Point-Spread Function Assessment of SG-DBR Based Swept Source for OCT Imaging

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Imaging

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Abstract

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Swept Source Optical Coherence Tomography (SS-OCT) is a medical imaging technique that requires high repetition rate, widely-tunable coherent laser sources. Sampled grating distributed Bragg reflector (SG-DBR) lasers are proven in telecom applications and are expected to fulfill the requirements for SS-OCT at a significantly lower cost than alternative solutions.

Constructed entirely on a semiconductor substrate, SG-DBR lasers require four synchronized waveforms to modulate the output wavelength and intensity. Because of this unique tuning mechanism, there are a number of systematic and noise sources that can affect the quality of the OCT point-spread function (PSF). Based on these noise sources, software is developed to simulate the waveforms in an SS-OCT system and determine the factors that limit width of the PSF central peak and the broadband skirts. Design curves are then created to specify the requirements to obtain a given performance.

Next, experimental tests are performed on a JDS Uniphase C-band SG-DBR laser to assess its performance limitations. Finally, by comparing theory and experiment, recommendations are made on acceptable systematic and random noise errors induced in the PSF.

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Chapter 1: Introduction

Swept Source Optical Coherence Tomography (SS-OCT) is a medical imaging technique that requires high repetition rate, widely-tunable coherent laser sources. Sampled grating distributed Bragg reflector (SG-DBR) lasers have been proven in telecom applications and are expected to fulfill the requirements for SS-OCT at a significantly lower cost than current solutions.

Previous work on SG-DBR lasers has been accomplished by graduate students working under Dr. Dennis Derickson. Shane O'Connor [16] was the first to investigate the SG-DBR laser for OCT by demonstrating a linear wavelength ramp. Brandon George [12] developed a linear frequency ramp and showed that it was possible to concatenate many tuning paths for larger coverage. Christopher Chiccone [5] developed an automated program to concatenate tuning paths in a linear frequency ramp with increased precision for an arbitrary SG-DBR laser. Dr. Jason Ensher at Insight Photonic Solutions in Lafayette Colorado is also a major contributor to the research on SG-DBR usage for OCT.

The main limitation of the SG-DBR approach is believed to be the way the tuning paths are concatenated. What is still unknown, and is the subject of this investigation is how exactly errors in concatenating tuning paths affect the resolution limits of an OCT image.

Chapter 2 begins with an introduction to the theory and design of the SG-DBR laser. Next, it shows the physical implementation of the laser package and explains the development of linear frequency ramps. The later sections are devoted to theory that will help the reader fully understand chapter 3.

Chapter 3 is an introduction to the theory of OCT. It begins with a comparison to other medical imaging procedures and explains the differences between time-domain and frequency-domain OCT. Next, it shows how the swept-source OCT system is implemented with the SG-DBR laser and discusses the non-idealities that lead to performance limitations.

Chapter 4 is a theoretical investigation into the performance limitations of an SG-DBR laser in OCT applications. Important parameters believed to be limiting the point-spread function of an OCT image are identified and a MATLAB program is developed to simulate a one-dimensional OCT trace. Based on the size and locations of the anomalies, performance curves are developed to aid in identifying the most important ways to improve OCT resolution.

Chapter 5 is an experimental investigation of the OCT system using a JDS Uniphase C-band SG-DBR laser. An unbalanced photodetector amplifier circuit is constructed and LabVIEW software written to record the optical signals. MATLAB software is used to process the signals and assess the resolution. Finally, based on the experimental evidence, further assumptions about the noise inherent to the sweeping laser are made and feed back into software to obtain an even more accurate prediction of the maximum resolution of the OCT system.

Chapter 6 summarizes the theoretical and experimental findings, suggests techniques to improve the quality of the PSF, and presents the conclusions.

Chapter 2: SG-DBR Laser Background and Technical Introduction

2.1 SG-DBR laser background

In much the same way a saxophone is a resonant cavity for sound waves, a laser is a resonant cavity for photons. Although many people today don't realize it, LASER stands for *Light Amplification by Stimulated Emission of Radiation*. *Stimulated emission* is the process by which an incident photon causes an electron in an excited state to transition to a lower state and emit a photon with the same frequency as the incident photon. Figure 2.1.1 shows the process of stimulated emission.

Figure 2.1.1: Stimulated emission in a two-state quantum system.

If many of the electrons are in the excited state before the incident photon causes the transition, many electrons will transition to the lower state at the same time. Most importantly, these photons will be *coherent* (in phase) with each other. In order for the photons to not be subsequently reabsorbed, the number of electrons in the excited state should greatly outnumber those in the lower state; this is known as a *population inversion*. The collection of quantum states that allow the population inversion to take place is known as the *gain medium*. The energy used to produce the population inversion is known as *pumping energy*.

Regardless of the technology used, all lasers utilize the same basic elements shown in Figure 2.1.2.

Figure 2.1.2: The basic elements of a laser [9].

Photons travel the length of the laser, are reflected by mirrors and return to their starting point in-phase so that they reinforce other photons at the same wavelength. This resonance condition can be exploited to produce lasers with very high Q (narrow frequency and wavelength) outputs.

Sampled grating distributed Bragg reflector (SG-DBR) lasers are solid state lasers developed for telecommunications applications that offer high-speed tuning through roughly 50 nm in the near-infrared (~ 1550 nm) band. The combination of fast sweep rate, high linearity and high coherence are important factors that benefit resolution in OCT. All elements of an SG-DBR laser are constructed on a semiconductor substrate. A cross section of the SG-DBR laser structure is shown in figure 2.1.3.

Figure 2.1.3: side-view of SG-DBR laser structure [7].

The gain section provides the gain medium for stimulated emission of the laser. The front mirror (FM) input controls the first reflecting surface and the back mirror (BM) the second reflecting surface. The phase adjust (PH, PM or Φ) is used to provide small changes in the effective length of the laser cavity and basically works as a fine trim. The semiconductor optical amplifier (SOA or amp) can be used to boost the output power of the laser or to act as a shutter providing high attenuation [7].

The main property separating an SG-DBR from other lasers is that it works by having mirrors that are also periodic structures. These periodic structures, called Bragg gratings, have reflectivity versus wavelength profiles that are also periodic, in much the same way as a Fabry-Perot interferometer. Figure 2.1.4 illustrates the periodic structure of the Bragg gratings.

Figure 2.1.4: The periodic Bragg grating in (b) leads to the periodic reflectivity versus wavelength profile in (a). Source: [7].

Adjusting the current to the mirror inputs can control the wavelength of the reflectivity peaks. However, the wavelength spacing between adjacent reflectivity peaks is different for front and back mirror. Therefore, by simultaneously adjusting the FM and BM currents, it is possible to align the front and rear mirrors so that only a single wavelength is reinforced, as shown in figure 2.1.5.

Figure 2.1.5: in (a) the front and rear mirrors are aligned to reinforce only a single wavelength. (b) Shows the corresponding laser output measured by an optical spectrum analyzer [7].

Unfortunately, adjusting only one mirror does not allow for continuous tuning of the laser. Tuning one mirror at a time results in discrete 5 nm wavelength transitions (called *mode hops*) because the laser is only able to maintain an output that satisfies one of its longitudinal cavity modes. Tuning both mirrors simultaneously results in 0.3 nm mode hops, and incorporating the phase mirror allows continuous tuning without the discrete wavelength transitions.

To accomplish continuous tuning of the laser, Brandon George [12] developed tuning maps of the laser that show the wavelength of the laser as a function of FM and BM currents. Figure 2.1.6 shows a sample tuning map of a C-band SG-DBR laser.

Figure 2.1.6: different tuning paths as a function of front and back mirror currents. Red represents longer wavelengths and blue shorter wavelengths [12].

Paths of consecutive wavelength are spread out over 2D space and are further divided into smaller sections that denote the 0.3 nm mode hops. Continuous tuning of the laser requires that FM and BM currents are rapidly changed at each tuning path transition and phase current changed many times per path. Although not shown on the figure, the SOA current must also be modulated to provide constant laser output power.

Linear frequency ramps for the laser were developed by Shane O'Connor [16]. Waveform files were created and loaded into four synchronized Agilent 33220A arbitrary waveform generators (*ARB's*). The *ARB's* generate a voltage signal but the inputs to the laser require a current source, so each input to the laser uses a custom-made resistor network. Additionally, this networks incorporate protective Zener diodes to prevent reverse biasing and to serve as an over-voltage (>5 V) protection. Most importantly, each input to the laser is constructed with semi-rigid coax to maintain 50Ω impedance right up to the laser package. Maintaining the proper system impedance is vital to obtaining high-speed performance. These laser package and all inputs are shown in figure 2.1.7.

Figure 2.1.7: Photo of C band SG-DBR laser input networks [5].

A Faraday Rotator optical isolator prevents reflections from the laser output traveling back into the laser cavity, which may cause undesirable effects in the laser output. From this point, when referring to the SG-DBR source, it can be considered to be the entire group of devices shown in figure 2.1.8.

Figure 2.1.8: SG-DBR laser package and devices required to produce a frequency sweep.

When the laser is set up to do a frequency sweep, it is generally assumed that it will be continually repeating itself. The *repetition rate* of the laser (expressed in Hz) is the number of times per second that a complete sweep is performed. Practical repetition rates in our laser are in the 1-10 kHz range.

The remaining portion of this chapter is devoted to an explanation of the physics and mathematics that are necessary to fully understand chapter 3.

2.2 Interferometry Background

Interferometry is the science of superimposing electromagnetic waves to extract some information about them. It relies on the basic principles of constructive and destructive interference studied in introductory physics courses. A practical, free space interferometry experiment can be constructed with only basic optics laboratory components, such as a laser, beam splitter, and mirrors. Figure 2.2.1 shows such an arrangement known as a *Mach-Zehnder interferometer* (MZI).

Figure 2.2.1: A Mach-Zehnder interferometer experiment [8].

Coherent light source is equally split by the beam splitter and each beam travels along a different optical distance. The beams are recombined by a second beam splitter and the resulting interference pattern observed with a detector. To observe interference fringes, either one of the path lengths can be varied, a transparent sample can be placed in the path of one of the beams, or the frequency of the source can be varied. The *Free Spectral Range* (FSR) is defined as the difference in optical frequency of the source between two

successive interference maxima or minima when the path length difference is held constant.

Similar interferometers exist for fiber optics applications. In this case, fused-fiber couplers are used to divide laser power between the two arms. A fused-fiber Mach-Zehnder interferometer is illustrated in figure 2.2.2.

Figure 2.2.2: Fused fiber Mach-Zehnder interferometer experiment [14].

Coherent light is equally split between at the input coupler, travels along the two arms and is recombined at the output coupler. The two outputs (+ and -) are 180° out of phase; that is, when + output reaches a peak, the – output reaches a minimum. The second input is not connected.

In general, the *temporal coherence* of the source is an important consideration because it limits the ability for the recombined beams to interfere. In simple terms, temporal coherence is a measure of how consistent the phase of the light wave is. After traveling a distance equal to the *coherence length*, the relative phase of an electromagnetic wave begins to randomize and it will no longer produce an interference pattern.

2.3 Balanced Photodetector Operation

A balanced or differential detector works by subtracting two input signals and outputting the difference. This technique eliminates any constant offset from the inputs (such as non-zero ground potential) but can also have the effect of removing intensity noise that is common to both signals. Figure 2.3.1 shows how a balanced detector eliminates noise and offset.

Figure 2.3.1: How a balanced detector can eliminate noise that is common to both inputs.

The figure of merit for a balanced detector's ability to remove common-mode noise is the Common-mode Rejection Ratio (CMRR). The ability to match components, both inside and outside the detector can be a limiting factor in obtaining a high CMRR.

Balanced detectors can be either electrical or optical. A functional block diagram of a balanced optical detector that uses photodiodes is shown in figure 2.3.2

Figure 2.3.2: Functional diagram of a balanced photodetector [17].

