Optoelectronic Frequency Stabilization Techniques

in Forced Oscillators

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ABSTRACT

Opto-electronic Frequency Stabilization Techniques in Forced Oscillators

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Prof. Afshin S. Daryoush, Supervisor

Forced frequency stabilization techniques of self-injection locking (SIL) and self-phase locked loop (SPLL) have been demonstrated to be effective for phase noise reduction. In SIL scheme, a portion of the oscillator output is directly injected back to the oscillator after passing through a long delay while in SPLL, the delayed signal is used to compare against a non-delayed signal to generate an error signal which will be used to control the oscillator frequency. The published literature and reported patent investigations has revealed that SIL and SPLL are only being used independently, whereas in this thesis SILPLL is introduced for the first time by simultaneously combining of SIL and SPLL; the control theory based modeling of SIL, SPLL, and SILPLL techniques are developed and the simulation results have indicated that phase noise of the oscillation signal is enhanced as injection locking removes phase noise in far-out offset frequency, while phase locking effectively reduces the close-in to carrier phase noise. Optimum delay length and delay parameters are identified for the most effective performance. Both SIL and SPLL techniques require a low noise figure and long fiber optic delay lines to provide substantial phase noise reduction, but the long delay generates undesirable sidemodes that are seen within 20kHz to 200kHz offset to carrier for fiber optic delays from 10km to 1km and are difficult to be filtered out by electrical filters. Therefore, forced oscillation technique employing short and long delay is proposed to suppress these sidemodes.

In this dissertation, experimental results of a dual self-injection locking (DSIL) and dual self-phase locked loop (DSPLL) employing short and long delays have been proposed for this sidemode suppression, while maintaining same amount of phase noise reduction provided by the long delay. As an example of DSIL, sidemode suppression of more than 20dB for fiber delay links of 1km and 5km have been experimentally
achieved compared to a single 5km long SIL with a phase noise reduction of 40dB (in reference to free running oscillator) at 10kHz offset from carrier for both standard OEO and a self-seeded structure with electrical 3 port oscillator at 10GHz; for a DSPLL fiber delay lines of 3km and 5km, a sidemode suppression of 29dB have also been experimentally achieved compared to SPLL of 5km with a phase noise reduction of 30dB (in reference to free running oscillator) at 10kHz offset from carrier. For the case of SPLL, phase locking performance of a 10GHz oscillation signal are experimentally evaluated as various methods of phase locking are compared. Experiment results that demonstrate the benefit of SILPLL incorporating dual delays have been reported for the first time corroborating analytical predictions. A dual SILPLL (DSILPLL) system with 3km and 5km fiber delay has been implemented, and measured phase noise reduction of 40dB provided by DSILPLL is the same as DSIL at 10kHz offset. However, at 1kHz offset, DSILPLL provides a phase noise reduction of 52dB which is 11dB better than DSIL; at 300Hz offset, DSILPLL provides 70dB reduction while DSIL provides only 42dB reduction.

In summary, DSILPLL is effective for sidemode suppression and phase noise reduction, where SPLL using tunable MZM with DSIL of a VCO provides the best performance improvement over other investigated topologies. Due to the advances in low noise electronics and broad bandwidth of the optical components used in the DSILPLL system, the DSILPLL technique has the potential to create highly stable RF oscillators approaching 100GHz.
## Acronyms and Symbols

### Acronyms

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<td>B</td>
<td>Injection locking loop gain</td>
</tr>
<tr>
<td>DRO</td>
<td>Dielectric resonator oscillator</td>
</tr>
<tr>
<td>DSIL</td>
<td>Dual loop Self-injection locking</td>
</tr>
<tr>
<td>DSILPLL</td>
<td>Dual loop Self-injection locked and phase locked loop</td>
</tr>
<tr>
<td>DSPLL</td>
<td>Dual loop Self-phase locked loop</td>
</tr>
<tr>
<td>G</td>
<td>Phase locking loop gain</td>
</tr>
<tr>
<td>IL</td>
<td>Injection locking</td>
</tr>
<tr>
<td>ILPLL</td>
<td>Injection locked and phase locked loop</td>
</tr>
<tr>
<td>MMS</td>
<td>Metamaterial Mobius Strip</td>
</tr>
<tr>
<td>OEO</td>
<td>Opto-electronic oscillator</td>
</tr>
<tr>
<td>PLL</td>
<td>Phase locked loop</td>
</tr>
<tr>
<td>SIL</td>
<td>Self-injection locking</td>
</tr>
<tr>
<td>SILPLL</td>
<td>Self-injection locked and phase locked loop</td>
</tr>
<tr>
<td>SLC</td>
<td>Sapphire loaded cavity</td>
</tr>
<tr>
<td>SPLLL</td>
<td>Self-phase locked loop</td>
</tr>
</tbody>
</table>

### Symbols

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_m$</td>
<td>Fourier frequency</td>
</tr>
<tr>
<td>$f_{osc}$</td>
<td>Oscillation frequency</td>
</tr>
<tr>
<td>$s_{\phi}(f_m)$</td>
<td>Power spectral density of phase</td>
</tr>
<tr>
<td>$L(f_m)$</td>
<td>Single sideband phase noise</td>
</tr>
<tr>
<td>$\phi_i$</td>
<td>Phase angle of input signal</td>
</tr>
</tbody>
</table>
\( \phi_o \)  
Phase angle of VCO

\( n_p \)  
Voltage of input noise to the forced oscillation system

\( n_q \)  
Voltage of oscillator phase noise

\( \phi_p \)  
Phase error due to \( n_p \)

\( \phi_q \)  
Phase error due to \( n_q \)

\( \phi_n \)  
Total phase error combining \( \phi_p \) and \( \phi_q \)

\( \zeta \)  
Damping factor of PLL loop

\( \omega_n \)  
Natural frequency of PLL loop

\( \tau \)  
Time delay of the fiber used in self-forced oscillation circuits
Chapter 1 INTRODUCTION

Low noise oscillators are at the heart of modern communication systems providing reference signal to establish or select particular transmission channels; they also enable clock signals for many other electronic systems ranging from microprocessors to wireless base stations, radar and satellite communication link. Unfortunately, performance of all practical oscillators suffers from phase noise. Phase noise refers to the short-term random fluctuation in the frequency (or phase) of an oscillator signal, and is of critical concern in the aforementioned systems.

For example, the phase noise requirement is stringent for a Doppler radar when it’s detecting small and slow moving object. Let’s assume a vehicle 5 miles away is moving directly toward the Doppler radar at a speed of 50 MPH, this corresponds to a Doppler frequency shift of about 1.5 kHz. In order to detect the vehicle, the phase noise at 1.5 kHz offset for a typical 10 GHz carrier needs to be lower than the signal reflected back from the vehicle, which requires an extremely clean spectrum of the LO. On the other hand, if a plane 5 miles away is traveling head-on towards the radar at a speed of 300 MPH then the Doppler frequency shift is approximately 9 kHz – 10 kHz. Although this relaxes the phase noise requirement for detection, a clean spectrum at 10 GHz is needed nonetheless. Commercial applications also impose challenging phase noise requirement for oscillators. As another example, in the direct digital microwave receivers, giga-samples/s (GSPS) analog to digital converters (ADC) require a low jitter clock signal at microwave frequencies. The ADC block diagram is shown in Figure 1.1, where a low noise oscillator serves as the clock signal to sample the input signal and the output signal is then being quantized into bits. The equivalent number of bits (ENOB) for the digitized signal is directly related to the jitter of the clock, shown in Figure 1.2a. For a 50 GSPS rate, the minimum aperture jitter is about 0.02 ps. And from Figure 1.2b, we find that this corresponds to a phase noise of about -160 dBc/Hz at 1 kHz offset for a 25 GHz carrier.
Both military and commercial applications impose challenging phase noise requirement for oscillators, and with the limited spectrum resources left in the L and S band, future applications yearn for oscillators with high spectral purity at X band and beyond. In the next section, techniques from the published literature for achieving high frequency ultra-stable oscillators will be examined, and the technique proposed in this thesis will also be presented.

This thesis is organized as follows: Chapter 2 is the review of literature, several previous works on the design of low phase noise oscillator using high quality factor resonators using long delays in both electrical and optical domains. Moreover, various frequency stabilization techniques reported in literature are examined to further reduce the achieved low phase noise. Chapter 3 focuses on frequency stabilization
technique of self-injection locking (SIL) of optoelectronic oscillator (OEO) and voltage controlled oscillators (VCO). Both analytical modeling and experimental results are presented for SIL oscillators using optical delay lines. Chapter 4 presents the analytical modeling and experimental results of another forced oscillation technique using self-phase lock loop (SPLL) using single and dual optical delays for OEO and VCO. In Chapter 5, technique of self-injection locked phase lock loop (SILPLL) is introduced for the first time by simultaneously combining SIL and SPLL is proposed, both analytical modeling and experimental results are provided to validate analytical predictions of the proposed structure. Chapter 6 summarizes this work and recommends for future work based on performance improvement of SILPLL using optimum length for a reduced noise figure fiber optic delay lines, narrowband tunable filtering using wavelength tunable optical transversal filter, and use of passively temperature stabilized OEO by Raman amplified photonic crystal fibers. A number of appendices in support of various parts of thesis are also provided in this thesis.
Chapter 2 REVIEW OF LITERATURE

In order to achieve low phase noise, various techniques and circuit topologies have been proposed and demonstrated. In general, all these different techniques for achieving low phase noise can be categorized into two groups: 1) using high quality factor (Q) resonators in oscillator circuits; 2) using external feedback control for the free running oscillator. The theoretical background for the first group is based on Leeson’s equation [2] which indicates that the oscillator phase noise is inversely proportional to the resonator Q. Therefore, great effort has been made in search of ‘perfect’ resonators. Sapphire loaded cavity (SLC) is capable of providing extremely high Q, and phase noise of -130dBc/Hz at 1kHz offset for a 10.24GHz SLC oscillator has been reported by Poseidon Scientific Instruments PTY Ltd [3].

Optoelectronic oscillator which utilizes extremely low loss fiber optic delay line as energy storage element has achieved unprecedented phase noise performance of -163dBc/Hz at 6kHz offset for a 10GHz carrier [4]. Recently, a novel oscillator that utilizes metamaterial Mobius strip (MMS) on a printed circuit board as its resonator has been reported by Synergy Microwave Inc. [5], phase noise of -120dBc/Hz at 10kHz offset is reported for a 6GHz carrier. The compact size and low cost feature of MMS based oscillator is attractive for commercialization. These oscillator circuits with high Q resonators will be briefly discussed in Section 2.2.

On the other hand, frequency stabilization using feedback control has been investigated extensively for phase noise reduction. A self-injection locking (SIL) topology is proposed by H.C. Chang [6]. The proposed structure is by passing the output of oscillator through an electrical delay line or a high Q resonator and feeding it back using a circulator. The experimentally verified modeling demonstrates that the overall oscillator phase noise is inversely proportional to the signal delay time or Q. However, due to the high loss and limited Q in electric circuits, the phase noise improvement is not significant. To bypass the limitation of electrical components, Lee et al. [7] employed a 2.4km long optical fiber in the feedback loop of the SIL topology and achieved a phase noise reduction of 27dB at 10kHz offset for a 30GHz oscillator. The concept of forced oscillation could also be extended to self-phase locked loop (SPLL)
demonstrated by R. Logan who is the first to construct a delay line frequency discriminator (DLFD) using fiber optic delay [8]. In [8], a 2.2km optic fiber is inserted in one arm of a frequency discriminator that generates an error signal to control the frequency of a VCO. The achieved phase noise reduction is limited to 12dB as the sidemodes associated with the 2.2km delay fiber degrades the spectrum purity. In order to reduce the impact of the sidemodes inherent with long fiber delay, Pillet et al. used dual DLFD for sidemode suppression where a combination of short and long delay is used [9]-[10]. The microwave signal generated from the beat note of a dual frequency laser (DFL) was sent into the dual DLFD whose output is used to stabilize the laser frequency for generating a more stable beat note. The analytical modeling provided explanation for this phase noise reduction. While IL phenomena are easy to implement, the phase noise in the close-in offset frequency range is degraded due to frequency offset detuning and limited locking range as explained in [11]. On the other hand, even though PLL has a longer pull-in time that results in a slow response, the high gain loop filter enables the PLL to remove the close-in phase noise significantly, while far away from carrier suffers from a higher noise. Sturzbecher et al. [12] demonstrated in externally forced oscillators, a better phase noise characteristics for both close-in and far-away from carrier and a wider locking range is achieved by combining IL and PLL (ILPLL) [12]. Forced stabilization techniques will be illustrated in Section 2.3

2.1 High Q Resonator Oscillator Circuits

2.1.1 Sapphire Loaded Cavity Oscillator (SLCO)

The prototype SLCO [3] is depicted in Figure 2.1. It consisted of the following components: SLC resonator, sustaining amplifier (AML with parallel HBTs), bandpass filter (combline 10.24 GHz ± 100 MHz), output coupler (-6 dB), custom varactor-controlled phase shifter ('VCP'), mechanical 'trombone' phase shifter (for coarse loop phase setting), isolators, and SMA adapters. The unloaded Q of the 'high-speed' SLC resonator was 133,000 at 10.24 GHz (operating at +35°C). The port couplings were set at $\beta_1 = \beta_2 = 0.28$, which gave the lowest phase noise while maintaining reasonable loop gain margin over the 0 to
+55° operating range (gain margin measured as 4.8 dB at +25°C). This coupling gave a resonator loaded Q = 85,200 and half-loaded bandwidth $f_{0.5} = f_0 / 2Q_L = 60$ kHz. Output power was +17.5 dBm. The measured phase noise of this SLCO is shown in Figure 2.2, phase noise is -130 dBc/Hz at 1 kHz offset and -155 dBc/Hz at 10 kHz offset for a 10.24 GHz carrier. A detailed review of SLCO is provided in [13].
2.1.2 OEO Basic Dynamics

The Optoelectronic Oscillator (OEO) utilizes the transmission characteristics of a modulator together with a fiber-optic delay line to convert light energy into stable, spectrally pure RF/microwave reference signals. Detailed descriptions of OEO principle can be found in [14]-[17], and are summarized in the following. A conceptual OEO block diagram is shown in Figure 2.3. In this configuration, light from a laser is introduced into an E/O modulator, the output of which is passed through a long optical fiber, and detected with a photodetector. After amplifying and filtering, the output of the photodetector is fed back to the electric port of the modulator. This configuration supports self-sustained oscillations at a frequency determined by the fiber delay length, bias setting of the modulator, and the bandpass characteristics of the filter.

![Figure 2.3 Schematic of OEO in [4].](image)

A regenerative feedback model is used to analyze the spectral properties of the OEO. The conditions for self-sustained oscillations include coherent addition of partial waves each way around the loop and a loop gain exceeding losses for the circulating waves in the loop. The first condition implies that all
signals that differ in phase by integer multiples of $2\pi$ from the fundamental signal can be sustained. The second condition implies that with adequate light input power, self-sustained oscillations may be obtained without the need for electrical amplifier. The phase noise of OEO based on this regenerative feedback model is given below,

$$S_{OUT}(f_m) = \left(1 + \left(\frac{f_{osc}}{2Qf_m}\right)^2\right)S_{IN}(f_m)$$

(2.1)

where $Q$ is the equivalent quality factor of the optical delay line; $f_{osc}$ is the oscillation frequency; $f_m$ is the offset frequency; $S_{IN}(f')$ is the input noise to signal ratio; $S_{OUT}(f')$ is the output noise spectrum.

The phase noise of an OEO with 16km fiber delay is shown in Figure 2.4. At 6kHz offset, the achieved phase noise is -163dBc/Hz. The superior performance results from the extremely low energy storage loss realized by incorporating long optical fiber. The optical fiber is also virtually free of any frequency-dependent loss, resulting in the same long storage time and high spectral purity signals for both low and high frequency oscillation.

![Figure 2.4 Experimental result of an OEO with 4km loop reported by D. Eliyahu et al in [4]](image)
Two variants of the standard OEO have also been studied extensively, and they are coupled OEO (COEO) [18]-[22] and the whispering gallery mode (WGM) based OEO [23]-[25]. In COEO, the long fiber delay served as high Q element in standard OEO is replaced by mode-locked laser (MLL). The benefit of using MLL is that the high Q operation originates from the regenerative feedback in the MLL thus eliminates the need for long fiber delays; in addition the mode-locked state in optical domain results in a very low timing jitter for the microwave signal, which translates into very low phase noise. For WGM OEO, the high Q operation is provided by WGM disk resonator. Due to the inherent high unloaded Q and the compact size of the resonator, WGM OEO is attractive for fabrication in miniature package while achieving low phase noise. However, the whispering gallery mode resonators as a class of dielectric resonators are spherical or toroidal geometry dependent and any stress caused strain in the dielectric resonators could cause microphonic noise in oscillators.

2.1.3 Metamaterial Mobius Strip Based Oscillator

Manipulating and tailoring the electromagnetic wave coupling properties of MMS leading to a reduction in size for a given operating frequency and suppression of spurious resonance modes, thereby improving the overall performance and reducing power consumption. Figure 2.5 shows a typical layout of metamaterial based MMS resonator and its equivalent circuit model. A layout of an oscillator using evanescent mode MMS resonator is shown in Figure 2.6. The measured phase noise of this oscillator is shown in Figure 2.7. The achieved phase noise is -120dBc/Hz at 10kHz for a 6GHz carrier.
Figure 2.5 A typical Metamaterial based Möbius strips resonator: (a) layout. (b) electrical equivalent lumped model circuits. Images adopted from [5]

Figure 2.6 A typical layout of 6 GHz oscillator (0.5x0.5 0.18 inches): The layout of oscillator using evanescent mode MMS resonator network realized using 20 mils RT/Duroid 5880 with $\varepsilon_r = 2.2$. Image adopted from [5]. (Copyright-Synergy Microwave Corp., NJ, Patent Pending)
2.2 Forced Frequency Stabilization Techniques

2.2.1 Self Injection Locking (SIL) using Electrical Feedback

Self-injection locking can be used to improve the phase noise of the existing oscillators. The analytical expression for SIL oscillator is given by,

$$L_{inj}(f') = \frac{L(f')}{(1 + \sqrt{P_{3dB}})^2}$$  \hspace{1cm} (2.2)

where $L_{inj}(f')$ is the phase noise of the self-injection locked oscillator; $L(f')$ is the phase noise of the free-running oscillator; $\rho$ is the injection power with respect to oscillation power; $f_{3dB}$ is the BW of the free-running oscillator and $\tau$ is the feedback delay. The analytical model indicates that if one increases the self-injection signal strength $\rho$ and the loop delay $\tau$ while keeping the loop phase to be 0 or $2\pi$, the phase noise of the oscillator at the noise offset frequency near the carrier can be reduced further. However, the longer...
loop delay $\tau$ requires the longer cable delay line, and it will be impractical to use a long electrical cable in the self injection locked oscillator to reduce the phase noise due to the high loss in electrical delay line. Figure 2.8 shows the block diagram of the self injection locked oscillator. The output of a MESFET oscillator passes through a circulator and is fed back to the oscillator through a delay element for self injection locking.

A coaxial cable is used to construct the feedback loop which provides a 15.7ns delay, as shown in Figure 2.9a. The experimental results of SIL oscillator are presented in Figure 2.9b. Different attenuators
are also selected to show the behavior of the phase noise under self-injection locking. The phase noise measurement results show qualitative agreement with the analytical expression. The phase noise departs from its ideal curves within some noise offset frequency ranges, which may be caused by the change of the oscillator output load.

