Improved RF Power Extraction from 1.55 μm Ge-on-SOI PIN Photodiodes with Load Impedance Optimization

A Thesis Presented to the
Electrical Engineering Department Faculty of
California Polytechnic State University, San Luis Obispo

In Partial Fulfillment
Of the Requirements for the
Master of Science Degree in Electrical Engineering

By
Andrew L. Huard

June 8, 2010

© 2010

Andrew L. Huard

ALL RIGHTS RESERVED

Committee Membership

Title: Improved RF Power Extraction from 1.55 μ m Ge-on-SOI PIN Photodiodes with Load Impedance Optimization

Author: Andrew L. Huard

Date Submitted: June 8, 2010

Committee Chair : Dr. Dennis J. Derickson

Committee Member: Dr. Samuel O. Agbo

Committee Member: Dr. John A. Saghri

Abstract

Title: Improved RF Power Extraction from 1.55 μ m Ge-on-SOI PIN Photodiodes with Load Impedance Optimization

Author: Andrew L. Huard

VLSI miniaturization has created the need for high-density, low-cost, monolithically-integrated optical interconnects. High output power photodetectors are needed to directly drive load circuitry, which improves the noise performance and dynamic range of optical communications links by eliminating a post amplifier stage. Elimination of the post amplifier also reduces circuit cost and complexity. A new Si-Ge PIN waveguide photodiode with 31GHz bandwidth and 93% quantum efficiency at 1550nm has been developed by Yin et al.⁷³, which was fabricated using standard CMOS processes on a Silicon substrate. This thesis demonstrates a method for improving the RF power extraction from these photodiodes by increasing the impedance of the load. An RF output power improvement of 5.5dB is obtained by increasing the load resistance from 50 Ω to 177 Ω with 15MHz modulation. The maximum obtainable RF power of all devices tested using 50 Ω and 100 Ω loads at 15MHz is 15.73dBm and 17.83dBm, respectively. The maximum obtainable RF power using a 177 Ω load for all devices tested is 17.67dBm, which is slightly smaller than that obtained with a 100 Ω load. A measurement procedure for RF power extraction at microwave frequencies is also described. Quarter-wavelength 70.71 Ω thin film coplanar waveguides are designed to transform 50 Ω to a higher impedance of 100 Ω for measurements of improved RF power extraction at 3GHz and 7GHz.

Keywords: Optoelectronics, RF power extraction, 1dB compression, compression current, optical interconnects, quantum efficiency, responsivity, bandwidth, P-I-N, Si-Ge photodiodes, waveguide photodiodes, signal integrity, thin film coplanar waveguide, quarter-wavelength transformer

Acknowledgements

I would like to express my sincerest gratitude to Dr. Dennis Derickson. His patience and enthusiasm are more than I could ever expect from a thesis advisor. The guidance, assistance, and encouragement he provided throughout our research and coursework have been invaluable toward my development as an engineer.

Much appreciation goes to Anand Ramaswamy and Molly Piels for their assistance with the high-power measurements. I would also like to thank Dr. John Bowers for allowing us access to his lab. Special thanks goes to Molly Piels for her quick and detailed replies to my many emails.

Finally, I would like to thank my family for their patience and support. Their unwavering encouragement and support were something I could always count on throughout my entire life.

Table of Contents

List of Figures	viii
List of Tables	xiii
1. Introduction.....	1
2. Optical Communications and Data Links	4
2.1 Signal Integrity of Electrical Interconnects	4
2.2 Need for Optical Interconnects.....	7
2.3 Developments in III-V Laser Sources	10
2.4 Transition from III-V to Silicon-Germanium in Waveguide Photodetectors.....	11
2.5 The Need for High-Bandwidth, High-Power Photodiodes	14
2.6 Monolithic Integration of Photodetectors on Silicon	16
3. Performance Limitations in High Power Photodiodes.....	19
3.1 Dark Current.....	19
3.2 Noise Sources	19
3.3 High Optical Power Considerations	22
3.4 Bandwidth Limitations	24
3.5 Quantum Efficiency and Responsivity.....	26
4. Comparison of Absorber Structure Types	27
4.1 Lateral versus Surface-Normal Photodetectors	27
4.2 Conventional P-I-N	31
4.3 Avalanche Photodiodes (APD)	32
4.4 Metal-Semiconductor-Metal (MSM) Photodiodes.....	34
4.5 Type-II Superlattice Photodiodes	35
4.6 Quantum Well Intersubband Photodetectors (QWIP).....	37
4.7 Partially Depleted Absorber (PDA)	43
4.8 Uni-Travelling Carrier (UTC)	45
4.9 Resonant-Cavity Enhanced (RCE).....	45
4.10 Travelling-Wave Photodiode (TWPD)	48
4.11 Velocity-Matched Distributed Photodetectors (VMDP).....	49
4.12 Survey of Photodetectors from the Past to the Present	52
5. Silicon-Germanium Evanescent P-I-N Waveguide Photodetector	58
5.1 Device Performance	59
5.2 Fabrication and Design Considerations.....	63
5.3 Sources of Optical Loss.....	65
5.4 Thermal Considerations	66
6. Device Characterization.....	68
6.1 IV-Characterization	68
6.2 Responsivity Measurement	85
6.3 Bandwidth Measurements	97
6.4 S11 Measurements	116
7. RF Power Compression Measurements Using a 50Ω Load	124
7.1 Construction of an Erbium-Doped Fiber Amplifier (EDFA).....	125
7.2 Oscilloscope Measurement of the Signal Compression	136
7.3 High Power RF Measurements at 15MHz	140

8. Compression Measurements Using Higher, Non-Standard Impedances	148
8.1 Low Frequency Load Construction	149
8.2 High Power RF Compression Measurements at 15MHz	162
9. Microwave-Frequency Power Compression	171
9.1 Microwave Frequency Load Design	171
9.2 High Power Microwave Compression Measurements at 3GHz and 7GHz	176
10. Summary	181
References	183
Glossary	187
Index	188

List of Figures

Figure 2.1: Eye diagrams showing (a) the ideal case and (b) an actual measurement with the strobe signal superimposed. Several important parameters related to signal integrity measurements are also shown. (Source: Hall, Hall, and McCall ²⁵)	5
Figure 2.2: Frequency performance regions of a stripline trace (Source: Johnson and Graham ³⁹)	7
Figure 2.3: Cross section of a monolithically integrated optoelectronic communications link using a SiO ₂ waveguide as the optical interconnect medium (Source: Huang ²⁹)	9
Figure 2.4: Laser power output drops with increasing ambient temperature (Source: Fang ²⁰)	11
Figure 2.5: Growth of Silicon-Germanium (SiGe) alloy on a silicon substrate produces “strained” silicon (Source: IBM ³⁰)	13
Figure 2.6: Block diagram of an optical receiver (Source: Jalali ³⁷)	14
Figure 2.7: Eye Diagram of the output of an optical link. (Source: Yin ⁷³)	15
Figure 2.8: Absorption coefficient for selected semiconductor materials (Source: Palik ⁴⁸)	16
Figure 2.9: Absorption coefficient for Ge _x Si _{1-x} alloys (Source: Jalali ³⁷)	17
Figure 3.1: Visible representation of increasing signal to noise ratio (Source: Razeghi ⁵²)	20
Figure 3.2: Discrete component diode model (Source: Chang ¹⁰)	22
Figure 3.3: Space charge effect shown by the space charge electric field negating part of the bias electric field, which reduces the overall net electric field across the intrinsic region	23
Figure 4.1: Comparison between (a) surface-normal and (b) lateral photodetectors (Source: Wu ⁷¹)	28
Figure 4.2: P-I-N photodiode (a) doping profile and (b) band diagram (Source: Razeghi ⁵²)	31
Figure 4.3: Surface Illuminated P-I-N photodiode showing (a) top-illumination and (b) bottom-illumination. (Source: Razeghi ⁵²)	32
Figure 4.4: Energy band diagram of an avalanche photodiode P-N junction (Source: Razeghi ⁵²)	33
Figure 4.5: Metal-semiconductor-metal (MSM) photodiode energy band diagram under the influence of an external bias and showing the photo-excitation of charge carriers (Source: Razeghi ⁵²)	35
Figure 4.6: Heterostructure energy band lineups for (a) type-I superlattice, (b) type-II superlattice staggered, and (c) type-II superlattice broken gap (Source: Razeghi ⁵²)	36
Figure 4.7: Energy band diagram of a type-II superlattice showing the overlapping wavefunctions of the majority carrier in each material (Source: Razeghi ⁵²)	36
Figure 4.8: Intersubband energy levels in a GaAs-AlGaAs quantum well. Showing a type-I superlattice heterostructure with both valance and conduction subbands. (Source: Razeghi ⁵²)	37
Figure 4.9: Quantization of the density of states, $g_{2D}(E)$, in a 2-dimensional quantum well. The three-dimensional density of states, $g_{3D}(E)$, is shown as a dashed line. (Source: Razeghi ⁵²)	38
Figure 4.10: Allowed energy states in a 2-dimensional quantum well with infinite potential barriers (Source: Razeghi ⁵²)	39
Figure 4.11: two-dimensional quantum well (a) conduction band energy diagram showing finite barrier potential and (b) wavefunction showing evanescent tails in the barrier region (Source: Razeghi ⁵²)	39
Figure 4.12: Number of intersubband energy levels is controlled by the potential barrier height and the width of the quantum well (Source: Razeghi ⁵²)	40
Figure 4.13: Active region of a quantum well intersubband photodetector (QWIP) showing (a) an intersubband transition in a wide well and (b) a transition from a quasi-bound state to the continuum conduction state in a narrow well. (Source: Razeghi ⁵²)	40
Figure 4.14: Propagation of a photogenerated carrier through multiple periods in a multiple quantum well region (Source: Liu ⁴⁴)	41

Figure 4.15: An MQW/SCH structure showing an exaggerated active region width. Optical confinement to the active region is typically between 1-5%. (Source: Razeghi ⁵²)	42
Figure 4.16: Structure of a device using Multiple Quantum Wells (MQW) and a Separated Confinement Heterostructure (SCH). 14 periods of $\text{Ge}_x\text{Se}_{1-x}$ to i-Si quantum wells are shown, which are surrounded a 1 μm thick cladding. (Source: Jalali ³⁷)	43
Figure 4.17: Comparison of (a) traditional P-I-N and (b) PDA doping profiles. Photogenerated carriers and fixed space charges are shown. (Source: Tulchinsky ⁵⁹)	43
Figure 4.18: Structure of an RCE photodiode showing a reflective resonant cavity surrounding an absorbing material (Source: Davidson ¹⁵)	46
Figure 4.19: 850nm RCE photodetector showing very high wavelength selectivity (Source: Davidson ¹⁵)	47
Figure 4.20: Diagram of a traveling-wave photodiode showing an impedance match to the external load (Source: Giboney ²³)	49
Figure 4.21: Schematic diagram of a VMDP showing electrical transmission line segments (Source: Davidson ¹⁵)	50
Figure 4.22: VMDP consisting of distributed P-I-N detectors (Source: Islam ³⁴)	50
Figure 4.23: Balanced distributed photodetector (Source: Islam ³⁴)	51
Figure 4.24: Photodiode survey showing the reported 1dB compression RF power versus bandwidth	53
Figure 5.1: Schematic and cross section of the Si-Ge Evanescent P-I-N Photodetector (Source: Ramaswamy ⁵¹)	58
Figure 5.2: (a) Photocurrent as a function of input optical power for several bias voltages and (b) DC responsivity for various optical powers and bias voltages (Source: Ramaswamy ⁵¹)	61
Figure 5.3: Comparison of several photodetectors by DC electrical power dissipation versus bandwidth. Waveguide photodetectors are shown in red and surface-normal detectors are shown in blue. (Source: Ramaswamy ⁵¹)	61
Figure 5.4: (a) Frequency response and (b) 3dB bandwidth, which shows excellent bandwidth linearity as a function of DC photocurrent (Source: Ramaswamy ⁵¹)	63
Figure 5.5: SEM cross section of a fabricated device (Source: Ramaswamy ⁵¹)	64
Figure 5.6: Thermal simulation showing 85°C temperature in the active region with 1W of DC power dissipated in the device (Source: Ramaswamy ⁵¹)	67
Figure 6.1: Measurement setup showing the VNA, high-bandwidth Rosenberger cable, and GS probe	69
Figure 6.2: Measurement setup showing the GS probe, photodetector chip, and lensed fiber assembly	70
Figure 6.3: Wirebond connection between the photodiode aluminum contact pads and a gold 50 Ω coplanar waveguide	71
Figure 6.4: Dimension parameters used by LineCalc's CPWG module (Source: Agilent)	72
Figure 6.5: Measurement setup for testing diodes without the G-S probe, which makes electrical connection using gold wirebond and a gold thin film coplanar waveguide	72
Figure 6.6: Block level diagram showing electrical and optical connections	73
Figure 6.7: IV characteristic block diagram	74
Figure 6.8: IV-Characteristic for Chip 1 Device 2	75
Figure 6.9: IV-Characteristic for all functioning devices on Chip 1	75
Figure 6.10: IV-characteristic for all functional devices on chip 2	76
Figure 6.11: IV-characteristic for all functional devices on chip 3	77
Figure 6.12: Discrete component diode model (Source: Chang ¹⁰)	78
Figure 6.13: Dark current performance regions for a PN junction diode (Source: Razeghi ⁵²)	79
Figure 6.14: Forward-bias IV-characteristic for C5D1, showing a series resistance of 24.8 Ω at 475°C	79
Figure 6.15: Extended IV characteristic for C2D2	81
Figure 6.16: Diagram showing locations of the series resistances R_{pad} and R_s	84

Figure 6.17: Microscope photograph of the lensed fiber, photodiode chip, and GS probe	86
Figure 6.18: Responsivity measurement block level diagram	86
Figure 6.19: System responsivity measurement of C2D2 at 6V reverse bias	88
Figure 6.20: System responsivity of C2D5 with 6V reverse bias and perfect lensed fiber alignment (0.23A/W), and with the lensed fiber misaligned (0.14A/W)	89
Figure 6.21: System responsivity measurement for C2D1 at several bias levels	90
Figure 6.22: System responsivity versus bias	90
Figure 6.23: System responsivity for all devices on chip 2, showing an average responsivity of 0.23A/W	91
Figure 6.24: Device responsivity for all devices on chip 2, showing an average responsivity of 0.70A/W	93
Figure 6.25: System responsivity of C5D1 showing 0.314A/W, or a device responsivity of 0.973A/W	94
Figure 6.26: Voltage dependent responsivity for C5D1 under high optical power conditions (+18dBm).....	95
Figure 6.27: Voltage-dependent responsivity for an InGaAs/InP photodiode operating at 50mA photocurrent (Source: Beling ⁶).....	96
Figure 6.28: Mach-Zehnder DC Bias Characterization	97
Figure 6.29: Mach-Zehnder DC Bias Characterization, Showing Optical Insertion Loss.....	98
Figure 6.30: Bandwidth measurement block diagram	99
Figure 6.31: S12 system measurement of C1D5 for various reverse biases and 0.6mA photocurrent....	100
Figure 6.32: S12 measurement of the Rosenberger coaxial cable	101
Figure 2.33: S12 response of the GS probe, which is recorded from the datasheet.....	101
Figure 6.34: S12 response estimated from the Mach-Zehnder interferometer datasheet	102
Figure 6.35: (a) System S12 measurement of C2D2, and (b) calculated S12 for the diode only	103
Figure 6.36: Measurement of the 3dB bandwidth for C2D2 at 0.2V reverse bias.....	104
Figure 6.37: 3dB Bandwidth of all devices on chips 2 and 3	105
Figure 6.38: Lensed fiber protrusion length illustration	106
Figure 6.39: (a) system S12 and (b) S12 referred to the diode (C1D5)	107
Figure 6.40: Bandwidth versus bias voltage for C1D5	107
Figure 6.41: Calculated junction capacitance for C1D5, showing a constant $C_j(V)$ of 1pF at a reverse bias greater than 2V	109
Figure 6.42: 3dB bandwidth calculation for C5D1 at 6V bias.....	111
Figure 6.43: Normalized S12 response for C3D1 with a wirebond electrical connection.....	112
Figure 6.44: S12 Measurement for C3D1 with system insertion loss correction	114
Figure 6.45: 3dB bandwidth measurement for C3D1 (wirebonded).....	114
Figure 6.46: Junction capacitance calculation for C3D1 showing RC and transit time limited bandwidth	115
Figure 6.47: Discrete component diode model (Source: Chang ¹⁰)	116
Figure 6.48: S11 measurement (blue) and curve fit (black) for C4D1 at 6V bias	117
Figure 6.49: Junction capacitance versus bias, measurement method comparison (C4D1)	118
Figure 6.50: Junction capacitance versus bias, measurement method comparison (C4D2)	118
Figure 6.51: Junction capacitance versus bias, measurement method comparison (C4D3)	119
Figure 6.52: Junction capacitance versus bias, measurement method comparison (C4D5)	119
Figure 6.53: Junction capacitance versus bias, measurement method comparison (C4D6)	120
Figure 6.54: Junction capacitance versus bias, measurement method comparison (C5D1)	120
Figure 6.55: Junction capacitance versus bias, measurement method comparison (C5D3)	121
Figure 6.56: Junction capacitance versus bias, measurement method comparison (C5D5)	121
Figure 6.57: Junction capacitance versus bias, measurement method comparison (C5D6)	122

Figure 6.58: Capacitance versus bias, showing low frequency detail; calculated using low-frequency reactance method (C5D3).....	123
Figure 7.1: Block diagram of an optical amplifier (Source: Derickson ¹⁷).....	125
Figure 7.2: Erbium energy band diagram (Source: Derickson ¹⁷)	126
Figure 7.3: Block diagram of an EDFA (Source: Derickson ¹⁷)	127
Figure 7.4: Simulation of the optical output power versus Erbium-doped fiber length	127
Figure 7.5: Total ASE versus Erbium-fiber length	128
Figure 7.6: ASE at 1550nm versus Erbium-fiber length	129
Figure 7.7: EDFA block level diagram.....	130
Figure 7.8: Final construction of our Erbium-doped Fiber Amplifier	131
Figure 7.9: Measurement of ASE for various levels of pump laser current. Pump laser power is 300mW at its rated maximum current of 600mA	132
Figure 7.10: Measured total ASE versus pump laser current	133
Figure 7.11: Measured 1550nm ASE versus pump laser current.....	133
Figure 7.12: Spectral characteristic of the EDFA input signal, showing +8.16dBm power at 1550nm.....	134
Figure 7.13: Spectral characteristic of the EDFA output signal, showing +18.61dBm power at 1550nm.....	135
Figure 7.14: Oscilloscope measurement block level diagram	136
Figure 7.15: RF compression waveform shown for C2D2 with 2V reverse bias and 20mA of photocurrent. No compression is present with 10mA of photocurrent	137
Figure 7.16: RF compression shown for C2D2 with 1V reverse bias and 10mA of photocurrent	140
Figure 7.17: Experimental setup for RF compression measurements under high optical power conditions	142
Figure 7.18: RF compression measurement setup showing the lensed fiber, photodiode, probe, attenuator load, coaxial RF output cable, and photodiode DC bias cable	143
Figure 7.19: RF compression composite plot for C2D2 with biases ranging from 1V to 6V (15MHz).....	144
Figure 7.20: Electrical RF power compression of C2D2 at 1V bias and 15MHz.....	145
Figure 8.1: Pi- network attenuator used to construct the 50 Ω , 100 Ω , and 177 Ω low frequency loads ...	150
Figure 8.2: Various RFC designs containing (a) all ferrite bead chokes and (b) combinations of two or three low-Q ferrite core inductors in series with three ferrite bead chokes	151
Figure 8.3: (a) Impedance and (b) S11 Smith Chart for the 9 ferrite bead RFC.....	151
Figure 8.4: (a) Impedance and (b) S11 Smith Chart for the 6 ferrite bead RFC.....	152
Figure 8.5: (a) Impedance and (b) S11 Smith Chart for the 3 ferrite bead RFC.....	152
Figure 8.6: (a) Impedance and (b) S11 Smith Chart for the 1 ferrite bead RFC.....	153
Figure 8.7: (a) Impedance and (b) S11 Smith Chart for the 1 ferrite coil RFC	153
Figure 8.8: (a) Impedance and (b) S11 Smith Chart for the 2 ferrite coil RFC	154
Figure 8.9: (a) Impedance and (b) S11 Smith Chart for the 1 ferrite coil and 3 ferrite bead RFC	154
Figure 8.10: (a) Impedance and (b) S11 Smith Chart for the RFC design with 2 ferrite coils and 3 ferrite beads.....	155
Figure 8.11: (a) Impedance and (b) S11 Smith Chart for the RFC design with 2 ferrite coils and 3 ferrite beads with an input wire loop.....	156
Figure 8.12: (a) Input impedance and (b) S11 Smith Chart of the pi-network attenuator seen from the device port without the RFC choke connected.....	157
Figure 8.13: (a) Input impedance and (b) S22 Smith Chart of the pi-network attenuator seen from the device port without the RFC choke connected.....	158
Figure 8.14: (a) Input impedance and (b) S11 Smith Chart of the final 177 Ω pi-network attenuator design with the RFC attached	159
Figure 8.15: (a) Input impedance and (b) S11 Smith Chart of the final 100 Ω pi-network attenuator design.....	159

Figure 8.16: (a) Input impedance and (b) S11 Smith Chart of the final 50 Ω pi-network attenuator design.....	160
Figure 8.17: 100 Ω pi-network attenuator load.....	160
Figure 8.18: 177 Ω pi-network attenuator load.....	161
Figure 8.19: RF compression composite plot for C2D1 using a 177 Ω pi-network attenuator load	163
Figure 8.20: RF Power versus the photocurrent at compression for several loads (C2D2).....	159
Figure 8.21: RF Power versus the photocurrent at compression for several loads (C2D4).....	161
Figure 8.22: RF Power versus the photocurrent at compression for several loads (C2D1).....	162
Figure 8.23: RF Power versus the photocurrent at compression for several loads (C2D5).....	163
Figure 8.24: RF Power versus the photocurrent at compression for several loads (C2D6).....	164
Figure 9.1: Microwave measurement setup showing the positions of the 100 Ω quarter-wavelength thin film coplanar waveguide, photodetector, and SMA connector	172
Figure 9.2: 3GHz quarter wavelength transformer matching 50 Ω to 100 Ω	174
Figure 9.3: 7GHz quarter wavelength transformer matching 50 Ω to 100 Ω	175
Figure 9.4: Quarterwave transformer simulation showing a real input impedance of 100 Ω at (a) 3GHz and (b) 7GHz using a 50 Ω load and a 70.71 Ω transformer characteristic impedance	177
Figure 9.5: 7-Segment transmission line approximation of an elliptical wirebond	178
Figure 9.6: 7-segment transmission line model showing the wirebond length needed for C5D1	179

List of Tables

Table 1.1: Summary of photodetectors tested and the measurements performed	2
Table 4.1: Survey of Selected Photodetectors from 1992-2010.....	55
Table 5.1: Summary of bandwidth results for various detector lengths and a detector width of 7.4 μ m (Sources: Ramaswamy ⁵¹ and Yin ⁷³)	60
Table 5.2: Sources of Optical Loss (Source: Yin ⁷³).....	66
Table 6.1: Dimensions for each photodiode and the device name for each	68
Table 6.2: ADS LineCalc parameters used to replicate the 50 Ω thin film coplanar waveguide dimensions	71
Table 6.3: 3dB bandwidth obtained from an S12 measurement of C2D5	104
Table 6.4: Calculated junction capacitance for all devices on chips 2 and 3 at 6V and 10V reverse bias	110
Table 6.5: Calculated junction capacitance for C3D1.....	115
Table 6.6: S11 curve fit parameters for C4D1	117
Table 6.7: Calculated junction capacitance at 6V for all devices on chips 4 and 5	123
Table 7.1: RF Compression Measurements using the input terminal of a spectrum analyzer as a 50 Ω load	147
Table 8.1: RFC impedances obtained for each design	156
Table 8.2: Resistor selection for each desired input impedance of the pi-network attenuator.....	157
Table 8.3: Calculated and measured input impedance for each attenuator design.....	160
Table 8.4: RF compression data for C2D2.....	165
Table 8.5: Maximum 1dB compressed RF power at 6V reverse bias and 15MHz modulation.....	170
Table 9.1: Coplanar waveguide simulation parameters for Agilent's ADS LineCalc CPWG module....	173
Table 9.2: Wirebond length selection summary	180

1. Introduction

A revolution is occurring in high-speed computing. In a February 2010 whitepaper titled “Using 10-Gbps Transceivers in 40G/100G Applications,” Altera Corporation⁴ writes that the recent developments in telecommunications equipment, specifically high-speed bridges and switches, has spurred a migration toward cutting-edge 100 Gbps interfaces that push the limits of bandwidth and throughput in modern computing devices.

