Low Power RF Transceiver Modeling and Design for Wireless Microsensor Networks

by

Andrew Yu Wang

Bachelor of Science in Electrical Engineering
University of Maryland College Park (1998)

Master of Science in Electrical Engineering and Computer Science
Massachusetts Institute of Technology (2000)

Submitted to the Department of Electrical Engineering and Computer Science in partial fulfillment of the requirements for the degree of

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Abstract

The design of wireless microsensor systems has gained increasing importance for a variety of civil and military applications. With the objective of providing short-range connectivity with significant fault tolerance, these systems find usage in such diverse areas as environmental monitoring, industrial process automation, and field surveillance. The main design objective is to maximize the battery life of the sensor nodes while ensuring reliable operations. To achieve this goal, the microsensor node has to be designed in a highly integrated fashion and optimized across all levels of system abstraction.

For microsensor networks, the RF transceiver dominates the power consumption. The concept of transceiver power efficiency is introduced, which defines the ratio of RF transmit power to transceiver electronics power, to show that short-range RF transceivers have low transceiver power efficiency. A system energy model is developed to show that the battery life of the transceiver not only depends on its power consumption, but more importantly, on its energy dissipation over the operation cycle. Both the transceiver power efficiency and the battery life can be improved significantly by increasing the data rate, reducing the start-up time, and improving the PA efficiency. Increasing the data rate drives down the fixed energy cost of the transceiver. Reducing the start-up time decreases the start-up energy overhead. Improving the PA efficiency lowers the energy per bit cost of the power amplifier.

The voltage controlled oscillator occupies a large fraction of the total energy budget in the operation of the microsensor transceiver. The design of integrated LC oscillators is investigated on both the system and the circuit design levels. On the system level, the phase noise requirement of the VCO as a function of the channel bandwidth and the data rate is derived. On the circuit level, the physical mechanisms of phase noise are examined and a low-power 5-GHz VCO is designed and fabricated in a 0.18-μm SiGe BiCMOS process. A technique is proposed to trade off phase noise for a lower bias current through the sizing of the switching transistors. This VCO demonstrates a phase noise of -125dBc/Hz at 7mA core bias current and -110dBc/Hz at 1.5mA, which exceeds the system phase noise requirement.

Thesis Supervisor: Charles G. Sodini
Title: Professor of Electrical Engineering and Computer Science
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Chapter 1

Introduction

The wireless communications market has experienced an explosive growth in the past decade. This rapid growth in the commercial market has generated a tremendous amount of research interest in radio frequency (RF) technology. In particular, as portable battery-powered devices become more ubiquitous, there is an ever increasing demand in low power and low cost design methodologies.

The ultra low power radio project at the Massachusetts Institute of Technology is a collaborative research effort whose goal is to investigate and develop novel system architectures and circuit techniques for short-range wireless microsensor systems. This research focuses on the design and implementation of low power RF transceiver front-end suitable for microsensor applications. The emphasis of the research is on the joint optimization of physical-layer system and circuit designs to maximize the battery life of the transceiver node.

1.1 RF Systems Landscape

Figure 1-1 shows the current RF systems landscape with application areas distinguished by distance and data rate requirements. From the distance perspective, RF systems are divided into short-range, local area, and wide area networks. From the data rate perspective, sensing and monitoring applications demand the least amount of data rate, voice networks occupy the middle range, and real-time multi-media applications require the highest data rate.

There is an inherent trade-off between data rate and transmission distance for RF sys-
tems because the amount of RF transmit power required scales with both data rate and transmission distance. Therefore, for limited RF transmit power due to either regulation or technology constraint, increasing the data rate capability of a system usually implies reduced transmission distance. The power consumption stays high to support either the high data rate or the long range.

Although sensor applications do not necessarily need to be limited to short transmission range and low raw data rate, as the sensing and the communication are independent from each other, we focus on short-range and low data rate sensor applications only. In this case, power consumption of these systems can be made potentially very small.

1.2 Research Background

Wireless microsensor networks have become an active research area at both the system and the circuit levels over the past 5-10 years. A wireless microsensor network distinguishes itself from a classic data network in two main ways. First, due to its mobile characteristic, a sensor’s position can change with regard to its neighbors, thus creating a spatial and temporal varying communications channel. This demands new techniques in looking at the capacity and control flow of such networks. Some seminal works in this area include Gupta [1] and Grossglauser [2], who established information theoretic bounds on the capacity of ad-
hoc networks. Techniques to improve capacity gains include using spreading and multi-user detection [3] and allocating transmit power based on channel conditions [4].

A second major characteristic that distinguishes wireless sensor networks from traditional data networks is the energy-constrained nature of the sensor nodes. Thus the complexity of the problem comes not only from finding the best routing and flow control algorithms, but also from contending with node failures due to deeply faded channels as well as limited energy resources [5]. From the system side, techniques to increase sensor network life are generally based on the principle that all nodes in the network should be regulated to maintain the same energy reserve [6]. One technique to accomplish this is by time-sharing the processing functions among all the sensors in the network [7].

On the circuit side, the power consumption of the transceiver depends heavily on both technology and design. For instance, Off-chip passive components can have quality factors several orders of magnitude better than their on-chip counterparts. Using off-chip inductors and capacitors, Rofougaran was able to reduce power consumption of a prototype 450-MHz front-end to below 1mW [8]. Similarly, Darabi implemented a 900-MHz pager receiver that consumes only 4.5mW [9]. However, off-chip components not only makes integration difficult, but may also increase cost and size. Thus integrated solutions are still favored today, and much research effort focuses on on-chip solutions [10]. Advances have been significant in this area. The power consumption of integrated Bluetooth transceivers have been brought down from 100’s of milli-Watts to 10’s of milli-Watts [11]. Transceivers designed for Zigbee, which is a new standard optimized for ad-hoc networks, have also been reported with power consumption figures in the same range [12].

To study the combined effect of system-level and circuit-level issues, several research groups have built actual microsensor networks. The μAMPS (Micro-Adaptive Multi-domain Power-aware Sensors) project at MIT focuses on developing a power-aware system that scales processing power based on demand [13]. The WINS (Wireless Integrated Network Sensors) project from UCLA focuses on the implementation of the RF front-end and achieves low power consumption using high-Q off-chip and MEMS devices [14]. The Berkeley SmartDust project implements an autonomous sensor that focuses on sensing functions such as temperature, humidity, and pressure [15]. Both off-the-shelf RF and optical front-ends have been experimented on the SmartDust nodes. The PicoRadio project at Berkeley aims to build an integrated sensor node that includes low power RF front-end as well as reconfig-
urable baseband microprocessor [16].

1.3 Microsensor Networks

Wireless microsensor networks can provide short-range connectivity with significant fault tolerances. These systems find usage in diverse areas such as environmental monitoring, industrial process automation, and field surveillance. Sensor networks fall into one of two general categories: self-assembled ad-hoc networks or centralized networks with base stations. This work focuses on a centralized network architecture which is common in an industrial environment. As shown in Figure 1-2, such a system is composed of numerous energy-constrained sensor nodes and a high-powered base station. The sensors collect data and send them to the base station for processing. More cells can be added to cover larger areas.

![Diagram of a centralized wireless microsensor network.](Image)

Figure 1-2: A centralized wireless microsensor network.

The wireless microsensor system is an emerging market technology that is quite distinctive from both conventional voice and data applications. The following section discusses its unique features and how they affect design choices.

- **High cell density** – A wireless sensor network contains as many as several thousand sensor nodes within a small area. Thus, they provide both extensive spatial coverage and significant fault tolerance. However, this imposes a challenge in the design of energy and bandwidth efficient multi-access schemes.

- **Ad-hoc distribution** – Spatial distribution is ad-hoc, and each sensor may have a
very different transmit path. This means some sensors could have line-of-sight (LOS) transmission while others might be totally obstructed from the base station. This not only creates difficulty in estimating the transmit power but also increases the dynamic range of the received signal.

- **Ease of deployment** – Sensor nodes should require minimal installation and virtually no maintenance. This implies that the protocols have to be simple as well as highly reconfigurable.

- **Low mobility** – Sensors are confined to a small area, so they are either static or are restricted in mobility. This means that a slow fading environment with low Doppler spread is expected.

- **Low data rate** – The data rate is typically as low as a few kilo-bits per second, and each data packet may contain only up to one hundred bits. This favors a duty-cycled bursty transmission scheme where the transmitter is turned off most of the time.

- **Low latency** – Packets are required to arrive at the base station within a small time delay. This puts a restriction on the maximum delay of the bursty transmission scheme. In addition, error correction protocols that require retransmission are clearly unfavored since they will increase delay.

- **Short transmission distance** – Typical transmission distance is less than ten meters. The transmit energy is small enough that the sensor node electronics become the dominant source of energy consumption. This characteristic plays a key role in our design approach.

- **Asymmetric data link** – The data link constitutes energy-constrained microsensor nodes and high-energy base stations. The base stations have high performance transceivers that can help to reduce the system complexity of the microsensor node transceivers.

- **Volume constraint** – The sensor is required to be compact, which imposes limits on both the amount of available energy source and on the complexity of the microsensor node.
The ultimate goal of the low power radio project is to maximize the battery life of the sensor nodes while complying with all the other requirements stated above. Sensor transmitter power consumption is the bottle-neck since the system lasts only as long as the sensors do. Table 1.1 shows specifications for a system that monitors machine operations in a factory environment [17]. This system is chosen as a design example because it presents some very interesting design challenges and trade-offs. In particular, the battery life span of great than one year is a very difficult requirement.

<table>
<thead>
<tr>
<th>Cell density</th>
<th>200 - 300 in 5mx5m area</th>
</tr>
</thead>
<tbody>
<tr>
<td>Range of link</td>
<td>&lt; 10m</td>
</tr>
<tr>
<td>Message rate (msg = 2bytes)</td>
<td>average: 20 msgs/sec maximum: 100 msgs/sec minimum: 2 msgs/sec</td>
</tr>
<tr>
<td>Error rate and latency</td>
<td>$10^{-3}$ after 5ms $10^{-6}$ after 10ms $10^{-9}$ after 15ms</td>
</tr>
<tr>
<td>Battery life size</td>
<td>&gt;1 years one coin-sized battery</td>
</tr>
</tbody>
</table>

Table 1.1: Wireless microsensor system specification for machine monitoring applications.

1.4 Research Scope and Contributions

As mentioned in the previous section, there is fruitful research taking place at both the system and the circuit levels. However, systems research and circuits research have been traditionally two independent entities, and there has been little effort that attempt to bridge the two together. On the system level, the lack of a good transceiver energy model has made modeling and prediction of network life-time difficult and often inaccurate. Therefore, a good energy model based on the physical electronics of the transceiver is critically important in producing sensible conclusions. Similarly, without a good understanding of the system trade-offs, circuit designers tend to over-design and subsequently lead to sub-optimal results.

In this research, the focus is on the joint optimization of system-level and circuit-level design issues in the physical layer. Specific research goals and contributions are:

- Develop a system energy model that takes into account the transceiver electronics characteristics.
- Determine system parameters that will meet performance specifications and minimize transceiver power consumption.

- Design critical RF front-end circuitry to determine necessary circuit performance versus power consumption trade-offs to be applied in the system model.

Figure 1-3 illustrates the research methodology undertaken in this project. A specific microsensor application, as the one depicted in Table 1.1, dictates a bit error rate (BER) and latency requirement. The goal of this research is to design a microsensor node transceiver that satisfies the BER and latency requirement and achieves the longest battery life. As will be discussed in Chapter 4, this entails not only the design of low power transceiver circuits, but more importantly, the joint optimization of both circuit and system. The design approach requires an understanding of the system block to determine a set of performance specifications. Similarly, it also requires an understanding of the trade-offs between the performance specifications with circuit power consumption and transceiver architecture. The performance specifications need then to be iteratively refined until the solution offers the desired long battery life.

![Diagram](image1)

**Figure 1-3:** Top-level system design approach.

### 1.5 Thesis Outline

The outline of this thesis is as follows. Chapter 2 presents an overview of the properties of a microsensor network. This includes the architecture of the microsensor node as well
as the microsensor transceiver operation. Chapter 3 presents the communications theory background, including detection theory in both Gaussian and fading channels. Chapter 4 incorporates the power consumption of the transceiver electronics into the system design. A transceiver energy model is developed and the concept of transceiver power efficiency is introduced. This chapter shows how the transceiver energy consumption can be minimized by combining system techniques and circuit design. Chapter 5 presents work on the design of RF building blocks. In particular, VCO power consumption versus phase noise performance is analyzed. Chapter 6 concludes the thesis work with a summary of results and provides future directions.
Chapter 2

Industrial Microsensor Networks

Microsensor networks can be used in a variety of applications. Depending on the particular application, the sensors can operate in very different ways. For example, some applications may require the sensors to operate at much higher duty cycle than others. Some applications may demand higher transmit power due to longer distance and/or more hostile environment. This means that the battery life can differ significantly from application to application. Although the energy optimization techniques derived in this thesis are general, a set of specific operating conditions will yield a different battery life. Therefore, it is important to understand the particular application that the microsensor network is used. This chapter describes the characteristics of industrial microsensor networks and their impact upon the operating conditions of the microsensors.

2.1 Microsensor Node Architecture

Industrial microsensor networks are used for sensing and machine monitoring purposes. Wireless transceivers are used to provide the communications link between the sensor nodes and the base stations. We confine the problem to a centralized wireless network where base stations provide timing and routing control to all the sensor nodes. There are several practical reasons why an ad-hoc network architecture is not used. In a factory environment, it is possible to install a number of base stations. Networking protocols, such as those to control contention and delay, are simpler to implement for a centralized network than for an ad-hoc network. This enables the research focus to be entirely placed on the design of the sensor node transceiver physical layer. In addition, power consumption of the microsensor
nodes can be reduced by taking advantage of the high-performance base stations.

