High-Order Compensation of Self-Phase Modulation in Laser Communication Systems

by Grant Falkenburg

S.B. EE., Massachusetts Institute of Technology (2016)

Submitted to the
Department of Electrical Engineering and Computer Science
In Partial Fulfillment of the Requirements for the Degree of
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Abstract

Self-phase modulation (SPM) creates a power and fiber-length dependent spectral broadening that reduces signal-to-noise ratio in free-space laser communication systems which use a fiber to connect their transmitter’s high-power optical amplifier to its telescope optics. The effects of SPM can be mitigated by using a phase modulator to down-chirp pulses before passing through the receiver’s matched filter. This thesis tests and evaluates a new SPM compensation technique—applying a phase modulation determined from a measurement of the SPM-distorted waveform’s optical intensity—and benchmarks it against sinusoidal phase modulation compensation. The spectra and throughput of the compensated signal are calculated and measured to determine the effectiveness of the new technique. It is found that the two techniques perform within 0.2 dB of each other for fiber lengths less than three times the nonlinear length, and it is expected that the new technique will outperform sinusoidal phase modulation for greater fiber lengths.

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Chapter 1

Introduction

In this thesis, I worked on reducing self-phase modulation (SPM) effects in free-space laser communication (lasercom) systems. I tested two compensation techniques—a standard and a novel approach—to mitigate self-phase modulation under a variety of conditions. The effectiveness of each technique to recover the original laser signal was measured and assessed. Chapter 1 provides the context of self-phase modulation in lasercom systems, the motivation for using compensation, and the scope of the research. Chapter 2 outlines the theory of SPM compensation and analytical basis of its benefit. Chapter 3 explains the design of the experiment to generate and compensate for SPM. Chapter 4 details the experimental procedure for measuring compensation performance. Chapter 5 evaluates the new compensation method based on the measurements. Finally, Chapter 6 concludes the research and proposes future experimental work.

1.1—Background and Motivation

Free space laser communication systems offer a high-efficiency, high bandwidth alternative to traditional radio frequency communication. Advances in component and sub-system technologies for fiber telecom applications can be leveraged to enable more rapid development of reliable and lower cost lasercom systems. Free space, as a communication channel, has minimal nonlinearities and dispersion, and optical signals have wide, unregulated bandwidth. Lasercom terminals can be located on the ground, on an airplane, or in space, making lasercom very versatile. Because lasers have much less beam divergence than radio waves, lasercom power transfer efficiency can be much greater than radio communication. For all these reasons lasercom has proven to be a useful technology and its applications will continue to expand in the future as new developments are made.
Figure 1. Simplified overview of a free-space lasercom system.

The basic architecture of a free-space lasercom system is shown in Fig. 1. A continuous wave (CW) laser source is encoded with data, typically modulated pulses. A high-power optical amplifier (HPOA) increases the power of the laser pulses to overcome power losses in the channel. The amplified modulated light then passes through a telescope that collimates the beam and points it at the receiver. At the receiver another telescope collects the light, and an optical low-noise amplifier (OLNA) pre-amplifies the received light for detection. Because the OLNA generates noise outside of the signal bandwidth, a matched filter is employed to optimally block this noise while transmitting the optical pulses. The detector converts the filtered optical signal to an electrical signal and additional processing decodes the pulses to usable data.

To deliver sufficient signal power through the channel for high data-rate operation, either high powers or large telescopes are needed. Lasercom terminal size constraints often preclude the use of large telescopes, thus requiring higher transmitter powers. However, as higher powers are used, self-phase modulation distorts the spectrum of the optical signal over the fiber coupling the HPOA to the transmitter telescope. SPM manifests itself as an intensity—and therefore time—dependent phase shift of the light, causing chirping of optical pulses [1]. SPM degrades the signal-to-noise ratio (SNR) since SPM-affected pulses have broader bandwidths than their transform-limited counterparts, incurring additional losses at the matched-filter in the receiver. Because of its intensity dependence, SPM makes an average-power limited system also peak-power limited [2]. Broadening the matched-filter bandwidth to accommodate the
SPM-broadened signal would not improve performance because additional noise would also be passed through the broader filter, degrading receiver performance. Chirped matched filters could be used to compensate for the SPM-induced chirp, but they are not readily available.

SPM is proportional to optical power density, distance propagated through an optical fiber, and the fiber’s nonlinear refractive index. In lasercom systems, most SPM effects occur in the fiber between the output stage of the HPOA and the transmitter telescope, where the optical power density is at its peak. Putting the HPOA physically close to the transmitter telescope can minimize this length of fiber and SPM-induced SNR loss, but not without other undesirable system modifications, such as thermal management requirements and increased size, weight, and power of the beam director assembly surrounding the telescope. These undesirable modifications can be minimized by mitigating SPM effects and allowing for greater physical separation between the HPOA and transmit telescope.

Mitigation methods for SPM currently include use of large-mode-area optical fibers, photonic bandgap/microstructure fibers, and negative nonlinear-index materials, but each of these approaches has its own drawbacks [3]. To date, the most promising mitigation approach has been compensation of SPM using an optical phase modulator synchronized to the pulse rate. This technique pre-chirps the pulses with the opposite slope to the SPM-induced chirp, yielding an approximately transform-limited signal incident on the matched filter. The phase modulator can be positioned in the transmitter before the HPOA or in the receiver before the matched filter. In short, the phase modulator effectively undoes the phase modulation of SPM so the original pulse spectrum is recovered at the receiver. If the optical intensity pulse shape is replicated and used to drive the phase modulator, then near-perfect compensation can be achieved [3]. With such compensation in place, many lasercom systems would be relieved of being peak-power limited.
1.2—Previous Work

Several experiments have demonstrated the use of a phase modulator to undo the spectral broadening caused by SPM. Ulmer used single-sinusoid correction of a 2.88 GHz RZ50 waveform (see Section 2.4) to reduce SPM related losses in a laser communication system from 5 dB to 2 dB [3]. Munroe demonstrated a bandwidth reduction from 10 pm to 1.2 pm in 1.5 ns pulses with $\phi_{max}(z) = 5\pi$ (see Section 2.1) by using a phase modulator driven by a pulse generator [4]. Xu, Mollenauer, and Liu used fundamental plus second harmonic correction for a 10 GHz RZ33 waveform (see Section 2.4) with $\phi_{max}(z) = 1.5\pi$ (see Section 2.1), which made the input and output spectra nearly identical [5]. Each of these demonstrations is a precedent for the research reported here. Ulmer’s demonstration concludes that sinusoidal phase modulation works in the context of lasercom and further improvement is possible by using more complex waveforms. Munroe and Xu’s demonstrations show that near-perfect spectral compensation using a phase modulator is possible. Based on these results, it is possible to negate the impairments due to SPM in a lasercom system.

1.3—Research Outline

Since compensation using sinusoidal phase modulation has been demonstrated, the goal for this research is to phase modulate using a higher-order compensation waveform, i.e., a waveform with frequency components beyond the pulse rate, to eliminate SPM-related spectral broadening. Here the compensation waveform is defined as the waveform used to drive the phase modulator. Ideally, the compensation waveform should match the optical intensity profile. The most promising methods of generating a higher-order compensation waveform are as follows: (1) using a saturating amplifier to shape a sinusoid; (2) generating a sinusoid and its harmonics and summing their weighted values; (3) detecting and replicating the optical intensity waveform. Of these, detecting and replicating the optical intensity waveform offers the most potential for improvement and is the easiest to implement. Phase modulation
with a perfect measurement of the optical intensity waveform will theoretically fully negate SPM effects while only requiring a high-bandwidth photodetector to generate the waveform. Thus, we consider detector-based compensation in this research.

This thesis benchmarks detector-based compensation and compares it to sinusoidal compensation. To do this, the power spectral densities and matched filter throughput powers are calculated, measured, and analyzed. We generate SPM-affected signals and compensate them using the two methods under different conditions. Points of concern include the bandwidth requirement of detector-based compensation and the effects of stimulated Brillouin scattering (SBS), another nonlinear optical phenomenon, on the measurements. To account for these concerns, compensation is applied to two systems with different data rates, and for each data rate, compensation is applied pre-fiber and post-fiber. The different data rates test the bandwidth limitations of detector-based compensation. Compensating pre-fiber versus post-fiber causes the amount of SBS to differ, allowing us to isolate the effects of SBS on SPM compensation.
Chapter 2

Theory

To understand how SPM affects a signal and how high-order compensation can mitigate SPM losses, we must develop a mathematical model of SPM and an intuition behind its effects. Section 2.1 presents the differential equation governing SPM and its simplified solution. Sections 2.2 and 2.3 demonstrate SPM’s effects on a Gaussian pulse and how we can compensate for them. Section 2.4 explains the model behind lasercom waveforms. Sections 2.5 and 2.6 calculate how much SNR loss can be recovered for an ideal lasercom waveform and the non-ideal waveform used in this thesis. Section 2.7 considers the impact of other deleterious optical effects on the experimental setup.

2.1—Self-Phase Modulation

The differential equation that describes propagation of light under SPM with negligible dispersion is

$$\frac{\partial U(z,t)}{\partial z} = \frac{i}{L_{NL}} \exp(-az) |U(z,t)|^2 U(z,t),$$

(1)

where \(U(z,t)\) is the dimensionless normalized electric field envelope, \(z\) is the propagation distance through the fiber, \(\alpha\) is the fiber’s attenuation coefficient, \(L_{NL}\) is the nonlinear length \([1]\), and propagation delay has been suppressed. \(\alpha\) is often negligible \((\alpha \approx 0)\) in free-space lasercom due to the short lengths of fiber used, but we will not assume such in this research. \(L_{NL}\) defines the fiber length at which SPM effects become significant and is equal to

$$L_{NL} = (\gamma P_0)^{-1},$$

(2)

where

$$\gamma = \frac{n_2 \omega_0}{c A_{eff}} = \frac{2\pi n_2}{\lambda A_{eff}},$$

(3)
with \( n_2 \) being the nonlinear coefficient of the refractive index, \( A_{eff} \) the fiber’s effective core area, \( P_0 \) the pulse’s peak power, \( c \) the speed of light, and \( \omega_0 \) and \( \lambda \) are the light’s center frequency and wavelength [1]. The solution to the propagation equation that includes the effects of SPM is [1]:

\[
U(z, t) = U(0, t) \exp[i\phi_{NL}(z, t)],
\]

where

\[
\phi_{NL}(z, t) = |U(0, t)|^2 \phi_{max}(z),
\]

\[
\phi_{max}(z) = z_{eff}(z) \gamma P_0 = \frac{z_{eff}(z)}{L_{NL}},
\]

\[
z_{eff}(z) = \frac{1}{\alpha} (1 - \exp(-\alpha z)).
\]

Here \( \phi_{max}(z) \) is the maximum phase shift caused by SPM, hence it is a convenient way to quantify the amount of SPM incurred by a pulse.

