The ability to accurately locate people and objects indoors will enable opportunities for control and automation of indoor environments. Current indoor localization solutions require dedicated hardware which must be custom tailored to each application. There are large technological, societal, and economic benefits in developing a more general purpose system that can be deployed as universally as the Global Positioning System (GPS). Localization systems based on mobile Wi-Fi communications show great promise toward meeting these needs because much of the infrastructure is already in place. To support this goal, a 5 GHz Wi-Fi front end receiver array for an indoor localization system based on a time difference of arrival (TDOA) multilateration algorithm is presented. The receiver consists of 4×80 MHz, highly synchronized channels to support the TDOA measurement. The receiver achieves propagation delay mismatch of less than 150 picoseconds. The full system is able to locate and track a mobile device to within 1 meter.
A 5 GHz Wi-Fi Receiver Front End with $4 \times 80$ MHz Channels for a High Accuracy Indoor Localization System

by

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I understand that my thesis will become part of the permanent collection of Oregon State University libraries. My signature below authorizes release of my thesis to any reader upon request.

Kyle L. Gray, Author
There are no endeavors in life that are accomplished individually and this thesis is no exception. I would like to mention a few of you who have contributed to this milestone in my academic career.

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A 5 GHZ WI-FI RECEIVER FRONT END WITH 4×80 MHZ CHANNELS
FOR A HIGH ACCURACY INDOOR LOCALIZATION SYSTEM

CHAPTER 1. INTRODUCTION

Recent advances in electronics and data processing have enabled the gathering of information previously obscured or unobtainable. For example, information about an environment can be gathered and processed in an effort to control or automate processes. One such useful piece of information is the position of a person or object within an environment. With information about a target’s position an endless number of applications can be imagined. Location-based services might provide businesses a method to offer smart incentives for shopping, or provide navigation in large buildings such as hospitals and airports. At home, one’s position could be used as a gesture recognition interface to monitor health and provide assistance with daily activities as well as provide entertainment. A lost item in the home could be located with ease. In emergency situations, the position of a fireman in a burning building, an agent who must remain silent while rescuing hostages, or a dog trained to locate explosives could all be of great value to responders on the outside. The location of a valuable asset could be used to trigger an alarm if it leaves an area without authorization.

For outdoor environments, the satellite-based Global Positioning System (GPS) can successfully track objects to within a few meters [1] making it the de facto standard for applications requiring positioning. However, for applications where it is desired to track the position of an object indoors, GPS is often unable to achieve the required accuracy. This is primarily because indoor environments contain many obstacles that attenuate and scatter the GPS signal making it difficult to accurately recover the necessary location information. The sheer variety of indoor environments creates a unique set of requirements for each application
and often relies on custom tailored hardware to meet the localization needs [2]. At present, there is no standard for indoor localization that is nearly as prolific as GPS for outdoor applications. The adoption of a standard system would be of great benefit to society as this allows the problem to be generalized and abstracted. This would expand the design space by encouraging the development of applications using the technology rather than the technology itself.

There are several methods employed in research that attempt to address the challenges associated with indoor localization. A general purpose system suitable for a large range of applications should be wireless and composed of two separate hardware units; namely a transmitter and a receiving measurement unit. The target object, whose position we desire, can assume the role of either transmitter or receiving measurement unit. In the first case the target transmits a signal which is received by one or more measurement units. In the second, the target receives a signal from one or more transmitter units and computes its own location. The communication signal can take on a wide variety of physical forms such as acoustic waves; ultraviolet, visible, or infrared light for camera-based imaging; radio signals like Wi-Fi, RFID, and Ultra-Wideband; magnetics; or even sensor data such as inertial, pressure, or temperature that is communicated wirelessly. A localization system can also be classified according to the signal processing algorithm that it uses. Parameters such as the time of arrival, the incidence angle, and the received signal strength can all be used in the position calculation. The preceding discussion demonstrates how many different combinations of systems can be arranged in order to satisfy the requirements of a given localization system [3].

Much of the current research is focused on localization using radio frequency electromagnetic waves largely due to the ubiquity of electronic devices. Nearly everyone carries some kind of personal mobile device capable of internet connectivity over a wireless local area network (WLAN). This fact in conjunction with the availability of WLAN access points (almost universally using the Wi-Fi standard) means there are several applications for which much of the localization infrastructure is already in place. This makes Wi-Fi an ideal tech-
nology medium with which to structure a low cost localization system for a diverse range of applications.

This work presents an indoor localization system based on the Wi-Fi communications between a personal mobile device and a wireless router. The localization algorithm is based on multilateration which uses time difference of arrival (TDOA) measurements between receivers to calculate distance and position. The primary contribution of this work is the design of a four channel, highly synchronous Wi-Fi receiver used as the analog front end (AFE) for a localization system [4] that was then used to demonstrate the system feasibility. This work will therefore focus on the design of the AFE but within the context of the localization system as a whole.
CHAPTER 2. THEORY OF OPERATION

2.1 Wi-Fi Communication Basics

The origins of the term “Wi-Fi” are debatable but today the term is synonymous with wireless internet access for phones, tablets, and PCs. Wi-Fi can be thought of as any wireless local area network (WLAN) that conforms to the IEEE 802.11 networking specifications. The Wi-Fi standard dictates everything involved in the communication link between a mobile device and the wireless router which serves as the central hub that connects that device to the internet or other network of devices. A WLAN with Wi-Fi connectivity is depicted in Fig. 2.1.

Figure 2.1: Wi-Fi is a set of specifications for connecting phones, tablets, and PCs to a wireless LAN. Modern routers combine access point, ethernet switch, router, and modem functions in a single unit.
The communication between a mobile device and a router takes place in either the 2.4 GHz or 5 GHz unlicensed bands. The 2.4 GHz band (2.400 GHz – 2.500 GHz) is widely used in numerous applications and is therefore subject to interference from microwave ovens, cordless telephones, Bluetooth devices, and other nearby Wi-Fi networks. By comparison, the 5 GHz band (5.150 GHz – 5.850 GHz) is less crowded but the coverage area is reduced because the larger operating frequency results in an increased attenuation rate.

The most recent amendment to the 802.11 specification is known as 802.11ac. A key improvement of the 802.11ac amendment is its expanded support for the 5 GHz band compared to previous amendments, for example 802.11a/n. Under 802.11ac, the 5 GHz band is divided into 20 MHz and 40 MHz channels, like 802.11a/n, but it also allows channels to be combined into 80 MHz and 160 MHz bands resulting in an unprecedented Wi-Fi bandwidth. The 802.11ac channel division plan is shown in Fig. 2.2.

A greater bandwidth translates to faster data rates but also has a distinct advantage with regards to indoor positioning. It will be shown in Section 4.2 that localization accuracy is improved with an increase in bandwidth so the largest available bandwidth is desirable for optimal localization accuracy. Unfortunately 160 MHz channels have not been widely adopted in commercial products due to regulatory concerns. Therefore this work will make use of the 80 MHz channels 42, 58, 106, 122, 138, and 155 as shown in Fig. 2.2.

Using an appropriate channel, information is sent back and forth from a mobile device to a router in the form of data packets. Wi-Fi packets often carry the desired internet data. However, many packet exchanges serve other purposes such as authentication, identification, routing, communication type, data order, power management, available spectrum, etc. [6]. The amount of data in each packet (and equivalently the duration of each packet) will depend on its purpose. In addition, the sequence and rate of these various packets depends on several factors and is essentially random from an external perspective. A typical communication signal between the mobile device and a router could look something like the time domain signal shown in Fig. 2.3(a). For the application of localization it is not necessary to decode the packet information but rather to detect whether a packet has arrived or not. For our
Figure 2.2: 5 GHz channel division plan for the 802.11ac specification. Reprinted from FCC channel plan document [5].
purposes the packet can be thought of as the binary envelope shown in Fig. 2.3(b) where the rising edge signifies the packet’s time of arrival.

![Figure 2.3: (a) Example Wi-Fi packets. (b) Simplified packet envelope conveying whether packet is present or not. The envelope’s rising edge indicates the packet’s time of arrival.](image)

2.2 System Overview

The illustration in Fig. 2.4 provides an overview of the proposed localization system. A mobile device capable of transmitting and receiving Wi-Fi signals according to the 802.11ac standard resides at some unknown position \((x_0, y_0, z_0)\) in a room. Once a Wi-Fi link is established between the mobile device and the wireless router, the device will begin sending packets to the router. The mobile device has no information about the position of the router so it must send the wireless data in every direction. Therefore it is possible to intercept the wireless signal using an antenna cluster. The Wi-Fi packets received at the antenna cluster are then sent to a signal processing system which outputs the mobile device’s position. Here it is assumed that the signal processing system is able to differentiate between packets originating at the mobile device from those originating at the wireless router.

The antenna cluster shown is composed of four, fixed-position antennas which serve as four independent input channels (Ch1, Ch2, Ch3, and Ch4) to the signal processing system.
A given data packet will arrive at each antenna at a different moment in time due to the different propagation distances to each antenna. Using one channel as a reference, the time difference of arrival (TDOA) between each channel is measured and processed in order to calculate the desired position coordinates \((x_0, y_0, z_0)\). This process of localization based on local time differences is known as multilateration. In contrast to trilateration, which uses absolute time of arrival to calculate position, multilateration has the advantage that the time of the original transmission does not need to be known. In other words, the mobile device and the signal processing system do not need a common clock to synchronize the time of arrival measurement [7].