In this arrangement both photodiodes are reverse biased. A transimpedance amplifier converts the current I_3 to a proportional voltage. Notice that if the current through both photodiodes is equal, then $I_3 = 0$ and there will be no output.

2.4 Fast Fourier Transform

A Fourier transform is an integral transformation that decomposes a signal into its constituent frequencies (or frequency spectrum). It is used extensively in electrical engineering, especially in signal processing where it is invaluable.

In order to quickly calculate Fourier transforms on computers, the *Fast Fourier Transform* (FFT) algorithm was developed in 1965 [6]. Given a signal that has been sampled at regular intervals, the FFT reconstructs an approximation to the signal's (continuous) Fourier transform. The FFT algorithm exploits symmetries in the mathematics of the calculation to exponentially reduce the computational complexity required. For a data-set of 10^6 points, the FFT is 50,000 times faster than the discrete Fourier transform [10]. Figure 2.4.1 shows the basics of the FFT process.

Figure 2.4.1 shows how a signal sampled at regular intervals in the time domain (left) is converted to frequency domain (right).

The horizontal scaling in the frequency domain is limited to $F_s/2$ because of the *Nyquist-Shannon Sampling Theorem*.

2.5 FFT Windowing

Because any (physical) signal must be finite in time, there exists noise due to the truncation of the signal. In other words, the Fourier transform of a finite periodic signal is not a delta function, but rather a sinc function. Not only that, but if the signal is not sampled at an integer number for each cycle of the waveform, noise also results from a periodic extension of the waveform. This is commonly referred to as *spectral leakage* because it is as if energy from the desired signal has leaked out to neighboring frequencies. Figure 2.5.1 shows an example of spectral leakage.

Figure 2.5.1: Sampling at an equal number of integers per sample results in a clearly defined peak (left) while sampling at a non-integer number of samples per cycle causes the signal peak to spread.

Fortunately, there is some help to be had in dealing with spectral leakage. Before computing an FFT, the signal can be scaled by a *window function* (or simply *window*) such that any effects from periodic extension are minimized. A common window is shown in figure 2.5.2. Figure 2.5.3 shows how a window function can improve resolution on an FFT.

Figure 2.5.2: Hann window in time domain (left) and frequency domain (right).

Figure 2.5.3: FFT of signal with Hanning window. Compared to figure 2.5.1 the window function offers a large increase in resolution.

2.6 Analog to Digital Conversion

An analog to digital converter (ADC) is a device that converts a continuous voltage signal into one that is both sampled in time and discretely quantized in amplitude.

Figure 2.6.1 shows the basic function of an ADC.

Figure 2.6.1: Sampling process of an ideal 4 bit ADC [11].

The continuous input signal is sampled every T_s and converted to an output word that is defined by n , the number of bits in the ADC. For a word of length n bits, there exist 2^n possible states, so the input of the ADC is quantized to one of 2^n values. The *full scale range* is the range of input voltages that an ADC can digitize without going over or under range.

The *least significant bit* (LSB) is best described as the quantization “bin width” and inputs to the ADC will be rounded to within $\frac{1}{2}$ LSB of their true value, provided they are within the full scale range. The noise created by this rounding process is known as *quantization noise* and can be directly calculated. For an ADC, the maximum possible signal to noise ratio of a sinusoidal input signal is determined by equation 2.6.1.

$$\text{Equation 2.6.1} \quad SNR_{\max} = 6.02n + 1.76 \text{ dB}$$

So for each extra bit in the ideal ADC, there is a 6.02 dB (four-fold) increase in $SNR_{\max}$.

If the input signal does not utilize the full scale range of the ADC, the *effective number of bits* of the ADC is reduced. For example, if a signal of $2.5 V_{p-p}$ is input to an 8-bit ADC

with a full scale range of 5 V, it only uses half of the available states of the ADC; that is, the ADC functions like a 7-bit ADC with a full scale range of 2.5 V.

2.7 Noise

Noise is any undesired signal existing in electronic components or an electrical signal. The source can be deterministic or random but all electrical and electronic devices are in some way a source of noise. For a frequency source such as a laser or oscillator, the results of these sources are amplitude and phase fluctuations in the desired signal. These fluctuations can be characterized by measurement as *amplitude modulation* (AM) and *phase modulation* (PM), the latter being directly related to the frequency stability of the source. The general equation of noise for a sinusoidal signal is

$$\text{Equation 2.7.1} \quad V(t) = [V_0 + \varepsilon(t)] \cos(2\pi f_0 t + \phi(t))$$

Where V_0 and f_0 are the nominal amplitude and frequency and $\varepsilon(t)$ and $\phi(t)$ are the instantaneous fluctuations in the signal [3]. The instantaneous fluctuations can be modeled as either deterministic (behaving predictably, as in a sinusoid) or random, for example a Gaussian random process.

AM is the mathematically simpler of the two and so will be discussed first. For a sinusoidal amplitude variation, $\varepsilon(t)$ takes the following form

$$\text{Equation 2.7.2} \quad \varepsilon(t) = \Delta V \cos(2\pi f_m t)$$

where f_m is the modulation frequency and ΔV the maximum amplitude deviation from the carrier. The maximum amplitude deviation relative to the carrier is known as the *amplitude modulation index* and is designated m in equation 2.7.3.

$$\text{Equation 2.7.3} \quad m = \frac{\Delta V}{V_0}$$

Using basic trigonometric identities one can show that the spectrum of a sinusoidal AM signal has the features in figure 2.7.1 (notice relative amplitudes of signals).

Figure 2.7.1: FFT of sinusoidal amplitude modulation showing carrier signal and sidebands.

In addition to the unaltered carrier signal, the sinusoidal modulating signal results in the appearance of two *sidebands* at ω_m above and below the carrier frequency, the height of which depends on the modulation index. Contrary to what one might guess, no signal appears at the modulation frequency. The height of the sinusoidal AM sidebands is given by equation 2.7.4.

$$\text{Equation 2.7.4} \quad V_{\text{sideband}} = \frac{mV_0}{2}$$

If the amplitude modulating signal is random it can be represented in the form of equation 2.7.5:

$$\text{Equation 2.7.5} \quad \varepsilon(t) = \sigma_{am} \text{rand}(0,1)$$

where $\text{rand}(0,1)$ is a zero-mean Gaussian random number with standard deviation 1 and σ_{am} the standard deviation of $\varepsilon(t)$. A Gaussian (or normal) distribution is usually chosen as the random process because a large number of independent random variables can

generally be approximated by a normal distribution. This property is known as the *Central Limit Theorem*.

For a signal that is sinusoidally modulated in phase, $\varphi(t)$ takes the form of equation 2.7.6.

$$\text{Equation 2.7.6} \quad \varepsilon(t) = \Delta\phi \cos(2\pi f_m t)$$

Where $\Delta\phi$ is the peak phase deviation in the signal, also called the *phase modulation index*. Because the modulation appears within the argument of the cosine function, the mathematics for describing this form of modulation are more complicated than AM. The FFT of sinusoidal PM is shown in figure 2.7.2.

Figure 2.7.2: FFT of Sinusoidal PM of Signal with modulation index $\Delta\phi \ll 1$.

With sinusoidal PM, there are actually an infinite number of sidebands, each offset from the carrier by integer values of ω_m . The height of the sidebands can be found exactly using Bessel functions of the first kind [18]. The number of sidebands that are resolvable in the FFT depends on the magnitude of the modulation index. Also, in contrast to AM, the amplitude of the central peak is not unity and will decrease with an increase in

modulation index. However, for a modulation index $\ll 1$, as is the case in most physical systems with small PM noise levels, only the first sidebands are resolvable and the central peak has amplitude of roughly V_0 (as is shown in figure 2.7.2). This is most often the case when analyzing oscillator or laser phase noise. Equation 2.7.7 gives the height of the sinusoidal PM sidebands for the case that the modulation index is much less than one.

$$\text{Equation 2.7.7} \quad V_{sideband} = \frac{\Delta\phi V_0}{2}$$

Mathematically, PM is very similar to *frequency modulation* (FM). The instantaneous frequency deviation of a signal is related to the phase modulation by the following equation

$$\text{Equation 2.7.8} \quad f(t) = \frac{1}{2\pi} \frac{d\phi(t)}{dt}$$

Where $f(t)$ is the instantaneous frequency deviation in Hz. The *frequency modulation index* is the ratio of maximum frequency deviation to the modulation frequency and is mathematically equal to the phase modulation index, as shown in the following equation

$$\text{Equation 2.7.9} \quad \Delta\phi = \frac{\Delta f}{f_m}$$

Where Δf is the peak frequency deviation in the signal.

If the PM signal is random it can (like AM) be represented in the following form

$$\text{Equation 2.7.10} \quad \phi(t) = \sigma_{pm} \text{rand}(0,1)$$

Where $\text{rand}(0,1)$ is a zero-mean Gaussian random number with standard deviation 1 and σ_{pm} the standard deviation of $\phi(t)$. The random variable can be thought of as containing a continuum of frequencies, each contributing to the modulation of the carrier. And unlike

sinusoidal modulation, the random variable also produces noise that is *white*, or distributed equally across all frequencies. Figure 2.7.3 shows an FFT of random PM noise.

Figure 2.7.3: FFT of random PM of signal showing broadband (white) noise and phase noise skirt close to the carrier signal.

In a physical system, there are many different sources of noise. Even when an optical detector is exposed to no light, there may be noise due to the *dark current* that flows through the photodiode. *Shot noise* is present due to the position-momentum uncertainty of the electrons in a current. Resistive elements, including those that are internal to the photodiodes, have *Johnson-Nyquist* (thermal) noise due to the thermal excitations of the charge carriers.

Flicker noise occurs at the junction between two materials, whether they are metal or semiconductor. The fundamental causes of flicker noise are not well described but it occurs due to resistance fluctuations between the junction. It occurs mainly in active devices but can also be found in passive devices such as carbon film resistors. Flicker noise is also known as *1/f noise* because the power spectral density is proportional to $1/f$.

Because of the frequency dependence, it is most important at DC and is overshadowed by other noise processes (thermal noise or white noise) at higher frequencies.

Chapter 3: Optical Coherence Tomography

3.1 OCT in General

Optical Coherence Tomography (OCT) is a medical imaging technique able to obtain high resolution three dimensional images of *in vivo* human tissue. OCT typically makes use of near infrared light to obtain images at a depth of roughly a few millimeters.

Unlike Computed Tomography (CT) scanning machines or traditional X-rays, both of which utilize ionizing radiation, OCT uses non-ionizing radiation which is free from exposure limits. Figure 3.1.1 is a comparison of the common (non-invasive) medical imaging procedures.

Figure 3.1.1: Comparison of medical imaging techniques applications and performance [15].

Like ultrasound, OCT determines the back reflection of a signal versus its depth. However, unlike ultrasound, it is not possible to directly measure the propagation time of

a beam of infrared light. Instead, OCT relies on the superposition (interference) of two electromagnetic waves to extract the depth information. By analyzing the resulting interference pattern, a profile of reflectivity versus distance can be determined and an image of the tissue sample can be reconstructed. Figure 3.1.2 is an example of what is available from a modern commercial OCT system for roughly 60k USD.

Figure 3.1.2: Three dimensional image of arterial wall in pig tissue using a Thorlabs swept-source OCT system [19].

3.2 Time Domain OCT

Early OCT systems were known as Time Domain OCT (TD-OCT) and used broadband, weakly coherent light sources such as LED's. Since the entire output power of the source is spread over a large frequency range, the resulting signal to noise ratio is usually poor and imaging depth is also be limited by the low coherence length of the source.

To vary the imaging depth, TD-OCT uses a movable mirror. Because of the requirements for mechanical precision and speed of the mirror (to achieve a practical

image refresh rate) such setups are less common, but are still used in some areas such as ophthalmology.

The advantage of TD-OCT is that the computer needs only to record the measured data—no post-processing is necessary to extract the reflectivity versus distance profile.

