Tuning of the Relative Phases of a Rotary Travelling Wave Oscillator

by

Raleigh Smith, B.Sc. Eng., P. Eng.

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Abstract

The work that follows describes the implementation and use of a simple phase detector for phase measurement, and an algorithm to adjust variable Transmission Line (TL) capacitance to tune the phases of an existing RTWO.

Symmetrical XOR (SXOR) phase detectors were designed, fabricated in SiGe BiCMOS and tested to resolve 1° of phase error over $45^\circ \pm 10^\circ$ and $90^\circ \pm 10^\circ$ up to 12 GHz. These phase detectors were used to measure and calibrate the phase of 45° transmission line sections of an 18-GHz RTWO fabricated in SiGe BiCMOS. This phase tuning was carried out by a simple phase alignment algorithm that can be implemented easily in verilog. This RTWO phase tuning system was shown to function correctly in simulation. Although the fabricated version did not converge correctly, the algorithm was able to adjust the phase of RTWO TL sections. The circuit problem preventing convergence is understood and can be corrected.

To Dolce and Dessi.

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Nomenclature

16PSK	16 Phase Shift Keying
32PSK	32 Phase Shift Keying
8PSK	8 Phase Shift Keying
ADC	Analogue to Digital Converter
ADPLL	All Digital Phase Locked Loop
AMS	Analog Mixed Signal
AMSD	Analog Mixed Signal Designer
BBPD	Bang Bang Phase Detector
BiCMOS	Bipolar/Complementary Metal-Oxide Semiconductor Technology
BPSK	Binary Phase Shift Keying
C	Capacitance
C_{μ}	Base to collector capacitance of a bipolar transistor
C_{CB}	Collector to Bulk Capacitance of a bipolar transistor
CDR	Clock and Data Recovery
CMOS	Complementary Metal-Oxide Semiconductor
CW	Continuous Wave
C_{π}	Base to emitter capacitance of a bipolar transistor
DAC	Digital to Analogue Converter
DCO	Digitally Controlled Oscillator
DTI	Deep Trench Isolation
DUT	Device Under Test
DVM	Digital Volt Meter
ECL	Emitter Coupled Logic
FCW	Frequency Control Word
FD	Fractional Divider

FPGA	Field-Programmable Gate Array
f_T	The frequency at which the current gain of a transistor is zero
FUS	Frequency Up-Sampling
GSG	Ground-Signal-Ground semiconductor probe
GSSG	Ground-Signal-Signal-Ground semiconductor probe
I_C	Collector Current of a bipolar transistor
I_E	Emitter Current of a bipolar transistor
IL	Insertion Loss
K_{phase}	Phase detector gain or transfer function [9]
L	Inductance
LDO	Low Drop-Out voltage regulator
LPF	Low Pass Filter
MD	Multimodulus Divider
mimCap	Metal-Insulator-Metal Capacitor
NPORT	Cadence® N port s-parameter block
NXOR	Exclusive NOR logical function
P1dB	The input power level of an amplifier at which its output power level, due to compression, is 1.0 <i>dB</i> below its expected small signal level.
PSK	Phase Shift Keying - digital modulation of carrier phase
PVT	Process, Voltage and Temperature variation
QAM	Quadrature Amplitude Modulation
QPSK	Quadrature Phase Shift Keying
R	Resistance
RF	Radio Frequency
RTWO	Rotary Travelling Wave Oscillator
SD	Standard Deviation
SPST	Single-Pole Single-Throw switch function
SWO	Standing Wave Oscillator
SXOR	Symmetrical Exclusive OR gate implementation [10]
TDC	Time to Digital Converter
TL	Transmission Line
TS	Time Sample
V _c	Control Voltage
VNA	Vector Network Analyser

V_T	Semiconductor thermal voltage $\approx 26\text{ mV}$ at room temperature
XOR	Exclusive OR function
α	Alpha - $\alpha = I_C/I_E$ for a bipolar transistor
τ_D	Propagation delay of a logic circuit [11]
τ_F	Transistor forward transit time [11]

Chapter 1

Introduction

1.1 Motivation

There is a growing demand for common use wireless digital communications systems to operate at high RF and now microwave frequencies. This trend has pushed oscillator and synthesizer design to where the Rotary Travelling Wave Oscillator (RTWO) is an appropriate implementation choice. One of the advantages an RTWO has over LC-oscillators is that it produces multiple phases of the operating frequency period. The availability of these phases allows system architects and designers to re-think conventional circuit implementations, some of which are discussed in Section 1.2.

The Rotary Travelling Wave Oscillator (RTWO) illustrated in Figure 1.1 consists of a continuous transmission line loop folded over on itself to form a crossover and a number of back-to-back inverter or gain elements. During steady-state oscillation a wave front travels around the loop in either a clockwise or anticlockwise direction. The instantaneous phases marked on Figure 1.1 (i.e., 0° , 45° , $90^\circ \dots$) show a possible path of a single wave front revolution around the loop. The frequency of oscillation is established by this electrical length, which includes the guided wavelength of the transmission lines and the gain element reactive parasitics.

Figure 1.1: Idealized Illustration of a Four Section RTWO [1]

As the wave travels around the closed loop the transmission lines store energy and then pass it to the gain elements. The gain elements replenish energy lost in the transmission lines and device resistances, and impose odd-mode or differential operation. The low loss and loaded-Q of the transmission lines allow the RTWO to have both relatively low power dissipation and low phase noise at microwave frequencies of oscillation.

Figure 1.1 illustrates that any phase relationship, from 0° to 360° , can be extracted from the RTWO by tapping the travelling wave at relative points on the Transmission Line (TL) structure. This is a significant advantage over other types of oscillators. However, the accuracy of the phase relationship required for various applications is very difficult to achieve due to the inherent asymmetry of the RTWO TL structure (i.e., length of inner versus outer loop, and cross-over discontinuities) and device mismatch. That is, for a phase error of less than 5° in a production environment with a satisfactory yield, lengthy design iterations and costly re-spins may be required [5], but in the end may not be enough.

The phase noise performance and power consumption of an RTWO is superior to that of a conventional LC-Oscillator as the frequency of operation increases past a

few GHz (i.e., microwave frequencies) [5, 12]. Additionally, as the frequency of operation increases the size of the RTWO decreases, which reduces the cost of fabrication. These as well as the phase tapping advantage mentioned earlier are significant motivation to justify research on finding ways to make RTWOs realizable in a production environment.

1.2 RTWO Phase Tuning Applications

The following paragraphs discuss some of the applications that can be improved by using multiple calibrated phases of an RTWO as a clock source. The two main sources of quantization error (non-linearity) in an All Digital Phase Locked Loop (ADPLL) are the Time to Digital Converter (TDC) and Digitally Controlled Oscillator (DCO). In Figure 1.2 the multiple phases of an RTWO DCO are used to reduce the quantization error in the feedback loop for a fractional divide ratio. A Bang-bang Phase Detector (BBPD) and Adaptive Loop Gain Controller replace the complexity and non-linearity of the TDC. Normally, fractional division ratios are created by alternating the ratio of the Multimodulus Divider (MD) between two consecutive integers, which can result in a large phase discontinuity in the divided clock. Here an RTWO DCO creates m phases that are used by a fractional divider to reduce the large phase error that results from the changes in divide ratios at the MD. Within the fractional divider there is an integer divide ratio of N . F phases are multiplexed on F consecutive clock cycles of f_0 . This reduces or increases the divider clock (f_{CLK}) time by $(F/N)(1/f_0)$. Increasing f_{CLK} cycle time will cause glitches that must be suppressed in the output clock. The Frequency Control Word (FCW) selects the f_0 . The $\Sigma\Delta$ /SS ($\Sigma\Delta$ /Spread Spectrum) block uses the FCW (integer + fraction) to modulate the divide ratios of the MD and Fractional Divider (FD) blocks to reduce and shape the

quantization error (nonlinearity) in the fractional divide ratio [2, 13].

Figure 1.2: Fractional-N ADPLL with RTWO DCO [2]

Figure 1.3 shows a Digital Radio Polar Transmitter with an RTWO DCO [3] and Outphasing Modulator for a Software-Defined Transmitter [14]. The DCO generates the carrier frequency f_{RF} and the FCW (Frequency Control Word) selects RF channel. QAM-16 signals are generated using polar variables (amplitude - ρ and phase - ϕ). Phase modulation signals (i.e., BPSK, QPSK, 8PSK, 16APSK, 32APSK) can be generated modulating ϕ only, while amplitude modulation (QAM) and power setting require amplitude (ρ) control. Grey Encoding can be used to reduce the bandwidth requirements for QAM signals. QAM-16 can be created using 12 phases and four amplitudes, while QAM-64 has 52 phases and 16 amplitudes.

Figure 1.3: RTWO DCO in a Digital Radio Polar Transmitter [3]

Figure 1.4 illustrates a 6-Port Network Receiver that requires four phases of a DCO [4]. Here a DCO similar to that used in Figure 1.3 can be used to generate these signals.

Figure 1.4: RTWO DCO in a 6-Port Network Receiver [4]

Other applications whose performance at high frequencies could be improved by phase tuning of RTWO clock signals are Clock Distribution Networks [15], Phase-to-Digital Converter for an ADPLL [16] and oversampling circuits or phase tracking loops used in clock and Data Recovery (CDR) circuits for data alignment [17].

1.3 Thesis Objectives and Scope

The primary objective of this thesis is to demonstrate that the relative phases of oscillating signals tapped from an RTWO can be tuned to have an accuracy of better than 1° . This accuracy was selected as a good standard for image rejection in a digital radio. That is, for an image rejection of 40 dB the phase error between the I and Q oscillator signal needs to be less than 1.2° and the amplitude error needs to be less than or equal to 2% [18].

It should be noted that this work does not include the development of an RTWO. The RTWO and single varactor elements for each transmission line section used in this thesis were developed by Kimia Ansari [6] and supplied as black-box component. The nominal operating frequency of this RTWO is ≈ 18 GHz.

This work also discusses some limitations of CMOS and Bipolar technologies for the implementation of high speed phase detectors used to phase tune an RTWO. That is, an RTWO is able to oscillate at very high frequencies (i.e., 18 GHz and 36 GHz [19]) relative to the current switching performance of transistors implemented in that same technology. This makes phase measurement on a cycle-to-cycle basis very challenging.

1.4 Contributions

The main contribution of this thesis is to demonstrate that fine tuning of relative phases of a RTWO operating at microwave frequencies can be carried out using simple phase detectors, voltage controlled varactors and a straight forward iterative calibration algorithm. In order to limit power dissipation, this calibration can be carried out during initial system power-on and thereafter periodically or after frequency changes.

Also contributed here are implementation details of RTWO phase calibration i.e., phase detection, technology selection, calibration of phase calibration circuits, impact of rotation direction on the calibration algorithm.

1.5 Thesis Organization

Chapter 1 (Introduction) discusses the reasons for carrying out this research, as well as research objectives, scope and some possible circuit applications. Contributions to industry are also briefly presented. Chapter 2 (RTWO Concepts) provides some technical background of the RTWO that is useful going forward and the basic RTWO phase tuning concept. Chapter 3 (RTWO Phase Calibration) presents system level proof of concept circuits and simulations of RTWO phase tuning, first using Matlab/Simulink and second using Cadence AMS Designer. Chapter 4 (Phase Measurement Circuit) describes the phase detector architecture selection and implementation, as well as the removal of unequal differential transit delays. Chapter 5 (RTWO Phase Tuning Verification) details the extracted simulation of both the SXOR phase detector and the phase tuning of an RTWO. Chapter 6 (Experimental Results) discusses the lab test set-ups, calibration procedures and test procedures used to gather measurement results from the SXOR phase detector test cards and the RTWO phase alignment test cards. A list of conclusions, contribution and recommendation is documented

in Chapter 7 (Conclusions and Recommendations). Appendix A describes the sub-blocks used to implement the Simulink RTWO 24 MHz block; Appendix B contains three versions of phase alignment algorithm verilog code (i.e., Cadence, FPGA and Matlab); and Appendix C displays the schematics, layout and BOMs for the two test cards used in this work.

Chapter 2

RTWO Concepts

2.1 Oscillator Fundamentals

All systems designed to oscillate must, at minimum, satisfy the two requirements of the Barkhausen criteria. First, the circuit must include a feedback path that adds a delayed version of the forward signal, the feedback signal, to the forward path. The feedback signal must be added constructively to the forward path. This is illustrated in Figure 2.1.

Figure 2.1: Classical Linear Feedback Model

Theoretically, a loop of wire or Transmission Line (TL) meets this criterion exactly, at every point along the loop, for a signal that oscillates with an integer number of guided wavelengths equal to the wire loop length. A Rotary Travelling Wave

Oscillator (RTWO) is made by creating a twist in the TL loop (i.e., a figure eight) and folding one of the resulting loops inside the other, see Figure 2.2. This creates a differential structure that is one half the length of the differential guided wave path.

Figure 2.2: [5] a) TL Loop; b) Twist; and c) Differential RTWO TL Structure

Secondly, after a circuit has reached steady state oscillation, the closed loop gain must be equal to 1.0. It should be noted that to ensure oscillation is initiated at start-up, grows to steady-state and is sustained, the open loop gain must be greater than 1.0. This allows a very small signal perturbation (i.e., circuit noise) to be amplified to overcome circuit losses and build into an oscillating signal. Once the closed loop system reaches steady state, the circuit operates in a non-linear fashion to suppress any gain in the circuit in excess of 1.0. Mathematically, this can be shown using (2.1) (i.e., the transfer function for the classical linear feedback model of Figure 2.1):

$$H(s) = \frac{Y(s)}{X(s)} = \frac{A(s)}{1 - B(s)A(s)} \quad (2.1)$$

Here, for the gain $B(s)A(s) = 1.0$, $H(s) = \infty$; therefore, the signal will be compressed into non-linearity.

Gain in the feedback path of an RTWO is achieved by inserting a number of back-to-back -Gm elements between differential points on the RTWO structure, as

illustrated in Figure 2.3. This structure of TL sections and gain elements (e.g., four -Gm sections, $N = 4$) also forces the system to operate differentially (odd mode). The RTWO rendering in Figure 2.3a) is somewhat theoretical since loops of finite TL width and equal circumference cannot fit inside one another. Figure 2.3b) is an RTWO layout with one cross-over, but a large difference between the inside loop and outside loop signal path lengths, and Figure 2.3c) is an RTWO layout with three cross-overs and a smaller loop signal path length difference than b). In a real implementation the placement of the gain stages in these layouts needs to be offset from what is shown below to maintain nominal 45° TL sections.

Figure 2.3: a) Conceptual RTWO [1]; b) Single Cross-Over; c) Three Cross-Over

2.2 The Distributed Oscillator

An RTWO can be considered a distributed amplifier special case (i.e., output fed back to input with a 180° twist in the loop and satisfy Barkhausen's criteria) where the drain and gate parasitic capacitances of the differential gain stages are absorbed into the RTWO transmission lines [5, 20, 21]. This significantly raises the gain-bandwidth product of the gain stages, which allows the RTWO to have gain and oscillate at higher frequencies than would be expected from a conventional LC-oscillator of the same process. Theoretically, the operating frequency is limited by only the f_T of the

amplifier circuit technology [15]. This is straight forward to simulate and has been realized in 130 nm CMOS at 18 GHz and 36 GHz in [19].

2.3 RTWO Frequency Calculation

The discussion that follows is a summary of concepts found in [1, 5], as well as other sources, and is included here to provide background understanding. After RTWO start up the steady-state wave front will propagate in the direction (i.e., either clockwise or anticlockwise) offering the lowest loss [7, 15]. The phase velocity of the wave is $V_p = 1/\sqrt{L_0 C_0}$, where L_0 and C_0 are the differential inductance and capacitance of the RTWO per unit length. In this thesis the RTWO is considered to be a differential structure and the signals that are extracted or tapped from it to be differential signals. If we select a differential position on the RTWO (e.g., differential tapping points $0^\circ/180^\circ$), then the range of the full physical circumference of the RTWO is 180° of phase relative to this physical starting point. The remaining 180° of phase is realized by reversing the tapping points of the same physical starting position (i.e., $180^\circ/0^\circ$) and travelling around the RTWO physical circumference again. Therefore, two laps of the RTWO are required to complete one cycle or wave length, λ , and for differential length l we have a boundary condition $2l = \lambda n$, where $n = 1, 3, 5 \dots$. The fundamental frequency tends to be dominant because it receives the largest loop gain and losses are much higher for the harmonics. Thus, the RTWO frequency f_0 is given by [5]:

$$f_0 = \frac{V_p}{2l} \quad (2.2)$$

$$= \frac{1}{2l\sqrt{L_o C_o}} \quad (2.3)$$

$$= \frac{1}{2\sqrt{LC}} \quad (2.4)$$

$$= \frac{1}{2N\sqrt{L_{diff} C_{diff}}} \quad (2.5)$$

where, V_p is the phase velocity of the travelling wave, L and C are the total inductance and capacitance of the RTWO structure, L_{diff} and C_{diff} are the total differential inductance and capacitance of a TL section and N is the total number of sections so that [5]:

$$L = lL_0 = NL_{diff} \quad (2.6)$$

$$C = lC_0 = NC_{diff} \quad (2.7)$$

These are the classical equations used to predict the fundamental oscillating frequency, f_0 , of an RTWO.

2.4 Superposition and Injection Locking

The startup and steady state oscillation of an RTWO can be described as the superposition of a series of standing wave oscillators (SWO) [22] that will injection-lock [23] with each other to create a stable travelling wave oscillator [1, 5, 24, 25]. Figure 2.4a) shows a $\lambda/4$ SWO. This SWO is realized by a $\lambda/4$ TL from the -Gm gain elements to the short circuit, the short produces a -180° reflection of the incident wave, and the $\lambda/4$ TL from the short circuit to the -Gm gain elements produces the in-phase positive feedback required by Barkhausen. The phase is fixed with position, but

the amplitude varies with position to produce a standing wave. Figure 2.4b) can be thought of as two $\lambda/4$ SWOs connected at the $-G_m$ elements to produce a $\lambda/2$ SWO of the same frequency as Figure 2.4a). Following this train of thought, Figure 2.4c) is also a $\lambda/2$ SWO formed by only considering a single set of $-G_m$ elements of an RTWO with shorts at the Nulls. This TL loop is formed by unfolding and untwisting the RTWO TLs (i.e., the reverse of Figure 2.2).

Figure 2.4: a) $\lambda/4$ SWO; b) $\lambda/2$ SWO [5]; and c) Loop SWO [5]

In Figure 2.4c) the $-G_m$ elements inject current into the TLs to produce a clockwise wave $V_a(t, Z)$ and an anticlockwise wave $V_b(t, Z)$ as follows [5]:

$$V_{a0}(t, Z) = A_0 \cos(\omega_0 t - \beta z) \quad (2.8)$$

$$V_{b0}(t, Z) = B_0 \cos(\omega_0 t + \beta z) \quad (2.9)$$

where A_0 and B_0 are peak voltage amplitudes, $\beta = 2\pi/\lambda$ is the propagation constant and z is the physical position around the ring. The distribution of voltage along the ring is derived in [5] and shows that:

$$V_{a0}(t, Z) + V_{b0}(t, Z) = 2B_0 \cos(\omega_0 t) \cos(\beta z) + (A_0 - B_0) \cos(\omega_0 t - \beta z) \quad (2.10)$$

where the first term describes the behaviour of a standing wave and the second term describes the behaviour of a travelling wave. When the circuit is perfectly symmetrical $A_0 = B_0$ only the standing wave will be present. The dashed lines in 2.4c) illustrate the motion of the standing wave oscillation.

Now we consider that the N TL sections of a RTWO have a temporal offset of $T_0/(2N)$, where $T_0 = 1/f_0$, and a positional offset of $\lambda/(2N)$. Using equation(2.10), the superposition of N standing waves (assuming perfect symmetry) gives the following equation [5]:

$$V_0(t, z) = \sum_{n=0}^{N-1} \left[2B_0 \cos \left\{ \omega_o \left(t - \frac{nT_o}{2N} \right) \right\} \cos \left\{ \beta \left(z - \frac{n\lambda}{2N} \right) \right\} \right] \quad (2.11)$$

A more complete derivation of the following traveling wave equations is provided in [5]. The positional offset is $2\lambda/N$ for anticlockwise equation:

$$V_0(t, z) = \sum_{n=0}^{N-1} B_0 \cos(\omega_o t - \beta z) \quad \text{clockwise rotation} \quad (2.12)$$

$$V_0(t, z) = \sum_{n=0}^{N-1} B_0 \cos(\omega_o t + \beta z) \quad \text{anticlockwise rotation} \quad (2.13)$$

Equations (2.12) and (2.13) both describe travelling waves. The current injected by one set of -Gm elements into the TLs and next set of -Gm elements injection locks the SWOs and produces a travelling wave.