![Figure 2.4: Full system architecture. The radio frequency Wi-Fi signal transmitted from the mobile device to the wireless router is intercepted by the antenna cluster and processed to produce the mobile device’s position.](image)

2.3 TDOA Localization Principle

Consider the simplified link arrangement shown in Fig. 2.5. The mobile device (object to be tracked) transmits an RF signal which is received at four separate receiver channels, Ch1 – Ch4. Due to the propagation distance from the device to a particular antenna $d_i$ and the finite velocity of RF signals $c$, a packet’s time of flight to a particular receiver channel $t_i$ will be given by Eq. (2.1).

$$t_i = \frac{d_i}{c}$$  \hspace{1cm} (2.1)

![Figure 2.5: System geometry for TDOA calculation.](image)

The first step is to define the 3D Cartesian coordinate system. For simplicity let the origin lie near the centroid of the antenna cluster plane. Then the fixed coordinates of the channel $i$ antenna can be defined as:

$$Ch_i : (x_i, y_i, 0)$$  \hspace{1cm} (2.2)
and the mobile coordinates of the device transmitter can be defined as:

\[ TX : (x_0, y_0, z_0). \] (2.3)

Using Euclidean geometry the distance from the mobile device to the channel \( i \) antenna can be written as:

\[ d_i = \sqrt{(x_0 - x_i)^2 + (y_0 - y_i)^2 + (z_0)^2}. \] (2.4)

It is actually time of arrival differences that are required by the system so the four receiver channels provide three difference measurements when one channel is chosen to be the reference. Using channel \( j \) as the reference \((j \neq i)\) and applying Eq. (2.1) to Eq. (2.4) yields the relationship between the measured TDOA and the target’s coordinates:

\[
\Delta t_{i-j} = \frac{d_i - d_j}{c} = \frac{1}{c} \left( \sqrt{(x_0 - x_i)^2 + (y_0 - y_i)^2 + (z_0)^2} - \sqrt{(x_0 - x_j)^2 + (y_0 - y_j)^2 + (z_0)^2} \right). \] (2.5)

The desired position coordinates \((x_0, y_0, z_0)\) can be calculated by solving Eq. (2.5) with numerical methods like the commonly used nonlinear least squares (NLLS) algorithm [8]. The solution can be thought of geometrically as three intersecting hyperboloids. In general, there can be multiple points that these hyperboloids intersect representing multiple location solutions. Typically there is only one feasible solution so this does not present a problem [9].
There are several important parameters to specify when designing a localization system. The goal of this project is to provide a proof of concept through the development of a working prototype. This chapter describes some of the universal parameters applicable to localization systems and develops the existing infrastructure with which the system must be built upon.

3.1 Functional Requirements

This section will define the primary functional requirements for a prototype system meant to serve multiple applications.

3.1.1 Accuracy

Accuracy can be defined as how precisely the target can be tracked in space. Accuracy typically refers to the percentage of measurements that fall within some statistical bounds. The measurements used in this definition may be either raw or processed using multiple raw estimations. In the latter case it becomes necessary to define the computation time for which the system can achieve a certain accuracy. For many applications it is desirable for the target’s location to be determined within a 1 meter radius which will be the accuracy goal for this work.

3.1.2 Update Rate

The update rate is analogous to the frame rate of a recorded video. A faster update rate will display the target’s movements fluidly and provide some perception of velocity whereas
a slower update rate will appear disjointed which makes velocity ambiguous.

Nearly all localization systems require some statistical processing on a large group of data to achieve appreciable performance. This implies an inexorable tradeoff between accuracy and update rate. A fast update rate will allow less time for computations which results in lower accuracy. Most human actions take place on the order of seconds. Therefore an update rate of at least 1 Hz will suffice to track expected movements.

3.1.3 Coverage

Often, the height dimension of people and objects in a room is irrelevant to specify. To this end, coverage will be defined as the two dimensional floor area with which the system can locate an object with the required accuracy. This work will target a coverage area of 10 meters $\times$ 10 meters.

3.1.4 Scalability

The coverage area should be scalable to perform equally well in a small office or a large warehouse. The simplest solution is to allow the system to operate with one or more area-proportional antenna clusters. In effect, the coverage area can be doubled by using twice as many antenna clusters.

3.1.5 Size

Ideally the system would be made as small as possible so as to fit in both small and large areas. However, as the spacing between antennas in the cluster decreases so does the signal TDOA. Smaller TDOA values are harder to measure precisely which leads to reduced accuracy. Thus the system size is constrained to that of the antenna cluster. In order to
be suitable in a variety of settings the antenna cluster geometry should be reconfigurable to allow the optimal cluster size for the desired localization region.

3.1.6 Latency

Latency refers to the delay between the time that a movement happens and the time that the movement is registered at the output of the system. This delay is mainly caused by the computations required by the system. The target applications take place on the order of seconds so a display delay of a second or two will be tolerable.

3.1.7 Number of Users

A general application might have any number of desired targets to track. This raises the question of how to associate a position with a particular target. The header of a Wi-Fi packet contains the unique IP address which can be used for identification. Thus it is crucial to preserve this information through the system so that it may be extracted and used to support multiple users.

3.1.8 Technology Used

The system should make use of the most current revision to the 802.11 specification, namely the 802.11ac. The receiver hardware should be designed for the 5 GHz Wi-Fi band so that the 80 MHz channel bandwidth may be taken advantage of. If possible, there needs to be a capability to expand to 160 MHz bandwidth when the technology becomes more available.
3.2 Signal Processing System Architecture

A diagram of the major signal processing blocks involved in the localization system is given in Fig. 3.1. The wireless signal received at the antenna cluster is first passed through the Analog Front End (AFE). The AFE’s primary functions are to downconvert the RF signal to baseband and scale it for sampling in the digital back end also known as the Data Acquisition Unit (DAU). The signal is sampled immediately at the input of the DAU. The digital signal is then processed and displayed back to the user. The primary focus of this work is to design and optimize the AFE block shown in Fig. 3.1 to function with the preexisting DAU and Data Processing blocks.

Figure 3.1: Signal processing system blocks.

The DAU is a complex system that can be further divided into the subsystems shown in Fig. 3.2. The four input channels are sampled by four independent – but synchronized – ADCs. This is important to note because the TDOA localization requires measuring a precise time value and the sampling rate determines the resolution of time measurements. Data is then collected by an FPGA and configured for network communication over an ethernet cable. The data can then be processed using the localization algorithm on a PC. The DAU board was developed previously for a slightly different application. However, its specifications are suitable for developing the first prototype making it desirable to reuse. One goal of this project was to get the system working without modifying the DAU.
Figure 3.2: Expanded view of the DAU system.
CHAPTER 4. TDOA MEASUREMENT AND SOURCES OF ERROR

It is very challenging to locate an object within a small spatial region because of the significant speed at which electromagnetic waves travel. Equation (2.1) shows how very small time differences must be accurately measured. For example, it is common for packets to arrive within a few hundred picoseconds of each other. Whenever it is desired to measure a very small quantity, noise in the measurement becomes a concern. This section describes how a TDOA measurement is performed and how small errors in the measurement can translate to position uncertainty.

4.1 TDOA Measurement Scheme

To understand how errors affect a measurement, it is first necessary to understand how measurements are made. A purely analog technique of measuring time differences on the order of nanoseconds would be very complex and difficult to calibrate. A much more attractive solution is to sample the received wireless signal and extract the time difference of arrival using digital methods. The target’s position can then be calculated according to the methods described in Section 2.3.

Consider the two Wi-Fi packet waveforms shown in Fig. 4.1(a). Assume that these signals represent the unprocessed waveforms incident at two of the four antennas in the cluster. Since the signal has to travel different distances to reach each antenna, the signal power level will be different for each channel due primarily to the signal expanding in space. The farther a signal travels, the more the wavefront will expand thus reducing the power density for a given antenna size. For distances more than a few centimeters in air, the resulting loss caused by spreading can be estimated using free space path loss (FSPL) which is related to the path distance $d$, the signal carrier frequency $f$, and the speed of light $c$. 
according to Eq. (4.1).

\[ FSPL = \left( \frac{4\pi df}{c} \right)^2 \]  \hspace{1cm} (4.1)

The desired quantity to be measured is the time difference of arrival (TDOA) between the given signal pair. The strategy here is to determine a common reference point for the two signals, measure the local time of arrival of the two signals using this reference point, and then compute the difference to find the TDOA.

The wireless signal is composed of an RF carrier frequency and a baseband envelope signal. When the signal arrives the envelope will begin to rise above the noise floor. The local time of arrival can be said to be the time at which the rising edge of the envelope crosses some voltage threshold. The signals may have different amplitudes because of the free-space path loss so the voltage threshold must be set individually for each signal. For example, one might use 50% of the average envelope voltage as the threshold. The TDOA is then simply the difference between the two local time measurements.

4.2 Effects of Quantization

It is desired that the signal be processed with digital methods so it must inevitably be quantized in both time and amplitude. This section describes how quantization inherently adds errors and how these errors translate to uncertainty in the system.

4.2.1 Time Quantization

Imagine two receivers in a 10 m × 10 m 2D plane with a transmitter at some location within the same plane as depicted in Fig. 4.2(a). The measured TDOA between the two receivers positions the transmitter on the hyperbolic curve shown in the figure. The magnitude of the TDOA will be the same at any point on this curve.
Figure 4.1: (a) Wi-Fi packet arrival time offset between Ch1 and Ch2. (b) Ch1 arrival time is measured as the moment Ch1 crosses the 50% voltage threshold. (c) Ch2 arrival time is calculated the same as Ch1. The TDOA is calculated as the difference between Ch1 and Ch2 arrival times.
Figure 4.2: (a) The TDOA measured between RX1 and RX2 positions TX somewhere on a hyperbola. (b) Finite time resolution leads to additional ambiguity in the position possibilities represented as a spreading in the solution space.