Figure 3.2.1 shows a typical TD-OCT experiment.

Figure 3.2.1: Time Domain OCT experiment [1].

Light from the source is split into two paths, one with a movable mirror and the other containing the sample (DUT). Light reflected from the reference and the sample is recombined at the splitter, detected and stored in a computer as amplitude (reflectivity) versus distance as the position of the mirror is changed.

A major disadvantage of such a configuration is the requirement for high-speed mechanical components to accomplish the depth scanning at the sort of refresh rate typically desired.

3.3 Frequency Domain OCT

To overcome some of the limitations of TD-OCT, frequency domain OCT (FD-OCT) was developed. Rather than encoding reflectivity versus time, reflectivity versus frequency is encoded instead.

Swept-source OCT (SS-OCT) uses a highly-coherent, narrow linewidth source that is linearly swept in frequency. This action replaces the movable mirror as in TD-OCT. The resulting time-domain signal is recorded by a photodetector and Fourier transformed to yield the reflectivity versus distance profile. Figure 3.3.1 shows how an SS-OCT measurement can be performed.

Figure 3.3.1: A swept-source OCT experiment. [1]

Light from the source is split into two paths, one containing a fixed mirror and the other the sample (DUT). Light reflected from the reference and the sample is recombined at the splitter, detected and stored in a computer. The recorded data is Fourier transformed to yield reflectivity versus position. Because of the narrow linewidth of the source, SS-OCT offers much higher resolution than TD-OCT. In modern systems, resolutions in

excess of 50 dB are possible. Figure 3.3.2 gives another view as to how a one-dimensional SS-OCT measurement is performed.

Figure 3.3.2: Procedure of a one-dimensional swept-source OCT measurement [12].

Light reflected from the device under test (DUT) is combined with the reference light to produce a beat signal which is then fast Fourier transformed (FFT) to extract a reflectivity versus distance profile. By repeating the process on another two axes and combining the data it is possible to obtain three-dimensional images.

3.4 Insight Photonic Solutions OCT System

The OCT system developed by Cal Poly and Insight Photonic Solutions is a form of SS-OCT. The swept-source, an SG-DBR laser, is coupled into single-mode (SM) optical fiber and is split 50/50 into two arms, each with a fused-fiber MZI. One arm creates a clock signal that triggers the sampling of the sample signal in the second arm. Software written in LabVIEW controls all devices and processes the signals. Figure 3.4.1 shows a diagram of the complete system.

Figure 3.4.1: Complete SS-OCT System developed by Insight Photonic Solutions performing a 1-D measurement.

The 7.1 GHz MZI serves as the reference arm and the 39 GHz MZI is used to measure the sample. One length of the 39 GHz MZI is connected to a three-port device known as an *optical circulator*. Light enters the circulator from the left, exits toward the sample, is reflected and then exits the right side of the circulator where it recombines to form the

interference pattern. The computer software then samples the interference pattern at equal intervals of laser frequency, performs the FFT and displays the data.

For our purposes, we are not so concerned with performing actual tissue measurements with the OCT system; rather, we are more concerned with the performance of the SG-DBR laser as it relates to OCT. A useful quantity in characterizing the laser's performance is known as the *point spread function* (PSF), an image of a single perfect reflection. In other words, the PSF is the system's *impulse response*. Notice that if the sample in figure 3.4.1 is replaced with a mirror of perfect reflectivity, the length of the MZI with the circulator acts like an ideal delay line. So, instead of dealing with the complexity of the circulator and mirror, they can simply be replaced with an interferometer of fixed FSR when assessing the PSF. The OCT system then takes the form shown in figure 3.4.2.

Figure 3.4.2 The SG-DBR SS-OCT experiment used to measure the PSF.

3.5 Creating the Clock Signal

In general, amplitude information for an OCT measurement needs to be encoded in some way so that it can be correlated to a depth measurement. In the OCT system of figure 3.4.2 it is necessary to know the spatial frequency of the laser as a function of time. Actually, not its exact value, just how much it changes so that it can be sampled at regular intervals for the FFT algorithm.

To do this a 7.1 GHz FSR MZI is used to produce a periodic interference pattern. This pattern repeats itself whenever the laser tuning passes through one FSR. If the laser tuning is linear (as it is designed to be), then the output of the interferometer will be a perfect sinusoidal signal; however, even if the laser tuning is not linear, the clock signal still serves as a reference of laser frequency. Because this signal is used to sample at equal intervals in frequency (equivalently in k -space) it is commonly referred to as the “kclock” (read as: KAY-clock).

In much the same manner as the clock, the DUT is represented by a 39 GHz FSR MZI. Because of the larger FSR, the DUT signal repeats itself at a proportionally lower rate than the kclock. Figure 3.5.1 shows the optical and electronic signals as they appear in the interferometry network when measuring the PSF.

Figure 3.5.1: Ideal input and output signals for the interferometry network.

The preceding figure doesn't actually show the complete story. Because the balanced detector requires two inputs (see section 2.3), the DUT arm actually uses both outputs of the interferometer (see section 2.2). The two signals are then differentially combined to yield the aforementioned voltage versus time profile. Figure 3.5.2 shows an expanded view of the DUT arm.

Figure 3.5.2: expanded view of DUT interferometry arm showing the two optical signals created by the MZI and the voltage signal at the output of the balanced detector.

The locations of the increasing zero-crossings of the clock signal are used to determine the locations of the sample points. From the sample locations a digital signal is created and saved as a text file. This electronic kclock or “ekclock” (see figure 3.4.2) is then loaded into an ARB and is used to trigger the PCI-6115 card to sample the DUT signal. Figure 3.5.3 shows how the kclock signal leads to sampling of the DUT.

Figure 3.5.3 Positive zero crossings of the kclock determine when to sample the DUT.

Once points on the DUT waveform have been sampled they are digitized, windowed and FFT'd to produce the PSF. Figure 3.5.4 shows the computer processing after the signal has been sampled and digitized.

Figure 3.5.4 Computer Processing of OCT signals in LabVIEW.

Let's take a moment to talk about the horizontal scaling of the PSF in figure 3.5.4. For a normal FFT, one would sample in time and scale the axis in frequency (see figure 2.4.1). Here the sampling is in k-space (reciprocal of wavelength), so what's needed is determine the k-space sampling interval, take the reciprocal and divide by two (because

of the Nyquist-Shannon Sampling Theorem) to get the max horizontal distance (the minimum is still zero). A tiny amount of algebra leads to the result of equation 3.5.1.

$$\text{Equation 3.5.1 } x_{\max} = \frac{c}{2 \Delta f_{\text{int}}}$$

Here $x_{\max}$ is the maximum horizontal distance, Δf_{int} the sampling interval (in this case one FSR) of the clock interferometer and c is the speed of light. Since the sampling interval is 7.1 GHz, d_x is equal to roughly 21.7 mm.

Possibly a subtle point, it is important to realize that the horizontal position of the central peak in the PSF is always the same, regardless of the laser repetition rate. This is because the two clock signals increase proportionally with the repetition rate. In other words, the sampling frequency is always about 5.5 (exactly $39/7.1$) times the DUT frequency.

3.6 Properties of the Actual PSF and Sampling Non-idealities

There are many imperfections in the laser source that cause the measured PSF to differ from the ideal case. Based on previous investigations, figure 3.6.1 shows the general structure of the PSF and the factors believed to be affecting its shape.

Figure 3.6.1: qualitative depiction of the measured PSF in our system and the noise sources believed to be responsible for its shape (not necessarily to scale). Source: [Jason Ensher at Insight Photonic Solutions].

The height of the central peak with respect to the noise floor is roughly 55 dB in the best results described by Insight Photonic Solutions. The FM and BM path transitions result in AM sidebands that are close to the carrier due to their occurrence once every path transition. Similarly, the phase mirror transitions also result in AM sidebands but at a greater offset from the carrier because they occur many times per path transition (see section 2.1 and figure 2.1.6). The noise floor that results from the laser being “parked” at a single output frequency could be due to a number of noise sources, but is suspected to be dominated by laser *Relative Intensity Noise* (RIN). With no input to the photodetectors, the noise floor is likely due to a combination of thermal and dark current noise.

The most important parameter believed to be affecting the shape of the PSF is how regularly the DUT is sampled. Ideally the DUT signal should be sampled roughly

5.5 times (exactly $39/7.1$, the ratio of the interferometer FSR's) per cycle without missing a point. However, when the laser transitions from path to path, there can be changes in the relative phase of the signals, amplitude modulation, or other undesirable effects. Moreover, for a small period of time between paths the laser output is essentially undefined and can consist of a large amount of noise. These path transitions are known as *stitch points* and are classified in one of two ways, as shown in figure 3.6.2.

Figure 3.6.2: classification of stitch errors as overlapping (left) or underlapping (right).

The stitch points are considered “overlapping” when laser frequency is repeated and “underlapping” when the laser skips over a range of frequencies. Figure 3.6.3 shows the corresponding interferometer output of an overlapping stitch point.

Figure 3.6.3: Overlapping stitch point where laser frequency is repeated, viewed through the 7.1 GHz interferometer.

Ideally, any phase changes in the clock and DUT signal should not affect the system resolution. This is because the two interferometers simultaneously provide a reference for laser frequency. However, phase changes can do damage if they cause a sample point to be double-counted (overlapping stitch point) or missed entirely (underlapping stitch point). It may even be possible to maintain a perfectly regular sampling interval, but a path transition that interrupts a sampling point can introduce a large amount of noise into the DUT signal, even if it was sampled at the “correct” location.

Deglitching, the process of removing repeated sampling points in the kclock, is vital to obtaining good PSF resolution. Figure 3.6.4 shows the PSF before and after deglitching.

Figure 3.6.4: PSF before (left) and after (right) sampling glitches are removed [5]

Software can automatically remove many glitches, but at this stage of development it is still necessary to manually deglitch the kclock.

In his research, Chiccone observed [5] that in total there are 35 stitch points, 25 are overlapping stitch points where laser frequency is repeated, and the rest are underlapping points where laser frequencies are skipped. This observation is now believed to be a mistake and that every stitch point is an overlapping point. In his defense, the author made the same mistake when first classifying the stitch points. To correctly classify the stitch points as under or overlapping, it is necessary to look at both interferometer signals simultaneously. Figure 3.6.5 shows a situation in which a stitch point could be easily misidentified if viewed through only the 7.1 GHz interferometer.

Figure 3.6.5: A stitch point that could be incorrectly classified.

From the figure, there is noise at the stitch point and by viewing the kclock signal it appears as though there is roughly zero change in laser frequency. But when viewing the DUT signal, the laser stitch point is clearly overlapping by a single kclock cycle. The stitch point with the least amount of overlap is shown in figure 3.6.6.

Figure 3.6.6: stitch point with the least amount of overlap.

Underlapping points are of much greater concern because they can result in a completely missed zero crossing point. Not only that, but if the laser does not output a complete wavelength range, it limits the resolution of the OCT system. Figure 3.6.7 illustrates a glitch introduced by an overlapping stitch and figure 3.6.8 illustrates a glitch by an underlapping point.

Figure 3.6.7: two sampling points are identified (circles), one before and one after an overlapping stitch point. To get the optimal PSF, only one sample point should be chosen.

Figure 3.6.8: an underlapping stitch point causes a positive zero crossing (circle) to be missed. In the actual waveform, the transition can be populated with noise, making it an unreliable measurement.

Sometimes glitches can be introduced by random noise near zero crossings. An easy way to prevent these errors is to filter the kclock signal with a *moving average filter*, a type of digital FIR filter, prior to analysis. The moving average filter can be thought of as roughly equivalent to an analog low-pass filter, but with zero phase delay. A filtered signal is compared to an unfiltered signal in figure 3.6.9.

Figure 3.6.9: Filtered kclock signal (red) versus original (blue).

Notice how the moving average filter smooths over the noise while leaving the outline of the sine wave unaffected.