VLSI architectures have become smaller and denser with an ever-increasing pin count. In the last decade, transistor density within integrated circuits has doubled every 18 months. Close packing and decreasing effective cross-sectional area of interconnects presents numerous problems in high frequency applications, most notably unwanted capacitive-mode crosstalk and ohmic heat generation.

Electrical interconnects are restricted by bandwidth limitations and other signal integrity issues when operating at extreme frequencies, typically around 10-20GHz. Recent advances in on-chip laser sources and monolithically integrated, high power Silicon-Germanium photodetectors make the use of optical interconnects an increasingly attractive alternative to copper traces.

This thesis explores the current state of optical interconnect technology and presents a new Si-Ge based P-I-N photodetector developed by Yin et al.⁷³ that is fabricated using standard CMOS compatible processes. The use of Si-Ge instead of III-V materials reduces fabrication costs and allows these detectors to be monolithically integrated onto a Silicon substrate.

This photodetector has a vertical P-I-N structure that consists of an intrinsic Germanium absorber positioned on top of a Silicon rib waveguide. The entire structure is fabricated on a Silicon-on-Insulator (SOI) wafer, which provides very high electrical insulation from the device to the bulk. In addition, the SOI oxide insulation layer improves optical confinement in the Si-waveguide due to the distinct contrast between the index of refraction for silicon dioxide (1.46) and silicon (3.5) at 1550nm.

The photodetectors tested are listed in table 1.1:

Device Name	Absorber Width (μm)	Absorber Length (μm)	Measurements Performed				
			RF Power	IV Char	Bandwidth	Responsivity	Input Impedance
C1D2	7.4	250		X			
C1D5	5.8	500		X	X		
C2D1	5.8	100	X	X	X	X	
C2D2	5.8	250	X	X	X	X	
C2D4	4.4	250	X	X	X	X	
C2D5	4.4	500	X	X	X	X	
C2D6	4.4	250	X	X	X	X	
C3D1	3.8	250		X	X		
C3D3	3.8	250		X	X		
C3D4	2.4	250		X	X		
C4D1	10.4	250					X
C4D2	10.4	250					X
C4D3	5.4	250					X
C4D5	5.4	250					X
C4D6	5.4	500					X
C5D1	9.0	250		X	X	X	X
C5D3	7.4	250					X
C5D5	7.4	500					X
C5D6	7.4	250					X

Table 1.1: Summary of photodetectors tested and the measurements performed

The primary goal of this work is to attempt to extract the maximum possible RF electrical output power from these photodiodes under very high optical power conditions. The maximum 1dB compressed electrical RF output power and photocurrent obtained using a 50Ω load with a 15MHz amplitude modulated optical signal at 1550nm and 6V reverse bias are 17.07dBm and 70.39mA, respectively. The optical power needed to sustain this compression current is approximately 18dBm.

By increasing the load resistance to 100Ω , 17.83dBm of output RF power is obtained with a photocurrent of 54.57mA. This RF power result, using the 100Ω load with 15MHz modulation, is over 3dB greater than that obtained by Ramaswamy et al.⁵¹ at 1GHz and 8V bias (14.35dBm). Consistent results were observed across all devices tested.

This paper is organized into 10 chapters. The first chapter is the introduction you are currently reading. Chapter 2 covers the fundamentals of optical links, outlines the benefits of using optical interconnects, and argues in favor of the transition from copper-based interconnects to optoelectronic systems based on Silicon-Germanium. Chapter 3 highlights the device limitations that are generally

present in high-power photodetectors. Chapter 4 discusses the solid-state device physics for several photodetector types and compares several photodetectors found in the literature. Its purpose is to lay the foundation for discussion on the device operation and performance of the photodetector by Yin et al.⁷³ in Chapter 5. Chapter 6 outlines the obtained responsivity, dark current, and bandwidth for several devices as well as the experimental methods used. Chapter 7 explores RF saturation under high optical power conditions into a 50Ω load at low frequency (15MHz). Chapter 8 repeats these low frequency measurements with a variety of resistive loads in the attempt to extract additional RF power. Chapter 9 is a theoretical chapter outlining a method for repeating the measurements of chapter 8 except at microwave frequencies.

2. Optical Communications and Data Links

This section studies the need for optical interconnects in high speed communications links. Recent developments in laser sources are discussed, which are fundamentally significant in forming an optical data link. The need for high power photodiode receivers is also discussed. This section then concludes with a discussion on the monolithic integration of these high-speed, high-power waveguide photodiodes.

2.1 Signal Integrity of Electrical Interconnects

Neighboring interconnects sometimes experience considerable coupling capacitance due to close packing. Among other detrimental effects, system bandwidth degrades as the RC time constant becomes larger with decreasing interconnect spacing and increasing coplanar capacitance. High frequency signal integrity suffers because the charging and discharging of interconnects does not have enough time to fully complete before the next subsequent cycle begins; therefore, the bandwidth of electrical interconnects is limited by the RC response of the copper trace.

Digital devices are especially vulnerable to RC bandwidth rolloff since square-wave clock signals have significant high frequency harmonic content. The rise- and fall-time of the clock degrades as the RC time constant is increased.

Logic errors may result if the rise-time or fall-time is large in comparison to the clock period. For example, consider the eye diagram of high frequency square wave with some harmonic degradation. As the signal becomes RC limited, the square-wave edges of the clock begin to appear rounded and eventually approach the appearance of a sinusoid with severe bandwidth limiting. If the fundamental frequency of the clock signal is greater than the 3dB cutoff frequency of the system, then the overall amplitude of the signal is reduced as well; illustrated on an eye diagram, this quasi-sinusoidal “eye” will appear to close. Stephen Hall, Garrett Hall, and James McCall²⁵ illustrate an ideal eye diagram along with an example of an actual eye diagram measurement in figure 2.1:

Figure 2.1: Eye diagrams showing (a) the ideal case and (b) an actual measurement with the strobe signal superimposed. Several important parameters related to signal integrity measurements are also shown. (Source: Hall, Hall, and McCall²⁵)

Crosstalk between electrical interconnects is also a significant problem. Several common VLSI architectures exist for reducing crosstalk. The most common techniques involve increasing the separation between traces and positioning sensitive striplines close to their underlying reference planes to reduce the reach of fringing electric and magnetic fields (Hall, Hall, and McCall²⁵ 67).

Methods for reducing crosstalk in an interconnect network typically consume valuable space. The current trend in VLSI miniaturization presents an interesting problem. Unwanted capacitive-mode and inductive-mode coupling, and heat generation will intensify as interconnect density increases. Therefore,

a fundamental limit exists in printed circuit board (PCB) trace density when considering the signal integrity requirements of a high-frequency system.

The phenomenal growth of the Internet and subsequently increasing need for higher bandwidth are spurring the need for ever-faster data transfer rates. Currently, electronic switching technology limits network routing speeds to below 50 Gbps. Dr. Hossin Abdeldayem of Frazier's optical technologies research group predicts that Terabit speeds will be required to contain the growth of the Internet and other bandwidth intensive applications (qtd. in Knier⁴⁰).

Resistive losses are also a problem for high-speed electrical interconnects. Skin effect losses force current away from the center area of the conductor and toward the outer rim. The result is a decreased effective cross sectional area of the metal trace. This effect increases with the square-root of the operating frequency, and it is modeled by a series resistor. Dielectric losses are also prevalent at very high frequencies, and increase linearly with the frequency of operation. This loss is modeled by a shunt resistor. Both skin effect and dielectric loss are inherent qualities of electrical interconnects and cannot be entirely avoided at high frequency. However, skin effect is commonly mitigated by using a larger trace circumference, and dielectric loss is reduced through the use of low dielectric constant substrates. Howard Johnson and Martin Graham³⁹ illustrate the performance regions versus trace length and frequency in figure 2.2:

Figure 2.2: Frequency performance regions of a stripline trace (Source: Johnson and Graham³⁹)

Figure 2.2 shows that even minimally small (1mm) trace lengths are subject to dielectric losses when operated above 8GHz. According to Johnson and Graham³⁹, the bandwidth of traces has been traditionally improved by decreasing the trace length. However, this is no longer the case as speed requirements have pushed the maximum tolerable trace length to miniscule dimensions.

A tradeoff exists between crosstalk and skin effect. Reducing skin effect losses by increasing the trace area has the detrimental effect of increasing capacitive-mode crosstalk. In addition, dielectric losses can only be mitigated by selection of a substrate material with sufficiently low dielectric constant. RT Duroid 5880, a common PCB substrate, has a relative dielectric constant of 2.2. The dielectric constant of air ($\epsilon_r=1.00059$) is very close to that of vacuum ($\epsilon_r=1$) and is the best dielectric material for high speed applications. Since existing substrate materials already have a low dielectric constant, it seems unlikely that further reduction will have any appreciable effect on performance.

2.2 Need for Optical Interconnects

Clearly, a paradigm shift is needed to achieve these feats in computer performance. Problems posed by interconnect miniaturization are fundamentally electrical; therefore, an alternative signal medium must be considered. Signal integrity issues inherent to electrical interconnects are nonexistent in

optical interconnects. A single optical interconnect may serve as a full-duplex transmission medium, allowing for transmission and reception of many simultaneous frequency channels. According to Knier⁴⁰, Photonic multiplexing offers significant cost savings as well as increased performance, scalability and reduced complexity.

Electrical interconnects are a significant bottleneck in the implementation of multi-chip modules (MCMs) at very high clock frequencies. In 1994, B. Jalali and A. F. J. Levi³⁷ write that microprocessors operating at frequencies up to 1GHz will likely contain memory packaged in a multi-chip module, with the need to communicate with other MCMs on other PCBs. The Intel 2010 Core i7-960, one of Intel Corporation's latest and fastest quad-core processors for consumer use at the time of writing this paper (2010), has a clock speed of 3.20GHz, which is more than a 3-fold increase in speed compared to similar microprocessors made 14 years ago in 1996.

This speed increase further exasperates the signal integrity issues inherent with using electrical interconnects. Although an active electrical bus may support these high frequencies, Jalali³⁷ adds, this method is expensive, and designers are disposed toward the development of alternative technologies. Jalali³⁷ concludes, "The use of high-performance single channel and multiple channel array optical data links based on packaging of electronic and photonic devices is just such an alternative technology." Therefore, the development of low-cost, scalable optical interconnects is necessary for the development of systems requiring high bandwidth performance. Tao Yin⁷³ writes, "As the aggregate data rate for several applications is approaching or exceeding 100 Gbps, it is increasingly desirable to develop integrated optical components with reduced size, complexity, and cost." Jalali³⁷ agrees with Yin's⁷³ assessment, "the present cost of optical interconnection techniques limits their application to ultrahigh performance or long distance transmission systems."

Beiju Huang²⁹ demonstrates a monolithically integrated optical data link, which uses only CMOS compatible processes. Monolithic integration of the photodetector with a Si substrate is an important goal for the realization of high speed optical data links. Park⁴⁹ explains, "Integration of the photodetector with the receiver is critical for lower capacitance and higher sensitivity." In the structure proposed by Huang²⁹,

a light emitting diode (LED) light source is used as the transmitter, and a waveguide photodiode is used as the receiver. The transmission medium is a SiO_2 waveguide, which is surrounded by metallization on all sides. The entire device is fabricated on a silicon substrate. This is shown in figure 2.3 below:

Figure 2.3: Cross section of a monolithically integrated optoelectronic communications link using a SiO_2 waveguide as the optical interconnect medium (Source: Huang²⁹)

Recently, Park⁴⁹ writes, Silicon-based Mach-Zehnder interferometer structures have been demonstrated on Silicon. Park⁴⁹ explains, “Strained silicon has been shown to break the inversion symmetry of silicon allowing silicon to exhibit electro-optic refractive index modulation.” The creation of a monolithic modulator on silicon is an important development toward the realization of low-cost, high-volume optoelectronic data links using standard CMOS processes.

The development of monolithically integrated optical gain elements is also an important research area, which could relax the power output requirements for laser sources. However, limitations in device physics for silicon are hampering research in this area. Park⁴⁹ explains, “The indirect bandgap of silicon has been a key hurdle for achieving optical gain elements. Although Raman lasers and amplifiers, and optical gain in nanopatterned silicon have been observed, an electrically pumped silicon waveguide gain element has been an unsolved challenge.”

The absence of effective optical gain elements on silicon requires the incorporation of prefabricated, high-power lasers in the design, Park⁴⁹ reasons. However, this alternative method has limits in expense and scalability. Park⁴⁹ asserts, “due to the tight alignment tolerances of the optical modes and the need to align each laser individually, this method has limited scalability, and it is difficult to envision die attaching more than a few lasers to each chip without prohibitive expense.” To satisfy these

requirements, Park⁴⁹ notes, laser sources with III-V epitaxial growth on silicon and an efficient laser to waveguide coupling scheme are needed. Recent progress in this research area, Park⁴⁹ reasons, “widen the possibility of building monolithically integrated on-chip laser sources on the silicon photonics platform.” Indeed, such a development would make silicon photonics an attractive medium to achieve low-cost, scalable, on-chip optical interconnects.

2.3 Developments in III-V Laser Sources

For optical interconnects to become a commercially viable substitute for electronic interconnects, a silicon-based, low-cost laser source must be developed. "Silicon Photonics is a critical part of tera-Scale computing," Intel Corporation CTO Justin Ratter explains. "We need the ability to move massive amounts of data on and off these very high performance chips" (qtd. in Intel³¹). Development of efficient III-V compound light sources has been a necessary step toward the ultimate goal of achieving inexpensive integration between optical and electronic components.

In 2007, researchers from Intel and the University of California, Santa Barbara (UCSB) have developed the world's first electrically pumped hybrid AlGaInAs-silicon evanescent laser using ordinary CMOS fabrication methods. The device is produced at low-cost and with high-volume, with the potential for hundreds of lasers to be fabricated in a single bonding step. The output frequency may be adjusted by varying the percent composition of the III-V layers, enabling the device to operate as the transmission component of a multi-channel transceiver for use in multiplexing applications. Physical separation between adjacent lasers is very small, on the order of 1-3.5 microns, allowing the incorporation of many waveguides on a single chip. A press release by Intel Corporation³¹ highlights the benefits of this work:

Researchers believe that with this development, silicon photonic chips containing dozens or even hundreds of hybrid silicon lasers could someday be built using standard high-volume, low-cost silicon manufacturing techniques. This development addresses one of the key hurdles to producing low-cost, highly integrated silicon photonic chips for use inside and around PCs, Servers, and future data centers.

Unfortunately, the Intel-UCSB hybrid laser suffers from relatively poor temperature tolerance due to poor heat extraction from the III-V active region. The laser has a maximum ambient lasing temperature of 40°C with the active region experiencing temperatures as high as 60°C (Fang et al.²⁰). Poor performance may result in ambient temperatures above 40°C with the laser exhibiting a high current threshold and poor output power under these conditions. This temperature dependence is shown in figure 2.4:

Figure 2.4: Laser power output drops with increasing ambient temperature (Source: Fang²⁰)

To maintain optimal operation, design measures must be considered to ensure that the laser package remains cool. The Intel-UCSB researchers acknowledge that heat-sinking the top side of the laser could theoretically reduce thermal resistivity in the device by as much as 24 °C/W, which is a considerable decrease (Fang et al.²⁰).

2.4 Transition From III-V to Silicon-Germanium in Waveguide Photodetectors

Heat extraction from the active region in III-V devices is significantly impeded by the thermal conductivity (K) of InGaAs, a meager 0.05W/cm·K, which is a factor of 30 times less than the thermal conductivity of Silicon (1.5W/cm·K). Replacement of III-V materials with a Silicon-Germanium (SiGe) alloy substantially improves the thermal dissipation of these devices. Increasing the thermal conductivity of the active region allows operation at higher output power before thermal failure, which enhances the usefulness of these devices for high-power applications.

One method of cost reduction is to ensure the use of mature CMOS technology in all steps of the fabrication process. Jalali³⁷ points out, “Cost reduction is a critical factor to ensure increasing market for optical data links. One way to achieve this is by use of the mature silicon technology in place of conventional III-V photonics.” Exotic III-V materials cannot be grown directly on Silicon due to a strong mismatch in the lattice parameter of both materials.

The III-V hybrid layer of the AlGaInAs-silicon evanescent laser is attached to the Silicon-on-insulator (SOI) substrate using a wafer bonding process at low temperature. Fang²⁰ writes, “The thermal expansion coefficient mismatch between Si K and InP K can introduce cracks in the III–V layers when bonding temperatures above 300 °C are used. Oxygen plasma assisted bonding is used to die attach the III-V layers to the silicon substrate, Fang²⁰ explains, which generates a thin oxide layer at the bonding interface. This oxide is transparent to light at 1550nm, and therefore does not interfere with the transmission of the optical signal from the III-V active region to the Silicon waveguide.

Direct epitaxial growth of III-V compounds on a Silicon substrate is unacceptable due to the mismatch in lattice atomic spacing between the two materials. The lattice parameter mismatch between III-V materials and Si leads to strain induced dislocation defects in the crystal lattice, which hinders device performance. However, the lattice parameter of SiGe is only slightly different from that of intrinsic Si, which makes sequential growth of these epitaxial layers possible; this is referred to as strained silicon.

Devices made of strained silicon benefit from increased electron mobility due to the stretched atomic spacing in the silicon lattice introduced by the surrounding layers of silicon-germanium. These electrons have fewer collisions, which results in higher mobility. This behavior can be explained using electron orbital theory. Celeste Biever⁴⁷ writes,

In ordinary silicon, all six orbitals have the same energy so there is no preferred direction of flow. But stretching the lattice decreases the energy of the two orbitals in that direction, letting electrons flow more easily along the aligned orbitals. Similarly, squeezing the lattice lets positive charges to flow more easily.

A diagram showing strained Silicon, courtesy of the International Business Machines (IBM³⁰) Corporation, is shown in figure 2.5:

Figure 2.5: Growth of Silicon-Germanium (SiGe) alloy on a silicon substrate produces “strained” silicon (Source: IBM³⁰)

Although hybrid laser sources such as the Intel-UCSB hybrid AlGaInAs-silicon evanescent laser still require III-V materials, this is not the case for SiGe-based photodiodes. Jalali³⁷ agrees, “Although the realization of practical light emitting diodes is impeded by the indirect energy bandgap of silicon, useful optical detection may be achieved. SiGe alloys are also advantageous as an optical absorber material; Jalali³⁷ writes, “the absorption coefficient can be extended to longer wavelengths by use of GeSi alloys.” Pure Germanium is not grown on top of Silicon because of the 4% lattice parameter mismatch between Silicon ($5.43\text{\AA}$) and Germanium ($5.64\text{\AA}$). Epitaxial growth of pure Germanium on Silicon would result in the introduction of dislocation defects into the crystal lattice, which relieves the strain between the two adjacent layers. When considered within the context of a SiGe photodetector, Tao Yin⁷³ explains, “These defects can lead to increased dark current which can degrade receiver sensitivity and reliability if the processing is not carefully handled.”

Replacement of III-V materials with SiGe has thermal and economic benefits. Thermal conductivity of the active region is increased by as much as a factor of 30. Although inappropriate for

laser sources due to an indirect energy bandgap, SiGe is a viable material for photodetectors. Cost is reduced by eliminating the III-V to Si bonding step, which is necessary due to the lattice parameter mismatch. Jalali³⁷ explains, “An all-Silicon receiver... represents a cost reduction as well as an enhancement in reliability compared to a conventional hybrid receiver consisting of a III-V pin photodetector and Si electronics.” Unlike III-V devices, monolithic integration of SiGe photodetectors on a Silicon substrate is possible. Reduced cost, improved thermal properties, and monolithic integration indicate that the SiGe-based photodetectors are a scalable and commercially-viable receiver candidate for optical interconnection applications. According to Davidson¹⁵, “The demand for gigabit-speed fiberoptic communication and the rising popularity of fiberoptics in RF/microwave systems have led to many advancements in high-speed photodetector technology.”

2.5 The Need for High-Speed, High-Power Photodiodes

Photodetector design is an important research area for improving the performance of optical links. Anand Ramaswamy⁵¹ points out, “High-performance analog optical links require photodiodes (PDs) that have high power handling capability as well as high linearity.” M. Saif Islam³⁴ proposes that logic circuits could be directly driven by such a high-power photodetector, which eliminates the need for high-bandwidth post amplifiers. Islam³⁴ writes, “In digital applications, high-speed and high-power detectors may become important for 40Gbps and beyond systems. Driving the decision circuit of a photo-receiver directly using a high power photodiode without post amplifiers for 40Gbps bandwidth applications could be a potential application.” A block diagram of the photodetector and transimpedance post amplifier is shown in figure 2.6:

Figure 2.6: Block diagram of an optical receiver (Source: Jalali³⁷)

Elimination of the post amplifier offers several improvements in optical link performance. Impedance mismatch between the output of the photodetector and input of the post amplifier causes reflections, which appear as double traces in an eye diagram of the amplifier output. Yin⁷³ illustrates such an eye diagram in the figure shown in figure 2.7:

Figure 2.7: Eye diagram of the output of an optical link. (Source: Yin⁷³)

The transimpedance amplifier is also a significant noise source. Jalali³⁷ writes, “the front end amplifier noise is dominated by the thermal noise of the feedback resistor at low bit rates and by the collector current shot noise at high bit rates.” Use of photodetectors with high optical saturation powers, Andrew Davidson and Robert Marsland¹⁵ explain, “allow analog fiber links to operate with less RF insertion loss, less noise, and larger spur-free dynamic ranges.” Islam³⁴ agrees, “In an externally modulated fiber optic link, high-speed photodetectors (PDs) with high saturation photocurrent can improve the overall link performance, including the link gain, noise figure, and spurious free dynamic range.” RF transimpedance post amplifiers are also expensive to integrate. Its removal makes these optical receivers more economical.