Next, assume that the time measurements are constrained to multiples of the sampling period $T_s$ of the analog to digital converter (ADC). The time difference of arrival calculation begins with the local time of arrival at each receiver measured as the moment in time that the signal crosses some threshold voltage. If the threshold is crossed at a time that is not a multiple of $T_s$ then the time of arrival will be set as the next sample, which could be up to one whole $T_s$ later. This situation is depicted in Fig. 4.3. This error in the measurement is random as it depends on the phase between the incoming signal and the ADC clock. The uncertainty in the measured time of arrival can be visualized as a spreading of the solution curve as shown in Fig. 4.2(b) where the thickness of the curve represents the random error.

Considering this error it is now possible to plot the set of 2D solution curves, spaced at the sampling period, to represent every possible solution for the 2 receiver geometry in the $10 \text{m} \times 10 \text{m}$ room. This is done in Fig. 4.4(a). Figure 4.4(a) gives an idea of the spatial accuracy of the TDOA localization relative to the position of a given receiver pair. The uncertainty is minimized in the region between the two receivers and steadily increases with the distance from the center.

To find the transmitter’s position it is necessary to introduce a third receiver in order to
Figure 4.3: (a) Time of arrival is measured as the first sample to cross the 50% voltage threshold in the rising edge of the packet envelope. (b) Expanded view of the boxed region in (a). The discrepancy between the first sample crossing and the continuous time crossing can be up to one sampling period.

find two TDOA equations to solve for the two unknown coordinates \((x_0, y_0)\) in this 2D space. If the two solution sets are plotted overlapping each other then it becomes apparent how the finite sampling resolution leads to imperfect location estimation. Figure 4.4(b) shows such a plot where each region corresponds to a given pair of TDOA measurements. The relationship between the uncertainty and the relative position of the target object can now be seen to be a complex function of the receiver geometry. However the macroscopic trend is still valid; uncertainty grows larger the further the target object moves from the receiver centroid.

4.2.2 Amplitude Quantization

The assumption made earlier about the time resolution being constrained to multiples of the sampling period \(T_s\) is not completely valid. In reality, the absolute time measurement can be made much more accurate than \(T_s\) by interpolating, or upsampling, the data in software. The preceding argument does however serve two purposes. The first is that it demonstrates how the localization accuracy depends on the position of the transmitter relative to the
Figure 4.4: (a) All TDOA possibilities between RX1 and RX2 for the given coverage area. Finite time resolution results in finite TDOA possibilities. (b) Adding a third receiver results in 2 sets of overlapping TDOA possibilities. The size of the blue regions indicate the localization accuracy at different positions in the room.
receiver geometry. The target regions shown in Fig. 4.4(b) can be made smaller through interpolation but the inaccuracy will still grow larger at greater distances from the receiver centroid. Thus the analysis provides a graphical method to design for the desired accuracy in a particular coverage area. The second purpose is that it will aid in understanding how the amplitude quantization translates to a timing error.

Extracting the local time of arrival measurement by means of interpolation relies on the voltage estimate of the rising edge to be accurate. How accurate the voltage estimate is will depend on the resolution of the ADC. An n-bit ADC will have $2^n$ discrete voltage levels with which to estimate the output.

Consider the voltage waveforms of Fig. 4.5. Let’s say the sampled waveform has discrete voltage levels but is continuous in time (or upsampled infinitely). It is apparent from Fig. 4.5 that no matter how upsampled the discrete waveform is there is still the potential for an error in the local time of arrival measurement because of the voltage quantization. The observation can be made that the magnitude of this error will depend on the slope of the rising edge as well as the number of bits utilized to quantize the voltage.

![Figure 4.5: (a) Rising edge of the envelope with discrete voltage levels but continuous in time. (b) Expanded view of the boxed region in (a). Even with infinite upsampling the time of threshold crossing can be delayed because of the discrete voltage levels.](image)

The effect of the rising edge slope is fairly straightforward to understand. The steeper
the slope, the faster the output will transition between quantization levels which minimizes the risk of error. Another way to state this is there will be fewer samples per quantization step therefore the maximum time deviation from the nominal threshold crossing is reduced. The rising edge slope of the envelope signal is dictated by the packet bandwidth originating at the transmitter. Thus, the error contribution of the slope is minimized by choosing the largest possible bandwidth for the packet.

The ADC’s resolution contributes to the timing error in a way analogous to the slope. If the quantized signal of Fig. 4.5 had twice as many voltage levels (i.e., one additional bit in the ADC) the output would transition between quantization levels faster reducing the timing error. Therefore the timing error is minimized by utilizing as many ADC bits as possible.

To make use of all ADC bits, it is required that the input signal voltage is scaled to the full-scale range of the ADC. For example, if the input signal is scaled to half of the ADC full-scale input range then the most significant bit will never change which is equivalent to operating with one less bit. For this reason the ADC must be preceded by a gain control system to ensure that the ADC input is scaled properly regardless of the received signal level.

The impact of quantization on the spatial localization error $\Delta d$ can be approximated with a few assumptions. An estimate of the slope can be made using a bandpass to lowpass response transformation. The output envelope of a bandpass filter with a stepped tone input is similar to the step response of a lowpass filter [10]. Therefore it can be assumed that the envelope is the same as the step response of a first order lowpass filter with a 3 dB frequency $f_{\text{max}}$. This step must also be scaled for the full-scale range of the ADC $V_{\text{pk}}$ (step amplitude is then $V_{\text{pk}}/2$). Assuming the envelope is the step response of a first order lowpass filter with a 3 dB frequency $f_{\text{max}}$. It can be shown that the rising edge slope of the envelope $m_{\text{env}}$ at the 50% voltage threshold crossing is given by Eq. (4.2).

$$m_{\text{env}} = \frac{2\pi f_{\text{max}} \cdot V_{\text{pk}}}{4} \quad (4.2)$$

Next, assume the ADC is operating with $n$ bits so the quantization steps are multiples
\[ V_{LSB} = \frac{V_{pk}}{2^n}. \] (4.3)

The time it takes to transition between a single quantization level is then

\[ \Delta t = \frac{V_{LSB}}{m_{env}} = \frac{1}{2\pi f_{max} \cdot 2^{n-2}}. \] (4.4)

This is the largest amount of time that could pass after the input signal has crossed the threshold and therefore the worst case time of arrival measurement error. This result is consistent with the earlier observation that the error only depends on the channel bandwidth and the resolution of the ADC. This can be translated to a spatial error using Eq. (2.1) and bearing in mind the TDOA is the difference between two time of arrival measurements which may result in up to twice the error.

\[ \Delta d = \frac{c}{2\pi f_{max} \cdot 2^{n-3}}. \] (4.5)

The choice to use the largest possible channel bandwidth is now clear. Using \( f_{max} = 80 \text{ MHz} \) and \( n = 6.5 \) (effective number of bits for DAU ADC at an operating frequency of 600 MHz) the magnitude of the spatial localization error can be calculated using Eq. (4.5) to be \( \Delta d = 53 \text{ mm} \).

### 4.3 Effects of Random Noise

Noise is an unavoidable limitation in RF communication systems. The main contributors of random noise for a Wi-Fi based indoor RF system are interference and multipath propagation [11]. Noise on the signal at the input to the ADC causes random fluctuations on the envelope rising edge which can have a considerable impact on the TDOA measurement for low SNR values. Since the rising edge fluctuations due to noise are random, the impact on position uncertainty must be analyzed statistically. In this section it will be determined how...
many packets must be gathered and averaged in order to be 95\% confident that the resulting calculation lies within a one meter radius of the nominal position in the presence of noise.

4.3.1 TDOA Measurement Bounds

The first step is to determine how much TDOA error can be tolerated in a measurement in order to maintain positioning accuracy within one meter. It was shown in Fig. 4.4(b) that the positioning accuracy depends on the receiver geometry as well as the position of the target transmitter in the room. There are an infinite number of choices for the receiver geometry and the transmitter position. Therefore, it becomes necessary to make some assumptions to limit the possible degrees of freedom. The following analysis will consider a single receiver geometry. Typical applications would have the receivers mounted on a wall with moderate spacing between receivers to balance the tradeoff between the size and the accuracy system requirements (see Section 3.1.5).

Figure 4.6 shows a coverage map for the assumed receiver geometry where receivers are centered on one wall and spaced 2 meters apart. Based on this coverage map it can be observed that the worst-case positioning occurs at the four corners of the room. With three receivers, two TDOA values will be measured. It is possible for either or both of these measurements to contain an error. However, if the sum of the two errors is held constant, the position deviation is always maximum when the TDOA error is fully contained in one of the measurements while the other remains correct.

In order to validate the claim that the corners present the worst-case positioning in the room, a fixed TDOA error of ±100 picoseconds was added to the TDOA value that resulted in the largest position deviation for the four transmitter positions (TX1-TX4) shown in Fig. 4.6. This was also done for a more typical transmitter position (TX5) for comparison. Table 4.1 gives the corresponding position error.

Table 4.1 shows that the four corners of the room (i.e. positions: TX1 - TX4) do
Figure 4.6: Coverage map for the assumed receiver geometry. Four potential, worst-case transmitter positions (TX1-TX4) and one typical position (TX5) were compared.

<table>
<thead>
<tr>
<th>TX Position</th>
<th>Position Error (TDOA + 100 psec)</th>
<th>Position Error (TDOA - 100 psec)</th>
</tr>
</thead>
<tbody>
<tr>
<td>TX1</td>
<td>0.89 m</td>
<td>1.00 m</td>
</tr>
<tr>
<td>TX2</td>
<td>0.91 m</td>
<td>1.00 m</td>
</tr>
<tr>
<td>TX3</td>
<td>0.91 m</td>
<td>1.00 m</td>
</tr>
<tr>
<td>TX4</td>
<td>0.89 m</td>
<td>1.00 m</td>
</tr>
<tr>
<td>TX5</td>
<td>0.24 m</td>
<td>0.24 m</td>
</tr>
</tbody>
</table>

Table 4.1: Position error for the five transmitter positions shown in Fig. 4.6.