3.7 Power Leveling

In the laser system of figure 3.4.2, *power leveling* refers to the flatness of laser power over an entire sweep. If the output power is not flat, then the PSF may suffer from amplitude modulation (AM) noise. Variations in output power can also potentially affect the locations of clock sampling points.

Probably the biggest advantage of the balanced detector over the unbalanced detector in this application is in how it deals with an increase in laser power. Figure 3.7.1 illustrates the difference:

Figure 3.7.1: comparison of how an unbalanced (left) versus balanced (right) detector deals with an increase in laser power. After the stitch point, the waveform in the unbalanced detector is no longer symmetric about zero Volts, whereas the waveform in the balanced detector is still symmetric.

From the figure, a stitch point causes a sudden increase in laser output power. In the unbalanced detector, this increases the peak value (when the detector sees maximum light) but does not affect the minimum value (when the detector sees no light). The output of the unbalanced detector then ceases to be symmetric and this leads to a problem where the zero crossing is a less reliable frequency reference. The laser frequency that previously would have been at 0 Volts is now at slightly more than zero and thus, the

sampling interval is affected. Contrast this with the balanced detector, where zero voltage is always a reliable reference for laser frequency, regardless of changes in output power.

Chapter 4: Simulation of the OCT System

4.1 Simulation Outline

The goal of this report is to simulate the operation of our SG-DBR OCT system in measuring the PSF. Most importantly, it is desirable to know given the sampling non-idealities and noise levels that are inherent to the devices used, what the best achievable PSF in this approach to OCT. Additionally, performance curves will be generated showing the level of anomalies that allow a certain PSF quality and noise floor.

By examining the general shape of the PSF (figure 3.6.1) a number of anomalies are identified for simulation. Figure 4.1.1 shows the related noise sources affecting the PSF and the way they are quantified.

Figure 4.1.1: Quantities to measure the simulated PSF. Simultaneously specified quantities are shown in same colors.

The FM-BM and phase transitions are considered *quasi-periodic* (QP) because they repeat at a somewhat regular interval (see figure 4.2.13).

To be entirely specific, the output signals from the two interferometers are simulated *after* they have been detected and acquired by the data acquisition card. Thus, the signals are proportional to laser power but are modeled as *voltage* waveforms. This is a very important point because the final signals are scaled as $20\log_{10}$ of the detected voltage, and not as $10\log_{10}$. Figure 4.1.2 shows the type of noise that can be added to the simulation and how it corresponds to measurable physical quantities.

Type of Noise Added to Simulation	Physical Effects Modeled
Sinusoidal AM	Sinusoidal Changes in Laser Power
Random AM	Instantaneous Power Variations in Laser (Intensity Noise), Optical Detector Noise (dark current, thermal noise)
Sinusoidal PM	Sinusoidal Phase Fluctuations Between Kclock and DUT Waveforms,
Random PM	Instantaneous Laser Frequency Instability (linewidth), Phase Fluctuations Between Kclock and DUT Waveforms,
Quasi-Periodic AM	Discrete Laser Power Changes at the FM-BM or Phase Transitions
Quasi-Periodic PM	Discrete Phase Offsets Between Kclock and DUT at FM-BM or Phase Transitions
Single-step PM	Discrete Change in Laser Frequency once per sweep

Figure 4.1.2: Physical meaning of noise introduced to simulation.

Figures 4.1.3 and 4.1.4 show how the noise sources in figure 4.1.2 are quantified mathematically.

Type of Noise Added to Simulation	Equation of Noise	Noise Parameters
Periodic AM	$V(t) = V_0 \left[1 + \frac{\Delta V}{V_0} \cos(2\pi f_m t) \right] \cos(2\pi f_0 t)$	Modulation index = $\Delta V/V_0$ (no units) Modulation Frequency = f_m (Hertz)
Random AM	$V(t) = V_0 \left[1 + \frac{\sigma_{am}}{V_0} \text{rand}(0,1) \right] \cos(2\pi f_0 t)$	Standard deviation relative to carrier = σ_{am}/V_0 ($V_{\text{noise rms}}/V_{\text{carrier peak}}$)
Periodic PM	$V(t) = V_0 \cos(2\pi f_0 t + \Delta\phi \cos(2\pi f_m t))$	Modulation index = $\Delta\phi$ (radians) Modulation Frequency = f_m (Hertz)
Random PM	$V(t) = V_0 \cos(2\pi f_0 t + \sigma_{pm} \text{rand}(0,1))$	Standard deviation of phase constant = σ_{pm} (radians)

Figure 4.1.3: mathematical explanation of simulated random and periodic noise parameters.

Type of Noise added to Simulation	Description of Noise	Noise Parameters
Quasi-Periodic AM Discrete amplitude variations that occur at the stitch points.		Nth amplitude = $V_n = V_0(1 + (\sigma_{am}/V_0) \text{rand}(0,1))$ (Volts) σ_{am}/V_0 = standard deviation of amplitude relative to carrier ($V_{\text{noise rms}}/V_{\text{carrier peak}}$)
Quasi-Periodic PM Discrete phase variations that occur at the stitch points.		Nth phase constant $\phi_n = \sigma_{pm} \text{rand}(0,1)$ (radians) σ_{pm} = standard deviation of phase constant (radians)
Single-step PM Single discrete phase transition that occurs only once.	Same as above, but happens only once per laser sweep	See above

Figure 4.1.4: mathematical explanation of simulated quasi-periodic noise parameters.

MATLAB was chosen to develop the simulations. The fundamental principle is that the kclock signal is used to trigger the sampling of the DUT signal, which is then Fast Fourier transformed to produce the PSF. To account for the stitch errors, a routine was developed to synthesize waveforms with errors at the FM-BM and/or Phase Mirror transition points. At these transition points the waveforms can change relative phase (laser frequency) and/or amplitude. Figure 4.1.5 shows the synthesized creation of two phase errors on the kclock signal.

Figure 4.1.5: example of intentionally introducing two phase errors (laser frequency steps) to the kclock signal.

Detection of zero crossings is determined by a subprogram, loosely based on the design of one found online [2]. It was modified to provide detection of positive zero crossings only and further optimized to always return the closest point to zero. The zero crossing detection is shown along with both signals in figure 4.1.6.

Figure 4.1.6: Kclock (red) and DUT (blue) signals shown with sample points.

Notice the inconsistent amplitude of the detected zero-crossing points on the kclock signal. This noise is due to the discrete nature of the waveforms; that is, with both waveforms defined by a finite number of points there is a limit in the accuracy of determining the zero-crossing points. Thus, a form of random *jitter* that may potentially cause degradation of the PSF has been introduced.

The quantization noise of the oscilloscope is simulated in software by scaling and rounding to the nearest integer. This method gives 2^n quantization levels, where n is the number of bits. The simulated waveforms are shown in figure 4.1.7 and the theoretical versus measured SNR is shown in figure 4.1.8.

Figure 4.1.7: Digitization of a ramp signal using software. The original signal is shown in red, the digitized signal is shown in yellow, and the error signal (the different between original and digitized) is shown in blue.

Figure 4.1.8: Measured versus theoretical (see equation 2.6.1) signal to noise ratio of simulated ADC.

The digitized signal displays the same characteristics that would be produced by an ideal ADC [11] and the experimental SNR is very close to the theoretical value (see section 2.6). Thus, it appears that the quantization model is functioning as designed.

A block diagram of the complete program structure is shown in figure 4.1.9.

Figure 4.1.9: Block diagram of complete MATLAB simulation program structure.

The parameters that are changed in the simulation are outlined in figure 4.1.10.

Simulation Parameters that Can Change (separately for DUT and kclock)
Periodic AM Modulation Index ($V_{\text{noise peak}}/V_{\text{carrier peak}}$)
Random AM Standard Deviation Relative to Carrier ($V_{\text{noise rms}}/V_{\text{carrier peak}}$)
Periodic PM Modulation Index (radians)
Random PM Standard Deviation (radians)
Quasi-Periodic AM Standard Deviation Relative to Carrier ($V_{\text{noise rms}}/V_{\text{carrier peak}}$)
Quasi-Periodic PM Standard Deviation (radians)
Single-step PM (radians)

Figure 4.1.10: list of MATLAB simulation parameters that are changed.

The parameters that are held constant throughout all simulations are shown in figure 4.1.11.

Unchanging Simulation Parameters
Array size that defines kclock and DUT waveforms = 50,500
Number of kclock zero crossings = 640
Kclock waveform frequency (in time domain) = 0.760 MHz
DUT waveform frequency (in time domain) = 0.760 MHz $\cdot$ (7.1/39) $\approx$ 0.138 MHz
Kclock modulation frequency = 50 kHz (periodic amplitude and phase modulation)
DUT modulation frequency = 50 kHz (periodic amplitude and phase modulation)
12 bits of digitization
Scale DUT by Hann window function before FFT
5-point moving average filter on kclock prior to zero-crossing detection

Figure 4.1.11: list of MATLAB simulation parameters that are held constant.

4.2 Fundamental Limits to Performance

Before looking at the complete system simulation, it is necessary to ask some basic questions about the simulations and whether any part of it tends to limit the resolution of the system.

With the chosen laser tuning bandwidth and choice kclock interferometer FSR, the number of sample points is roughly 640; this is therefore the number of points in the DUT signal. After applying the window function to the sampled data, the FFT algorithm goes to work and produces the PSF. So, assuming a perfect quality sine wave of roughly 640 points, just how good is the resolution of the FFT algorithm? Figures 4.2.1 and 4.2.2 show the results.

Figure 4.2.1: result of an FFT of a pure sine wave of 640 points.

Figure 4.2.2: close-up of an FFT of a pure sine wave of 640 points.

From the figures, this is the fundamental limit of the FFT process with the total number of sample points and the sampling interval (samples per cycle/ratio of interferometer FSR's) used in the experiment.

Now we measure the amplitude of the central peak as a function of offset to gauge the resolution of the FFT process. Figure 4.2.3 shows the result.

Figure 4.2.3: PSF amplitude of central peak as a function of offset (of PSF produced in figure 4.2.1)

Now, what is the effect if the pure sine wave above is digitized at 12 bits (same as the PC-6115 oscilloscope)? Figure 4.2.3 shows the result.

Figure 4.2.4: Pure sine wave with 12 bits of digitization.

From the figure we see that with 12 bits applied to the sample signal, there is a white noise floor below -90 dBc. The amplitude of the central peak as a function of offset is shown in figure 4.2.5.

Figure 4.2.5: PSF amplitude of central peak as a function of offset (of PSF produced in figure 4.2.4)

Comparing figures 4.2.3 and 4.2.3, there is no distinguishable difference between the resolution of the central peak with and without digitization. This leads to the assumption that digitization at 12 bits has a negligible effect on system resolution.

Next, how is the resolution of the system affected when the kclock is used to trigger sampling of the DUT? Here both signals are perfectly sinusoidal. Figure 4.2.6 shows the result.

Figure 4.2.6: The result of pure sine kclock and DUT signals, no digitization.

The noise created by the kclock sampling is especially interesting because of the pattern that it creates. There are spikes that are spaced approximately every two mm, the height which varies from -75 to -50 dBc. Since the main interest is in obtaining a maximum of 50-60 dB, the spikes of larger amplitude may be a limiting factor in obtaining a high resolution. Once again, the amplitude of the central peak is measured as a function of offset; the result is shown in figure 4.2.7.

Figure 4.2.7: PSF amplitude of central peak as a function of offset (of PSF produced in figure 4.2.6)

Compared to figure 4.2.3, figure 4.2.7 has no noticeable difference in terms of phase noise. Therefore, it is theorized that although the sampling noise results in the appearance of spurious peaks, it does not contribute any noise to the central peak.