With the elimination of the transimpedance amplifier and the inclusion of a photodetector of sufficiently high saturation power, Islam³⁴ writes, “the performance is limited by the relative intensity noise (RIN) of the laser source and the amplified spontaneous emission noise (ASE) from erbium-doped fiber amplifiers (EDFA).” Furthermore, Islam³⁴ adds, “laser RIN and EDFA-added noise can be

suppressed by balanced receivers.” Clearly, high-speed data links could benefit greatly by the development of an improved high power photodetector.

2.6 Monolithic Integration of Photodetectors on Silicon

Monolithic integration of photodetectors on a silicon substrate is a necessary step toward the realization of low cost optical interconnects. Silicon is effectively transparent for wavelengths greater than 1100nm, which makes it an appropriate low-loss waveguide material at the communications wavelengths of 1300nm and 1550nm. Germanium is also an excellent absorption material with an absorption coefficient of 460 cm^{-1} at 1550 nm. Silicon is non-absorbing at 1550nm, which makes it a good transmission medium at this wavelength. Edward D. Palik⁴⁸ outlines the absorption coefficient versus wavelength for several materials (Si, GaAs, InP, Ge, and InGaAs), which is shown in the figure on the following page:

Figure 2.8: Absorption coefficient for selected semiconductor materials (Source: Palik⁴⁸)

SiGe is sometimes chosen as an absorption material for two reasons. First, the 4% lattice parameter mismatch between pure Silicon and Germanium causes dislocation defects at the boundary between the two materials. Second, the absorption coefficient of Ge can be extended to longer

wavelengths by increasing the percentage of Si in the absorber material. This strained silicon absorber, according to Jalali³⁷, has a sufficiently high absorption coefficient while maintaining a small lattice parameter mismatch between the absorber and waveguide material that is less than 4%. It is this low lattice parameter mismatch that makes the combination of SiGe absorber and Si waveguide compatible with standard CMOS processes. A pure Ge absorber grown on top of a Si waveguide, according to Yin⁷³, would result in dislocations in the crystal lattice that relieve stresses caused by the parameter mismatch. Yin⁷³ writes, “These defects can lead to increased dark current which can degrade receiver sensitivity and reliability if the processing is not carefully handled.” To mitigate these defects in Ge-on-Si epitaxy, Yin⁷³ explains, a thin 0.1 μm thick buffer layer of intrinsic Ge is grown directly on Si at low temperature. Further growth of Ge on top of this buffer layer is then conducted at temperatures generally around 700°C.

The absorption coefficient of $\text{Ge}_x\text{Si}_{1-x}$ is dependent on the germanium concentration (x) in the SiGe alloy, and resides between that of pure Si and Ge, as shown in the figure 2.9:

Figure 2.9: Absorption coefficient for $\text{Ge}_x\text{Si}_{1-x}$ alloys (Source: Jalali³⁷)

The transparent Silicon waveguide allows 1550nm light to travel down the waveguide to the opaque SiGe absorber, where it decays evanescently with increasing depth into the absorbing material. The absorber converts this energy into an electrical current, which then serves as the output of the photodetector. The waveguide material should have a high index of refraction to ensure confinement of the optical signal. Surrounding layers should consist of a lower index of refraction to act as the waveguide cladding. In Silicon devices, this cladding usually consists of Silicon Oxide, which has an index of refraction of 1.444 at 1550nm. Pure Silicon has an index of refraction nearly twice as high: 3.476 at 1550nm. Pure Germanium has an index of refraction of 4.36 at 1550nm, which is higher than that of pure Silicon. The refractive index for $\text{Ge}_x\text{Si}_{1-x}$ at 1550nm is expected to be between that of pure Silicon (3.476) and pure Germanium (4.36), with a value dependent on the atomic composition (x) of Germanium in the SiGe alloy. This refractive index, according to Jalali³⁷, is obtained by a linear interpolation between the indices of pure Silicon (n_{Si}) and pure Germanium (n_{Ge}) based on the lattice percentage of Ge (x). The high refractive index of Silicon gives it excellent light routing capabilities.

In addition to the excellent light coupling properties of pure-Ge, its absorption coefficient is also desirably high at 1550nm. In fact, the absorption coefficient of pure-Ge is higher than that of any SiGe alloy at this wavelength, which justifies its use as an absorber material despite the risk of additional dislocation defects. “Strained silicon” results if the Ge layer is kept thin, which eliminates these defects. It is for these reasons that recent advances in vertical P-I-N rib waveguide photodiodes consist of a thin, intrinsic Ge absorption layer on top of a doped Si waveguide.

3. Performance Limitations in High Power Photodiodes

Photodetectors of all types suffer from similar limitations, including the following: dark current, noise sources, series resistance, resistive heating, thermal runaway, space charge effects, transit time, recombination lifetime, and RC bandwidth limitations. This section serves as a brief overview of these limitations.

3.1 Dark Current

When a diode is reverse biased, charge carriers are emptied from the junction area and a space charge layer, also known as the depletion region, results. The depletion region is highly resistive, preventing most current flow under non-illuminated conditions except for a small leakage current called the dark current.

Dark current is undesirable for several reasons. First, it sets a current floor that always flows through the device, regardless of illumination conditions. This current is in the nano-ampere (nA) range when operated at a low reverse bias and typically increases to a few micro-amperes (μA) for moderate bias. The dark current is also device dependent and increases with the number of lattice defects present in the Ge active region. Ideally, this current is approximately equal to the saturation current (I_s) shown in the diode equation:

$$I = I_s(e^{Vq/nkT} - 1)$$

Second, because the dark current is always present, its associated shot noise will also be present.

Therefore, the dark current acts as a constant noise source that is independent of illumination level.

3.2 Noise Sources

Noise is an important consideration in detectors since these signals are often processed using a post amplifier. These amplifiers amplify both signal and noise alike, resulting in unacceptable levels of noise output power. The effect of noise on the signal to noise ratio (SNR) can be clearly seen in figure 3.1:

Figure 3.1: Visible representation of increasing signal to noise ratio (Source: Razeghi⁵²)

Thermal noise, or Johnson noise, is bandwidth and temperature dependent. Johnson noise, according to Manijeh Razeghi⁵², results from the random motion of thermally-generated electron-hole pairs within the semiconductor device. It is modeled as a white noise source and has the following equation in terms of Boltzmann's constant, k_b , temperature, T , and bandwidth, Δf :

$$P_{Johnson} = k_b T \Delta f$$

Another significant noise source is shot noise, which is dependent on device bandwidth, time-average current level, and load impedance. This noise, Razeghi⁵² points out, arises due to the quantization of charge carriers entering and leaving the device. It has the following equation, where q is the Coulomb constant, I_{DC} is the time-average current, R is the load impedance, and Δf is the bandwidth of the device:

$$P_{shot} = 2qI_{DC}R\Delta f$$

When operated in high reverse bias, generation-recombination (G-R) noise becomes significant. For a nearly intrinsic semiconductor with a moderate bias, Razeghi⁵² writes, this noise is the following, where V_b is the bias voltage, l is the detector length, w is the detector width, t is the absorber thickness, b is the ratio of mobilities for electrons and holes (μ_e/μ_h), n is the electron concentration, p is the hole concentration, τ is the carrier lifetime, ω is the angular frequency of operation, R is the load impedance, and Δf is the bandwidth of the device:

$$P_{G-R} = \frac{\left[\frac{2V_b}{(lwt)^{1/2}} \left(\frac{1+b}{bn+p} \right) \left(\frac{np}{n+p} \frac{\tau \Delta f}{1+\omega^2 \tau^2} \right)^{1/2} \right]^2}{R}$$

G-R noise tends to decrease at higher modulation frequency and for larger load impedances. However, this noise increases at higher bias voltage and larger measurement bandwidth.

At high reverse bias, the diode enters the avalanche performance region. Operating near avalanche breakdown allows greater detector gain, but at the expense of increased noise level due to the avalanche process. Sergio Franco²¹, author of *Design with Operational Amplifiers and Integrated Circuits*, points out that shot noise and avalanche noise both model current spikes, but for different reasons. Shot noise occurs due to the quantization of current, but avalanche noise occurs due to the cascaded electron-hole pair generation that results from high energy carrier collision with atoms in the crystal lattice. According to Franco²¹, the avalanche process is notoriously noisy, and usually dominates other noise sources, including shot noise. Avalanche breakdown and avalanche noise is discussed in more detail in sections 3.2 and 4.3.

Temperature noise, according to Razeghi⁵², is an important consideration for temperature sensitive devices. It occurs from temperature fluctuations in the environment and from ohmic heating during brief episodes of high photocurrent. A thermoelectric cooler (TEC) can be used to stabilize device temperature and reduce the temperature noise.

At low modulation frequencies, $1/f$ noise may become dominant over other noise sources. According to Razeghi⁵², $1/f$ noise results from carrier trapping and re-emission from defects at the contact pads and along the crystal surface. The power of this noise source is proportional to the inverse of the operating frequency, which is where its name is derived.

The last noise source of significance is the so-called photon noise. This noise results from the quantization of photons, which subsequently results in the quantization of photocurrent. This noise source is actually the shot noise stated in terms of photon flux. Razeghi⁵² illustrates an expression for photon

noise using the following equation, which is dependent on the photon flux Φ_{photon} , absorber area A , frequency of operation ν , bandwidth Δf , quantum efficiency η , and Planck's constant h :

$$(P_{NEP})_{photon} = h\nu \left[\frac{2A\Phi_{photon}\Delta f}{\eta} \right]^{1/2}$$

Photon noise becomes significant at higher optical power since the photon flux is increased. The absorber cross-sectional area also impacts the photon noise. Increased quantum efficiency allows more photons to be absorbed from the available flux, which diminishes the effect of quantization.

3.3 High Optical Power Considerations

Under high optical power conditions, the series resistance of the photodiode plays a major role in determining the cutoff behavior of the device. It is responsible for starving the diode of bias voltage when operating at high photocurrents. This series resistance, R_s , is best understood by observing the complete photodiode model, illustrated by Kai Chang¹⁰ in his book titled *Microwave Solid-State Circuits and Applications*, which is shown below:

Figure 3.2: Discrete component diode model (Source: Chang¹⁰)

When operated under reverse bias and no illumination, the junction region becomes completely devoid of charge carriers, and subsequently the dynamic junction resistance $R_j(V)$ becomes very large. If the absorption region is then illuminated, the dynamic resistance adjusts to allow current to flow.

If the optical signal does not vary with respect to time, then the junction capacitance $C_j(V)$ and series inductance L_s may be ignored. These components are important when considering the bandwidth and input impedance of the device.

The series resistance results from the ohmic resistance of the metal contact pads and contact diffusion layer, and due to the resistivity of the semiconductor layers leading up to the active region. Small series resistance is desirable, which is usually on the order of a few ohms and is device dependant.

Although small, the series resistance can limit high power operation when the photocurrent is very high. Illumination by high optical power creates a large potential drop across this series resistance that can negate the external reverse bias. When the edge of conduction (EOC) is reached, the diode can no longer produce any additional photocurrent.

High levels of photocurrent also have a detrimental effect on detector performance that is independent of the series resistance. This current consists of photogenerated electron-hole pairs, which move apart under the influence of an external bias electric field. Holes move toward the anode, and electrons move toward the cathode. Therefore, an electric field is setup by the photocarriers themselves, which opposes the bias electric field. Eventually, at high photocurrent levels, these carriers completely shield the junction and the bias electric field collapses. This phenomenon is called the space charge effect. The space charge effect is illustrated in figure 3.3 for a P-I-N photodiode:

Figure 3.3: Space charge effect shown by the space charge electric field negating part of the bias electric field, which reduces the overall net electric field across the intrinsic region

Efforts to reduce the space charge effect can be seen in the uni-traveling carrier³⁶ (UTC) and partially depleted absorber⁵⁹ (PDA) structures. UTC and PDA are discussed in sections 4.8 and 4.7, respectively.

Heat is also an important consideration. Ohmic heating through the series resistance can result in thermal runaway, which eventually leads to thermal failure. Thermal runaway occurs because of the thermionic generation of carriers in the depletion region. As temperature in the device rises, the thermal generation of electron-hole pairs is increased. These thermally-generated charge carriers in turn provide their own contribution to the current, which raises the total device current ever-so-slightly. Because more current is flowing through the device, ohmic heat generation is increased current rises yet again. The result is thermal runaway, which is a positive feedback loop that causes the temperature and current to rise simultaneously until thermal failure results.

3.4 Bandwidth Limitations

Bandwidth is an important consideration in optical communication links because it determines the maximum speed that data can be transmitted. Three factors limit the bandwidth in photodetectors: the transit time of carriers through the absorption material, the recombination lifetime of photocarriers, and the RC time constant of the device. This section addresses these limitations in detail.

The most fundamental bandwidth limitation is transit time. Electrons and holes each have different mobilities in the absorber material, and therefore move at different rates. The slowest carrier velocity determines the time necessary to clear the absorbing region of carriers, which is the definition of transit time. Electrons are usually faster than holes, and absorber structures have been developed that block the movement of the holes. Uni-traveling carrier (UTC) absorbers reduce transit time by blocking the movement of holes using a quantum well heterostructure. UTC photodiodes are discussed in section 4.8. Another method is to simply make the absorbing region thinner. However, a thinner absorption region decreases the volume available to absorb photons, which subsequently reduces the optical responsivity. Nevertheless, it is possible to design an absorber thickness that reduces transit time without impacting the responsivity. The concentration of photons decreases exponentially with increasing

absorber depth. Therefore, most of the incident light can be captured, while minimizing transit time, if the absorber thickness is carefully selected.

Recombination lifetime also plays a role in determining the bandwidth. As charge carriers traverse the space charge region, a small possibility exists that these carriers can recombine. In direct band-to-band recombination, the electron and hole annihilate each other to produce a photon. Recombination in indirect bandgap materials, such as Germanium, is slightly different. This process usually dissipates excess energy through the release of a phonon, or crystal lattice vibration. Charge carriers must have a sufficient recombination lifetime to traverse the space charge region and contribute to the current at the contact terminals. The absorber thickness should be kept thin to allow a sufficient number of charge carriers to reach the contact terminals without recombining.

The RC time constant is another factor limiting device bandwidth. The device capacitance consists of capacitance due to the contact pads and absorber cross sectional area, and due to the junction capacitance under reverse bias. These two sources are arranged in parallel, which sums their contribution to the total capacitance. Sources of resistance include the diode series resistance, contact resistance, and load resistance. Each is connected in series, which sums their contribution to the total resistance. This forms a low pass filter with break frequency equal to the following:

$$\omega_c = \frac{1}{RC}$$

Reducing device capacitance is of paramount importance if a high bandwidth is desired. This can be accomplished by reducing the cross sectional area of the absorber region. Bandwidth is also improved though the reduction of series resistance.

Although all of these factors contribute to the overall bandwidth, one factor is usually dominant over the others. Devices with large absorber cross sectional area are likely to be RC limited if the absorber thickness is thin enough.

3.5 Quantum Efficiency and Responsivity

Since a single photogeneration event requires the absorption of a single photon, the amount of current a photodiode can produce is limited by the total photon flux coupled into the absorption region of the device. According to Razeghi⁵², photodiode responsivity (R_i) is proportional to the quantum efficiency (η), wavelength (λ), electron charge (q), Plank's constant (h), and the speed of light (c) through the following equation:

$$R_i = \frac{\eta q \lambda}{hc}$$

The ideal responsivity has a sharp cutoff at the wavelength corresponding to an energy (hc/λ) equal to the bandgap of the active region material. According to Razeghi⁵², sources of non-ideality in the device responsivity include the following:

- Incomplete absorption of incident photons
- Frequency-dependent optical losses
- Surface traps and other recombination mechanisms
- Photogeneration of carriers in areas outside of the active region

An upper limit is placed upon the obtainable responsivity by the wavelength of operation assuming a quantum efficiency of 100%. At 1550nm, the maximum obtainable responsivity is 1.25A/W.

4. Comparison of Absorber Structure Types

Photodetectors with active regions consisting of III-V semiconductor material or SiGe alloys have been demonstrated using a variety of structures. Light coupling into the active region is accomplished by means of a surface-normal or lateral illumination, and conversion of this coupled light into an electrical signal is achieved using one of many absorbing region types. This section outlines many of the absorber structures that are used in photodetectors, including the following: conventional P-I-N, avalanche photodiodes (APD), metal-semiconductor-metal (MSM) photodiodes, type-II superlattice photodetectors, partially depleted absorber (PDA), resonant cavity enhanced (RCE) absorber, uni-traveling carrier (UTC) absorber, quantum well intersubband photodetectors (QWIP), travelling wave photodiodes (TWPD), and velocity-matched distributed photodetectors (VMDP).

4.1 Lateral versus Surface-Normal Photodetectors

The simplest and most fundamental photodetector type is the surface-illuminated photodetector. Light is normally incident on the absorber surface, which is then converted into an electrical current. Incident light intensity (W/m^2) decays evanescently as it passes through the active region, which causes the generation of electron and hole pairs.

Surface-normal photodetectors exhibit a bandwidth-efficiency tradeoff due to the thickness and cross-sectional area of the absorbing region. The absorber thickness and carrier mobility determine the transit time (TT), which is reduced by making the absorber region as thin as possible. However, since the incident radiation evanescently decays as it transits through the absorber material, reducing the absorber thickness denies some of this radiation the opportunity to be absorbed. Therefore, efforts to improve the transit time of surface-normal devices by making the absorption region thinner inherently cause the detection efficiency to decrease. Furthermore, efforts to increase the detector cross-sectional area will introduce additional device capacitance, which increases the RC time constant and decreases bandwidth. These two factors produce a bandwidth-efficiency tradeoff in surface-normal detectors.

Comparison of Absorber Structure Types

Laterally-illuminated photodetectors break the bandwidth- efficiency tradeoff found in surface-normal detectors and are easier to monolithically integrate. In addition, Yin⁷³ states, the absorption coefficient of pure Ge is not large enough to sustain the high-bandwidth and high-efficiency requirements found in modern optical communication technologies. Yin⁷³ explains, “The most obvious way around this problem is to use a waveguide-based detector which breaks the trade-off between bandwidth and responsivity present in surface normal detectors, and is much easier to integrate with other types of devices.” Laterally illuminated photodetectors are rectangular and are typically positioned directly above a waveguide composed of a material with similar refractive index to the absorber. Wu⁷¹ contrasts lateral versus surface-normal illumination in figure 4.1 below:

Figure 4.1: Comparison between (a) surface-normal and (b) lateral photodetectors
(Source: Wu⁷¹)

In the lateral structure, light evanescently couples into the absorber region from the waveguide, which produces an electrical current as the light is absorbed. The waveguide photodetector breaks the bandwidth-efficiency tradeoff by reducing the absorber thickness to decrease transit time and boost device bandwidth while increasing the absorber length to allow more light to be absorbed and improve device efficiency. This simultaneous improvement in both transit time and bandwidth presents an enormous advantage of lateral illumination over surface-normal structures. Yin⁷³ agrees with this assertion:

By increasing the detector length, the responsivity of Ge waveguide detectors can be greatly increased at 1550nm even though the material absorption coefficient is small at

this wavelength. The detector speed can also be simultaneously optimized by using a thin Ge layer to reduce transit time limitations. The overall reduction in area and capacitance in going from a normal incident to a waveguide-based photodetector helps to improve receiver sensitivity and enable greater than 20 GHz operation which will be critical in the increasingly important application of short distance communications.

The lateral absorber must have a sufficient length to absorb the entire amount of incident light for maximum optical to electrical conversion efficiency. Coupling from the waveguide to absorption region as well as the absorption coefficient determine the required length for this condition. A silicon waveguide photodetector with an offset quantum well structure has been demonstrated by Park et al.⁴⁹ to exhibit a quantum well coupling coefficient of 2-4% from the silicon waveguide to the absorber region. Yin⁷³ argues that the coupling coefficient can be increased by reducing the waveguide height, but at the expense of greater polarization dependence:

Coupling to the Ge layer can be improved by optimizing Ge layer thickness and reducing the rib waveguide height, allowing for even shorter detectors. However the trade-off for reducing the rib height is increased polarization sensitivity, which would reduce the viability of such waveguide detectors.

The coupling coefficient between the waveguide and active region is dramatically higher for absorbers positioned toward the center of the waveguide cross section. Davidson¹⁵ reports a coupling coefficient as high as 88%, with a theoretical limit of 98%. Nevertheless, the low coupling coefficient shown by Park⁴⁹ (2-4%) is overcome by increasing the detector length. Park⁴⁹ writes, “Ge waveguide detectors with a width of 4.4 μ m and length of 250 μ m exhibit no further increase in photocurrent compared to their 100 μ m long counterparts, indicating that virtually all the light was absorbed within the first 100 μ m of Ge.” Jalali³⁷ provides a mathematical expression for the transverse optical confinement factor, which is shown below:

$$\Gamma_T = \frac{\int_0^h |E(y)|^2 dy}{\int_{-\infty}^{\infty} |E(y)|^2 dy} \approx \frac{V^2}{2 + V^2}$$

$$V = \frac{2\pi}{\lambda} h(n_{SL}^2 - n_s^2)^{1/2}$$

In this equation, V is the normalized frequency and film thickness, h is the thickness of the absorber core layer, n_{SL} is the average refractive index of the absorber structure, and n_s is the refractive index for the substrate layer. This equation points out several interesting properties for optically-symmetric layer structures. The confinement factor increases with greater absorber height, shorter wavelength, and greater index of refraction in the absorber material. The transverse optical confinement factor approaches unity for large V . These equations also give the external quantum efficiency, Jalali³⁷ points out, yielding the following expression:

$$\eta_{ext} = C(1 - R)(1 - e^{-\alpha_0 r \Gamma_T L})$$

Where R is the reflectivity of the waveguide front facet, C is the fiber-to-waveguide coupling efficiency, α_0 is the bulk absorption coefficient of the absorber, r is the quantum well duty cycle (set to 1 for non-quantum well absorbers), and L is the detector length. The fiber-to-waveguide coupling efficiency, C , according to Jalali³⁷, is given by the overlap integral of the waveguide and fiber transverse electric fields, which is mathematically expressed by the following equation:

$$C = \left[\iint_{-\infty}^{\infty} E_g(x, y) \cdot E_f(x, y) dx dy \right]^2$$

Another feature of waveguide photodetectors is the ability to operate at lower bias voltage due to the thinner depletion region. The bias electric field is strengthened by a thinner depletion region.

Davidson¹⁵ explains the advantage this gives lateral detectors over surface-normal detectors:

Because the depletion layer can be made thinner, designs requiring a lower voltage to deplete the absorbing layer are possible. This is especially helpful in situations where a long absorption length would otherwise require thick depletion layers and high bias voltages. Low voltage operation can be important for battery-operated devices.