Indeed result in reduced accuracy compared with the position of TX5 in the center of the room. Thus, it has been determined that the TDOA must be measured to within ±100 picoseconds in order to achieve the one meter accuracy requirement for all locations in the 10 m × 10 m room.
4.3.2 Statistical Analysis

The TDOA value is determined as the difference between two TOA measurements. Therefore, it is actually TOA measurement errors that need to be analyzed. The permissible TOA error is half that of the TDOA because the TDOA is the difference between two TOA measurements. To understand how random noise affects the TOA measurement consider the noisy Wi-Fi packet depicted in Fig. 4.7(a). It will be shown in Chapter 5 that the minimum required signal-to-noise ratio (SNR) is 11.4 dB. This is the most limiting case and is therefore the SNR considered in Fig. 4.7(a) and for the subsequent analysis. The impact of the noise can be observed by simulating the packet several times with different sets of random noise. Figure 4.7(b) shows how the threshold crossing, indicating the TOA, is blurred in the presence of noise for 1000 trials.

![Figure 4.7](image_url)

Figure 4.7: (a) An incoming Wi-Fi packet with the minimum allowable SNR. (b) Zoomed in view of the packet’s TOA. Random noise on the signal results in TOA error.

The goal is to determine how many packets must be gathered and averaged in order to be 95% confident that the resulting calculation lies within a one meter radius of the nominal position in the presence of noise. This is equivalent to finding a 95% confidence interval for which the TOA measurement is within ±50 picoseconds. Assuming the TOA measurement
has a Gaussian distribution, a 95% confidence interval is defined as [12]

\[
\left(\overline{TOA} - 1.96 \cdot \frac{\sigma_{TOA}}{\sqrt{n}}, \overline{TOA} + 1.96 \cdot \frac{\sigma_{TOA}}{\sqrt{n}}\right).
\] (4.6)

where \(\overline{TOA}\) is the nominal time of arrival measurement, \(\sigma_{TOA}\) is the standard deviation of the time of arrival with minimum SNR, and \(n\) is the number of packets averaged to calculate the time of arrival. The value of \(\sigma_{TOA}\) can be determined by gathering several TOA simulations (as shown in Fig. 4.7(b)) until the standard deviation changes by less than 0.1%. This resulted in

\[
\sigma_{TOA} = 440 \ psec.
\] (4.7)

Therefore, the number of packets required to position with a one meter accuracy 95% of the time requires

\[
n = \left(1.96 \cdot \frac{\sigma_{TOA}}{50 \ psec}\right)^2 \approx 300 \ packets.
\] (4.8)

This limits the update rate to at least the amount of time it takes to gather 300 packets.

Figure 4.8 shows the distribution of the simulated TOA for 1000 packets with minimum SNR. The mean of this data approaches the actual, nominal TOA. Figure 4.9 shows a comparison of how the running mean, median, and mode approach the nominal TOA. In Fig. 4.9(a) the TOA mean fluctuates until about 300-400 packets have been received and the mean settles very close to the nominal TOA value. The value of \(\sigma_{TOA}\) is nearly nine times larger than the 50 picoseconds TOA tolerance indicating that outliers should be expected. The median of a dataset is less sensitive than the mean to outliers. The TOA median in Fig. 4.9(b) provides support to this claim. The TOA median settles to the nominal TOA value about 100 packets earlier than the TOA mean.
Figure 4.8: Histogram of the TOA for 1000 packets in the presence of random noise.
Figure 4.9: Comparison of the (a) TOA mean and (b) TOA median considering 1000 received packets.
CHAPTER 5. ANALOG FRONT END REQUIREMENTS

The analog front end (AFE) block conditions the signal presented to the DAU board. In keeping with the system requirements laid out in Section 3.1, the main AFE requirements include:

1. Removing interference and the image frequency [11] so that only the desired Wi-Fi signal is sampled and processed.

2. Minimizing AFE noise contribution in order to maximize system sensitivity.

3. Down-converting the signal frequency to one that can be sampled at a 3 GHz sampling frequency without aliasing.

4. Amplifying the received signal adequately to utilize the full resolution of the ADC.

5. Preserving the time difference of arrival through each channel.

6. Capable of using 160 MHz channel bandwidth when the technology becomes available.

7. Configurable to allow the use of all 6 × 80-MHz channels in the 5-GHz Wi-Fi band.

These requirements can be achieved using the standard 4-channel superheterodyne architecture shown in Fig. 5.1. This chapter will describe the key parameters and design choices for meeting the listed AFE requirements.

5.1 Sensitivity and Noise Figure

In the context of the localization system the input sensitivity can be defined as the minimum input signal power $P_{\text{sig,in}}$ such that the output signal to noise ratio $SNR_{\text{out}}$ exceeds the minimum SNR required by the 802.11 standard. In order to be Wi-Fi compliant the AFE
must have a bit error rate (BER) of less than $10^{-5}$. Considering the orthogonal frequency division multiplexing (OFDM) channel encoding used for Wi-Fi, this BER translates to an output SNR requirement given in Eq. (5.1) [13].

$$SNR_{out} \geq 11.4 \text{ dB} \quad (5.1)$$

The minimum expected input signal power (i.e., the sensitivity) is defined by the 802.11 standard as $P_{\text{sig,in},\text{min}} = -65$ dbm [6]. Allowing for a 5 dB margin, the system should operate with

$$P_{\text{sig,in}} \geq -70 \text{ dBm.} \quad (5.2)$$

Assuming the input is matched, the input noise power is given by $P_{n,in} = kTB$ where $k$ is the Boltzmann constant, $T$ is the temperature in Kelvin, and $B$ is the system bandwidth in Hz. The total system noise figure $NF_{\text{tot}}$ is then given by Eq. (5.3).
\[ NF_{\text{tot}} = SNR_{\text{in}} - SNR_{\text{out}} = P_{\text{sig,in}} - 10 \log (kTB) - SNR_{\text{out}} \quad (5.3) \]

Therefore, a constraint on the system noise figure can be calculated using Eqns. (5.1)-(5.3).

\[ NF_{\text{tot}} \leq 9.6 \, \text{dB} \quad (5.4) \]

The impact of this noise figure constraint on the individual block requirements can be seen using the Friis formula [14]. If the block gain is designated \( G_i \) and the block noise factor is designated \( n_i \) then the total system noise factor \( n_{\text{tot}} \) is given by

\[ n_{\text{tot}} = n_{\text{BPF}1} + \frac{n_{\text{LNA}} - 1}{G_{\text{BPFF}}} + \frac{n_{\text{LNA}} - 1}{G_{\text{LNA}}} + \frac{n_{\text{BPFF}} - 1}{G_{\text{BPFF}} \cdot G_{\text{LNA}} \cdot G_{\text{MIX}}} + \frac{n_{\text{AGC}} - 1}{G_{\text{BPFF}} \cdot G_{\text{LNA}} \cdot G_{\text{MIX}} \cdot G_{\text{BPFF}}} \quad (5.5) \]

It is safe to assume that both of the bandpass filters will be passive because of the frequencies at which they will be operating. It can be shown that any passive block has a noise factor equal to its loss which is equal to the inverse of its gain [11]: \( n_{\text{passive}} = L_{\text{passive}} = (G_{\text{passive}})^{-1} \). The total noise factor expression now becomes

\[ n_{\text{tot}} = L_{\text{BPFF}1} \cdot n_{\text{LNA}} + \frac{L_{\text{BPFF1}} (n_{\text{MIX}} - 1)}{G_{\text{LNA}}} + \frac{L_{\text{BPFF1}} (L_{\text{BPFF2}} - 1)}{G_{\text{LNA}} \cdot G_{\text{MIX}}} + \frac{L_{\text{BPFF1}} \cdot L_{\text{BPFF2}} (n_{\text{AGC}} - 1)}{G_{\text{LNA}} \cdot G_{\text{MIX}}} \quad (5.6) \]

Another fair assumption is that the LNA gain \( G_{\text{LNA}} \) and the mixer conversion gain \( G_{\text{MIX}} \) will be relatively large compared to the passband loss of the filters \( L_{\text{BPFF1}} \) and \( L_{\text{BPFF2}} \) as well as the noise factor of the AGC system \( n_{\text{AGC}} \) so the last two terms of Eq. (5.6) can be safely neglected. Thus a good approximation of the total noise factor is given by

\[ n_{\text{tot}} = L_{\text{BPFF1}} \left( n_{\text{LNA}} + \frac{n_{\text{MIX}} - 1}{G_{\text{LNA}}} \right) \quad (5.7) \]

Equation (5.7) plainly shows that in order to minimize the total system noise factor the LNA gain should be large while its noise factor is kept small. The mixer noise factor must also be small but its gain value does not appear in Eq. (5.7). As long as \( G_{\text{MIX}} \) is moderate
to large the earlier assumption is validated and the absolute value does not affect the noise factor.

Assuming a conservative estimate of the RF filter loss as $L_{BP F1} = 3 \, dB$ the design relationship between $G_{LNA}$, $n_{LNA}$, and $n_{MIX}$ can be derived by combining Eq. (5.4) with Eq. (5.7).

$$n_{LNA} + \frac{n_{MIX} - 1}{G_{LNA}} \leq 7.94 \quad (5.8)$$

Thus the LNA and mixer components must be selected such that Eq. (5.8) is satisfied to ensure the AFE meets the SNR, sensitivity, and noise figure requirements.

### 5.2 Interference and Image Rejection

The signal received at each antenna is assumed to be at a fairly low level (around -70 dBm to -30 dBm depending on the distance from a mobile device to the antenna cluster). In addition, the antennas are optimized for both the 2.4 GHz and 5 GHz Wi-Fi bands so it is likely that the signal is corrupted by interference; especially from the 2.4 GHz band. The RF filters are responsible for passing the entire 5 GHz Wi-Fi band (5.15 GHz – 5.85 GHz) while removing the 2.4 GHz band (2.4 GHz – 2.5 GHz).