Figure 2-1 shows the architecture of the sensor node. The sensor block converts physical data into digital bits. A conventional transducer or a MEMs based sensor is at the core of the sensor block. The type of sensors available today include thermal, pressure, temperature, magnetic, radiation, and many more [18]. These sensors can have very low power consumption. For example, MEMs based temperature and pressure sensors have power consumption on the order of a few to a few hundred micro-watts [19,20].

![Figure 2-1: Microsensor node architecture.](image)

The DSP block performs essential signal processing functions, which include data analysis, compression, and baseband modem operations. If a general processor is used to implement this block, then the power consumption can be prohibitively high. An Intel StrongARM micro-controller unit consumes 400mW running at 206MHz [21]. Fortunately, the power consumption can be scaled down significantly for low speed ASIC solutions. For example, a digital processor designed for microsensor applications consumes 0.5mW at 4.4MHz [12]. Since this work focuses on microsensor applications with low raw data rate and minimum signal processing requirement, the power consumption of the DSP block can be assumed to be small.

The RF transceiver block includes both a transmitter and a receiver — the transmitter sends data to the base station, and the receiver receives commands from the base station to control the sensor node operations. Within the transceiver, the TX block modulates and up-converts the digital data to the RF carrier frequency. The PA block provides the required RF output power to the antenna. The RX block performs down-conversion and demodulation. The LO block provides a stable reference frequency for both the transmit and the receive paths.

RF transceivers operating at Giga-Hertz frequencies consume 10's to 100's of milli-Watts [11,12]. This is more than an order of magnitude greater than that of either the
sensor block or the DSP block. What this means is that the energy cost of the microsensor node is dominated by the RF transceiver, or equivalently, by the communications cost. For this reason, the focus of this work is on the design of the RF transceiver. The energy cost of the sensor and the DSP are assumed to be negligible. This assumption applies to industrial applications where the primary function of the sensor node is to send short bursts of messages periodically.

2.2 Data Rate and Latency

As illustrated in Table 1.1 in Chapter 1, the maximum message rate is 100 messages per second, with each message occupying 2-Bytes. This leads to a maximum data rate of 1.6kbps. Since the RF transceiver dominates the energy cost, it makes sense to buffer this data and transmit it in burst mode. The transceiver should rest in the off state for as long as possible to conserve battery life.

A special characteristic of an industrial microsensor system is its stringent latency requirement. The data from the sensor is required to arrive at the base station within a short amount of time to enable real-time data management. This latency is determined by how fast the base station needs to react to the sensor data message. In addition, the sensor is required to send an “alive” message periodically even when there is no data. This is to ensure the base station knows that the sensor is operational.

The industrial application being studied in this work has a 5ms latency requirement. If there is data to be sent, it needs to arrive at the base station in 5ms; if there is no data, then the sensor needs to send an “alive” message. Therefore, the sensor transceiver operates in burst mode every 5ms as shown in Figure 2-2. The duty cycle of the transceiver is

\[ D = \frac{t_{op}}{T_{lat}} \]  

(2.1)

The operation time of the transceiver, \( t_{op} \), depends on the data rate of the transceiver as well as on the size of the data packets sent and received in each operation cycle. The transceiver data rate is the rate at which the transceiver operates, which can be different from that of the raw data rate generated by the sensor. The transceiver data rate must be higher than the raw data rate for \( D < 1 \).

The data packet size is different from the message size. For short messages, the packet
size is dominated by the additional header bits, which are used to perform functions such as synchronization and error correction coding. For the industrial microsensor application considered in this work, the message size is 2-Bytes and the data packet size is 100-bits. Since the packets are short and infrequent, we want burst transmissions.

2.3 Multi-access Protocol

The type of multi-access protocol influences the battery life of the transceiver node. In this work, a simple hybrid TDM-FDM multi-access technique is proposed [22]. A contention-based MAC is avoided since it cannot guarantee the latency requirement due to packet collisions and retransmissions. In addition, the efficiency of such a scheme decreases when there are a large number of sensors transmitting frequently [23].

The TDM-FDM technique is shown in Figure 2-3. Each sensor node transmits every 5ms with operation time $t_{op}$. There is a guard time $t_{guard}$ between the transmission of adjacent time slots. Therefore, the number of time slots available are

$$N = \frac{T_{lat}}{T_{op} + T_{guard}} \quad (2.2)$$

For a $T_{op}$ of 100µs and a $T_{guard}$ of 20µs, $N$ is 41. In other words, 41 sensors can transmit within the same frequency channel without collisions. To accommodate more sensors in the same cell, multiple frequency channels must be used. Therefore, the demand for bandwidth is large for densely populated sensor networks.

This TDM-FDM technique is chosen for its effectiveness and simplicity. Since this work focuses on the design of the physical layer, the MAC layer is not investigated in detail. The TDM-FDM technique can be implemented in ways that improve its robustness without affecting the operation of the physical layer. For example, the base station can select the
frequency channels based on the quality of these channels, or the base station can hop among the available channels to improve diversity. The physical layer will operate in the same way independent of these enhancements.

2.4 Carrier Frequency

The carrier frequency affects both the power consumption of the transceiver electronics and the range of RF transmission. Low carrier frequency enables low circuit power consumption [24] and long range [25]. However, the amount of spectrum at low frequency is scarce, which can be a bottle-neck for densely populated sensor networks. In addition, low carrier frequency is not amenable to integration since off-chip inductors and capacitors are required. For these reasons, more and more short-range transceivers are implemented in the 2GHz and 5GHz unlicensed bands. This work focuses on the design of short-range transceivers in the UNII band, which occupies 300MHz unlicensed spectrum at 5GHz [26].

2.5 Battery Capacity

Modeling battery capacity is difficult because the actual battery capacity depends on the specific way the battery is discharged. In this work, we assume a simple linear battery capacity model where the battery is treated as a linear storage of current. If a battery has capacity $C$, which is expressed in milli-Amp-hours (mAh), then the battery life is

$$T_{batt} = \frac{C}{I}$$  \hspace{1cm} (2.3)
where $\bar{I}$ is the average current throughout the life time of the battery. When the transceiver is operated with a duty cycle $D$, the battery life can be conveniently expressed as

$$T_{\text{batt}} = \frac{C}{I_{op} \cdot D}$$

(2.4)

where $I_{op}$ is the average current consumption in one operation cycle. In this work, $C$ is assumed to be 1000mAh, which is the nominal capacity of a standard CR2477 lithium coin battery [27].

Figure 2-4 shows the battery life requirement as a function of $I_{op}$ and duty cycle. To achieve a one-year battery life at 1% duty cycle, for example, the average current must be kept under 11mA, which translates to approximately 20mW at 1.8V supply. With a 5ms latency, a 1% duty cycle implies that the transceiver is only allowed to operate for 50$\mu$s every cycle. If the transceiver transmits and receives 100-bit packets during each operation cycle at a data rate of 1Mbps, then the transceiver operates for at least 200$\mu$s, which increases the duty cycle to 4%. This places very difficult requirements on the transceiver power consumption.

![Figure 2-4: Limit on the average transceiver current consumption per operation cycle.](image)

The linear battery model is a very simplified battery capacity model. The actual battery life depends on both the discharge rate and the battery relaxation effect [28]. Battery capacity decreases when the battery is discharged at a high rate, which is above a couple of
milli-Amperes for a lithium coin battery [27]. Since an RF transceiver draws significantly more current than that, the actual battery life will be degraded. However, when a transceiver operates in burst mode, the battery will recover some of the capacity lost at high discharge rate during the time the transceiver is off. This is called the relaxation effect. Therefore, the actual battery life depends on both the peak current drawn from the battery and the pulse rate of the burst-mode transceiver operation. Due to the complexity in modeling these effects, which often can only be captured through measurement, the linear battery model is used to give a first order estimate of the battery life. Although this model is optimistic, it is accurate in comparing the battery life of different transceivers provided that peak current drawn and pulse rate are similar for these transceivers.

2.6 Summary

This chapter describes the operating characteristics of industrial microsensor networks including sensor node architecture, data rate and latency requirements, multi-access protocols, and the choice and impact of RF carrier frequency. The design of the microsensor node transceiver needs to take advantage of these characteristics to reduce the high energy cost of the communications link. In order to extend the sensor node battery life beyond one-year, the supply current drawn by the sensor needs to be kept below 11mA if the sensor operates at a duty cycle of 1%. This is a difficult target for Giga-Hertz RF transceivers.
Chapter 3

Detection in AWGN and Rayleigh Channels

Modern communication systems use digital modulation techniques, which have many advantages over their analog counter-parts [25]. Some of these advantages include increased channel capacity, greater noise immunity, and robustness against channel impairment. This chapter provides the background information on detection theory in Gaussian and fading channels. Modulation techniques are discussed and their advantages are compared in terms of modulation power efficiency and bandwidth efficiency. Link budget is analyzed in the context of the microsensor network and the required microsensor RF transmit power is determined.

3.1 Uplink and Downlink

The communications link between the sensor node transceiver and the base station transceiver is shown in Figure 3-1. In an ideal communications system, the important blocks are the modulator (Mod) and the demodulator (Demod). The signal to noise ratio per bit, $E_b/N_0$, at the demodulator front-end determines the bit error rate (BER). An ideal radio performs linear amplification and frequency translation functions and does not affect the SNR of the modulated signal. Therefore, from a pure communications perspective, the radio is transparent to the system designer. In reality, the noise and nonlinearity of the radio front-end degrades the SNR at the demodulator and increases the bit error rate of the received data.
The uplink consists of the sensor node transmitter and the base station receiver, so the design specification of the sensor node transmitter is influenced by the quality of the base station receiver. Due to the asymmetric nature of the link, the sensor node transmitter should be simple and low power, while the base station receiver can afford the complexity and should be designed to achieve the best sensitivity and linearity to relax the design of the sensor node transmitter. Similarly, the downlink consists of the base station transmitter and the sensor node receiver, and the base station transmitter should be designed to accommodate as much non-ideality as possible in the sensor node receiver.

The duplexer (DUP) block on the base station transceiver enables the base station to operate in full duplex mode. The sensor node transceiver operates in half-duplex mode only.

### 3.2 Detection in AWGN Channel

Figure 3-2 illustrates the concept of the AWGN channel. The signal $s(t)$ is the continuous-time modulated waveform, and $r(t)$ is the output signal distorted by the channel noise. In this model, the channel response is assumed to be flat, i.e., no distortion, and the only noise present is the thermal noise $n(t)$ generated by the receiver antenna. In many applications, such as deep space communications, where thermal noise is the dominate source of noise, the AWGN channel model is extremely accurate.

The thermal noise has a flat power spectrum density (PSD) up to 100GHz, as shown in Figure 3-3. Its one-sided PSD, $N_0$, is defined as the noise power transferred into a matched
Figure 3-2: The Additive White Gaussian Noise (AWGN) channel.

load per hertz, and is given by:

\[ N_0 = kT \]  

(3.1)

where \( k \) is the Boltzmann's constant and \( T \) is the absolute temperature in Kelvin. At a noise temperature of 300K, which is typical for receivers in the Giga-Hertz range, \( N_0 \) is approximately equal to -174dBm/Hz.

Figure 3-3: Autocorrelation function and power spectrum density of white noise.

3.2.1 Optimal and Suboptimal Detection

Given that the input comes from a set of pre-defined waveforms \( \{s_k(t), 0 \leq k \leq M - 1\} \), the optimal receivers are shown in Figures 3-4 and 3-5. The received signal, \( r(t) \), is matched (or correlated) to each of the possible input waveform \( s_k(t) \), and the branch that produces the highest SNR is chosen as the most likely input. This type of receiver is called a maximum likelihood detector. The difference between the two receivers is that the matched filter receiver is linear while the correlator receiver is non-linear due to the multiplier.

Matched filter and correlator receivers require exact phase synchronization at the carrier frequency. Consider a passband input signal, \( r(t) \), written as the following,

\[ r(t) = s(t)e^{j\omega t} \]  

(3.2)
where $s(t)$ is the complex baseband signal, and $e^{j\omega_c t}$ is the carrier frequency (i.e., both I and Q carriers).

In order to recover $s(t)$, an exact copy of the carrier signal, $e^{j\omega_c t}$, needs to be produced at the receiver. Since the receiver does not know the exact phase of the transmitted carrier, it must be able to track the received carrier. This is called carrier synchronization, or carrier recovery, which requires the use of a phase-locked loop (PLL).

Due to the limited bandwidth and the non-idealities of a PLL, carrier tracking can be difficult in an environment where the phase of the received carrier varies rapidly. For instance, in a fading channel where a random phase is introduced by multipath fading, the carrier recovery loop must be fast enough to track this phase error. In addition, phase and fre-
frequency errors at the transmitter frequency synthesizer cause instability in the carrier phase, which can potentially cause the carrier recovery loop to false-lock. Due to these problems, sub-optimal detection techniques are often used in practice to avoid carrier synchronization. These techniques belong to a general category called noncoherent detection.

Figure 3-6 shows a generalized M-ary noncoherent receiver. The received signal first goes through either a correlator or a matched filter. Since the receiver carrier is not synchronized to the transmitted carrier, a phase error $e^{j\theta}$ is produced. The correlator output goes through a complex magnitude block which eliminates the phase error. However, any phase information in the input signal $s_k(t)$ is also lost. This means that any modulation scheme that relies on carrying information in the phase component, such as Phase Shift Keying (PSK) or Quadrature Amplitude Modulation (QAM), can not be detected. In addition, the performance of noncoherent detection will not be as good as coherent detection since the phase information in the input signal is ignored in the detection process (i.e., noncoherent receiver only does partial detection). Performance for noncoherently detected On-Off Keying (OOK) and Frequency Shift Keying (FSK) signals are discussed in the next section.