The simulated signals of Figure 2.5 are from the TL sections of an $N = 4$ RTWO (similar to Figure 2.3) implemented in a 130 nm CMOS process using CMOS buffers as -Gm elements and coupled TL models. The coupled TL sections have a length = $55 \mu m$, width = $8 \mu m$, separation = $16 \mu m$, and the RTWO $f_0 = 24 GHz$. Here start-up was initiated by a 1.0 V initial condition applied to the 0° TL section, so

the 0° and 180° points begin to oscillate first as standing waves with the diagonally opposite points (i.e., 90° and 270°) as nulls. Just prior to 4 ns this superposition of SWO behaviour injection locks into the travelling wave behaviour of an RTWO. With oscillators that do not require this initial condition at start-up this incipient standing wave mode is very short and the RTWO will achieve its travelling wave mode almost immediately. It should be noted that equation (2.10) predicts the peak amplitude of the standing wave mode oscillations should be twice that of the travelling wave mode oscillations. In this design the -Gm CMOS buffers will limit this amplitude to 1.2 V .

Figure 2.5: 24 GHz, 1.2-V CMOS RTWO TL Signals w.r.t. Ground

2.5 Phase Adjustment Concepts

Phase adjustment of the RTWO is achieved by tuning the variable capacitance of individual line sections, which is realized with varactor banks connected across the TLs within each line section. This concept was identified in [1] and is summarized below with the aid of Figure 2.6.

Figure 2.6: Lumped Element RTWO with Proposed Phase Calibration Method [1]

In Figure 2.6 the four TL sections of an RTWO have been modelled as lossless simplified lumped LC elements. L_{Diff} is the total differential inductance of each TL section. Each capacitive element represents the sum of the mutual capacitance of the differential TL section, the total associated driver NPN transistors capacitances (i.e., base to collector C_μ , base to emitter C_π and collector to bulk C_{cB} [6]), and the nominal variable capacitance of varactor banks installed in that TL section. The varactor capacitance of an individual TL section can be controlled independently of other TL sections to locally adjust the phase velocity of the travelling wave front. The example of Figure 2.6 shows that the travelling wave phase velocity is either slowing down or speeding up relative to the previous TL section for a wave front travelling from left to right. Here equal amounts of varactor capacitance (ΔC) are added to where the phase is leading and subtracted from where the phase is lagging so that phase is corrected, but the overall frequency of the RTWO remains constant.

It should be noted that in this example both phase measurement and adjustment are independent of the direction of wave front rotation. A side effect of this method of phase tuning is that the amplitude of the travelling wave will vary around the RTWO as the impedance of individual TL sections is de-tuned.

If differential phase detectors are installed to measure the relative phase across each TL section, then an algorithm can be created to iteratively adjust the phases of TL sections as required and wait for the RTWO to settle. These iterations would continue until the phase across all TL sections was within a predefined tolerance, then the algorithm would stop. It is assumed that the purpose of the phase alignment algorithm is to tune all TL sections of an RTWO to have equal phase measurements. Any single phase detector measurement is relative to the other phase detector measurements carried out on a particular RTWO device. There is no requirement for any phase measurement to be absolutely accurate to an external standard.

2.6 Phase Tuning Limits

Reference [1] demonstrates mathematically that the clockwise or anticlockwise travelling wave modes of oscillation are the only stable states of the RTWO and it will not revert back to the standing wave mode, which is an unstable mode. However, this assumes that each SWO is injection locked at a common frequency. That is, as with any injection locked system there is both a minimum injection current requirement and injection frequency bandwidth beyond which locking will not occur. Adjusting the ΔC of individual TL sections changes the V_p and frequency of those TL sections. As the frequency of adjacent TL sections approaches the injection locking limit, the phase noise will degrade and spurs will appear.

Reference [1] also documents a simulated injection lock range of $\Delta C = \pm 6 \%$

on the phase tuning of individual TL sections. Additionally, smaller changes of the V_p (ΔC) of individual TL sections produces signals of slightly different frequencies on the RTWO, which increases the phase noise of the system. Their fabricated prototype demonstrated 4 dB of phase noise degradation for a tuning range of more than 10° (i.e., $\Delta C \approx \pm 2.5\%$). The study of these limitations is beyond the scope of this thesis.

2.7 Phase Change Calculation

The lumped inductance L_{Diff} and lumped capacitance C_{Diff} of Figure 2.6 form a Low Pass Filter (LPF) model of an RTWO TL section and are used as a starting point for deriving an equation that relates ΔC to phase. An ABCD parameter description of this LPF is as follows:

$$\begin{vmatrix} A & B \\ C & D \end{vmatrix} = \begin{vmatrix} 1 & Z_L \\ 0 & 1 \end{vmatrix} \begin{vmatrix} 1 & 0 \\ Y_C & 1 \end{vmatrix} = \begin{vmatrix} 1 + Z_L Y_C & Z_L \\ Y_C & 1 \end{vmatrix} \quad (2.14)$$

where $Y_C = j\omega C_{Diff}$ and $Z_C = j\omega L_{Diff}$. Using the conversion between two port network parameters (i.e., ABCD to s-parameter) found in [26], s_{21} can be found from equation (2.14) to give:

$$\begin{aligned} s_{21} &= \frac{2}{1 + Z_L Y_C + Z_L/Z_o + Y_C Z_o + 1} \\ &= \frac{2}{2 - \omega^2 LC + j\omega L/Z_o + j\omega C Z_o} \end{aligned} \quad (2.15)$$

substituting $L = Z_o^2 C$ into equation (2.15) we get:

$$\begin{aligned} s_{21} &= \frac{2}{2 - \omega^2 Z_o^2 C^2 + j2\omega Z_o C} \\ &= \frac{2(2 - \omega^2 Z_o^2 C^2 - j2\omega Z_o C)}{(2 - \omega^2 Z_o^2 C^2)^2 + (2\omega Z_o C)^2} \end{aligned} \quad (2.16)$$

Solving s_{21} for phase we get:

$$\theta = -\arctan\left(\frac{2\omega Z_o C}{2 - \omega^2 Z_o^2 C^2}\right) \quad (2.17)$$

Therefore, the change in phase for a change in variable capacitance in an RTWO TL section can be determined from:

$$\Delta\theta = \arctan\left(\frac{2\omega Z_o(C + \Delta C)}{2 - \omega^2 Z_o^2(C + \Delta C)^2}\right) - \arctan\left(\frac{2\omega Z_o C}{2 - \omega^2 Z_o^2 C^2}\right) \quad (2.18)$$

Equation (2.18) is useful for estimating the change in phase of an RTWO TL section for a change in varactor capacitance in that RTWO TL section. A change in varactor capacitance is due to a change in varactor control voltage. Armed with these two relationships an estimate of the required correction voltage step size can be made, see Section 5.2. However, it should be noted that this equation gives, at best, a first order approximation. This is because an RTWO is a dynamic system where a single parameter change creates a new steady state. This equation only considers

the starting point of a single TL section; not the resulting new steady state of the whole RTWO. That is, assuming TL sections are not isolated from one another, a capacitance change made in one TL section will change the electrical phase length in every TL section of a particular RTWO. An example of this behaviour is shown in Figure 3.14, where for any single phase adjustment the varactors of only two TL sections are adjusted. Yet the phase of all four TL sections changes to give the appearance that all four TL sections are being adjusted.

Additionally, when a TL section capacitance is changed its characteristic impedance also changes, which changes signal amplitude and the impedance transition between TL sections. The parameter levels under which injection locking to produce a travelling wave system (injection current and bandwidth) have also changed. Taken to an extreme, TL capacitance adjustment can cause frequency domain spurs to appear and degeneration to SWO operation (i.e., failure to injection lock) as discussed in Section 2.6.

Chapter 3

RTWO Phase Calibration

3.1 The System Concept

Top-Down development has proven to be the design methodology for systems and circuits that results in predictable performance in the shortest development cycles. This begins with system implementation and simulation at the highest level of abstraction. System architects and designers will define the system as a group of functional blocks and interconnecting signals so they may write comprehensive specifications for these blocks and signals. Each block at a particular level of abstraction is made up of blocks and signals from the next lower level of abstraction. This approach promotes logical partitioning, as well as a complete and effective understanding of the system at any level of abstraction while design details at other levels are hidden until required. It also provides a starting point for bottom-up iterations that will help to refine the system specification to a practical and cost-effective realizable point.

This methodology may also be used to demonstrate a proof of concept. That is, to find the answers to questions like will the system work logically and mathematically, and does a particular implementation achieve the original goals. This chapter first discusses the Matlab[®]/Simulink[®] circuits and then the Cadence[®] Virtuoso[®] AMS

Designer circuits that were created and simulated to demonstrate the concept of RTWO phase tuning.

3.2 Proof of Concept - Matlab[®]/Simulink[®]

Figure 3.1 is a functional block diagram of the system used to demonstrate the concept of RTWO phase tuning. The block RTWO 24 MHz contains the RTWO divided into eight 45° TL sections tapped to provide differential outputs $\phi 1/\phi 2$, $\phi 3/\phi 4$, $\phi 5/\phi 6$ and $\phi 7/\phi 8$. The inputs DlyCtl1 to DlyCtl4 are the variable delay control signals associated with each tapped differential TL section. The simulation frequency was reduced to 24 MHz to shorten the real time duration of individual simulation runs. Parasitic components are not implemented in this model so this change in operating frequency does not detract from the demonstration.

Figure 3.1: Simulink® System Level RTWO Phase Calibration Circuit

Point A of Figure 3.1 is a sample of the rail-to-rail differential output clock signals ($\phi 7$ and $\phi 8$) of the RTWO after it has reached steady-state oscillation, see Figure 3.2.

Figure 3.2: 1.2 V Rail-to-Rail RTWO Differential Output Clock Signals (Point A)

These 1.2 V rail-to-rail (i.e., CMOS voltage levels) clock signals are then converted to differential Current Mode Logic (CML) voltage levels (i.e., $0.85V \pm 0.350mV$) using CML buffers. This conversion to CML was necessary because in a realizable circuit CMOS gates will not support signals operating at a frequencies in the order of 18 GHz. The CML clock signals are fed to differential to single ended buffers (Buff) and to XOR gates. The Buff block outputs, point F, show the clock phases with phase skew intentionally programmed in the RTWO as a starting point for the phase alignment simulation, labelled Before Calibration in Figure 3.3. Specifically, the phases marked $\Delta 1 = 59.6^\circ$, between $\phi 1/\phi 2$ and $\phi 3/\phi 4$, $\Delta 2 = 43.6^\circ$, between $\phi 3/\phi 4$ and $\phi 5/\phi 6$, $\Delta 3 = 47.2^\circ$, between $\phi 5/\phi 6$ and $\phi 7/\phi 8$, and $\Delta 4 = 29.6^\circ$, between $\phi 7/\phi 8$ and $\phi 2/\phi 1$. It should be noted that these four phase deltas will always sum to 180° in an RTWO that has reached steady-state travelling wave oscillation. The signals of Figure 3.3, labelled After Calibration, show all these phase differences equal to $45^\circ \pm Limit$, where *Limit* is a predefined resolution, see RTWO Phase Alignment Flow Chart Figure 3.9. Here the calibration algorithm has iteratively adjusted the phase deltas, determined that the phase error between all the measured signals is

within a predefined resolution and has stopped the algorithm as no further adjustment was required.

Figure 3.3: Output Phases (Point F) Before Calibration - Top; After Calibration - Bottom

3.2.1 RTWO Phase Measurement

In Figure 3.1 the differential clock signals representing the relative phases of the four differential TL sections of the RTWO are applied consecutively to the inputs of four XOR phase detectors. The XOR gates produce a duty cycle signal that is proportional to the phase taken up by each differential TL section. In Figure 3.4 we can see that the phase difference between signals In1+ and In2+ is represented by the duty cycle of the Out signal.

Figure 3.4: CML XOR Phase Detect - Phase in to Duty Cycle Out (Points B & C)

It should be noted that the phase of the differential inputs feeding the bottom XOR from the top CML gate (i.e., ϕ_1, ϕ_2) is reversed so the measurement will be over the single remaining TL section rather than the three TL sections already measured. In fact, this fourth XOR phase detector is redundant as this phase difference at steady-state can be determined by subtracting the sum of the other three measured phases from 180° as follows:

$$\Delta N = 180^\circ - \sum_{n=1}^{N-1} \Delta n \quad (3.1)$$

where for this RTWO $N = 4$.

The outputs of these XOR phase detectors are single-ended CML (i.e., $0.85V \pm 0.350mV$) clock signals of twice the frequency of the input signals with output duty cycles proportional to the input phase differences. These duty cycles are averaged through a Low Pass Filters block (LPF), which consists of two cascaded single order LPFs with cut-off frequencies of 500 kHz followed by 2.0 MHz. Next the Buff blocks applies a dc offset of $-0.6V$ and a gain of five to produce the signals at point D,

i.e., Out1/4 of Figure 3.5. The amplitudes of these signals are proportional to the measured phases at each iteration of the phase alignment algorithm.

Figure 3.5: Phase Measurement and Corresponding RTWO Delay Corrections Signals (Points D & E)

Next these signals are converted to binary numbers with a 4-Channel 12-Bit Analogue-to-Digital Converter (ADC). These numbers are passed to the Phase Alignment Algorithm block, which finds the TL section that has the largest and the TL section has the smallest phase measurement. It then uses this information to adjust variable delay signals Dly1/4o, which control the phase delay of each differential TL section of the RTWO. These functions are triggered every $20 \mu s$ by timer T1. This cycle time is set to be longer than the settling time of the RTWO (i.e., $< 13 \mu s$) after a change in phase delay. Figure 3.5 illustrates these signals at point E for each cycle of T1 (i.e., iteration of the Phase Alignment Algorithm). In Figure 3.1 the $3.6 \mu s$ Step block holds off Phase Alignment Algorithm at simulation start until after the

RTWO 24 MHz block has been initialized i.e., the initial TL section delays have been programmed, see Table 3.1.

The Out1/4 signals of Figure 3.5 show numbers that represent the measured phase at the beginning through to the end of the simulation. After the algorithm makes the first phase measurement, i.e., $Out1 = 2244$, $Out2 = 1188$, $Out3 = 1425$ and $Out4 = 262.6$, it will adjust the delay of the largest and smallest phase delays using signals Dly1/4o. The algorithm (and simulation) will stop when all these numbers are within seven of each other, which is illustrated by the measurement results $Out1 = Out2 = Out3 = Out4 = 1280$.

Each channel of the 4-Ch 12-Bit ADC block of Figure 3.1 has a full range of 1.2 V. The 12-Bit ADC plus the coding gain of 2.322 *bits* (i.e., a gain of 5 or 14 *dB* from the previous analogue blocks) gives an effective 14.322-Bit ADC, which results in the following measurement resolution:

$$1.2/(2^{14.322} - 1) = 58.6 \mu V \quad (3.2)$$

$$180/(2^{14.322} - 1) = 8.79 m^\circ \quad (3.3)$$

$$58.6 \mu V / 8.79 m^\circ = 6.67 mV/^\circ \quad (3.4)$$

This result can be stated more simply from the full-scale voltage and full-scale phase as follows:

$$1.2 V / 180^\circ = 6.67 mV/^\circ \quad (3.5)$$

The algorithm stop resolution of seven mentioned earlier is equivalent to $7 \times 8.79 m^\circ = 61.53 m^\circ$. At the same time this coding gain reduces the maximum measurement range to $180^\circ / 5 = 36^\circ$.

3.2.2 The Simulink[®] RTWO Block

Figure 3.6 is the RTWO 24 MHz block at the next lower level of abstraction. Here we see four distinct TL sections made up of Gain, Delay and 45° Section blocks each delineated by differential clock signal tapping points. Also visible are the TL section delay control inputs (i.e., DlyCtl1/4), which determine the phase delay of blocks Delay 1 to Delay 4. Gain 1 to 4 provide the back-to-back or differential gain that supports oscillation in the RTWO, see Figure 2.3. 45° Section 1 to 4 insert nominal loss, nominal 45° phase delay and a LPF characteristic into the TL sections. The 0.2 V, 0.2 ns Step block provides a signal perturbation to disrupt the steady-state condition of the RTWO and initiate the build up of an oscillating signal in the loop at simulation start.

Figure 3.6: Simulink[®] Implementation of a Nomially 24 MHz RTWO

Prior to running the Simulink[®] model a Matlab[®] file, `setup.m`, must be run to initialize the simulation Time Step (TS) and other variables used in the model (i.e., FUS, Freq, Loss, alpha, slew and offset_n). The 0.2 V, 0.2 ns Step block in the Simulink[®] model, Figure 3.6, is where this TS is established and other blocks in the system inherit this parameter from this block. The TS is calculated follows:

$$TS = 1/(FUS \times Freq) \quad (3.6)$$

where, $FUS = 3600$ and $Freq = 24 \times 10^6$.

Variable Frequency Up Sampling (FUS) establishes that there will be 3600 TSs per cycle at the nominal operating frequency of 24 MHz. It should be noted that the frequency of the RTWO is not determined by these variables. The frequency of a particular RTWO simulation is controlled by the delay elements in its TL paths.

The delay of the four differential TL sections is initialized to produce the phase error or skew that will be corrected by the phase alignment algorithm during simulation. The initial phase skew is set up in the 45° Section blocks of Figure 3.6 using `setup.m`, i.e., $Z^{-(FUS/8)+offset_n}$ in Tables 3.1 and 3.2. In Figure 3.5 we see that after the calibration algorithm begins it initializes the variable Delay blocks to 250 TSs so both increases and decreases in the delay tuning can be made to this value. The rows across Table 3.1 show the total delay and resulting skewed phases for each TL section at initialization, i.e., $Variable\ Delay\ Block + FUS/8 + Offset_n = Total\ (TS)$ for a TL section.

Table 3.1: Transmission Line Phase Delay at Simulation Start

TL Section	Initial 45° TL Section (TS)				Δ Phase (°)
	Variable Delay Block	FUS/8	Offset_n	Total	
1	250	450	-104	596	59.6
2	250	450	-264	436	43.6
3	250	450	-228	472	47.2
4	250	450	-404	296	29.6
Total Delay/Phase for One Revolution				1800	180.0

The frequency of oscillation, f_0 , can be calculated from the total delay of one

revolution, Table 3.1, of the RTWO as follows:

$$f_0 = 1/(2 \times 1800 \times TS) = 24.0 \text{ MHz} \quad (3.7)$$

With this starting point the simulation ran for $23.5 \mu s$, which included a $3.5 \mu s$ initialization period and 10 correction iterations of $2.0 \mu s$ each. The total delay for each TL section after the simulation was completed is shown in Table 3.2. We see that the variable delay block delays were adjusted to make the totals approximately equal, see right hand side of Figure 3.5 signals Dly1o to Dly4o. Therefore, the phases have been adjusted to be equal.

Table 3.2: Transmission Line Phase Delay at Simulation End

TL Section	Final 45° TL Section (TS)				Δ Phase (°)
	Variable Delay Block	FUS/8	Offset_n	Total	
1	104.4	450	-104	450.4	45.015
2	264.5	450	-264	450.5	45.025
3	228.1	450	-228	450.1	44.985
4	404	450	-404	450.0	44.975
Total Delay/Phase for One Revolution				1801	180.00

The total delay for one revolution is approximately the same as it was at initialization; therefore, using equation (3.7) $f_0 = 23.987 \text{ MHz}$ at simulation end and has been constant throughout phase calibration. This is because the phase adjustment algorithm adjusts the largest and smallest phase delays errors by the same amount during each of its iterations. Figure 3.7 shows the FFT spectrum of the rail-to-rail signal, ϕ_8 of Figure 3.2, to be similar to this calculated value, approximately 23.90 MHz. This simulated FFT frequency value did not change throughout the simulation.

Figure 3.7: Simulink® FFT at RTWO Output (Point A)

Additional sub-blocks of the RTWO 24 MHz Block are described in Appendix A.

3.2.3 The Phase Alignment Algorithm

The Phase Alignment Algorithm and its associated blocks are illustrated by Figure 3.8. The code for this block is implemented using a subset of Matlab® commands and has been included in Appendix B.3. The two timer blocks have been explained previously. The Stop output indicates when the algorithm has adjusted the RTWO phases to be within $45^\circ \pm \text{Limit}$ and has halted. The In1 to In4 signals of the Phase Alignment Algorithm block are the digitized phase measurements and the Dly1/4o outputs are the resulting RTWO delay adjustments (i.e., Point E of Figure 3.1 corresponds to Point E of Figure 3.8). The remaining input/output signals connect to a series of single TS delays contained in the Integer Delay Store block. This implements storage so the algorithm can base its current state on data from its previous state. The Sampling Rate Conversion block contains four Rate Transition blocks to convert the continuous signals from the algorithm block to discrete signals required by the Variable Integer Delay blocks shown in Figure A.2.