Waveguide photodetectors, due to their high bandwidth, high responsivity, and ease of monolithic integration, are superior to surface-normal detectors in many applications, and they are the candidate of choice for use in modern optical data links.

4.2 Conventional P-I-N

Conventional P-I-N photodiodes consist of an intrinsic, undoped silicon region sandwiched between p- and n-doped silicon regions. Photogenerated carriers are produced when light excites silicon atoms in the intrinsic region (i-layer), producing electron hole pairs, which drift toward the n- and p-doped regions. Electrons and holes diffuse into the p- and n-doped regions from the external terminals to maintain charge neutrality. The structure of a conventional P-I-N photodiode is shown in figure 4.2:

Figure: 4.2: P-I-N photodiode (a) doping profile and (b) band diagram (Source: Razeghi⁵²)

Razeghi⁵² writes, “Because of the very low density of free carriers in the i-region and its high resistivity, any applied bias is dropped almost entirely across this i-layer, which is then fully depleted at low reverse bias.” Therefore, Razeghi⁵² argues, the depletion layer width is controlled by the size of the intrinsic region, “which can be tailored to meet the requirements of photoresponse and frequency bandwidth.”

Practical applications of P-I-N photodiodes require an opening in the ohmic contacts for light to pass through and strike the intrinsic layer, which is the active region of the photodetector. Top-illumination and back-illumination are two surface-normal methods can be used. Both methods are shown in figure 4.3 below. Back-side illumination, according to Razeghi⁵², along with total depletion of the intrinsic layer, and especially the use of a low reverse bias, is “important for digital operation and for a low-noise performance.” The latter condition may be important in avoiding cascaded avalanche carrier generation, which is known to produce tremendous noise.

Figure 4.3: Surface Illuminated P-I-N photodiode showing (a) top-illumination and (b) bottom-illumination. (Source: Razeghi⁵²)

The response speed and bandwidth of P-I-N photodetectors are limited by the transit time and the resistance and capacitance of the surrounding external circuit. The transit time is determined by the width of the intrinsic layer and velocity of carriers within. Bandwidth can be improved by shortening the width of the i-layer, but doing so reduces the amount of light that is absorbed and consequently lowers the device responsivity.

4.3 Avalanche Photodiodes (APD)

Avalanche Photodiodes (APD) operate through the cascaded ionization of electron-hole carriers. These diodes must be strongly reverse-biased to ensure a large potential drop across the P-N junction. As

Comparison of Absorber Structure Types

light strikes the junction region, electron-hole pairs are generated. Electrons are excited by the electric field across the space charge region and accelerate toward the n-type layer; likewise, holes accelerate toward the p-type layer. This acceleration is constantly interrupted by collisions within the crystal lattice, which decreases the net acceleration of these carriers. The rate of lattice collisions becomes greater as the carrier velocity increases. The result is an average saturation velocity, which has a kinetic energy that obeys the following equation:

$$KE = \frac{1}{2}mv^2$$

Razeghi⁵² writes, “When the electron can gain enough kinetic energy to ionize an atom, it creates a second electron-hole pair. This is called impact ionization.” An energy band diagram of an avalanche photodiode P-N junction is shown in figure 4.4 below:

Figure 4.4: Energy band diagram of an avalanche photodiode P-N junction (Source: Razeghi⁵²)

Point 1 in figure 1.8 shows the initial photogenerated electron-hole pair excitation due to an incident photon of light. The electron is excited toward the n-type layer, where it accelerates to an energy level sufficient to cause impact ionization of an atom in the crystal lattice and creation of another electron-hole pair. Impact ionization can occur many times across the P-N junction, resulting in an “avalanche” of electrons reaching the n-type region. This cascaded generation of carriers from a single excitation event is called avalanche multiplication.

Avalanche photodiodes, Razeghi⁵² points out, have high speed and extreme sensitivity, with the ability to count individual photons. However, since impact ionization is a random process, these devices also exhibit tremendous noise at their output.

All diodes experience increased dark current at very high reverse-bias due to the thermal excitation of electron hole pairs and subsequent avalanche multiplication. This current increases with greater bias until the device reaches avalanche breakdown, which allows a tremendous amount of current to flow under relatively high voltage conditions. The power output from avalanche breakdown may reach levels significant enough to cause thermal failure. Although operation in the avalanche region is required for APD photodiodes, very high reverse-bias is undesirable in other P-N devices due to increased noise from random impact ionization events. For this reason, Razeghi⁵² explains, P-N junction photodetectors are typically operated under low-bias to ensure a thin depletion region and to minimize the possibility of localized uncontrolled avalanches. However, operation at extremely low bias prevents carriers from reaching their saturation velocity, which increases the transit time of the device and ultimately lowers the bandwidth.

4.4 Metal-Semiconductor-Metal (MSM) Photodiodes

A metal-semiconductor-metal (MSM) photodiode consists of two Schottky-barrier junctions across an undoped semiconductor material. The layers of an MSM diode reside on a single plane with the semiconductor material almost completely depleted.

An electric field is established across the semiconductor material by an external bias voltage applied to the metal contacts. The resulting energy band diagram is shown in figure 1.9 below:

Figure 4.5: Metal-semiconductor-metal (MSM) photodiode energy band diagram under the influence of an external bias and showing the photo-excitation of charge carriers (Source: Razeghi⁵²)

As figure 4.5 shows, photo-excitation of electron-hole pairs occur within the depleted semiconductor material. These carriers are separated and drift toward opposite metal contacts under the influence of the bias electric field. Like charges are injected at the opposite metal interface once a photocarrier leaves the semiconductor material in order to maintain overall charge neutrality.

According to Razeghi⁵², the frequency response of MSM photodiodes is limited by the transit time of carriers across the semiconductor layer and by the capacitive charge-up time of the diode. MSM photodiodes, Razeghi⁵² contends, have a low dark current and exhibit a large bandwidth. The planar structure of these photodiodes makes these devices receptive to simple integration schemes.

4.5 Type-II Superlattice Photodiodes

Type-II superlattice heterostructures consist of a PN junction between two materials with a broken energy gap alignment. The type-II superlattice energy band diagram is compared against other types in figure 4.6:

Comparison of Absorber Structure Types

Figure 4.6: Heterostructure energy band lineups for (a) type-I superlattice, (b) type-II superlattice staggered, and (c) type-II superlattice broken gap (Source: Razeghi⁵²)

A broken energy gap, consistent with the type-II superlattice shown in figure 4.6 (c), is used to construct a photodetector from layers of InAs and GaSb. As figure 4.7 shows, the valance band of GaSb is at a higher energy than the conduction band of InAs. Electrons prefer to reside at low energy levels, and the electron wavefunction shows a local maximum within the lower conduction energy band of the InAs layer. Likewise, holes prefer to reside at high energy levels, and the hole wavefunction shows a local maximum within the higher valance energy band of the GaSb layer.

Figure 4.7: Energy band diagram of a type-II superlattice showing the overlapping wavefunctions of the majority carrier in each material (Source: Razeghi⁵²)

The energy band structure, shown in figure 4.7, results in a quantum well structure with finite barrier potential. Razeghi⁵² points out, “Although the electrons and holes are mostly confined in different layers, their wavefunctions can extend into the thin superlattice barriers. As a result, the overlap of the electron and hole wavefunctions is not strictly zero.” Therefore, Razeghi⁵² argues, the active region must

reside at the superlattice intersections where the wavefunctions overlap. Although the very small width of the superlattice barrier limits the optical absorption to the layer interfaces, Razeghi⁵² admits, “The spatial separation of electrons and holes is advantageous in reducing the Auger recombination rate, which results in a longer lifetime of the photogenerated electron-hole pairs.”

Photogenerated carriers quantum-mechanically tunnel through any superlattice barriers that are blocking their path. A single quantum well and its associated energy barrier is called a period. Increasing the number of periods improves the detector sensitivity by enhancing the total number of active regions in the device. However, the addition of more cascaded quantum wells slows the transit time of carriers through the active region because these carriers are trapped for some time in each of the quantum wells. A stacked device consisting of successive, periodic heterojunctions is said to contain multiple quantum wells (MQW), which provides an enhanced absorption area over several stacked active regions.

Type-II superlattice photodiodes, according to Razeghi⁵², exhibit high detectivity at room temperature, with reported wavelengths as long as 32 μm with 50 superlattice periods, mainly due to a reduced Auger recombination rate.

4.6 Quantum Well Intersubband Photodetectors (QWIP)

Quantum well intersubband photodetectors (QWIP) operate through the intersubband transition of carriers in a quantum well. These subbands are shown as discrete energy levels in figure 4.8:

Figure 4.8: Intersubband energy levels in a GaAs-AlGaAs quantum well. Showing a type-I superlattice heterostructure with both valance and conduction subbands. (Source: Razeghi⁵²)

These subbands result from a quantized density of states, $g(E)$, which is shown in figure 4.9:

Figure 4.9: Quantization of the density of states, $g_{2D}(E)$, in a 2-dimensional quantum well. The three-dimensional density of states, $g_{3D}(E)$, is shown as a dashed line. (Source: Razeghi⁵²)

The density of states is quantized because the wavefunction is forced into what is essentially a two-dimensional volume, which limits the allowed wavelength to discrete multiples of the well width for an infinitely strong barrier potential. This is shown in figure 4.10:

Figure 4.10: Allowed energy states in a 2-dimensional quantum well with infinite potential barriers (Source: Razeghi⁵²)

The incorporation of finite barrier potential allows some of the wavefunction, $\Psi(x)$, to leak through, which creates the possibility of quantum mechanical tunneling through the barrier if it is thin enough. A quantum well with finite barrier potential, U_0 , is shown in figure 4.11 below:

Figure 4.11: two-dimensional quantum well (a) conduction band energy diagram showing finite barrier potential and (b) wavefunction showing evanescent tails in the barrier region (Source: Razeghi⁵²)

The number of intersubband energy states can be controlled by varying the potential barrier height and width of the quantum well. Increasing the barrier height allows a greater number of intersubband energy to appear before reaching the conduction band energy of the barrier material, and increasing the quantum well width decreases the energy spacing between subbands. This is illustrated by figure 4.12:

Figure 4.12: Number of intersubband energy levels is controlled by the potential barrier height and the width of the quantum well (Source: Razeghi⁵²)

Carriers that are excited into higher intersubband energy states have a greater chance of tunneling out of the quantum well and contributing to the photocurrent. This process, according to Razeghi⁵², is the reverse of that used by the quantum cascade laser. The energy diagram and intersubband absorption of a photon in a quantum well are illustrated in figure 4.13 below:

Figure 4.13: Active region of a quantum well intersubband photodetector (QWIP) showing (a) an intersubband transition in a wide well and (b) a transition from a quasi-bound state to the continuum conduction state in a narrow well. (Source: Razeghi⁵²)

Quantum wells can be made narrow enough to exhibit only a single subband energy level. Excitation of an electron in this quasi-bound energy state, as shown in figure 4.13(b), will elevate it to the continuum conduction state. As the electron propagates through the MQW region, it may be trapped and reemitted by the quantum wells in its path. This process is shown in figure 4.14 for a SiGe/Si multiple quantum well structure:

Figure 4.14: Propagation of a photogenerated carrier through multiple periods in a multiple quantum well region (Source: Liu⁴⁴)

The intersubband transition shown in figure 4.13(a) requires quantum tunneling to traverse any quantum well potential barriers in its path. According to Razeghi⁵², “For photoconductive devices utilizing intersubband absorption, the photogenerated carriers travel out of the quantum wells and contribute to the photocurrent.”

“QWIPs,” Razeghi⁵² reasons, “have a narrow absorption spectrum that can be tailored to match optical transitions in the 3~20 μm wavelength range by adjusting the quantum well width and barrier height or barrier layer composition.” More importantly, Razeghi⁵² points out, QWIPs “can be made using mature III-V semiconductors based on gallium arsenide (GaAs) or indium phosphide (InP) substrates.”

The quantum well-based active region and surrounding waveguide layer may be enclosed in an undoped, high-bandgap cladding material to confine electrical carriers and optics into separate regions. This separated confinement heterostructure (SCH), Razeghi⁵² explains, helps reduce carrier absorption in the waveguide layer and helps increase the internal quantum efficiency. Because the cladding material may have a substantially lower index of refraction than the waveguide and multiple quantum well layers,

Comparison of Absorber Structure Types

optical confinement into the active region is improved. Unfortunately, Razeghi⁵² acknowledges, the confinement factor for the active region remains quite small (1~5%). This is due to the small width of the active region when compared with the waveguide and cladding. This is typically overcome by increasing the number of periods in the MQW structure. An MQW/SCH structure composed of GaAs/AlGaAs layers is shown in figure 4.15:

Figure 4.15: An MQW/SCH structure showing an exaggerated active region width. Optical confinement to the active region is typically between 1-5%. (Source: Razeghi⁵²)

However, Jalali³⁷ points out, the number of MQW periods is limited by the equilibrium critical thickness of the strained layers. Jalali³⁷ demonstrates a MQW/SCH device with multiple $\text{Ge}_x\text{Se}_{1-x}$ to Si quantum wells sandwiched between two thick layers of intrinsic Si, which connect to a p^+ contact pad and to an n^+ substrate. This is shown in figure 4.16 below:

Comparison of Absorber Structure Types

Figure 4.16: Structure of a device using Multiple Quantum Wells (MQW) and a Separated Confinement Heterostructure (SCH). 14 periods of $\text{Ge}_x\text{Se}_{1-x}$ to i-Si quantum wells are shown, which are surrounded a $1\mu\text{m}$ thick cladding. (Source: Jalali³⁷)

4.7 Partially Depleted Absorber (PDA)

Partially depleted absorber (PDA) photodiodes show a significant reduction in reducing space charge effects in comparison to the conventional P-I-N photodiode. The doping profile of a PDA photodiode is shown compared alongside a traditional P-I-N photodiode in figure 4.17 below:

Figure 4.17: Comparison of (a) traditional P-I-N and (b) PDA doping profiles. Photogenerated carriers and fixed space charges are shown. (Source: Tulchinsky⁵⁹)

In a conventional P-I-N photodiode, photogenerated carriers are created as electron-hole pairs inside the intrinsic region. These carriers then drift to opposite sides of the photodiode under the electric field influence of the P-N junction built in potential and any applied reverse bias. Electrons drift toward the n-type material and holes drift toward the p-type material. The depletion region is composed of

stationary space charges, with positive charge in the n-type layer and negative charge in the p-type layer.

As photogenerated electrons and holes drift apart in the intrinsic region, an electric field between these charges is created that opposes the electric field of the stationary space charges and external-bias.

Therefore, given sufficient photocurrent, it is possible for these carriers to completely screen out the bias electric field, which, according to Tulchinsky⁵⁹, would result in the loss of the RF signal. This space charge effect, Tulchinsky⁵⁹ explains, is one of the main factors limiting a photodiode's output power since it restricts the optical power a photodiode can withstand before loss of the RF signal. Space charge effects can be mitigated in traditional P-I-N photodiodes by increasing the bias voltage. High reverse-bias voltages are undesirable due to increased power dissipation, dark current, and shot noise associated with the device.

The partially depleted absorber (PDA) structure reduces space-charge effects by injecting extra carriers into the intrinsic layer to reinforce the bias electric field within this region⁵⁹. This is accomplished by extending the doping concentration of the contact layers into the intrinsic region. The doping concentration in the intrinsic layer is tapered, with a depth on either side that is proportionate to the drift velocity of holes and electrons. "In both the InGaAs and InP semiconductor material systems," Tulchinsky⁵⁹ notes, "the velocity of electrons is higher than that of the holes, so the depletion region charge is dominated by slow moving holes." Consequently, InGaAs/InP PDAs have a p-type doping profile that extends further into the intrinsic region than for the n-type doping profile.

In addition to changes in the doping profile, Tulchinsky⁵⁸ notes, "the design of these diodes includes a thinned *i*-layer to further reduce space-charge effects and minimize thermal loading of the depletion layer." Although reducing the width of the *i*-layer reduces the absorption area and results in a lower responsivity, Tulchinsky⁵⁸ writes, "This limitation is mitigated in the PDA photodiode design by using a graded doping of the optically absorbing regions on either side of the *i*-layer to increase optical absorption." Thinning the intrinsic layer also improves the series resistance of the device. Tulchinsky notes, "We believe the extremely large photocurrents generated by these detectors are a direct result of

the low series resistance of these photodiodes. Above a bias of 1-2V, Joule heating (I^2R) limits the saturated photocurrent.”

PDA photodiodes show exceptionally large optical power at high RF frequencies. Photocurrents as high as 153mA at 5GHz and 107mA at 10GHz for a 34 μ m diameter diode have been demonstrated. Tulchinsky⁵⁸ observes, “These photocurrents are five to ten times higher than commercially available detectors at these frequencies.” PDA photodiodes are also competitive with other diode types. In 2005, Tulchinsky⁵⁸ demonstrated a 34 μ m diameter photodiode with a 3dB bandwidth of 7GHz and 23.5dBm of output power, which was the highest reported RF output power from a photodetector at that time.

4.8 Uni-traveling Carrier (UTC)

The space charge effect can be further reduced in photodiodes by restricting the photocurrent to a single charge carrier, which is usually the electron. Islam³⁴ explains, “The overshoot velocity of electrons is one order of magnitude larger than that of the hole and this helps shorten the carrier transit time and reduces the space charge effect which is a common issue in any surface illuminated photodiode.” Restriction of hole movement is achieved by addition of a heterostructure layer that has a large energy level difference in the valance band (ΔE_V) and small energy level difference in the conduction band (ΔE_C). An example of such a heterostructure (figure 4.14) is discussed in the section on QWIP photodetectors (section 4.6).

Uni-traveling carrier (UTC) photodiodes achieve high-speed and high saturation photocurrent simultaneously. A UTC photodiode operating at 1550nm achieved 180mA of photocurrent with a 3dB bandwidth of 65 GHz, and the same device attained an RF output power of 11dBm at 100GHz modulation (qtd. in Islam³⁴).

4.9 Resonant Cavity Enhanced (RCE)

The absorption layer thickness in surface-normal devices is designed around a tradeoff between bandwidth and responsivity. Waveguide structures have broken the bandwidth-responsivity tradeoff by increasing the absorption layer length to capture more of the incident light while keeping the absorption

Comparison of Absorber Structure Types

layer thickness small in order to minimize transit time; however, the waveguide to absorber coupling coefficient is small (2-4%) for absorber structures that are offset from the center of the waveguide and the absorber length must be very long to compensate. The large increase in absorber length, according to Islam³⁴, has a detrimental effect on RF bandwidth because the electrical capacitance is increased. Resonant cavity enhanced (RCE) photodiodes sidestep this limitation by combining high responsivity with small absorber dimensions.

The basic structure of an RCE photodiode consists of a thin absorption layer sandwiched between two reflective regions. The width of the absorption region is matched to a specific wavelength, which creates a resonant cavity. This arrangement is illustrated by Davidson¹⁵, which is shown in figure 4.18:

Figure 4.18: Structure of an RCE photodiode showing a reflective resonant cavity surrounding an absorbing material (Source: Davidson¹⁵)

RCE photodiodes have internal efficiencies that approach 100% absorption of incident light. Davidson¹⁵ writes, “The internal quantum efficiency of the device approaches unity, significantly exceeding that obtained with a single pass through the absorbing layer.” Since the device structure is a Fabry-Perot resonant cavity, the wavelength selectivity of the device is very sharp. Davidson¹⁵ illustrates this selectivity in a plot of detector efficiency versus wavelength for an 850nm photodetector, which is shown in figure 4.19 below:

Figure 4.19: 850nm RCE photodetector showing very high wavelength selectivity (Source: Davidson¹⁵)

This high wavelength selectivity requires a laser source that has a perfectly matched wavelength with the detector. Islam³⁴ writes, “Although quantum efficiency is high for these kinds of devices, they don’t operate in wide frequency ranges. However, for a RF photonic application, this limitation will not affect any system performance if the laser source and the photodiodes’ response wavelength are designed to have a perfect match.” High wavelength selectivity has the added benefit of filtering optical noise such as ASE and RIN.

Another useful feature of RCE photodiodes is their ability to achieve low-voltage device operation. Davidson¹⁵ explains, “Because the depletion layer can be made thinner, designs requiring a lower voltage to deplete the absorbing layer are possible. This is particularly helpful in situations where a long absorption length would otherwise require thick depletion layers and high bias voltages.” Operation at lower voltages, Davidson¹⁵ points out, is important for power conservation in battery-operated devices.

RCE detectors, Davidson¹⁵ notes, come in P-I-N, Schottky, and avalanche photodiode structures. Davidson¹⁵ lists three examples of common materials used by RCE detectors: InP/InGaAs/InAlAs, AlAs/GaAs/Ge, and Si/SiGe. GaAs/AlAs/Ge is an attractive combination due to high contrast in the index of refraction between GaAs (3.5) and AlAs (3), which allows high reflectivity in the Bragg reflectors, Davidson¹⁵ points out. According to Davidson¹⁵, an InGaAs/InP/InAlAs P-I-N RCE photodetector operating at 1550nm achieved a mirror reflectivity of 97% with a peak quantum efficiency of 65%. This

performance, Davidson¹⁵ emphasizes, “represents an enhancement factor of four in responsivity over a conventional P-I-N diode with the same 200-nm InGaAs absorbing layer.

In 2002, Islam³⁴ notes, an RCE photodiode with a linear photocurrent of 5mA and 20GHz bandwidth was reported. Surface-illuminated RCE photodiodes have great potential in reaching high levels of linearity and ultra-fast response, Islam³⁴ argues.

4.10 Travelling-Wave Photodiode (TWPd)

The waveguide photodetector (WGPd) described in section 4.1 is a lumped-element device with a bandwidth limited to the RC time constant of the device and the transit time of carriers in the absorption region. Although the transit time limitation is addressed by decreasing the width of the depletion region, the RC bandwidth limitation remains. The bandwidth is defined as the reciprocal of the RC time constant, where R is the series resistance and:

$$B = \frac{1}{2\pi RC}$$

The capacitance model follows that of a parallel plate capacitor $C = \epsilon A/d$, where ϵ is the dielectric constant, A is the absorber cross sectional area, and d is the width of the absorber layer.

The lumped-element behavior of WGPd, Kirk Giboney²³ explains, arises due to the strong impedance mismatch between the photodetector and the load. According to Giboney²³, “Multiple electrical reflections in the device cause the entire junction area to participate in the response. This is why WGPds are best represented by a lumped element model, and suffer from an RC bandwidth limitation in which the device capacitance is determined by the total junction area.”