It is important to keep in mind that the signal sent by the sensor node is detected at the base station. The additional complexity added by coherent detection is not an issue at the base station receiver. This is not true for the downlink, since the detection occurs at the sensor node, so noncoherent detection should be considered. However, from the power consumption point of view, complexity doesn’t necessarily imply high power consumption.
For example, the carrier recovery circuit operates at base band, so the power consumption can be low. The complexity versus power consumption trade-off is dependent upon the particular circuit.

### 3.3 Classes of Modulation

This section examines several general classes of modulation and shows the trade-offs among them. It will be shown that for each modulation the BER is determined by the $E_b/N_0$. Each modulation technique differs in its power efficiency, which is the $E_b/N_0$ requirement at a particular BER, and bandwidth efficiency, which is the data rate supported per unity bandwidth.

#### 3.3.1 On-Off Keying

On-off keying is the simplest binary modulation system. Its signal waveforms are of the form

$$
\begin{align*}
  s_0(t) &= \sqrt{\frac{2E_b}{T}} \cos(\omega_c t), \quad 0 < t < T \\
  s_1(t) &= 0 \quad 0 < t < T
\end{align*}
$$

where $E_b$ is the energy per bit, $T$ is the symbol period, and $\omega_c$ is the carrier frequency.

### Coherent Detection

Figure 3-7 shows the signal constellation for OOK. The variable $d_{\min}$ is the minimum distance between the constellation points and is a function of the signal amplitude. In order for the average bit energy to be $E_b$, the two constellation points need to be $(0,0)$, and $(\sqrt{2A}, 0)$, where $A = \sqrt{E_b}$. The bit error rate is

$$
P_b(E) = Q\left(\frac{d_{\min}}{2\sigma}\right) = Q\left(\frac{\sqrt{2A}}{2\sqrt{\frac{N_0}{2}}}\right) = Q\left(\sqrt{\frac{E_b}{N_0}}\right) \quad (3.4)
$$

where the function $Q(x)$ is the tail probability of a normal Gaussian distribution,

$$
Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt \quad (3.5)
$$
Noncoherent Detection

Noncoherent detection of OOK can be performed through an envelope detector as shown in Figure 3-8.

\[
\text{Pe} = \frac{1}{2} e^{-\frac{E_b}{2N_0}}
\]  

Despite its simplicity, we do not consider OOK in our system. The amplitude of a signal is typically corrupted more severely than either the frequency or the phase by man-made noise and by multipath fading, which can result in significant BER degradation. For this reason, most communication systems today rely on PSK, QAM, or FSK.

3.3.2 Phase Shift Keying

Phase shift keying is one of the most popular modulation schemes used in modern communication systems. Its signal waveform is given by
\[ s_k(t) = \sqrt{\frac{2E_b}{T}} \cos \left[ \omega_c t + \frac{2\pi k}{M} \right], 1 \leq k \leq M - 1 \]  

(3.7)

where \( M \) is the number of input symbols. The signal constellations for 2-PSK (BPSK), 4-PSK (QPSK), and 8-PSK are shown in Figure 3-9.

![Signal constellations of BPSK, QPSK, and 8-PSK.](image)

Figure 3-9: Signal constellations of BPSK, QPSK, and 8-PSK.

The probability of error computation is straightforward and is given as

\[
P_b(E) \approx \begin{cases} 
Q \left( \sqrt{\frac{2E_b}{N_0}} \right), & M = 2 \\
\frac{2}{r} Q \left( \sqrt{\frac{2rE_b}{N_0}} \sin \frac{\pi}{M} \right), & M \neq 2 
\end{cases}
\]  

(3.8)

where \( r \) is the number of bits per symbol: \( r = \log_2(M) \). In the case of BPSK \((M = 2)\), the equation is different because there is only one nearest neighbor as opposed to 2 for every other value of \( M \).

PSK is a bandwidth efficient modulation scheme because the bandwidth required does not increase with \( M \). Assuming an ideal brick-wall shaping filter, the bandwidth is \( 1/T \). Since the symbol rate is also \( 1/T \), the bandwidth efficiency is defined as the data rate \( R \), over the bandwidth \( W \), and has the unit of bits/s/Hz,

\[
\frac{R}{W} = \frac{\text{symbol rate}}{\text{bandwidth}} = \frac{\text{bits}}{\text{symbol}} = r
\]  

(3.9)

so the bandwidth efficiency is simply the number of bits per symbol. Therefore, it can be improved by using higher level modulations.

The cost of improving bandwidth efficiency is a reduction in power efficiency. For higher level PSK modulation, it takes larger \( E_b/N_0 \) to achieve the same BER. In fact, as the bandwidth efficiency improves linearly, \( E_b/N_0 \) rises exponentially to produce impractical transmit power requirement as \( M \) becomes large.
3.3.3 Quadrature Amplitude Modulation

Quadrature Amplitude Modulation is similar to Phase Shift Keying except that information is encoded in both phase and amplitude, as illustrated in Figure 3-10. Since QAM constellations use space more efficiently than PSK, they require less power to achieve the same BER. Thus for large $M$ ($M > 16$), QAM is usually used in place of PSK. The problem with QAM is that automatic gain control must always be employed to reduce I/Q mismatch. This can be difficult if the signal amplitude fluctuates due to channel impairments.

![Figure 3-10: M-QAM Constellation for $M = 4, 16, 64$.](image)

The probability of error for QAM is [29]

$$P_b(E) \approx \frac{4}{r} Q \left( \sqrt{\frac{3rE_b}{(M-1)N_0}} \right)$$

(3.10)

3.3.4 I/Q Modulation

PSK and QAM belong to a general category of modulation techniques called I/Q modulation, where the signal constellation can be represented by an in-phase component $I$

$$I = A_I \cos (\omega_c t)$$

(3.11)

and a quadrature component $Q$

$$Q = A_Q \sin (\omega_c t)$$

(3.12)

where $A_I$ and $A_Q$ are the in-phase and quadrature amplitudes.

I/Q transmitter and receiver architectures are shown in Figure 3-11. In the transmitter, the $I$ and $Q$ components are generated within the modulator (Mod). They go through the digital-to-analog converter (DAC), are up-converted to the carrier frequency by the RF mixers, and are combined and sent out via the power amplifier (PA). In the receiver, the
received signal is first amplified by a low-noise amplifier (LNA) and is down-converted by the mixers into $I$ and $Q$ components. The IF signal is then digitized through the analog-to-digital converter (ADC) and demodulated by the Demod block.

The I/Q architecture is the most versatile transceiver architecture today. Its advantage is that the modulation and demodulation functionalities are decoupled from the RF radio front-end, which makes it easy to generate arbitrary waveforms and data rate. This is not the case for all transceiver architectures. However, this architecture is relatively complex and its circuit power consumption is high. We will show that FSK transmitter and receiver can have lower power consumption than the I/Q transceiver.
3.3.5 Frequency Shift Keying

Frequency Shift Keying is a type of nonlinear modulation for which the output signal does not scale with the input signal in a linear fashion. The signal waveforms of binary FSK are given by

\[
\begin{align*}
    s_0(t) &= \sqrt{\frac{E_b}{T}} \cos[(\omega_c + 2\pi \frac{\Delta f}{2})t], \quad 0 \leq t \leq T \\
    s_1(t) &= \sqrt{\frac{E_b}{T}} \cos[(\omega_c - 2\pi \frac{\Delta f}{2})t], \quad 0 \leq t \leq T
\end{align*}
\]

(3.13)

where \( \Delta f \) is the separation between the two input signals. For M-FSK, additional signals are added at \( \Delta f \) apart.

Orthogonal FSK

The performance of FSK depends on the correlation among the signals \( s_i(t) \). Figure 3-12 shows the correlation between two sinusoids separated by \( \Delta f \). The normalized separation, \( m = \Delta f T \), where \( T \) is the symbol period, is called the modulation index. FSK signals used in practice are almost always orthogonal, which occurs at \( \Delta f = i/2T \), where \( i \) is an integer. In this case, the bit error rate is given by

\[
P_b(E) = \frac{M}{2} Q\left( \frac{r E_b}{N_0} \right)
\]

(3.14)

One distinct difference between FSK and PSK/QAM is that FSK requires less signal power than PSK/QAM to achieve the same bit error rate at large \( M \). In PSK and QAM, if more constellation points are added with the requirement that \( d_{\text{min}} \) stays the same (to keep the same bit error rate), the constellation must be expanded in the radial direction. PSK must use a larger circle, and QAM must add additional constellation points outside of the existing ones. Either way, the average symbol energy is increased. The average symbol energy for FSK, on the other hand, stays constant regardless of \( M \). This is because \( d_{\text{min}} \) in FSK does not depend on the amplitude, but rather, it depends only on the frequency separation. For this reason, \( E_b/N_0 \) actually decreases for large \( M \).

The cost in the improved power efficiency is bandwidth efficiency. Since each additional signal must occupy a frequency separation of \( \Delta f \), the bandwidth efficiency for FSK is

\[
R/W \approx \frac{r/T}{M \cdot \Delta f} = \frac{\log_2 M}{M \cdot m}
\]

(3.15)
where \( m \) is, again, the modulation index.

The bandwidth requirement of high-order FSK can grow large very fast at large \( M \). For example, assuming that the symbol rate is 1Mbps, then 2-FSK uses 1MHz of bandwidth. 32-FSK, on the other hand, requires more than 3MHz of bandwidth to achieve the same bit rate for an \( E_b/N_0 \) savings of about 5dB. Tripling the bandwidth requirement is a significant cost for many communication systems. In fact, channel coding can achieve the same \( E_b/N_0 \) reduction without incurring nearly as much penalty on the bandwidth. For this reason, high-order FSK is rarely used.

**Minimum Shift Keying**

Minimum Shift Keying (MSK) is a special case of binary FSK where \( \Delta f = 1/2T \), which is the minimum frequency separation required to produce two orthogonal signals. MSK is a popular modulation scheme for mobile channels due to the following desirable properties: constant envelope, good spectral efficiency, and good BER performance.

MSK has a simple interpretation as a form of FSK. If a symbol \( a_k \) is sent, where \( a_k = \pm 1 \), the phase change during one symbol period is

\[
\Delta \phi = a_k 2\pi \frac{\Delta f}{2} T = a_k \frac{\pi}{2}
\]

\[ (3.16) \]
Thus, the phase advances by 90° if a one is sent and decreases by 90° if a zero is sent. The amplitude of MSK signal always stays constant.

In light of this result, MSK can be modulated by controlling the local oscillator through either directly modulating the VCO or dithering the divide value in the frequency synthesizer inside the LO [30]. This direct modulation architecture is shown in Figure 3-13. As compared to the generic I/Q transmitter shown in Figure 3-11(a), this architecture has eliminated several circuit building blocks including the two power-hungry RF mixers. This architecture is chosen for the sensor node transmitter due to its low power consumption [31].

![Figure 3-13: Direct modulation of MSK signaling.](image)

An MSK signal can also be easily detected using a noncoherent frequency discriminator circuit, as shown in Figure 3-14. The received signal is first down-converted to an IF frequency, then the signal amplitude near the two tones at $\omega_0$ and $\omega_1$ are compared. Note as compared to the I/Q receiver shown in Figure 3-11(b), this architecture eliminates one power-hungry down-conversion mixer. This architecture is chosen for the sensor node receiver.

This circuit is not as good as the generic noncoherent detection circuit shown in Figure 3-6 because the narrow bandpass filter, which is used in place of the correlator, does not match to the input signal perfectly. The loss in SNR can be compensated by increasing the RF transmit power of the base station transmitter.

![Figure 3-14: MSK receiver with frequency discriminator.](image)
Figure 3-15 shows the simulated output eye diagrams of noncoherently and coherently detected GMSK signals. Figures 3-15(a) and 3-15(b) shows the noncoherent frequency discriminator outputs with the bandwidth of the bandpass filters equal to $0.5/T$ and $0.3/T$, respectively. When $BW=0.3$, the eye looks half-way closed, while the coherently detected MSK signal, as shown in Figures 3-15(c) and 3-15(d), still has a wide eye opening. This shows that noncoherent detection is inferior to coherent detection in terms of BER. The bit error rate of noncoherently detected MSK is

$$P_e = \frac{1}{2}e^{-E_b/2N_O}$$

(3.17)

### 3.3.6 Modulation Comparisons

Figure 3-16 shows the modulation power efficiency versus bandwidth efficiency trade-off of M-PSK, M-QAM, and M-FSK. The y-axis is the $E_b/N_O$ required to achieve a bit error rate of $10^{-3}$, and the x-axis is the bandwidth efficiency. QAM is complex and power hungry, so it is only popular in high data rate applications where power efficiency is sacrificed for bandwidth efficiency. PSK is bandwidth efficient and has a good balance between complexity and performance for small $M$. FSK is easy to implement and has good power efficiency.

The $E_b/N_O$ shown in Figure 3-16 only affects the RF transmit power. As described in Chapter 2, the sensor transceiver power consumption is dominated by the electronics power and not by the RF transmit power. Therefore, it is not enough just to consider the $E_b/N_O$ improvement from the modulation techniques. The effect of the modulation techniques on the power consumption of the sensor node transceiver circuits must be evaluated.

In Chapter 4, a comparison is made between binary FSK and binary PSK transmitter power consumption by including the contribution of both the RF transmit power and the circuit electronics power. Even though binary PSK is more efficient in terms of $E_b/N_O$ as shown in Figure 3-16, its electronics consume more power as compared to a binary FSK transmitter, so its total power consumption is higher for low to intermediate data rates. This shows that in order to evaluate the power consumption of transceivers with different communication protocols, both the communication protocols and the transceiver electronics need to be considered.
(a) Frequency discriminator output eye diagram with Gaussian filter (BW=0.5/T).

(b) Frequency discriminator output eye diagram with Gaussian filter (BW=0.3/T).