Figure 3.8: Simulink[®] Algorithm Blocks

Figure 3.9 is a flow chart representation of the Matlab[®]/Simulink[®] Phase Alignment Algorithm. During time T_0 (i.e., initial $3.6 \mu s$ of simulation) the RTWO variables are initialized, the RTWO is started and allowed to reach steady-state. After T_0 expires, T_1 generates a pulse every $2.0 \mu s$ until the simulation stops. On each pulse of T_1 the ADC phase measurements are read for all four differential TL sections, and the minimum and maximum phase delays are found. If the maximum phase delay minus the minimum phase delay is greater than a predefined limit (i.e., seven has been arbitrarily selected, other values will also work), then the TL delay skew is too

large and must be corrected. Otherwise, stop.

Next, the delay adjustment step size is determined (i.e., $Step = Limit \times 0.05$), which will be added to the minimum delay TL section and subtracted from the maximum delay TL section. The size of this step must be small enough so limit cycles will not occur that will prevent the algorithm stop condition from being met. The condition of multiple maximum and minimum TL sections is handled by dividing the step size by the number of equal length maximum and minimum TL sections to give Max_step and Min_step. Finally, the new variable delay values for the TL sections requiring adjustment are determined by adding Max_step and Min_step as required. After these new delays are programmed into the RTWO, the RTWO reaches a new steady-state after a settling time of $< 13 \mu s$. This procedure is repeated on the next rising edge of timer T1.

Figure 3.9: Simulink[®] RTWO Phase Alignment Algorithm Flow Chart

3.3 Proof of Concept - Cadence[®] Virtuoso[®] AMS Designer

Figure 3.10 is a functional block diagram of the system used to demonstrate the concept of RTWO phase tuning in the Cadence[®] Virtuoso[®] AMS Designer mixed signal circuit simulator environment.

Figure 3.10: Cadence® Virtuoso® AMS Designer System Level RTWO Phase Calibration Circuit

The RTWO 18 GHz block can be described as an implementation of Figure 1.1, where nodes $0^\circ/180^\circ$ are in the bottom left corner; nodes $45^\circ/225^\circ$ are in the bottom right corner; nodes $90^\circ/270^\circ$ are in the top left corner and nodes $135^\circ/315^\circ$ are in the top right corner of the structure. Similarly, for an anticlockwise travelling wave Outp/n.L is the end node of the left side; Outp/n.B is the end node of the bottom; Outp/n.R is the end node of the right side and Outp/n.T is the end node of the top

of the structure.

The TLs of the RTWO 18 GHz block have been implemented in several ways: initially, using a 130 nm CMOS kit microstrip TL models; then as 2.5D s-parameter data, generated from a TL layout using ADS-Momentum, in an NPORT block; and finally using ideal lumped R, L and C elements derived from ADS-Momentum simulation as described in [6]. The 130 nm CMOS kit TL models worked well; however, these models were not available in the NXP BiCOMS kit. Using the NXP BiCMOS kit, the NPORT s-parameter block using spline interpolation gave mixed results in a transient simulation. That is, after minor circuit changes the transient simulation would often function for a short period and then become unstable. The ideal lumped RLC RTWO implementation was more stable and ran more quickly, so it was used in many of circuit simulations that follow.

The Phase Alignment Algorithm block of Figure 3.10 is implemented in Verilog and is described in detail in section 3.3.3. The timer input T0i holds off the Phase Alignment Algorithm function for the first 35.0 *ns* of simulation while the initial RTWO phase error is set, the RTWO starts and reaches stead-state oscillation. After T0i is high, each rising edge of the 10 *ns* periodic clock signal at T1i initiates an iteration of the Phase Alignment Algorithm. Both these clock signals oscillate between 0 to 1.2 *V* and are generated from independent sources available in the simulator analogue library.

Figure 3.11 shows a sample of the differential output clock signals (Outp_T and Outn_T) at Point A, Figure 3.10, of the RTWO after it has reached steady-state oscillation.

Figure 3.11: RTWO Differential Output Clock Signals (Point A)

These $\pm 300 \text{ mV}$ differential clock signals are then converted to differential CML voltage levels (i.e., $0.85 \text{ V} \pm 0.350 \text{ mV}$) using CML buffers implemented in Verilog-A. This conversion to CML was chosen to support signals operating at a frequencies in the order of 18 GHz. The CML clock signals are fed to differential to single ended buffers (Buff) and to CML XOR gates. The Buff block outputs, Point D, show the clock phases with phase skew intentionally programmed in the RTWO as a starting point for the phase alignment simulation, labelled Before Calibration in Figure 3.12. Specifically, the phases marked $\Delta 1 = 48.08^\circ$ (Top), between $Outp_T/Outn_T$ and $Outp_R/Outn_R$, $\Delta 2 = 39.19^\circ$ (Right), between $Outp_R/Outn_R$ and $Outp_B/Outn_B$, $\Delta 3 = 42.11^\circ$ (Bottom), between $Outp_B/Outn_B$ and $Outp_L/Outn_L$, and $\Delta 4 = 50.61^\circ$ (Left), between $Outp_L/Outn_L$ and $Outn_T/Outp_T$. It should be noted that these four phase deltas will always sum to 180° in an RTWO that has reached steady-state travelling wave oscillation. The signals of Figure 3.12, labelled After Calibration, show all these phase differences equal to $45^\circ \pm Limit$, where $Limit$ is a predefined resolution, see

RTWO Phase Alignment Flow Chart Figure 3.17. Here the calibration algorithm has iteratively adjusted the phase deltas, determined that the phase error between all the measured signals is within a predefined resolution and has stopped the algorithm as no further adjustment was required.

Figure 3.12: Output Phases (Point D) Before Calibration - Top; After Calibration - Bottom

The labelled signals of Figure 3.12 show that the RTWO travelling wave front

rotation is clockwise, i.e., 135m315 precedes 90m270, which precedes 45m225, which precedes 0m180. The steady-state frequency of oscillation is approximately 17.95 GHz before calibration and approximately 17.86 GHz after calibration.

3.3.1 RTWO Phase Measurement

In Figure 3.10 the differential clock signals representing the relative phases of the four differential TL sections of the RTWO are applied consecutively to the inputs of four XOR phase detectors implemented in Verilog-A. The XOR gates produce a duty cycle signal that is proportional to the phase taken up by each differential TL section. In Figure 3.13 we can see that the phase difference between input signals is represented by the duty cycle of the output signal.

Figure 3.13: CML XOR - Phase in to Duty Cycle Out (Points B & C)

It should be noted that the phase of the differential inputs feeding the bottom XOR from the top CML gate (i.e., Outp.L and Outn.L) is reversed so the measurement will be over the single remaining TL section rather than the three TL sections already measured. In fact, this fourth XOR phase detector is redundant as this phase difference at steady-state can be determined by subtracting the sum of the other three measured phases from 180° , see equation (3.1).

The outputs of these XOR phase detectors are single-ended CML (i.e., $0.85V \pm 0.350mV$) clock signals of twice the frequency of the input signals with output duty cycles proportional to the input phase differences. These duty cycles are averaged through LPF blocks, which consists of two cascaded single order LPFs with cut-off frequencies of 79.6 MHz followed by 318.3 MHz. Next the single-ended input to single-ended output Buff block, implemented in Verilog-A, is used to provide gain and bias control. Here the gain of this block is set to one with a bias of $-0.5V$ to remove the $0.5V$ offset shown in the top plot of Figure 3.13, which passes directly through the LPF. The amplitudes of these signals (i.e., Right, Bottom, Left and top on Figure 3.10) are proportional to the measured phases at each iteration of the phase alignment algorithm.

Next these signals are converted to binary numbers using an ideal 4-Channel 12-Bit Analogue-to-Digital Converter (ADC) implemented in Verilog-A. These numbers are passed over 12-bit buses RPi[11:0], RPi[11:0], RPi[11:0] and RPi[11:0] to the Phase Alignment Algorithm block, which finds the TL section that has the largest and the TL section has the smallest phase measurement. It then uses this information to adjust variable capacitance shown in Figure 3.16, through 9-bit buses TCo[8:0], LCo[8:0], BCo[8:0] and RCo[8:0], which control the phase of each differential TL section of the RTWO. These functions are triggered every 10 ns by the periodic clock signal at T1i. This cycle time is set to be the same as the settling time of the

RTWO after a change in phase delay. Figure 3.14 illustrates these signals at Right, Bottom, Left and Top of Figure 3.10 for each cycle of T1i (i.e., iteration of the Phase Alignment Algorithm). In this figure 175 mV represents 45°. At $time = 35\text{ ns}$ the initial phase error, programmed into the RTWO by the Phase Alignment Algorithm (see Appendix B.1) is marked; at $time = 560\text{ ns}$ the final phase error is marked. After $time = 560\text{ ns}$ the Phase Alignment Algorithm ceases to make further phase corrections as the *Limit* requirement has been met.

Figure 3.14: Phase Measurement Output Voltages

The Buff blocks of Figure 3.10 have a maximum output signal of 0.7 V and the ADC reference voltage is 1.2 V, which results in the following measurement resolution:

$$(0.7/1.2)(2^{12} - 1) = 2388.75 \text{ bits} \quad (3.8)$$

$$0.7 \text{ V}/2388.75 \text{ bits} = 293.0 \text{ } \mu\text{V}/\text{bit} \quad (3.9)$$

$$180^\circ/2388.75 \text{ bits} = 75.35 \text{ } m^\circ/\text{bit} \quad (3.10)$$

$$293.0 \text{ } \mu\text{V}/\text{bit}/75.35 \text{ } m^\circ/\text{bit} = 3.\bar{8} \text{ mV}/^\circ \quad (3.11)$$

This result can be stated more simply from the full-scale voltage and full-scale phase as follows:

$$0.7 \text{ V}/180^\circ = 3.\bar{8} \text{ mV}/^\circ \quad (3.12)$$

The algorithm stop resolution or *Limit* was set to six for the above simulation to give $6 \times 75.35 \text{ } m^\circ/\text{bit} = 0.452^\circ$ i.e., $Limit = 0.452^\circ$. Table 3.3 shows the marked points of Figure 3.14 in degrees, i.e., by dividing the marker voltage by the result of equation (3.12).

Table 3.3: Phase Alignment Simulaiton Results

TL Section	Before Calibration (35.0 ns)	After Calibration (560.0 ns)
Left	50.61°	$\geq 44.826^\circ / \leq 45.217^\circ$
top	48.08°	45.217°
Bottom	42.11°	44.826°
Right	39.19°	$\geq 44.826^\circ / \leq 45.217^\circ$

Here we can subtract Bottom from Top to get the final phase error, i.e., $45.217^\circ - 44.826^\circ = 0.391^\circ$, which is less than the $Limit = 0.452^\circ$.

3.3.2 The Virtuoso[®] AMS Designer RTWO Block

The block RTWO 18 GHz of Figure 3.10 contains an RTWO implemented using the ideal lumped RLC elements of Figure 3.15, which are divided into eight 45° TL sections tapped to provide differential outputs at ports P1/P2, P3/P4, P5/P6

and P7/P8. The back-to-back inverters, marked -G_m, provide oscillator gain, and the output buffers provide isolation between the oscillator and load of any external circuitry. These gain elements were implemented using the NXP BiCMOS kit by K.T. Ansari of [6] and have been use here without modification or analysis.

Figure 3.15: Lumped Element RTWO based on Equations from [6, 7]

The digital control outputs TCo[8:0], LCo[8:0], BCo[8:0] and RCo[8:0] from the Phase Alignment Algorithm tune the variable capacitance of RTWO TL sections Top, Left, Bottom and Right, see Figure 3.10. Each 9-bit bus is connected to the [C256:C1] control signals shown in Figure 3.16 (i.e., C256 to bus signal 8 ... C1 to bus signal 0). Here $C = 500 \text{ aF}$ was determined through simulation.

Figure 3.16: Digitally Controlled Ideal Capacitor Tuning Array

An updated version of [6] provides the RLC values that were used as a starting point for the RTWO in this simulation. These values were determined by 2.5D ADS-Momentum simulations performed on the RTWO TLs. In this paper equation (3.13), a refinement of equation (2.2), was used to determine the operating frequency (f_o) of an RTWO with four TL sections, i.e., $n = 4$.

$$f_0 = \frac{1}{2l_{total}\sqrt{L_{perlen}\left(C_o + \frac{C_{seg}}{l_{seg}}\right)}} \quad (3.13)$$

where, $l_{total} = 1.488 \text{ mm}$ is the total physical length of the loop, $L_{perlen} = 423 \text{ nH/m}$ is the per unit length differential inductance of the loop, $C_o/n = 101 \text{ pF/m}/4 = 25.25 \text{ pF/m}$ is the per unit length differential capacitance of the interconnect, $C_{seg} = 254.55 \text{ fF}$ is the total differential capacitance of each TL segment and $l_{seg} = l_{total}/n = 1.488 \text{ mm}/4 = 372 \text{ }\mu\text{m}$ is the length of each TL segment. This gives:

$$\begin{aligned} f_0 &= \frac{1}{2(1.488 \times 10^{-3})\sqrt{423 \times 10^{-9}\left(25.25 \times 10^{-12} + \frac{254.55 \times 10^{-15}}{372 \times 10^{-6}}\right)}} \quad (3.14) \\ &= 19.40 \text{ GHz} \end{aligned}$$

which is consistent with [6]. This frequency was reduced to approximately 18 GHz after additional parasitics were included in the implemented design.

C_{seg} of equation (3.13) can be described as follows:

$$C_{seg} = C_{Var} + C_{-Gm} + C_{buff} \quad (3.15)$$

where, for each TL section C_{Var} is the variable tuning capacitance; C_{-Gm} is the

total capacitance associated with the back-to-back gain elements; and C_{buff} is the capacitance of the signal tapping buffers. In the system simulation of Figure 3.10 C_{Var} was set to a nominal value and the additional capacitor tuning array of Figure 3.16 was used for phase tuning. The starting capacitance of this array was $256 \times C = 256 \times 500 \text{ aF} = 128 \text{ fF}$. The nominal value of $C_{Var} = 46 \text{ fF}$; therefore, the RLC components were adjusted through simulation to compensate for this change. The simulated values are: $R = 50 \text{ m}\Omega$ (starting value $1977 \text{ }\Omega/m \times l_{total}/2n = 367.7 \text{ m}\Omega$); $C_{Diff} = 15.0 \text{ fF}$ (starting value $C_o \times l_{total}/2n = 18.786 \text{ fF}$) and $L_{Diff/2} = 54.0 \text{ pH}$ (starting value $L_{perlen} \times l_{total}/2n = 78.678 \text{ pH}$) to produce an operating frequency of approximately 17.9 GHz. It should be noted that values of $R \geq 80 \text{ m}\Omega$ caused the AMSD mixed signal simulation to become unstable.

3.3.3 The Phase Alignment Algorithm

Figure 3.17 is a flow chart representation of the Cadence[®] Virtuos[®] AMS Designer Phase Alignment Algorithm. When $T0i = 0 \text{ V}$ (i.e., initial 35 ns of simulation) the RTWO variables are initialized, the RTWO is started and allowed to reach steady-state. After $T0i = 1.2 \text{ V}$ the T1i receives a rising edge every 10 ns until the simulation stops. On each rising edge of T1i the ADC phase measurements are read for all four differential TL sections, and the minimum and maximum phase delays are found. The maximum phase delay minus the minimum phase delay is compared to three limits to determine the size of the correction step that will be made to the tuning varactors voltages. If the max/min phase delay difference is smaller than the smallest limit, then the algorithm will stop.

The size of the varactor tuning step must be small enough so limit cycles will not occur that will prevent the algorithm stop condition from being met. The TL sections with maximum and minimum phase delay are then determined and the tuning step is

subtracted from the minimum (varactor voltage decreased) and added to maximum (varactor voltage increased) phase TL sections. After these new varactor settings are programmed into the RTWO, the RTWO reaches a new steady-state after a settling time of $< 10 \text{ ns}$. This procedure is repeated on the next rising edge of timer T1i.

Figure 3.17: Cadence RTWO Phase Alignment Algorithm Flow Chart

The Verilog implementation of the flow chart of Figure 3.17 has been included in Appendix B.1. Using the result of equation (3.11) the Max Limit is $63 \times 75.35 \text{ m}^\circ/\text{bit} = 2.7126^\circ$ and the Medium Limit is $15 \times 75.35 \text{ m}^\circ/\text{bit} = 1.13025^\circ$, the Min Limit was previously calculated to be $Limit = 0.452^\circ$. The capacitance array tuning steps are $max_step = 11 \times 500 \text{ aF} = 5.5 \text{ fF}$, $medium_step = 3 \times 500 \text{ aF} = 1.5 \text{ fF}$, and $min_step = 1 \times 500 \text{ aF} = 500 \text{ aF}$.

Chapter 4

Phase Measurement Circuit

4.1 Phase Detector Selection

An RTWO can be considered a distributed amplifier special case (i.e., output signals flipped 180° and fed back to the input) where the drain and gate parasitic capacitances of the differential gain stages are absorbed into the RTWO transmission lines. This significantly raises the gain-bandwidth product of the gain stages, which allows the RTWO to have gain and oscillate at higher frequencies than would be expected from a conventional LC-oscillator of the same process. This is straight forward to simulate and has been realized in 130 nm CMOS at 18 GHz and 36 GHz in [19].

A Current Mode Logic (CML) XOR gate was selected as the best phase detector candidate based on its high speed operation [27], differential architecture and simplicity. That is, it has only two transistors in its switching (charging path) path; therefore, it will have fewer parasitic circuit elements and a greater potential for high speed operation when compared to more complicated implementations (i.e., sample and hold, flip-flop ...).

An additional advantage of an RTWO is that its structure is intrinsically differential. This advantage can be maintained in oscillator signal paths outside the

RTWO by tapping its relative phases differentially. With this approach the maximum relative phase difference between any two differentially tapped signals along the RTWO is 180° . Therefore, a simple phase detector like an XOR (XNOR) with a gain of $K_{phase} = \pm V_{range}/\pi$ (V/rad) [9] has an adequate measurement range for this application.

Figure 4.1 shows a differential XOR phase detector realized using CML. Tapped points along the RTWO are connected to the differential inputs A+/- and B+/- to produce outputs Out+/-, which are averaged through Low Pass Filters (LPF) to produce voltage outputs proportional to an input phase difference.

Figure 4.1: CML XOR Function with Averaged (LPF) Outputs

The XOR phase detector produces an output signal that is twice the input signal frequency with a duty cycle that is linearly proportional to and has a monotonic relationship with the phase difference between the two input signals through 180° .

Figure 4.2 illustrates the XOR phase detector transfer function after the duty-cycle of the output signal is averaged through a LPF to produce a constant level signal between V_{CC} and $V_{CC} - \Delta V$. Both outputs of Figure 4.2 are shown and the equations for average output voltages are as follows:

$$\begin{aligned}
V_{Out_Avg+} &= (V_{CC} - \Delta V) + K_{phase}(\phi A - \phi B) \quad (V) \\
&= (V_{CC} - \Delta V) + (\Delta V/\pi)(\phi A - \phi B) \quad (V) \quad (4.1)
\end{aligned}$$

$$V_{Out_Avg-} = (V_{CC} - \Delta V) + (\Delta V/\pi)(\phi B - \phi A) \quad (V) \quad (4.2)$$

Figure 4.2: CML XOR Transfer Function Including LPF

4.2 CML Circuit Implementation

CML was chosen as the XOR structure for this application because of the differential signal and speed requirements. Initial simulations were done using a 130 nm NMOS

differential transistor pair, but these transistors did not switch current completely on or completely off at microwave frequencies. The result was a non-linear transfer function that changed with frequency and would require a complex calibration scheme to linearize. This problem was mitigated by moving to a SiGe BiCMOS process. Fortunately, an 18 GHz RTWO had been implemented previously in this process, so the objective was changed to phase tune this RTWO.

