Introduction

Wireless systems have become a dominant presence in our lives. From voice communication to data transmission and general connectivity, wireless options existed that allow greater mobility and freedom than their wired counterparts. As more time and money was invested in the wireless industry, products began to show up on the market ranging from cell phones to wireless LAN devices to ad hoc Bluetooth type devices. With the growing need for more bandwidth and increased data rates, in combination with the saturation of existing wireless channels, the FCC is considering opening new spectrums centered at several tens of Gigahertz. While technology has allowed us to push the envelopes of the frequency spectrum and utilize ever-increasing frequencies for communication, our understanding of the propagation characteristics of such high frequency channels has lagged behind.

The task proposed for this undergraduate research project was the design and construction of a Gigahertz sliding correlator channel sounder, an instrument used for the characterization and analysis of wireless channels [1]. This project greatly mirrors the work done by Chris R. Anderson as detailed in his Master's thesis, which has served as an excellent reference for much of the information in this document [1].

The Multipath Problem

At very low frequencies, wireless transmission may approach the textbook concept of transmitting and receiving a single signal [1]. However, as clock rates go up, so does the complexity of wireless transmission. Electromagnetic waves reflect, diffract, and scatter, alleviating the need for direct line of sight between transmitter and receiver, but also creating the complicated multipath problem. Multipath is the nearly infinite number of paths that a signal can take from transmitter to receiver, reflecting off smooth objects, scattering off rough objects, and diffracting around sharp corners [1]. When a high frequency signal is transmitted, it is received not once, but several times at the receiver. Signals that bounce around in the environment and have longer overall path lengths take a longer to reach the receiver. The more the signal bounces off and around objects along a given path, the longer it takes to reach the receiver and, generally speaking, the more it is attenuated.

Introduction to Spread Spectrum

The channel sounder works by sending a broadband signal over a given channel using a technique called "spread spectrum" [1]. Spread spectrum originated as a means of secure communication for the military, spreading a signal out in the frequency domain to give it a very low peak power. To observers, a spread spectrum signal looks similar to white noise, and thus has a Low Probability of Intercept (LPI). The noise-like properties of a spread spectrum signal are achieved by modulating the signal with a pseudo-random noise (PN) code, which is a pseudo-random binary sequence. PN codes are generated by shift registers configured with linear feedback taps, and this will be detailed in a later section. The PN code exhibits some interesting properties, which will be noted for the sake of completeness.

1) The Maximal Length Linear Shift Register (MLSR) sequence has a period of L chips, with $L = 2^N - 1$. (That is, an N bit shift register can generate a PN code of L bits, where the maximum value of L is $2^N - 1$.)

2) The statistical distribution of "1's" and "0's" is the same as in a random sequence with the exception that the total number of "1's" is always one larger than the total number of "0's", independent of the length of the code. (Note that this implies only codes of odd length are possible.)

3) The modula-2 sum of any m-sequence with a shifted version of itself produces another shifted version of the same sequence.

4) All possible *N*-bit words will appear in the sequence exactly once, except for the all-zeros combination. (The all-zeros combination should never occur because the shift register will become locked in this state.)

5) The autocorrelation of the PN sequence is given by $R_{xx}(\tau)$:

$$R_{xx}(\tau) = -\frac{1}{L}\sum_{\ell=-\infty}^{\ell=\infty}\left(1+\frac{1}{L}\right)\Delta\left(\frac{\tau-\ell L T_c}{T_c}\right)$$

T_c is the clock rate (also known as the chip rate)

$$\Delta\left(\frac{t}{\tau}\right) = \begin{Bmatrix} 1-\frac{|t|}{\tau}, & |t| \le T_c \\ 0, & |t| > T_c \end{Bmatrix}$$

The spectrum of the PN code follows a $\text{sinc}^2(f)$ envelope, with the nulls occurring at integral multiples of the clock frequency [1]. The spectrum is made up of discrete peaks between envelope nulls. The number of peaks between nulls is given by $L - 1$ where L is the length of the PN sequence. Figures 1 and 2 show the spectrum and autocorrelation of the PN code.