### 2.2—SPM on a Gaussian Pulse

To understand the effects of SPM on a signal, we can apply (4) and (5) to a Gaussian pulse and examine the resulting behavior in the time and frequency domains. A normalized Gaussian pulse with an \( e^{-1} \) intensity half-width of \( \Delta T \) is

\[
U(0, t) = \exp\left(- \frac{t^2}{2(\Delta T)^2}\right).
\]

Plugging (8) into (4) and (5), we then get

\[
\phi_{NL}(z, t) = \phi_{max}(z) \exp\left(- \frac{t^2}{(\Delta T)^2}\right),
\]

\[
U(z, t) = \exp\left(- \frac{t^2}{2(\Delta T)^2}\right) + i \phi_{max}(z) \exp\left(- \frac{t^2}{(\Delta T)^2}\right).
\]

Many characteristics of the pulse can be calculated from the time domain response. For example, knowing that
\[ \Delta \omega(z, t) = -\frac{\partial \phi_{NL}(z, t)}{\partial t} \]

is the instantaneous frequency shift created by SPM, we can calculate its maximum value

\[ \Delta \omega_{\text{max}}(z) = \max \left( -\frac{\partial \phi_{NL}(z, t)}{\partial t} \right) = \max \left( \frac{2t}{(\Delta \Gamma)^2} \phi_{\text{max}}(z) \exp \left( -\frac{t^2}{(\Delta \Gamma)^2} \right) \right) \]

\[ \Delta \omega_{\text{max}}(z) = \Delta \omega \left( z, \frac{\Delta \Gamma}{\sqrt{2}} \right) = \frac{\sqrt{2}}{\Delta \Gamma} \phi_{\text{max}}(z) e^{-\frac{1}{2}} \approx 0.86\phi_{\text{max}}(z) \Delta \omega_0 \]  

where \( \Delta \omega_0 \) is the initial \( e^{-1} \) frequency half-width of the Gaussian pulse. While seemingly useful, this instantaneous frequency shift does not provide much insight into the pulse’s frequency spectrum. A more sophisticated calculation by Pinault and Potasek [6] analytically determines the ratio of the SPM root-mean-square bandwidth to the original bandwidth of a Gaussian pulse:

\[ \frac{\Delta \omega_{\text{SPM}}(z)}{\Delta \omega_0} = (1 + \frac{4}{3\sqrt{3}} (\phi_{\text{max}}(z))^2)^{\frac{1}{2}} \approx (1 + (0.88\phi_{\text{max}}(z))^2)^{\frac{1}{2}} \]

For further insight we have plotted in Fig. 2a the time dependence of the real part of the electric field as \( \phi_{\text{max}}(z) \) increases. To calculate the electric field of an optical pulse, we take the real part of the product of the electric field envelope \( U(z, t) \), the optical carrier frequency \( e^{-i\omega_0 t} \), and the peak electric field \( E_0 \):

\[ E(z, t) = \text{Re}\{E_0 U(z, t)e^{-i\omega_0 t}\}. \]  

Here, because our presentation has suppressed propagation delay, Eq. (15) has omitted the space-dependent phase shift \( e^{ik_0z} \) where \( k_0 = \frac{2\pi}{\lambda} \) is the wave number.
Figure 2. Electric field of a Gaussian pulse under SPM. The left panel (2a) shows the electric field of a Gaussian pulse over time for varying $\phi_{\text{max}}(z)$ with $\Delta T = 100$ ps and $\frac{\omega_0}{2\pi} = 20$ GHz. The right panel (2b) shows the corresponding spectrum. We have chosen $\frac{\omega_0}{2\pi} = 20$ GHz for illustration purposes, even though this value is in the microwave rather than the optical spectrum.

Although the Fourier transform of a signal that includes SPM effects generally does not have an analytic expression, a numerical approximation can be made. Figure 2b shows the frequency spectrum for a Gaussian pulse as $\phi_{\text{max}}(z)$ increases. From Fig. 2a, we see that both the intensity envelope and the number of peaks remain constant in the time domain. However, in the frequency domain, Fig. 2b shows that both frequency bandwidth and the number of peaks increase linearly with $\phi_{\text{max}}(z)$. It can be analytically shown [1] that for large $\phi_{\text{max}}(z)$ the number of frequency domain peaks, $M$, is approximately

$$M(z) \approx \frac{\phi_{\text{max}}(z)}{\pi} + \frac{1}{2}. \quad (16)$$

From our numerical expression for the electric field, we can calculate the response of the pulse through a matched filter at various $\phi_{\text{max}}(z)$ values. The matched filter has an impulse response envelope of $H_0(t) = U^*(0, -t)$ to optimally detect the SPM-unaffected pulse in the presence of additive Gaussian white noise, where $U^*(0, -t)$ denotes the complex conjugate of $U(0, -t)$. The effective SPM-induced SNR loss is approximately the pulse’s peak power reduction through the filter, since the noise is assumed to be unaffected by SPM. To calculate the peak power reduction for $\phi_{\text{max}}(z) > 0$, we convolve $H_0(t)$
with \( U(z, t) \), evaluate at \( t = 0 \), take the squared magnitude of this quantity to model the photodetection process, and normalize it to the value at \( \phi_{\text{max}} = 0 \). The result of this calculation for a Gaussian pulse is shown in Fig. 3. The SPM penalty increases significantly for \( \phi_{\text{max}}(z) > 2 \), at which point an increase in signal power provides diminishing benefits to the SNR.

![Match for a Gaussian Pulse under SPM](image)

Figure 3. Filter match of a Gaussian pulse under SPM.

### 2.3—SPM Compensation of a Gaussian Pulse

One way to compensate for SPM in a Gaussian pulse is to linearly down-chirp the pulse. A linear chirp is a linear variation of the instantaneous frequency over time, and thus the phase contribution of a linear chirp \( \phi_c \) is quadratic in time

\[
\phi_c(t) = k_c t^2
\]  

(17)

where \( k_c \) is the chirp rate in radians per second squared. For each \( \phi_{\text{max}}(z) \), there is a value of \( k_c \) that optimizes pulse transmission through the matched filter. Since this value of \( k_c \) is a function of \( \phi_{\text{max}}(z) \), we’ll call the optimal chirp rate \( k_c(\phi_{\text{max}}) \). One way to approximate \( k_c(\phi_{\text{max}}) \) is by taking the Taylor series of \( \phi_{NL}(z, t) \) and cancelling the \( t^2 \) term with \( k_c \):

\[
\phi_{NL}(t) + \phi_c(t) = \phi_{\text{max}}(z)(1 - \left( \frac{t}{\Delta T} \right)^2 + \frac{1}{3} \left( \frac{t}{\Delta T} \right)^4 \cdots) + k_c t^2
\]  

(18)

\[
k_c(\phi_{\text{max}}) \approx \frac{\phi_{\text{max}}(z)}{\Delta T^2}.
\]  

(19)
This approximation reasonably predicts the value of \( k_c(\phi_{max}) \) that maximizes the matched filter response. To improve the result, we need to calculate the matched filter response over a range of \( k_c \) and \( \phi_{max}(z) \) values and then determine the value of \( k_c \) that locally maximizes the filter response for each \( \phi_{max}(z) \). Figure 4a plots the matched filter response for \( k_c \in [0, \frac{10}{\Delta T^2}] \) and \( \phi_{max}(z) \in [0,10] \), and Fig. 4b plots \( k_c(\phi_{max}) \) with the corresponding filter match.

Figure 4. Filter match of a chirped Gaussian pulse under SPM. The left panel (4a) shows the filter match for varying \( k_c \) and \( \phi_{max}(z) \). The right panel (4b) plots the maximum of the filter match as a function of \( \phi_{max}(z) \) and the corresponding \( k_c(\phi_{max}) \).

From Fig. 4b, we can see that the optimal amount of chirp is about half of our estimate \( \frac{\phi_{max}(z)}{\Delta T^2} \). With linear chirp compensation, the SNR loss at \( \phi_{max}(z) = 10 \) was reduced to 1.7 dB from 9.6 dB. A linear chirp corrects for most of the SPM-related filter loss in Gaussian pulses but raises the question of how much correction is possible. Noting that the SPM phase shift is proportional to the intensity of the optical signal, perfect compensation can be achieved if the optical signal is oppositely phase-modulated by the scaled intensity of the optical pulse. This methodology is the reason for using the photodetected intensity profile of the optical pulse to drive a phase modulator—it will theoretically fully undo SPM effects.
2.4—RZ Waveforms and Mach-Zehnder Modulators

In lasercom systems, Gaussian pulses are rarely used. Instead, return-to-zero (RZ) waveforms are used, which are carved from a CW laser source with a Mach-Zehnder intensity modulator (MZM). An RZ waveform returns to zero in between each communication symbol, and has the benefit of reduced inter-symbol interference and lower bit-error rates over non-return to zero waveforms [7]. Common RZ waveforms are depicted in the top right of Fig. 5.

