technology uses a small bandwidth of 180 KHz from LTE networks. Although this can cover large number of devices thereby improving the coverage however, the narrow bandwidth restricts the position accuracy to be below 50 meters [27]. High cost and high-power consumption are some of the the disadvantages. Moreover, it is dependent on LTE signal coverage.

Recent Industrial Demands

The use of different technologies depends mainly on the requirement of the application. The recent industrial demands for indoor positioning and navigation systems imposes following requirements:

1. Robust system, catering the dense multi-path affects as well as good penetration through structures.
2. Low-cost low-power system requirement for wireless sensor networks as it involves deployment of large number of sensor nodes.
3. Energy efficient system for mobile nodes.
4. Multi-user scalability of the system. The design of the efficient scheduling access schemes which can accommodate multiple users, and which guarantee QoS.
5. Precise clock synchronization schemes for distributed WSN.
6. Localization accuracy of few centimeter or few millimeters in an indoor environments.
7. Self-calibration mechanism among the anchor nodes.
8. Preferably the use of UWB as wireless technology is due to its operation in unlicensed frequency band and due to its high temporal resolution, as well as its coexistence with other radio technologies.

In this thesis, ultra-wide band impulse radio (UWB-IR) is used as a wireless technology. UWB-IR has high temporal resolution due to its high bandwidth and thus, has the ability to resolve large number of multi-paths in an indoor environment. Therefore, UWB-IR is a potential candidate for obtaining precise ranging and high accuracy in localization. However, to recover the UWB-IR signal at the receiver, high sampling rate analog to digital converter (ADC) is required. In addition to this, high speed processors, such as Field Programmable Gate Array (FPGA) and Digital Signal Processors (DSP) are required to process the recovered samples. These requirements may not be feasible for low-cost and power constrained wireless sensor nodes. The time-based ranging methods such time of arrival (TOA) and time difference of arrival (TDOA) can be used to find the ranging and range difference between the nodes. These ranging techniques require some level of synchronization among the nodes. To avoid the synchronization requirements among the nodes differential time difference of arrival (DTDOA) can be used. Moreover, for practical implementation of timing based schemes, the timing measurements are carried out using clocks at the nodes. These clocks are implemented as counters triggered by the rising or falling edge of an oscillator output signal. These counters are discrete values with the granularity of an oscillator period. Moreover, these clocks are affected by the oscillator variations termed as clock skew and/or clock offset. To avoid high sampling requirements and to obtain precise ranging in the order of few centimeters, equivalent time sampling (ETS) technique is useful for low-cost and low-power wireless sensor nodes.

1.1 Motivations

The main motivations of the thesis are as follows:

1. To sample a UWB-IR signal in real-time, which are of the order of few GHz, a very high sampling rate ADC and high-speed processors are required. High sampling rate ADCs provide high time resolution and positioning accuracy but are very costly for practical deployment in low-cost applications. In context of wireless sensor networks, a cost-efficient sampling technique such as ETS is required which provides high resolution and accuracy.
2. In the current literature, the TOA modeling is restricted for real time sampling (RTS) scheme for indoor localization. Thus, it is important to model the TOA ranging equations for the ETS-based system, referred to as E-TOA in this work.

3. In the current literature, a continuous time clock modeling is assumed to obtain the TOA ranging measurements. However, in practice, the clocks are implemented in discrete-time counters which does not provide same level of granularity as compared to continuous time. Therefore, to analyze a practical indoor localization system a discrete-time clock modeling is required.

4. Clock Synchronization is one of the crucial and important parameters in the design of precise localization system. Synchronization of wireless sensor nodes in the order of few nanoseconds is challenging and costly for practical deployment. Thus, a practical scheme without stringent synchronization requirement among the nodes is needed.

5. E-TOA measurements are corrupted by the variations in the clock i.e. clock skew and clock offset at both the transmitter and receiver. A scheme is required to correct the variations due to the clock and obtain the correct E-TOA measurements.

6. Experimental verification and demonstration of localization scheme using ETS-based system using low-cost architecture for asynchronous wireless sensor nodes have not been carried out in the prior works.

1.2 Objectives

The main objectives of this thesis are as follows:

1. To obtain high resolution TOA, in the order of sub nanoseconds, while still using low sampling rate ADC's with sampling rate in the order of few MHz.
2. To mathematically model high-resolution low sampling rate ETS technique for obtaining the precise TOA.
3. To develop an indoor localization system using asynchronous wireless sensor network nodes.
4. To verify and demonstrate the proposed high time resolution scheme experimentally for asynchronous wireless sensor nodes using in house designed components.
5. To experimentally demonstrate the multi-tag localization scheme for wireless sensor nodes.

1.3 Contributions

Motivated from the above requirements, in this thesis, we have developed and presented a practical indoor localization system which caters some of the mentioned

requirements. Below are the major contributions of the thesis:

1. We use ETS technique to extract high resolution channel impulse response (CIR) which is used to obtain very precise TOA information. A novel analytical equation is proposed to determine the equivalent time of arrival (E-TOA) for achieving sub-ns resolution, with much reduced ADC sampling rate (in the order of 2-3 MHz). The proposed E-TOA equation is different from the conventional real-time sampling equation due to the presence of additional transmitter clock drift, and thus sensitive to both the transmitter and the receiver clock drift variations. The effect of timing uncertainties relating to the transmitter clock start time and the receiver clock offset is analyzed with variations in the transmitter and receiver clock drift. The validation of the proposed E-TOA equation is carried out numerically using simulations along with experimental validation.

2. Reconstructed signal obtained by ETS technique, that require periodic transmission of the same signal repetitively, is severely affected by the variation in transmitter and receiver clock drift as against the real time sampling where the variations are due to receiver node clock drift only. Thus, it requires a protocol to precisely estimate the transmitter and receiver clock parameters. A scheme, which uses a novel mathematical equivalent time of arrival (E-TOA) model for ETS based system, for clock drift estimation is presented. Based on clock drift estimation parameters, receiver nodes are tuned to the same frequency.

3. Since, the reconstructed signal obtained by ETS technique is severely affected by the variation in transmitter (or target node) clock drift, a frame level drift estimation is required to accommodate different/multiple target node at each frame and to obtain the precise E-TOA information. Thus, a novel scheme which estimates the drift at frame level is proposed and demonstrated for obtaining the precise range information.

4. E-TOA measurements are further used to propose an equivalent differential time difference of arrival (E-DTDOA) based ranging algorithm, which relaxed the time offset synchronization requirement between the wireless nodes, and still achieving high time resolution. The E-DTDOA range measurements are subsequently used to obtain precise localization of the target node/s. The feasibility of the algorithm proposed is demonstrated experimentally using in house designed wireless sensor nodes.

5. Utilizing the proposed E-DTDOA mechanism, this thesis further demonstrates, experimentally, a hybrid sub-1 GHz and UWB-IR based indoor positioning system.

The system integrates sub-1 GHz technology for medium access control (MAC) and communication along with UWB-IR technology for the purpose of ranging. The presented positioning system employs energy efficient tags using UWB-IR in transmitter mode and the DTDOA as a measurement technique. A static experiment was performed for the system with 10 tags and 4 anchor nodes in an indoor environment. During the experimentation, the accuracy in tags positioning was found to be less than 30 cm in an area of 16m x 5.3m.

6. Low costs and easy deployment. Receiver nodes are designed using low-cost off the shelf components with the sampling rate of around 2-3 MHz, unlike systems using high sampling rate and costly oscilloscopes. We also highlight that the scheme proposed in this work allows efficient use of low-cost anchor nodes and target nodes with poorer crystal clock and larger time drift for precise localization. Moreover, the energy efficiency is maintained, by using an architecture where UWB-IR tags are operated in transmit only mode to save the power at the mobile/target sensor nodes by exploiting the fact that reception of UWB-IR is more power consuming than transmission.

1.4 Organization of the thesis

The remainder of the thesis is organized as follows:

Chapter 2 discusses the work done in this field and state-of-the-art methodology. An introduction on Ultra-wideband signal along with the localization scheme is discussed. The discussion on different aspects of synchronization is provided. The introduction on real time sampling and equivalent time sampling is given followed by comparing some commercially off the shelf available UWB based platforms with state-of-the-art literature available..

In Chapter 3 a novel analytical equation is proposed to determine the E-TOA for achieving sub-nanosecond resolution, with much reduced ADC sampling rate (in the order of 2-3 MHz). A detailed validation of the proposed E-TOA equation is provided along with the sensitivity analysis with variations in transmitter and receiver clock parameters. The validation of the E-TOA equation is further demonstrated using real experiments.

In Chapter 4 the proposed E-TOA equations are utilized to propose a linear clock skew estimation method. The E-TOA is further used to propose a novel E-DTDOA algorithm for localization of a target node using UWB-IR. The verification of the

clock estimation method is demonstrated experimentally. The feasibility of the target node localization using the proposed algorithm is shown experimentally.

In chapter 5 we present experimental results of localization of a target node using ETS technique for a wired architectures. The wired architecture is extended to a wireless architecture and presents a multi tag hybrid sub-1 GHz and UWB-IR based indoor positioning system. Chapter 6 concludes the thesis.

2 Literature Review of Related Work

2.1 Ultra-Wide Band Signals

As per the Federal Communications Commission (FCC) guidelines a signal is considered as Ultra-Wide Band (UWB) signal if it satisfies the following criteria:

1. The 10-dB fractional bandwidth is greater than 0.20. The fractional bandwidth is defined as

$$B_{frac} = \frac{f_h - f_l}{f_c} \geq 0.2 \quad (2.1)$$

where, f_h and f_l are the upper and lower frequency of the -10 dB emission point and f_c is the carrier frequency.

2. The 10-dB absolute bandwidth is equal to or greater than 500 MHz, irrespective of the fractional bandwidth i.e.

$$B_{abs} = f_h - f_l \geq 0.5 \quad (2.2)$$

UWB signal spectrum is shown in Fig.2.1. UWB spectrum ranges between 3.1 to 10.6 GHz [28]. With such a large bandwidth spectrum, UWB based systems use the spectrum that have already been assigned to other systems. Therefore, the regulatory bodies like FCC has imposed strict regulatory condition on signal transmission power. Average equivalent isotropically-radiated power (EIRP) limit for UWB transmission in different countries is shown in Fig.2.2. We discuss some of the UWB signaling techniques in the following subsection.

2.1.1 Ultra-Wide Band Impulse Radio (UWB-IR)

The narrow pulses which occupy very large bandwidth and are like impulses in time domain, in the order of sub-nanoseconds, are referred to as impulse radio. The work in [29] was a pioneering work where time-hopped UWB impulse radio was introduced. They introduced a carrier-less impulse radio scheme which do not require

Figure 2.1: Ultra Wide-band signal spectrum

up-conversion and down-conversion of the transmitted and received signal, respectively. It can also be used with some of the modulation techniques. The most common modulation schemes are pulse position modulation (PPM) and pulse amplitude modulation (PAM). The information is embedded in the position or in amplitude in PPM and PAM, respectively. The main advantage of using such impulses are interference reduction. The interference from other sources can be reduced as receivers listen to the channel for small fraction of periods since the duty cycle for these pulses are very low. These narrow pulse has another advantage that transmission of such pulses consumes very less energy. Therefore, using such narrow pulses can be a step towards designing energy efficient systems [30]. Moreover, these narrow pulses provide very high time domain resolution. In this work, we use UWB-IR based signals, as it provides very high time resolution and thus very useful in obtaining precise range information. This in turn can result in accurate position estimation for indoor positioning system. The advantages and benefits of UWB-IR is summarized in Table.2.1. The commonly used pulse shapes include Gaussian, Laplacian and Hermitian. The pulse shape of UWB-IR is assumed to be Gaussian. The Gaussian pulse can be expressed as

$$\Psi(t) = Ae^{-\frac{t^2}{2\sigma^2}} \quad (2.3)$$

where A controls the energy of the signal, while σ controls the pulse width. The derivatives of this Gaussian pulse is another Gaussian pulse of higher order. It is

Figure 2.2: FCC UWB spectrum mask for indoor environment

interesting to note that 4^{th} and 5^{th} derivatives of Gaussian pulse is closely distributed within the defined range limits of FCC spectrum mask. Thus, Gaussian pulse of 5^{th} order is a suitable candidate for practical implementation of systems using UWB as the transmission technology. The Gaussian pulse used in this thesis is shown in Fig.2.3.

2.1.2 Multi-band UWB

The idea of using multi-band schemes is to divide the whole UWB band into multiple bands. This allows efficient use of the overall spectrum by allowing multiple UWB signal at different frequency. The overall band is divided into small spectrum bands, with frequency greater than 500 MHz so as to satisfy the FCC emission regulations. The main advantages of using multi-band is 1) Co-existence with different wireless technologies. 2) Interoperability, since the allocation of the UWB band is different in different parts of the world, by having different bands the UWB devices can be used in different parts as per the regulations of that region. 3) Ability to avoid Narrow-band interference, the transmission over the frequency where other narrow band devices (such as 802.11a compliant WLAN devices) operate can be avoided.

Table 2.1: Advantages and Benefits of UWB-IR

Advantages	Benefit
It can coexist with other radio technologies	Exploits use of unlicensed band
Large capacity of the channel	High data rate
Can work in low SNR scenarios	Comparatively perform better in noisy environments.
Transmit power is low	Provides high degree of security with low probability of detection and intercept.
Simple transceiver architecture	Enables ultra-low power, smaller form factor, and better mean time between failures, all at a reduced cost.

2.1.3 Multi-carrier UWB

In this approach UWB symbols are used to modulate multiple carriers for transmitting information. Multiple non-overlapping orthogonal carriers can be used to implement such a technique. Orthogonal Frequency Division Multiplexing (OFDM) is another way which can be used for implementation. The advantages of using multi-carrier scheme is that it is robust against inter symbol interference (ISI), however, it has several disadvantages: 1) Low spectral efficiency, use of multiple carriers result in reduction of spectral efficiency of the system 2) High complexity of the transmitter and receiver, as it uses many mixers and filter banks. Some of the related works employing Carrier based (CB) UWB and Impulse based (IB) UWB techniques are given in Table.2.2.

2.2 Standardization

IEEE 802.15.3c and IEEE 802.15.4a are the two UWB physical layer standards to cater for high data rate and low data rate applications, respectively [40]. The former supports 110 Mbps up to 10 m with OFDM technology. The possible applications foreseen by the task group include multimedia streaming, imaging etc. IEEE 802.15.4a standard supported potential applications include wireless sensor networks, home automation, etc. IEEE 802.15.4a task group provides various channel models for different indoor environments. Saleh and Valenzuela (SV) channel

Figure 2.3: Ultra Wide-band impulse radio

model is the best suited model for indoor propagation [41]. It is adopted in IEEE 802.15.4a for channel modeling. According to SV model, the UWB channel model has clusters, and within the clusters has multi-path components (MPCs), known as rays. The cluster follows a Poisson arrival rate with arrival rate denoted by Λ , while rays following a Laplacian distribution with ray arrival rate λ . The mean cluster energy decays exponentially with parameter Γ . Different values of the parameters Λ , λ and Γ is selected for different propagation environments. The channel model is demonstrated in Fig.2.4. A channel model can be expressed as

$$h(t) = \sum_{i=1}^C \sum_{j=1}^R a_{i,j} \delta(t - t_i - \gamma_{i,j}), \quad (2.4)$$

where, C denotes the number of clusters with R rays, $a_{i,j}$ is the channel coefficient of the j^{th} ray in the i^{th} cluster, t_i denotes the arrival time of the clusters and $\gamma_{i,j}$ denotes the arrival of j^{th} ray of the i^{th} cluster. Moreover, this standard models the variations in the amplitude of the channel coefficient $|a_{i,j}|$ as Nakagami distribution. Nakagami distribution is mathematically given by

$$p_{|\alpha|}(x) = \frac{2}{\tilde{\Gamma}(m)} \left(\frac{m}{\tilde{\Omega}} \right)^m x^{2m-1} e^{-mx^2/\tilde{\Omega}}, \quad (2.5)$$

Table 2.2: Some of the related works based on UWB

Research Group	Technology Used	Frequency Band (GHz)	Accuracy	Operating Range (Indoors)
[31]	CB UWB	5.4-10.6	2-5 mm (3-D)	5 m
[32]	CB UWB	7-8	1.7 cm (1-D)	10 m
[33]-[34]	CB UWB	22.58-25.7	0.1-2 mm (1-D)	8 m
[35]	CB UWB	5.8	< 2 mm (2-D)	10 X 10 cm
[36]	IB UWB	2-7	5-10 mm (1-D)	5 m
[37]	IB UWB	3.7 - 5	< 20 cm (2-D)	8 m
[38]	IB UWB	3.2 - 5.2	1 cm (1-D)	8 m
[39]	IB UWB	0.01 - 5	1.5 cm (2-D)	2 m

for $x \geq 0$, $m \geq 0.5$ and $\tilde{\Omega} \geq 0$, where m is the Nakagami m -factor, $\tilde{\Omega}$ is the mean power of the channel coefficient, and $\tilde{\Gamma}(m)$ is the gamma function given by

$$\tilde{\Gamma}(z) = \int_0^{\infty} t^{z-1} e^{-t} dt \quad (2.6)$$

for a complex number z with a positive real part. Since IEEE 802.15.4a has mandated the UWB-IR as the preferred localization techniques and its advantages over other techniques, in this work, we adopt this standard with UWB-IR to demonstrate the proposed localization schemes. In the next section we discuss some of the commonly used localization schemes.

2.3 Localization Schemes

The location of an unknown node can be determined using the available information between the unknown position node and the known position nodes. The information between the two nodes includes: 1. Timing information, categorized as TOA and TDOA, 2. Received Signal Strength (RSS), 3. Angle of Arrival (AOA) and 4. Location dependent known as location fingerprinting. We will discuss these techniques in more detail in the subsequent sections. In general, different radio based localization algorithms are categorized as:

- 1. Scene Dependent:** The characteristics of the signal received from the node

Figure 2.4: Channel model for indoor environments

to be localized is searched in the database and the closest match is obtained. Corresponding to the searched signal characteristics, location is obtained from the database. The database contains the statistical information of the signal characteristics along with the corresponding location information stored during the real experiments.

2. Proximity: In case of proximity based techniques, many nodes are distributed in an environment with a particular Cell-ID. The target nodes identifies its location based on the Cell-ID received from a particular cell.

3. Geometrical: The geometrical approach is further divided into sub categories as:

a) Angulation based: This requires the angle information of the incoming or outgoing signal from one node to another. This further includes,

- Angle of Arrival (AOA)

b) Lateralation based: This require the range information to be recovered between two nodes. This is further categorized as:

- Received Signal Strength (RSS)
- Time of Arrival (TOA)
- Time Difference of Arrival (TDOA)
- Phase of Arrival (POA)

2.3.1 Angulation Based

This approach exploits the use of signal angle of arrival to localize the target node. Signal angle of arrival is obtained by using multiple antennas. Using an array of antennas increases the cost and complexity of the hardware. However, the advantages of using such technique is that they do not require any clock synchronization and moreover, they do not require many nodes for localization.

1. Angle of Arrival (AOA)

AOA measurement provides information about the direction of an incoming signal, hence the angle between the two nodes. The method is illustrated in Fig.2.5. The combination of antenna arrays, with N_a antennas element, are used to measure the AOA of a signal. A simple method to find the AOA is to obtain the phase difference between the signals received at two different antennas. The phase difference can be

Figure 2.5: Illustration of angle of arrival. Signal coming on d spaced antenna array.

written as

$$\phi_1(\theta) - \phi_0(\theta) = 2\pi f_c \frac{d}{c} \sin\theta \quad (2.7)$$

where $\phi_i(\theta)$ is the phase seen at element i , f_c is the carrier frequency and c is the speed of light. Another method to determine the AOA of the signal is using the beam-forming techniques. In such techniques, antenna arrays direct their beam towards a particular direction to detect the incoming signal [42]-[45]. In a broader

sense, AOA technique can be divided into the following categories: 1) Conventional AoA technique, this technique direct their beam in all the direction and find the maximum power peaks [46]. Some of the common techniques include delay-and sum method [44], and Minimum Variance Distortion less Response (MVDR) [47]. However, the bandwidth plays an important role in the resolution of the angle and thus requires many antenna arrays. 2) Subspace-based techniques, these sub-optimal techniques decompose the received signals into signal subspace and noise subspace. Some of the classical sub-space based methods are MUSIC [48] and ESPRIT [49]. MUSIC is based on the Eigen value decomposition of the input data. The disadvantage of such techniques is that they do not perform well in low SNR. 3) Maximum likelihood techniques, these are the optimal techniques and perform better than the subspace based methods even in low SNR scenarios. However, they have high computational complexity to be implemented in real scenarios.

Figure 2.6: Angle of arrival at anchor nodes

The position of the target node can be determined by obtaining the AOA using the techniques discussed above. Assuming the position of the target node as (x, y) and from Fig.2.6, the x-coordinate and y-coordinate, assuming the known anchor node coordinates with d meters as the distance between the two anchor nodes, is given as:

$$x = \frac{d \tan(\alpha_2)}{\tan(\alpha_2) - \tan(\alpha_1)}, \quad y = \frac{d \tan(\alpha_1) \tan(\alpha_2)}{\tan(\alpha_2) - \tan(\alpha_1)} \quad (2.8)$$

These approaches are very costly and computationally complex when applied for high bandwidth signal such as UWB-IR, since the detection of such a narrow pulse with multi paths is very challenging [50]. In such cases time based approaches are efficient and low-cost.

2.3.2 Lateration based

Lateration is a technique to compute the position of a target node (TN) after obtaining its distances or ranges from different fixed non-collinear points or anchor nodes (AN). When three non-collinear points are involved, it is termed as trilateration. The illustration is shown in Fig.2.7. The distance equation between AN and TN is the equation of a circle. The distance of the TN from each of the non-collinear points or ANs forms a circle which has a center as the known location of the AN and radius as the distance between AN and TN. The point of intersection of three of these circles then yields the coordinates of the object of interest. A solution to the trilateration problem in 3D is presented in [51], which is considered as computationally most efficient method for the trilateration problem.

Figure 2.7: Trilateration-based localization of a target node (TN)

Trilateration based technique requires the measurement of distances between the TN and ANs. This can be achieved by measuring either of the following:

1. Phase shift Or Phase of Arrival
2. Received signal strength (RSS)
3. Time of flight (TOF)

1. Phase Shift or Phase of Arrival (POA)

It exploits the use of carrier phase difference for obtaining the distance between

the nodes. The transmitted carrier signal is received at the receiver where it is phased locked and sent back to the same transmitter. Difference in the phase of the reflected signal and originally transmitted signal is converted to the distance between the transmitter and receiver using the below relation [52]

$$d = \lambda \left(\frac{\Delta\phi}{2\pi} + n \right) \quad (2.9)$$

where, d is the distance between the transmitting and receiving nodes, $\Delta\phi$ is the phase difference, λ is the wavelength of the carrier signal, and n is the number of cycles passed between the transmission and reception of the signal. However, the measurement of phase difference can only be made modulo of 2π , which leads to an ambiguity in determining the phase difference of the source [53]. To solve the phase ambiguity, certain methods such as modulo conversion method [54], subarray grouping and ambiguity searching [55] was proposed.

2. Received signal strength (RSS) based positioning

It provides the range information between the transmitting and receiving node based on the relation between distance and the received signal strength (RSS) at the receiver [56]. The RSS measurements are used for this purpose. Ideally, average RSS over a sufficiently long time interval can be modeled as [57]

$$P(d) = P_0 - 10n\log_{10}(d/d_0), \quad (2.10)$$

where n is the path loss exponent which is dependent on the environment, $P(d)$ is the average received power (dB) at a distance d and P_0 is the received power (dB) at a reference distance d_0 . The $P(d)$ can be obtained as

$$P(d) = \frac{1}{T} \int_0^T |r(t)|^2 dt, \quad (2.11)$$

where the amplitude of the multi-path components are represented by $r(t)$ as observed in the channel impulse response. The above formulation however excludes the effect of shadowing and multi-path fading. If the effect of shadowing and multi-path is included, the received power (in dB) can be modeled as a Gaussian random variable with mean $P(d)$ and variance σ_{sh}^2 ; i.e.

$$10\log_{10}P(d) \sim N(P(d), \sigma_{sh}^2).$$

For the power model described above, the lower bound of the estimated distance can be given as [58]

$$\sqrt{\text{Var}(\hat{d})} \geq \frac{\ln 10}{10} \frac{\sigma_{sh}}{n} d,$$

where, $\hat{d}$ represents an unbiased estimate of d . The drawback of RSS based scheme is that it is affected by the variation in the multi-path noise, shadowing, scattering and interference. These variations may result in high error in the RSS value, due to which this technique is used with other techniques in combination, to improve the ranging accuracy [59].

3. Time of flight (TOF) based Positioning schemes

The distance between two nodes A and B can be obtained by measuring the TOF information of a signal propagating from node A (or transmitter) and node B (receiver). The TOF information is given by

$$t_{pd} = d/c \quad (2.12)$$

where d is the actual distance between the nodes and c is the speed of light. The matched filter or correlation-based TOA estimation technique is used to combat the noise from the received signal [60]. For example, let the received signal at a node be expressed as

$$r(t) = \alpha s(t - t_{pd}) + n(t) \quad (2.13)$$

where t_{pd} is the propagation delay or TOF estimate, α is the channel coefficient, and $n(t)$ is white Gaussian noise with zero mean and a spectral density of $N_0/2$. The correlation is performed between the received signal $r(t)$ and a template signal $s(t - \hat{\tau})$, for different delay $\hat{\tau}$. The output of the correlation can be used to estimate the TOA. Similarly, in matched filter scheme, a filter is used at the receiver which is matched to the transmitted signal. This matched filter technique estimate the peak of the output of the matched filter [61]. The relation between the bandwidth and the theoretical limits for TOA estimation is given by the CRLB which can be expressed as [62], [63]

$$\sqrt{\text{Var}(\hat{t}_{pd})} \geq \frac{1}{2\sqrt{2\pi}\sqrt{SNR}\beta}, \quad (2.14)$$

where $\hat{t}_{pd}$ represents the estimated TOA, $SNR = \alpha^2 E/N_0$ is the signal-to-noise ratio, E denotes the energy of the signal, and β denotes the bandwidth of the signal. Notice that the denominator of (2.14) is proportional to the energy in the

signal and β depends on the shape of the pulse. Contrary to RSS-based methods, the ranging capability depends on the signal structure. This shows that large value of β is beneficial for ranging. In the following sub-sections we discuss some of the time based ranging schemes. The notations used in the following sub-sections is given in the box below.

Notations: The variables indicated in small letters denote the reference time (nominal value). The variable in small letter with $(\tilde{\bullet})$ denotes the skewed time with respect to the reference time (referred as actual value). The variable $(\overset{\circ}{\bullet})$ denotes the local time. The variables indicated in capital letter denote the equivalent time having both nominal and actual values. The superscript and the subscript on the variables denote either transmitter (tx) and receiver (rx).

1. Time of Arrival (TOA): TOA is used to obtain the range information between the transmitting and receiving node. Range or distance is obtained using (2.12). This forms the basis for TOA technique commonly used in positioning. Assuming the transmitter as the TN and receiver as the AN. The distance between the TN and AN is the radius of the circle with AN position coordinates as the center of the circle. To locate a TN, TOA technique requires at least three such ANs which make circles with respect to the TN. The intersection point of the circles is the location of the TN. The illustration of this technique is shown in Fig.2.8. There are

Figure 2.8: Time of arrival illustration

many TOA variants available in the literature. One of the strict requirement while using TOA is that the TN and ANs must have synchronized clocks. Moreover, the AN must know the transmit start time of the TN to obtain the TOA between the nodes. The timing information can be exchanged via certain protocols given below:

a) One-way ranging: At a certain start time t_{start}^{tx} , node tx transmits a signal which is received by the node rx. The timestamp at node rx is given by

$$t_{rx} = t_{start}^{tx} + t_{pd,rx}^{tx} \quad (2.15)$$

where $t_{pd,rx}^{tx}$ is the propagation delay between node tx and node rx. In this scenario, nodes are assumed to be synchronized and t_{start}^{tx} is known at the node rx, then $t_{pd,rx}^{tx}$ can be obtained. After obtaining $t_{pd,rx}^{tx}$, the distance between the node tx and node rx or the ranging can be obtained as $d_{rx}^{tx} = t_{pd,rx}^{tx} \times c$, where c is the speed of light. This distance information between the nodes is subsequently used to find the position of the target node (or node tx), whose position is given by (x, y) and anchor node (or node rx) with fixed position (x_n, y_n) . The distance between node tx and node rx can be given by

$$(x - x_n)^2 + (y - y_n)^2 = (d_{rx}^{tx})^2 \quad (2.16)$$

Above equation can be solved in 2-dimensions by having two independent equations which can be obtained by having two anchor nodes in the system. However, solving two independent circle equations gives two ambiguous points. Therefore, at-least three anchor nodes are required to find a unique solution. The nonlinear equations of three circles with center at (x_1, y_1) , (x_2, y_2) and (x_3, y_3) and radius d_1 , d_2 and d_3 , respectively, are given as

$$\begin{aligned} (x - x_1)^2 + (y - y_1)^2 &= d_1^2 \\ (x - x_2)^2 + (y - y_2)^2 &= d_2^2 \\ (x - x_3)^2 + (y - y_3)^2 &= d_3^2 \end{aligned} \quad (2.17)$$

The solution of the above equations is a point of intersection of the three circles. It is important to note that presence of clock imperfections in the transmitting and receiving node affect the timing equations. However, for simplicity we have discussed the case without clock offset and clock skew. We will discuss about the variations in clock parameters in the later sections.

b) Two-way ranging: This scheme does not require any stringent synchronization between the nodes. In this scheme, node tx transmits a signal to node rx which is transmitted back to node tx after a certain delay. Total time taken by the signal to reach back at node tx is known as round trip time (RTT). At node tx, RTT is calculated, which is given by

$$RTT = 2t_{pd,rx}^{tx} + delay \quad (2.18)$$

where $t_{pd,rx}^{tx}$ is the propagation delay which can be calculated as $delay$ is known at node tx. Further, the position of a node can be obtained in a similar manner as discussed in one-way ranging scheme.

For the narrow band signal and in indoor scenarios, the multi-path effect will be a factor which will influence the estimation of TOA. However, this effect can be reduced by using an Ultra wide bandwidth signal owing to its narrow pulses. Thus, it is preferable to use UWB signaling scheme while using time based ranging methods. For the practical implementation of such schemes, to measure the TOA with high precision in the order of few nanoseconds, very high frequency clock rates with high granularity or precision is required.