(c) GMSK coherent detection I-channel eye diagram (BT=0.3).

(d) GMSK coherent detection Q-channel eye diagram (BT=0.3).

Figure 3-15: Noncoherently and coherently detected GMSK eye diagrams.
Figure 3-16: SNR versus bandwidth efficiency in AWGN with BER=10^{-3}.

3.4 Detection in Multipath Fading Channel

In an AWGN channel, the primary source of performance degradation is thermal noise, and the main signal distortion is caused by bandlimited filtering. However, in a realistic wireless mobile environment, the above assumptions are no longer sufficient. Since the signal travelling from the transmitter to the receiver comes from multiple reflective paths due to motions and obstructions, the received signal experiences variations in both amplitude and phase. This propagation model is called multipath propagation, and the fading effect is called multipath fading.

In statistical terms, the multipath propagation model can be separated into two types of fading effects: large-scale fading and small-scale fading. Each of these affect the communications system in different ways.

3.4.1 Large-Scale Fading

Large-scale fading predicts the mean signal strength for large transmitter-receiver separation distances. The local received power is computed by averaging signal measurement within a radius of several wavelengths or greater [25]. The average received power, as a function of
transmitter-receiver separation $d$, is given by the following equation [29],

$$P_R(d) = \frac{P_T G_T G_R \lambda^2}{(4\pi)^2 d^n L}$$  \hspace{1cm} (3.18)

where each of the variables is defined in Table 3.1.

<table>
<thead>
<tr>
<th>$P_R$</th>
<th>received signal power</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_T$</td>
<td>transmitted signal power</td>
</tr>
<tr>
<td>$G_T$</td>
<td>transmitter antenna gain</td>
</tr>
<tr>
<td>$G_R$</td>
<td>receiver antenna gain</td>
</tr>
<tr>
<td>$\lambda$</td>
<td>carrier wavelength</td>
</tr>
<tr>
<td>$d$</td>
<td>transmitter receiver separation distance</td>
</tr>
<tr>
<td>$n$</td>
<td>path loss exponent</td>
</tr>
<tr>
<td>$L$</td>
<td>system loss factor not related to propagation: transmission line attenuation, filter losses, antenna losses, etc.</td>
</tr>
</tbody>
</table>

Table 3.1: Summary of variables for Equation (3.18).

The variable $n$ in Equation (3.18) is the path loss exponent, which ranges from $n = 2$ in free space to $n > 4$ in obstructed areas. Some typical values of $n$ are summarized in Table 3.2 [25].

<table>
<thead>
<tr>
<th>ENVIRONMENT</th>
<th>$n$</th>
</tr>
</thead>
<tbody>
<tr>
<td>free space</td>
<td>2</td>
</tr>
<tr>
<td>obstructed in factory</td>
<td>2-3</td>
</tr>
<tr>
<td>urban area cellular radio</td>
<td>2.7-3.5</td>
</tr>
<tr>
<td>obstructed in building</td>
<td>4-6</td>
</tr>
</tbody>
</table>

Table 3.2: Summary of typical path loss exponent values.

The average path loss $PL(d)$ is defined as

$$PL(d)[dB] = 10 \log \frac{P_T}{P_R} = 10 \log \left[ \frac{(4\pi)^2 d^n}{G_T G_R \lambda^2} \right]$$  \hspace{1cm} (3.19)

In actual measurements, average path loss is determined at a reference distance $d_o$, which is taken to be 1m in indoor channels and 1km for large cells. Path loss at an arbitrary distance $d > d_o$ is interpolated with the following formula

$$PL(d)[dB] = PL(d_o)[dB] + 10n \log \left( \frac{d}{d_o} \right) + X_\sigma$$  \hspace{1cm} (3.20)
where $X_\sigma$ is a zero-mean Gaussian random variable with variance $\sigma^2$, which models the variation in the mean path loss.

### 3.4.2 Indoor Environment

An indoor factory environment is considered in this project. Thus, it is essential to characterize the propagation characteristics in such a setting. An indoor environment differs from the traditional mobile channel in two aspects. First, the distances covered are much smaller. Second, the variability of the environment is much greater. Propagation in buildings is strongly influenced by specific features as layout, construction materials, and building types, etc.

Equation (3.20) is still a valid model for indoor environment. Some typical data on $n$ and $\sigma$ is given in the following table [32–34]

<table>
<thead>
<tr>
<th>Building</th>
<th>Frequency (MHz)</th>
<th>$n$</th>
<th>$\sigma$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Factory, LOS</td>
<td>1300</td>
<td>1.8</td>
<td>4.6</td>
</tr>
<tr>
<td>light cluttered</td>
<td>1300</td>
<td>1.8</td>
<td>4.4</td>
</tr>
<tr>
<td>heavy cluttered</td>
<td>1300</td>
<td>2.38</td>
<td>4.67</td>
</tr>
<tr>
<td>Factory, obstructed</td>
<td>1300</td>
<td>2.81</td>
<td>8.09</td>
</tr>
<tr>
<td>light cluttered</td>
<td>1300</td>
<td>3.0</td>
<td>7.0</td>
</tr>
<tr>
<td>heavy cluttered</td>
<td>1500</td>
<td>2.6</td>
<td>14.1</td>
</tr>
<tr>
<td>office, hard partition</td>
<td>2450</td>
<td>1.3</td>
<td>6.0</td>
</tr>
<tr>
<td>office, NLOS</td>
<td>2450</td>
<td>2.5</td>
<td>5.7</td>
</tr>
<tr>
<td>office, LOS (hallway)</td>
<td>5250</td>
<td>2.6</td>
<td>5.7</td>
</tr>
<tr>
<td>office, NLOS</td>
<td>5250</td>
<td>1.8</td>
<td>5.8</td>
</tr>
</tbody>
</table>

Table 3.3: Summary of typical path loss data for indoor environment.

The variation $\sigma$ can be quite large depending on different settings. This is why an accurate prediction of large scale path loss is difficult to obtain. Fortunately, it has been shown that in an indoor environment the path loss index is very close to 2 if there are no walls in the transmission path [35]. In addition, the path loss variation is small due to short transmission distance. In such an environment, the small-scale path loss is a more serious concern.
### 3.4.3 Small-Scale Fading

Small-scale fading models the rapid fluctuation of the received signal strength as a result of very small changes in the spatial separation between a transmitter and receiver. This change is on the order of a few wavelengths and can be as small as half a wavelength. Small-scale fading is categorized into delay spreading of the signal, which is a function of spatial characteristics, and time variance of the channel, which is manifested in Doppler shift and spectrum broadening.

At frequencies in the multi-GHz regime (i.e., UHF and SHF), the Doppler spread is around 10Hz for a relatively stationary environment [29]. The coherent bandwidth is reported to be between 5-10MHz at both 2.4GHz [36] and 5GHz [35] for a heavily obstructed indoor environment. Therefore, for channel bandwidth less than the coherent bandwidth, the channel can be assumed to be frequency-nonselective and slowly-fading. Frequency-nonselectivity implies that equalization for cancelling channel induced ISI is not necessary. Slowly-fading means that the amplitude of the transmitted signal can be assumed to be constant during a symbol period. Consequently, the channel response, $C(\tau; t)$, is a complex constant during one symbol interval.

$$C(\tau; t) = \alpha e^{-j\phi}, \quad (k-1)T < t < kT$$

(3.21)

where $\alpha$ and $\phi$ are random processes that change value every symbol interval.

Assuming that there is no line-of-sight component and that many multipath signals exist, then by the central limit theorem, $C(\tau; t)$ can be modeled as a zero-mean complex Gaussian process. It is well-known that the amplitude of a complex Gaussian process is Rayleigh distributed, and the phase is uniformly distributed in $[-\pi, \pi]$ [37]. This model is called the *Rayleigh fading* model. The PDF of $\alpha$ is given as

$$f(\alpha) = \frac{\alpha}{\sigma^2} e^{-\alpha^2/(2\sigma^2)}$$

(3.22)

More complex models of $C(\tau; t)$ exist. For instance, if there is a line of sight component, then $C(\tau; t)$ is modeled as a complex Gaussian process with a non-zero mean. The amplitude in this case follows a *Rician* distribution. Fortunately, it has been shown that in obstructed sites the amplitude distribution is close to Rayleigh [36], which simplifies the analysis and
modeling process.

The received signal, \( r(t) \), can be written as

\[
r(t) = \alpha e^{-j\phi} s(t) + n(t)
\] (3.23)

In addition, assuming the fading is slow enough that the phase \( \phi \) can be tracked by the carrier synchronization loop, which is usually the case, the effect of the Rayleigh channel is an amplitude scaling on the input signal \( s(t) \). This translates into a scaling in the SNR \( \gamma_b = E_b/N_0 \) as follows,

\[
\gamma_b = \frac{E_b}{N_0} \alpha^2
\] (3.24)

If \( \alpha \) is known, the probability of error can be computed the same way with the new \( \gamma_b \) as the variable of interest. The overall error probability is then computed by averaging over all possible \( \gamma_b \), i.e.,

\[
P_e = \int_0^\infty P_e(\gamma_b) f(\gamma_b) d\gamma_b
\] (3.25)

where \( f(\gamma_b) \) is the PDF of the SNR,

\[
f(\gamma_b) = \frac{1}{\tilde{\gamma}_b} e^{-\gamma_b/\tilde{\gamma}_b}
\] (3.26)

and \( \tilde{\gamma}_b \) is the average signal-to-noise ratio defined as

\[
\tilde{\gamma}_b = \frac{E_b}{N_0} \cdot E[\alpha^2]
\] (3.27)

The above formulas provide the tools for deriving the error probabilities of various modulation schemes. Probability of error for binary FSK in AWGN and Rayleigh fading channels are shown in Figure 3-17. Unlike the water-fall BER curve of the AWGN channel, the BER decreases much more slowly as a function of \( E_b/N_0 \) in a Rayleigh channel. Closed form solutions for higher level modulations are complicated even if they can be derived [38]. Simulation is usually used to determine BER for more complex modulation systems.
### 3.5 Transmitter Output Power Requirement

Given a required bit error rate for a communications system, the formulas derived in the previous sections allow us to find the $E_b/N_O$ that satisfies the BER. Once we know the $E_b/N_O$ requirement at the receiver, the transmitter power requirement can be derived as shown in Figure 3-18.

$$P_T = \frac{(4\pi)^2 d^8}{G_T G_R \lambda^2} \cdot \frac{F \cdot R \cdot N_O \cdot \left(\frac{E_b}{N_O}\right)}{P_{sens}}$$

![Figure 3-18: Determining the transmitter Output power requirement.](image)

Given the required $E_b/N_O$, the energy per bit $E_b$ can be computed. The required signal power, then is $E_b$ multiplied by the data rate $R$. Since the noise of the receiver electronics degrades the SNR of the received signal, the required signal power needs to be multiplied by
the noise factor $F$. This signal power is called the receiver sensitivity, which is the minimum detectable signal level that achieves the required bit error rate at the receiver.

A more intuitive way of looking at the problem is that given an $E_b/N_0$, the $SNR$ requirement is

$$SNR = \left(\frac{E_b}{N_0}\right) \cdot \frac{R}{BW}$$

(3.28)

where $E_b \cdot R$ is the signal power and $N_0 \cdot BW$ is the noise power. Since the available noise power at the receiver antenna is $N_0 \cdot BW$, the required receiver sensitivity needs to be

$$P_{sens} = SNR \cdot N_0 \cdot BW \cdot F$$

(3.29)

which simplifies to the same expression.

Once the receiver sensitivity is known, the transmitter output power can be computed by adding in the average path loss $PL(d)$, as in shown in Equation (3.19). The final equation for the required transmitter output power is

$$P_T = \frac{(4\pi)^2d^2}{G_TG_R\lambda^2} \cdot F \cdot R \cdot N_0 \cdot \left(\frac{E_b}{N_0}\right)$$

(3.30)

Figure 3-19 shows the link budget for a coherent binary FSK system with a carrier frequency of 5.8GHz, a path loss exponent of 3, a receiver noise figure of 6dB, a data rate of 1Mbps, and an $E_b/N_0$ of 21dB. For a BER level of $10^{-3}$, the required SNR is 27dB for FSK in a Rayleigh channel, as shown in Figure 3-17. This SNR can be reduced to 21dB with coding [39]. It is assumed that the antennas are isotropic so that $G_T$ and $G_R$ are unity.

In Figure 3-19, it can be seen that the required transmit power is -10dBm at a distance of 10 meters. To combine the link loss variation $\sigma$ listed in Table 3.3, a link margin should be added. Adding a link margin of 10dB pushes the transmit power up to 0dBm, which is still quite small. For a log-normal distributed large-scale path loss with a standard deviation of 4-5dB, a 10dB link margin ensures a better than 98% coverage probability [37].

### 3.6 Summary

In this chapter, detection and propagation theories are discussed and appropriate models are given. In an obstructed factory environment, large scale fading follows a log-normal distribution, while non-line-of-sight small scale fading can be closely approximated by a
Rayleigh distribution. OOK, M-PSK, M-QAM, and M-FSK are compared in terms of modulation power efficiency and bandwidth efficiency. M-PSK provides the best modulation power efficiency and bandwidth efficiency trade-off. M-FSK has the advantage of simpler transmitter and receiver circuitry, which makes it more suitable for the microsensor application considered in this work. The modulation power efficiency only affects the RF transmit power and does not take into account the power consumption of the RF circuits, which depends on the modulation type as well. The combined effect of RF transmit power and circuit electronics power is investigated in Chapter 4.
Chapter 4

RF Transceiver System Optimization

This chapter presents design techniques that maximize the battery life of the microsensor transceiver. We will show that maximum battery life is achieved by optimizing the design of the system and the circuits together. In this approach, the central idea is to develop a transceiver energy model that takes into account circuit power consumption as a function of system parameters. This approach not only maximizes battery life but also provides insights into the design of the transceiver circuitry. It shows clearly which circuit building blocks are the key ones to focus on to reduce energy cost.