![MZM Transfer Function and RZ Waveforms](image)

Figure 5. MZM transfer function and RZ waveforms. Bottom left – Drive voltage over time for common RZ waveforms. Top left – Voltage to intensity transfer function of an MZM. Top right – Common RZ intensity waveforms [2].
An MZM modulates intensity by splitting an optical signal into two paths, applying phase modulation to one or both paths, and combining the two paths as shown in Fig. 6. A single-drive modulator only applies phase modulation to one path, whereas a dual-drive modulator applies a phase modulation to both paths. The phase difference between the two paths causes the electric field to constructively or destructively interfere at the MZM’s output beam splitter and thus modulate the output intensity. As a result, the input voltage $V_{MZM}$ to output electric field attenuation $H_{MZM}$ relation is approximately sinusoidal [2]. For a dual-drive modulator, the transfer function is generally of the form

$$H_{MZM,DD}(V_{MZM}) = \sqrt{R_1 T_2} e^{j\left(\frac{nV_{MZM}}{2v_\pi} + \theta_{os1}\right)} + \sqrt{T_1 R_2} e^{j\left(-\frac{nV_{MZM}}{2v_\pi} + \theta_{os2}\right)}, \quad (20)$$

where $R_1$, $T_1$ and $R_2$, $T_2$ are the reflection and transmission coefficients of the input and output beam-splitters, $v_\pi$ is the voltage needed to go from an intensity peak to a null, and $\theta_{os}$ is the offset phase contributed by each path of the MZM. For a single-drive modulator, the transfer function is generally

$$H_{MZM,SD}(V_{MZM}) = \sqrt{R_1 T_2} e^{j\left(\frac{nV_{MZM}}{v_\pi} + \theta_{os1}\right)} + \sqrt{T_1 R_2} e^{j\theta_{os2}}, \quad (21)$$

when the phase modulator is on the top path of the MZM. When the beam splitters are perfectly balanced and the offset phases are equal, Eq. (20) simplifies to

$$H_{MZM,DD}(V_{MZM}) \approx \cos\left(\frac{nV_{MZM}}{2v_\pi}\right) e^{j\theta_{os}} \quad (22)$$

for a dual-drive modulator and Eq. (21) simplifies to
\[ H_{MZM,SD}(V_{MZM}) = \cos(\frac{\pi V_{MZM}}{2V_\pi})e^{i\theta_{\text{os}} + \frac{\pi V_{MZM}}{2V_\pi}} \] (23)

for a single-drive modulator. Taking the squared magnitude of either of these expressions gives us the intensity transfer function

\[ |H_{MZM}(V_{MZM})|^2 \approx \cos^2(\frac{\pi V_{MZM}}{2V_\pi}) \] (24)

as shown in the top left panel of Fig. 5. The RZ waveforms are generated by driving the MZM with a sinusoidal voltage, whose amplitude and bias determine the type of waveform obtained, as shown in the bottom left panel of Fig. 5.

In an actual MZM, a few non-idealities need to be considered when evaluating the performance of a lasercom system, including extinction ratio, chirp, and phase drift. Extinction ratio describes the MZM’s ability to fully null a signal and is equal to the minimum intensity transmission divided by the maximum intensity transmission of the MZM. Extinction ratio affects the shape of the RZ waveform by increasing the fundamental frequency component and decreasing the harmonic components, and is determined by the MZM’s reflection and transmission coefficients. Chirp is always added to a signal in a single-drive MZM, and can also be present in a dual-drive MZM if its extinction ratio is poor. While chirp does not affect the intensity profile of an RZ waveform, it affects the frequency spectrum and thus the power of the waveform through a matched filter. Phase drift occurs due to thermal gradients and charge buildup, and causes the offset phases to vary slowly [2]. Phase drift changes the effective bias point of the input sinusoid, thus the RZ intensity waveform must be monitored to ensure its quality.

While each RZ waveform in Fig. 5 has its pros and cons, there is one waveform of particular interest for use in lasercom. The RZ50 waveform requires a smaller drive voltage than RZ33 and RZ66, and is less sensitive to drive amplitude and bias variations, making MZM drift not as harmful. Additionally, there is no chirp penalty caused by crossing a transmission null [2]. While the RZ50 waveform does not have the best match with a Gaussian filter (see Section 2.5), these benefits make the
RZ50 a more practical waveform to implement. Because of the RZ50’s widespread use, we will perform our calculations and measurements using RZ50 waveforms.

Once an RZ waveform is generated, data can be encoded in a variety of formats. For example, in an on-off-keyed modulation scheme, the presence of an RZ pulse represents a 1 and the absence of a pulse represents a 0. On-off keying is simple to implement but does not support power-efficient communication. Lasercom typically employs differential phase-shift keyed modulation, where the carrier wave within each pulse is phase modulated by an amount corresponding to the symbol to be sent, and a delay interferometer demodulates the signal by comparing the phases of two consecutive pulses. In this experiment, we do not modulate the optical pulses, so details of modulation formats will not be discussed further.

2.5—Ideal SPM Compensation in Laser Communication

Compensating SPM for an RZ waveform is similar to that for a Gaussian pulse. However, since RZ waveforms are periodic we need to apply a periodic compensation waveform, the simplest of which is a sine wave. A sine wave corrects for most of the SNR loss, but there is still room for improvement. On the other hand, driving a phase modulator with a perfectly reconstructed RZ waveform would correct for all of the SPM, but is difficult to achieve due to the requirements for wide-bandwidth RF signals. A good compromise is to include the fundamental and a few harmonics of the intensity waveform. This can be accomplished several different ways, as discussed in Section 1.3, but we are most interested in detecting the optical intensity.

Ideally, we would employ a matched filter that has an impulse response of the RZ waveform used, but this is not possible to implement. A Gaussian filter is well matched with an RZ waveform and is commonly available; we use this filter to evaluate the SPM penalty for an RZ waveform. Because the Gaussian filter is not perfectly matched to the RZ waveform, an additional mismatch penalty occurs. This
mismatch is on the order of 0.1 dB and is omitted in our calculations since we are only interested in the penalty due to SPM [2]. The ratio \( \beta \) of the intensity waveform’s full width at half-maximum (FWHM) to the filter’s squared impulse response FWHM can be calculated using the optimal matched filter condition. A table containing \( \beta \) for common RZ waveforms is provided in [2]. Knowing \( \beta \) and the Gaussian filter’s squared-frequency-response FWHM \( f_{FWHM} \), we can calculate the RZ waveform’s frequency to be

\[
f_0 = f_{FWHM} \frac{\pi r_{RZ}}{2 \ln(2) \beta},
\]

where \( r_{RZ} \) is the unitless ratio of the RZ intensity waveform’s FWHM to the waveform’s period.

We face a difficulty when measuring the SPM penalty in practice. A high-bandwidth photodetector and oscilloscope are required to measure the filtered RZ waveform’s peak power (the signal power). In the context of this thesis, it was more practical to measure the average power of the filtered RZ waveform since a standard optical power meter is sufficient to make this measurement. Figure 7 demonstrates that the average power loss and peak power loss of a filtered RZ50 waveform are well correlated for \( \phi_{max}(z) < 5 \), making the average power measurement a reasonable approximation for determining the peak power and corresponding SPM-induced SNR penalty. For a compensated RZ50 waveform, the average and peak power losses are nearly equivalent, implying that peak power loss can be estimated from the average power loss.

The peak and average power of an RZ waveform through a filter can be calculated in the time or frequency domains. The corresponding measurements for these calculations are more practical to make in the frequency domain because an optical spectrum analyzer’s bandwidth is much broader than that of a photodetector. Additionally, the electric field’s phase and the filter’s impulse response cannot be measured easily. Therefore, we will perform the calculations in the frequency domain. The peak power at the matched filter’s output is proportional to

\[
P_0 \propto \left| \int_{-\infty}^{\infty} E_0 \cdot \tilde{U}(z, \omega) \cdot \tilde{H}(\omega) \, d\omega \right|^2,
\]

where...
where $\tilde{U}(z, \omega)$ is the Fourier transform of $U(z, t)$, and $\tilde{H}_0(\omega)$ is the Fourier transform of $H_0(t)$. The average power is proportional to

$$\langle P(t) \rangle \propto \int_{-\infty}^{\infty} |E_0 \tilde{U}(z, \omega)\tilde{H}_0(\omega)|^2 d\omega. \quad (27)$$

Note that since $U(z, t)$ is periodic, the peak and average powers can be expressed in terms of discrete summations rather than integrals.

To estimate the amount of phase modulation required for compensating SPM in an RZ waveform, we follow a procedure similar to that used for the chirped Gaussian pulse. Instead of taking a Taylor series of the intensity waveform, we decompose the normalized RZ intensity waveform into its Fourier series. This returns the amplitudes and phases of the frequency components of SPM that we wish to cancel. The peak power is then maximized by varying the amplitudes and phases of the compensating waveform around the estimates, as with the chirped Gaussian pulse.

We evaluated the SPM-induced power loss of an RZ50 waveform transmitted through a Gaussian matched filter with sinusoidal compensation and sinusoidal plus third harmonic compensation. An RZ50 waveform is half-wave symmetric after removing the zero frequency offset, i.e., a shift of the waveform by half a period is equal to the negative waveform, thus the RZ50 intensity waveform does not have even harmonics and does not require second harmonic compensation.
Figure 7. SNR penalty for a RZ50 pulse waveform through a Gaussian filter under SPM. Two sinusoidal compensation methods are compared: fundamental and fundamental plus third harmonic. Additionally, the peak power is compared to the average power of the waveform through the filter.

Figure 7 shows the results of the calculation for an RZ50 waveform, indicating that for low $\phi_{\text{max}}(z)$ a single sinusoid provides excellent compensation and a sinusoid plus its third harmonic provides nearly perfect compensation. This analysis shows that when doing detector-based compensation, the bandwidth of the photodetector should be at least three times the fundamental frequency. With this condition given, we expect that near perfect compensation can be achieved.