The theory for the cause of the spikes is due to the nature of the way the kclock is sampled. Because of the discrete nature of the signals, the sampling interval can never be at perfect intervals of frequency (refer back to figure 4.1.6). Another way to visualize what is happening is to detect the zero crossing points on the kclock and then use these points to sample the kclock instead of the DUT signal. In other words, use the kclock to sample itself at the zero crossing points. Ideally, the zero crossing points should occur at exactly zero, and the resulting power spectrum should be null. Instead, there is the pattern shown in figure 4.2.8.

Figure 4.2.8: result of sampling the kclock with itself.

Figure 4.2.8 shows that sampling the kclock with itself produces a periodic pattern that is very similar to what was observed in figure 4.2.6. With the perfect sine wave generated by the computer program, there is a certain amount of periodicity in the way that errors are introduced because of the discrete nature of the signals. One would not expect this type of noise to be present in typical analog zero-crossing circuitry.

Theoretically, it should be possible to improve the result by increasing the array size of the kclock and DUT signal points (while keeping the number of samples the same). If 1 million, or roughly twenty times the normal array size is used, the result is shown in figure 4.2.9.

Figure 4.2.9: Result of sampling the DUT signal with array size of 1.01 million points.

From the figure, we can see that the spurious peaks are decreased in amplitude to a maximum of roughly -80 dBc. This makes sense because with a larger array size, the zero-crossing detection should be more accurate. In general, it would be beneficial to know what array size to ensure that this anomaly does not affect the measured PSF.

Figure 4.2.10 shows a plot of the maximum amplitude of the anomaly versus the array size.

Figure 4.2.10: Amplitude of ‘spikes’ versus array size.

It is theorized that these “spikes” may not be visible in an actual PSF because of random effects. That is, the random nature of a physical signal destroys any periodicity that would otherwise result from sampling an ideal sine wave. Sure enough, if a small amount of random phase noise is added to the kclock signal, the spikes seem to go away. Figure 4.2.11 shows the result.

Figure 4.2.11: Enough random phase modulation applied to kclock signal removes the spikes. Standard deviation of phase noise is 0.05 radians, equivalent to a standard deviation of 56.5 MHz for laser frequency.

To determine the noise associated with the detectors, data was taken through a National Instruments PCI-5105 oscilloscope with both detectors exposed to no light. Figure 4.2.12 shows the resulting FFT of the data (refer to figure 5.1.1 for system block diagram).

Figure 4.2.12: FFT of optical detectors when their inputs have been blacked out. Unbalanced (left) and balanced (right).

From the resulting FFT we can see that the noise from both detectors is random (white) noise at a level of roughly -80 dB. So to simulate this noise, we use a Gaussian random variable.

To measure the noise inherent to the laser, data is taken from the interferometer network while the laser output is “parked” at a single frequency (not sweeping). Figure 4.2.13 shows the results.

Figure 4.2.13: Laser output noise viewed through unbalanced detector (left) and balanced detector (right). The inputs to the SG-DBR laser are held at: FM = 18.2 mA, BM = 24.4 mA, PM = 0 mA

From the figure it appears that the laser noise is roughly equivalent to white noise, although the spectrum is not completely flat and there is a higher amount of noise in the low frequency range. Figure 4.2.14 shows the laser noise at lower frequencies.

Figure 4.2.14: Laser noise through unbalanced (left) and balanced (right) detectors below 1.5 MHz (close-up of figure 4.2.10).

Because both noise sources are roughly equivalent to white noise, the decision was made to simulate them using a Gaussian random variable at amplitudes of -80 and -70 dB. The standard deviation of a Gaussian noise source needed to produce the required -80 dB and -70 dB was determined by simulation in MATLAB.

Based on the maximum voltage swing of our optical detectors, it is assumed that this voltage is now equivalent to the maximum amplitude that should be possible to achieve. Now we find the difference between this and the previously measured noise levels, so that the kclock and DUT signals can be referred to in terms of dBc, or “decibels relative to the carrier signal.” This is the most convenient way to specify noise levels and is made possible because the amplitudes of the two waveforms are never directly compared against each other and only “interact” in terms of sampled points.

To determine exactly where the stitch points occur, we look at the output of the laser using the final C-band calibration files [5]. Historically, the laser sweeps have been

recorded as roughly 50k data points. This is because the ARB's are limited to 64k total points and the person who first developed [16] the sweeping laser thought that only using 50k of them would give enough "headroom." Initially, the author decided on an array size of 50.5k points in a simulated sweep and measured the location of the glitches in reference to this number, although the stitch points can be scaled to any multiply of 50.5k. Figure 4.2.15 shows the locations of stitch points on a complete laser sweep and figure 4.2.16 a histogram of the time between stitch points:

Stitch Number	Location (midpoint)
1	1216
2	3039.5
3	4871
4	6715.5
5	7885
6	9700.5
7	11500.5
8	13337
9	13460.5
10	14652
11	16430
12	18235.5
13	20041.5
14	20282
15	21445.5
16	23225
17	25002
18	26811
19	27413
20	28850
21	30636
22	32421.5
23	34194.5
24	35075.5
25	36285.5
26	38041.5
27	39818.5
28	41590
29	42498.5
30	43633
31	44172
32	44238
33	45355.5

34	47210
35	48969.5

Figure 4.2.15: (approximate) location of stitch points in 50.5k total points. FM-BM transitions are bolded.

The time between stitch points is useful to know because it is a gauge of the periodicity of the phase transitions. The time between stitch points (again with an array size of 50.5k points) is shown in figure 4.2.16.

Figure 4.2.16: (unscaled) time between stitch errors. Total time is 50.5k points.

From the figure one can see that the time between stitch points is somewhat regular, with an average of roughly 1700 points.

4.3 Simulation Results

The “best possible” PSF results from only the noise sources that cannot be removed; i.e. it is assumed that the laser sweep is completely linear (perfectly sinusoidal kclock and DUT) and has added to it only random detector and “non-sweeping” laser

noise, both of which are quantified in section 4.2. We use an array size of 50.5k points because it closely matches the number of points in an actual measurement. Figure 4.3.1 and 4.3.2 show the results.

Figure 4.3.1: Simulated “best possible” PSF result with only detector and laser noise. Note that the two spikes are from the previously discussed “sampling noise” and not from a source of modulation.

Figure 4.3.2: Close-up of “best possible” simulated PSF.

Overall, the central peak in the PSF looks very smooth, until about -55 dBc. Below - 55 dBc, both sides of the peak seem to become slightly distorted. This is the result of the random noise (of nominal amplitude -70 dBc) introduced into the simulation. Notice in figure 4.3.1 that there are two prominent peaks on either side of the central peak. These are the result of the previously mentioned (see figure 4.2.4) sampling noise, which is not observed in the physical system.

Now the performance curves are presented for the simulations with varying levels of noise. Some of the noise sources should be correlated between both waveforms (such as discrete/sinusoidal increases in laser power) while others should not (random amplitude and phase noise). Because of that, the design curves for the OCT system are broken into three sections (kclock only, DUT only, or noise on both waveforms at the

same time) and then by the type of noise added. Please refer back to figures 4.1.1 through 4.1.3 for an illustration and mathematical description of the noise quantities.

All performance curves are presented with an array size of 50.5k points because it closely matches the number of points in the physical measurement. For curves that show noise floor, the noise floor is estimated by window averaging the last 70 points in the FFT plot, so that the calculation is not influenced by the characteristics of the central peak, as shown in figure 4.3.3. For a complete list of parameters and definitions used in the simulations, please see section 4.1.

Figure 4.3.3: the noise floor is calculated by averaging the final 70 points in the PSF.

4.3 Simulation Results

4.3a Noise Applied only to DUT

Here are the design curves based on the noise sources that are introduced to the DUT signal only. The only noise sources applied to the kclock are the detector and laser noise and are equal for all simulations in this subsection.

Figure 4.3.4: PSF noise floor versus amplitude modulation standard deviation relative to carrier applied to DUT signal. Noise floor is calculated by a 70 point window average.

Figure 4.3.5: PSF noise level of carrier signal as a function of offset and DUT random amplitude standard deviation relative to carrier.

Figure 4.3.6: PSF noise floor versus phase modulation standard deviation applied to DUT signal. Calculated by a 70 point window average.

Figure 4.3.7: PSF noise level of carrier signal as a function of offset and DUT random phase modulation standard deviation.

Figure 4.3.8: PSF max sideband amplitude versus sinusoidal phase modulation index applied to DUT signal.

Figure 4.3.9: Noise floor and max sideband amplitude versus quasi-periodic phase modulation standard deviation applied to DUT signal at FM-BM transitions. Noise floor is calculated by a 70 point window average. Each data point is average of 10 measurements. Error bars show +/- standard deviation.

Figure 4.3.10: Noise floor and max sideband amplitude versus quasi-periodic phase modulation standard deviation applied to DUT signal at phase mirror transitions. Noise floor is calculated by a 70 point window average. Each data point is average of 10 measurements. Error bars show +/- standard deviation.

4.3b Noise Applied only to Kclock

Here are the design curves based on the noise sources that are introduced to the kclock signal only. The only noise sources applied to the kclock are the detector and laser noise and are equal for all simulations in this subsection. It should be noted that unlike the DUT signal, the kclock signal is filtered with a 5 point moving-average filter prior to zero-crossing detection (see figure 3.6.9). This causes the kclock signal to be less sensitive to the effects of random noise.

Figure 4.3.11: PSF noise floor versus amplitude modulation standard deviation relative to carrier applied to kclock signal. Calculated by a 70 point window average.

Figure 4.3.12: PSF noise level of carrier signal as a function of offset and kclock random amplitude standard deviation relative to carrier.

Figure 4.3.13: PSF noise floor versus phase modulation standard deviation applied to kclock signal. Calculated by a 70 point window average.

Figure 4.3.14: PSF noise level of carrier signal as a function of offset and kclock random phase modulation standard deviation.

Figure 4.3.15: PSF max sideband amplitude versus sinusoidal phase modulation index applied to kclock signal.

Figure 4.3.16: Noise floor and max sideband amplitude versus quasi-periodic phase modulation standard deviation applied to kclock signal at FM-BM transitions. Noise floor is calculated by a 70 point window average. Each data point is average of 10 measurements. Error bars show +/- standard deviation.

Figure 4.3.17: Noise floor and max sideband amplitude versus quasi-periodic phase modulation standard deviation applied to kclock signal at phase mirror transitions. Noise floor is calculated by a 70 point window average. Each data point is average of 10 measurements. Error bars show +/- standard deviation.

4.3c Noise Applied simultaneously to kclock and DUT Signals

Here we present the design curves based on the noise sources that are introduced to both the kclock and DUT signals. Increases in laser power will appear across both signals, but the waveforms will behave differently do to the difference in operation between unbalanced (kclock) and balanced (DUT) detectors (see section 3.7 and especially figure 3.7.1).

Figure 4.3.18: Max sideband amplitude versus sinusoidal amplitude modulation index applied to kclock and DUT signals.

Figure 4.3.19: Noise floor and max sideband amplitude versus quasi-periodic amplitude modulation standard deviation relative to carrier applied to kclock and DUT signals at FM-BM transitions. Noise floor is calculated by a 70 point window average. Each data point is average of 10 measurements. Error bars show +/- standard deviation.

Figure 4.3.20: Noise floor and max sideband amplitude versus quasi-periodic amplitude modulation standard deviation relative to carrier applied to kclock and DUT signals at phase mirror transitions. Noise floor is calculated by a 70 point window average. Each data point is average of 10 measurements. Error bars show +/- standard deviation.

4.4 General Equations for Noise Levels and Resolution

The performance curves shown in sections 4.3a-4.3c are unique to the FSR of the interferometers used in the experiment. To generalize the results to arbitrary FSR, we present the following equations and results.

If for example one has some amount of noise on the DUT signal, and a noise-free kclock (or perfect sampling interval), it is a simple task to calculate the effects of noise on the PSF by using the noise equations outlined in section 2.7 (specifically equations 2.7.4 and 2.7.7). However, in the opposite scenario, with a noisy kclock and an ideal DUT, what are the effects on the PSF? The general answer is now more complex, but in some cases it is still possible to derive analytical expressions for the “virtual” noise on the DUT signal.