4.1 Transceiver Power Consumption

This section derives the proper modeling of transceiver power consumption. We show that the transceiver power consumption is only a weak function of the RF output power for short transmission range and low to moderate data rate. For this reason, the transceiver power consumption does not scale with either data rate or distance, as it is commonly assumed [40].

4.1.1 Power Consumption of Short-Range Transceivers

Figure 4-1 shows a simplified block diagram of a generic RF transceiver. There are four major circuit building blocks within the transceiver. The TX block is responsible for modulation and up-conversion, the RX block is for down-conversion and demodulation, the LO
block generates the required carrier frequency, and the PA amplifies the signal to produce the required RF output power $P_T$.

![RF transceiver](image)

**RF transceiver**

Figure 4-1: Generic transceiver architecture.

For a single-carrier system with low to intermediate data rate, the RF circuit blocks consume most of the power. This includes the up-conversion mixers in the TX, the LNA and the down-conversion mixers in the RX, the VCO and the divider in the LO, and the output stage in the PA. However, estimating the power consumption of a transceiver is difficult because it is heavily dependent upon the transceiver architecture, the process technology, and the circuit design.

Figure 4-2 shows power consumption of Bluetooth transceivers reported in ISSCC 2001. The power consumption numbers vary from as low as 20-30mW to as high as 200mW. The differences come from choices of architecture, technology, and design. On the chart, the lowest power transceiver reports a power consumption of 30mW in the receive mode and 21mW in the transmit mode [11]. This design is aimed at low power operations and so its design choices are made accordingly. It uses a silicon germanium BiCMOS process that provides high gain and low noise bipolar devices. The modulation is performed through direct modulation of the local oscillator, which reduces transmitter power consumption by eliminating the TX block. Because this is one of the lowest-power short-range transceivers reported to date, the power consumption figures of this transceiver is used as a starting point for energy analysis in this chapter.
4.1.2 Short-Range versus Long-Range Transceivers

Table 4.1 shows typical power consumption numbers for GSM [41], 802.11b [42], and Bluetooth transceivers [11]. $P_{RX}$ is the power consumption of the RX block, $P_{TX}$ is the power consumption of the TX block, $P_{LO}$ is the power consumption of the LO block. In the receive mode, both RX and LO need to be on, so $P_{RX} + P_{LO}$ is the power consumption in the receive path. In the transmit mode, TX, LO, and PA need to be on, so $P_{TX} + P_{LO}$ is the power consumption in the transmit path excluding the PA. $P_T$ is the RF transmit power, and $P_{PA}$ is the PA DC power consumption assuming a 40% power added efficiency. PA efficiency depends on PA type, output power level, technology, and design. A fixed realizable efficiency is assumed here for the ease of comparison. The modeling of PA efficiency is discussed in more detail in Section 4.3.

<table>
<thead>
<tr>
<th></th>
<th>$P_{RX} + P_{LO}$ (mW)</th>
<th>$P_{TX} + P_{LO}$ (mW)</th>
<th>$P_T$ (mW)</th>
<th>$P_{PA}$ (mW)</th>
</tr>
</thead>
<tbody>
<tr>
<td>GSM</td>
<td>240</td>
<td>360</td>
<td>1000</td>
<td>2500</td>
</tr>
<tr>
<td>802.11b</td>
<td>60</td>
<td>100</td>
<td>100</td>
<td>250</td>
</tr>
<tr>
<td>Bluetooth</td>
<td>30</td>
<td>12</td>
<td>1</td>
<td>2.5</td>
</tr>
</tbody>
</table>

Table 4.1: Power consumption of short and long range transceivers.

The GSM transceiver has a transmission range greater than 1km and has the most
stringent system specifications. Its transceiver electronics power and RF transmit power are the highest. The 802.11b transceiver has an intermediate range on the order of 10’s of meters. The Bluetooth transceiver has the shortest transmission range and the simplest system specifications. Its power consumption is the lowest.

The RF transmit power of the Bluetooth transceiver is three orders of magnitude lower than that of the GSM transceiver due to the reduction in transmission range. However, the reduction in transceiver electronics power is only an order of magnitude. What this means is that in long-range transceivers the power is dominated by the RF transmit power, but in short-range transceivers the power is dominated by the electronics power.

Since the power consumption of short-range transceivers is dominated by the RF electronics, system design techniques, including coding and diversity, do not reduce the overall power consumption of the transceiver. As in the case of the Bluetooth transceiver, further reduction of the RF transmit power, which is 1mW, does not impact the power consumption of the transceiver, which is an order of magnitude higher.

This implies that focus should be placed upon minimizing the transceiver electronics power. However, achieving significant power reduction by circuit design alone is also very difficult due to fundamental limits in transconductance and noise. A literature survey shows that the lowest-power Giga-Hertz transceivers are on the order of 10’s of milli-Watts [11,12,43]. Reducing the power consumption significantly below this level is difficult.

4.2 Transceiver Power Efficiency

In the following sections, we will show that significant battery life improvement can be achieved by combining system techniques and circuit design together. In order to do this, a new concept, called transceiver power efficiency, needs to be introduced. The transceiver power efficiency is defined as

\[ \eta_P = \frac{\alpha_T \cdot P_T}{\alpha_R \cdot (P_{RX} + P_{LO}) + \alpha_T \cdot (P_{TX} + P_{LO} + P_{PA})} \] (4.1)

In simplest terms, \( \eta_P \) is the ratio of RF transmit power to RF electronics power. \( P_T \) is the RF transmit power, \( P_{RX} + P_{LO} \) is the power consumption in the receive path, and \( P_{TX} + P_{LO} + P_{PA} \) is the power consumption in the transmit path. The terms \( \alpha_T \) and \( \alpha_R \)
are the transmitter and receiver activity factors defined as

\[ \alpha_T = \frac{\bar{t}_{tx}}{\bar{t}_{tx} + \bar{t}_{rx}} \quad (4.2) \]

\[ \alpha_R = \frac{\bar{t}_{rx}}{\bar{t}_{tx} + \bar{t}_{rx}} \quad (4.3) \]

where \( \bar{t}_{rx} \) is the average receive time and \( \bar{t}_{tx} \) is the average transmit time in one receive-transmit cycle. This takes into account systems that have asymmetric transmit and receive requirements.

In our application, the sensor node transmits and receives equal-sized packets every 5ms, so \( \alpha_T = \alpha_R = 0.5 \). Equation (4.1) can then be simplified to

\[ \eta_P = \frac{P_T}{(P_{RX} + P_{LO}) + (P_{TX} + P_{LO} + P_{PA})} \quad (4.4) \]

When the power amplifier dominates the transceiver power consumption, \( \eta_P \) can be approximated as

\[ \eta_P \approx \frac{P_T}{P_{PA}} = \eta_{PA} \quad (4.5) \]

where \( \eta_{PA} \) is the PA efficiency. In this case, the transceiver power consumption can be reduced through either better circuit design that improves the PA efficiency or more efficient system techniques that reduce the RF transmit power.

When the PA power consumption is small relative to the rest of the transceiver electronics, \( \eta_P \) is approximately

\[ \eta_P \approx \frac{P_T}{(P_{RX} + P_{LO}) + (P_{TX} + P_{LO})} \quad (4.6) \]

In this case, PA design and efficient system techniques do not affect the transceiver power consumption. Instead, the focus should be placed upon the design of LO, RX, and TX.

Table 4.2 shows the transceiver power efficiency for the GSM, 802.11b, and Bluetooth transceivers. The maximum efficiency achievable for these systems is 40%, which is the PA efficiency assumed in the previous section. Both GSM and 802.11b transceivers have respectable efficiencies. The Bluetooth transceiver, on the other hand, has an efficiency of 2%, which means that the RF transmit power is a very small fraction of the total transceiver power consumption.
Table 4.2: Comparison of transceiver power efficiency for GSM, 802.11b, and Bluetooth transceivers.

<table>
<thead>
<tr>
<th>Transmitter</th>
<th>Efficiency</th>
</tr>
</thead>
<tbody>
<tr>
<td>GSM</td>
<td>32%</td>
</tr>
<tr>
<td>802.11b</td>
<td>24%</td>
</tr>
<tr>
<td>Bluetooth</td>
<td>12%</td>
</tr>
</tbody>
</table>

Classic communications theory only considers the RF transmit power $P_T$. $P_T$ can be reduced via two main techniques: coding and diversity. Coding, which includes modulation as a special case, increases the minimum distance between constellation points of the coded symbols. Diversity, which can be applied in frequency, time, or space, combats fading in a time-variant multi-path channel. For systems such as GSM where the transceiver power efficiency is high, these techniques can reduce the transceiver power consumption effectively. However, this does not apply to short-range transceivers, and these system techniques tend not to be taken as seriously, from the perspective of reducing the total transceiver power consumption, since they are not as effective.

In the following sections, we will show that the transceiver power efficiency for short-range transceivers can be improved. When this happens, the battery life of the transceiver improves and system techniques become more effective in the overall design.

### 4.3 Transceiver Energy Model

The battery life of the microsensor node is determined by the transceiver energy consumption, which depends on both the power consumption of the transceiver and on the specific way the transceiver is operated. Transceiver energy modeling is first applied by Cho [44] to investigate the effect of multi-access protocols on the battery life of microsensor networks. It was later extended by Wang [22] to the study of the effect of modulation techniques on transceiver energy consumption. This research builds upon the previous results, and in this section we present a simple energy model that takes into account both circuit and system parameters.
4.3.1 Transceiver Operation

The microsensor transceiver is operated in burst mode. The amount of delay between each burst is determined by the latency requirement, which is 5ms for this application. Figure 4-3 shows power consumption versus time using the low-power Bluetooth transceiver described in Section 4.1. The transceiver starts up, receives 100-bits at 1Mbps, transmits 100-bits at 1Mbps, and then shuts off. Each mode of operation is described in detail below.

![Burst receive/transmit cycle for Bluetooth transceiver.](image)

4.3.2 Start-up Mode

When the transceiver is first turned on, it takes some time for the frequency synthesizer and the VCO to lock to the carrier frequency. The start-up energy can be modelled as follows:

$$E_{\text{start}} = P_{LO} \cdot t_{\text{start}}$$

(4.7)

where $P_{LO}$ is the power consumption of the synthesizer and the VCO. The term $t_{\text{start}}$ is the required settling time, which is determined by the synthesizer loop bandwidth. Larger loop bandwidth reduces the start-up time, but it trades off with power consumption and
stability issues. This topic is discussed in detail in [44].

Other RF building blocks such as LNA and mixer have negligible start up time. Power amplifiers may have a significant start-up time at high levels of output power, but this is not a problem for the output power range of microsensor networks. Therefore, these circuit blocks can remain in the off-state during the start-up mode.

4.3.3 Receive Mode

The active components of the receiver include the low noise amplifier (LNA), down-conversion mixers, frequency synthesizer, VCO, IF amplifier, and demodulator (Demod). The receiver energy consumption can be modeled as follows:

\[ E_{rx} = (P_{LO} + P_{RX}) \cdot t_{tx} \]  

(4.8)

where \( P_{RX} \) includes the power consumption of the LNA, mixer, IF amplifier, and demodulator. The receiver power consumption is dictated by the carrier frequency and the noise and linearity requirements. Once these parameters are determined, to the first order the power consumption can be approximated as a constant for data rates up to 10’s of Mbps. In other words, the power consumption is dominated by the RF building blocks that operate at the carrier frequency. The IF demodulator power varies with data rate, but it can be made small by choosing a low IF.

4.3.4 Transmit Mode

The transmitter includes the modulator (Mod), frequency synthesizer and VCO (shared with the receiver), and power amplifier (PA). The data modulates the VCO and produces an FSK signal at the desired data rate and carrier frequency. A simple transmitter energy model is shown in Equation (4.9). The modulator consumes very little energy and therefore can be neglected.

\[ E_{tx} = (P_{LO} + P_{PA}) \cdot t_{tx} \]  

(4.9)

\( P_{LO} \) can be approximated as a constant. \( P_{PA} \) depends on the PA efficiency, the link-budget, and the sensitivity requirement. Specifically,

\[ P_{PA} = \frac{1}{\eta_{PA}} \cdot P_{T} \]  

(4.10)
The PA efficiency depends on factors such as PA type, output power level, process technology and design. Although PA design is not the focus of this work, the PA power consumption is important enough that it deserves an investigation on the system level. Figure 4-4 plots the ideal PA DC power consumption as a function of PA output power for three different types of PA's: Class-A, Class-B, and Class-E.

Class-A amplifier has a constant bias and therefore the DC power is invariant to the output power. It’s maximum achievable efficiency is 50%. Class-B amplifier is biased at a 50% conduction angle and has an ideal efficiency of 78%. Class-AB amplifier, which is not shown in the figure, has a conduction angle of greater than 50% and its performance stands in-between that of the Class-A and Class-B. Class-AB amplifier is often used in practice due to its balance between linearity and efficiency. Class-E amplifier is a switching type of amplifier which can have a theoretical efficiency of 100%. The price that comes with better efficiency is worsened linearity. Class-E PA is non-linear due to its switching operation and can only be used in constant-envelope modulations such as FSK [29].

The challenge of PA design increases as a function of both the output power and frequency. Typical achievable PA efficiencies at Giga-Hertz frequencies and below-Watt-level
output power are about 30% for Class-A and 70% for Class-E \cite{45,46}. Since the RF output power level is low for short-range applications, PA design becomes easier and higher efficiency numbers are likely more achievable.

The Bluetooth transceiver with power consumption numbers shown in Figure 4-3 has an efficiency of approximately 10% at 1mW RF output power. This efficiency is very low but is not uncommon among Bluetooth transceivers, where the design focus is usually not on the PA since the PA power consumption is relatively low as compared to other circuit blocks. As we show in Section 4.4, the PA efficiency makes little difference to the total energy cost of the transceiver at the data rate of 1Mbps, but it becomes important at high data rate and can greatly improve the battery life if a highly efficient PA is used.

4.3.5 Switch Mode

When the sensor transceiver switches from the receive mode to the transmit mode, the local oscillator frequency needs to change. For a heterodyne receiver and a direct-conversion transmitter, the amount of frequency adjustment is equal to the IF frequency, which is several Mega-Hertz for a low-IF receiver. The energy expended during this mode is

\[ E_{\text{switch}} = P_{LO} \cdot t_{\text{switch}} \quad (4.12) \]

4.3.6 Total Energy Consumption

Define the total energy consumption during each duty cycle as \( E_{op} \), then \( E_{op} \) can be expressed as

\[ E_{op} = E_{\text{start}} + E_{rx} + E_{\text{switch}} + E_{tx} \quad (4.13) \]

Note that \( E_{op} \) is the total area in Figure 4-3. Therefore, from the energy perspective, the objective is to minimize \( E_{op} \). Careful circuit design reduces the power consumption, which is the y-axis in Figure 4-3. What’s not apparent is that system techniques reduce the time of operation, which is the x-axis. Since the time of operation and the power consumption are correlated, an understanding of both circuit design and system design is essential in order to optimize \( E_{op} \).

Figure 4-5 shows how much energy is used in each mode. Receive mode consumes the most amount of energy (46%), while the transmit mode and the start-up mode take in 32%
and 20% respectively. The switch mode energy is negligible due to the short switch time. The transceiver operates 330μs every 5ms, which makes the duty cycle at 6.6%. With a 1000mAh battery, the life time for this sensor transceiver is around 2-months.

![Energy allocation per duty cycle for Bluetooth transceiver.](image)

Figure 4-5: Energy allocation per duty cycle for Bluetooth transceiver.

### 4.4 Energy Optimization

The microsensor system discussed in Chapter 1 requires a battery life of one year or better. Although the Bluetooth transceiver described in the last section falls short of this requirement, it serves as a starting point for making improvements. This section examines $E_{\text{op}}$ in detail and suggests ways to increase the battery life by considering both circuit and system improvements.

#### 4.4.1 Start-up Energy

The start-up energy can be a significant part of the total energy consumption, especially when the transceiver is used to send short packets in burst mode [44]. Figure 4-6 shows the normalized start-up energy cost as a function of the start-up time. It is clear that the start-up overhead increases as a function of the start-up time.

The start-up overhead is a function of the transmit and receive times. If the transmit and receive times are long, then the start-up overhead is relatively small. A simple calculation
Figure 4-6: Start-up energy cost as a function of start-up time.

shows that the start-up energy becomes negligible if the following condition is held true:

\[ t_{\text{start}} \ll \left( 1 + \frac{P_{RX}}{P_{LO}} \right) \cdot t_{rx} + t_{\text{switch}} + \left( 1 + \frac{P_{PA}}{P_{LO}} \right) \cdot t_{tx} \]  

(4.14)

This equation shows that the start-up time requirement reduces if the transmit and receive times are increased. In addition, the start-up time requirement depends on the ratio between the power consumption of the local oscillator and that of the RX and of the PA. If the RX and the PA powers dominate, then the start-up energy, which is a function of the LO power only, is less significant. Unfortunately, \( P_{RX} \) and \( P_{PA} \) are typically on the same order as \( P_{LO} \), so Equation (4.14) needs to be evaluated.

For the receive/transmit scheme shown in Figure 4-3, the right hand-side of Equation (4.14) is evaluated to be approximately 450\( \mu \)s. To keep \( E_{\text{start}} \) an order of magnitude below \( E_{op} \), it is desirable to have a start-up time of less than 45\( \mu \)s. Cho has demonstrated a 6GHz frequency synthesizer implementation with a start-up time under 20\( \mu \)s [31], which would keep the start-up overhead under 5% of \( E_{op} \).
4.4.2 Data Rate

As the data rate of the system is increased, the transmitter RF output power needs to be increased accordingly in order to keep the same transmit distance and $E_b/N_0$. However, the power consumption of the other RF building blocks, including the LNA, the mixer, the VCO, and the frequency synthesizer, do not scale with data rate up to tens of Mega-bits per second. The reason is that on-chip passives have Q's on the order of 10-20, which leads to large bandwidth on the order of hundreds of Mega-Hertz. This bandwidth is much greater than that of the data bandwidth, and therefore increasing the data rate does not affect the operation of the RF circuitry to the first order. The IF demodulator power consumption does scale with the data rate, but its power consumption is small due to the low operating frequency. What this implies is that data rate does not scale linearly with transceiver power consumption, and this leads to optimization opportunities.

Assuming that the sensor node transceiver receives a packet of length $L_{pkt}$ and transmits a packet of the same length during each duty cycle. Then the transmit and receive times are

$$t_{tx} = t_{rx} = \frac{L_{pkt}}{R}$$  \hspace{1cm} (4.15)

Now $E_{op}$ can be re-written as

$$E_{op} = P_{LO} \cdot (t_{start} + t_{switch}) + (P_{LO} + P_{RX}) \cdot \frac{L_{pkt}}{R} + (P_{LO} + P_{PA}) \cdot \frac{L_{pkt}}{R}$$  \hspace{1cm} (4.16)

What this equation shows is that the transmit and receive powers are inversely proportional to the data rate. Since the PA power scales with the data rate, we can re-write the PA power consumption as

$$P_{PA} = \frac{1}{\eta} \cdot \gamma_{PA} \cdot d^\alpha \cdot R$$  \hspace{1cm} (4.17)

where $\gamma_{PA}$ is a constant for a fixed carrier frequency and $E_b/N_0$. At very high data rate, $E_{op}$ approaches to

$$E_{op} = P_{LO} \cdot (t_{start} + t_{switch}) + \frac{1}{\eta} \cdot \gamma_{PA} \cdot d^\alpha \cdot L_{pkt}$$  \hspace{1cm} (4.18)

Two important observations come out of this equation. First, note that at high data rate, the energy does not depend on the data rate. Second, the two dominant energy

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components are that from the start-up transient and that from the power amplifier. This is because as the data rate increases, the transmit and receive times decrease. Since the power of RF building blocks such as LNA, mixer, VCO, and the frequency synthesizer do not scale with data rate to the first order, their contribution to $E_{op}$ scales down linearly with the transmit and receive times. The PA power consumption does scale with the data rate, so its contribution to $E_{op}$ is a constant regardless of the data rate.

Figure 4-7 shows $E_{op}$ as a function of data rate. The two solid curves have start-up time $120\mu s$ and PA efficiencies $10\%$ and $70\%$, respectively. The two dotted curves have start-up time $20\mu s$ and efficiencies $10\%$ and $70\%$, respectively. At low data rate, $E_{op}$ is dominated by the fixed cost, and optimized designs of the start-up energy and the PA do not make a difference. At high data rate, the start-up energy and the PA energy dominates, so in order to increase the battery life, good circuit design techniques need to be applied to minimize the start-up time and to maximize the PA efficiency.

![Figure 4-7: $E_{op}$ as a function of data rate.](image)

Figure 4-8 shows the impact of PA efficiency on the battery life at a data rate of 10Mbps. At $t_{\text{start}} = 120\mu s$, the start-up energy is so large that the battery life is limited to 7-month even if the PA reaches $100\%$ efficiency. At $t_{\text{start}} = 20\mu s$, the battery life is much improved. The PA efficiency needs to be higher than about $30\%$ to have a 1-year or better battery life. This is easily achievable as discussed in Section 4.3.4.
The analysis carried out in this section shows two important observations. First, increasing the data rate reduces the fixed energy cost of the transceiver. This may seem straightforward but it is not obvious from a pure communications system perspective. It is commonly believed that low data rate should be used in low-power systems since the SNR requirement scales linearly with the data rate. For this reason, low power systems such as Zigbee has a maximum data rate of 250-kbps and a typical data rate on the order of 10's of kilo-bits per second. The results in this section show that increasing the data rate reduces the fixed energy cost of the power-hungry RF circuit building blocks and cuts down the overall energy cost significantly.

The second point is that this analysis offers insights into circuit design. At high data rate, the important circuit building blocks to design are the local oscillator and the power amplifier since both dominate the power consumption. RF building blocks such as LNA and mixer are not as important to optimize since their energy cost are driven down by the data rate.

Unfortunately, the data rate of the system can not be increased indefinitely. The main reason for this comes not from the circuit limitations but rather from the system limitations. Due to limited coherence bandwidth available at Giga-Hertz carrier frequencies, single carrier systems can not sustain data rate much higher than 10Mbps without experiencing
frequency selective fading. In this case multi-carrier systems may be used to increase the data rate, but the additional complexity in the digital signal processing as well as peak-to-average ratio problems inevitably increase power consumption. The trade-off between data rate and power consumption for multi-carrier systems is much more difficult to evaluate.

4.4.3 Transceiver Power Efficiency

The above results can also be explained via the transceiver power efficiency concept. The Bluetooth transceiver we started out with has a transceiver power efficiency of 2%. With the new design at a data rate of 10Mbps and a PA efficiency of 70%, the transceiver power efficiency increases to 18%, as shown in Figure 4-9. This means that more power is used in RF transmission rather than in transceiver electronics overhead. Therefore, the new design is significantly more power efficient. In addition, note that the PA efficiency makes a difference only at high data rate.

![Figure 4-9: Transceiver power efficiency as a function of data rate.](image)

From communications theory, we know that the BER performance depends on $E_b/N_o$, which is not a function of data rate. For this reason, at high transceiver power efficiency, increasing the data rate further does not improve the power efficiency. This can be seen from Figure 4-7, where at high data rate $E_op$ levels off.

Figure 4-10 shows the transceiver activity with a 20μs start-up time and a 10Mbps data
rate. The maximum RF output power is set at 10mW to accommodate the higher data rate, and the PA efficiency is assumed to be 70%. The switching time is reduced to below 5μs with the faster frequency synthesizer. The $E_{op}$ of this transceiver is 8x lower than that of the Bluetooth transceiver. The battery life-time extends from 2-months to 16-months, which is a significant improvement and can not be achieved by either system or circuit optimization alone.

![Burst receive-transmit cycle for improved transceiver.](image)

**Figure 4-10:** Burst receive-transmit cycle for improved transceiver.

### 4.4.4 Receiver Power as Fixed Cost

At high transceiver efficiency, the fixed cost of the transceiver electronics is small, so the power consumption requirement of these transceiver circuit building blocks can be relaxed without affecting the battery life significantly. This is illustrated in Figure 4-11, which plots $E_{op}$ as a function of the power consumption of the RX block, which includes the LNA, the mixer, and the IF demodulator. Assuming that $P_{RX}$ increases due to relaxed design, $E_{op}$ at 10Mbps rises at a more gradual slope as compared to $E_{op}$ at 1Mbps. This implies that at high data rate the power consumption of the receiver electronics is less significant, and so the design constraints on its power consumption can be relaxed without affecting the total energy cost.
4.5 Transceiver Power Figure of Merit

In Chapter 3, we have shown that binary phase shift keying has better modulation power efficiency than frequency shift keying. BPSK requires $3\text{dB}$ less $E_b/N_0$ as compared to coherent FSK and $6\text{dB}$ less $E_b/N_0$ as compared to noncoherent FSK. Previously we have also discussed that FSK transceivers have lower electronics power than PSK transceivers. So which technique is better?

The concept of transceiver power efficiency can offer some intuition in making the choice. Intuitively, an FSK transceiver should be used when the transceiver power efficiency is low, while a PSK transceiver should be used when the transceiver power efficiency is high. This is because at low transceiver power efficiency, overhead dominates, while at high transceiver power efficiency, RF transmit power dominates.

Unfortunately, transceiver power efficiency alone is not enough to make the decision. This is because the transceiver power efficiency only defines the ratio of RF transmit power to transceiver electronics power and does not allow the comparison of $E_b/N_0$ across different modulation and coding techniques. To do so we introduce the concept of transceiver power
figure of merit (FOM), which is defined as

\[ \gamma_P = \frac{E_b}{N_0 \eta_P} \]  

(4.19)

For a transceiver with high power efficiency, the FOM approaches the theoretical \( E_b/N_O \) requirement. For a transceiver with low power efficiency, the transceiver power efficiency is essentially an overhead cost on top of the \( E_b/N_O \).

Figure 4-12 shows the transceiver power FOM comparison between binary FSK and binary PSK transceivers as a function of data rate. At low data rate, the transceiver power efficiency is low, so both transceivers have a large overhead cost relative to \( E_b/N_O \). BFSK is a little bit better than BPSK due to its lower electronics power consumption. As the data rate increases, the transceiver power efficiency increases, and the FOM improves. At high data rate, the FOM approaches the theoretical \( E_b/N_O \). Note that the BPSK transceiver becomes superior for data rates above 30Mbps because \( E_b/N_O \) is the determining FOM at high transceiver power efficiency. The cross-over point between BFSK and BPSK depends on the particular system. For example, for a system with longer range, which means greater RF transmit power, and better transceiver power efficiency, the cross-over point would occur earlier.

![Figure 4-12: BFSK and BPSK transceiver power FOM comparison.](image)
Figure 4-12 shows clearly that most of the gains in transceiver power FOM comes from increasing the data rate rather than from choosing between FSK and PSK. Thus, the transceiver power FOM provides a means to allow the designer to evaluate the combined effectiveness of system, architecture, and circuit techniques.

4.6 Modeling Assumptions and Limitations

Any model has assumptions and limitations and it is important to understand them. The most important assumption made in this work is that the system under investigation has short transmission range. Because of this, the RF transmit power is low as compared to the transceiver electronics power, and this allows optimization opportunities. The result of the optimization will vary from system to system depending on the transmission range and the RF transmit power requirement.

Another important assumption is that the system transmits and receives fixed amount of data (i.e. 100-bit packets) during each operation cycle regardless of the transceiver data rate. The problem would have been different if the sensor transceiver was required to operate continuously.

As the data rate increases, the power consumption of the transceiver will increase. So what are the limitations on the data rate? Here we discuss the effect of data rate on the major circuit building blocks.

Increasing the data rate requires the PA to produce higher RF output power. The RF output power scales linearly with the data rate and is equal to 1mW per 1Mbps for the microsensor network. Currently, integrated Giga-Hertz PAs are capable of producing RF output power levels on the order of a few hundred milli-Watts or greater [47,48], which implies that they can handle data rates up to 100Mbps for a 10-meter range. Therefore, the PA RF output power requirement is not a bottleneck. What can be a challenge is maintaining high PA efficiency at high output power levels.

As the data rate increases, noise in the transceiver electronics may increase to a level that the system is no longer thermal-noise limited. In particular, the integrated in-band local oscillator phase noise needs to be kept below the SNR requirement of the system, which is 21dB. The synthesizer integrated phase noise for 802.11a transceivers are consistently reported to be below -35dBc for a bandwidth of 10MHz around the carrier [49,50], which
is well below the required SNR.

The power consumption of the DSP in general scales linearly with the frequency of operation. However, note that in our application, the amount of data to be processed is fixed, so the DSP can operate at a lower rate as compared to the RF transceiver. Therefore, the speed of the DSP does not need to change with the data rate of the transceiver.

From above, we can conclude that the limitation from the circuit electronics is not severe and the data rate can go up to several 10's of Mega-bits per second without incurring excessive increase in power consumption. The limitations come first from the system side, in particular from the coherence bandwidth, which is between 5-10MHz for a heavily obstructed indoor environment [36]. Extending the bandwidth much beyond the coherence bandwidth can incur severe BER degradation due to frequency selective fading [29]. Therefore, a bandwidth of 10MHz is already at the upper limit for a single-carrier system.

4.7 Summary

There are several interesting results that have come out of this chapter. First, the concept of transceiver power efficiency is introduced. The transceiver power efficiency is more general than the modulation power efficiency in that it takes into account the power consumption of the transceiver electronics. It is shown that short range transceivers have low transceiver power efficiency due to fixed energy cost from the RF building blocks required to perform up and down conversions. The transceiver power efficiency can be improved significantly by increasing the data rate, reducing the start-up time, and improving the PA efficiency. Increasing the data rate drives down the fixed energy cost of the transceiver. Reducing the start-up time decreases the start-up energy overhead. Improving the PA efficiency lowers the energy per bit cost of the power amplifier.
Chapter 5

Voltage Controlled Oscillator Design

In the previous chapter, it was shown that the fixed energy cost due to the RF electronics can be driven down by increasing the data rate. At high data rate, the two important circuit building blocks are the local oscillator and the power amplifier. Since the local oscillator is on during the start-up, receive, and transmit modes, it occupies a large fraction of the overall energy budget. Within the local oscillator, the two most power hungry components are the voltage controlled oscillator and the frequency divider, with the VCO accounting for more than 50% of the total power consumption. Therefore, designing a low power VCO is an important task.

This chapter focuses on the design of low power voltage controlled oscillators. First, the system requirement of VCO phase noise is analyzed. We show the dependencies of phase noise on data rate and bandwidth. Then the design and measurement results of a 5-GHz VCO is presented. We show that the power consumption of the VCO can be reduced at the lower phase noise requirement.

5.1 Phase Noise Definition

An ideal oscillator produces an output sinusoidal signal

\[ V(t) = A \cos(\omega t + \phi) \]  (5.1)
In a practical oscillator, however, noise corrupts both the amplitude and the phase of the signal. The amplitude noise can be effectively eliminated by amplitude limiting, but there is no good way to remove the phase noise $\phi(t)$. Therefore, minimizing the phase noise has been the most important design criterion for a VCO.

The phase noise is defined as

$$\mathcal{L}(f_m) = 10 \cdot \log \left[ \frac{P_{\text{noise}}(f_o + f_m, 1\text{Hz})}{P_{\text{sig}}} \right]$$  \hspace{1cm} (5.2)$$

where $P_{\text{noise}}(f_o + f_m, 1\text{Hz})$ is the single side-band power at a frequency offset of $f_m$ from the carrier in a unit bandwidth of 1Hz [51]. $P_{\text{sig}}$ is the signal power and is equal to

$$P_{\text{sig}} = \left( \frac{A/\sqrt{2}}{R} \right)^2$$  \hspace{1cm} (5.3)$$

where $A$ is the signal amplitude and $R$ is the tank impedance at resonance. Since the phase noise is measured with respect to the signal power in an unity bandwidth, it has a unit of dBC/Hz.

A semi-empirical formula for phase noise is first derived by Leeson [52]

$$\mathcal{L}(f_m) = 10 \cdot \log \left\{ \frac{2kTRF}{A^2} \left( 1 + \frac{f_o}{2Qf_m} \right)^2 \left( 1 + \frac{\Delta f_{1/f^3}}{|f_m|} \right) \right\}$$  \hspace{1cm} (5.4)$$

As shown in Figure 5-1, the phase noise has three distinctive regions: a $1/f^3$ region due to up-conversion of device $1/f$ noise, a $1/f^2$ region due to thermal noise from both the resonate tank and the active devices, and a noise floor. Leeson’s formula does not provide means of determining the noise factor $F$ or the flicker noise corner frequency $\Delta f_{1/f^3}$.

### 5.2 VCO Phase Noise Requirement

VCO phase noise affects system performance in two ways. The phase noise that falls within the channel may degrade the SNR of the transmitted signal. Fortunately, this in-band phase noise can be suppressed by the loop response of the frequency synthesizer. The more difficult problem is the far-off phase noise, which can degrade the SNR of an adjacent channel.

This problem is illustrated in Figure 5-2. Assuming that two sensor nodes are transmitting at the same time on channels 1 and 2, and that sensor 1 is farther away from the base
station than sensor 2, then the base station receives more power from sensor 2. If the power received from sensor 2 is much greater than that from sensor 1, then the far-off phase noise of sensor 2 can corrupt the SNR of sensor 1.

Assuming that the far-off phase noise coming from channel 2 is the dominate noise source, then the SNR of channel 1 can be expressed as

\[
SNR = \frac{\int_{\Delta f} S_1(f) df}{\int_{\Delta f} P_{S2} C_2(f) df}
\approx \frac{P_{S1}}{P_{S2} \cdot C_2(f_m) \cdot \Delta f}
\]

\[ (5.5) \]
\[ (5.6) \]

Figure 5-2: Effect of adjacent channel phase noise on SNR.
\[ L_2(f_m) - \Delta P \cdot \Delta f \]

where \( S_1(f) \) is the power spectral density of the channel 1 signal, \( P_{S_1} \) and \( P_{S_2} \) are the signal powers, and \( L_2(f_m) \) is the normalized phase noise of channel 2 at \( f_m \), which is the frequency separation between channels 1 and 2. Simulation shows that the approximation made to the phase noise integration in Equation (5.6) has an error of less than 2dB.

Equation (5.7) makes intuitive sense: the SNR is inversely proportional to the amount of phase noise, the size of the interferer, and the channel bandwidth. Given an SNR, the required phase noise can be computed as follows

\[
L_2(f_m) = -(SNR|_{dB} + \Delta P|_{dB} + \Delta f|_{dB})
\]

In the above equation, \( \Delta P \) depends on the multi-access protocol used for the network. Assuming that the base station has a high quality VCO, then we only need to worry about the phase noise of the sensor node transmitters. This phase noise could cause SNR degradation when the sensor nodes are transmitting/receiving at the same time on adjacent channels. Figure 5-3 illustrates the possible scenarios.

![Figure 5-3: Multi-access protocol affects phase noise requirement.](image)

When sensors A and C are both transmitting to base station BS1, the phase noise of node C can degrade the signal from node A due to the near-far problem. A similar problem
occurs when node A is transmitting while node B is receiving. These scenarios can be avoided if pure time division multiplexing is used inside the cell where only one sensor is active at any instant of time. However, when the number of sensors are greater than the number of available time slots, multiple frequency channels must be used. Even if pure TDMA could be used, it does not solve the problem of interferences caused by adjacent cells. As shown in the figure, this can occur when sensor B is transmitting to BS1 while sensor D is attempting to receive from BS2. The SNR of the signal that sensor D is receiving can be corrupted by the phase noise of the signal that sensor B is transmitting.

Consider the adjacent cell scenario described above, and assume that sensor B transmits at an output power level $P_B$, the signal power that arrives at sensor D is

$$P_{S2} = P_B|dB - PL(\Delta d)|dB$$

(5.9)

where $PL(\Delta d)$ is the link loss from sensor B to sensor D and $\Delta d$ is the distance between the two sensors. The power that sensor D receives from BS2 is

$$P_{S1} = P_{BS2}|dB - PL(d)|dB$$

(5.10)

where $P_{BS2}$ is the output power from BS2, $PL(d)$ is the link loss between sensor D and BS2 with $d$ being the distance between them. Therefore, the fading margin $\Delta P$ can be express as

$$\Delta P = P_{S2}|dB - P_{S1}|dB$$

(5.11)

For a $\Delta d$ of 1 meter, a path loss exponent of 3, and similar $P_B$ and $P_{BS2}$, $\Delta P$ is $30dB$.

The phase noise requirement is usually specified at a fixed reference offset frequency from the carrier regardless of the channel bandwidth. For example, a reference offset frequency of 1MHz is commonly used. For a data rate of 1Mbps, a channel bandwidth of 1MHz, and an SNR of $21dB$, the phase noise requirement at 1MHz offset is

$$L(1MHz) = -(SNR|dB + \Delta P|dB + \Delta f|dB) = -111dBc/Hz$$

(5.12)

If the phase noise is specified at a different offset frequency $f_m$, then the phase noise at
1MHz can be approximated as

\[ \mathcal{L}(1\text{MHz}) = \mathcal{L}(f_m) + k \cdot 10 \cdot \log\left( \frac{f_m}{1\text{MHz}} \right) \tag{5.13} \]

where \( k \) equals to 2 in the \( 1/f^2 \) region and equals to 3 in the \( 1/f^3 \) region.

For binary modulations, the channel bandwidth \( \Delta f \) is approximately equal to the data rate \( R \). In addition, if the channels are packed as tightly as possible, then the channel bandwidth \( \Delta f \) is also equal to the channel separation \( f_m \). If the data rate within the system (i.e., of the sensor node and the base station) is increased, then \( \Delta f \) and \( f_m \) need to be increased correspondingly. However, \( \Delta P \) does not change because both \( P_B \) and \( P_{BS2} \) are increased in proportion to the data rate to maintain a constant SNR and BER. By examining Equations (5.8) and (5.13), it is clear that increasing the data rate decreases \( \mathcal{L}(f_m) \) at 10dB per decade, but the phase noise at \( \mathcal{L}(1\text{MHz}) \) increases at \( k \cdot 10\text{dB} \) per decade relative to \( \mathcal{L}(f_m) \). The net result is that the phase noise increases at \( (k-1) \cdot 10\text{dB} \) per decade. Figure 5-4 plots the phase noise requirement as a function of the data rate assuming that the phase noise between 1MHz and \( f_m \) lies within the \( 1/f^2 \) region. As seen in the figure, the phase noise requirement is reduced from about -111dBc/Hz to -101dBc/Hz as the data rate is increased from 1Mbps to 10Mbps. Note that if the phase noise lies within the \( 1/f^3 \) region, then the reduction in the phase noise requirement is even greater.

### 5.3 Integrated LC Oscillators

The previous section derived the phase noise requirement for the voltage controlled oscillator on the sensor node. This section investigates the design of integrated oscillators. The focus is on oscillators that rely on an LC tank as the resonate element. LC oscillators tend to produce good phase noise and are relatively easy to integrate at Giga-Hertz carrier frequencies.

#### 5.3.1 Operation Principles

An integrated LC oscillator is based on the oscillation of an LC tank, as shown in Figure 5-5. The core of the tank consists of an inductor \( L \) and a capacitor \( C \). Given a non-zero
Figure 5-4: Increasing the data rate reduces VCO phase noise requirement.

initial condition, the tank oscillates at the natural frequency

$$\omega_0 = \frac{1}{\sqrt{LC}}$$  \hspace{1cm} (5.14)

A common rule of thumb is that a 1nH inductor and a 1pF capacitor resonates at 5GHz. Fortuitously, these inductor and capacitor values fall within the capabilities of modern fabrication technologies.

Figure 5-5: LC oscillator with negative resistance.

Since a real L or C is always associated with a parasitic resistance R, this parasitic resistance would cause the oscillation amplitude to decay over time. Therefore, a negative-R circuit is needed to replenish the energy lost due to the parasitic resistance. A simple and
An elegant technique to generate the negative resistance is shown in Figure 5-6. The LC tank consists of a pair of inductors and varactor diodes. The negative resistance is generated through a cross-coupled pair of transistors with a tail current source. A simple calculation shows that the impedance of the negative resistance seen by the tank is $-2/g_m$, where $g_m$ is the transconductance of the cross-coupled pair transistors.

![Figure 5-6: negative-gm VCO.](image)

The basic operation of the negative-gm oscillator can also be explained via the feedback model. The output voltages $V_{o+}$ and $V_{o-}$ are related by

$$V_{o+} = -g_m \cdot Z \cdot V_{o-}$$  \hspace{1cm} (5.15)

and

$$V_{o-} = -g_m \cdot Z \cdot V_{o+}$$  \hspace{1cm} (5.16)

where $Z$ is the single-ended impedance seen at either $V_{o+}$ or $V_{o-}$. Therefore, the feedback model of the circuit is simply as shown in Figure 5-7, and the expression for $V_{o+}$ is

$$V_{o+} = \frac{1}{1 - (g_m \cdot Z)^2}$$  \hspace{1cm} (5.17)
This circuit oscillates when $Z = 1/g_m$, which only occurs at the resonate frequency where the impedance $Z$ is purely real. In practice, it is desirable to make $g_m$ of the cross-coupled pair transistors 2-3 times larger than the desired negative conductance to ensure oscillation start-up. The non-linearity of the transistors will keep the average negative resistance at the desired value.

$$\begin{align*}
V_{o+} & \quad -g_m Z \\
& \quad -g_m Z \\
& \quad V_{o-}
\end{align*}$$

Figure 5-7: VCO feedback loop.

### 5.3.2 Phase Noise Mechanisms

Although the physical mechanisms of oscillator phase noise are still not completely understood, recent advances in oscillator research have generated much insight and provided practical guidelines to the design of LC tuned oscillators. There are three main contributors to the phase noise of the negative-gm VCO shown in Figure 5-6. The first comes from the thermal noise of the tank impedance, $4kTR$. This noise is shaped by the tank impedance and modulates the zero-crossing instants of the differential pair [53].

The second noise source comes from the tail transistor. This noise is commutated by the switching pair like in a single-balanced mixer and contributes to phase noise via two mechanisms. Noise close to DC is up-converted to the carrier frequency in the form of AM noise, which then modulates the phase delay of the positive feedback loop and appears as phase noise through the process of AM-PM conversion [54]. Jerng showed that this noise can be reduced by increasing both the linear range and $f_T$ of the switching pair, which can be accomplished by decreasing the width of the switching device [55]. In addition, the noise close to the second harmonic can be significant since it is down-converted to the fundamental like in a mixer [56].

The third noise source comes from the differential pair. The thermal noise of the switching pair is converted to phase noise in a similar manner as in the case of the tank impedance. The flicker noise of the switching pair appears as phase noise through AM-PM conversion
in a similar manner as in the case of the tail transistor [55].

Figure 5-8 shows phase noise and tank amplitude as a function of bias current $I$. According to Equation (5.4), the phase noise is inversely proportional to the square of the tank amplitude $A$. At low bias current, $A$ is proportional to the bias current, so the phase noise improves as the bias current is increased. However, the tank amplitude is eventually limited by the supply voltage, and increasing the bias current beyond this point does not change the tank amplitude. Since the noise still increases as a function of $I$, the phase noise degrades. This implies that the best phase noise is achieved at the bias current that is just enough to drive the tank amplitude to its maximum [53].

![Figure 5-8: Qualitative plot of phase noise and tank amplitude as a function of bias current.](image)

### 5.4 LC Oscillator Design

An LC oscillator is designed and fabricated in a 0.18-μm SiGe BiCMOS process. The VCO core with device parameters is shown in Figure 5-9. This VCO is designed to operate at low supply current, and its phase noise versus bias current characteristics is investigated.

#### 5.4.1 LC Tank

The LC tank consists of a pair of spiral inductors and p-n junction varactors. According to simulation, the varactors have Q values on the order of 100 at 5GHz, so the Q of the tank is limited by the inductors. The Q of the inductors is shown in Figure 5-10. As expected, the
Q increases linearly as a function of frequency at low frequencies, but it degrades at high frequencies due to several mechanisms. First, the skin effect pushes the current toward the outside surface of the conductor, which effectively reduces the cross-sectional area of the conductor. Second, an inductor with multiple spirals have the current flowing in the same direction in each spiral. However, the magnetic field associated with the current in each spiral induces a current on the surface of the adjacent spiral that moves in the opposite direction as the signal current. Therefore, the signal current that goes through the inductor is reduced. This effect can be mitigated by increasing the distance of separation between the adjacent spirals. Third, the dielectric between the top-metal inductor and the bottom substrate looks like a capacitor, which means that electric loss occurs through the substrate at high frequencies. This effect can be minimized by providing either a low (substrate ground) or high (deep trench) impedance termination above the substrate [57]. Fourth, the magnetic field associated with the inductor induces eddy currents in the substrate and creates energy loss in the inductor. This effect can be reduced by slotting the substrate. Yue showed that this can also be accomplished by placing a slotted ground metal pad under
The parallel parasitic resistance of an inductor $L$ is approximately

$$ R = \omega_0 \cdot L \cdot Q $$

Substituting this expression into Equation (5.4), we see that the phase noise is related to $L$ and $Q$ in the $1/f^2$ region as

$$ \mathcal{L}(f_m) \propto 10 \cdot \log \left( \frac{1}{L \cdot Q^3} \right) $$

Thus the quantity $L \cdot Q^3$ should be maximized to achieve the best phase noise. Note that this is not equivalent to maximizing the amplitude of the signal $A$, which is proportional to $L \cdot Q$.

Figure 5-10 shows that the $Q$ value of the 1.1nH inductor is about 20 at 5GHz. This yields a $R$ of roughly 700$\Omega$, which means the $g_m$ of the crossed-couple pair has to be greater than 1.4mA/V. However, this only takes into account the unloaded $Q$ of the inductor. The loaded $Q$ will be invariably worse due to other parasitics in the tank. Adding a factor of 2-3 to ensure proper start-up condition, the lower bound of the switching pair $g_m$ should be set at 6mA/V or higher.
5.4.2 Cross-Coupled Pair

The sizing of the cross-coupled pair devices demands a trade-off between phase noise and \( g_m \). Figure 5-11 shows phase noise and the switching device \( g_m \) as a function of the switching device width at a bias current of 1.5mA. The phase noise improves at smaller device width because the phase noise originating from the tail transistor through AM-PM conversion is reduced. However, the \( g_m \) decreases as the width scales down as well. In order to ensure proper start-up condition, the \( g_m \) needs to be greater than 6mA/V, which means that the devices need to have a width of 20\( \mu \)m or larger.

![Figure 5-11: Simulated phase noise and \( g_m \) as a function of switching device width.](image)

5.4.3 Tail Transistor

The tail transistor provides two functions. First, it supplies the supply current to the VCO core. Second, it presents a high impedance node to the LC tank. Due to the large signal switching action of the cross-coupled pair, each of the switching transistors alternately enters the triode region. When this happens, the LC tank sees a low impedance into the switching transistor, and the Q of the tank could be degraded without the stiff tail transistor.
5.4.4 Phase Noise Performance

Figure 5-12 shows measured phase noise performance as a function of the core bias current. The data shows good agreement with theory. At low current levels, the phase noise improves at 6dB/Octave as the current is increased. This is the current-limited regime where the signal amplitude grows in proportion to the bias current. At high current levels, however, the signal amplitude is limited by the supply voltage, and the phase noise degrades due to the rise in the noise factor. The best phase noise performance is measured at -125dBc/Hz at a bias current of 7mA. The lowest supply current that allows the proper start-up of the VCO is 1.5mA, and the phase noise is better than -110dBc/Hz. The oscillation frequency is measured to be at 5.3GHz.

![Figure 5-12: Measured phase noise as a function of bias current.](image)

In Section 5.1, we have shown that the phase noise required for a data rate of 10Mbps is only -101dBc/Hz, so the phase noise at the lowest supply current still far exceeds this requirement. This means that there are still room to push the supply current further down. At the expense of higher phase noise, the cross-coupled devices can be made wider to achieve the same $g_m$ at lower supply current.
5.4.5 Die Photo

Figure 5-13 shows the die photo of the oscillator. The die is measured at 0.75mm x 1mm and is bondpad limited. Other than the bondpads, most of the die area is taken up by the two inductors at the bottom-center of the chip.

Figure 5-13: Die photo of the oscillator.

5.5 Summary

In this chapter, voltage controlled oscillators are studied at both the system and circuit levels. At the system level, the VCO phase noise requirement is derived. It is shown that increasing the data rate reduces the phase noise requirement of the VCO. At the circuit design level, a 5GHz LC VCO is designed and measured. This VCO has a phase noise of -125dBc/Hz at 7mA bias current and -110dBc/Hz at 1.5mA. A technique is proposed to trade off optimal phase noise for a lower start-up bias current through the sizing of the switching transistors. This design shows that there is still enough phase noise headroom to lower the supply current even further.
Chapter 6

Conclusions

This work describes the modeling and design of a short-range RF transceiver for industrial microsensor applications. In communication systems, the signal-to-noise ratio per bit, $E_b/NO$, is used to characterize the received signal power required to achieve a desired bit-error rate. The RF transmit power requirement can then be derived based on both $E_b/NO$ and link budget. However, this does not take into account the transceiver electronics power, which is dominant in short-range RF transceivers. In this case, standard communication techniques, such as modulation, coding, and diversity, have negligible effect on the total transceiver power consumption. There is no clear strategy how to minimize the physical layer energy cost and no metric to evaluate the energy cost of different transceivers.

A new concept, the transceiver power efficiency, is introduced in this work to characterize the overhead energy cost due to the transceiver electronics. It is defined as a ratio between the communication cost and the circuit electronics cost. We show that short-range low data rate transceivers have very low transceiver power efficiency, which means that the energy cost of the electronics dominates over the energy cost required for RF transmission. We propose a clear strategy to lowering the energy cost of the transceiver: the transceiver power efficiency should be maximized.

The transceiver power efficiency is improved through increasing the transceiver data rate. This drives down the fixed energy cost of the transceiver electronics. At high data rate, reducing the start-up time and improving the PA efficiency further increase the battery life. As compared to a Bluetooth transceiver that has a data rate of 1Mbps, a start-up time of 120μs, and a class-A PA with 10% efficiency, increasing the data rate to 10Mbps, reducing
the start-up time to 20μs, and using a class-E PA with 70% efficiency will increase the battery life by a factor of 8.

A simple transceiver energy model is developed to study the transceiver energy cost. This model takes into account the energy consumption of the transceiver in the start-up, receive, switch, and transmit modes. This model confirms that, at high data rate, the transceiver energy is dominated by the start-up and the PA energy costs, and therefore circuit design emphasis should be placed on the local oscillator and the power amplifier. This result is not apparent without the proper modeling of the transceiver energy.

The result that high data rate improves the battery life also comes as a surprise. It is commonly believed that low power transceivers should employ low data rate to reduce the SNR requirement. For example, the recent Zigbee/IEEE802.15.4 standard, which is aimed toward low power low duty-cycle applications, chose a maximum data rate of 250kbps despite the fact that multi-Mbps data rate is easily achievable today. Our work suggests that standards such as Zigbee should choose a high data rate to minimize energy cost.

The transceiver power efficiency is a relative figure of merit and can not be used to compare systems with different communication protocols. We develop a general figure of merit, the transceiver power figure of merit, that includes the effect of both communication protocols and transceiver electronics. This simple figure of merit can be used to compare the energy cost of RF transceivers with different system and circuit architectures.

In recent years, Giga-Hertz RF transceivers have reported power consumption figures as low as a few 10's of milli-Watts [11, 12]. It is therefore increasingly difficult to reduce this power consumption figure by a significant amount. In this work, it is shown that substantial performance improvement can still be attained through joint optimization of system and circuit design. This approach can reduce the energy cost of the transceiver significantly without placing demanding specifications on the RF transceiver circuitry. In fact, system modeling should make circuit design more focused on the important circuit blocks, if it doesn’t make the specifications easier to achieve.

A power hungry transceiver circuit building block is the voltage controlled oscillator. A part of this work investigates the design of integrated LC oscillators on both the system and circuit levels, focusing on the low power end. On the system level, the phase noise requirement of the VCO is derived. On the circuit level, the physical mechanisms of phase noise are examined and a 5GHz VCO is designed and measured. It is shown that the phase
noise performance of the VCO far exceeds the phase noise requirement, and therefore power consumption can be reduced by further increasing the switching transistor width.

6.1 Future Works

There are several challenging and relevant design areas not covered in this work. First, the limitation on data rate comes not from circuit design but from system constraint. The coherence bandwidth at Giga-Hertz frequency is reported to be only between 5-10MHz in a heavily obstructed indoor environment, so the first problem that a high data rate system encounters is BER degradation due to frequency selective fading. For this reason, a data bandwidth of 10MHz is already at the upper limit for a single carrier system. This problem could be mitigated through either multi-carrier modulation or multi-level PSK/QAM, both of which would increase the power consumption of the transceiver. The overall energy cost could conceivably be reduced if the data rate increases faster than the power consumption. Due to the increased complexity, the analysis on the power consumption of these transceivers could be difficult.

A second challenge for industrial microsensor networks is to attain adequate diversity. In general, diversity can be realized through frequency, time, or space. The microsensor node transmits and receives very short bursts of data, so frequency hopping within packet transmission is not possible. For applications with stringent latency requirement, such as the one described in this work, time diversity may cause too much delay to be acceptable. This leaves only spatial diversity. Placing multiple antennas on the base station can certainly improve the quality of the uplink, but placing multiple antennas on the microsensor receiver could increase the power consumption of the sensor node significantly. Therefore, finding a power-efficient diversity technique for the downlink is an interesting and important problem.

In the circuit design arena, in addition to the local oscillator and the power amplifier, there are two other circuit blocks that deserve close attention. The first is the microsensor receiver demodulator. Demodulation is more involved than modulation, so the demodulator must be designed with care in order to keep the power consumption low. In addition, the complexity of the demodulator could grow fast for more sophisticated modulation techniques. The second is the DC/DC converter. It is important to realize that the load profile of the battery is determined by the operation of the DC/DC converter rather than by the
current drawn from the transceiver chip. Thus the DC/DC converter needs to be designed to have both high efficiency and low instantaneous discharge rate.
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