2.2.2 Self Injection Locking using Optical Feedback

The electrical delay line length is limited by high loss for frequencies beyond several gigahertz. Optical fiber delay line, on the other hand, has very low transmission loss and is suitable for building self-injection locked oscillators. A self-injection locked oscillator using optical feedback is reported by Lee et al. [7], as can be seen in Figure 2.10. The electrical loop has sufficient gain to oscillate by itself. A part of output signals from the electrical oscillator, is injected into the oscillator after passing through a long optical delay line and it locks the electrical oscillator, achieving self-injection locking. The phase noise of this oscillator can still be expressed by (2.2).

![Figure 2.10 Schematic of Self Injection Locked Oscillator using Optical Delay in [7]](image-url)
Measured phase noise of this oscillator is presented in Figure 2.11. A 2.4 km long optical fiber is used in the loop, which will provide 12 μs delay. The oscillation frequency is at 30GHz. With the help of a long delay, a small injection power ($\rho = \sqrt{P_i/P_o} = 1.6 \times 10^{-3}$) is already providing a significant phase noise reduction of about 20 dB. Further phase noise reduction is also observed with higher injection power ($\rho = 2.5 \times 10^{-1}$).

![Figure 2.11 Phase noise of SIL Oscillator under Different Injection Power (adopted from [7])](image)

It is worth mentioning that in the case of $\rho = 2.5 \times 10^{-1}$, the measured phase noise -119dBc/Hz at 10kHz offset does not match up with the analytical result -137dBc/Hz obtained from (2.2). A possible explanation is the interaction between electrical oscillation and optoelectronic oscillation may degrade the overall system performance. When there is no power or the power is small in the feedback loop, the sidemode level is low, as can be seen in Figure 2.12a and 2.12b. When the injection power is strong, the circulating power in the optical loop is strong enough to sustain self-oscillation, shown in Figure 2.12c. This oscillation appears at offset frequencies given by,

$$\Delta f = \frac{1}{\tau} = \frac{c}{nl}$$  (2.3)
where $\tau$ is the loop delay time; $c$ is the speed of the light; $n$ is the effective refraction index of the fiber; $l$ is the length of the fiber. The interaction between the electrical oscillation and the optical oscillation may have a negative effect on spectral purity and this phenomenon will be studied more thoroughly in the later chapters.

![Figure 2.12 Sidemodes of SIL Oscillator under Different Injection Power (adopted from [7])](image)

### 2.2.3 Self Phase Locked Loop (SPLL) Oscillator using Fiber Optic Delay

The use of a long, stable, fiber optic delay at RF/microwave frequencies provides much greater sensitivity to frequency fluctuations of the electronic oscillator, compared to conventional coaxial delay-line implementations, resulting in greatly improved phase stability. The self-phase locked oscillator configuration is depicted in Figure 2.13a. Phase fluctuations of the voltage controlled oscillator (VCO) are converted to baseband voltage fluctuations by the fiber optic discriminator. This baseband signal is amplified, filtered, and fed back to the phase control port of the VCO to reduce the phase fluctuations. In the fiber optic discriminator, shown in Figure 2.13b, the RF input signal is split and compared to a delayed version of itself in a phase detector.
Analytical modeling of phase noise for SPL oscillator is given by,

\[ S_\phi(f') = \frac{S_\phi(f')}{(2\pi f')^2} \]  \hspace{1cm} (2.4)

where \( S_\phi(f') \) is the phase noise of the free running VCO; \( \tau \) is the feedback delay time; \( f' \) is the offset frequency. Similar to the self-injection locking case, a long delay is again needed for better phase noise reduction. Measured phase noise is shown in Figure 2.14. A phase noise reduction of about 12dB is achieved at 40kHz offset for a 100MHz carrier, for a commercially available frequency synthesizer when a 2.2km long fiber is used in the frequency discriminator. Once again, we can see the secondary peak associated with the long delay.

Figure 2.14 Measured Phase Noise of SPL Oscillator (adopted from [8])
2.2.4 Self Phase Locking with Multiple Optical Delays

In section 2.3.3, we see that SPLL is an effective way to minimize the phase noise of an oscillator. However, the long delay not only brings in the sidemode but also has a limited phase locking range. This is reported experimentally by Pillet et al. In their experiment, a SPLL (Figure 2.15) is constructed to stabilize the microwave output from a dual-frequency laser whose optical spectrum is shown in Figure 2.16. The DFL has 2 optical output frequencies that are 12 GHz apart and the beat frequency is the microwave output. The electrical error signal of the FD will control the optical phase of these 2 modes to eventually stabilize the microwave output.

![Figure 2.15 SPLL for Dual Frequency Laser](image1)

**Figure 2.15** SPLL for Dual Frequency Laser (adopted from [10]).

![Figure 2.16 Optical Spectrum of Dual Frequency Laser](image2)

**Figure 2.16** Optical Spectrum of Dual Frequency Laser (adopted from [10]).
Output microwave phase noise of DFL using self-phase locked loop is shown in Figure 2.17. When 1km long delay is used in the frequency discriminator, the PLL loop bandwidth is about 4kHz. To increase the locking range, SPLL structure of multiple frequency discriminators is also proposed. The configuration of multiple frequency discriminators is depicted in Figure 2.18. Two pieces of optical fiber is inserted into one arm of a standard frequency discriminator. One has a length of 100m which provides 0.5μs delay while the other is 1km long which provides 5μs delay. The overall output of the FD is equivalent to the sum of 2 individual FD with different delays.

![Figure 2.17 Phase Noise of the Microwave Output from the DFL. 1-phase noise of the self phase locked DFL; 2-noise originated from the optics; 3-noise originated from the electronics; 4-phase noise of the free running DFL. Results are adopted from [10].](image1)

![Figure 2.18 Configuration of Double Frequency Discriminator (τ₁ = 0.5μs; τ₂ = 0.5μs). Results are adopted from [10].](image2)
Experimental result of double frequency discriminator is given in Figure 2.19 as well as the results for single FD with 100m delay and 1km delay. The phase noise of double FD is lower than single FD with 100m delay and the locking range of double FD is better than single FD with 1km delay.

Figure 2.19 Phase Noise of: 1-Free Run; 2-100m Alone; 3-1km Alone; 4-100m and 1km Combined. Results are adopted from [10].

2.2.5 ILPLL Oscillator

Injection locking removes the phase noise in far away offset frequency while phase locking effectively reduces close-in to carrier phase noise. By combining IL and PLL, a clean spectrum in a wider range of offset frequency is expected. An Oscillator with ILPLL is reported by Sturzebecher et al [12]. Its configuration is depicted in Figure 2.20. The oscillator consists of two transistors with a feedback path. The reference signal, from a laser, is detected by a photo detector and fed to the oscillator through matching network M1 for injection locking. For phase locking, the reference signal is also compared with LO signal to create a low frequency signal to change the varactor diode bias in the feedback loop to provide phase correction.
The analytical modeling for phase noise of ILPLL oscillator is given below,

\[ S_{ILPLL}(f') = \frac{S_{ref}(f')\Delta f^2 \cos^2(\phi_{detune}) + f'^2 S_{PLL}(f')}{f'^2 + \Delta f^2 \cos^2(\phi_{detune})} \]  \quad (2.5) 

where \( S_{ref}(f') \) is the phase noise of the reference signal; \( \Delta f \) is the injection locking range. \( \phi_{detune} \) is the phase detuning between the LO signal and the reference signal; \( f' \) is the offset frequency; \( S_{PLL} \) is the phase noise of the oscillator when it is phase locked to the reference, and is given by

\[ S_{PLL}(f') = \frac{f_{n0}^4 S_{ref}(f') + f'^4 S_0(f')}{(f'^2 + f_{n0}^2)^2} \]  \quad (2.6)

where \( f_{n0} \) is the natural resonant frequency in the phase locked loop. Simulation results of the ILPLL oscillator phase noise at 10kHz offset using (2.5) and (2.6) are shown in Figure 2.21. The simulation results indicate that small phase detuning, high natural resonant frequency and high injection locking range is preferred to achieve a substantial phase noise reduction. More details of the impact of these parameters are to be discussed in chapter 4. Different implementation of ILPLL technique can also be found in [26]-[30].
2.3 Patents Relevant to Forced Frequency Stabilization Techniques

2.3.1 Controlled Frequency Signal Source Apparatus Including A Feedback Path For the Reduction of Phase Noise


The block diagram is depicted in Figure 2.22, a voltage controlled oscillator (VCO) 12 which is connected for supplying a periodic signal to a noise discriminator 16 that is connected with a gain unit 18. More specifically, the noise discriminator provides an error signal \( V_d = K_d \cos(\omega \tau + \phi_s) \), where \( K_d \) is a gain constant associated with the phase detector, \( \omega \) is the angular frequency of the VCO, \( \tau \) is the time delay exhibited by time delay unit 30 and \( \phi_s \) is the phase shift provided by variable phase shifter 32. A portion of the error signal \( V_d \) is amplified by the gain unit, whose output is \( V_n \). A voltage adder 22 adds \( V_n \) to an external dc voltage \( V_i \) to provide the control signal for the VCO. The phase shifter control 36 is configured to provide a control signal, \( V_p \), to cause the variable phase shifter 32 to supply the amount of phase that is necessary to establish the output signal provided by phase detector equal to zero at the VCO output frequency. This results in the center frequency of the discriminator being held precisely at the oscillator operating frequency.
The phase noise of the free running VCO is given as \( \phi_0(\omega) = \frac{V_\phi K_v}{\omega} \), where \( V_\phi \) represents a noise voltage having a flat frequency spectrum and a unit of volts/\( \sqrt{\text{Hz}} \), \( K_v \) is the VCO conversion gain normally in radians/(volts\times seconds). Using the phase noise reduction technique presented in this patent, the phase noise of the system is given as \( \phi(\omega) = \frac{V_\phi K_v G_0}{\omega} \), where \( G_0 = K_v K_S \tau A_0 + 1 \) represents the gain of the overall feedback system and \( A_0 = \frac{R_{46}}{R_{48}} \). \( R_{48} \) and \( R_{46} \) are the resistance in the gain unit depicted in Figure 2.22. The phase noise performances are simulated in Figure 2.23. We can see that between frequency range \( \omega_2/G_0 \) and \( \omega_0 G_0 \), the phase noise of the system is lower than the free running VCO. \( \omega_2 = \frac{K_0 K_S}{R_{46} C_{42}} \) where \( K_5 \) is a constant associated with the phase shifter and \( \omega_0 = \frac{1}{R_{48} C_{50}} \).
2.3.2 Widely Tunable Oscillator Stabilization Using Analog Fiber Optic Delay Line

(US Patent 5,204,640; Date: Apr. 20, 1993; Inventor: Ronald T. Logan; Assignee: California Institute of Technology, Pasadena, Calif.)

A circuit for stabilizing an oscillator having an oscillator output signal includes a long optical fiber delay line which receives an optical version of the oscillator output signal and a phase detector sensing the oscillator output signal and the delayed optical output signal through the long optical fiber. The output of the phase detector is fed back to the oscillator’s frequency control input to stabilize the frequency. The phase detector and the delay line are a delay line discriminator, and the length of the optical filter is selected so as to optimize the discriminator’s sensitivity against signal attenuation in the optical fiber, the optical fiber length typically being on the order of 10 kilometers. An adjustable phase shifter which maintains phase quadrature at the phase detector inputs at equilibrium is controlled by a tuning microprocessor responsive to an externally controlled change in the frequency at which the oscillator is to be stabilized, so as to follow changes in the selected oscillator frequency. As a result, an oscillator may be tuned in the circuit of the invention to any frequency in the general range of D.C. to optical frequencies.
2.3.3 Microwave Oscillator with Noise Degeneration Feedback Circuit


A microwave oscillator is shown to include an oscillator having an output and a control port and a feedback circuit disposed between the output and the control port of the oscillator. The feedback circuit includes a modulated laser, having an input and an output, the input responsive to a portion of a signal from the output of the oscillator and a photo detector having an input and an output, the input of the photo detector responsive to a signal from the output of the modulated laser delayed by a predetermined amount of time. The feedback circuit further includes a detector having a first and a second input and an output, the first input of the detector responsive to a signal from the output of the photo detector, the second input responsive to a portion of the signal from the output of the oscillator shifted in phase to be in phase quadrature with the signal at the first input of the detector and the output of the detector coupled to the
control port of the oscillator. With such an arrangement, a microwave oscillator having improved FM noise performance than known microwave oscillators is provided.

Figure 2.25 Block Diagram of the Invention from [33].

2.4 Problem Statement and Motivation for This Thesis Work

Even though sapphire loaded cavity oscillator (SLCO) provides excellent phase noise performance, SLC is sensitive to temperature change and mechanical vibration. In many cases, SLC is cooled at cryogenic temperature for optimum performance but cumbersome apparatus is required to achieve the cooling. If SLC oscillator is used in an airborne or ship based radar system, anti-vibration suspension is required to maintain the low phase noise, which increases the complexity of the system. OEO has higher immunity to
temperature variation and mechanical vibration, but due to the size and the high cost, it is more suitable to be used as bench-top unit for metrology applications rather than field applications. Metamaterial Mobius strip based oscillator is competitive for its compact size and low cost, but the phase noise at the moment is not as good as the reported SLCO and OEO. Frequency stabilization techniques using forced oscillations, such as SIL and SPPLL, have been demonstrated to be effective for phase noise reduction. However, search of the published literature and reported patent has revealed that SIL and SPPLL are only being used independently. As a result, SIL suffers from phase noise degradation in the close-in to carrier offset while SPPLL suffers degradation in the far-out offset.

In this thesis – to the best of my knowledge – SIL and SPPLL are proposed to be combined as SILPLL for the first time. Based on the reported ILPLL forced oscillations, it is proposed in this thesis that phase noise of the oscillation signal is enhanced as injection locking process of SIL removes phase noise in far-out offset frequencies, while the SPPLL process of forced oscillator effectively reduces the close-in to carrier phase noise. In SILPLL scheme, it is anticipated that the phase noise reduction is achieved utilizing optical fiber as energy storage element, which is similar to an OEO, in principle, SILPLL phase noise should reach that of an OEO. Dual loop SILPLL incorporating short and long delay is also proposed to suppress the side-modes associated with the long delay while maintaining the phase noise reduction provided by the long delay. Due to the advances in low noise electronics and broad bandwidth of the optical components used in the DSILPLL system, the DSILPLL technique has the potential to create highly stable RF oscillators approaching 100GHz. The goal in this thesis is to:

1. Evaluate forced SIL to achieve phase noise reduction of free running oscillators (VCO, OEO)
2. Identify optimum methods of SPPLL in terms of loop filter amplifier behavior in terms of natural frequency and damping factor of a second order type II loop
3. Compare performance of conventional electrical phase shifter based phase locking process with optical tuning of Mach Zehnder modulators
4. Explore analytical and experimental performance of SILPLL, while side-mode reduction techniques are also introduced using dual optoelectronic delay elements
5. Design fiber optic delay based frequency stabilization structure targeted for reaching the forced oscillator’s SSB phase noise of better than -140dBc/Hz at 10kHz offset for carrier frequencies of 10GHz and beyond.

The thesis has documented steps to develop understanding of forced oscillation dynamics and methods that will lead to optimum phase noise performance of oscillators. These results are presented in chapters 3, 4, and 5, while a number of appendices provide technical support for the opto-electronic oscillator behavior as well as forced oscillation techniques.
Chapter 3 Forced Oscillations Using Self-Injection Locking

Frequency stability of local oscillators is paramount in a number of coherent detection systems. Phase noise of oscillators can be reduced by forced oscillation processes, such as injection locking to an external low noise source [34]. The lowest phase noise achievable is determined by the noise of the external source in the case of conventional injection locking. However, in many cases we are reaching limits of stability of stable sources to lock free-running oscillators; and ultra-high stability and low noise sources are not readily available at microwave frequencies and beyond for future instrumentation systems. Self-injection locking (SIL) has been developed and demonstrated to be an effective method for phase noise reduction in [6] and [35]. SIL can be implemented by feeding part of the past oscillator output signal back to itself after passing through a delay line or resonator. It has also been shown that long delay or high quality factor (Q) is crucial for substantial phase noise reduction. Although it is possible to have phase noise reduction using an electrical delay, the improvement is poor because the delay length is limited due to high loss in the electrical delay lines or a limited Q of resonators at microwave frequencies [36]. To overcome the loss limitations of electrical delay lines, low loss fiber optic delay lines are proposed for the realization of SIL [7]. However, the side-modes associated with the long optical delay lines become undesirable since they appear as spurious oscillations at offset frequencies very close to carrier frequency and are hard to be removed using standard electrical filtering. Hence, a multi-loop configuration is proposed in opto-electronic oscillators (OEO) as a side-mode suppression scheme. There are two different topologies reported in multi-loop OEO: i) oscillation is not established in individual loops since the loop gain is kept to be smaller than unity in each loop, while the combined loop gain of all the loops is equal to or greater than one for joint oscillation [37]-[41]; ii) the other approach develops oscillation in each individual loops and the side-mode suppression is achieved using a coupled oscillation scheme [42]-[43]. However, in our novel approach the multi-loop implementation is different from the above mentioned topologies, as in our SIL approach using multi-loops we require oscillation in one loop using positive feedback and just coupling in other loops using negative feedback.
Even though dynamic modeling of injection locked oscillators (ILO) is studied by many authors [44]-[49], only conventional injection locking topologies are considered and the focus is on numerical computation of locking range and power spectrum. In this thesis, a system level analysis is presented for the phase noise of ILO within locking range, and topologies of both external IL and SIL with delays of at least $\mu$s are addressed in this paper. Experiments are performed to demonstrate the concept of SIL using fiber optic delay lines. This paper provides for the first time a comprehensive analytical modeling and experimental verification results of self-injection locking of an electrical oscillator in terms of close-in to carrier phase noise and performance of spurious oscillations using two optical delay loops. Moreover, this paper provides for the first time analytical modeling and experiment results for self-injection locked OEO using one and two optical delay lines in terms of close-in to carrier phase noise and spurious oscillation power levels. Discussions are also provided in terms of physical limitations of SIL technique.

3.1 Analysis of IL Phase Noise

3.1.1 Analysis of Conventional IL

In this section, a system level modeling is used as a unified model for phase noise modeling of oscillators with injection scheme. Since the phase dynamics of injection locking process is equivalent to that of first order type I phase-locked loop (PLL) [50], it is intuitive for us to derive the phase noise expression of ILO using PLL model. This approach is preferred as there is opportunity to extend modeling to PLL and implement a unified modeling of injection locked phase locked loop (ILPLL) oscillators [12], [26]-[30]. As part of the modeling, let $y_i$ and $y_o$ be the injecting signal and the output signal of the oscillator in the free running case, respectively:

$$y_i = \cos(\omega t + \varphi_i(t))$$  \hspace{1cm} (3.1)

$$y_o = \cos(\omega t + \varphi_o(t))$$  \hspace{1cm} (3.2)
Using (3.1) and (3.2) with the famous Adler’s equation [44], then phase dynamics of IL can be written as

\[
\frac{d\varphi_o(t)}{dt} = \rho \omega_{3dB} \sin(\varphi_i(t) - \varphi_o(t)). \tag{3.3}
\]

\(\rho = \sqrt{(P/P_\circ)}\) is the injection strength, \(\omega_{3dB}=\omega_r/2Q\) is half the 3dB bandwidth of the oscillator resonator, and \(\omega_r\) is the center frequency of the resonator. When the frequency difference between the injecting signal and the free running oscillator is small, the phase difference between them is also small. Thus we can linearize (3.3) to have

\[
\frac{d\varphi_o(t)}{dt} = K_I(\varphi_i(t) - \varphi_o(t)) \tag{3.4}
\]

where \(K_I = \rho \omega_{3dB}\). Equation (3.4) is in the same form of the phase dynamics of the first order type I PLL. Performing the Laplace transform of the above time domain variable, (3.4) can be expressed in s-domain as

\[
s\varphi_o(s) = K_I(\varphi_i(s) - \varphi_o(s)) \tag{3.5}
\]

The transfer function of phase of IL can be found as

\[
H_I = \frac{\varphi_o}{\varphi_i} = \frac{K_I}{s+K_I} = \frac{G_I}{1+G_I} \tag{3.6}
\]

where \(G_I = K_I/s\). From now on, the quantities in the s-domain will be expressed without showing explicitly the dependency on ‘s’ if no confusion is caused. For example, \(\varphi_o(s)\) will be written as \(\varphi_o\). The block diagram representing (3.6) is depicted in Figure 3.1a. We can see that IL resembles a negative feedback control loop that contains an integrator. The integrator is usually a voltage controlled oscillator.
(VCO), which is suitable for integration of PLL function. The loop behavior in the presence of noise sources is systematically presented in Figure 3.1b, where the major noise contributors in IL are $n_p$ at the injection point that contains injecting signal noise and residual noise of the system, and $n_q$, which is the oscillator phase noise. Note that in Figure 3.1b, the total output phase is the nominal phase $\phi_o$ and the phase error $\phi_n$ due to the presence of $n_p$ and $n_q$.