### 2.6—SPM Compensation of Non-Ideal RZ Waveforms

The calculations we have performed in Section 2.5 assume that an ideal MZM transfer function is used to generate the RZ waveform. While this assumption holds true for a dual-drive modulator with an excellent extinction ratio, the MZM used in this thesis is single-drive with a non-negligible extinction ratio. For convenient analysis of the non-ideal RZ waveform, we can rewrite Eq. (21) to be

$$H_{\text{MZM,SD}}(V_{\text{MZM}}) = \sqrt{A}(\sqrt{B}e^{j \left(\frac{\pi V_{\text{MZM}}}{V_{\pi}}\right)} + \sqrt{1-B})$$

(28)
where $A$ is a scaling factor and $B$ is a balance factor such that $A = R_1 T_2 + T_1 R_2$, $B = \frac{R_1 T_2}{R_1 T_2 + T_1 R_2}$ and the offset phases are zero. We can now analyze the transfer function using these factors. The squared magnitude and phase of the transfer function is equal to

$$|H_{MZM,SD}(V_{MZM})|^2 = A(1 + 2\sqrt{B(1-B)} \cos \left( \frac{\pi V_{MZM}}{V_{\pi}} \right)),$$  

(29)

$$\angle H_{MZM,SD}(V_{MZM}) = \tan^{-1} \left( \frac{\sqrt{B} \sin \left( \frac{\pi V_{MZM}}{V_{\pi}} \right)}{\sqrt{B} \cos \left( \frac{\pi V_{MZM}}{V_{\pi}} \right) + \sqrt{1-B}} \right).$$

(30)

While we cannot measure $R_1$, $T_1$, $R_2$ or $T_2$ directly, we can calculate $B$ from the extinction ratio, which can be measured or read from the MZM’s datasheet. From Eq. (29), we can derive

$$B = \frac{1}{2} + \frac{\sqrt{ER}}{1+ER},$$

(31)

where $ER$ is the extinction ratio. Because there are two possible balance factors for a given extinction ratio, knowing only the extinction ratio is not enough information to determine the expression for the RZ waveform generated. These two balance factors generate the same RZ intensity waveform, but have different phase contributions due to the asymmetry of $B$ and $(1-B)$ in Eq. (30).

There is an additional complication when generating RZ50 and RZ40 waveforms using a single-drive MZM. Figure 5 depicts the bias point of the RZ50’s input sinusoid to be $0.25 V_{\pi}$, but the bias point can also be $-0.25 V_{\pi}$. These two bias points again have equal intensity characteristics but opposite phase responses, which can be seen if $V_{MZM}$ is made negative in Eqs. (29) and (30). The two bias points combined with the two balance factors suggest that an RZ50 waveform can have any of four phases with our model of an MZM. For an extinction ratio of -13 dB, the typical value listed in the datasheet of the MZM used in this thesis, and $A = 0.5$, the magnitude and four possible phases of an RZ50 waveform are displayed in Fig. 8.
Figure 8. Non-ideal RZ50 magnitude and phase. The waveform can have any of four phases. The waveform here has a -13 dB extinction ratio.

Each of these four RZ50 waveforms has a different response through a matched filter, and thus not knowing which waveform is used when taking measurements can lead to erroneous results. The type of waveform used can be determined by measuring the filter match under SPM. Whether $B$ is greater than 0.5 can be determined by the power spectral density (PSD), although the PSD cannot determine whether the overall phase shift is positive or negative as shown in Fig. 9. If the waveform has a positive phase, the phase contribution partially cancels SPM and makes the waveform pre-compensated, and vice-versa if the phase is negative. This effect is displayed in the top row of Fig. 10. As with the ideal RZ50 waveform, the filtered SPM-induced waveform’s peak power is generally greater than the same waveform’s average power, making the average power loss a reasonable lower-bound for the SNR penalty. The middle and bottom rows of Fig. 10 show the power loss of a non-ideal RZ50 waveform with sinusoidal compensation and bandlimited detector-based compensation that includes the fundamental and third harmonic but none others to emulate the experimentally detected waveform. Note that the peak power and average power in Fig. 10 are normalized to the ideal RZ50 no-SPM filter match, and some of the non-ideal RZ50 waveforms are better matched to a Gaussian filter than the ideal RZ50. The normalization is done so the filter matches for each non-ideal RZ50 can be compared. From Fig. 7 and the middle and bottom rows of Fig. 10, we observe that the power losses of the
compensated non-ideal RZ50 waveforms are similar to the power losses of a compensated ideal RZ50. This observation is important because we are most interested in how well compensation will improve an ideal RZ50 under SPM. While we cannot directly calculate the compensation improvement of an ideal RZ50 from non-ideal RZ50 measurements, the similarity between the two waveforms’ power losses imply that the power loss of the ideal RZ50 should be comparable to the power loss of a non-ideal RZ50 in practice.

![Non-Ideal RZ50 Frequency Spectrum](image.png)

Figure 9. Non-ideal RZ50 power spectral density. Here a stem plot is used to represent impulses in the spectrum.
Figure 10. SNR penalty for a non-ideal RZ50 waveform. Top row—No compensation is applied. Middle row—Sinusoidal compensation is applied. Bottom row—Bandlimited detector-based compensation is applied. All waveforms are normalized to the ideal RZ50 no-SPM filter match.
2.7—Other Optical Effects

Self-phase modulation is not the only deleterious optical effect that occurs in laser communication systems. Group velocity dispersion is a frequency-dependent phase shift that is exacerbated by SPM due to SPM’s frequency broadening. Dispersion is negligible relative to SPM when the nonlinear length $L_{NL}$ is much less than the dispersion length $L_D$, which is equal to

$$L_D = \frac{T_0^2}{|\beta_2|},$$

(32)

where $T_0$ is the $\frac{1}{e}$ half-width time of the pulse, and $\beta_2$ is the dispersion parameter of the fiber [1]. For telecom-standard fiber, such as Corning SMF-28, $\beta_2 = 20.7 \text{ ps}^2/\text{km}$, and for 5 GHz RZ50 pulses, $T_0 = 42$ ps, making $L_D = 85$ km. On the other hand, $\gamma = 1.43 \text{ rad W}^{-1} \text{km}$ in Corning SMF-28 fiber, and for a peak power of $P_0 = 0.1 \text{ W}$, $L_{NL} = 7$ km. With these conservative pulse rates and powers, we can safely ignore the effects of dispersion on a SPM-affected signal for the purpose of this research.

Stimulated Brillouin scattering (SBS) is a phenomenon in which fiber-coupled light is reflected backward because of interactions between the optical signal and acoustic modes of the fiber material. Forward-propagating light generates acoustic waves due to electrostriction, a process by which an electric field deforms a material. These acoustic waves form a Bragg diffraction grating, reflecting light toward the source. Because the acoustic wave is propagating, the reflected light undergoes a Doppler shift of about 10 GHz with a linewidth of about 10 MHz [1]. At sufficient power, the majority of light propagates back toward the source. The power threshold for the onset of SBS is

$$P_{0,SBS} \approx \frac{21\lambda_{eff}}{L_{eff} \beta_B},$$

(33)
where $g_B$ is the Brillouin gain, and $L_{\text{eff}}$ is the effective interaction length equal to $z_{\text{eff}}(L)$ as defined in Eq. (7), where $L$ is the length of the fiber [1]. In standard silica fiber, $g_B \approx 5 \times 10^{-11} \frac{\text{m}}{\text{W}}$, and for Corning SMF-28 fiber, $A_{\text{eff}} = 85 \, \mu\text{m}^2$, making the SBS length threshold $L_{\text{eff}} \approx 0.4 \text{km}$ for $P_0 = 0.1 \, \text{W}$.

From these calculations, we expect that SBS will corrupt the signal before the onset of SPM. Fortunately, this equation only holds true for narrowband light. If the propagating light has a linewidth greater than the Brillouin linewidth, SBS is suppressed significantly. Most lasercom systems operate at bandwidths far greater than 10 MHz, so the data modulation in these systems broadens the spectrum enough to suppress SBS. In a system with no data modulation, such as one used in this research, phase modulation can be applied to broaden the spectrum beyond the Brillouin linewidth. One may assume that the use of an RZ waveform broadens the spectrum sufficiently to suppress SBS, but doing so generates harmonics in the frequency spectrum of the signal without increasing the linewidth of each frequency component. Hence, using an RZ waveform alone reduces but does not completely suppress SBS.

Similar to SBS, stimulated Raman scattering (SRS) is a process in which light interacts with optical phonons and generates redshifted light that propagates forward. Spontaneous Raman scattering generates small amounts of redshifted light that the optical signal amplifies, eventually depleting the optical signal’s power. The peak power threshold at which the majority of the signal’s power becomes redshifted is approximately

$$P_{0,SRS} \approx \frac{16A_{\text{eff}}}{L_{\text{eff}}g_R},$$  \hspace{1cm} (34)

where $g_R$ is the Raman gain coefficient [1]. Fortunately, $g_R \approx 10^{-13} \frac{\text{m}}{\text{W}}$ in silica fiber, making the SRS length threshold $L_{\text{eff}} \approx 136 \text{ km}$ when $P_0 = 0.1 \, \text{W}$ and $A_{\text{eff}} = 85 \, \mu\text{m}^2$ as before. We can conclude that SRS is negligible in the context of this thesis.
Chapter 3

Experimental Design

To fulfill our objective of demonstrating a new technique for SPM compensation, we need a system capable of generating SPM, applying compensation, and measuring spectra and matched filter throughput using readily available parts while suppressing SBS and minimizing dispersion. The calculations in Section 2.7 provide a set of guidelines for parts selection—the HPOA should be able to generate pulses with at least 0.1 W peak power, the SPM-producing fiber must be longer than 7 km and shorter than 85 km, the pulse rate must be less than 5 GHz, and a phase modulator must be available to suppress SBS. A few additional constraints are that the laser source’s center frequency must equal that of the matched filter, and the pulse rate must be tuned to match the matched filter’s bandwidth. Section 3.1 explains the system used to generate SPM and the sections that follow cover the modifications needed for SPM compensation and some variations on which compensation will be tested.

3.1—SPM Generation

Figure 11. System used to generate and measure SPM.
Figure 11 shows a block diagram of a system to generate and evaluate SPM. A laser source generates the optical signal at the carrier frequency required for the system’s operation. The laser source should generally have stable output power and wavelength with a narrow linewidth; a distributed-feedback (DFB) laser is suitable for this application. DFB lasers are typically discrete components that require a laser diode driver with a current and temperature controller for stable operation. The temperature controller interfaces with an on-chip thermistor and thermoelectric cooler to sense and adjust the laser’s temperature. Adjusting the temperature will coarsely tune the laser’s frequency, and adjusting the current changes the laser’s power and frequency. With this in mind, we can align the laser’s frequency with the matched filter’s center frequency.

The CW laser source is connected to an MZM, which carves out an RZ waveform. The MZM is driven by a sine wave generated by the amplified output of an RF tone synthesizer. A DC voltage adjusts the bias point of the sine wave on the MZM transfer function. The RF amplifier used in this experiment has a built-in bias tee to set the MZM bias point. The DC voltage is provided by a standard adjustable power supply. The output of the RF synthesizer is split for triggering the oscilloscope with a waveform locked to the synthesizer frequency.