2. Time Difference of Arrival (TDOA): TDOA is based on calculating the difference of TOA of the signal received by different anchor nodes. In case of TDOA, we do not need any synchronization between the TN and AN which is an advantage over TOA based range measurement where synchronization is required between TN and AN. In a case with two ANs, TDOA measurement can be obtained by estimating TOA at each AN and then by taking the difference between the two estimates. For example, $\hat{\tau}_1$ is the estimated TOA at AN₁(or rx₁) from the received signal $r_1(t)$, and $\hat{\tau}_2$ is the estimated TOA at AN₂(or rx₂) from the received signal $r_2(t)$. The TOA estimates are given by

$$\begin{aligned} \hat{\tau}_1 &= t_{start}^{tx} + t_{off,rx_1} + t_{pd,rx_1}^{tx}, \\ \hat{\tau}_2 &= t_{start}^{tx} + t_{off,rx_2} + t_{pd,rx_2}^{tx}. \end{aligned} \quad (2.19)$$

where, t_{start}^{tx} is the transmit start time of TN, t_{off,rx_1} , t_{off,rx_2} are the clocks offsets and t_{pd,rx_1}^{tx} and t_{pd,rx_2}^{tx} are the propagation delay at AN₁ and AN₂, respectively. After taking the difference between the two estimates, the TDOA measurement can be

written as

$$\begin{aligned}\hat{\tau}_{TDOA} &= \hat{\tau}_1 - \hat{\tau}_2 \\ \hat{\tau}_{TDOA} &= t_{off,rx_1} + t_{pd,rx_1}^{tx} - t_{off,rx_2} - t_{pd,rx_2}^{tx}.\end{aligned}\quad (2.20)$$

It is observed that t_{start}^{tx} gets canceled due to which there is no synchronization requirement between TN and AN. If the ANs are synchronized, the timing offset is the same for each TOA estimation (i.e. $t_{off,rx_1} = t_{off,rx_2}$), then TDOA measurement can be further obtained as

$$\hat{\tau}_{TDOA} = t_{pd,rx_1}^{tx} - t_{pd,rx_2}^{tx}. \quad (2.21)$$

In terms of distance, this can be written as

$$\frac{\hat{d}_{TDOA}}{c} = \frac{d_{rx_1}^{tx}}{c} - \frac{d_{rx_2}^{tx}}{c}, \quad (2.22)$$

Equation (2.22) results in hyperbola as shown in Fig.2.9. Sufficient number of an-

Figure 2.9: Illustration of hyperbolic TDOA

chor nodes are needed to get independent TDOA equations. These equations are subsequently used to obtain the target node coordinates. As given by eq. (2.14), the accuracy of TOA measurement increases with the bandwidth and SNR of the signal, same conclusion holds true for TDOA measurement. Moreover, TDOA measurements can also be obtained by cross correlation method. The cross correlation between the received signal $r_1(t)$ and $r_2(t)$ is performed. The delay is calculated by finding the largest value out of the cross correlated values. The cross-correlation

function can be expressed as in [64]

$$\phi_{1,2}(\tau) = \frac{1}{T} \int r_1(t)r_2(t + \tau)dt, \quad (2.23)$$

where T is the observation interval, and the TDOA estimate is given by

$$\hat{\tau}_{\text{TDOA}} = \arg \max_{\tau} |\phi_{1,2}(\tau)|.$$

In general, the performance of cross-correlation based TDOA degrades over multi-path channel and colored noise as compared to that of single-path channel and white noise. Depending on the nodes transmission architecture, TDOA is classified in the

Figure 2.10: Illustration of TDOA scheme

following categories:

a) Unilateral TDOA: In this category, anchor nodes (AN) act as a transmitter and a target node (TN) receives the signal from the different ANs and use this information to localize itself. Since, the TN is receiving the signal from the ANs, this type of system can accommodate many TNs. This improves the scalability of the system. Therefore, this type of scheme is useful in applications which require localization and navigation of many TNs at minimum implementation cost. In this case, ANs must be synchronized which can either be done using wires or wirelessly. The disadvantage of using this scheme while using UWB-IR is that, the target node which is a mobile device and restricted by power is operated in reception mode and reception of UWB-IR consumes more power than transmission. Thus, this

Figure 2.11: Unilateral TDOA

Figure 2.12: Unilateral TDOA timing diagram

architecture may not be feasible in low power target nodes. Systems using this architecture are detailed in [65]-[67], and its classes such as asynchronous TDOA, sequential TDOA are explained in [68] and [69], respectively. The illustration of the architecture and the timing diagram of this technique is shown in Fig.2.11 and Fig.2.12.

b) Multilateral TDOA: In this category of TDOA mechanism, TN transmits the signal and ANs receives the signal to perform the TDOA. The AN needs to be fully synchronized. The range difference is obtained between different pair of ANs which can be used at the common server to find the localization of the TN as demonstrated in Fig.2.13 and Fig.2.14. This architecture can accommodate many TNs with a certain update rate by utilizing proper transmission scheme such as time division multiple access (TDMA). It is also important to note that the transmission of UWB-IR consumes less energy than its reception. Therefore, the main advantage of this scheme over unilateral scheme is its energy efficiency at the TN, which is

Figure 2.13: Multilateral TDOA

Figure 2.14: Multilateral TDOA timing diagram

very important as the TNs are battery operated. In this thesis, we have utilized this approach where TNs are operated in transmit-only mode.

We list some of the contributions related to TDOA mechanism. Authors in [70] provide the basic method to solve the TDOA equations for the localization of the node. Since TDOA equations are non-linear equations, authors in [71]-[72] provide a simplified method to solve these equations for localization. While solving the TDOA equations, solutions can have local minima which is avoided in [73]. Moreover, authors in [74] provide the lower bound on the accuracy of TDOA. Some of the applications which utilizes the idea of TDOA for indoor localization are given in [75]. In [76], some of the TDOA measurement mechanisms in multi-path conditions are demonstrated. While applying these techniques in the real world scenarios, it is necessary to tackle the outliers produced due to some random errors in the system. Therefore, authors in [77] propose ways to eliminate the outliers in the

TDOA measurements. In most of the literature pertaining to TDOA, either full clock synchronization (i.e. clock skew and clock offset synchronization) or partial clock synchronization (i.e. with clock offset only synchronization) among the anchor nodes is assumed. Based on anchor nodes synchronization level, TDOA can be classified in different categories: First, where anchor nodes are assumed to be fully synchronized, with clock drift and offset synchronization, and TDOA can be obtained across the anchor nodes [78], [79]. Second, where anchor nodes are partially synchronized, with clock offset only synchronization [80]. Third, where anchor nodes are asynchronous and which is referred to as asynchronous-TDOA (A-TDOA). In third category, TDOA is obtained at a node from either two different signal sources or two signals from same signal source but delayed in time. In [81], authors calculate the TDOA at only one anchor node, between the signal directly received from other anchor node and the same signal relayed and forwarded by the target node. In [82], time difference measurement is obtained at an anchor node between the signal from the same target node at different locations. Sufficient locations are used to formulate enough equations and all the locations are estimated simultaneously. However, synchronization of clocks in the order of picoseconds, which is required for precise ranging, is challenging and costly task over wireless radio links [83]-[86]. Moreover, it is a repetitive process because the clock error accumulates over time due to the presence of clock drift. Therefore, it is practically infeasible and challenging to repetitively synchronize wireless sensor nodes in real world scenarios. We use DTDOA to overcome these challenges which is discussed below.

3. Differential Time Difference of Arrival (DTDOA): To overcome the challenges of synchronization among the receiver (or anchors) nodes, DTDOA, that works on un-synchronized ANs, is a practical consideration [87]-[89]. DTDOA is based on adding a transmitting reference node (RN) in the system, and, the overall DTDOA mechanism works in the following steps. First, ANs receive the signals from two transmitting nodes (i.e. from a TN, whose position is to be determined and a RN) and recording the TOA information from each of the transmitting node. Second, at each participating AN, a TDOA is obtained between the two TOAs recorded in step 1. Third, DTDOA is obtained by finding the difference between the TDOAs across the anchor nodes. Step 1 cancels the clock offset which is same in each TOA recorded from different TN, and step 2 further cancels the start time of the two transmitting nodes. The illustration is given in Fig.2.15. This will be more clear with the detailed equations given below. From the obtained DTDOA

measurement values, TNs can be located by solving the set of equations.

We illustrate the mathematical details of the DTDOA mechanism using two transmitting nodes (TN denoted by tx_1 and RN denoted by tx_2) and two ANs (denoted by rx_1 and rx_2). The TOA of the signal transmitted from the TN tx_1 at time instant t_1 , which is received by the AN rx_1 is given as

$$\tau_{rx_1}^{tx_1, (t_1)} = t_{start}^{tx_1, (t_1)} + t_{pd, rx_1}^{tx_1, (t_1)} + \phi_{rx_1} + n_{rx_1}^{(t_1)}, \quad (2.24)$$

where $t_{start}^{tx_1, (t_1)}$ is the start time of tx_1 at time instant t_1 , $t_{pd, rx_1}^{tx_1, (t_1)}$ is the propagation delay between tx_1 and rx_1 , ϕ_{rx_1} is the clock offset of rx_1 and $n_{rx_1}^{(t_1)}$ is the measurement noise at rx_1 at time instant t_1 . Similarly, the TOA of the signal transmitted from

Figure 2.15: DTDOA Illustration

the target node tx_1 at time instant t_1 which is received by the anchor node rx_2 is given as

$$\tau_{rx_2}^{tx_1, (t_1)} = t_{start}^{tx_1, (t_1)} + t_{pd, rx_2}^{tx_1, (t_1)} + \phi_{rx_2} + n_{rx_2}^{(t_1)}, \quad (2.25)$$

where $t_{pd, rx_2}^{tx_1, (t_1)}$ is the propagation delay between tx_1 and rx_2 , ϕ_{rx_2} is the clock offset of rx_2 and $n_{rx_2}^{(t_1)}$ is the measurement noise at rx_2 at time instant t_1 . The TOA of the signal transmitted from the target node tx_2 at time instant t_2 which is received by

the anchor node rx₁ is given as

$$\tau_{rx_1}^{tx_2, (t_2)} = t_{start}^{tx_2, (t_2)} + t_{pd, rx_1}^{tx_2, (t_2)} + \phi_{rx_1} + n_{rx_1}^{(t_2)} \quad (2.26)$$

where $t_{start}^{tx_2, (t_2)}$ is the start time of tx₂ at time instant t_2 , $t_{pd, rx_1}^{tx_2, (t_2)}$ is the propagation delay between tx₂ and rx₁, and $n_{rx_1}^{(t_2)}$ is the measurement noise at rx₁ at time instant t_2 . Similarly, the TOA of the signal transmitted from the target node tx₂ at time instant t_2 which is received by the anchor node rx₂ is given as

$$\tau_{rx_2}^{tx_2, (t_2)} = t_{start}^{tx_2, (t_2)} + t_{pd, rx_2}^{tx_2, (t_2)} + \phi_{rx_2} + n_{rx_2}^{(t_2)}, \quad (2.27)$$

where $t_{pd, rx_2}^{tx_2, (t_2)}$ is the propagation delay between tx₂ and rx₂, ϕ_{rx_2} is the clock offset of rx₂ and $n_{rx_2}^{(t_2)}$ is the measurement noise at rx₂ at time instant t_2 . The measurement error $n_{rx_i}^{(t_i)}$ is modeled with zero mean and $\sigma_{n_i}^2$ where $i \in 1, 2$. The TDOA at time instant t_1 is given as

$$\Gamma_{rx_1, rx_2}^{tx_1, (t_1)} = \tau_{rx_1}^{tx_1, (t_1)} - \tau_{rx_2}^{tx_1, (t_1)} \quad (2.28)$$

Similarly, the TDOA at time instant t_2 is given as

$$\Gamma_{rx_1, rx_2}^{tx_2, (t_2)} = \tau_{rx_1}^{tx_2, (t_2)} - \tau_{rx_2}^{tx_2, (t_2)} \quad (2.29)$$

Further, taking the difference of (2.28) and (2.29), DTDOA is given as

$$D_{rx_1, rx_2}^{tx_1, tx_2, (t_1, t_2)} = \Gamma_{rx_1, rx_2}^{tx_1, (t_1)} - \Gamma_{rx_1, rx_2}^{tx_2, (t_2)} \quad (2.30)$$

By substituting the values from (2.28) and (2.29) into (2.30), we get

$$\begin{aligned} D_{rx_1, rx_2}^{tx_1, tx_2, (t_1, t_2)} &= t_{start}^{tx_1, (t_1)} + t_{pd, rx_1}^{tx_1, (t_1)} + \phi_{rx_1} + n_{rx_1}^{(t_1)} - \left(t_{start}^{tx_1, (t_1)} + t_{pd, rx_2}^{tx_1, (t_1)} + \phi_{rx_2} + n_{rx_2}^{(t_1)} \right) \\ &\quad - \left(t_{start}^{tx_2, (t_2)} + t_{pd, rx_1}^{tx_2, (t_2)} + \phi_{rx_1} + n_{rx_1}^{(t_2)} - t_{start}^{tx_2, (t_2)} + t_{pd, rx_2}^{tx_2, (t_2)} + \phi_{rx_2} + n_{rx_2}^{(t_2)} \right) \end{aligned} \quad (2.31)$$

Simplifying (2.31), the start time $t_{start}^{tx_1, (t_1)}$ and $t_{start}^{tx_2, (t_2)}$ and the clock offsets ϕ_{rx_1} and ϕ_{rx_2} gets canceled¹ and we get

¹In TOA, TDOA and DTDOA equations, the effect of clock skew is not taken into consideration. Only clock offset is considered which shows that clock signal is partially synchronized. Ignoring clock skew simplifies the equations and makes above mentioned schemes easy to understand. However, it is not practical to ignore the effect of clock skew and therefore, in our work we have considered fully unsynchronized clock containing the effect of both clock offset and clock skew. We will discuss about it in the subsequent section related to clock modeling.

$$D_{rx_1, rx_2}^{tx_1, tx_2}(t_1, t_2) = t_{pd, rx_1}^{tx_1, (t_1)} - t_{pd, rx_2}^{tx_1, (t_1)} + t_{pd, rx_2}^{tx_2, (t_2)} - t_{pd, rx_1}^{tx_2, (t_2)} + n_{rx_1}^{(t_1)} - n_{rx_1}^{(t_2)} + n_{rx_2}^{(t_2)} - n_{rx_2}^{(t_1)} \quad (2.32)$$

Assuming tx_2 as a reference node (RN) with known position, or known term

$\left(t_{pd, rx_2}^{tx_2, (t_2)} - t_{pd, rx_1}^{tx_2, (t_2)}\right)$, eq. (2.32) can provide the range difference between tx_1 and ANs, $\left(t_{pd, rx_1}^{tx_1, (t_1)} - t_{pd, rx_2}^{tx_1, (t_1)}\right)$. To find the coordinates of the target node, more number of ANs can be added to the system and the range difference between the TN and other ANs can be obtained to form sufficient independent equations which can be solved to find the unique solution.

In asynchronous systems, using low-power transmit-only TN where start time of the TN is not known, a DTDOA scheme as proposed in [87]-[89] is more feasible. In [90], [91], real world implementation of DTDOA is presented for asynchronous nodes. In [92], authors assume an artificially generated signal from one of the base node or reference node which is received by other ANs, to find DTDOA across the ANs. Authors in [93] use DTDOA approach to localize the TN. They use additional beacons to first estimate the clock parameters and then use the TN measurements to localize the TN. In [94], authors use DTDOA approach along with clock drift compensation mechanism in large scale asynchronous WLANs. In Table.2.3, various ranging mechanisms have been compared for different criteria.

2.4 Clock Synchronization

Clock synchronization is an important factor for different fundamental operations of wireless sensor networks (WSN) such as signal processing applications/data fusion, spectral efficiency, transmission scheduling [95], power management [96], and scalability [97] etc. In this thesis, however, we will consider the clock synchronization aspects used for positioning of WSN. Positioning of nodes in a WSN includes numerous sensor nodes dispersed in an environment with few or all of them associated with separate clocks. To get the basic range data between the nodes, it is essential for all the sensor nodes to have common clock or time frame. In this context, clock synchronization of the sensor nodes is of great significance in positioning applications as it gives a common time scale to all the participating nodes. Clock synchronization means each participating node's local clock can be coordinated by exchanging the local time information. We illustrate the problem of synchronization with the help of Fig.2.16, Fig.2.17, Fig.2.18. We define $t_i(n)$ as the time of the n^{th} tick ($n = 0, 1, 2, \dots$) of the i^{th} clock ($i = 1, 2, \dots, N$, where N is the total

number of nodes). The synchronization of the clocks for different sensor nodes can be categorized as follows:

Table 2.3: Comparison of ranging mechanism

Criteria	AOA	TOA	TDOA	DTDOA	RSS
2D-3D	At least 2 AN are needed for 2D and 3 AN for 3D.	At least 3 AN are needed for 2D and 4 AN for 3D.	At least 3 AN are needed for 2D and 4 AN for 3D.	At least 3 AN are needed for 2D and 4 AN for 3D.	At least 3 AN are needed for 2D and 4 AN for 3D.
LOS & NLOS	1. Sender and receiver needs clear LOS, 2. Large position error due to multi-path, as they change the phase of a clock.	Has the ability to work in both LOS and NLOS	Has the ability to work in LOS and NLOS	Has the ability to work in LOS and NLOS	LOS is preferred to reduce multi-path effect, walls and obstacles badly effects the performance
Synchronization	Clock sync is not required to a large extent	Sender and AN must be synchronized	AN must be synchronized	No Sync is required.	No Syncn is required.
Other issues	Costly and bulky as it requires large antenna array	Clock drift and offset are the main issues	Degrades the performance due to greater noise	Clock drift is the main issues	Sensitive to channel inconsistency. Short distance is preferred.

1. Unsynchronized Clocks: Clock in the nodes are fully asynchronous i.e. each node has a different local clock frequency (clock skew) and clock phase (clock offset) i.e. the clocks remain asynchronous with generally different local periods $t_i(n) - t_i(n - 1) = T_i$, and phases $t_i(n)$. The illustration of this category is shown in Fig.2.16. Thus, for unsynchronized clocks, we can write from the Fig.2.16

$$\begin{aligned}
t_i(n + 1) - t_i(n) &\neq T, i = 1, \dots, N. \\
t_i(n) &\neq t_j(n) \text{ for each } i \neq j.
\end{aligned} \tag{2.33}$$

Figure 2.16: Unsynchronized clocks

2. Frequency synchronization: In this category, clock is partially synchronized and only clock frequency (clock skew) among the nodes is same but clock offset is different. This state is achieved when nodes exchange the information and attain common frequency. We say that a condition of frequency synchronization to a common frequency $1/T$ is achieved if the local periods $t_i(n) - t_i(n-1) = T$ are the same for all clocks. An Illustration is shown in Fig.2.17. Thus, for frequency synchronization, we have the following condition

$$t_i(n+1) - t_i(n) = T, i = 1, \dots, N. \quad (2.34)$$

Figure 2.17: Frequency synchronized clocks

3. Frequency and Phase synchronization: Full synchronization between the nodes is achieved when both the phase and frequency are same. Full synchronization is attained if clocks tick at the same times, i.e., $t_i(n) = t_j(n), i = j$. An illustration

of this category is shown in Fig.2.18. For frequency and phase synchronization, we can write

$$t_i(n) = t_j(n) \text{ for each } i = j. \quad (2.35)$$

Figure 2.18: Frequency and phase synchronization

2.4.1 General Clock Concepts

Here, we discuss some of the general concepts involved in clock modeling.

2.4.1.1 Continuous Clock model for sensor nodes

A crystal oscillator is used to implement the clock of any sensor node. Each individual node in a sensor network has its own clock. Let t be the true time or the nominal time. However, the clock frequency varies randomly due to various physical effects such as temperature and crystal aging. Therefore, irrespective of whether the clock is tuned perfectly, it may drift away from the ideal time. Two different clock models are:

1. The offset only model: In this model, local clock time at node B, t_B° , is given as

$$t_B^\circ = t + \theta_B, \quad (2.36)$$

where θ_B is the absolute offset of node B's clock with respect to a reference clock. For two nodes B and C, the relative clock equation between two nodes with t_C°

Figure 2.19: Clock model displaying clock skew and clock offset

denoting the local time at node C can be written as

$$t_C^\circ = t_B^\circ + \theta_{B \rightarrow C}, \quad (2.37)$$

where $\theta_{B \rightarrow C}$ represents the relative offset relating the time at node B to that of node C.

2. Skew-offset model: In this model, the local clock at node B is modeled as

$$t_B^\circ = \alpha_B t + \theta_B, \quad (2.38)$$

where $\alpha_B = 1 + \delta_B$ is the absolute clock skew with δ_B is the variations in parts per million (ppm). θ_B is the absolute offset of node B. The relative equation between the node B and node C can be given as

$$t_C^\circ = \alpha_{B \rightarrow C} t_B^\circ + \theta_{B \rightarrow C} \quad (2.39)$$

where, $\alpha_{B \rightarrow C}$ and $\theta_{B \rightarrow C}$ represents the relative offset and skew, respectively. The clock model is graphically illustrated in Fig.2.19. The effect of clock skew is shown in Fig.2.20 over a period of time. It is demonstrated using Fig.2.20 that during the interval of 20 milliseconds, the clock skew effect can be as high as 1.2 microseconds, even for a smaller time duration of 3 milliseconds the error can be approximately 0.18 microseconds, for the clock variation of 60 ppm. In localization applications, this error in the clock is significant because it requires accuracy in the order of few

Figure 2.20: Error in time of a crystal for different clock skew values.

nanoseconds. Thus, there is a need to synchronize the clocks for precise ranging and localization.

2.4.1.2 Discrete Clock model for sensor nodes

We discussed the continuous time model in the above section. However, in digital hardware events are triggered by the rising or falling edge of the driving oscillator. The discrete clock model is shown in Fig.2.21. The discrete valued local clock

Figure 2.21: Discrete time clock model

counter t_A° is given by

$$t_A^\circ = \left\lceil \frac{t}{t_A} \right\rceil t_A \quad (2.40)$$

where, $\lceil \cdot \rceil$ symbolizes the ceiling function and $t_A = \frac{1}{f_A}$ is the nominal clock period of the oscillator. Nominal clock period is the same for both the nodes A and B (i.e. $t_A = t_B = t_N$). This means that in the local time, each of the nodes will count t_N seconds per clock period. However, as observed from the real time frame, the actual clock period of node B is given by $\frac{t_N}{\alpha_B}$, where $\alpha_B = 1 + \delta_B$ is the absolute clock skew. The discrete valued local clock counter t_B° is thus given by

$$t_B^\circ = \left\lceil \frac{\alpha_B t - \phi}{t_N} \right\rceil t_N + \hat{\theta}_B, \quad (2.41)$$

where the offsets are related as $\theta_B = \hat{\theta}_B - \phi$, $\hat{\theta}_B = \left\lceil \frac{\theta_B}{t_N} \right\rceil t_N$ is in the full clock periods and $\phi = \hat{\theta}_B - \theta_B$ is lower than one clock period. The use of the discrete time clocks allows the transmit node to transmit the signal in integer multiples of the rising or falling edge of the clock. Similarly, the signal received at the receiver will be captured at the integer multiple of the rising or falling edge of the clock. This leads to the rounding effects in the timestamps and which subsequently results in low accuracy of the timing measurements. We discuss the time rounding issues below.

Time Rounding Issues in Discrete Clock Model: In practice, clocks are

Figure 2.22: Rounding effect of clock

implemented as counters triggered by the rising or falling edge of an oscillator output signal. These counters are discrete values with the granularity of an oscillator period [98]. Hence, TOA recordings can only be performed with the granularity of an oscillator period [99]. We understand the time rounding issues with the help of illustration in Fig.2.22. Consider two nodes, a transmitter (tx) and a receiver (rx). tx transmits the signal at the rising edge of the clock and rx receives the signal at the rising edge after the propagation delay. The loss time as observed from Fig.2.22 can be given as

$$t_{loss} = \left\lceil \frac{t_{start}^{tx} + t_{pd,rx}^{tx}}{t_{clock}} \right\rceil t_{clock} - (t_{start}^{tx} + t_{pd,rx}^{tx}). \quad (2.42)$$

For example, assume the nominal pulse repetitive frequency of the oscillator to be 100 MHz ($t_{clock} = 0.01$ microsecond). If the start time of the transmitter is assumed to be $t_{start}^{tx} = 0.04$ microsecond and assuming the distance between the tx and rx to be 2 meters (equivalent to $t_{pd,rx}^{tx} = 6.67$ nanoSeconds). The calculated loss time using (2.42) is given by $t_{loss} = 3.33$ nanoseconds which corresponds to 1 m of error. Now, if the clock frequency is reduced to 1 GHz ($t_{clock} = 1$ nanosecond) i.e. clock with higher granularity corresponds to 10 cm of loss time error. In order to minimize the error to 3 cm, the clock oscillation frequency must be around 5 GHz. It is well known that UWB-IR is a promising approach for time based ranging schemes, however, it also suffers with clock rounding problems if implemented with low frequency crystal oscillators. Having a clock with high frequency is a costly approach for low-cost and low-power wireless sensor nodes. Therefore, in this thesis, we will demonstrate how we can utilize the low frequency crystal to obtain high precision ranging.

Some of the recent works which has adopted the discrete clock model for TOA measurements using costly time to digital converter (TDC) are given in [98]-[100]. Authors in [100] consider a discrete clock model and propose a round trip time measurement mechanism in a master-slave network. Authors in [101] present a mathematical modeling of discrete clocks and use them for synchronizing a master-slave network. It is noted that these works demonstrate the time measurements using very high resolution TDC's (in the order of sub-picoseconds) [98] which are very costly devices to be used in a low-cost wireless sensor nodes.

So far we have discussed some of the basic clock concepts. We now discuss some of the important work done related to clock synchronization. Synchronization can be done either through wired or wireless communication between the nodes. The

popular Network Time Protocol (NTP) and Precision Time Protocol (PTP) are used for synchronization of the wired networks. Due to the existence of many wireless technologies and increasing demand of wireless sensor networks, has resulted in various wireless synchronization protocols. Many wireless synchronization protocols has been proposed in synchronization literature. Based on the execution of synchronization mechanism, these protocols can be broadly divided into two categories: pairwise based and fully distributed based. In the former category, the synchronization is performed between a pair of nodes which is extended to many nodes by forming a hierarchical network structure either in the form of spanning tree or clusters. Some of the protocols which are included in the pairwise category are Time synchronization Protocols for Sensor Networks (TPSN) [102], Reference Broadcast Synchronization (RBS) [103], Flooding Time Synchronization Protocol (FTSP) [104], Lightweight Tree based Synchronization (LTS) [105], Tiny Sync [106], Pairwise Broadcast Synchronization (PBS) [107], Delay Measurement Time Synchronization [103] and Hierarchy Referencing Time Synchronization (HRTS) [108]. A detailed statistical modeling for using these protocols can be found in [109]. The main disadvantage of these approaches are the accumulation of synchronization error as the network size grows with the increase in the distance from the reference node. However, in case of fully distributed synchronization algorithms, nodes communicate with the neighboring nodes to help each other for synchronization. The main advantage of using a distributed approach is that they are robust to the dynamic environment and moreover they can be scaled to many nodes. The wireless synchronization algorithms can be further classified into two main categories:

1) **Packet coupling:** In packet coupling, a periodic exchange of packets is accompanied by the sender's local timestamp. The main sources of error in such techniques are the random delays present at different processing steps of a packet. The error source consists of the following components: a) Send time: It includes the message building time spent at the sender usually at the application layer, it also includes the time taken in transmitting the message to the network layer. b) Access time: Time spent in waiting to access the channel (at the MAC layer). One of the most significant component and highly variable. c) Transmission time: Message transmission time at the physical layer. d) Propagation time: Time spent over the channel. e) Reception time: Time required for receiving the message at the physical layer, same as transmission time. f) Receive time: Time to construct the received signal and send it to the application layer at the receiver. The synchronization accuracies

in the order of milliseconds to microseconds can be achieved in the packet based techniques. However, the disadvantages of using these techniques are that they are computationally complex, since large number of packets are exchanged. In addition to this, this method is energy consuming as well as it shows poor performance when the network is scaled. Some of the related works in this category includes average consensus principle based clock synchronization [110], [111], belief propagation [112], [113], and Kalman filtering approach [114], [115]. In consensus based methods message delays are not considered, whereas in belief propagation based methods only clock offsets are adjusted while discarding the clock skew. This requires frequent synchronization. Kalman filtering is used to track clock skew and clock offset [114], [115]. Moreover, fully distributed Kalman filtering based algorithm to track both clock skew and clock offset is proposed in [116]. In addition, recently graphical models based on message passing mechanism are also proposed to track clock offsets in the presence of exponential family distributed random variables [117], [118].

2) **Pulse coupling:** Physical layer scheme in which the local timing information is encoded with the transmission times of the given waveform. A node transmits its pulses as per its local clock. Other nodes in its vicinity can track or adjust their clock based on the signal received from the node. The advantage of this type of scheme is it is easily scalable. Moreover, this kind of coupling is less complex as it just requires simple base band level processing. Some of the works based on pulse coupling are discussed in [119]-[120].

In most of the techniques discussed in the literature for clock synchronization, a continuous clock model is assumed. The continuous clock model assumes very high time precision which can only be achieved using costly timing devices such as time to digital converter (TDC) or high resolution oscilloscopes. Moreover, in standard wireless hardware, clocks are implemented as discrete counters triggered by the rising edge of an oscillator output signal. The granularity obtained is equal to the granularity of an oscillator period [121]. Thus, TOA measurements can only be performed with the granularity of an oscillator period. Recently, authors in [100] demonstrated the time measurements technique using discrete clock model. They use UWB signal to measure the transmit and receiver time while using high resolution measurement time to digital converter device. In addition, authors also demonstrated the clock synchronization scheme using discrete clock behavior [122]. Moreover, a mathematical model using discrete clocks is proposed to determine the TOA measurements [101].

A discrete clock is an important consideration from implementation point of view. Unfortunately, there has been no work which demonstrate the use of discrete clocks and use it to determine the high precision localization system. In this thesis, we use the discrete clock model and demonstrates how it affects the measurement error. In the next section we briefly discuss about some of the implementation challenges related to clock synchronization.