![Conceptual block diagram representation of IL using control theory representation; a) without noise sources present; b) with noise sources of $n_p$ and $n_q$ added in the loop, and $\phi_n = \phi_p + \phi_q$.](image)

Assuming the noises are a small perturbation to the steady locked oscillation, thus linearity still holds in the IL system. Then we can find the oscillator output due to noise using superposition principle. We first use standard loop analysis to find out the output phase error $\phi_p$ due to input noise $n_p$ only as

$$\phi_p = G_i (\phi_i - \phi_o) - G_i \phi_p + G_i n_p \quad (3.7)$$

In (3.7), the term $(\phi_i - \phi_o)$ represents a static phase error due to the frequency detuning between the reference signal and the oscillator. When the static phase error is large (e.g., $\phi_i - \phi_o > 80^\circ$), its impact on phase noise has to be accounted for. However in most of the cases, the static phase error is small due to limited injection locking range and it does not have significant impact on phase noise [11]-[12] therefore this term can be neglected in the remaining phase noise analysis. Rearranging (3.7) in terms of $\phi_p$ after neglecting the static phase error $(\phi_i - \phi_o)$, results in
\[ \varphi_p = \frac{g_I}{1 + g_I} n_p = H_I n_p. \] 

(3.8)

Similarly, the output phase error \( \varphi_q \) due to oscillator phase noise \( n_q \) only is expressed as

\[ \varphi_q = \frac{1}{1 + g_I} n_q = E_i n_q. \] 

(3.9)

\( E_i \) is termed as error transfer function of the loop. Then the total output phase error \( \varphi_n \) is found by combining both \( \varphi_p \) and \( \varphi_q \), and is given as

\[ \varphi_n = \varphi_p + \varphi_q = H_I n_p + E_i n_q. \] 

(3.10)

The power spectral density of the output phase error \( \varphi_n \) at offset frequency \( f_m \) becomes

\[ S_I(f_m) = |H_I(s)|^2 S_p(f_m) + |E_i(s)|^2 S_q(f_m) \] 

(3.11)

Where \( s = j2\pi f_m \) \( S_I(f_m) \) is the power spectral density of phase error \( \varphi_n \). \( S_p(f_m) \) and \( S_q(f_m) \) are the power spectral densities of \( n_p \) and \( n_q \) respectively. When the phase fluctuation is small, the single side-band (SSB) phase noise (denoted using \( \mathcal{L} \)) has a simple relationship with the power spectral density, and is given below as

\[ \mathcal{L}_I(f_m) = \frac{1}{2} S_I(f_m). \] 

(3.12)
3.1.2 Analysis of Self-Injection Locking (SIL)

a) Derivation of Oscillator Phase Noise with SIL

The conceptual block diagram of SIL using control theory representation is shown in Figure 3.2. A portion of the oscillator output signal experiences a time delay of $\tau$ and this delay is expressed in $s$-domain as $e^{-s\tau}$. The phase of the delayed signal is then compared against that of non-delayed signal to generate an error signal for self-injection to the oscillator similar to the one shown in Figure 3.1b. Note that terms of $\varphi_i$ and $\varphi_o$ are neglected for simplicity provided that they do not affect the overall phase noise.

![Image of conceptual block diagram of SIL](image)

Figure 3.2 Conceptual block diagram of SIL using control theory representation with a self-feedback time delay of $\tau$. Terms of $\varphi_i$ and $\varphi_o$ are neglected for simplicity.

The phase noise of the SIL can be found by using the same procedure as presented in section 3.1 for conventional IL. The only difference is that $n_p$ in this case does not contain the injecting signal noise but only the residual noise of the system. We can first find the output $\varphi_p$ due to $n_p$ as

$$\varphi_p = G_1(e^{-s\tau} - 1)\varphi_p + G_1n_p$$  \hspace{1cm} (3.13)

and hence after regrouping

$$\varphi_p = \frac{G_1}{1+(1-e^{-s\tau})G_1}n_p = H_{SI}n_p.$$ \hspace{1cm} (3.14)
$H_{SI}$ is the loop transfer function of the SIL system. Phase error $\varphi_q$ due to oscillator phase noise $n_q$ can be found in a similar fashion as:

$$\varphi_q = \frac{1}{1+e^{-sT}G_i} n_q = E_{SI} n_q.$$  \hfill (3.15)

$E_{SI}$ is the error transfer function of the SIL system. Hence the overall output phase error $\varphi_n$ is found as

$$\varphi_n = \varphi_p + \varphi_q = H_{SI} n_p + E_{SI} n_q.$$  \hfill (3.16)

Note that $H_{SI}$ and $E_{SI}$ have resonant peaks at offset frequencies of $f_m'$, where $f_m'$ is the integer multiple of $1/T$. The power spectral density of $\varphi_n$ in (3.16) becomes

$$S_{SI}(f_m) = |H_{SI}(s)|^2 S_p(f_m') + |E_{SI}(s)|^2 S_q(f_m)$$  \hfill (3.17)

The SSB phase noise for SIL can then be found as

$$L_{SI}(f_m) = \frac{1}{2} S_{SI}(f_m)$$  \hfill (3.18)

$S_p(f_m)$ is the system residual noise expressed as:

$$S_p(f_m) = \frac{kTBF}{P_s} \left( \frac{f_c}{f_m} + 1 \right)$$  \hfill (3.19)
$S_q(f_m)$ is expressed using Leeson’s equation [2]

$$S_q(f_m) = \frac{kTB\cdot F_o}{P_s} \left[ \frac{1}{f_m^2} \left( \frac{f_m^2}{4Q_L^2} \right) + \frac{1}{f_m^2} \left( \frac{f_m^2}{4Q_L^2} \right) + \frac{f_c^2}{f_m + 1} \right]$$  \hspace{1cm} (3.20)

where $k = 1.38 \times 10^{-23} \text{J/K}$ is the Boltzmann constant; $T = 290^\circ \text{K}$ is the room temperature in Kelvin; $B = 1 \text{Hz}$ is the noise bandwidth being considered; $F$ is the noise figure associated with the system residual noise; $F_o$ is the noise figure associated with the oscillator circuit; $P_s$ is the carrier power level; $f_c = 1 \text{MHz}$ is the flicker noise corner frequency, $f_o$ is the oscillation frequency; $Q_L$ is the loaded $Q$ of the oscillator resonator. For oscillator phase noise $S_q(f_m)$, a roll-off rate of 30 dB/decade is expected when $f_m < f_c$ and 20 dB/decade for $f_m > f_c$. A higher $Q_L$ provides lower SSB phase noise in the region where $f_m > f_c$. Advantage of opto-electronic oscillators is that $Q_L$ could be enhanced by increasing the fiber delay length.

**b) Simulated Phase Noise of Electrical Oscillator with and without SIL**

Single side-band phase noise simulation are provided below for an electrical oscillator with SIL using (3.18). In Figure 3.3a, the black dotted curve shows the SSB phase noise of a free running electrical oscillator (cf. section 3.1), whose practical values of loaded quality factor, $Q_L = 500$ and noise figure, NF=32dB are considered, even though dielectric resonator oscillators (DRO) with $Q_L = 2000$ and NF=18dB have already been developed, achieving an SSB phase noise of -111dBc/Hz at 10kHz offset [51]. The phase noise drops at a rate of 30dB/decade at offset frequencies till flicker corner frequency of about 500kHz. Other curves in various colors show the phase noise of SIL with various optical delays of 1km to 8km. From the simulation results, a 1km delay in SIL improves the phase noise by 22dB at 1kHz offset and the slope rolls off at a rate of 30dB/decade till corner frequency of 100kHz. A number of spurious oscillations are also manifested as side-modes of every 200kHz. When the delay increases to 8km the improvement is about 39dB with a slope of 30dB/decade till corner frequency of 10kHz and side-modes of every 25kHz. We can see that long delay is crucial for phase noise reduction. Figure 3.3b
shows the simulation results for 8km optical delay with various injection strengths. The green, blue and red curves show the phase noise of SIL with ρ=0.00316, 0.01 and 0.0316 (i.e., injection ratio of -50dB, -40dB, and -30dB), respectively. The best phase noise is achieved when the injection strength is the strongest at a level of -30dB.

Figure 3.3  a) Simulated SSB phase noise of electrical oscillator without (Black: free running electrical oscillator) and with SIL for ρ=0.0316 with different delays (Magenta: 1km; Green: 3km; Blue: 5km; Red: 8km). b) Simulated phase noise of 8km long SIL with different injection strengths. (Green: ρ=0.00316; Blue: ρ=0.01; Red: ρ=0.0316.) In all these plots the oscillator output power is $P_S=16\text{dBm}$, with fiber optic link NF=60dB, and $P_N=-114\text{dBm}$.

c) Simulated Phase Noise of OEO with and without SIL

An electrical oscillator is replaced by an opto-electronic (OEO) oscillator with fiber delay line of 1km for achieving high loaded quality factor. By using (3.14), phase noise of OEO with SIL is also simulated. Figure 3.4 shows the phase noise of a standard OEO and SIL OEO with various feedback delays when ρ=0.0316. The phase noise of this standard OEO is shown in black dashed curve; other colored curves show the phase noise of SIL OEO with different optical delays in the feedback loop. The simulated phase noise is the lowest in the case of 8km delay. Phase noise performance for OEO with 8km SIL delay remains the same under injection strengths of 0.00316, 0.01 and 0.0316. The spikes that appear in Figure
3.4 are the poles associated with the transfer functions of different delays, and they may not represent the actual location and level of the spurious signals.

![Simulated phase noise of a standard OEO realized using a 1km long fiber delay without (Black: standard OEO) and with SIL (Green: standard OEO with 3km SIL; Blue: standard OEO with 5km SIL; Red: standard OEO with 8km SIL). The SIL is accomplished for injection ratio of \( \rho = 0.0316 \) for different fiber delay lengths. (\( P_s = 16\text{dBm}, \text{fiber optic link NF}=60\text{dB}, P_N=-114\text{dBm} \)).](image)

3.1.3 Analysis of Dual Loop Self-Injection Locking (DSIL)

From the previous simulations, longer delay is required for better phase noise reduction in SIL configuration. However, the side-modes originated from the spurious oscillation in long delay lines appear as prominent noise sources at offset frequencies very close to the carrier. These side-mode levels are very high because the frequency selectivity of practical RF filtering is not high enough to remove these side-modes. A feasible way to suppress these spurious signals is by implementing two different non-harmonically related delays in the feedback path. Because the two delays have different mode spacing, only those modes that are common to both delays will survive. Other modes are hence being suppressed. This alternative filtering is similar to transversal optical filters reported by others [52]-[55].
a) Derivation of Oscillator Phase Noise using DSIL

The control theory representation of a dual loop SIL (DSIL) configuration is shown in Figure 3.5. In this case, the feedback is split into two separate paths, one passes through a long delay and another passes through a short delay. Phases of two different delayed signals are compared with the current signal separately and error signals are injected to oscillator. When the loop lengths are not harmonically related, then the resultant output becomes the sum of individual loop actions as periodic resonances are suppressed in strength.

The system level modeling is employed to derive the phase noise of DSIL. Assumption is made that the two injection loops do not interact with each other, so that the output is a superposition of the actions of individual loops. The residual noise of the system is also going to be the superposition of the noises in each loop (i.e. \( n_p = n_1 + n_2 \)). Then, the output \( \Phi_{o1} \) due to noise \( n_1 \) is expressed as:

\[
\Phi_p = -G_I \Phi_p + e^{-\varepsilon \tau_1} G_I \Phi_p - G_I \Phi_p + e^{-\varepsilon \tau_2} G_I \Phi_p + G_I n_p
\]

(3.21)

When deriving (3.21), the signal levels are assumed to be the same in both paths so that the loop gain \( G_I \) is identical for both paths. When the signal levels are not identical, different loop gains have to be used for each path, which complicates the analysis. Fortunately, it is easy to adjust the power level in the delay paths using a variable attenuator hence the equal power assumption is valid in most of the cases.

Rearrange (3.21), we have

\[
\Phi_p = \frac{G_I}{1 + G_I(1 - e^{-\varepsilon \tau_1}) + G_I(1 - e^{-\varepsilon \tau_2})} n_p = H_{DSIL} n_p
\]

(3.22)

where \( H_{DSIL} \) is the transfer function of the DSIL system. Similarly the phase error \( \Phi_q \) due to VCO phase noise \( n_q \) is found as
where \( E_{DSIL} \) is the error transfer function of the DSIL system. Finally, total phase error \( \varphi_n \) for DSIL is expressed as:

\[
S_{DSIL}(f_m) = |H_{DSIL}(s)|^2 S_p(f_m) + |E_{DSIL}(s)|^2 S_q(f_m)
\]  

(3.24)

where \( S_p(f_m) \) and \( S_q(f_m) \) are defined as in (3.19) and (3.20).

---

**Figure 3.5** Conceptual block diagram of DSIL using control theory representation. \( n_i \) for \( i=1, 2 \) are noise sources associated with each delay line being fed back to injection port of the oscillator and \( n_2 \) is output noise power.

---

*b) Simulated Phase Noise of Electrical Oscillator with DSIL*

Simulation using (3.24) for electrical oscillator with DSIL is depicted in Figure 3.6(a). The parameters for the electrical oscillator are kept the same as in the previous simulation. The two delays in the DSIL configuration are 1km and 8km. By combining two delays, the phase noise is maintained at the same level of 8km SIL while the spurious level is reduced by about 25dB compared to 8km SIL. Phase noise performances for DSIL electrical oscillator with various length combinations are also simulated, and the simulated results are provided in Figure 3.6b. The SSB phase noise level of DSIL is determined by the longer delay in the loop and the spurious level is determined by proper selection of length combinations. The simulated SSB phase noise result is superior for a combination of 5km and 8km delay elements.
c) Simulated Phase Noise of OEO with DSIL

Simulations for DSIL in a standard OEO are provided in Figure 3.7. The combinations of extra delay with a longer delay of 8km provide superior phase noise than the combination with shorter delay of 5km. This result is also intuitively expected.

![Figure 3.6](image1.jpg)  ![Figure 3.7](image2.jpg)

**Figure 3.6** a) Comparison of the simulated phase noise of an electrical oscillator using different injection topologies (Red: Dual-SIL 1km and 8km; Green: SIL 1km; Blue: SIL 8km). b) Simulated phase noise of DSIL with various combinations. Black: DSIL 100m and 8km; Magenta: DSIL 500m and 8km; Green: DSIL 1km and 8km; Blue: DSIL 3km and 8km; Red: DSIL 5km and 8km. \((P_S=16\,\text{dBm},\,\text{fiber optic link}\,NF=60\,\text{dB},\,P_N=-114\,\text{dBm})\).

**Figure 3.7** Simulated phase noise and spurious signal levels of a standard 1km OEO with DSIL. Red: 3km + 5km; Blue: 3km + 8km; Green: 5km + 8km. \((P_S=16\,\text{dBm},\,\text{fiber optic link}\,NF=60\,\text{dB},\,P_N=-114\,\text{dBm})\).
3.2 Experiment Results of SIL Electrical Cavity Oscillator

3.2.1 Electrical Cavity Oscillator Realization

The electrical oscillator consists of a band-pass filter (BPF) constructed using a metallic cylindrical resonant cavity with \( Q = 2500 \) at 10GHz and a power amplifier (Amp) from B&Z (BZ3-09801050-602422-102020) with small signal gain of 27dB and 1dB compression level of 24dBm. The metallic cylindrical resonant cavity unloaded Q factor of 2500 was estimated from the injection locking of this electrical oscillator. The oscillation power is coupled to an extremely low phase noise spectrum analyzer (R&S FSUP26) for SSB phase noise measurement (cf. Figure 3.8). The oscillation frequency is 9.818GHz and the carrier power level is 16dBm. The measured phase noise of this oscillator is shown in the black curve of Figure 3.9a. The phase noise of the electrical oscillator is -58dBc/Hz at 1kHz offset and -81dBc/Hz at 10kHz offset with a roll off rate of about 30dB/decade after 10kHz offset carrier. The flicker corner frequency is estimated to be about 1MHz and the noise figure is approximately 32dB. The loaded Q factor is about \( Q_L = 500 \) based on the measured phase noise. The free running phase noise for this oscillator is poor because of relatively low Q resonator characteristics of this metallic cylindrical resonant cavity.

3.2.2 SIL Phase Noise

The block diagram for electrical oscillator with SIL is depicted in Figure 3.8. The oscillator is controlled using an opto-electronic delay line by driving a Mach-Zehnder modulator (MZM) from JDSU (MN21024083) with electrical oscillator signal of 16dBm at 10GHz. The optical input power of 16dBm to the MZM is provided by a DFB laser from Mitsubishi (FU-68PDF-510M67B) whose output is amplified by an EDFA from Nuphoton (NP2000). The modulated optical output of the MZM passes through an optical delay and is detected by a photodetector from Discovery Semiconductor (DSC50S), and the received RF signal is amplified by a 24dB low noise amplifier from Kuhne Electronic (101A & 101B) and fed back to the oscillator for SIL. The measured NF is 58dB while the calculated NF is 60dB
which matches well with the measured data and these values are used for analytical predictions. The experimental results of the close-in to carrier phase noise of the SIL agree well with the analytical modeling predictions. The impact of various delay lengths on the close-in to carrier phase noise was evaluated and the measured phase noise of the 10GHz electrical oscillator is shown in Figure 3.9a. In the experiment setup, the injection strength $p$ is kept at 0.0316 for different delays of 1km to 8km. The phase noise for 1km delay improves by 20dB to -101dBc/Hz at 10kHz offset; in the case of 8km delay, the phase noise is -94dBc/Hz at 1kHz offset and -118dBc/Hz at 10kHz offset, corresponding to an improvement of 37dB with respect to the free running case. The diamonds in Figure 3.9a are the actual spurious levels for the first dominant spurious side-mode associated with the different delays. Note that the diamonds are measured in unit of dBc using super-heterodyning function of R&S FSUP26 while the solid curve are measured in unit of dBc/Hz using PLL function of R&S FSUP26. As the delay becomes longer, the spurious side-mode moves closer to the carrier frequency and the level becomes higher. For 8km delay, the spurious signals are located at offset frequencies of every 25kHz, and the level is -42dBc for the first side-mode. This spurious level is undesirable for a signal generator and is to be reduced. The effect of different injection strengths is shown in Figure 3.9b. In this case, the delay is fixed at 8km. Three different levels of injection strength with increment of 10dB are considered. The strongest injection strength is -30dB compared to carrier power, and we can see that the phase noise under the strongest injection is 20dB lower than the case of weakest injection of -50dB. However, the spurious signal level for weakest injection is 10dB lower than the case of strongest injection.
Figure 3.8 Block diagram of electrical cavity oscillator with SIL.