In Section 5.1 it was determined that the minimum input signal power in the desired 5 GHz band would be $P_{5 GHz,min} = -70 \, dBm$. In order to estimate the maximum expected power in the interfering 2.4 GHz band, a simple experiment was performed. By placing a wireless router very close to a Wi-Fi channel power measurement device, it was determined that the maximum power in the 2.4 GHz band was $P_{2.4 GHz,max} = -30 \, dBm$. Therefore, the RF BPF loss at the edge of the 2.4 GHz band $L_{BP F1} (2.4 GHz)$ should be sufficient to attenuate the maximum 2.4 GHz level to less than the minimum desired 5 GHz level.

$$L_{BP F1} (2.4 GHz) \geq P_{2.4 GHz,max} - P_{5 GHz,min} \approx 40 \, dB \quad (5.9)$$
The RF filters will also aid in reducing the image frequency before downconversion. If the desired 80 MHz RF channel is centered at $f_{RF}$, the desired intermediate frequency is centered at $f_{IF}$, and if low-side injection is assumed such that the local oscillator frequency $f_{LO} = f_{RF} - f_{IF}$ is less than $f_{RF}$ then the image band $B_{img}$ is given by

$$B_{img} = [(f_{RF} - 40 \text{ MHz}) - 2f_{IF}] - [(f_{RF} + 40 \text{ MHz}) - 2f_{IF}].$$ \hspace{1cm} (5.10)

Equation (5.10) shows that for most IF’s in the hundreds of MHz range the image band will probably lie somewhere between 3 GHz – 5 GHz. The FCC regulated applications in this band deal primarily with space-to-earth satellite communications and aeronautical radionavigation which are very weak in most locations so the image frequency is not expected to cause any issues [15].

In Fig. (2.2) it can be seen that many of the 80 MHz channels are spaced one directly after the other in frequency. The potential that an adjacent channel is occupied exists and can cause significant problems if the adjacent channel signal is allowed to pass through to the localization algorithm. The purpose of the channel select BPF is to remove this adjacent channel interference. The main design parameters are the IF center frequency $f_{IF}$ and the bandwidth $B_{IF}$. These parameters should be chosen in order to achieve the sharpest filter rolloff possible.

To begin the choice of where to place the IF consider the following. It is possible to design sharper filters when they are centered at lower frequencies meaning the channel select filter can attenuate interference more if a lower IF is used. On the other hand Eq. (5.10) shows that the larger the IF, the farther the image frequency will be from the desired RF frequency. This allows the input RF filters to attenuate the image frequency more before mixing occurs. Thus there is a tradeoff between the interference rejection and image rejection [11]. It has been argued that image rejection is not a concern so it is expected that nearby interferers should take priority. Therefore, it would be prudent to use the lowest IF possible.

Unfortunately the input matching network to the ADC on the DAU is designed for
relatively high frequencies so – in keeping with the requirement not to alter the DAU board – a compromise has to be made. The input network to the ADC and its measured impedance are shown in Fig. 5.2. To achieve the lowest IF possible while still maintaining a good 50Ω match the lower bound of the IF should be larger than 400 MHz. Besides the filter rolloff limitations, the upper bound also needs to be limited in order to avoid aliasing by the 3 GHz ADC. The Shannon sampling theorem states that information is preserved as long as the sampling rate is at least double the max signal frequency $f_s > 2f_{max}$ [16]. Realistically it is better to have $f_s$ much larger than $2f_{max}$ to relax the specification for the data recovery filter. Using a conservative choice of $f_s > 4f_{max}$, the upper bound of the IF should be less than 750 MHz.

Due to the limited options for filters centered in the $400 MHz \leq f_{IF} \leq 750 MHz$ range, the IF was chosen to be

$$f_{IF} = 600 MHz.$$ \hspace{1cm} (5.11)

In order to maximize the envelope rise time, it is desirable to use 80 MHz channels. However, the 802.11ac standard allows the use of 160 MHz channels as well. It would be best to design the system to allow for 160 MHz channels when the technology supports it. Therefore the channel select BPF should have bandwidth equal to or slightly larger than 160 MHz. Due to the limited filter options, the bandwidth was chosen to be

$$B_{IF} = 200 MHz.$$ \hspace{1cm} (5.12)

With $f_{IF}$ and $B_{IF}$ selected the filter should then be chosen to have the largest rolloff available in order to minimize the effects of adjacent channel interference.

It should be noted that with a 200 MHz bandwidth filter using 80 MHz channels there is 120 MHz of unused frequency band. Therefore, referring to Fig. 2.2, 40 MHz of both adjacent channels will be passed. The U-NII-2 band (channels 58, 106, 122, and 138 in Fig. 2.2) is typically not used in commercial products because the band is not exclusively
dedicated to Wi-Fi and the FCC has placed restrictions on the maximum power transmitted in the band [5]. As long as testing is restricted to channels 106 and 122 the adjacent channel will be unoccupied and not cause interference issues with testing the prototype.

![AFE schematic](image)

Figure 5.2: (a) Single-ended to differential conversion and impedance matching network at ADC input. (b) Measured and simulated input impedance to the DAU.

### 5.3 Gain Range

It is important that the AFE output signal is scaled properly to utilize the full resolution of the ADC. The AFE input signal power level will vary because of factors like transmitter power, FSPL, antenna directivity and polarization, multipath fading, etc. [11]. Since it is desired that the output signal power level remain constant for a range of input levels the AFE must have a variable gain. The range of the AFE gain will need to equal or exceed the expected range of input signal power levels. In Section 5.1 it was determined that the minimum input signal power would be $P_{\text{sig,in,min}} = -70$ dBm. To estimate the maximum expected signal power would require many difficult assumptions about the influencing factors listed above. A better method of determining the maximum expected input signal power is to measure the signal power a short distance from the mobile device that will be used to
test the system. Performing such an experiment it was found that $P_{\text{sig,in,\,max}} = -30 \text{ dBm}$. Therefore the gain range required by the AFE is

$$\text{Gain Range} \geq P_{\text{sig,in,\,max}} - P_{\text{sig,in,\,min}} = 40 \text{ dB} \quad (5.13)$$

To achieve this variable gain, the AFE incorporates a gain control system at its output as shown in Fig. 5.1. The system works by sensing the input signal strength via a received signal strength indicator (RSSI) which then controls the gain of a variable gain amplifier (VGA). All other system blocks have a constant gain. It is the gain of the VGA that must satisfy the gain range requirement. It is also important to note that the RSSI system must remain linear with respect to the input over this 40 dB range. A more detailed description of the gain control system can be found in Section 8.2.

### 5.4 Tuning Range and Phase Noise

The local oscillator (LO) is unique in the system of Fig. 5.1 in that there is only one LO that is shared between the four receiver channels. Simply put, this is because it is required to match the four receiver channels as closely as possible. In other words, each receiver channel should manipulate the signal in precisely the same way. A more detailed view of the local oscillator system is shown in Fig. 5.3.

The parameters of interest for the LO are its tuning range and phase noise. The full tuning range is straightforward to specify. Since the channel select filter is centered at a fixed frequency $f_{\text{IF}}$ and it is desired to possess the ability to use any of the available channels in the 5 GHz band, the tuning range must be as wide as the 5 GHz band.

$$\text{LO Tuning Range} \geq (5.15 GHz - f_{\text{IF}}) - (5.85 GHz - f_{\text{IF}}) = 700 \text{ MHz} \quad (5.14)$$

The 802.11 standard requires the LO phase noise to be less than -111 dBc/Hz at 25 MHz offset to mitigate the interference caused by additional oscillator tones mixing a portion
of the adjacent channel to the desired channel \cite{13}, \cite{17}. The phase noise will be contributed almost exclusively by the voltage controlled oscillator (VCO) in the programmable oscillator so this component must be selected to meet the requirements. The 802.11 standard also specifies that the frequency drift must be less than 25 ppm \cite{18}. Assuming the programmable oscillator makes use of a phase locked loop (PLL) then the frequency drift is almost entirely determined by the reference oscillator drift. It should be noted that the localization system is meant to operate indoors so temperature variation in the reference oscillator can be considered negligible.

In order to switch between various channels it is necessary for there to be some kind of user interface to the LO. This is most easily done with a digital controller which means it is also important to specify the minimum frequency step size. The smallest tuning step required is equal to the smallest adjacent channel spacing which is 20 MHz.

\begin{equation}
\text{LO Frequency Steps} < 20 \text{MHz}
\end{equation}
5.5 Propagation Delay Matching

Since TDOA localization requires precise time difference measurements, it is critical that the timing information is preserved through the system. Another way to state this is that the delay through the analog system – between the antenna and the analog-to-digital-converter – must be matched for all four channels. The degree with which the receiver channels must be matched is dictated by the sampling rate of the ADC since this sets the minimum time resolution required of the analog system. The difference in the delay between any two channels $\Delta \tau_d$ should be much less than the sample period $T_s$.

$$\Delta \tau < \frac{T_s}{10} \quad (5.16)$$

If the printed circuit board (PCB) transmission lines have an effective dielectric constant $\varepsilon_{eff}$ then the propagation velocity is $c/\sqrt{\varepsilon_{eff}}$ and a constraint on the permissible difference in the channel lengths $\Delta L$ can be derived.

$$\Delta L < \frac{c}{\sqrt{\varepsilon_{eff}}} \cdot \frac{T_s}{10}$$

A standard FR4 dielectric PCB with microstrip transmission lines will have $3 \leq \varepsilon_{eff} \leq 3.25$ and the sampling period is set at $T_s = (3 GHz)^{-1}$ by the DAU design therefore it is required that $\Delta L < 5.5 \text{mm}$. This can be achieved with a careful design of the AFE PCB.
CHAPTER 6. ANALOG FRONT END DESIGN

The analog front end has been defined and specified. This chapter focuses on some of the nuances involved in moving from the specifications to a hardware implementation.