2.4.2 Implementation Challenges

1. Fast Drifting: Due to the fast drifting of the clock, the offset only scheme is not much reliable in applications where high precision synchronization is required. The drifting of the clock makes the estimated offset out of date over a short period and thus, requires frequent synchronization. Frequent synchronization is a costly approach in wireless sensor networks. Therefore, it is preferable to jointly estimate the clock skew along with offset to ensure the long period synchronization.

2. Sensor Node Limitations: In literature, most of the mathematical analysis and simulation has been carried out by using tools such as MATLAB and/or other optimization tools. However, sensor nodes are constrained by limited processing capability with limited memory. Therefore, techniques which rely on estimating the parameters based on large sample of data may not be feasible in real time implementation of sensor nodes. This issue need to be tackled using practical evaluations.

2.5 Real Time Sampling Technique

The UWB signal has a large bandwidth. Due to the large bandwidth of the UWB signal, it is difficult and costly to sample the received signal. Hence, it is one of the most critical issue in the design of UWB based receivers. Sampling the UWB signal at or above the Nyquist rate requires high sampling frequency which is not feasible in low-cost and low-power wireless sensor nodes and thus restricts the practical implementation of applications based on UWB technology. In addition to this, a high sampling rate requires very powerful signal processing capabilities to process the large sampled data. To handle such a high processing requirement field programmable gate arrays (FPGA) and high speed digital signal processors (DSP) are needed which is again a costly solution for the implementation of UWB technology for low-cost low-power wireless sensor nodes. We can recover the UWB

Figure 2.23: Real time sampling

signal by performing real time sampling or equivalent time sampling. In real time sampling, ideally, the sampling method captures all the samples from a single pulse. The real time sampling is illustrated in Fig.2.23. In the figure, t_{start}^{tx} is the start time of the transmitted UWB pulse, $t_{cap,rx}$ is the rx clock offset and $t_{tx,rx}^{pd}$ is the propagation delay, t_{rx}^{cd} is the circuit delay when the received signal is passed through the receiver chain, and t_{rx} is the nominal sampling interval of the receiver clock with δ_{rx} as the variations in the clock, in parts per million and thus, the actual sampling interval is $\tilde{t}_{rx} = \frac{t_{rx}}{1+\delta_{rx}}$. Therefore, the total time of arrival, in discrete time, at the receiver is given as

$$\text{Total time of arrival} = \left\lfloor \frac{(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})(1 + \delta_{rx})}{t_{rx}} \right\rfloor \tilde{t}_{rx} \quad (2.43)$$

For very high sampling rate i.e. for very small sampling interval t_{rx} , eq. 2.43 can be approximated in continuous time as

$$\text{Total time of arrival} \approx (t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})(1 + \delta_{rx}) \quad (2.44)$$

This reduces to a conventional affine model of the clock which is given by

$$\begin{aligned} \text{Total time of arrival} &\approx (t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})\alpha_{rx} \\ &\approx \alpha_{rx}(t_{start}^{tx} + t_{tx,rx}^{pd} + t_{rx}^{cd}) + \beta_{rx} \\ &\approx \alpha_{rx}t + \beta_{rx} \end{aligned} \quad (2.45)$$

where $\beta_{rx} = -\alpha_{rx} \times t_{cap,rx}$ and $t = (t_{start}^{tx} + t_{tx,rx}^{pd} + t_{rx}^{cd})$. This takes the form as illustrated in Fig.2.19. In the next section, we discuss some of the basics of equivalent time sampling which is used in this thesis for precise recovery of time of arrival.

2.6 Equivalent Time Sampling Technique

Equivalent time sampling (ETS) is a technique in which a waveform is created over time by using a series of samples taken from repetitive waveforms. This technique is applicable only for periodic signals, i.e. it requires multiple occurrences of the same signal to reconstruct a waveform [123]. Hence, the signal must repeat identically each time or the recreated waveform will get distorted. ETS can help in achieving high sample rates. In ETS technique, samples are captured from many repetition of the identical pulses. The number of samples acquired from a single pulse can be one or many [123], [124]. The general idea behind equivalent time sampling is described through Fig.2.24. The equivalent-time digitizer acquires one sample per period signal. In the first period of the periodic signal, one sample is taken. For the second period, again another sample is taken, but the sampling begins after a very short, but well defined time delay. For the third period, the sampling is again delayed by equal amount of time, compared to the second period. This process is repeated many times, and after n signal periods, n samples are captured to recover the waveform of the signal. This is illustrated in Fig.2.24, where the samples acquired from all the periods are combined to recreate a complete waveform. Because the processing time of the samples does not affect the time resolution, the ETS method can provide a very high bandwidth and an extremely high virtual sample rate at a moderate

hardware effort.

Figure 2.24: Equivalent time sampling

2.6.1 Benefits of Equivalent time sampling over other schemes

Precise ranging and multi path resolution capacity of UWB-IR, due to its large bandwidth, is an advantage over other narrow band techniques. However, the large bandwidth of UWB-IR requires Nyquist sampling rate of several GHz. Sampling at such high rate is not feasible for low cost practical implementation [125], [126]. In literature, authors have considered Nyquist sampling rate to achieve high ranging accuracy. While some authors used analog correlators or energy detectors with sub-Nyquist sampling. Authors in [127] have employed energy detection technique while using sub-Nyquist sampling rate. For energy detection based schemes, the incoming or received signal is squared and integrated over a certain interval. TOA is estimated from the signal recovered where the energy is exceeded to certain threshold. The drawback of this technique is that it involves squaring of the signal, which results in squaring the noise and thus, causing the signal to noise ratio penalty [128]. Recently, authors in [129], [130] utilize compressive sensing technique to reduce the sampling rate of the receiver. These compressive sensing techniques work on the principle of sparsity of the data [131]. However, in most practical cases using UWB-IR, these

techniques do not perform well due to the high density of multi-path. For example, according to the IEEE 802.15.4a standard, the cluster arrival rate and the ray arrival rate of CM1 model is 0.0233 ns^{-1} and 2.5 ns^{-1} , respectively. This means we have 0.0233 clusters in 1 ns^{-1} or 1 cluster length is approximately 43 ns . Since each cluster in this channel model has rays, so 43 ns has approximately 107 distinct multi-path components as per the ray arrival rate of 2.5 ns^{-1} . We need a sampling rate of $f_s \geq 107/(43 * 10^{-9}) = 2.5 \text{ GHz}$. In this case the sparsity rate is 1(out of 1). To have the sparsity rate of 0.1 or channel to be sufficiently sparse, we need to sample the channel at the sampling rate of $f_s \geq 107/0.1/(43 * 10^{-9}) = 25 \text{ GHz}$. Therefore, to have the sparsity rate of 0.1 or less we need to sample the channel at sampling rate of 25 GHz or more, which is practically infeasible for low cost systems. So, the use of compressed sensing for an indoor environment will not provide efficient solution with UWB-IR signals. Moreover, in [132], authors have designed a dictionary of parameterized waveform so that sparse representation of the received signal can be presumed. Using this approach along with compressed sensing approach can help in reducing the sampling rate as low as 1/3 of the Nyquist rate. However, the reduction is substantial but this is still an expensive solution for applications involving low-cost wireless sensor nodes. In [133], authors formulated a harmonic retrieval problem to get the channel impulse response which can achieve up to 1/8 of the sampling rate reduction. However, harmonic retrieval problem is blind to the circular shifts and hence it is unable to estimate the timing offsets. Equivalent time sampling technique proves to be a feasible solution for practical implementation of precise indoor localization solution using low-cost wireless sensor nodes with low speed signal processors. Moreover, different from earlier mentioned energy detector schemes, the frame samples obtained by equivalent time sampling scheme can be averaged to further suppress the noise. Therefore, this scheme can enjoy the benefit of an increased signal to noise ratio. Some of the disadvantages of equivalent time sampling are: 1) Long preamble to recover the signal at the receiver. This causes lower update rate. 2) The target needs to be in the same position for many periods.

2.7 State-of-the-art

Many research groups have utilized technologies such as Wi-Fi, ZigBee, UHF etc. for indoor localization [134]. However, all these technologies are restricted by their

inability to resolve multi-paths in the indoor environment. In such scenario, UWB-IR is a promising technology providing high accuracy with fine time resolution [135].

Despite many theoretical results and analysis on UWB ranging and localization, less work is devoted to the practical implementation of the protocols. A few research groups have implemented some of the prototypes and detailed the measurement results. A detailed comparative study of the works which uses UWB as a wireless technology along with their architecture is presented in Table.2.4. Due to the high bandwidth of the UWB signal, high sampling rate is required to precisely recover the signal. Thus, it requires high sampling devices at the receiver. Authors in [79], [136] use high speed oscilloscopes to receive the signal for the proposed TDOA based systems. In [137], FPGA is used to trigger the receiver oscilloscope used for signal recovery. The authors in [138] demonstrated the round trip time ranging mechanism using UWB. The received signal is sampled at the sampling rate of 41.3 ps to obtain the precise range measurement. Similarly, authors in [139], demonstrated a precise UWB based system for warehouse application where the presence of racks affects the LOS. They obtain the blocking distribution and use them to improve the accuracy of the system. They demonstrated the system using vector network analyzer (VNA) to obtain the frequency response of the UWB channel. Authors in [99] have demonstrated a real-world prototype for node synchronization and ranging using UWB-IR. The prototype system uses an equivalent time sampling oscilloscope to capture the channel impulse response (CIR) from several consecutive pulses transmitted from the transmitter. CIR is used to estimate TOA which is eventually used for clock synchronization. Several authors have suggested mechanisms to obtain the high precision TOA [109]. They either use high end oscilloscopes or costly devices like time to digital converters (TDC) to extract the precise timing information. Recently, a precise range measurement scheme using UWB has been proposed in [140] which resolves the timing information up-to the resolution of few pico-seconds by using costly pico-second resolution TDC. Same authors in [122] propose a precise time synchronization scheme between the master and slave nodes using high resolution timing devices. Authors in [81] have practically realized the TDOA based localization system where all the target nodes receives and re-transmits the UWB signal. This avoids the need of synchronized anchor nodes as needed for TDOA architecture, however the receive and re-transmit architecture, while using UWB-IR, at the target node makes the target node consume more energy. Furthermore, this system is based on energy detectors to recover the signal, which generally gives lower

accuracy as compared to the high sampling rate architectures. In addition, systems based on energy detector is comparatively less efficient in cases when signal is affected by multi-path fading. Moreover, very few results, utilizing the signals with high bandwidth, have been verified experimentally in the literature. They either use high end oscilloscopes, costly analog to digital converters (ADC) or costly devices like time to digital converters (TDC) to extract the precise timing information [81], [137], [141], [142], [143].

With high speed oscilloscope and high sampling rate devices, although it is easier to identify the critical timing of the input signal, it is not suitable for commercialization and for practical TOA based localization system. Thus far, researchers have developed many techniques to reduce the sampling rate of the system. In [144], authors use an ETS technique to reduce the ADC sampling rate. [144] mainly focus on designing the architecture for precise positioning while considering a synchronized system. They did not focus on different synchronization aspects while using ETS technique. In [145], authors use ETS technique to estimate the CIR using the bounded data uncertainty model. They consider a model under the assumption that the receiver has perfect knowledge of the time instant of the first arrival of the first pulse in the pulse train. Additionally, under the same assumption, authors in [146] further proposed a sub-Nyquist sampling rate-based mechanism to estimate accurate TOA of the signal. In [146], the focus is on estimating the clock drift due to the longer preamble used in ETS technique. The clock drift is estimated by observing the transition between frame phases due to the accumulated drift. Based on the observed transition path, drift is estimated, then calibrated and the accurate TOA of the signal is obtained. However, the algorithm proposed may not be feasible for real time practical environments involving unsynchronized wireless sensor nodes.

In addition to the above, obtaining ranges between the nodes from unsynchronized TOA is not straightforward [147]. The main parameters that corrupt conventional one-way TOA measurements are receiver clock offset, receiver clock drift and target node start time [148]. The corruption is worse in techniques using ETS due to the longer transmission time and additional variations due to transmitter clock drift. A strict synchronization mechanism is required to synchronize the nodes and obtain precise ranging, which is challenging and costly for practical implementation. To avoid stringent clock synchronization requirements two-way ranging is commonly used in the literature [100], [149]. Recently, authors in [100], [101] propose a round trip time measurement mechanism in a master-slave network using UWB-IR with

costly time to digital converters (TDC). However, two-way ranging may not be feasible for low-power, low-cost *transmit only* target nodes. In these cases, time difference of arrival (TDOA) technique can be used to obtain the range difference between the target and anchors [150]. However, these techniques require anchors to be synchronized to a common reference clock [79], [136], [137]. Providing a common clock may not be feasible for *easy-to-deploy* wireless anchors. To avoid this situation, an approach utilizing difference of TDOAs, i.e. differential time difference of arrival (DTDOA), can be used [87], [88]. Authors in [91], [92], [93], [94] utilize this mechanism to propose different localization schemes. However, it is important to highlight that unlike conventional DTDOA, in E-DTDOA there is an additional transmitter clock drift term arising due to the ETS. This additional clock drift introduces a significant error in the accuracy of ranging, as demonstrated in this paper. We highlight that most of the results published in literature [79], [151] and [152] use costly equipment such as high-speed oscilloscope, field programmable gate array (FPGA), and TDC at the receiver, which gives the opportunity to precisely determine the pulse's TOA.

Moreover, using UWB-IR is an expensive solution in terms of energy efficiency if both the source node and anchor nodes are transceivers, since the reception of UWB-IR signals consumes more energy than transmission [153]. Therefore, a hybrid scheme is required where communication can be achieved using different wireless technology while UWB can only be used for ranging. In [154], UWB is used along with infrared technology for robot tracking. A target robot is equipped with a UWB tag with UWB anchor nodes installed at the known position. A TDOA approach is used to find the position of the robot. However, FPGA is used to process the large bandwidth UWB signal. Thus, a hybrid approach is utilized in this work where sub-1 GHz [155] is used for command and control purposes, whereas UWB IR is used in transmit-only mode for ranging purposes. The sub-1 GHz is used over 2.5 GHz technology due to its advantages like longer range capability, lower cost, and reliable and robust link connectivity.

Many commercial and experimental systems are available in the market which use UWB technology for ranging and positioning. Some of these systems are based on IEEE 802.15.4a [156], [157], whereas other use proprietary technology [158]. DWM1000 is a commercially available UWB radio communication chip provided by Decawave. Commercially available IEEE 802.15.4a compliant UWB transceivers DWM1000 are used by authors in [156], [157]. The chip is based on IEEE 802.15.4a

standard. The chip utilizes TOA/TDOA for ranging purpose. UWB is used for both ranging as well as for medium access control. The ranging accuracy of less than 10 cm is reported using this system. Authors in [158], uses a commercial UWB localization system from Ubisense to evaluate the localization performance of the system. The Ubisense can attain a positioning accuracy of few cm. Moreover, Ubisense also provides real time localization system (RTLS). This system follows proprietary technology for positioning and medium access control. Ubisense uses UWB for positioning purpose whereas it uses 2.4 GHz ISM band for the purpose of medium access control. The work in [159] also uses 2.4 GHz band for controlling the tags while implementing time division multiple access scheme. In addition, the UWB real-time location system (RTLS) offered by Zebra Enterprise Solutions integrates with other RTLS which utilize technologies such as global positioning system (GPS), RFID, and WLAN [160]. To this end, in Fig.2.25, we provide the comparisons of some of the commercial off the shelf UWB platforms along with the hardware used in this thesis.

Comparisons with some Real time sampling (RTS) based system

In this section, we have compared the proposed equivalent time sampling based system with state of the art systems based on real time sampling. For comparison, we consider works which use UWB as a signaling technique. The results of the experiment performed in the previous works are summarized in Table.2.4. Most of these systems occupy large bandwidth (>2 GHz), where either a high-speed oscilloscope was often used as a receiver, or receiver employing high sampling rate ADC was used. This gives an opportunity to precisely determine the pulses' time of arrival (peak position) at the receiver. Specifically, in [142], a non-coherent energy detector based receiver with 20 GSps sampling rate was used with all the anchor nodes synchronized. In [151], [161], a non-coherent UWB system with two 8 bit-1.5 Gsps ADCs were used to sample the signal and a Virtex4 Xilinx FPGA was used to process the sampled data. Similarly, in [152] a real time sampling receiver with ADC sampling rate at 2.5 GSps was used along with the FPGA. Moreover, in [162], a receiver with 8 bit-2.88 GS/s real time sampling ADC was used along with FPGA. In addition to the above mentioned systems which uses high sampling ADC's, in [79] and [135] high speed 4 channel real time oscilloscope was used as a receiver to determine the time of arrival of the signal. Furthermore, [163] uses Lecroy SDA 816 Zi real time oscilloscope to recover the UWB signal. All the presented systems are based on real time sampling with GHz sampling frequency, which uses costly equip-

Figure 2.25: Comparison of commercially available UWB platforms

ment's to obtain precise time of arrival (TOA) while using high bandwidth UWB signal. The accuracy achieved by these systems are mentioned in Table. 2.4. However, the system developed in this thesis is equipped only with a low cost ADC with sampling frequency between 2-3 MHz while still achieving sub-nanosecond resolution accuracy. Considering the localization performance, our developed system is significantly better than the ones found in the literature, with the key advantage of having a fully integrated receiver and a processor instead of using a high-speed oscilloscope and high sampling rate ADC. Moreover, a comparable localization performance has been achieved using asynchronous nodes, whereas the other listed systems assume synchronous nodes.

Table 2.4: Comparison of Ranging Mechanism

Ref	Method	Transmitter	Receiver	Synchronization	Localization Error	Test Area	Remarks
[164]	TDOA	2 nS pulse	-	Synchronized	18 cm	8x8 m ²	Static Scenario
[165]	TDOA	300 pS pulse 2 GHz BW	-	Synchronized	< 15cm	10x10m ²	Multi cell
[162]	TDOA	-	LNA + ADC 8 bit 2.88 GS/s real time sampling + FPGA	Synchronized	< 30 cm	6x8m ²	Static Scenario
[158]	TDOA	Pulse Gen. 2.5 GHz BW	Commercial product from Ubisense	Synchronized	13 cm	3x3m ²	Moving robot
[161]	TDOA	2 nS pulse $f_c = 3.5$ GHz	Non-coherent UWB system, ADC 8 bits 1.5Gsps, Virtex4 Xilinx FPGA	Synchronized	20 cm	5x8m ²	Mobile Robot
[31]	TDOA	300 pS pulse	Envelope detection + I/Q signals are sub sampled using UWB sub sampling mixer	Synchronized	RMSE (3-D) 5.24 mm	50x150x 250mm ³	Mobile Robot
[135]	TDOA	1 nS pulse $f_c = 4$ GHz	Envelope detector and oscilloscope	Synchronized	8cm RMS	4.8x6m ²	Static Scenario
[79]	TDOA	1 nS pulse $f_c = 4$ GHz	LNA + High speed 4 channel oscilloscope as a receiver	Synchronized	8 cm RMS	12x10m ²	Static Scenario
[142]	TDOA	1.5 nS pulse	Non-coherent Energy Detector + ADC Sampling rate at 20GS/s and 2 GS/s	Synchronized	10 cm	6x6m ²	-

Ref	Method	Transmitter	Receiver	Synchronization	Localization Error	Test Area	Remarks
[137]	TOA, 2D, RTT	1 nS pulse	Oscilloscope + XC3S50AN FPGA	Synchronized	mean= 7.40 cm	1.8x2.4m ²	Static Scenario
[151]	TDOA	3 nS pulse $f_c = 3.5$ GHz	Non-coherent UWB system, ADC 8 bits 1.5Gsps and a Virtex4 Xilinx FPGA	Synchronized	sub 20 cm	6x8m ²	Mobile Robot
[163]	TDOA	240 pS pulse 2-11 GHz BW	Lecroy SDA 816 Zi real time oscilloscope	Synchronized	10 cm	4x19m ²	Static
[137]	TOA, 2D, RTT	1 nS pulse	Oscilloscope, XC3S50AN FPGA [Use FPGA to sequentially trigger four transmitters to emit pulses and provide trigger to the receiver which is an oscilloscope]	Synchronized	mean= 7.40 cm	1.8x2.4m ²	Static Scenario
[166]	TDOA	Pulse generator	Non coherent Energy detection, LNA +Correlator+ Comparator+TDC with 27pS resolution	Synchronized	3.6 cm average	6.3x5.9x3 m ³	-
[167]	TOA, 2D, RTT	1 nS pulse $f_c = 5.6$ GHz	Squarer + Comparator, TDC with 50 ps resolution	No Synchronization	mean error = 5.303 cm	1.1x2.1m ²	Static Scenario

Ref	Method	Transmitter	Receiver	Synchronization	Localization Error	Test Area	Remarks
[168]	TOA, RTT	300 pS pulse 0.3-1 GHz BW	Energy detection + TDC with 65 pS resolution + ML505 Virtex-5 FPGA	-	13 cm RMSE	12x12m ²	Static Scenario
[99]	TOA	-	FPGA (Stratix II EP2S60F672C3 Altera) to implement the TPU and the ABC	mean = 10-15 nS	-	-	-
[92]	D- TDOA	-	-	Unsynchronized	< 30 cm	9x9x4 m ³	Static Scenario
[146]	TOA	Simulation Only	Equivalent time sampling with no experimental Demonstration	5-50 nS for different SNR and thresholds	-	-	Drfit compen- sation
[145]	-	Simulation Only	Equivalent time sampling	-	-	-	Channel Estima- tion
[169]	2 way Rang- ing	DecaWave EVK1000	DecaWave EVK1000	Synchronized	LOS= 11 cm	5.3x11.5m ²	-
[143]	TDOA	250 pS pulse 7 GHz BW	Squarer + Comparator + TDC with 25 ps resolution	Synchronized	mean(2D): 3.4 cm; mean(3D): 11.4 cm; RMS(2D): 7.3 cm; RMS(3D) 18.5 cm	9.5x10m ²	Static Scenario
[152]	TOA	Gigabit Transceivers 1.2 nS pulse 800 MHz BW	Real time sampling with ADC at 2.5 GSps + FPGA	5 nSec	-	-	-

Ref	Method	Transmitter	Receiver	Synchronization	Localization Error	Test Area	Remarks
[100]	2 way RTT	300 pS pulse 0.3-1 GHz BW	Energy detection + TDC with 65 pS resolution + ML505 Virtex-5 FPGA	< 1 Hz for frequency estimation; 1 ns for phase estimation	Ranging between 2 nodes, 17 cm RMSE	1 m to 4 m distance	Two nodes
[156]	2 way ranging	DecaWave DWM1000	DecaWave DWM1000	Synchronized	10 cm 2D; 20 cm 3D	-	Moving robot
[122]	2 way ranging	-	Real time using TDC with 100 pS resolution	sub-nanosecond	-	-	-
[81]	TDOA	995 ps pulse	Energy detection + TDC with 50 pS RMS resolution; receive and re-transmit	Unsynchronized	20.7m; 15.2m; 21.3m	LOS 8X8; LOS 6x6; NLOS 5X5	Static Scenario
This work	E-DTDOA	1 nS pulse, BW = 2 GHz, $f_c = 4$ GHz	Correlator + Equivalent time sampling ADC with sampling freq approx 3 MHz + ARM Cortex A8 microprocessor on BeagleBone Black (BBB)	Unsynchronized	ranging < 30 cm	10 m	Static Scenario

3 A Novel E-DTDOA Based One-way Ranging Using UWB-IR with Unsynchronized Anchors

3.1 Introduction

Processing of channel impulse response (CIR) for the precise TOA of the direct path is the basis for accurate positioning in indoor cluttered environments [135], [170]-[172]. This requires a receiver which is capable of measuring the CIR with high time resolution in sub-nS as well as large bandwidth signals, such as ultra wide-band impulse radio (UWB-IR), for efficient separation of multi-paths. However, the Nyquist sampling rate of several GHz is required at the receiver to recover such large bandwidth signals. Sampling at such a rate is an expensive solution and thus practically limited by cost and computational complexity of the required hardware [79], [126], [136], [137], [173], [174]. Many techniques have been developed in the past to reduce the sampling rate [132], [133], [175], [176]. Despite many techniques available, it is challenging to achieve sub-ns resolution required for centimeter level ranging accuracy in low-cost, low-power sensor nodes [79], [135]- [137]. Equivalent time sampling (ETS) is a promising mechanism to obtain precise CIR in the order of sub-nS at much reduced analog to digital converter (ADC) sampling rate at the cost of repetitive signal transmission [124]. It is important to highlight that CIR obtained from ETS mechanism distinguish itself from real time sampling (RTS). In RTS, a single transmitted pulse is require to recover a single frame capture, whereas many transmitted pulses are require to recover a single frame using ETS mechanism. Thus, the measured TOA in case of RTS is mainly affected by the imperfections in the receiver clock (i.e. clock drift), while in case of ETS it is affected by the imperfections of transmitter and receiver clock drift. To the best of authors knowledge, there is no work which presents an analytical model that can capture the complete ETS process

and applicable to different clock synchronization system architectures. In addition, the practical implementation and analysis of work utilizing the ETS technique for ranging is missing in the literature.

In this chapter, we first propose a set of analytical equations to determine sub-ns resolution equivalent time of arrival (E-TOA) for an ETS based system. The timing information is extracted from the CIR of the recovered equivalent time sampled signal in an equivalent time frame, and, hence the name E-TOA. The effect of the transmitter and receiver variation is reflected in the proposed E-TOA equation and also in ranging errors, as will be discussed in the chapter. In addition to this, keeping the strict synchronization of ranging systems in mind and compared with other time based ranging schemes such as time difference of arrival (TDOA), which often use customized hardware and software to reduce the error in synchronization and TDOA measurement, in this chapter, we follow the principle of differential time difference of arrival (DTDOA), and further propose an equivalent differential time difference of arrival (E-DTDOA) ranging scheme. The main contributions of the chapter are highlighted as: 1) We propose an analytical equation to determine E-TOA of an ETS based system. It is shown that the equation is sensitive to the variation in transmitter clock drift and thus, different from conventional RTS TOA equations. The proposed equation is validated using numerical simulations and experimental demonstrations. 2) Using these E-TOA measurements, a practical low cost one-way E-DTDOA based ranging scheme is further proposed and demonstrated for unsynchronized anchors. 3) High time resolution, in the order of sub-ns, is achieved from low sampling rate ADC (as low as in the order of few MHz) through repetitive sampling. Receiver nodes are designed using low cost off the shelf ADC's with the sampling rate of around 3 MHz, unlike high sampling rate and costly oscilloscopes.

The rest of the chapter is organized as follows: Sec. 3.2 details the system overview, the basic operation involved for ETS and the proposed E-TOA equations. Sec. 3.3 discusses the verification of the proposed equation along with the sensitivity analysis using numerical simulations. Sec. 3.4 validates the proposed equation experimentally. Sec. 3.5 further discusses the proposed ranging mechanism using E-TOA equations along with simulation and experimental results. The chapter is concluded in Sec. 3.6.

Notations: The variables indicated in small letters denote the reference time (nominal value). The variable in small letter with $(\bullet)$ denotes the skewed time with respect to the reference time (referred as actual value). The variables indicated in capital letter denote the equivalent time having both nominal and actual values. The superscript and the subscript on the variables denote either transmitter (tx) and receiver (rx). For better readability, the comprehensive list of notations used in the below sections are given in Table.3.1.

3.2 System Overview

This section details the basic operation of an ETS technique and the proposed analytical equations for determining the E-TOA in an ETS based system.

3.2.1 Basic Operation

Consider a transmitting node tx and a receiving node rx. Node tx transmits wide bandwidth ranging¹ signal at each rising/falling edge of the digital clock running at tx. The edge of the clock repeats at an actual pulse repetition interval (PRI), $\tilde{t}_{tx} = \frac{1}{\tilde{f}_{tx}}$, where $\tilde{f}_{tx}$ is the actual pulse repetition frequency (PRF). Node rx, on other hand, captures the transmitted signal at each digital tick of the ADC clock, which operates at an actual sampling interval of $\tilde{t}_{rx} = \frac{1}{\tilde{f}_{rx}}$, where $\tilde{f}_{rx}$ is the ADC sampling frequency. Due to the wide bandwidth of the signal, it is difficult and costly to sample the received signal in real time with current ADC technology. Thus, ETS technique is used to recover the precise CIR of the signal [124]. The ETS technique requires the transmitter to transmit the same pulse repetitively. At the receiver, only a few interleaved points are sampled from each transmitted pulse. The samples are then properly combined in order to reconstruct a single equivalent time pulse, which is equivalent to that sampled at a high rate. Fig.3.1 illustrates the ETS mechanism. Mathematically, assume that n_s points are taken from each transmitted pulse, and the actual equivalent sampling interval is $\tilde{T}_{eq,rx}^{tx}$, then the actual ADC sampling frequency, $\tilde{f}_{rx}$, is given by

¹The fundamental requirement of the equivalent time sampling technique is the repetitive transmission of an identical pulses. Therefore, the scheme is applicable, in general for all the systems employing equivalent time sampling technique. However, in this work, for demonstration, we are using UWB-IR as the transmission signal with 2 GHz bandwidth ranging between 3 - 5 GHz with 4 GHz as the center frequency.