Figure 3.9 a) Experimental measurement of SSB phase noise and spurious signal levels of SIL for $\rho=0.0316$ with different delays (Magenta: 1km; Green: 3km; Blue: 5km; Red: 8km. Magenta Diamond: 186699Hz, -60dBc; Green Diamond: 66145Hz, -50dBc; Blue Diamond: 40072Hz, -45dBc; Red Diamond: 25147Hz, -42dBc). b) Experimental measurement of SSB phase noise and spurious signal levels of SIL for 8km delay with different injection strengths (Green: $\rho=0.00316$; Blue: $\rho=0.01$; Red: $\rho=0.0316$. Green Diamond: 22576Hz, -53dBc; Blue Diamond: 24733Hz, -50dBc; Red Diamond: 25147Hz, -42dBc). Oscillator characteristics are $P_S=16$dBm, $NF=32$dB and optical link $NF=60$dB for both cases.

3.2.3 DSIL Phase Noise

The block diagram for the electrical cavity oscillator with DSIL is depicted in Figure 3.10. For DSIL, the output of the MZM is split into two paths with different delays using 50% optical coupler from Newport (F-CPL-S12155). The optical signals are detected by two identical optical receivers from Discovery Semiconductor (DSC50S); the received signals of two different delays are then combined and fed back to the oscillator as dual loop SIL after amplification by low noise amplifier of 24dB gain from Khune.
Electronic (101A & 101B). The estimated fiber optic noise figure is 60dB and corresponds to noise power level of -114dBm.

![Figure 3.10 Block diagram of electrical cavity oscillator with dual loop SIL.](image)

Experiment result for electrical oscillator with DSIL is reported for the first time (cf. Figure 3.11). Three different optical delay line combinations are used, the phase noise for ‘3km+8km’ case and ‘5km+8km’ case are practically the same while ‘1km+8km’ case suffers a 5dB degradation at 1kHz offset. One interesting phenomenon for DSIL is that the location and level of the first major spurious mode are all different in three cases. Neither do they appear at 25kHz (mode spacing related to 8km) nor do they appear at 200kHz (mode spacing related to 1km). In fact, they appear at 175kHz, 126kHz and 75kHz offsets respectively. The spurious suppression is significant in DSIL as we can see all the spurious signal moves away from the carrier and the levels are dropped to below -60dBc which is more than 20dB lower than SIL. Moreover, the spurious signals associated with 8km delay that would have appeared at offset frequencies about 25kHz, 50kHz and 75kHz with high power level are being filtered-out and somewhat suppressed by the shorter loops. The dominant side-modes of 25 kHz, 50kHz and 75kHz now appear as humps with a reduced side mode level as seen in the measurement results of Figure 3.11. Therefore a more dominant side mode is observed at 126 kHz. Nonetheless, the mechanism for mode selection in DSIL...
needs to be further explored analytically for optimum delay length selection in each fiber optic delay lines of dual loop SIL systems.

![Figure 3.11](image)

Figure 3.11 Experimental measurement of SSB phase noise and spurious signal levels of DSIL with different fiber length combinations (Green: 1km and 8km; Blue: 3km and 8km; Red: 5km and 8km; Green Diamonds: 175447Hz, -73dBc; Blue Diamonds: 125753Hz, -68dBc; Red Diamonds: 75879Hz, -64dBc). Oscillator characteristics are $P_s=16\text{dBm}$, $NF=32\text{dB}$ and optical link $NF=60\text{dB}$.

### 3.3 Experiment Results of SIL OEO

#### 3.3.1 OEO Realization

The block diagram for standard OEO configuration is depicted in the dashed black box of Figure 3.12, where the metallic resonant cavity is employed as a narrowband band-pass filter. Two low noise amplifiers (LNA) and two power amplifiers (Amp) are used to compensate for the RF signal loss in the MZM link. The power amplifiers employed here have very high 1dB compression points to achieve higher oscillator output power while in the current form of the OEO a 1 dB compression is experienced in MZM at a lower power level than the power amplifiers. Hence the saturation is due to the MZM but not the amplifiers. The fiber optic delay line of 1km long is selected for the OEO and the RF filtering is performed using the metallic cylindrical resonant cavity with unloaded $Q$ of 2500. The measured oscillation is at 10GHz and output power of this standard OEO is 16dBm. Phase noise performance for this standard OEO is shown in the black curve in Figure 3.13. The measured phase noise is -83dBc/Hz.
and -109dBc/Hz at 1kHz and 10kHz offset, respectively. The black diamond shows the actual location of the first spurious signal at 197kHz offset and the level is -40dBc. The measured performance is significantly better than the electrical feedback oscillator presented in section 3.1 and somewhat resembles the performance of electrical oscillator with SIL using a 1km long fiber delay line in section 3.2 (cf. Figure 3.9).

3.3.1 SIL Phase Noise

The block diagram of SIL OEO is depicted in Figure 3.12. The output of the MZM is split into two parts; one passes through a 1km delay and another passes through a longer delay and a 10dB optical attenuator. The RF gain is sufficient to compensate for the loss in the 1km loop, but not the second delay; hence optoelectronic oscillation will take place in the 1km loop and not at the longer delay. Because of the optical attenuator, the longer delay will not get enough gain to oscillate thus it forms a SIL to the OEO. Phase noise for SIL OEO is shown in Figure 3.13 for various long delays. Delays of 3km, 5km and 8km are selected, and the lowest phase noise is achieved using an 8km delay as predicted (cf. Figure 3.4 at levels of -96dBc/Hz at 1kHz offset (13dB lower than standard OEO) and -118dBc/Hz at 10kHz offset (9dB lower than standard OEO). The spurious level of -69dBc for 8km delay is also the lowest which is 29dB lower than standard OEO.
Figure 3.12 System block diagram of 1km long standard OEO with SIL using various fiber delay lengths. Optical attenuation provided by block of “Atten.” assures a negative feedback using the longer delay and the 10dB optical attenuation results in $\rho \approx 0.06$.

Figure 3.13 Experimental measurement results of SSB phase noise of a standard OEO (Black: standard 1km OEO) with various SIL lengths (Green: 3km; Blue: 5km; Red: 8km) and spurious signal levels (Black Diamond: 196627Hz, -40dBc; Green Diamond: 199645Hz, -60dBc; Blue: Diamond: 200485Hz, -63dBc; Red Diamond: 201002Hz, -69dBc). The OEO electrical characteristics are $P_S = 16\text{dBm}$, $P_N = -114\text{dBm}$ and $NF = 60\text{dB}$. 
3.3.2 DSIL Phase Noise

The concept of dual loop SIL is employed to reduce the spurious signal levels. Experimental setup for OEO with DSIL is conceptually depicted in Figure 3.14. The OEO portion of the setup is the same as in the SIL case. In the feedback path, the optical signal is further split into two branches using 50% optical coupler from Ascentta (CP-S-15-20-22-XX-S-L-10-FA). The delay lengths in the branches are both selected to be longer than the delay in the standard OEO. The signals from the two branches are combined at the same photodetector using another 50% optical coupler from Ascentta (CP-S-15-20-22-XX-S-L-10-FA). Optical attenuator is omitted since the insertion loss of optical couplers is high enough to prevent optoelectronic oscillation in the feedback path. In order to limit the injection strength, RF signal in the OEO is amplified before combining with the feedback signal.

Phase noise for OEO with DSIL is shown in Figure 3.15 as different lengths of delay lines are used in DSIL loops. Various length combinations are selected to experimentally evaluate the phase noise behavior of DSIL OEO and the level of spurious signals. The phase noise for different combinations remains almost the same in close-in offset range but there is a noticeable noise floor variation among the cases. For example, the noise floor for the ‘5.5km+8km’ case is about -108dBc/Hz while it is -122dBc/Hz for the ‘3.5km+8km’ case. The first spurious signal for different length combinations are tabulated in Table 3.1. Spurious level of ‘3.5km+8km’ combination is significantly lower than those of other combinations as the performance is also better in terms of close-in to carrier phase noise.
Figure 3.14 System block diagram of 1km long standard OEO with DSIL optical delay lines. Optical power levels for short optical delay compared to the longest and longer delay lines are indicated in the figure.

Figure 3.15 Experimental results of SSB phase noise of the OEO with various fiber length combinations of DSIL for $P_s=16$dBm.

In order to better understand the behavior of DSIL OEO, the side-mode spacing for different delays are tabulated in Table 3.2. We can see that 1km, 3km, 5km and 8km long fiber all have common modes at 200kHz while 3.5km and 5.5km long fiber don’t have modes at 200kHz, hence the spurious level for these 2 cases are lower than other cases, especially for 3.5km fiber. The phase noise of 5.5km fiber is poor; possible reason is that the modes of 5.5km fiber are too close to those of 8km fiber.
### Table 3.1  Locations and signal levels of the first spurious for different length combinations of DSIL

<table>
<thead>
<tr>
<th>DSIL Delay Combinations</th>
<th>3km + 5km</th>
<th>3km + 8km</th>
<th>3.5km + 8km</th>
<th>5km + 8km</th>
<th>5.5km + 8km</th>
</tr>
</thead>
<tbody>
<tr>
<td>Spurious Freq. (Hz)</td>
<td>197,658</td>
<td>196,791</td>
<td>196,644</td>
<td>197,147</td>
<td>196,215</td>
</tr>
<tr>
<td>Spurious Level (dBc)</td>
<td>-34</td>
<td>-36</td>
<td>-48</td>
<td>-35</td>
<td>-38</td>
</tr>
</tbody>
</table>

### Table 3.2  Approximate side-mode spacing for different fiber delay lengths.

<table>
<thead>
<tr>
<th>Fiber Length</th>
<th>1km</th>
<th>3km</th>
<th>3.5km</th>
<th>5km</th>
<th>5.5km</th>
<th>8km</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mode Spacing</td>
<td>200kHz</td>
<td>66.7kHz</td>
<td>57.1kHz</td>
<td>40kHz</td>
<td>36.4kHz</td>
<td>25kHz</td>
</tr>
</tbody>
</table>

### 3.4 Conclusion and Discussions

This paper provided analytical modeling and experimental measurements of SSB phase noise of forced electrical cavity oscillator and OEO using SIL and DSIL techniques. These results provide insights into understanding of forced oscillation behavior in general and are very attractive in performance for optically realized stable clock signals. These stable clocks will play an important role in many coherent communication and target tracking radar systems. Comparison of SSB close-in to carrier phase noise for different electrical cavity oscillator configurations is rendered in Table 3.3. The simulated results agree well with the actual measurement results. Moreover, it is clear from Table 3.3 that SIL and DSIL are effective for phase noise reduction in electrical oscillator. Comparison for the spurious levels is provided in Table 3.4. The DSIL provided a 22dB reduction in terms of the spurious levels, while it has pushed to the 3rd harmonic at 75kHz.

### Table 3.3  Comparison of SSB phase noise for the electrical oscillator with different circuit configurations.

<table>
<thead>
<tr>
<th>Phase Noise Comparison</th>
<th>Simulated (dBc/Hz)</th>
<th>Measured (dBc/Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Offset-frequency( f_m)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>1kHz</td>
<td>1kHz</td>
<td>1kHz</td>
</tr>
<tr>
<td>10kHz</td>
<td>10kHz</td>
<td>10kHz</td>
</tr>
<tr>
<td>Cavity Osc.</td>
<td></td>
<td></td>
</tr>
<tr>
<td>-52</td>
<td>-82</td>
<td>-58</td>
</tr>
<tr>
<td>-81</td>
<td></td>
<td></td>
</tr>
<tr>
<td>SIL 8km</td>
<td></td>
<td></td>
</tr>
<tr>
<td>-93</td>
<td>-120</td>
<td>-94</td>
</tr>
<tr>
<td>-118</td>
<td></td>
<td></td>
</tr>
<tr>
<td>DSIL 5km+8km</td>
<td></td>
<td></td>
</tr>
<tr>
<td>-93</td>
<td>-120</td>
<td>-92</td>
</tr>
<tr>
<td>-116</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
Table 3.4 Measured dominant spurious signals for the electrical oscillator with different circuit configurations.

<table>
<thead>
<tr>
<th>Spurious Comparison</th>
<th>Freq. (Hz)</th>
<th>Level (dBc)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SIL 8km</td>
<td>25147</td>
<td>-42</td>
</tr>
<tr>
<td>DSIL 5km+8km</td>
<td>75879</td>
<td>-64</td>
</tr>
</tbody>
</table>

On the other hand, comparison of phase noise performance for OEO with different circuit configurations is rendered in Table 3.5. For the case of DSIL case, the simulated results do not agree well with the measured results. Possible reason is that the modes in the two feedback branches interact with each other hence the noise floor becomes much higher. Comparison for spurs of different OEO configuration is given in Table 3.6. The higher spurious level in the DSIL case is understandable since DSIL has much higher noise floor. The mode competition between the side-modes of 8km and 5km delays degrades the spectrum purity of forced oscillation for OEO. If so, approaches that mode-lock these spurious frequencies have to be explored [56]-[60], which will reduce the mode-partition noise contributions [61].

Table 3.5 Comparisons of phase noise for OEO with different circuit configurations

<table>
<thead>
<tr>
<th>Phase Noise Comparison</th>
<th>Simulated (dBc/Hz)</th>
<th>Measured (dBc/Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1kHz</td>
<td>10kHz</td>
</tr>
<tr>
<td>Standard OEO</td>
<td>-92</td>
<td>-112</td>
</tr>
<tr>
<td>SIL 8km</td>
<td>-92</td>
<td>-120</td>
</tr>
<tr>
<td>DSIL 5km+8km</td>
<td>-92</td>
<td>-118</td>
</tr>
</tbody>
</table>

Table 3.6 Measured Spurious for OEO with different circuit configurations.

<table>
<thead>
<tr>
<th>Spurious Comparison</th>
<th>Freq. (Hz)</th>
<th>Level (dBc)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Standard OEO</td>
<td>1,966,627</td>
<td>-40</td>
</tr>
<tr>
<td>SIL 8km</td>
<td>201,002</td>
<td>-69</td>
</tr>
<tr>
<td>DSIL 5km+8km</td>
<td>196,644</td>
<td>-48</td>
</tr>
</tbody>
</table>

To further improve the overall performance of SIL and DSIL for electrical cavity oscillator or OEO with long fiber optic delay lines, it is important to reduce the overall amplitude noise of the fiber optic delay lines. From the measured data, the noise floor of the optoelectronic system is around -114dBm/Hz,
corresponding to a NF of about 60dB, which is typical for an external modulated optical link with the
commonly experienced optical power of 10dBm, \( V_x = 6V \) of Mach-Zehnder modulator (MZM), and RIN
of \(-140\text{dB/Hz}\). However, the NF could be reduced by increasing the optical power, using a lower \( V_x \)
MZM, and limiting the optical source laser RIN\[62\]-[63]. Figure 3.16 shows the phase noise of SIL OEO
with different NF values, where the best performance is observed at a minimum reported noise figure of
4dB\[64\]-[66]. It is clear from the figure that phase noise is reduced because of a lower noise floor in the
system. On the other hand, the RF link loss of the fiber optic link is about 47dB and to compensate for the
excessive loss, more than one stages of amplification are required to maintain operation at appropriate
power level. Even though low noise amplifiers are used in the amplification chain, an increase in the
overall NF of the system is experienced. A proper selection of amplifiers will definitely reduce AM-PM
conversion of the OEO. Moreover, the need for electronic amplifiers can be minimized by incorporating a
higher optical power source, where a better phase noise is expected [67].

![Figure 3.16 Simulated SSB phase Noise for injection power ratio of 10dB of SIL OEO 8km with different values of noise figure for fiber optic link delays (Black: NF=60dB; Green: NF=30dB; Blue: NF=15dB; Red: NF=4dB).](image)

Another mechanism for further phase noise improvement is phase locking. The phase of the modes
associated with the long delay is not locked through injection locking and only the frequencies are locked
to one another. In chapter 4, self-phase locked loop is explored to improve the close-in to carrier phase
noise.
Chapter 4 Forced Oscillations Using Self-Phase Locked Loop

SPLL was first proposed by Underhill [68] using a phase alternating line (PAL) delay line for a low frequency oscillator working in 3-6 MHz. Shortly after, Aitchison and Batliwala [69] built a stabilization circuit using WG16 waveguide to provide the delay and demonstrated that the principle established by Underhill could be extended to microwave frequencies. Contrary to conventional PLL where error signal is obtained by comparing against an external reference, error signal in SPLL is generated from a delay line frequency discriminator (DLFD) [70]. Longer delay provides higher discriminator sensitivity that is crucial for substantial phase noise reduction. However in [68]-[69], the delay is limited due to the loss of the cable and the dimension of the waveguide, therefore the overall reduction is limited. To bypass the limit, optical fiber delay lines with extremely low loss have been used to implement the SPLL, resulting in improved performance of close-in to carrier phase noise [8]-[10], [71].

In this chapter, SPLL using optical delay has been applied for phase noise reduction of microwave oscillations in an OEO and an electrical VCO. To fulfill the SPLL function, a tunable element is required to provide the frequency adjustment. Conventional approach for frequency tuning of OEO can be achieved by inserting electrical phase shifter in the loop to change the electrical phase [72]-[74]. In contrast to the conventional approach, optical phase shifting is considered in this thesis and it provides frequency tuning of OEO by changing the operation point of the MZM away from quadrature biasing point. Phase noise performance of SPLL OEO using the electrical phase control and optical phase control is compared and evaluated for the first time. In this chapter a comprehensive analytical modeling for various methods of self-phase locking are introduced in section 4.1 for both single and multiple self-phase locked loops. Moreover, performance improvements are studied in terms of loop filter performance, natural resonance frequency and damping factors of type II phase locking circuits. The experimental evaluation of self-phase locking are presented in section 4.2 using a VCO, in section 4.3 using a conventional tunable phase shifter, and in section 4.4 for a MZM based optical phase shifting. The performance comparisons of the experimental and analytical results are summarized in section 4.5.
4.1 Phase Noise Analysis of Self-Phase Locking

4.1.1 Conventional PLL

Derivation of the modeling is based on the time domain equation of a PLL, and it is given in [75] as

\[
\frac{d\varphi_o(t)}{dt} = K_d K_o K_f(t) \sin(\varphi_i(t) - \varphi_o(t))
\]  

(4.1)

where \( \varphi_i(t) \) and \( \varphi_o(t) \) are the phases of input signal and output signal respectively; \( K_d \) is the phase detector gain in V/rad, \( K_o \) the VCO tuning sensitivity in rad/V, and \( K_f(t) \) the impulse response of the low-pass filter. The symbol ‘\(*\)’ represents the convolution product. The static phase error \( \varphi_i(t) - \varphi_o(t) \) can be assumed to be a small quantity without loss of generality, then (4.1) can be linearized as

\[
\frac{d\varphi_o(t)}{dt} = K_p(t) [\varphi_i(t) - \varphi_o(t)]
\]  

(4.2)

where \( K_p(t) = K_d K_o K_f(t) \). The above time domain equation is converted into s-domain by Laplace transform, resulting in

\[
s \varphi_o = K_p (\varphi_i - \varphi_o)
\]  

(4.3)

Then we can find the transfer function of the PLL as

\[
H_p = \frac{G_p}{1 + G_p}
\]  

(4.4)

where \( G_p = \frac{K_p}{s} \).
From the system transfer function, we can represent the PLL system in frequency domain as shown in Figure 4.1. Assuming the noise is a small perturbation to the steady state solution, thus the linearity still holds, then overall output phase error can be found using superposition principle similar to the procedure used in section 3.1. Let’s first find out the output phase error \( \varphi_p \) due to the input noise \( n_p \). Using basic loop analysis, we have

\[
\varphi_p = G_p(\varphi_i - \varphi_o) - G_p\varphi_p + G_p n_p
\]

(4.5)

The static phase error \( \varphi_i - \varphi_o \) is usually very small for a second order type II PLL system [75], which is also the particular PLL system being considered in this thesis, therefore the static phase error can be neglected. Thus (4.5) can be simplified and regrouped as

\[
\varphi_p = \frac{G_p}{1+G_p} n_p = H_p n_p
\]

(4.6)

Similarly, we can find the output phase error \( \varphi_q \) due to the VCO noise \( n_q \) as

\[
\varphi_q = \frac{1}{1+G_p} n_q = E_p n_q
\]

(4.7)

where \( E_p \) is the error transfer function of PLL system. The overall phase error \( \varphi_n \) can be found by combining \( \varphi_p \) and \( \varphi_q \) as

\[
\varphi_n = \varphi_p + \varphi_q = H_p n_p + E_p n_q
\]

(4.8)
The overall output noise spectrum becomes

\[ S_p(f_m) = |H_p(s)|^2 S_p(f_m) + |E_p(s)|^2 S_q(f_m) \]  

(4.9)

where \( S_p(f_m) \) and \( S_q(f_m) \) are defined in (3.19) and (3.20). Then the SSB phase noise is given as

\[ L_p(f_m) = \frac{1}{2} S_p(f_m) \]  

(4.10)

Figure 4.1 Control theory representation of conventional PLL in the presence of noise.