6.1 Component Selection

Table 6.1 lists the components and their relevant parameters chosen to meet the requirements laid out in Chapter 5. All components were selected for an off-the-shelf implementation. Note the local oscillator system parameters are considered separately.

The RF and channel select filters were limited in options and were selected based on the bandwidth requirements discussed in Section 5.2. The VGA was implemented on a board separate from the AFE because it was already available and capable of meeting the gain range requirement of Eq. (5.13). The LNA and mixer were then chosen to satisfy the noise figure requirement given by Eq. (5.7). Equation (5.7) contains three variables $G_{LNA}$, $n_{LNA}$, and $n_{MIX}$ so a simple, low cost LNA was chosen thus constraining the mixer to

$$N_{FMIX} \leq 21 \text{ dB}$$

The mixer was selected to meet this requirement and also for its compatibility with the local oscillator component.

Considering the component values listed in Table 6.1 the total system gain is simply the sum of the individual block gains and the total noise figure is calculated by Eq. (5.7).

$$G_{tot} = 5 \text{ dB} - 65 \text{ dB}$$

$$N_{Ftot} = 5.5 \text{ dB}$$
<table>
<thead>
<tr>
<th>Component</th>
<th>Gain (dB)</th>
<th>Noise Figure (dB)</th>
<th>Freq. Range (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>RF Filter</td>
<td>-1.8</td>
<td>1.8</td>
<td>5150 - 5825</td>
</tr>
<tr>
<td>LNA</td>
<td>13.0</td>
<td>1.5</td>
<td>4900 - 5900</td>
</tr>
<tr>
<td>Mixer</td>
<td>3.9</td>
<td>12</td>
<td>N/A</td>
</tr>
<tr>
<td>Channel Select Filter</td>
<td>-1.5</td>
<td>1.5</td>
<td>500 - 700</td>
</tr>
<tr>
<td>VGA</td>
<td>(-8.6, 51.4)</td>
<td>8.5</td>
<td>DC - 1800</td>
</tr>
</tbody>
</table>

Table 6.1: Selected components and their block parameters.

The selected components clearly meet the requirements for sensitivity (Eq. (5.4)) and gain range (Eq. (5.13)).

<table>
<thead>
<tr>
<th>Component</th>
<th>Tuning Range</th>
<th>Frequency Steps</th>
<th>Phase Noise</th>
<th>Frequency Drift</th>
</tr>
</thead>
<tbody>
<tr>
<td>Local Oscillator</td>
<td>4540 - 5330 MHz</td>
<td>10 MHz</td>
<td>-150 dBc @ 25 MHz</td>
<td>20 ppm</td>
</tr>
</tbody>
</table>

Table 6.2: Selected local oscillator block parameters.

Table 6.2 lists the relevant local oscillator parameters. The phase noise for the local oscillator is significantly less than the -111 dBc required by the 802.11 specification. Because of the exceptional phase noise performance, the tuning range quoted in Table 6.2 is not defined by phase noise but rather by phase jitter. The selected LO component contains circuitry to measure the phase jitter and an output pin to communicate whether the jitter falls within a specified range. The quoted tuning range is defined as the range in which the phase jitter status bit is asserted under the strictest conditions offered by the component. The LO parameters are well within the specifications detailed in Section 5.4.

6.2 Impedance Matching

To maximize the power transferred from a source to a load, or equivalently from one RF block to another, the input and output impedances must be matched over the desired
frequency range. Almost all of the selected components were predesigned with 50 Ω input and output impedances for the desired frequency range. This is useful because it allows components to be connected together with guaranteed impedance matching. The only exception to this is the non-50 Ω mixer output. The chosen mixer component is specified to operate from DC to 1 GHz so it may be used for any intermediate frequency band in this range. An output matching network is required to match the output over the desired IF band.

6.2.1 Mixer Output

![Matching network diagram](image)

Figure 6.1: Matching network used to connect the mixer to the channel select bandpass filter.

The matching network at the mixer output is shown in Fig. 6.1. The output model was provided in the mixer datasheet. Inductors $L_1$ and $L_2$ resonate with $C_m$ at the output frequency $f_{IF}$. This removes the reactive component at $f_{IF}$ so the mixer output impedance is equal to the resistive component $R_m$. This is accomplished by choosing $L_1$ and $L_2$ values according to Eq. (6.4).

$$L_1 = L_2 = \frac{1}{(2\pi f_{IF})^2 (2C_m)}$$  \hspace{1cm} (6.4)

This calculation relies on the assumption that the inductance looking into the primary side of the transformer is much larger than $L_1$ and $L_2$ so that the transformer’s contribution
to the total inductance is negligible.

The mixer output is differential but the input to the channel select filter is single-ended. Therefore, the transformer \(X_1\) functions as a balun as well as transforms the relatively large mixer output closer to the 50\(\Omega\) resistance required from the perspective of the filter. This is accomplished by choosing the turns ratio \(T\) of the transformer (defined primary/secondary) as:

\[
T = \sqrt{\frac{R_m}{|Z_L|}} = 2.4.
\] (6.5)

Transformers typically only come with integer turn ratios so a turn ratio of 2.4 is not practical for this prototype. Available options for \(T\) at 600 MHz are 2 or 4. It can be shown that the power delivered to the filter \(P_L\) depends on \(R_m\), \(Z_L\), and \(T\) according to [19]:

\[
P_L \propto \frac{R_m^2 |Z_L| T^2}{(R_m + |Z_L| T^2)^2}.
\] (6.6)

For the available options of \(T\):

\[
T = 2 : \quad P_L \propto 70.1 \Omega
\] (6.7)

\[
T = 4 : \quad P_L \propto 56.6 \Omega.
\] (6.8)

It is desired to maximize the power throughput. Therefore, the best option is to use \(T = 2\). This results in 96% power transfer compared to the case when \(T = 2.4\) so the slight resistive mismatch leads to 0.2 dB less gain through the mixer which is permissible.

6.2.2 Transmission Lines

Now that it is established that all components have matching input and output impedances, it is necessary to consider the effects of the board interconnect. The metal traces used to connect one component to another can significantly alter the impedances at either end of the
line as the trace length approaches the wavelength of the signal it carries. Erring on the side of caution and assuming that transmission line effects become significant at one sixteenth of a wavelength, then we can calculate the maximum trace length above which it becomes necessary to control the impedance \[20\].

\[
L_{\text{max}} = \frac{\lambda}{16} = \frac{c}{16 f_{\text{max}} \sqrt{\varepsilon_{\text{eff}}}} = 1.8 \text{ mm} \quad (6.9)
\]

Due to the physical size of components and soldering considerations, it is unreasonable to use traces shorter than 1.8 mm. Therefore, careful transmission line design is required for optimal matching between components.

Printed circuit board (PCB) traces are typically layered with thin strips of copper separated by a dielectric. This geometry makes the microstrip transmission line, shown in Fig. 6.2, a natural choice for component interconnect. Assuming the strip width, \( W \), is greater than the strip height above the ground plane, \( h \), then the characteristic impedance, \( Z_0 \), depends on the microstrip dimensions, \( W/h \), and the relative dielectric permittivity, \( \varepsilon_r \), according to \[19\]:

\[
\varepsilon_{\text{eff}} = \frac{\varepsilon_r}{2} + \frac{\varepsilon_r - 1}{2} \cdot \frac{1}{\sqrt{1 + 12 \cdot \frac{h}{W}}} \quad (6.10)
\]

\[
Z_0 = \frac{120\pi}{\sqrt{\varepsilon_{\text{eff}}}} \cdot \frac{1}{\left[ \frac{W}{h} + 1.393 + 0.677 \cdot \ln \left( \frac{W}{h} + 1.444 \right) \right]} \quad (6.11)
\]

Most PCB manufacturers have a standard spacing between metal layers for conventional builds. The PCB manufacturer used for this work uses \( h = 9.3 \) mils. Knowing \( \varepsilon_{r,FR4} = 4.3 \) allows the required microstrip width to be calculated for \( Z_0 = 50 \Omega \).

\[
W = 18.5 \text{ mils} \quad (6.11)
\]
6.3 PCB Layout

It was necessary to design the PCB layout such that all four receiver channels had the same path length. In addition, the LO path must also be matched for each channel. The pin out for the mixer highlighting the main 3 ports is given in Fig. 6.3(a), the PCB layout geometry used to accomplish the channel matching goals is shown in Fig. 6.3(b), and a photo of the final populated PCB is shown in Fig. 6.4.
Figure 6.3: (a) Orientation of the mixer signal ports. (b) Block diagram layout plan for matching channel lengths and LO paths.
Figure 6.4: PCB layout demonstrating the symmetric channel length design from Fig. 6.3(b).
CHAPTER 7. MEASURED RESULTS

This chapter describes the measured performance of the AFE as well as the measured performance of the complete localization system incorporating the AFE. The final system meets the basic requirements of a general purpose indoor localization system and the system achieves the benchmark goal of a 1 meter localization accuracy.

7.1 Analog Front End Results

7.1.1 Gain and Bandwidth

The AFE gain and bandwidth for each channel were measured by applying a swept input frequency and measuring the output power using a spectrum analyzer. The AGC gain was set to 0 dB for this measurement. The result is shown in Fig. 7.1. The gain in the center 80 MHz band is measured at about 7 dB and the bandwidth is 500 MHz – 700 MHz as set by the channel select filter. This gain is 6 dB lower than the expected value. It will be shown in Section 7.1.2 that this is due to a mismatch at the input.

7.1.2 S-Parameters

The input and output matching were evaluated using an S-parameter network analyzer. Figure 7.2 shows the input and output reflection S-parameters.

Figure 7.2 shows the impedance is well matched at the output however the input match is less than ideal. The source of this mismatch is likely a combination of the interface between the SMA connector and the PCB trace interface along with the input impedance of the RF filter. This mismatch will introduce the reduction in gain apparent in Fig. 7.1. The additional loss, compared to the case when the input is matched, is given by the mismatch loss, $ML$. 
Figure 7.1: Measured gain and bandwidth for each channel in the AFE.

(a) (b)

Figure 7.2: (a) The ratio of the input voltage that is reflected at the AFE input. (b) The ratio of a voltage applied at the output that would be reflected back to the source.
defined in Eq. 7.1 [19]. The mismatch loss is plotted across frequency in Fig. 7.3.

\[ ML = -10 \log \left( 1 - |s_{11}|^2 \right) \]  

(7.1)

Figure 7.3: Loss introduced by the input mismatch.

7.1.3 Noise Figure and Output SNR

The loss introduced by the input mismatch will have a noticeable impact on the system noise figure. The loss can be modeled as a passive attenuator placed before the receiver system as shown in Fig. 7.4. The total system noise figure for this system can be derived using the Friis equation.

\[ NF_{tot} = ML + NF_{RX}. \]  

(7.2)

In other words, the noise figure is increased by the amount of the mismatch loss shown in Fig. 7.3. At the nominal RF frequency of 5530 MHz the input mismatch loss is 4.8 dB. Accounting for this loss a system level simulation was performed with the resulting system
noise figure

\[ NF_{tot} = 11.2 \text{ dB} \quad (7.3) \]

which is consistent with the prediction of Eq. (7.2). The resulting noise figure is somewhat degraded from the nominal noise figure predicted by Eq. (6.2). Using Eq. (5.3) the output SNR can be calculated as shown in Eq. (7.4).

\[ SNR_{out} = P_{\text{sig,in}} - NF_{tot} - 10 \log (kTB) = 9.8 \text{ dB} \quad (7.4) \]

\[
\begin{align*}
\text{Input} & \quad \text{Mismatch} \\
\text{Loss} & \\
\text{RF System} & \quad \text{RF System}
\end{align*}
\]

\[ NF = ML = NF_{RX} \]

Figure 7.4: Equivalent stage 1 for noise figure degradation estimate.

This output SNR translates to a bit error rate of \( BER = 30 \times 10^{-5} \) [13] which is 30 times larger than that required for Wi-Fi compliance. However, it is not necessary to demodulate the Wi-Fi data in order to test the localization system so the AFE is still perfectly suitable for validating the prototype. The reduced SNR will also have an effect on the random noise analysis of Section 4.3. Considering a minimum output SNR of 9.8 dB results in a TOA measurement standard deviation of

\[ \sigma_{TOA} = 650 \text{ psec} \quad (7.5) \]

which means it is now required to gather
\[ n = \left( 1.96 \cdot \frac{\sigma_{TOA}}{50 \text{ psec}} \right)^2 \approx 650 \text{ packets} \quad (7.6) \]

in order to achieve the required one meter accuracy. Another way of stating this is that the update rate must be reduced by a little less than half in order to maintain the desired accuracy in the presence of the mismatch.

### 7.1.4 Propagation Delay Mismatch

The propagation delay must be well matched for each channel in the AFE to preserve the TDOA. To measure the propagation delay mismatch, a Wi-Fi signal was transmitted to a single antenna which was split four ways to be used as a phase-matched input signal. A diagram of the measurement setup is given in Fig. 7.5. The output signal was then measured with an oscilloscope sampling at a rate of 10 GS/sec. Figure 7.6 shows a zoomed in view of the IF carrier phase offset demonstrating the delay spread. The spread is less than one sample of the oscilloscope but the delay can be observed to be about 100 picoseconds in this region. Using linear interpolation a more precise estimate of the propagation delay mismatch was calculated for each zero crossing in Fig. 7.6(a). This result is given in Fig. 7.7 with a maximum calculated delay mismatch of 145 picoseconds. Thus, it can be concluded that the AFE preserves the TDOA to less than 150 picoseconds which is less than one half of the sampling period.

![Figure 7.5: Test setup for measuring the channel propagation delay mismatch.](image-url)
Figure 7.6: (a) Snapshot of a few carrier cycles in the received Wi-Fi packet. (b) Expanded view of the boxed region in (a). The delay mismatch at the zero crossing is about 100 picoseconds.

Figure 7.7: Channel delay at each zero crossing in Fig. 7.6(a).
7.2 Localization System Results

The first test for the localization system was to measure the TDOA measurement variation for a fixed transmitter position. Ideally the TDOA calculation should not change from one packet to the next, however slight variations will occur due to random measurement errors (see Chapter 4) and system propagation delay mismatch. The calculated TDOA for a four antenna cluster as a function of packet number is shown in Fig. 7.8(a). The vertical axis is converted to distance (using Eq. (2.1)) to give a better feel for the localization accuracy. It can be observed from this figure that the variations occur around a central TDOA value. In fact, with enough packets, the statistical mean gives an accurate estimate of the TDOA value as indicated by the Gaussian distribution of the data shown in Fig. 7.8(b). Hence, the more measurements the system is allowed to take for a TDOA calculation the more accurate that calculation will be. This fact demonstrates the tradeoff between localization accuracy and update rate.

Figure 7.8: Measured TDOA distribution for a fixed transmitter position and 100 received packets. Time has been converted to distance to provide a better understanding of the spatial uncertainty.
Figure 7.9 provides a better way to visualize how the TDOA measurement variation affects the position calculation. Figure 7.9 gives a bird’s eye view of the networking lab where the localization experiments were performed with two antenna clusters. Each plot in the figure shows a different transmitter location while the two antenna cluster locations remain fixed. A set of 5000 position calculations (red dots) is shown for each transmitter position and the ratio of the total number of calculations that fall within 1 meter of the transmitter is given in the lower right corner. It can be concluded from Fig. 7.9 that a single position calculation falls within 1 meter 60% of the time.

By applying some statistical averaging on a set of received packets, the location estimation can be improved. The system performance including averaging is demonstrated in Fig. 7.10. The mobile device was carried on the trajectory shown by the dashed black line. The number of samples used to calculate the position was chosen such that the system could refresh about once every second. The resulting calculated trajectory is shown by the green line. This verifies that the localization system is capable of real time tracking to within 1 meter.
Figure 7.9: Position accuracy of 5000 packets for 4 transmitter locations: (a) transmitter location 1, (b) location 2, (c) location 3, (d) location 4. The transmitter location is represented by the solid blue square in the center of the blue circle which designates a one meter radius from the transmitter. The red circles represent the location estimation for a single packet. The two fixed antenna locations are shown as blue outlined squares connected by a blue line on the left and bottom of each figure.
Figure 7.10: Localization accuracy of the system after averaging for a moving transmitter. The trajectory is shown by the dotted black line and the system calculated trajectory is shown as the solid green line. The mobile device was carried on (a) a rectangular trajectory, and (b) a linear trajectory.
A primary goal of this project is to demonstrate a proof of concept for a TDOA, Wi-Fi based localization approach. While this was achieved, there are a number of design considerations that must be addressed to commercialize this technology. This chapter addresses some of the main requirements in moving toward this goal.

### 8.1 System Improvements

#### 8.1.1 System Integration

The current system consists of an antenna cluster, an analog front end board with external gain control, a data acquisition unit, as well as a PC for performing most of the signal processing. Significant advantages in mobility and aesthetic appeal could be obtained by integrating hardware and software components into a single unified system. The antenna cluster must remain large for performance reasons. However the electronics system size should be made insignificant by comparison. Eventually the electronics could be housed in a compartment within the cluster.

Integrating the current electronic systems will also result in improved localization accuracy. With full control over the design of the RF blocks a designer could make use of the current state of the art to develop a custom solution. In addition the time delay mismatch between receiver channels would be greatly reduced owing to the improved matching capabilities of IC manufacturing.

The main bottleneck in processing speed is the data communication rate between the DAU and the PC. The data sampled at the input of the DAU is stored in memory and then sent to the PC in packets according to the ethernet protocol. A considerable improvement in processing speed could be achieved if the PC and networking protocol were eliminated from
the system by processing the input data locally in the digital back end system. The system would then only need to output data relevant to the position display.

8.1.2 Data Acquisition Unit Renovation

Many of the current performance limitations stem from the requirement to use the preexisting digital back end data acquisition unit (DAU). The DAU was developed to capture ultrawideband (UWB) signals which typically occupy a bandwidth from 1.5 GHz to 4 GHz [13]. As such, the requirements for the ADC in an UWB application are not necessarily aligned with the requirements for localization. For example, it was determined that this application would benefit from using a low intermediate frequency for baseband processing, therefore, a somewhat low speed, high resolution ADC would be optimal. However UWB signals contain much larger frequencies which require a very high speed ADC which forces the use of a lower resolution.

Furthermore, the DAU was designed with four single-ended input channels. Three dimensional TDOA localization requires at least four channels to operate so this forces the AFE outputs to be single-ended. This results in wasted performance in the AFE as most components are intended for differential operation for inherent noise rejection properties.

8.1.3 IncorporatingAutomaticGain Control

The necessity of automatic gain control was described in Sections 4.2 and 5.3. Due to time constraints the gain control is currently accomplished with a separate board external to the AFE. This gain control board was designed with a manual interface for gain adjustment thus eliminating the possibility of any automatic gain control. This requires the gain to be recalibrated and manually set each time the transmitter is moved to a new location and severely limits the coverage area for trajectory measurements.
8.2 Automatic Gain Control System Design

To support future progress in meeting stricter requirements, an Automatic Gain Control (AGC) system was designed at the system level. The performance of the system was validated by simulation.