Table 3.1: List of Notations Used

Notations	Definition
$t_{tx} = \frac{1}{f_{tx}}$	Nominal PRI of the tx transmitter clock
$t_{rx} = \frac{1}{f_{rx}}$	Nominal ADC sampling interval of the rx receiver clock
$T_{eq,rx}^{tx} = (t_{rx} - t_{tx})$	Nominal equivalent time resolution between tx node and the rx node
$\alpha_{tx} = 1 + \delta_{tx}$	δ_{tx} is tx clock drift in ppm
$\tilde{t}_{tx} = \frac{1}{\tilde{f}_{tx}} = \frac{t_{tx}}{\alpha_{tx}}$	Actual PRI of the tx clock
$\alpha_{rx} = 1 + \delta_{rx}$	δ_{rx} is rx receiver clock drift in parts per million (ppm)
$\tilde{t}_{rx} = \frac{1}{\tilde{f}_{rx}} = \frac{t_{rx}}{\alpha_{rx}}$	Actual ADC sampling interval of the rx receiver clock
$\tilde{T}_{eq,rx}^{tx} = \left(\frac{t_{rx}}{\alpha_{rx}} - \frac{t_{tx}}{\alpha_{tx}} \right)$	Actual equivalent time resolution between tx node and the rx node
t_{start}^{tx}	Start time of the tx transmitter's first UWB-IR pulse
$t_{cap,rx}$	Clock offset at the rx receiver clock or start of the ADC sample capture window
$t_{tx,rx}^{pd}$	Propagation delay from tx node to rx node
t_{rx}^{cd}	Circuit delay at the rx receiver
$\hat{K}_{E-TOA}$	Equivalent time samples corresponding to equivalent time of arrival (E-TOA)
Ψ_{tx}	Equivalent time error due to the tx transmitter clock drift
Ψ_{rx}	Equivalent time error due to the rx receiver clock drift
$\Psi_{rx}^{tx} = \Psi_{rx} - \Psi_{tx}$	Equivalent time error due to the transmitter and receiver clock drift

$$\tilde{f}_{rx} = \frac{1}{\tilde{t}_{rx}} = \frac{n_s}{\tilde{T}_{eq,rx}^{tx} + \tilde{t}_{tx}}, \quad (3.1)$$

The ETS mechanism follows a relation $\tilde{t}_{rx} > \tilde{t}_{tx}$ and $\tilde{t}_{rx} \gg \tilde{T}_{eq,rx}^{tx}$. The condition to be satisfied to ensure $\tilde{t}_{rx} > \tilde{t}_{tx}$ is $(|\delta_{rx}| - |\delta_{tx}|) < (\approx \frac{\tilde{T}_{eq,rx}^{tx}}{\tilde{t}_{tx}})$ i.e. the difference between the clock drift of tx and rx must be less than the ratio of the equivalent time resolution and the nominal tx PRF. Fig.3.1 illustrates the case where one point is taken from each received pulse (i.e. $n_s = 1$)² to construct equivalent time pulse/s at the receiver. From (3.1), this technique reduces the ADC rate to the order of pulsing rate at the cost of repetitive signal transmission. The actual number of

²Throughout in this thesis, we consider the case when one point is taken from each received pulse to reconstruct the equivalent time sampled pulse i.e. we consider $n_s = 1$ in (3.1).

Figure 3.1: Illustration of equivalent time sampling

transmitted pulses ($\tilde{F}_{rx}^{tx}$) at the transmitter to reconstruct a single equivalent pulse at the receiver is given by

$$\tilde{F}_{rx}^{tx} = \frac{\tilde{t}_{tx}}{\tilde{T}_{eq,rx}^{tx}}. \quad (3.2)$$

From (3.2), for a given $\tilde{t}_{tx}$, the number of repetitive pulses at the transmitter $\tilde{F}_{rx}^{tx}$ is inversely proportional to $\tilde{T}_{eq,rx}^{tx}$. To reduce the total number of repetitive transmissions, it is desirable to increase n_s and $\tilde{T}_{eq,rx}^{tx}$. From (3.1), increasing n_s is equivalent to increasing the ADC rate. Hence the value of n_s is mainly constrained by the available ADC rate. The maximum equivalent sampling interval, $\tilde{T}_{eq,rx}^{tx}$, should satisfy the Nyquist-Shannon sampling theorem such that $\frac{1}{\tilde{T}_{eq,rx}^{tx}} \geq 2f_{bw}$, where f_{bw} is the bandwidth of the signal. For example, if $f_{bw} = 1 \text{ GHz}$, $\tilde{T}_{eq,rx}^{tx}$ is taken to be less than 0.5 ns. In other words, the equivalent sampling rate at the receiver is 2 Giga samples per second (GS/s). The above description forms the basis to recover CIR of the signal with much reduced ADC sampling rate. To achieve a certain nominal ETS resolution ($T_{eq,rx}^{tx}$), the nominal tx PRF and the nominal ADC sampling frequency can be set as per the relation in (3.1). For example, to achieve the equivalent time resolution of 0.11 nanosecond, we adjust the tx PRF as $f_{tx} = 3 \text{ MHz}$ and rx

nominal ADC sampling frequency $f_{rx} = 2.9990103 \text{ MHz}$. The illustration of one such captured frame is shown in Fig.3.2.

Figure 3.2: Equivalent time sampled frame at the receiver obtained by sampling multiple transmitted frames.

3.2.2 Assumptions

This section summarizes the assumptions required to follow the mechanism proposed in this work. The following assumptions are thus required:

- 1) Channel remains static during the measurement period during a frame period.
- 2) That nominal values of the transmitter and receiver clock frequencies or the nominal periods t_{tx} and t_{rx} , are typically available at the receiver.
- 3) The nominal periods t_{tx} and t_{rx} can be selected based on the frequency tolerance specifications provided by the manufacturer and the desired nominal equivalent time sampling interval.
- 4) However, the selection of these values are restricted to certain conditions i.e. the maximum actual equivalent sampling interval, $\tilde{T}_{eq,rx}^{tx}$, should satisfy the Nyquist-Shannon sampling theorem such that $\frac{1}{\tilde{T}_{eq,rx}^{tx}} \geq 2f_{bw}$, where f_{bw} is the bandwidth of the signal. In addition, the ETS mechanism follows a relation $\tilde{t}_{rx} > \tilde{t}_{tx}$ and $\tilde{t}_{rx} \gg \tilde{T}_{eq,rx}^{tx}$.

The condition to be satisfied to ensure $\tilde{t}_{rx} > \tilde{t}_{tx}$ is $(|\delta_{rx}| - |\delta_{tx}|) \lesssim \frac{\tilde{T}_{eq,rx}^{tx}}{\tilde{t}_{tx}}$ i.e. the difference between the clock drift of tx and rx must be less than the ratio of the equivalent time resolution and the nominal tx PRF.

5) As previously illustrated in Fig. 3.1, while performing equivalent time sampling technique, we are taking only one sample (i.e. we consider $n_s = 1$ in (3.1)) per received pulse in order to reconstruct the equivalent time sampled signal at the receiver.

3.2.3 Signal Model

The following sections describe signal modeling and the proposed E-TOA equation for UWB-IR based system employing ETS. At node tx, a train of UWB-IR signal, $s(t)$, can be generated by convolving a higher order derivative Gaussian pulse shape, $g_{tx}(t)$, with a delta function having actual PRI of $\tilde{t}_{tx}$ i.e.

$$s(t) = g_{tx}(t) * \sum_{n=0}^{\lfloor \frac{t_{trans}}{\tilde{t}_{tx}} \rfloor} a_n \delta(t - t_{start}^{tx} - n\tilde{t}_{tx}), \quad (3.3)$$

where, a_n can hold values $\{-1, 0, 1\}$, t_{trans} is the total transmission time of the transmitter. The notations t_{start}^{tx} , $\tilde{t}_{tx}$ in (3.3) are defined in Table 3.1. The transmitted pulse sequence, $s(t)$, propagates through the L -path fading channel with impulse response $h(t) = \sum_{\ell=1}^L \alpha_\ell \delta(t - \tau_\ell)$, where α_ℓ is the channel gain and τ_ℓ is the channel delays with $\tau_1 (= t_{tx,rx}^{pd})$ denotes the TOA of the first ray or the propagation delay. The received signal at the receiver is the result of the convolution between the transmitted sequence and the channel response and thus given by

$$r(t) = s(t) * h(t), \quad (3.4)$$

substituting (3.3) in (3.4), we get

$$r(t) = g_{tx}(t) * h(t) * \sum_{n=0}^{\lfloor \frac{t_{trans}}{\tilde{t}_{tx}} \rfloor} a_n \delta(t - t_{start}^{tx} - n\tilde{t}_{tx}), \quad (3.5)$$

The received signal $r(t)$ is passed through the receiver chain with circuit delay t_{rx}^{cd} and thus (3.5) can also be represented as

$$r(t) = \sum_{n=0}^{\lfloor \frac{t_{trans}}{\tilde{t}_{tx}} \rfloor} \sum_{\ell=1}^L a_n \alpha_{\ell} g_{tx}(t - t_{start}^{tx} - \tau_{\ell} - t_{rx}^{cd} - n\tilde{t}_{tx}), \quad (3.6)$$

$r(t)$ is then passed through the ADC where an ETS is performed with actual sampling period of $\tilde{t}_{rx}$. ETS can be modeled by multiplying the received signal $r(t)$ with a delta function having time period $\tilde{t}_{rx}$, as below

$$y(t) = r(t) \times \sum_{m=0}^M \delta(t - t_{cap,rx} - m\tilde{t}_{rx}), \quad (3.7)$$

where the notations $t_{cap,rx}$, $\tilde{t}_{rx}$ in (3.7) are defined in Table 3.1.

Fig.3.3 illustrates the tx and rx clock mechanism. It can be seen from Fig.3.3 that ADC starts capturing the received signal at $t_{cap,rx}$, which is much before t_{start}^{tx} . Therefore, the captured signal consists of two parts: i) samples collected before the reception of the first pulse at the receiver where only noise (σ^2) is sampled denoted by Term A in Fig.3.3, and ii) samples after the first pulse is received, denoted by Term B, and which are obtained by substituting (3.6) into (3.7). The captured signal is thus represented by

$$y(m\tilde{t}_{rx}) = \begin{cases} \sigma^2; & \text{for } m = m' \leq \left\lfloor \frac{t_{rx}^{cd} - t_{cap,rx} + t_{start}^{tx} + \tau_{\ell}}{\tilde{t}_{rx}} \right\rfloor \\ \sum_{m=m'+1}^M \sum_{n=0}^{\lfloor \frac{t_{trans}}{\tilde{t}_{tx}} \rfloor} \sum_{\ell=1}^L a_n \alpha_{\ell} g_{tx}(v) + \tilde{\sigma}^2; & \text{otherwise} \end{cases} \quad (3.8)$$

where $v = (t_{cap,rx} + m\tilde{t}_{rx} - t_{start}^{tx} - \tau_{\ell} - t_{rx}^{cd} - n\tilde{t}_{tx})$, l' is the index of the sample captured at $m' + 1$ clock sample, $\tilde{\sigma}^2$ is the noise term which is different from before the UWB tx signal arrive at the receiver due to changes in the gain of the receiver. For the ranging signal, we consider $a_n = 1$. We also consider the ADC sampling time to be sufficiently larger than the total transmission time to receive all the transmitted pulses i.e $M\tilde{t}_{rx} > \lfloor \frac{t_{trans}}{\tilde{t}_{tx}} \rfloor \tilde{t}_{tx}$.

The samples captured in (3.8) are combined to reconstruct the equivalent time pulse/s and, the time frame changes from real time to equivalent time with actual equivalent time resolution of $\tilde{T}_{eq,rx}^{tx}$. The illustration of the recovered equivalent time

Figure 3.3: Illustration to show the clock mechanism. Term A denotes the integer number of clock cycles and Term B is the remainder term. Term B is slided in the equivalent time sampling.

sampled frame³ with p pulses is shown in Fig.3.4, where $p = 0, 1, 2, 3, \dots, P$ is the index of recovered equivalent time pulse obtained after ETS.

3.2.4 Equivalent Time of Arrival (E-TOA)

It is observed from Fig.3.3 and (3.8), that the total time of arrival of the first pulse at the receiver rx is given by $\left(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd}\right)$ after replacing $\tau_\ell = \tau_1 = t_{tx,rx}^{pd}$ in (3.8). This total time can also be represented by an integer number of the receiver clock cycles, m' (denoted by Term A in Fig.3.3) plus a remainder term, modeled by a modulus operation (denoted by Term B in Fig.3.3). This is mathematically expressed as

$$\begin{aligned} \left(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd} \right) = & \left[\left(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd} \right) / \tilde{t}_{rx} \right] \tilde{t}_{rx} \\ & + \left[\text{mod} \left(\left(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd} \right), \tilde{t}_{rx} \right) \right], \quad (3.9) \end{aligned}$$

³It is to be noted that recovered equivalent time frame consists of many equivalent time pulses. For example, Fig.3.4 shows 2 such pulses in a frame.

Figure 3.4: Captured equivalent time sampled frame with two equivalent time pulses

where modulus function in (3.9) is defined as $a = kb + \text{mod}(a, b)$; where $k \in \mathbb{Z}$ and $0 \leq \text{mod}(a, b) < b$. (3.9) can also be represented in terms of receiver clock drift δ_{rx} after substituting $\tilde{t}_{rx} = \frac{t_{rx}}{(1+\delta_{rx})}$, and given as

$$\begin{aligned} \left(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd} \right) &= \left[(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})(1 + \delta_{rx}) / t_{rx} \right] \tilde{t}_{rx} \\ &+ \left[\text{mod} \left((t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd}), t_{rx} / (1 + \delta_{rx}) \right) \right]. \end{aligned} \quad (3.10)$$

Since the signal is recovered using ETS, the time interval i.e. Term B is slid in the equivalent time portion and corresponding equivalent time samples are obtained. Due to this sliding, the recovered samples are thus affected by the imperfections of both the tx and rx clock drift. The equivalent time samples, S , observed due to this time interval (Term B) is thus given by dividing it with an actual equivalent time resolution $\tilde{T}_{eq,rx}^{tx}$

$$S = \frac{\text{mod} \left((t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd}), t_{rx} / (1 + \delta_{rx}) \right)}{\tilde{T}_{eq,rx}^{tx}} \quad (3.11)$$

Therefore, the total equivalent time samples observed before the time of arrival in an equivalent time frame is equal to the integer number of samples (as obtained in Term A) plus the equivalent time samples observed in (3.11). These total samples are denoted by $\tilde{K}_{E-TOA}$, and thus represented as

$$\tilde{K}_{E-TOA} = \left\lfloor \frac{(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})(1 + \delta_{rx})}{t_{rx}} \right\rfloor + S \quad (3.12)$$

Further, from the relation in (3.1), $\tilde{T}_{eq,rx}^{tx}$ for $n_s = 1$ can be rewritten as

$$\tilde{T}_{eq,rx}^{tx} = \left(\frac{t_{rx}}{\alpha_{rx}} - \frac{t_{tx}}{\alpha_{tx}} \right), \quad (3.13)$$

A simple manipulation of (3.13) gives us the following representation,

$$\begin{aligned} \left(\frac{t_{rx}}{\alpha_{rx}} - \frac{t_{tx}}{\alpha_{tx}} \right) &= \left(t_{rx} - t_{tx} + \frac{t_{rx}}{\alpha_{rx}} - t_{rx} + t_{tx} - \frac{t_{tx}}{\alpha_{tx}} \right), \\ \tilde{T}_{eq,rx}^{tx} &= T_{eq,rx}^{tx} (1 + \Psi_{rx} - \Psi_{tx}), \\ \tilde{T}_{eq,rx}^{tx} &= T_{eq,rx}^{tx} (1 + \Psi_{rx}^{tx}), \end{aligned} \quad (3.14)$$

where $\Psi_{rx}^{tx} = \Psi_{rx} - \Psi_{tx}$, and Ψ_{tx} and Ψ_{rx} is given by

$$\begin{aligned} \Psi_{tx} &= \frac{\left(\frac{t_{tx}}{\alpha_{tx}} - t_{tx} \right)}{T_{eq,rx}^{tx}} = \frac{-\Delta f_{tx}/f_{tx}^2}{\left(\frac{1}{f_{rx}} - \frac{1}{f_{tx}} \right)} = \frac{-\delta_{tx} \times t_{tx}}{T_{eq,rx}^{tx}}, \\ \Psi_{rx} &= \frac{\left(\frac{t_{rx}}{\alpha_{rx}} - t_{rx} \right)}{T_{eq,rx}^{tx}} = \frac{-\Delta f_{rx}/f_{rx}^2}{\left(\frac{1}{f_{rx}} - \frac{1}{f_{tx}} \right)} = \frac{-\delta_{rx} \times t_{rx}}{T_{eq,rx}^{tx}}. \end{aligned} \quad (3.15)$$

Substituting (3.14) into (3.11) and (3.12), and manipulating using Taylor series expansion, we get the total equivalent time samples as

$$\begin{aligned} \tilde{K}_{E-TOA} &= \left\lfloor \frac{(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})(1 + \delta_{rx})}{t_{rx}} \right\rfloor \\ &+ \left[\text{mod} \left((t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd}), t_{rx}/(1 + \delta_{rx}) \right) \right] \frac{1}{T_{eq,rx}^{tx}} \left(1 + \sum_{l=1}^{\infty} (-1)^l (\Psi_{rx}^{tx})^l \right), \end{aligned} \quad (3.16)$$

where, $\text{mod } (a,b)$ in (3.16) represents a modulo b , m is the index of recovered equivalent time pulse in the captured frame (c.f. Fig.3.4), and

$$\Psi_{rx}^{tx} = \Psi_{rx} - \Psi_{tx} = \left(\frac{\delta_{tx} \times t_{tx}}{T_{eq,rx}^{tx}} - \frac{\delta_{rx} \times t_{rx}}{T_{eq,rx}^{tx}} \right) \quad (3.17)$$

is the error due to ETS technique. The equation for total equivalent time samples, given in (3.16), is further generalized for the p^{th} pulse and given as

$$\begin{aligned} \tilde{K}_p^{E-TOA} = & \left\lfloor \frac{(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})(1 + \delta_{rx})}{t_{rx}} \right\rfloor \\ & + \left[\text{mod} \left((t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd}), t_{rx}/(1 + \delta_{rx}) \right) + p\tilde{t}_{tx} \right] \frac{1}{T_{eq,rx}^{tx}} \left(1 + \sum_{l=1}^{\infty} (-1)^l (\Psi_{rx}^{tx})^l \right), \end{aligned} \quad (3.18)$$

Proposition 1. *For a system employing ETS technique, the noise free equivalent time of arrival (E-TOA), $\tilde{\zeta}_{rx}^{tx}$, of the signal from a transmitting node tx to a receiving node rx, for an equivalent time frame of the p^{th} pulse is given by*

$$\tilde{\zeta}_{rx}^{tx} = \tilde{K}_p^{E-TOA} \times \tilde{T}_{eq,rx}^{tx}, \quad (3.19)$$

where, $\tilde{K}_p^{E-TOA}$ is given by (3.18) and $\tilde{T}_{eq,rx}^{tx}$ is the actual equivalent time resolution.

Remark 1. It is important to interpret the E-TOA error Ψ_{rx}^{tx} , appearing in the expression of $\tilde{K}_p^{E-TOA}$ in (3.19). Since $\frac{t_{tx}}{T_{eq,rx}^{tx}}$ is approximately equal to $\frac{t_{rx}}{T_{eq,rx}^{tx}}$ and is in the order of a few thousands (assuming t_{tx} and t_{rx} in the order of few microseconds), the difference of the clock drift of the two nodes is degraded thousands of times i.e. for the system employing ETS the clock drift effect is magnified by $\Psi_{rx}^{tx} \approx (\delta_{tx} - \delta_{rx}) \frac{t_{tx}}{T_{eq,rx}^{tx}}$.

Remark 2. We remark here that equation (3.19) distinguish itself from the RTS equation mainly due to the influence of transmitter clock drift. Specifically, it is the magnified transmitter and receiver drift difference by an equivalent time PRF (i.e. a few thousand times). We further state here that this difference is an important aspect to analyze the systems which uses TOA, TDOA, D-TDOA approaches for ranging purpose. The experimental demonstration of such a scheme is presented in section V.

Remark 3. As can be seen from the expression of $\tilde{K}_p^{E-TOA}$ in (3.19), and because

Figure 3.5: Simulation model and simulation parameters

of the huge time difference $(t_{start}^{tx} - t_{cap,rx})$, in the actual ranging system there must be a mechanism to know, from the measurements, which part is from Term A and which part is from Term B. However, this can also be resolved by applying techniques such as TDOA and DTDOA across the receivers/(or anchors) involved in ranging or positioning. We therefore present an E-DTDOA based algorithm to utilize the proposed E-TOA equation for finding the range difference of a target node between two anchor nodes. The details and demonstrations of these ranging algorithms is investigated in section V.

Since calculation of $\tilde{T}_{eq,rx}^{tx}$ in (3.19) requires estimation of precise tx and rx clock drift, and which is not known a-priori, therefore we approximate E-TOA in (3.19) by multiplying the recovered equivalent time samples $\tilde{K}_p^{E-TOA}$ with the nominal equivalent time resolution $T_{eq,rx}^{tx}$ instead of actual $\tilde{T}_{eq,rx}^{tx}$ i.e. we assume (3.19) as

$$\hat{\zeta}_{rx}^{tx} = \tilde{K}_p^{E-TOA} \times T_{eq,rx}^{tx}, \quad (3.20)$$

and perform sensitivity analysis with changes in transmitter and receiver clock drift. However, in the subsequent sections we will propose a mechanism to estimate $\tilde{T}_{eq,rx}^{tx}$ and obtain the actual E-TOA as in (3.19).

3.3 Validation and Sensitivity Analysis of the Proposed E-TOA Equations using Simulations

We use (3.20) to validate the proposed analytical E-TOA equation. Numerical simulations are performed using the simulation model described below.

3.3.1 Simulation Setup

The block diagram of the simulation model with simulation parameters are shown in Fig.3.5. The signal model as given by (3.3)-(3.8), is adopted for simulations using MATLAB programming. The channel responses $h(t)$ are generated according to models in CM1 and CM3 [177]. The simulation parameters such as t_{start}^{tx} , δ_{tx} , $t_{cap,rx}$ and δ_{rx} are selected randomly at the beginning of each cycle of a frame transmission, with frame transmission period of 2 mS. The range of values of these parameters are shown along with Fig.3.5. It is notable that these parameters then remain constant over one frame transmission period i.e for 2 mS. In addition, the propagation delay, $t_{tx,rx}^{pd}$, between tx and rx is assumed to be constant for each cycle with $t_{tx,rx}^{pd} = 47.133$ nS. The received signal at rx node, $r(t)$, is passed through the receiver chain (with circuit delay approximated to be zero i.e. $t_{rx}^{cd} = 0$) as shown in Fig.3.5. The signal is further sampled at the ADC using ETS technique to get the equivalent time sampled signal $y(m\tilde{t}_{rx})$ as given by equation (3.8). The voltage of the received signal is converted to the power unit (i.e. $\frac{voltage^2}{Impedance}$) after the down-conversion stage where the signal is passed through a multiplier or mixer. After ADC, the amplitude of the signal, which has power unit, is quantized to an integer number of ADC quantization levels. These quantization levels further squares up after passing through the correlator, which uses the time shifted version of the same signal. The resultant amplitude is termed as correlated amplitude. The raw data as captured by the ADC and its correlated output is shown in Fig. 3.6a and Fig. 3.6b, respectively. A robust peak searching algorithm is used to estimate the time of arrival from the recovered equivalent time sampled signal. The algorithm finds the first local peak which exceeds some threshold. A simple method which is dependent on environment data is used to find the threshold. The threshold is adjusted until the variance in TOA is lowered. In the next section we verify the mathematical formulation of the proposed E-TOA equation in (3.20) using this simulation model.

3.3.2 Impact of SNR on E-TOA

Simulations have been run to investigate the performance of signal to noise ratio on the proposed E-TOA. The block diagram of the simulation model with simulation parameters are shown in Fig.3.5. The signal model as given by (3.3)-(3.8), is adopted for simulations using MATLAB programming. The channel responses $h(t)$ are generated according to models in CM3 [177]. The simulation parameters such

Figure 3.6: Experimental Data of the recovered equivalent time sampled signal

as t_{start}^{tx} , $t_{cap,rx}$ are kept to a same constant value, while for this simulation no clock drift is considered i.e. $\delta_{tx} = 0$ and $\delta_{rx} = 0$. The frame transmission period is equal to 2 mS. The transmit power = 22.53 dBm with peak voltage of 3 volts, distance between tx and rx = 14.14 meter, computed path loss = 70.51 dB, attenuation at the rx = 5 dB and noise power level = -74 dBm. The equivalent time resolution is adjusted as per the relation in (3.1). We consider different combination of transmitter and receiver frequency to get the desired equivalent time resolution. The combination of frequencies selected were $[f_{tx} = 3 \text{ MHz}, f_{rx} = 2.999010 \text{ MHz}]$, $[f_{tx} = 2 \text{ MHz}, f_{rx} = 1.999500 \text{ MHz}]$, $[f_{tx} = 2 \text{ MHz}, f_{rx} = 1.999200 \text{ MHz}]$ to get the equivalent time resolution of 0.11 nanoseconds, 0.125 nanoseconds and 0.20 nanoseconds, respectively. Fig.3.7 shows the SNR vrs range error for different equivalent time sampling resolution. It is observed from the figure that the error decreases in all the cases as SNR increases. However, the smoothness of the plot is affected by the limited resolution points used to generate the UWB signal at the transmitter during simulation.

3.3.3 Validation of the Proposed E-TOA Equations

The mathematical formulation of an E-TOA equation in (3.20) is validated by comparing the simulated E-TOA value, $\hat{\zeta}_{rx}^{tx}(sim)$, and calculated E-TOA value, $\hat{\zeta}_{rx}^{tx}(cal)$. The difference between $\hat{\zeta}_{rx}^{tx}(sim)$ and $\hat{\zeta}_{rx}^{tx}(cal)$, in nano-seconds, is plotted in Fig.3.8 for different order of summation l using different combinations of δ_{tx} and δ_{rx} , while keeping other parameters fixed. In addition to the values of the parameters mentioned in Fig.3.5, value of other parameters considered for plotting Fig.3.8 are: $t_{start}^{tx} = 282380.77 \text{ nS}$, $t_{cap,rx} = 50.1 \text{ nS}$, $t_{rx}^{cd} = 0$ and $p_m = 0$. These values are se-

Figure 3.7: SNR vrs Range error

lected to consider the effect of resolution in the order of sub-nanoseconds. Different combinations of δ_{tx} and δ_{rx} are selected which are chosen to be the representative of practical implementations of low cost crystal oscillators. Fig.3.8 shows the convergence between the $\hat{\zeta}_{rx}^{tx}(sim) - \hat{\zeta}_{rx}^{tx}(cal)$ after $l = 5$. Although it is observed, that the expression of $\tilde{K}_p^{E-TOA}$ in the proposed E-TOA equation is non-linear in Ψ_{rx}^{tx} , it is useful to linearize it by assuming $l = 1$. For precision ranging, the TOA error should be less than 1 nS and hence clock drift between rx and tx nodes should be less than 15 ppm for the linearized model as indicated in Fig.3.8 inset.

Assuming the linearized model with $l = 1$, Fig.3.9 is plotted to further verify the floor and the modulus part appearing in $\tilde{K}_p^{E-TOA}$ term of eq. (3.20). Fig.3.9 plots the error between nominal E-TOA, denoted by $\zeta_{rx}^{tx} = \hat{\zeta}_{rx}^{tx} |_{\delta_{tx}=\delta_{rx}=0}$ (i.e. without tx and rx clock skew variations), and actual E-TOA, $\hat{\zeta}_{rx}^{tx}$ vs the variations in $(t_{start}^{tx} - t_{cap,rx})$. It is observed from Fig.3.9 that when the value of $(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})$ is less than actual sampling rate ($\tilde{t}_{rx}$) the floor term is zero and error is mainly

Figure 3.8: The plot shows the error between the calculated $\hat{\zeta}_{rx}^{tx}(cal)$ value and the simulated $\hat{\zeta}_{rx}^{tx}(sim)$ value for different values of l .

contributed by the modulus part of the equation. This error increases with the increase in $(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})$ to a certain value until it drops down suddenly as shown in the Fig.3.9. This sudden drop occur at the instants when the value of $(t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})$ crosses the integer number of sampling rate, $\tilde{t}_{rx}$, and thereafter, it starts having the floor term/s. Moreover, the error values at each data points in the plot can be justified from the mathematical expression of E-TOA in (3.20). It can be seen that the error shown in Fig.3.9 varies according to the difference between the clock drift of the tx and rx node as mentioned in remark 2.

3.3.4 Sensitivity Analysis of E-TOA equation with variation in t_{tx} and t_{rx} clock drift

From the expression of $\hat{\zeta}_{rx}^{tx}$ in (3.20), we observe that $\hat{\zeta}_{rx}^{tx}$ is sensitive to the variations of the local clock oscillator parameters of tx and rx node. Thus, for the purpose of sensitivity analysis with variations in tx and rx clock drift, we differentiate the error $(\hat{\zeta}_{rx}^{tx} - \zeta_{rx}^{tx})$ w.r.t δ_{rx} in (3.20) to get

$$\frac{\partial(\hat{\zeta}_{rx}^{tx} - \zeta_{rx}^{tx})}{\partial \delta_{rx}} = \left(X - \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor t_{rx}(1 - \delta_{rx}) \right) \left(\sum_{l=1}^{\infty} (-1)^{l+1} l \left(\delta_{tx} \frac{t_{tx}}{T_{eq,rx}^{tx}} - \delta_{rx} \frac{t_{rx}}{T_{eq,rx}^{tx}} \right)^l \right. \\ \left. + \frac{t_{rx}}{T_{eq,rx}^{tx}} + \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor \left(1 + \sum_{l=1}^{\infty} (-1)^l \left(\delta_{tx} \frac{t_{tx}}{T_{eq,rx}^{tx}} - \delta_{rx} \frac{t_{rx}}{T_{eq,rx}^{tx}} \right)^l \right) \right). \quad (3.21)$$

3.3 Validation and Sensitivity Analysis of the Proposed E-TOA Equations using Simulations

Figure 3.9: The plot shows the error in the nominal (ζ_{rx}^{tx}) and actual E-TOA ($\hat{\zeta}_{rx}^{tx}$) vs $(t_{start}^{tx} - t_{cap,rx})$, for different δ_{tx} and δ_{rx} . The figure shows 2 cases: case 1 (Top figure) when $f_{tx} = 2$ MHz and $f_{rx} = 1.999500125$ MHz, and case 2 (Lower figure) when $f_{tx} = 3$ MHz and $f_{rx} = 2.999010025$ MHz.

However, to simplify it further, asymptotically we can write (3.21) as

$$\frac{\partial(\hat{\zeta}_{rx}^{tx} - \zeta_{rx}^{tx})}{\partial \delta_{rx}} \approx (A' \delta_{rx} - B' \delta_{tx} + C') \quad (3.22)$$

where, $A' = \frac{2t_{rx}^2}{T_{eq,rx}^{tx}} \left\lfloor \frac{X}{t_{rx}} \right\rfloor$, $B' = \frac{t_{rx}t_{tx}}{T_{eq,rx}^{tx}} \left\lfloor \frac{X}{t_{rx}} \right\rfloor$, $C' = \left(\left\lfloor \frac{X}{t_{rx}} \right\rfloor t_{rx} + X \frac{t_{rx}}{T_{eq,rx}^{tx}} - \left\lfloor \frac{X}{t_{rx}} \right\rfloor \frac{t_{rx}^2}{T_{eq,rx}^{tx}} \right)$ and $X = (t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})$. Similarly, when differentiating w.r.t δ_{tx} we get

$$\begin{aligned} \frac{\partial(\zeta_{rx}^{tx} - \hat{\zeta}_{rx}^{tx})}{\partial \delta_{tx}} = & - \left(- \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor t_{rx} + X + \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor t_{rx} \delta_{rx} \right) \\ & \sum_{l=1}^{\infty} (-1)^l l \left(\delta_{tx} \frac{t_{tx}}{T_{eq,rx}^{tx}} - \delta_{rx} \frac{t_{rx}}{T_{eq,rx}^{tx}} \right)^{l-1} \frac{t_{tx}}{T_{eq,rx}^{tx}}. \end{aligned} \quad (3.23)$$

However, to simplify it further, we can write (3.23) as

$$\frac{\partial(\zeta_{rx}^{tx} - \hat{\zeta}_{rx}^{tx})}{\partial \delta_{tx}} \approx (A'' \delta_{rx} - C'') \quad (3.24)$$

where $A'' = \frac{t_{rx}t_{tx}}{T_{eq,rx}^{tx}} \left\lfloor \frac{X}{t_{rx}} \right\rfloor$ and $C'' = \left(\left\lfloor \frac{X}{t_{rx}} \right\rfloor \frac{t_{rx}t_{tx}}{T_{eq,rx}^{tx}} - X \frac{t_{tx}}{T_{eq,rx}^{tx}} \right)$.