### 4.1.2 Analysis of Self-Phase Locked Loop (SPLL)

The control theory representation is shown below in Figure 4.2 for SPLL. A portion of the VCO output is being delayed and the phase of the delayed signal is compared against the phase of the current signal. Again, we can use superposition principle to find out the overall noise of the SPLL system. Using standard loop analysis, we first find out the output phase error \( \varphi_p \) due to the input noise \( n_p \) as

\[ \varphi_p = G_p(e^{-st} - 1)\varphi_p + G_p n_p \]  

(4.11)
Rearrange (4.11) to become

\[ \varphi_p = \frac{g_p}{1 + (1 - e^{-sT})g_p} n_p = H_{SP} n_p \]  

(4.12)

where \( H_{SP} \) is termed as transfer function of SPLL. Phase error \( \varphi_q \) due to VCO noise \( n_q \) can be found in a similar fashion as

\[ \varphi_q = \frac{1}{1 + (1 - e^{-sT})g_p} n_q = E_{SP} n_q \]

(4.13)

where \( E_{SP} \) is error transfer function of SPLL. Then the overall phase error \( \varphi_n \) is found as

\[ \varphi_n = \varphi_p + \varphi_q = H_{SP} n_p + E_{SP} n_q \]

(4.14)

Then the power spectrum of \( \varphi_n \) becomes

\[ S_{SP}(f_m) = |H_{SP}(s)|^2 S_p(f_m) + |E_{SP}(s)|^2 S_q(f_m) \]

(4.15)

The SSB phase noise is expressed as

\[ \mathcal{L}_{SP}(f_m) = \frac{1}{2} S_{SP}(f_m). \]

(4.16)
Simulation results of SPLL OEO using (4.16) is shown below in Figure 4.3. The black dashed curve is measured phase noise of an OEO with 100m delay in the free running case. Colored curves represent phase noise of OEO with SPLL for different delays. From the simulation results, phase noise decreases as the delay in SPLL increases.

Simulation is also performed to investigate the impact of different filter components. In the simulation, SPLL delay is fixed at 5km while different ‘Mixer Filter’ boards are used. We can see from Figure 4.4 that board #2 provides best phase noise reduction due to its large filter bandwidth. It should also be noted that when constructing SPLL, loop stability is of practical concern. A combination of large filter
bandwidth and a long fiber delay may result in an unstable loop as the long fiber delay will introduce additional poles (sidemodes) in the system.

Figure 4.4 Simulated phase noise of SPLL with 5km delay for different ‘Mixer+LPFA’ boards using RoF Link 2.Kd=0.01V/\text{rad} and Kp=2\pi \times 200\text{kHz/V}

4.1.3 Analysis of Dual Loop Self-Phase Locked Loop (DSPLL)

Control theory representation is shown in Figure 4.5 for DSPLL. Phase noise expression of DSPLL can be found using loop analysis in a similar fashion to single loop SPLL, and is given as

\[ S_{DSP}(f_m) = |H_{DSP}(s)|^2 S_p(f_m) + |E_{DSP}(s)|^2 S_q(f_m) \]  \hspace{1cm} (4.17)

where

\[ H_{DSP}(s) = \frac{G_p}{1 + G_p(1 - e^{-st}) + G_p(1 - e^{-st})} \]

is the transfer function of the DSPLL system, and

\[ E_{DSP}(s) = \frac{1}{1 + G_p(1 - e^{-st}) + G_p(1 - e^{-st})} \]
is the error transfer function of the DSPLL system. The final SSB phase noise expression is

\[ L_{DSP}(f_m) = \frac{1}{2} S_{DSP}(f_m) \]  \hspace{1cm} (4.18)

Figure 4.5  Control theory representation of DSPLL

Phase noise simulation of DSPLL using (4.18) is shown in Figure 4.6. From the simulation results, DSPLL provides similar phase noise reduction compared to SPLL, but the side-mode level is greatly reduced in DSPLL due to additional loop.

Figure 4.6 Simulated phase noise of DSPLL with Circuit 1 (medium loop BW) for different combinations of delays using RoF Link 2.
Impact of different filter components on phase noise is simulated for a length combination of 3km and 5km to identify the optimum filter parameters, simulation results are shown in Figure 4.7. Once again, we can see that the best phase noise is provided by board #3. However, in terms of physical implementation, loop stability has to be considered in order for the PLL to function properly.

![Figure 4.7 Simulated phase noise of SPLL with delay combination of 3km and 5km for different circuits using RoF Link 2.](image)

### 4.2 Experimental Results of SPLL VCO

#### 4.2.1 VCO Realization

The VCO circuit is shown in Figure 4.8. It consists of a tunable filter and an amplifier (Avantek AMT 9634) with small signal gain of 28dB and 1dB compression of 12dBm referred to the output. The tunable filter is constructed using an open circuit microstrip line terminated with two varactor diodes (Aeroflex MGV125-08). The tunability of the filter is achieved by changing the reverse bias voltage of the diodes. The VCO free running frequency is at 8.5GHz with an output power of 0dBm, and the tuning sensitivity is about 200kHz/V at 5V bias voltage. The measured phase noise is shown in the black dashed curve of Figure 4.10. The phase noise slope is about 30dB/decade up to 10MHz offset which indicates a flicker
noise dominated system. This high flicker noise is usually associated with an FET based amplifier, and flicker noise could be reduced if HBT based amplifier is used.

![Image of the VCO](image.png)

**Figure 4.8** Picture of the VCO (a) Top view (b) Bottom View

### 4.2.2 Phase Noise of SPLL VCO

Experiment setup of SPLL is depicted in Figure 4.9. Output of the VCO is amplified and drives the MZM. The modulated optical signal passes through a long fiber delay of 3km and is converted to electrical signal by a photodetector. The delayed signal is sent to the RF port of the ‘Mixer+Filter’ board #1 for comparison with the non-delayed signal coupled directly from the VCO output. Measured phase noise of SPLL VCO is shown in blue curve of Figure 4.10. The phase noise of the SPLL VCO is reduced from -26dB of free running case to -71dB at 1kHz offset corresponding to an improvement of 45dB; at 10kHz offset, phase noise is improved by 20dB reaching -78dBc/Hz. Simulation result using (4.19) is also depicted in red curve of Figure 4.10. We can see the excellent agreement between analytical and experimental results. SPLL with 5km delay is attempted to achieve further phase noise reduction, but the
long delay introduces strong side-modes that are spaced every 40kHz away from the carrier causing the loop to be unstable, and the PLL failed to acquire a locked state. In the next section, DSPLL has been demonstrated to be effective for side-mode suppression leading to a stable loop operation and eventually a better phase noise reduction.

Figure 4.9 Experiment Setup of SPLL VCO

Figure 4.10 Phase noise of SPLL with 3km delay and Circuit 1 using RoF Link2. Blue curve is measured phase noise for $K_d=0.01V/\text{rad}$, $K_o=2\pi \times 1\text{MHz}/V$; Red curve is simulated phase noise.
4.2.3 Phase Noise of DSPLL VCO

Experiment setup of DSPLL is depicted in Figure 4.11. The circuit diagram of DSPLL is similar to that of SPLL. The difference is output of the MZM is split into two: one passes through a 5km delay while the other passes through a 3km delay. The two delayed signals are received by two diodes independently, and are combined in a 3dB coupler. The combined signal is amplified and sent to the ‘Mixer+Filter’ board to be compared with a non-delayed signal from the VCO output. Experiment result is shown in blue curve of Figure 4.12. The phase noise of the DSPLL VCO is reduced from -26dB of free running case to -91dB at 1kHz offset corresponding to an improvement of 65dB; at 10kHz offset, phase noise is improved by 42dB reaching -100dBc/Hz. Simulation result of DSPLL using (4.20) is also provided in red curve of Figure 4.12, and it matches up well with the measurement result. Phase noise comparison between SPLL with 3km delay and DSPLL with 3km and 5km delay is shown in Figure 4.13. We can see that DSPLL provides 20dB more reduction than SPLL in offset frequencies from 1kHz to 50kHz.

Figure 4.11  Experiment Setup of DSPLL for VCO
Figure 4.12 Phase noise of DSPLL with delay combination of 3km+5km and Circuit 1 using RoF Link2. Blue curve is measured phase noise for $K_d=0.01V/rad$, $K_o=2\pi \times 1MHz/V$; Red curve is simulated phase noise.

Figure 4.13 Phase noise comparison of SPLL and DSPLL.
4.3 Experimental Results of SPLL OEO based on Tunable BPF

4.3.1 OEO realization

The block diagram for standard OEO configuration is depicted in the dashed black box of Figure 4.14, where the 5 port BPF is employed as a tunable band-pass filter that determines the OEO oscillation frequency. One low noise amplifiers (Kuhne Electronic 101A) and a power amplifiers (B&Z BZ3-09801050) shown as ‘LNA1’ and ‘PowerAmp1’ in Figure 4.14 are used to compensate for the RF signal loss in the MZM link. The fiber optic delay line of 100m long is selected for the OEO. The measured oscillation is at 8.5GHz and output power of this standard OEO is 6dBm. Phase noise performance for this standard OEO is shown in the black curve in Figure 4.15. The measured phase noise is -53dBc/Hz and -81dBc/Hz at 1kHz and 10kHz offset, respectively.

4.3.2 Measured Phase Noise of SPLL OEO using BPF Control

Block diagram of SPLL OEO is shown in Figure 4.14. In addition to a standard OEO, a portion of the optical power is coupled out and passes through a longer fiber delay of 3km. The delayed signal is amplified by a low noise amplifier (Kuhne Electronic 101A) followed by a power amplifier (Miteq AMF-3D) shown as ‘LNA2’ and ‘PowerAmp2’ in Figure 4.14 respectively; the delayed signal is then compared with the non-delayed OEO output at the ‘Mixer+LPFA’ board to generate an error signal to control the bias voltage of the varactor diode for frequency adjustment of the OEO. The measured phase noise of SPLL OEO is provided in Figure 4.15. The phase noise of SPLL OEO at 1kHz is -91dBc/Hz, which is 38dB lower than free running OEO; and at 10kHz, -108dBc/Hz is achieved corresponding to a 27dB improvement.
Figure 4.14 Experiment Setup of SPLL OEO based on BPF

Figure 4.15 Comparison of SPLL phase noise between measured and simulated. RoF Link 2 is used and the loop parameters are $K_d=0.01\text{V/rad}$ and $K_o=2\pi\times200\text{kHz/V}$. 
4.3.3 Impact of Different ‘Mixer+LPFA’ Boards on phase noise for a fixed delay.

Different ‘Mixer+LPFA’ boards are used to investigate the impact of PLL loop bandwidth on phase noise for an optical fiber delay of 3km. Experimental setup is the same as depicted in Figure 4.14. From the measurement results, the best phase noise performance is provided by Board #1 with a loop bandwidth of about 100kHz rather than Board #2 with a loop bandwidth of about 1MHz, which contradicts the analytical prediction. Possible reason for the discrepancy is that the 5km delay introduces too many sidemodes within the PLL loop bandwidth, and the interaction between the sidemodes degrades the performance of SPLL. Measured phase noise performance is tabulated in Table 4.1 for comparison.

| Table 4.1 Comparison of Different Boards for a fixed delay |
|-----------------|-----------------|-----------------|
|                 | 1kHz            | 10kHz           |
| Board 1         | -89.2           | -105.5          |
| Board 2         | -78.9           | -95.7           |
| Board 3         | -68.1           | -96.9           |

Figure 4.16 Measured phase noise of SPLL OEO for 3km delay with different ‘Mixer+LPFA’ Boards. RoF Link 2 is used and the loop parameters are $K_d=0.01\text{V/\text{rad}}$ and $K_o=2\pi \times 200\text{kHz/V}$. 
4.4 Experimental Results of SPLL OEO based on MZM

4.4.1 OEO Realization

The block diagram for standard OEO configuration is depicted in the dashed black box of Figure 4.17, it is similar to the one depicted in Figure 4.17 of section 4.3. The differences are: 1) power amplifiers with 30dB gain between 8.5 – 9.6GHz from Avantek (AMT 9634) are used in blocks of ‘Amp1’, ‘Amp2’, and ‘Amp3’. 2) The error signal is sent to the MZM bias port rather than 5 port BPF bias port to achieve the frequency control of the OEO. The MZM control is advantageous over BPF control since the optical phase deviation introduced by MZM will be increased as the light propagates along the fiber, hence more tuning sensitivity. On the other hand, MZM has much wider bandwidth as opposed to electrical BPF. The measured oscillation is at 9.6GHz with output power of 6dBm. Phase noise performance for this standard OEO is shown in the black curve in Figure 4.18 The measured phase noise is -69dBc/Hz and -96dBc/Hz at 1kHz and 10kHz offset, respectively.

![Standard OEO Block Diagram](image.png)

Figure 4.17 Experiment setup of SPLL OEO based on MZM
4.4.2 Measured Phase Noise of SPLL OEO using MZM Control

Experimental phase noise of SPLL OEO with MZM control is shown in Figure 4.18. The phase noise reduces as the delay in SPLL increases as expected from the analytical modeling. Best result is obtained using a 5km delay in SPLL, and the noise level is reduced by 20dB reaching -89dBc/Hz at 1kHz offset and by 23dB reaching -119dBc/Hz at 10kHz offset. Note that in section 4.3 where a BPF control is used for SPLL operation, 5km delay results in an unstable loop operation. In contrast, SPLL is functioning for MZM control with 5km delay and provides better phase noise reduction. Figure 4.19 shows the excellent agreement between simulation (red curve) and measured (blue curve) results of SPLL OEO using MZM control for 5km delay.

![Figure 4.18 Measured phase noise of SPLL with Circuit 1 (medium loop BW) for different delays using RoF Link 2. The loop parameters are $K_d=0.01\text{V/rad}$ and $K_o=2\pi\times200\text{kHz/V}$. Black curve: phase noise of free running 100m OEO; Red curve: phase noise of SPLL with 1000m delay; Blue curve: phase noise of SPLL with 3000m delay; Green curve: phase noise of SPLL with 5000m delay.](image)
Figure 4.19 Comparison of SPLL phase noise between measured and simulated. RoF Link 2 is used and the loop parameters are $K _ { d } = 0.01 \text{V/rad}$ and $K _ { o } = 2 \pi \times 200 \text{kHz/V}$.

4.5 Summary

This chapter dedicated to analysis, design, and experimental performance evaluation of self-phase locked loop oscillators using long optical delay lines. Tables 4.1, 4.2, and 4.3 summarizes the experimentally achieved phase noise improvement results at 1kHz and 10kHz offset frequencies using MZM link in comparison to analytically predicted performance.

Table 4.1 Comparison of SSB phase noise for VCO with different circuit configurations

<table>
<thead>
<tr>
<th>Phase Noise Comparison</th>
<th>Measured (dBc/Hz)</th>
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<tbody>
<tr>
<td></td>
<td>1kHz</td>
</tr>
<tr>
<td>VCO w/ 5port VCO</td>
<td>-26</td>
</tr>
<tr>
<td>SPLL 3km</td>
<td>-71</td>
</tr>
<tr>
<td>DSPLL 3km+5km</td>
<td>-91</td>
</tr>
</tbody>
</table>

Table 4.2 Comparison of SSB phase noise for free running OEO and SPLL OEO with BPF Control

<table>
<thead>
<tr>
<th>Phase Noise Comparison</th>
<th>Measured (dBc/Hz)</th>
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<tr>
<td></td>
<td>1kHz</td>
</tr>
<tr>
<td>OEO 100m</td>
<td>-53</td>
</tr>
<tr>
<td>SPLL 3km</td>
<td>-91</td>
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</tbody>
</table>
Table 4.3  Comparison of SSB phase noise for free running OEO using Avantek Amplifiers and SPLL OEO with MZM Control

<table>
<thead>
<tr>
<th>Phase Noise Comparison</th>
<th>Measured (dBc/Hz)</th>
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<tbody>
<tr>
<td></td>
<td>1kHz</td>
</tr>
<tr>
<td>OEO 100m</td>
<td>-69</td>
</tr>
<tr>
<td>SPLL 5km</td>
<td>-89</td>
</tr>
</tbody>
</table>

As demonstrated for ILPLL oscillators both close-in and far away from carrier phase noise could be reduced by combing IL and PLL function; therefore, one could employ forced self-injection and self-phase locking to build an SILPLL oscillator for improved phase noise performance [76]-[80]. The merits of the SILPLL techniques are to be investigated in chapter 5.
Chapter 5 Forced Oscillations Using Self-Injection Locked and Phase Locked Loop

In previous chapters, it has been demonstrated that SIL and SPLLL provide excellent phase noise reduction in far-out and close-in offset respectively. Combining SIL and SPLLL simultaneously, low phase noise level in wider offset range is expected. A control theory based modeling of SILPLL is developed in section 5.1 to model the phase noise of SILPLL in locked state for the first time, and this linear model is later shown to be accurate as analytical prediction corroborates with measurement results. Experimental result that demonstrates the benefit of SILPLL over SIL in close-in to carrier offset frequencies for a noisy electrical VCO is reported in section 5.2 for the first time. Performance of dual loop SILPLL with various delay combinations are also reported in section 5.2 for the first time. In section 5.3, a stable DRO is used to replace the noisy VCO in order to achieve lower absolute phase noise. Analysis of phase noise performance for SILPLL technique is also provided to identify the limiting factor in current SILPLL system. Then the SILPLL technique is extended to an OEO in section 5.4 to show that this technique is applicable for various oscillators for phase noise reduction. Section 5.5 summarizes the experimental results of different circuit topologies and concludes the chapter.