Once the RZ waveform has been generated, a phase modulator is used to suppress SBS that would otherwise occur in the fiber. This phase modulator is driven by a standard signal generator, whose output amplitude, frequency, and waveform can be adjusted to best suppress SBS. The optical signal is then amplified by an erbium-doped fiber amplifier (EDFA), boosting the power of the pulses to reach a peak of at least 0.1 W. EDFA’s often have adjustable output power, making it convenient to control the pulse’s peak power rather than adjusting the DFB laser’s current, which would also affect the carrier frequency.

A circulator is added to the output of the EDFA to prevent backward-propagating light from entering and being amplified by the EDFA. A circulator is a three-port optical device that passes the input.
of port 1 to the output of port 2, and the input of port 2 to the output of port 3. The circulator serves the secondary function of allowing us to measure the amount of back-reflected SBS.

Once the high-power pulse waveform has been generated, SPM is produced by passing the waveform through several kilometers of fiber. A standard part that fits the constraints is a 20-km-long reel of SMF-28 fiber with connectors spliced on the ends. With an attenuation of $\frac{0.18\, \text{dB}}{\text{km}}$, we can convert this to $\alpha = \frac{0.18 \ln(10)}{10} = 0.041\, \text{km}^{-1}$, making $L_{\text{eff}} = 13.6\, \text{km}$. We expect to see $\phi_{\max}(L) \approx 2$ with this configuration.

The SPM-affected signal at the output of the fiber reel is split to simultaneously observe the spectrum on an optical spectrum analyzer (OSA) and pass the signal through the matched filter. An OSA sweeps a narrowband filter and measures the power transmitted through it, allowing us to obtain the power spectral density of the SPM-affected signal. The frequency response of this narrowband filter can be measured by sweeping the OSA with a CW laser as the input. Using that filter response measurement, we can obtain a high-resolution spectrum by deconvolving an OSA-measured spectrum with the narrowband filter’s frequency response.

The matched filter is a fiber-Bragg grating (FBG) with a Gaussian frequency response. The FBG must have a bandwidth that optimally matches the pulse rate as described in Section 2.5. The power transmitted through the matched filter can be measured with an optical power meter, allowing us to fully characterize the SPM-power penalty. The FBG reflects the frequency band of interest, so a circulator is used to isolate the input signal from the desired reflected output. Because the circulator is essential to the functionality of the FBG, it is implied that a circulator is included with every FBG used in this experiment.
3.2—Sinusoidal Compensation

Sinusoidal compensation is the simplest type of phase modulator-based SPM compensation, and will serve as the reference for the performance of detector-based compensation. Figure 12 shows a block diagram of the overall system used to evaluate the performance of sinusoidal SPM compensation. We use an RF splitter with the synthesizer to obtain the extra sinusoid signal for phase modulation. The output level of the RF synthesizer is adjusted to compensate for the loss incurred by the splitter so that the input amplitude to the MZM remains the same.

The phase of the sinusoid to be used for SPM compensation must be phase matched to the pulse shape. This matching is accomplished with a variable delay line. Without this phase matching, the sinusoidal phase modulation could exacerbate, rather than compensate, SPM effects. Male-to-female SMA port savers were used to provide a fixed phase shift when the variable delay line reached its limit. An RF amplifier amplifies the signal for driving the compensator’s phase modulator. The amount of phase modulation can be adjusted by changing the gain of the RF amplifier. For the RF amplifier used in this experiment, the gain was adjusted by applying a voltage to the gain control pin of the device. However, this gain control comes with a drawback—reducing the gain tends to saturate the amplifier’s...
output. Care must be taken to provide a signal of proper amplitude to the amplifier so that minimal saturation occurs.

### 3.3—Detector-Based Compensation

Figure 13. System used to apply sinusoidal or detector-based compensation. Components used specifically for detector-based compensation are darkened, and X’s are used to mark alternate connections to components.

Figure 13 shows an overview of the system used to generate SPM and apply either sinusoidal or detector-based compensation. To perform detector-based compensation, we first need a copy of the RZ waveform to measure. A beam splitter is inserted after the MZM to take 5% of the RZ waveform’s power for measuring, passing 95% to the rest of the system. Using a beam splitter of this ratio allows a sufficient amount of power to reach the high-power EDFA, because EDFAs require a certain amount of input power to work properly. The only downside is that 5% of the signal does not provide sufficient voltage to drive the phase modulator. Therefore, an additional optical amplifier must be inserted after the 5% split. A subsidiary EDFA provides optical amplification and additionally serves as a power control for the compensating waveform. With the EDFA in place to control the power, the RF amplifier can be kept at maximum gain to reduce saturation effects.

Following amplification by the EDFA, the optical RZ signal is converted to an electrical intensity waveform by a photodetector. The photodetector used in this experiment has an 11 GHz bandwidth,
making it usable for second and third harmonic compensation on waveforms up to 3.7 GHz. Now that we have the compensating waveform, we can pass it through the same components as we did for sinusoidal compensation—a variable delay and RF amplifier—before driving the phase modulator. With this system in place, we can easily switch between detector-based compensation and sinusoidal compensation by connecting either the split RF tone or the photodetector to the variable delay. When not being used for compensation, the measured intensity can be displayed on an oscilloscope to inspect the RZ intensity waveform.

3.4—Post-Fiber Compensation

To this point we have discussed our setups for sinusoidal and detector-based compensation as pre-compensation techniques, i.e., compensation applied before the signal propagates through the fiber. Thus, Figs. 12 and 13 model how compensation would be applied in a lasercom transmitter. However, because of the complicated interaction between SPM and SBS, it is beneficial to consider post-fiber compensation. Pre-fiber compensation changes the frequency distribution of the signal before it passes through the fiber, which means that the amount of SBS changes depending on the amount and type of compensation applied. This difference makes it difficult to compare compensation methods because SBS’s effect on each compensation method is not obvious. Furthermore, the interaction between SBS and SPM with pre-fiber compensation is chaotic. In particular, spectral peaks tend to be reflected due to SBS, reducing the peak intensity and amount of SPM. This reduction of SPM means too much compensation is applied, broadening the signal spectrum. The broadened spectrum reduces SBS, increasing the forward-propagating power and the amount of SPM. An increase in the amount of SPM then narrows the spectrum as the correct amount of compensation is applied, causing the cycle to repeat. All of this occurs in an unstable equilibrium when applying pre-fiber compensation, resulting in power fluctuations.

Post-fiber compensation avoids the preceding issue because changing the amount of compensation does not change the frequency spectrum in the fiber, keeping SBS constant regardless of
the type of compensation used. Moreover, SPM without pre-compensation is shown to increase the SBS threshold due to its spectral broadening [8]. One may wonder why pre-fiber compensation would cause any more SBS than no compensation. The difference is that in pre-fiber compensation, the spectrum starts broadened and becomes narrowed as SPM is applied through the length of the fiber. The narrow spectrum causes higher amounts of spontaneous Brillouin scattering at the far end of the fiber that gets amplified through the length of the fiber.

![Figure 14. Post-fiber compensation system. Components modified from the pre-compensation system are darkened.](image)

Figure 14 shows a block diagram indicating how we implemented post-fiber compensation. Post-fiber compensation requires one component in addition to the pre-compensation system. The phase modulator requires a polarized input, and the 20 km reel of fiber randomizes the signal’s polarization. Thus, a polarization stabilizer is needed to align the randomly-polarized signal with the polarization needed by the phase modulator. A conventional polarizer would also solve this issue, but is lossy and would cause power variations with changing polarization states. The polarization stabilizer is positioned after the reel of fiber, and the SPM-compensating phase modulator is positioned after the polarization stabilizer. After compensation is applied, the spectrum and filter match are measured as done for our pre-fiber compensation techniques.
3.5—Low-Rate Compensation

The pre-fiber and post-fiber compensation systems are sufficient for testing sinusoidal and detector-based compensation, but one of the primary concerns with detector-based compensation is whether the photodetector, RF amplifier, and phase modulator have sufficient bandwidth to replicate the higher-order harmonics. Testing compensation with a lower pulse rate would relax the bandwidth limits on the devices. The rate of the entire system can be adjusted by changing the RF synthesizer’s frequency and changing the FBG to match the pulse rate. The main constraint when adjusting the pulse rate is finding an FBG that matches the new lower rate.

Fortunately, because FBG filters are Gaussian, a cascade of two filters is also Gaussian and can be used as a single matched filter. Thus, as long as the center wavelengths of the two FBG filters are reasonably well matched, we can cascade two FBG filters to obtain a matched filter at a lower pulse rate that is not excessively lossy. The equivalent center frequency (which can be converted to a center wavelength) and bandwidth of the cascaded filter is

\[
\mu_{eq} = \frac{\mu_1 \sigma_1^2 + \mu_2 \sigma_2^2}{\sigma_1^2 + \sigma_2^2},
\]

\[
\sigma_{eq}^2 = \frac{\sigma_1^2 \sigma_2^2}{\sigma_1^2 + \sigma_2^2},
\]

where \( \mu_1 \) and \( \mu_2 \) are the individual filters’ center frequencies, and \( \sigma_1^2 \) and \( \sigma_2^2 \) are the corresponding squared bandwidths. Additionally, there is a maximum electric field transmissivity of the equivalent filter, equal to

\[
a_{eq} = a_1 a_2 \exp\left(-\frac{(\mu_1 - \mu_2)^2}{2(\sigma_1^2 + \sigma_2^2)}\right),
\]

where \( a_1 \) and \( a_2 \) are the maximum electric field transmissivities of the two filters. Thus, by simply cascading two FBG filters, we can perform low-rate compensation either post-fiber or pre-fiber.
We have now described all the experimental setups used in this research. Table 1 lists the manufacturer and model number of each component employed in Figs. 11-14.

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<td>AQ8201-21</td>
</tr>
<tr>
<td>OSA</td>
<td>Ando</td>
<td>AQ6317</td>
</tr>
<tr>
<td>Optical 95/5 splitter</td>
<td>generic</td>
<td>--</td>
</tr>
<tr>
<td>Low power EDFA</td>
<td>Oprel</td>
<td>OFA17D1241S</td>
</tr>
<tr>
<td>Photodetector</td>
<td>HP</td>
<td>11982A</td>
</tr>
<tr>
<td>RF variable delay</td>
<td>Sage</td>
<td>6705K-2</td>
</tr>
<tr>
<td>RF variable delay</td>
<td>Macom</td>
<td>2054-6102-00</td>
</tr>
<tr>
<td>Polarization stabilizer</td>
<td>General Photonics</td>
<td>POS-103A</td>
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<tr>
<td>Oscilloscope</td>
<td>Agilent</td>
<td>86100A, 86112A</td>
</tr>
<tr>
<td>Power meter for SBS</td>
<td>HP</td>
<td>8153A, 81532A</td>
</tr>
<tr>
<td>Precision wavelength meter</td>
<td>Bristol</td>
<td>228A</td>
</tr>
<tr>
<td>Power supply for RF amplifier</td>
<td>Agilent</td>
<td>E3647A</td>
</tr>
<tr>
<td>Power supply for current and temperature controller, gain control for RF amplifier</td>
<td>Agilent</td>
<td>E3631A</td>
</tr>
</tbody>
</table>

Table 1. List of components and equipment used in Figs. 11-14.
Chapter 4

Experimental Procedure

Each piece of equipment in this experiment has imperfections that must be considered for measurements to be properly made. We must carefully prepare a procedure to account for each imperfection as much as possible and take unbiased measurements. Section 4.1 discusses the initialization of various components in the system to ensure consistency throughout the measurement process. Section 4.2 discusses the procedure for measuring sinusoidal and detector-based compensation performance pre-fiber at a high data rate, and Section 4.3 expands on the procedure for post-fiber and low data rate compensation. Figure 15 depicts an overview of the procedure.

4.1—System Initialization

With the components in Table 1 selected, we initialized the components for taking measurements. The DFB laser was the first component we initialized because its wavelength drifts for several days after initially powered. The key settings for the DFB laser are wavelength and power. The power was set to be half of the maximum rating, 10mW, by adjusting the current setting on the current controller; this allowed...
us to fine tune the DFB laser’s frequency later without the risk of overpowering it. The temperature was then adjusted to align the DFB’s frequency with the FBG filter. We measured the FBG filter’s center frequency by sending broadband light through the FBG and using the OSA to measure the frequency of maximum transmission. The broadband light was generated by the Oprel EDFA with no input connected. We then monitored the DFB laser’s center frequency on the OSA while tuning it to the FBG’s center frequency. With the frequency approximately tuned, any further frequency adjustments could be made by changing the DFB laser’s current. If the FBG filter’s FWHM is not known, it can be measured for later use by sweeping the DFB laser’s frequency to find the -3 dB power transmission frequencies using the wavelength meter with its built-in power meter. In this thesis, the FWHM of the FBG was calculated to be 5.3 GHz using the frequency response. After re-centering the DFB laser, it was left powered on to stabilize until measurements are ready to take.

The next step we took was to configure the SBS-suppressing phase modulator and waveform generator. Our goal was to minimize the amount of SBS, which was measured by connecting the third port of the EDFA’s circulator to a power meter. With the phase modulator, EDFA, and fiber reel connected to the DFB laser, we set the EDFA’s output power to the maximum setting of 100mW and adjusted the waveform generator’s amplitude, frequency, and waveform until the back-reflected power was minimized. This setting was experimentally determined to be a 20 MHz sine wave with an amplitude of 10 V peak-to-peak. This is the maximum frequency and amplitude that the Agilent 33220A waveform generator can supply, suggesting that SBS could be further suppressed with a stronger, higher-frequency sine wave. At the time of taking measurements this setting was believed to be sufficient, but as will be seen in Chapter 5, SBS caused some of the measurements to be unusable. It is thus recommended for the SBS threshold to be set well below the maximum power, i.e., only a small fraction of the transmitted power should be back-reflected at the maximum power setting.

The third component we set up was the MZM. We need to set the MZM’s input sinusoid’s frequency, amplitude and offset to generate an RZ50 waveform. The frequency was set to 4.3 GHz to
optimally match the FBG filter as calculated using Eq. (25). A few methods of calibrating the RZ50’s shape were tested: (1) measuring the transmitted time domain intensity waveform; (2) measuring the transmitted frequency spectrum; (3) measuring the frequency components of the time domain intensity waveform. Of these methods, the third was the most consistent, although the OSA deconvolution approach for achieving high resolution spectra described in Section 3.1 had not yet been implemented at the time of testing. We took advantage of the fact that the RZ50 intensity waveform is half-wave symmetric to center the sinusoid’s zero frequency offset at the MZM’s 50% transmissivity point. By displaying the intensity waveform along with a half-period-delayed and inverted copy on the oscilloscope, we adjusted the bias until the two waveforms were equal. With the bias set, we adjusted the RF synthesizer’s amplitude until the ratio of the waveform’s average power to fundamental frequency power equaled that of an RZ intensity waveform. The system used for this setup is displayed in Fig. 16.

![Figure 16. System used to set the MZM’s input amplitude and bias.](image-url)

The MZM has a non-negligible drift of its offset phases that causes the RZ waveform to change shape. During the course of experimenting, it was observed that large changes in the MZM bias voltages caused drifts that would continue for hours. While it is favorable to bias the MZM to generate an RZ waveform as close to ideal as possible, it is more favorable to minimize drifts because any change in shape of the RZ waveform will directly cause a change in the average power of the waveform. Thus, we
chose to forego precise MZM biasing in favor of bias stability by setting the MZM several hours before successive measurements and not in between measurements.

Because the MZM used in this experiment is single-drive, there are four possible RZ50 waveforms that can be generated as discussed in Section 2.6. We are interested in using an RZ50 waveform with negative phase shift (see Fig. 8), because the negative phase shift will add with SPM to broaden the spectrum and further suppress SBS. To determine which RZ50 our MZM generated, we measured the SPM average power penalty for both positive and negative phase shifts and then qualitatively fit the measurements to the curves in Fig. 10. The SPM average power penalty was measured by configuring the system as displayed in Fig. 13 and recording the average power through the matched filter at thirteen EDFA power levels between 1 mW and 100 mW. We then switched the MZM voltage bias by $\frac{V_{pp}}{2}$ to negate the phase shift and repeated the SPM power penalty measurement for the new bias. The power penalty for each bias was compared, and we set the MZM to the bias with a greater SPM-induced power loss, since it has the desired negative phase shift. From these measurements we also determined whether the balance factor $B$ was less or greater than 0.5. Knowing the waveform phase will aid in calculating $\phi_{max}(L)$ from the uncompensated power loss and PSD.

4.2—High Rate, Pre-Fiber Measurement Procedure

The system diagramed in Fig. 13 was assembled to evaluate sinusoidal compensation and detector-based compensation. An additional power meter was connected to the third port of the EDFA’s circulator to measure any SBS. It is beneficial to test sinusoidal and detector-based compensation using a system that can perform both, so that any power variations from switching connectors and adding splitters is minimized across the two configurations, giving us the best basis of comparison. After verifying that the DFB laser source’s center frequency was aligned with the matched filter and the MZM was appropriately biased to generate an RZ waveform with negative phase shift, we were ready to take
measurements. If the MZM bias needed to be adjusted, we waited several hours after adjusting the bias for the bias to stabilize before taking measurements.

The first measurements we took were the uncompensated SPM-affected PSD and uncompensated filtered SPM average power. For these measurements, we terminated the variable delay’s input and the photodetector and RF splitter’s outputs with 50-ohm loads. The EDFA’s power was set to a value well below the SPM threshold, i.e., 1 mW, and stepped up to the maximum power, 100 mW, in thirteen power levels of 1, 2, 5, 10, 20, 30, 40, 50, 60, 70, 80, 90, and 100 mW. For each power level, the power through the filter and the power propagating backward toward the EDFA were recorded, and the OSA was swept to take a PSD measurement. Each PSD was saved to a file containing frequency and power density samples. All power meters were set to average the power for about a hundred samples or ten seconds. Taking uncompensated measurements serves two purposes—to give a basis of improvement for the compensated waveforms, and to calibrate the amount of SPM for each measurement. Because there are frequency drifts in the FBG, bias drifts in the MZM, and thermal variations that affect nearly every component in the system, we need to calibrate the amount of SPM to obtain precise measurements.

After the preceding baseline calibration was completed we performed sinusoidal compensation. We connected the output of the second splitter from the synthesizer to the variable delay. The EDFA’s power was set to the maximum first because it was easier to apply compensation with higher amounts of SPM. The compensating RF amplifier’s gain and the variable delay were coarsely adjusted to maximize the power through the matched filter. The gain and delay were then alternatingly fine-tuned over several iterations until the power was maximized. Pre-fiber compensation has the instability issue with SBS, so the power meters must have averaging present while making the gain and delay adjustments. With the power maximized we recorded the filter throughput, backward-propagating power, the RF amplifier’s gain, and the PSD. This measurement was repeated for each of the thirteen power levels of the EDFA, this time descending from 100 mW to 1 mW. Finally, we measured the OSA impulse response by setting the EDFA’s power to 10mW and briefly powering off the frequency synthesizer while sweeping the OSA
and saving the PSD. We repeated the uncompensated measurements and sinusoidal compensation measurements for a total of two sets of pre-fiber, high-rate sinusoidal compensation measurements.

To perform detector-based compensation, we disconnected the split synthesizer signal from the variable delay, connected the output of the photodetector to the variable delay, and connected an RF terminator to the disconnected RF splitter. The gain of the RF amplifier was turned all the way up to reduce saturation effects. Now the subsidiary EDFA was used to control the amount of detector-based compensation. With the power of the high-power EDFA set to maximum, we turned up the power of the subsidiary EDFA and adjusted the variable delay to maximize the filter throughput power, as we did for sinusoidal compensation. The filter throughput, backward-propagating power, subsidiary EDFA power, and PSD were recorded and the process was repeated for the thirteen primary EDFA power levels. We then measured the OSA impulse response. We alternated between taking uncompensated measurements and compensated measurements two times, although in the next measurements this was done three times.

4.3—Low Rate, Post-Fiber Measurement Procedure

Once sinusoidal compensation and detector-based compensation had been tested for the high rate, pre-fiber system, we performed post-fiber compensation. Performing post-fiber compensation is fundamentally similar to that of pre-fiber compensation. We assembled the system as depicted in Fig. 14, and repeated the measurements as done for pre-fiber compensation with a few minor changes. First, because SBS is suppressed due to SPM and remains consistent throughout all of our measurements, SBS measurements were only necessary to ensure that the SBS threshold was not exceeded. Second, compensating post-fiber comes at a cost in convenience—there is a slowly varying phase shift added by the 20 km of fiber that makes it difficult to keep the compensating waveform’s phase synchronized with the signal’s phase. Therefore, instead of averaging the power meter for ten seconds, we set the power meter to average for a few seconds. We can do this without compromising the measurements’ integrity because without the pre-fiber SBS instability there are smaller power fluctuations at the output of the fiber
reel. With the slowly varying phase shift in mind, we performed sinusoidal and detector-based compensation three times each.

Once sinusoidal compensation and detector-based compensation had been tested for the post-fiber high-rate system, we performed low-rate compensation by adding an additional FBG filter after the initial one. The FWHM of the equivalent filter can be calculated from Eq. (36) but should be measured using the procedure in Section 4.1. In this thesis, previous measurements of the FBGs’ frequency responses allowed the cascaded frequency response and FWHM to be accurately calculated. With the new FWHM of 4.0 GHz, the synthesizer’s frequency was changed to 3.2 GHz and the RZ50 waveform was re-calibrated with a sinusoid of proper bias and amplitude. Note that as long as the bias does not change by more than $\frac{V_m}{4}$, the phase shift of the RZ50 waveform will remain negative and will not have to be re-measured. The procedure for performing sinusoidal and detector-based compensation was then repeated with the additional FBG filter in the system.

The last set of measurements was taken with low rate, pre-fiber compensation. We re-assembled the system in Fig. 13, leaving in the cascaded FBG as it was connected for post-fiber, low rate compensation. Three sets of each type of compensation were measured.

While all of the measurements taken are sufficient for basic analysis, we can gain further insight by measuring the signal used to drive the phase modulator. This measurement allows us to calculate the maximum achievable compensation using a model of the system. Since the system was already configured for low-rate compensation, we measured the output of the RF drive amplifier as if we were performing sinusoidal and detector-based compensation. The drive signal was measured at sixteen drive gains for sinusoidal compensation and thirteen EDFA powers for detector-based compensation. We then re-configured the system for high-rate compensation, re-calibrated the MZM bias, and took more measurements of the drive signal.
Chapter 5
Experimental Results

5.1—Average Power Loss Measurements

We are primarily interested in how well detector-based compensation is able to reduce SPM-related losses in comparison to sinusoidal compensation. Thus, it is the average power through the filter—our proxy for SNR loss—that we will present first. The results for pre-fiber and post-fiber, low rate and high rate compensation are displayed in Figs. 17–20. Figure 21 presents the average of all of the low rate measurements and the average of all of the high rate measurements. The data shown is the output power through the matched filter divided by the EDFA power and normalized to the maximum uncompensated penalty for each power sweep, where each curve represents a single power sweep from 1 mW to 100 mW. Minimal processing is done to faithfully represent the data collected. The distinction between the uncompensated trials for sinusoidal compensation and detector-based compensation is made to provide a better representation of the improvement due to compensation. In all of the trials, there is a sharp increase in normalized average power at input powers from 1 mW to 5 mW, attributable to a disparity between the EDFA’s power setting and its actual output power. Note that the loss is given in terms of the EDFA’s average power instead of $\phi_{\max}(L)$; this is because $\phi_{\max}(L)$ depends on many parameters for which we have estimates but do not know the exact value. A more sophisticated analysis is performed in Sections 5.2 and 5.3, where we use our model of the system to estimate $\phi_{\max}(L)$ and various other parameters.
Figure 17. Average power penalty for a low rate, pre-fiber compensated signal.

Figure 18. Average power penalty for a high rate, pre-fiber compensated signal.
Figure 19. Average power penalty for a low rate, post-fiber compensated signal.

Figure 20. Average power penalty for a high rate, post-fiber compensated signal.
5.2—Average Power Loss Model

To help determine why the results of the experiment turned out the way they did, we parametrically modeled the system based on the uncompensated measurements in an attempt to replicate the results of the compensated measurements. The model was seeded with typical values for the parameters and put into an optimization algorithm to best fit the uncompensated average power loss. After the parameters were calculated, the maximum compensated throughput for each power setting was calculated using the measured phase modulator drive signal. Here the best fit is achieved when the sum of the squared error between the measured average powers in dBm and the calculated average powers in dBm is minimized. The model’s parameters are the MZM drive sinusoid’s bias for each power sweep, the MZM drive sinusoid’s amplitude, the input loss between the EDFA and the fiber reel, the output loss between the fiber reel and the power meter, and the MZM’s extinction ratio. This model assumes that the MZM drive sinusoid’s bias changes for the six power sweeps (three for sinusoidal compensation, three for detector-based compensation) due to MZM bias drift, but the relative amplitude and extinction ratio is stable. The input loss from the EDFA’s output allowed us to calculate the power going into the fiber reel and thus $\phi_{\text{max}}(L)$, and the lumped output loss accounts for the various losses in the fiber reel, polarization stabilizer, optical splitter, phase modulator, and FBG. If the calculation of the maximum
compensated average power equals our measurements, we can analyze the results of the best fit model to ascertain why the compensation techniques did not perform as we predicted in Section 2.6.

We have taken many precautions to suppress SBS, but unfortunately these precautions were not sufficient to fully suppress SBS when we tested pre-fiber compensation. In the high rate and low rate pre-fiber tests, the back-reflected power was measured to be 7 dBm (5 mW) on average at an input power of 100 mW. Considering that about 25 mW is received at the end of the fiber at this power level and assuming that the back-reflected power would have propagated forward without SBS, SBS caused an additional loss of −0.8 dB, a significant amount compared to the accuracy of the measurements we require. Since we do not model SBS in this thesis, we cannot predict how this error affects our measurements. Thus, we will not attempt to model the pre-fiber compensation measurements.

The best fit model parameters for the low rate, post-fiber average power measurements are presented in Table 2. All of the parameters are within reason: the drive bias was experimentally set with 0.01 V precision where $V_\pi$ is approximately 4 V as stated in the datasheet; the drive amplitude could have been erroneously set by a factor of 1.374; the extinction ratio is not far off the expected −13 dB; the input loss was measured to be about −1 dB; and the output loss is at the least −7 dB from fiber loss and the optical splitter.

<table>
<thead>
<tr>
<th>MZM Drive Bias Deviations</th>
<th>MZM Drive Amplitude Deviation Ratio</th>
<th>Extinction Ratio</th>
<th>Input Loss</th>
<th>Output Loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.014 V/V_\pi</td>
<td>1.374 V/V</td>
<td>-19.0 dB</td>
<td>-1.03 dB</td>
<td>-10.81 dB</td>
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<tr>
<td>0.000 V/V_\pi</td>
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<td></td>
</tr>
<tr>
<td>0.007 V/V_\pi</td>
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<tr>
<td>0.009 V/V_\pi</td>
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<td></td>
</tr>
<tr>
<td>0.027 V/V_\pi</td>
<td></td>
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<tr>
<td>0.017 V/V_\pi</td>
<td></td>
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</tr>
</tbody>
</table>

Table 2. Model parameters used to calculate the average power loss that best fits the measured power loss in the low rate, post-fiber compensated system.
Using the parameters of the model and the measured phase modulator drive signal, we then calculated the maximum compensated filter throughput. While we did not measure the drive signal until the end, we recorded the amplifier gain or subsidiary EDFA power and can interpolate between the signals we measured to obtain approximations of the drive signals. The best fit average power loss and maximum compensated filter throughput are presented in Figs. 22 and 23. These figures normalize the average power penalty to the maximum average power achieved by the model. The model fits the uncompensated average power loss nearly perfectly; however, the maximum calculated compensated filter throughput only crudely matches the measurements. While the maximum difference between the measured and calculated compensated average power loss is only 0.5 dB, the qualitative trend of the calculated power does not fit the measured power for either sinusoidal or detector-based compensation.

Figure 22. Modeled average power penalty for a low rate, post-fiber sinusoidal compensated signal.
The same model was tested on the high rate post-fiber measurements. However, the best fit parameters did not converge unless the MZM drive sinusoid’s amplitude was fixed, and small variations in the amplitude caused the fitting parameters to vary greatly. The inconsistent nature of fitting the high rate, post-fiber, average power measurements was expected to lead to a faulty analysis, thus we did not proceed with the analysis.

5.3—Power Spectral Density Model

Just as we made a model to best fit the uncompensated average power through the filter, we also made a model that best fits the PSD of the uncompensated SPM affected signal. The spectra measured by the OSA was deconvolved with the OSA filter’s frequency response to return the squared magnitude of the RZ waveform’s harmonics. To verify that the deconvolved PSD was valid, we calculated how much power would pass through the FBG filter based on the PSD. The result of this calculation for the post-
fiber PSD measurements is presented in Fig. 24. This calculation replicates the major trends we see in the measured post-fiber average power.

![Graph showing high and low rate post-fiber penalty from PSD measurements.](image)

**Figure 24.** Calculation of the average power penalty using PSD measurements.

After verifying that the PSD measurements were valid, we used our model and an optimization algorithm to best fit the PSD using the same parameters as before. Now the optimum criterion is achieved when the squared error between the calculated uncompensated PSD in mW and the measured uncompensated PSD in mW is minimized. This calculation was done using mW instead of dBm to add weight to higher energy harmonics which are less prone to noise. The best-fit model parameters for low rate post-fiber compensation using uncompensated PSD measurements are presented in Table 3, one set of PSD measurements and best-fit calculations is shown in Fig. 25, and the measured and modeled average power penalties are shown in Fig. 26. The parameters are all reasonable except for the extinction ratio, which is quite different from its typical value. Regardless, the calculated PSD appears to fit the measurements up to the third harmonic, and the uncompensated average power penalty calculation follows the trend of the measurements. Unfortunately, this model does not replicate the average power penalty of the compensated waveform any better than the previous model.
Table 3. Model parameters used to calculate the PSD that best fits the measured PSD in the low rate, post-fiber compensated system.

