Micropower Impulse Radio for Remote Controlled Insect Flight

by Keith G Lyon

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MICROPOWER IMPULSE RADIO FOR REMOTE CONTROLLED INSECT FLIGHT

A Dissertation
Presented to the Faculty of the Graduate School
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Doctor of Philosophy

by
Keith G. Lyon
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Insects have remarkable strength and stamina compared to their body mass and fly and maneuver effortlessly in ways that are impossible for present day robotic flyers. Therefore, efforts to control and direct flying insects for our own purposes have a huge potential payoff. One such effort, discussed in this dissertation, concerns the control of a *Manduca Sexta* moth by sending commands by radio to neural probes implanted in the thorax. The electronics hardware represents a major challenge in itself because the moth can carry only 700 milligrams, most of which is occupied by a small watch-battery.

Ultimately, the moth must carry not only a radio receiver to pick up commands sent by the controller, but also a transmitter to return gathered information and fulfill its mission. Commercial “low-power”, burst-mode radios have proven inadequate because the battery cannot satisfy their peak power consumption. Instead, this dissertation focuses on the development of an alternative “impulse radio”, which consumes power only during the ~100 picosecond interval required to generate a microwave pulse.

The specific transmitter architecture presented here uses a nonlinear transmission line to directly convert digital signals provided by a microcontroller into microwave pulses broadcast by an antenna. This dissertation discusses (1) the background and theory of impulse-radios and (2) nonlinear transmission lines, (3) circuit board prototypes and (4) a CMOS implementation of the trans-
mitter, (5) a study of the wireless link between the moth and its controller, as well as (6) efforts to implement the radio using light-weight, inexpensive plastic and polymer materials, before (7) reflecting on the potential of the new transmitter and possible directions for future work.
BIOGRAPHICAL SKETCH

Keith Lyon hails from the Southwest of the United States. He grew up in New Mexico and attended the University of Wyoming in Laramie where he received the BS in Computer Engineering. His personal quest to understand how computers work took him on a tour through software, digital electronics, analog circuits, and semiconductor device physics before arriving at microwave electronics and ultimately the nebulous field of systems engineering. He is a dedicated runner with a taste for trail races, a chef and gourmand with a flare for chiles, and a literatus who finds infinite amusement from Wodehouse.
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CHAPTER 1
INTRODUCTION

The recent proliferation of mobile devices has generated significant interest among the RF and Microwave communities in radios with small size, reduced power consumption and increased bandwidth. These requirements are often contradictory because circuits operating at higher frequencies, and thus higher bandwidths, require larger quiescent currents to mitigate the effects of noise and parasitics. In response, designers have pursued a class of radios known as Ultra-Wideband Impulse Radio (UWB-IR), which exploit aggressive duty cycle scaling and novel signal synthesis techniques to reduce power. In turn, the FCC has designated spectrum from 3.1-10.6 GHz, as well as a band at 60 GHz for use by unlicensed, low-power UWB devices. A growing body of work devoted to UWB circuits, as well as synchronization, multi-user access, and channel modeling appears in the engineering literature.

The applications targeted by most UWB researchers include cellphone and computer peripherals, as well as household gadgets. In general, such devices are several centimeters in size, with batteries capable of supplying peak currents in the milliamp range. However, there is another class of mobile applications where small size, tight integration, and device weight are of critical importance. The example application considered in this dissertation is a living Manduca Sexta moth whose flight is controlled remotely by a human operator. In this case, the insect can lift no more than one gram of additional mass in sustained and stable flight, placing severe constraints of the energy capacity and output resistance of the battery. While low-power sensors and microprocessors have existed for
some time, the wireless hardware still dominates the power budget.

1.1 Remote Controlled Insect Flight

Micro-air vehicles have recently become a large, forward-looking subfield in the aerospace and robotics communities. The goal of such research is the demonstration of bird or insect sized robots capable of autonomous flight and able to execute “missions” commonly of a military intelligence gathering, or search and rescue nature. Presently, the field is quite advanced, with textbook-level material now becoming available [1, 2, 3]. However, despite this progress, further developments are hindered by incomplete understanding of the biomechanics of flapping-wing flight. Since birds and insects have mastered flight in this regime (high Reynolds number) where fixed-wing aircraft fail, a logical shortcut is to “co-opt” these creatures into fulfilling the missions intended for micro-air vehicles. This approach frees the system designer from the complexity of flight, but still requires the payload electronics to be built with severe power, weight, and size constraints.

From a system perspective, an such an insect cyborg consists of sensors and neural probes which interface the electronics with the moth’s physiology, a controller, and a radio to communicate with the remote “operator”. Fig.1.1 shows a overview of the different subsystems; the biological interface and flight control is described elsewhere [4, 5], while this dissertation focuses on the radio architecture. The electronics are powered by a small battery (which consumes most of the weight budget), possibly in combination with energy-harvesting components. Because the output resistance of these power sources is typically more than 100Ω, supply current is very limited and the radio must operate with low
Figure 1.1: System block diagram for the remote-controlled moth (a) - the focus of this dissertation is on the radio and antenna. The inset shows an adult moth with an unripe tomato for approximate scale. The asymmetrical nature of the radio link is shown in (b).
peak power consumption.

1.2 Existing Low-Power Radio Technology

The first wireless designs for the moth were based on commercially available integrated microprocessor/radio components. While average power calculations indicated that the design should operate for several hours with a 1 mA-hr, 500 mg watch battery, the devices failed when attempting to initialize the wireless link. Some investigation revealed that the radios, based on the commercial ZigBee standard, were intended to operate in bursts, requiring startup, transmit, and shutdown times of several milliseconds each, while consuming close to 50 mW of peak power. Unfortunately, the watch battery was unable to supply the required current without the supply voltage dropping to unacceptable levels. These early experiments underscored the need for a pulsed radio architecture operating with low peak power.

Ultra-Wideband Impulse Radios are ideal for the moth because they consume power only during pulse synthesis and transmission, and have no startup-shutdown transition states. Most low-power architectures are based on digital circuits, either using specially timed chains of inverters to synthesize the pulse, or by generating a large amplitude glitch - a wide bandwidth impulse - with a logic gate. In both cases, the switching time of digital circuits limits the maximum achievable center frequency. While aggressive transistor scaling and careful layout have resulted in small-signal Silicon CMOS circuits operating near 100 GHz, large-signal digital circuits are more limited and suffer increasing power dissipation due to leakage currents as well as higher costs of fabrication.
Therefore, one should not expect scaling alone to enable existing UWB circuits to be extended to ever-higher frequencies. In order to use an efficient antenna under 1 cm in size, the moth’s transmitter must operate above 7.5 GHz, preferably at even higher frequencies to relax design constraints.

Recent UWB transmitters presented in literature focus on the use of digital circuits to reduce power consumption. One such example uses a digitally controlled ring-oscillator to generate bursts of pulses in the 500 MHz bands defined by IEEE 802.15.4a [6]. The energy consumption is about 40 pJ/pulse for bands near 3.5 GHz, but more than doubles to 87 pJ/pulse at 10 GHz. A more recent proposal reduces the energy consumption significantly, to 17 pJ/pulse, but focuses on the lower frequency bands between 3.0 and 5.7 GHz [7]. Other digital pulse generators [8, 9] use chains of specially timed inverters or logic gates to generate a predetermined waveform. The low standby power and fast duty-cycling of these devices are very advantageous, but the response of digital circuits at higher frequencies is poor, making their use difficult at or above 10 GHz.

Another approach [10] uses a switchable microwave oscillator to generate pulses with a well controlled spectrum. While the energy-per-pulse is small, around 18 pJ/pulse, the pulse amplitude is limited by the slow growth from noise of the oscillations. The oscillator may buildup faster by providing more bias current, but the bias current must remain constant for the pulse duration of several nanoseconds so this method is more likely to have peak-power issues than other approaches. Additionally, the minimum current required by an oscillator increases significantly with its center frequency.

A more recent proposal uses edge-combining to generate a short impulse,
followed by a bandpass filter to determine the final pulse spectrum \[11\]. The energy consumption of this device is estimated to be as low as 9 pJ/pulse, but the bandpass filter is challenging to implement on-chip and the bandwidth is ultimately limited by the logic gate used for edge-combining. The circuit also requires about 3 mW of static power, making it impractical for ultra-light applications.

There have been few proposals for UWB-IR devices operating above 10 GHz. In \[12\] an impulse radio based on a switchable oscillator with a fast startup circuit is shown at 30 GHz. This device is based on a compound semiconductor pHEMT technology and has very high power consumption.

Devices operating in the 60 GHz UWB band generally focus on very high data-rates and require power consumption on the order of 100 mW \[13, 14\]. Circuit design is quite challenging at this frequency, making it difficult to justify the engineering effort for applications with sub-Gbps bitrates. However, this band also has the advantage of allowing small, lightweight antennas; a half-wave dipole would be about 2.5 mm in length. A simple, low-power transmitter operating near 60 GHz would be very advantageous for ultra-light wireless applications.

In an attempt to satisfy the conflicting requirements of high frequency and low-power, this dissertation presents an Impulse Radio architecture based on frequency multiplication rather than pulse synthesis. At the core of this architecture is the Nonlinear Transmission Line (NLTL), which is a passive but nonlinear device, allowing processes such as frequency conversion to occur without quiescent power dissipation. The power required by such a transmitter is determined, in the ideal case, by the capacitive charging and discharging of the
NLTL, and decreases as the center frequency increases.

NLTLs have previously been proposed as UWB signal sources because of their ability to generate very short falltimes, 4-10 ps in modern devices \[15\]. Such systems use GaAs-based edge-compression NLTLs which have been engineered to eliminate Bragg ringing and provide a smooth falling edge with minimal overshoot. They are primarily of interest to the UWB radar and Ground-Penetrating radar communities where users seek the maximum possible pulse bandwidth, and the required voltage swings (6 V or larger) are advantageous.

In the context of communications however, regulatory requirements, receiver complexity, and channel modeling must also be considered, so the widest bandwidth pulse is no longer the best solution. In this case, NLTLs with Bragg ringing have been shown to concentrate signal energy in the frequency domain \[16,17\], allowing more control over the output spectrum.

1.3 Concept for a UWB Transmitter

The frequency multiplication, or equivalently, edge-sharpening properties of an NLTL make it a promising interface between low-speed, baseband circuits and the high frequency antenna. Once the characteristics of the driver and antenna are determined, the transmitter is constructed with an NLTL of sufficient length to sharpen the slow, low-frequency content falling edge from the driver to a fast, high-frequency content edge capable of exciting the antenna. The shape of the final transmitted pulse is then determined by the bandpass characteristic of the antenna itself.
Figure 1.2: Diagrams for CMOS-driven NLTL-based transmitters with single-ended (a), and differential (b) configurations.

The basic architecture is shown in Fig. 1.2 with both single-ended and differential configurations. The differential NLTL appears advantageous because it allows both positive and negative voltages to be applied easily, however the additional inductors roughly double the layout area. Further, in a bulk CMOS technology, the varactor’s nwell-to-bulk capacitance must be (dis)charged by the driver, limiting the rise/fall times and preventing the line from ringing. Therefore, we refer to the single-ended configuration in the remainder of this paper. In order to generate the negative voltages from a simple inverter driver, the varactor nwells are biased at $V_{DD}/2$. The driver circuit is a CMOS inverter sized such that its dynamic output impedance is matched to the NLTL.

The complete transmitter broadcasts a pulse on the rising-edge of its input, which is compatible with the digital-logic levels of the same technology. Data may be encoded with On-Off Keying (OOK), Pulse-Position Modulation (PPM), or similar schemes by low-power digital circuits prior to the NLTL driver. For example, Return-to-Zero (RZ) coded data will be transmitted as an OOK pulse.
sequence. The designer has considerable flexibility to select the coding and pulse rate in a final design. The remainder of this dissertation describes the theory, design, and testing of the NLTL-based UWB-IR transmitter.
CHAPTER 2
NLTL THEORY

The nonlinear transmission line can be understood from the circuit model of a conventional transmission line. Here, a ladder of identical series inductances and shunt capacitances is used to represent the signal delay of a lossless, dispersion-less line. In the case of the NLTL, the capacitance or inductance is allowed to vary as a function of the voltage or current, respectively. Such a transmission line is “non-linear” in the sense that two input waveforms, identical except for their amplitude, may result in drastically different output waveforms. For the purposes of constructing a UWB transmitter, the most important properties of the NLTL are the edge-sharpening effect and Bragg cutoff frequency.

2.1 Lumped-Element Model

The lumped-element model considers the NLTL as a ladder of discrete inductors and capacitors, as shown in Fig.2.1a. Following the classic derivation of wave propagation on normal transmission lines, the NLTL has the parameters

\[ Z_0 = \sqrt{\frac{L}{C(V)}} \]  
\[ \beta = \omega \sqrt{LC(V)} \]

Figure 2.1: Circuit models for lumped-element and semi-discrete NLTLs.
in the low-frequency regime, where \( Z_0 \) is the characteristic impedance, \( \beta \) is the propagation constant, and \( \omega \) is the frequency of interest. Because the capacitance varies with voltage, it is clear that the characteristic impedance and phase velocity do so as well. Thus the waveform will be distorted because its crests and troughs propagate at different speeds. The process in Fig. 2.2 shows the trough of the wave at \( t_2 \) propagating faster than the crest of the wave at \( t_1 \). The falltime, \( t_2 - t_1 \), is reduced, “sharpening” the edge.

The reduction in falltime can be estimated by considering the total difference in velocity of the maxima and minima of a given waveform. Neglecting dispersion, the velocity is given by

\[
\nu = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC(V)}}
\]  (2.3)

and therefore the total sharpening is

\[
T_s = \sqrt{LC(V_H)} - \sqrt{LC(V_L)}
\]  (2.4)
for a single “section” or unit-cell of lumped elements. The resulting falltime after \( N \) sections is therefore

\[
\tau = \tau_0 - N T_s
\]  

(2.5)

where \( \tau_0 \) is the input falltime.

Given a long NLTL, Eq. (2.5) predicts an infinitely sharp edge (zero falltime), and even negative falltimes, which cannot be physically realized. This difficulty is resolved by considering the low-pass filter characteristic of the NLTL, which limits the ultimate falltime to some non-zero value. In the case of a continuous line, this low-pass characteristic follows from high-frequency losses and dielectric relaxation. However, for an “artificial” line constructed from an actual lattice of discrete inductors and capacitors, the cutoff frequency can be derived without such considerations.