5.1 Modeling of SILPLL

Recall from sections 3.1 and 4.1 that linearized IL and PLL phase dynamics are represented as:

\[
\frac{d\varphi_o(t)}{dt} = K_I[\varphi_i(t) - \varphi_o(t)]
\]

\[
\frac{d\varphi_o(t)}{dt} = K_p(t)[\varphi_i(t) - \varphi_o(t)]
\]

where \(K_I = \rho \omega_{3db}\) and \(K_p(t) = K_d K_o K_f(t)\). \(\rho\) is the injection strength, \(\omega_{3db}\) is half the 3 dB bandwidth of the oscillator, \(K_d\) is the phase detector sensitivity in V/rad, \(K_o\) is the VCO tuning sensitivity in rad/V, and \(K_f(t)\) is the impulse response of the loop filter. Under the condition of small phase error (i.e. \(\varphi_i(t)\) —
\( \phi_o(t) \ll 1 \), then linearity still holds when the system is in locked state, and the ILPLL phase dynamics can be obtained by adding IL and PLL phase dynamics as

\[
\frac{d\phi_o(t)}{dt} = K_i[\phi_i(t) - \phi_o(t)] + K_p(t)[\phi_i(t) - \phi_o(t)] \tag{5.1}
\]

The above time domain equation is converted into s-domain by performing Laplace transform on both sides of (5.1), resulting in

\[
s\phi_o = (K_i + K_p)[\phi_i - \phi_o] \tag{5.2}
\]

Then the transfer function of the ILPLL is found as

\[
H_{IP} = \frac{\phi_o}{\phi_i} = \frac{G_{IP}}{1+G_{IP}} \tag{5.3}
\]

where \( G_{IP} = G_i + G_p \). From the system transfer function, we can represent the ILPLL system in frequency domain in a control theory based block diagram, as shown in Figure 5.1. Note that the static phase error is assumed to be small, therefore they can be omitted from the block diagram for simplicity from now on since they do not have significant impact to the output phase error.

![Figure 5.1 Control theory representation of ILPLL system in the presence of noise.](image)
Assuming that noise is a small perturbation to the steady state solution, thus the superposition principle of linear system could be used to estimate the overall output phase error $\varphi_n$, and eventually lead to the ILPLL noise spectrum $S_{IP}(f_m)$. Let’s first find out the phase error $\varphi_p$ due to the input noise $n_p$, as depicted in Figure 5.1. Using basic loop analysis, relationship between the phase error $\varphi_p$ and input noise $n_p$ can be found as

$$\varphi_p = \frac{G_{IP}}{1 + G_{IP}} n_p = H_{IP} n_p$$  \hfill (5.4)

Similarly, we can find the output phase error $\varphi_q$ due to oscillator phase noise $n_q$ as

$$\varphi_q = \frac{1}{1 + G_{IP}} n_q = E_{IP} n_q$$  \hfill (5.5)

where $E_{IP}$ is the error transfer function of ILPLL system. Then the total phase error $\varphi_n$ is found as,

$$\varphi_n = H_{IP} n_p + E_{IP} n_q$$  \hfill (5.6)

The noise spectrum of $\varphi_n$ becomes

$$S_{IP}(f_m) = |H_{IP}|^2 S_p(f_m) + |E_{IP}|^2 S_q(f_m)$$  \hfill (5.7)

Finally, the SSB phase noise is given as

$$L_{IP}(f_m) = \frac{1}{2} S_{IP}(f_m)$$  \hfill (5.8)
In (5.7), $S_p(f_m)$ represents the oscillator phase noise; $S_p(f_m)$ represents the system residual noise. In our experiments, the system residual noise $S_p(f_m)$ is dominated by laser relative intensity noise (RIN) at far away offsets and system flicker noise at close in offsets. Expression of $S_p(f_m)$ at frequency offset of $f_m$ is given in [81] as

$$S_p(f_m) = \frac{N_{RIN} I_{ph} R}{\epsilon_s} (1 + \frac{f_c}{f_m}) \quad (5.9)$$

where $N_{RIN}$ is the laser RIN, $I_{ph}$ is the photo current at the photodetector, $R$ is the load resistance of the photodetector, and $f_c = 1$ MHz is the flicker corner frequency associated with the RF power amplifiers (c.f. section 3.1).

We can easily extend the ILPLL analysis to the case of self-ILPLL whose control theory representation is shown below in Figure 5.2. A portion of the oscillator output is being delayed and the phase of the delayed signal is compared against the phase of the current signal. As we can see, a time delay $\tau$ is expressed as $e^{- \tau}$ in s-domain. Note that the delay is shared by both the SIL function and SPLL function. Again the superposition principle of linear systems is applied to find out the overall phase error of the SILPLL system. Using standard loop analysis, the transfer function of output phase error $\phi_p$ due to input noise $n_p$ is given as

$$\phi_p = \frac{G_{IP}}{1+(1-e^{-\tau})G_{IP}} n_p = H_{SIL} n_p, \quad (5.10)$$

and $H_{SIL}$ is the transfer function of SILPLL. Similarly, phase error $\phi_q$ due to $n_q$ can be found as

$$\phi_q = \frac{1}{1+(1-e^{-\tau})G_{IP}} n_q = E_{SIL} n_q, \quad (5.11)$$
and $E_{SIP}$ is the transfer function of SILPLL. Total phase error $\varphi_n$ can be found as

$$\varphi_n = H_{SIP}n_p + E_{SIP}n_q$$  \hspace{1cm} (5.12)

The overall output noise spectrum of $\varphi_n$ becomes

$$S_{SIP}(f_m) = |H_{SIP}|^2S_p(f_m) + |E_{SIP}|^2S_q(f_m)$$  \hspace{1cm} (5.13)

Finally, the SSB phase noise of SILPLL is given as

$$\mathcal{L}_{SIP}(f_m) = \frac{1}{2}S_{SIP}(f_m)$$  \hspace{1cm} (5.14)

Figure 5.2 Control theory representation of SILPLL.

The control theory representation of single loop SILPLL can be readily extend to N-loop SILPLL, as shown in Figure 5.3. Phase noise of N-loop SILPLL can be found using loop analysis in a similar fashion to single loop SILPLL, and is given as

$$\mathcal{L}_{NSIP}(f_m) = \frac{1}{2}S_{NSIP}(f_m) = \frac{1}{2}[|H_{NSIP}(s)|^2S_p(f_m) + |E_{NSIP}(s)|^2S_q(f_m)]$$  \hspace{1cm} (5.15)
where

\[ H_{NSLP}(s) = \frac{G}{s + G \sum_{i=1}^{N}(N - e^{-s\tau_i})} \]

is the transfer function of N-loop SILPLL system, and

\[ E_{NSLP}(s) = \frac{s}{s + G \sum_{i=1}^{N}(N - e^{-s\tau_i})} \]

is the error transfer function of N-loop SILPLL system. Note that in Figure 5.3, \( \tau_i \) is the delay time in the \( i^{th} \) path, and \( n_p \) is the sum of the input noise in each loop. Again, the delay paths are shared by SIL and SPLL.

![Figure 5.3 Control theory representation of N-loop SILPLL.](image)

**5.2 SILPLL VCO with EAM Link**

In this section, a VCO with high tuning sensitivity and an EAM link with less loss are used to construct the SILPLL function for achieving higher PLL BW and eventually improving the SILPLL performance.
5.2.1 Single Loop SILPLL VCO

Block diagram of experimental setup of SILPLL VCO using EAM link is depicted below in Figure 5.4. The EAM link is shown in Figure 5.4 with blue blocks. A laser diode is integrated with an EAM in a single package (Fujitsu FLD5F10NP), denoted as ‘EAM-LD’ in the block diagram. Output of the ‘EAM-LD’ is amplified by an EDFA (NuPhoton NP2000) whose output passes through a long fiber delay. The delayed optical signal is converted to electrical signal by a photodetector (PD). The EAM link has a RF loss of 12 dB and a NF of 40 dB. Other parameters associated with the EAM link are: $N_{RIN} = -135$ dB/Hz, $I_{ph} = 2$ mA and $R_{PD} = 50$ Ω. These parameters are used throughout all experiments and simulations when EAM link is used.

To construct the SILPLL system, the VCO (c.f. section 4.5) drives the EAM-LD, and the delayed signal is sent to ‘Mixer+LPFA’ board #3 with power level of 13dBm after amplification, resulting in an improved phase detector sensitivity of about 0.1V/rad as opposed to 0.01V/rad in MZM link. The delayed signal is compared with non-delayed signal for phase comparison, and the generated error signal will control the reverse bias voltage of the varactor diodes on the BPF to adjust the VCO frequency. The tuning sensitivity is increased to 1MHz/V at 5V compared to 200kHz/V in an OEO. Note that there isn’t a dedicated path for SIL, but the power at Mixer RF port will leak into the VCO due to finite isolation between RF and LO ports of the mixer. The VCO is very prone to injection locking due to the very low Q BPF, and this leakage power of about -25dBm is strong enough to practically form an SIL path.
Measured phase noise of SILPLL with 1km delay is plotted as blue curve in Figure 5.5. The achieved phase noise is -69dBc/Hz at 300Hz corresponding to a reduction of 58dB and is -87dBc/Hz at 10kHz corresponding to a reduction of 29dB. Note that when the SPLL is functioning, the SIL due to power leakage from mixer RF port is also present, the overall phase noise is really due to the combination of SIL and SPLL. For comparison, phase noise of SIL alone with 1km delay is also provided in green curve of Figure 5.5. We can see that SILPLL phase noise is superior to SIL phase noise up to 3kHz offset, beyond 3kHz SILPLL phase noise follows SIL phase noise which is expected from the analytical modeling. Simulated phase noise using (5.14) for VCO employing SILPLL is also shown in Figure 5.5 as red curve. We can see the simulation result agrees well with the measured result in a wide offset range except for offset frequencies under 700 Hz. Possible reason for the discrepancy is that the op-amp gain increases much faster than that predicted by the linear model, providing additional phase noise reduction. As a result, the phase noise slope below 1 kHz offset becomes flat. Phase noise of different frequency stabilization techniques for VCO using EAM link are tabulated in Table 5.1 to quantitatively show the benefit of SILPLL over SIL.
Figure 5.5  Experiment Results of VCO with SILPLL. Mixer Board #3 is used, 1km delay, Black: VCO free run; Red: SIL 1km; Blue: SILPLL 1km; $K_d=0.1\text{V/rad}$, $K_w=2\pi\times5\text{MHz/V}$.

### Table 5.1 Comparison of different circuit topologies utilizing single fiber delay for VCO with FR4 BP

<table>
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<tr>
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<th>300Hz</th>
<th>1kHz</th>
<th>10kHz</th>
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<tbody>
<tr>
<td>Measured VCO free run</td>
<td>-11</td>
<td>-29</td>
<td>-58</td>
</tr>
<tr>
<td>Measured SIL 1 km</td>
<td>-42</td>
<td>-58</td>
<td>-86</td>
</tr>
<tr>
<td>Measured SILPLL 1 km</td>
<td>-69</td>
<td>-68</td>
<td>-87</td>
</tr>
<tr>
<td>Simulated SILPLL 1 km</td>
<td>-61</td>
<td>-70</td>
<td>-90</td>
</tr>
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</table>

Fiber delays longer than 1km are also attempted to provide more phase noise reduction. However, the sidemodes of the long delay makes the PLL loop to be unstable. Dual loop SILPLL is proposed to suppress the sidemode level of long delay for improved loop stability and to eventually further reduce phase noise.
5.2.2 Dual Loop SILPLL VCO

The block diagram of dual loop SILPLL is similar to that of single loop SILPLL (cf. subsection 5.2.1), and is shown in Figure 5.6. The difference is the output of the EDFA is split into two paths: one with a 5km delay and another with a 3km delay. Signals from the two delayed path are picked up by two photodetectors independently, and the converted electrical signals are combined in a 90° hybrid. The combined signal is amplified and then is sent to the ‘Mixer+LPFA’ to complete the SPLL function. Once again, the leakage from mixer RF port will induce SIL function onto the VCO, resulting in a combined operation of SIL and SPLL.

Figure 5.6 Block diagram of dual loop SILPLL with VCO.

Measured phase noise of DSILPLL with various delay combinations are shown in Figure 5.7. From the measured results, combination of 3 km and 5 km delays yields the best phase noise of -82 dBC/Hz at 300 Hz offset resulting in a 71 dB improvement and -98 dBC/Hz at 10 kHz offset resulting in a 40 dB improvement. Phase noise performance in the case of 1 km and 5 km delay is inferior to 3 km and 5 km delay even though the length of the longer delay is the same. This could be due to the sidemodes of 1 km (every 200 kHz) and 5 km (every 40 kHz) being harmonically related therefore the sidemode suppression is not so effective as in the case of 3 km and 5 km delays, where the sidemodes are non-harmonically
related. In particular, 66.7 kHz for 3 km delay and 40 kHz for 5 km delay. Measured phase noise of DSILPLL and DSIL are shown in Figure 5.8 for comparison. Same delay combination of 3km and 5km is used for both stabilization techniques. We can see that phase noise of DSILPLL (blue curve) is 29dB lower than that of DSIL (green curve), which demonstrates the advantage of DSILPLL over DSIL alone. Simulated phase noise using (5.15) for VCO employing DSILPLL with 3 km and 5 km delays is shown as red curve in Figure 5.8. Again, the simulation results corroborates with the measured results. Spot noise with different circuit topologies are also tabulated in Table 5.2 for comparison of different techniques.

Figure 5.7 Phase Noise of DSILPLL VCO. Mixer Board #3 is used, Black: VCO free run; Red: DSILPLL 1km+3km; Blue: DSILPLL 1km+5km; Green: DSILPLL 3km+5km. $K_d=0.1V/rad$, $K_o=2\pi \times 1$ MHz/V.
Figure 5.8 Phase Noise of DSILPLL VCO. Mixer Board #3 is used, 1km delay, Black: VCO free run; Red: SIL 1km; Blue: SILPLL 1km; \( K_d = 0.1 \text{V/rad}, K_o = 2\pi \times 1 \text{MHz/V} \).

Table 5.2 Comparison of different circuit topologies utilizing single fiber delay for VCO with FR4 BPF

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<th>300Hz</th>
<th>1kHz</th>
<th>10kHz</th>
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<tbody>
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<td>-58</td>
</tr>
<tr>
<td>Measured DSIL 3km+5km</td>
<td>-53</td>
<td>-69</td>
<td>-97</td>
</tr>
<tr>
<td>Measured DSILPLL 3km+5km</td>
<td>-82</td>
<td>-80</td>
<td>-98</td>
</tr>
<tr>
<td>Simulated DSILPLL 3km+5km</td>
<td>-80</td>
<td>-88</td>
<td>-107</td>
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</table>

Phase noise performances of single loop and dual loop SILPLL are also compared in Figure 5.9, the blue and green rectangles represent the actual level of the first spurious signals of SILPLL and DSILPLL, respectively. By comparison of the side-mode levels, DSILPLL demonstrated better side-mode suppression as opposed to single loop SILPLL. For SILPLL, the first spurious signal appears at offset frequency of 186 kHz with a level of -40 dBc (blue rectangle), which is relatively high; but for DSILPLL, the actual first spurious signal appears at 75 kHz offset with a level of -62 dBc (green rectangle), corresponding to a suppression of 22 dB compared to that in SILPLL configuration.
Figure 5.9 Comparison of VCO phase noise employing different stabilization techniques. Black dashed: VCO free run; Red: SIL 1km; Blue: SILPLL 1km; Magenta: DSIL 3km+5km; Green: DSILPLL 3km+5km. SIL parameters: $p = 0.003$, $\omega_{3dB} = 60$ MHz. SPLL parameters: $K_d = 0.1V/\text{rad}$, $K_o = 2\pi \times 1\text{MHz}/V$.

5.3 SILPLL DRO with EAM Link

In the previous section, the lowest achievable phase noise is limited by the noisy electrical VCO. In this section, a very clean dielectric resonator oscillator (DRO) from Synergy Microwave Corp. (DRO 100) is used to replace the VCO, resulting in a lower phase noise.

5.3.1 System overview of DRO with SILPLL

The optical delay in SILPLL system is provided by the same EAM link used in the previous section, also shown in Figure 5.10 with blue blocks. For the SILPLL operation, the EAM-LD is driven by the DRO, and the modulated light passes through two different optical delay lines, termed as dual loop SILPLL (DSILPLL). The delay configuration shown in Figure 5.10 represents a total delay length of $L_A + L_B$ in path 1 while in path 2 the delay length is $L_B$. The delayed signal is first amplified by two amplifiers (Avantek AMT 9634), and a small portion of the amplified signal is coupled out and directly fed back to the DRO through a circulator, forming the SIL function (green path in Figure 5.10); another portion is compared against the non-delayed DRO output to generate an error signal for frequency adjustment, completing the SPLL function (purple path in Figure 5.10). The error signal is generated by the ‘Mixer +
LPFA’ board #1 with a low pass cutoff frequency of 100 kHz, and it outputs a DC voltage of 4.6 V when the PLL is in locked state. The DRO (DRO100) used in the experiments outputs a 10GHz signal with 10dBm power. The phase noise of this DRO is shown as the black dashed curve in Figure 5.11 with -84 dBc/Hz at 1 kHz offset and -111 dBc/Hz at 10 kHz offset. The loaded Q of the resonator is found to be about 2000 by curve fitting the measured phase noise using Leeson’s equation [2]. The tuning sensitivity is 280 kHz/V measured at a tuning voltage of 4.6 V that matches the ‘Mixer+LPFA’ board DC output.

![Figure 5.10 Block diagram of SILPLL DRO system.](image)

### 5.3.2 Experiment results of DRO with SILPLL

In the DSILPLL experiment, the injection strength is kept at $\rho=0.3$. For the SPLL portion, a power level of 13 dBm results in a phase detector sensitivity $K_d$ of 0.1 V/ rad, and the DRO tuning sensitivity $K_o$ is $2\pi \times 280$ kHz/V. Phase noise performance of DSILPLL with different delay combinations is shown in Figure 5.11. The blue curve represents the case for 3 km and 5 km delays, and the red curve for 5 km and 8 km delays. Again, we can see from Figure 5.11 that longer delay provides higher phase noise reduction, as expected. In the case of 5 km and 8 km delays, phase noise of -100 dBc/Hz at 1 kHz offset is achieved,
resulting in a 16 dB improvement; at 10 kHz offset, the phase noise is -125 dBc/Hz, resulting in a 14 dB improvement. Phase noise comparison between DRO employing SIL and DSILPLL is also depicted in Figure 5.12. From the measured results, DSILPLL provides only a slight reduction over SIL. We will show through analytical modeling in subsection 5.3.3 that the performances of SIL and DSILPLL in our experiments are limited by the system residual noise.

![Figure 5.11](image1.png)

**Figure 5.11** Phase noise of DRO employing DSILPLL with different delay combinations. Black: VCO free run; Blue: DSILPLL 3 km and 8 km; Red: DSILSPLL 5 km and 8 km. SIL parameters: SIL for $\rho = 0.3$ and $\omega_{3dB} = \pi \times 5$ MHz. SPLL parameters: $K_d = 0.1$ V/rad, $K_o = 2\pi \times 1$ MHz/V.

![Figure 5.12](image2.png)

**Figure 5.12** Phase noise comparison of DRO employing SIL and DSILPLL. Black: DRO free run; Blue: SIL 8km; Red: DSILSPLL 5 km and 8 km. SIL parameters: SIL for $\rho = 0.3$ and $\omega_{3dB} = \pi \times 5$ MHz. SPLL parameters: $K_d = 0.1$ V/rad, $K_o = 2\pi \times 1$ MHz/V.
5.3.3 Analysis of DRO with SILPLL

Analysis of performance limitation in our experiments is provided, followed by a prediction of potential improvement by reducing system residual noise. By observing phase noise expression of DSILPLL given in (5.15), the overall phase noise consists of two terms: the first term $T_a = |H_{NSIP}|^2 S_{n_p}(f_m)$ represents noise contribution from the system residual noise $S_{n_p}(f_m)$; the second term $T_b = |E_{NSIP}|^2 S_{n_q}(f_m)$ represents contribution from the DRO phase noise $S_{n_q}(f_m)$. Phase noise simulation using (5.15) with experimental DSILPLL parameters are performed for DRO employing DSILPLL with 5km and 8km delays. Simulation results are shown in Figure 5.13: the red curve shows the overall noise of DSILPLL DRO; contributions from system residual noise and DRO phase noise are shown as blue and green dashed curves, respectively. We can see the contribution of system residual noise dominates in close in offsets until 3 MHz; beyond 3 MHz, DRO phase noise starts to take over. From Figure 5.13, the DRO phase noise (green dashed curve) is reduced to -166 dBc/Hz at 10 kHz offsets due to the combined operation of SIL and SPLL but it is buried under the system residual noise. Measured phase noise of DSILPLL DRO is shown as magenta curve in Figure 5.13, and it corroborates with the simulation results.