Traveling-wave photodiodes (TWPd) overcome this issue by matching the characteristic impedance of the photodetector to the resistance of the load. As a distributed element, Giboney²³ points out, TWPds retain the optical bandwidth, temperature range, efficiency, and integrability of lumped-element WGPds without suffering from the RC bandwidth limitation. The bandwidth limitation in TWPds is independent of device length as long as this length is significantly longer than the absorption length, Islam³⁴ contends. A basic TWPd is illustrated by Giboney²³ in figure 4.20 below:

Figure 4.20: Diagram of a traveling-wave photodiode showing an impedance match to the external load
(Source: Giboney²³)

Bandwidth in TWPDs is limited by the velocity mismatch between the electrical output traveling down the absorber transmission line and the optical wave present in the underlying waveguide, Islam³⁴ explains. As of 2003, Islam³⁴ observes, “a good number of TWPd’s were reported with ultra high bandwidth although none could achieve a good current level to be able to become an effective device in RF photonic links.”

4.11 Velocity-Matched Distributed Photodetectors (VMDP)

Velocity-matched distributed photodetectors (VMDP), Islam³⁴ explains, overcome the bandwidth limitations of TWPDs by slowing the electrical propagation speed to match the speed of propagation in the optical waveguide. Photodetectors are placed along a 50Ω coplanar transmission line to provide periodic capacitance and slow down the electrical signal. By adjusting the length and separation of the photodiodes, it is possible to achieve a perfect match between the propagation speed of the electrical signal to that of the optical signal, thereby removing the velocity-mismatch bandwidth limitation. Davidson¹⁵ explains, “Making the electrical and optical wave velocities equal ensures that the photocurrents add in phase, leading to efficient combining of the RF signals.” Davidson¹⁵ illustrates this concept in figure 4.21 below:

Figure 4.21: Schematic diagram of a VMDP showing electrical transmission line segments (Source: Davidson¹⁵)

The distributed arrangement of VMDPs, Davidson¹⁵ emphasizes, is capable of extending the saturation powers into the 100mW level because the overall absorption volume is made larger than a single device on its own. VMDPs benefit from high bandwidth, large saturation power, and ease of integration simultaneously, which makes these devices very attractive for use in RF photonic links.

A schematic diagram of a VMDP and 50Ω coplanar stripline is illustrated by Islam³⁴ in figure 4.22:

Figure 4.22: VMDP consisting of distributed P-I-N detectors (Source: Islam³⁴)

A distributed balanced photodetector can be constructed from VMDP photodiodes. According to Islam³⁴, the structure of such a device consists of a pair of optical waveguides underneath two arrays of high-speed P-I-N photodiodes distributed along a 50Ω ground-signal-ground (GSG) coplanar stripline.

Comparison of Absorber Structure Types

The spacing of the detectors is velocity-matched to the propagation speed of the optical waveguide.

Islam³⁴ illustrates this structure in figure 4.23 below:

Figure 4.23: Balanced distributed photodetector (Source: Islam³⁴)

At high photocurrents, Wu⁷¹ explains, the receiver noise is dominated by the relative intensity noise (RIN) of the laser source and amplified spontaneous emission (ASE) from the erbium doped fiber amplifier (EDFA). Balanced distributed photodetectors are of significant interest due to their ability to cancel out RIN and ASE, he notes. The noise performance, Wu⁷¹ acknowledges, is limited only by the shot noise of the individual photodetector elements.

Because the number of photons decays exponentially with each photodetector on the waveguide, the first diode experiences the most absorption and therefore the highest photocurrent. This serial technique wastes the performance potential of photodiodes further down the chain, and places the highest power requirements on the first diode. Illuminating the photodiodes in a parallel fashion solves this problem by spreading the optical signal out to all photodiodes evenly. Splitting the optical signal N -times, Islam³⁴ notes, could potentially increase the photocurrent N -fold. He reports that a 1×4 split, parallel-fed, velocity-matched TWPD has achieved a linear photocurrent of 52.2mA with a photocurrent limited only by the output power of the EDFA.

4.12 Survey of High-Power Photodetectors from the Past to the Present

Research into the development of high-power photodiodes has spanned nearly two decades. One of the earliest PIN InGaAs waveguide photodiodes was created by Williams et al.⁶⁵ in 1992, which had a bandwidth of 20GHz and a responsivity of 0.21A/W. Since then, three photodetector types have dominated researchers' focus: partially depleted absorber (PDA), uni-traveling carrier (UTC), and traveling wave photodiode (TWPD). In 2010, a PDA photodiode developed by Itakura et al.⁷⁴ produced 29dBm of electrical RF power with a bandwidth of 5GHz, a compression current of 315mA, and a responsivity of 0.55A/W. UTC photodiodes are typically illuminated through the backside and show a very high bandwidth. In 2010, Shi et al.⁵⁴ demonstrated a UTC photodiode with 10.7dBm RF power at compression, a compression current of 37mA, a responsivity of 0.15A/W, and a bandwidth of 180GHz.

Other photodiode types are also used in high-power optical links. A quantum well intersubband photodetector (QWIP) was developed in 2005 by Cellek et al.⁹, which produced a very high responsivity (2.9A/W). QWIPs typically operate at longer wavelengths due to the relatively small energy level transitions allowed in the intersubband quantum wells. Although QWIPs have high responsivity, they have poor saturation output power. Part et al. reports a QWIP photodetector with a responsivity of 5.7A/W and a saturation photocurrent of 25mA. Chiu et al.¹², in 2009, created a resonant cavity enhanced (RCE) photodiode with 6.4dBm of electrical RF power at 1dB compression, a narrowband operating frequency of 40 GHz, a compression current of 18mA, and a responsivity of 0.8A/W. RCE photodiodes are inherently narrowband since these use a resonant cavity to increase their quantum efficiency. Avalanche photodetectors (APD) are typically not used in high-power photonic links because of their very high gain. APD diodes have applications in single-photon detection, and generally do not compare well with other high-power photodetector types.

Figure 7.24 shows a comparison between the reported bandwidth and RF power at 1dB compression for several photodiodes:

Comparison of Absorber Structure Types

Figure 4.24: Photodiode survey showing the reported 1dB compression RF power versus bandwidth

Table 4.1 shows the reported 1dB compression RF power, bandwidth, compression current, and responsivity for each photodiode surveyed:

Comparison of Absorber Structure Types

Survey of Photodetectors: Past to the Present							
Author	Year	Type	Material	RF Power (dBm)	Bandwidth (GHz)	Photocurrent (mA)	Responsivity (A/W)
Itakura et al. ⁷⁴	2010	PDA	InGaAs/InP	29	5	315	0.55
Shi et al. ⁵⁴	2010	Backside-Illuminated UTC	InGaAs/InP/InAlGaAs	10.7	180	37	0.15
Ang et al. ⁵	2010	PIN, Si Waveguide, i-Ge Absorber	Ge/Si	Not Given	11.3	Not Given	0.92
Xue et al. ⁷²	2009	PIN, Si Waveguide, i-Ge Absorber	Ge/Si	Not Given	13.3	Not Given	0.32
Yin et al. ⁷³	2009	PIN, Si Waveguide, i-Ge Absorber	Ge/Si	14.35	31	70	0.89
Chtioui et al. ¹⁴	2009	Backside-Illuminated UTC	InGaAs/InP	17.5	29	80	0.9
Li et al. ⁴²	2009	Balanced Photodetector	InGaAs/InP	22	11	136	0.82
Beling et al. ⁸	2009	TWPD Array	InGaAs/InP	13	17	114	0.55
Wang et al. ⁶³	2008	PIN, Si Waveguide, i-Ge Absorber	Ge/Si	Not Given	5.5	Not Given	0.29
Park et al. ⁴⁹	2008	MQW/SCH, Si-Waveguide	Si/InP/InGaAsP/AlGaInAs	Not Given	3	25	5.7
Williams et al. ⁶⁷	2008	PDA	InGaAs/InP	26	6	620	Not Given
Wu et al. ⁷¹	2008	Backside-Illuminated UTC	InGaAs/InP	6.28	120	24.6	Not Given
Vivien et al. ⁶²	2008	PIN, Si Waveguide, i-Ge Absorber	Ge/Si	Not Given	42	Not Given	Not Given
Vivien et al. ⁶¹	2007	MSM Ge-on-Si, Si Waveguide	Ge/Si	Not Given	25	Not Given	1.0
Williams et al. ⁶⁹	2007	PDA	InP/InGaAs	29	2	500	Not Given
Wang et al. ⁶⁴	2007	UTC	InGaAs/InP	25	14	130	0.75
Beling et al. ⁷	2007	TWPD	InGaAsP/InP	10	10	27	0.3
Liu et al. ⁴⁵	2006	PIN, Si Waveguide, i-Ge Absorber	Ge/Si	Not Given	30	Not Given	1.0
Chiu et al. ¹²	2006	Resonant Cavity Enhanced	InGaAs/InP	6.4	40	18	0.8
Duan et al. ¹⁹	2006	UTC	InGaAs/InP	24.5	10	150	Not Given
Tulchinsky et al. ⁵⁸	2005	PDA	InP/InGaAs	23.5	7	130	Not Given
Demiguel et al. ¹⁸	2005	PDA	InGaAs/InP	4.5	40	17	Not Given
Cellek et al. ⁹	2005	QWIP	InGaAs/InP	Not Given	Not Given	Not Given	2.9
Pauchard et al. ⁵⁰	2004	PIN, InGaAs on Si	Si/InGaAs/InP	Not Given	11	76.5	0.8
Shi et al. ⁵³	2004	PDA	InGaAs/InP	6.5	40	23	1.01
Umbach et al. ⁶⁰	2001	PIN	InP/InGaAsP/InGaAs	12.5	50	10	0.7

(CONTINUED) Survey of Photodetectors: Past to the Present							
Ito et al. ³⁶	2000	Backside-Illuminated UTC	InGaAs/InP	11	65	180	Not Given
Takeuchi et al. ⁵⁷	2000	PIN, InGaAs Waveguide	InGaAs/InP	Not Given	40	10	0.71
Ito et al. ³⁵	2000	Backside-Illuminated UTC	InGaAs/InP	10	60	20	Not Given
Shimizu et al. ⁵⁵	2000	Backside-Illuminated UTC	InGaAs/InP	9	40	35	0.27
Williams et al. ⁶⁸	1999	Surface-Illuminated PIN	InGaAs/InP	20	1	65	Not Given
Alles et al. ³	1998	TWPD	InGaAs/InGaAsAl	-15	60	Not Given	0.3
Chau et al. ¹¹	1998	VMDP (MSM)	InGaAs/InAlAs/InP	Not Given	18	Not Given	0.424
Lin et al. ⁴³	1997	VMDP (MSM)	AlGaAs/GaAs	Not Given	49	56	Not Given
Giboney et al. ²⁴	1996	TWPD	GaAs/AlGaAs	Not Given	190	Not Given	0.295
Harari et al. ²⁶	1995	PIN, InGaAs Waveguide	InGaAs/InP	5	60	Not Given	Not Given
Jasmin et al. ³⁸	1995	PIN, InGaAs Waveguide	InGaAs/InP	-2	10	8	1
Giboney et al. ²³	1995	TWPD	GaAs/AlGaAs	Not Given	172	Not Given	0.295
Williams et al. ⁶⁵	1992	PIN, InGaAs Waveguide	InGaAs/InP	Not Given	20	Not Given	0.21

Table 4.1: Survey of Selected Photodetectors from 1992-2010

Table 4.1 shows an interesting trend in photodetector design. UTC photodiodes provide the best bandwidth of any photodetector type. UTC photodiodes also offer high output power. The highest electrical RF power for a UTC photodiode is 24.5dBm with a bandwidth of 10GHz (Duan et al.¹⁹). Recent efforts have been made to increase the bandwidth in UTC diodes even further. Shi et al.⁵⁴ (2010) reports a UTC photodiode with an RF power of 10.7dBm with a 3dB bandwidth of 180GHz and responsivity of 0.15A/W. UTC photodiodes exhibit enormous bandwidth performance due to a hole-blocking heterostructure, which only allows electrons to transit the active region. Electrons have a faster mobility than holes, which allows them to transit through the active region faster. Bandwidth can be increased even further by decreasing the cross sectional area of the device. Transit time and RC loading are the primary bandwidth limiters in most photodiodes.

PDA photodiodes also show an interesting trend. In 2010, Itakura et al.⁷⁴ produced a PDA photodiode with 29dBm RF power and 5GHz bandwidth. PDA photodiodes have lower bandwidths than UTC type devices because both holes and electrons participate in the conduction process. PDA type devices overcome space charge effects by infusing the intrinsic region with a tapered doping profile. Space charge effects are the primary factor limiting the maximum obtainable RF power.

TWPDs show inherently low power and relatively poor bandwidth. Beling et al.⁸ created a TWPD array with 13dBm output RF power, 17GHz bandwidth, 114mA compression current, and 0.55A/W responsivity. The purpose of TWPDs is to increase the absorption volume without sacrificing bandwidth. Bandwidth in TWPD, Islam³⁴ writes, “is limited by the optical absorption coefficient and the velocity mismatches between the optical wave and the forward- and reverse-traveling electrical photocurrent waves rather than an RC bandwidth limitation determined by the total junction area.” For this reason, Islam³⁴ concludes, TWPDs can obtain larger bandwidth-efficiency products than what is possible in lumped element photodetectors. In 1996, a TWPD by Giboney et al.²⁴ yielded a 3dB-bandwidth of nearly 190GHz, which is a phenomenal result.

Nearly all the photodiodes described in table 4.1 use III-V materials for the waveguide and absorber. The photodiode produced by Yin et al.⁷³ has the highest bandwidth-efficiency product

Comparison of Absorber Structure Types

(27.3GHz) for any Ge P-I-N photodetector operating at 1550nm as of 2007. Production costs are lowered by the use of Si/Ge, which is compatible with standard CMOS processing. In addition, the thermal conductivity of Silicon is nearly 30 times greater than InGaAs, which improves heat dissipation from the active region and allows operation at higher DC photocurrents. These factors are motivating the transition from III-V to Si/Ge based photodetectors.

5. Silicon-Germanium Evanescent P-I-N Waveguide Photodetector

The structure and performance of the Silicon-Germanium Evanescent Waveguide Photodetector is described by Yin et al.⁷³ in the paper, *31GHz Ge n-i-p Waveguide Photodetectors on Silicon-on-Insulator Substrate*. Ramaswamy et al.⁵¹ further describes this device.

An illustration of this photodetector is shown in figure 5.1 below:

Figure 5.1: Schematic and cross section of the Si-Ge Evanescent P-I-N Photodetector
(Source: Ramaswamy⁵¹)

An intrinsic Ge layer absorbs light that is coupled into it by the doped Si waveguide underneath. This P-I-N photodetector is arranged vertically, with the cathode positioned above the i-Ge region and the anode positioned below. The top of the absorption region is heavily n-doped to provide good ohmic

contact between the pad metallization and the intrinsic Ge layer underneath, which forms the diode cathode. The Si waveguide is p-doped to form the diode anode. The sidewalls of the rib waveguide are not doped past the ground terminals, which provide electrical insulation between devices on adjacent rib waveguides. Heavy p-doping is used at the ground metallization interface to create a low resistance ohmic contact to the underlying waveguide. SiO₂ (oxide) is formed above the waveguide region to provide very high resistance insulation between the ground and signal terminals. The rib waveguide is grown on top of a Silicon-on-Insulator (SOI) wafer, which provides very high resistance between the device and the underlying Silicon bulk.

5.1 Device Performance

These detectors use a Si-rib waveguide structure. Light is evanescently coupled from the waveguide into the Ge absorber where it is converted into an electrical current. Surface normal photodetectors suffer from a bandwidth-responsivity tradeoff. This limitation exists because the size of the absorption region simultaneously controls the responsivity and RC bandwidth of the device.

Waveguide photodiodes break the responsivity-bandwidth tradeoff by incorporating a long lateral absorber length while keeping the thickness of the absorption region small. Decreasing the absorber thickness simultaneously decreases the transit time and absorption efficiency of an infinitesimal segment while increasing the capacitance. Despite this decrease in absorption efficiency over a small segment, the total absorption is nearly 100% because the absorber is very long. This results in higher transit time limited bandwidth without sacrificing responsivity. A thin absorption region improves bandwidth by lowering the transit time of photocarriers moving through the absorber material. In addition, the RC time constant is reduced by decreasing the cross sectional area of the depletion region since this acts like a parallel plate capacitor. According to Yin⁷³, the capacitance of the depletion region is calculated to be 66.7fF at 2V reverse bias with a 0.8μm absorber thickness and a length of 50μm, and the pad capacitance is measured to be 23fF.

Efforts to reduce the device capacitance focus on decreasing the cross sectional area of the absorber. The absorber width is made as small as possible, but the length can extend a great distance—as

much as 1000 μm for these devices. Even though the absorption coefficient of Ge is poor, nearly all of the light is absorbed for devices with a length of 100 μm or greater, Yin⁷³ writes. Ge detectors with 250 μm length exhibit no further increase in responsivity than 100 μm long devices, Yin⁷³ observes. Therefore, Yin⁷³ concludes that receiver sensitivity and bandwidth are simultaneously increased by switching from a surface normal to waveguide structure.

Yin⁷³ reports a very high responsivity (1.16A/W) and bandwidth (29.4GHz) for the 100 μm x 4.4 μm device at 2V reverse bias and 1550nm wavelength with a quantum efficiency of 93%. A device with half the length (50 μm x 4.4 μm) showed a responsivity of 0.89A/W, an optical bandwidth of 31.3 GHz, and a quantum efficiency of 71%. When operating in the photovoltaic mode with no external bias (0V), the bandwidth of the 50 μm and 100 μm detectors decreases to 15.7GHz and 10GHz, respectively. For devices of longer length, the bandwidth is expected to drop. According to Ramaswamy⁵¹, the 500 μm long devices have a 3dB bandwidth as low as 4.38GHz at 5V reverse bias, which remains constant for photocurrents up to 50mA. These bandwidth results are summarized in table 1.9.1 below:

<i>Bandwidth Summary</i>			
Reverse Bias	50μm length	100μm length	500μm length
0V	15.7 GHz	10 GHz	--
2V	31.3 GHz	29.4 GHz	--
5V	--	--	4.38 GHz

Table 5.1: Summary of bandwidth results for various detector lengths and a detector width of 7.4 μm

(Sources: Ramaswamy⁵¹ and Yin⁷³)

An effective figure of merit for high-power photodiodes is the bandwidth-efficiency product. “This bandwidth-efficiency product of 27.3GHz,” Yin⁷³ writes, “is the highest reported for any Ge photodetector at 1550nm.” Communication links require both high speed and high quantum efficiency. This photodiode provides both high bandwidth and high efficiency simultaneously.

Ramaswamy⁵¹ further describes the high power performance of the Si-Ge PIN photodiodes. Power dissipation of 1.003W is accomplished at 125.49mA photocurrent and 8V bias. Such high electrical output power is necessary to drive high frequency loads without the need for a post amplifier stage. Figure 5.2 shows the DC photocurrent obtained for various input optical powers and bias voltages and the responsivity obtained for each case:

Figure 5.2: (a) Photocurrent as a function of input optical power for several bias voltages and (b) DC responsivity for various optical powers and bias voltages (Source: Ramaswamy⁵¹)

Ramaswamy⁵¹ illustrates the DC power and bandwidth performance of these devices in comparison to other recent advances in waveguide and surface-normal photodetectors in figure 5.3:

Figure 5.3: Comparison of several photodetectors by DC electrical power dissipation versus bandwidth. Waveguide photodetectors are shown in red and surface-normal detectors are shown in blue. (Source: Ramaswamy⁵¹)

The move from III-V to Si-Ge absorber material is motivated in part by the improved thermal conductivity of Si-Ge (1.5W/cmK) over that of InP (~0.05W/cmK). A Si-Ge structure, with a doped silicon waveguide and intrinsic germanium absorber, is able to transfer heat away from the active region nearly 30 times faster than a comparable InP structure (InGaAsP waveguide and InGaAs absorber). This makes the SiGe platform attractive for high temperature operation despite having an indirect bandgap. Indirect bandgap materials, such as SiGe, have lower optical absorption efficiency than direct bandgap materials, such as InP and its complex ternary or quaternary compounds. Photodetectors that can dissipate more heat away from the active region can output higher electrical powers before thermal failure.

Radio Frequency (RF) testing of these Ge-Si PIN photodetectors under high optical power conditions is the main focus of this thesis paper. Ramaswamy⁵¹ demonstrated a 1dB compression RF power of 14.35dBm at 1GHz with 60mA of photocurrent and 8V reverse bias for a 7.4 μ m x 1000 μ m device under 100% optical modulation. As a comparison, Tulchinsky⁵⁸ developed a PDA photodiode that holds the current 1dB compression RF power record at 26dBm; however, this device has a much lower bandwidth (300MHz) than the Ge-Si PIN devices (4.38GHz-31.3GHz).

Responsivity does not change even for wavelengths as long as 1560-1570nm, which is outside the absorption edge of bulk Ge. This occurs because of tensile-strain induced bandgap narrowing in the germanium material. Therefore, the absorption linearity of modulation sidebands is not a concern because the wavelength absorption edge is extended far away from the operating wavelength (1550nm).

The bandwidth is limited by compression of the RF signal at high optical power. The 3dB bandwidth remains constant despite an increasing optical power, which proves that space charge effects have little effect on the bandwidth at high bias, Ramaswamy⁵¹ asserts. This is true for optical powers all the way up to the 1dB compression point. At low bias, the 3dB bandwidth drops due to a voltage dependent responsivity of the photodiode, Ramaswamy⁵¹ claims. At extreme biases of 8V or higher, the diode has additional electrical gain due to avalanche effects. At 10 volts, the output RF power should have significantly more gain than the same RF signal at a lower bias, such as 6V. The frequency response and bandwidth for this device are illustrated by Ramaswamy⁵¹ in Figure 5.4 below:

Figure 5.4: (a) Frequency response and (b) 3dB bandwidth, which shows excellent bandwidth linearity as a function of DC photocurrent (Source: Ramaswamy⁵¹)

In addition to high quantum efficiency and high bandwidth, a low dark current is needed to enable the use of these detectors in next generation communication applications and optical interconnects, Yin⁷³ asserts. The 50 μ m and 100 μ m long devices, each possessing a 7.4 μ m width, have a dark current of 169nA and 267nA, respectively, at 2V reverse bias. According to Yin⁷³, contributions to the dark current come from the bulk material and from the sidewalls of the germanium absorber. These factors are addressed in the design and fabrication of these devices.

5.2 Fabrication and Design Considerations

Yin⁷³ describes the fabrication process of these detectors. This section serves as a summary of this process and discusses the design considerations that are made during fabrication. An SEM cross section of the device is illustrated by Ramaswamy⁵¹, which is shown in figure 5.5:

Figure 5.5: SEM cross section of a fabricated device (Source: Ramaswamy⁵¹)

The fabrication process is done on a 100Ω-cm silicon on insulator (SOI) wafer, which has a 1μm thick buried oxide layer and 1.5μm thick intrinsic silicon layer. Boron is injected only at the sites that the photodetectors will reside in, which minimizes parasitic capacitance between the substrate and metal contact pads and minimizes free carrier loss. A silicon rib is formed for the waveguide by etching the substrate 0.6μm deep on the sides with a rib width of 1.4μm. This forms a rectangular rib-waveguide with a height of 1.5μm and width of 1.4μm, which supports single mode operation. A 4.5μm wide lateral taper is patterned at the waveguide facet to improve the facet to waveguide coupling. A similar taper structure is patterned at the detector end, which has a width equal to that of the absorber.