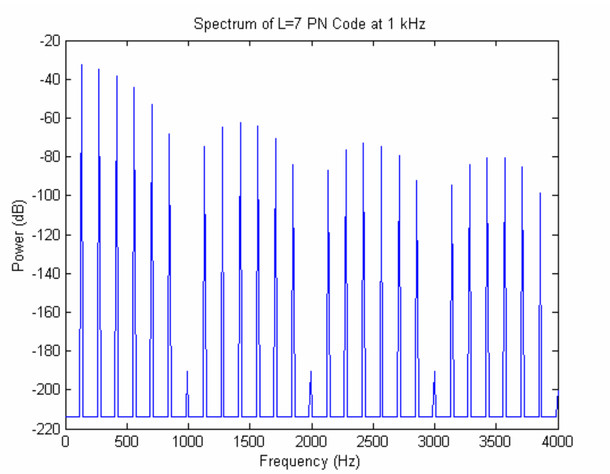

Figure 1. Spectrum of an L = 7, f_c = 1 kHz PN sequence.

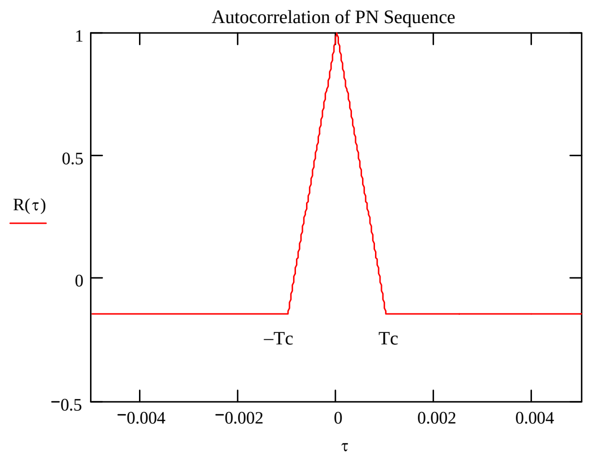

Figure 2. Normalized autocorrelation of an $L = 7$, $f_c = 1$ kHz PN sequence.

Channel Sounder Fundamentals

The sliding correlator channel sounder clocks the PN codes at very high frequencies to fill a wireless channel with noise-like content [1]. The PN codes are modulated onto the carrier frequency and transmitted across the channel. Some distance away, a receiver demodulates the signal and performs the "sliding correlation". The received PN code, which is clocked by the transmitter at some frequency f_T, is mixed with an identical PN code clocked by the receiver at some slightly slower frequency f_R. Because the received PN code is clocked at a higher frequency than the PN code generated at the receiver, the received code slides past the slower receiver-generated code in time.

When the faster code slides past the slower code such that they are momentarily perfectly aligned, the crosscorrelation will be very large. At points of poor alignment, the crosscorrelation will be $-1/N$. This was indicated by the autocorrelation of the PN code. A series of alignments due to the multipath environment will generate a series of triangular peaks in the crosscorrelation. This effectively is the impulse response of the channel. By taking the Fourier Transform of this impulse response, the frequency response of the channel may be found.

PN Generator Overview

The PN generator is based around a shift register configured with linear feedback [2]. Certain registers on the shift register are selected as feedback taps. The bits at these taps are XOR-ed together (modulo-2 addition). The output of the modulo-2 addition is sent to the input of the shift register. This configuration is the Fibonacci architecture for a linear feedback shift register [3]. Another configuration, called the Galios architecture, is functionally equivalent but performs operations in parallel, rather than in series. The Galois PN generator is generally faster than the Fibonacci PN generator due to its parallel architecture. Figure 3 compares a Galois and Fibonacci implementation.