The following steps were used to design the XOR circuit:

1. Base Resistance (R) and Capacitance (C) have been identified as the major parasitic components limiting the switching speed of SiGe HBT device [28, 29]. Next, in descending order of importance, is load RC, forward transit delay (τ_F), collector RC and finally emitter RC. In order to mitigate these parasitics minimum size, high speed, high current transistors were selected for this circuit (i.e., $width/length/emitters = 0.4 \mu m/1.3 \mu m/4$) [30].
2. Next a minimum size differential pair, implemented with these bipolar transistors, with resistive loads was verified to switch current on and off completely at more than 20 GHz.
3. Collector current (I_C) was selected for high speed switching. This was done by selecting an I_C that approaches maximum f_T (minimum τ_F from equation (4.7)) operation [29, 31, 32]. At maximum f_T I_C is approximately 18.0 mA and since $I_E = I_C/\alpha$ this current was selected as the bias or tail current for the differential transistor pairs.
4. The output voltage swing (ΔV) was selected next. ΔV_{min} needs to be large enough to switch the differential pair on or off completely i.e., $> 5V_T = 130 mV$ at room temperature [11, 33]. ΔV_{max} should be low enough to not impact circuit

speed (i.e., $\tau_D \propto R_L C_{EFF} \approx \Delta V / I_C$, so total gate delay is proportional to the voltage swing) and to allow easy biasing of stacked transistors between V_{CC} and ground. ΔV_{max} should be high enough so that the stacked differential transistor pairs of successive circuits will continue to switch current correctly over PVT. The range of $400 \text{ mV} \leq V_{max} \leq 500 \text{ mV}$ was chosen to meet these requirements [32].

5. The CML load resistors were chosen to be 24Ω . Q2 of Figure 4.3 will drive 18.0 mA through to the load resistors in the XOR topology, which gives $\Delta V = 18 \text{ mA} \times 24 \Omega = 0.432 \text{ V}$ to meet the requirements of item 4 in this list.

4.3 Removal of Unequal Transit Delays

Consider two differential signals, tapped from an RTWO, that have a phase difference of 90° . One differential signal is applied to the A+/- inputs of the XOR gate of Figure 4.3 and the other differential signal is applied to the B+/- inputs.

Figure 4.3: Bipolar CML XOR Schematic

If we assume that these signals are biased correctly, the transistors are all the same size, and if we ignore any impedance difference between the A and B inputs, and ignore delay paths that are common, the A inputs to outputs V0+/- path length ($delay = \tau_A$) is different from the B inputs to outputs V0+/- path length ($delay = \tau_B$) as follows:

$$\tau_{\Delta} = \tau_A - \tau_B \quad (4.3)$$

$$\tau_A = \tau_{BC} + \tau_{Trace} + \tau_F \quad (4.4)$$

$$\tau_B = \tau_{BC} \quad (4.5)$$

$$\begin{aligned} \tau_{\Delta} &= \tau_{BC} + \tau_{Trace} + \tau_F - \tau_{BC} \\ &= \tau_{Trace} + \tau_F \end{aligned} \quad (4.6)$$

where: τ_{Δ} is the time difference between the two paths; τ_{BC} is the transit delay from transistor base to collector; τ_{Trace} is the path delay through the traces from the collectors of transistors Q3 and Q4 to the emitters of transistors Q5 to Q8; and τ_F is the forward transit delay from transistor emitter to collector.

From equation (4.6) we can see that the XOR gate will have an output duty cycle error of τ_{Δ} , which will result in an output offset error in a phase detector application. This error can be minimized by keeping τ_{Trace} very short and operating the transistors at a current density for maximum f_T [11], where:

$$f_{TMAX} \approx 1/2\pi\tau_F \quad (4.7)$$

A more direct solution is to never introduce the problem in the first place by

constructing an XOR gate that is absolutely symmetrical. The Symmetrical XOR (SXOR) gate [10] is illustrated in Figure 4.4.

Figure 4.4: Bipolar CML SXOR Schematic

Here a replica of the basic XOR is added to the original XOR circuit to allow the A and B inputs to be applied to both the first and second levels of transistors so that the aggregate delays, τ'_A and τ'_B , are now equal. That is, τ'_A and τ'_B are a result of both the first and second level to output delays. For example, when inputs A+ and B+ are high, then in Figure 4.4 transistors Q3, Q5, Q10 and Q12 are on to pull V_{o+} low. If the same bias current is used with the SXOR circuit as was used in the XOR circuit, then the load resistors must be halved to produce the same output ΔV as the current passing through them is doubled.

Q1, Q2 and Q9 form a basic current mirror circuit to bias the switching transistors of the SXOR circuit. The emitter length of these transistors was increased to $10.3 \mu m$ (i.e., $width/length/emitters = 0.4 \mu m/10.3 \mu m/4$). This extended the constant

current operation of the mirror transistors Q2 and Q9 from a $V_{ce} = 0.5\text{ V}$ down to $V_{ce} = 0.2\text{ V}$.

Figure 4.5: SXOR and XOR Output Duty Cycles for 90° Inputs v. Frequency

Figure 4.5 shows the simulated output duty cycles of the XOR of Figure 4.3 and the SXOR of Figure 4.4 for an input phase difference of 90° over a frequency range of 2.0 GHz to 20.0 GHz. All transistors are minimum size high speed, high current bipolar devices described earlier. The input clock signals are square waves with an amplitude of 0.5 V and rise/fall times of 1 ps . The transient simulation output signals are averaged over the period of 4 ns to 6 ns for each point plotted. The SXOR plot of Figure 4.5 illustrates a significant improvement in accuracy over the XOR plot. The 0.78% duty cycle error at 2.0 GHz corresponds to 1.4° of phase error and the 7.51% duty cycle error at 20.0 GHz corresponds to 13.6° of phase error. This increase in error with frequency is nearly linear, which indicates that this error is almost completely due an increase in transit delay (τ_F) with frequency. The interconnect trace delay (i.e., τ_{Trace}) of equation (4.6) is dependent on microstrip TL geometry and the thickness and relative permittivity of the TL dielectric, and is independent

of frequency [34]. Therefore, in a realization of the XOR circuit this delay would increase the error of Figure 4.5 by a fixed amount. The SXOR topology will also compensate for this fixed error.

Figure 4.6 is similar to Figure 4.5 except that the input phase difference is 45° , which should result in a duty cycle of 25%. Here the error at 2.0 GHz is 0.78% as it was for the 90° case of Figure 4.5. However, as the frequency increases past 12.0 GHz both SXOR and XOR duty cycle outputs deviate from their expected linear paths. As stated previously, if the frequency of the input clock signals is f , then the SXOR/XOR transistors must switch at $2f$. As the phase relationship between the input clocks approaches 45° and output duty cycle approaches 25%, the switching of the SXOR/XOR transistors must occur in half the time it did for the 90° input phase relationship. This effectively doubles the switching frequency again to $4f$. Unfortunately, at an effective frequency of 48.0 GHz the transistors do not completely switch. This results in the non-linear deviation in the curves of Figure 4.6.

Figure 4.6: SXOR and XOR Output Duty Cycles for 45° Inputs v. Frequency

4.4 Input Biasing and ECL Outputs

Figure 4.7 shows the SXOR biasing or input level shifting circuit for inputs A and B (Figure 4.4) from the RTWO. The base signals A2+/- and B2+/- are connected to the bases of Q12 to Q15 and Q5 to Q8 of Figure 4.4. Q16 to Q19 provide emitter follower outputs to drive the bases of transistor Q3, Q4, Q10 and Q11 of Figure 4.4. The emitter area of these transistors was increased to $width/length/emitters = 0.4 \mu m/10.3 \mu m/12$ in order to provide enough current to maintain adequate voltage swing at the bases of Q3, Q4, Q10 and Q11. The emitter-follower loads are diode connected minimum size high speed, high current bipolar transistors (i.e., $width/length/emitters = 0.4 \mu m/1.3 \mu m/4$) Q20 to Q23 connected to resistors $Rbn = 10.2 k\Omega$.

Figure 4.7: SXOR Input Signal Bias Network Schematic

Q24 and Q25 of Figure 4.8 are emitter-follower transistors that isolate the CML outputs of the SXOR from the load. This is important so the switching characteristics of the SXOR are maintained over a range of loads while current is independently provided to any circuit that follows. Additionally, Q24 and Q25 reduce capacitive loading on the CML output node to maintain the speed of the circuit [35]. These transistors change the SXOR from CML at outputs Vo+/- to Emitter-Coupled Logic

(ECL) at outputs $V_{oe+/-}$.

Figure 4.8: SXOR ECL Output Transistor Schematic

4.5 SXOR Device Layout

Figure 4.9 illustrates the layout of the SXOR phase detector described in the previous section. This version 2 of the SXOR layout is $162 \mu m \times 130 \mu m$ at its largest points, which is less than half the size of the version 1 device. The CML load resistors, $R_L/2$ of Figure 4.4, are NXP kit type resPD poly $W/L/M = 10 \mu m \times 7 \mu m \times 4$ to make 12Ω . The current limit for unsilicided poly is $1 \mu m/mA$ for 10 year reliability [36,37]. Therefore, the total resistor *width* $= 4 \times W = 4 \times 10 \mu m = 40 \mu m$ gives a $40 mA$ limit, which is more than adequate for the $36 mA$ flowing through these devices. The common centroid layout (provides some immunity to process mismatch) of these resistor elements is labelled ABBA/BAAB in Figure 4.9 [38]. Additionally, there are four dummy resistors labelled R_{Dum} across the top of the layout that help with fabrication consistency of these and the emitter-follower output load resistors R_{ECL} . As a precaution, all R_{Dum} resistors are connected to ground on both sides. Resistors R_{ECL} are NXP kit type resPB [36] of size $W/L/M = 5 \mu m \times 9.4 \mu m \times 2$ to make 300Ω . The total resistor *width* $= 2 \times 5 \mu m = 10 \mu m$ gives a $10 mA$ reliability limit, which is more than adequate for the $6.3 mA$ of rms current flowing through these

devices.

While the SXOR topology equalizes the signal paths running vertically across the device, it does not address differences in the horizontal signal path lengths. This is achieved by making the signal paths lengths from input A2+ to Q12/Q15 = input A2- to Q13/Q14 = input B2- to Q6/Q7 = input B2+ = Q5/Q8; and input A2+ to Q3 = input A2- to Q4 = input B2- to Q11 = input B2+ to Q10. Some examples are illustrated in Figure 4.9. Care was taken to include via height when calculating the total length of these paths.

Dummy Resistors were not considered necessary for the input biasing network resistors R_{bn} as their specific values are not critical. They were realized using NXP kit type resPZ [36] of size $W/L/M = 1 \mu m \times 5.1 \mu m \times 1$ to make $10.2 k\Omega$. The total resistor width of $1.0 \mu m$ gives a $1.0 mA$ reliability limit, which is more than adequate for the $100 \mu A$ of rms current flowing through these devices.

Most signal paths within the SXOR layout were implemented using metal layers M1, M2 or M3 (nominally 737 nm thick) as they connect more directly to the transistors than layers M5 (nominally 2055 nm thick) and M6 (nominally 3055 nm thick). These traces were made wider than minimum to improve their I^2R losses. The trace capacitance to substrate ground was not considered an issue because significant charging currents were being switched. All traces cross each other at 90° to minimize inductive coupling. M5 was used to create a ground ring, which is tied down to substrate ground around the SXOR device. Ibias and Vcc, as well as Voe+/- were implemented using M6 [37].

Figure 4.9: Layout of SXOR with ECL Outputs

4.6 Phase Detector Rev1 versus Rev2

Two revisions of the differential phase detector test device, Figure 6.3, were fabricated. Both these die measure $1.0 \text{ mm} \times 1.0 \text{ mm}$. The inset box shown in Figure 4.10 (revision one, *Phdetect_1*) marks the boundary of the phase detector portion of the circuit layout, which is $296.4 \mu\text{m} \times 154.075 \mu\text{m} = 45,667.8 \mu\text{m}^2$. The inset box shown in Figure 4.11 (revision two, *Phdetect_2*) marks the boundary of the phase detector portion of the circuit layout, which is $161.175 \mu\text{m} \times 128.675 \mu\text{m} = 20,739.2 \mu\text{m}^2$.

Figure 4.10: SXOR Phase Detector Version One Die (*Phdetect_1*)

This is a layout area reduction of 56%, which was achieved by repositioning transistors to allow shorter and more symmetrical trace routing.

Figure 4.11: SXOR Phase Detector Version Two Die (*Phdetect_2*)

Chapter 5

RTWO Phase Tuning Verification

5.1 Extracted RTWO Phase Detector Verification

Figure 5.1 is a functional block diagram of the simulation test bench used to verify the performance of the extracted ECL SXOR phase detector. This verification has two objectives: first, to verify the monotonicity of the output voltage $V\phi$ for an Input Phase Δ ramp; and second, to determine the phase detector resolution.

Figure 5.1: Simulation Test Bench for Phase Detector

The inputs of the Single-Ended In to Differential Out Buffers of Figure 5.1 come from an 18 GHz square wave source that produces 410 mV_{pp} at the buffer inputs through 50Ω source resistances. One input is delayed by the variable Input Phase Δ . The Single-Ended In to Differential Out Buffer circuit implementation is illustrated by Figure 6.1; the layout of the ECL SXOR circuit is illustrated by Figure 4.9; and the LPF circuit implementation is illustrated by Figure 6.2. These three circuit elements are the main components of the device shown in Figure 6.3. The output Op Amp is shown here to represent the equation $Out_{Avg+} - Out_{Avg-} = V\phi$. The Monte Carlo simulations that follow were carried out on extracted versions of these circuits at the corners defined in Table 5.1 over process and mismatch.

Table 5.1: Monte Carlo Simulation Corner Definitions

Corner	Temperature	Voltage (Nominal 2.5 V)
Fast	0°	+5% (2.625 V)
Slow	100°	-5% (2.375 V)

The plots of Figure 5.2 were generated in the following manner. 200 nominal Monte Carlo simulations were run, over process and mismatch, with *Input Phase* $\Delta = 45^\circ$, *Temperature* $= 0^\circ$ and $V_{CC} = 2.625 \text{ V}$ (i.e., +5%), then runs were selected that include 98.5% (excluding outliers) of the results between them i.e., Run 10 and Run 136. Parametric sweeps were then carried out using Runs 10 and 136 from 35° to 55° in 1.0° steps. The results are illustrated by Figure 5.2.

Figure 5.2: Monte Carlo Results for $45^\circ \pm 10^\circ$, Fast, Extracted

The plots of Figure 5.2 show that the the Output Voltages are clearly monotonic for input increments of 1.0° . They also exhibit a gentle curve or non-linearity caused by incomplete transistor switching around *Input Phase* $\Delta = 45^\circ \pm 10^\circ$. This was discussed in Section 4.3.

Table 5.2 shows the data that was plotted in Figure 5.2, as well as ΔV_ϕ , which is the difference between consecutive V_ϕ voltages over a single run or the 1.0° output voltage step size. The Range shows how much the output voltage step size varies, due to the non-linearity discussed earlier, over the Input Phase Δ range. The Average step size and associated Standard Deviation (SD) are also calculated for each run. Here the SD values indicate that the averages are a good representation of the data. The + and - Tolerances show how much the Average output voltage step sized varies over the range of 35° to 55° . That is, for Run 136 the average step size is $3.6 \text{ mV} +15.70\%/-20.11\%$ and for Run 10 the average step size is $4.4 \text{ mV} +14.81\%/-19.63\%$.

These results show that using these SXOR phase detectors to make absolute phase measurements (i.e., measurements to an absolute standard) is not possible without a calibration scheme. For example, Figure 5.2 shows that the $V\phi$ voltage for an *InputPhase* Δ of 35° with Run 136 is approximately the same as the $V\phi$ voltage for an *InputPhase* Δ of 50° with Run 10. However, if one makes the assumption that the temperature, voltage, process and mismatch variation of a single Monte Carlo Run will be consistent over the physical area of an RTWO on a single die, it is possible to make accurate relative phase measurements using these SXOR phase detectors. That is, a number of SXOR phase detectors can be used to measure the phases difference across TL sections, that should have nominally equal electrical length, around an RTWO. The tolerances of the two extreme runs shown in Table 5.2 are very similar and are less than $\pm 50\%$; therefore, relative phase tuning to within 1.0° should be possible with this implementation.

Table 5.2: Summary of Monte Carlo Results for $45^\circ \pm 10^\circ$, Fast, Extracted

Input Phase $\Delta(^{\circ})$	$V\phi$ (V) Run 136	$\Delta V\phi$ (V)	$V\phi$ (V) Run 10	$\Delta V\phi$ (V)
35	0.2351	—	0.2994	—
36	0.2322	0.0029	0.2959	0.0035
37	0.2291	0.0031	0.2922	0.0037
38	0.2260	0.0031	0.2884	0.0038
39	0.2228	0.0032	0.2845	0.0039
40	0.2195	0.0033	0.2806	0.0039
41	0.2162	0.0033	0.2765	0.0041
42	0.2127	0.0035	0.2724	0.0041
43	0.2092	0.0035	0.2682	0.0042
44	0.2057	0.0035	0.2639	0.0043
45	0.2020	0.0037	0.2595	0.0044
46	0.1983	0.0037	0.2550	0.0045
47	0.1946	0.0037	0.2505	0.0045
48	0.1908	0.0038	0.2460	0.0045
49	0.1869	0.0039	0.2413	0.0047
50	0.1830	0.0039	0.2367	0.0046
51	0.1790	0.0040	0.2319	0.0048
52	0.1749	0.0041	0.2271	0.0048
53	0.1709	0.0040	0.2222	0.0049
54	0.1667	0.0042	0.2173	0.0049
55	0.1625	0.0042	0.2123	0.0050
Range (V)		0.0013	—	0.0015
Average (V)		0.0036	—	0.0044
SD (V)		380×10^{-6}	—	430×10^{-6}
+ Tolerance (%)		15.70	—	14.81
- Tolerance (%)		20.11	—	19.63

The plots shown in Figure 5.3 were generated with the same process used to generate the results of Figure 5.2 except $Temperature = 100^\circ$ and $V_{CC} = 2.375 V$ (i.e., -5%). Runs were selected that include 97.5% (excluding outliers) of the results between them i.e., Run 52 and Run 136.

The plots of Figure 5.3 are monotonic for input increments of 1.0° and show less curve or better linearity than the plots of Figure 5.2. This is also indicated by the smaller Range values of Table 5.3. The maximum run, Run 52, of Figure 5.3 is similar to the minimum run, Run 136, of Figure 5.2, which further indicates the difficulty in using this SXOR phase detector for absolute measurements without a calibration scheme.

Figure 5.3: Monte Carlo Results for $45^\circ \pm 10^\circ$, Slow, Extracted

The SD values of Table 5.3 indicate that the averages are a good representation of the data. The average step size for Run 136 is $2.7 mV +15.03\%/-18.37\%$ and for Run 10 the average step size is $3.6 mV +13.26\%/-14.36\%$. As was stated previously, the tolerances of the two extreme runs shown in Table 5.3 are very similar and are

less than $\pm 50\%$; therefore, relative phase tuning to within 1.0° should be possible with this implementation.

Table 5.3: Summary of Monte Carlo Results for $45^\circ \pm 10^\circ$, Slow, Extracted

Parameter	Run 136 ΔV_ϕ (V)	Run 10 ΔV_ϕ (V)
Range (V)	0.0009	0.0010
Average (V)	0.0027	0.0036
SD (V)	256×10^{-6}	317×10^{-6}
+ Tolerance (%)	15.03	13.26
- Tolerance (%)	18.37	14.36

The procedure for generating Figures 5.4 and 5.5 and Table 5.4 is the same as the procedure discussed earlier in this section except that the *Input Phase* $\Delta = 90^\circ \pm 10^\circ$.

Figure 5.4: Monte Carlo Results for $90^\circ \pm 10^\circ$, Fast, Extracted

Runs 183 and 110 of Figure 5.4, and runs 186 and 173 of Figure 5.5 encompass 100% of the Monte Carlo results, i.e., no outliers. There is very little discernible curve

in these plots, which indicates that at a nominal 90° phase difference the transistors are switching completely on and completely off at 18 GHz. This improvement in linearity over the 45° cases is quantified by the relatively small Range of 0.0002 V listed in Table 5.4.

Fast Run 183 crosses $V_\phi = 0$ V at 94° and Fast Run 110 crosses $V_\phi = 0$ V at 85° , while Slow Run 186 crosses $V_\phi = 0$ V at 95° and Fast Run 110 crosses $V_\phi = 0$ V at 84° . This indicates that there is significant overlap in the Fast and Slow Monte Carlo results when compared to the 45° simulations.

Figure 5.5: Monte Carlo Results for $90^\circ \pm 10^\circ$, Slow, Extracted

The SD values of Table 5.4 indicate that the averages are a good representation of the data. The + and - Tolerances show how much the Average output voltage step sized varies over the range of 80° to 100° . These values show a significant improvement in accuracy over the 45° simulations. Unfortunately, there is significant variation in the average output voltage step size between the fast and slow corners. Therefore, as with the 45° simulations, absolute phase measurement requires a calibration scheme.