Once the waveguide is formed, the germanium absorber material is deposited on top of it. A 1.3μm thick Ge film is grown using a selective epitaxial process. A 0.1μm thick buffer Ge layer is grown at low temperature to minimize the formation of dislocation defects. The remainder (1.2μm) is grown at 700°C. Chemical-mechanical polishing is used to reduce the absorber thickness to its final size (0.8μm). The wafer is then annealed to decrease the threading dislocation density.

Beam propagation modeling reveals a local maximum in absorption when the Ge thickness is 0.8μm, Yin⁷³ explains. This absorption maximum occurs due to the coupling of light from the waveguide

to the overlying absorber, which is sensitive to the absorber dimensions. This absorber thickness, Y_{in}^{73} asserts, is the optimal result for both high responsivity and high bandwidth. The waveguide height also affects absorber coupling. According to Y_{in}^{73} , better coupling is achieved using a thinner waveguide. However, thinner waveguides introduce significant polarization sensitivity. Waveguide dimensions of $1.5\mu\text{m} \times 1.4\mu\text{m}$ are chosen to reduce the polarization sensitivity. Due to the roughly square dimensions of the rectangular waveguide, these detectors show a polarization sensitivity of less than 1dB.

Phosphorous is implanted into the top of the Ge region, which forms the n-doped cathode of a P-I-N junction. The silicon waveguide has been doped with Boron atoms in a previous step, which serves as the device anode. A high concentration of Phosphorous is injected at the very top of the Ge absorber to form an ohmic contact. Similarly, high concentrations of Boron are injected at the base of the Si-rib waveguide to form the ohmic contacts for the anode. These high density injections are required to prevent Schottky contacts from forming after metallization of the contact pads.

Dopants are activated and implant lattice damage is corrected by performing a Rapid Thermal Annealing (RTA) step at 650°C for 10 seconds. Finally, $1\mu\text{m}$ of Aluminum is deposited and patterned to form the metal contacts of the device.

To reduce dark current in the device, design considerations are made to improve the dark current contributions of the bulk and sidewalls. The bulk contribution is reduced through the reduction of dislocation defects, either by growth or annealing conditions. The sidewall contribution is reduced by pulling in the n^{+} phosphorous implantation $0.4\mu\text{m}$ away from the Ge sidewalls. This reduces the E-field that is allowed to access the trap-laden interfaces at the sidewalls, which reduces the surface leakage current.

5.3 Sources of Optical Loss

Consideration of optical loss sources is critical in the calculation of the device responsivity. Several sources of loss exist, including the following: scattering loss, lensed-fiber to facet coupling loss, transmission loss, and free carrier loss. Each of these sources is discussed.

A 4.5 μm wide taper is used to collect incident light at the waveguide facet. Coupling is accomplished using a lensed fiber (Corning OptiFocus Lensed Fiber) with a 3.3 μm spot size, which has a facet coupling loss of 4.5dB. Using a smaller spot size of 2 μm , Yin⁷³ is able to achieve a 3dB facet coupling loss. This loss increases as the fiber becomes misaligned from the center of the waveguide.

According to Yin⁷³, The transmission loss of the Si-rib waveguide is very low, on the order of 0.55 dB/cm. Scattering losses, Yin⁷³ explains, account for 2-5% of the total transmission loss. The majority of the optical mode is confined to the waveguide, but a small portion occupies the bulk substrate. This occurs because of the relatively close index of refraction between the Si waveguide (3.5) and SiO₂ (1.4) at 1550nm. Light prefers to travel in the material with the highest index of refraction, but portions may leak out into surrounding materials that have a lower index of refraction.

Another source of loss is the free carrier loss, which is introduced by the heavy doping of the Si-rib waveguide. According to Yin⁷³, this loss is approximately 0.1dB.

Loss due to absorber coupling is not considered because, according to Yin⁷³, all of the light is absorbed for detectors of at least 100 μm length.

These sources of optical loss are summarized in table 5.2 below:

Source of Optical Loss	Magnitude
Lensed-fiber to Facet Coupling	4.5dB
Waveguide Transmission Losses	0.55dB/cm
Free Carrier Loss	0.1dB

Table 5.2: Sources of Optical Loss (Source: Yin⁷³)

5.4 Thermal Considerations

Heat dissipation restricts the maximum obtainable RF power and compression current, which limits the dynamic range of the device. According to Ramaswamy⁵¹, the absorber region heats up to temperatures as high as 85°C when dissipating 1W of DC power. A thermal simulation of the device under these conditions is illustrated by Ramaswamy⁵¹, and is shown in figure 5.6:

Figure 5.6: Thermal simulation showing 85°C temperature in the active region with 1W of DC power dissipated in the device (Source: Ramaswamy⁵¹)

From figure 5.6, it is seen that heat is dissipated mainly through the signal and ground metallization. The oxide layers to either side and below the device have poor thermal conductivity. The bulk Si dissipates nearly no heat because the heat is blocked by the SOI oxide layer. To allow heat to flow into the bulk Si layer, vias can be placed directly below the ground terminals and Si waveguide. Because the bias potential at these sites is already ground, piercing the oxide layer with a via is expected to have no electrical effect on the underlying bulk.

6. Device Characterization

A basic characterization of the silicon waveguide devices is carried out to gain a better understanding of its capabilities. First, an IV characterization is performed to find the avalanche breakdown voltage and series resistance of the photodiodes. Then, after accounting for sources of optical loss, the responsivity of each device is found. Finally, this section concludes with a measurement of device bandwidth under low optical powers.

The devices tested are summarized in table 6.1 below:

Device Name	Absorber Width (μm)	Absorber Length (μm)	Cross Sectional Area (10^{-12} m^2)
Chip 1, Device 2 (C1D2)	7.4	250	1850
Chip 1, Device 5 (C1D5)	5.8	500	2900
Chip 2, Device 1 (C2D1)	5.8	100	580
Chip 2, Device 2 (C2D2)	5.8	250	1450
Chip 2, Device 4 (C2D4)	4.4	250	1100
Chip 2, Device 5 (C2D5)	4.4	500	2200
Chip 2, Device 6 (C2D6)	4.4	250	1100
Chip 3, Device 1 (C3D1)	3.8	250	950
Chip 3, Device 3 (C3D3)	3.8	250	950
Chip 3, Device 4 (C3D4)	2.4	250	600
Chip 4, Device 1 (C4D1)	10.4	250	2600
Chip 4, Device 2 (C4D2)	10.4	250	2600
Chip 4, Device 3 (C4D3)	5.4	250	1350
Chip 4, Device 5 (C4D5)	5.4	250	1350
Chip 4, Device 6 (C4D6)	5.4	500	2700
Chip 5, Device 1 (C5D1)	9.0	250	2250
Chip 5, Device 3 (C5D3)	7.4	250	1850
Chip 5, Device 5 (C5D5)	7.4	500	3700
Chip 5, Device 6 (C5D6)	7.4	250	1850

Table 6.1: Dimensions for each photodiode and the device name for each

6.1 IV-Characterization

Electrical contact between the measurement setup and the metal contact pads on the chip is made using a ground-signal (GS) probe purchased from Cascade Microtech (ACP50-150). Current is measured using a precision ammeter (Keithley 485), which is placed in series between a voltage source (Topward 3302D) and the DC bias port of an HP 8720B Vector Network Analyzer (VNA). This bias port on the VNA serves as both a bias-T and as a BNC to SMA adapter. An SMA male-to-male coaxial cable,

purchased from Rosenberger, connects the GS probe to the output port of the VNA. A photograph showing the locations of the SMA cable, VNA, and probe is shown in figure 6.1:

Figure 6.1: Measurement setup showing the VNA, high-bandwidth Rosenberger cable, and GS probe

Figure 6.2 shows the GS probe and a lensed fiber, which is used to couple light into the waveguide facet on the chip. The lensed fiber is discussed in section 6.2.

Figure 6.2: Measurement setup showing the GS probe, photodetector chip, and lensed fiber assembly

The probe terminals may introduce an artificial heat dissipation pathway to these devices that would not generally be present in a typical application. To explore the probe's effect on heat dissipation, we seek to eliminate the probe from the measurement setup altogether. To do this, C3D1 has been wirebonded to a 50Ω coplanar waveguide, which is electrically connected to an SMA connector by a small blob of silver filled epoxy (Epotek H20E). The wirebond material is gold, which makes a good connection between the aluminum pads and the gold surface of the 50Ω coplanar waveguide. This wirebond connection is shown in figure 6.3:

Figure 6.3: Wirebond connection between the photodiode aluminum contact pads and a gold 50Ω coplanar waveguide

The thin film coplanar waveguide is plated with gold, which rests on a 25mil thick, sapphire substrate. The dimensions of this thin film chip are 170mil x 170mil. The center conductor width and spacing from the coplanar ground plane is measured using Vernier calipers, which are then verified as for 50Ω impedance waveguide using the LineCalc program from the Advanced Design System (ADS) by Agilent Corporation¹. Table 6.2 summarizes the parameters entered into LineCalc's CPWG module to simulate a 50Ω impedance waveguide with similar dimensions to those measured using the calipers:

Parameter	Value	Parameter	Value
Center conductor width (W)	12.3 mil	Ground plane spacing (G)	7 mil
ϵ_r	9.8	μ_r	1
Substrate height (H)	25 mil	Metallization thickness (T)	0.125 mil
Conductivity	4.52E7	Loss Tangent	0.0002
Frequency	3GHz	Length (L)	Any

Table 6.2: ADS LineCalc parameters used to replicate the 50Ω thin film coplanar waveguide dimensions

The W, G, H, and T parameters from table 6.2 are shown graphically in figure 6.4:

Figure 6.4: Dimension parameters used by LineCalc's CPWG module (Source: Agilent¹)

The SMA connector is held in place by a metal vice, which is secured to the optics bench using screws. The lensed fiber is positioned using an XYZ positioner, as shown in figure 6.5 below:

Figure 6.5: Measurement setup for testing diodes without the G-S probe, which makes electrical connection using gold wirebond and a gold thin film coplanar waveguide

Electrical connection is accomplished using a G-S probe and optical coupling to the silicon waveguide is done using a lensed optical fiber. This is shown by the block level diagram in figure 6.6:

Figure 6.6: Block level diagram showing electrical and optical connections

IV characteristic measurements do not require the lensed fiber, and therefore this has been eliminated from the measurement setup. A block level diagram showing the IV characteristic measurement setup is shown in figure 6.7:

Figure 6.7: IV characteristic measurement block diagram

We first establish a baseline IV-curve under reverse bias conditions by performing a detailed characteristic measurement at 0.1V intervals. This measurement is shown in figure 6.8. From this plot, a soft reverse breakdown is observed to occur around 12.8 volts, which produces over 1mA of dark current. In contrast, a 10 volt bias only produces 0.1mA of dark current. At 6V, the dark current is 15.16 μ A. The median voltage between zero-bias and reverse breakdown appears to be around 6V. This median voltage is the optimum DC bias for obtaining maximum RF power output from the device. RF measurements are described in more detail in chapters 7-9.

Figure 6.8: IV-Characteristic for Chip 1 Device 2

This process is repeated for device 5 on chip 1. The result gives the IV-Characteristic for all functioning devices on chip 1:

Figure 6.9: IV-Characteristic for all functioning devices on Chip 1

Similarly, an IV-characteristic curve is obtained for all the devices on chip 2. This is shown in figure 6.10. Devices with similar dimensions appear to have similar IV-characteristic curves. For example, devices 4 and 6 appear to have IV-characteristic curves that nearly overlap, and this makes sense since both devices have the same dimensions ($4.4\mu\text{m} \times 250\mu\text{m}$). Devices with greater cross sectional area, such as device 5, have greater dark current and experience reverse breakdown at lower bias. Likewise, devices with lower cross sectional area, such as device 1, have less dark current and can be biased at higher levels before breakdown.

Figure 6.10: IV-characteristic for all functional devices on chip 2

The IV-characteristic for chip 3 is shown next:

Figure 6.11: IV-characteristic for all functional devices on chip 3

The IV characteristic curves shown in figure 6.11 are very similar with little deviation from each other. An explanation for this can be found in the device dimensions. Each of these devices has the same length ($250\mu\text{m}$) with small variations in width ($3.8\mu\text{m}$ and $2.4\mu\text{m}$). Reverse breakdown occurs at a bias of 12.5V for all devices.

The series resistance, discussed in section 3.3, can be found through an analysis of the forward- and reverse-bias IV-characteristic. When operating at high forward bias, greater than 1.5V , the diode is completely conducting, which corresponds to a very low dynamic junction resistance (R_j). The result is an I-V slope that equals the conductance of the series resistance. The inverse of this slope is the series resistance. The discrete component model for a diode is explained in section 3.3, which is illustrated again for the reader's convenience in figure 6.12:

Figure 6.12: Discrete component diode model (Source: Chang¹⁰)

The ohmic resistance (R_s) is a constant resistance that is determined by the contact resistance of the contact pads and resistivity of the semiconductor layers. R_s does not change with bias voltage, unlike the dynamic resistance (R_j). The dynamic resistance is very low in forward bias (a few ohms) and very high in reverse bias (several mega ohms).

According to Razeghi⁵², current is limited at very high forward bias by the series resistance of the diode (R_s). This is shown by the dark current performance regions of the diode, which are illustrated in figure 6.13. In forward-bias, these regions are dominated by recombination, diffusion, high-injection, and series resistance effects. Current in reverse-bias is dominated by thermal generation and surface leakage.

Figure 6.13: Dark current performance regions for a PN junction diode (Source: Razeghi⁵²)

A forward-bias characteristic is obtained for C5D1 with an active region temperature of 475°C, which is shown in figure 6.14 below:

Figure 6.14: Forward-bias IV-characteristic for C5D1, showing a series resistance of 24.8Ω at 475°C

The measurement shown in figure 6.14 has a current knee that is too broad to be explained by the ideal diode equation at room temperature. Attempts to curve fit to the ideal diode equation at room temperature yield an ideality factor greater of 2.16, which is not theoretically valid. The ideality factor must have a value between 1 and 2 for this model. Razeghi⁵² gives an expression for the forward bias characteristic that accounts for the recombination current and temperature directly instead of relying on an ideality factor. This is the model that is used to curve fit the measurement in figure 6.14, where I_0 is the saturation current and I_{R0} is the recombination current. I_{R0} is a material constant that depends on the minority carrier recombination lifetimes and depletion layer width.

$$I = I_0 \exp\left(\frac{qV}{k_b T}\right) + I_{R0} \exp\left(\frac{qV}{2k_b T}\right)$$

where

$$V = V_{bias} - IR_S$$

Although the curve fit in figure 6.14 is not exactly precise at biases ranging from 0.3V to 1.5V, it shows a distinctly linear region past 1.5V, which is due to the series resistance of the diode. Adjusting the series resistance in the model changes the slope of this region. Changing other factors, such as the temperature T , saturation current I_0 or the recombination current I_{R0} changes the turn on voltage of the diode, but does not change the slope of the series resistance region. Increasing the device temperature causes the current knee to smooth out compared to room temperature, with temperatures as high as 475°C needed to obtain a close curve fit to the actual measurement. This temperature is reasonably high since the surrounding oxide layers have very poor thermal conductivity and tend to trap the heat.

The series resistance is the inverse of the linear slope at biases above 1.5V. This yields a series resistance (R_S) of 24.80Ω for this device. Deviation from this model may be due to surface recombination current, elevated temperature in the active region, and lattice defects in the germanium absorber, which all tend to increase the dark current. Exact curve fitting could be accomplished by adjusting the active region temperature and recombination and saturation current coefficients.

To obtain the series resistance of C2D2, an extended IV characteristic measurement is performed.

This is shown in figure 6.15:

Figure 6.15: Extended IV characteristic for C2D2

The forward characteristic in figure 6.14 shows a maximum slope of 0.08256A/W , which corresponds to a resistance of 12.11Ω . Similarly, the reverse characteristic shows a smaller slope and larger resistance (29.7Ω). The device experienced thermal failure at 3.9V forward bias and 211mA of dark current, which corresponds to a DC power of 0.82W . The slope starts to drop off around the failure point because series resistance is added to the device due to melting of the intrinsic Ge layer.

The reverse characteristic resistance in figure 6.15 should equal the forward characteristic resistance. The dielectric oxide layer may have been damaged during reverse bias testing, which would result in additional leakage current under forward bias. The maximum current limit for reverse bias testing should be 10mA , which limits the power dissipation to 0.2W . In addition, the forward

characteristic failed at a lower DC power level than the reverse characteristic. This indicates that damage to the oxide dielectric is present during forward bias testing. Therefore, the series resistance is taken as the reverse bias value (29.7Ω), which is higher. The forward bias resistance is too small and does not make sense when compared to other diodes.

Contact resistance between the pad metallization and probe contributes nearly 16Ω to the total series resistance which is the outcome of an aluminum oxide patina that forms on the metal surface. A patina is a naturally occurring, protective oxide layer that forms on the surface of metals and prevents further oxidation of metal below the patina layer. This oxide is deposited on the probe tip after making contact and must be removed by an abrasive cleaning process. However, this abrasive cleaning creates surface roughness on the probe tip that can increase the contact resistance even further. It is difficult to measure this resistance directly because the probe breaks through the oxide when shorting the terminals to a solid metal object, such as brass.

In addition, the probe must be perfectly level to make good electrical contact to both ground and signal pads simultaneously. The probe is very sensitive to rotation misalignment. Even a slight misalignment (1 deg) may result in an open circuit when the probe is pressed onto the contact pads. Often, this misalignment is smaller than the eye can see, and the probe angle has to be adjusted many times before the probe will conduct. Applying more pressure to the probe will cause one lead to bend and dig into the aluminum pad, which lowers the other lead far enough to make contact. Digging into the pad metallization allows the pad tips to break through the oxide patina, which lowers the contact resistance to an acceptable level around 16Ω .

The pad resistance can be found by measuring the total series resistance of two devices and considering a system of equations. This total resistance includes the series resistance due to the absorber cross sectional area and the contact resistance to the metal pads. The total series resistance of C5D1 and C2D2 are 24.8Ω and 29.7Ω , respectively.

The series resistance of other photodiodes can be approximated by the size of their cross sectional area in relation to C5D1, which is $9.0\mu\text{m} \times 250\mu\text{m}$, assuming a negligible probe contact resistance. An equation to relate the series resistance of C5D1 to that of other devices can be found from the definition of series resistivity:

$$\rho = \frac{A}{L} R_S$$

Assuming the resistivity and absorber length between devices is held constant, the series resistance (R_S) of one device can be related to another by the ratio of their cross sectional areas (A):

$$R_{S(C2D2)} = \frac{A_{C5D1}}{A_{C2D2}} R_{S(C5D1)}$$

The total series resistance is then the sum of the absorber resistance and contact pad resistance:

$$R_{S, Total} = R_{pad} + R_S$$

Unfortunately, forward and reverse bias IV characteristic data for the other devices was taken for a maximum current of 2mA. This is not enough current to see the full IV characteristic and obtain the total series resistance. These devices were destroyed during other tests before the IV characteristic data could be retaken for higher biases. Since taking this data is risky and could result in increased leakage current. Therefore, only C5D1 and C2D2 have been tested at such high biases.

Using the resistivity and series resistance equations, the pad and absorber resistance calculation is as follows. The total series resistance of C5D1 (24.8Ω) is the sum of the pad and absorber series resistances:

$$\begin{aligned} R_{total(C5D1)} &= R_{pad} + R_{S(C5D1)} \\ R_{pad} &= 24.8 - R_{S(C5D1)} \dots\dots\dots (1) \end{aligned}$$

The same condition applies to the total series resistance of C2D2 (29.7Ω), except the absorber resistance is expressed in terms of C5D1 using the absorber cross sectional area ratio:

$$\begin{aligned} R_{total(C2D2)} &= R_{pad} + R_{S(C2D2)} \\ R_{S(C2D2)} &= \frac{A_{C5D1}}{A_{C2D2}} R_{S(C5D1)} \end{aligned}$$

$$R_{total (C2D2)} = R_{pad} + \frac{A_{C5D1}}{A_{C2D2}} R_{S (C5D1)} \dots\dots\dots (2)$$

Substituting equation (1) into equation (2) yields the absorber series resistance of C5D1:

$$29.7 = 24.8 - R_{S (C5D1)} + \frac{2250}{1450} R_{S (C5D1)}$$

$$R_{S (C5D1)} = 8.877 \Omega$$

Likewise, the absorber series resistance of C2D2 is the following:

$$R_{S (C2D2)} = \frac{A_{C5D1}}{A_{C2D2}} R_{S (C5D1)} = \frac{2250}{1450} (8.877) = 13.77 \Omega$$

The contact pad resistance is found by subtracting the absorber resistance from the total series resistance. This resistance is assumed to be the same across all devices.

$$R_{pad} = 24.8 - R_{S (C5D1)} = 24.8 - 8.877 = 15.92 \Omega$$

The contact pad resistance (R_{pad}) and ohmic resistance (R_S) are shown in figure 6.16 below:

Figure 6.16: Diagram showing locations of the series resistances R_{pad} and R_S

Table 6.3 summarizes the absorber cross sectional area and predicted series resistance for each device on chips 1-5:

Device	Width (μm)	Length (μm)	Cross Sectional Area (10^{-12} m^2)	Predicted Absorber Series Resistance, R_S (Ω)	Contact Pad Resistance, R_{Pad} (Ω)	Predicted Total Series Resistance (Ω)
C1D2	7.4	250	1850	10.80	15.92	26.72
C1D5	5.8	500	2900	6.89	15.92	22.81
C2D1	5.8	100	580	34.44	15.92	50.36
C2D2	5.8	250	1450	13.77	15.92	29.69 (actual)
C2D4	4.4	250	1100	18.16	15.92	34.08
C2D5	4.4	500	2200	9.08	15.92	25.00
C2D6	4.4	250	1100	18.16	15.92	34.08
C3D1	3.8	250	950	21.02	15.92	36.94
C3D3	3.8	250	950	21.02	15.92	36.94
C3D4	2.4	250	600	33.29	15.92	49.21
C4D1	10.4	250	2600	7.68	15.92	23.60
C4D2	10.4	250	2600	7.68	15.92	23.60
C4D3	5.4	250	1350	14.80	15.92	30.72
C4D5	5.4	250	1350	14.80	15.92	30.72
C4D6	5.4	500	2700	7.40	15.92	23.32
C5D1	9.0	250	2250	8.88	15.92	24.80 (actual)
C5D3	7.4	250	1850	10.80	15.92	26.72
C5D5	7.4	500	3700	5.40	15.92	21.32
C5D6	7.4	250	1850	10.80	15.92	26.72

Table 6.3: Calculated series resistance

6.2 Responsivity Measurement

A lensed fiber is used to focus laser light onto a $3.3\mu\text{m}$ spot size on the waveguide facet. This light is then coupled into the silicon waveguide, which has a $4.5\mu\text{m}$ taper structure for efficient light coupling. A microscope photograph of the probe, chip, and lensed fiber is shown in figure 6.17:

Figure 6.17: Microscope photograph of the lensed fiber, photodiode chip, and GS probe

A block level diagram showing the measurement setup for taking responsivity data is shown in figure 6.18:

Figure 6.18: Responsivity measurement block level diagram

The system responsivity is obtained by measuring the current produced by a known optical power level. A precision ammeter (Keithley 485) is used to measure currents up to 2mA. The optical power launched into the lensed fiber is measured using a 2% splitter with the tap terminal connected to an Agilent 8604B Optical Spectrum Analyzer (OSA).