For example, consider a sinusoidal kclock waveform with added periodic phase modulation. The phase modulation creates a variation in the kclock zero crossings which then change the sample points on the DUT waveform. Intuitively, one can visualize that the same effect could be generated by removing the sinusoidal modulation from the kclock, scaling it appropriately and inserting it into the DUT waveform. Figure 4.4.1 shows the relationships between phase modulation on the kclock and the equivalent virtual phase modulation that appears on the DUT signal.

Given a noiseless DUT and noisy kclock in the time domain:	
Kclock waveform	$V_{clock}(t) = V_{clock} \cos(2\pi f_{clock} t + \Delta\phi_{clock} \cos(2\pi f_m t))$
DUT Waveform	$V_{DUT}(t) = V_{DUT} \cos(2\pi f_{DUT} t)$
The waveforms can be converted to an equivalent “virtual” noise source on the DUT as follows:	
New kclock waveform	$V_{clock}(t) = V_{clock} \cos(2\pi f_{clock} t)$
New DUT Waveform	$V_{DUT}(t) = V_{DUT} \cos(2\pi f_{DUT} t + \Delta\phi_{DUT} \cos(2\pi f_m t))$
Kclock peak frequency error $= \Delta f_{clock}$	
Kclock peak time error $= \Delta t_{clock}$	
$\Delta t_{clock} = \frac{\Delta f_{clock}}{2\pi f_m f_{clock}}$	
DUT peak phase error	
$\Delta\phi_{DUT} = 2\pi (\Delta t_{clock} f_{DUT}) = \left(\frac{\Delta f_{clock} f_{DUT}}{f_m f_{clock}} \right) = \Delta\phi_{clock} \left(\frac{FSR_{clock}}{FSR_{DUT}} \right)$	
Periodic sideband height relative to carrier (in dBc; approximation given a low modulation index)	
$L(f) = 20 \log \left(\frac{\Delta\phi_{DUT}}{2} \right)$	

Figure 4.4.1: relationships between kclock waveform phase noise levels and the “virtual” noise level as it appears on the DUT waveform.

For example, say that using a 10 GHz kclock and a 40 GHz, the peak kclock phase error $\Delta\phi_{clock}$ is observed to be 1 radian. Then the equivalent peak phase error on the DUT signal $\Delta\phi_{DUT}$ is $1 \cdot (10 \text{ GHz} / 40 \text{ GHz}) = 0.25$ radians.

Also important for general analysis is the horizontal axis scaling on the PSF. In an OCT measurement the horizontal axis of the PSF is shown in units of distance (see equation 3.5.1). But it can also be important to consider the horizontal axis in terms of temporal frequency when analyzing the effects of modulation and jitter on the kclock and DUT waveforms. For example, this viewpoint may be useful to identify the modulation frequency. Figure 4.4.2 shows the equations that pertain to the scaling of the horizontal axis in the PSF.

Kclock sampling interval (Hertz) $= \Delta f_{\text{int}}$ Time between kclock samples (seconds) $= \Delta t_{\text{sample}}$ Maximum horizontal distance in PSF (equation 3.5.1, c is speed of light in free space) $x_{\text{max}} = \frac{c}{2(\Delta f_{\text{int}})}$
Number of points in PSF $= N_{\text{PSF}}$ Number of sample points $= N_{\text{samples}}$ $N_{\text{PSF}} = \frac{N_{\text{samples}}}{2} + 1$
Spatial resolution of PSF $\Delta x_{\text{PSF}} = \frac{x_{\text{max}}}{N_{\text{PSF}}} = \frac{c}{(\Delta f_{\text{int}})(N_{\text{samples}} + 2)}$
Time frequency resolution of PSF (Hertz) $\Delta f_{\text{PSF}} = \frac{1}{(N_{\text{samples}} + 2)\Delta t_{\text{sample}}}$

Figure 4.4.2: relationships between horizontal axis scaling of the PSF

Chapter 5: Experimental Investigation

5.1 Experimental Setup

With the software and techniques developed from simulating the OCT system, it would be desirable to know what kind of performance one can expect from the actual devices. Also, observations on the signals can be made and then implemented into the simulation. Then a final simulation can be run, compared to the observed FFT and the differences described.

The experimental setup is similar to (figure 3.4.2), except that the Cal Poly group no longer possesses the PCI-6115 card. It has been replaced with a PCI-5105 card, which offers superior performance (60 MS/s versus 10 MS/s) and 16 MB of onboard memory. The new experimental setup is shown in figure 5.1.1.

Figure 5.1.1: Current SG-DBR OCT demonstration system at Cal Poly.

At this point, an ekclock (see section 3.5 for description) is not used to trigger sampling of the DUT as in the earlier experimental setup, but this configuration will still provide the raw data needed to evaluate the performance of the laser.

The unbalanced detector was designed and built by the author. It uses an AD8047 high speed voltage-feedback operational amplifier in transimpedance configuration and a Perkins-Elmer C30617BQC-07 photodiode collimated to optical fiber. The schematic is shown in figure 5.1.2.

Figure 5.1.2: Unbalanced Detector Schematic. It utilizes a 10V, single-sided power supply and 5V is applied to the non-inverting input to reverse bias the photodiode.

The photodiode acts as a current source that is proportional to the optical power incident on its surface. Any current not required to bias the opamp (an insignificant amount) will flow across the feedback resistor R_f , producing a voltage that is proportional to the photocurrent. The feedback capacitor C_f limits the bandwidth of the amplifier for stability and noise rejection. With this choice of R_f and C_f , the 3 dB bandwidth of the amplifier is almost 8.5 MHz (see AD8047 datasheet).

The 68 uF capacitors are standard electrolytic, the 0.1 uF are metal film, and the 2 pF (two 1 pF in parallel) ceramic. The output pin is connected to 50Ω semi-rigid coax. The entire device is constructed on copper-clad circuit board in a technique known as *ugly construction*. Components are soldered directly to each other and the large copper area provides a low impedance ground plane. Figure 5.1.3 shows a photo of the completed device.

Figure 5.1.3: photo of unbalanced photodetector transimpedance amplifier.

Not including the photodiode, total cost of the detector is less than \$15. For comparison, the Thor Labs balanced detector cost more than \$1300.

5.2 Measurement Results

With the completed detector it is now possible to simultaneously acquire both interferometer signals during a complete laser sweep. To record the interferometer signals, the author wrote a LabVIEW VI that waits for a trigger signal on and saves them to file. Figure 5.2.1 shows a picture of both signals as measured.

Figure 5.2.1: both interferometer signals acquired throughout multiple sweeps. Array size of 150k points each, 60 MS/s sample rate, 1 kHz laser sweep repetition rate.

From the preceding figure it is obvious that laser power is not constant throughout the entire sweep. Figures 5.2.2 and 5.2.3 show the DUT and kclock in detail.

Figure 5.2.2: Time detail of measured DUT signal.

Figure 5.2.3: Time detail of measured kclock signal.

To calculate the PSF, the DC level is removed from the signals and they are loaded them into the MATLAB program used to generate the performance curves in chapter 4. Figures 5.2.4, 5.2.5 and 5.2.6 show the results, first without deglitching, then with manual deglitching and finally a close up of the central peak.

Figure 5.2.4: PSF calculated with measured signals without deglitching (removing repeated sample points).

Figure 5.2.5: PSF calculated with measured signals after deglitching.

Figure 5.2.6: Close-up of PSF calculated with measured signals after deglitching.

As the figures demonstrate, deglitching the kclock drastically improves the resolution of the PSF, from less than 30 dB to roughly 50 dB. To get the maximum resolution, any repeated sampled points (that occur at overlapping stitch points) must be removed. Even a single repeated sample point can drastically compromise the resolution of the PSF.

Figure 5.2.7 shows the effect of one through four repeated sampling points.

Figure 5.2.7: PSF with one (upper left) two (upper right) three (lower left) and four (lower right) repeated sample points.

Repeated sampling points result in broadening of the central peak and an increase in the overall noise floor. With only two repeated sampling points, the noise floor rises to -40 dBc and the peak width below -30 dBc is almost 2 mm. With more than two repeated sample points missed, the effects of additional repeated sample points are less noticeable. It should be noted that due to the shape of the window function, extra sample points at the

beginning or end of the waveform will have a smaller effect than those that occur in the middle.

By observing figure 5.2.1, it is obvious that the kclock is subject to a large amount of amplitude modulation from peak to peak. This could theoretically lead to errors in determining the sampling interval (see section 3.8 for explanation). However, the dark fringes appear to be much more consistent in their amplitude than the light fringes. Therefore, using the dark fringes (valleys) to determine the sampling interval should result in greater accuracy. Figures 5.2.8 and 5.2.9 shows the result of using the dark fringes to calculate the PSF.

Figure 5.2.8: Plot of the PSF using dark fringe (valley) detection.

Figure 5.2.9: Close-up of the PSF using dark fringe (valley) detection.

Although not a drastic improvement compared to the PSF with the zero-crossing algorithm (figure 5.2.6), valley-detection results in a PSF that has lower amplitude sidebands and fewer spurious responses near the central peak.

5.3 Results Using Insight Photonic Solutions Developed Software

As mentioned in section 3.4, Insight Photonic Solutions developed a program to generate a PSF using LabVIEW software that interfaces directly with an oscilloscope. It has a graphical user interface that allows the user to analyze the sampling points that the program has selected and change them if they do not optimize the resolution of the PSF.

Rather than using a zero-crossing threshold detector on the kclock signal, it instead looks for peaks and valleys and determines the midpoint between them to determine the sample point. This is an advantage because in the unbalanced detector, variations in laser power can affect the sampling interval (see section 3.8). Figure 5.3.1 shows the results of the initial PSF and Figure 5.3.2 after glitches have been removed manually.

Figure 5.3.1: PSF without manual deglitching of the kclock.

Figure 5.3.2: PSF after manual deglitching of the kclock.

Figure 5.3.3: Close-up of the PSF using Insight Photonic Solutions LabVIEW software.

As is shown earlier, the PSF is drastically improved by the deglitching procedure.

Compared to the original software algorithm developed in this paper, the PSF appears to be more defined and the heights of the sidebands are reduced as well. The quality of this PSF seems to be comparable to the PSF that uses valley detection.

5.4 PSF Based on Measured Noise Levels

To get an even better prediction of the theoretical performance of the laser system one can make assumptions about the measured noise levels in section 5.2 and plug them back into the simulation. Based on the interferometer waveforms, it appears that most of the noise is random AM; specifically, AM that is most noticeable when detected laser power is at its peak (peak value for unbalanced detector, peak or minimum value for balanced detector). This type of noise was already implemented in the simulation, so it is a simple matter to vary the magnitude until it matches visually what is measured. Figure 5.4.1 shows a sample of the DUT signal with the random AM noise implemented (compare to figure 5.2.2).

Figure 5.4.1: Simulated random amplitude noise on DUT signal.

Also an important factor affecting the laser output is modulation due to power leveling (see section 3.8). It is easy to see by looking at (figure 5.2.1) that laser power is not constant over an entire sweep. To incorporate effect into the simulation, the interferometer was disconnected and the balanced detector was used to directly measure the laser. The result of this measurement is shown in figure 5.4.2.

Figure 5.4.2: Laser power versus time of the C-band SG-DBR laser through one sweep.

The vertical axis is scaled based on the responsivity of the balanced detector. However, the important part of the figure is not the absolute value of the laser power but the fact that is changing over the duration of the laser sweep. This “power leveling envelope” can be used to modulate the signals used in the simulation.

With power leveling noise and the “heuristic” random AM noise implemented we can finally look at the final simulations of the PSF. Figures 5.4.3 and 5.4.4 show the PSF (without the power leveling envelope).