Table 5.4: Summary of Monte Carlo Results for $90^\circ \pm 10^\circ$, Extracted

Parameter	Fast $V\phi$ (V) (Fig. 5.4)		Slow $V\phi$ (V) (Fig. 5.5)	
	Run 183	Run 110	Run 186	Run 173
Range (V)	0.0002	0.0002	0.0002	0.0002
Average (V)	0.0055	0.0056	0.0037	0.0038
SD (V)	42×10^{-6}	45×10^{-6}	50×10^{-6}	64×10^{-6}
+ Tolerance (%)	1.08	0.77	2.68	2.11
- Tolerance (%)	2.00	2.43	2.66	3.47

Additional nominal Monte Carlo simulations show that the extracted phase detectors performed well in the ranges $45^\circ \pm 10^\circ$ and $90^\circ \pm 10^\circ$ up to 20 GHz. While the performance of these phase detectors with $90^\circ \pm 10^\circ$ input phase differences is superior to the performance with $45^\circ \pm 10^\circ$ input phase differences, this improvement was not considered to be significant enough to rule out using these phase detectors to measure $45^\circ \pm 10^\circ$ phase differences. Also, additional circuit components (i.e., $\lambda/8$ transmission lines and transistor switches) would be needed to implemented, characterized and calibrated to use $90^\circ \pm 10^\circ$ input phase differences in this application.

5.2 RTWO System Simulation Set-up

Figure 5.6 is a functional block diagram of the test bench used to simulate the implementation of the RTWO phase alignment circuit that was fabricated. The 250 fs , 500 fs and 750 fs Delay blocks were used to create a phase error for the system to correct. This was necessary as the RTWO 18 GHz block sections needed to be identical for the simulation to be stable. The four SXOR blocks include two extracted differential input buffers and an extracted SXOR circuit. In order to reduce the number of die pads only one of the filtered signals was used as an output. The

complementary unused outputs were terminated in $50\ \Omega$ resistors. Extracted versions of the LPF blocks were used in this simulation.

The 4-Channel, 12-Bit, Parallel A/D block and Phase Alignment Algorithm block are reused from Figure 3.10. Here the A/D reference voltage is $2.5\ V$ and the resolution is given by equation (5.1). A 1° single-ended transfer function step is $\approx 2\ mV$; therefore, the calculated resolution is adequate for this simulation.

$$2.5/(2^{12} - 1) = 610.5\ \mu V \quad (5.1)$$

The 4-Channel, 9-Bit, Parallel D/A block has been added to allow the digital control outputs of the phase alignment algorithm to adjust the analogue control inputs of the four varactors (i.e., Vtune_Top/Vtune_Left/Vtune_Bottom/Vtune_Right). The D/A reference voltage is $2.5\ V$ and its resolution is given by equation (5.2).

$$2.5/(2^9 - 1) = 4.892\ mV \quad (5.2)$$

Using equation (2.18), as well as the circuit parameter values of Figure 5.7 and [6], it was determined that an $8\ fF$ change in varactor capacitance results in $\approx 0.5^\circ$ of change in the electrical phase of a single TL section. The varactor differential capacitance versus control voltage plot of Figure 5.8 shows that, to a first order approximation, the voltage resolution calculated in equation (5.2) will never result in a phase step of more than 0.5° .

The four $5\ k\Omega$ resistors connected to the Vtune signals allow these signals to be controlled individually for phase tuning or as a single frequency tuning control signal.

The SXOR phase detector inputs at point A of Figure 5.6 are the same polarity as the other phase detector connections and are reversed from Figure 3.10. If we use the Out- signal of this SXOR, as has been done with the other three SXOR phase

detectors, the phase measurement would be the sum of the measurements made by those phase detectors. This error has been avoided by using Out+ from this SXOR phase detector.

Figure 5.6: AMSD System Model of Correctly Working RTWO Phase Calibration Circuit

The RTWO 18 GHz block transmission lines are realized with lumped element components as shown in Figure 5.7. The component values were determined from [6, 7]. The oscillator gain and buffer elements, as well as varactors were implemented

using kit components. These component designs were used in the fabricated RTWO circuit.

Figure 5.7: Lumped Element RTWO LT [6,7] with Kit Varactors [8]

The VcapDNMOSNW component of Figure 5.7 is a differential varactor from the NXP BiCMOS kit [8] with the characteristics illustrated in Figure 5.8. This plot was generated using differential s-parameters.

Figure 5.8: Differential C, Q and R for VcapDnMOSNW Kit Varactor

This VcapDnMOSNW component was selected for this application because of its

large capacitance range and the slope of the capacitance versus control voltage curve is better suited to analogue voltage control (i.e., smaller change in C for a change in V) than others available in the kit. Also, its symmetry about the half capacitance point marked on the figure will give consistent tuning over its range. The size of this varactor (i.e., $W/L = 5 \mu m \times 0.5 \mu m$ with eight stripes) was selected to optimize Q and capacitance for this application.

5.3 RTWO System Simulation - Typical Results

The filtered phase detector results of the extracted AMS Designer system model of Figure 5.6 are illustrated in Figure 5.9. Delay elements of 250 fs, 500 fs, 0 fs and 750 fs were used to create a phase error for the phase alignment algorithm to correct. The system functioned correctly and converged up to the point where the varactor control voltage range was exceeded (i.e., 490 ns).

Figure 5.9: Simulated Left, Bottom, Right and Top Signals of Figure 5.6

Figure 5.10 shows the nominal varactor control voltages at each iteration of the phase alignment algorithm. At 490 ns the algorithm tries to reduce the Left varactor Control signal voltage ($V_{\text{tune_Left}}$) below zero, which is outside its range of 0 V to 2.5 V. The digital word that determines the $V_{\text{tune_Left}}$ voltage then flips over to a near maximum voltage. This problem can be resolved by adding additional differential varactor circuits to each TL section to increase the tuning range and by adjusting the RTWO TL lengths to compensate for the resulting operating frequency reduction due to the increase in minimum total varactor capacitance.

Figure 5.10: Simulated Varactor Control Voltages of Figure 5.6

5.4 RTWO System Simulation - Updated from Fabrication

Figure 5.11 is an AMS Designer model of the circuit that was fabricated to produce the die of Figure 6.11. This circuit differs from Figure 5.6 as follows:

1. The system of Figure 5.6 uses a lumped element TL RTWO implementation, Figure 5.7. Here, Figure 5.11, the transmission lines of the TL sections have been implemented in an eight port s-parameter module (i.e., N-Port Network). This module points to the Touchstone[®] file RTWK_ideal_mlin.s8p, which was generated using ADS Momentum[®] and the kit substrate to simulate the layout of the RTWO transmission lines. This eight port s-parameter 2.5D simulation

was run from 0 Hz to 140 GHz to include at approximately eight harmonics. The s-parameter was used because it improved the stability of this simulation.

2. The portion of the AMS Designer model in the dashed box shown in Figure 5.11, labelled Fabricated Die and excluding the N-Port Network, is a single extracted block. In the model of Figure 5.6 each illustrated circuit block was extracted separately. It should be noted that previous attempts to simulate using this single extracted block failed. This was because the layer one ground plan under the RTWO and oscillator signal distribution transmission lines failed the layout design rules when it was continuous. A significant number of small holes or cut-outs were inserted in the ground plane to satisfy these design rules. Unfortunately, these holes made the ground plane equivalent circuit extraction very complicated, which in turn prevented practical simulation. A second extracted circuit was created with a continuous ground plan and has been used here.
3. The final difference is in the filtered phase detector output signals at points A, B and C of Figure 5.11. It was explained earlier that for single-ended operation if the negative output signals were used from the first three differential SXOR circuits, the positive output would have to be used from the fourth SXOR to ensure that all four phases are measured. The filtered Out+ signal of the first SXOR and filtered Out- of the fourth SXOR were what was actually fabricated.

Figure 5.11: AMSD System Model of Implemented RTWO Phase Calibration Circuit

The simulated filtered phase detector results of Figure 5.11 are illustrated in Figure 5.12. Here no artificial delay elements were needed as the circuit layout variation provided finite phase errors. We can see that the system did not converge as it has in previous examples. This is because the phase detector output, point B of Figure 5.11, is measuring a nominal 135° instead of 45°, refer to the XOR transfer function of Figure 4.2. Additionally, the corresponding varactor control will need to correct in

the opposite direction from the other varactor controls.

All four phase detector outputs show various amounts of RF feed-through - the most significant on the signal marked Bottom. This feed-through will be reduced by using the phase detector outputs differentially and could be mitigated further by improvements in the layout to reduce RF coupling.

Figure 5.12: Simulated Left, Bottom, Right and Top Signals of Figure 5.11

The varactor control voltage ($V_{\text{tune_Top}}$, $V_{\text{tune_Left}}$, $V_{\text{tune_Bottom}}$ and $V_{\text{tune_Right}}$) simulation results corresponding to Figure 5.12 are shown in Figure 5.13. Here $V_{\text{tune_Left}}$ needs to correct in the opposite direction.

Figure 5.13: Simulated Varactor Control Voltages of Figure 5.11

The simulation results of Figure 5.14 are the phase detector output voltages using the AMS Designer model of Figure 5.6 with the phase detector output signals configured as in Figure 5.11. The objective here was to see if similar results could be obtained using a circuit that had previously given good results. Here we can see that the Left phase detector voltage is diverging as in the previous simulation. In fact, the markers show that all these signal are increasing in voltage with time. These signals are spread further apart as a result of the artificial delay elements.

Figure 5.14: Simulated Left, Bottom, Right and Top Signals of Modified Figure 5.6

The varactor control voltage ($V_{\text{tune_Top}}$, $V_{\text{tune_Left}}$, $V_{\text{tune_Bottom}}$ and $V_{\text{tune_Right}}$) simulation results corresponding to Figure 5.14 are shown in Figure 5.15. Here $V_{\text{tune_Left}}$ needs to correct in the opposite direction. Also, we see the result of running out of varactor capacitance at around 250 ns. In a practical application multiple circuit elements would be required to extend the tuning range.

Figure 5.15: Simulated Varactor Control Voltages of Modified Figure 5.6

Chapter 6

Experimental Results

6.1 SXOR Phase Detector Test Set-up and Procedure

Characterization of the SXOR phase detector is required to determine its accuracy and thus, its suitability for measuring the phase error of an RTWO. In order to provide differential stimulus signals to the SXOR from a single-ended lab signal generator, single-ended to differential buffers were implemented. This buffer circuit is shown in Figure 6.1. It consists of a minimum size transistor differential pair with emitter-follower outputs. The single-ended input to Q3 is dc blocked by a 3 pF mimCap so the input can be biased optimally and independently of the external signal generator (i.e., no external bias tee requirement). This configuration is duplicated for Q4 to balance the conversion from single-ended input to differential output. The emitter-follower outputs are also capacitor isolated to allow these differential signals to be biased optimally for the phase detector inputs. The output mimCaps are 10 pF to support wideband testing (i.e., 2.0 GHz to 20.0 GHz) of the phase detector. Transistors Q1 and Q2 are similar to the current mirror transistors of Figure 4.4.

Figure 6.1: Single-Ended Input to Differential Output Buffer

Figure 6.2 is the two-stage LPF used to average the output duty cycle signals of the phase detector. A phase detector output signal will charge the 2 pF capacitor through the $1.1\text{ k}\Omega$ resistor to produce an average dc level proportional to the duty cycle. The second $1.1\text{ k}\Omega$ resistor and 200 fF capacitor filter the feed-through signal that is present at twice the input operating frequency. This feed-through is caused by the frequency doubling effect of the XOR phase detector function.

Figure 6.2: Low Pass Filter (LPF)

The circuits of Figures 4.9, 6.1 and 6.2 were combined to produce the test chip illustrated in Figure 6.3. The SXOR phase detector is located in the centre of the Phdetect_2 test device layout. The outputs of the SXOR feed two LPF devices, which in turn feed the pads $V\phi o+/-$. Around the SXOR phase detector is a V_{CC} ring that supplies the phase detector and two input buffers. This ring is connected to the

V_{CC} pad and is decoupled by a bank of mimCaps totalling 22 pF [39]. Three other 22 pF mimCaps are used to decouple Ibias0 for the phase detector, as well as Ibias1 and Ibias2, which are associated with the input single ended to differential output buffers. The high frequency inputs of these buffers are connected to the gnd/sig/gnd (GSG) pads for probing. The Sense_ V_{CC} and Sense_Ibias0 pads are connected to V_{CC} and Ibias0 so the voltage at these points can be more directly measured with a high impedance volt meter. This allows the adjustment of these supplies to compensate for I^2R losses in the electrical supply path. The buffers and phase detector are enclosed within a Deep Trench Isolation (DTI) rectangle to isolate them from the supply decoupling [40].

Figure 6.3: Differential Phase Detector (Phdetect_2) Test Device Layout

A phase detector test card was designed and fabricated to provide a test platform for the Phdetect_1 and Phdetect_2 Devices Under Test (DUT), see Appendix C.1 for PCB schematic, layout and BOM. A single DUT was mounted on a phase detector test card ground pad using conductive epoxy so the the high speed signals (i.e., gnd/sig/gnd) could be probed and the remaining low frequency signals could be wire

bonded, see Figure 6.3. This PCB also included four adjustable Low Drop-Out (LDO) linear voltage regulators and filter circuits to supply $I_{bias0/1/2}$ and V_{CC} so they could be controlled independently. These regulators together with LPF components provide attenuation of supply noise from less than 100 Hz to 1.0 GHz. Attenuation of supply noise above 1.0 GHz is provided by the inductance of the bond wires (i.e., typically 1.0 nH/mm ; $> 3.0 \text{ mm}$ or $\approx 3.0 \text{ nH}$) that connect these chip pads to pads on the PCB. The LDO regulators are supplied by a $+4.5 \text{ Vdc}$ battery; therefore, any relevant noise on this supply line will originate from the SXOR phase detector switching. The outputs of Phdetect_1 and Phdetect_2, $V_{\phi o+/-}$, are filtered to generate outputs VC-_OUT and VC+_OUT. The voltage of these outputs is measured using two Agilent® 34440A 6 1/2 digit Digital Volt Meters (DVM). Figure 6.4 is a functional diagram of the test set-up used to control the phase inputs and measure the corresponding voltage outputs over phase and frequency.

Figure 6.4: Differential Phase Detector Lab Test Set-up

In Figure 6.4 an Agilent E4432B ESG is used to generate a CW tone (RF Out), which is split by an Agilent 11667A Resistive Power Splitter ($IL = 7 \text{ dB}$; $\pm 3^\circ$). One branch of this split signal is connected to a Narda[®] 4203-10 10 dB Directional Coupler through a Narda 3753B Phase Shifter (3.5 GHz to 12.4 GHz $\pm 0.5^\circ$) and the other branch is connected directly to another Narda 4203-10 10 dB Directional Coupler. The outputs of these Directional Couplers is connected to 40-GHz GSG probes through SMA male-to-male adaptors. The GSG probes supply CW tones to each of

the inputs of the DUT. This provides controllable phase Δ input signals at frequencies up to ≈ 12.0 GHz. The 10-dB down Directional Coupler outputs are connected to two channels of a 16-GHz Agilent Infiniium DSA-X 91604A Digital Signal Analyzer (i.e., high speed oscilloscope) to measure the phase difference between the resulting CW tones fed to the DUT. These connections are made using cables that are phase matched over temperature and torsion from 100 MHz to 18.0 GHz. The stability of these matched cables greatly improves our ability to measure phase increments of 1° at 12 GHz (i.e., 231.5 fs) for a long enough duration for the measurement SD to fall below ≈ 80 fs. An HP[®] 54750A 50-GHz Digitizing Oscilloscope was also used to verify these test results.

Eight phase detector test cards were assembled - boards one to four included the version one phase detector die (Phdetect_1, Figure 4.10), boards five to eight included the version two phase detector die (Phdetect_2, Figure 4.11). Table 6.1 lists the voltage setting and the resulting current consumption and power dissipation for each board. Unfortunately, the die probe pads on boards five and eight became contaminated at some point during assembly and could not be probed; therefore, results for these boards were not included in the table below.

Table 6.1: SXOR Test Card Voltage, Current and Power Dissipation

Parameter	Card 1	Card 2	Card 3	Card 4	Card 6	Card 7
V_{cc} (V)	2.5697	2.5686	2.5705	2.5680	2.5613	2.5683
V_{bias0} (V)	2.5014	2.5038	2.5015	2.5085	2.5084	2.5060
V_{bias1} (V)	2.5018	2.5027	2.5034	2.5094	2.5026	2.5051
V_{bias2} (V)	2.5020	2.5028	2.5035	2.5095	2.5026	2.5053
Sense_V_{cc} (V)	—	—	—	—	2.5019	2.5020
Sense_I_{bias0} (V)	—	—	—	—	0.7864	0.7807
I_{cc} (mA)	120.6	120.4	124.1	120.6	125.6	126.8
I_{bias0} (mA)	17.38	17.23	17.79	17.41	17.89	18.13
I_{bias1} (mA)	7.128	7.010	7.257	7.089	7.241	7.336
I_{bias2} (mA)	7.083	6.990	7.260	7.048	7.267	7.323
Power (mW)	388.9	387.4	399.8	388.9	402.9	407.8

The V_{cc} voltage of boards six and seven was adjusted so that the Sense_V_{cc} voltage was approximately 2.50 V. The V_{cc} voltage of boards one through four was then set to approximately the same value as the die on these boards did not have Sense_V_{cc} signals. Sense_I_{bias0} voltage levels were verified to be close to the simulation values. The current consumption of Phdetect_2 die is marginally larger than the current consumption of Phdetect_1 die, which may be due to lower path loss in the more compact Phdetect_2 layout.

6.2 SXOR Phase Detector Test Set-up Calibration

In order to take accurate measurements using the test set-up of Figure 6.4 we must first determine the propagation delays associated with the test set-up equipment and cables so they can be de-embedded from the measurement stimulus (i.e., the input phase Δ). This calibration was carried out using both direct and indirect methods

described in the following two subsections.

6.2.1 Direct Test Set-up Calibration

The test set-up of Figure 6.4 can be used to directly determine the correction that needs to be applied to the input phase difference stimulus. When the SXOR phase detector input phase difference is 90° the phase detector output voltages VC-_OUT and VC+_OUT will be equal if operating correctly. Therefore, the differential output voltage should be $VC+_OUT - VC-_OUT = 0$ V. The test set-up calibration procedure is as follows:

1. Power-up the test equipment for at least 45 minutes before it will be used.
2. Set the E4433B ESG to generate a 12.0 GHz CW tone. This frequency is limited by the tuning resolution of the Narda 3753B phase adjuster.
3. Set the E4433B ESG CW tone output level to 11.4 dBm. This will produce signal levels of ≈ 600 mV when measured at the output port of either Narda 4203-10 directional coupler using the DSA-X 91604A oscilloscope ($50\ \Omega$ termination). These output ports are connected to the probes through SMA adapters in Figure 6.4. This level also ensures that a large enough CW tone can be measured at the -10 dB coupled ports.
4. Re-connect the test set-up as in Figure 6.4 and power up the DUT.
5. Connect the 34440A DVMs to VC-_OUT and VC+_OUT and adjust the 3753B phase shifter until these outputs are equal.
6. Measure the input phase difference at the zero-crossing points using the DSA-X 91604A oscilloscope. This phase correction factor is the difference between

the measured input phase difference and the 90° phase difference that is being applied to the DUT.

This correction factor was used to adjust the input phase difference for all the phase detector measurements included in this section.

This calibration method is simple and direct, and was carried out at the beginning of every test run to minimize drift error during the experiment period. However, this calibration method assumes that the DUT is functioning without any phase offset problems. The indirect test set-up calibration described in the next subsection will mitigate this issue.

6.2.2 Indirect Test Set-up Calibration

The set-up illustrated in Figure 6.5 shows how this test set-up calibration procedure was carried out.

Figure 6.5: Test Cable Delay De-Embedding Set-up

The first three steps in the Direct Test Set-up Calibration procedure were carried out. At this point the phase difference between the signals applied to the DUT was measured and recorded with the DSA-X 91604A oscilloscope for a nominal phase at a frequency of 12 GHz. That is, at the directional coupler outputs with the 10-*dB* coupled outputs terminated in 50 Ω loads. Next, the directional coupler outputs were terminated in 50 Ω loads and the phase difference between the 10-*dB* coupled outputs

through matched cables was measured with the DSA-X 91604A oscilloscope at 12 GHz for the same phase setting. The first measurement value was then subtracted from the second measurement value to determine the phase delay in degrees that was used to adjust the input phase stimulus. This procedure moved the measurement plane to the input of the probes.

Figure 6.6: Probe Delay De-Embedding Set-up

The delay through the probes was determined using an Agilent 8720ES Vector Network Analyzer (VNA). First, single port VNA calibrations (i.e., s_{11} and s_{22}) at $12 \text{ GHz} \pm 100 \text{ MHz}$ were carried out to the ends of the VNA cables that will connect to the probes (i.e., Calibration Plane of Figure 6.6). Next, the VNA cables were connected to the GSG probes and the probes were landed on the open pads of a Calibration Substrate. The s_{11} and s_{22} delay was measured and recorded.

This measurement is the round-trip delay from the calibration plane to the probe tips, across the open offset delay (i.e., 0.327 ps) and back from the probe tip to the calibration plane. The delay from calibration plane to probe tip for GSG probe V92, with measured delay = 182.1 ps, is give by equation (6.1).

$$(182.1/2) - 0.327 = 90.723 \text{ ps} \quad (6.1)$$

The delay from calibration plane to probe tip for GSG probe W31, with measured delay = 178.8 ps, is give by equation (6.2).

$$(178.8/2) - 0.327 = 89.073 \text{ ps} \quad (6.2)$$

The difference between these two delays, $90.723 - 89.073 = 1.65 \text{ ps}$, was used to adjust the input stimulus phase delay value.

The input stimulus phase delay values determine using the Indirect Test Set-up calibration procedure were found to be consistent with the input stimulus phase delay values found using the Direct Test Set-up Calibration procedure. This showed that there were no significant input phase offsets between the two differential phase detector inputs. Further, the Direct Test Set-up Calibration procedure was used to generate the test results presented in this document.

6.3 SXOR Phase Detector Test Results

Testing was focused on boards five to eight because significant improvements had been made to the Phdetect_2 die (i.e., improved layout, phase detector sense line on VCC and Ibias0 and additional ground pads). Board four, including the Phdetect_1, was also tested to verify the operation of this device.

The ECL outputs at Out-/ + of Figure 6.7 will have an offset error that is approximately equal to the error in V_{CC} from its nominal voltage. This error can be removed by subtracting Out_Avg- from Out_Avg+ to get a differential voltage, $V\phi$ Measured.

Figure 6.7: Op Amp Removal of V_{CC} Error

The results plotted in Figures 6.8 and 6.9 and listed in Tables 6.2 and 6.3 were obtained by measuring the steady state voltage at $Out_Avg-/+$ w.r.t. ground and subtracting the results. Figure 6.7 shows how this function could be implemented with a CMOS Op Amp. R_1 can be used to remove any significant dc offset that exists in the Op Amp.

Figure 6.8: Phase Detector Transfer Function - 12.0 GHz, Measured

Figure 6.8 shows the transfer function measurements of Boards 4, 6 and 7 for input phase steps of 5° . These results compare well with the ideal transfer function shown in Figure 4.2 except for the Output Phase extremes. The curves are monotonic for input phase steps of 5° .

Also, plotted are Monte Carlo simulation results that were carried out on the extracted version of the tested circuit of Figure 6.3. These Monte Carlo results

(i.e., over process and mismatch) were worst case of 200 runs at 50°C . The 50°C temperature value was estimated by matching nominal simulation results at elevated temperatures with the measured results. Here we see that for input phase differences around 45° and 90° the test results fall between the Monte Carlo simulations.

Figure 6.9: Phase Detector Transfer $45^\circ \pm 10^\circ$ and $90^\circ \pm 10^\circ$ - 12.0 GHz, Measured

Figure 6.9 shows the transfer function measurements of Boards 4, 6 and 7 for input phase steps of 1° over the ranges of $45^\circ \pm 10^\circ$ and $90^\circ \pm 10^\circ$. The curves are

monotonic for input phase steps of 1° . The variation in monotonicity shown in this plot is mostly due to the phase tuner. The Narda 3753B is a mechanical tuner that is adjusted by hand. It has problems with repeatability and resolution at 12 GHz.

The test results for the version two phase detectors (i.e., Phdetect_2) on Boards 6 and 7 fall within the Monte Carlo simulations. The test results for the version one phase detector, Board 4, around 45° fall below the lower Monte Carlo simulation result. This excursion is not surprising as the Monte Carlo simulations were carried out on the extracted version two phase detector.

Table 6.2 lists the data plotted in Figure 6.9 for the range of $45^\circ \pm 10^\circ$. We can see from the Input Phase Δ values the difficulty in adjusting the mechanical phase tuner. The important results of this table are the slope values 4.2 mV for Board 7, 3.8 mV for Board 6 and 3.8 mV for Board 4. These values indicate the average differential output voltage for an input phase step of 1° .

Table 6.3 lists the data plotted in Figure 6.9 for the range of $90^\circ \pm 10^\circ$. Here the slope values are 5.0 mV for Board 7, 4.9 mV for Board 6 and 4.4 mV for Board 4. These values indicate the average differential output voltage for an input phase step of 1° . It should be noted that there is approximately 1 mV difference in voltage between these two ranges. This is likely due to the improved linearity of the phase measurements taken around 90° versus those taken around 45° .

Table 6.2: Summary of Measured Results for $45^\circ \pm 10^\circ$ - 12.0 GHz, Measured

Board 7 ($45^\circ \pm 10^\circ$)		Board 6 ($45^\circ \pm 10^\circ$)		Board 4 ($45^\circ \pm 10^\circ$)	
Input Phase $\Delta(^\circ)$	$V\phi$ (V) Measured	Input Phase $\Delta(^\circ)$	$V\phi$ (V) Measured	Input Phase $\Delta(^\circ)$	$V\phi$ (V) Measured
35.28	0.2551	34.93	0.2594	35.33	0.2376
35.80	0.2539	35.93	0.2555	36.05	0.2350
36.98	0.2490	36.96	0.2524	37.23	0.2313
38.33	0.2443	37.91	0.2481	37.98	0.2286
38.68	0.2417	38.75	0.2456	39.15	0.2247
40.35	0.2371	39.73	0.2419	39.99	0.2218
41.46	0.2321	41.06	0.2341	41.25	0.2172
41.71	0.2301	41.79	0.2316	41.90	0.2145
43.28	0.2230	43.14	0.2258	42.65	0.2125
44.08	0.2199	44.04	0.2231	43.80	0.2081
45.02	0.2160	44.85	0.2207	45.29	0.2027
46.00	0.2126	46.02	0.2192	46.05	0.1991
46.72	0.2087	46.75	0.2165	47.25	0.1939
48.14	0.2035	48.43	0.2113	47.86	0.1914
48.97	0.1997	49.22	0.2055	49.02	0.1870
49.92	0.1947	49.85	0.2019	50.20	0.1823
50.97	0.1898	50.97	0.1976	51.10	0.1784
51.72	0.1869	52.25	0.1945	51.55	0.1770
52.85	0.1818	52.96	0.1893	53.32	0.1702
54.36	0.1750	53.86	0.1850	53.78	0.1679
55.07	0.1717	55.34	0.1812	55.62	0.1600
Range (V)	0.0834	—	0.0782	—	0.0776
Slope (V/ $^\circ$)	0.0042	—	0.0038	—	0.0038

Table 6.3: Summary of Measured Results for $90^\circ \pm 10^\circ$ - 12.0 GHz, Measured

Board 7 ($90^\circ \pm 10^\circ$)		Board 6 ($90^\circ \pm 10^\circ$)		Board 4 ($90^\circ \pm 10^\circ$)	
Input Phase $\Delta(^\circ)$	$V\phi$ (V) Measured	Input Phase $\Delta(^\circ)$	$V\phi$ (V) Measured	Input Phase $\Delta(^\circ)$	$V\phi$ (V) Measured
79.82	0.0489	79.80	0.0491	80.01	0.0464
80.81	0.0438	80.79	0.0438	81.27	0.0417
82.32	0.0353	82.06	0.0381	81.96	0.0386
83.03	0.0304	82.71	0.0351	83.09	0.0350
84.23	0.0254	84.04	0.0291	84.24	0.0301
85.12	0.0213	84.47	0.0260	84.84	0.0267
85.80	0.0185	85.52	0.0223	86.23	0.0213
87.29	0.0105	86.58	0.0166	87.24	0.0170
88.06	0.0081	87.78	0.0113	87.94	0.0126
89.05	0.0024	88.75	0.0065	89.15	0.0098
90.00	0	90.00	0.0003	90.11	0.0042
91.13	-0.0076	91.21	-0.0063	91.29	-0.0005
92.37	-0.0136	92.15	-0.0115	92.26	-0.0055
93.06	-0.0175	92.88	-0.0149	92.78	-0.0075
94.27	-0.0230	94.12	-0.0213	94.49	-0.0147
95.10	-0.0279	95.14	-0.0257	95.29	-0.0192
96.04	-0.0314	95.97	-0.0304	95.67	-0.0212
96.82	-0.0364	97.17	-0.0367	97.05	-0.0258
97.96	-0.0420	98.11	-0.0402	98.36	-0.0340
98.76	-0.0455	99.01	-0.0454	99.47	-0.0389
100.05	-0.0523	99.96	-0.0500	100.45	-0.0426
Range (V)	0.1012	—	0.0991	—	0.0890
Slope (V/ $^\circ$)	0.0050	—	0.0049	—	0.0044

6.4 Discussion of Results - SXOR Phase Detector

Simple Symmetrical XOR (SXOR) phase detectors were designed, fabricated in SiGe BiCMOS and tested to resolve 1° of phase error over $45^\circ \pm 10^\circ$ and $90^\circ \pm 10^\circ$ up to 12 GHz. This 12 GHz limit is the maximum operating frequency of the test set-up (i.e., limited by the Narda 3753B Phase Shifter), not the maximum operating frequency of the DUT. The following was also demonstrated:

1. The measured SXOR transfer functions of multiple devices show good agreement with the ideal transfer function.
2. The measured SXOR transfer functions of multiple devices were monotonic for input phase steps of 1° over the ranges of $45^\circ \pm 10^\circ$ and $90^\circ \pm 10^\circ$.
3. The measured SXOR phase detector average output voltage change for 1° input phase steps over the ranges of $45^\circ \pm 10^\circ$ and $90^\circ \pm 10^\circ$ was $\approx 4\text{ mV}$ and $\approx 5\text{ mV}$. These output voltages were consistent over these ranges.
4. The measured SXOR transfer functions of multiple devices show good agreement with simulation at 50°C . That is, these results are between the worst case extracted simulation Monte Carlo curves.
5. The differential transit delay through an XOR circuit can be equalized by using a symmetrical XOR or SXOR architecture.
6. The simple ECL outputs of these phase detectors are very sensitive to V_{cc} variation in single-ended operation. This problem can be mitigated by using the SXOR outputs differentially.
7. The SXOR phase detector and buffer power dissipation values listed in Table 6.1 are very high. This is because the transistors of these components were biased

for a current density close to f_T , which is equivalent to an $I_c = 14$ to 18 mA. The objective here was to demonstrate RTWO phase tuning and the optimization of power dissipation was considered to be beyond the scope of this work.

The following errors were addressed during the testing of the fabricated SXOR phase detector.

Systematic Error

1. Two test measurement calibration methods (Direct and Indirect) were developed and used to remove phase error in the test set-up. These methods were carried out immediately before every test run and the results were used to calibrate the input phase stimulus.
2. The phase delay difference between the two GSG probes was determined and removed from the input phase stimulus.
3. Cables matched over torsion and temperature up to 18 GHz were used to measure phase difference. These cables have a loss of ≈ 1.4 dB at 18 GHz and significantly improved measurement stability and repeatability.
4. Care had to be taken to ensure the GSG probes were landed solidly on the die pads. That is, so that normal vibration in the test area did not move the probes on the pads. Probe to pad movement caused the phase detector output voltage to jump significantly.

Random Error

1. The manual phase adjustment mechanism of the Narda phase adjuster was not smooth or continuous. Generally, the phase would change initially in the

opposite direction of adjustment and then jump to a new phase position in the intended direction of adjustment. A significant amount of fine tuning was required for each input phase Δ . Unfortunately, no other phase adjusters were available and the purchase of an electronic phase adjuster with adequate range and resolution was prohibitively expensive.

2. The high speed oscilloscope used to measure phase was configured to take a moving average of 256 measurements. After the input phase adjustment was made for each measurement point, the moving average was reset and a phase value was recorded after the SD was stable and less than 80 fs.

Drift Error

1. Test equipment was turned on and allowed to warm up to an operating temperature for at least 45 minutes before each test run.
2. All test runs were carried out without pauses. The objective here was to minimize the effects of room temperature variation and other test equipment drift mechanisms over time.
3. At the beginning of each test the system was calibrated to 90° . Measurements were then taken from this point in the direction of decreasing input phase delta (i.e., 89° , 88° ...). After measurements were taken in this direction, measurements were taken in the direction of increasing input phase delta (i.e., 89° , 88° ...). In the time between these two measurement sequences drift error occurred that can be seen in the test result graphs, Figures 6.8 and 6.9.

6.5 RTWO Phase Alignment Test Set-up and Procedure

Figure 6.10 shows the test set-up used to evaluate the RTWO phase alignment test die. An RTWO tuning test card was designed and fabricated to provide a platform for this DUT, see Appendix C.2 for PCB schematic, layout and BOM. A single DUT was mounted on an RTWO phase alignment test card ground pad using conductive epoxy so the high speed RTWO output signals (i.e., GND/OSC-/OSC+/GND) could be probed, and their amplitude and frequency measured using an HP 8593E spectrum analyzer. This test card supports nine adjustable LDO power supplies, which receive power from an Agilent E3546A DC supply through nine separate power supply filters. These regulators together with LPF components provide attenuation of supply noise from frequencies of less than 100 Hz to 1.0 GHz. Attenuation of supply noise above 1.0 GHz is provided by the inductance of the bond wires (i.e., typically 1.0 nH/mm ; $> 3.0 \text{ mm}$ or $\approx 3.0 \text{ nH}$) that connect these chip pads to pads on the PCB. The RTWO gain stage amplifiers were supplied directly from an Agilent E3646A power supply through VccOSC as this voltage was adjusted beyond the range of the LDO regulators.

The voltage levels of the filtered single ended SXOR phase detector output signals, VOELLB/VOETL/VOERT/VOEBR, were measured using three Agilent 34401A and one Keithley[®] 2010 6 1/2 digit DVM. These signals were also sampled and digitized using the Maxim[®] MAX126 4-channel, 14-bit ADC of a MAX126 evaluation board. This board provided a 16.0 MHz oscillator, LDO regulators and 2.5 V voltage reference required by the MAX126. This sets the analogue input voltage range to be -2.5 V to $+2.5 \text{ V}$. The input voltage range required in this application is 0 to $+2.5 \text{ V}$; therefore, the most significant bit of the two's complement digital outputs will not be

used, making the ADC effectively 13-bits.

$$2.5/(2^{13} - 1) = 305.2 \mu V \quad (6.3)$$

Equation (6.3) shows the resolution of the ADC to be $305.2 \mu V$. The $\pm 12.0 Vdc$ required for this card were supplied by an Agilent E3646A DC supply.

The digitized phase detector outputs were converted from $5.0 V$ logic to $3.3 V$ logic and fed to the hardware implementation of the phase calibration algorithm, which was realized in a Diligent Atlys (Xilinx®) Spartan-6 FPGA board. The algorithm timer input zero (T0i, Figure 3.10) was connected to a Single-Pole Single-Throw (SPST) switch to allow the algorithm to be reset manually. The algorithm timer input one (T1i, Figure 3.10), which initiates an iteration of the phase alignment algorithm on every low-to-high state change, was connected to a SPST momentary switch to allow single-stepping through RTWO phase correction process. Both these switches were debounced within the algorithm, see Appendix B.2.

After the state of T0i is changed from reset, the phase calibration algorithm will initialize the varactor voltages to be mid range (i.e., $\approx 1.25 Vdc$) and the Maxim MAX126 ADC is initialized to carry out a four channel conversion sequence at every iteration. After every T1i low-to-high state change the algorithm will carry out four consecutive reads of the MAX126 and uses these digital voltage levels to generate new varactor voltages.

The results of each algorithm iteration are digital representations of newly generated varactor control voltage levels, which are presented to an Analog Devices® AD7305 4-channel, 8-bit DAC mounted on the RTWO tuning test card. This communication occurs over a 12-signal parallel A/D/C (Address/Data/Control) bus that

cycles through the four digital number to analogue voltage conversions that are generated by every algorithm iteration. The DAC output voltage range is 0 to +2.5 V, which is determined by the +2.5 V reference voltage generated on the Maxim MAX126 ADC card. Equation (6.4) shows the resolution of the ADC to be 9.804 mV.

$$2.5/(2^8 - 1) = 9.804 \text{ mV} \quad (6.4)$$

The varactor control voltages, VCL/VCT/VCR/VCB, were measured using four Wavetech[®] 3 1/2 digit DVMs.

Figure 6.10: Tuneable RTWO Lab Test Set-up

Figure 6.11 is a die photograph of the RTWO (centre) with differential buffers and SXOR phase detectors (at the corners) that is the DUT of Figure 6.10. This RTWO was created by Kimia Ansari who also assembled this layout and submitted it for fabrication. All signals except for the GND/OSC-/OSC+/GND were wire bonded to the RTWO Tuning Test Card (Figures C.5 and C.6). Signals GND/OSC-/OSC+/GND were probed as shown in Figure 6.10.

Figure 6.11: RTWO Phase Alignment Test Die (DUT Figure 6.10)

VCL/VCB/VCR/VCT of Figure 6.11 are the varactor control signals for the Left, Bottom, Right and Top sections of the RTWO. VccOSC supplies power to the RTWO amplifiers and VccBO supplies power to the buffers that tap the oscillator phases directly from the RTWO sections. VOETL/VOELB/VOEBR/VOERT are single ended filtered phase detector output signals from the Top-Left, Left-Bottom, Bottom-Right and Right-Top corners of the RTWO. VccBTL/VccBLB/VccBBR/VccBRT

supply power to the differential buffers that drive the inputs of the differential SXOR phase detectors with oscillator phases from the tap buffers. $V_{ccPDTL}/V_{ccPDLB}/V_{ccPDBR}/V_{ccPDRT}$ supply the differential SXOR devices at each corner.

Table 6.4 lists the supply voltages, currents and total power dissipation for the two RTWO Tuning Test Cards used to evaluate the RTWO phase alignment test die. Here $V_{ccBO} \approx 2.5 V_{dc}$ and $V_{ccOSC} \approx 4.5 V_{dc}$ were set based on testing done by Kimia Ansari. The differential buffer supply voltages were set to $\approx 2.56 V$ based on the phase detector testing results shown in the V_{cc} row of Table 6.1. The SXOR phase detector supply voltages were adjusted so the filtered outputs would have similar initial values.

Table 6.4: RTWO Tuning Test Card Voltage, Current and Power Dissipation

Parameter	Card 1	Card 2	Parameter	Card 1	Card 2
V_{ccBO} (V)	2.489	2.505	I_{ccBO} (mA)	202.4	200.3
V_{ccOSC} (V)	4.500	4.500	I_{ccOSC} (mA)	21.0	21.0
V_{ccBTL} (V)	2.565	2.585	I_{ccBTL} (mA)	84.1	85.1
V_{ccPDTL} (V)	2.437	2.423	I_{ccPDTL} (mA)	53.6	53.2
V_{ccBLB} (V)	2.561	2.571	I_{ccBLB} (mA)	82.1	81.8
V_{ccPDLB} (V)	2.682	2.661	I_{ccPDLB} (mA)	59.4	58.7
V_{ccBBR} (V)	2.559	2.569	I_{ccBBR} (mA)	83.2	83.3
V_{ccPDBR} (V)	2.650	2.565	I_{ccPDBR} (mA)	58.9	56.1
V_{ccPDRT} (V)	2.888	2.640	I_{ccPDRT} (mA)	115.2	100.0
V_{ccBRT} (V)	2.683	2.571	I_{ccBRT} (mA)	46.0	43.0
Power (W)				2.139	2.044

6.6 RTWO Phase Alignment Test Results

The test set-up of Figure 6.10 with the DUT of Figure 6.11 were used to generate the results of Figures 6.12 to 6.16.

Figure 6.12: Test Board 1 Phase Detector Output Signals

Figure 6.12 shows the single-ended SXOR phase detector output voltages for test Board 1. The 16 iterations of the phase alignment algorithm adjust the varactor control voltage between 0 V and 2.5 V. On iteration 17 varactor control voltage VCB overflows (Figure 6.13) and the algorithm is initialized on iteration 18. As was described in Section 5.4, we can see that the system did not converge. This is because the phase detector output, point B of Figure 5.11, is measuring a nominal 135° instead of 45° , refer to the XOR transfer function of Figure 4.2. Additionally,

the corresponding varactor control will need to correct in the opposite direction from the other varactor controls.

Figure 6.13: Test Board 1 Phase Detector Output Signals

Figure 6.14 shows the single-ended SXOR phase detector output voltages for test Board 2. These signals clearly diverge in a manner that is different than Board 1. This may be due to manufacturing variation or direction of RTWO rotation.

Figure 6.14: Test Board 2 Phase Detector Output Signals

Figure 6.15 shows the varactor control voltages that are attempting to correct the diverging phases of Figure 6.14. Here we see that the VCT and VCB control voltages run out of varactor range in 16 iterations and overflow on iteration 17. The system is initialized at iteration 18.

Figure 6.15: Test Board 2 Phase Detector Output Signals

During the testing of Boards 1 and 2 the RTWO frequency of oscillation was monitored. The frequency variation over the course of tuning is ≈ 150 MHz. This was expected and is similar to simulation results. This is clearly less than the frequency tuning range of ≈ 1.1 GHz.

The phase alignment algorithm will minimize the change in RTWO average frequency during phase tuning. That is, at each algorithm iteration a tuning voltage increase is made to one TL section and the tuning voltage is decreased by this same amount in another TL section. The objective here is to add and remove equal amounts of capacitance to maintain a constant average frequency of oscillation. However, the slope of the varactor capacitance curve changes with control voltage (see Figure 5.8); therefore, the total capacitance of the RTWO TL sections will change with phase

tuning and the oscillating frequency will drift. Also, the frequency must be measured at a specific point on the RTWO and this frequency will change during phase tuning.

Figure 6.16: Test Board 2 Phase Detector Output Signals

6.7 Discussion of Results - RTWO Phase Alignment

The SXOR phase detectors were used to measure and calibrate the phase of 45° transmission line sections of an 18-GHz RTWO fabricated in SiGe BiCMOS. Although the convergence of the phase tuning was unsuccessful, the phase alignment algorithm was able to change the phase of the RTWO TL sections. The following was also demonstrated:

1. The phase alignment algorithm implemented in verilog functioned correctly as can be seen in Figures 6.13 and 6.15.
2. The phase alignment algorithm will minimize the change in RTWO average frequency during phase tuning. This change was $\approx 150 \text{ MHz}$ for the devices tested.
3. The reason the fabricated RTWO could not be tuned is understood. That is, the phase detector output signals are not the correct signals for the existing phase alignment algorithm. The phases being measured are $\Delta 1 + 180^\circ$, $\Delta 2$, $\Delta 3$, and $(\Delta 1 + \Delta 2 + \Delta 3)$. The voltage levels and the tuning direction for these signals need to be corrected by a revised version of the existing phase alignment algorithm. This work is currently under way and because these simulations take an extended period of time, are continuing on past the deadline for this document.
4. Additional varactor components should be added to each TL section to extend the phase tuning range. The length of the RTWO TL sections would then need to be reduced to compensate for the increase in total nominal capacitance in each TL section. That is, to maintain the same operating frequency.
5. In order to minimize power consumption, this RTWO phase calibration should be carried out during initial system power-on and thereafter, periodically or after frequency changes.

The following errors were addressed during the testing of the fabricated RTWO phase alignment test die.

Systematic Error

1. No calibration was required for this test set-up.

Random Error

1. The switch state changes used to control the FPGA code were debounced in the verilog code.

Drift Error

1. Test equipment was turned on and allowed to warm up to an operating temperature for at least 45 minutes before each test run.
2. All test runs were carried out without pauses. The objective here was to minimize the effects of room temperature and other drift mechanisms within the test equipment.

Chapter 7

Conclusions and Recommendations

7.1 Summary of Work

Simple Symmetrical XOR (SXOR) phase detectors were designed, fabricated in SiGe BiCMOS and tested to resolve 1° of phase error over $45^\circ \pm 10^\circ$ and $90^\circ \pm 10^\circ$ up to 12 GHz. This 12 GHz limit is the maximum operating frequency of the test set-up (i.e., limited by the Narda 3753B Phase Shifter), not the maximum operating frequency of the DUT. The following was also demonstrated:

1. In simulation this extracted phase detector can resolve 1° of phase error over $45^\circ \pm 10^\circ$ and $90^\circ \pm 10^\circ$ up to a frequency of 20.0 GHz.
2. The differential transit delay through an XOR circuit can be equalized by using a symmetrical XOR or SXOR architecture.
3. The simple ECL outputs of these phase detectors are very sensitive to V_{cc} variation in single-ended operation. This problem can be mitigated by using the SXOR outputs differentially.
4. If multiple differential SXOR phase detectors are used to make phase measurements, these measurement values can be compared and used to make accurate

RTWO phase adjustment. That is, relative phase measurements can be made using SXOR gates without the need of complex calibration functions. This has been demonstrated in simulation and is certainly possible in a fabricated circuit, although it has not been fully demonstrated here.

These phase detectors were used to measure and calibrate the phase of 45° transmission line sections of an existing 18-GHz RTWO fabricated in SiGe BiCMOS. Although the convergence of the phase tuning was unsuccessful, the phase alignment algorithm was able to change the phase of the RTWO TL sections. The following was also demonstrated:

1. A simple phase alignment algorithm was implemented to phase tune an RTWO. This was successfully demonstrated using both Matlab/Simulink and Cadence AMS Designer.
2. The phase alignment algorithm will minimize the change in RTWO average frequency during phase tuning. That is, at each algorithm iteration a tuning voltage increase is made to one TL section and the tuning voltage is decreased by this same amount in another TL section. The objective here is to add and remove equal amounts of capacitance to maintain a constant average frequency of oscillation. However, the slope of the varactor capacitance curve changes with control voltage (see Figure 5.8); therefore, the total capacitance of the RTWO TL sections will change with phase tuning and the oscillating frequency will drift. Also, the frequency must be measured at a specific point on the RTWO and this frequency will change during phase tuning.
3. The phase alignment algorithm described in this work will converge the TL section phases of an RTWO to be within an acceptable tolerance value regardless of the RTWO direction of rotation.

4. The reason the fabricated RTWO could not be tuned is understood. That is, the phase detector output signals are not the correct signals for the existing phase alignment algorithm. Also, the tuning direction for this incorrect signal will need to be reversed.
5. Equation (2.18) was derived and used to estimate the change in phase of an RTWO TL section for a change in varactor capacitance in that RTWO section.

7.2 Contributions

Simple Symmetrical XOR (SXOR) phase detectors were designed, fabricated in SiGe BiCMOS and tested to resolve 1° of phase error over $45^\circ \pm 10^\circ$ and $90^\circ \pm 10^\circ$ up to 12 GHz. These phase detectors were used to measure and calibrate the phase of 45° transmission line sections of an existing 18-GHz RTWO fabricated in SiGe BiCMOS. This phase tuning was carried out by a simple phase alignment algorithm that can be implemented easily in a hardware definition language, in this case verilog. This RTWO phase tuning system was shown to function correctly in simulation. Although the fabricated version did not function correctly, the circuit error is understood and can be corrected.

7.3 Improvement and Future Work

1. The current RTWO die was layed out in great haste to make a tape out date. The following improvements could be made to this circuit.
 - (a) The phase detector connection problems identified previously (Section 5.4, Figures 5.11, 5.12 and 5.13) could be corrected with a die revision.

- (b) Additional capacitive tuning elements components should be added to each TL section to extend the phase tuning range. The length of the RTWO TL sections would then need to be reduced to compensate for the increase in total nominal capacitance in each TL section. That is, to maintain the same operating frequency.
 - (c) The layout should be assessed and steps taken to reduce signal coupling and improve signal isolation (i.e., reduce feed-through at f_o and $2f_o$).
 - (d) Both positive and negative SXOR outputs should be brought out to pads.
 - (e) Sense lines should be installed to accurately measure V_{cc} and I_{bias} voltage levels for individual devices.
2. Op Amp circuits should be added to the differential SXOR outputs to produce phase measurement signals that are immune to V_{cc} variation.
 3. In order to minimize power consumption, this RTWO phase calibration should be carried out during initial system power-on and thereafter, periodically or after frequency changes.
 4. Cycle-to-cycle phase detection at 18 GHz dissipates significant power. Additional methods of phase detection that consume less power should be explored.

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Appendix A

Additional Simulink RTWO Block Descriptions

A.1 Sub-Blocks of the RTWO 24 MHz Block

The gain stages of Figure 3.6 are illustrated in Figure A.1 at the next level of abstraction down. Here there are two summing nodes that add the input signals from the RTWO TL sections to their amplified complementary signals delayed by 180° , i.e., Integer Delay $Z^{-(FUS/2)}$. The Amplifier blocks have enough gain to produce a 1.2 V square wave output from the attenuated input signal. The Saturation block prevents the amplified signal from exceeding 1.2 V as the power supply would in a CMOS inverter.

The sum of the incoming RTWO signal and its amplified complement will produce a signal that is larger than 1.2 V so it is limited by a Saturation block. The Transfer fcn blocks are LPFs that reduce the amplifier gain as frequency increases to mimic the gain bandwidth product of a real amplifier. Here $\alpha = 1/(2\pi Freq \times N)$, where N is the number of harmonics to be included. It was found that $N \geq 20$ was required for the simulation to work well, and $N = 60$ was used for the simulation results shown

in this document. Rate Transition blocks are required by Simulink[®] where ever a signal path changes form continuous to discrete.

Figure A.1: RTWO Back-to-Back Gain Block

The Amplifier block is made of two sub-blocks, Rate Limiter and Discrete Derivative. The Rate Limiter block output is a positive slew rate signal when the input signal is high (i.e., $> 0.6 V$) and a negative slew rate signal when the input signal is low or $< 0.6 V$. The gain or slew rate for both positive and negative ramps is $24 MV/s$. The Discrete Derivative block differentiates these input positive and negative going ramps to produce a square wave output. The gain of this block is 5×10^{-8} , so the cascaded gains produce a square wave of amplitude $24 \times 10^6 \times 5 \times 10^{-8} = 1.2 V$. At this point, the question can be asked: why not implement this function with a simple amplifier followed by a Saturation block to limit the output signal level? There

are two problems with this approach:

1. In a circuit simulator amplifier gain and output signal level are limited by the compression point of the amplifier (P1dB). The Simulink[®] amplifiers have no compression point; therefore, if there is more open-loop gain in the system than loss (a Barkhausen criterion), the RTWO travelling wave signal will accumulate gain and its level will head off to infinity as it rotates around the loop over and over again. The input signal can be mixed with itself, amplified and combined to create a third order component that will compress the gain. Unfortunately, this does not limit the output signal level and the travelling wave signal level heads off to infinity as before.
2. Simulink[®] amplifiers are simple multiply functions. That is, any signal level at the amplifier input is multiplied by the selected gain to produce an output signal. Therefore, if the low signal level is not exactly zero, but some very small positive value, it too will head to infinity as it continuously travels around the RTWO. In a circuit simulator the transistor based amplifiers drive the output signal to high and low levels that are limited by the power supply.

As long as the attenuation of any TL section is less than 6.0 *dB* so the input high signal level of the Amplifier blocks of Figure A.1 is greater than 0.6 *V*, the output signal will have a high level of 1.2 *V* and a low level of 0 *V*.

The controllable delay blocks of Figure 3.6 (i.e., Delay 1 to Delay 4) are shown in Figure A.2. These Variable Integer Delay blocks are programmed to provide discrete delay from input to output. Here the output signal is the input signal delayed by *i* TSs. These blocks are initialized to 250 TS by the Phase Alignment Algorithm.

Figure A.2: RTWO Variable Integer Delay Block

Figure A.3 shows the details of the 45° Section block of Figure 3.6. Here, from input to output, we have TL section Loss and LPF elements followed by a Rate Transition block to convert the continuous signal to a discrete signal so a nominal 45° delay can be applied (i.e., $FUS/8$) with block Integer Delay. The offset_n portion of this variable is present in each TL section (i.e., offset_1 to offset_4) and is used to insert a initial system phase skew so the Phase Alignment Algorithm (Figure 3.9) will have something to correct. The setup.m file initializes $Loss = 0.6$ and $alpha = 1/(2\pi Freq \times N)$, where $N = 60$ is the number of harmonics to be included.

Figure A.3: RTWO 45 Degree Section

A.2 Gain Blocks

The implementation of the CML buffers of Figure 3.1 is illustrated in Figure A.4. The differential input signals levels are halved and the negative input is subtracted from the positive input. The result is applied to two gain stages, ± 0.5833 , to produce the CML amplitude swing of 350 mV . A 0.85 V bias is then applied to these signals to mimic a CML signal at the outputs. That is, $V_{OH} = 1.2\text{ V}$ and $V_{OL} = 0.5\text{ V}$.

Figure A.4: CML Buffer for RTWO Phase Taps

The differential in to single-ended out buffer of Figure 3.1 is shown in Figure A.5. The negative differential CML input is subtracted from the positive differential CML input and the resulting amplitude is halved and biased at 0.85 V to produce a single-ended output, $V_{OH} = 1.2\text{ V}$ and $V_{OL} = 0.5\text{ V}$.

Figure A.5: Differential In to Single-Ended Out Buffer

In A.6 the single-ended buffer at the filtered output of the XOR phase detector of Figure 3.1 is described. The buffer biases the input voltage at -0.6 V to remove its dc offset so the signal can be amplified by five without exceeding the 1.2 V input

range of the ADC. This can be considered an effective coding gain of 14.0 *dB* for the ADC function, which will improve the tuning resolution of the system.

Figure A.6: Single-Ended Buffer

Appendix B

Phase Calibration Algorithm Code

B.1 Cadence Verilog Code

The following Verilog code is resident in the Phase Alignment Algorithm block of Figure 3.10.

```
// Verilog HDL for 4 Section RTWO Phase Calibration.
```

```
module Phase_Cal_1(T1,T0,CCW,TCo,LCo,BCo,RCo,BPi,RPi,TPi,LPi,Status);
```

```
input      T1,T0,CCW;
input      [11:0]  BPi,RPi,TPi,LPi;
output     [11:0]  Status;
output     [8:0]   TCo,LCo,BCo,RCo;
```

```
wire       T1,T0,CCW;
wire       [11:0]  BPi,RPi,TPi,LPi;
wire       [11:0]  Status;
wire       [8:0]   TCo,LCo,BCo,RCo;
```

```
reg        [11:0]  MaxVal,MinVal,Limit,Outsig;
reg        [8:0]   sum1,sum2,sum3,sum4;
reg        [2:0]   Maxi,Mini;
reg        [3:0]   Delta;
```

```

assign      TCo = sum1;
assign      LCo = sum2;
assign      BCo = sum3;
assign      RCo = sum4;
assign      Status = Outsig;

function [23:0] max_min;    //Returns the max and min phase measurements.
    input [11:0] A,B,C,D;
    reg [11:0] max,min;
    begin
        if ((A >= B) && (A >= C) && (A >= D))
            max = A;
        else if ((B >= A) && (B >= C) && (B >= D))
            max = B;
        else if ((C >= A) && (C >= B) && (C >= D))
            max = C;
        else if ((D >= A) && (D >= B) && (D >= C))
            max = D;
        else
            max = 12'b00000000000000;
        if ((A <= B) && (A <= C) && (A <= D))
            min = A;
        else if ((B <= A) && (B <= C) && (B <= D))
            min = B;
        else if ((C <= A) && (C <= B) && (C <= D))
            min = C;
        else if ((D <= A) && (D <= B) && (D <= C))
            min = D;
        else
            min = 12'b00000000000000;
        max_min = {max,min};
    end
endfunction

```

```

always @(posedge T1 or posedge T0)
begin
    if (!T0)          //Initializes the initial RTWO tuning and sets the stop limit.
    begin
        sum1 <= 9'b101101000;
        sum2 <= 9'b1000000000;
        sum3 <= 9'b010011000;
        sum4 <= 9'b1000000000;
        Limit <= 12'b0000000000110;
        Outsig <= 12'b0000000000000;
    end
    else
    begin
        {MaxVal,MinVal} <= max_min(BPi,RPi,TPi,LPi);
        Outsig <= MaxVal - MinVal;
        if ((MaxVal - MinVal) > Limit)      //If phase error > limit.
        begin                               //Determine the min and max TL sections.
            if (TPi == MaxVal)
                Maxi <= 3'b001;
            else if (LPi == MaxVal)
                Maxi <= 3'b010;
            else if (BPi == MaxVal)
                Maxi <= 3'b011;
            else if (RPi == MaxVal)
                Maxi <= 3'b100;
            else
                Maxi <= 3'b000;

            if (TPi == MinVal)
                Mini <= 3'b001;
            else if (LPi == MinVal)
                Mini <= 3'b010;
            else if (BPi == MinVal)
                Mini <= 3'b011;
        end
    end
end

```

```

else if (RPi == MinVal)
    Mini <= 3'b100;
else
    Mini <= 3'b000;

if ((MaxVal - MinVal) > 6'b111111)    //Set phase adjustment.
    Delta = 4'b1011;
else if ((MaxVal - MinVal) > 4'b1111)
    Delta = 4'b0011;
else
    Delta = 4'b0001;

case (Maxi)    //Make appropriate phase adjustment.
    3'b001:
        sum1 = sum1 - Delta;
    3'b010:
        sum2 = sum2 - Delta;
    3'b011:
        sum3 = sum3 - Delta;
    3'b100:
        sum4 = sum4 - Delta;
    default: Outssig = 12'b111111111111;
endcase
case (Mini)
    3'b001:
        sum1 = sum1 + Delta;
    3'b010:
        sum2 = sum2 + Delta;
    3'b011:
        sum3 = sum3 + Delta;
    3'b100:
        sum4 = sum4 + Delta;
    default: Outssig = 12'b111111111111;
endcase

```

```

        end
    end
end
endmodule

```

B.2 FPGA Verilog Code

The following Verilog code is resident in the Phase Alignment Algorithm block of Figure 6.10. The main program is DAC_ADC_prog_2, which calls IBUFG to buffer the Spartan-6 FPGA oscillator input; debounce to qualify the initialization and iteration external switch inputs; and Phase_Cal_Mult_2 to calculate the new varactor voltages from the current phase detector output voltages.

```

`timescale 1ns / 1ps
/////////////////////////////////////////////////////////////////
// Company: Carleton University
// Engineer: Raleigh Smith
//
// Create Date:    18:54:04 05/14/2014
// Module Name:    DAC_ADC_prog_2
// Project Name: RTWO Phase Alignment Lab Testing.
// Target Devices: Xilinx Spartan-6
// Tool versions: ISE 14.7
//
/* Description: At initialization sets up Maxim MAX126 ADC for
4-channel operation. At every iteration reads the 4-channels
of the ADC, calls Phase_Cal_Mult_2 to process these phase
detector output values and writes newly calculated varactor
voltage levels to the AD7305 4-channel DAC. */
//
// Dependencies: IBUFG, debounce and Phase_Cal_Mult_2.
//
// Revision: 4.2
// Additional Comments:

```

```
//
////////////////////////////////////////////////////////////////
module DAC_ADC_prog_2(CONVSTb, CSb, RDb, WRb, INTb, A, T0, T1, AD_WRb,
Add, Data, LDACb, clock);

input    T0, T1, clock, INTb;
inout    [13:0] A;
output    CONVSTb, CSb, RDb, WRb, AD_WRb, LDACb;
output    [1:0] Add;
output    [7:0] Data;

wire      clock, clock_buffered, T1, T1_deb, T0, T0_deb, INTb;
wire      [13:0] A;
wire      [7:0] TCo, LCo, BCo, RCo;
reg        CONVSTb, CSb, RDb, WRb, AD_WRb, LDACb, setup, clock_1, update;
reg        [7:0] wr_count;
reg        [13:0] BPi_reg, RPi_reg, TPi_reg, LPi_reg, sequence;
reg        [1:0] Add;
reg        [7:0] Data;
reg        [3:0] count;
reg        [10:0] count_1;

assign    A = sequence;

    IBUFG globalClockBuffer (.I(clock), .O(clock_buffered));
    debounce debounce_1(0, clock_buffered, T1, T1_deb);
    debounce debounce_2(0, clock_buffered, T0, T0_deb);
    Phase_Cal_Mult_2 Phase_Cal(T0_deb,TCo,LCo,BCo,RCo,BPi_refg,RPi_reg,
    TPi_reg,LPi_reg,
    update,clock_1);

always @(posedge clock_buffered)
```



```

begin
    count_1 <= count_1 + 1;
    if (count_1 == 50)
        begin
            clock_1 <= 1'b1;
count_1 <= 0;
        end
    else
        clock_1 <= 1'b0;
    end

always @(posedge clock_1)
begin
    if (!T0_deb)
        begin
            count <= 4'b0000;
            sequence <= 14'bzzzzzzzzzzzzzzzz;
            CONVSTb <= 1'b1;
            RDb <= 1'b1;
            CSb <= 1'b1;
            WRb <= 1'b1;
            setup <= 1'b0;
            AD_WRb <= 1'b1;
            LDACb <= 1'b1;
            wr_count <= 8'b01000000;
/*      TCo <= 8'b01111111;
            LCo <= 8'b01111111;
            BCo <= 8'b01111111;
            RCo <= 8'b01111111;  */
            update <= 1'b0;
            Data <= 8'bzzzzzzzz;

        end

    else if ((T0_deb) && (count <= 4'b1100))

```

```
begin
    count <= count + 1;
    case (count)
    4'b0001:
        begin
            CSb <= 1'b0;
            sequence <= 14'bzzzzzzzzzz0011;
            WRb <= 1'b1;
        end
    4'b0010:
        begin
            CSb <= 1'b0;
            WRb <= 1'b0;
        end
    4'b0111:
        begin
            CSb <= 1'b0;
            WRb <= 1'b1;
        end
    4'b1000:
        begin
            CSb <= 1'b1;
            WRb <= 1'b1;
        end
    4'b1100:
        begin
            CSb <= 1'b1;
            sequence <= 14'bzzzzzzzzzzzzzz;
            WRb <= 1'b1;
        end
    setup <= 1'b1;
    end
endcase
end
```

```

if (setup)
begin
  if (T1_deb)
  begin
    CONVSTb <= 1'b1;
    CSb <= 1'b1;
    RDb <= 1'b1;
    WRb <= 1'b1;
    AD_WRb <= 1'b1;
    LDACb <= 1'b1;
    wr_count <= 8'b00000000;
    Add <= 2'b00;
    Data <= 8'bzzzzzzzz;
  end
  else if ((INTb) && (wr_count <= 8'b00000111))
  begin
    wr_count <= wr_count + 1;
    case (wr_count)
      8'b000000001: //10 nsec
    begin
      WRb <= 1'b0;
      CONVSTb <= 1'b0;
    end
      8'b000000111: //70 nsec
    begin
      WRb <= 1'b1;
      CONVSTb <= 1'b1;
    end
    endcase
  end
  else if ((!INTb) || ((wr_count >= 8'b00001001) && (wr_count <=
8'b00101111)))
  begin
    wr_count <= wr_count + 1;

```

```

case (wr_count)
  8'b00001001:      //90 nsec
  begin
    CSb <= 1'b0;
    RDb <= 1'b0;
  end
  8'b00001110:      //140 nsec
  begin
    CSb <= 1'b1;
    RDb <= 1'b1;
  //    BPi_reg <= 14'b01110010100011;
    BPi_reg <= A;
  end
  8'b00010100:      //200 nsec
  begin
    CSb <= 1'b0;
    RDb <= 1'b0;
  end
  8'b00011001:      //250 nsec
  begin
    CSb <= 1'b1;
    RDb <= 1'b1;
  //    RPi_reg <= 14'b01110010100011;
    RPi_reg <= A;
  end
  8'b00011111:      //310 nsec
  begin
    CSb <= 1'b0;
    RDb <= 1'b0;
  end
  8'b00100100:      //360 sec
  begin
    CSb <= 1'b1;
    RDb <= 1'b1;

```

```

//      TPi_reg <= 14'b01110010100011;
      TPi_reg <= A;
    end
    8'b00101010:      //420 nsec
    begin
      CSb <= 1'b0;
      RDb <= 1'b0;
    end
    8'b00101111:      //470 nsec
    begin
      CSb <= 1'b1;
      RDb <= 1'b1;
//      LPi_reg <= 14'b01110010100011;
      LPi_reg <= A;
    end
  endcase
end
else if ((!INTb) || ((wr_count >= 8'b00110000) && (wr_count <=
8'b00111111)))) //480 nsec to 1420 nsec.
  begin
    wr_count <= wr_count + 1;
case (wr_count)
  8'b00110011:
    update <= 1'b1;
  8'b00110100:
    update <= 1'b0;
endcase
end
else if ((!INTb) || ((wr_count >= 8'b01000000) && (wr_count <=
8'b10001110)))) //660 nsec to 1420 nsec.
  begin
    wr_count <= wr_count + 1;
    case (wr_count)
      8'b01000011:      //670 nsec

```

```

begin
    Add <= 2'b00;
    Data <= TCo;
    AD_WRb <= 1'b0;
end
8'b01001101:          //770 nsec
    AD_WRb <= 1'b1;
8'b01010010:          //820 nsec
begin
    Add <= 2'b01;
    Data <= LCo;
    AD_WRb <= 1'b0;
end
8'b01011100:          //920 nsec
    AD_WRb <= 1'b1;
8'b01100001:          //970 nsec
begin
    Add <= 2'b10;
    Data <= BCo;
    AD_WRb <= 1'b0;
end
8'b01101011:          //1070 sec
    AD_WRb <= 1'b1;
8'b01110000:          //1120 nsec
begin
    Add <= 2'b11;
    Data <= RCo;
    AD_WRb <= 1'b0;
end
8'b01111010:          //1220 nsec
    AD_WRb <= 1'b1;
8'b01111111:          //1270 nsec
    Data <= 8'bzzzzzzzz;
8'b10000100:          //1320 nsec

```

[illegible]

```

module debounce (reset, clock, noisy, clean);

    input reset, clock, noisy;
    output clean;

    parameter NDELAY = 650000;
    parameter NBITS = 20;

    reg [NBITS-1:0] count;
    reg xnew, clean;

    // IBUFG globalClockBuffer (.I(clock), .O(clock_buffered));

    always @(posedge clock)
        if (reset) begin xnew <= noisy; clean <= noisy; count <= 0; end
        else if (noisy != xnew) begin xnew <= noisy; count <= 0; end
        else if (count == NDELAY) clean <= xnew;
        else count <= count+1;

endmodule

'timescale 1ns / 1ps
/////////////////////////////////////////////////////////////////
// Company: Carleton University
// Engineer: Raleigh Smith
//
// Create Date:    20:21:51 05/19/2014
// Module Name:    Phase_Cal_Mult_2
// Project Name:   RTWO Phase Alignment Lab Testing.
// Target Devices: Xilinx Spartan-6
// Tool versions:  ISE 14.7
/* Description: This program creates new varactor voltage levels
from the current phase detector voltage levels.*/

```



```

//
// Dependencies: A buffered clock and valid phase detector data.
//
// Revision: many.
// Additional Comments:
//
////////////////////////////////////////////////////////////////
module Phase_Cal_Mult_2(T0,TCo,LCo,BCo,RCo,BPi,RPi,TPi,LPi,update,
clock);

//inout      update;
input        T0,update,clock;
input        [13:0] BPi,RPi,TPi,LPi;
output       [7:0]  TCo,LCo,BCo,RCo;

wire         T0,update,clock;
wire         [13:0] BPi,RPi,TPi,LPi;
wire         [7:0]  TCo,LCo,BCo,RCo;

reg          [13:0] Limit;
reg          [7:0]  sum1,sum2,sum3,sum4;
reg          [2:0]  Maxi,Mini,Delta;
reg [13:0]    max,min;

assign       TCo = sum1;
assign       LCo = sum2;
assign       BCo = sum3;
assign       RCo = sum4;

always @(BPi, RPi, TPi, LPi)
begin
    if ((BPi >= RPi) && (BPi >= TPi) && (BPi >= LPi))
        max <= BPi;
    else if ((RPi >= BPi) && (RPi >= TPi) && (RPi >= LPi))

```

```

    max <= RPi;
else if ((TPi >= BPi) && (TPi >= RPi) && (TPi >= LPi))
    max <= TPi;
else if ((LPi >= BPi) && (LPi >= RPi) && (LPi >= TPi))
    max <= LPi;
else
    max <= 14'b0000000000000000;
if ((BPi <= RPi) && (BPi <= TPi) && (BPi <= LPi))
    min <= BPi;
else if ((RPi <= BPi) && (RPi <= TPi) && (RPi <= LPi))
    min <= RPi;
else if ((TPi <= BPi) && (TPi <= RPi) && (TPi <= LPi))
    min <= TPi;
else if ((LPi <= BPi) && (LPi <= RPi) && (LPi <= TPi))
    min <= LPi;
else
    min <= 14'b0000000000000000;
end

```

```

always @(posedge clock)
begin
    if ((max - min) > Limit)
    begin
        if (TPi == max)
            Maxi <= 3'b001;
        else if (LPi == max)
            Maxi <= 3'b010;
        else if (BPi == max)
            Maxi <= 3'b011;
        else if (RPi == max)
            Maxi <= 3'b100;
        else
            Maxi <= 3'b000;
    end
end

```

```

    if (TPi == min)
        Mini <= 3'b001;
    else if (LPi == min)
        Mini <= 3'b010;
    else if (BPi == min)
        Mini <= 3'b011;
    else if (RPi == min)
        Mini <= 3'b100;
    else
        Mini <= 3'b000;

    if ((max - min) > 6'b111111)
        Delta <= 3'b111;
    else if ((max - min) > 3'b111)
        Delta <= 3'b011;
    else
        Delta <= 3'b001;
end
end

always @(posedge clock)
begin
    if (!T0)
        begin
            Limit <= 14'b000000000000011;
            sum1 <= 8'b01111111;
            sum2 <= 8'b01111111;
            sum3 <= 8'b01111111;
            sum4 <= 8'b01111111;
        end
    else if ((T0) && (update))
        begin
            case (Maxi)

```

```

    3'b001:
        sum1 <= sum1 + Delta;
    3'b010:
        sum2 <= sum2 + Delta;
    3'b011:
        sum3 <= sum3 + Delta;
    3'b100:
        sum4 <= sum4 + Delta;
endcase
case (Mini)
    3'b001:
        sum1 <= sum1 - Delta;
    3'b010:
        sum2 <= sum2 - Delta;
    3'b011:
        sum3 <= sum3 - Delta;
    3'b100:
        sum4 <= sum4 - Delta;
endcase
end
end

endmodule

```

B.3 Matlab Code

The following is Matlab code is resident in the Phase Alignment Algorithm block of Figure 3.8.

```

function [RST,stp,No,TOTo1,TOTo2,TOTo3,TOTo4,Out1,Out2,Out3,Out4,
Dly1,Dly2,Dly3,Dly4] = fcn(T0,T1,In1,In2,In3,In4,Ni,TOTi1,TOTi2,
TOTi3,TOTi4,In1_,In2_,In3_,In4_,D1,D2,D3,D4)
% This block supports an embeddable subset of the MATLAB language.

```

```

delta = 0;
delta_min = 0;
delta_max = 0;
total = 0;
sz = [0 0 0 0];
stp = 0;

if T0 == 1                                %After T0 expires execute this code
    No = Ni + 1;                          %Update variables from last correction iteration.
    TOTo1 = TOTi1 + In1;
    TOTo2 = TOTi2 + In2;
    TOTo3 = TOTi3 + In3;
    TOTo4 = TOTi4 + In4;
    Dly1 = D1;
    Dly2 = D2;
    Dly3 = D3;
    Dly4 = D4;
    Out1 = In1_;
    Out2 = In2_;
    Out3 = In3_;
    Out4 = In4_;

    if No == 18000
        Out1 = TOTo1/No;
        Out2 = TOTo2/No;
        Out3 = TOTo3/No;
        Out4 = TOTo4/No;
        TOTo1 = 0;
        TOTo2 = 0;
        TOTo3 = 0;
        TOTo4 = 0;
        No = 0;
    end
end

```

```

if (T1 ==1)
    RST = 1;
    In_vec = [In1_ In2_ In3_ In4_];    %Assemble phase measurement vector.
    diff = (max(In_vec)-min(In_vec));  %Find phase error range.
    if diff > 7                        %If its > limit,
        delta = 0.05*diff;           %set TL section adjustment value.

        if In_vec(1) == max(In_vec)  %Find max/min TL sections.
            sz(1) = 3;
            total = 3;
        elseif In_vec(1) == min(In_vec)
            sz(1) = 1;
            total = 1;
        else
            sz(1) = 2;
            total = 2;
        end

        if In_vec(2) == max(In_vec)
            sz(2) = 3;
            total = total + 3;
        elseif In_vec(2) == min(In_vec)
            sz(2) = 1;
            total = total + 1;
        else
            sz(2) = 2;
            total = total + 2;
        end

        if In_vec(3) == max(In_vec)
            sz(3) = 3;
            total = total + 3;
        elseif In_vec(3) == min(In_vec)

```

```
        sz(3) = 1;
        total = total + 1;
    else
        sz(3) = 2;
        total = total + 2;
    end

    if In_vec(4) == max(In_vec)
        sz(4) = 3;
        total = total + 3;
    elseif In_vec(4) == min(In_vec)
        sz(4) = 1;
        total = total + 1;
    else
        sz(4) = 2;
        total = total + 2;
    end

    if total == 6
        delta_min = delta/3;
        delta_max = delta;
    elseif total == 7
        delta_min = delta/2;
        delta_max = delta;
    elseif total == 9
        delta_max = delta/2;
        delta_min = delta;
    elseif total == 10
        delta_max = delta/3;
        delta_min = delta;
    else
        delta_min = delta;
        delta_max = delta;
    end

    %Adjustment is spread over more
    %than one minimum or maximum TL
    %section?
```

```

        if sz(1) == 3                                %Apply delay adjustment.
            Dly1 = D1 - delta_max;
        elseif sz(1) == 1
            Dly1 = D1 + delta_min;
        else
            Dly1 = D1;
        end

        if sz(2) == 3
            Dly2 = D2 - delta_max;
        elseif sz(2) == 1
            Dly2 = D2 + delta_min;
        else
            Dly2 = D2;
        end

        if sz(3) == 3
            Dly3 = D3 - delta_max;
        elseif sz(3) == 1
            Dly3 = D3 + delta_min;
        else
            Dly3 = D3;
        end

        if sz(4) == 3
            Dly4 = D4 - delta_max;
        elseif sz(4) == 1
            Dly4 = D4 + delta_min;
        else
            Dly4 = D4;
        end
    else
        stp = 1;
    end

```



```
        end

    else
        RST = 0;
        total = 2;
    end
else %Initialize variables until T0 expires.
    RST = 0;
    No = 0;
    Dly1 = 250;
    Dly2 = 250;
    Dly3 = 250;
    Dly4 = 250;
    Out1 = 0;
    Out2 = 0;
    Out3 = 0;
    Out4 = 0;
    TOTo1 = 0;
    TOTo2 = 0;
    TOTo3 = 0;
    TOTo4 = 0;
end
```

Appendix C

Test and Evaluation Printed Circuit Boards

C.1 Phase Detector PCB

This section describes the PCB that was used to evaluate both versions of the SXOR phase detector. Figure C.1 is the schematic diagram of the phase detector test card.

Figure C.1: Phase Detector Test Card Schematic

Figure C.2 is the bill of materials of the phase detector test card.

Item	Reference	Quantity	Manufacturer	Description	Part Number
1	PCB1	15	SpeedyPCB.com	Phase Detector Test PCB	Phdetect_001
2	C5, C6, C7, C8	4	AVX	0603, X7R, 1uF, 16V, 10%	0603YC105KAT2A
3	R1, R2, R3, R4	4	Panasonic	0603, 10K, 1%	ERJ-3EKF1002V
4	R1, R2, R3, R4	4	Panasonic	0603, 20K, 1%	ERJ-3EKF2002V
5	R5, R6, R7, R8	4	Panasonic	0603, 18K, 1%	ERJ-3EKF1802V
6	X1	1	Molex	87832-16 Circuit Header, Shrouded	0878321620
7	R9, R10, R11, R12	4	Bourns	10K, 10 Turn Pot, SMT, Vertical	3224X-1-103E
8	U1, U2, U3, U4	4	ST-Microelectronics	SOT23-5L, Low Noise, Adj, LDO, 1.8 - 5.5V	LDCL015MR
9	C1, C2, C3, C4	4	AVX	1206, X7R, , 10uF, 10V,	1206ZC106KAT2A
10	C9, C10	2	AVX	0603, NPO, 2pF, 50V, 0.25pF	06035A2R0CAT2A
11	R17, R18	2	Panasonic	0603, 1K, 1%	ERJ-3EKF1001V
12	R13, R14, R15, R16	4	Panasonic	0603, 1 Ohm, 1%	ERJ-3RQF1R0V
13	L1, L2, L3, L4	4	Murata Electronics NA	0603, 43nH, 2%, 380mA, 2.7GHz	LQW18AN43NG00D
14	F1, F2, F3, F4	4	Johanson Dielectrics	0603, X2Y, 100nF	100X14W104MV4T
15		50	Molex	Terminals, 24 - 30 AWG	503948054
16		4	Molex	Housing, 2mm, 16 Pos, Polarized	511101651

Figure C.2: Phase Detector Test Card BOM

Figure C.3 is the primary side of the fabricated phase detector test card, which shows the component placement including the DUT.

Figure C.3: Phase Detector Test Card Layout - Primary Side

Figure C.4 is the secondary side of the fabricated of the phase detector test card.

Figure C.4: Phase Detector Test Card Layout - Secondary Side

C.2 RTWO Phase Alignment PCB

This section describes the PCB that was used to evaluate the tuning of the RTWO provided by Kimia Ansari. Figure C.5 is page one of the schematic diagram of the RTWO tuning test card.

Figure C.5: RTWO Tuning Test Card Schematic - Page 1

Figure C.6 is page two of the schematic diagram of the RTWO tuning test card.

Figure C.6: RTWO Tuning Test Card Schematic - Page 2

Figure C.7 is the bill of materials of the RTWO tuning test card.

Item	Reference	Quantity	Manufacturer	Description	Part Number
1	PCB1	5	SpeedyPCB.com	RTWO/Phase Detector Test Card	RTWK3_PD_Tune
2	C7, C8, C9, C10, C11, C12, C18, C19, C20, C21, C22	11	AVX	0603, X5R, 2.2uF, 10V, 10%	0603ZD225KAT2A
3	C23	1	AVX	1210, X5R, 100uF, 6.3V, 20%	12106D107MAT2A
4	C24	1	AVX	0402, X7R, 100pF, 16V, 10%	0402YC101KAT2A
5	R1, R2, R3, R4, R5, R6, R25, R26, R27, R28, R29	11	Panasonic	0603, 10K, 1%	ERJ-3EKF1002V
6	R7, R8, R9, R10, R11, R12, R30, R31, R32, R33, R34	11	Panasonic	0603, 18K, 1%	ERJ-3EKF1802V
7	X1	1	Molex	87832-08 Circuit Header, Shrouded	0878320820
8	X2	1	Molex	87832-24 Circuit Header, Shrouded	0878322420
9	R13, R14, R15, R16, R17, R18, R35, R36, R37, R38, R39	11	Bourns	10K, 10 Turn Pot, SMT, Vertical	3224X-1-103E
10	U1, U2, U3, U4, U5, U6, U7, U8, U9, U10, U11	11	ST-Microelectronics	SOT23-5L, 300mA, Low Noise, Adj, LDO, 1.8 - 5.5V	LDK130M-R
11	C1, C2, C3, C4, C5, C6, C13, C14, C15, C16, C17	11	AVX	1206, X7R, , 10uF, 10V,	1206ZC106KAT2A
12	R19, R20, R21, R22, R23, R24, R40, R41, R42 R43, R44	11	Panasonic	0603, 1 Ohm, 1%	ERJ-3RQF1R0V
13	L1, L2, L3, L4, L5, L6, L7, L8, L9, L10, L11, L12	12	Murata Electronics NA	0603, 43nH, 2%, 380mA, 2.7GHz	LQW18AN43NG00D
14	F1, F2, F3, F4, F5, F6, F7, F8, F9, F10, F11, F12	12	Johanson Dielectrics	0603, X2Y, 100nF	100X14W104MV4T
15		150	Molex	Terminals, 24 - 30 AWG	503948054
16		4	Molex	Housing, 2mm, 24 Pos, Polarized	51110-2451
16		4	Molex	Housing, 2mm, 8 Pos, Polarized	51110-0851

Figure C.7: RTWO Tuning Test Card BOM

Figure C.8 is the primary side of the fabricated RTWO tuning test card, which shows the component placement including the DUT.

Figure C.9: RTWO Tuning Test Card Layout - Secondary Side