The optical power ratio between the tap and output terminals of the 2% splitter is determined experimentally. An optical power of 13.63mW at 1550nm is launched into the input port, which produces 0.239mW of optical power at the tap port and 12.81mW of power at the output port. This corresponds to an insertion loss between the input and output ports of 0.27dB and a tap ratio of 1.866% with respect to the output port. Therefore, the optical power launched into the output port can be found using the following relation:

$$P_{output}(mW) = \frac{P_{tap}(mW)}{0.01866}$$

$$P_{output}(dBm) = P_{tap}(dBm) + 17.29dB$$

Once laser power is coupled into the facet of the photodetector waveguide, measurements of the current versus optical power are made for several optical power levels. The slope of this data gives the system responsivity, which is shown for C2D2 at 6V reverse bias in figure 6.19 below:

Figure 6.19: System responsivity measurement of C2D2 at 6V reverse bias. A responsivity of 0.24A/W is obtained from the linear slope

To analyze the effect of lensed fiber misalignment on system responsivity, this measurement is again conducted except with the fiber positioned slightly away from the center of the waveguide (0.5-1 μ m misalignment). The result of this measurement is shown in figure 6.20. The measured responsivity is nearly half of the original value, which indicates that the fiber must be perfectly aligned in order to achieve accurate responsivity measurements. To achieve the best alignment, the lensed fiber position is adjusted until the ammeter shows a steady, maximum photocurrent. If the fiber is misaligned slightly, this current will appear to drift noisily because of natural but subtle movement in the fiber due to air circulation in the room and thermal expansion and contraction of the metal apparatus holding the lensed fiber in place.

Figure 6.20: System responsivity of C2D5 with 6V reverse bias and perfect lensed fiber alignment (0.23A/W), and with the lensed fiber misaligned (0.14A/W)

Responsivity is also characterized for several bias levels for C1D2. Figure 6.21 shows the photocurrent versus the optical power input into the lensed fiber for reverse bias levels between 2 and 8 volts. Figure 6.22 shows the responsivity for each bias level. This result shows that the responsivity of the system is independent of bias for voltages above 2V and holds a constant value around (0.2A/W).

Figure 6.21: System responsivity measurement for C2D1 at several bias levels

Figure 6.22: System responsivity versus bias

The system responsivity for each device on chip 2 is tested in this fashion at 6V reverse bias. The result is summarized in figure 6.23:

Figure 6.23: System responsivity for all devices on chip 2, showing an average responsivity of 0.23A/W. The absorber length is labeled for each device.

The responsivity values shown in figure 6.23 should have consistent values. Measurements by Tao Yin⁷³ indicate that nearly 100% absorption occurs for devices that are 100 μ m or longer. Variation in measured responsivity could be due to dust or other particles on the waveguide facet. C2D4 and C2D6 show responsivities that are separated by nearly 0.02A/W, despite having identical absorber lengths (250 μ m).

The system responsivity can be mathematically converted into device responsivity by taking into account sources of optical loss and subtracting these from the optical power reaching the absorber. Optical loss sources are summarized in table 5.2. The devices on chip 2 have a waveguide length of approximately 80% of the chip length (3.3mm), or 2.64mm. The waveguide transmission loss is

calculated using this length and the centimeter transmission loss (0.55dB/cm) found experimentally by Tao Yin⁷³:

$$TL(dB) = \frac{0.55dB}{cm} \frac{cm}{10mm} (0.8)(3.3mm) = 0.145dB$$

The free carrier loss in the waveguide is approximately 0.1dB total, and the facet coupling loss is approximately 4.5dB. Therefore, the total optical loss spanning from the lensed fiber to the absorber is 4.75dB.

To calculate the device responsivity, the equivalent optical power reaching the absorber is calculated. Assuming 1W of optical power, the equivalent power reaching the absorber is the following:

$$P_{absorbed} = 1W - P_{loss} = 30dBm - 4.75dB = 335mW$$

The 1W optical power assumption is mathematically convenient because it produces a photocurrent equal to the system responsivity. For example, if 1W of optical power is launched into the lensed fiber of a system with responsivity of 0.24A/W, then 0.24A of photocurrent is subsequently produced. If optical losses are taken into account, then that same current is produced by an optical power at the absorber that is 4.75dB less than the system optical power. The device responsivity (R_{device}) in this example is now the following:

$$R_{device} = \frac{(R_{system})(1W)}{P_{absorbed}} = \frac{(0.24A/W)(1W)}{335mW} = 0.716 A/W$$

Figure 6.24 shows the device responsivity at 6V bias, which uses this correction factor:

Figure 6.24: Device responsivity for all devices on chip 2, showing an average responsivity of 0.70A/W

To compare the devices on chip 2 to other chips, the responsivity of C5D1 is measured. This is shown in figure 6.25:

Figure 6.25: System responsivity of C5D1 showing 0.314A/W, or a device responsivity of 0.973A/W

Yin⁷³ was able to produce a device responsivity of 1.16A/W with a 100 μ m long absorber. The device responsivity obtained for C2D1 (0.64A/W), also a 100 μ m long device, is nearly half the value obtained by Yin⁷³. The optical loss needed to yield Yin's⁷³ device responsivity (1.16A/W) is 6.8dB, which is 2dB greater than the loss value used to calculate the device responsivity.

The voltage dependence of the responsivity is measured for C5D1 at low optical power (0.17mA photocurrent). As shown in figure 6.25, the responsivity remains constant (0.31A/W) even at 0V external bias. This result occurs because the photocurrent is not high enough to oppose the built in potential by dropping voltage across the diode series resistance. The effect of high photocurrent will become apparent when testing under high optical power conditions.

An Erbium Doped Fiber Amplifier (EDFA), described in section 7.1, is used to test the voltage dependent responsivity with high optical power (18dBm). The system responsivity is tested by measuring

the optical power with a 2% splitter and recording the photodetector current. The ratio of the current and optical power is the responsivity, which is obtained for several bias levels. The fiber position is corrected for each data point to account for drift in the fiber-chip coupling. This is shown for C5D1 in figure 6.26:

Figure 6.26: Voltage dependent responsivity for C5D1 under high optical power conditions (+18dBm) at 1550nm. Below 3V, the diode is limited by R_S . Above 3V, the voltage dependent responsivity is shown.

Figure 6.26 shows a steadily decreasing responsivity when a reverse bias of less than 2V is placed across the diode. If this line is extended into the forward bias region, zero responsivity is obtained at approximately 0.3V forward bias. Above 3V bias, a relatively constant system responsivity of 0.28A/W is obtained. This corresponds to a diode responsivity of 0.836A/W. The responsivity decreases at biases less than 2.5V because the large current passing through the diode (32.7mA) drops significant potential across the series resistance of C5D1 and because of space charge effects. Dark current is subtracted from the total current when calculating the responsivity. An active region temperature of 475°C is assumed. According to Razeghi⁵², the total current is the sum between the photocurrent and the temperature-dependent dark current, which is shown by the following expression:

$$I = I_0 \left(e^{\frac{-Ea}{kT}} - 1 \right) - I_{ph}$$

A 0.3eV activation energy (E_a) is used, Boltzmann's constant (k) is expressed in units of eV/K, and temperature is in units of Kelvin. I_0 is found by measuring the dark current at room temperature. This room temperature dark current is 14.27 μ A, which results in an I_0 of 2.064A. At 475°C, this dark current increases to 19.65mA and becomes the dominant contributor to the total current.

This result is somewhat closer to Yin's⁷³ responsivity measurement (1.16A/W) for a 7.4 μ m x 100 μ m device. Yin⁷³ contends that waveguide photodetectors having a length of 100 μ m or greater will absorb all of the light that passes through it. Differences in responsivity could be due to fiber positioning and differences in the manufacturing quality of the silicon waveguide. Surface roughness in the silicon waveguide could introduce additional scattering loss that may be absent in Yin's⁷³ device.

C5D1 shows a 6V system responsivity of 0.281A/W. The responsivity at higher bias gradually increases, which is a source of non-linearity. This can be compared with the voltage-dependent responsivity behavior at high reverse bias described by Beling et al.⁶ for an InGaAs/InP photodiode operating at 50mA photocurrent, which is shown in figure 6.27 below:

Figure 6.27: Voltage-dependent responsivity for an InGaAs/InP photodiode operating at 50mA photocurrent (Source: Beling⁶)

Although figure 6.27 does not further our understanding of the SiGe photodetectors tested, it does highlight the shape and appearance of voltage dependent responsivity in general for several wavelengths.

6.3 Bandwidth Measurements

A Mach-Zehnder interferometer is inserted into the optical path to allow RF modulator of the optical signal. This modulator is the Mach 10 (model 023) purchased from the Codeon Corporation. Its datasheet shows that this modulator has a 4.09dB optical insertion loss at -3.89V DC bias and an S21 bandwidth of 10.65GHz.

The Mach-Zehnder modulator is characterized by applying a constant optical input power and measuring the output power using the OSA for DC biases ranging from -7V to 6.5V in 0.1V increments. The result is shown in figure 6.28 below:

Figure 6.28: Mach-Zehnder DC Bias Characterization.

Figure 6.28 can also be expressed in terms of the optical insertion loss. This is shown in figure 6.29:

Figure 6.29: Mach-Zehnder DC Bias Characterization, Showing Optical Insertion Loss

Figures 6.28 and 6.29 provide details on the bias and modulation power needed to achieve 100% modulation of the optical signal. Figure 6.29 shows that the maximum and minimum insertion losses occur at 4.6V and -3.8V, respectively. A 100% modulated optical signal can be achieved by DC biasing at the center of these voltages, 0.4V, and providing a peak voltage swing of 4.2V. This corresponds to an input AC power of 22.5dBm into a 50Ω load.

Figure 6.28 indicates that the linear region of the modulator (<5% deviation) occurs between -2.1V bias and 1.7V bias. For maximally linear optical modulation, the DC bias should be set to -0.2V with a peak AC voltage swing of 1.9V. An AC power of 15.6dBm is required.

Since the maximum power output from the VNA is -10dBm, the modulator can be set to 0V DC bias without introducing non-linearity by operating in the center of the modulator's linear region. Port 2 of the VNA is connected to the AC terminals of the Mach-Zehnder modulator, and the electrical output

from the photodetector is connected to Port 1. A block level diagram showing the bandwidth measurement setup is shown in figure 6.30 below:

Figure 6.30: Bandwidth measurement block diagram

The bandwidth of this system can be obtained from an S12 measurement. Such a measurement is shown in figure 6.31:

Figure 6.31: Raw S12 system measurement of C1D5 for various reverse biases and 0.6mA photocurrent, no S12 correction

Figure 6.31 illustrates that device bandwidth can be improved by increasing the applied reverse bias. This can be explained through the voltage dependent junction capacitance of the diode. Reverse biases of 0V and 0.1V show significantly less transmission at frequencies above 1GHz than at 0.5V bias. Increasing the device bias further produces a progressively diminishing return.

To find the bandwidth response of the diode, the S12 response of each surrounding system component is subtracted from the measurement. This data is shown in the following plots: the measured S12 response of the Rosenberger coaxial cable is shown in figure 6.32, S12 data from the GS probe datasheet is shown in figure 6.33, and S12 data from the Mach-Zehnder modulator datasheet is shown in figure 6.34.

Figure 6.32: S12 measurement of the Rosenberger coaxial cable

Figure 2.33: S12 response of the GS probe, which is recorded from the datasheet

Figure 6.34: S12 response estimated from the Mach-Zehnder interferometer datasheet

The Mach-Zehnder S12 response (figure 6.34) is estimated from a plot shown on the datasheet. It is expressed in units of dBe, which are electrical decibels. These units are used because a photodiode is required to detect the modulation response of the optical signal.

The 20GHz VNA has a technical problem that allows no more than a few plots from being obtained without requiring a manual restart and recalibration. This is an intermittent problem, which causes the machine to permanently lock up. This can only be resolved by powering down the machine, which makes collecting consistent data by GPIB very time consuming. Manually recording data by hand using the screen markers was actually found to be a more efficient method for collecting S-parameter data using this machine. In addition, increasing the number of collected data points from 401 to 1601 results in the VNA locking up every time the data is captured over GPIB. The use of 1601 data points is necessary because it significantly smoothes out combing artifacts in the S12 response and produces a sharp, clean curve on the VNA display. Although the same result is produced using averaging and aperture smoothing

in the 401 data point result, the GPIB capture program deactivates these settings during the capture process. The problem may be a hardware issue or a fault in the capture software. Unfortunately, this is the only instrument available at our institution that is capable of measuring S-parameter data for frequencies higher than 3GHz.

S12 data is recorded for all devices on chips 2 and 3 using this point-by-point collection method. Figure 6.35 (a) shows the system S12 measurement for C2D2, and Figure 6.23 (b) shows the calculated device S12, which is obtained by subtracting the RF insertion loss of the modulator, probe, and coaxial cable from the system S12 measurement:

Figure 6.35: (a) System S12 measurement of C2D2, and (b) calculated S12 for the diode only

The 3dB bandwidth is obtained by measuring the difference in frequency between the point of least insertion loss and the frequency where S12 drops 3dB below this value. Linear correlation is used to approximate the frequency between two points where S12 drops by 3dB. To show this graphically, a line was drawn in MATLAB at this point that is exactly 3dB tall, proving that the point obtained is truly the 3dB cutoff point. This plot is shown in figure 6.36 for C2D2 under 0.2V bias:

Figure 6.36: Measurement of the 3dB bandwidth for C2D2 at 0.2V reverse bias

This process is repeated for each diode on chips 2 and 3. The result is shown in table 6.3 below:

	0.2V Bias	1V Bias	6V Bias	10V Bias
Chip 2, Device 1	0.883 GHz	2.387 GHz	5.944 GHz	5.216 GHz
Chip 2, Device 2	2.203 GHz	3.207 GHz	3.706 GHz	3.695 GHz
Chip 2, Device 4	1.120 GHz	2.969 GHz	3.821 GHz	6.148 GHz
Chip 2, Device 5	1.598 GHz	2.048 GHz	3.163 GHz	3.595 GHz
Chip 2, Device 6	1.422 GHz	3.431 GHz	5.930 GHz	6.296 GHz
Chip 3, Device 1	1.176 GHz	2.426 GHz	3.325 GHz	4.958 GHz
Chip 3, Device 3	2.996 GHz	3.725 GHz	6.604 GHz	9.385 GHz
Chip 3, Device 4	3.009 GHz	6.531 GHz	7.186 GHz	10.710 GHz

Table 6.3: 3dB bandwidth obtained from an S12 measurement of C2D5

At 6V reverse bias, the worst bandwidth is obtained for C2D5 (3.2GHz), which also has the highest absorber cross sectional area ($2200 \times 10^{-12} \text{ m}^2$). Higher bandwidth is obtained for C3D4 (10.7GHz), which indicates that this device has a lower RC time constant. This device has one of the lowest absorber cross sectional areas when compared to the other devices ($600 \times 10^{-12} \text{ m}^2$). Only C2D1 has a smaller cross sectional area ($580 \times 10^{-12} \text{ m}^2$), but the contact pads of this diode were probed significantly more than any other device on the chip, which could have resulted in damage to the metallization over time and raised the contact resistance. Indeed, an increase in pad contact resistance raises the RC time constant, which lowers the obtainable bandwidth.

Contact pad damage may result from probe digging, from scratching of the metal surface, and from foreign objects being deposited into the metal surface. These foreign objects include scratched off aluminum oxide from previous tests and sandpaper grit from the cleaning process. An occasional abrasive cleaning of the probes is performed to remove oxide buildup on the probe tips. However, even with severe contact pad damage, C2D1 is able to produce nearly 6GHz of bandwidth at 6V bias, which is nearly double that of C2D5 (3.1GHz).

Figure 6.37 graphically summarizes the bandwidth data shown in table 6.3:

Figure 6.37: 3dB Bandwidth of all devices on chips 2 and 3

Drift in position of the XYZ fixture can cause a visible fluctuation in S12 when viewed on the VNA display. To address this, four measurement sweeps are averaged internally by the VNA, and the trigger is subsequently deactivated. The XYZ fixture and lensed fiber move slightly over time due to thermal expansion of the metal base and air circulation in the room. This effect can be mitigated by

reducing the length of lensed fiber that is allowed to protrude from the metal alignment block, which sits on top of the XYZ positioning fixture. The lensed fiber protrusion length is illustrated in figure 6.38 below:

Figure 6.38: Lensed fiber protrusion length illustration

Additionally, the algorithm used to calculate the 3dB bandwidth loses accuracy due to the coarse interval of the data collected. An interval of 1GHz is used to collect 20 data points using markers on the VNA for frequencies ranging from 1GHz to 20GHz. In addition, two datapoints are collected showing the 3dB system bandwidth: one usually at 130MHz, the lowest measurement frequency available, and one with an S21 value that is 3dB lower. Better accuracy could be obtained if more data points are collected. However, we elected to collect only 22 data points per measurement in the interest of time.

Bandwidth data for C1D5 can also be obtained. The raw system S12 measurement was shown in figure 6.31, but has been copied to figure 6.39 (a) for reader convenience. This measurement is processed by subtracting the insertion loss of the each component in the optical link until only the S12 value for the diode remains. The resulting device S12 measurement is shown in figure 6.39 (b):

Figure 6.39: (a) system S12 and (b) S12 referred to the diode (C1D5)

Bandwidth is also voltage dependent and drops sharply at low bias. Figure 6.40 shows the 3dB bandwidth versus bias voltage for C1D5:

Figure 6.40: Bandwidth versus bias voltage for C1D5

The measured bandwidth appears to be consistent for measurement at biases higher than 2V. Figure 6.40 shows significant measurement error, even between bias levels that are 0.1V apart. For example, the bandwidth at 0.3V (1.5GHz) is nearly 1GHz larger than the measured bandwidth at 0.4V (0.5GHz). Even with 401 data points, the measurement shown in Figure 6.32 does not have a smooth appearance. Increasing the number of data points to 1601 smoothes out the frequency response considerably, but capturing this data using GPIB is difficult due to an assumed technical problem with our VNA. The VNA locks up every time a 1601 data point capture is attempted, which requires a manual restart to regain control of the instrument. Capturing data at 201 data points also triggers this problem, but does so less frequently. Obtaining a smooth S12 response is critical in performing accurate bandwidth measurements. Averaging and smoothing aperture are deactivated by the capturing program.

At low bias and high modulation frequency, bandwidth is limited by the transit time in the intrinsic region. Charge carriers must be accelerated by a sufficiently large junction electric field to reach their saturation velocity and transit through the intrinsic region before the next modulation cycle begins. This junction electric field is strengthened by the applied reverse bias. Therefore, transit time is the dominant factor limiting bandwidth at low bias.

For higher biases, the bandwidth in these devices is RC limited. The junction capacitance, diode series resistance, and load resistance each contribute to the RC time constant. From table 6.3, the series resistance calculated for C1D5 is 4.33Ω.

A slight change in bandwidth may be attributed to the voltage dependent junction capacitance, $C_j(V)$. This junction capacitance is dependent on the depletion region width, which models a parallel plate capacitor. However, the depletion region width in P-I-N diodes is dominated by the thickness of the intrinsic region, and this width becomes only slightly larger as the reverse bias voltage increases. Therefore, the intrinsic region width mainly determines the width of the depletion region and subsequently fixes the junction capacitance to a near-constant value.

The junction capacitance is found using the following equation for reverse-biases greater than 2V, where R_S is the series resistance, R_L is the load resistance, and $C_j(V)$ is the junction capacitance, and B is the measured bandwidth in Hz:

$$C_j(V) = \frac{1}{2\pi B(R_S + R_L)}$$

Figure 6.41 shows the calculated junction capacitance versus the applied reverse bias:

Figure 6.41: Calculated junction capacitance for C1D5, showing a constant $C_j(V)$ of 1pF at a reverse bias greater than 2V. This capacitance calculation is inconclusive at biases below 2V.

Figure 6.41 shows a constant junction capacitance of 1pF at a bias greater than 2V for C1D5.

This is consistent with the model of a P-I-N diode, since the width of the depletion region mostly depends on the intrinsic region width. The junction capacitance below 2V bias is experimentally inconclusive; however, this capacitance is expected to remain constant even at very low bias. Elevated junction capacitances below 2V bias are measurement artifacts that indicate transmit time limited bandwidth.

The junction capacitance can also be found for other diodes on chips 2 and 3. Table 6.4 shows the calculated junction capacitance for these devices at 6V and 10V reverse bias:

Diode	Area (10^{-12} m^2)	BW, 6V (GHz)	C_j , 6V (pF)	BW, 10V (GHz)	C_j , 10V (pF)
C1D5	2900	2.694	1.090	3.230	0.907
C2D1	580	5.944	0.373	5.216	0.426
C2D2	1450	3.706	0.732	3.695	0.734
C2D4	1100	3.821	0.678	6.148	0.421
C2D5	2200	3.163	0.903	3.595	0.795
C2D6	1100	5.930	0.439	6.296	0.411
C3D1	950	3.325	0.756	4.958	0.508
C3D3	950	6.604	0.381	9.385	0.268
C3D4	600	7.186	0.312	10.710	0.209

Table 6.4: Calculated junction capacitance for all devices on chips 2 and 3 at 6V and 10V reverse bias

The junction capacitance should increase linearly with increasing cross sectional area of the absorber layer. As expected, the junction capacitance at 6V is greater than that at 10V, which is due to a slight expansion of the depletion region with increasing bias. Deviation of data points from the linear slope may be due to variation in the contact resistance between each of the measurements.

The total series resistance, including the load and contact resistances, is used in the calculation of the junction capacitance. Variation in contact resistance could result from deposition of aluminum oxide from the pad metallization to the probe tips. Additionally, contact resistance may increase if the probe is not adequately level to the pads on the chip. Any variation in contact resistance between measurements will alter both the bandwidth obtained and the junction capacitance calculated.

An S12 measurement is performed for C5D1 at 6V bias, which results in a bandwidth of 6.046GHz and a junction capacitance of 0.352pF. Figure 6.42 shows this measurement, which corrects for the insertion loss of each system component:

Figure 6.42: 3dB bandwidth calculation for C5D1 at 6V bias

The junction capacitance for C5D1 at 6V bias can be calculated from this bandwidth measurement. Assuming a diode series resistance of 24.80Ω and a load resistance of 50Ω , the junction capacitance from this 3dB bandwidth calculation (6.046GHz) is the following:

$$C_{j(C5D1)}(6V) = \frac{1}{2\pi B(R_S + R_L)} = \frac{1}{2\pi(6.046\text{GHz})(24.8 + 50)} = 0.352\text{pF}$$

An S12 measurement for C3D1 is taken for several bias voltages and photocurrents. A composite of these measurements is shown in figure 6.43:

Figure 6.43: Normalized S12 response for C3D1 with a wirebond electrical connection

The S12 response in figure 6.43 has been normalized to 0dB at the frequency of lowest insertion loss, which was 130MHz. This was done to allow the reader to visually classify the S12 responses by bias voltage and photocurrent. In addition, the coupling from lensed fiber to waveguide varies over time, which makes obtaining a consistent transducer gain nearly impossible. However, poor coupling merely results in an offset in the S12 data and does not affect the overall bandwidth response.