Figure 5.13  Phase noise simulation of DSILPLL DRO using (5.15). $N_{RIN} = -135$ dB/Hz, $I_{ph} = 2$ mA, $R = 50$ Ω, and $f_c = 1$MHz.
The simulation has shown that the current limitation of our DSILPLL system is due to a high RIN level of -135 dB/Hz. If we can reduce the RIN level to -170 dB/Hz by using a better laser source \[82\], phase noise of -155 dBc/Hz is predicted for DRO employing DSILPLL with 5 km and 8 km delays, as shown in Figure 5.14. Table 5.3 summarizes measured results for SIL and DSILPLL, and compares simulated and measured results for DSILPLL.

Figure 5.14 Predicted phase noise of DRO employing DSILPLL with 5 km and 8 km delays when RIN is reduced to -170 dB/Hz.

Table 5.3 Comparison of phase noise for DRO employing SIL and DSILPLL

<table>
<thead>
<tr>
<th></th>
<th>Phase Noise at 1kHz (dBc/Hz)</th>
<th>Phase Noise at 10kHz (dBc/Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Measured DRO Free-Run</td>
<td>-69</td>
<td>-112</td>
</tr>
<tr>
<td>Measured SIL 8km</td>
<td>-99</td>
<td>-123</td>
</tr>
<tr>
<td>Measured DSILPLL 5km+8km</td>
<td>-100</td>
<td>-125</td>
</tr>
<tr>
<td>Simulated DSILPLL 5km+8km</td>
<td>-103</td>
<td>-133</td>
</tr>
</tbody>
</table>
5.4 SILPLL OEO using MZM as SPLL control

The technique of SILPLL is applied to a standard OEO for phase noise reduction, and the experimental results has been reported for the first time. The block diagram of SILPLL OEO is shown in Figure 5.15. A standard OEO with 100m fiber delay is shown within the black dashed box. A portion of the optical signal of the OEO is coupled out and is being delayed by a longer fiber of 5km. The delayed optical signal is converted to electrical signal by a photodetector, half of the photodetector output is sent back to the standard OEO directly to form an SIL path (shown in a green curve); another half is amplified and is sent to the ‘Mixer+LPFA’ board #1 for comparison against the non-delayed signal to generate an error signal for frequency adjustment of the OEO by changing the MZM bias voltage. The SPLL portion is shown in purple in Figure 5.15.

Figure 5.15 Block diagram for SILPLL OEO.

The measured SILPLL phase noise with 5 km delay is shown in blue curve of Figure 5.16. The achieved phase noise is -96 dBc/Hz at 1 kHz offset and is -120 dBc/Hz at 10 kHz offset, which demonstrates a reduction of 27 dB and 24 dB at 1 kHz and 10 kHz offset, respectively. Simulated phase noise of SILPLL using (5.13) is also provided as red curve in Figure 5.16, which agrees well with the measurement.
Phase noise of OEO with different frequency stabilization techniques are plotted in Figure 5.17, the spot noise at 1kHz and 10kHz are also tabulated in Table 5.4. From the measured results, the distinction between different technologies are insignificant even though SILPLL is expected to achieve lower phase noise than SIL in the close-in to carrier offset region while maintaining same noise level of SIL in the far-out offsets. The reason for the limitation is similar to the case of DRO employing SILPLL: noise contribution from system residual noise dominates over the oscillator phase noise in the close-in to carrier offset frequencies.
Table 5.4 Comparison of SSB phase noise for free running OEO and SPLL OEO with MZM Control

<table>
<thead>
<tr>
<th></th>
<th>Phase Noise at 1kHz (dBc/Hz)</th>
<th>Phase Noise at 10 kHz (dBc/Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>OEO Free-Run</td>
<td>-69</td>
<td>-96</td>
</tr>
<tr>
<td>SPLL 5 km</td>
<td>-89</td>
<td>-119</td>
</tr>
<tr>
<td>SIL 5 km</td>
<td>-91</td>
<td>-119</td>
</tr>
<tr>
<td>SILPLL 5 km</td>
<td>-96</td>
<td>-120</td>
</tr>
</tbody>
</table>

5.5 Summary

This chapter is dedicated to analysis and experimental performance evaluation of self-injection locked phase locked loop (SILPLL) oscillators using long optical delay lines. Table 5.5 summarizes the lowest achieved phase noise in various oscillator circuits employing different configurations of SILPLL technique. It can be seen from the table that low noise oscillators (DRO or OEO) and long delays are crucial to achieve low phase noise. However, the system residual noise becomes the bottle neck of the system performance. Hence it is crucial to use low noise components when building SILPLL system; on the other hand optimum delay length to achieve the lowest possible phase noise will be discussed in Chapter 6.

Table 5.5 Comparison of different circuit topologies utilizing single fiber delay for VCO with FR4 BPF

<table>
<thead>
<tr>
<th></th>
<th>1kHz</th>
<th>10kHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>VCO DSILPLL 3km+5km w/ EAM</td>
<td>-80</td>
<td>-98</td>
</tr>
<tr>
<td>DRO DSILPLL 5km+8km w/ EAM</td>
<td>-100</td>
<td>-125</td>
</tr>
<tr>
<td>OEO SILPLL 5km w/ MZM</td>
<td>-96</td>
<td>-120</td>
</tr>
</tbody>
</table>
Chapter 6 CONCLUSION AND FUTURE WORK

6.1 Summary of this Thesis

Experimental results of a dual self-injection locking (DSIL) and dual self-phase locked loop (DSPLL) employing short and long delays have been proposed for sidemode suppression, while maintaining same amount of phase noise reduction provided by the long delay. As an example of DSIL, sidemode suppression of more than 20dB for fiber delay links of 1km and 5km have been experimentally achieved compared to a single 5km long SIL with a phase noise reduction of 40dB (in reference to free running oscillator) at 10kHz offset from carrier for both standard OEO and a self-seeded structure with electrical 3 port oscillator at 10GHz; for a DSPLL fiber delay lines of 3km and 5km, a sidemode suppression of 29dB have also been experimentally achieved compared to SPLL of 5km with a phase noise reduction of 30dB (in reference to free running oscillator) at 10kHz offset from carrier. For the case of SPLL, phase locking performance of a 10GHz oscillation signal are experimentally evaluated as various methods of phase locking are compared. The phase locking methods are based on of a 5 port bandpass filter as tunable electrical phase shifter, a tunable three port electrical VCO, tunable Mach-Zehnder modulator (MZM) as optical phase shifter, and a tunable VCO using electro-absorption (EA) modulator. Experiment results that demonstrate the benefit of SILPLL incorporating dual delays have been reported for the first time corroborating analytical predictions. A dual SILPLL (DSILPLL) system with 3km and 5km fiber delay has been implemented, and measured phase noise reduction of 40dB provided by DSILPLL is the same as DSIL at 10kHz offset. However, at 1kHz offset, DSILPLL provides a phase noise reduction of 52dB which is 11dB higher than DSIL; at 300Hz offset, DSILPLL provides 70dB reduction while DSIL provides only 42dB reduction.

The best achieved phase noise is -120dBc/Hz at 10kHz offset for a 9.6GHz carrier. Current limitation of the system performance is due to the high noise level of the system (cf. Appendix B). If low flicker noise HBT based amplifier, low $V_{\pi}$ MZM and low RIN laser are to be used in the SILPLL system,
simulation result indicates that SILPLL phase noise of -155dBc/Hz at 10 kHz offset can be achieved for a 10GHz carrier.

In summary, DSILPLL is effective for side-mode suppression and phase noise reduction, where SPLL using tunable MZM with DSIL of a VCO provides the best performance improvement over other investigated topologies. Due to the advances in low noise electronics and broad bandwidth of the optical components used in the DSILPLL system, the DSILPLL technique has the potential to create highly stable RF oscillators approaching 100GHz.

6.2 Recommendations for Future Work

6.2.1 Using Optimum Delay Length and Improved Fiber Optic Link

In section 5.4, the limitations of SILPLL system are identified. In this subsection, methods to improve the overall performance are discussed and can be used as a guideline for future work.

From the previous simulation and experimental results, longer delays are crucial to achieve low phase noise. This could be explained by observing the transfer function and error transfer function of SILPLL system, shown in Figure 6.1. The blue curves represent transfer functions of SILPLL, and the red curves are error transfer functions of SILPLL. We can see that transfer function behaves like a low pass filter while error transfer function behaves like a high pass filter. For a stable oscillator with a clean spectrum, transfer function has much higher impact on phase noise in close-in offset region than error transfer function. Recall from (5.14) that the system residual noise will be multiplied by transfer function and the product becomes the dominant close-in to carrier phase noise. The system residual noise in SILPLL is essentially the noise to signal ratio at the output of the photodetector. When the delay length is not very long (e.g. below 10km) the ratio is largely dependent on the optical and RF hardware and is not a strong function of the delay length. On the other hand, the magnitude of the transfer function is inversely proportional to the delay length as can be seen from Figure 6.1. The magnitude of the transfer function for
8 km delay (blue dashed curve) is 18 dB lower than that for 1 km delay (blue solid curve) below 10 kHz offset. As a result, the contribution from system residual noise is reduced due to the long delay. We can see a similar decrease of error transfer function magnitude in Figure 6.1 when the delay length increases, which also provides reduction to the oscillator phase noise. But this reduction may not be observable as the system residual noise dominates in the close-in offset region.

![Figure 6.1 Simulated loop transfer function (blue) and error transfer function (red) for SILPLL for two delays of 1km and 8km. ‘Mixer+LPFA’ board 1 is used for simulation. Other parameters used for simulation are: $\rho = 0.3$ and $\omega_{3dB} = \pi \times 5 \text{ MHz}$ for SIL function; $K_D = 0.1 \text{ V/rad}$, $K_O = 2\pi \times 1 \text{ MHz/V}$ for SPLL function.](image)

Even though it is desirable to use longer delay to achieve better phase noise reduction, there is a maximum length beyond which the phase noise starts to degrade because the noise to signal ratio (or system residual noise) at the output of the photodetector increases due to high loss introduced by long delay. In the experiment setup, a laser with RIN level of about -135 dB/Hz is used, resulting in a RIN dominated system. For optical power ($P_{in}$) of 1 mW at input of the fiber, the noise and signal power as a function of delay length are calculated based on Eqs. (1) and (2) in [66], and the results are tabulated in Table 6.1. For length below 20 km with typical attenuation of 0.3 dB/km, the link loss is relatively low resulting in a moderate photo-current level hence the output noise at the photodetector is dominated by
RIN noise. Since both RIN noise and signal power have square dependency on photo-current, they decrease at the same rate as the delay length increases. When the delay increases to beyond 20 km, the photo-current diminishes due to higher link loss then shot noise becomes the dominant one. In this case, the signal power reduces faster than the shot noise as the length increases hence the noise to signal ratio rises up, degrading the phase noise performance. In the calculation for Table 6.1, the input noise source is assumed to be a matched load at room temperature (i.e. -174 dBm/Hz). In reality, input noise power could be higher than -174 dBm/Hz due to other noise sources. Nonetheless, this simplified calculation provides insights into the impact of delay length on system residual noise.

Table 6.1 Noise and Signal Power as a Function of Delay Length for RIN = -135 dB/Hz and $P_N = 1$ mW

<table>
<thead>
<tr>
<th>Delay Length (km)</th>
<th>$P_s$ (dBm)</th>
<th>$N_{out}$ (dBm/Hz)</th>
<th>$S_{RES}$ (dBc/Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>-33</td>
<td>-159</td>
<td>-126</td>
</tr>
<tr>
<td>10</td>
<td>-36</td>
<td>-162</td>
<td>-126</td>
</tr>
<tr>
<td>15</td>
<td>-39</td>
<td>-165</td>
<td>-126</td>
</tr>
<tr>
<td>20</td>
<td>-42</td>
<td>-167</td>
<td>-125</td>
</tr>
<tr>
<td>25</td>
<td>-45</td>
<td>-169</td>
<td>-124</td>
</tr>
<tr>
<td>30</td>
<td>-48</td>
<td>-171</td>
<td>-123</td>
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<tr>
<td>35</td>
<td>-51</td>
<td>-172</td>
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<td>40</td>
<td>-54</td>
<td>-173</td>
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</tr>
<tr>
<td>50</td>
<td>-60</td>
<td>-174</td>
<td>-114</td>
</tr>
<tr>
<td>60</td>
<td>-66</td>
<td>-174</td>
<td>-108</td>
</tr>
</tbody>
</table>

In order to achieve lowest phase noise in close-in offset region, the optimum length needs to be identified such that the product of transfer function and system residual noise would be minimum. Figure 6.2 depicts the simulated results of this product at 1 kHz offset as a function of delay length. The phase noise reduces to -122 dBc/Hz as the delay increases to 30 km. Above that, the phase noise starts to degrade due to an increased noise to signal ratio. The simulation indicates a length of 30 km will provide the best phase noise at 1 kHz, but care must be taken when such a long delay is to be used since the side-modes are only 6.7 kHz apart. Multi-loop implementation is necessary to suppress the side-modes, otherwise the system could be unstable.
Figure 6.2 Simulated phase noise at 1 kHz offset as a function of delay length for RIN = -135 dB/Hz and $P_{IN} = 1$ mW. ‘Mixer+LPFA’ board 1 is used for simulation. Other parameters used for simulation are: $\rho = 0.3$ and $\omega_{3dB} = \pi \times 5$ MHz for SIL function; $K_D = 0.1 V/rad$, $K_O = 2\pi \times 1 MHz/V$ for SPLL function.

As discussed in section 5.3, the RIN level of -135 dB/Hz is the major noise source, and it limits the lowest phase noise of the SILPLL system. If a better laser with RIN level of -170 dB/Hz is used [82], then the RIN noise reduces and shot noise becomes dominant. In a shot noise dominated system, it is desirable to increase the optical power ($P_{IN}$) to reduce the noise to signal ratio. Figure 6.3 shows the phase noise at 1 kHz offset when the RIN is reduced to -170 dB/Hz and $P_{IN} = 10$ mW. In this case, phase noise of -143 dBc/Hz is predicted when a delay of 20 km is used.

Figure 6.3 Simulated phase noise at 1 kHz offset as a function of delay length RIN = -170 dB/Hz and $P_{IN} = 10$ mW. ‘Mixer+LPFA’ board 1 is used for simulation. Other parameters used for simulation are: $\rho = 0.3$ and $\omega_{3dB} = \pi \times 5$ MHz for SIL function; $K_D = 0.1 V/rad$, $K_O = 2\pi \times 1 MHz/V$ for SPLL function.
6.2.2 Optical Filtering

Optical transversal filters, or Mach-Zehnder interferometers (MZI’s), are proposed as another alternative to further suppress side-modes present in OEO’s. MZI’s filter RF signals in the optical domain, and have been reported in [83]. The advantage of using optical transversal techniques over equivalent electrical techniques lies in their inherent narrow bandwidth, small size and low loss for very high order, and ease of tunability.

The feed-forward 1st order MZI consists of two ideal 3-dB fiber couplers having a delay arm and reference arm as shown in Figure 6.4. Light is injected into port 1 or 2 and received at port 3 or 4. The filter transfer function is given at operating RF angular frequency of \( \omega \) by

\[
H(\omega) = \frac{1 + \cos[\omega(\tau_d + \tau_D)]]}{2}
\]  

(6.1)

where \( \tau_d \) is due to the fiber delay at source wavelength of \( \lambda_o \). This delay is related to refraction index \( n(\lambda_o) \) of fiber core at \( \lambda_o \), speed of light in free space \( c \), and \( \Delta L \) the fiber length difference between reference and delayed arms as \( \tau_d = \Delta L n(\lambda_o)/c \). The term \( \tau_D \) in (6.1) is due to fiber dispersion and provides option of narrowband filter tuning by adjusting the optical source wavelength and is represented by \( \tau_D = D \Delta L (\Delta \lambda) \), where \( \Delta \lambda \) is wavelength tuning away from \( \lambda_o \) and \( D \) is the dispersion parameter in unit of ps/nm/km. For simplicity, (6.1) assumes 3-dB equal split in the couplers. It is apparent that the filter frequency response is dependent on delay length and optical wavelength tuning. The optical wavelength can be varied to determine the overall tunability of the filter and is specified in units of kHz/nm.
1st order MZI was characterized. The filter consisted of Corning SMF-28E fiber and Ascentta 2x2 3 dB couplers for C-band. A Eudyna FLD5F10NP DFB laser is biased by a Lightwave LDC3900 controller and externally modulated by a Gigatronics GT9000 Microwave Synthesizer. The operating wavelength is 1554 nm as measured with an Anritsu MS9710 optical spectrum analyzer. The MZI output is fed to a Discovery DSC50S PIN photodiode, which produces an RF signal to be amplified by a FET based Avantek 30 dB amplifier. The amplified RF signal is monitored using a Rohde & Schwarz FSUP signal source analyzer. The filter transfer function is measured for a 150 m MZI and results are depicted in Figure 6.5, where the max-hold feature of spectrum analyzer is employed, while sweeping frequency of the RF synthesizer in 5 kHz increments. Tuning capability was also measured for a 100 m 1st order filter. In this case, the DFB laser was replaced by a widely tunable laser (IDPhotnics CoBrite DX4) to vary optical wavelength. The 100 m 1st order filter was tested from 1534 nm to 1565 nm and frequency response is shown in Figure 6.6. A tuning of 33 kHz/nm is experimentally measured that is comparable to the simulated performance.
Figure 6.5 Transfer function of 1\textsuperscript{st} order MZI with 150m delay, as an optoelectronic transversal filter.

Figure 6.6 Experimental results of frequency tuning of a 1\textsuperscript{st} order optical transversal filter using a 100m long fibers at wavelengths of 1534nm (blue) and 1565nm (green).

In order to test the phase noise performance of the optical transversal filter, the MZI is placed in an opto-electronic oscillator, as depicted in Figure 6.7. The experimental results of opto-electronic transversal filter are reported for the first time here. The MZI follows the optical delay to provide side-mode suppression, as an alternative to multi loop ILPLL techniques reported in [84]-[86]. Output ports 3 and 4 are electrically combined after detection by photodiode (PD) using a Wilkinson power combiner (i.e., a 3dB coupler), as opposed to optical combination to avoid higher phase noise generated due to the optical phase fluctuations for optical interference. For differential lengths longer than the coherent length of laser diode, the optical phase fluctuation is insignificant for optical MZI. Phase noise measurements are
performed using the Rohde & Schwarz FSUP. A 3km delay is selected for the OEO delay. The 1st order transversal filters with the lengths of 30m, 100m, 150m, and 500m are inserted after the 3km delay (cf. Figure 6.7); phase noise measurement results are shown in Figure 6.8. The experimental results with the filters demonstrate phase noise roll-off slope of -30dB/decade at close-in to carrier until 10kHz offset. Lowest achieved phase noise at 10kHz offset is -133dBc/Hz for the case of 30m, while about -125dBc/Hz is measured for other lengths. Possible reason for the phase noise improvement using 30m filter is that mode-locking between a larger number of coupled modes in the transversal filter based OEO reduces the mode-partition noise [56]-[61].