<table>
<thead>
<tr>
<th>MZM Drive Bias Deviations</th>
<th>MZM Drive Amplitude Deviation Ratio</th>
<th>Extinction Ratio</th>
<th>Input Loss</th>
<th>Output Loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.098 V/V&lt;sub&gt;π&lt;/sub&gt;</td>
<td>0.957 V/V</td>
<td>-28.0 dB</td>
<td>-2.17 dB</td>
<td>-10.4 dB</td>
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<td>0.096 V/V&lt;sub&gt;π&lt;/sub&gt;</td>
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<td>0.081 V/V&lt;sub&gt;π&lt;/sub&gt;</td>
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</table>

Figure 25. Measured and best fit PSDs for a low rate, uncompensated signal. Lines are used to show contours for each input power and do not represent a continuous spectrum.

Figure 26. PSD modeled average power penalty for a low rate, post-fiber compensated signal.
We repeated the same calculation for the high rate post-fiber PSD measurements. The model parameters, arranged in Table 4, are surprisingly similar to those for the low rate post-fiber PSD model with the exception of the MZM drive biases. Figure 27 shows that a best fit PSD for a single power sweep matches the corresponding measured PSD up to the second harmonic. Note that the measured SPM-affected spectrum shows significantly more broadening than the best fit spectrum. The modeled and measured average power losses are presented in Fig. 28. As with the low rate post-fiber PSD model, the calculated uncompensated average power penalty fits the measurements but the calculated compensated average power penalty does not.

<table>
<thead>
<tr>
<th>MZM Drive Bias</th>
<th>MZM Drive Amplitude Deviation Ratio</th>
<th>Extinction Ratio</th>
<th>Input Loss</th>
<th>Output Loss</th>
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<td>0.186 V/V_π</td>
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<td>-27.5 dB</td>
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<td>-10.1 dB</td>
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<tr>
<td>0.187 V/V_π</td>
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<td>0.181 V/V_π</td>
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<td>0.135 V/V_π</td>
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<td>0.168 V/V_π</td>
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<td>0.193 V/V_π</td>
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</tbody>
</table>

Table 4. Model parameters used to calculate the PSD that best fits the measured PSD in the high rate, post-fiber compensated system.

Figure 27. Measured and best fit PSDs for a high rate, uncompensated signal. Lines are used to show contours for each input power and do not represent a continuous spectrum.
5.4—Analysis and Discussion

The detector-based compensation technique performed comparably to sinusoidal compensation with both techniques recovering about 4 dB of SPM-incurred losses at $\phi_{\max}(L) \approx 3$. Detector-based compensation reduced SPM losses up to 0.1 dB more than sinusoidal compensation for $\phi_{\max}(L) < 2$ and sinusoidal compensation performed up to 0.2 dB better than detector-based compensation for $\phi_{\max}(L) > 2$. These results are generally consistent throughout all of the measurements and they demonstrate that detector-based compensation is viable for mitigating SPM losses. The compensated PSDs, presented in Appendix B, show how both techniques are able to reverse SPM’s spectral broadening. However, both techniques faced a 1 dB penalty at $\phi_{\max}(L) \approx 3$ that was not predicted by the calculations in Section 2.6. Of all the models we generated, the uncompensated average power penalty model best predicted the performance of both compensation techniques, but not well enough to use its parameters as a justification for the 1 dB penalty. This suggests that there are other effects we did not account for in our model that are affecting our measurements.

We would like to know how much SBS had an impact on our measurements. One way to determine how much power is lost due to SBS and other optical effects is to examine the signal’s power before passing through the filter. The pre-filter power is not affected by SPM, thus any power degradation is caused by other optical effects. We estimated the pre-fiber power by integrating the PSD
measurements, and from here we calculated the loss in the fiber as a function of input power. Figure 29 displays the pre-filter loss for each set of measurements. This pre-filter loss explains why both compensation techniques performed a few tenths of a decibel worse than the calculations—some of the power loss occurred before the signal even passed through the filter. It is likely that the pre-filter loss at higher powers is caused by SBS, given that the penalty increases with increasing power. Note that the pre-filter power loss from 1 mW to 5 mW is caused by the disparity between the EDFA’s power setting and output power.

![Graphs of Pre-filter Power Loss](image)

Figure 29. Pre-filter power loss. The loss is normalized to the maximum uncompensated power loss and averaged over the three power sweeps (two for high rate, pre-fiber compensation).

Calculations using the back-reflected power report similar losses in our post-fiber compensation measurements. For $\phi_{\max}(L) \approx 3$, an average of 2.2 dBm (1.7 mW) was back-reflected toward the EDFA and approximately 15 dBm (31.6 mW) propagated to the output of the fiber, assuming an input loss of -1
dB and a fiber absorption loss of -4 dB. If all of the back-reflected power had instead propagated forward, the output would increase by 0.2 dB. With an input loss of -2 dB, the output would increase by 0.3 dB. This calculation supports the idea that SBS is causing some, if not all, of the power loss before the filter.

There is also evidence that other optical effects are present. There is an asymmetry of the uncompensated PSD measurements that does not occur with only SPM. This could be the result of SRS, four-wave mixing, or dispersion interacting with SPM. It is possible that our measurements differed from the calculations by a few tenths of decibels because of these effects.

One of the main objectives of this experiment was to evaluate how detector-based compensation compares to sinusoidal compensation. In making this comparison, we must not neglect the saturating effects of the RF amplifier when performing sinusoidal compensation. The phase modulator drive signals returned by the analysis in Section 5.2 are shown in Appendix A. The amount of second and third harmonic content in the sinusoidal compensation waveform is quite significant for low EDFA powers. It is likely that this extra harmonic content improved sinusoidal compensation in our measurements, thus making detector-based compensation appear to be less beneficial.

The results of this experiment may not be applicable to all lasercom systems. The conditions of the experiment are different from that of a typical lasercom system in five noteworthy ways. First, data is encoded in the pulses of a lasercom system, affecting the frequency spectrum and thus the throughput of the signal with the matched filter. Second, in this experiment average power was used in lieu of peak power, giving us only an approximation of the SNR benefit of both compensation techniques. Third, a single-drive MZM was used, causing additional phase shifts that would not be present in a dual-drive MZM. Fourth, SBS and presumably other optical phenomena had subtle effects on the measurements. These optical phenomena are likely to be different when peak powers are much higher than 200 mW and fiber lengths are much shorter than 20 km. Finally, we were only able to achieve $\phi_{\text{max}}(L) \approx 3$ in our measurements, and at this level, the theory predicts that detector-based compensation will perform 0.1 dB
better than sinusoidal compensation. Detector-based compensation is expected to be most beneficial for higher $\phi_{max}(L)$ which we did not test in this experiment.
Chapter 6
Conclusion and Future Work

We tested sinusoidal compensation and detector-based compensation of SPM-affected RZ50 pulses at 4.3 GHz and 3.2 GHz pulse rates. With $\phi_{max}(L) < 3$, detector-based compensation performed within 0.2 dB of sinusoidal compensation across all of the test conditions. At $\phi_{max}(L) \geq 3$ both techniques were able to recover about 4 dB of SPM-incurred losses, but there was still a 1 dB loss. We used the uncompensated measurements to generate models of the system in an attempt to explain the 1 dB loss; however, none of the models were able to accurately do so. Based on the SBS and PSD measurements, the best explanation is that other optical effects—for which the model did not account—caused additional power loss at higher input powers. It is likely that detector-based compensation will outperform sinusoidal compensation in an actual lasercom system where more SPM and less SBS occur.

Perhaps the most surprising result of the experiment is that both compensation techniques did not perform as well as the theory suggested. A simple experiment to determine whether this result was caused by an inaccurate compensation waveform or by other optical effects is to perform compensation using the addition of multiple RF sinusoids. This technique allows the amplitudes and phases of the compensating waveform’s harmonics to be independently adjusted to completely mitigate SPM losses. With SPM completely mitigated, the loss from other nonlinear effects can be measured. If this compensation technique works as well as sinusoidal or detector-based compensation, then it can be concluded that other optical effects are present in the fiber.

To ultimately determine how well detector-based compensation performs, it has to be tested on a lasercom testbed. The conditions of the experiment performed in this thesis are different from a lasercom system in a few ways. There are higher powers, shorter fiber lengths and data modulation in a lasercom system, and a bit-error-rate test is used to calculate SNR instead of average power. In this setting,
detector-based compensation shows great promise for fully mitigating SPM where sinusoidal compensation cannot.
Appendix

Appendix A—Phase Modulator Drive Signals

To demonstrate the effects of saturation caused by the RF amplifier and the effect of taking a band-limited measurement of the optical intensity, we have plotted the phase modulator drive signal for sinusoidal compensation and detector-based compensation in Figs. 30 and 32. Note that we did not directly measure the phase modulator drive signal while performing compensation; we used the recorded amplifier gain or subsidiary EDFA power and interpolated between the drive signals we measured to obtain approximations of the drive signals. The model in Section 5.2 is used to convert the voltage to a phase shift. These measurements are for low rate, post-fiber compensation. The corresponding Fourier transform of each waveform, normalized to the amplitude of the fundamental, is shown in Figs. 31 and 33. The RZ intensity waveform returned by the model is also shown in Figs. 30-33 for comparison.

Figure 30. Phase modulator drive signal for low rate, post-fiber sinusoidal compensation.
Figure 31. Normalized harmonic content of the low rate, post-fiber phase modulator drive signal for sinusoidal compensation.

Figure 32. Phase modulator drive signal for low rate, post-fiber detector-based compensation.
Appendix B—Compensated Power Spectral Densities

We have presented the uncompensated power spectral densities for low rate and high rate compensation in Section 5.3. While we did not analyze the compensated PSDs, they may be of interest to the reader. The compensated PSDs for post-fiber compensation are shown in Fig. 34. The EDFA powers displayed here are 1 mW, 5 mW, 20 mW, 40 mW, 60 mW, 80 mW, and 100 mW.
Figure 34. Post-fiber compensated PSDs.
Bibliography