The complete dispersion and impedance relations for the lumped element model can be found from the ABCD matrix of the unit cell in Fig. 2.1b,

\[
\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & j \omega L/2 \\ 0 & 1 \end{bmatrix} \cdot \begin{bmatrix} 1 & 0 \\ j \omega C & 1 \end{bmatrix} \cdot \begin{bmatrix} 1 & j \omega L/2 \\ 0 & 1 \end{bmatrix}
\]  

(2.6)

Seeking traveling wave solutions, the determinant

\[
\begin{vmatrix} A - e^{-\gamma} & B \\ C & D - e^{-\gamma} \end{vmatrix} = 0
\]  

(2.7)

giving the general dispersion and impedance relations as

\[
\cosh \gamma = \frac{A + D}{2}
\]  

(2.8)

\[
Z_B = \sqrt{\frac{B}{C}}
\]  

(2.9)

which in the case of the lumped element cell reduce to

\[
\cos \beta = 1 - \frac{1}{2} \omega^2 LC
\]  

(2.10)
\[ Z_B = \sqrt{\frac{L}{C} - \frac{1}{4} \omega^2 L^2} \]  

(2.11)

The cutoff frequency can be found from the conditions \( \beta = \pi \) and \( Z_B = 0 \), which gives

\[ \omega_B = \frac{2}{L C} \]  

(2.12)

where \( \omega_B \) is the Bragg frequency, so called because of the half-wavelength per section condition. One may therefore expect a lumped-element NLTL to support falltimes no shorter than

\[ \tau_{\text{min}} = \pi \sqrt{L C} \]  

(2.13)

being the reciprocal of the Bragg frequency.

The lumped element model captures the phenomena essential to the UWB transmitter design, and has been the dominant model used in literature due to its mathematical tractability.

### 2.2 Semi-Discrete Model

Despite the success of the lumped-element model, in practice, NLTLs often use a section of conventional transmission line as the series inductance. This configuration may be regarded as a transmission line periodically loaded with voltage-variable capacitors (varactors), forming the unit cell in Fig 2.1b. Thus, only the varactors are lumped, or discrete, elements, while the rest of the device is a distributed, periodic structure.

These NLTLs can be modeled with the lumped-element method by replacing the sections of conventional transmission line with their lumped-element
where $\xi = d/\nu$ is the delay time of the transmission line section. The approximation is valid for long wavelengths, specifically $d < \lambda/8$. However, at higher frequencies, significant discrepancies occur between the lumped-element and semi-discrete models, in particular the observed Bragg frequency is greatly underestimated by the lumped-element case [18].

A more accurate model can be derived by returning to the ABCD matrix for the semi-discrete unit cell, which is the matrix product,

$$
\begin{bmatrix}
\cos \frac{1}{2} \xi \omega & j Z_0 \sin \frac{1}{2} \xi \omega \\
-j Y_0 \sin \frac{1}{2} \xi \omega & \cos \frac{1}{2} \xi \omega
\end{bmatrix}
\begin{bmatrix}
1 & 0 \\
-j \omega C & 1
\end{bmatrix}
\begin{bmatrix}
\cos \frac{1}{2} \xi \omega & j Z_0 \sin \frac{1}{2} \xi \omega \\
-j Y_0 \sin \frac{1}{2} \xi \omega & \cos \frac{1}{2} \xi \omega
\end{bmatrix}
$$

where $Y_0 = 1/Z_0$. Proceeding as in Section 2.1 results in the dispersion relation and Bloch impedance

$$
\cos \beta = \cos \xi \omega - \omega C Z_0 \sin \xi \omega \\
Z_B = Z_0 \sqrt{\frac{2 \sin \xi \omega + \omega C Z_0 (\cos \xi \omega - 1)}{2 \sin \xi \omega + \omega C Z_0 (\cos \xi \omega + 1)}}
$$

Previous authors resort to simulation at this point, which is sufficient for analysis, but not for the design of a semi-discrete NLTL with a specific cutoff frequency. Further simplification is possible by considering the two different time-scales on the NLTL: the transmission line delay, and the varactor charging time,

$$
\xi = d/\nu \\
\tau = C Z_0
$$
The above equations now have the dimensionless parameters

\[ k = \frac{\tau}{\xi} \quad (2.21) \]
\[ x = \xi \omega \quad (2.22) \]

and reduce to the simpler forms

\[ \left(\frac{Z_0}{Z}\right)^2 = 1 + k \quad (2.23) \]
\[ \frac{1}{2} k x = \cot \frac{x}{2} \quad (2.24) \]

To design an NLTL with a given Bragg frequency and impedance, one need only choose the parameter \( k \); the capacitance, spacing, and base impedance are given by

\[ C = \frac{k x}{Z_0 \omega_B} \quad (2.25) \]
\[ d = \frac{x}{\omega_B \nu} \quad (2.26) \]
\[ Z_0 = Z \sqrt{k + 1} \quad (2.27) \]

which provides a simple and accurate design procedure for this type of NLTL.

**Example Design 1:** As an example, a large but realizable \( k \) might be close to 10, requiring a 165.8 \( \Omega \) base transmission line to make a 50 \( \Omega \) NLTL. The parameter \( x \) can be computed immediately from Eq. (2.24) once \( k \) has been chosen, in this case \( x \approx 0.622 \). The required capacitance and delay, given a Bragg frequency of 10 GHz are then

\[ C' = \frac{10 \cdot 0.622}{50 \Omega \cdot 2\pi \cdot 10 \text{ GHz}} \approx 1.979 \text{ pF} \]
\[ \xi = \frac{x}{\omega_B} = \frac{0.622}{2\pi 10 \text{ GHz}} \approx 9.899 \text{ ps} \]
so the capacitor spacing is

\[
\frac{d}{\lambda} = \xi f_B \approx 0.09899
\]

slightly less than a tenth of the wavelength.

**Example Design 2:** Similarly, a low value for \( k \) might be 0.1, although this is unlikely to be result in a useful device. The required base transmission line must be

\[
Z_0 = 50 \Omega \sqrt{1 + 0.1} \approx 52.44 \Omega
\]

and \( x \approx 2.858 \). The required capacitance and delay are

\[
C = 90.97 \text{ fF}
\]

\[
\xi = 45.48 \text{ ps}
\]

assuming that the Bragg frequency is again 10 GHz. If the base transmission line has a propagation velocity of 0.75 \( c \), the capacitor spacing would be

\[
d = \xi \nu \approx 10.22 \text{ mm}
\]

which is nearly 46% of the wavelength on the base line (\( \lambda \approx 22.48 \text{ mm} \)), resulting in a very lightly and sparsely loaded structure.

The semi-discrete model can be further extended to lossy NLTLs by allowing the transmission line to have a complex propagation constant, \( \gamma = \alpha + j \beta \), and placing a resistance \( R \) in series with the varactor. The resulting dispersion relation, given for completeness, is

\[
\cosh \gamma_{NLTL} = F_{\text{even}} \cosh \alpha + j F_{\text{odd}} \sinh \alpha
\]  

(2.28)
\[
F_{\text{even}} = \cos x - \frac{1}{2}k x \left( \frac{1}{1 + \rho^2 k^2 x^2} - j \frac{\rho k x}{1 + \rho^2 k^2 x^2} \right) \sin x \quad (2.29)
\]
\[
F_{\text{odd}} = \sin x + \frac{1}{2}k x \left( \frac{1}{1 + \rho^2 k^2 x^2} + j \frac{\rho k x}{1 + \rho^2 k^2 x^2} \right) \cos x \quad (2.30)
\]
where \( \rho = R/Z_0 \) is the dimensionless varactor series resistance.

### 2.3 Model Convergence

Because the lumped-element model is successful at low frequencies, it should occur as a special case of the semi-discrete model in the long-wavelength limit. Making a small angle approximation in (2.17), the right-hand side becomes

\[
1 - \frac{1}{2} \omega^2 \left( 2 \xi^2 - \xi \tau \right) \quad (2.31)
\]

which, substituting (2.14) and (2.15), becomes

\[
\cos \beta = 1 - \frac{1}{2} \omega^2 LC \left( 1 - \frac{2}{k} \right) \quad (2.32)
\]

which reduces to the lumped-element case in the limit of \( k \to \infty \). By a similar argument, the Bloch impedance becomes

\[
Z_B = Z_0 \sqrt{\frac{\tau \omega / k}{\tau \omega (1 + 1/k)}} \quad (2.33)
\]

under the small angle approximation, which yields

\[
Z_B = \sqrt{\frac{L}{C} \left( 1 + \frac{1}{k} \right)} \quad (2.34)
\]

which is equivalent to the lumped-element case in the limit of large \( k \). Since the regime of large \( k \) implies that transmission line delay is negligible, it makes intuitive sense to recover the lumped-element model from the distributed model in this case.
2.4 Bragg Frequency Ringing

Although edge-sharpening is a useful phenomenon for generating high-speed signals, it is not sufficient for creating a UWB transmitter. The sharpened falling-edge has an extremely large bandwidth, most of which lies outside the passband of a practical antenna, so only a small fraction of the signal energy is able to contribute to the transmitted pulse. The NLTL must therefore also concentrate the signal energy in the frequency domain.

Fortunately, mechanisms for narrowing the bandwidth of the output signal are readily available. Earlier designers noted that the NLTL’s Bragg frequency must be significantly higher than the varactor cutoff, or equivalently, dielectric relaxation frequencies to avoid signal distortion [18,19]. In this regime, the minimum falltime is determined by losses which damp out the higher frequency components of the waveform. However, in the case that losses remain small up to the Bragg frequency, the output waveform shows large amplitude, long duration ringing. Intuitively, this effect is due to the low-pass filter characteristic of the NLTL: the frequency multiplication process can generate frequencies no higher than $f_B$, because these frequencies cannot propagate on the lattice, but after the minimum falltime is reached, frequency multiplication continues to transfer energy from low frequencies to higher ones. Therefore, a peak appears in the output spectrum near $f_B$, corresponding to ringing which grows in amplitude as the wave propagates on the NLTL.

An ideal, lossless NLTL can be easily simulated with conventional circuit simulation software, or general differential equation solvers. To investigate pulse formation in the general case we construct a lossless lumped element
Figure 2.3: Signal spectrum at various points along an ideal NLTL for a Gaussian monocycle input waveform (inset). A significant peak in the spectrum appears at the Bragg frequency, even outgrowing the input peak for a sufficiently long NLTL.

NLTL with a large number of sections and unit Bragg frequency. The C-V curve for the varactors uses a tanh function to approximate a MOS capacitor.

The input to the NLTL is an inverted Gaussian monocycle with amplitude large enough to saturate the C-V curve. The monocycle is chosen because it is a DC-balanced waveform with a relatively simple spectrum. A square-pulse waveform, on the other hand, has significant energy at DC and therefore a lower conversion efficiency on the NLTL. The monocycle’s energy is initially concentrated around $\frac{1}{4} f_B$, but the NLTL pushes it to higher frequencies and concentrates it near the Bragg frequency. Fig. 2.3 shows that a large wideband signal centered at $f_B$ has formed after 30 sections and the low frequency energy has been significantly depleted. Relatively little energy remains in the intermediate frequencies, demonstrating both frequency conversion and concentration by the NLTL. By sections 20 and 30, about 42% and 68% of the total signal energy is
Figure 2.4: Relative amplitudes of the input monocycle and output ringing for an ideal NLTL with a quadratic fit. Both scales are normalized to the simulation in Fig. 2.3. The power spectral density at Section 30 (inset) is shown normalized to the maximum input monocycle amplitude.

found in the band between $0.7 f_B$ and $1.3 f_B$, suggesting that high conversion efficiencies are possible with a sufficiently long line.

The conversion efficiency may also be expected to scale with the amplitude of the input signal, as is the case for narrowband harmonic generation. For example, it is well known that, for NLTLs in the small-signal regime, the second-harmonic output amplitude increases with the square of the input amplitude [20]. We find a similar relation on the Bragg ringing line by scaling the amplitude of the input monocycle and comparing the ringing magnitude at the 30th section. At small input amplitudes the waveform does not reach its minimum falltime so little or no ringing is present. But once the monocycle exceeds the threshold for ringing to form, the output (ringing) amplitude scales quadratically with the input (Fig. 2.4). The observation of a threshold again illustrates the interaction between the harmonic generation properties of the NLTL and its...
low-pass filter effect. In general, we do not expect the NLTL to encounter input amplitudes which significantly saturate the C-V characteristic since a MOS-varactor’s C-V curve is closely related to the choice of supply voltage, nor are large amplitudes germane to low-power operation. However, in the case of amplitudes which moderately saturate the C-V curve, we observe output amplitudes which continue to grow quadratically, as the power in the individual harmonics grows.

2.5 Next-Neighbor Coupling

Next-neighbor coupling provides an additional method to determine the ringing frequency on the NLTL. Additional capacitors are placed on the NLTL connecting every second section, with the coupling coefficient being

$$\gamma_{NNN} = \frac{C^*}{C_0}$$

where $C^*$ is the additional capacitor and $C_0$ is the original NLTL shunt capacitance. The additional coupling lowers the group velocity compared to the phase velocity, increasing the minimum falltime and reducing the ringing frequency while preserving the Bragg frequency. The complete dispersion relation from
Figure 2.6: Group (dashed) and phase (solid) velocities for NNN-coupled lines after [21].

([16, 21]) is given by

\[
\frac{\omega^2}{\omega_B^2} = \frac{\sin^2 \beta/2}{1 + 4\gamma \sin^2 \beta}
\]  

(2.35)

from which one can derive the group and phase velocities

\[
\nu_p = \frac{1}{\beta} \frac{\sin \beta/2}{\sqrt{1 + 4\gamma \sin^2 \beta}}
\]

(2.36)

\[
\nu_g = \frac{\cos \beta/2}{2\sqrt{1 + 4\gamma \sin^2 \beta}} - \frac{4\gamma \cos \beta \sin \beta/2 \sin \beta}{(1 + 4\gamma \sin^2 \beta)^{3/2}}
\]

(2.37)

for the case of a lumped-element line, however the results may be qualitatively extended to semi-discrete lines with large \( k \) values. When the NNN coupling is present, the group and phase velocities have local minima which disrupt the NLTL’s edge sharpening, causing ringing to occur below the Bragg frequency. When the coupling is sufficiently strong, as in the case of \( \gamma = 0.5 \) in Fig. 2.6, group and phase velocities are matched at specific frequencies, leading to large amplitude ringing.

The primary advantage of this technique is that the Block impedance, \( Z_B \),
approaches zero near $f_B$, making the structure particularly susceptible to series losses, which are proportional to $1/Z_B$. The use of NNN coupling pushes the ringing to a lower frequency where $Z_B$ is closer to its design value, reducing attenuation and improving impedance matching at the NLTL’s output. This makes NNN particularly attractive for semi-discrete NLTLs in CMOS, where the long transmission line segments are particularly lossy.

2.6 Power Requirements

The energy required to create a pulse on the NLTL has three main contributions: energy consumed by the driver circuit, losses on the NLTL, and the capacitive charging energy of the NLTL. The first two are very implementation dependent, however the charging energy is intrinsic to the NLTL. The transmitter is expected to have a slow input, with a wavelength longer than the total length of the line, and be terminated by an antenna, which appears as an open-circuit at low frequency. Therefore, the voltage is constant everywhere on the line before $t_1$ in Fig.2.2. Since the current vanishes under these conditions, no energy is stored in the inductors and the entire structure appears to the driver as a single nonlinear capacitance of $N \cdot C(V)$ where $N$ is the total number of sections. To generate a pulse the capacitance is charged to a positive voltage, $V_H$, and then (dis)charged to a negative voltage, $V_L$. Following the pulse, the capacitance must be discharged to 0 V to complete the cycle. The total charging energy per pulse is therefore given by

$$E_{Ch} = N \left[ C(V_H) \cdot V_H^2 + C(V_L) \cdot V_L^2 \right]$$  \hspace{1cm} (2.38)
for an ideal NLTL. This result immediately suggests a scaling rule for the pulse generator. Power consumption should increase linearly with the number of sections and decrease with the inverse of the Bragg frequency. Specifically, the charging energy scales as

\[ E_{Ch} = N \left( V_L^2 + \eta V_H^2 \right) \frac{1}{\pi Z_0} \frac{1}{f_B} \]  

(2.39)

where \( \eta \) is the ratio \( C(V_H)/C(V_L) \), and \( Z_0 \) is the characteristic impedance of the NLTL at \( V_L \). Although losses in the inductors also contribute to the required energy, the inductor size also scales as \( 1/f_B \). The overall energy consumption can therefore be expected to scale by \( 1/f_B \) until the dissipation in the driving inverter becomes significant.

Finally, the maximum pulse rate is limited by the time required to recharge the line to \( V_H \). After some manipulation, we obtain an upper-bound on the pulse repetition frequency (PRF) of

\[ PRF = \frac{\pi f_B}{2 N \eta} \]  

(2.40)

allowing two RC time constants to recharge the line. In practice, since it is not possible to perfectly impedance-match the NLTL, the maximum PRF must also allow time for source and load reflections to dissipate. Therefore, Eq. (2.40) does not represent a hard limit, rather, the quality of the output waveform degrades at high PRFs, allowing some compromises to achieve the maximum data rate.
To confirm the theoretical results and determine whether the NLTL can form a feasible transmitter, semi-discrete circuit board prototypes were constructed. The primary purpose of this work is the demonstration of Bragg ringing in a semi-discrete model NLTL with and without NNN coupling. Our implementation uses conductor-backed Coplanar Waveguides (CPW) on FR-4 laminate. The CPW allows low parasitics for shunt connected elements (unlike microstrip), and offers good noise isolation and low loss. However, due to the layered nature of fiberglass laminates, the effective dielectric constant for edge coupled structures like the CPW can be difficult to control. The conductor backing creates a hybrid CPW-Microstrip mode which sees an effective dielectric constant closer that of the bulk. However, the bulk dielectric constant of FR-4 is notoriously variable. We therefore construct several test CPWs to verify $Z_0$ and $\nu$ before completing the NLTL design. Our final design has a $k$ parameter of 1.07, using $Z_0 = 72\ \Omega$, $d = 4.0\ cm$, and $C_o = 3.9\ pF$.

To introduce NNN coupling while minimizing parasitics, we fold the CPW into a serpentine pattern and place varactors at the bends. We use radial bends and place vias to minimize coupling to the Coplanar Slotline (CPS) mode, since the influence of the varactors is drastically reduced for this mode. Additional vias are placed in the CPW center conductors and the coupling capacitors are mounted on the backside with traces kept as short as possible to minimize parasitic inductance. The final board is shown in Fig.3.1.

We examine cases of moderate coupling to create a large shift in $f_R$ while minimizing potential alteration of the NLTL’s small-signal characteristics. Based
on $C_o$ above, the coupling capacitors are 1pF and 2pF, corresponding to coupling strengths of 0.25 and 0.5, respectively.

3.1 Testing Procedure

3.1.1 Background on S-Parameters

An essential set of quantities for characterizing microwave structures such as NLTLs are the scattering parameters (S-parameters), which relate the incoming and outgoing wave amplitudes at the both the input and output ports. An essential review can be found in [22]. Because the connections in the test setup are long compared to most wavelength of interest, it is essential to remove the effects of the test setup from the final measurement by calibrating the network analyzer.

Two widely used calibration methods are Short-Open-Load-Thru (SOLT) and Thru-Reflect-Line (TRL), which relay on different sets of calibration standards and may be appropriate in different circumstances. SOLT relies on a set
of four perfect, or at least perfectly known, reference terminations placed at the same reference plane, making it suitable for measuring devices with coaxial connections where excellent reference standards can be realized. On the other hand, TRL uses only 3 reference standards, which need not be perfectly known. However, it does place restrictions on the minimum and maximum phase between the Thru and Line standards, requiring multiple Line standards when covering a broad frequency range and very long Lines when measuring at low frequencies. For these reasons TRL may also be referred to as Line-Reflect-Line (LRL, referring to the nonzero phase shift in the Thru) and Line-Reflect-Match (LRM, using a matched load to simulate an infinitely long line). TRL is well suited to on-chip measurements where the precision of the standards and consistency of reference planes cannot be guaranteed.

Often, obtaining an accurate measurement of on-chip devices with SOLT calibration is extremely challenging. An example is shown in Fig. 3.2 where the S-parameters of an on-chip transmission line are measured following both a TRL and SOLT calibration and the line’s characteristic impedance is extracted using the method from [23]. The TRL calibrated measurement shows a 100 Ω impedance with no complex component, whereas the for the SOLT calibrated measurement, both the real and imaginary components oscillate significantly. The likely cause is that the effective reference planes are not consistent between the different SOLT standards on the chip.

When measuring circuit boards, however, SOLT is usually sufficient, as the coaxial reference standards are precisely constructed. As long as the coax-to-PCB transitions are well impedance matched, SOLT is actually preferred for circuit boards as the use of TRL requires fabrication of additional (and often
Figure 3.2: Characteristic impedance of an on-chip transmission line extracted from S-parameters measured with TRL and SOLT calibrations. The TRL measurements show a constant real impedance, while the SOLT measurements are nearly useless.

large) boards for use as the reference standards.

### 3.1.2 NLTL Measurements

We first measure the 1MHz C-V characteristic for the diodes in order to predict the NLTL properties. We then measure the NLTL S-parameters at DC biases from 0V to 6V and deembed with SOLT calibration. As shown in Fig.3.3, there is a large discontinuity in the first derivative of $S_{21}$ phase which identifies the Bragg frequency. Additionally the NLTLs with NNN coupling clearly show a dip and peak in their group velocity, similar to Fig. 2.6. These features are not as
pronounced as expected, likely because the low \( k \) parameter (more distributed line) reduces the effective influence of the NNN capacitors.

We measure the NLTL impedance over the same bias range using Time Domain Reflectometry (TDR). The variations in Bragg frequency and impedance are shown in Fig. 3.4 compared with the expected values computed from the diode C-V. The C-V characteristic predicts both parameters reasonably well at low biases. Note that the addition of NNN coupling has very little impact on either the Bragg frequency or impedance. At 2V bias the diodes have capacitance 3.9pF and we accurately achieve \( f_B = 1 \text{GHz} \) and \( Z_{NLTL} = 50 \Omega \) in agreement with the design procedure.

To directly measure the edge sharpening and onset of ringing we must provide a variable falltime at the NLTL’s input. Since this is not possible with com-
Figure 3.4: Analytical $f_B$ and $Z_{NLTL}$ calculated from the diode C-V, and the measured results.

To measure the distortion of the sine wave, we use a CW RF source and examine the distortion of the sine wave, taking advantage of the fact that the 10-90 falltime for a sine wave of period $T$ is

$$
\tau_{\text{sin}} = \frac{\arcsin \frac{4}{5}}{\pi} T
$$

(3.1)

Fig. 3.5 shows the output falltime measured for a sine wave from 1.5V to 2.5V compared with the analytical falltime. It is clearly seen that the output falltime has been compressed by about 800ps for longer falltimes, but is unchanged for short falltimes. The transition occurs at the NLTL’s critical falltime, which is related to the ringing frequency by

$$
f_R = \frac{1}{2 \tau_{\text{crit}}}
$$

(3.2)

Finally, we use a high bandwidth pulse generator and risetime filter to generate a controlled falltime of 1.2ns at the NLTL’s input. The time-domain response
Figure 3.5: All NLTLs show a constant 800ps edge sharpening for long falltimes and no sharpening for short falltimes.

For the normal NLTL is shown in Fig.3.6. The output shows ringing of about 200mV amplitude only on the falling edge. The gain spectrum shows a significant concentration of energy in a 500MHz-600MHz band. The 1pF-coupled NLTL has $\tau_{crit} > 1.2\text{ns}$, so there is significant ringing on both edges despite the risetime filter (Fig.3.7). The amplitude at the falling edge is greater than at the rising edge by about 100mV, which suggests the same edge-sharpening/low-pass filter effect is enhancing the ringing. The addition of NNN coupling has successfully reduced $f_R$ by about a factor of 2 (to 285MHz) in this case. For the 2-pF coupled line the ringing on both edges is very severe and becomes distorted by the NLTL’s edge-sharpening, rendering this line unusable without a slower risetime filter.
Figure 3.6: Time-domain ringing in the Normal NLTL occurs only on the falling edge. The DC bias voltage has been subtracted for clarity. The gain spectrum shows significant concentration of energy in a narrow bandwidth.

Figure 3.7: Ringing in the NLTL with 1pF coupling occurs on both edges due to the cutoff of the risetime filter. The ringing at the falling edge has been enhanced by about 100mV.
3.2 Conclusion

The circuit board NLTLs provide excellent validation for the theoretical results and design procedure presented in Chapter 2: the cutoff frequency and characteristic impedance are predicted precisely, overcoming one of the major concerns discussed in [18] for a semi-discrete line. Further, the group velocity clearly shows local minima under NNN coupling, demonstrating that the dispersion relation presented in Section 2.5 applies qualitatively to semi-discrete NLTLs as well. Finally, the NLTLs are high-quality, with small-signal losses of only ~6 dB near the Bragg frequency. The low losses result in strong ringing, which may be tuned by the NNN capacitance.

Despite its success, the circuit board design is particularly large (30 cm) and requires nearly 40 sections. A direct scaling of the Bragg frequency of 1 GHz to 10 GHz would result in a device still 3 cm in size and impractical for most purposes. Instead, an RFIC version must employ miniaturization techniques, such as high $k$ parameters, strong NNN coupling, and floating shielding (creating a slow-wave effect), which make the design more challenging and tend to increase loss.
CHAPTER 4
RFIC NLTLS

Ideally, the NLTL transmitter should be integrated on a high-frequency Silicon integrated circuit (RFIC), along with other application circuitry. This can be a particularly challenging environment due to the small dimensions and close proximity to the conductive Silicon substrate.

4.1 Passives on RFIC

Implementing an NLTL on an integrated circuit requires an understanding of the behavior of the passive components used to construct the line. Transmission lines, varactors, and inductors are all available in modern RF-CMOS processes, but their characteristics are far from ideal. This section reviews the relevant characteristics of each.

4.1.1 Transmission Lines

While passive transmission lines are possible on an RFIC, the construction of low-loss, electrically long, high-impedance lines is very difficult. Based on the semi-discrete NLTL model, the transmission line’s equivalent inductance can be used as a figure of merit.

\[ L_{TL} = \xi Z_0 = \frac{d}{\nu} Z_0 \] (4.1)

Therefore, the ideal line will have simultaneously high impedance and low velocity. The most common types of on-chip transmission line are microstrip,
Microstrip Coplanar Waveguide Coplanar Stripline

Figure 4.1: Common transmission lines in an RFIC environment with electric field configuration for the dominant mode of propagation. CPW and CPS lines have a hybrid Microstrip mode due to the proximity of the substrate ground plane.

coplanar waveguide (CPW), and coplanar stripline (CPS), which are shown in Fig 4.1 with electric field lines for their fundamental modes.

Both coplanar lines control their characteristic impedance through the combination of conductor width and spacing, both parameters available to an RFIC designer. The impedance of a microstrip is controlled by the conductor height (a maximum of 18 µm in the 8-metal process used for the NLTLs) and width, making high impedance lines very narrow. The CPS line offers the highest possible $Z_0$, 100-150 Ω being possible on the RFIC. The maximum possible impedance for both coplanar lines is limited by the presence of the substrate ground plane, which causes a hybrid microstrip-coplanar mode when the conductor spacing is on the order of the conductor height [24, 25].

Losses are a major issue for all transmission lines; the main mechanisms are conductor losses, due to skin-effect, dielectric losses, and radiation losses. Radiation losses are neglected here because the dimensions of the transmission line are much smaller than the wavelength. Radiative coupling to dielectric slab modes is also unlikely because the dielectric contrast at the Inter-metal
dielectric-Air interface is small (4.1:1 for SiO\textsubscript{2}, compared with 13.1:1 for a GaAs MMIC) and the presence of metal fill tends to disrupt the slab mode. Instead, dielectric loss is the main concern due to the proximity and conductivity of the Silicon substrate.

The electric and magnetic fields from the transmission line penetrate into the substrate, causing conduction and eddy currents which dissipate energy. These losses become more severe as the conductivity of the Silicon increases and are difficult to model because the three-dimensional doping concentration is not known by the designer. A modern RFIC process starts with 1-2 Ω-cm wafers, with the conductivity increased further by various dopant implants. To prevent substrate losses, transmission lines are placed on the top metal layers, farthest from the substrate, with metal shielding on lower layers, and lightly-doped substrate regions underneath.

All three techniques were used in designing CPS lines for the NLTLs, with extensive electro-magnetic simulations in HFSS to predict the propagation and attenuation constants. However, in order to achieve a high-impedance CPS line, a conductor spacing of 25-50 μm, larger than the conductor to substrate spacing, was used. While good results have been reported for CPW lines and inductors with floating metal shields \cite{26}, our CPS lines still showed significant high-frequency attenuation beyond that predicted in simulation. Ultimately, the metal shielding proved ineffective because the thickness of the available shield layers was less than the skin depth, allowing fields to penetrate the substrate. The wide-spacing CPS lines thus had large substrate losses, as well as conduction losses in the thin floating shield.

The high-impedance CPS lines used here are likely close to the limit of fea-
Figure 4.2: Attenuation ($\alpha$), and phase ($\beta$) constants for a 100 $\Omega$ CPS line in the 8-metal RFIC process used for NLTLs. Attenuation is much higher than predicted by EM simulation, but measured $\beta$ matches closely.

sibility in a modern CMOS process. The conductor spacings were significantly greater than the distance to the substrate, and the lengths, 540-790 $\mu$m between varactors, consume significant die area. These dimensions require large areas of the chip free of the metal fill used to meet pattern density rules required by modern processing techniques (specifically, chemical-mechanical polishing) and are thus difficult to fabricate. The difficulties with designing high-impedance, low-loss transmission lines on RFICs largely contributed to the poor performance of the NLTLs presented in Section 4.4.
4.1.2 Varactors

Two types of voltage-variable capacitors are available on an RFIC: reverse biased diodes and MOS-capacitors. Diodes have been used in most NLTLs to date because very fast devices can be easily made on GaAs. The capacitance-voltage characteristic for a diode depends on the doping profile of its junction and falls between $1/V^{1/3}$ and $1/V^2$ for a practical device. This means that the capacitance varies somewhat slowly with voltage, requiring signal amplitudes of several volts for effective edge-sharpening. Such large swings are at odds with the low power requirement for the transmitter, due to the dependence of energy on $V^2$ in Eq. (2.38), and the need for driver circuitry to generate and use voltages larger than the supply voltage.

MOS-capacitors are unique in that their capacitance changes significantly for a very small voltage swing. The device structure (Fig 4.3) is identical to an n-type Mosfet, but the well or body doping is changed to n-type as well. The capacitance varies as majority carriers are accumulated or depleted from the region under the gate. As no sources of minority carriers (p-type regions) are found in the device, an inversion layer will not form (except for very low frequencies in the kilohertz range where thermal generation becomes significant). The C-V characteristic (Fig 4.4) is therefore determined by the same parameters (gate-oxide thickness, channel doping, etc) as the Mosfet threshold voltage. The varactor can reach deep-depletion (minimum capacitance) at a voltage around $-V_T$, the threshold voltage for the p-fet. For the CMOS technology used for the NLTLs, the Mos-varactors can achieve capacitances of $C_{\text{max}}$ and $\sim 0.3 C_{\text{max}}$ at 0.5V and -0.5V, respectively. Although careful design is required to achieve both positive and negative gate voltages, the total voltage swing on the NLTL is less
Figure 4.3: Device structure for an n-type Mosfet (left) and accumulation-mode Mos-varactor (right). The Mosfet requires minority electrons to transit the p-doped channel, whereas the varactor accumulates and depletes a layer of majority carriers.

Figure 4.4: Capacitance voltage characteristic for a typical MOS-varactor at two different frequencies. At 60 GHz the device has a larger effective capacitance as it approaches its self-resonant frequency. Inset shows the Capacitance-frequency characteristics at -0.5, 0, and +0.5 V.

than the supply voltage of 1.2 V.

As majority carrier devices, Mos-varactors can also operate at very-high frequencies, higher than the cutoff frequency of the transistors in the same technology. Because the Mosfet requires minority carriers to cross its channel region,
its intrinsic speed is limited by the transit time,

\[ \tau_t = \frac{L^2}{\mu_{\text{eff}} V_{DS}} \]  

(4.2)

where \( L \) is the channel length and \( \mu_{\text{eff}} \) is the effective channel mobility. By contrast, the varactor need only accumulate and deplete majority carriers under its gate, a process governed by the dielectric relaxation time,

\[ \tau_d = \varepsilon \rho = \frac{\varepsilon}{q\mu n} \]  

(4.3)

where \( \rho \) is the bulk resistivity, \( \varepsilon \) is the dielectric constant, and \( \mu \) and \( n \) are, respectively, the mobility and majority carrier concentration in the bulk. This time constant is about a picosecond for 1Ω-cm Silicon, much faster than the transit time. Using varactors with very high cutoff frequencies, NLTLs can generate signals faster than the transistor cutoff frequency [27, 28]. However, practical varactors will be limited by the parasitic inductance and resistance connecting them to the NLTL, imposing a self-resonant frequency (due to inductance) and extrinsic cutoff frequency (due to resistance) significantly lower than the intrinsic dielectric relaxation frequency. This is a particular problem in modern CMOS, where 6 metal layers may lie between the NLTL and the varactor. Varactors in these technologies may have self-resonant frequencies in the range of 50 GHz.

### 4.1.3 Spiral Inductors

Spiral inductors are typically large, distributed structures used in planar circuits such as CMOS. Due to their size (over 100 µm across), and closely spaced windings, they suffer from a variety of loss mechanisms such as coupling to
Figure 4.5: Simulated quality factors for the inductors used in 10, 20, 30 and 60 GHz NLTLs. The 160 pH device is a single-turn spiral and has a significantly higher Q due to the reduced proximity-effect losses.

The spiral inductors for use in NLTLs are designed to have a peak-Q as close to the Bragg frequency as possible, with the designs and models taken from the foundry’s standard-cell library. In some cases, the peak-Q may fall significantly below $f_B$, as for the 160 pH and 510 pH inductors in Fig. 4.5 (intended for 30 and 60 GHz NLTLs, respectively), but fully-custom spiral design may be able to further optimize their performance.
4.1.4 Substrate Effects

Substrate effects and modeling are another area to which a large body of literature is dedicated. Again, a full review is beyond the present scope, but in general, the doped Silicon substrate used in CMOS is neither a good conductor nor a good insulator and can contribute substantially to losses and lack of isolation across large areas of an RFIC. Predicting and modeling these effects is difficult because the length scales can be very large compared to the size of circuit elements. Additionally, the doping of the Silicon varies with depth, with the profile usually not available to the designer. The most common strategies for dealing with substrate effects for RFIC passives such as transmission lines and inductors are to increase the distance between the device and the substrate, and to insert metal shielding to screen the substrate. An excellent discussion of these strategies can be found in [26].

4.2 Semi-Discrete NLTLs and Layout

To investigate the characteristics of on-chip NLTLs compared to circuit board versions, we fabricated NLTLs in TSMC’s 0.25µm mixed-mode process. The varactors are 200fF nmos capacitors spaced at 675µm intervals along 62 Ω coplanar waveguides. Versions were constructed with and without shielding by floating metal strips 2µm wide and 2µm apart placed on the metal layer 1µm beneath the CPW (referred to here as GbCPW). The expected per-section CPW delay, is computed from HFSS simulations and given in Table [4.1]. The designed $k$ parameters for the CPW NLTLs and GbCPW NLTLs are 2.5 and 1.2, respectively. Both tight- and loose-folded serpentine patterns are implemented with 10 NLTL
sections using both conventional CPW and GbCPW. A six-section straight NLTL (no grid) is used for reference. Fig.4.6 shows the die with all NLTLs and varactor locations.

The phase constants, Fig.4.7, extracted from S-parameter measurements, show good agreement with the analytical model of Section 2.2 up to about 25 GHz. Here, the analytical model uses the design parameters extracted from simulation, giving a systematic error noticeable in the different slopes of the unfolded NLTL and the model. Overall, the unfolded and tight-folding lines have lower than expected propagation velocities, while the loose-folding lines are faster than expected. Fig.4.8 shows the extracted attenuation constants along with the analytical model. The cutoff is accurately predicted by considering varactor loss alone, while transmission line loss dominates at low frequency. The loosely folded NLTLs have slightly higher cutoff frequencies than the model predicts.
Figure 4.7: Extracted phase constants compared with the analytical model due to their higher than expected propagation velocities. Conversely, the unfolded and tightly folded NLTLs have lower than predicted cutoff frequencies due to their larger transmission line delay. All of the folded lines have relatively low stopband attenuation due to coupled line effects.

The edge sharpening performance of the NLTLs is examined by sine-wave distortion measurements. The line is excited by a CW RF signal with the RF power adjusted to maintain a constant amplitude (2V peak-to-peak) at the NLTL output. The output falltime is then measured on a high-speed oscilloscope. Since the 10-90 falltime of a sine-wave varies with frequency, the edge sharpening and large signal cutoff can be measured by a simple frequency sweep. Fig. 4.9 shows the measured output falltimes. The sharpening is roughly constant at low and intermediate frequencies while the output falltime saturates at high frequencies. The saturated falltime is 20ps for all NLTLs, correspond-
Figure 4.8: Extracted attenuation constants. The cutoff frequencies are well predicted by varactor losses alone.

The large-signal cutoff frequency is 25GHz, which roughly agrees with the small-signal behavior.

Finally, the per-section edge sharpening is extracted and shown in Fig.4.10. The figure of merit for an edge-sharpening line is the normalized compression: $\kappa = \Delta \tau_f / \xi$, where $\Delta \tau_f$ is the per-section change in falltime. The normalized compression at low frequencies is estimated from the average (mean) sharpening in the flat region of Fig.4.10, corresponding to input falltimes of about 164ps to 87ps. The results are summarized in Table 4.1. A typical value for $\kappa$ in GaAs NLTLs is around 0.75 [18]. The tightly folded CPW NLTL and unfolded NLTL show excellent performance in this regard. Both GbCPW NLTLs have significantly reduced performance due to their smaller $k$ parameters. Since the $k$ parameter describes the strength of the capacitive loading, smaller values imply that the varactor’s influence, and thus edge sharpening capability, is de-
Figure 4.9: Compressed Falltimes. For clarity, only the errorbars for the loose-folding GbCPW NLTL are shown; the other measurements have similar uncertainty.

Table 4.1: Summary of initial on-chip NLTLs

<table>
<thead>
<tr>
<th>Layout</th>
<th>CPW Delay, $\xi$</th>
<th>Compression, $\kappa$</th>
<th>Layout dimensions</th>
<th>$k$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Loose GbCPW</td>
<td>6.9ps</td>
<td>0.32</td>
<td>315$\mu$m x 1365$\mu$m</td>
<td>1.25</td>
</tr>
<tr>
<td>Loose CPW</td>
<td>4.9ps</td>
<td>0.55</td>
<td>315$\mu$m x 1365$\mu$m</td>
<td>2.53</td>
</tr>
<tr>
<td>Tight GbCPW</td>
<td>6.9ps</td>
<td>0.49</td>
<td>345$\mu$m x 1050$\mu$m</td>
<td>2.50</td>
</tr>
<tr>
<td>Tight CPW</td>
<td>4.9ps</td>
<td>0.82</td>
<td>345$\mu$m x 1050$\mu$m</td>
<td>5.06</td>
</tr>
<tr>
<td>Unfolded CPW</td>
<td>4.9ps</td>
<td>0.97</td>
<td>36$\mu$m x 2650$\mu$m</td>
<td>5.06</td>
</tr>
</tbody>
</table>

creased. Finally, we see that the use of folding also reduces $\kappa$, with the largest degradation in loose-folding NLTLs. Here, coupled line effects create a transmission path that bypasses the varactors, improving the small signal response at the expense of edge sharpening.
4.3 Semi-Discrete NLTLs Transmitters

The first attempt at an NLTL transmitter focused on scaling the semi-discrete circuit board version to dimensions compatible with an RFIC. In this case, an NLTL with a 10 GHz Bragg frequency would require a varactor spacing of several millimeters, using an impractical amount to layout area, so NNN coupling must be used to reduce the ringing frequency.

Several NLTLs are designed for a 30 GHz Bragg frequency and have NNN parameters of 0, 0.5 and 0.75. They are fabricated in IBM’s 0.13\(\mu\)m process using high-impedance coplanar striplines in top-metal and NMOS loading varactors. Each line has a varactor spacing of 540\(\mu\)m and a \(k\) parameter of 7.2. Due to die area constraints and the need to test multiple devices, the length of each NLTL is limited to 10 sections. The on-chip NLTLs are shown in Fig.4.11.
Figure 4.11: Partial die photo showing two on-chip NLTLs, both normal (top) and NNN-coupled (bottom).

The edge compression performance of the NLTLs is measured using the sine wave distortion technique from Section 3.1. The lines have a nominal edge compression of about 45 ps with minimum falltimes of about 15 ps, 30 ps, and 40 ps for the normal, \( \gamma = 0.5 \) NNN, and \( \gamma = 0.75 \) NNN, respectively. While the NNN coupling at least doubles the minimum falltime, coupling coefficients larger than 0.5 appear to have diminishing returns.

Pulse formation on the NLTL can be observed using a high speed oscilloscope. The line is driven by a square-wave from an Agilent 8133A with a measured falltime of 60 ps, meaning all NLTLs should achieve their minimum falltime. The response of the normal NLTLs shows small amplitude ringing superimposed on the square wave falling edge. Both NNN coupled lines show lower
Figure 4.12: Sine distortion measurements on all NLTLs. Nominal edge-compression is about 45 ps, with minimum falltimes around 15 ps, 30 ps, and 40 ps.

frequency ringing with an amplitude of 100-200 mV. While the ringing pulse is fairly broadband due to its short duration, its main frequency component can be estimated from the time between its maxima or minima. The pulse frequencies are therefore about 40 GHz for the normal NLTL and 10-15 GHz for both NNN coupled NLTLs.

To separate the high frequency pulse from the remaining base-band signal, the antenna itself is used as a bandpass filter. We use home-built TEM horn antennas with peak transmission efficiency around 10-14 GHz to test the transmitter over a range of 70 cm, using the oscilloscope as the receiver. The NNN coupled NLTLs were driven with short duration square pulses from the 8133A. The pulse transmitted from the NNN NLTL \((\gamma = 0.5)\) is shown in Fig.4.14. The pulse shape is largely independent of the input signal. The large 8 mV peak
Figure 4.13: Time-domain response to a square-wave input with 60 ps falltime.

Figure 4.14: Impulse transmitted between TEM horn antennas at 70 cm. The main peak is from the square-wave falling edge, with trailing peaks from the NLTL ringing.
is due to the sharp falling edge, with a smaller 2 mV peak and trailing oscillations from the NLTL ringing. These measurements demonstrate that despite the low conversion efficiency of the square-wave input, an NLTL transmitter is still adequate over moderate distances.

4.3.1 Measurement of NLTL Charging Energy

Finally, we attempt to verify the power requirements derived in Eq. (2.38) by directly measuring the power supplied to the NLTL by a low frequency driver. The NLTL is driven by a 100 kHz sine wave from an SRS 345 function generator with a series connected current amplifier (SRS 570). The time-dependent voltage and current are then captured on an oscilloscope (Fig. 4.15). The energy stored on the NLTL is simply

$$ E_{IV}(t) = \int_0^t i(\tau) v(\tau) d\tau \quad (4.4) $$

and the energy consumed in a complete cycle is the sum of the two peak energies,

$$ E_p = E_{IV}(v = -0.5) + E_{IV}(v = 0.5) \quad (4.5) $$

From the oscilloscope measurements, the current and voltage are deskewed to eliminate the time delay of the cables and amplifier, and a background power of 26 nW is subtracted from the results of (4.4) to account for resistive loss in the cables and probes.

For comparison, we measure the C-V characteristic of the varactors, and apply $1/2 C(V) V^2$ to the driving voltage to recover the result of Eq. (2.38). Both methods of computing the charging energy are shown in Fig. 4.16 and compared with circuit simulation of the I-V method. The simulation results closely follow
the C-V method. However, the I-V results show that the energy stored in the varactors does not go to zero during the rising-edge zero-crossing of the input waveform. This effect is due to the decreasing width of the depletion region in the MOS capacitor. As the capacitance increases and the depletion region narrows, a portion of the charge stored on the capacitor “plates” goes to compensate the space charge in the depletion region, rather than contributing to the externally measured discharge current. This charge is recovered again on the falling-edge, as the capacitance decreases and the depletion region widens to its original width. Intuitively, the additional energy can be considered as the work required to move the capacitor plates.

The energy per pulse obtained from Eq. (4.5) is 0.55 pJ and 0.83 pJ for the C-V and I-V methods, respectively, which is extremely small compared to existing UWB technologies.
Figure 4.16: Measured energy stored in the NLTL as a function of time obtained from Eq. (4.4), compared with simulation, and the measured C-V method.

4.3.2 Outlook

Although the energy consumption of integrated semi-discrete NLTLs is very small, their measured performance does not quite fulfill the design goals. The transmitted waveforms in Fig 4.14 are primarily impulses due to the sharp falling-edge, and the output waveforms in Fig 4.13 show that, while ringing occurs, it is nearly masked by slower capacitive discharging. The NLTLs therefore fail to concentrate the signal into an appropriate bandwidth.

Additionally, the devices presented here represent the limits of what is likely to be achieved in a commercial CMOS process. The large varactor spacing (more than 500 µm) makes it very difficult to meet pattern density rules required by modern CMOS foundries due to the need to exclude metal fill from the transmission line region. The result is large areas of low density, surrounded by areas of very high density. Even at these dimensions, the Bragg frequency, 40 GHz,
is close to the maximum supported by the foundry device models, making it difficult to guarantee performance of the varactors. Finally, the loss of the base transmission line is much higher than predicted, likely due to substrate losses. Attempts to shield the CPS lines from the substrate were not effective because the metal layers available for shielding are significantly thinner than the skin-depth at 40 GHz.

Scaling the semi-discrete NLTLs to a lower Bragg frequency would drastically increase the layout area, unless a slow-wave structure can be used to compact the transmission lines. However, in absence of magnetic materials, a slow-wave structure must use increased capacitance to raise the effective dielectric constant, thus decreasing the impedance of the line. Therefore, the figure of merit in Eq. (4.1) cannot be improved by using slow-wave structures. In fact, it becomes difficult to maintain an NLTL impedance of 50 $\Omega$, inviting significant losses and reducing the NLTL performance further. The solution is instead to return to lumped-element NLTLs, which can be integrated in CMOS using spiral inductors or bondwires.

4.4 Lumped-Element NLTLs in RF-CMOS

As the substrate and transmission line losses are very high for the semi-discrete NLTLs, lumped-element NLTLs were constructed in the same CMOS technology using spiral inductors. The main advantages of spiral inductors over bondwires are the lack of additional post-fabrication steps, and the availability of mature, measurement-based foundry models.

The improvement of the lumped-element design over the previous semi-
discrete designs is evident from S-parameter transmission measurements taken from both structures. Fig 4.17 compares the magnitude transmission from the semi-discrete design with the 20 GHz lumped-element design. The semi-discrete design has losses which increase rapidly with frequency, up to 15 dB near the Bragg cutoff. By contrast, the lumped-element design has as much as a 6 dB advantage in the 10-15 GHz range and losses no greater than -6 dB near cutoff. Additionally, the lumped-element design has a significantly reduced layout area, and meets pattern density rules quite easily.

Figure 4.17: S-parameter transmission for CPS-line (semi-discrete) and spiral-inductor (lumped-element) NLTLs on RFIC. The lumped-element design has a sharp Bragg cutoff and 6 dB reduction in high-frequency loss.
Table 4.2: Nominal L and C values for each NLTL at $V_{gs} = -0.5$ V.

<table>
<thead>
<tr>
<th>Design frequency [GHz]</th>
<th>Inductance [pH]</th>
<th>Capacitance [fF]</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>3220</td>
<td>1300</td>
</tr>
<tr>
<td>10</td>
<td>1390</td>
<td>530</td>
</tr>
<tr>
<td>15</td>
<td>1050</td>
<td>420</td>
</tr>
<tr>
<td>20</td>
<td>770</td>
<td>300</td>
</tr>
<tr>
<td>25</td>
<td>640</td>
<td>260</td>
</tr>
<tr>
<td>30</td>
<td>510</td>
<td>200</td>
</tr>
<tr>
<td>60</td>
<td>159</td>
<td>115</td>
</tr>
</tbody>
</table>

4.4.1 Simulation and Modeling

To examine the performance of realistic lumped-element NLTLs, we first model complete UWB transmitters based in IBM’s 0.13$\mu$m RF-CMOS process. The NLTLs are designed to have Bragg frequencies between 5 GHz and 30 GHz - in 5 GHz steps - and a characteristic impedance of 50 $\Omega$ at -0.5 V bias. An additional NLTL with a Bragg frequency above 60 GHz is designed to represent the extreme high-frequency case. The nominal inductor and capacitor values are shown in Table 4.2.

The NLTLs are driven by CMOS inverters in the same technology, sized to have an output impedance near 50 $\Omega$ during switching (i.e. $R_{on-N} || R_{on-P} \approx 50\Omega$). To understand the performance of the inverters, we simulate their small-signal characteristics (without layout parasitics) with a 1.5V supply and 0.75 V gate bias. The resulting s-parameters of the driver, Fig. 4.18, show a 3-dB frequency of 8.4 GHz and a unity-gain frequency near 22 GHz. These parameters allow us to estimate the intrinsic minimum falltime of the driver as

$$\tau_{90-10} \approx \frac{0.35}{BW_{3dB}}$$

(4.6)

which is about 42 ps. The internal energy dissipation of the driver is expected to be much smaller than the charging energy of the NLTLs; to verify this assump-
Figure 4.18: Simulated s-parameters for the CMOS driver circuit indicating its small-signal forward gain and output matching. The unity-gain frequency is approximately 22 GHz.

In this section we simulate the inverter under large-signal switching in the time-domain (output open-circuit, 100 ps input falltime). Integrating the supply current over one full cycle shows 950 fJ energy dissipation, which is small, but not insignificant compared to the NLTL energies from Eq. (2.38).

Because loss reduces the conversion efficiency and damps high-frequency signals in a practical NLTL, each device has an optimal number of sections resulting in the largest ringing amplitude. While the issue of loss on NLTLs is analytically complicated in the large-signal case, the optimal length can be estimated by simulating a sufficiently long device and finding the maximum ringing amplitude along the line. Fig. 4.19 shows the peak-to-peak ringing amplitude at the first 17 sections for the 10, 20 and 30 GHz NLTLs. In each case, the amplitude reaches a peak and decays slowly. Interestingly, the peak amplitudes are close to the amplitude of the input waveform, suggesting the NLTL reaches
Figure 4.19: Peak-to-peak ringing amplitude at each section of the 10, 20, and 30 GHz NLTLs showing an optimal length around 9 sections.

A balance between different harmonics, rather than being limited by loss. By 9 sections, both the 10 and 20 GHz devices have reached their peak amplitudes, whereas the 30 GHz device has reached ~79% of its final amplitude. Therefore 9 sections is chosen as the length for all NLTLs in further simulations as a compromise between output amplitude and charging energy, as well as the layout area required for implementation.

The complete transmitters (NLTL with driving inverter) are simulated in the time-domain using the Spectre simulator and using a 1.5 V square-wave with 100 ps rise/fall times as input. The resulting output waveform shows strong ringing on the falling edge despite the small number of sections. Waveforms at the input and output of the NLTLs with 10 and 60 GHz Bragg frequencies are shown in Fig.4.20 when the output is terminated with a 50 Ω load. The ringing frequencies are about 9.96 and 59.8 GHz, respectively, estimated from the peak-to-peak spacing and the initial amplitude is over 700 mV for the 10 GHz device.
The output amplitude is somewhat smaller than predicted in Fig. 4.19 due to the impedance mismatch at the end of the NLTL, as the NLTL’s impedance varies significantly with both bias and frequency. Some ringing is also seen at the input due to the rapid sharpening of the waveform and the output reflection. However, an absence of any overshoot on the rising edge indicates that the ringing is generated by the edge sharpening of the NLTL.

To estimate the transmitter’s energy consumption, the simulation must include an antenna load. For simplicity, we use a one-port S-parameter model of a planar-elliptical dipole UWB antenna scaled such that the antenna’s 10 dB cutoff lies just below the Bragg frequency. This type of antenna is chosen for its simple geometry and broadband response; it has a high return-loss (well matched) half-wave dipole resonance just above its 10 dB cutoff, and a flat broadband response with 12-15 dB return loss at higher frequencies [30]. The S-parameters are computed in HFSS and scaled in frequency to provide the appropriate low-
With each NLTL connected to the antenna load, we integrate the current supplied to the driving inverter over one transmit cycle to obtain the energy consumed per pulse. The resulting energy consumption follows our proposed scaling rule (Table 4.3), but differs from the ideal case, Eq. (2.38), by roughly a factor of three. We simulate the 10 GHz NLTL again with an open-circuit termination and find that the energy decreases only slightly, by about 1.5 pJ. Therefore, we conclude that the additional energy consumption is dominated by process-dependent losses, such as the inductors’ Q-factor, rather than radiation by the antenna.

### 4.4.2 Experimental Devices

NLTL transmitters targeting 10 GHz, 20 GHz, and 30 GHz were fabricated in IBM’s 0.13 µm RF-CMOS technology. The total area of each transmitter is about 1mm x 0.25mm. All devices are driven by on-chip CMOS inverters powered from a 1.5V supply. Input data is provided by a pulse-pattern generator with 60ps nominal rise/fall times. The 10 GHz device is shown in Fig. [4.21](#). When characterizing the NLTL and transmission properties, the input waveform is a 10 ns square pulse from 0 V to 1.5 V with a 32 MHz repetition rate.

The pulse generation properties of the transmitters are verified by connecting the NLTL output to a high-speed sampling oscilloscope. Each device shows significant ringing on the falling edge. The ringing frequencies can be estimated by the peak-to-peak interval. These frequencies are 9.37 GHz, 16.1 GHz, and 22.5 GHz for the 10, 20, and 30 GHz designs, respectively. The lower observed
ringing frequencies may be explained by the reduction in the pulse amplitude, resulting in a higher $V_L$ and therefore larger $C(V_L)$. Similarly, the peak amplitudes are 243 mV, 158 mV, and 120 mV, respectively. The reduced amplitudes for the higher frequency designs are mostly due to the falltime of the driving inverter. These NLTLs require more edge-sharpening to reach their minimum falltime, so fewer sections are available to build ringing. Additionally, propagation loss in the test-cables connecting to the oscilloscope further reduces the observed amplitudes.

The wireless transmission properties are then tested by connecting a small monopole antenna, consisting of the extended center conductor of a coax adapter, to the wafer probe at the NLTL output. It should be noted again that the transmission properties of the antenna can heavily influence the radiated pulse shape. The present monopole is approximately 4 mm long, with a large width/length ratio. Return loss measurements suggest it provides broadband radiation above 10 GHz (defined by $RL > 10$ dB), making it appropriate for the 10 GHz transmitter. Attempts to fashion similar antennas for the 20 and 30 GHz devices by trimming the center conductor resulted in unacceptable return loss.
Figure 4.22: Time-domain waveforms for each NLTL transmitter measured with a sampling oscilloscope (50 Ω termination). The 10 GHz line shows significant ringing; reduced amplitude and duration for the higher frequency lines is mostly due to the fixed input falltime.

The transmitted waveform is received by a home-built TEM horn antenna and recorded with the high-speed sampling oscilloscope. At short distances, about 20 cm, the received waveform is sufficiently strong to be recorded without amplification. The waveform received from the 10 GHz transmitter is shown in Fig 4.24 and consists of a 500 ps pulse centered around 10 GHz. The power-spectral density of the pulse is measured by a spectrum analyzer with the transmitter running at 32 Mpulse/s and the resulting comb spectrum is shown in Fig 4.25. The pulse spectrum has a maximum at 10.7 GHz and a 10 dB bandwidth of nearly 2.1 GHz.
Figure 4.23: Photograph of the experiment setup showing the monopole antenna attached to the wafer probe.

Figure 4.24: Pulse waveform received from the 10 GHz transmitter at a distance of 20 cm by the TEM horn. Center frequency and duration are about 10-11 GHz and 500 ps, respectively.
4.4.3 Power Consumption & Scaling

Since the NLTLs are driven by on-chip inverters, the total energy required for pulse transmission can be estimated from the supply current draw at a fixed pulse-repetition rate. At a PRF of 32 MHz, the 10, 20, and 30 GHz devices draw 472, 390, and 340 \( \mu A \), respectively, from a 1.5 V supply. The supply current and output waveforms are observed to be independent of the input pulse width for pulses in the range of 1-50 ns. Similarly, for PRFs up to around 100 MHz, the ringing and transmitted pulses are unchanged; the current increases linearly with the PRF, as expected.

The resulting energy consumption of each transmitter is shown in Table 4.3 in terms of the energy per pulse, with the ideal values calculated with Eq. (2.38). The figures are roughly 2x larger than the simulated values, owing to the high-frequency accuracy of the foundry models and any additional loss mechanisms.
Table 4.3: Energy-per-pulse requirements for the NLTL transmitters.

<table>
<thead>
<tr>
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</thead>
<tbody>
<tr>
<td>5</td>
<td>12.02</td>
<td>35.34</td>
<td>22.1</td>
</tr>
<tr>
<td>10</td>
<td>4.85</td>
<td>15.84</td>
<td>22.1</td>
</tr>
<tr>
<td>15</td>
<td>3.90</td>
<td>11.67</td>
<td></td>
</tr>
<tr>
<td>20</td>
<td>2.98</td>
<td>9.58</td>
<td>18.3</td>
</tr>
<tr>
<td>25</td>
<td>2.36</td>
<td>2.36</td>
<td></td>
</tr>
<tr>
<td>30</td>
<td>1.84</td>
<td>6.79</td>
<td>15.9</td>
</tr>
<tr>
<td>60</td>
<td>0.982</td>
<td>4.69</td>
<td></td>
</tr>
</tbody>
</table>

The best performing device in terms of energy is the 30 GHz transmitter with 15.9 pJ/pulse.

The large discrepancy between the ideal NLTL and simulation (Fig.4.26) is likely due to losses in the spiral inductors and underscores the importance of technology dependent factors in the transmitter performance. The required series inductance also scales with $1/f_B$, so the scaling law in Eq. (2.39) is preserved, although with a different slope, when inductive losses are considered. Similarly, the difference between the simulated and measured devices grows with increasing Bragg frequency as the lumped element models lose accuracy. One contribution is the resistance in the wiring connecting the inductors and varactors, which becomes increasingly significant (and geometry-dependent) at higher frequencies. Additionally the IBM device models used in our simulations tend to overestimate the Q-factor of small MOS varactors under negative bias [31]. This limitation is particularly relevant to our NLTLs because the ringing occurs when the line has reached negative gate-source voltages. However, the Q vs frequency trend is consistent between IBM’s models and measurements, so these errors are not expected to cause significant deviations from the $1/f_B$ energy scaling characteristic.

65
Figure 4.26: Energy-per-pulse scaling with frequency for lumped-element NLTLs.

4.4.4 Simulated 60 GHz NLTL

A closer examination of the 60 GHz design is warranted as it operates far beyond the cutoff frequency of the driver and beyond the frequencies typically used for passives in the 0.13 µm technology. The s-parameters for the NLTL are shown in Fig. 4.27, showing a cutoff frequency close to 60 GHz with insertion loss around 15 dB. In this case, both the inductors and varactors have self-resonant frequencies (SRF) slightly above \( f_B \), which turns out to be somewhat beneficial. The inductors are chosen from the standard cell library to have the maximum possible peak-Q frequency, however their nominal inductance is too small for a 50 Ω, 60 GHz line. But because their SRF is slightly above 60 GHz, the effective inductance at \( f_B \) increases, at the expense of loss, resulting in the correct cutoff frequency. Fortunately, the inductor geometry is a single turn spiral, which eliminates the proximity effect losses between the windings of multi-turn spirals, resulting in a much larger peak-Q. Although the inductors
are operated above their peak-Q frequency, the value of Q shown in Fig. 4.5 at 60 GHz is approximately equal to the peak-Q of the inductors used in the lower frequency NLTLs.

Similarly, the varactors see a high-frequency enhancement due to self-resonance. In simulation the varactors are given a series inductance of 10 pH to approximate their wiring inductance (which would require careful 3D-EM simulation to predict accurately), however their SRF changes with DC bias over a wide range (due to the change in capacitance). This results in a significant enhancement to their nonlinearity, shown in Fig. 4.4 as $C_{\text{max}}$ nearly doubles (low SRF), while $C_{\text{min}}$ remains nearly constant (high SRF). Therefore falltimes may be compressed in a shorter length of NLTL than otherwise expected (as long as the falltime is short enough to see the enhancement), allowing the 60 GHz design to develop strong ringing comparable to the 10 GHz NLTL, despite its relatively slower input.
The initial success of the 60 GHz design suggests an interesting strategy for enhancing NLTL performance by engineering the SRFs of the inductors and varactors. Care should be taken with this approach, not only because the SRF is difficult to predict accurately (particularly for varactors where it is highly layout dependent), but because the existence of a resonance necessarily implies, by the Kramers-Kronig relations, a large peak in the loss near the resonant frequency. Therefore careful design and measurement should be used to verify the practicality of the 60 GHz transmitter.

4.5 Outlook

Although the current devices have slightly higher energy consumption than other recent proposals, the scaling rule suggests that NLTL-based transmitters are strongly competitive at frequencies above the current FCC defined UWB window. In particular, extrapolating from our measured devices suggests an NLTL transmitter could operate at 60 GHz with only 13 pJ/pulse, assuming that technology factors are controlled. While the output power is low for our present devices, it is not impractical for a reasonable receiver at short to medium distances. This may offer an acceptable trade-off versus battery life and antenna size for ultra-light applications. Additionally, the transmitter is relatively straightforward to design and offers the possibility of exceeding the driving transistor’s $f_T$, which minimizes engineering and fabrication costs.
CHAPTER 5
WIRELESS LINK

The ultimate goal of this research is to demonstrate a complete wireless link between the Manduca Sexta moth and a basestation receiver.

5.1 Link Requirements

The endpoints of the communications link are the electronics package on the moth, and the basestation control center. The moth can carry about 700 mg in flight and has available electrical power in the range of 500-750 $\mu$W \cite{4}. By contrast, the basestation is expected to be carried by a person or a vehicle, allowing several kilograms of mass and potentially several watts of available power. The link is therefore highly asymmetric and many of the constraints placed on the transmitter can be satisfied by increasing the complexity of the receiver. While the complete basestation is beyond the scope of this dissertation, it is worthwhile to reflect on its design at the system level.

The moth uses the NLTL and a Super-regenerative receiver (SRR, described later in this chapter) for its transmit and receive components, respectively. Both devices conserve power by aggressive duty cycle scaling: the NLTL transmits pulses of 500 ps duration; the SRR “listens” for incoming signals during a similarly short window. Therefore, synchronization between the moth and basestation is essential, but the difficulty is easily resolved by taking advantage of the asymmetry in the link (Fig.5.1). Because the basestation is relatively unconstrained in power, it can transmit and receive with a long symbol duration. The basestation transmits an Amplitude Shift Keyed (ASK) or On-Off Keyed (OOK)
Figure 5.1: Conceptual diagram for the complete moth-basestation wireless link. Plots show the operation of each receiver, as intended to relax synchronization requirements at the expense of basestation power. The moth uses a short time window to receive a long-duration pulse, while the basestation uses a long time window for a short duration pulse.
signal to facilitate the moth’s use of an SRR. Similarly, to ease the synchronization burden, the basestation uses a high-gain energy collection receiver which measures the total energy received in a given bandwidth and time window.

A more quantitative system design should include Bit-Error Rate and component requirements for different basestation architectures. However, the operating environment of the moth is unknown, so a range estimate and channel model of radio propagation are not available. Therefore, the system architecture proposed above may not be optimal, but the requirements placed on the basestation are expected to be achievable by conventional microwave engineering.

5.2 Transmit Link Measurements

The moth to basestation transmission can be demonstrated in the laboratory environment. The basestation receiver is constructed from available off-the-shelf microwave components, including a frontend amplifier and lowpass filter from Picosecond Pulse labs, a passive diode mixer, and a benchtop realtime oscilloscope and signal generator (as the local-oscillator). The receive antenna is a home-built TEM-horn with peak on-axis gain near 10 GHz. The lumped-element NLTL transmitter described in Section 4.4 is connected to the monopole antenna, and a pulse train of all ones is provided as the baseband data by a 32 MHz square wave. The basestation is mounted on a equipment cart which can be moved around the lab to vary the link distance.

Under these conditions, pulses from the NLTL can be identified by the receiver and detected on the oscilloscope with a reasonable signal-noise ratio out to a distance of roughly 2 meters. The received pulse train is shown in Fig 5.2.
Figure 5.2: Pulse train received from the NLTL transmitter at 1 meter by the direct conversion receiver. Relatively long bursts can be received despite the lack of coherence with the transmitter.

with a block diagram of the receiver inset. While this validates the NLTL transmitter, it should be noted that this test in no way indicates the best possible performance of the link. The frontend amplifier provides only 13 dB gain with a noise figure as high as 6 dB, leaving significant room for improvement even among off-the-shelf microwave amplifiers. Further, no baseband amplifier (between the filter and oscilloscope) was used; and the oscilloscope has a much higher noise floor than would be present in a well-designed detector. The success of this test, despite the crude receiver design indicates strong potential performance from the existing NLTL transmitter.
5.3 Moth Receiver Design

Design of low-power receivers is often much more challenging than that of transmitters. Fundamentally, a receiver must operate at full power to “listen” for a longer time window because it cannot know the exact timing of the incoming signal. Most of the power budget is consumed by the high-gain, low noise-figure frontend amplifier. For example, a receiver recently proposed for *Manduca Sexta* is based on energy-collection and requires at least 8.4 mW when active, of which 70% represents the LNA [32]. In designing a new receiver for the moth, the super-regenerative architecture is used for its large gain to power ratio and negligible idle-state power (i.e. power consumption is proportional to the symbol-rate).

5.3.1 Super-Regenerative Receiver Theory

The super-regenerative receiver consists of an RF oscillator which is tuned to the frequency of the incoming signal and started/stopped by a control signal called the quench. When the oscillator starts from noise, in absence of a received signal, its amplitude will grow slowly. However, when a weak received signal is injected into the oscillator during startup, the amplitude will grow quickly and the initial phase will be determined by the injected signal. The concept of operation is illustrated in Fig. 5.3, where the quench signal $\zeta(t)$ controls the gain of the oscillator. When the quench becomes negative, the oscillator’s gain exceeds unity and oscillations begin to grow. The sensitivity to injected signals is described by a sensitivity function $s(t)$, which peaks as $\zeta(t)$ crosses zero. Sometime later, when $\zeta(t)$ becomes positive again, the oscillator’s amplitude will reach a
Figure 5.3: Block diagram of a generic superregenerative receiver (a), and its operation in the time-domain (b). Note the slight timing skew between the sensitivity function $s(t)$ and the peak of the incoming pulse.
maximum and begin to decay. The envelope can be extracted by a detector and compared to a threshold value to determine the presence of absence of an RF pulse. Based on these principles, super-regenerative receivers (SRRs) can have gains exceeding 50 dB while consuming sub-milliamp bias currents [33, 34].

The complete theory of SRR operation is quite involved [35, 34] and beyond the scope of this dissertation. The architecture has fallen out of use, in favor of the super-heterodyne receiver, for most applications, primarily because SRRs tend to have poorer frequency selectivity. However, when designing a UWB receiver low frequency selectivity can be advantageous because it allows the receiver to capture the full energy of the UWB pulse [36]. In general, the frequency selectivity is proportional to the Fourier transform of \( s(t) \), so it can be tuned by the quench waveform [33]. Based on this advantage and their low power consumption, SRRs have seen renewed interest from the UWB community [36].

5.3.2 Receiver Circuit Design

The SRR in implemented in the same 130 nm RF-CMOS technology as the NLTLs. The core (Fig.5.4) is a negative-gm oscillator consisting of a cross-coupled n-fet pair with spiral inductors and MOS varactors as the frequency selective tank. The bias current in the oscillator is controlled by the p-fet Quench transistor, which sets the gain of the n-fet pair. Finally, a pair of common-gate p-fets are used to inject external signals into the oscillator.

The oscillator’s envelope is detected and compared to a threshold value by a latched current comparator. Here, two cross-coupled inverters (red in Fig.5.4) are pre-charged to Vdd by the Reset signal (blue). When Reset is deactivated
Figure 5.4: Circuit diagram of the Super-regenerative Receiver. The core is a negative-gm oscillator with bias control and signal injection. The detector is a latched comparator (green) made from cross-coupled inverters (red), with a reset (blue).

(logic high) each inverter attempts to discharge its output node to ground, however, the discharge current is limited by the pass transistors (green). Should the oscillator’s envelope exceed the threshold value, the left-hand inverter in Fig. 5.4 will “win”, discharging its output node faster, and pulling the detect signal high. The main advantages of this detector architecture are that it consumes no power in the idle state, and that the cross-coupled inverters act as a sense-amplifier, where positive feedback allows the circuit to very quickly respond to small differences between the envelope and threshold.

The final layout of the 10 GHz SRR is shown in Fig. 5.5. The circuit occupies approximately 0.25 mm$^2$ of die area. Two spiral inductors with opposite windings are used to prevent spurious coupling from the environment to the oscillator. The fully-differential injected signal pad appear at the top, with the differential oscillator output (with open-drain buffers) at the bottom.
5.3.3 Measured Receiver Performance

The large number of I/O signals required by the receiver requires a more involved test setup than for the NLTLs, shown in Fig.5.6. The Keithley 2410 is used for power, and to provide the precise measurements of supply current needed to estimate the SRR’s power consumption. The remaining DC control signals, which tune the oscillator’s frequency, and set the detector’s threshold, are provided by the Agilent supply. An RF signal generator capable of low-level output power (down to -110 dBm) is used to provide a single-ended inject signal, while the high-speed pattern generator is used to provide precise timing of the Quench and Reset signals. The Detect and Oscillator output signals are measured with an Agilent DCA-86100 sampling oscilloscope.

Initially, the response of the oscillator is tested at various inject power levels to determine the sensitivity receiver. With the output measured on the sam-
Figure 5.6: Instrument configuration for the SRR. The intended Quench/Reset timing is generated by the 8133a pattern generator. Outputs are monitored with a sampling and realtime oscilloscopes.

Sampling scope, several startup cycles are averaged together, giving a rough estimate of the envelope amplitude. Comparing the envelopes measured with a 50 ns quench time in Fig.5.7, the receiver appears to operate in the logarithmic mode (envelope amplitude saturates), while dropping to the linear mode (amplitude depends on inject power) for lower power levels. The ultimate sensitivity, where the amplitude is indistinguishable from oscillations building from noise (self-start), appears to be about -65 dBm. The best observed performance is 890 $\mu$A current from a 1.2 V supply, equivalent to 107 pJ/pulse.

Unfortunately, the first generation SRR circuit proved to have several problems, including poor long-term stability, as well as significant pulse-to-pulse variation. The apparent cause is a floating node in the detector circuit (the above
measurements were taken with the detector disabled). Due to this condition, more detailed measurements of the sensitivity and Bit-error rate could not be obtained. However, the observed sensitivity of -65 dBm is considerably poorer than previously reported designs, which have sensitivities of -80 to -90 dBm \cite{37, 33}. Considering future designs, the latched comparator should be replaced with a squaring circuit. The clocked nature of the comparator presents too many opportunities for back-coupling from the detector to the oscillator, in addition to being prone to the floating node problems which plague the current design.

Regardless, of its failings, the present circuit shows a strong potential for super-regenerative receivers for the *Manduca Sexta* project, as well as similar applications. Its power dissipation is considerably lower than previously proposed energy collection receivers while maintaining similar or better sensitivity \cite{32}. The power dissipation is similar to the current state-of-the-art found in the
literature [38], which is a coherent receiver operating at sub-1 GHz frequencies.

5.4 Outlook

The crucial components of the moth to basestation radio link have been successfully demonstrated. Although the output power of the NLTL transmitter is low, it is clearly interoperable with a simple, unoptimized receiver at indoor distances. Simple improvements to the direct conversion receiver in Section [5.2] such as increasing the LNA gain to 30 dB and adding a baseband amplifier of 15 dB (both commercially available) would increase the range to 100 meters. Before that point, however, more detailed considerations apply, including sources of noise and interference. Development of a realistic channel model for the moth’s environment should be a priority in future system integration work.

Additionally, a very promising receiver has been presented, with applications mobile devices in general. The SRR circuit will require a redesign and tape-out to be completely functional. However, the power consumption measured in the first generation device is the lowest in the 10 GHz frequency range. If this figure can be confirmed in the revised design, and the sensitivity improved, the SRR will complete the basestation to moth link and allow low-power, high data rate, bidirectional communication with the insect.
CHAPTER 6
POLYMER & FLEX NLTLS

The experiments with CMOS NLTL transmitters have shown some potential for this technology, but the low output power levels and difficulty controlling the pulse envelope make it unlikely to be a strong competitor against more conventional UWB circuits. The biggest limiting factor for the CMOS NLTLs has been losses due to the conductive Silicon substrate. However, NLTLs have a fundamental advantage in that edge-sharpening can generate signals faster than the $f_T$ of the transistors in the same technology [27]. In CMOS, however, increasing high-frequency losses will prevent higher frequency NLTLs from being effective. Instead, the NLTL transmitter will be most advantageous in a technology where transistor $f_T$ is low, but the quality factor of passives is still high. The field of organic and flexible electronics presents such an opportunity: the low mobilities of such materials limit the transistors to frequencies on the order of 10 MHz, while low dielectric losses allow passive inductors and capacitors to be fabricated with self-resonant frequencies as high as 10 GHz [39].

In designing an NLTL for flexible electronics applications, one requires a high-quality, high-frequency, voltage-variable capacitor. Organic and amorphous semiconductors should not be relied on, as their carrier mobilities vary by orders of magnitude, depending on processing methods; low mobility and low carrier density can increase the dielectric relaxation time into the nanosecond range for these materials. At the same time, any material used must be processable at the low temperatures (150º C) required by flexible plastic substrates. One candidate is the ferroelectric polymer poly-vinylidene fluoride (PVDF). This material is frequently used for ultrasound transducers due to its
strong piezoelectric coefficient and ability to support large strains \[40, 41\].

As a ferroelectric material, PVDF supports a permanent electric polarization in absence of an applied field, which implies a nonlinear Polarization vs Field (P-E) response, and thus a voltage variable capacitance \[42, 43, 44\]. Fig. 6.1 shows the molecular structure of this material, along with the origin of its ferroelectric and piezoelectric properties. A dipole moment exists between the electro-positive Hydrogen and electro-negative Fluorine on opposite sides of the polymer chain; when many chains crystallize into similar orientations, a net polarization exists in the material. Similarly, the application of an electric field changes the distance, \(d\), between the Hydrogen and Fluorine, causing the material to expand and contract. Applying a sufficiently strong electric field will cause the chain to rotate, reversing the direction of polarization.

PVDF has been previously used in thin-film capacitors \[45, 46, 47, 48\], and the dielectric properties of PVDF have been well studied by the ultrasound community over the frequency range of kHz - 100 MHz. These samples, which are typically several microns thick and composed of a blend of crystalline and amorphous phases, show large dielectric loss tangents in the range of 1-10 MHz.
However, these loss tangents, as well as the relaxation processes in the microwave range, are associated with grain boundaries and crystalline-amorphous boundaries in the material [49, 50]. At the same time, optical second harmonic generation experiments show that the nonlinear P-E characteristics are still present at optical frequencies [51], although the ferroelectric polarization cannot reverse that quickly.

Because the dielectric losses are related to grain boundaries, single grain PVDF shows a very square, low-loss hysteresis loop. To isolate single ferroelectric grains, PVDF must be patterned into isolated islands not much larger than the characteristic grain size, about 100 nm. Arrays of single grain PVDF islands have been previously demonstrated by nano-imprint lithography and shown to have very low piezoelectric losses [52].

### 6.1 Device Concept

The target device is an NLTL using an array of Metal-Insulator-Metal (MIM) PVDF capacitors its nonlinear elements. The fabrication should require three steps: patterning the bottom metal, imprinting the PVDF islands, and patterning the top metal to form the NLTL. The metal layers can be deposited by evaporating Chromium and Gold, and patterned with standard photolithography.

The key step, imprint lithography onto PVDF requires a template to be fabricated on a Silicon wafer. Imprint lithography, diagrammed in Fig. 6.2 works by pressing the template into a PVDF thin film at elevated temperature and pressure. After the PVDF undergoes plastic deformation, the template is removed and the isolated islands remain on the original substrate.
Figure 6.2: Diagram of nano-imprint lithography. The rigid template is pressed into the polymer film at elevated temperature and removed once the pattern has been transferred.
6.2 Fabrication Procedure

Arrays of PVDF islands, and the nano-imprint template are fabricated in CNF, with some steps (spin coating PVDF) performed in the NBTC laboratories. In all experiments, the material used is PVDF-TrFE 70/30 copolymer from Piezotech.

6.2.1 Template Fabrication

The template wafer is fabricated first using electron beam lithography to pattern an array of ridges in HSQ e-beam resist. HSQ is a negative resist which transforms into SiO2 upon exposure to an electron beam. The fabrication process is outlined in Table 6.1. The pattern in the HSQ consists of lines 20-50 nm wide, and 100-200 nm apart, covering an area of 100 µm by 100 µm, and requires a high-contrast development process to resolve such dense, small features. Initial attempts showed that the standard dilute TMAH developer does not provide enough contrast to resolve the pattern. A more aggressive developer (KOH) was used, but required high e-beam doses (higher than 8000 µC/cm²) which greatly limited the size of the patterns which could be exposed in a reasonable time. In order to reduce the dose, a concentrated TMAH developer was heated to 50º C prior to use, resulting in a very high-contrast pattern at low doses (1500 µC/cm²). Finally, the underlying Silicon is anisotropically etched with a Chlorine-chemistry RIE and the HSQ is removed, resulting in a flat top surface and smooth sidewalls.
Table 6.1: Process sheet for template fabrication.

<table>
<thead>
<tr>
<th>Step</th>
<th>Equipment</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wafer Clean</td>
<td>SC1/SC2 Hamatech</td>
<td></td>
</tr>
<tr>
<td>Pre-bake</td>
<td></td>
<td>170°C, 2 min</td>
</tr>
<tr>
<td>Spin Resist</td>
<td>HSQ, 6%</td>
<td>100 nm thick</td>
</tr>
<tr>
<td>Bake</td>
<td></td>
<td>170°C, 2 min</td>
</tr>
<tr>
<td>E-beam exposure</td>
<td>JEOL 9300</td>
<td>Dose 1500 µC/cm²</td>
</tr>
<tr>
<td>Develop</td>
<td>MIF-312</td>
<td>50°C, 1 min</td>
</tr>
<tr>
<td>Silicon etch</td>
<td>Plasmatherm 720</td>
<td>100 nm deep</td>
</tr>
<tr>
<td>Remove HSQ</td>
<td>Buffered HF</td>
<td></td>
</tr>
<tr>
<td>Template Anti-Adhesion</td>
<td>MVD-100</td>
<td>Vapor phase FOTS</td>
</tr>
</tbody>
</table>

Figure 6.3: SEM images of a complete nanoimprint template

6.2.2 PVDF Fabrication

PVDF is highly chemically resistant, but it can be solution processed using an appropriate polar solvent. The most common solvent used is N,N-dimethylformamide (DMF), which preferentially deposits PVDF in the ferroelectric phase [53]. A dilute solution of PVDF in DMF is used to spin coat thin films between 50 nm and 200 nm onto Silicon wafers. However, because DMF is very hygroscopic, spin-coating must be done under less than 20% relative humidity or
Table 6.2: Process sheet for single grain PVDF islands

<table>
<thead>
<tr>
<th>Step</th>
<th>Equipment</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wafer Clean</td>
<td>SC1/SC2 Hamatech</td>
<td></td>
</tr>
<tr>
<td>Adhesion Layer</td>
<td>MVD-100</td>
<td>Vapor phase APTMS</td>
</tr>
<tr>
<td>Spin PVDF</td>
<td></td>
<td>Closed bowl, &lt;20% RH</td>
</tr>
<tr>
<td>Imprint</td>
<td>NX-2500</td>
<td>350 bar, 140°C, 6 min</td>
</tr>
</tbody>
</table>

Water absorption will cause the polymer to precipitate from the solution, resulting in a cloudy and unusable film. After some experimentation, it was found that PVDF must be spun outside the cleanroom (which is maintained at 40% RH), and with a closed bowl, creating a solvent-vapor saturated atmosphere.

Nanoimprint lithography is performed on the spin-coated PVDF films without further processing. The template and sample wafers are placed face-to-face and imprinted with the Nanonex-2500 imprint lithography tool at a pressure of at least 350 bar. The sample temperature is raised to 140°C, corresponding to the softening point of PVDF, during the imprint. A short imprint time of 6 minutes is sufficient to transfer the pattern, but a longer time (up to 30 minutes) can be used to completely anneal the islands into single ferroelectric grains.

### 6.3 Results

After some experimentation, mostly due to the difficulty in obtaining high quality PVDF films with the lack of humidity control in NBTC, good PVDF nano-islands were obtained on a Silicon substrate. Islands 200 nm wide with 50 nm spacing are shown in Fig 6.3(a). Smaller islands with dimensions of 100 nm require an adhesion layer of monolayer aminopropyltrimethoxy-silane prior to
Figure 6.4: SEM images of PVDF nano-islands. (a) shows a portion of a 100 µm by 100 µm array on Silicon. (b) shows a wider area view of a similar imprint. The island dimensions are about 200 nm with 50 nm spacing in both cases.

spin coating to prevent them from being torn away when the template is separated from the sample. An additional experiment imprinted PVDF islands on top of a 100 nm gold layer, resulting in small patches of good islands, surrounded by unimprinted PVDF film. A typical patch of islands is shown in Fig 6.3(b).

6.3.1 Piezo-Response Force Microscopy

While the ferroelectric properties continuous thin films can be investigated direct electrical measurements, the isolated PVDF nano-islands require a technique with high spatial resolution to verify their characteristics before integration into a macroscopic device. Scanning probe microscopy techniques allow an electrode on the order of 10 nm size to be placed in contact with a sample to measure the conductivity or mechanical properties. Fortunately, in PVDF ferroelectricity also gives rise to the converse piezoelectric effect, coupling the
mechanical and electronic responses. The ferroelectric properties can therefore be indirectly probed via the material’s mechanical response to an applied field.

The experimental technique Piezo-response force microscopy (PFM) is readily available on modern atomic force microscopes (AFMs) without additional hardware [54, 55, 56]. In this technique, shown in Fig. 6.5, an AC voltage is applied to the (conductive) AFM tip during scanning, and the vertical deflection signal is sent to an additional Lock-in amplifier, which provides the magnitude and phase of the piezoelectric response. The Lock-in amplifier is required to isolate the cantilever deflection due to piezoelectricity from that due to the sample topography, as well as to recover very small deflections (typical piezoelectric coefficients are on the order of 10-100 pm/V) in the presence of noise. The resulting amplitude and phase data provide estimates of the out-of-plane piezoelectric coefficient, $d_{33}$, and ferroelectric domain orientation, respectively.
To understand the PFM phase response, consider the polarisation vs electric field (P-E) hysteresis loop of a typical ferroelectric material, shown in Fig. 6.6. The piezoelectric response will have a “butterfly” hysteresis characteristic (insets) with minimal deflection, $\delta$, when the polarization is zero (i.e., at the coercive field, $E_C$), while above or below $E_C$ the material will be seen to expand. However, the coercive voltage may be positive or negative, depending on the direction of the ferroelectric remnant polarization. Thus, applying a small-signal stimulus about zero volts, the material will either expand or contract during the positive half-cycle, depending on its polarization state, resulting in either a $0^\circ$ or $180^\circ$ phase shift of the deflection signal, respectively. By monitoring the PFM phase signal, one can therefore detect the boundaries and orientations of
ferroelectric domains in the sample

6.3.1.1 PFM Imaging Results

The PFM measurements were carried out using a Dimension 3100 AFM with a Nanoscope IV controller and MESP cantilevers \( k \approx 5.0 \text{ N/m} \) from Veeco. The microscope was calibrated with a periodically-poled Lithium Niobate reference (Veeco) and used to scan a 160 nm continuous PVDF film on Silicon, as well as 200 nm islands imprinted into a similar PVDF film. The continuous film was annealed at 145ºC for 30 minutes to promote formation of \( \beta \)-phase grains.

Initially, both the continuous film and the islands are randomly polarized, exhibiting minimal piezoelectric response. To confirm the ferroelectric response, the AFM is set to scan a 2 \( \mu \text{m} \times 2 \mu \text{m} \) region with a DC tip voltage of -12V. At the conclusion of this scan, the AFM scans a 1 \( \mu \text{m} \times 1 \mu \text{m} \) area with a DC voltage of +12V, yielding concentric regions of opposite polarization. A PFM scan is then performed on the 4 \( \mu \text{m} \times 4 \mu \text{m} \) area containing this region, giving a high contrast PFM phase image. The resulting image from the continuous film in Fig 6.7a shows square regions of strong positive and negative polarization.

The polarization experiment is repeated on the PVDF islands, yielding regions of oppositely poled islands. In this case, however, the scans axes are oriented at roughly 45º with respect to the island edges. Upon performing the PFM scan, the boundaries of the polarized regions follow a zig-zag pattern coincident with the island edges, best visible at the bottom and right edges in Fig. 6.7b, suggesting that each island is a single ferroelectric domain which responds to the tip voltage as a unit.
Figure 6.7: Piezo-response phase images of a continuous PVDF film (a), and 200 nm PVDF islands (b). Significant crosstalk between the topography and PFM signals is evident in (b), however, the single-domain nature is evident in the islands in the bottom right of the poling region.
Unfortunately, the PFM contrast is limited when scanning the PVDF islands; despite the narrow bandwidth of the Lock-in amplifier, crosstalk from the topography signal nearly dominates the image in Fig. 6.7b. In this case, the topography signal consists of a reasonably flat surface (island tops) separated by narrow, deep trenches, which can be reasonably approximated by Dirac delta functions. Therefore, it is impossible to eliminate crosstalk from the topology regardless of Lock-in bandwidth. Further, the depth of the trenches is greater than 50 nm - two orders of magnitude larger than the expected piezo-response deflection (single angstroms). The nature of this particular topography therefore poses a significant challenge for isolating the PFM signal.

6.3.1.2 Material Properties from PFM

More quantitative information about specific points on the sample can be obtained by leaving the AFM tip in a fixed location and applying the voltage stimulus. The piezoelectric response can be directly measured by sweeping the tip voltage while monitoring the deflection signal. These sweeps are performed on both the continuous film and islands using a softer cantilever (SCM-PIC, $k \approx 0.2$ N/m) to improve the measurement sensitivity. (The stiffer MESP cantilever was found reduce the displacement of the PVDF film.) In both cases, the deflection response is near the noise floor of the instrument despite averaging up to 16 sweeps, but the $d_{33}$ coefficient and coercive field strength can be estimated.

The deflection responses from both samples are post-processed by fitting a linear characteristic to the measured deflection at higher voltages. In this case, a threshold of $\pm 4V$ is used to prevent the fitting from being affected by the noise near zero deflection. Combining the linear fits of the negative and positive seg-
Figure 6.8: Cantilever deflection vs. voltage for the continuous film (a), and 200 nm islands (b).
ments of the forward and reverse sweeps approximately recovers the “butter-
fly” hysteresis characteristic expected for piezoelectric displacement. Data for
the continuous film and islands are shown in Fig. 6.8. The coercive voltage
is taken to be the intersection of the positive and negative segments, while \(d_{33}\)
is the slope of the linear fit. The extracted \(d_{33}\) values are in the range of 30-50
pm/V, consistent with the commonly accepted value of 38 pm/V for PVDF-
TrFE. The minimum displacement points are consistent with a coercive field
around 10 MV/m, somewhat lower than expected, but still consistent with the
literature [52].

Finally, a qualitative comparison between the continuous film and islands
can be drawn from the noise levels present in Fig. 6.8. The high noise level
observed for the continuous film suggests some randomization of the effective
ferroelectric polarization near the AFM tip, and thus the presence of multiple
domains and amorphous-crystalline boundary regions. By contrast, this ran-
domization is significantly reduced in the islands sample, suggesting a more
square P-E hysteresis loop. Therefore, one may conclude that the formation of
isolated islands improves the ferroelectric response and reduces the influence of
relaxation processes associated with the amorphous-crystalline boundaries.

6.4 Capacitor Fabrication

The most common topologies for microwave capacitors are metal-insulator-
metal (MIM) or metal-insulator-semiconductor (MIS), which differ by the mate-
rial used for the bottom contact. Many authors in the literature also distinguish
between an insulator and a ferroelectric, so a Au/PVDF/Au structure would be
called an MFM capacitor to distinguish it from an MIFS or MFIS capacitor where an additional (linear) dielectric separates the PVDF from one of the electrodes. Since the process development of PVDF islands is done on silicon wafers, but MFM and MFIS capacitors are made, but ultimately only the MFM is applicable to flexible electronics because the ferroelectric-semiconductor interface significantly alters the characteristics of the device [57].

Various MFM capacitors were made with continuous PVDF films and islands with thicknesses in the range of 50-160 nm roughly following the procedure in Table 6.3. Bottom metal was evaporated onto cleaned wafers in various combinations of Ti/Au, Cr/Au, Cr/Au/Cr, and Cr. PVDF was then applied by spin-coating and a top-metal layer consisting of Au or Cr/Au was thermally evaporated after annealing the PVDF at 140°C for 30 minutes. The top-metal was then patterned by traditional photolithography and wet-etching to form individual 200 µm by 200 µm capacitors.

One major difficulty encountered during fabrication was removal of the photoresist. While PVDF is inert to the exposure and development steps, it is also an organic polymer and tends to be susceptible to the same attacks at the resist. The most common methods of resist stripping are soaking in Acetone, a “strip bath” of heated N-methyl-2-pyrrolidone (NMP), and O₂ plasma ashing. Both Acetone and NMP are known to be effective solvents for PVDF and were observed to remove it from the wafer. In fact Acetone was already used to strip defective PVDF films from old wafers earlier in the project. Similarly, PVDF films placed in the resist strip bath were detached from the wafer by NMP, even where covered by metal. Plasma was thought to leave PVDF unaffected when covered by metal, however further measurements of these devices showed large
Figure 6.9: PVDF islands on gold (a-d) for various island sizes and spacings, and chromium (e,f) for 200 nm islands with 50 nm spacing. The imprint on Au formed clusters due to deformation of the metal, while the imprint on Cr formed islands over a large area with some defects. Individual dark and light islands in (e) and (f) indicate spontaneous polarization of the PVDF.
Table 6.3: Process Sheet for MFM capacitors.

<table>
<thead>
<tr>
<th>Step</th>
<th>Equipment</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wafer Clean</td>
<td>SC1/SC2 Hamatech</td>
<td></td>
</tr>
<tr>
<td>Bottom Electrode</td>
<td>SC4500 Evaporator</td>
<td>Au/Cr 10/100 nm, e-beam</td>
</tr>
<tr>
<td>Spin PVDF</td>
<td></td>
<td>Closed Bowl, &lt;20% RH</td>
</tr>
<tr>
<td>Imprint [Optional]</td>
<td>NX-2500</td>
<td>350 bar, 140ºC, 6 min</td>
</tr>
<tr>
<td>Anneal [if no imprint]</td>
<td></td>
<td>145ºC, 30 min</td>
</tr>
<tr>
<td>Top Electrode</td>
<td>SC4500 Evaporator</td>
<td>Thermal Au, 100 nm</td>
</tr>
<tr>
<td>Spin Photoresist</td>
<td></td>
<td>P-20/S-1813, 1 µm</td>
</tr>
<tr>
<td>Pre-bake</td>
<td></td>
<td>90ºC, 1 min</td>
</tr>
<tr>
<td>Expose</td>
<td>AS-200 Stepper</td>
<td></td>
</tr>
<tr>
<td>Develop</td>
<td>Hamatech</td>
<td>MIF-300, 1 min</td>
</tr>
<tr>
<td>Post-bake</td>
<td></td>
<td>115ºC, 1 min</td>
</tr>
<tr>
<td>Etch top metal</td>
<td>Wet etch</td>
<td>Type TFA (Iodide based), &lt; 1 min</td>
</tr>
<tr>
<td>Remove Resist</td>
<td>YES-200 Asher</td>
<td></td>
</tr>
</tbody>
</table>

leakage currents for MFM capacitors (but not for MFIS), suggesting that material at the pattern edges was indeed damaged.

Further difficulties were encountered concerning the composition of the bottom electrode. Gold is the most commonly used metal for commercial PVDF films because it adheres easily and does not create “dead-zones” at the interface (as Aluminum has been reported to do). However, when wet-etching the top metal it was found that PVDF is permeable to the etchant, causing the bottom metal to be removed as well. Chromium barrier layers and bottom electrodes were used to preserve the bottom metal with some success.

Finally, the use of gold in the bottom electrode caused the imprint to fail, preventing the patterning of islands. For a Cr/Au 10/100 nm metal, the im-
print formed islands only in small clusters, leaving a continuous PVDF film in the remaining areas; a typical cluster is shown in Fig. 6.9b. This defect suggests that the hardness of the underlying layers play a significant role in the imprint. Chromium, Silicon, and SiO$_2$ are highly resistant to plastic deformation with hardnesses in the range of 1-10 GPa. Gold, by contrast is much softer, with a Vickers hardness of only ~216 MPa. Therefore it likely that the gold was deformed during the imprint, preventing pattern transfer. While the problem was mostly resolved by using a 50 nm Cr-only electrode (Figs. 6.9e and 6.9f), the issue is likely to prevent imprinting onto a flexible substrate such as Kapton.

### 6.5 Electrical Measurements

Finally, the electrical characteristics of the PVDF capacitors reveal the effects of processing and the Silicon substrate. The polarization of the PVDF film can be measured using the Sawyer-Tower circuit shown in Fig. 6.10 inset; the circuit places the test capacitor in series with a known reference capacitor. The reference capacitor therefore stores the same total charge as the ferroelectric capacitor, so the voltage $V_{\text{pol}}$ is proportional to the polarization of the ferroelectric.

Polarization hysteresis curves were measured for continuous film capacitors with both the MFIS and MFM configurations. For the MFM devices, no hysteresis was seen, but a large (nanoamp) leakage current was present. For the MFIS devices, a constant current was observed when reversing the polarization, shown in Fig. 6.10. The lack of saturation and the constant charging current indicate that the PVDF has been damaged during processing. The PVDF can still be polarized however, and unlike the MFM case, the compensating charge in
the semiconductor blocks reverse conduction until the polarization has been reversed, allowing a hysteresis loop to be observed.

The effects of the semiconductor interface charge are confirmed by C-V measurements of the same devices. When the device is illuminated, rapid carrier generation in the semiconductor allows the interface charge to be inverted quickly - effectively shorting out the semiconductor capacitance. The capacitance of the PVDF then shows the expected “butterfly” hysteresis loop with little effect from the Silicon substrate (Fig. 6.11). However, when the light is removed the PVDF surface charge causes a thick depletion region to form at the semiconductor surface, dominating the ferroelectric capacitance.

The effects of the ferroelectric-semiconductor interface suggest that useful varactors must use the MFM configuration. However, all such devices tested showed conductivities large enough to prevent useful capacitance or polariza-
Figure 6.11: Typical C-V characteristics from a continuous film PVDF MFIS capacitor. Under illumination the capacitor displays a “butterfly” hysteresis loop, but in darkness the semiconductor capacitance dominates.

as the MFIS devices also show strong conduction, it is likely that the PVDF has been damaged during the patterning of the top-metal layer.

6.6 Outlook

Once the appropriate process conditions were determined, the PVDF islands proved quite easy to fabricate, with the main difficulty being the creation of the template. Both the PFM imaging and PFM deflection measurements suggest that the crystallinity and ferroelectric response have been improved by the formation of islands. While these measurements alone are not sufficient, they do suggest that nano-patterned PVDF is a promising direction leading to high-
frequency, flexible electronics varactors. Ultimately, the metal deposition and patterning will need to be resolved, as present methods appear to damage the PVDF during photoresist removal. An effective metallization process will allow direct electrical measurements of the material at gigahertz frequencies, as well as the formation of PVDF-based NLTLs.
CHAPTER 7
CONCLUSION

This dissertation has investigated the possibility of using a nonlinear transmission line as a UWB-IR transmitter, with a remote controlled *Manduca Sexta* moth as the driving application. The Bragg-ringing NLTL represents a new mode of operation for nonlinear transmission lines in general, and a new transmitter architecture for UWB ratios in particular. The primary results presented here are:

1. Experimental demonstration of NLTL wireless transmitters with energy consumption in the range of 10-20 pJ/pulse.
2. Theoretical and experimental demonstration of the energy-consumption vs frequency scaling for NLTL transmitters.
3. Experimental demonstration of a wireless link using the NLTL transmitter.
4. Feasibility study of NLTLs using ferroelectric polymers for flexible electronics.

Additionally, a number of secondary results have been presented:

1. Experimental demonstration of a mostly functional Super-regenerative receiver with energy-consumption in the 110 pJ/pulse range.
2. A tractable theory and design method for semi-discrete NLTLs in the small-signal regime.
3. Development of a fabrication process for ferroelectric polymer NLTLs.
The primary advantages of the NLTL transmitter technique are the reduced hardware complexity and the nature of the energy scaling with frequency. NLTLs are unique in that the energy required to generate a microwave pulse (by charging and discharging the line) scales down with the inverse of the pulse center-frequency. By contrast, in a conventional LC-oscillator, power consumption increases with the square of frequency \[58\]. Similarly, UWB transmitters which synthesize pulses using digital circuits increase device switching and thus, energy consumption, as the center-frequency increases.

7.1 Application to Controlled Insect Flight

The radio transmitter presented in Section 4.4 meets the primary requirements of the insect cyborg system. Then energy consumption of \(~20\) pJ/pulse, while modest compared with the state-of-the-art presented in literature, requires only \(2.2\) \(\mu\)W at the target data rate of 100 kbps, and can achieve \(~225\) kbps at the target power of \(5\) \(\mu\)W. Similarly, improvements to the basestation receiver should allow the link to function at a distance of 100 meters or more. In order to effectively test the transmitter in-flight, it must be integrated with the microprocessor and related circuitry in either a single die or system-in-package configuration, both of which require additional RFIC tapeouts. This system design and test is the focus of later phases of the cyborg moth project.
7.2 Outlook for Future Work

Future research into the radio system should concentrate on low-loss CMOS NLTLs operating at 60 GHz, where their energy consumption (extrapolated from the measurements in Section 4.4) is expected to be more competitive with existing technologies. Additionally, work should continue toward the demonstration of ferroelectric polymer NLTLs. The flexible electronics environment is uniquely suited to this technology thanks to its low transistor cutoff frequencies and high-quality passives. The potential benefits of a successful transmitter are large, since flexible electronics faces significant challenges interfacing power and information with the outside world. In conclusion, these developing areas of technology present strong opportunities for NLTL low-power radio transmitters.


[51] N. Tsutsumi, T. Ono, and T. Kiyotsukuri, “Internal Electric Field and