Figure 5.4.3: PSF with added AM noise (no power leveling). This represents the theoretical performance limit if laser power is kept perfectly flat.

Figure 5.4.4: Close up of PSF with added AM noise (no power leveling).

Based on the figures, the theoretical maximum obtainable resolution is between 50 and 60 dB if laser power is kept perfectly flat. Compared to the almost 50 dB of resolution generated from the measured PSF, this is a very encouraging result because it shows that the PSF is close to reaching the peak performance of the system. The same simulation with the addition of power leveling generates the result in figure 5.4.5 and 5.4.6.

Figure 5.4.5: Simulated PSF with random AM and power leveling noise.

Figure 5.4.6: Close up of simulated PSF with random AM and power leveling noise.

With power leveling implemented there is a noticeable increase in noise. Noise peaks reach levels of between -40 and 45 dBc and there are also sidebands that form around the central peak. Overall, figure 5.4.6 has some of the same features that are observed on the measured PSF (5.2.6) such as spurious peaks at a height of roughly -40 dB.

Chapter 6: Noise Observations and Conclusion

6.1 Observations about Noise in Our System

Deglitching the kclock is absolutely vital. One needs to ensure that in the case of overlapping stitch points, there are no repeated sampling points (see figure 3.6.7).

Failure to capture every single sample point without repetition leads to a huge increase in the noise floor and width of the central peak that dominate over all other noise sources.

Missed sampling points are an equal concern to repeated sampling points because they affect the quality of the PSF in the same way. However, the measurements demonstrate that the automated calibration program is able to generate a laser sweep such that every stitch point has at least some amount of overlap. Therefore, missed sampling points with the C-band laser investigated are no longer a concern.

Besides deglitching, amplitude modulation at the stitch points is likely the most important source of noise in our system. The calibration done by Chiccone [5] states that the laser has approximately 3% variation in output power, but according to current measurements laser power varies by more than 10%, mainly through discrete transitions at the stitch points, though there are also gradual changes in between the stitch points. The change in responsivity of the photodiode over the working range of the laser is less than 5%, so it should not account for the discrepancy.

The automated laser calibration software used an optical spectrum analyzer to validate the output power. It is my belief that the spectrum analyzer is unable to handle measurements at such a high repetition rate (1 kHz) as used in our laser. At a 1 kHz repetition rate, the electrical signals at the output of the detector are almost 0.8 MHz. My

recommendation would be to instead use the Thor Labs PDB130C balanced detector (or other similar high speed detector) connected to an oscilloscope.

The effects of AM noise lead to a very important point: assessing the linearity of the frequency sweep is rather unimportant because we can always use the kclock to precisely determine the change in output frequency (notwithstanding small, random variations or jitter due to the finite size of the waveform arrays). However, variations in laser power lead to distortion in the PSF and will ultimately limit the accuracy of the complete OCT system when measuring samples at different depths.

When observing the PSF, the sidebands around the central peak are not symmetric as one would see through periodic/sinusoidal modulation. Also, they are very consistent in amplitude and location regardless of the method used to determine the sample points, leading to the theory that they are a direct result of the power variation in the laser and not due to variations in the sampling interval.

Random AM noise on the DUT signal appears to set the limit on the noise floor of the PSF to less than 60 dB and is at least 10 dB higher than either the noise of the detectors or the laser when it is fixed at a single frequency. This source of noise is inherent to the laser and cannot be removed like AM noise at the stitch points can. It's theorized that this source of noise could be reduced by using a higher precision current source for the gain section, optimally a purpose-built laser-diode power supply, rather than the switching-mode bench power supply that was used here.

Measuring changes in the width of the PSF is very difficult because it is so sharply defined. In a PSF of 321 pixels, the width of the central peak below 30 dBc is

only 8 pixels. Therefore, without some sort of interpolation it is virtually impossible to say exactly how wide the central peak is at given amplitude.

Using the kclock dark fringes to determine the sampling interval, or alternatively the Insight Photonic Solutions software that interpolates between peaks and valleys, both seem to produce slightly better PSF's than the original idea of using the zero crossings. This result is likely caused by the amplitude modulation on the kclock signal. However, even with a perfect kclock/sampling interval, amplitude modulation on the DUT still degrades the quality of the PSF.

Random noise tends to be the greatest when the interferometer is centered on a bright fringe. For the unbalanced detector, this occurs at a voltage minimum but for the balanced detector it occurs at either peak. Therefore, it is theorized that zero-crossing detection is more beneficial when using a balanced detector, whereas the unbalanced detector should rely on minimum detection. However, these observations should be weighed against the fact that the interferometer is most sensitive to changes in frequency at the zero-crossings.

6.2 Conclusion

This thesis presented a method for characterizing the PSF of a SS-OCT system using a JDS Uniphase C-band SG-DBR laser. Based on previous observations of the PSF, possible noise sources were identified and simulated using MATLAB. Performance curves were generated to show how the shape of the PSF changes in response to the strength of the noise levels.

Next, the PSF was measured using a 39 GHz FSR interferometer to simulate the DUT and a 7.1 GHz interferometer to serve as the sampling interval reference. Comparisons were made between different techniques in determining the sampling interval, and it was determined that better performance can be achieved by min/max detection as opposed to threshold detection. Additionally, these measurements identified large variations in the output power of the laser that resulted from the automated SG-DBR calibration system developed in reference [5].

Based on the observed noise levels the theoretical spurious-free dynamic range of the PSF should be almost 60 dB and limited almost entirely by the white noise floor from the intensity noise of the laser. Reducing the intensity noise could likely be achieved by using a low noise current source to drive the gain section of the laser. However, non-random variations in output power lead to spurious responses that limit the maximum resolution to roughly 40 dB. As a general observation, with the application of random noise, the shape of the central peak does not change noticeably until roughly 15 dB above the average white noise floor.

The linearity of the frequency sweep is not an important consideration because the kclock always provides a consistent reference for laser frequency. Also, based on the

linewidth of the laser (roughly 20 MHz, measured in [5]), the effects of random laser frequency variation are very small, and should limit the noise floor of the PSF to roughly 70 dB. Therefore, it is determined that the most important noise source is non-random laser output power variation because it creates spurious responses that reach well above the white noise floor. Only once the output power is stabilized would it be beneficial to begin mitigating the effects of random laser power variation.

Figure 6.2.1 summarizes the approximate noise levels required to produce a spurious-free dynamic range (SFDR) of at least 50dB, a reasonable benchmark considering past developments in OCT.

↓Noise Type ↓	SFDR Required (dBc)	DUT Noise Level Required	Kclock Noise Level Required	Noise Quantity (noise units)
Sinusoidal AM	50	<0.0075	<0.0075	Modulation index ($V_{\text{noise peak}}/V_{\text{carrier peak}}$)
Random AM	50	<0.020	<0.30	SD relative to carrier ($V_{\text{noise rms}}/V_{\text{carrier peak}}$)
Sinusoidal PM	50	<0.0065	<0.036	Modulation index (radians)
Random PM	50	<0.020	<0.30	SD (radians)
Quasi-periodic AM	50	<0.010	<0.010	SD relative to carrier ($V_{\text{noise rms}}/V_{\text{carrier peak}}$)

Figure 6.2.1: Outline of the noise parameters necessary to achieve a 50 dBc SFDR in the PSF. Noise quantities are specified individually.

A very important point about figure 6.2.1 is that the noise quantities specified are individually; i.e., that they assume only one source of noise at time. If two or more noise sources are present, the level of each noise source will need to be lower than the maximum specified in figure 6.2.1 in order to achieve the stated 50 dB SFDR.

Works Cited

- [1] Brezinski, Mark E. *Optical Coherence Tomography: Principles and Applications*. Amsterdam [u.a.: Elsevier Academic, 2006. Print.
- [2] Brueckner, Steffen. *Matlab Central*. Web. 3 Mar. 2011.
<<http://www.mathworks.com/matlabcentral/fileexchange/2432>>.
- [3] Button, Kenneth John. *Infrared and Millimeter Waves*. Vol. 11. New York [u.a.: Academic, 1984. 239-89. Print.
- [4] Cheever, Erik. "A (Re)Introduction to MATLAB." *Swarthmore College*. Web. 1 Mar. 2011.
<<http://www.swarthmore.edu/NatSci/echeeve1/Class/e71/E71L1/E71L1.html>>.
- [5] Christopher L. Chiccone, "Automated SG-DBR Tunable Laser Calibration Optimized for Optical Coherence Tomography Applications," Electrical Engineering Department, California State Polytechnic University, San Luis Obispo. June 2011.
- [6] Cooley, James W.; Tukey, John W. (1965). "An algorithm for the machine calculation of complex Fourier series". *Math. Comput.* **19** (90): 297–301.
- [7] Dennis Derickson, Mike Bernacil, Andrew DeKelaita, Ben Maher, Shane O'Connor, Matthew N. Sysak, and Leif Johanssen, "SGDBR single-chip wavelength tunable lasers for swept source OCT," Electrical Engineering Department, California Polytechnic State University, San Luis Obispo.
- [8] "Encyclopedia of Laser Physics and Technology - Interferometers, Mach-Zehnder, Michelson, Fabry-Pérot, Common-Path." *RP Photonics Consulting - Laser and Amplifier Design, Nonlinear Optics, Fiber Optics, Fiber Lasers and Amplifiers, Ultrashort Pulses, Software, Optoelectronics, Consultant, Training*. Web. 06 Sept. 2011. <<http://www.rp-photonics.com/interferometers.html>>.
- [9] "Encyclopedia of Laser Physics and Technology - Lasers, Principle of Operation, Resonator, Cavity, Laser Beam, Stimulated Emission." *RP Photonics Consulting - Laser and Amplifier Design, Nonlinear Optics, Fiber Optics, Fiber Lasers and Amplifiers, Ultrashort Pulses, Software, Optoelectronics, Consultant, Training*. Web. 06 Sept. 2011. <<http://www.rp-photonics.com/lasers.html>>.
- [10] Essick, John. *Advanced LabVIEW Labs*. Upper Saddle River, NJ: Prentice Hall, 1999. Print.
- [11] Franco, Sergio. *Design with Operational Amplifiers and Analog Integrated Circuits*. New York: McGraw-Hill, 2002. Print.

- [12] Brandon J. George, "Swept-Frequency Sampled Grating Distributed Bragg Reflector Lasers Optimized for Optical Coherence Tomography Applications," Electrical Engineering Department, California Polytechnic State University, San Luis Obispo. December 2009.
- [13] Brandon George, Dennis Derickson, "High-Speed Concatenation of Frequency Ramps Using Sampled Grating Distributed Bragg Reflector Laser Diode Sources for OCT Resolution Enhancement," Electrical Engineering Department, California Polytechnic State University, San Luis Obispo.
- [14] Hecht, Jeff. *Understanding Fiber Optics*. Upper Saddle River, NJ: Pearson/Prentice Hall, 2006. Print.
- [15] "OBEL: Introduction to Optical Coherence Tomography (OCT)." *Optical Biomedical Engineering Laboratory, The University of Western Australia*. Web. 09 Apr. 2012. <<http://obel.ee.uwa.edu.au/research/oct/intro/>>.
- [16] Shane O'Connor, "High Speed Wavelength Tuning of SG-DBR Lasers for Light Detection and Ranging and Optical Coherence Tomography," Electrical Engineering Department, California Polytechnic State University, San Luis Obispo. June 2008.
- [17] "Operation Manual Thorlabs Instrumentation PDB100 Series Balanced Amplified Photodetectors." *Thor Labs*. Web. 15 Oct. 2011. <<http://www.thorlabs.us/Thorcat/12400/12405-D02.pdf>>.
- [18] Pozar, David M. . *Microwave and RF Design of Wireless Systems*. New York: John Wiley and Sons, 2001. Print.
- [19] *Thorlabs OCT Gallery*. Web. 6 Mar. 2011. <<http://www.thorlabs.com/laserimaging/index.cfm?pageref=30&page=OCT-Blood-Vessel>>.

Appendix A – MATLAB Code

MATLAB Main Program

```
clear
clf

%final version of fft simulation that allows for every possible type of
%anomaly to be inserted.

%k is used to scale the total number of points (k = 1 means 50.5k
points)
k = 1.0;

%length of signal (array size) is 50*k points
L = 50500*k;

%"fundamental" sampling time is sampling rate of the oscilloscope (60
MS/s)
ts = 1/(k*60e6);

%"master" time interval vector (for total # of points/array size)
T = (0:L-1)*ts;

%clock and dut angular frequencies. (Roughly 640 sample points in the
real
%data set)
wc = (2*pi)*760000;
wd = wc*(7.1/39);

%~~~~~sources of noise that cannot be removed~~~~~

%mu is average value, sd is standard deviation

%~~detector amplitude noise~~

%random AM detector noise at -83 dBc for unbalanced detector (sd =
0.0025)

mu_det = 0;
sd_det = 0.0025;

r_det = mu_det + sd_det.*randn(L,1);
r_det = r_det';

%random AM detector noise at -81 dBc for balanced detector (sd =
0.00103)

mu_bdet = 0;
sd_bdet = 0.00103;
```

```

r_bdet = mu_bdet + sd_bdet.*randn(L,1);
r_bdet = r_bdet';

%~~laser amplitude noise~~

%random AM laser noise at -73 dBc for unbalanced detector (sd = 0.0078)

mu_las = 0;
sd_las = 0.0078;

r_las = mu_las + sd_las.*randn(L,1);
r_las = r_las';

%random AM laser noise at -81 dBc for balanced detector (sd = 0.0030)

mu_blas = 0;
sd_blas = 0.0030;

r_blas = mu_blas + sd_blas.*randn(L,1);
r_blas = r_blas';

%~~~~~

%Noise sources that can be changed

%values of the quasi-periodic noise sources

%random amplitude offsets (at path transitions), appears on both kclock
and dut signal
mu_amp_path = 1;    %nominally 1
sd_amp_path = 0.0;

r_amp_path = mu_amp_path + sd_amp_path.*randn(L,1);    %change the
dimensions of this?
r_amp_path = r_amp_path';

%random phase offsets (at path transitions), separate for each signal
%kclock signal random phase offsets (radians)
mu_phase_path_k = 0;    %nominally zero
sd_phase_path_k = 0.0;

r_phase_path_k = mu_phase_path_k + sd_phase_path_k.*randn(L,1);
r_phase_path_k = r_phase_path_k';

%DUT signal random phase offsets (radians)
mu_phase_path_d = 0;    %nominally zero
sd_phase_path_d = 0.00;

r_phase_path_d = mu_phase_path_d + sd_phase_path_d.*randn(L,1);
r_phase_path_d = r_phase_path_d';

```

```

%now the constantly varying noise sources

%kclock sinusoidal phase modulation
pm_kclock_mod_index = 0.0;

wpm = (2*pi)*(50e3);

%DUT sinusoidal phase modulation
pm_dut_mod_index = 0.0;

%sinusoidal amplitude modulation (kclock AND DUT)
am_mod_index = 0.0;

wam = (2*pi)*(50e3);

%kclock RANDOM phase modulation
mu_k_phase = 0; %nominally zero
sd_k_phase = 0.00;
%DUT RANDOM phase modulation
mu_d_phase = 0; %nominally zero
sd_d_phase = 0.0;

%NOTE: rest of the phase modulation is in the kclock loop since the
vector for
%the phase noise must be inside the argument of the cosine

%NOTE: in real life, random phase offsets appear on kclock signal and
by a SCALED ammount on
%the DUT signal. If this is true, then the PSF will be unchanged.
%This is the ideal action of the kclock as it provides a reference of
laser
%frequency.

%kclock RANDOM amplitude modulation
mu_ram_k = 0;
sd_ram_k = 0.0; %roughly 0.012 for simulated noise level

r_ram_k = mu_ram_k + sd_ram_k.*randn(L,1);
r_ram_k = r_ram_k';

%DUT RANDOM amplitude modulation

mu_ram_d = 0;
sd_ram_d = 0.015; %roughly 0.015 for simulated noise level

r_ram_d = mu_ram_d + sd_ram_d.*randn(L,1);
r_ram_d = r_ram_d';

%~~~generate the quasi-periodic modulation~~~

```

```

%initial (constant) phase offset for the signal (to adjust relative
phase
%difference between clock and DUT signals). Changing this value does
not
%appear to have an effect on the PSF.
kclock_phase_i = -pi;

%read the phase error file and transpose. "zeroes.txt" means no phase
error is
%introduced at the path transitions (in this part of the program).
phase_error = dlmread('zeros.txt');

%read the time error file, round to nearest integer and transpose
%fm_bm_transitions.txt (low frequency) or phase_transitions.txt (high
frequency)
time_error = k*round(dlmread('phase_transitions.txt'));

%add a zero to the front of the phase error vector (needed for the loop
to function properly)
phase_error = [0 phase_error];

%initialise empty clock vector for loop
kclock = [];

for i=1:length(time_error)+1

%R_p is the ith component of the phase error (from file)
R_p = phase_error(i);

%now, define a new t vector for each iteration so that in the absence
of
%any phase randomness, the clock signal will be continuous (no stitch
errors)

%starting t value
%must be 0 in the case of the first iteration (i = 1)
%second iteration must begin after first has ended, and so on...

%if i is 1, then t_start = 0, otherwise t_start = 1 + A(i-1), one more
than
%the position of the previous error

if i==1
    t_start = 0;
else
    t_start = 1+time_error(i-1);
end

%ending t value
%ends 1 before the next will begin, A(i+1)
%unless it is the final stitch point, then the final point is at L-1

if i <= length(time_error)
    t_end = time_error(i);

```

```

else
    t_end = L-1;
end

%form the time vector for the segment
t = (t_start:t_end)*ts;

%random phase modulation vector
r_k_phase = mu_k_phase + sd_k_phase.*randn(length(t),1);
r_k_phase = r_k_phase';

%compute one segment of the waveform
kclock_s = (r_amp_path(i))*(1 + am_mod_index*sin(wam*t)).*(cos(wc*t +
pm_kclock_mod_index*cos(wpm*t) + r_phase_path_k(i) + R_p + r_k_phase +
kclock_phase_i)+1)-1;

%append to previously calculated arrays
kclock = [kclock kclock_s];
end

%add random noise to the kclock
kclock = (kclock + 1).*(r_ram_k + 1) - 1 + (r_det + r_las);

%now build the dut signal

%initial phase for dut signal
dut_phase_i = 0;

%initialise empty dut vector for loop
dut = [];

for i=1:length(time_error)+1

    %R_p is the ith component of the phase error
    %(use this only if simulating the actual phase transitions as measured)
    R_p = phase_error(i);

    %now, define a new t vector for each iteration so that in the absence
    of
    %any phase randomness, the clock signal will be continuous (no errors)

    %~~starting t value~~

    %must be 0 in the case of the first iteration (i = 1)
    %second iteration must begin after first has ended, and so on...

    %if i is 1, then t_start = 0, otherwise t_start = 1 + time_error(i-1),
    one more than the
    %position of the previous error

    if i==1
        t_start = 0;
    else

```

```

    t_start = 1+time_error(i-1);
end

%ending t value
%ends 1 before the previous will begin, A(i+1)
%unless it is the final stitch point, in which the sinewave continues
until
%the ending time

if i <= length(time_error)
    t_end = time_error(i);
else
    t_end = L-1;
end

%form the time vector for the segment
t = (t_start:t_end)*ts;

%random phase modulation
r_d_phase = mu_d_phase + sd_d_phase.*randn(length(t),1);
r_d_phase = r_d_phase';

%compute one segment
dut_s = (r_amp_path(i))*(1 + am_mod_index*sin(wam*t)).*cos(wd*t +
pm_dut_mod_index*cos(wpm*t) + r_phase_path_d(i) + R_p + r_d_phase +
dut_phase_i);

%append the segment to previously calculated array
dut = [dut dut_s];
end

%add random noise to the dut
dut = dut.*(r_ram_d + 1) + (r_blas + r_bdet);

%power leveling noise
power_leveling_profile = dlmread('better_power_leveling_profile.txt');

kclock = (kclock+1).*power_leveling_profile-1;
dut = dut.*power_leveling_profile;

%filter the kclock with a 5 point, zero-delay, moving-average filter
a = [1 0 0];
b = [1 1 1 1 1]*0.2;
filtered_kclock = filtfilt(b,a,kclock);
filtered_dut = filtfilt(b,a,dut);

%test for number of zero crossings
zero_indices = pcrossingA(filtered_kclock);

%sample the dut at the zero crossing points
sampled_dut = dut(zero_indices);

%digitize the signal (12 bits)
sampled_dut = digitizer(sampled_dut,12);

```

```

%now do the FFT

%create the window function
window = 2*hann(length(sampled_dut));

%scale the dut by the window function (colon forces column vectors)
wdut = window(:).*sampled_dut(:);

%do not zero pad the sampled DUT as it degrades the quality of the
signal.
%NFFT = 2^nextpow2(length(sampled_dut)); <-- zero padding code

NFFT = length(sampled_dut);

%perform FFT on wdut with NFFT steps. Rescale by length.
fdut = fft(wdut,NFFT)/length(sampled_dut);

%"frequency" vector for x axis of FFT plot (in millimeters)
Fs = 21.7*2;
f = Fs/2*linspace(0,1,NFFT/2+1);

%intensity vector in dB scale
H = 20*log10(2*abs(fdut(1:NFFT/2+1)));

plot(f,H)
xlabel('position (mm)')
ylabel('amplitude (dBc)')
title('Simulated PSF, array size of 50,500 points, Hann window')
%axis([0 21.7 -100 0]) %plot from 0 to 21.7 mm, -100 to 0 dB
axis([6.5 9.5 -60 0])

%calculation of the average noise floor that does not include the area
%around the central peak
noise_floor = sum(H(250:end))/(length(H)-250);

```

Zero-Crossing Detector

```
function [indz] = pccrossingA(S)
%finds the positive derivative zero crossings (as index locations) of a
signal S

%first, find all zero crossings

%find exact zeros
ind0 = find(S == 0);

%multiply signal values by previous signal values
S1 = S(1:end-1).* S(2:end);

%find negative values of S1, indicating zero cross
ind1 = find(S1<0);

%concatenate exact and inexact zeros and sort
ind = sort([ind0 ind1]);

%now find positive zeros only

%subtract signal values at subsequent times
Sk = S(ind+1)-S(ind);

%find indices of values that Sk is greater than zero (S is increasing)
indp = Sk > 0;

%positive zero indices of S
indz = ind(indp);

%Now, the previous code finds the points to the LEFT of zero, and not
%the closest point to zero (as is optimal) so we must correct for this

%indz is the index left of zero and indz+1 the index to the right of
zero
indc = S(indz+1)+S(indz);

%negative values means the magnitude of the value to the left of zero
is
%greater than the magnitude of the value to the right of zero
indj = find(indc < 0);

%if so, then increase the index value of the corresponding point
for i=1:length(indj)
    indz(indj(i)) = indz(indj(i))+1;
end
```


Software Digitizer

```
function [d_signal] = digitizer(S,b)
%Digitizes a signal S with b bits.  Works best if signal amplitude is
%[-1,1].

%make max value of signal slightly less than 1 (for rounding purposes,
eps = 2^-52)
S(S>=1)=(1-eps);

%shift signal to Y between [0,2]
S = S + 1;

%scale by 2^(n-1)
Sq = S*2^(b-1);

%floor
Sq = floor(Sq);

%rescale again
Sq = Sq/(2^(b-1));

%subtract 1 to shift to [-1 1] and add 1/2 LSB
Sq = Sq - 1 + 2^(-b);

d_signal = Sq;
```