Figure 6.7 Experimental set-up of OEO with optoelectronic transversal filter. Various lengths of fiber are considered both in terms of OEO and also opto-electronic transversal filter implementations.

Figure 6.8 Measured close-in to carrier phase noise of a 3km OEO using various transversal filter lengths of 30m (red), 100m (blue), 150m (black), 500m (magenta).
6.2.3 Passively Temperature Compensated OEO Incorporating Raman Amplification

The high quality factor associated with long standard fiber optic delays is the primary advantage of OEOs. However, the refractive index (n) and length (L) of standard fiber used in these optoelectronic systems vary with environmental factors [87]-[88]. In particularly, temperature (T) fluctuation in harsh environments causes the oscillation frequency (f) to drift, resulting in a poor thermal stability. The frequency variation due to temperature of an OEO can be expressed as:

\[
\frac{\Delta f}{f \Delta T} = -k \left[ \left( \frac{\Delta n}{n \Delta T} \right) + \left( \frac{\Delta L}{L \Delta T} \right) \right]
\] (6.2)

Here \(k = \left( 1 + \frac{\tau_{\text{fix}}}{\tau_{\text{Fiber}}} \right)\) where \(\tau_{\text{fix}}\) is the delay associated with the fixed lengths and \(\tau_{\text{Fiber}}\) is the controllable delay of the OEO. Active temperature compensation technique is proposed in standard fibers to reduce frequency fluctuations that most fiber based OEOs suffer from, where fluctuation rate of 8 ppm/C is reduced to 0.1 ppm/C [89]. To achieve a reduced temperature sensitivity of optical fiber delay lines, Kaba et al. used solid core photonic crystal fiber [90] and demonstrated a factor of 3 reduction in temperature sensitivity compared to SMF–28 fiber, but this reduction is still not sufficient and hence the system begs for alternate temperature stabilization techniques. Photonic crystal fibers with air voids are proposed [91]-[92] as index of refraction of air experiences negative temperature sensitivity [93]. Daryoush et al. report on a new technique of passive temperature compensation by utilizing a composite fiber structure consisting of silica based fiber cascaded with a hollow core photonic crystal fiber (HC-PCF) in [91]. The negative temperature sensitivity slope of effective refractive index of HC-PCF and positive temperature sensitivity slope of silica based fibers (e.g., Corning’s SMF-28) were combined in appropriate length ratios into a composite fiber delay line to passively mitigate the frequency sensitivity of the OEO system.

Although this passive temperature compensation technique was useful for the OEO in achieving thermal stability, it comes at the cost of degradation in the close-in to carrier phase noise of the 10 GHz OEO, as shown in Figure 6.9. A 65 dB degradation is measured in the phase noise of the OEO, which was
attributed to additional electrical and optical amplifiers required to compensate for the excessive optical losses, but this paper identifies it as primarily due to excessive losses (2 dB/m) of photonic crystal fibers.

Figure 6.9 Measured single data point at 10 kHz offset is fitted to Leeson’s close-in to carrier phase noise model of OEO using 1 km standard fiber (red) and using composite fiber (blue) of 30 m HC-PCF and 970 m standard fiber.

that results in reduction of an effective Q of composite resonant structure. A technique of compensating optical losses in the photonic crystal fibers by taking advantage of Raman amplification using nonlinear Raman scattering in silica portion of HC-PCF is proposed. The presented analytical modeling describes practical conditions, where reduced optical losses results in delay lines with effectively high fiber delay line Q factor. This increase in the quality factor of the composite fiber is also maintained with passive temperature compensation, where excellent thermal stability and extremely low close-in to carrier phase noise are predicted in the OEO.

To mitigate the effect of degradation of phase noise due to increased attenuation of composite fiber delay line, Raman amplified composite fiber delay line is proposed by the author for the first time [94]. More specifically, by taking advantage of nonlinear Stimulated Raman Scattering (SRS) in the silica portion of the HC-PCF, the effective fiber attenuation is reduced and hence resulting in increased quality.
factor of the composite structure. With respect to the other nonlinear optical fiber amplification processes (e.g., stimulated Brillouin scattering, Erbium doped fiber amplifier), Raman amplification (RA) is the most attractive technique due to its benefits like flat gain over broad bandwidths, improved noise performance and reduced nonlinear penalty allowing for longer fiber lengths, and closer channel spacing. Raman amplification results from the nonlinear scattering process in optical fibers wherein a part of the pump optical energy at a higher optical frequency is transferred to the signal at a lower frequency. The optical gain spectrum is peaked, when optical signal spectrally coincides with the Stokes lines at 13.2 THz lower than the pump frequency. In the proposed approach, Raman amplifier gain is designed to compensate for losses associated with the PCF in the composite fiber such that the signal output power remains very close to the input optical power. Therefore, the overall loss/gain of the fiber would be approaching 0 dB, and the effective loss is expressed as $\alpha_{\text{effective}} = \frac{P_S(L) - P_S(0)}{L}$, where $P_S(L)$ and $P_S(0)$ are the signal power levels at length of L and 0 km respectively. The pump laser source is judiciously selected with the optical signal source wavelength and it can be either forward (along the direction of signal) or reverse direction (opposite to the direction of the signal) propagating, but reverse pumping is preferred over forward pumping because the Pin Photodiode based optical receivers would not suffer from a higher shot noise and nonlinear limitations of the forward propagating systems. The location of the pump source would be at the beginning of the PCF fiber in forward pumping so that the signal and pump wavelengths enter the fiber at the same point and travel in the same direction and at the end of the fiber for reverse pumping so that the signal and pump wavelengths enter the fiber at its two ends and travel in the opposite directions.

To achieve effective loss/gain of 0 dB over a fiber length in a practical OEO, pump power levels have to be appropriately selected for the fiber length of choice. For example for an extremely long standard optical fiber delay line of length of 80km with 0.33 dB/km attenuation (e.g., Corning SMF-28 fiber), a pump power of 20dBm(22dBm) is required for 5dBm (10dBm) input signal power. Table 6.2 depicts the comparison of the quality factor of standard SMF-28 fiber with that of the composite fiber before and
after Raman amplification. On the other hand, the required laser pump power is around 35dBm for a 1km length of commercial HC-PCF fiber with 15 dB/km attenuation (e.g., NKT Photonics HC-1550-04) for signal power level of 5dBm. As technology of HC-PCF has advanced and now commercial products are readily available (though at a very high cost) with attenuation levels of 15dB/km compared to 2dB/m of circa 2005, we anticipate that attenuation levels could be reduced from 15 dB/km gradually to 1dB/km in the next few years. Therefore, for two practical signal powers of 5dBm and 10dBm prediction of the required pump power and associated maximum length are depicted in Figure 6.10, as an effective optical attenuation 0.33dB/km is used as an attenuation of SMF-28 fiber while the optical attenuation of HC-PCF fiber is reduced from 10dB/km in gradual steps to values of 5dB/km, 3dB/km, and eventually to 1dB/km.

In Table 6.2 performance comparison of OEO in terms of effective Q and close-in to carrier phase noise at 10kHz offset are made for cases of with (shaded column) and without (unshaded columns) Raman Amplifier (RA). Note passive temperature compensation is attained with HC-PCF to SMF-28 fiber length ratio of 38:1; moreover, the achieved maximum length of composite fiber increases from 1.23 km to 3km as attenuation of the HC-PCF fibers are reduced from 15dB/km to 10dB/km for the same pump power of 35dBm. Moreover, the optimum length can be increased to 10km for a pump power of only 30dBm. The calculated effective optical attenuation of the composite fiber with RA are about 0.1dB/km for the 3dB/km, 0.16dB/km for 10dB/km, and even approaching 0.14dB/km for commercially available HC-PCF (NKT Photonics HC-1550-04) with loss of 15dB/km. Note that with RA there is a Q enhancement for the composite fiber, while without RA a loss dominated by HC-PCF causes a significant reduction in the effective Q. The resulting achieved close-in to carrier phase noise at two offset frequencies of 1 kHz and 10kHz from 10GHz carrier are also rendered in Table 6.2. A significant reduction in phase noise in composite fibers with RA with an effective loss of 0.1dB/km and 10 kHz/$C^0$ drift is estimated due to the enhancement of about 7 dB compared to OEO with SMF-28 fiber.
Figure 6.10 Minimum optical pump power versus optical fiber lengths to overcome excessive loss in the composite fiber using HC-PCF with optical attenuation of 1dB/km, 3dB/km, 5dB/km, and 10dB/km for signal power levels of 5dBm(solid lines) and 10dBm(dashed lines).
<table>
<thead>
<tr>
<th></th>
<th>Phase Noise (dBc/Hz) @ 10kHz with RA</th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>-143</td>
<td>-150</td>
<td>-146</td>
<td>-147</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>Phase Noise (dBc/Hz) @ 10kHz without RA</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Phase Noise (dBc/Hz) @ 1kHz with RA</td>
<td>-123</td>
<td>-130</td>
<td>-126</td>
<td>-111</td>
<td>-107</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Phase Noise (dBc/Hz) @ 1kHz without RA</td>
<td>-</td>
<td>-7</td>
<td>-3</td>
<td>4</td>
<td>87</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Phase Noise Degradation with RA</td>
<td></td>
<td>Q_2</td>
<td>Q_4</td>
<td>Q_6</td>
<td>Q_8</td>
<td>Q_10</td>
<td>Q_12</td>
</tr>
<tr>
<td>Phase Noise Degradation without RA</td>
<td>-</td>
<td>22</td>
<td>5</td>
<td>3</td>
<td>6</td>
<td>4</td>
<td></td>
</tr>
<tr>
<td>Relative Q with RA</td>
<td>0.10</td>
<td>0.16</td>
<td>0.14</td>
<td>0.16</td>
<td>0.14</td>
<td>0.16</td>
<td></td>
</tr>
<tr>
<td>Relative Q without RA</td>
<td>0.10</td>
<td>3</td>
<td>10</td>
<td>3/35</td>
<td>1.23/35</td>
<td>Composite fibers with passive Temperature Compensation of Lp:Ls = 38:1</td>
<td></td>
</tr>
<tr>
<td>Effective Att. of composite fiber with RA ((\alpha_s) in dB/Km, (\alpha_c=0.33)dB/k)</td>
<td>-</td>
<td>3</td>
<td>10</td>
<td>3/35</td>
<td>1.23/35</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Effective Att. of HC-PCF without RA ((\alpha_s) in dB/Km)</td>
<td>-</td>
<td>3</td>
<td>10</td>
<td>3/35</td>
<td>1.23/35</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Fiber Delay Length (km)/ Pump power (dBm)</td>
<td>10/NA</td>
<td>10/30</td>
<td>10</td>
<td>3/35</td>
<td>1.23/35</td>
<td>Composite fibers with passive Temperature Compensation of Lp:Ls = 38:1</td>
<td></td>
</tr>
</tbody>
</table>
REFERENCES


111


[70] Hewlett-Packard Product Note 11792C-2


**APPENDICES**

**Appendix A. Radio over Fiber (RoF) Link Characterization**

*a) Characterization of MZM Link*

![Fig. A1. Experimental Setup for Open Loop Characterization of RoF link using MZM](image)
Fig. A2. MZM Output Optical Power as a Function of Bias Voltage

Table A1. Link Loss of Different MZM Bias Point

<table>
<thead>
<tr>
<th>MZM Link</th>
<th>MZM Bias</th>
<th>Link Loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>MZM Link 1</td>
<td>0V</td>
<td>47dB</td>
</tr>
<tr>
<td>MZM Link 2</td>
<td>-0.5V</td>
<td>42dB</td>
</tr>
</tbody>
</table>

b) Characterization of EAM Link

Fig. A3. Experimental Setup for Open Loop Characterization of RoF link
Fig. A4. RF Power vs EAM Bias (Optical Input = 7dBm, RF Driving = 10dBm)

Table A2. Link Loss of EAM Link w/ and w/o EDFA

<table>
<thead>
<tr>
<th>EAM Bias</th>
<th>Link Loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>w/o EDFA</td>
<td>1V</td>
</tr>
<tr>
<td>w/ EDFA</td>
<td>1V</td>
</tr>
</tbody>
</table>
Appendix B Noise Estimation of RoF Link

a) Analysis of MZM Link

<table>
<thead>
<tr>
<th>Noise Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>RIN = -140dBc</td>
<td>Laser RIN</td>
</tr>
<tr>
<td>$I_{\text{photo}}=2\text{mA}$</td>
<td>DC photo current of the photo detector</td>
</tr>
<tr>
<td>R=50Ω</td>
<td>Load resistance of photo detector</td>
</tr>
<tr>
<td>$f_C=1\text{MHz}$</td>
<td>Amplifier flicker frequency</td>
</tr>
<tr>
<td>$G_A=10\text{dB}$</td>
<td>Amplifier Large Signal Gain</td>
</tr>
</tbody>
</table>

Noise Spectral Density

$$S_N(f_m) = \frac{(N_{\text{RIN}} + N_{\text{Shot}}) \times G_A}{P_S} \times \left(\frac{f_c}{f_m} + 1\right)$$

SSB phase noise

$$\mathcal{L}_N(f_m) = 10 \log_{10} S_N(f_m) - 3\text{dB}$$

where

$$N_{\text{RIN}} = I_{\text{photo}}^2 \times R \times RIN \times G_A = -137\text{dBm/Hz} \quad \text{and} \quad N_{\text{Shot}} = 2e \times I_{\text{photo}} \times R \times G_A = -165\text{dBc/Hz}$$
Fig. A5. Simulated noise floor and measured phase noise of different circuit topologies

*b) Analysis of EAM Link*

<table>
<thead>
<tr>
<th>Noise Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>RIN = -120dBc</td>
</tr>
<tr>
<td>I_{photo} = 2mA</td>
</tr>
<tr>
<td>R = 50Ω</td>
</tr>
<tr>
<td>f_c = 1MHz</td>
</tr>
<tr>
<td>G_A = 10dB</td>
</tr>
</tbody>
</table>
Fig. A5. Simulated noise floor and measured phase noise of different circuit topologies
Appendix C OEO characterization

\( f=8.8\text{GHz}, P_s=5\text{dBm} \)

<table>
<thead>
<tr>
<th>Delay (m)</th>
<th>1kHz</th>
<th>10kHz</th>
<th>100kHz</th>
<th>1MHz</th>
<th>10MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>100m</td>
<td>-69</td>
<td>-101</td>
<td>-121</td>
<td>-134</td>
<td>-144</td>
</tr>
<tr>
<td>500m</td>
<td>-80</td>
<td>-112</td>
<td>-124</td>
<td>-135</td>
<td>-143</td>
</tr>
<tr>
<td>1000m</td>
<td>-91</td>
<td>-120</td>
<td>-130</td>
<td>-135</td>
<td>-144</td>
</tr>
<tr>
<td>5000m</td>
<td>-101</td>
<td>-127</td>
<td>-131</td>
<td>-121</td>
<td>-140</td>
</tr>
</tbody>
</table>
Appendix D Loop Filter Characterization

An active low pass filter is constructed using an op-amp. The circuit of the active filter is shown in Figure D1. The active filter transfer function is shown in equation (D.1). Different resistors and capacitors are used to achieve different PLL loop bandwidth. The calculated loop bandwidth is given in Table D.1. Experiments are performed to verify the PLL loop bandwidth estimation. In the experiment, a DRO (Synergy DRO100) is locked to a synthesizer (HP 8340B), and the loop bandwidth is deduced from the measured phase noise of the DRO under locked condition, experiment data is shown in Figures D.2, D.3, and D.4.

![Active Filter Circuit Topology](image)

Figure D.1. Active Filter Circuit Topology

<table>
<thead>
<tr>
<th>Board</th>
<th>Resistor and Capacitor Values</th>
<th>Loop BW for ( K_d = 0.1 ) and ( K_o = 200 \text{kHz/V} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>( R_1 = 51 \Omega, R_2 = 1k\Omega, C = 4.7nF )</td>
<td>about 200kHz</td>
</tr>
<tr>
<td>2</td>
<td>( R_1 = 51 \Omega, R_2 = 3.3k\Omega, C = 0.27nF )</td>
<td>about 700kHz</td>
</tr>
<tr>
<td>3</td>
<td>( R_1 = 51 \Omega, R_2 = 100\Omega, C = 470nF )</td>
<td>about 20kHz</td>
</tr>
</tbody>
</table>
Circuit 1: \( R_1 = 51\, \Omega, R_2 = 1\, k\Omega, C = 4.7\, nF \)

Figure D.2 Phase noise of DRO when it is locked to HP8340B. PLL Loop bandwidth about 200kHz

Circuit 2: \( R_1 = 51\, \Omega, R_2 = 3.3\, k\Omega, C = 0.27\, nF \)

Figure D.3 Phase noise of DRO when it is locked to HP8340B. PLL Loop bandwidth about 700kHz
Circuit 3: $R_1=51\,\Omega$, $R_2=100\,\Omega$, $C=470\,\text{nF}$

Figure D.4 Phase noise of DRO when it is locked to HP8340B. PLL Loop bandwidth about 20kHz
VITA

Academics:

*Southwest Jiaotong University, China*

2009 – 2011  **M.S. – Electrical Engineering**  
*Drexel University, Philadelphia, PA*

2011 – Present  **Ph.D – Electrical Engineering**  
*Drexel University, Philadelphia, PA*

Experiences:

1.  Research Assistant at ECE Department Drexel University (Advisor: Dr. Afshin Daryoush)
   - Characterized performance of DFB-Laser, Mach-Zehnder Modulator, Polarization Controller, High Speed Photodiode using Optical Spectrum Analyzer and Optical Power Meter
   - Measured frequency drift of an 8GHz VCO with 5-port Band Pass Filter (BPF) over a 15 minute period using Spectrum Analyzer and Characterized injection locking range and Q factor of the VCO by locking it to an HP Frequency Synthesizer
   - Designed microwave matching networks, filters, couplers, phase shifters, switches and LNA using Agilent ADS
   - Fabricated microwave circuits on PCB using LPKF milling machine for frequencies between 850MHz – 1.5GHz and characterized the circuits using Agilent 40GHz Network Analyzer, Spectrum Analyzer and Power Meter.

2.  Teaching Assistant at ECE Department Drexel University
   - ECEE 471/518 – 472/519 – 473/517, ECEC 356, ECE 201, ENGR 202, ECES 302

3.  Internship at China Telecomm (Guangdong, China)
   - Took part in the maintenance of add-drop multiplexers in SDH system
Projects:

1. Frequency Stabilization of 10GHz Optoelectronic Oscillator (Advisor: Dr. Afshin Daryoush)
   - Analytical modeling for phase noise of VCO in standard Phase-Locked Loop (PLL) and standard Injection Locking(IL) and standard Injection Locked Phase Locked Loop
   - Analytical modeling for phase noise of VCO in Self-PLL, Self-IL and Self-ILPLL

2. Calculation of Q factor in Long Optical Fiber
   - Improved modeling of Q factor by taking into account both the attenuation and the length of the delay
   - Approximated the total energy stored in optical fiber using a Gaussian mode profile

3. Phase Noise Characterization of a Feedback Oscillator
   - Constructed system level oscillator circuits in the Agilent ADS
   - Performed simulation using Leeson’s equation and ADS built-in circuit level modeling to compare performance accuracy and identify the modeling limitation of the Agilent ADS program.

4. CMOS Push-Pull Amplifier
   - Analysis of a push-pull amplifier in ADS using ‘Harmonic Balance’ frequency-domain simulator and ‘Transient’ time-domain simulator
   - The design utilizes the proprietary IBM 180nm SiGe CMOS technology for use between 8 GHz – 12 GHz with the best possible phase noise characteristics with over 1Watt of output power.

Publications:


