DESIGN AND TESTING OF A SELF-POWERED WIRELESS HYDROGEN SENSING PLATFORM

By

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by

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DESIGN AND TESTING OF A SELF-POWERED WIRELESS HYDROGEN SENSING PLATFORM

By

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May 2006

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Major Department: Electrical and Computer Engineering

Within the ongoing interdisciplinary hydrogen research at the University of Florida, a self–powered wireless hydrogen sensor node has been designed and developed. By using multi-source energy harvesting circuitry designed and developed at the University of Florida, scavenged or “reclaimed” energy from light emitting and vibrational sources serve as the source of power for commercial low power microcontrollers, amplifiers, and RF transmitters. After system power up, the sensor node is capable of conditioning and deciphering the output of hydrogen sensitive ZnO nano-rods sensors also developed at the University of Florida. Upon the detection of a discernible amount of hydrogen, the system will “wake” from an idle state to create a wireless data communication link to relay the detection of hydrogen to a central monitoring station. The system’s sensitivity is on the order of parts per million, and hydrogen concentrations starting as low as 10 PPM can be detected.
The thesis will discuss the performance of the self-powered wireless hydrogen sensor node, and also focus on the design and optimization of the detection circuitry, digital processing considerations, modulation scheme and wireless communications link to maintain an accurate and reliable system, while expending a minimal amount of energy scavenged from ambient light or vibration for very long lifetime operation.
CHAPTER 1
INTRODUCTION

Motivation

Self-powered wireless sensors themselves are becoming a popular topic for implementation in future systems. The idea of inexpensive sensor devices with very long-lifetime operation involving minimal maintenance, is poised to make a very strong impact on the engineering community. Because these devices can be deployed in very harsh and dangerous environs, environments can be monitored remotely in real-time to update the end-user of ambient conditions, and report any slight deviations from safe and normal operating conditions. This can help prevent endangerment to the quality of life for surrounding bodies, without placing expensive machinery and humans in harm’s way.

As fossil fuel supplies are depleted, alternative fuel supplies such as hydrogen which can be quickly replenished, are swiftly growing in popularity, and with a self-powered wireless hydrogen sensing platform capable of sensitivity on orders of parts per million (PPM), the applications of these sensors can include monitoring of hydrogen powered machines, combustion gas detection in spacecrafts, and solid oxide fuel cells with proton-exchange membranes [1,2].

With a self-powered wireless hydrogen sensor, and the current advances in system-on-chip technology and MEMS processes, an integrated circuit capable of harvesting energy supplies through ambient conditions such as lighting and vibration, can be realized in a physically small package to control and report the dangers involved in the development of hydrogen as a viable fuel source for the future. The overall block
Contributions

The objective of this research is to develop a self-powered wireless hydrogen sensor node powered through energy harvesting techniques and capable of PPM hydrogen sensitivity to report and transmit data, via a low-power wireless communications link, to a central monitoring station. The specific goals of the research provided in this dissertation, are to:

- Present limitations involved in the design of a low-power wireless sensing platform given current restrictions placed by government requirements, and those restrictions involved in using commercial off-the-shelf analog components.

- Develop a low-power sensor interface to convert the reaction of the hydrogen sensitive mechanism into a conditioned signal which can be accurately represented in digital form, while consuming minimal scavenged energy.

- Develop, test, and optimize a wireless communications link for communication between the self-powered wireless sensor node and a central monitoring station, while meeting the regulations imposed by governmental agencies.
• Develop a source coding scheme to minimize the required power to transmit a single message.

• Perform a full system integration and test system with a live hydrogen source

**Thesis Organization**

In chapter 2, the limitations involved in the design of a self-powered wireless hydrogen sensor are presented. Chapter 3 introduces the hydrogen sensing mechanism, and the design and testing of a differential detection circuit to interface the hydrogen sensing mechanism to a digital back-end. Chapter 4 iterates the microcontroller system, and the operation of the controller system with regard to the sensor platform, while Chapter 5 emphasizes the design, testing, and optimization of the wireless communications link. A minimum energy source coding scheme is discussed in chapter 6, and a full system implementation between energy harvesting devices, RF front-end, sensing mechanism and sensor interface, and controller system are described in Chapter 7. Finally, conclusions, current work, and future work are detailed in Chapter 8.
CHAPTER 2
WIRELESS SENSOR SYSTEM LIMITATIONS

There cannot be the successful design of a fully wireless sensor platform without the designation of limitations to the system. Once a thorough analysis of those limitations which would hinder or stop the progress of the design is completed, design orientated specifications that the system must meet can be set to assist in the realization of the system.

To be a truly self-powered wireless hydrogen sensor node, a study of the potential short-comings of the energy harvesting devices serving as the power sources for the system must be completed. By using the available power provided from energy harvesting techniques, the design of the sensor system must strive for both accuracy and ultra-low power operation. With the requirement for a low-power sensor design, also includes those problems associated with available commercial low-power and low voltage analog and digital components. A thorough investigation into the problems and solutions to the issues expected to arise in the design of the low-power analog and digital blocks of the system will be required. In addition to a power analysis, an analysis of the limitations that will make a direct impact on the design of a wireless transmitter and receiver will also be discussed.

Energy Harvesting and Reclamation

This section reviews the solar and vibration harvesting devices and circuits designed, fabricated, and tested at the University of Florida. The objective of these designs was for an end product capable of extracting energy from both photovoltaic, and
piezo-electric (PZT) energy harvesting devices. This energy would then be primed by efficient power converters for use by the hydrogen sensor, sensor interface, microcontroller, and transmitter. Interested parties should refer to the original references [3,4,5,6,7]. For the purposes of this dissertation, a system level implementation of the energy harvesting devices can be found in Figure 2-1.

Figure 2-1. System Level Implementation of Energy Harvesting Devices

**Solar Energy Harvesting**

Photovoltaic devices are a mature commercial product, and offer the attractive availability of high energy density per area. They are however, limited to real-time lighting and temperature conditions. The IXOLAR XOD17-04B Solar Cell seen in Figure 2-2 was used as the solar harvesting device, and the energy yielded from this device was conditioned by a Pulse-Resonant Power Converter designed at the University of Florida for use by the electronics of the wireless sensor system.

The Pulse Resonant Power Converter, whose functional block diagram can be seen in Figure 2-3(a) was designed to be self-powered and self-controlled for maximum power point tracking, low switching loss, and to convert an input voltage of 0.8 – 1.2V to a
steady 2V output voltage to be used by the sensor, sensor interface, microcontroller, and RF transmitter. The bare die photo of this power IC can be seen in Figure 2-3 (b).

Figure 2-2. IXOLAR XOD17-04B Solar Cell

Figure 2-3. Pulse Resonant Power Converter. (a) Functional block diagram (b) Bare die photo of Pulse Resonant Power Converter IC

Vibrational Energy Harvesting

For the harvesting of vibrational energy, piezo-electric devices are attractive as sources in that no light is required, and the collection of energy is proportional to the volume of the devices. The limiting factor however, is the magnitude and frequency of...
the vibrations. As MEMS PZT devices are currently being developed, for a proof of concept design, commercial PZT bimorph beams were purchased and used. Four PSI D220-A4-203YB Double Quick Mounted Y-Pole Bender seen in Figure 2-4 (a), were selected as the PZT devices, and were connected to a direct charging circuit (full-bridge rectifier and shunt capacitance) seen in Figure 2-4 (b) constructed at UF.

![Figure 2-4. Vibration Energy Devices. (a) Four mounted PSI D220-A4-203YB Double Quick Mounted Y-Pole Bender (b) Direct Charging Circuit](image)

The vibration energy harvesting system was tested under lab conditions and used a mechanical shaker tuned to 1 g\textsubscript{rms} @ 130 Hz (the resonant frequency of the bimorph beam), as the source of vibrations. This proof of concept design was able to deliver 250
uW of power. The efficiency and power transfer to battery versus the mechanical input power of the PZT bi-morph beams can be seen in Figure 2-5.

![Graphs showing Power to Battery vs. Mechanical Input Power and Direct Charging Efficiency vs. Mechanical Input Power](image)

**Figure 2-5. Performance of Vibration Energy Harvesting Devices (a) Power to Battery vs. Mechanical Input Power (b) Direct Charging Efficiency vs. Mechanical Input Power**

**Design Limitations of Low-Power and Low-Voltage Discrete Components**

System designers working with discrete commercial integrated circuits (IC) are moving towards single – supply, low - power, and low - voltage designs. Such designs are becoming more popular within design communities due to their reduction of cost, component count, and power consumption. An innate feature of using low-voltage single-supplies is the reduction of quiescent current, which is necessary for battery-powered systems, or in the case of self-powered operation, scavenged energy.

Another not-so-obvious reasoning for using a single-supply system is the increase of the reliability of the system. Due to the desire for a long-lifetime sensor node requiring minimal maintenance, by designing a system with discrete components operating at levels much lower than their maximum ratings, this inherently increases the lifetime of the device. The trade-offs of designing a system with low – power and low – voltage operation however, comes the adverse affects associated with the slew rate,
bandwidth, and “head-room” problems [8]. These trade-offs must be considered in the
design of a sensor interface with the necessity to obtain an accurate real-world signal.
This following section provides insight into the careful design of a sensor interface.

**Dynamic Range**

As previously mentioned, “Headroom” is the associated dynamic range of a low-
oltage system. Since the system’s usable resolution is dependent upon the signal to
oise ratio, the dynamic range of a system is perhaps one of the most significant trade-
offs in low-voltage single-supply design[8]. If one considered conventional u741 +/- 15
V dual supply operational amplifiers, due to the architecture of this device, the
input/output has a “fixed” head room of 2 V. This 2V denotes that the maximum
input/output swing of the u741 is between +/- 13 V--this swing is the dynamic range of
the operational amplifier and can be seen in Figure 2-6 from [8],

![Figure 2-6. Dynamic Range of u741 Operational Amplifier](image)

If one were to then translate this fixed headroom to a single-supply operational
amplifier operating from 0 to 3 V, the input/output maximum swing does not exist
because the headroom required is 4 V, and the maximum voltage span provided is only 3
V. The dynamic range of this device is absolutely unacceptable and will prove to be
disastrous in the final design. Thus, the single-supply design community is constantly increasing the demand for rail-to-rail amplifiers with very low headroom requirements to increase the dynamic range of the system. [8]

**Input Offset Voltage**

Due to the unbalances of the transistors and resistors within an operational amplifier, it is often the case that an input offset voltage can be found between the input terminals of an operational amplifier. Unfortunately, input offset voltages do not scale down with supply voltages, and by reducing the power supply voltage, there may be a disproportional shift in input offset voltages [9]. At heart, an operational amplifier is simply a differential amplifier which will amplify the difference between two input terminals. If one were to consider a system with high large-signal gain, along with the signal, the output would also include the amplified error of the input offset voltage.

Conventional operational amplifiers such as the u741 usually include offset-null trimming points. Although these points will compensate for input offset voltage in a per case basis, the trimming potentiometers, as well as the offset voltage drift of the operational amplifier, are plagued with their dependence to temperature. Although most operational amplifiers are factory trimmed, a system designer should also place a heavy emphasis on searching for operational amplifiers with low input offset voltage, and low offset drifts. Often more times than not, the system designer will place more importance on the selection of a device based on errors caused by drift due to temperature or time than due to absolute magnitudes of an error. This is due to microprocessor computing horsepower becoming fairly inexpensive and allowing system designers to easily linearly correlate data using software such as a crude lookup-table approach to compensate for absolute magnitudes of error. Such software techniques however, cannot compensate for
effects due to time and temperature, which makes the design of a good sensor system that much more important for a device with a desired long lifetime of operation with minimal maintenance.

**Least Significant Bit (LSB)**

With a wavering voltage span, is the relative precision required to acquire a signal. As compared to a +/- 15 V system, given an ADC of 12-bits, a 3 V span would require up to eight times of precision as compared to that of a +/- 15 V system. A comparison seen in Figure 2-7 from [9] graphically displays the increasing precision of the LSB to varying system voltage supplies.

---

**Figure 2-7. The Shrinking LSB - LSB for Multiple Voltage Supply Spans**
Limitations of a Wireless System

Once a system is successful in overcoming the requirement for low power, and has the ability to accurately obtain the real world signal of interest, if a wireless sensor node is unable to communicate with a central monitoring station, all efforts made in the design of power harvesting and of signal processing are wasted. Alike to the design of an accurate analog to digital interface, a RF wireless system suffers just as many if not more limitations. In addition to the physical limitations of a reliable wireless front end, include those limitations imposed by the United States Federal Communications Commission’s (FCC) Code of Federal Regulations (CFR) to further aggravate a system designer. Such limitations will be further discussed in this section.

Wireless Channel Estimation Techniques

Due to the effects of multi-path, and signal power-loss, the accurate modeling of the wireless channel within which a transmitted message will propagate through becomes a fairly difficult and complex parameter to model. An understanding of the wireless channel is essential in the build of a link budget, to estimate the data range of a wireless system. A link budget can be expressed as [10];

$$P_{\text{Loss}} dB = P_{dB} - P_{RdB} + G_{TB} + G_{RdB}$$

Where ($P_T$) is the power of the transmitter, ($G_T$) and ($G_R$) are the gains of the transmitter and receiver antennas respectively, ($P_R$) is the received power, or receiver sensitivity in the case of a link budget, and Path loss is the attenuation due to the propagation of a RF signal. Once we can verify the output power, gains of antennas, and receiver sensitivity, we can use an appropriate path loss model to extract a theoretical estimate of our distance. Unfortunately, an accurate path loss model is difficult to
realize, however, there are certain techniques one can use to estimate and calculate a theoretical data range for our wireless system.

**Free – space path loss**

Free-space path loss is the attenuation of the electromagnetic waves of a radio signal traveling from a transmitter to receiver. [11] Free-space loss is usually the first unknown parameter to be estimated, and alike to its moniker, free-space path loss is the attenuation based strictly on the ideal propagation condition for the spherical expansion of the RF wave front [12] through “free-space” where there is only one clear line-of-sight path between the transmitter and receiver, and does not include those losses incorporated with reflections from objects, diffraction, refraction, absorption and any other variables. This of course results in a path loss with significantly less attenuation that can be found in real-world practice. [12]

The principle behind the free-space path loss is-- signal attenuation is proportional to both the square of the distance as well as the square of the frequency while taking into account the expanding spherical wave-front of the electromagnetic signal. As a electromagnetic signal travels from a transmitter to a receiver, the signal spreads in all three-dimensions creating an expanding spherical surface. A sphere with radius R has a surface area of:

\[ S_{\text{Surface Area Sphere}} = 4 \cdot \pi \cdot (R)^2 \]

If R, the radius is doubled, then the surface area is increased by a factor of four. This relationship is known as the “Inverse Square Law”. Given these details, we can derive that the free-space path loss of a wireless channel is:
Where $\lambda = \frac{c}{f_c}$, c is the speed of light ($3 \times 10^8$), $f_c$ is the fundamental frequency of interest, and R is the distance between the receiver and transmitter.

Using this path loss estimate, H. T. Friis presented his channel model; the Friis Free Space Path Loss Model which is used to calculate the receiver signal power ($P_R$) with respect to antenna gain of the transmitter ($G_T$) and receiver ($G_R$) antennas, attenuation due to path loss, and transmitted power ($P_T$). The Friis Free Space Equation is [13]:

$$P_R(R) = \frac{P_T \cdot G_T \cdot G_R \cdot \lambda^2}{(4 \cdot \pi \cdot R)^2}$$

The Free-Scale Path loss model makes the assumptions of the most ideal of conditions for the transmission of data. Unfortunately, in the real world, these assumptions are rarely, if ever true. Thus, more sophisticated modeling is required.

**Two-ray ground reflection model**

A more common approach for the estimation of propagation signal attenuation is the plane earth propagation model which models the average attenuation associated with the distance between a stationary transmitter and receiver with direct line-of-sight, while taking into account the ground reflection path. This method is considered to be more accurate of a model and can be used to roughly estimate the attenuation for fundamental
frequencies within the ultra-high-frequency bands between 200 MHz and 5 GHz. [10] The path loss for the two-ray ground reflection model can be seen in the following equation, where \((R)\) is the distance between the transmitter and receiver, and \((h_t)\) and \((h_r)\) are the heights of the transmitter and receiver from the ground, respectively.

\[
\text{TwoRayGroundLoss} = \frac{R^4}{h_t^2 \cdot h_r^2}
\]

From this path loss equation, it is assumed that \(R\) is much greater than the heights of either antenna. From this path loss equation, we determine the power received \((P_R)\) given the transmitter power \((P_T)\), the heights \((h_t, h_r)\) and gains \((G_T, G_R)\) of the transmitter and receivers respectively.

\[
P_R(R) = \frac{P_T \cdot G_T \cdot G_R \cdot h_t^2 \cdot h_r^2}{R^4}
\]

Up to now, both models of propagation have been using the dependence of distance of the propagation path to estimate the received power for the wireless system. This technique of estimation represents the communication range as that of a sphere. We have yet to consider the effects due to random multi-paths, which is a more credible model for channel estimation.

**Shadowing model**

Both free-space and two-ray ground models neglect the reality that the power received at a given receiver is more of a random variable due to multiple path effects, or fading effects. [13] The use of a shadowing model extends the ideal sphere of the previous methods, where the predicted received power is more of a “mean” of received
signals, and creates a richer more statistical model based on the environment or “terrain” of the propagation path. The shadowing model consists mainly of two parts. One part, like previous modeling techniques, calculates a mean path loss for the system within the path of propagation, while the second part reflects variations of power at certain distances [13]. Table 2-1 and Table 2-2 show typical values for channel estimation using the shadowing model.

Table 2-1. Typical Values for Path Loss Exponent $\beta$

<table>
<thead>
<tr>
<th>Environment</th>
<th>$\beta$</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Outdoor</strong></td>
<td></td>
</tr>
<tr>
<td>Free Space</td>
<td>2</td>
</tr>
<tr>
<td>Shadowed Urban Area</td>
<td>2.7 to 5</td>
</tr>
<tr>
<td><strong>In Building</strong></td>
<td></td>
</tr>
<tr>
<td>Line-of-Sight</td>
<td>1.6 to 1.8</td>
</tr>
<tr>
<td>Obstructed</td>
<td>4 to 6</td>
</tr>
</tbody>
</table>

Table 2-2. Typical Values of Shadowing Deviation $\sigma_{dB}$

<table>
<thead>
<tr>
<th>Environment</th>
<th>$\sigma_{dB}$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Outdoor</td>
<td>4 to 12</td>
</tr>
<tr>
<td>Office, hard partition</td>
<td>7</td>
</tr>
<tr>
<td>Office, soft partition</td>
<td>9.6</td>
</tr>
<tr>
<td>Factory, line-of-sight</td>
<td>3 to 6</td>
</tr>
<tr>
<td>Factory, obstructed</td>
<td>6.8</td>
</tr>
</tbody>
</table>

In the path loss portion of the shadowing model, the mean received power ($P_R$) at distance ($R$) is found by referencing ($R$) to the power received at a distance closer in, ($R_0$). The mean received power for a given distance $R$ is computed through the following equation where $\beta$ is a typical path loss exponent found in Table 2-1.

$$\frac{P_R(R_0)}{P_R(R)} = \left(\frac{R}{R_0}\right)^\beta$$

This can also be expressed in terms of dB.
For the variations of received power at certain distances, it can be modeled as a log-normal random variable, or a Gaussian random variable $X_{dB}$ with a mean of zero, and a standard deviation of $\sigma_{dB}$. Thus, the shadowing model can be fully represented by:

$$\frac{P_R(R)}{P_R(R_0)}^{dB} = -10 \cdot \beta \cdot \log \left( \frac{R}{R_0} \right) + X_{dB}$$

The shadowing model gives a more authentic estimation of a wireless channel. However, a common radio practice for interior wireless channel estimation is to use any path loss models presented in previous sections, and to assume an addition 15 to 20 dB of fade margin in the link budget calculation. This margin should account for multi-path phenomena, shadows, reflections, system losses, and other divergences from an ideal system model. [10]

**FCC Part 15 Regulations**

Within the United States of America, there are strict impingements placed on the use of the radio spectrum between 9 kHz up to 3 THz. The responsibility of regulating the radio spectrum falls in the hands of the FCC who administrates the spectrum band for non-federal governmental use, and the National Telecommunications and Information Administration (NTIA), a unit under the Department of Commerce, who regulates the spectrum for use by the federal government. Since this design, is for non-Federal Government use, only the details of restrictions placed from the FCC will be assessed in detail. Interested parties should visit [www.fcc.gov](http://www.fcc.gov) for more information.
The Federal Communications Commission

The FCC, established by the Communications Act of 1934, is an independent government agency whose duties and responsibilities are directly decided by Congress. The FCC consists of five commissioners appointed by the president and confirmed by the senate, and are slated for five year terms except when filling an unexpired term. One commissioner is appointed chairperson by the president, and only a maximum of three commissioners may be allowed to be of the same political party [14].

Figure 2-8. FCC Organizational Chart

The FCC is maintained by a staff organized by function, and has jurisdiction over the entire 50 states, the District of Columbia, as well as all U.S. territorial possessions. Broken up into six operating Bureaus, and ten Staff Offices, the Bureaus’ are responsible for processing license applications, analyzing complaints, conducting investigations, developing and implementing regulatory programs, and taking part in hearings while the
Offices provide support services. An organizational chart of the different operating offices and bureaus of the FCC can be seen in Figure 2-8.

**FCC rules, regulations, and safety**

The rules and regulations of devices capable of emitting radio waves within the radio spectrum are located in title 47 of the code of federal regulations (CFR). Although the Office of Engineering and Technology (OET) maintains and is responsible for parts 2, 5, 15, and 18 of title 47, the official rules are published and maintained in the Federal Register. [15] The OET is an office within the FCC and is responsible for allocation of the radio spectrum for public, non government use, and provides advice on technical and policy issues governed by title 47 of the CFR. The OET is also responsible for the maintenance of the Table of Frequency Allocations. The Rules of title 47 of the CFR are divided into part 0 through 101, and organized into four sub chapters. Rules applicable to the design of a sensor are outlined as follows:

- **Part 2:** Frequency Allocations and radio Treaty Matters; general rules and regulations
- **Part 5:** Experimental Radio Service(other than broadcast)
- **Part 15:** Radio Frequency Devices
- **Part 18:** Industrial, Scientific, and Medical equipment

The regulations set by the FCC to police the spectrum and mitigate unfair interference between intentional and unintentional radiating devices (such as computer monitors), ultimately limits the radiated power from radio systems. Thus, these regulations directly affect both the RF transmitter and any gains associated with antennas. Given the premise mentioned earlier where distance is proportional to the output power of the transmitter/antenna pair, the FCC creates further difficulties in the design of a
wireless front end. In addition to those rules and restrictions set forth by the FCC, additional insight into the human-electrical interaction serves as the basis for another regulatory standard.

As required by the National Environmental Policy Act of 1969, the effect on the quality of the human environment from the emissions of transmitters regulated by the FCC needs to be evaluated by the FCC. Although there are currently no federally mandated standards for RF exposure, several non-governmental organizations such as the American National Standards Institute (ANSI), the Institute of Electrical and Electronics Engineers, Inc. (IEEE) and the National Council on Radiation Protection and Measurements (NCRP) have recommended limitations for human and RF electromagnetic field exposure. Several of these recommended limitations are [15]

- Limitations set for the span of 3 kHz to 300 GHz.
- Controlled environments (where energy levels can be accurately determined and every person on premise is aware of the presence of EM fields) allow for higher power than that of uncontrolled environments (where energy levels are unknown and where personnel on facilities may be unaware of presence of EM fields)
- Lowest E-field exposure limits occur at frequencies between 30 and 300 MHz. (1 mW/cm² (61.4 V/m) controlled, 0.2 mW/cm² (27.5 V/m) uncontrolled)
- Lowest H-field exposure limits occur at 100-300 MHz. (1 mW/cm² (0.163 A/m) controlled, 0.2 mW/cm² (0.0728 A/m) uncontrolled)
- Above 100 MHz, safety limits for E and H fields remain the same.
- Below 100 MHz, E-field radiation has lower power density limits than do H-field.
The reasoning behind setting more stringent limitation on power densities within the frequency band of 30 MHz to 300 MHz, is that the natural resonant frequencies of the human body occur between 30 to 300 MHz, and at frequencies above and below, the
human body should absorb less energy. [16] A more graphic representation of IEEE RF Safety Guidelines can be seen in Figure 2-9.

![Figure 2-9. Proposed IEEE RF Safety Guidelines](image)

The guidelines set by the FCC and the safety standards recommended by IEEE, along with the uncertainty of the wireless channel between the transmitter and receiver, show that the design of a reliable wireless front end is not a menial task; instead, it requires an intensive selection of proper components to optimize an already limited system.
CHAPTER 3
SENSOR INTERFACE DEVELOPMENT

The proper development of the hydrogen sensing mechanism and the analog to
digital interface for this device is essential to the success of the design of a robust
hydrogen sensing wireless sensor. Extensive efforts must be made to fully understand the
hydrogen sensing mechanism before a proper design can be made for the interface
between the sensing mechanism and an Analog to Digital converter (ADC). The
hydrogen sensing mechanisms used were ZnO nano-rods developed and fabricated at the
University of Florida. These nano-sensors proved themselves to be a robust solution for
the missing role of a hydrogen sensing mechanism.

The nano-rods were interfaced to an analog to digital converter with careful
considerations as previously described in Chapter 1. After careful planning, a differential
detection circuit consisting of an instrumentation amplifier topology is used to ease in the
transition from a real world signal to a digitized representation. This topology showed
strong immunity to error sources as detailed from before, and served as a successful
interface between the reactions due to hydrogen on the ZnO nano-rod and the analog-to-
digital converter of the digital realm.

The background, analysis, design, and implementation of the ZnO nano-rod and
instrumentation amplifier are detailed within this chapter.
ZnO Nano-Rods

The Zinc-Oxide (ZnO) nano-rods developed at the University of Florida between the collaboration of students and faculty of the Department of Materials Science and Engineering and the Department of Chemical Engineering, were used as the hydrogen sensing mechanisms of the system. These lightweight hydrogen sensors were designed with the goal of achieving high sensitivity, rapid response to stimuli, reversibility, and low power consumption, all within a physically small and light package.

The unique characteristics of ZnO nano-tubes and nanorods make them fundamentally appropriate candidates for the sensing of hydrogen. ZnO is a material currently used in the detection of humidity, UV light, and gas, and has shown to change resistance with respect to both temperature and hydrogen exposure. Because of its wide bandgap of 3.2eV, the ease of synthesizing nanostructures, the availability of heterostructures, and the bio-safe characteristics of this material, ZnO is a most attractive material for the specific sensing application at hand [1,2]. With ZnO nano-rods placed in an array, as a gas sensor, they are able to create a large chemically sensitive surface-to-volume ratio which is needed for high sensitivity in hydrogen sensing. ZnO nano-rods can also be produced cheaply, and are highly compatible with other microelectronic devices.

To enhance the device’s sensitivity to hydrogen, catalytic coatings or dopings of platinum (Pt) or palladium (Pd) were used to further increase the ZnO nano-rods hydrogen sensing mechanisms [1,2].
ZnO Nano-Rod Fabrication Process

The processes of growing and packaging the ZnO Nano-Rods for use as the hydrogen sensing mechanisms in the sensor system are presented. Interested parties should read [1,2].

ZnO Nano-Rod site selective growth was accomplished by nucleating the Nano-Rods on discontinuous Au islands of nominal thickness 20 Å coated with a substrate. It was previously shown that synthesis of ZnO Nano-Rods on wide assortment of substrates is a fairly uncomplicated task, increasing the ease of Nano-rod synthesis [1]. Deposition of ZnO Nano-Rods were achieved using molecular-beam epitaxy with a base pressure of 5·10⁻⁸ mbar using high purity (99.999%) Zn metal and O₂/O₃ plasma discharge as the source chemical. Growth time was approximately 2 h at 600 °C. Resultant ZnO Nano-Rods grew to a typical length of 2 – 15 um, with diameters in the range of 30-150 nm. A schematic of a multiple ZnO Nano-Rod Sensor can be seen Figure 3-1 (a), while Figure 3-1(b) shows a scanning electron micrograph of the home-grown ZnO Nano-Rods. Selected area diffraction patterns showed the ZnO Nano-Rods to be of single crystalline form.

Additional coatings of Pt or Pd were deposited by sputtering in some cases, forming Pd thin films coatings of approximate thickness of 80 Å or Pt thin film coating of approximate thickness of 10 Å. Electrodes on multiple ZnO Nano-Rods were created by using a shadow mask to pattern sputtered Al/Ti/Au electrodes contacting both ends of multiple ZnO Nano-Rods on Al₂O₃ substrates. Electrodes were separated through a spacing of approximately 400 um. Au wires were then bonded to the contact pads of the package for transportation and current-voltage measurements. The packaged hydrogen sensing mechanism can be seen in Figure 3-2.
Figure 3-1. ZnO Nano-Rods. (a) Schematic of Multiple ZnO Nano-Rods (b) Scanning Electron Micrograph of ZnO Nano-Rods.

Figure 3-2. Close-Up of Packaged ZnO Nano-Rod Sensor

**Performance of ZnO Nano-Rods**

The fabricated and packaged ZnO Nano-Rods were tested under laboratory conditions where their reactions to hydrogen can be monitored under a controlled hydrogen environment. This controlled environment is within a hydrogen chamber.
located at the University of Florida, and allows for an accurate and quick assessment on the performance of the fabricated ZnO Nano-Rod sensors. Figure 3-3 shows the schematic of the hydrogen chamber, and a more detailed description on the operation of the chamber will be included in a later chapter.

![Schematic of Hydrogen Chamber](image)

**Figure 3-3. Simple Schematic of Hydrogen Chamber Used for ZnO Nano-Rod Testing**

![Schematic of Biasing](image)

**Figure 3-4. Schematic of Biasing for ZnO Nano-Rod Hydrogen Sensitivity Testing**

For the purpose of this section, the only information pertinent to understanding the performance of the ZnO Nano-Rods are the chamber’s functions of injecting controllable amounts of N₂ and H₂ to create different concentrations of hydrogen on the orders of parts per million (PPM) within the chamber. All measurements were performed using a
HP 4156B Semiconductor Parameter Analyzer with the packaged ZnO Nano-Rods biased with a 0.5 V supply on one terminal, and ground on the other as seen in the schematic of Figure 3-4.

**Uncoated ZnO nano-rods**

The relative transient response of uncoated ZnO Nano-Rods as the gas ambient is switched from pure N₂ to concentrations of hydrogen in nitrogen ranging from 10 to 500 PPM can be seen in Figure 3-5.

![Figure 3-5. Uncoated ZnO Nano-Rod - Relative Resistance Change for Various Hydrogen Concentrations](image)

This shows a relative resistance change of approximately 0.70% for 500 PPM of H₂ in N₂ after 10 minutes of exposure, with inconsistent results for lower concentrations. The gas-sensing mechanisms include the desorption of adsorbed surface hydrogen and grain boundaries in poly-ZnO, exchange of charges between adsorbed gas species and the ZnO surface, leading to changes in depletion depth and changes in surface or grain boundary conduction by gas adsorption/desorption [2]. Thus, it is shown the ZnO Nano-Rods are a suitable candidate for the sensing of hydrogen. The performance of un-
coated ZnO Nano-Rods serves as a starting point for comparison to the increased hydrogen sensitivities of Pt or Pd coated ZnO Nano-Rods.

**Pd coated ZnO nano-rods**

Pd coated ZnO Nano-Rods have a relative transient response shown in Figure 3-6. Once again, this shows the reaction of the Pd coated ZnO Nano-Rods as the gas ambient is switched from N₂ to increasing concentrations of H₂ in N₂ starting from 10 PPM to 500 PPM.

![Figure 3-6. Pd-coated ZnO Nano-Rod - Relative Resistance Change for Various Hydrogen Concentrations](image)

By comparison, there is an approximate five-fold increase in response to hydrogen as compared to that of uncoated devices. This shows that the addition of Pd appears to be effective in catalytic dissociation of the H₂ to atomic hydrogen. The relative response of Pd-coated Nano-Rods can be seen as a function of H₂ concentration in N₂, where Pd-coated ZnO Nano-Rods were capable of detecting hydrogen down to less than 10 PPM with a relative response of greater than 2.6% at 10 PPM. They have also shown to have greater than 4.2% relative resistance change to 500 PPM after 10 minutes of exposure.
Additionally, as seen in Figure 3-7, ZnO Nano-Rods show no response to the presence of O2 at room temperature.

![Graph](image)

**Figure 3-7. Pd-coated ZnO Nano-Rod. Absolute Resistance Change for Various Hydrogen Concentrations**

The reversible chemi-sorption of reactive gases at the surface of metal oxides such as ZnO can provide a large and reversible variation in conductance of the material. Thus, the recovery of the initial resistance upon removal from the hydrogen ambient is quick (less than 20 seconds). The sputtered Pd-ZnO Nano-Rod also shows to have an increased effective conductivity due to the presence of metal. Palladium has proved to be a catalyst which can enhance the sensitivity of the hydrogen sensing mechanisms.

**Pt coated ZnO nano-rods**

Similar fabrication and testing procedures were done with Pt Coated ZnO Nano-Rods, except for the replacement of Pd with Pt as the sputtered metal. Figure 3-8 shows the relative transient response of the Pt-ZnO Nano-Rods as the ambient gas is switched from N2 to various concentrations of H2 in N2 ranging from 10 to 500 PPM.
Figure 3-8. Pt-coated ZnO Nano-Rod - Relative Resistance Change for Various Hydrogen Concentrations

Similar transient characteristics can be seen between Pd-ZnO Nano-Rods and Pt-ZnO Nano-Rods such as rapid recovery of initial resistance, with a 90% recovery within 20 seconds upon removal of the hydrogen from the ambient with the replacement of O₂ or air, and no response to the presence of ambient O₂ or N₂ at room temperatures. However, the change from initial resistance after 10 minutes of exposure to H₂ has an almost two fold increase as compared to Pd coated ZnO Nano-Rods. This shows that as a hydrogen sensing mechanism, Pt – coated ZnO Nano-Rods have increased sensitivity to hydrogen, and would prove as a better sensing device than both Pd and uncoated ZnO Nano-Rods.

Detection Interface for Hydrogen Sensitive Devices

The challenge in designing the interface between a sensor and the Analog-to-Digital (A/D) converter of a system is found in the necessity to obtain an accurate real world signal with the limitations of low power and reduced voltage swings. These limitations were previously discussed in chapter 1, and will be reiterated as the design of
the differential detection circuit is described. Furthermore, for accomplishing a long-
lifetime mode of operation, this sensor system is faced with the limitations of analog
development, as well as the requirement for low-power operation.

One of the most important objectives of the sensor interface design is for the
compatibility between the sensing mechanism to the detection circuitry and ADC.
Thankfully, ZnO Nano-Rods were chosen due to their high compatibility with
microelectronic devices. The design will focus on meeting the demands of the ZnO
Nano-Rods, power requirements, and meeting those requirements set by the resolution of
the ADC input of the microcontroller. Additionally, the differential detection interface
should only detect reactions due to hydrogen, and void all changes caused by other
variables such as temperature. Given that the ZnO Nano-Rod’s initial response to any
exposure of hydrogen is distinct and immediate, this intrinsic characteristic will serve as
an ally for the successful detection of hydrogen within the sensor system.

This section will detail in depth the design of a differential detection sensor
interface—from theoretical concept, to considerations of component selection, and finally
an evaluation on the performance of the fabricated and packaged differential detection
circuit.

**Wheatstone Resistive Bridge**

The intrinsic characteristic of ZnO Nano-Rods that makes them suitable as
hydrogen sensing mechanisms is their change in resistance with respect to how much and
how long the device has been exposed to hydrogen. To accurately detect the presence of
hydrogen in the ambient, since the nominal resistance of the ZnO Nano-Rods is a
function of hydrogen concentrations, the precise measurement of the change in resistance
of the ZnO Nano-Rods can be used to correlate the change in resistance to the concentration of hydrogen in the ambient.

Bridge circuits, although old and primitive, continue to commonly serve as the best solution for the measurement of resistance, capacitance, and inductance. The resistive bridge instrument used for the measurement of an unknown electrical resistance is known as a Wheatstone bridge. The Wheatstone resistive bridge as seen in Figure 3-9, invented by Samuel Hunter Christie in 1833 and popularized and improved by Sir Charles Wheatstone in 1843 [17], illustrates the concept of using a difference measurement for obtaining the value of an unknown electrical component.

![Wheatstone Resistive Bridge](image)

Figure 3-9. Wheatstone Resistive Bridge

A difference measurement is taken across $V_g$ seen in Figure 3-9, where $V_g$ is dependent upon the nominal resistance values of $R_1$, $R_2$, $R_3$ and $R_4$. With $R_1$ and $R_3$ resistance values set and non-changing, the voltage divisor combinations of $R_3/R_4$ and $R_1/R_2$ respectively, form the voltage $V_g$ across $V_2$ and $V_1$. $V_g$ can be equated as:

$$V_g = V_2 - V_1$$

So,
\[
V_g = \left( \frac{R_4}{R_3 + R_4} - \frac{R_2}{R_1 + R_2} \right) \cdot V_s \quad \text{Where } V_s \text{ is the supply voltage}
\]

Thus, it can be seen that if \( R_4 \) and \( R_2 \) are of equal values, the voltage across \( V_g \) should be 0 V, and any variation of \( R_2 \) with respect to \( R_4 \) will create a voltage drop across \( V_g \). By measuring this voltage drop, assuming the reference resistor \( R_4 \) has remained constant, we can decipher the change in resistance of \( R_2 \).

Because the ZnO Nano-Rods exhibit a change to both temperature and hydrogen concentrations, the relationship between \( R_4 \) and \( R_2 \) serves to be advantageous in providing a way for making the system impervious to the variable of temperature. By using a passivated ZnO Nano-Rod device encased in glass as the reference resistor \( R_4 \) and an un-covered and exposed ZnO Nano-Rod device as \( R_2 \), additional changes in resistance for the exposed ZnO Nano-Rod due to effects of temperature will be compensated by \( R_4 \), and so the only changes seen at \( V_g \) are those caused by the exposed ZnO Nano-Rod reacting to the presence of hydrogen gas in the ambient.

For the mode of low power operation, by setting resistors \( R_1 \) and \( R_3 \) as bias resistors much larger magnitude in comparison to that of \( R_4 \) and \( R_2 \), these resistors can be used to limit the power consumed by the bridge circuit. Using the relationship of Voltage = Current \( \cdot \) Resistance, and Power = Voltage \( \cdot \) Current, using very large resistance values for \( R_1 \) and \( R_3 \) should limit the current passing through the two legs of the bridge circuit, and since power is directly proportional to current, limit the overall power consumption of the Wheatstone Bridge.

This limiting of current however, lowers the dynamic voltage swings of \( V_2 \) and \( V_1 \), which in turn makes the output voltage \( V_g \) very small in comparison to the supply.
To successfully detect the change in resistance, the voltage $V_g$ must be amplified to meet the resolution requirements of the ADC. This requires the addition of an amplifier stage to buffer and amplify the signal before processing by the ADC of the sensor system. Since the Wheatstone resistive bridge uses the concept of differential measurement, the amplification stage must adhere to the design of a differential detection interface.

**Differential Detection Interface**

The design of a differential detection interface must meet the requirements of several issues noted in the previous section. Firstly, the design of the interface must remain steadfast to the original concept of using differential measurements to determine the detection of hydrogen, and lastly, the interface must have high large-signal gain to amplify the output of the current-limited Wheatstone bridge.

**Difference amplifier**

To remain a differential measurement instrument, a difference amplifier as seen in Figure 3-10 from [18] is employed. The architecture of an operational amplifier with no feedback by itself is already a differential amplifier with an uncontrollable gain. Typical inverting or non-inverting amplifiers are restrictive in that there is the practical loss of one of the two inputs; however, by using a difference amplifier topology, a nominal gain can be set, with both inputs remaining intact. By keeping both inputs intact, the internal differential architecture of an operation amplifier can be exploited to be used as the differential measurement inputs.

In a difference amplifier configuration, the analysis of the circuit is essentially the same as that of an inverting amplifier. The only difference is the non-inverting input of
the operational amplifier being set to a voltage that is a fraction of $V_2$ rather than ground, where $V_2$ is dependent upon the resistance values of $R_3$ and $R_2$.

![Difference Amplifier](Image)

Figure 3-10. Difference Amplifier

The output of the operational amplifier, $V_{OUT}$, can be found using the following equation:

$$V_{OUT} = \frac{R_3}{R_2} \cdot (V_2 - V_1)$$

Unfortunately, as compared to a non-inverting amplifier configuration, the input to the difference amplifier has fairly low impedance. Because $V_2$ and $V_1$ of the Wheatstone resistive bridge is to be connected to the differencing inputs of the interface, if the inputs of the interface are of low input impedance, this will pose a serious problem to the accuracy of Wheatstone bridge due to the presence of a separate path for current to cause an inaccurate representation of output voltage at the output nodes of the Wheatstone
bridge. Fortunately there is a simple solution. All that is needed is a non-inverting buffer, or a voltage follower to be added as seen in Figure 3-11.

![Difference Amplifier with Non-Inverting Buffer to Differential Inputs](image)

Figure 3-11. Difference Amplifier with Non-Inverting Buffer to Differential Inputs

An improvement to a simple voltage follower can be seen in Figure 3-12. This topology is known as an instrumentation amplifier.

![Instrumentation Amplifier](image)

Figure 3-12. Instrumentation Amplifier
**Instrumentation amplifier**

Alike to the addition of a voltage follower as seen in Figure 3-11, Figure 3-12 shows an improved version of Figure 3-11 through the addition of three resistors connecting the two input buffer circuits to the difference amplifier. Using this topology, there can be the establishment of a gain stage before the large-signal gain of the difference amplifier, while maintaining a high input impedance to isolate the Wheatstone resistive bridge from the feedback resistor network of the difference amplifier.

With the introduction of the three resistors connecting the high impedance input buffers to the difference amplifier, additional gain can be provided before the gain of the difference amplifier. This gain is achieved by creating a voltage drop across $R_g$ from the isolated input voltages at $V_1$ and $V_2$. This voltage drop induces a current through $R_g$, and since the feedback loops of the input buffers draws little to no current, the same amount of current is drawn through the two resistors labeled $R_1$. This produces a voltage drop across nodes 3 and 4 equal to:

$$V_{3-4} = (V_2 - V_1) \cdot \left(1 + \frac{2 \cdot R_1}{R_g}\right)$$

With the combination of the input buffer stages, three connecting resistors, and difference amplifier, the total large-signal gain of the instrumentation amplifier is found to be equal to:

$$V_{OUT} = (V_2 - V_1) \cdot \left(1 + \frac{2 \cdot R_1}{R_g}\right) \cdot \frac{R_3}{R_2}$$

The schematic for the full differential detection circuit can be seen in Figure 3-13. Here we see the Wheatstone resistive bridge serving as the input to the instrumentation
amplifier, with the inclusion of the exposed ZnO Nano-Rod, passivated ZnO Nano-Rod, Wheatstone current limiting bias resistors, and the feedback network of resistors, $R_g$, $R_1$, $R_2$, and $R_3$ of the instrumentation amplifier. At this point, the design of the differential detection interface is completed, and the selection of the proper components for the fabrication of the differential detection interface can start.

![Full Schematic for Differential Detection Circuit](image)

Figure 3-13. Full Schematic for Differential Detection Circuit

**Realization and Testing of Differential Detection Circuit**

Thorough considerations from the limitations of commercial discrete components noted in Chapter 2 remained in mind for the selection of components. Instrumentation amplifiers intrinsically include very low DC offset, low drift, low noise, high open-loop gain, high common-mode rejection, and high input impedances, making them highly accurate, and stable circuits for long and short term use [19]. From the discussions of
chapter 2, heavy emphasis should be put on the selection of an operational amplifier with requirements of low supply voltage, low supply current, low DC input offset, low drift, high Common-mode rejection, and rail to rail voltage swing capabilities.

**Selection of operational amplifier**

A thorough internet search of available low power operational amplifiers was conducted, and the results of this search can be summarized in Table 3-1.

<table>
<thead>
<tr>
<th>NAME</th>
<th>$V_{\text{supply (MIN)}}$</th>
<th>$I_{\text{supply (uA)}}$</th>
<th>$V_{\text{OS (MAX) (uV)}}$</th>
<th>$V_{\text{OS (TYP) (uV)}}$</th>
<th>Rail to Rail</th>
</tr>
</thead>
<tbody>
<tr>
<td>MAX4289</td>
<td>1</td>
<td>9</td>
<td>2000</td>
<td>200</td>
<td>Output</td>
</tr>
<tr>
<td>MAX406</td>
<td>2.5</td>
<td>1.2</td>
<td>500</td>
<td>250</td>
<td>Output</td>
</tr>
<tr>
<td>MAX478</td>
<td>2</td>
<td>17</td>
<td>70</td>
<td>30</td>
<td>no</td>
</tr>
<tr>
<td>INA321</td>
<td>2.7</td>
<td>40</td>
<td>500</td>
<td>200</td>
<td>Output</td>
</tr>
<tr>
<td>OPA336</td>
<td>2.3</td>
<td>20</td>
<td>125</td>
<td>60</td>
<td>Output</td>
</tr>
<tr>
<td>TLV2401</td>
<td>2.5</td>
<td>0.88</td>
<td>1200</td>
<td>390</td>
<td>Input Output</td>
</tr>
</tbody>
</table>

From Table 3-1, Maxim IC’s MAX4289 was chosen as the operational amplifier of choice. At the time, the MAX4289 showed to have the best input offset as compared to power consumption. The most attractive part of the MAX4289, is the minimum voltage range it was rated for. Out of all the other operational amplifiers studied, none were able to operate from a single-supply voltage as low as 1 V, while draining only 9uA and maintaining an input offset voltage of 200 uV.

**Simulation of differential detection circuit**

The assembly of all components involved both the components required for the instrumentation amplifier, as well as the resistive bridge. For use as the exposed ZnO Nano-Rods, new Pt-Coated ZnO Nano-Rods were grown and packaged for use. The nominal resistance change to the injection of 500 PPM of H$_2$ in N$_2$ into the ambient can be seen in Figure 3-14.
Figure 3-14 New Pt-coated ZnO Nano-Rod Grown and Packaged for Differential Detection Circuit

Because current testing only involves testing at room temperature, the requirement for a passivated ZnO Nano-Rod to compensate for temperature deviations is not required. This simplifies the design in allowing the passivated ZnO Nano-Rod to be replaced by a resistor. In order to tune the gains of the amplifier, the system was simulated using Agilent ADS, with the operational amplifier’s parameters entered with the values seen in Figure 3-15, which are the parameters found on the datasheet for the MAX4289.

After gain tuning, the Differential Detection circuit seen in Figure 3-13, was simulated with the component values found in Table 3-2. In the simulation, the exposed ZnO Nano-Rod’s resistance value was swept from 1565 ohms to 1461 ohms, which correlates to the resistance span of the Pt-ZnO Nano-Rods with 500PPM of H₂ in the ambient seen in Figure 3-14. The simulation set up of the Differential Detection circuit including parameters of the MAX4289 can be seen in Figure 3-15. The output of the
The instrumentation amplifier was plotted against the swept resistance span, and can be seen in Figure 3-16.

### Table 3-2. Differential Detection Circuit Component Values

<table>
<thead>
<tr>
<th>PART</th>
<th>VALUE</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_g$</td>
<td>470 kohms</td>
</tr>
<tr>
<td>$R_1$</td>
<td>2 Mohms</td>
</tr>
<tr>
<td>$R_2$</td>
<td>39 kohms</td>
</tr>
<tr>
<td>$R_3$</td>
<td>2 Mohms</td>
</tr>
<tr>
<td>$R_{bias}$</td>
<td>270 kohms</td>
</tr>
<tr>
<td>Passivated ZnO</td>
<td>1565 ohms</td>
</tr>
<tr>
<td>Operational Amplifier</td>
<td>MAX4289</td>
</tr>
</tbody>
</table>

---

Figure 3-15. Agilent ADS 2003 Simulation Setup for Differential Detection Circuit
Figure 3-16. Agilent ADS 2003 Simulation: Output Voltage to Swept ZnO Nano-Rod Resistance

Given the supply voltage of 2 V, the designed output power of the power IC power converters designed for the sensor system, and assuming a 10-bit ADC, the A/D has a resolution of about 2mV, with 1024 voltage levels between 0 and 2V, and The output of the interface must be able to meet this requirement and provide at least a 2mV per ohm (output voltage to ZnO resistance change) output. Fortunately, from the graph of Figure 3-16, we can see that this requirement of resolution can be achieved, with an approximate 4mV/ohm falling slope. The simulation also reveals the effects of the input offset voltages of the operational amplifiers. From Figure 3-16, it can be seen that for the case where both the exposed ZnO Nano-Rod and passivated ZnO Nano-Rod is matched, there exists an approximate 200 uV DC offset. This is unfortunate, but can be compensated
through software coding. The next step is for the design of a printed circuit board (PCB) for use of fabricating and testing of the differential detection system.

Fabrication of differential detection circuit

The next step was to layout the schematic of the instrumentation amplifier onto PCB, and assemble the device for testing. The designed PCB layout can be seen in Figure 3-17. Same component list as seen in Table 3-2 were used in the assembly, and the final assembled device with a packaged ZnO Nano-Rod device can be seen in Figure 3-18. Initial measurements for the fabricated sensor interface can be seen in Table 3-3.

Figure 3-17. Protel PCB Top and Bottom Layout

Figure 3-18. Fabricated and Assembled Differential Detection Interface Board with Packaged ZnO Nano-Rod Sensor
Table 3-3. Initial Measurements of Differential Detection Interface

<table>
<thead>
<tr>
<th>Supply Voltage</th>
<th>ZnO Resistance</th>
<th>H₂ PPM</th>
<th>Supply Current</th>
<th>Power</th>
<th>Sensor Output</th>
</tr>
</thead>
<tbody>
<tr>
<td>2V</td>
<td>1565</td>
<td>0</td>
<td>42uA</td>
<td>84uW</td>
<td>30uV</td>
</tr>
<tr>
<td>2V</td>
<td>1522</td>
<td>10</td>
<td>44.3uA</td>
<td>88.4uW</td>
<td>152mV</td>
</tr>
<tr>
<td>2V</td>
<td>1500</td>
<td>500</td>
<td>44.3uA</td>
<td>88.6uW</td>
<td>210mV</td>
</tr>
</tbody>
</table>

Given the A/D resolution of 2mV, from Table 3-3, we can see that the sensor and sensor interface is capable of meeting the resolution requirements of the system while consuming minimal power, and remaining sufficiently impervious to the limiting effects described in chapter 2.

A study was then conducted to measure the linearity of the whole assembled differential detection device. For this experiment, the exposed ZnO Nano-Rod was replaced by discrete chip resistors ranging from 1460 to 1562, while the passivated ZnO Nano-Rod was replaced by a chip resistor with nominal resistance of 1562. This procedure is similar to the simulation detailed earlier in the chapter. 1562 ohms was used instead of 1565 due to the available value of resistors. A plot of the performance for the linearity of the instrumentation amplifier with respect to the swept value of the exposed ZnO Nano-rod bridge resistance is seen in Figure 3-19. From these results, it is shown that the performance of the differential detection circuit is actually better than simulated results.

The detection circuit shows to have good large-signal linearity with an approximate -4mV/ohm slope, which matches the slope of the simulated system. In addition, for the case when both exposed ZnO Nano-Rods and passivated ZnO Nano-Rods are matched, the DC offset is only a mere 30 uV. This shows that the instrumentation amplifier is
indeed impervious to the effects of input offset voltage, and the architecture has managed to compensate for the absolute input offsets of the MAX4289 operational amplifiers. Software compensation of input offset voltages mentioned earlier is unnecessary for the differential detection circuit. The next process would be to integrate the sensor interface to that of the digital signal processing portions of the wireless hydrogen sensor node.

Figure 3-19. Measured Output Voltage vs. ZnO Nano-Rod Resistance Sweep for Fabricated Differential Detection Circuit
CHAPTER 4
MICROCONTROLLER DEVELOPMENT

Microcontroller Selection

The proper selection of a microcontroller was essential to the success of the design. The system required the microcontroller to include an onboard ADC with enough resolution to track the changes of the sensors and be capable of conditioning and processing the data received from the sensor interface. Enough onboard memory is needed to retain both the runtime code as well as store the data sampled by the ADC. There is also the requirement for our microcontroller to have the ability to encode and send this data to the transmitter via a serial output. The system would be optimized if the microcontroller included a serial output port capable of sourcing enough power to drive and power the transmitter. There also exists the requirement for the microcontroller to be easily reprogrammable and consume a minimal amount of power. Interested parties should reference to [20].

Because the initial goal is for a truly self-powered system, an emphasis must be placed on the assumption that the bulk of power will be consumed in the active states of the microcontroller and transmitter. However, because the time required for the ZnO Nano-rods to saturate is on the order of minutes, the system can operate with a very low duty cycle, and so, idle or standby current also becomes a significant factor in the decision of which microcontroller to select.

Eventually, Texas Instrument’s MSP430F1232IPW was selected. This specific microcontroller was chosen because of its many features, large searchable
knowledgebase, and the quality of assistance and samples given by TI. Table 4-1 highlights all the pertinent features of our microcontroller selection.

Table 4-1. Features of Texas Instruments’ MSP430F1232IPW

<table>
<thead>
<tr>
<th>Type of Program Memory</th>
<th>Flash</th>
</tr>
</thead>
<tbody>
<tr>
<td>Program Memory</td>
<td>8 kB</td>
</tr>
<tr>
<td>RAM</td>
<td>256 Bytes</td>
</tr>
<tr>
<td>I/O Pins</td>
<td>22 pins</td>
</tr>
<tr>
<td>ADC</td>
<td>10-bit SAR</td>
</tr>
<tr>
<td>Interface</td>
<td>1 Hardware SPI or UART, Timer UART</td>
</tr>
<tr>
<td>Supply Voltage Range</td>
<td>1.8 V – 3.6 V</td>
</tr>
<tr>
<td>Active Mode</td>
<td>200uA @ 1 MHz, 2.2 V supply</td>
</tr>
<tr>
<td>Standby Mode</td>
<td>0.7 uA</td>
</tr>
<tr>
<td># of Power Saving Modes</td>
<td>5</td>
</tr>
</tbody>
</table>

**Modes of Operation**

Currently the microcontroller is programmed to run as a state machine, and has two different reprogrammable modes of operation. In each mode of operation, the microcontroller operates within the following states: initialize, collect data, transmit data, and sleep. The first mode of operation is for the level monitoring of hydrogen. This mode runs through each state until a discernable threshold of hydrogen is detected. This threshold is set so that although hydrogen is present, the level of hydrogen is not enough to pose any serious danger. Once hydrogen is detected, the microcontroller forces the RF front-end to transmit an emergency pulse to the central monitoring station, and returns back to an idle mode.

The second mode of operation is of data transmission. In this mode, the microcontroller collects data from the sensor interface, and queues this data to the RF front-end to be transmitted to the central monitoring station. This mode is for a constant tracking of hydrogen levels, while the level monitoring mode is to alert the end user that
hydrogen has indeed been detected. The state flow diagram for the Level Monitoring Mode and Data Transmission Mode can be seen in Figure 4-1.

Figure 4-1. Microprocessor State Flow Diagrams. (a) Level Monitoring State Flow Diagram. (b) Data Transmission State Flow Diagram

**Power Requirements of Microcontroller**

To analyze the power consumption of the microcontroller during various stages of operation, a 383 ohm resistor was connected in series between the power supply, and microcontroller. Differential probes were used to measure the voltage across the 383 ohm resistor. Calculation for the average power consumption is as follows:

- Total Area (TA) under the measured curve is calculated (units of V·sec)
- Peak power is calculated as:
  \[ P_{peakPower} = \frac{V_{MAX}}{383} \cdot V_{SUPPLY} \]
  Where VMAX is the maximum point of the measured curve
- Average power is calculated as
  \[ AvgPower = \left( \frac{T_{TotalArea}}{\tau_{duration} \cdot 383} \right) \cdot V_{SUPPLY} \]
  Where \( \tau_{duration} \) is the duration of the measured curve
A power analysis was done by David Johnson at Cisco to examine the power requirements of the controller system. From this analysis, it is observed that the power consumption for the microcontroller to remain idle, output data via serial power (either high or low bit), and to scan the ADC’s input, is a constant 2.5 uW. The most power consumed by the microcontroller at any time, is in the microcontroller’s initialization state, which occurs only once during initial power up of the microcontroller. The initialization time for the microcontroller is only for 12.5ms, where average power consumption is 3.07mW with a peak power of 7.3mW (as seen in Figure 4-2).

Figure 4-2. Initialization Power Required for MSP430F1232IPW
CHAPTER 5
LOW-POWER WIRELESS COMMUNICATION LINK

In the design of wireless sensors, the most power consuming component is often found in the wireless front-end. To make matters worse, components within a RF transceiver/transmitter/receivers such as power amplifiers or oscillators, at best, only have efficiencies slightly better than 50% [10]. This means at best performance, to transmit 100mW of power, the device will require 200mW of power. However, the effects of low efficiency can be mitigated through several techniques. Because typical sensor nodes remain in idle states much longer than in active states, the sensor nodes themselves are of very low duty-cycle. By using low duty-cycle and low data rates, components within the transmitter for the wireless sensor can be turned off when no data is present to be transmitted, and the entire transceiver/transmitter/receiver can be placed into a low-power sleep mode. The considerations for the design and implementation of the wireless communications link are detailed in this chapter.

Selection of a Modulation Technique

From the previously mentioned reality of both the nature of the sensor system, as well as the limitations of the components within a RF transceiver, transmitter, and receiver, there exists a modulation technique which can take advantage of all the mentioned limitations and requirements of the system. Firstly, the system must be able to obtain power from scavenged sources, and allow for long life-time operation with minimal maintenance, which reduces the complexity of the wireless front end from that of a transceiver to that of a lone transmitter. Additionally, because the system itself is of
low duty-cycle and lower data rate, and because of the low efficiencies of RF discrete components, a modulation scheme is required which can exploit the concept of consuming power only when transmitting data, and requires a simple transmitter architecture, with few discrete RF parts counts. By using a modulation scheme of low complexity, the depth of modulation can be realized through a transmitter architecture with fewer discrete power consuming components. Because these components intrinsically show poor efficiency, a lower component count will further reduce the power consumption for the RF front end.

The simplest modulation scheme available is that of a “carrier present, carrier absent” technique, also known as “On-Off Keying” (OOK). What made OOK an appealing modulation scheme was the intrinsic premise that an OOK transmitter would only be “on” and consuming power when the RF front end was transmitting a “high” or a “1”, and that the transmitter has the advantage of going into an “idle” state so that little to no power would be consumed on the transmission of a “low” or “0”. A comparison of the transmitter architecture and performance for $\pi/4$ DQPSK as a reference to OOK [21]
can be seen in Figure 5-1. It is shown that the $\pi/4$ DQPSK transmitter architecture is significantly more complex than that of an OOK transmitter. This complexity is a trade off between performance and complexity (correlating to power consumption) for the selection of a modulation scheme.

The disadvantage of OOK modulation however, is found in the error caused by the presence of unwanted or undesirable signals. Typically, OOK is an unappealing modulation scheme for networks of heavy traffic, but because a sensor system rarely will transmit with a duty cycle of more than 25% of the time, OOK modulation is suitable for use as the modulation scheme of the wireless sensor node. OOK differs from Amplitude Shift Keying (ASK) in that there is no carrier present during the transmission of a zero. This allows for additional power reduction on the transmitter side, however, it allows OOK to be more susceptible to an interfering signal making detection by the receiver more difficult as compared to ASK. This is a performance trade-off between ASK and OOK.

![Figure 5-2](image)

Figure 5-2. Signal Constellations. (a) OOK(b) ASK(c) and as a reference, FSK

The signal constellations for ASK and OOK can be seen in Figure 5-2 from [22], with the signal constellation for FSK as a reference for comparison. For the design of the
wireless sensor node, because power requirements out-weigh any other requirements, OOK is selected as the choice of modulation due to the characteristics of OOK which make it a highly power conserving form of modulation.

**Selection of Operating Frequency**

Once a modulation scheme is selected, the selection of which frequency band to operate within is needed. The considerations for the selection of which frequency band are heavily dependent on the limitations stated in Chapter 2 considering impairments of the wireless channel, FCC regulations, and the modulation scheme derived in the previous section. Because OOK is highly sensitive to interference signals, the selection of the operating frequency is very important for the success of the wireless system.

![Path Loss Attenuation(dB) with Respect to Carrier Frequency](image)

**Figure 5-3. Path Loss Attenuation(dB) with Respect to Carrier Frequency.**

As seen in Figure 5-3, regarding free space path loss, it can be seen that as operating frequency is increased, for a given distance, the attenuation of the propagation path increases. Thus, with a lower frequency, longer transmission distances can be
achieved with lower power. Since lower power is required to attain a particular distance, less output power is needed, which in turn reduces the power consumption requirements of the transmitter.

This however comes at a cost. From the discussion of Chapter 2, it was also mentioned that the most stringent output powers were placed on the frequency band of 30 to 300 MHz. To reiterate, this was done to limit RF power absorption by the human body, because the human body is naturally resonant between those frequencies. Thus, a trade-off exists in that although limited in output power, using a lower frequency increases the transmission distance, but decreases the required output power to transmit a certain distance.

Another factor to consider is the traffic involved in each frequency band. Common operating frequencies such as the 902-928 MHz band, are constantly being used with continuous transmissions of voice, data, video, and offer high level interference from microwave ovens and spread spectrum devices. Because an OOK modulation scheme is highly susceptible to interference, the overcrowding of frequency bands within 900 MHz, 2.4 GHz, and 5 GHz can prove to be treacherous to a low complexity modulation scheme such as OOK. Other lower frequency bands such as the 260 to 470 MHz bands are much more open and less crowded. Typical frequencies within these bands such as 315 MHz, 418 MHz, and 433.92 MHz only compete with garage door/keyless entry systems, or interference from amateur radio users [23].

Due to these factors, and the availability of commercial products, the operating frequency of 300 MHz to 315 MHz within the 260-470 MHz FCC frequency band was selected. This operating frequency offers a fairly interference free band, but is shown to
have some unusual restrictions specific to the frequency band of 260 to 470 MHz set by the FCC under part 15.231 which will be detailed later on in the chapter.

**Selection and Performance of a RF Transmitter**

To simplify the design of the wireless sensor node, commercial RF transmitters were chosen to be used as the RF front end of the sensor system. After selection of modulation scheme and operating frequency, an internet search was performed to find available transmitter/receiver pair packages capable of OOK within the 315 MHz operating frequency band. A brief listing detailing available transmitters with their corresponding performance specifications can be seen in Table 5-1.

<table>
<thead>
<tr>
<th>MODEL</th>
<th>$V_{\text{MIN}}$</th>
<th>$V_{\text{MAX}}$</th>
<th>$I_{\text{SUPPLYMAX}}$</th>
<th>$I_{\text{SUPPLYMIN}}$</th>
<th>Output Power(min)</th>
<th>Output Power(max)</th>
<th>Freq</th>
<th>TYPE</th>
</tr>
</thead>
<tbody>
<tr>
<td>LINX TXM-315-LR</td>
<td>2.1</td>
<td>3.6</td>
<td>5.1mA</td>
<td>1.8mA</td>
<td>-4 dBm</td>
<td>8 dBm</td>
<td>315 MHz</td>
<td>SAW</td>
</tr>
<tr>
<td>MAX1472</td>
<td>2.1</td>
<td>3.6</td>
<td>9.1mA</td>
<td>1.5mA</td>
<td>3.3 dBm</td>
<td>6 dBm</td>
<td>315 MHz</td>
<td>Crystal</td>
</tr>
<tr>
<td>MAX1479</td>
<td>2.1</td>
<td>3.6</td>
<td>6.7mA</td>
<td>2.9mA</td>
<td>2.7 dBm</td>
<td>5.3 dBm</td>
<td>315 MHz</td>
<td>Crystal</td>
</tr>
<tr>
<td>Ming TX-99</td>
<td>?</td>
<td>5</td>
<td>1.6mA</td>
<td>?</td>
<td>?</td>
<td>?</td>
<td>300 MHz</td>
<td>LC</td>
</tr>
<tr>
<td>Atmel U2741B</td>
<td>2</td>
<td>5.5</td>
<td>12.5mA</td>
<td>?</td>
<td>1.5 dBm</td>
<td>5 dBm</td>
<td>315 MHz</td>
<td>Crystal</td>
</tr>
</tbody>
</table>

The list found in Table 5-1 was whittled down to two choices for a RF front-end—the TX-99 manufactured by Rayming Corporations, and the TXM-315-LR manufactured by Linx Technologies. These specific transmitters were selected due to their low power consumption, low component count, and low complexity for the ease of rapid prototyping and development while meeting the requirements for a low-power OOK transmitter operating within the 260 to 470 MHz frequency band.

**Rayming Corporation – TX-99 300 MHz AM Transmitter/RE-99 Receiver Pair**

From a previous project, there was a Ming TX-99 transmitter/ RE-99 receiver pair available for use for this project. The difficulties of the Ming TX-99 were that little
documentation was provided for both the company, and the device. The datasheet for the Ming TX-99 offered very few maximum and minimum operating conditions. However, what made up for the difficulties of the Ming TX-99 was the simplicity of the design.

**Ming Tx-99 transmitter**

The architecture of this transmitter is based on a colpitts oscillator design seen in Figure 5-4 and consists of a single high frequency NPN BJT transistor and a LC tank to tune the transmitter to oscillate at a specific frequency.

![Figure 5-4. Schematic of Ming TX-99 Taken from Datasheet](image)

Initial performance tests show that when biased at 0.6 V, the transmitter drains 850uA, which translates a power requirement of 510uW to transmit a constant 50% duty cycle 580mV peak to peak square pulse train of 1 kHz. The LC tank included a variable capacitance with a range of 2-7 pF to allow for the tuning of the operating frequency. Another advantage of the Ming TX-99 transmitter was the onboard antenna. The onboard antenna served as the inductor for the LC tank. With the Supply Voltage and Data nodes tied together as seen in the schematic of Figure 5.5, the transmitter can be used as an OOK transmitter. The whole transmitter module can be seen in Figure 5.6 with the onboard antenna highlighted for detail.
Additionally, the printed micro-strip inductor which serves as the onboard antenna can be tapped for attachment of an external antenna. For output power measurements, the signal pin of a SMA connector was soldered and tapped to the micro-strip inductor where an external antenna would be attached to, with ground of the SMA connector tied to ground of the Ming TX-99. The SMA was then attached to one end of a 1ft SMA cable (FLX402#1), and the other end to a DC blocker before finally attached to the input of a HP 8563E 9kHz to 26.5 GHz Spectrum Analyzer. The DC blocker was used to prevent damage to the spectrum analyzer by blocking DC current from directly entering the input of the spectrum analyzer. Both the supply voltage and data node of the Ming
TX-99 were connected to a power supply (Agilent E3631A) set to 2 V to send a 100% duty cycle signal (transmitter continuously on). Test setup can be seen in Figure 5-7, and the transmitter with attached SMA connector can be seen in Figure 5-8.
The result was an output power of approximately -4.67 dBm while draining 1.95 mA continuously from a 2 V supply. To test for minimum voltage operation, for 100% duty cycle, the transmitter was capable of operating at 1.2 V while draining 290 μA at an output power of -21.17 dBm. Assuming efficiency can be calculated as:

$$Efficiency = \frac{P_{Output\ Power}}{V_{Supply} \cdot I_{Supply}}$$

The data found in Table 5-2 can be tabulated.

<table>
<thead>
<tr>
<th>$V_{SUPPLY}$</th>
<th>$I_{SUPPLY}$ (mA)</th>
<th>Output Power (dBm)</th>
<th>Power (mW)</th>
<th>Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.2</td>
<td>0.29</td>
<td>-21.17</td>
<td>0.007638358</td>
<td>2.194930413</td>
</tr>
<tr>
<td>2</td>
<td>1.95</td>
<td>-4.5</td>
<td>0.354813389</td>
<td>9.097779211</td>
</tr>
</tbody>
</table>

**Ming RE-99 receiver**

Since a central monitoring station can be assumed to provide as much power as needed, the power requirements are not as stringent on the receiver side, making the RE-99, the receiver complement of the TX-99 a suitable receiving unit. Alike to the TX-99 transmitter, the RE-99 also lacks in documentation. From the schematic found in the datasheet as seen in Figure 5-8, it is seen that the RE-99 is an envelope detection circuit which is typical for AM/ASK/OOK receivers.

![Figure 5-8. Schematic of Ming RE-99 Taken from Datasheet](image-url)
Alike to the TX-99 the RE-99 provides an onboard antenna which also serves as the LC tank resonating at 300 MHz as the input for the receiver. It differs from the TX-99 in that it is no longer a micro strip antenna, but a loop antenna of 2 turns, and diameter of 5mm. It also offers a tap for an external antenna, and as recommended, a quarter wave monopole antenna was created by cutting a 22 gauge copper wire down to a length of 9.36 inches, which is approximately ¼ the wavelength for 300 MHz, and soldered to the antenna tap. Unfortunately, the sensitivity of the receiver was unable to be measured, but instead, a distance measurement was performed. A picture of the RE-99 receiver without the external antenna can be seen in Figure 5-9.

Figure 5-9. Ming RE-99 Receiver

**Ming distance measurements**

Because the sensitivity of the receiver is unknown, it was decided to perform an experiment to find the maximum transmission distance. The experiment was conducted in the atrium on the first floor of the New Engineering Building at the University of Florida. The floor plan of the atrium can be seen in Figure 5-11. The setup is detailed as follows:

- Transmitter was tied to serial output (USART) of a microcontroller (MSP430) outputting a constant data stream. The serial output of the microprocessor was tied to both the supply voltage and data node of the transmitter, forcing an OOK modulation scheme, while providing power to the transmitter. The microprocessor was powered with a 2V supply.
- The transmitter remained stationary and the receiver was attached to a cart for mobility. The height of the transmitter and receiver were 0.45m and 0.55m respectively. The receiver was powered via another power supply. A diagram of this setup can be seen in Figure 5-12.

- The output of the receiver was tied to the input of a Tektronix TDS210 Two Channel Digital Real-Time Oscilloscope

- For received power measurements, a 22 gauge copper quarter wave monopole antenna soldered to a SMA connector was connected to an Agilent E4448A PSA Series Spectrum Analyzer

![Figure 5-11. Floorplan of First Floor Atrium of New Engineering Building](image1)

![Figure 5-12. Experimental Setup for Distance Measurements](image2)

Within the testing, the placement of a 22 gauge copper quarter wave monopole antenna was placed on the transmitter, receiver, or both. The maximum transmission
distances for these test cases can be seen in Table 5-3. From this data, it shows that with a quarter wave monopole antenna on both the transmitter and receiver, the maximum distance for the successful detection of the original data stream serially outputted by the microprocessor, was found to be 19.4 m.

Table 5-3. Maximum Transmission Distances with Varying Antenna Locations

<table>
<thead>
<tr>
<th>Antenna Location</th>
<th>Maximum Distance</th>
</tr>
</thead>
<tbody>
<tr>
<td>Receiver Only</td>
<td>14.5 m</td>
</tr>
<tr>
<td>Transmitter Only</td>
<td>16.8 m</td>
</tr>
<tr>
<td>Transmitter &amp; Receiver</td>
<td>19.4 m</td>
</tr>
</tbody>
</table>

In addition to this maximum transmission distance experiment, measurements for received power were taken as well. These measurements were taken from an Agilent E4448A PSA Series Spectrum Analyzer connected to a quarter wave monopole antenna. For all measurements, the transmitter also had a matched quarter wave monopole antenna. Figure 5-13 shows the received power versus distance, with reference to the layout of the testing environment, the atrium of the New Engineering Building.

Figure 5-13. Received Power vs. Distance With Reference to Room Shape

From this Figure, it can be seen that at around 10m, the hallways of the floor plan began to act as a sort of waveguide. This caused the received power to increase after 10
meters was reached. From the maximum distance experiment, it is shown that the maximum distance achieved was set to be around 19.4m. At that distance, the received power was approximately -70 dBm. From this, it can be concluded that the sensitivity of the receiver is approximately -70 dBm. Figure 5-14 shows the power spectrum taken from the spectrum analyzer at 1m, and 8m respectively.

![Figure 5-14. Screen Capture of Received Power Spectrum. (a) at 1m (b) and 8m](image)

**Linx Technologies LR series Transmitter and Receiver**

Unfortunately, Rayming Corporation no longer exists. As time goes by, locating the Ming TX-99 or RE-99 for purchase becomes increasingly difficult as suppliers have depleted their stocks with no new shipments coming in to replenish their supply.

This prompted the selection of the Linx LR series transmitter and receiver, which is an update from their LC series line of transmitters and receivers. With these transmitters and receivers, comes a plethora of application notes and documentation to aide in the design of a wireless communications link. Linx Technologies also provides pre-fabricated low-profile antennas for use with the transmitters and receivers. The outstanding product provided by Linx Technologies, as well as the plethora of
documentation notes, makes rapid prototyping and development with the LR series transmitter and receiver fairly painless.

**Linx Technologies TXM-315-LR**

The LR series transmitter from Linx Technologies is a high performance synthesized ASK/OOK transmitter which has the ability to reach a serial data rate of 10 kbps. The transmitter consists of a PLL synthesized architecture offering low-power consumption, accurate operating frequency, and power-down functions with an antenna serving as the only external part needed. The system level architecture of the transmitter can be seen in Figure 5-15 from the TXM-315-LR datasheet.

![Figure 5-15. System Level Architecture of LINX TXM-315-LR](image)

The components of the transmitter consist of a Voltage Controlled Oscillator (VCO) locked through a phase locked loop (PLL) which is referenced to a high precision crystal. The output of the VCO is then amplified and buffered by a power amplifier before the carrier is filtered to attenuate and suppress harmonics and spurious emissions to within legal limits. The carrier is then output to free space via the 50 ohm antenna port of the transmitter. The pin-out of the TXM-315-LR transmitter can be seen in Figure 5-16.
Several unique features such as a Power Down line (PDN) which is used to power down the transmitter’s power amplifier when no data is present for transmission, and a Level Adjust line (LADJ) which is used to limit the output power for the transmitter, can be used in unison to even further reduce the power consumption of the transmitter. The LADJ line can prove to be even more useful during FCC testing and verification to compensate for antenna gains. By tying the PDN, supply voltage, and Data input node together, and driving this node with the serial output of the microprocessor, the transmitter, like the Ming TX-99, can be used as an OOK transmitter.

Initial test setup where the antenna port is soldered to a 50 ohm micro-strip, and connected to a spectrum analyzer (Agilent E4448A PSA series Spectrum Analyzer) via a 1 ft long SMA cable show the minimum bias of 1.6 V to send a 100% duty cycle signal (transmitter continuously on) requires 6 mA to output 0.34 dBm of output power. Additionally, at 2 V bias, the transmitter can output approximately 3.08 dBm while draining 8 mA of current. A list of power specifications, with efficiency calculated in the same fashion as for the Ming TX-99 analysis detailed previously, is seen in Table 5-4.

<table>
<thead>
<tr>
<th>( V_{\text{SUPPLY}} )</th>
<th>( I_{\text{SUPPLY}} ) (mA)</th>
<th>Output Power (dBm)</th>
<th>Power (mW)</th>
<th>Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.6</td>
<td>6</td>
<td>0.34</td>
<td>1.081433951</td>
<td>11.26493699</td>
</tr>
<tr>
<td>2</td>
<td>8</td>
<td>3.08</td>
<td>2.032357011</td>
<td>12.70223132</td>
</tr>
</tbody>
</table>

Table 5-4. Performance of LINX TXM-315-LR
As compared to the Ming TX-99, although output power and efficiency are both better than the TX-99, the minimum power required to turn the transmitter on is significantly higher, and may pose a serious problem when trying to obtain power from scavenged energy. This shows that although Linx Technologies’ TXM-315-LR is a more robust and mature commercial product, the most important requirement of minimal power expenditure is not met, and may not serve as a suitable RF transmitter for the wireless hydrogen sensor node.

**Linx Technologies RXM-315-LR**

Alike to the assumption made for the Ming RE-99, the same assumption that a central monitoring station can provide as much power as deemed necessary is also made for the RXM-315-LR. Unlike the products by Ming, and alike to the TXM-315-LR, there exists a plethora of documentation for the RXM-315-LR, as well as several application notes.

The system level architecture of the RXM-315-LR is seen in Figure 5.17, and unlike the envelope detection circuit of the RE-99, the RXM-315-LR receiver modules employs a single-conversion super-heterodyne architecture to demodulate the received signal.

![Figure 5-17. System Level Architecture for RXM-315-LR](image)
The RF signal entering from the 50 ohm matched antenna is band-pass filtered before being amplified by an NMOS cascade, Low Noise Amplifier (LNA). The amplified signal is then down-converted to a 10.7 MHz Intermediate Frequency (IF) which is done by mixing the amplified signal with a VCO controlled by a PLL, referenced to a high precision crystal. The mixer stage, which down-converts the signal, consists of a pair of double balanced mixers, and includes an image rejection circuit. Once down-converted to an IF frequency, the signal is further amplified, filtered and demodulated to recover the original base-band data bit-stream. This baseband signal is squared by a data slicer and output to the DATA pin of the receiver.

This architecture, along with the high IF frequency and ceramic IF filters, helps reduce the susceptibility to interference, which is a problem associated with OOK modulation. The pin-out of the receiver can be seen in Figure 5.18. Due to the architecture and components of this receiver, it is able to achieve a very high sensitivity of -112 dBm, while remaining unsusceptible to interfering signals which plague OOK communication links.

<p>| | | | | |</p>
<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>NC</td>
<td>ANT</td>
<td>16</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>NC</td>
<td>GND</td>
<td>15</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>NC</td>
<td>NC</td>
<td>14</td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>GND</td>
<td>NC</td>
<td>13</td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>VCC</td>
<td>NC</td>
<td>12</td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>PDN</td>
<td>NC</td>
<td>11</td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>RSSI</td>
<td>NC</td>
<td>10</td>
<td></td>
</tr>
<tr>
<td>8</td>
<td>DATA</td>
<td>NC</td>
<td>9</td>
<td></td>
</tr>
</tbody>
</table>

Figure 5-18. Pin-Out of RXM-315-LR Transmitter

Compared to the Ming RE-99 receiver, Linx Technologies RXM-315-LR receiver shows to have the best performance out of both receivers, while the Ming TX-99 shows
to have the most favorable performance between the two transmitters. To optimize the wireless communication link, the combination of the Ming TX-99 and Linx RXM-315-LR should be used together. Unfortunately both components are set to operate at different frequencies. As stated before, one of the attractive characteristics of the Ming TX-99 is the tunable variable capacitor which can change the resonant frequency of the LC tank. If the LC tank were to be tuned to 315 MHz, the Ming TX-99 can be used in conjunction with the RXM-315-LR to create an optimized wireless link for the wireless hydrogen sensor node.

**Wireless Link Optimization**

The optimization of the wireless link includes not only the combination of the Ming- TX-99 transmitter with the Linx Technologies RXM-315-LR receiver, but also the minimization of power consumption by the Ming TX-99 transmitter, development of a low-profile antenna to reduce the size and increase the compactness of the overall wireless sensor package, and work on the receiver side to gather data from the transmitter for use as a central monitoring station. Additionally, studies should be done to develop a wireless node operating within the restrictions set by FCC part 15.231, which regulates operation within the 260-470 MHz range. Prior to all the additional work, a starting ground should be set by performing an initial power analysis of the system.

**Ming TX-99 Power Analysis**

A power analysis on the Ming TX-99 was performed by David Johnson, at Cisco Systems in Bradenton, FL. Similar to the power analysis performed for the microcontroller in Chapter 4. To reiterate the test setup, a 383 ohm resistor was connected in series between the power supply, and microcontroller/RF transmitter. The USART output of the microcontroller was used to power and drive the Ming TX-99
transmitter. P6248 Differential probes attached to a Tektronix TDS5104B Digital Phosphor Oscilloscope were used to measure the voltage across the 383 ohm resistor. Calculation for the average power consumption is as follows:

- Total Area (TA) under the measured curve is calculated (units of V·sec)
- Peak power is calculated as:
  \[ P_{\text{peak power}} = \frac{V_{\text{MAX}}}{383} \cdot V_{\text{SUPPLY}} \]
  Where \( V_{\text{MAX}} \) is the maximum point of the measured curve
- Average power is calculated as
  \[ \text{AvgPower} = \left( \frac{\text{TotalArea}}{\tau_{\text{duration}} \cdot 383} \right) \cdot V_{\text{SUPPLY}} \]
  Where \( \tau_{\text{duration}} \) is the duration of the measured curve

It was previously shown in Chapter 4, that for any task other than for initialization, the MSP430 microcontroller only consumed a low 2.5uW of power. To test the average power consumption for the transmission of a 500 uS bit length pulse, the procedure above is applied, and the measured curve taken from a screen dump of the Tektronix TDS5104B Digital Phosphor Oscilloscope can be seen in Figure 5-19.

![Figure 5-19. Tektronix TDS5104B Digital Phosphor Oscilloscope Screen Capture of Power Analysis Performed for RF Transmission of One Bit](image-url)
From this curve, it can be calculated that for the transmission of a 500 uS pulse, the transmitter consumes an average power of 261 uW with a peak of 522 uW. As expected, the RF transmitter does not consume power for the RF transmission of a logical “low” or “0”. The charging and discharging characteristics of Figure 5-19 may be due to the LC resonant tank which is used to set the operating frequency of the RF transmitter to 300 MHz.

Another experiment was performed to analyze the power consumption for the RF transmission of multiple bits. The screen capture from the Tektronix TDS5104B Digital Phosphor Oscilloscope can be seen in Figure 5-20. From Figure 5-20, it can be seen that all rising slopes and falling slopes are equal.

Figure 5-20. Tektronix TDS5104B Digital Phosphor Oscilloscope Screen Capture of Power Analysis Performed for RF Transmission of Multiple Bits

By grouping all rising slopes together at the front, and grouping all falling slopes together in the back to arrange a simple triangle, and assuming that the power is not completely discharged between “high” bits, it can be gathered that the worst case
average power consumption (maximum average power consumption) to send N bits, can be equated as:

\[ \text{WorstCasePower} = \frac{1}{2} \cdot P_{\text{peakPower}} \cdot N_{\text{bits}} \]

Where \( P_{\text{peakPower}} \) is the peak power required to transmit a single bit (261 uW), and the worst case data input to the transmitter is a continuous train of “1” or “high” pulses, one after the other with no “low” or “0” bits in between.

**Low Profile Antenna**

To increase the compactness of the wireless sensor, Linx Technologies provides the ANT-315-SP, or “SPLATCH” antenna. The features of this antenna are an ultra-compact package and good resistance to proximity effects. The SPLATCH uses a grounded-line technique to achieve a quarter wave type antenna centered at 315 MHz, with a bandwidth of 5 MHz. The “SPLATCH” antenna with dimensions can be seen in Figure 5-21.

![Figure 5-21. LINX ANT-315-SP SPLATCH Antenna From Datasheet](image-url)
Unfortunately, unlike the TXM-315-LR transmitter or RXM-315-LR receiver, the SPLATCH antenna does not provide much in documentation. The only details given are the antenna’s requirements for a 1.5” x 3.0” ground plan, and a 50 Ohm micro-strip line between RFIN of the antenna, and the RF output node of the transmitter. An assembled FR4 testing board for the SPLATCH antenna can be seen in Figure 5-22.

![Testing Board for SPLATCH Antenna. (a) Front (b) and Back](image)

Figure 5-22. Testing Board for SPLATCH Antenna. (a) Front (b) and Back

Initial tests to obtain the gain of the SPLATCH antenna were performed with an Agilent E8316A 10 Mhz to 6 GHz PNA series Network Analyzer, and an Agilent E4448A PSA Series Spectrum Analyzer. An antenna test structure was tied to the Agilent E8254A 250 kHz to 40 GHz PSG-A Series Signal Generator via a FLX402#1 SMA cable to serve as the transmitting antenna, and an identical antenna structure was tied to the Agilent E4448A PSA Series Spectrum Analyzer outputting a RF power of 10 dBm at 315 MHz, via a FLX402#1 SMA cable to serve as a receiving antenna. With respect to the free space path loss equation, where
\[ P_R(R) = \frac{P_T \cdot G_T \cdot G_R \cdot \lambda^2}{(4 \cdot \pi \cdot R)^2} \]

Assuming frequency, distance, and transmitted and received power are all controlled parameters, since the transmitter and receivers antennas are identical, their respective gains can be calculated. For these calculations, cable losses were also taken into account. Initial gain measurements from different distances showed the data found in Table 5-5. As a comparison, the gains of the 22 gauge copper quarter-wave monopole antennas were also tested to serve as a comparison to the SPLATCH antennas.

### Table 5-5. Antenna Gain Measurements

<table>
<thead>
<tr>
<th>Distance (m)</th>
<th>Received Power (dBm)</th>
<th>Transmitted Power (dBm)</th>
<th>Path Loss (dB)</th>
<th>Cable Loss (dB)</th>
<th>Gain (dB)</th>
<th>Antenna</th>
</tr>
</thead>
<tbody>
<tr>
<td>4</td>
<td>-58.11</td>
<td>10</td>
<td>34.44918309</td>
<td>0.41</td>
<td>-16.6254</td>
<td>SPLATCH</td>
</tr>
<tr>
<td>5</td>
<td>-60</td>
<td>10</td>
<td>36.38738335</td>
<td>0.41</td>
<td>-16.6013</td>
<td>SPLATCH</td>
</tr>
<tr>
<td>4</td>
<td>-26</td>
<td>10</td>
<td>34.44918309</td>
<td>0.41</td>
<td>-0.57041</td>
<td>Monopole</td>
</tr>
<tr>
<td>5</td>
<td>-28</td>
<td>10</td>
<td>36.38738335</td>
<td>0.41</td>
<td>-0.60131</td>
<td>Monopole</td>
</tr>
</tbody>
</table>

From these measurements, it can be seen that the “SPLATCH” antennas exhibited poor performance with an approximate gain of -16.6 dB, as compared to the -0.6 dB gain of the monopole antennas. Clearly, there is a distinct performance loss of the SPLATCH antenna over the monopole antennas, and a more comprehensive analysis of the SPLATCH antenna should be conducted.

To measure the resonant frequency of the SPLATCH antenna, a 1 Port S-parameter measurement was taken using an antenna, and an Agilent E8316A 10 Mhz to 6 GHz PNA series Network Analyzer. Additionally, the antenna test structure was simulated in Ansoft Designer. From the S-Parameter measurements taken by Agilent E8316A 10 Mhz to 6 GHz PNA series Network Analyzer shown in Figure 5-23, the SPLATCH antennas
showed to have a resonant frequency located at 330 MHz with a bandwidth of 5 MHz, rather than 315 MHz as shown in the specifications of the datasheet.

Figure 5-23. S-Parameter for SPLATCH Antenna

Since the fundamental operating frequency is 315 MHz, this may be the reason for the poor performance of the antenna. A planar EM simulation in Ansoft Designer was performed, and showed similar results compared to the measurements taken from the E8316A 10 Mhz to 6 GHz PNA series Network Analyzer as seen in Figure 5.24(a). Also seen in Figure 5.24(b) is the corresponding antenna test structure model within Ansoft Designer.

To shift the resonant frequency of the SPLATCH antenna down to 315 MHz, a matching circuit consisting of a shunt and series capacitor were simulated in Ansoft Designer, and then realized and tuned on the SPLATCH antenna test structure. The
frequency shifted matched antenna S parameters can be seen in Figure 5-25 with
comparison to the original un-matched antenna. Figure 5-26 shows that the matched
antenna retains an approximate -10 dB bandwidth of 5 MHz.

Figure 5-24. S-Parameters of Measured (red) and Simulated (blue) in Ansoft Designer.
   (a) S11 (b) Simulation Setup

Figure 5-25. Matched Antenna (Red) vs. Unmatched Antenna (blue)
Once matched, the SPLATCH antenna was again tested for gain. The experiment setup is identical to the previously mentioned set up, and the data for this experiment can be found in Table 5-6.

Table 5-6. Gain Measurements for Matched Antenna

<table>
<thead>
<tr>
<th>Distance (m)</th>
<th>Received Power (dBm)</th>
<th>Transmitted Power (dBm)</th>
<th>Path Loss (dB)</th>
<th>Cable Loss (dB)</th>
<th>Gain (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>-49</td>
<td>4</td>
<td>28.42858318</td>
<td>0.11</td>
<td>-12.2307</td>
</tr>
<tr>
<td>3</td>
<td>-50</td>
<td>4</td>
<td>31.95040836</td>
<td>0.11</td>
<td>-10.9698</td>
</tr>
<tr>
<td>4</td>
<td>-48</td>
<td>4</td>
<td>34.44918309</td>
<td>0.11</td>
<td>-8.72041</td>
</tr>
</tbody>
</table>

These results show an approximate 6 dB increase in performance, but the resulting gain is still lower than that of the quarter wave 22 gauge copper monopole antenna. This may be due to the poor radiation efficiency stemming from electrically small characteristics of the antenna as compared to the 22 gauge copper monopole antennas. A trade-off thus exists in the reduction of gain for a compact surface mount antenna. For
optimization purposes, since size is an unlimited parameter at the central monitoring station, the quarter wave 22-gauge monopole antenna can be used as the antenna for the receiver, while the SPLATCH antenna can be used as the antenna for the wireless sensor node.

**RF Transmitter Optimization**

As previously mentioned, an attractive feature of the Ming TX-99 transmitter is the simplicity of the design, and the exposed discrete components that can be changed and replaced to tune both frequency, and power consumption. All components were extracted and measured, with the micro-strip inductance measured by an Agilent E8316A 10 Mhz to 6 GHz PNA series Network Analyzer. The impedance measurement of the inductor can be seen in Figure 5-27.

![Figure 5-27. Microstrip Inductance Measurement for Ming TX-99 Onboard Antenna](image-url)
By placing the value of these components into Agilent ADS 2003, the discrete components can be varied to reduce the minimum power requirements of the RF transmitter. A list of all values relating to the discrete components of the Ming TX-99 can be found in Table 5-7.

Table 5-7. Component Values of Ming TX-99 Transmitter

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>C1</td>
<td>2-7 pF</td>
</tr>
<tr>
<td>C2</td>
<td>12 pF</td>
</tr>
<tr>
<td>C3</td>
<td>3.3 pF</td>
</tr>
<tr>
<td>L1</td>
<td>28.22 nH</td>
</tr>
<tr>
<td>L2</td>
<td>1 uH</td>
</tr>
<tr>
<td>R1</td>
<td>47 kOhms</td>
</tr>
<tr>
<td>R2</td>
<td>100 Ohms</td>
</tr>
<tr>
<td>Transistor</td>
<td>MMBTH10</td>
</tr>
</tbody>
</table>

By simulating the circuit found previously in Figure 5-5 in Agilent ADS 2003, the discrete components directly correlating to output power and power consumption can be found. The simulation setup in Agilent ADS can be found in Figure 5-28. From these simulations, it is seen that the major components that control output power and power consumption were related to the resistors found at the base and emitter of the transmitter. The emitter resistance increases the stability of the transmitter, so the component with the most effect on the output power and power consumption is the resistor in the base. Also discovered, is the output power can be increased by the removal of the diode also found at the base of the transistor.

Another component that had an influence on the power characteristics of the transmitter was the high-frequency NPN BJT transistor. By choosing a different transistor capable of high frequency operation, and had lower power requirements, the
overall power of the transmitter can be further reduced. Table 5-8 shows the results of an internet search for high-frequency NPN BJT transistors.

<table>
<thead>
<tr>
<th>NPN Transistor</th>
<th>Ic(mA)</th>
<th>PTOT(mW)</th>
<th>$f_r$</th>
<th>Noise Figure</th>
</tr>
</thead>
<tbody>
<tr>
<td>MMBTH10</td>
<td>50</td>
<td>1250</td>
<td>650 MHz</td>
<td>N/A</td>
</tr>
<tr>
<td>BFS19</td>
<td>25</td>
<td>500</td>
<td>260 MHz</td>
<td>N/A</td>
</tr>
<tr>
<td>BFT25A</td>
<td>6.5</td>
<td>32</td>
<td>5 GHz</td>
<td>1.8 @ 1 GHz</td>
</tr>
<tr>
<td>BFW92A</td>
<td>25</td>
<td>375</td>
<td>3.2 GHz</td>
<td>2.5 @ 800 MHz</td>
</tr>
<tr>
<td>BFS17A</td>
<td>25</td>
<td>300</td>
<td>3 GHz</td>
<td>2.5 @ 800 MHz</td>
</tr>
</tbody>
</table>

Agilent ADS was used to find the minimal operating point for the transmitter to turn on and transmit present a carrier. The simulation setup within ADS 2003 was seen previously in Figure 5-28. To actually test the performance changes, a manual sweep of these values was performed by de-soldering and re-soldering different components onto the transmitter board, and measuring the output power and current consumption given a 2V supply, in a similar setup as previously described. However, instead of using an
Agilent E4448A PSA Series Spectrum Analyzer, a HP 8563E spectrum analyzer was used. Table 5-9 shows the results of this trial and error experimentation with the transmitter.

Table 5-9. Performance of Various Transistors and Resistors for Ming TX-99 Transmitter

<table>
<thead>
<tr>
<th>Transistor</th>
<th>Base Resistance (kOhms)</th>
<th>Emitter Resistance (Ohms)</th>
<th>Supply Voltage</th>
<th>Supply Current</th>
<th>Output Power (dBm)</th>
<th>Output Power (mW)</th>
<th>Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>MMBTH10</td>
<td>47</td>
<td>100</td>
<td>2</td>
<td>2.75</td>
<td>-3</td>
<td>0.5012</td>
<td>9.1124952</td>
</tr>
<tr>
<td>MMBTH10</td>
<td>47</td>
<td>300</td>
<td>2</td>
<td>1.77</td>
<td>-13</td>
<td>0.0501</td>
<td>1.2529681</td>
</tr>
<tr>
<td>MMBTH10</td>
<td>100</td>
<td>100</td>
<td>2</td>
<td>0.911</td>
<td>-11</td>
<td>0.0794</td>
<td>4.35965</td>
</tr>
<tr>
<td>MMBTH10</td>
<td>100</td>
<td>200</td>
<td>2</td>
<td>0.996</td>
<td>-9</td>
<td>0.1259</td>
<td>6.3199067</td>
</tr>
<tr>
<td>MMBTH10</td>
<td>200</td>
<td>100</td>
<td>2</td>
<td>0.923</td>
<td>-11.83</td>
<td>0.0656</td>
<td>3.5544164</td>
</tr>
<tr>
<td>MMBTH10</td>
<td>200</td>
<td>200</td>
<td>2</td>
<td>0.86</td>
<td>-13</td>
<td>0.0501</td>
<td>2.9138793</td>
</tr>
<tr>
<td>MMBTH10</td>
<td>200</td>
<td>300</td>
<td>2</td>
<td>1.73</td>
<td>-1.33</td>
<td>0.7362</td>
<td>21.277662</td>
</tr>
<tr>
<td>BFT25A</td>
<td>47</td>
<td>100</td>
<td>2</td>
<td>1.52</td>
<td>-3.33</td>
<td>0.4645</td>
<td>15.280108</td>
</tr>
<tr>
<td>BFT25A</td>
<td>47</td>
<td>200</td>
<td>2</td>
<td>1.095</td>
<td>-5.5</td>
<td>0.2818</td>
<td>12.869328</td>
</tr>
<tr>
<td>BFT25A</td>
<td>100</td>
<td>100</td>
<td>2</td>
<td>0.8</td>
<td>-8.5</td>
<td>0.1413</td>
<td>8.8283597</td>
</tr>
<tr>
<td>BFT25A</td>
<td>150</td>
<td>100</td>
<td>2</td>
<td>0.65</td>
<td>-10.93</td>
<td>0.0826</td>
<td>6.3541381</td>
</tr>
<tr>
<td>BFS17A</td>
<td>47</td>
<td>100</td>
<td>2</td>
<td>2.1</td>
<td>-2.67</td>
<td>0.5408</td>
<td>12.875103</td>
</tr>
</tbody>
</table>

In Table 5-9, the resistance combination which produces the highest efficiencies for each transistor are boxed in red. It is found that the Philips’ BFT25A high-frequency NPN BJT transistor served as the best transistor for use in the RF transmitter. By using the same nominal resistances from the original Ming Tx-99 transmitter, it was capable of increasing the efficiency by increasing the output power, while lowering the overall power consumption for the transmission of a constant 315 MHz carrier, as compared to the original Ming TX-99 tuned to 315 MHz.

**FCC Part 15.231**

The operating frequency of 315 MHz lies in the FCC frequency operating band of 260 MHz to 470 MHz, which is regulated by Part 15.231 of CFR47. The specific
regulations of part 15.231 are rather unusual in that for many bands, the FCC specifies only fundamental power, harmonic levels, and allowed bandwidth, while for the frequency band of 260 to 470 MHz, the FCC regulates this spectrum based on the intended function and form of the transmitted data. Part 15.231 is broken into paragraphs A through D, while paragraph E applies only if the rules specific to paragraph A are broken. Due to the complexity and application specific regulations of part- 15.231, the limitations of part 15.231 are best illustrated in a flowchart form as seen in Figure 5-29 taken from [23]. Interested parties should visit the FCC website at http://wireless.fcc.gov/rules.html.

Figure 5-29. Flowchart for FCC Part 15-231 Requirements
The limitations of part 15.231 section A are expressed in sections B through D, and can be found in Table 5-10, while the limitations of section E can be found in Table 5-11 and illustrated in Figure 5.30 from [24].

Table 5-10. Limitations under FCC Part 15.231 (a-d) **Linear Interpolations

<table>
<thead>
<tr>
<th>Fundamental Frequency (MHz)</th>
<th>Field Strength of Fundamental (uV/m)</th>
<th>Field Strength of Spurious Emission (uV/m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>40.66 - 40.70</td>
<td>2250</td>
<td>225</td>
</tr>
<tr>
<td>70 - 130</td>
<td>1250</td>
<td>125</td>
</tr>
<tr>
<td>130 - 174</td>
<td>1250 to 3750**</td>
<td>125 to 375**</td>
</tr>
<tr>
<td>174 - 260</td>
<td>3750</td>
<td>375</td>
</tr>
<tr>
<td>260 - 470</td>
<td>3750 to 12500**</td>
<td>375 to 1250**</td>
</tr>
<tr>
<td>Above 470</td>
<td>12500</td>
<td>1250</td>
</tr>
</tbody>
</table>

Table 5-11. Limitations under FCC Part 15.231 (e) **Linear Interpolations

<table>
<thead>
<tr>
<th>Fundamental Frequency (MHz)</th>
<th>Field Strength of Fundamental (uV/m)</th>
<th>Field Strength of Spurious Emission (uV/m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>40.66 - 40.70</td>
<td>1000</td>
<td>100</td>
</tr>
<tr>
<td>70 - 130</td>
<td>500</td>
<td>50</td>
</tr>
<tr>
<td>130 - 174</td>
<td>500 to 1500**</td>
<td>50 to 150**</td>
</tr>
<tr>
<td>174 - 260</td>
<td>1500</td>
<td>150</td>
</tr>
<tr>
<td>260 - 470</td>
<td>1500 to 5000**</td>
<td>150 to 500**</td>
</tr>
<tr>
<td>Above 470</td>
<td>5000</td>
<td>500</td>
</tr>
</tbody>
</table>

Figure 5-30. Graphical Representation of Field Strength Limitations for Part 15.231 Section e.
For the frequency of 315 MHz, the field strength of the fundamental in section A can be calculated as uV/m at 3 meters = 41.6667(F) – 7083.3333, while for section E, the field strength can be calculated as uV/m at 3 meters = 16.6667(F) = 2833.3333, where F is the operating frequency in MHz. This shows that the requirements for section E of part 15.231 are lowered by almost 40% as compared to the limitations for part 15.231 section A. Since multiple accessing schemes are currently being developed, and an ID tag will be assigned to every transmitted message, The RF transmitter should be designed to meet the requirements of Part 15.231 section a found in Table 5-10.

For an operating frequency of 315 MHz, the field strength fundamental is calculated to be 6.041mV/m referenced to 3 meters. Assuming an isotropic antenna of gain 1, the allowed transmitted power can be calculated as:

\[ P_{Transmitted} = \frac{1}{30} \cdot \left( E^2 \cdot d^2 \right) \]

Where E is the field strength and d is the reference distance of 3m. From this equation, it can be found that the transmitted power, assuming an isotropic antenna of gain 1, is approximately 10 uW, or -19.6 dBm. Given the best case gain from the SPLATCH antenna of -8.72 dB, an output power less than -10.88 dBm, or approximately -11 dBm is required from the transmitter to meet the FCC part 15.231 requirements for radiated power. From Table 5-9, it is shown that the Ming TX-99, with a base resistance of 200k ohms, emitter resistance of 100 ohms, and the NPN BJT transistor replaced with the Philips BFT25A, fulfills the –10.88 dBm output power requirement with an output power equal to -10.93 dBm. Thus, by using this specific transmitter configuration, the requirements of FCC part 15.231 can be fulfilled.
Given a receiver sensitivity of -112 dBm, receiver antenna of approximately -1 dB, transmitter antenna of -9 dB, output power of -10.93 dBm, and the two-ray ground reflection model, assuming the height of both the transmitting and receiving antenna is 1 meter, the theoretical distance of the wireless sensor node can be calculated to be approximately 189 meters. If we include an estimated 15 dB path loss associated with the loss due to random variables such as multi-path fading effects, this distance is lowered to approximately 79.7 meters.

**Central Monitoring Station**

The transmitters of the wireless hydrogen sensor eventually will be required to communicate to a central monitoring station. For the design of the central monitoring station which consists of a RF receiver module, a Data Acquisition (DAQ) device, and a laptop or desktop computer, it is assumed that power consumption is of little worry for the development of the receiving unit. The central monitoring station was originally designed for use with the Ming RE-99. Since the RE-99 is of an envelope detection topology, the RE-99 is especially sensitive to interference and noise when a carrier is absent. To remedy this problem, considerations into software decoding should be emphasized in the design of the central monitoring station to filter and remove the effects of noise and interference.

The central monitoring station currently consists of the Ming RE-99 receiver, and a National Instrument USB-6008 DAQ device. The Ming receiver is powered and outputs data through an analog output and input node on the USB-6008 DAQ device. The USB-6008 DAQ device itself is powered via an USB port from a laptop running LabVIEW 7.1. The USB-6008 provides basic data acquisition functionality for the purpose of data logging, and making portable measurements, while being fairly affordable, but powerful
enough for sophisticated measurement applications. The USB-6008 consists of eight 12-bit analog input channels, 12 digital I/O lines, 2 analog outputs, and 1 counter, with a maximum analog sampling rate of 10 k Samples/sec and an analog output range of 0 to +5 V which is sufficient for powering and gathering received data from the Ming RE-99 receiver.

When a carrier is absent, due to the envelope detection topology of the Ming RE-99, the receiver becomes susceptible to noise and interference. To mitigate the effects of the noise and interference, software coding was developed to distinguish data pulses from noise and interference. The proposed solution is to use the maximum sampling rate of 10kHz for the DAQ, and continuously sample, taking one analog sample every 100 μS. Then these samples are buffered, and a moving average filter is used to differentiate pulses from noise and interference. An example of a moving average filter can be seen in Figure 5.31.

![Figure 5-31. Moving Average Filter Example](image)

From Figure 5.31, it shows that the middle four samples are the pulse, and the rest is noise. If the samples are continuously filtered over a 4 sample window, and if the average of those 4 samples is equal to, or greater than the expected amplitude, then a
pulse has been detected. This moving average filter was implemented in LabVIEW code, and can be seen in Figure 5-32(a). To provide a graphical interface, the LabVIEW front panel user gui can be found in Figure 5-32(b).

The considerations, design, and optimization of a wireless communications link have been described in this chapter. A modulation scheme of OOK was selected, and a carrier frequency of 315 MHz was picked as the operating frequency of the wireless communications link. Additionally, a low-profile antenna was tested and matched, and the transmitter set to operate within the confines of FCC part 15.231. The development of a central monitoring station was also described. To reiterate the selection of an OOK modulation scheme, OOK is attractive in that power is only consumed for the transmission of a “1” or “high” bit, and consumes little to no power for the transmission of a “0” or “low”. By exploiting these characteristics of OOK, perhaps a source coding scheme can be used to further reduce the power consumption required to send a fixed length message. A source coding scheme for minimum energy expenditure is described in the following chapter.
CHAPTER 6
MINIMUM REDUNDANCY MINIMUM ENERGY CODING

Within the system level design of the wireless sensor, are the trade-offs involved with system optimization. Current development in wireless technologies involve new coding and modulations schemes which can bring about better performance for power, bandwidth, data-rate, and error. With these new coding and modulation schemes however, come the complexities of the correlating circuits and system topologies. The trade-off exists in that a system like the wireless hydrogen sensor, can reduce data-rate and bandwidth, reduce circuit and system complexity, and opt for a modulation scheme that consists of a more simple design for the sake of lower power consumption. This is why OOK was selected as the modulation scheme of choice. But why should system level power optimization just stop at the circuit level implementation? Perhaps a coding technique can be used to sacrifice data rate, but lower the power consumption of the system, which is the most important factor in the design of a wireless hydrogen sensor node. The development of a source coding technique to reduce the power required to send a message is reported in this chapter.

Minimum Energy Coding

Currently, an On-Off Keying (OOK) modulation approach is used for RF data transmission. OOK was selected as the modulation technique due to the intrinsic property of OOK, where power is consumed at a “1” or high, and no power is consumed for a “0”. It is this intrinsic property that can be exploited to further minimize the current power consumption for data transmission. If the number of “1” or high bits in the
transmitted message can be reduced, then the overall power needed for the successful relay of a message will also be reduced. Thus, if a Minimum Energy (ME) coded message has less “high” bits as compared to the original source message, the result is a reduction of power consumption. A ME coding scheme is reported in detail by Erin and Asada [25], where the same properties of an OOK modulation scheme are exploited for their Minimum Energy Coding scheme.

The Minimum Energy Coding scheme developed by [25] uses a source coding method to convert a source code of a specified length, to a codeword sequence of fixed length with a maximum of one high bit in the entire codeword sequence. A minimum energy code table for the coding scheme proposed by [25] can be seen in Table 6-1.

<table>
<thead>
<tr>
<th>Source Bits</th>
<th>Codeword</th>
<th>Source Bits</th>
<th>Codeword</th>
</tr>
</thead>
<tbody>
<tr>
<td>ME(3,2)</td>
<td>ME(3,2)</td>
<td>ME(7,3)</td>
<td>ME(7,3)</td>
</tr>
<tr>
<td>00</td>
<td>000</td>
<td>000</td>
<td>0000000</td>
</tr>
<tr>
<td>01</td>
<td>001</td>
<td>001</td>
<td>0000001</td>
</tr>
<tr>
<td>10</td>
<td>010</td>
<td>010</td>
<td>0000010</td>
</tr>
<tr>
<td>11</td>
<td>100</td>
<td>011</td>
<td>0000100</td>
</tr>
<tr>
<td></td>
<td>100</td>
<td>100</td>
<td>0001000</td>
</tr>
<tr>
<td></td>
<td>101</td>
<td>010</td>
<td>0010000</td>
</tr>
<tr>
<td></td>
<td>110</td>
<td>011</td>
<td>0100000</td>
</tr>
<tr>
<td></td>
<td>111</td>
<td>100</td>
<td>1000000</td>
</tr>
</tbody>
</table>

Although the concept behind this source coding scheme is concrete, and will definitely would lower the power required for the transmission of a message with fixed length, this coding scheme lowers the power consumption at an expense to codeword length. To represent all the symbols possible for a source message of bit of length n within a coded message of only one “high” bit, a coded message length of $2^n - 1$ is required. So, to transmit a source code of length 4 bits, a coded message of length 15 is required. This adds 9 redundant bits to the original source message, and is not an
efficient coding scheme given an OOK modulation scheme. Because OOK modulation consists of data transmission over a single carrier, the longer a single transmitter occupies the fundamental carrier, the higher the probability of a collision caused by other transmitters attempting to transmit at the same time. This collision would result in the corruption or total loss of transmitted data from all sensor nodes attempting to transmit at the same time.

Minimum Redundancy Minimum Energy Coding

A technique proposed for use with the wireless hydrogen sensor, is one of both minimum energy, and minimum redundancy. This technique differs from previous work on this matter due to the consideration of minimizing energy without sacrificing the codeword length, redundant bits, and amount of time required for a transmitter to send a single message. Previous work in ME Coding [25] has suffered from the concatenation of many redundant bits--for example, requiring a coded message of 32 bits in length to send a 5 bit source message.

Because of the use of an OOK modulation technique on a single frequency carrier, as mentioned before, when two sensors are trying to transmit data on the channel at the same time, this collision results in a corruption or complete loss of data. By reducing the need for redundant bits, and lowering the message length, there is a reduction in the time one transmitter is occupying the channel, and reduces the chances for collision from multiple transmissions.

There are two different schemes that are currently being considering. One scheme is that of a delay based method with only one “high” bit per message, and another is the mapping of (n) source bits to a message with a maximum of two or three “high” bits per coded message.
For the delay based method, consider a (n) bit source message and assume that:

- Each message has equal probability of occurrence
- All bit lengths are of equal durations
- Power is only expended on the transmission of a “high” bit.

For the delay based source coding, the coded message consists of a single high bit with a specific pre-determined number of “zeros” appended before the “high” bit. These predetermined zero bits will look like a “delay” at the receiver side, and since all bit lengths are equal, there can be a mapping of this “delay” to the original source code as seen in Table 6-1.

Another source coding scheme involves the coding of a message with a pre-defined number greater than one, which translates to the maximum number of “high” bits in the coded message. This coding scheme assumes the same assumptions previously presented, and maps the source message to a coded message with a maximum of only two or three high bits per message. For example:

Consider a source message of 6 bits long. To find a coded message of minimum length (m) that can allow for $2^6$ or 64 symbols, while using only a maximum of three “high” bits, solve for,

$$2^n = 2^6 = 64 = mC_3 + mC_2 + mC_1 + mC_0$$

So, the minimum (m) needed to satisfy the above equation is 7.

This shows that a coded message of 7 bits long is required to be able to express all 64 symbols of the source message of 6 bits long. Thus, by the addition of 1 redundant bit,
64 symbols can be achieved while only using a maximum of three “high” bits as seen in Table 6-1.

Table 6-2. Proposed Source Coding Technique with Comparison to Technique of Erin and Asada.

<table>
<thead>
<tr>
<th>Source</th>
<th>CODED – 1 “high” (Previous Work)</th>
<th>CODED – 1 “high-delay”</th>
<th>CODED -2 “high”</th>
</tr>
</thead>
<tbody>
<tr>
<td>0000</td>
<td>000000000000000000000000000000</td>
<td>000000000000000000001</td>
<td>00000000000000000000</td>
</tr>
<tr>
<td>0001</td>
<td>000000000000000000000000000001</td>
<td>000000000000000000001</td>
<td>00000000000000000000</td>
</tr>
<tr>
<td>0010</td>
<td>000000000000000000000000000001</td>
<td>000000000000000000001</td>
<td>00000000000000000000</td>
</tr>
<tr>
<td>0011</td>
<td>000000000000000000000000000001</td>
<td>000000000000000000001</td>
<td>00000000000000000000</td>
</tr>
<tr>
<td>0100</td>
<td>000000000000000000000000000001</td>
<td>000000000000000000001</td>
<td>00000000000000000000</td>
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<tr>
<td>0101</td>
<td>000000000000000000000000000001</td>
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<tr>
<td>0110</td>
<td>000000000000000000000000000001</td>
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<td>0111</td>
<td>000000000000000000000000000001</td>
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<td>000000000000000000000000000001</td>
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<tr>
<td>1001</td>
<td>000000000000000000000000000001</td>
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<tr>
<td>1010</td>
<td>000000000000000000000000000001</td>
<td>000000000000000000001</td>
<td>00000000000000000000</td>
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<tr>
<td>1011</td>
<td>000000000000000000000000000001</td>
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</tr>
<tr>
<td>1100</td>
<td>000000000000000000000000000001</td>
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<td>1101</td>
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</tr>
<tr>
<td>1110</td>
<td>000000000000000000000000000001</td>
<td>000000000000000000001</td>
<td>00000000000000000000</td>
</tr>
<tr>
<td>1111</td>
<td>000000000000000000000000000001</td>
<td>000000000000000000001</td>
<td>00000000000000000000</td>
</tr>
</tbody>
</table>
comparison of different coding schemes for different source bit lengths, to see where the power reduced per additional redundant bit is maximized.

![Power Consumption Reduction per Additional Redundant Bit](image)

**Figure 6-1. Power Consumption Reduction per Additional Redundant Bit Comparison for Multiple Source Coding Schemes**

From this graph, the maximum power reduction per redundant bit can be found in the delay based scheme with source bit length of 3 bits. Also, it can be seen that for the case of 3 source bits, with a coded maximum of two “high” bits, and the case of 6 source bits with the coding scheme of a maximum of three “high” bits are maximum points for each coding technique.

Another consideration for the source coding scheme, is the multiple accessing scheme required to reduce collisions caused by transmitters attempting to transmit on the same carrier at the same time. A stop and random wait interval scheme, taken from
current RFID technologies, is being considered for the multiple accessing scheme. Every sensor node, will be assigned its own random number. This random number will serve as both the stop and wait interval, as well as the ID of the sensor, which will be appended before or after the data bits of the transmitted message. By setting the stop and wait interval to be the random number times the maximum amount of time required to transmit a single message, this will help reduce the number of transmitters transmitting at the same time, since all stop and wait intervals are randomly generated, and pre-assigned.
CHAPTER 7
FULL SYSTEM INTEGRATION TESTING

To combine all efforts made for the development of a self-powered wireless hydrogen sensor, a full system integration test of the energy harvesting devices, sensor mechanism and sensor interface, microcontroller, RF transmitter, and RF receiver was conducted at the University of Florida on October 20, 2005. The set up of the system integration, including procedure for the introduction of a controlled level of hydrogen into the ambient, is detailed in this chapter. Most figures in this chapter were taken from screen captures of a video made to document this experiment.

Hydrogen Chamber Equipment Setup

A full system integration test was successful in detecting and transmitting via a wireless link, the presence of a controlled amount of hydrogen in the ambient, while obtaining power through scavenged energy using energy harvesting techniques mentioned in previous chapters. This full system integration and test was performed with a hydrogen source provided by the hydrogen chamber in the basement of the New Physics Building. The schematic of the hydrogen chamber is re-drawn in Figure 7.1. The procedure for the injection of 500 PPM of hydrogen is as follows:

- There are two large gas tanks where, one tank is 500 PPM of compressed hydrogen, and other is 99.99% compressed nitrogen. By using different flow rates for hydrogen and nitrogen, different concentrations of hydrogen in nitrogen can be achieved. This is how 10 PPM, 100 PPM, and 200 PPM of hydrogen in nitrogen were tested.

- The furnace seen in the figure is used to run tests at high temperatures. For the purposes of this experiment, the Pt ZnO Nano-Rods remained at room temperature.
The Pt-ZnO Nano-Rod was attached to the glass support beam using two metal leads which can be accessed from outside the hydrogen chamber for connection to the sensor interface via alligator style clips.

The Pt-ZnO is first measured with air in the ambient by the HP4156B Semiconductor Parameter Analyzer.

Sample is loaded into hydrogen chamber, and opening is sealed tight by 3 screws which clamps a rubber o-ring at the base of the glass support beam to the opening of the hydrogen chamber.

Mechanical pump is turned on, and Valve 1 and Valve 2 are opened. Valve 3 is slowly opened to vacuum the chamber until pressure reaches less than 0 atm, or .010 torr.

Valve 3 is closed, and Nitrogen is pumped at maximum flow rate until pressure gauge reaches 760 torr, or 1 atm. Once pressure has reached 1 atm, or 760 torr, cut nitrogen flow

Set hydrogen flow to maximum flow rate, and flow H₂ into system. Open valve 5 so that there can be a constant flow of hydrogen, and pressure within hydrogen chamber remains at 1 atm or 760 torr.

Figure 7-1. Redrawn Schematic with Higher Detail of Hydrogen Chamber Components
Full System Integration Test

As mentioned in the procedure, the ZnO Nano-Rod sensor was connected to the sensor interface via alligator clips. The output of the sensor interface was connected via alligator clips and a BNC cable to the ADC input of the microcontroller. The RF transmitter is connected to the USART output of the microcontroller as seen in Figure 7-2.

Figure 7-2. Microcontroller with RF Transmitter Attached to Microcontroller USART PORT

Power supply nodes were tied together via BNC cables, and were attached to the output of the energy harvesting devices. Additionally, to see that the Pt-ZnO Nano-Rod was indeed changing to the introduction of hydrogen, and to see that the ADC of the microcontroller was capable of detecting these changes, the microcontroller was set to run in level monitoring mode, where there is a constant transmission of data. Since this mode necessitates for a higher power requirement due the increase in duty cycle for data transmission, as well as more “high” bits per message as compared to threshold detection, if the energy scavenging techniques are successful for this mode of operation, then both
modes of operation can operate through the use of energy harvesting. On the central
monitoring / RF receiver side, the receiver was powered by an analog output node of the
USB-6008 DAQ device connected to the USB port of a Hewlett Packard laptop running
LabVIEW 7.1. The Output of the Ming RE-99 Receiver was connected to a Tektronix
TDS210 Two Channel Digital Real Time Oscilloscope to clearly see the data pattern
from the ADC of the microcontroller received by the Ming RE-99 receiver, showing the
Pt-ZnO Nano-Rods reaction to the introduction of 500 PPM of hydrogen into the
ambient. Figure 7-3 shows a data bit-stream pattern received by the Ming RE-99 receiver
during the full system integration test.

![Image of data bit-stream pattern](image)

Figure 7-3. Output Data Bit-Stream Pattern of Received Data From Ming RE-99 During
System Integration Testing

This same setup was used with the solar energy harvesting circuitry, and vibration
harvesting circuitry separately, and was successful in powering the system for each
scenario. The solar and vibration harvesting circuitry were detailed earlier in chapter 2,
and can be seen in the experimental setup in Figures 7-4 and 7-5, respectively.

For the solar harvesting technique, since the basement of the New Physics Building
provides little light in the ambient for solar harvesting, in the case of the solar power
reclamation system, a lamp was used to provide ambient light for the solar cells and
power IC. For the case of the vibration energy reclamation circuit, a shaker tuned to the resonant frequency of the four PSI D220-A4-203YB Double Quick Mounted Y-Pole Bender PZT devices connected to a direct charging circuit, was used as the vibration source for the vibration harvesting circuitry.

Figure 7-4. Solar Cells (left) with Solar Power IC (right)

Figure 7-5. Vibration Energy Harvesting Components (a) PSI PZT Beams (b) Direct Charging Circuit

A block diagram of the system level integration testing with corresponding screen caps from the video made to document this test can be found in Figure 7-6.
Figure 7-6. System Integration Testing Block Diagram with Screen Caps From Video
CHAPTER 8
CONCLUSION AND FUTURE WORK

Conclusion

In conclusion, a fully self-powered wireless hydrogen sensor was successfully designed and tested at the University of Florida. Three contributions made to this project, and detailed in this thesis, are the design of a low-power sensor interface for the Pt-ZnO Nano-Rod hydrogen sensing mechanism, the design and optimization of the RF communications platform for the Self-Powered Wireless Hydrogen Sensor and Central Monitoring Station, and a proposed minimum redundancy, minimum energy coding scheme to both provide multiple accessing schemes for future deployment of a network of self-powered wireless hydrogen sensors, and to minimize energy without expense to transmitted message length to lower possibility of data transmission collisions, and lower required energy to transmit a given message.

For the case of the differential detection circuit, as a comparison to current technologies, is MAXIM-IC’s new for 2006, MAX4208/4209 instrumentation amplifier package claiming to be the world’s best instrumentation amplifier. A comparison of this amplifier to the instrumentation amplifier of the differential detection circuit designed for this project can be seen in Table 8.1.

Table 8-1. Comparison of Performance Between Designed Differential Detection Circuit, and Other Commercially Available Designs.

<table>
<thead>
<tr>
<th>PART</th>
<th>$V_{\text{MIN}}$</th>
<th>$V_{\text{MAX}}$</th>
<th>$I_{\text{SUPPLY}}$</th>
<th>$V_{\text{OS}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>This design</td>
<td>1 V</td>
<td>5.5 V</td>
<td>27uA</td>
<td>30uV</td>
</tr>
<tr>
<td>MAX4208/4209</td>
<td>2.85 V</td>
<td>5.5 V</td>
<td>750uA</td>
<td>15uV</td>
</tr>
</tbody>
</table>
It is seen that for almost similar performance, the MAX4208 and MAX4209 require a higher supply voltage, and much higher supply current to power the instrumentation amplifier.

**Future Work**

Although a fully self-powered wireless hydrogen sensor was successfully designed and tested, the work doesn’t end there. Currently, a single module, integrating all components (energy harvesting, microcontroller, sensing mechanism and sensor interface, and RF transmitter) of the original self-powered wireless hydrogen sensor, is currently being designed and tested. Additionally, this new module will contain the minimum redundancy, minimum energy coding scheme and multiple accessing schemes as detailed in a previous chapter, so that multiple sensor nodes can be realized.

Also, rather than using the Linx technologies ANT-315-SP surface mount SPLATCH antenna, and a separate MING Tx-99 transmitter board, an onboard microstrip antenna is being currently designed, and a transmitter similar to the MING TX-99 has already been designed and integrated onto the single board module. The PCB trace of the newly designed single module board can be seen in Figure 8-1, with the assembled board shown in Figure 8-2. The assembled board seen in Figure 8-2, has a SMA connector, which will eventually be replaced by a LINX low profile SPLATCH antenna, or the antenna currently being designed.

Additionally, new operational amplifiers were made available after the full system integration test, which require even lower power without cost to performance. These operational amplifiers are currently being tested and fitted into the differential detection circuit design to replace the MAX4289. One of these operational amplifiers, the
MAX4039 operational amplifier outputs a buffered 1.232 V reference voltage, which can be used to power the resistive bridge, and isolate the bridge from the supply voltage, making it even more insusceptible to any voltage shifts of the supply power. The MAX4039 itself, also only requires a minimum of 1.8 supply voltage, and 2uA supply current to achieve an internal 200uV input offset voltage.

Another operational amplifier currently being considered is the MAX991X line. These operational amplifiers also exhibit similar performance of input offset voltages on the order of 200uV, with a low bias requirement of 1.8 V, and 4uA supply per amplifier. Additionally, these operational amplifiers have an available shutdown/ idle state which drains a mere 1 nA during shutdown state. With this amplifier, even more power can be saved by strobing this amplifier so that it will only be turned on when the collection of data by the ADC is needed. For the receiver side, LabVIEW code is also currently being re-coded for the replacement of the Ming RE-99 with the Linx Technologies RXM-315-LR. Eventually, another full system demonstration, including MEMS scale PZT vibration energy harvesting devices currently being fabricated, will need to be done.

Figure 8-1. Protel PCB Top and Bottom Layout For Fully Integrated Board
Figure 8-2. Fully Assembled Single Module
LIST OF REFERENCES


BIOGRAPHICAL SKETCH

Jerry Chun-Pai Jun finished his undergraduate coursework in electrical engineering at the University of Florida, and entered graduate school in electrical engineering at the University of Florida in Fall of 2004. Jerry is interested in the design and applications of wireless sensors, specifically in the field of RFID systems, implantable bio-sensors, and ambient monitoring sensors. As a master’s graduate student, Jerry worked under Dr. Jenshan Lin as a research assistant in the RF System on Chip (RFSOC) laboratory at the University of Florida. After graduation, Jerry will be working at Motorola in Plantation, FL, and hopes to continue his career in RF/microwave technologies.