Refer Appendix for proof of (3.22) and (3.24). (3.22) and (3.24) are validated in Fig.3.10 and Fig.3.11, respectively using numerical simulations. Fig.3.10 shows the plot between the error ($\hat{\zeta}_{rx}^{tx} - \zeta_{rx}^{tx}$) vs δ_{rx} , while keeping other parameters fixed. The rate of change of the plot in Fig.3.10 is justified with the help of theoretical expression in (3.22). It is seen that the rate of change of the error w.r.t δ_{rx} is a function of the timing parameters, δ_{rx} and δ_{tx} . Similarly, Fig.3.11 is plotted with variations in δ_{tx} keeping $\delta_{rx} = 0$. The slope of Fig.3.11 can be justified from the expression in (3.24). For example, the slope observed from Fig.3.11 at $\delta_{tx} = 7$ ppm for the case when $t_{start}^{tx} = 150.1$ nS is ($\frac{2.060}{7} nS/ppm = 0.2942857 \times 10^{-3}$). The slope obtained from the expression in (3.24) by substituting the value of $X (= 97.133$ nS) and $\delta_{rx} = 0$ is equal to $0.29431299 \times 10^{-3}$. We have demonstrated in Fig.3.9,

Figure 3.10: The plot shows the error in the nominal E-TOA ($\hat{\zeta}_{rx}^{tx}$) and actual E-TOA (ζ_{rx}^{tx}) vrs δ_{rx} for different t_{start}^{tx} , while keeping $\delta_{tx} = 0$.

Fig.3.10 and Fig.3.11, that error can be few nanoseconds even for small variation in the clock drift, which is significant for precise ranging with cm level accuracy. It is thus shown, that error caused due to the imperfections in both the tx and rx clock

Figure 3.11: The plot shows the error in the nominal (ζ_{rx}^{tx}) and actual E-TOA ($\hat{\zeta}_{rx}^{tx}$) vrs δ_{tx} for different t_{start}^{tx} , while keeping $\delta_{rx} = 0$.

drift is an important aspect for precision ranging while using ETS based schemes. Moreover, it is important to precisely estimate the tx and rx clock drift to estimate the actual equivalent time resolution $\tilde{T}_{eq,rx}^{tx}$. We will see in the later sections, how to estimate $\tilde{T}_{eq,rx}^{tx}$ and how we can exploit the use of these E-TOA technique to find the precise range difference between the nodes.

3.4 Experimental Validation of the Proposed E-TOA Equation

In this section, we demonstrate the validation of the proposed E-TOA equation by comparing the experimental and calculated value of $\tilde{K}_{E-TOA}$, as appearing in (3.20), using hardware experiments. The details of the experimental setup used for the verification purpose is discussed below.

Figure 3.12: Experimental setup used for the verification of the proposed E-TOA equations

3.4.1 Experimental Setup

Fig.3.12 shows the experimental test bed used for the verification of E-TOA equation. It consists of a transmitting node tx and a receiving node rx, an oscilloscope and a server to read E-TOA values from the rx node. We use in house designed tx and rx node and the description of the nodes is as follows:

3.4.1.1 Transmitting Node (tx)

The UWB transmitter has two major components, one is an impulse generator and another a pulse forming circuit. A high power sub-nanosecond impulse is generated from the impulse generator. This impulse is used in pulse forming network which translates the generated impulse to form a desired Gaussian UWB pulse. The measured -10 dB bandwidth of the pulse generator is about 2 GHz. The generated pulse has a high peak power of over +20dBm. The tag operates at a frequency between 3-5 GHz with center frequency at 4 GHz. The tx transmits UWB-IR ranging signal with a nominal PRF, f_{tx} and data sequence through a micro-controller unit. The PRF can be adjusted as per the desired requirements, under the constraint of the FCC defined emission regulations for UWB signals [178]. In this work, we adjusted the nominal PRF, f_{tx} , at 3 MHz.

3.4.1.2 Receiving Node (rx)

The block diagram of the rx node is shown in Fig.3.13. The receiver node rx adopts a non-coherent correlation-based approach. Node rx consists of a UWB front end receiver, followed by a variable gain amplifier, a low sampling rate ADC, a micro-

Figure 3.13: Block diagram of the tx and rx node.

controller unit, the BeagleBone Black (BBB) embedded computer and a battery bank. The front end of UWB receiver consists of a UWB antenna, band-pass filter, low noise amplifier (LNA), and a mixer stage. The UWB receiver detects the envelope of the transmitted UWB pulses. The baseband amplifier in turn, amplifies the UWB envelope before it is sampled by the ADC. Since UWB signals have a large bandwidth, an ADC with high sampling frequency is required to follow the Nyquist sampling criteria. This will eventually lead to a need for expensive ADC chips in the system. To provide a cost effective solution, we employ ETS technique in our system by using a low cost ADC to sample the repetitive UWB envelopes. Since the sampling of UWB envelope is based on ETS, a lower sampling clock frequency is used, which is generated by the software adjustable digital direct synthesizer (DDS) [179] that utilizes 75 MHz temperature compensated oscillator (TCXO). The nominal sampling frequency of the receiver was adjusted at $f_{rx} = 2.9990103$ MHz. Moreover, using low sampling rate ADC can further help in using processor which can easily be used in low-power and low-cost sensor nodes. In our system, we use low processing unit, BBB embedded Linux device which uses ARM Cortex-A8 1GHz processor.

The ARM Cortex A8 microprocessor on BBB integrates two on-chip Programmable Real-time Units (PRU) clocked by a 200 MHz oscillator. This provides the fast real-time response with each PRU instruction taking 5 ns. The ADC channel data that connects to the front-end receiver board is fed to the PRU via special PRU pins. The PRU's firmware then records the 16 KB frame data into the device memory allocated on Dual Data Rate (DDR) external RAM memory. On the Linux user space, our software program is running two different processes. The first process grabs the data from the device memory before it gets overwritten by subsequent data frame. Based on the data stored by the first process, the second process runs the TOA detection algorithm in python. The second process performs the corre-

lation and then process the data to estimate experimental value of E-TOA sample index, denoted by $\tilde{K}_p^{E-TOA}(exp)$, of the recovered equivalent time sampled UWB signal. These E-TOA objects, $\tilde{K}_p^{E-TOA}(exp)$, are transferred to the central server via UHF channel. Fig.3.14 shows the illustration of the recovered equivalent time sampled signal, as stored in BBB embedded computer, at node rx. The cursors in Fig.3.14 displays the sample index of the E-TOA, $\tilde{K}_p^{E-TOA}(exp)$. The values $X = 3673$ and $X = 6639$ in the cursor represents $\tilde{K}_p^{E-TOA}(exp)$ for pulse index $p = 0$ and $p = 1$, respectively.

Figure 3.14: Recovered equivalent time sampled signal after correlation displaying the sample index of observed E-TOA i.e. $\tilde{K}_p^{E-TOA}(exp)$. The X values in the cursor displays $\tilde{K}_p^{E-TOA}(exp)$ for pulse index $p = 0$ and $p = 1$.

Note: The frequency tolerance limit is mentioned in the specifications of the crystal unit given by the manufacturer. In the presented system, TSX-3225 [180] is used as the crystal unit for the transmitter clock. The frequency tolerance for this unit is mentioned as ± 10 ppm. Likewise, at the receiver XG-1000CA [181] is used as a crystal unit, which has a frequency tolerance of ± 100 ppm. Now, given the range of frequency tolerance specifications, the suitable values for nominal $f_{tx} = \frac{1}{t_{tx}}$ and $f_{rx} = \frac{1}{t_{rx}}$ are selected. For example, to consider the sub-nanosecond

Figure 3.15: Snapshot of the signal observed on the oscilloscopes for experimental verification of E-TOA equation. Specifically used to find τ_{diff} .

resolution of 0.11 ns, the nominal transmitter pulse repetitive frequency f_{tx} and the nominal receiver ADC sampling frequency f_{rx} must be adjusted to 3 MHz and 2.9990103 MHz, respectively. With the given specifications, the maximum variation in transmitter clock, corresponding to ± 10 ppm (or ± 30 Hz), f_{tx} can be $f_{tx} - \Delta f_{tx} = 3 \text{ MHz} - 30 \text{ Hz} = 2999970.0 \text{ Hz}$ and $f_{tx} + \Delta f_{tx} = 3 \text{ MHz} + 30 \text{ Hz} = 3000030 \text{ Hz}$ i.e. the range of true $\tilde{f}_{tx}$ is between $[2999970.0 - 3000030] \text{ Hz}$. Similarly, the maximum variation in the receiver clock, corresponding to ± 100 ppm (or ± 300 Hz), f_{rx} can be $f_{rx} - \Delta f_{rx} = 2999010.3 \text{ Hz} - 300 \text{ Hz} = 2998710.3 \text{ Hz}$ and $f_{rx} + \Delta f_{rx} = 2999010.3 \text{ Hz} + 300 \text{ Hz} = 2999310.3 \text{ Hz}$ i.e. the range of true $\tilde{f}_{rx}$ is between $[2998710.3 - 2999310.3] \text{ Hz}$. Thus, we observe, with the selected nominal values and the given frequency tolerance specifications, the range of true $\tilde{f}_{tx}$ is always greater than the range of true $\tilde{f}_{rx}$. Therefore, the condition $\tilde{f}_{tx} > \tilde{f}_{rx}$ or $\tilde{t}_{tx} < \tilde{t}_{rx}$ always holds true for the selected nominal values of f_{tx} and f_{rx} .

3.4.2 Validation Methodology and Experimental Results

For validating the E-TOA equation experimentally, the experimental E-TOA sample index, $\tilde{K}_p^{E-TOA}(\text{exp})$ as obtained in the above section, is compared with the calculated E-TOA index, denoted by $\tilde{K}_p^{E-TOA}(\text{cal})$. Expression in (3.16), is used to

calculate $\tilde{K}_p^{E-TOA}(cal)$. However, to calculate $\tilde{K}_p^{E-TOA}(cal)$ using (3.16), it is required to find the corresponding experimental values of the parameters t_{start}^{tx} , $t_{tx,rx}^{pd}$, t_{rx}^{cd} , $t_{cap,rx}$, δ_{rx} as used in (3.16). During the experiments these parameters were obtained from the receiver board using Tektronix MDO 3104 oscilloscope. The snapshot of one such signal realization observed on the oscilloscope is shown in Fig.3.15. Fig.3.15 shows the trigger signal which initiates the start of the ADC capture point, ADC clock signal and UWB-IR signal. The parameter, $t_{cap,rx}$, was obtained by probing the oscilloscope channel at the triggering point which initiated the start of the ADC capture point. The sum of other parameters, $(t_{start}^{tx} + t_{tx,rx}^{pd} + t_{rx}^{cd})$, was obtained by probing the oscilloscope channel at the point where signal is sent to the ADC for sampling. The difference, $\tau_{diff} = (t_{start}^{tx} + t_{tx,rx}^{pd} + t_{rx}^{cd}) - t_{cap,rx}$, is then obtained from the oscilloscope. In addition, the other parameter, δ_{rx} (or $\Delta f_{rx}/f_{rx}$), is obtained from the PRF of the recovered equivalent time sampled signal, and the relation in (3.1) and (3.2) by assuming δ_{tx} (or $\Delta f_{tx}/f_{tx}$) = 0. For example, for the case shown in Fig.3.14, total no. of PRF sample points obtained is given by $\tilde{F}_{rx}^{tx} = (6639 - 3673) = 2966$. From the relation in (3.1) and (3.2), we can write $\frac{1}{f_{rx} + \Delta f_{rx}} - \frac{1}{f_{tx}} = \left(\frac{1}{\tilde{F}_{rx}^{tx}}\right)$. Substituting f_{rx} , f_{tx} and $\tilde{F}_{rx}^{tx}$ we get $\Delta f_{rx} = -21.449$ Hz. These parameters are substituted in (3.16) to get $\hat{K}_{E-TOA}(cal)$. The experimental

Table 3.2: Validation of the proposed E-TOA equation $d = 0.5$ m

Run	$\tilde{K}_p^{E-TOA}(exp)$ ($p = 0$)	$\tilde{K}_p^{E-TOA}(cal)$ ($p = 0$)	$\tau_{diff}(uS)$	$\Delta f_{rx}(Hz)$
1	3673	3673	287.74620	-21.45
2	3176	3177	286.69020	-20.77
3	3372	3372	293.04570	-21.45
4	3140	3140	289.3529	-21.45
5	3341	3342	292.7091	-21.79

value of $\tilde{K}_p^{E-TOA}(exp)$ and corresponding calculated value $\tilde{K}_p^{E-TOA}(cal)$ for different runs, when the distance between tx node and rx node is 0.5 m and 1 m are notified in Table. 3.2 and Table. 3.3, respectively. The close match between $\tilde{K}_p^{E-TOA}(exp)$ and $\tilde{K}_p^{E-TOA}(cal)$ further validates the analytical expression of the proposed E-TOA equation, experimentally. For the completeness of the analysis, we further demonstrate how the proposed equation can be utilized for determining the ranges among the nodes.

Table 3.3: Validation of the proposed E-TOA equation $d = 1\text{m}$

Run	$\tilde{K}_p^{E-TOA}(exp)$ ($p = 0$)	$\tilde{K}_p^{E-TOA}(cal)$ ($p = 0$)	$\tau_{diff}(uS)$	$\Delta f_{rx}(Hz)$
1	3627	3628	287.74110	-21.45
2	3708	3709	287.08300	-20.08
3	2993	2995	292.00340	-21.79
4	1144	1144	294.79520	-20.43
5	3622	3622	285.07310	-19.74

Remark 4. It is important to highlight here, that an oscilloscope in the above experiment was used to find the numerical value of $\tilde{K}_p^{E-TOA}(cal)$ with the help of expression in (3.16). This was compared with $\tilde{K}_p^{E-TOA}(exp)$ as obtained by BBB embedded within the receiver. For the real ranging experiments, we use $\tilde{K}_p^{E-TOA}(exp)$ from BBB to obtain the ranges between the anchor nodes and thus, do not require the need of an oscilloscope.

3.5 Proposed Ranging Mechanism Using E-TOA Equations

This section illustrates the use of E-TOA measurements to propose E-DTDOA based ranging mechanism between the nodes.

3.5.1 Equivalent Differential Time Difference of Arrival (E-DTDOA) Operation

The overall E-DTDOA mechanism is demonstrated with the help of Fig.3.19. Consider two transmitting nodes, tx_1 and tx_2 , and two receiving nodes, rx_1 and rx_2 . Transmitting nodes transmits the signal at different time slots with tx_1 transmitting first followed by tx_2 . The equivalent time sampled signal from tx_1 and tx_2 is received at rx_1 and rx_2 , and which is illustrated in Fig.3.19. E-TOA is obtained from the signal transmitted from tx_1 and received at receivers rx_1 and rx_2 , which is given by (3.20), and denoted by $\hat{\zeta}_{rx_1}^{tx_1}$ and $\hat{\zeta}_{rx_2}^{tx_1}$, respectively. Equivalent time difference of arrival (E-TDOA), denoted by $\hat{T}_{rx_1, rx_2}^{tx_1}$, and calculated across rx_1 and rx_2 is thus

given as (c.f. Fig.3.19)

$$\hat{T}_{rx_1,rx_2}^{tx_1} = \hat{\zeta}_{rx_1}^{tx_1} - \hat{\zeta}_{rx_2}^{tx_1} \quad (3.25)$$

Similarly, E-TDOA, $\hat{T}_{rx_1,rx_2}^{tx_2}$, obtained across rx_1 and rx_2 from the signal received from tx_2 is given by

$$\hat{T}_{rx_1,rx_2}^{tx_2} = \hat{\zeta}_{rx_1}^{tx_2} - \hat{\zeta}_{rx_2}^{tx_2} \quad (3.26)$$

From the E-TDOA's in (3.25) and (3.26), we obtain E-DTDOA, denoted by $\hat{D}_{rx_1,rx_2}^{tx_1,tx_2}$ as

$$\hat{D}_{rx_1,rx_2}^{tx_1,tx_2} = \hat{T}_{rx_1,rx_2}^{tx_1} - \hat{T}_{rx_1,rx_2}^{tx_2} \quad (3.27)$$

We use this approach to find the range difference of arrival of the target node tx_1 , assuming tx_2 as the known position reference node. It is important to interpret the significance of differencing in (3.25), (3.26) and (3.27). In an ideal case when there is no clock drift at tx and rx clocks i.e. when $\delta_{tx_1} = \delta_{tx_2} = \delta_{rx_1} = \delta_{rx_2} = 0$, differencing across the rx's, as in (3.25), will get rid of the start time of tx_1 node. Similarly, differencing across the rx's, as in (3.26), will get rid of the start time of tx_2 node. Further differencing and calculating E-DTDOA, as in (3.27), will cancel the offset term introduced by different rx's. However, in practical scenarios, the range difference accuracy is affected by the imperfections of the clock drift of the nodes. To overcome this we propose a practical ranging algorithm for an ETS based system.

Algorithm 3.1 is suggested to find the E-DTDOA, $\hat{D}_{rx_1,rx_2}^{tx_1,tx_2}$ for the cases when rx nodes are unsynchronized. The rx nodes are first calibrated by tuning the frequency of the nodes i.e. by making $\delta_{rx_1} = \delta_{rx_2} = 0$. Subsequently, the actual equivalent time resolution, $\tilde{T}_{eq,rx}^{tx}$, is estimated with the help of eq. (3.2)

$$\tilde{T}_{eq,rx}^{tx} = \left(\frac{t_{rx}}{\alpha_{rx}} - \frac{t_{tx}}{\alpha_{tx}} \right) \approx \frac{\left(\frac{1}{f_{tx}} \right)}{\tilde{F}_{rx}^{tx}}, \quad (3.28)$$

where, $\tilde{F}_{rx}^{tx}$ in (3.28) is obtained from the total number of observed samples between the recovered equivalent time pulses obtained from the measurements (c.f. Fig.3.19). Algorithm steps are summarized below.

Algorithm 3.1 E-DTDOA Estimation

Input: $\hat{\zeta}_{rx_1}^{tx_1}, \hat{\zeta}_{rx_2}^{tx_1}, \hat{\zeta}_{rx_1}^{tx_2}, \hat{\zeta}_{rx_2}^{tx_2}$ for p=0 pulse after correlation

Output: $\hat{D}_{rx_1, rx_2}^{tx_1, tx_2}$

- 1: $\delta_{rx_1} = \delta_{rx_2} \approx 0$; Readers tuned to the same frequency
 - 2: Find $\tilde{T}_{eq, rx_1}^{tx_1}, \tilde{T}_{eq, rx_2}^{tx_1}, \tilde{T}_{eq, rx_1}^{tx_2}, \tilde{T}_{eq, rx_2}^{tx_2}$ using (3.28)
 - 3: Find $\hat{T}_{rx_1, rx_2}^{tx_1} = \left(\hat{\zeta}_{rx_1}^{tx_1} \times \frac{\tilde{T}_{eq, rx_1}^{tx_1}}{\tilde{T}_{eq, rx}^{tx_1}} \right) - \left(\hat{\zeta}_{rx_2}^{tx_1} \times \frac{\tilde{T}_{eq, rx_2}^{tx_1}}{\tilde{T}_{eq, rx}^{tx_1}} \right)$
 - 4: Find $\hat{T}_{rx_1, rx_2}^{tx_2} = \left(\hat{\zeta}_{rx_1}^{tx_2} \times \frac{\tilde{T}_{eq, rx_1}^{tx_2}}{\tilde{T}_{eq, rx}^{tx_2}} \right) - \left(\hat{\zeta}_{rx_2}^{tx_2} \times \frac{\tilde{T}_{eq, rx_2}^{tx_2}}{\tilde{T}_{eq, rx}^{tx_2}} \right)$
 - 5: Find $\hat{D}_{rx_1, rx_2}^{tx_1, tx_2} = \hat{T}_{rx_1, rx_2}^{tx_1} - \hat{T}_{rx_1, rx_2}^{tx_2}$
-

Figure 3.16: Simulation model

3.5.2 Ranging Simulation Results and Discussions

For simulations, we consider 2 tx nodes (tx_1 and tx_2) and 2 rx nodes (rx_1 and rx_2). tx_1 is assumed to be the target node while tx_2 is considered as a reference node whose position is known. The simulation model for tx_1 , tx_2 , rx_1 , and rx_2 with their coordinates is shown in Fig.3.16. For simplicity, we consider placing tx_1 and tx_2 at the same location. The two tx nodes transmit one after the other with tx_1 transmitting first followed by tx_2 . Each of the receiving nodes rx_1 , and rx_2 run on its own clock and thus has different ADC sampling rate $\tilde{t}_{rx}$. For each (tx – rx) pair, E-TOA is obtained as per (3.19), and which is further used to find the E-TDOA using (3.25), (3.26) and E-DTDOA using (3.27).

Fig. 3.17a shows the sensitivity of the range difference of the target node with variations in the transmitter and receiver local clock. Since two rx nodes are not synchronized, there is a non-zero frequency difference between two nodes. In practice, two rx nodes can be tuned to the same frequency, however, the frequency adjustment resolution between two receivers is limited by the available hardware

Figure 3.17: δ_{tx} on x-axis represents both δ_{tx_1} and δ_{tx_2} . $\delta_{tx} = [x]$ represents the random variation between $[-x \ x]$ ppm. $\delta_{rx_1} = 0$ and δ_{rx_2} is randomly varied between 0.3 Hz and 0.6 Hz.

such as AD9834 [179]. For AD9834 the frequency adjustment resolution of 0.28 Hz can be achieved. Thus, two nodes (rx_1 and rx_2) will always have the minimum frequency difference of 0.28 Hz. Moreover, in practice, this difference will never be a fixed value and it will be varying to a value close to 0.28 Hz. This scenario is modeled in simulation by keeping the difference $|\Delta f_{rx_1} - \Delta f_{rx_2}| \approx 0.30$ Hz and to realize the variations we randomly vary this difference between $[0.3 - 0.6]$ Hz. Other simulation parameters like start time of the transmitter (t_{start}^{tx}) was fixed to a certain value, while the ADC capture time ($t_{cap,rx}$) was randomly varied between $[0 - 20]$ μ S for both the rx nodes. Moreover, both the tx nodes clock frequency was randomly varied up-to the range of $[-10 \ 10]$ ppm. Fig. 3.17a demonstrated the distribution of the range difference error ($D_{actual} - D_{nominal}$) between the nominal and the actual E-DTDOA values for different variation in tx nodes clock frequency (or drift). It is observed from Fig. 3.17a that the error is high when the variation in δ_{tx_1} and δ_{tx_2} is high. It is seen that the error can be as high as 1.5 meters when δ_{tx_1} and δ_{tx_2} is varied between $[-10 \ 10]$ ppm which is high for the precise ranging system with cm level accuracy. To improve the accuracy of the system we applied the proposed algorithm 3.1 to find the range difference of the target node, and the results of the error in the range difference is displayed in Fig. 3.17b. It is seen that the accuracy

Figure 3.18: Experimental ranging model

Figure 3.19: Illustration of the recovered equivalent time sampled signal at nodes rx_1 and rx_2

was improved and the error is within 10 cm for 50% of the cases. In the next section, we further demonstrated the feasibility of the proposed algorithm experimentally.

3.5.3 Ranging Experimental Results and Discussions

Algorithm 3.1 is implemented using in house designed tx and rx nodes. The proposed ranging system consists of two wireless anchor nodes, as shown in Fig.3.18. These anchor nodes were designed by collocating 2 tx nodes, tx_1 and tx_2 , on 2 rx nodes, rx_1 and rx_2 , i.e. node tx_1 was collocated with node rx_1 , while node tx_2 was collocated with node rx_2 . The system works when one of the anchor node (or master anchor node) initiates the process by broadcasting a beacon via UHF link to tx nodes. The

Figure 3.20: Range error distribution for different distances between two nodes

tx nodes receives the beacon and sends the acknowledgment along with its ID to the master anchor nodes. Subsequently, the tx node generates and transmits UWB-IR signal in their respective time slots. The nodes follow the transmission pattern with node tx_1 transmitting first followed by node tx_2 . Both the nodes transmits UWB-IR ranging signal for 2 mS one after the other. On the other hand two rx node's ADC sampling clock frequency was adjusted by tuning the DDS AD9834 with the frequency adjustment resolution of 0.3 Hz. The DDS tuning can be done by a software program running in a BBB linux user space. The anchor nodes acquire the UWB signals from the tx nodes and compute the signal E-TOA, $\tilde{K}_p^{E-TOA}(exp)$, using the BBB embedded computer. The recovered correlated equivalent time sampled signal at nodes rx_1 and rx_2 is shown in Fig.3.19. This E-TOA data is then transferred to the central server using UHF link. The server runs MATLAB to compute the range difference, $\hat{D}_{rx_1,rx_2}^{tx_1,tx_2}$, as given in algorithm 3.1.

Ranging results are plotted in Fig. 3.20a and Fig. 3.20b for different range of distances between the anchor nodes. Around 200 realizations were obtained to plot the figures. Fig. 3.20a and Fig. 3.20b shows the distribution of the measured ranging error in indoor LOS environment for the actual node distance of 5.411 m and 11.128 m, respectively. The ground truth value of the range between the nodes is measured using a handheld laser distance meter. Fig. 3.20a and Fig. 3.20b shows that ranging error is less than 10 cm for 50% of the time while it is less than 20 cm for 90% of the time in both cases. The high accuracy is achieved due to the recovery

of sub-nanosecond resolution CIR, which results in an improvement in the range estimation as demonstrated in this work. However, the assumption of not knowing the precise transmitter frequency as in (3.28) can lead to errors in the ranging, since the ETS mechanism is significantly affected by the variation in transmitter clock drift. Thus, by precisely estimating the transmitter frequency and obtaining the precise equivalent time resolution will further lead to significant improvements in the ranging results, which will be demonstrated in our future work. Moreover, the ranging errors can be further reduced by improving the ETS resolution.

3.6 Conclusions

Accurate range information can be obtained from the precise recovery of CIR. However, to obtain CIR using UWB-IR, high sampling rate ADC is required. To avoid such a requirement, ETS is employed to obtain high resolution CIR even with low sampling rate. To analyze such a system, we first propose an analytical technique to determine the E-TOA, which is generic for different time synchronization system architecture. The theoretical and experimental validation of analytical E-TOA has been demonstrated in the chapter. The effect of timing uncertainties relating to the transmitter clock start time and the receiver clock offset is analyzed with variations in the transmitter and receiver clock drift. A ranging scheme has been proposed using E-TOA measurements. It is thus demonstrated that high time resolution is achieved using the low-cost off the shelf components with low sampling rate ADC to recover high bandwidth signal, it is thus practically feasible and easily deploy-able for low-cost wireless sensor nodes.

3.7 Appendix

From the expression of $\hat{\zeta}_{rx}^{tx}$ in (3.20), we observe that $\hat{\zeta}_{rx}^{tx}$ is sensitive to the variations of the local clock oscillator parameters of tx and rx node, which is denoted by a variable Ψ_{rx}^{tx} . Thus, for the purpose of sensitivity analysis with variations in receiver clock drift, we differentiate the error $(\hat{\zeta}_{rx}^{tx} - \zeta_{rx}^{tx})$ w.r.t δ_{rx} in (3.20) to get

$$\frac{\partial(\hat{\zeta}_{rx}^{tx} - \zeta_{rx}^{tx})}{\partial\delta_{rx}} = \left(X - \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor t_{rx}(1 - \delta_{rx}) \right) \frac{\partial f(\delta_{rx}, \delta_{tx})}{\partial\delta_{rx}} + \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor f(\delta_{rx}, \delta_{tx}) \quad (3.29)$$

where, $X = (t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})$ and

$$f(\delta_{rx}, \delta_{tx}) = \left(1 + \sum_{l=1}^{\infty} (-1)^l (\Psi_{rx}^{tx})^l \right). \quad (3.30)$$

where $\Psi_{rx}^{tx} = (\delta_{tx} \frac{t_{tx}}{T_{eq,rx}^{tx}} - \delta_{rx} \frac{t_{rx}}{T_{eq,rx}^{tx}})$, and thus, we can write (3.30) as

$$f(\delta_{rx}, \delta_{tx}) = 1 + \sum_{l=1}^{\infty} (-1)^l \left(\delta_{tx} \frac{t_{tx}}{T_{eq,rx}^{tx}} - \delta_{rx} \frac{t_{rx}}{T_{eq,rx}^{tx}} \right)^l \quad (3.31)$$

Differentiating $f(\delta_{rx}, \delta_{tx})$ w.r.t δ_{rx} we get

$$\frac{\partial f(\delta_{rx}, \delta_{tx})}{\partial \delta_{rx}} = \sum_{l=1}^{\infty} (-1)^{l+1} l \left(\delta_{tx} \frac{t_{tx}}{T_{eq,rx}^{tx}} - \delta_{rx} \frac{t_{rx}}{T_{eq,rx}^{tx}} \right)^l \frac{t_{rx}}{T_{eq,rx}^{tx}} \quad (3.32)$$

Substituting (3.31) and (3.32) in (3.29), the final expression of the slope of the error

$$\begin{aligned} \frac{\partial(\hat{\zeta}_{rx}^{tx} - \zeta_{rx}^{tx})}{\partial \delta_{rx}} = & \left(X - \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor t_{rx} (1 - \delta_{rx}) \right) \left(\sum_{l=1}^{\infty} (-1)^{l+1} l \left(\delta_{tx} \frac{t_{tx}}{T_{eq,rx}^{tx}} - \delta_{rx} \frac{t_{rx}}{T_{eq,rx}^{tx}} \right)^l \frac{t_{rx}}{T_{eq,rx}^{tx}} \right) \\ & + \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor \left(1 + \sum_{l=1}^{\infty} (-1)^l \left(\delta_{tx} \frac{t_{tx}}{T_{eq,rx}^{tx}} - \delta_{rx} \frac{t_{rx}}{T_{eq,rx}^{tx}} \right)^l \right). \end{aligned} \quad (3.33)$$

Assuming⁴ $l=1$, the simplified form of (3.33) is given as

$$\begin{aligned} \frac{\partial(\hat{\zeta}_{rx}^{tx} - \zeta_{rx}^{tx})}{\partial \delta_{rx}} \approx & \left(\left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor t_{rx} - \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor \frac{t_{rx}^2}{T_{eq,rx}^{tx}} + \right. \\ & \left. X \frac{t_{rx}}{T_{eq,rx}^{tx}} - \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor \frac{t_{rx} t_{tx} \delta_{tx}}{T_{eq,rx}^{tx}} + \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor \frac{t_{rx}^2 2 \delta_{rx}}{T_{eq,rx}^{tx}} \right). \end{aligned} \quad (3.34)$$

Since, $\delta_{rx} \ll 1$, we approximate $\left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor \approx \left\lfloor \frac{X}{t_{rx}} \right\rfloor$. Therefore, asymptotically we can write (3.34) as

$$\frac{\partial(\hat{\zeta}_{rx}^{tx} - \zeta_{rx}^{tx})}{\partial \delta_{rx}} \approx (A' \delta_{rx} - B' \delta_{tx} + C'), \quad (3.35)$$

⁴The value of $l=1$ is used to obtain the linear form of the E-TOA equation. Therefore, for the equation to be valid the difference of δ_{tx} and δ_{rx} must be chosen to be less than 10 ppm as observed from Fig.3.8.

where, $A' = \frac{2t_{rx}^2}{T_{eq,rx}^{tx}} \left\lfloor \frac{X}{t_{rx}} \right\rfloor$, $B' = \frac{t_{rx}t_{tx}}{T_{eq,rx}^{tx}} \left\lfloor \frac{X}{t_{rx}} \right\rfloor$, $C' = \left(\left\lfloor \frac{X}{t_{rx}} \right\rfloor t_{rx} + X \frac{t_{rx}}{T_{eq,rx}^{tx}} - \left\lfloor \frac{X}{t_{rx}} \right\rfloor \frac{t_{rx}^2}{T_{eq,rx}^{tx}} \right)$ and $X = (t_{start}^{tx} - t_{cap,rx} + t_{tx,rx}^{pd} + t_{rx}^{cd})$. Similarly, to perform the sensitivity analysis with variations in transmitter clock drift, we differentiate the error $(\hat{\zeta}_{rx}^{tx} - \zeta_{rx}^{tx})$ w.r.t δ_{tx} , we get

$$\frac{\partial(\zeta_{rx}^{tx} - \hat{\zeta}_{rx}^{tx})}{\partial\delta_{tx}} = - \left(- \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor t_{rx} + X + \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor t_{rx} \delta_{rx} \right) \frac{\partial f(\delta_{rx}, \delta_{tx})}{\partial\delta_{tx}}. \quad (3.36)$$

Differentiating $f(\delta_{rx}, \delta_{tx})$ w.r.t δ_{tx} , we get

$$\frac{\partial f(\delta_{rx}, \delta_{tx})}{\partial\delta_{tx}} = \sum_{l=1}^{\infty} (-1)^l \left(\delta_{tx} \frac{t_{tx}}{T_{eq,rx}^{tx}} - \delta_{rx} \frac{t_{rx}}{T_{eq,rx}^{tx}} \right)^{l-1} \frac{t_{tx}}{T_{eq,rx}^{tx}}. \quad (3.37)$$

Substituting (3.37) into (3.36), we get

$$\begin{aligned} \frac{\partial(\zeta_{rx}^{tx} - \hat{\zeta}_{rx}^{tx})}{\partial\delta_{tx}} = & - \left(- \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor t_{rx} + X + \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor t_{rx} \delta_{rx} \right) \\ & \times \sum_{l=1}^{\infty} (-1)^l \left(\delta_{tx} \frac{t_{tx}}{T_{eq,rx}^{tx}} - \delta_{rx} \frac{t_{rx}}{T_{eq,rx}^{tx}} \right)^{l-1} \frac{t_{tx}}{T_{eq,rx}^{tx}}. \end{aligned} \quad (3.38)$$

and simplifying (3.38) by assuming $l=1$, we get

$$\frac{\partial(\zeta_{rx}^{tx} - \hat{\zeta}_{rx}^{tx})}{\partial\delta_{tx}} \approx - \left(\left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor \frac{t_{rx}t_{tx}}{T_{eq,rx}^{tx}} - X \frac{t_{tx}}{T_{eq,rx}^{tx}} - \left\lfloor \frac{X(1 + \delta_{rx})}{t_{rx}} \right\rfloor \frac{t_{rx}t_{tx}\delta_{rx}}{T_{eq,rx}^{tx}} \right). \quad (3.39)$$

Since, $\delta_{rx} \ll 1$, we can approximate $\left\lfloor \frac{X(1+\delta_{rx})}{t_{rx}} \right\rfloor \approx \left\lfloor \frac{X}{t_{rx}} \right\rfloor$. Therefore, asymptotically, (3.39) will result in

$$\frac{\partial(\zeta_{rx}^{tx} - \hat{\zeta}_{rx}^{tx})}{\partial\delta_{tx}} \approx (A''\delta_{rx} - C''), \quad (3.40)$$

where $A'' = \frac{t_{rx}t_{tx}}{T_{eq,rx}^{tx}} \left\lfloor \frac{X}{t_{rx}} \right\rfloor$ and $C'' = \left(\left\lfloor \frac{X}{t_{rx}} \right\rfloor \frac{t_{rx}t_{tx}}{T_{eq,rx}^{tx}} - X \frac{t_{tx}}{T_{eq,rx}^{tx}} \right)$.

4 E-DTDOA based Localization for Wireless Sensor Networks with Clock Drift Compensation

4.1 Introduction

In chapter 3, we proposed a novel analytical E-TOA equations for systems employing equivalent time sampling (ETS) technique. The validations of E-TOA equations was carried out numerically using simulations as well as by experimental demonstrations. It was demonstrated that the equivalent time error which is the difference of the clock drift of the two nodes (between transmitter and receiver node) is degraded thousands of times i.e. for the case of ETS the clock drift affect is magnified by $\Psi_{rx}^{tx} \approx (\delta_{tx} - \delta_{rx}) \frac{t_{tx}}{T_{eq,rx}^{tx}}$ (refer chapter 3). However, while using these E-TOA equations for finding the range difference of arrival of a target node we assume the nominal transmitter frequency to find the equivalent time resolution. Since the clock drift error is magnified to thousands of times, it is desirable to estimate the precise transmitter PRF as well as receiver ADC sampling frequency in the order of sub-nanoseconds. Moreover, since the reconstructed signal obtained by ETS technique is severely affected by the variation in transmitter (or target node) clock drift, a frame level drift estimation is required to accommodate different/multiple target node at each frame and to obtain the precise E-TOA information.

In this chapter, we therefore use a novel mathematical E-TOA model to propose a linear clock skew estimation methodology which is subsequently used to find the accurate equivalent range difference of arrival. The precise receiver clock drift estimate is first used to tune the readers to a common frequency. Additionally, the precise estimate of transmitter clock drift is further used to find the actual equivalent time resolution of the recovered equivalent time sampled signal. This is used to correct the E-TOA and which is further utilized to estimate the actual equivalent time dif-

ference of arrival (termed E-DTDOA as obtained from E-TOA) which can be further used to estimate the position of the target node. The transmitter and receiver clock estimate is verified using simulations and, experimentally using in house designed components. We show here that very precise clock frequency synchronization, in the order of sub-nanoseconds, can be achieved. We highlight that the precise synchronization is possible while still using low cost crystal oscillators thus avoiding the need of costly equipment and devices. Moreover, the clock offset is canceled by performing E-DTDOA mechanism. Further, it is shown using simulations and experimental demonstrations that using the estimated clock parameters accurate positioning of the target node can be obtained. To this end, we also highlight that the scheme proposed in this work allows efficient use of low cost anchor nodes and target nodes with poor crystal clock and larger time drift for precise localization. The rest of the chapter is organized as follows. Sec.4.2 details the system model and the proposed algorithm. Sec.4.3 discusses the simulations results followed by experimental results in Sec.4.4 and Sec.4.5. Sec.4.6 discusses the comparisons with some state-of-the-art methods and systems. The chapter is concluded in Sec.4.7.

Notations: The variables indicated in small letters denote the reference time (nominal value). The variable in small letter with $(\tilde{\bullet})$ denotes the skewed time with respect to the reference time (referred as actual value). The variables indicated in capital letter denote the equivalent time having both nominal and actual values. The superscript and the subscript on the variables i.e. i and j denote transmitter and receiver. For better readability, the comprehensive list of notations used in the below sections are given in Table 4.1.

4.2 System Model

The system model is shown in Fig.4.1. We consider a WSN having n_j anchor nodes with index $j = 1$ reserved for master anchor node and $j \in \{2, 3, \dots, s\}$ for slave anchor nodes. We have $tx_i, i \in \{1, 2\}$ transmit only target nodes. All the nodes operate asynchronously with their respective local clock. The master anchor node is assumed to have an accurate (stable) internal clock oscillators and serve as a timing reference. All other node's, i.e. slave anchor nodes and target nodes, local clock is affected by the imperfection in the clock parameters i.e. clock drift (frequency imperfections) and clock offset (phase imperfections).

The target nodes tx_i transmits UWB-IR ranging signal at each rising/falling edge of the clock. The rising/falling edge of the clock repeats at a nominal pulse repetitive

Figure 4.1: System model with anchor nodes and target nodes

frequency (PRF), f_i . The receiver node (or anchor node) on the other hand captures the transmitted UWB-IR at each digital tick of the clock of analog to digital converter (ADC), which operates at a sampling frequency of $f_j (< f_i)$. The receiver (or anchor nodes) uses ETS technique to reconstruct the signal. The illustration of the CIR of the recovered signal, in an equivalent time frame¹, from two transmitters (tx_1 and tx_2) is shown in Fig.4.2.

4.2.1 E-TOA Modeling for ETS based System

The equivalent time of arrival (E-TOA) of the signal, from transmitter node i , to the receiver node j , for k^{th} equivalent time frame, $\hat{\zeta}_{i,j}^{(k)}$, can be assumed to be as [Chapter 3 eq. (3.20)]

$$\hat{\zeta}_{i,j}^{(k)} = \hat{K}_{i,j}^{(k)} * T_{eq,j}^i + q_{i,j}^{(k)}, \quad (4.1)$$

where the corresponding equivalent time samples, $\hat{K}_{i,j}^{(k)}$, is denoted by

¹It is to be noted that a recovered equivalent time frame consists of many equivalent time pulses. In this work we consider a frame which consists of signal from two transmitters transmitting one after the other at different time slots. We follow the transmission pattern with node tx_1 transmitting first followed by node tx_2 . A frame in Fig.4.2 shows 2 equivalent time pulses from both the transmitters.

Table 4.1: List of Notations Used ($()^{(k)}$ superscript k in the table denotes k^{th} equivalent time frame)

Notations	Definition
$t_i = \frac{1}{f_i}$	Nominal pulse repetitive interval (PRI) of the i^{th} transmitter clock
$t_j = \frac{1}{f_j}$	Nominal ADC sampling interval of the j^{th} receiver clock
$T_{eq,j}^i = (t_j - t_i)$	Nominal equivalent time resolution between tx node i and the rx node j
$\alpha_i^{(k)} = 1 + \delta_i^{(k)}$	$\delta_i^{(k)}$ is i^{th} transmitter clock drift in parts per million (ppm)
$\tilde{t}_i^{(k)} = \frac{1}{\tilde{f}_i^{(k)}} = \frac{t_i}{\alpha_i^{(k)}}$	Actual pulse repetitive interval (PRI) of the i^{th} transmitter clock
$\alpha_j^{(k)} = 1 + \delta_j^{(k)}$	$\delta_j^{(k)}$ is j^{th} receiver clock drift in parts per million (ppm)
$\tilde{t}_j^{(k)} = \frac{1}{\tilde{f}_j^{(k)}} = \frac{t_j}{\alpha_j^{(k)}}$	Actual ADC sampling interval of the j^{th} receiver clock
$\tilde{T}_{eq,j}^i = \left(\frac{t_j}{\alpha_j^{(k)}} - \frac{t_i}{\alpha_i^{(k)}} \right)$	Actual equivalent time resolution between tx node i and the rx node j
$t_{start,i}^{(k)}$	Start time of the i^{th} transmitter's first UWB-IR pulse
$t_{cap,j}^{(k)}$	Clock offset at the j^{th} receiver clock or start of the ADC sample capture window
$t_{i,j}^{(k)}$	Propagation delay from node i to node j
$t_{c,j}^{(k)}$	Circuit delay at the j^{th} receiver
$\hat{K}_{i,j}^{(k)}$	Equivalent time samples corresponding to equivalent time of arrival (E-TOA)
$\Psi_i^{(k)}$	Equivalent time error due to the i^{th} transmitter clock drift
$\Psi_j^{(k)}$	Equivalent time error due to the j^{th} receiver clock drift
$\Psi_{ji}^{(k)} = \Psi_j^{(k)} - \Psi_i^{(k)}$	Equivalent time error due to the transmitter and receiver clock drift

$$\hat{K}_{i,j}^{(k)} = \left\lfloor \frac{(t_{start,i}^{(k)} - t_{cap,j}^{(k)} + t_{i,j}^{(k)} + t_{c,j}^{(k)})(1 + \delta_j^{(k)})}{t_j} \right\rfloor + \left[\text{mod} \left((t_{start,i}^{(k)} - t_{cap,j}^{(k)} + t_{i,j}^{(k)} + t_{c,j}^{(k)}), t_j / (1 + \delta_j^{(k)}) \right) + p \tilde{t}_i^{(k)} \right] \times \frac{1}{T_{eq,j}^i (1 + \Psi_{ji}^{(k)})} + w_{i,j}^{(k)}, \quad (4.2)$$

with $q_{i,j}^{(k)} = w_{i,j}^{(k)} \times T_{eq,j}^i$ is the measurement error with zero mean and variance σ_q^2 , with $w_{i,j}^{(k)}$ denotes the measurement error in samples. $t_{start,i}^{(k)}$, $t_{cap,j}^{(k)}$, $t_{i,j}^{(k)}$, $t_{c,j}^{(k)}$, $\delta_j^{(k)}$, t_j , $\tilde{t}_i^{(k)}$, $T_{eq,j}^i$, $\Psi_{ji}^{(k)}$ in (4.2) are given in Table. 3.1. p is the index of recovered equivalent time pulse in the captured frame (Fig.4.2 shows the recovered equivalent time sampled signal with $(p = 0)$ and $(p = 1)$ for the nodes tx₁ and tx₂). $\Psi_{ji}^{(k)} = \Psi_j^{(k)} - \Psi_i^{(k)}$, $\Psi_{ji}^{(k)}$ is the E-TOA error due to the transmitter and receiver clock drift [182]. $\Psi_i^{(k)}$ and

$\Psi_j^{(k)}$ is given by

$$\begin{aligned}\Psi_i^{(k)} &= \frac{\Delta t_i}{(t_j - t_i)} = \frac{(\tilde{t}_i^{(k)} - t_i)}{T_{eq,j}^i} = \frac{-\Delta f_i^{(k)} / f_i^2}{\left(\frac{1}{f_j} - \frac{1}{f_i}\right)} = \frac{-\delta_i^{(k)} \times t_i}{T_{eq,j}^i} \\ \Psi_j^{(k)} &= \frac{\Delta t_j}{(t_j - t_i)} = \frac{(\tilde{t}_j^{(k)} - t_j)}{T_{eq,j}^i} = \frac{-\Delta f_j^{(k)} / f_j^2}{\left(\frac{1}{f_j} - \frac{1}{f_i}\right)} = \frac{-\delta_j^{(k)} \times t_j}{T_{eq,j}^i}\end{aligned}\quad (4.3)$$

where the notations used in (4.3) are given in Table. 3.1. For notational convenience in the subsequent sections we represent the nominal and the actual equivalent time resolution $T_{eq,j}^i$, $\tilde{T}_{eq,j}^{i(k)}$, as equation

$$T_{eq,j}^i = (t_j - t_i); \quad \tilde{T}_{eq,j}^{i(k)} = \left(\frac{t_j}{\alpha_j^{(k)}} - \frac{t_i}{\alpha_i^{(k)}} \right) = \frac{\left(\frac{1}{\tilde{f}_i^{(k)}} \right)}{\tilde{F}_j^{i(k)}}, \quad (4.4)$$

where, $\tilde{F}_j^{i(k)}$ is the actual pulse repetitive sample points between two recovered equivalent time pulse from node i to node j . Moreover, for ETS $\frac{t_j}{\alpha_j^{(k)}} > \frac{t_i}{\alpha_i^{(k)}}$, $t_j > t_i$ & $\left(\frac{t_j}{\alpha_j^{(k)}} - \frac{t_i}{\alpha_i^{(k)}} \right) \leq \frac{1}{2F_p}$, where F_p is the bandwidth of transmitted pulse [183]. In addition, we assume constant clock drift over a period of a frame capture.

Remark: In (4.1), to represent the equation in time we multiply it with the nominal equivalent time resolution ($T_{eq,j}^i$). However, the actual equivalent time resolution will be different due to the variations in the nominal clock frequencies. In the subsequent sections, we will present the mechanism to estimate the actual frequencies and subsequently the actual equivalent time resolution to find the actual E-TOA.

4.2.2 Clock Frequency Synchronization Mechanism

From the observed k^{th} equivalent time frame (c.f. Fig.4.2), the pulse repetitive interval (PRI) between two consecutive pulses, obtained from the i^{th} transmitter node to the j^{th} receiver node can be given by subtracting two E-TOA's

$$\hat{\xi}_{i,j}^{(k)} = \left(\hat{\zeta}_{i,j}^{(k)}(p=1) - \hat{\zeta}_{i,j}^{(k)}(p=0) \right), \quad (4.5)$$

Figure 4.2: Representation of one ranging cycle and reconstructed k^{th} equivalent time signal at a receiver node/anchor node j with signal from tx_1 ($i = 1$) and tx_2 ($i = 2$)

Substituting (4.1) in (4.5) and rearranging, we get a equation in terms of the nodes clock parameters, and which is given by

$$\hat{\xi}_{i,j}^{(k)} = t_i(1 - \delta_i^{(k)}) \times \frac{1}{(1 + \Psi_j^{(k)} - \Psi_i^{(k)})} + v_{i,j}^{(k)}, \quad (4.6)$$

where, $v_{i,j}^{(k)} = (q_{i,j}^{(k)} - \tilde{q}_{i,j}^{(k)})$ is the noise term and assumed to be a zero mean Gaussian RV with variance σ_v^2 . Substituting $\Psi_j^{(k)}$ and $\Psi_i^{(k)}$ from (4.3) and rearranging, eq. (4.6) can be further written in terms of $\delta_i^{(k)}$ and $\delta_j^{(k)}$ as

$$\left(\hat{\xi}_{i,j}^{(k)} \frac{t_i}{T_{eq,j}^i} + t_i \right) \delta_i^{(k)} - \frac{\hat{\xi}_{i,j}^{(k)} t_j}{T_{eq,j}^i} \delta_j^{(k)} = (t_i - \hat{\xi}_{i,j}^{(k)}) + \tilde{v}_{i,j}^{(k)}, \quad (4.7)$$

where, $\tilde{v}_{i,j}^{(k)} = v_{i,j}^{(k)}(1 + \Psi_j^{(k)} - \Psi_i^{(k)}) = v_{i,j}^{(k)} \left(1 + \frac{\delta_i^{(k)} \times t_i}{T_{eq,j}^i} - \frac{\delta_j^{(k)} \times t_j}{T_{eq,j}^i} \right)$. Since, the term $v_{i,j}^{(k)} \left(\frac{\delta_i^{(k)} \times t_i}{T_{eq,j}^i} - \frac{\delta_j^{(k)} \times t_j}{T_{eq,j}^i} \right)$ is very small compared to $v_{i,j}^{(k)}$, for simplicity we ignore $v_{i,j}^{(k)} \left(\frac{\delta_i^{(k)} \times t_i}{T_{eq,j}^i} - \frac{\delta_j^{(k)} \times t_j}{T_{eq,j}^i} \right)$

$$\underbrace{\begin{bmatrix} \hat{\xi}_{i,j}^{(k)} - t_i \\ \hat{\xi}_{i,j}^{(k)} - t_i \\ \hat{\xi}_{i,j}^{(k)} - t_i \\ \hat{\xi}_{i,j}^{(k)} - t_i \\ \hat{\xi}_{i,j}^{(k)} - t_i \\ \hat{\xi}_{i,j}^{(k)} - t_i \\ \hat{\xi}_{i,j}^{(k)} - t_i \\ \hat{\xi}_{i,j}^{(k)} - t_i \end{bmatrix}}_{x^{(k)}} = \underbrace{\begin{bmatrix} \hat{\xi}_{1,1}^{(k)} \frac{t_j}{T_{eq}} & 0 & 0 & 0 & -(\hat{\xi}_{1,1}^{(k)} \frac{t_i}{T_{eq}} + t_i) & 0 \\ \hat{\xi}_{2,1}^{(k)} \frac{t_j}{T_{eq}} & 0 & 0 & 0 & 0 & -(\hat{\xi}_{2,1}^{(k)} \frac{t_i}{T_{eq}} + t_i) \\ 0 & \hat{\xi}_{1,2}^{(k)} \frac{t_j}{T_{eq}} & 0 & 0 & -(\hat{\xi}_{1,2}^{(k)} \frac{t_i}{T_{eq}} + t_i) & 0 \\ 0 & \hat{\xi}_{2,2}^{(k)} \frac{t_j}{T_{eq}} & 0 & 0 & 0 & -(\hat{\xi}_{2,2}^{(k)} \frac{t_i}{T_{eq}} + t_i) \\ 0 & 0 & \hat{\xi}_{1,3}^{(k)} \frac{t_j}{T_{eq}} & 0 & -(\hat{\xi}_{1,3}^{(k)} \frac{t_i}{T_{eq}} + t_i) & 0 \\ 0 & 0 & \hat{\xi}_{2,3}^{(k)} \frac{t_j}{T_{eq}} & 0 & 0 & -(\hat{\xi}_{2,3}^{(k)} \frac{t_i}{T_{eq}} + t_i) \\ 0 & 0 & 0 & \hat{\xi}_{1,4}^{(k)} \frac{t_j}{T_{eq}} & -(\hat{\xi}_{1,4}^{(k)} \frac{t_i}{T_{eq}} + t_i) & 0 \\ 0 & 0 & 0 & \hat{\xi}_{2,4}^{(k)} \frac{t_j}{T_{eq}} & 0 & -(\hat{\xi}_{2,4}^{(k)} \frac{t_i}{T_{eq}} + t_i) \end{bmatrix}}_H \underbrace{\begin{bmatrix} \delta_{j=1}^{(k)} \\ \delta_{j=2}^{(k)} \\ \delta_{j=3}^{(k)} \\ \delta_{j=4}^{(k)} \\ \delta_{i=1}^{(k)} \\ \delta_{i=2}^{(k)} \end{bmatrix}}_{\theta^{(k)}} \quad (4(A))$$

$-\frac{\delta_j^{(k)} \times t_j}{T_{eq,j}^i}$) and assume $\tilde{v}_{i,j}^{(k)} \approx v_{i,j}^{(k)}$. The clock drift ratios $\delta_i^{(k)}$ and $\delta_j^{(k)}$ have been randomly introduced in the nominal frequencies of the j anchor nodes and i transmitting nodes, respectively, and follows uniform distribution $u(a, b)$. Moving all the unknowns ($\delta_i^{(k)}$ and $\delta_j^{(k)}$) to one side and rearranging (4.7), we get

$$\hat{\xi}_{i,j}^{(k)} - t_i = \frac{\hat{\xi}_{i,j}^{(k)} t_j}{T_{eq,j}^i} \delta_j^{(k)} - \left(\hat{\xi}_{i,j}^{(k)} \frac{t_i}{T_{eq,j}^i} + t_i \right) \delta_i^{(k)} + v_{i,j}^{(k)}. \quad (4.8)$$

Equation (4.8) is formulated for each $i^{\text{th}} - j^{\text{th}}$ node pair at all the participating anchor nodes. Thus, for the k^{th} equivalent time frame recorded at all the anchor nodes, observed PRI values using (4.8) are combined and a generalized linear model can be formulated which is given by

$$x^{(k)} = H^{(k)} \theta^{(k)} + v_{i,j}^{(k)}, \quad (4.9)$$

where, $x^{(k)} = (\hat{\xi}_{i,j}^{(k)} - t_i) \in \mathbb{R}^{(tx_i n_j) \times 1}$, and $H^{(k)} \in \mathbb{R}^{(tx_i n_j) \times (tx_i + n_j)}$ is the coefficient matrix of $\theta^{(k)}$. $\theta^{(k)}$ is a $(tx_i + n_j) \times 1$ random vector of parameters $[\delta_i^{(k)}, \delta_j^{(k)}]^T \in \mathbb{R}^{(tx_i + n_j) \times 1}$, whose realization is to be estimated. We show, $x^{(k)}$ and $H^{(k)}$, for 2 target node ($i \in \{1, 2\}$) and 4 anchor nodes ($j \in \{1, 2, 3, 4\}$), in eq. 9(A) given on next page. However, the generalization for more number of nodes is straightforward.

Remark: The linear equation in (4.9) is solvable if $H^{(k)}$ is a full rank matrix and invertible. This can be achieved by imposing a constraint by assigning the anchor node 1 as a timing reference with $\delta_{j=1}^{(k)} = 0$. The linear equation in (4.9) can be

solved for $\theta^{(k)}$ by using the solution as

$$\hat{\theta}^{(k)} = (H^{(k)T} H^{(k)})^{-1} H^{(k)T} x^{(k)}. \quad (4.10)$$

Subsequently, from the estimated $\hat{\theta}^{(k)}$, change in clock frequency $\Delta f_i^{(k)}$ and $\Delta f_j^{(k)}$ is obtained by using the relation in (4.3). Thus, actual transmitter and receiver frequencies can be obtained precisely using the method proposed. The proposed scheme is further utilized to obtain the precise ranging and positioning of the target nodes which is explained and demonstrated in the subsequent sections.

Figure 4.3: Clock Mechanism

4.2.3 Ranging and Positioning

Algorithm 4.1, as shown on the next page, is proposed to find the actual equivalent differential time difference of arrival, $\hat{D}_{n_1, n_l}^{tx_1, tx_2(k)}$ with $l \in \{2, 3, \dots, s\}$, for the k^{th} equivalent time frame between transmitting nodes (tx_i), and the participating anchor nodes (n_j). We assume master anchor node (n_1) as the base anchor node to find the differencing across the nodes. Moreover, we assume one of the transmitting node (tx_2) as a reference node whose position is known, and which is subsequently used to find the range difference of a target node (tx_1) from the obtained $\hat{D}_{n_1, n_l}^{tx_1, tx_2(k)}$. The obtained range difference of the target node is subsequently used to determine the position of the target node. We detail the description of finding the range difference

of the target node in this section. From (4.2), the sample index of total equivalent time of arrival between node i and node j can be simplified as

$$\hat{K}_{i,j}^{(k)} = C_{i,j}^{(k)} + \frac{\phi_{i,j}^{(k)}}{\tilde{T}_{eq,j}^{i(k)}} + \frac{t_{i,j}^{(k)}}{\tilde{T}_{eq,j}^{i(k)}} + \frac{t_{c,j}^{(k)}}{\tilde{T}_{eq,j}^{i(k)}} + \frac{p\tilde{t}_i^{(k)}}{\tilde{T}_{eq,j}^{i(k)}} + w_{i,j}^{(k)}, \quad (4.11)$$

The term $\left\lfloor \frac{(t_{start,i}^{(k)} - t_{cap,j}^{(k)} + t_{i,j}^{(k)} + t_{c,j}^{(k)})(1 + \delta_j^{(k)})}{t_j} \right\rfloor$ in (4.2) is an integer number of samples recovered before transmitter tx_i starts transmitting (refer Fig.4.3), and which is denoted by a variable $C_{i,j}^{(k)}$ in (4.11). Thus we have

$$C_{i,j}^{(k)} = \left\lfloor \frac{(t_{start,i}^{(k)} - t_{cap,j}^{(k)} + t_{i,j}^{(k)} + t_{c,j}^{(k)})(1 + \delta_j^{(k)})}{t_j} \right\rfloor. \quad (4.12)$$

$\phi_{i,j}^{(k)}$ in (4.11) is the remaining non-integer offset part as demonstrated in Fig.4.3. While simplifying eq. (4.11), we assume that there is no integer number of ambiguity i.e. the sum of $(\phi_{i,j}^{(k)} + t_{i,j}^{(k)} + t_{c,j}^{(k)})$ lies within one actual PRI ($\tilde{t}_j^{(k)}$) of the receiver and do not cross over to the next PRF cycle. This integer ambiguity can be reduced to minimum by assuming the PRF to be sufficiently large as in our case where f_j is taken as 2 - 3 MHz.

After obtaining the precise tx and rx frequencies as obtained in section II (B), the receivers are tuned to the same frequency, thereby making $\delta_{j=1}^{(k)} = \delta_{j=l}^{(k)} \approx 0$ for all $l \in \{2, 3, \dots, s\}$. Moreover, we get the actual equivalent time resolution ($\tilde{T}_{eq,j}^{i(k)}$) using the relation in (4.4). Thus, after obtaining the actual equivalent time resolution ($\tilde{T}_{eq,j}^{i(k)}$) and using (4.11), we get the actual equivalent time of arrival (E-TOA) between node i and node j as

$$\hat{E}_j^{i(k)} = \left(\hat{K}_{i,j}^{(k)} \times \tilde{T}_{eq,j}^{i(k)} \right), \quad (4.13)$$

substituting $\hat{K}_{i,j}^{(k)}$ from (4.11) in (4.13) to get

$$\hat{E}_j^{i(k)} = \left(C_{i,j}^{(k)} + \frac{\phi_{i,j}^{(k)}}{\tilde{T}_{eq,j}^{i(k)}} + \frac{t_{i,j}^{(k)}}{\tilde{T}_{eq,j}^{i(k)}} + \frac{t_{c,j}^{(k)}}{\tilde{T}_{eq,j}^{i(k)}} + \frac{p\tilde{t}_i^{(k)}}{\tilde{T}_{eq,j}^{i(k)}} + w_{i,j}^{(k)} \right) \times \tilde{T}_{eq,j}^{i(k)}, \quad (4.14)$$

For simplicity, we illustrate the case further assuming $p = 0$ (i.e. we take the first equivalent time pulse) for two transmitting nodes, $i = 1, 2$ and two anchor nodes, $j = 1, 2$. Equivalent time difference of arrival (E-TDOA) at the same anchor node,

$j = 1$, with signals from transmitter $i = 1$ and $i = 2$ is given by

$$\begin{aligned}\hat{T}_{j=1}^{i=1,i=2(k)} &= \left(\hat{E}_1^{1(k)} - \hat{E}_1^{2(k)} \right), \\ &= \left(C_{1,1}^{(k)} \times \tilde{T}_{eq,1}^{1(k)} + \phi_{1,1}^{(k)} + t_{1,1}^{(k)} + t_{c,1}^{(k)} + w_{1,1}^{(k)} \times \tilde{T}_{eq,1}^{1(k)} \right) \\ &\quad - \left(C_{2,1}^{(k)} \times \tilde{T}_{eq,1}^{2(k)} + \phi_{2,1}^{(k)} + t_{2,1}^{(k)} + t_{c,1}^{(k)} + w_{2,1}^{(k)} \times \tilde{T}_{eq,1}^{2(k)} \right). \quad (4.15)\end{aligned}$$

Similarly, equivalent time difference of arrival (E-TDOA) at the anchor node, $j = 2$, with signals from transmitter $i = 1$ and $i = 2$ is given by

$$\begin{aligned}\hat{T}_{j=2}^{i=1,i=2(k)} &= \left(\hat{E}_2^{1(k)} - \hat{E}_2^{2(k)} \right), \\ &= \left(C_{1,2}^{(k)} \times \tilde{T}_{eq,2}^{1(k)} + \phi_{1,2}^{(k)} + t_{1,2}^{(k)} + t_{c,2}^{(k)} + w_{1,2}^{(k)} \times \tilde{T}_{eq,2}^{1(k)} \right) \\ &\quad - \left(C_{2,2}^{(k)} \times \tilde{T}_{eq,2}^{2(k)} + \phi_{2,2}^{(k)} + t_{2,2}^{(k)} + t_{c,2}^{(k)} + w_{2,2}^{(k)} \times \tilde{T}_{eq,2}^{2(k)} \right). \quad (4.16)\end{aligned}$$

Eq. (4.15) and (4.16) are used to find equivalent differential time difference of arrival (E-DTDOA), which is given by

$$\hat{D}_{j=1,2}^{i=1,i=2(k)} = \left(\hat{T}_{j=1}^{i=1,i=2(k)} - \hat{T}_{j=2}^{i=1,i=2(k)} \right), \quad (4.17)$$

Substituting (4.15) and (4.16) in (4.17) and rearranging, we get

$$\begin{aligned}\hat{D}_{j=1,2}^{i=1,i=2(k)} &= \left(C_{1,1}^{(k)} \times \tilde{T}_{eq,1}^{1(k)} - C_{1,2}^{(k)} \times \tilde{T}_{eq,2}^{1(k)} \right) + \left(t_{1,1}^{(k)} - t_{1,2}^{(k)} \right) \left(C_{2,1}^{(k)} \times \tilde{T}_{eq,1}^{2(k)} - C_{2,2}^{(k)} \times \tilde{T}_{eq,2}^{2(k)} \right) \\ &\quad - \left(t_{2,1}^{(k)} - t_{2,2}^{(k)} \right) + \left(w_{1,1}^{(k)} \times \tilde{T}_{eq,1}^{1(k)} - w_{1,2}^{(k)} \times \tilde{T}_{eq,2}^{1(k)} \right) - \left(w_{2,1}^{(k)} \times \tilde{T}_{eq,1}^{2(k)} - w_{2,2}^{(k)} \times \tilde{T}_{eq,2}^{2(k)} \right) \quad (4.18)\end{aligned}$$

Since, the anchor nodes are tuned to the same frequency, i.e. we have $\delta_{j=1}^{(k)} = \delta_{j=2}^{(k)} \approx 0$. Thus, from (4.3), we have $\Psi_j^{(k)} = 0$. Therefore, substituting $\tilde{T}_{eq,j}^{i(k)} = T_{eq,j}^i(1 - \Psi_i^{(k)}) = T_{eq,j}^i \left(1 + \frac{\delta_i^{(k)} \times t_i}{T_{eq,j}^i} \right)$ in (4.18), and rearranging to get

Algorithm 4.1 E-DTDOA Range difference estimation

Input: $\hat{K}_{i,j}^{(k)}$

Output: $\hat{D}_{j=1,l}^{tx_1,tx_2(k)} : l \in \{2, 3, \dots, s\}$

```

1: Estimate  $\hat{\theta}^{(k)}$  using (4.10)
2: Tune the readers to the same frequency;  $\delta_{j=1}^{(k)} = \delta_{j=l}^{(k)} \approx 0$ 
3: for  $i \leftarrow 1$  to 2 do
4:   for  $j \leftarrow 1$  to  $s$  do
5:      $\tilde{T}_{eq,j}^i(k) = \left( \frac{t_j}{\alpha_j^{(k)}} - \frac{t_i}{\alpha_i^{(k)}} \right)$ 
6:      $\hat{E}_j^i(k) = \left( \hat{K}_{i,j}^{(k)} * \tilde{T}_{eq,j}^i(k) \right)$ 
7:     if  $j \geq 2$  then
8:        $l = j$ 
9:        $\hat{T}_{j=1,l}^{i=1(k)} = \left( \hat{E}_{j=1}^i(k) - \hat{E}_l^i(k) \right)$ 
10:      if  $i == 2$  then
11:         $\hat{D}_{j=1,l}^{i=1,i=2(k)} = \left( \hat{T}_{j=1,l}^{i=1(k)} - \hat{T}_{j=1,l}^{i=2(k)} \right)$ 
12:      end if
13:    end if
14:  end for
15: end for

```

$$\begin{aligned}
\hat{D}_{j=1,2}^{i=1,i=2(k)} &= \left(C_{1,1}^{(k)} - C_{1,2}^{(k)} \right) \times T_{eq,j}^i \left(1 + \frac{\delta_{i=1}^{(k)} \times t_1}{T_{eq,j}^i} \right) \\
&\quad - \left(C_{2,1}^{(k)} - C_{2,2}^{(k)} \right) \times T_{eq,j}^i \left(1 + \frac{\delta_{i=2}^{(k)} \times t_2}{T_{eq,j}^i} \right) + \left(t_{1,1}^{(k)} - t_{1,2}^{(k)} \right) \\
&\quad - \left(t_{2,1}^{(k)} - t_{2,2}^{(k)} \right) + \left(w_{1,1}^{(k)} \times \tilde{T}_{eq,1}^1(k) - w_{1,2}^{(k)} \times \tilde{T}_{eq,2}^1(k) \right) - \left(w_{2,1}^{(k)} \times \tilde{T}_{eq,1}^2(k) - w_{2,2}^{(k)} \times \tilde{T}_{eq,2}^2(k) \right),
\end{aligned} \tag{4.19}$$

To further simplify (4.19), we revisit (4.12). Since $\delta_j^{(k)} \approx 0$, and the sum $(t_{i,j}^{(k)} + t_{c,j}^{(k)})$ is in the order of few nanoseconds i.e. $(t_{i,j}^{(k)} + t_{c,j}^{(k)} < t_j)$, and for practical cases, the difference $(t_{start,i}^{(k)} - t_{cap,j}^{(k)})$ is assumed to be in the order of few tens of micro-seconds, thereby making $(t_{start,i}^{(k)} - t_{cap,j}^{(k)}) > t_j$ (since t_j is few sub-microseconds). Therefore, (4.12) can be approximately written as $C_{i,j}^{(k)} \approx \left\lfloor \frac{(t_{start,i}^{(k)} - t_{cap,j}^{(k)})}{t_j} \right\rfloor$. Moreover, since the signal transmitted from the two transmitters are captured in the single equivalent time frame, the difference between the integer number of sample across the anchor nodes when calculated from the same transmitter node, i.e. $(C_{1,1}^{(k)} - C_{1,2}^{(k)})$, is equal to the integer number of sample across the same set of anchor nodes when calculated

from the another transmitter node, i.e. $(C_{2,1}^{(k)} - C_{2,2}^{(k)})$. Mathematically, this can be written as

$$(C_{1,1}^{(k)} - C_{1,2}^{(k)}) = (C_{2,1}^{(k)} - C_{2,2}^{(k)}) = C^{(k)} \approx \left[\frac{(t_{cap,1}^{(k)} - t_{cap,2}^{(k)})}{t_j} \right], \quad (4.20)$$

Using (4.20), and since the nominal pulse repetitive duration of the transmitting nodes is same, i.e. $t_1 = t_2$, (4.19) can be written as

$$\begin{aligned} \hat{D}_{j=1,2}^{i=1,i=2(k)} = & C^{(k)} \times \left((\delta_{i=1}^{(k)} - \delta_{i=2}^{(k)}) \times t_1 \right) + (t_{1,1}^{(k)} - t_{1,2}^{(k)}) - (t_{2,1}^{(k)} - t_{2,2}^{(k)}) \\ & + (w_{1,1}^{(k)} - w_{1,2}^{(k)}) \left(1 + \frac{\delta_{i=1}^{(k)} \times t_1}{T_{eq,j}^i} \right) - (w_{2,1}^{(k)} - w_{2,2}^{(k)}) \left(1 + \frac{\delta_{i=2}^{(k)} \times t_2}{T_{eq,j}^i} \right). \end{aligned} \quad (4.21)$$

From (4.21), the range difference of the target node, $(t_{1,1}^{(k)} - t_{1,2}^{(k)})$, is obtained assuming one node as a reference node with known position. In addition, we observe that $C^{(k)}$ in (4.21) is mainly affected by the variation in $t_{cap,j}^{(k)}$ (from equation (4.20)). It is also noted that the difference between the start of the capture frame at different anchor nodes, $(t_{cap,1}^{(k)} - t_{cap,2}^{(k)})$, can be in the order of few tens of microseconds. In this work, we assume this difference to be anywhere between $[1 - 20]$ μ S, which implies $C^{(k)} = [3 - 60]$ samples for $f_j = 3$ MHz. With the maximum, $C^{(k)} = 60$, samples, and assuming the variation in transmitters clock, $\delta_{i=1}^{(k)}$ and $\delta_{i=2}^{(k)}$ to be ± 10 ppm, the maximum error by the term, $(C^{(k)}) \times ((\delta_{i=1} - \delta_{i=2}) \times t_1)$, in (4.21), can be calculated as $(60) * ((\pm 20 \text{ ppm}) \times (1/3 \text{ MHz})) = \pm 0.4 \text{ nS}$. Thus, the maximum ranging error caused due to the variation in start of the capture frame, within $[1 - 20]$ μ S, assuming tx clock variation of ± 10 ppm, can be $\pm 12 \text{ cm}$ in addition to the measurement error as given in (4.21).

4.3 Simulation Results & Discussions

A simulation of a network with 2 transmitting nodes (1 target and 1 reference node) and 4 anchor nodes is considered. The transmitting nodes transmits the UWB ranging signal one after the other. Each transmitting node transmits signal for the sufficient time (which varies depending on the transmission PRF) so as to recover at-least 2 equivalent time pulses at the receiver from the same transmitter. A frame is observed when two transmitters stop transmitting (c.f. Fig.4.2). The simulation has been performed using the topology shown in Fig.4.4. We simulate for two

Figure 4.4: Scatter plots of the error ellipses for the target node at different positions. The anchor nodes are denoted by (■), the reference node by (△) and the true location of target node (●). For both the cases the plot shows the position for different measurement errors, i.e. $\sigma_z = 0.11$ nS denoted by dash line (---), 0.33 nS denoted by dot line (.....), and 0.55 nS denoted by dash-dot line (-...)

different cases: Case 1 with nominal frequencies of the nodes, i.e. $f_i = 2$ MHz and $f_j = 1.999500$ MHz resulting in nominal equivalent time resolution $T_{eq,j}^i = 0.125$ nS. Case 2, with $f_i = 3$ MHz and $f_j = 2.999010$ MHz resulting in $T_{eq,j}^i = 0.110$ nS. However, these nominal frequencies are affected by the drift of the local oscillators and hence, introduces the clock drift. In simulations, the clock drift ratios $\delta_i^{(k)}$ and $\delta_j^{(k)}$ have been randomly introduced in the nominal frequencies of the nodes and uniformly distributed within $[-100\ 100]$ ppm and $[-10\ 10]$ ppm for the j anchor nodes and i transmitting nodes, respectively. These cases are chosen to be representative of practical implementations of low cost crystal oscillators [90]. At each k^{th} realization, the start time, $t_{start,i}^{(k)}$, of the transmitters tx_1 and tx_2 is randomly selected between $[100\ 150]$ μ S and $[3\ 4]$ mS, respectively. Moreover, $t_{cap,j}^{(k)}$, for all the anchor nodes is randomly selected anywhere between 50 nS – 20 μ S. The root mean square error

(RMSE) of frequency estimate is used to evaluate the performance of the proposed estimator in (4.10). This is given as $RMSE = \sqrt{\frac{1}{N}(\sum \|\Delta f^{(k)} - \hat{\Delta f}^{(k)}\|^2)}$, where N is the number of runs. We use $N = 1000$ realizations for the simulations. For two cases, Fig.4.5 shows RMSE of the estimated frequencies of the transmitting and the anchor nodes vrs. different variations in the receiver clock drift while keeping $\delta_i^{(k)} = [-10 \ 10]$ ppm with measurement noise level $\sigma_q = 0.11$ nS . It is seen that the proposed estimation method achieves good accuracy for different clock drift variations. The proposed method is used to estimate the precise frequencies of the nodes which is used in algorithm 4.1. Algorithm 4.1 is employed, and Fig.4.4 is plotted after calculating the position of the target nodes. 4 different static position of the target node is demonstrated in the figure for both the cases. The blue and black ellipses shows the error in the positioning without correction (i.e. using the nominal frequencies of the nodes) for case 1 and case 2, respectively. While, the green and red ellipses indicate the performance of the positioning employing algorithm 4.1 for case 1 and case 2, respectively. The standard deviation of the estimated x and y coordinates (σ_x, σ_y), in meters, for different σ_q (in nanoseconds) for case 2 is displayed in Table. 4.2.

Table 4.2: Standard deviation of the estimated x and y coordinates for different target node positions as in Fig.4.4

Target Node Position	σ_w nS	σ_x (m)	σ_y (m)
$(x, y) = (8, 8)$	0.11	0.039	0.044
	0.33	0.114	0.116
	0.55	0.187	0.194
$(x, y) = (6, 3)$	0.11	0.035	0.043
	0.33	0.102	0.113
	0.55	0.170	0.180
$(x, y) = (2, 2)$	0.11	0.036	0.042
	0.33	0.106	0.108
	0.55	0.177	0.182
$(x, y) = (2, 8)$	0.11	0.038	0.044
	0.33	0.108	0.118
	0.55	0.176	0.187

Figure 4.5: RMSE of clock frequency estimate (in Hz), or clock drift, for the anchors and transmitting nodes with $\delta_i = [-10 \ 10]$ ppm. The value ‘ x ’ of δ_j on x-axis denotes uniform distribution between $[-x \ x]$ ppm.

4.4 Experimental Demonstration

In this section we experimentally demonstrated and verify the proposed algorithm 4.1 using in house designed hardware components. The details of the hardware components are given in [90]. The experimental setup is shown in Fig.4.6 with four anchor nodes (AN), one target node (TN) and one reference node (RN). The RN and the TN is collocated at anchor node 1 and the anchor node 3, respectively, as shown in Fig.4.6. All the anchor nodes, target node, and a reference node are running on its local clock. The specifications of the target node and reference node crystal oscillator is given in [180]. To provide a cost effective solution, we employ an ETS technique at the receiver (anchor nodes) by using a low cost ADC to sample the repetitive UWB envelopes. Since the sampling of UWB envelop is based on ETS, a lower sampling clock frequency is used, which is generated by the digital direct synthesizer (DDS) [179] that utilizes 75 MHz crystal oscillator [181]. The nominal sampling frequency

Figure 4.6: Experimental test-bed measurement environment and system. Figure shows the placement of anchor nodes (AN), target node (TN) and reference node (RN).

used for the experiments was $f_j = 1.999500$ MHz. During the experiments, the TX nominal pulse repetitive clock frequency was adjusted at $f_i = 2$ MHz which results in a nominal equivalent time resolution of 0.125 nS. Moreover, the target node and the reference node were programmed to transmit the UWB ranging signal for the period of 4 mS. The transmission scheme used in this work is with target node transmitting the signal first followed by the reference node. The equivalent time frame is captured from both the nodes as shown in Fig.4.2. From the recovered k^{th} equivalent time frame, the sample index as in eq. (4.2) is obtained for all the target-anchor and reference-anchor pairs, and this is subsequently sent to the central processing unit for further processing.

4.5 Experimental Results

In the following section, clock frequency synchronization mechanism, as discussed in section III (B), followed by the positioning of the target node is demonstrated using the proposed Algorithm-4.1

4.5.0.1 Clock Synchronization Steps and Results

The steps below demonstrates the verification of the proposed synchronization mechanism.

Step 1: The anchor nodes clock frequency was adjusted to the same value at 1999500

Table 4.3: Experimental demonstration of the estimated parameters and the clock frequency adjustment for different runs

Run	Anchor Nodes	Clock Frequency (MHz)	Mean Estimated Clock Parameters (Hz)	Mean Estimated Clock Parameters (Hz)
Run 1	AN 1	1.999500	0	$TX_1 = -8.10$
	AN 2	1.999500	8.97	
	AN 3	1.999500	2.36	$TX_2 = -8.54$
	AN 4	1.999500	1.38	
Run 2	AN 1	1.999500	0	$TX_1 = -7.79$
	AN 2	1.999509	0.067	
	AN 3	1.999503	-0.088	$TX_2 = -8.19$
	AN 4	1.999501	0.651	
Run 3	AN 1	1.999492	0	$TX_1 = 0.46$
	AN 2	1.999501	0.084	
	AN 3	1.999495	-0.311	$TX_2 = -0.06$
	AN 4	1.999493	0.418	

Hz by tuning the DDS². Column III of Table.4.3 shows the adjusted frequency in different runs of the experiment. After the anchor nodes are set to this frequency, the transmitting nodes TX_1 and TX_2 transmits the UWB-IR ranging signal. The change in clock frequency, i.e. $\Delta f_i^{(k)}$ and $\Delta f_j^{(k)}$, by using the method proposed in section III (B) was estimated. We repeat the experiment with the same settings multiple times and the mean value of the estimated clock frequency is obtained. Table.4.3 shows the mean value of the estimated clock parameters, in Hz, for the anchor nodes, in Column 4, and for the target node (TX_1) and a reference node (TX_2), in Column 5. We consider anchor node 1 as the reference AN. 4.7a and 4.7b shows the distribution of the difference in the PRF sample points between a pair of anchor nodes as obtained from TX_1 and TX_2 , respectively. For instance, a difference of 72 sample points between $AN_1 - AN_2$ w.r.t TX_1 is observed from 4.7a. These sample point difference corresponds to the timing error of $72 \times 0.125E - 09 = 9E - 09$ sec (assuming the nominal equivalent time resolution), which equals to 2.7 meter of

²For DDS AD9834 the frequency adjustment resolution of 0.28 Hz can be achieved. However, for the experiments we adjusted the frequency of the nodes with 1 Hz frequency resolution.

Figure 4.7: Experimental results of sample point difference between the anchor nodes without clock drift compensation.

ranging error.

Step 2: In run 2, the anchor nodes clock frequency was tuned by the corresponding estimated parameter value as obtained in run 1. The AN2 clock frequency was adjusted to 1.999509 MHz since the observed change in frequency during run 1 for AN2 was 8.97 Hz (refer Table.4.3 column 4) . Similarly, it was done for AN3 and AN4 and the newly adjusted values are displayed in column 3, run 2 of Table.4.3. The experiment was again repeated with the changed frequency values to obtain $\hat{\theta}^{(k)}$, $\Delta f_i^{(k)}$ and $\Delta f_j^{(k)}$. The corresponding estimated mean values for run 2 are reported in column 4 and column 5 of Table.4.3. For run 2, the distribution of the difference in the PRF sample points, calculated between the pair of anchor nodes, with respect to TX₁ and TX₂ is shown in 4.8a and 4.8b, respectively. It is observed that the difference in the sample points between the pair of nodes is reduced. The average sample point difference between all the pair of anchor nodes, as seen from 4.8a and 4.8b, is now 7 sample points and which is equivalent to $7 \times 0.125E-09 = 0.875$ nS. The reduction in the sample point helps verifying the estimated clock parameters

Figure 4.8: Experimental results of sample point difference after anchor nodes clock drift is tuned.

for the anchor nodes. 4.7c and 4.8c is plotted to show the variation in the TX node clock frequency while using ETS technique. Ideally, when master anchor node 1 is set at frequency 1.999500 MHz, and if transmitting PRF of TX node is exactly 2 MHz, the nominal pulse repetitive sample points between two recovered equivalent time pulses is 4000 samples (from relation in (3.13)). Thus, from 4.7c and 4.8c, we can see the variation in the pulse repetitive sample points at the reference anchor node 1 w.r.t the TX₁ and TX₂, when compared to the nominal pulse repetitive sample points. These variations are accounted due to the variations in the TX clock actual frequency. Step 3: To verify the observed variation in TX node clock frequency and the estimated TX clock parameters, we adjusted the anchor nodes frequency with respect to one of the transmitter node and then observe the pulse repetitive sample points. For instance, it was observed that the estimated variation in frequency for TX₁ and TX₂ in run 2 was TX₁ = -7.79 Hz and TX₂ = -8.19 Hz. Therefore, we adjusted the anchor nodes frequency by rounding it to -8 Hz to the frequency set in run 2. For run 3, the new adjusted frequency of the anchor nodes

Figure 4.9: Frequency synchronized experimental results after target node estimated drift is adjusted.

are shown in column 3 of Table. 4.3. We repeat the experiment and estimate the change in the clock frequency $\Delta f_i^{(k)}$ and $\Delta f_j^{(k)}$. The distribution of the sample point difference is shown in Fig.4.9. It is observed from 4.9a, 4.9b and 4.9c, that the mean value of the sample point difference is reduced to between 0-5 sample points. This corresponds to the node to node frequency synchronization accuracy of less than $5 \times 0.125E - 09 = 0.625$ nS.

4.5.0.2 Localization Results

The clock frequency of the anchor nodes were adjusted to the values as displayed in run 3, and an equivalent differential time difference of arrival (E-DTDOA) method, as mentioned in algorithm 4.1, is used to find the range difference of the target node (TX₁). This is subsequently used to find the position of the target node (TN). Fig.4.10 demonstrates the localization accuracy of the target node (TN) using the

Figure 4.10: Experimental demonstration of a target node (TN) localization using proposed algorithm 4.1. AN, RN refers to anchor nodes and reference node.

method proposed. The experimental result shows the accuracy of less than 30 cm with frequency adjustment resolution of 1 Hz between the anchor nodes.

4.6 Comparison with State of the Art Systems

In this section, we have compared the proposed system with state of the art systems. For comparison, we consider works which use UWB as a signaling technique. The results of the experiment performed in the previous works are summarized in Table.4.4. The synchronization and the localization accuracy of different existing works are compared with the proposed method. In most of the results published in literature costly equipment such as high speed oscilloscope was often used as a receiver, which gives the opportunity to precisely determine the pulse's TOA. Our proposed system uses low cost components with very low ADC sampling rate while

Table 4.4: Comparison with the existing works

Ref	Synchronization Accuracy	Localization Accuracy	Implementation Architecture	Cost
[184]	2 nSec	-	ASIC based Real Time sampling using ADC at 4 GSps	High
[185]	synchronized	30 cm	FPGA based Real Time sampling using ADC at 2.88 GSps	High
[99]	10 - 15 nSec	-	FPGA based using Adder based clock	High
[152]	5 nSec	-	FPGA based Real Time sampling using ADC at 2.5 GSps	High
[140]	1 nSec	-	Real time using TDC with 100 psec resolution	High
[143]	synchronized	-	Real time with comparator and TDC with 25 psec resolution	High
[101, 122]	sub-nanosecond	-	Real Time using TDC with 100 psec resolution	High
[144]	synchronized	10 cm	FPGA based Equivalent Time sampling	Moderate
[146]	2-10 nSec	-	Equivalent time sampling with no experimental demonstration	-
This work	sub-nanosecond	<30 cm	Equivalent time sampling using ADC at sampling frequency of 2 MHz	Low

still using high bandwidth UWB-IR signals. The performance of the proposed system, i.e. synchronization and localization accuracy, is better as compared to the systems found in the literature [99], [101], [122], [140], [143], [152], [184], [185] which use high cost and high precision oscilloscopes and time to digital converters (TDC). The presented systems in [99], [101], [122], [140], [143], [152], [184], [185] are based on real time sampling, which require costly equipment to obtain precise time of arrival (TOA) while using high bandwidth UWB signal. It is important to note that references [144] and [146] are based on equivalent time sampling (ETS) technique which offers low-cost solution with good synchronization and localization accuracy. [144], however uses a synchronized non-coherent receiver architecture and demonstrates the system using costly equivalent time sampling oscilloscopes. Moreover, no analytical modeling is presented in this work, which is important to validate the experimental results. [146], on the other hand uses simulation without experimental demonstration, to obtain precise TOA using equivalent time sampling technique.

Moreover, this work also lacks in terms of mathematical modeling for precise localization of target node/s. Similar to [144] and [146], our proposed work uses ETS technique. However, we would like to highlight that we present and experimentally demonstrate the equivalent differential time difference of arrival (E-DTDOA)-based system using E-TOA analytical modeling. To the best of authors knowledge, such a system has never been investigated in the literature. Another important benefit of proposed system is that it uses low-cost components with low sampling rate and achieves synchronization accuracy in the order of sub-nanoseconds and localization accuracy less than 30 cm.

4.7 Conclusion

A high time resolution scheme for localization of low-cost asynchronous wireless sensor nodes is proposed using an ETS technique. The proposed scheme achieves accurate drift estimates for the transmitting and receiving nodes. Moreover, clock offset is canceled by using an approach of E-DTDOA. The E-DTDOA approach is used to find the range difference between the target node and anchors, which is subsequently used for finding the position of a target node. The positioning accuracy of less than 30 cm is achieved with 1 Hz clock frequency adjustment resolution. Thus, we envision that the scheme proposed is beneficial in designing precise indoor localization system using high bandwidth signal with efficient use of low cost anchor nodes and target nodes with poor crystal clock and larger time drift.

5 Experimental Evaluation of UWB-IR based Indoor Localization System with low ADC sampling rate

5.1 Introduction

In the previous chapters we have proposed a novel analytical model for obtaining E-TOA for systems employing ETS technique. We have validated and demonstrated the feasibility of the proposed model. We have also seen how we can use these E-TOA to further get E-DTDOA based range measurements. In this chapter, we experimentally demonstrate, using in house designed components, the proposed range measurement scheme for UWB-IR based indoor localization system. The system is specifically designed and developed for low-cost low-power UWB-IR transmit-only tags (target nodes) for ranging. The processing of channel impulse response (CIR) for the precise time of arrival (TOA) of the direct path is the basis for accurate positioning in indoor cluttered environments. In the previous works, following points are observed:

- 1) To measure the precise CIR using impulse-radio UWB signals, bulky devices like oscilloscopes and high sampling rate ADC's were used [31], [143]. In [143], a performance comparison among ten impulse-radio UWB localization systems is given. These systems are laboratory set-ups that use an oscilloscope or squarer/comparator for capturing the emitted signals and estimating the TOA or TDOA.

- 2) With the help of the technological advancement, devices like time to digital converter (TDC) are used to obtain the precise TOA [143]-[101] using time-stamping technique. TDC can measure time-stamps with a precision of sub nano-seconds [98]. However, TDC is not useful to precisely estimate the CIR. The most recent systems

in [143]-[101], demonstrated round trip time measurement technique using TDC at the level of discrete clock ticks.

The use of oscilloscopes and TDC is infeasible for wireless sensor networks due to the bulkiness and high cost of oscilloscopes and expensive TDCs, respectively.

3) Moreover, commercially available systems from Decawave and Bespoon are based on time-stamping based techniques to find the TOA. Such systems can also determine the CIR. However, a very high preamble length (1024 symbols as compared to 128 symbols required for LOS scenario) is required to be able to find the multi path components from CIR to improve the performance in NLOS [186]. Moreover, in such systems both the communication and ranging is performed using UWB signals.

4) As per the best of authors knowledge, the implementation of the ETS technique for precise positioning using CIR for wireless sensor networks has never been demonstrated in the literature.

Below points are highlighted for the implementation of the ETS technique for wireless sensor network:

1) The basis of ETS technique is to reconstruct a single pulse at the receiver using multiple transmitted pulses. This is the first time that the ETS technique is used and modeled for the positioning system to find the precise E-TOA. This technique leverages us to use the ADC with low sampling rate (in the order of 2-3 MHz) at the receiver, which still allows to achieve precise time resolution in the order of pico-seconds. ETS technique used in this work and oscilloscopes used in previous works process CIR to determine precise TOA. However, it is important to note that the utilized low sampling rate ADC in our implementation are quite inexpensive compared to costly oscilloscopes.

2) The implementation of E-TOA and E-DTDOA scheme using impulse radio UWB allows us to use a non-coherent receiver architecture with off the shelf components and some well-known and available software. The implemented receiver architecture uses the conventional receiver components instead of any technologically advanced components.

3) The system integrates sub-1 GHz with impulse radio UWB technology, where former is used for medium access control (MAC) and communication and later is used for the ranging purpose. It is interesting to note that the reception of IR-UWB is more power consuming than transmission [153]. Using this information, the energy efficiency is maintained by operating IR-UWB based target nodes in transmit-only

mode and the reception of the signal is carried out using sub-1-GHz technology. To the best of author's knowledge, this is the first time that the demonstration of a multi target node hybrid system using E-TOA and E-DTDOA technique is presented.