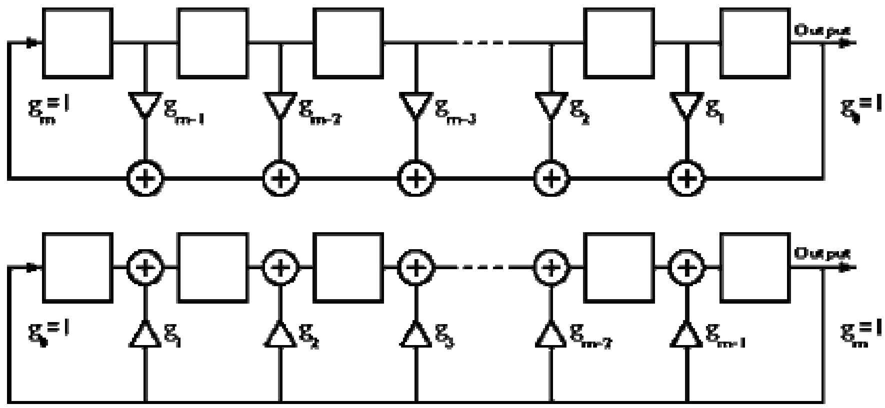

Figure 3. Fibonacci (top) and Galois (bottom) linear feedback. [3]

Note that the Galois architecture requires parallel I/O for the shift register, while the Fibonacci implementation requires shift in / parallel out for the shift register. One other item of interest is that the "output" of a PN generator can be taken from basically anywhere in the circuit. All registers and XOR gates will output the same PN code, though they will all be out of phase from each other.

Selecting the correct registers to use as feedback taps is an important in designing the PN generator. Only certain feedback configurations give rise to what is called a maximal-length sequence, or m-sequence for short. M-sequences are the longest possible codes that a PN generator can produce before repeating. For an N-bit shift register, its m-sequence length is given by L in the Equation 1.

$$L = 2^N - 1 \qquad \text{(Eq. 1)}$$

All other feedback tap configurations will result in a sequence length that is less than L. Therefore, to make the best use of the shift register, it is important to choose feedback taps that give rise to an m-sequence. This is not a difficult task, as data tables abound that provide the proper feedback taps for a given shift register size [3]

In order to begin outputting the PN code, the shift register must be initialized to some non-zero value. Typically, the shift register is filled with 1's via the use of an OR gate between the final XOR gate and the shift register [2]. One of the OR gate inputs is the XOR gate output. The other OR gate input is used to switch the PN generator between initialization mode and normal operation. A diagram of this setup is shown in Figure 4.

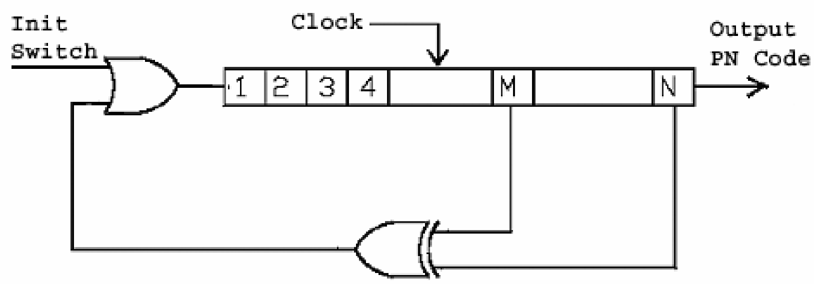

Figure 4. Diagram of a two-tap linear feedback shift register. [1]

PN Generator Design Choices

There are two principle variables that had to be determined before design could begin. These were the clock frequency f_c of the PN generator and the size of the shift register *N*. The choice for these variables has a tremendous effect on the performance of the sliding correlator channel sounder, as is indicated in Table 1.

Table 1. System Parameter Dependencies [1]

System Parameter	PN Generator Property	Dependence
Time Domain Resolution of Multipath Signals	Clock frequency, f_c	$T_{\mathrm{Res}} = \frac{1}{f_c}(\mathrm{sec})$
Dynamic Range	Code length, L_{PN}	$D_R = 20\log(L_{PN})(\mathrm{dB})$
Maximum Resolvable Mulitpath Delay Time	Code length and Clock frequency	$T_{Max} = \frac{L_{PN}}{f_c}(\mathrm{sec})$
Process Gain	Slide factor, *k*, and Clock frequency	$G_P = 10\log\left(\frac{0.88 f_c}{k}\right)(\mathrm{dB})$
Maximum Doppler Shift Resolution	Slide factor, Clock frequency, and Code length	$f_{D_{MAX}} = \frac{f_c}{2kL_{PN}}(\mathrm{Hz})$
RF Bandwidth	Clock frequency	$BW = 2f_c\,(\mathrm{Hz})$

Many of the system parameters are beyond the current scope of this project. However, the dynamic range and RF bandwidth can be addressed here. The dynamic range is related to the peak value of the autocorrelation and thereby, the length of the sequence. When a longer sequence is autocorrelated, its peak value will be greater than that of a shorter sequence. Greater dynamic range allows for more accurate analysis of the channel and a greater SNR of the output impulse response.

The RF bandwidth refers is directly related to the clock frequency. The sliding correlator channel needs to produce a wide band signal to determine the frequency response of the wireless channel. Therefore, the PN generator must be clocked at a very high frequency. Recall that the PN code spectra resembles $\mathrm{sinc}^2(f)$ with nulls occurring at the integer multiples of the clock frequency. The RF bandwidth is a measure of the size of the main hump in the envelope, as much of the spectral power is in the range of $\pm f_c$. Thus, the faster the PN generator is clocked, the larger the bandwidth of the signal. It is desired to sound the channel with signals that have a bandwidth of 1 GHz or more, so the PN generator must be able to operate at extremely high frequencies. Clocking at these frequencies requires a special family of integrated circuits called Emitter-Coupled Logic (ECL).

Circuit Design

The PN generator was designed using ECL devices. There were five main components of the PN generator design: the 3 Volt sink, the shift register clock, the bias voltage, the initialization circuit, and the linear feedback shift register (LFSR).

3 Volt Current Sink

It was mentioned earlier that the Thevenin equivalent termination scheme seemed to be the preferred method, but for some reason this scheme had implementation problems that resulted in the destruction of several devices. Therefore, after much difficulty, the 3 V supply scheme was chosen. However, this presented another problem, because with the exception of the ground terminal, most power supply terminals do not sink current very well. Tests showed that voltage would build up on the 3 V supply terminal if one tried to sink too much current into the terminal, leading to an actual voltage greater than 3 V. To resolve this problem, it was found that by connecting a resistor from the 3 V terminal to ground allowed the power supply to divert any current going into the 3 V terminal to ground. The value of this resistor should be chosen so that the current from 3 V to ground is always equal to or greater than the current being sunk into 3 V.

Of course, this is not a very elegant scheme, and still requires an extra supply. To avoid this, a 3 Volt sink was designed using an op-amp and a PNP transistor. The op-amp was configured as a voltage-follower to produce 2.3 V at the base of the transistor. The transistor was operated in the forward-active region to ensure approximately a 0.7 V drop from the emitter to the base. By grounding the collector terminal, the voltage drop from emitter to collector was approximately 3.0 V. Looking into the transistor from the emitter, the transistor has a very low impedance and thereby acts as an excellent current sink.

To generate the 2.3 V at the op-amp output, a simple voltage divider was used. The actual voltage generated by the voltage divider was 2.27 V. However this was acceptable, as it was better for the 3 Volt sink to be a slightly less than 3 V rather than greater. This would mean that the ECL outputs would source slightly more current than nominal, but also makes it less likely for V_{TT} to ever be greater than V_{OL}. Another caveat of the current sink was that, in order for the transistor to operate in the forward-active region and thereby have $V_{EB} = 0.7$ V, the transistor had to sink at least 20 mA of current. When an ECL output is low, it sources 0.25/50 = 5 mA, and when high, 1.05/50 = 21 mA. Obviously it is not difficult to meet this requirement simply by connecting devices to the current sink. However, to ensure that the 3 Volt sink is always approximately 3 Volts, a 100 ohm resistor to V_{CC} was used to sink 20 mA of current into the transistor's emitter. The final 3V current sink circuit is shown in Figure 5.