8.2.1 Architecture

The AGC system is critical in meeting the requirements of the localization system (see Section 5.3). It is responsible for amplifying an input signal, with up to 40 dB of amplitude variation, up to the full scale range of the ADC. The proposed system is shown in more detail in Fig. 8.1. The input power level to the AGC block mainly depends on the distance separating the transmitter from the receiver. The input power level is sensed by the peak detector which outputs a voltage proportional to the logarithm of the input voltage. In other words, the output voltage is linear with respect to the input power measured in decibels. Typically, peak detectors work in the log domain as it allows the peak detector to operate linearly over a greater dynamic range [21]. The peak detector output is then sampled and processed by a microprocessor (μP) which controls the gain of a variable gain amplifier (VGA).

![Figure 8.1: Block diagram of the proposed automatic gain control system.](image)

Recall from Section 2.1 that there are several packet types involved in Wi-Fi com-
communication. These packets will contain varying amounts of data and will therefore differ slightly in their duration. The duration of a typical packet is on the order of about 10 µs. In addition, the rate at which packets are sent and received depends on the availability of the client device and the available spectrum. In an environment free from interference one might expect around 1000 packets per second resulting in a packet duty ratio of 1%. However in a room with a single router serving Wi-Fi to several clients this rate could drop to 100 packets per second or less which results in a sub 0.1% duty ratio. This means that it is reasonable to expect less than 1% of the received signal to carry actual data. This makes the signal strength difficult to detect for two reasons. First, the duty ratio is so low that the average signal strength is essentially equal to the noise floor. Second, there is no way to predict when the signal should arrive due to its intermittency. Thus it is necessary to design the system with a logic element to detect when the signal is present and adjust the gain accordingly. This logic takes place in the microprocessor block (µP) of Fig. 8.1.

To solidify this point, consider an ideal AGC system capable of producing the correct output level for any input level. If the gain were allowed to adjust continuously, the VGA would spend the majority of the time at a very large gain value in an effort to amplify the noise to the desired output level. This noise would likely trigger a false packet detection by the localization algorithm and what’s worse, the VGA would have to ramp down from full gain with each packet arrival; even when the packet strength had not changed. This situation would significantly degrade the system performance and reinforces the need for the microprocessor.

8.2.2 Algorithm

In order to support future revisions of this work, an algorithm was developed with the purpose of detecting the received signal strength from the output signal of the peak detector. A diagram of the basic algorithm is shown in Fig. 8.2. Incoming samples of the serial peak
detector output $V_p[n]$ are stored in a shift register of length $k$. If $k$ is chosen to be sufficiently large then the low duty cycle of the Wi-Fi signal can be exploited by our earlier observation that the average received signal strength is approximately equal to the noise floor. This allows the noise floor to be periodically measured and compared with the current signal strength. Thus we now have a way to only adjust the gain if a Wi-Fi packet is present. When the received packet strength is larger than the noise floor the required gain is calculated as a function of the packet strength.

$$\text{noise}_{\text{floor}} \approx \text{mean}(V_p)$$

- Only adjust gain when input is greater than the noise floor
- Keep track of how fast the input is changing
- If input is changing too fast compress the rate of change of the gain

$$\text{gain}_{\text{out}} = \text{mean}(\cdot)$$

Figure 8.2: Basic algorithm used to only adjust the gain when a signal packet is present.

There is a small issue with this logic: the rise time of the packet envelope is not equal to zero. Consider the example incoming Wi-Fi packets depicted in Fig. 8.3(a). The packets produce the pulsed waveform at the output of the peak detector as shown in Fig. 8.3(b). The AGC algorithm would analyze the peak detector output and determine the noise floor to be slightly greater than 0.8 V. During the rise and fall time of the peak detector output signal the voltage is above the noise floor and, therefore, triggers the gain to change. It can be seen in Fig. 8.3(c) that the gain continues to change until the full packet is received and the input
falls below the noise floor. Thus, the gain setting during the time between packets is set much too high. If the VGA has sufficient gain this will result in the noise levels near the full scale range of the ADC. Figure 8.3(d) shows the impact of this on the AGC output. When a packet arrives there is a brief moment before the gain has had time to adjust which results in a voltage spike at the output. Eventually the gain settles but it is difficult to distinguish between the signal and the noise as they are the same voltage level at the output.

![Figure 8.3](image)

Figure 8.3: (a) Incoming Wi-Fi packets. (b) The peak detector produces output pulses proportional to the log of the input packet peak voltage. (c) The microprocessor computes the required gain when the input exceeds the noise floor. (d) This results in a nearly constant amplitude output as the gain is adjusted such that the signal and noise are the same voltage level at the output.

A solution to this problem is to restrict the gain from changing too rapidly. To accom-
plish this, every time the gain is adjusted, the percent change in the input signal, \( \Delta[n] \), is also calculated. If \( \Delta[n] \) exceeds some factor, \( \Delta_{tol} \), then the gain is adjusted until it is acceptably close to the previous calculated value. Finally, to further reduce the rate of change of the gain setting, the actual gain setting passed to the VGA is equal to the mean of the previous \( k \) gain calculations. This windowed average operation is synonymous with low pass filtering and, therefore, the length of window will determine the settling time of the AGC gain.

In an effort to add a degree of robustness to the algorithm, a tuning parameter for the gain, \( g_{fix} \), was incorporated. This parameter is necessary because the peak detector is linear with respect to the input in the log domain. At some point in the system, the gain setting must be converted back to a dimensionless power gain (units of W/W) for the VGA. Any systematic error, such as nonlinearity in the peak detector or rounding errors in the gain calculation, will be exacerbated by this conversion.

If the desired power gain is \( P_G = g = 10^{G_{dB}/10} \), where \( g \) is measured in W/W and \( G_{dB} \) is measured in dB, but due to some nonlinearity, the output gain, \( G_{dB} \), is larger by a fraction \( \varepsilon \) then the conversion process will look like:

\[
P_G = 10^{(1+\varepsilon)G_{dB}/10} = 10^{G_{dB}/10} \cdot (1+\varepsilon) = g^{(1+\varepsilon)}.
\]

After the conversion the output gain is the desired gain, \( g \), raised to the error, \( (1+\varepsilon) \), whereas the gain, \( G_{dB} \), was only multiplied by the error, \( (1+\varepsilon) \). Thus the output gain setting is made more sensitive to errors by processing in the log domain because a small error in the log domain translates to a larger error in the linear domain. For example, if the desired gain is \( P_G = 10 \) W/W = 10 dB but the calculated gain is 10.5 dB then the final output gain would be 11.2 W/W. A 5% increase in the dB gain leads to a 12% increase in the W/W gain! Therefore a system calibrated for a -40 dBm input signal might not be calibrated for a -20 dBm signal and would require some correction factor.

This gain correction factor, \( g_{fix} \), is also considered in the gain \([n]\) calculation shown in Fig. 8.2. In order to be fully automated, \( g_{fix} \) should be set based on the output packet voltage.
level. Thus the output should be fed back to the ADC and the packet voltage extracted in a manner analogous to the input extraction. This voltage can then be compared to a reference voltage in order to set the optimal value for $g_{fix}$.

8.2.3 Simulation Results

The full automatic gain control block was simulated at the system level. The system performance was verified by testing a variety of realistic input conditions. Figure 8.4 shows all intermediate signals for the AGC system in Fig. 8.1. For this test case the first four input packets have a voltage level around 10 mV but, from the fifth packet onward, the voltage level doubles. The peak detector senses this change and increases the output voltage slightly. The peak detector signal is the input to the gain calculation algorithm. The output of the algorithm is the required gain which is provided to the VGA. Finally the AGC output is plotted after applying the calculated gain. Figure 8.4(d) shows how the AGC system reacts properly to an abrupt level change at the input. After the fourth packet, the gain begins to reduce and settles at the proper level in two or three packet’s time.

Figure 8.5 shows a variety of extreme input signals to provide an idea of the algorithm’s robustness. Case (a) shows a situation similar to that shown in Fig. 8.4 except the level change is much larger at 26 dB. Case (b) shows the reverse situation where the input is suddenly reduced by 26 dB. Case (c) shows three level transitions and each time the output is able to adapt. Case (d) shows a long gap between packet bursts. Notice that the gain is not adjusted during the packet-free time and when the next packet arrives the gain remains right where it was last set. These simulation results are a good indication of the proposed automatic gain controller’s suitability for use in the localization system. The system is able to detect the incoming packet signal strength and amplify the packet to the proper level for analog-to-digital conversion. Additionally, the AGC system is able to respond fast so very few packets are wasted while the VGA gain is being adjusted.
Figure 8.4: Signal flow for the AGC system. (a) Input signal packets with a sudden change in voltage level. (b) The peak detector produces output pulses proportional to the log of the input packet peak voltage. (c) The microprocessor computes the necessary gain reduction based on the peak detector output. (d) After a brief time the AGC output settles to the desired voltage level.
Figure 8.5: Variety of simulated test cases used to test the AGC system design. (a) A 26 dB increase in input level. (b) A 26 dB decrease in input level. (c) Multiple increases and decreases. (d) A long gap in the packet burst.
CHAPTER 9. CONCLUSION

An indoor localization system based on the Wi-Fi standard has been presented. The system is able to locate the target mobile device to within 1 meter 60% of the time based on a single measurement. Through the analysis of several measurements, the system is shown to track the target to within 1 meter at a real time refresh rate of 1 Hz. The system has been tested with two clusters but is capable of incorporating more in order to increase the accuracy and coverage area. These attributes make the localization system suitable for a wide variety of general purpose applications.

An analog front end receiver was developed to function compatibly with a pre-existing system designed for ultrawideband based localization. In many ways the two systems are similar however there are many design choices that limit the performance of the system based on the Wi-Fi implementation. To this end, an evaluation of the issues and a detailed revision plan was presented to support further development of the system.
BIBLIOGRAPHY