We start by demonstrating the experimental results using the wired architecture where anchor nodes were connected to the central server using wires, thereby relaxing the need to synchronize the clocks. Subsequently, we demonstrate the use of hybrid sub-GHz UWB-IR based wireless indoor positioning system employing ETS technique. We demonstrate the low-cost and low-power high accuracy localization system with support of multi-tag operation using hybrid sub-1 GHz and UWB-IR technology. The scheme presented here integrates the UWB transmitter with an ultra low-power sub-1 GHz transceiver. We have developed a non-coherent indoor UWB positioning system which targets high precision applications. Moreover, to provide a cost effective solution, in our system, as also discussed in previous chapters, we employ ETS technique at the receiver by using a low cost ADC to sample the repetitive UWB envelopes. The advantage of sampling the UWB envelop using ETS technique is that it allows the usage of a lower sampling clock frequency. Such low frequency can be generated by the digital direct synthesizer (DDS) that utilizes 75 MHz crystal oscillator [187]. The sampling frequency used in our application is approximately 2 MHz in wired architecture case while it is 3 MHz for wireless architecture.

The major contributions of this chapter lies in designing and demonstrating the feasible system which can be used for low-cost sensor node. Some of the salient features of the system are: we practically demonstrate a low-cost indoor localization solution using much reduced ADC sampling rate. High time resolution (10 pS–100 pS) is achieved from low sampling rate ADC (as low as in the order of few MHz) through repetitive sampling. Anchor nodes (or receiver nodes) are designed using low cost off the shelf ADC's with sampling between 2-3 MHz, unlike high sampling rate and costly oscilloscopes. The experimental evaluation of the system is presented where the precise ranges are obtained between the target and anchor nodes, and these ranges are further used for calculating the position of the target nodes.

The rest of the chapter is organized as follows: Sec. 5.2 details the wired system architecture and protocol description of DTDOA based range measurement techniques along with the experimental setup and experimental results. Sec. 5.3 details the wireless system architecture. Sec. 5.4 explains the experimental setup with re-

sults and discussions, followed by comparison in Sec. 5.5 and conclusion in Sec. 5.6.

Figure 5.1: Simplified block diagram of the mobile tag and the anchor node

5.2 Wired System Architecture

The architecture of the proposed indoor positioning system consists of a network of fixed anchor nodes, a reference node and a mobile nodes (or target nodes), whose location needs to be determine. The reference node is a stationary device with known location. A simplified block diagram of the target node and anchor nodes is shown in Fig.5.1 and its functionality can be explained as follows. The UHF link initiates the process by broadcasting a beacon to the mobile and a reference nodes. The mobile node and the reference node receives the beacon and sends the acknowledgment along with its ID to anchor nodes. Subsequently, both the nodes start transmitting UWB-IR signal simultaneously. This signal is used for range calculation at the anchor nodes. The anchor nodes acquire the UWB-IR signals from the mobile node and the reference node using ETS technique. The equivalent time sampled signal is then processed at the fusion center to calculate the E-TOA for each received signal and then E-DTDOA. The central server or fusion center runs MATLAB to estimate the location of mobile nodes/tags using TDOA range values computed from the received TOA data. Different components of the system are shown in Fig.5.2.

5.2.1 Hardware Description

5.2.1.1 Mobile Tag

The mobile tag primarily consists of the UWB transmitter and TI RF micro-controller unit (MCU). The TI MCU acts as a controller in the tag hardware. The UWB transmitter consists of two stages, namely an impulse generator stage and a pulse forming stage. The impulse generator is designed to generate a high power sub-nanosecond impulse. The pulse forming network then translates the generated impulse to construct the desired Gaussian UWB pulse with a high peak power of over +20 dBm [188]. The tag transmits the UWB-IR signal with a pulse repetition frequency of around 2 MHz for a duration of 4 ms.

Figure 5.2: Various components of indoor positioning system

5.2.1.2 Anchor Node

The anchor nodes comprises of a UWB front end receiver, followed by a variable gain amplifier and a low cost and low sampling rate ADC which is connected to a Desktop/PC. The UWB front end receiver consists of a UWB antenna, band-pass

Figure 5.3: Experimental setup of indoor positioning system

filter, low noise amplifier and a mixer stage. The locator box contains a base-band amplifier, a low cost ADC and a MCU. The UWB receiver detects the envelop of the UWB pulses transmitted by the mobile node and the reference node. The base-band amplifier in turn, amplifies the UWB envelop before it is sampled by the ADC. Since UWB signals have a large bandwidth, an ADC with a high sampling frequency is required in order to follow the Nyquist sampling theorem. This will eventually lead to a need for expensive ADC chips in the system. To provide a cost effective solution, we employ the ETS technique in our system by using a low cost ADC to sample the repetitive UWB envelops. Since the sampling of UWB envelop is based on ETS, a lower sampling clock frequency is used, which is generated by the digital direct synthesizer (DDS) that utilizes 75 MHz temperature compensated oscillator (TCXO). The sampling frequency used in our application is about 2 MHz. Apart from its higher costs, a high sampling ADC generates a deluge of data which is difficult to handle by the general purpose micro-controller. Thus, using low sampling rate ADC can further help in using low cost micro-controller to process the ADC data and which can be easily used in low-cost low-power sensor nodes.

5.2.2 Experimental Setup and Results

An E-DTDOA is used to find the range difference between the nodes. The localization system comprises of few anchor nodes (AN), mobile nodes (MN) and a reference nodes (RN). The AN's are deployed on the walls of a room as shown in Fig. 5.3a. These AN's are used for localization of a MN (or target node) placed in the room. Some of these anchor nodes are connected, through a UTP cable, to a common ADC locator box 1 (LB1), while remaining others are connected to a similar, but having different ADC, locator box 2 (LB2). Since, two locator boxes (LB1 and LB2) are

Figure 5.4: Illustration of the captured signal from different anchor nodes (receivers)

using different ADC's, they are running on different clock oscillator, thereby having unsynchronized clocks. Thus, AN's connected to the same LB will have the same clock parameters whereas the AN's connected to different LB's will have different clock parameters. These anchor nodes are randomly selected, based on the signal to noise ratio (SNR), for finding the E-TOA of the nodes. The problem however is in synchronizing the AN's selected from the different locator boxes. This synchronization is an important step in determining the accurate TDOA range measurements and to accurately determine the location of the target node. For synchronization of the AN selected from different LB's, we use a common reference node (RN). The position of the RN is predetermined. The RN is within the range of all asynchronous AN's participating in the localization of target node/s. The transmitted signal from the RN and MN signal is received at the respective anchor nodes of different LB. For illustration we have shown the captured signal in Fig.5.4. We perform two time difference, first, between the signals received from two transmitters (i.e. mobile/target node denoted as tx_1 and, reference node denoted as tx_2) at the same anchor

Figure 5.5: Range error distribution plot between AN_2 and AN_4

node (denoted by rx_1) i.e. $TDOA_{rx_1}^{tx_1,tx_2}$ and given by¹

$$TDOA_{rx_1}^{tx_1,tx_2} = TOA_{rx_1}^{tx_1} - TOA_{rx_1}^{tx_2} \quad (5.1)$$

Similarly, we find time difference of arrival at another anchor node (denoted by rx_2) i.e. $TDOA_{rx_2}^{tx_1,tx_2}$ and given by

$$TDOA_{rx_2}^{tx_1,tx_2} = TOA_{rx_2}^{tx_1} - TOA_{rx_2}^{tx_2} \quad (5.2)$$

and second, the difference between the TDOA's so obtained at different receivers which is given by

$$DTDOA_{rx_1,rx_2}^{tx_1,tx_2} = TDOA_{rx_1}^{tx_1,tx_2} - TDOA_{rx_2}^{tx_1,tx_2} \quad (5.3)$$

The detailed mechanism of differential time difference of arrival is explained in Chapter 4. Assuming the position of one transmitter (reference node) is known, the range difference of other transmitter (mobile node) from two receivers can be obtained. These ranges can be used to calculate the position of the mobile tag using commonly known recursive least square methods.

¹The mathematical modeling of equivalent differential time difference of arrival is detailed in Chapter 4.

Figure 5.6: Range error distribution plot between AN_5 and AN_4

The experimental set up of the indoor positioning system is shown in Fig. 5.3a. The coverage area for the experiment is approximately about 10 m x 15 m. Six different anchor nodes, AN_1 , AN_2 , AN_3 , AN_4 , AN_5 , AN_6 are used with one mobile/target node (MN) and one reference node (RN) for the experiment. Fig. 5.3b shows the true location of the anchor nodes, mobile node and a reference node. The figure displays the actual coordinates of all the anchor nodes. The actual distances from the mobile node and reference node to all the anchor nodes are also displayed in Fig. 5.3b. Four different anchor nodes, i.e. AN_2 , AN_4 , AN_5 and AN_6 , are selected out of 6 available. These AN are selected based on the best SNR of the received signal among all other nodes. $AN_2 - AN_4$ and $AN_5 - AN_6$ are connected to locator box 1 and locator box 2, respectively.

The experiment was conducted in 2 dimensional where all the anchor nodes were deployed at the same height. As shown in Fig. 5.4, the reference node has a unique waveform as compared to the mobile node. This is implemented in such a way so as to identify the simultaneously received signal from RN and the MN on a same frame. In the experiment, AN_2 was preferred over AN_1 so as to measure the effect of the farthest anchor node from the mobile node. We further note that signal to AN_6 is partially blocked by a metallic pillar as shown in Fig. 5.3b. The UWB-IR signal is transmitted from the mobile node. The pulse repetitive frequency (PRF) of

Figure 5.7: Range error distribution plot between AN_6 and AN_4

the signal was adjusted at 1.625MHz. On the other end, equivalent time sampling is used to sample the UWB-IR signal and for signal reconstruction purposes at each anchor node. During an experiment we set the ADC sampling frequency of 1.6245 MHz. The set PRF and the ADC sampling frequency will result in an equivalent time sampling resolution of around 0.1763 ns. Hence, with uncertainty of 10 sample points, it can be translated to be about 50 cm errors. However, resolution can be adjusted with manipulation of the sampling clock frequency.

Furthermore, for simplification of E-DTDOA calculations, AN_4 is considered as the base anchor node. AN_4 is chosen since it is in direct line of sight and is nearest to the target/mobile node. Thus, it receives the highest SNR. $TDOA_{24}$ is denoted as the range difference between AN_2 and AN_4 . Range difference between $AN_5 - AN_4$ and $AN_6 - AN_4$ is denoted by $TDOA_{54}$ and $TDOA_{64}$, respectively. Fig.5.5, Fig.5.6 and Fig.5.7 shows the error distribution between the measured distance difference and the actual physical distance difference of the mobile node from different anchor nodes. The result is obtained through 1000 measurement data collection. Fig.5.5 shows the mobile range difference error (R_{24}) between AN_2 and AN_4 . Fig.5.6 shows the mobile range difference error (R_{54}) between AN_5 and AN_4 , similarly Fig.5.7 shows the mobile range difference error (R_{64}) between AN_6 and AN_4 . The observed range differences are justified by the Gaussian shape error distribution. The mean

Figure 5.8: Measured position of the mobile node

error computed for R_{24} , R_{54} and R_{64} are 15cm, 53cm and 13cm, respectively. Moreover, the observed error standard deviation obtained for R_{24} , R_{54} and R_{64} is 29.69cm, 33.75cm and 33cm, respectively. We note here the range difference error for R_{54} is 53cm. This is accounted by the fact that the line of sight between receiver anchor node 6 and mobile node is blocked by a metallic pillar as shown in Fig. 5.3b. However, this can be mitigated by using existing NLOS mitigation techniques. Moreover, the error achieved can be reduced to an extent by increasing the sampling resolution of the ADC clock frequency which is demonstrated in the later sections. We further calculate the position of the mobile/target node using the obtained range information. Fig.5.8 shows the measured localization accuracy of the mobile node in an indoor environment. Fig.5.9 further shows the cumulative distribution plot of x-axis and y- axis of the measured position of the mobile nodes. It can be seen that more than 80% of the position points are within 50 cm of accuracy for both the coordinates.

5.3 Wireless System Architecture

The system discussed above was extended to a complete wireless system. The system architecture, as shown in Fig.5.10, consists of reader/anchor nodes incorporated

Figure 5.9: Cumulative distribution of mobile node coordinates

with reference tags, mobile tags and data fusion center. A block diagram with main system components is shown in Fig.5.11. The fabricated tag and the anchor nodes used in this work are shown in Fig.5.12. The overall system functioning can be explained as follows. The master anchor node starts the process by sending a beacon to tags using sub-1 GHz wireless link. After receiving the beacon, the tags transmit the UWB-IR signal in their respective time slots. The anchor nodes upon receiving UWB-IR signal from tags computes the E-TOA ranging information and sends this information to data fusion center through sub-1 GHz wireless link. At the data fusion center, the E-TOA information is processed and E-DTDOA technique is used to compute range difference. With the known positions of anchor nodes along with computed ranges of tag from anchor nodes, the position of tag is determined using recursive least squares method. The position of anchor node is determined through self-calibration mechanism as discussed in our previous work [90].

The architecture for tags used in this system consists of UWB-IR transmitter board, sub-1 GHz ultra low power RF transceiver from Texas Instruments CC1310 and power management circuit. The UWB transmitter consists of an impulse generator which exploits peak power limit of low duty cycle UWB system and a pulse forming network [188] that shapes the UWB pulses to fit in the spectrum as defined in FCC regulations [178]. In order to achieve the low power design requirement for

Figure 5.10: System architecture of Sub-1 GHz UWB-IR based indoor positioning system

Figure 5.11: Block diagram of the tx and rx node.

coin cell battery powered tags, the operation of duty cycling is performed through CC1310 and thus providing the function for power management within tag. Also, to meet the requirement for low power design, the tags were designed to operate in UWB-IR transmit mode as UWB-IR receiver is likely to consume more power during operation. Apart from providing power management functionality within tags, CC1310 also maintains sub-1 GHz wireless link with anchor nodes.

We use anchor nodes with UWB transmitter along with sub-1 GHz transceiver. The UWB envelop signal constructed by the front end UWB receiver is amplified through a base-band amplifier before analog to digital conversion (ADC). The ADC uses the sampling clock of around 3 MHz which is generated by the digital direct synthesizer (DDS). The DDS utilizes 75 MHz crystal oscillator for deriving the ADC sampling clock. Through repetitive sampling, high time resolution (in the order of

Figure 5.12: Fabricated tag and anchor nodes used for the experiment

picoseconds) can be achieved from a low sampling frequency ADC clock. The time of arrival estimation is done from the recovered equivalent time sampled signal (c.f. Fig.5.14) using Programmable Real Time Unit (PRU) in BeagleBone Black (BBB) embedded Linux device [90], [187]. Since our scheme uses reference tag for the positioning purpose, we collocate the reference tag at the anchor nodes. The overall specifications of the hybrid sub-1 GHz-UWB based multi-tag system is shown in Table. 5.1. It is notable that the power consumption of the tag is dependent on the update rate. Power consumption of the sub-1 GHz transceiver is less than 25 mW and 80 mW in TX and RX mode, respectively. Whereas, the power consumption of the UWB transmitter is less than 9 mW in active mode and less than 0.3 mW at 1Hz update rate.

5.3.1 Multi-Tag Access Scheme

The different steps of the access schemes are summarized in Fig.5.13. The overall system works in two different modes:

- 1) MAC mode: A time division multiple access scheme is utilized in the system. The master anchor nodes initiates the process by transmitting a beacon using a sub-1 GHz band with frequency (F_1) of 922 MHz. The beacon on the other hand is received by the active target tags in the network. The tags send an acknowledgment packet appending the randomly generated transmission slot number along with the corresponding tag ID using a different frequency (F_2) of 922.7 MHz. The master tag upon receiving the acknowledgment verifies the slot number generated by the tags

Table 5.1: Specifications of hybrid sub-1 GHz and UWB IR system.

UWB TRANSMITTER		SUB-1 GHz TRANSCEIVERS	
UWB Bandwidth	3 - 5 GHz	sub-1 GHz Bandwidth	779 - 928 MHz
UWB Peak Power	100 mW	sub-1 GHz Transmit Power	+14 dBm
Power Consumption (in active mode)	< 9 mW	sub-1 GHz Power Consumption (in Tx mode)	< 80 mW
Power Consumption (1 Hz refresh Rate)	< 0.3 mW	sub-1 GHz Power Consumption (in Rx mode)	< 25 mW
Battery Lifetime (1 Hz refresh Rate)	90 days	Receiver Sensitivity	-112 dBm

and re-allocates the slot number in case of any discrepancy. The tags are registered at the master node in this manner. The master node sends the updated tag slot information in the next cycle beacon. This process continues and any new tag can join the sensor network in similar manner in subsequent master beacon cycle.

2) Positioning mode: The positioning mode starts after MAC mode. During the positioning mode, the registered tag upon receiving the master beacon in subsequent cycle, finds its tag ID along with slot number in the beacon. After finding its ID, it sends acknowledgment back to anchor nodes and starts UWB signal transmission for ranging during assigned time duration. This process repeats in every master beacon cycle of 1 Hz, leading to 1 Hz as update rate in tag positioning.

In this work, we follow the scheme with target tag ($A_1, A_2, \dots$) transmitting followed by a reference tag (F_0). The reference tag is co-located at one of the anchor nodes and whose position is known. The signal frame at the receiver is captured by combining the UWB signals from two tags together. The demonstration of the captured signal at the receiver is shown in Fig.5.14.

Figure 5.13: Different steps in the transmission mechanism for master node and target tags.

5.4 Experimental setup with Results and Discussions

To perform the experiment, we consider 4 anchor nodes and 10 target tags ($A_1, A_2, A_3, A_4, A_5, A_6, A_7, A_8, A_9, A_A$). The ground positions of the anchor nodes and the target nodes are shown in Fig.5.15. The coverage area for the experiment is approximately about 16 m x 5.3 m. The experiment was conducted at Delta-EEE joint lab on Internet of things at Nanyang Technological University, Singapore. The tags were placed at three different tripods adjusted at the same height. Tags A_1, A_2, A_3 were placed at tripod 1, tags A_4, A_5, A_6, A_7 were placed at tripod 2, and tags A_8, A_9, A_A were placed at tripod 3. For the experiment, F_0 , the reference tag, was collocated at the anchor node 1. All the anchor nodes were deployed at the same height. We follow the transmission scheme pairing one tag with one reference tag as: $[(A_1F_0), (A_2F_0), (A_3F_0), (A_4F_0), (A_5F_0), (A_6F_0), (A_7F_0), (A_8F_0), (A_9F_0), (A_A F_0)]$ i.e. UWB-IR signal is transmitted from mobile tag followed by the UWB-IR signal from reference tag. As equivalent time sampling is used to sample the UWB-IR signal and for signal reconstruction purposes at each anchor node, pulse repetition

Figure 5.14: Illustration of the channel impulse response of the recovered signal at the anchor nodes with a target tag (Tag_1) and a reference node (F_0).

frequency (PRF) for UWB signal is set at 3 MHz. During the experiment we set the ADC sampling frequency of 2.999010327 MHz. The set PRF and the ADC sampling frequency will result in an equivalent time sampling resolution of around 0.11 nS. Hence, with uncertainty of 1 sample point in TOA values, the error in ranges can be translated to about 3.3 cm. Equivalent differential time difference of arrival (E-DTDOA) is used to find the range difference between the nodes as described in Sec.5.2.2. Assuming the position of one transmitter (i.e. a reference node, tx_2) is known, the range difference of other transmitter (mobile node) from two receivers can be obtained. These ranges can be used to calculate the position of the mobile tag using commonly known recursive least square methods. The estimated positions of all the 10 tags is shown in Fig.5.16. The inset shows the zoomed position of the tags. The spread of the tag can be clearly seen in the inset. The corresponding data point shows the mean value of the position of the tags. From the experimental results, it can be seen that the positioning accuracy of less than 30 cm is achieved for all 10 tags in an indoor area of 16 m x 5.3 m.

5.5 Comparison with State of the Art Systems

Table.5.2 shows the comparison of this work with some of the closely related works [31], [146] and [145]. Reference [31] uses ETS technique to process the received signal.

Figure 5.15: Ground position of 10 tags and 4 anchor nodes

Signal from sub-sampler are then fed to the FPGA where leading edge detection algorithm locates the pulse position. In this work, all the anchor nodes (or base stations) are given the same clock and are thus synchronized. Moreover, reference [146] further proposed a sub-Nyquist sampling rate-based mechanism to estimate accurate time of arrival (TOA) of the signal. In [146], the focus is on estimating the clock drift due to the longer preamble used in ETS technique. In [145], authors use ETS to estimate the CIR using the bounded data uncertainty model. All of the above works consider a model under the assumption that the receiver has perfect knowledge of the time instant of the first arrival of the first pulse in the pulse train. In addition, no analytical modeling of TOA and experimental demonstration utilizing ETS for ranging of unsynchronized nodes is presented in these works. To the best of authors knowledge, ETS based ranging system for fully unsynchronized nodes using analytical modeling has never been investigated in the literature. For the sake of comparisons, we have also presented few systems based on real time sampling techniques [79], [151] and [152]. In most of the results published in literature, costly

Figure 5.16: Experimental results of positioning of 10 mobile tags and 4 anchor nodes. Insets in the figure shows the zoomed spread of the target tags. Data point in the figure shows the mean value of the spread.

equipment such as high-speed oscilloscope, FPGA was often used as a receiver, which gives the opportunity to precisely determine the pulse's TOA. Our system benefits in the sense that no complicated and costly instruments are used and can be implemented with commercially off the shelf components and some well-known software with good ranging accuracy with a precision of sub nanoseconds.

5.6 Conclusions

In this chapter, the architecture and implementation of a low power tag hybrid sub-1 GHz and UWB-IR scheme for an indoor positioning system is presented. The hardware and software building blocks involved in the system have been discussed. The medium access scheme for multi tag operation is also demonstrated in the chapter. Furthermore, a differential time difference of arrival technique is employed to find the location of a mobile tag. Experimental results are presented to verify the performance of the positioning system which consists of 10 mobile tags and 4 anchor nodes. The accuracy of the system is demonstrated by placing different tags at different static locations. The positioning accuracy of less than 30 cm is achieved in an indoor area of 16 m x 5.3 m. Since the system uses low cost and low sampling rate ADC to recover high bandwidth signal, it is practically feasible and easily deploy-able for low-cost wireless sensor nodes. Moreover, the scheme further

avoids the need of high-speed processors required for processing high sampling rate data, as experimentally demonstrated in this chapter.

Table 5.2: Comparison with the existing works. % error is w.r.t the largest range

Ref	Method	Transmitter	Receiver	Synchronization	Test Area	Ranging Error (%) error)	Remarks	Cost
[31]	TDOA	300 pS pulse	Envelope detection + I/Q signals are sub sampled using UWB sub sampling mixer with FPGA	Synchronized	50x150x 250 m^3	5.24 mm (2.096%)	Mobile Robot	High
[146]	TOA	Generic, Not Available	Equivalent time sampling with no experimental Demonstration	5-50 nS	-	-	Drfit compensation	-
[145]	-	Generic, Not Available	Equivalent time sampling	-	-	-	Channel Estimation	-
[79]	TDOA	1 nS pulse $f_c = 4$ GHz	LNA + High speed 4 channel oscilloscope as a receiver	Synchronized	12x10 m^2	8 cm (0.6%)	Static Scenario	Moderate
[151]	TDOA	3 nS pulse $f_c = 3.5$ GHz	Non-coherent UWB system, ADC 8 bits with 1.5Gsps and a Virtex4 Xilinx FPGA	Synchronized	6x8 m^2	20 cm (2.5%)	Mobile Robot	High
[152]	TOA	1.2 nS pulse 800 MHz BW	Real time sampling with ADC at 2.5 GSps + FPGA	5 nSec	-	-	Gigabit Transceivers	High
This work	EDTDOA	1 nS pulse, BW = 2 GHz, $f_c = 4$ GHz	Correlator + Equivalent time sampling ADC with sampling freq approx 3 MHz + ARM Cortex A8 microprocessor on BeagleBone Black (BBB)	Unsynchronized	16x5.6 m^2	<30 cm (1.8%)	Static Scenario	Low

6 Conclusions & Future Work

6.1 Conclusions

In this thesis, a novel analytical equation is proposed to determine the equivalent time of arrival (E-TOA) for systems employing equivalent time sampling (ETS) technique. The obtained E-TOA achieves sub nanosecond resolution while using much reduced ADC sampling rate (in the order of 2-3 MHz). The timing information is extracted from high resolution channel impulse response (CIR) which is obtained using ETS technique. The proposed E-TOA equation is different from conventional real-time sampling equation due to the presence of additional transmitter clock drift, and thus sensitive to both the transmitter and the receiver clock drift variations. The validation of the E-TOA equation is carried out numerically using real simulations along with experimental verification. The effect of timing uncertainties relating to the transmitter clock start time and the receiver clock offset is analyzed with variations in the transmitter and receiver clock drift. It is shown that the error due to the ETS is the magnified transmitter and receiver drift difference by an equivalent time transmitter pulse repetitive frequency. With E-TOA measurements, an equivalent differential time difference of arrival (E-DTDOA) based one-way ranging scheme for unsynchronized anchors is further proposed. The presence of transmitter clock drift makes E-DTDOA different from the conventional DTDOA scheme in which the ranging accuracy is affected only by the receiver clock drift. It is thus demonstrated that achieved ranging accuracy using E-DTDOA scheme is adversely affected by the transmitter's clock drift. The proposed E-TOA equations are further utilized to find the accurate drift estimates for the transmitting and receiving nodes. These drift estimates are used to find the precise pulse repetitive frequency of the transmitter and ADC sampling frequency of the receiver. The precise receiver frequency estimates are used to tune the readers or anchor nodes to the same frequency. In addition, the transmitter frequency estimates are used to find the precise equivalent time resolution of the recovered equivalent time sampled signal. The precise equivalent

lent time resolution is used to find the actual or corrected equivalent time of arrival. In addition the other parameters which corrupt the E-TOA such as the start time of the transmitter and clock offset of the anchor nodes are canceled by using an approach of equivalent differential time difference of arrival (E-DTDOA). Thus, start time of the transmitter and the clock offset of the anchor nodes is canceled by using E-DTDOA across the anchor nodes. The E-DTDOA approach is used to find the range difference between the target node and anchors, which is subsequently used for finding the position of a target node. We show that using the proposed E-DTODA algorithm, high-ranging accuracy can be achieved without stringent synchronization requirement between the anchor nodes.

The feasibility of the proposed scheme for an asynchronous indoor localization system is experimentally demonstrated which uses UWB-IR as a ranging signal. For practical demonstrations initially wired architecture which connects anchor nodes using wires was used. The system uses low cost and low sampling rate (few MHz) ADC's to recover high bandwidth signal, which is practically feasible for low cost sensor nodes. Moreover, the system further avoids the need of high speed processing devices required for processing high sampling data. We demonstrated high accuracy localization of a target node using the proposed algorithm.

We further extend our work from a wired architecture to wireless architecture, where all the anchor nodes are wireless. Additionally, we also present the architecture and implementation of a hybrid sub-1 GHz and UWB-IR scheme for an indoor positioning system. The hardware and software building blocks involved in the system have been discussed. The medium access scheme which is used for multi tag operation is also demonstrated. Furthermore, we use equivalent differential time difference of arrival technique to cancel the affect of clock offsets of the anchor node and subsequently used target node range difference of arrival to find the location of a target node. Experimental results are presented to verify the performance of the wireless positioning system which consists of 10 mobile tags and 4 anchor nodes. The accuracy of the system is demonstrated by placing different tags at different static locations. The positioning accuracy of less than 30 cm is achieved in an indoor experimental area of 16 m x 5.3 m. Thus, we envision that the scheme proposed is beneficial in designing precise indoor localization system using high bandwidth signal with efficient use of low cost anchor nodes and target nodes with poor crystal clock and larger time drift.

6.2 Future Work

1) Optimized selection of the pair of target nodes.

As observed from the final expression of equivalent differential time difference of arrival (E-DTDOA) in chapter 4, the equation has additional biased term which is caused due to the difference in the different start of the ADC capture time of different anchor nodes and the difference in the clock skew of the target nodes. This may result in the bias error. Therefore, to reduce this error an optimized selection of the combination of target nodes is needed which can minimize the affect of this bias error. The target node skew estimate as proposed in this work can be beneficial in identifying the pair of target which can minimize this error.

2) Reducing the transmission time and still achieving the same performance.

The current work focuses on the mathematical formulation of the ETS technique and its experimental demonstration. However, the drawback of ETS is it requires many transmitted pulses to recover a single pulse at the receiver. This in turn increases the ranging time. Therefore, a scheme which can reduce the number of transmitted pulses while still achieving the same performance can be a possible direction for future exploration.

3) Energy efficiency analysis

Since ETS technique as used in this work involves transmission of multiple pulses at the transmitter, this results in high power consumption at the transmitter as compared to the real time sampling based system, however the power consumed at the receiver is less in ETS as compared to the real time sampling. Therefore, a comparison of the total power consumption of the overall system for the cases when real time sampling is applied and when equivalent time sampling technique is applied is required to optimize the performance of both the systems.

4) Optimum position of the reference node and the anchor nodes

The position of the reference node as well as anchor nodes plays an important role in determining the accurate position of the target node, thus a self calibrating scheme which can place the anchor nodes at an optimal location along with an optimal combination of the reference can be seen as a possible future work.

5) Device free human detection and enumeration

The estimation of precise channel impulse response along with its mathematical modeling is demonstrated in this work. The precise channel information so obtained can be combined with machine learning algorithms to investigate the device free

human detection and enumeration as required in certain applications.

Author's Publication

- **A. Vashistha** and C. L. Law, “A Novel E-DTDOA Based One-way Ranging Using UWB-IR with Unsynchronized Anchors,” in IEEE transactions on Industrial Informatics, Doi: 10.1109/TII.2020.2992620.
- **A. Vashistha** and C. L. Law, “E-DTDOA based Localization for Wireless Sensor Networks with Clock Skew Compensation,” in IEEE Sensors Journal, 2019, Doi: 10.1109/JSEN.2019.2953811.
- **A. Vashistha**, A. Gupta and C. L. Law, “ Low Power Tag Hybrid Sub-GHz UWB-IR based Indoor Positioning System using DTDOA-Experimental Demonstration”, accepted for publication in 2019 IEEE 10th RFID -TA, Pisa, Italy, 2019
- **A. Vashistha**, R. W. C. Ling, and C. L. Law, "A TDOA Measurement Technique for Asynchronous Indoor Localization System using UWB-IR," Indoor positioning and Indoor navigation (IPIN), 2016, Madrid, Spain.
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