Three distinct bandwidth performance regions are observed from figure 6.43. At high bias, 3V or greater, the bandwidth remains relatively large and stays constant across a wide bias range. For biases that are 2V and lower, the S12 response follows a different characteristic, which tends to make the bandwidth sharply smaller. At 0V bias, the bandwidth drops significantly. The S12 response is fairly consistent between the 1V and 2V biases. Above 2.5V, the bandwidth is RC limited. Below 2.5V, the bandwidth is limited by the transit time of the intrinsic Ge region. When operating at 17mA photocurrent and 2.5V

bias, the bandwidth shows an RC limitation. A sharp transition to transit time limited bandwidth is observed when increasing the current from 17mA to 25mA at the same bias (2.5V). This transition from RC limited to transit time limited bandwidth is due to the increasing voltage drop by the series resistance with greater levels of photocurrent. In fact, transit time limited bandwidth is seen even at 1V bias and 25mA photocurrent.

An important observation of the bandwidth behavior in figure 6.43 is the close spacing of the S12 curves at high bias (2.5-6V) and the distant spacing of S12 curves at low bias (0-2.5V). At high bias, junction capacitance is roughly constant. Therefore, the RC time constant should also remain roughly constant, which results in a bandwidth that remains constant with increasing junction potential. When operating in low bias, the junction capacitance increases sharply as the junction potential is reduced. This is best seen by comparing the S12 curve at 1V bias to the curve at 0V bias, which is noticeably steeper. The RC time constant and transit time both partially contribute to the decrease in bandwidth at low bias.

The normalized S12 response for C3D1, shown in figure 6.39, is corrected for the insertion loss of the coaxial cable, Mach Zehnder modulator, and G-S probe. The corrected S12 response is shown in figure 6.44:

Figure 6.44: S12 Measurement for C3D1 with system insertion loss correction

Figure 6.44 shows inductive peaking between 4-6GHz. This results in 2dB gain after the S12 correction.

The bandwidth is measured from figure 6.44 by determining the frequency range between the first data point (130MHz) and to a point -3dB less in magnitude. This is shown in figure 6.45 below:

Figure 6.45: 3dB bandwidth measurement for C3D1 (wirebonded)

The 3dB bandwidth, calculated series resistance, and calculated junction capacitance are summarized in table 6.5 below. The diode reverse bias is calculated by subtracting the voltage drop across R_S :

External Reverse Bias (V)	Current (mA)	Diode Reverse Bias (V)	Bandwidth (GHz)	R_S (Ω)	C_J (pF)
0	3.60	-0.13	0.398	36.94	4.60
1	16.88	0.38	0.463	36.94	3.95
2	24.67	1.09	0.512	36.94	3.58
2.5	25.04	1.58	5.296	36.94	0.35
2.5	16.87	1.88	7.873	36.94	0.23
3	21.27	2.21	8.038	36.94	0.23
4	25.55	3.06	8.032	36.94	0.23
6	29.39	4.91	8.779	36.94	0.21

Table 6.5: Calculated junction capacitance for C3D1

The calculated junction capacitance is plotted versus the diode reverse bias in figure 6.46 below:

Figure 6.46: Junction capacitance calculation for C3D1 showing RC and transit time limited bandwidth

Figure 6.46 shows two distinct bandwidth performance regions. At high bias, the junction electric field is sufficiently high to accelerate electrons across the depletion region with a low transit time. As the bias is decreased, the bandwidth abruptly lowers by nearly 8GHz to approximately 500MHz. This is the result of transit time limiting. The actual junction capacitance is nearly flat until the bias drops to about 0.2V, which is observed in section 6.4 using S11 measurements.

6.4 S11 Measurements

Measurements of the diode input impedance (S11) can reveal information about the junction capacitance using two calculation methods. The first method calculates the capacitance directly using the impedance at low frequency and the following capacitive-reactance equation:

$$X_c = \frac{-1}{2\pi fC}$$

The frequency chosen for low frequency capacitance calculation is 66MHz, which yields a junction capacitance of 0.8983pF at 6V bias. The second method is to directly curve fit the S11 measurement to the model illustrated by Cheng in figure 6.47 below:

Figure 6.47: Discrete component diode model (Source: Chang¹⁰)

Table 6.6 shows the curve fit parameters used to fit the S11 model of C4D1. The contact pad resistance has been included with R_s :

BIAS (V)	Cj (pF)	Rj (MΩ)	Rs (Ω)	Ls (pH)
0.5	1.126	11.6	61.39	1.966
1	1.027	2.13	74.82	140.4
2	0.9707	1.40	68.15	209.7
3	0.9429	42.9	73.15	136.5
4	0.9211	13.3	75.41	120
5	0.9046	3.39	77.18	112.4
6	0.8879	28.3	76.77	95.07
7	0.8759	9.92	81.32	104.60

Table 6.6: S11 curve fit parameters for C4D1

The S11 input impedance of C4D1 at 6V bias is shown in the Smith Chart below. The curve fit (table 6.6) and raw S11 data are superimposed on this chart (figure 6.48):

Figure 6.48: S11 measurement (blue) and curve fit (black) for C4D1 at 6V bias

The curve fit method does not work well at 0V bias. However, it matches the low-frequency reactance calculation very well up to 0.5V. Figure 6.49 compares the junction capacitance result from either method versus bias for C4D1:

Figure 6.49: Junction capacitance versus bias, measurement method comparison (C4D1)

This is repeated for all devices on chips 4 and 5. The junction capacitance for each device is shown in figures 6.50-6.57:

Figure 6.50: Junction capacitance versus bias, measurement method comparison (C4D2)

Figure 6.51: Junction capacitance versus bias, measurement method comparison (C4D3)

Figure 6.52: Junction capacitance versus bias, measurement method comparison (C4D5)

Figure 6.53: Junction capacitance versus bias, measurement method comparison (C4D6)

Figure 6.54: Junction capacitance versus bias, measurement method comparison (C5D1)

Figure 6.55: Junction capacitance versus bias, measurement method comparison (C5D3)

Figure 6.56: Junction capacitance versus bias, measurement method comparison (C5D5)

Figure 6.57: Junction capacitance versus bias, measurement method comparison (C5D6)

The calculated junction capacitance of C5D1 has a discrepancy between the S11 and bandwidth measurement methods. The junction capacitance at 6V is found to be 0.352pF using the device bandwidth and a diode series impedance of 24.8Ω connected to a 50Ω load. Calculating this capacitance using the S11 method yields a result that is nearly double (0.7855pF). This discrepancy could be due to an inaccurate diode series resistance. The contact pad contact resistance changes each time the probe is touched down onto the metal pads. It could be that the contact resistance is greater than the 16Ω resistance accounted for in the diode series impedance. Assuming that the junction capacitance really is 0.7855pF, this yields a contact pad resistance of 24.5Ω .

The S11 curve fit method seems to be the most accurate because the shape of the curve fits better to the expected result. The junction capacitance should remain roughly constant as the frequency is increased. The low-frequency reactance method has a capacitance rolloff above 5V. Regardless of the method, the y-axis is exaggerated to the point where these capacitance differences appear large. In fact,

these methods produce nearly the same capacitance. This is evident when viewing the full junction capacitance characteristic up to 0V bias:

Figure 6.58: Capacitance versus bias, showing low frequency detail; calculated using low-frequency reactance method (C5D3)

The junction capacitance at 6V bias is of particular interest for RF measurements. Table 6.7 shows this capacitance for all devices using both calculation methods:

Device	6V Calculated Junction Capacitance (pF)	
	Low-Frequency Reactance Method	S11 Curve Fit Method
C4D1	0.8983	0.9086
C4D2	0.8699	0.9003
C4D3	0.6037	0.6467
C4D5	0.8826	0.9787
C4D6	0.6968	0.6437
C5D1	0.7479	0.7855
C5D3	0.7068	0.7455
C5D5	1.143	1.181
C5D6	0.7148	0.7431

Table 6.7: Calculated junction capacitance at 6V for all devices on chips 4 and 5

A parallel resonator is used to eliminate the effect of this junction capacitance at a narrowband microwave frequency in chapter 9.

7. RF Power Compression Measurements Using a 50Ω Load

RF power compression occurs for two main reasons: diode series resistance and space charge effects. Each of these factors is mitigated by increasing the applied external bias, but at the expense of increased power dissipation by the device. The high power Silicon waveguide photodetector developed by Yin et al.⁷³ uses a germanium absorber and silicon rib waveguide structure to improve thermal dissipation from the active region and allow operation at higher output power with greater photocurrent.

The voltage swing of the diode is limited by the turn-on voltage at one end and the avalanche breakdown voltage at the other. The maximum obtainable RF power at compression is essentially limited by the turn-on voltage, which is alleviated by operating with a higher external bias voltage. Before and during RF compression, a large photocurrent causes a significant potential drop across the diode series resistance, which starves the depletion region of bias voltage. This has the dual effect of increasing the charge carrier density and removing bias potential, which both contribute to limiting caused by space charge effects.

To test the RF response at high optical powers, an Erbium-doped Fiber Amplifier is needed. Several pre-built optical amplifiers are available in our lab. The best of these amplifiers is capable of 5dBm optical power output at 1550nm with 200mA of laser current. This is not enough optical power to saturate these photodiodes at any meaningful bias level.

We elected to construct another EDFA using a 300mW pump laser (Bookham LC94K74) and a 19 meter long Er-fiber (OFS HP980X), which has an optimized length that was found using the OASiX Erbium fiber simulator available from OFS. This optical amplifier is capable of outputting up to +18.61dBm of optical power under saturation conditions and can saturate our photodiodes for external biases up to 3V. Even more optical power is needed to achieve saturation at higher bias voltages.

To obtain even greater optical power, we use an even higher power EDFA from the University of California Santa Barbara (UCSB). This optical amplifier is capable of output power exceeding 30dBm, which results in 1dB compression for biases up to 6V. The EDFA itself (IPG EAD-2K-C) is capable of up

to 2W of output power, but this level is reduced using a programmable optical attenuator (Agilent 81577A Attenuator Power Control).

This section begins by describing the measurement setup, a review of the pre-built EDFAs available in our laboratory, and the construction of an EDFA. The section concludes with a description of the RF power compression measurements using a standard 50Ω load and a discussion of the results.

7.1 Construction of an Erbium-Doped Fiber Amplifier (EDFA)

Optical amplifiers accept an optical signal as an input, and they output an optical signal with the same wavelength at higher power. Douglas Baney illustrates a block level diagram of an optical amplifier in figure 7.1 below (Derickson¹⁷):

Figure 7.1: Block diagram of an optical amplifier (Source: Derickson¹⁷)

Erbium-doped fiber amplifiers operate by storing laser energy in a metastable state within Erbium atoms that are interspersed throughout a single mode glass fiber, and releasing this energy when light of a similar wavelength and energy as the metastable energy level induces emission of this stored energy.

The pump laser is usually 980nm, which corresponds to the energy level that is just above the metastable state. Alternatively, pump lasers can operate at any energy state above the metastable state, including the following wavelengths: 1480nm, 980nm, 810nm, 660nm, 543nm, 520nm. Baney illustrates the energy band diagram for Erbium, which is shown in figure 7.2 (Derickson¹⁷):

Figure 7.2: Erbium energy band diagram (Source: Derickson¹⁷)

According to Baney, excited Erbium atoms dissipate energy through non-radiative processes, usually the emission of lattice phonons (heat), until the metastable state is reached (Derickson¹⁷, 522). Baney writes, “The tendency to radiate a photon when transitioning to a lower energy level increases with the energy gap” (qtd. in Derickson¹⁷, 522). Baney explains, “the 1480nm pump wavelength is used in EDFAs for a number of reasons including... good power efficiency since there is a small energy difference between 1480nm and 1550nm” (qtd. in Derickson¹⁷, 523). When pumping at 980nm, the Erbium atoms quickly decay non-radiatively to the metastable state within, which takes approximately 2μs, Baney writes. In contrast, Baney emphasizes, decay from the metastable to the ground state has a time constant of roughly 10ms (Derickson¹⁷, 525).

Spontaneous emission from the metastable to ground states has a random phase and direction. Approximately 1% of this spontaneous emission is captured by the optical fiber mode where it can excite Erbium atoms to discharge stored energy. Once one Erbium atom discharges its energy, other Erbium atoms are also stimulated to emit their energy as well. This cascaded energy discharge results in a noise

source called amplified spontaneous emission (ASE). ASE, Baney explains, causes degradation of the amplifier SNR (Derickson¹⁷, 525).

An EDFA design is illustrated by Baney and is shown in figure 7.3 below (Derickson¹⁷, 521):

Figure 7.3: Block diagram of an EDFA (Source: Derickson¹⁷)

Our EDFA is constructed with a 19m length erbium doped fiber, which is the optimum length obtained from an output power simulation using OASiX 4.0, available from Lucent Technologies.

Figure 7.4: Simulation of the optical output power versus Erbium-doped fiber length

The 19m Erbium-fiber length is selected to simultaneously maximize gain while minimizing ASE. A simulation of total ASE versus Erbium-fiber length is shown in figure 7.5 below:

Figure 7.5: Total ASE versus Erbium-fiber length

From figures 7.4 and 7.5, the anticipated output power and total ASE for a 19m length fiber are 107mW and 1mW, respectively. The ASE at 1550nm also shows good noise characteristics with 19m fiber length, which is shown in figure 7.6:

Figure 7.6: ASE at 1550nm versus Erbium-fiber length

An optical filter centered at 1550nm can be used to reduce the total ASE to the local value at this wavelength. According to figure 7.6, the ASE power at 1550nm is approximately 0.02mW.

Our EDFA is constructed using a 300mW pump laser operating at 980nm, a 980/1550nm wavelength-division multiplexer (WDM), a 20% splitter, a 2% splitter, and a 1550nm optical isolator. Our design differs from that shown in figure 7.3 since we only had one isolator available. We excluded the isolator on the input terminal, which did not degrade the performance of the amplifier to any observable degree. The 20% splitter is used directly at input terminal, and the 2% splitter is used after the optical isolator on the output terminal. A block diagram of the constructed EDFA is shown in figure 7.7:

Figure 7.7: EDFA block level diagram

Each component of the EDFA is tested individually prior to being fusion spliced into the system. The insertion loss of the WDM, Isolator, and 2% splitter are 0.16dB, 0.6dB, and 0.16dB, respectively. Fusion splices have an insertion loss of approximately 0.1dB, and the insertion loss of the FC/APC angled fiber connectors is about 0.3dB.

A photograph of the constructed EDFA is shown in figure 7.8 below:

Figure 7.8: Final construction of our Erbium-doped Fiber Amplifier

ASE data is taken for this amplifier at a range of pump laser currents. These correspond to varying levels of pump laser power. This data is shown in figure 7.9. At the maximum rated current of the pump laser, 600mA, the ASE reaches a local maximum of approximately 8dBm at 1530nm.

Figure 7.9: Measurement of ASE for various levels of pump laser current. Pump laser power is 300mW at its rated maximum current of 600mA

The total ASE shown in figure 7.9 has a linear relationship to the pump laser current. This indicates that ASE increases linearly with the laser pump power. A plot of the total ASE versus pump laser current is shown in figure 7.10. This plot indicates that 47mW of total ASE power is produced at the maximum pump laser power.

Figure 7.10: Measured total ASE versus pump laser current

Figure 7.11 shows the measured ASE at 1550nm versus pump laser current:

Figure 7.11: Measured 1550nm ASE versus pump laser current

The total ASE is very high (47mW). Filtering the amplifier output at 1550nm can reduce the ASE to 1mW at the rated maximum pump laser power. This filtering is only necessary for 100% modulated optical input with modulation periods that approach the spontaneous emission time constant for the Erbium metastable state (10ms), which corresponds to frequencies of 100Hz and lower. Higher input modulation frequencies tend to stimulate many of the Erbium atoms to release their energy before spontaneous emission has a chance to occur. This is only an issue for signals with amplitude modulation that tends to lower the input power to very low levels for an amount of time comparable to the metastable time constant. Applying laser power at an appropriately high modulation frequency should eliminate much of this ASE and negate the need for an optical filter.

To test the saturation output power of the EDFA, an optical source with +8.16dBm power at 1550nm is input into the amplifier. The spectral power of this optical source is shown in figure 7.12:

Figure 7.12: Spectral characteristic of the EDFA input signal, showing +8.16dBm power at 1550nm

Figure 7.13 shows the spectral characteristic of the EDFA output signal, with the input shown in figure 7.12 applied:

Figure 7.13: Spectral characteristic of the EDFA output signal, showing +18.61dBm power at 1550nm

The spectral content shown to either side of 1550nm in figure 7.13 is due to the combined effect of the amplified input laser characteristic and the ASE due to the EDFA. The maximum local power of this spectral noise is approximately -15dBm at 1555nm, which is an acceptably low value. Figure 7.13 shows the worst characteristic the EDFA can possibly produce, since the input power is saturating the amplifier. Applying lower input power will likely reduce this side-spectral content to even lower levels.

This EDFA is used to amplify the modulated optical signal and provide adequate power levels to compress the RF output signal from the photodiodes under test. The EDFA is saturated at +18.61dBm of optical power with the highest input power available applied (+8.16dBm). The output tap terminal shows a 15.50dB tap ratio with the power at the output terminal. As explained in section 7.3, this power is only

adequate enough to cause RF compression with diode biases of 3V and lower. To measure the compression characteristics at higher bias, more optical power amplification is needed.

7.2 Oscilloscope Measurement of the Signal Compression

To understand the compression mechanism, the waveform is viewed in an oscilloscope (HP54501A Digitizing Oscilloscope) under RF compression conditions. The coaxial cable leading from the probing station is connected to the 1MΩ terminal of the oscilloscope and a 50Ω load is placed in shunt. An SMA to BNC adapter and BNC t-connector are used to connect the coaxial cable to the oscilloscope and shunt load. An EDFA is used to amplify an incident optical signal with approximately 63% amplitude modulation. The modulation frequency (15MHz) is kept low to make measurement with the oscilloscope easier. Bias is applied to the diode using a bias-T (HP 11612A), which is connected in series with the coaxial cable. This is shown by the block level diagram in figure 7.14:

Figure 7.14: Oscilloscope measurement block level diagram

Figure 7.15 shows two cases of RF output from C2D2 with a time average photocurrent of 10mA and 20mA. The photodiode is reverse biased at 2V for each measurement.

Figure 7.15: RF compression waveform shown for C2D2 with 2V reverse bias and 20mA of photocurrent. No compression is present with 10mA of photocurrent.

The RF response with 10mA photocurrent shows no compression, but the response at 20mA photocurrent shows clipping at the bottom of the waveform, which is responsible for RF power compression of these devices at high optical powers. The output signal at 20mA photocurrent is exactly twice as large as the signal at 10mA, which is the result of the optical modulation being doubled in magnitude by changing the gain of the EDFA. The pump laser current controls the EDFA amplification, which changes the time-average optical power incident on the photodiode.

Clipping of the RF response is due to the combined effect of the diode series resistance, probe contact resistance, voltage dependent responsivity, and space charge effects. The minimum voltage of the trough, where clipping is observed, is -0.578V. This voltage is the result of an AC signal superimposed on the DC bias level, with the DC voltage blocked by the series capacitor present in the bias-T. Therefore, the voltage present across the diode and its series resistance is the superposition of this AC signal and the DC bias. This voltage has a minimum value of 1.422V under these testing conditions, which is the bias voltage (2V) added to the minimum value of the trough voltage (-0.578V). Similarly, a peak voltage of

2.640V is obtained by adding the AC peak value (0.64V) to the bias voltage (2V). The peak and trough currents are obtained by a superposition of a 20mA DC photocurrent and an AC component flowing through a 50Ω load. At the voltage peak, the total current is 7.2mA. During the voltage trough, the total current is 31.6mA.

The series resistance of C2D2 (12.11Ω) reduces the reverse bias voltage available to the active layers of the diode. At the time of the AC peak, 2.55V reverse bias is available to the photodiode. During the AC trough, the photodiode bias is reduced to 1.04V by the series resistance. This voltage rises and falls due to an AC current contribution that depends on an incident amplitude modulated signal. The responsivity at the waveform peak is 0.9864A/W, and the responsivity at the trough is 0.5748A/W. This corresponds to a peak and trough optical input of 7.2mW and 54.9mW, respectively. From these optical inputs, an amplitude modulation index of 62.9% is calculated:

$$\mu = \frac{P_{PK}}{P_{AVG}} = \frac{\frac{P_{trough} - P_{peak}}{2}}{\frac{P_{trough} + P_{peak}}{2}} = \frac{54.9 - 7.2}{54.9 + 7.2} = 62.9\%$$

The analysis procedure is as follows:

1. Determine the voltage drop across the series resistance due to the peak and trough current, and subtract it from the system peak and trough voltages
2. Determine the device responsivity
3. From the current and responsivity, calculate the incident optical power during the peak and trough conditions
4. Calculate the amplitude modulation index and compare with a characterization of the Mach-Zehnder modulator insertion loss versus bias voltage. The DC and AC bias conditions of the Mach-Zehnder modulator are already known.

The Mach-Zehnder modulator is biased at 0V DC with +17dBm AC added by a signal generator operating at 15MHz. The Mach-Zehnder characteristic was shown in figure 6.28. The peak amplitude of a

17dBm AC signal is 2.24V into a 50Ω load. To find the modulation index from the Mach-Zehnder characterization, the following procedure is used:

1. Determine the insertion loss at the peak and trough of the Mach-Zehnder bias waveform ($\pm 2.24\text{V}$).
2. Assume a 1W input power and calculate the output power for each of the insertion losses calculated in step 1.
3. From the hypothetical optical output values, determine the amplitude modulation index.

Assuming a hypothetical 1W input optical power, the Mach-Zehnder could produce maximum and minimum optical powers of 932mW and 164mW, respectively. This corresponds to a theoretical modulation index of 70%:

$$\mu = \frac{P_{PK}}{P_{AVG}} = \frac{\frac{P_{trough} - P_{peak}}{2}}{\frac{P_{trough} + P_{peak}}{2}} = \frac{932 - 164}{932 + 164} = 70.0\%$$

This theoretical modulation index (70%) is nearly 7% higher than the modulation index that was experimentally calculated (63%). This occurs for two reasons. First, the Mach-Zehnder transfer characteristic is not symmetrical around zero volts bias. The optical signal swings lower when biased above 0V than when it is biased below 0V. This tends to exaggerate the bottom half of the signal, which lowers its effective modulation index. To show this, the hypothetical output power at 0V is 560mW, which is 372mW below the maximum power (932mW) and 396mW above the minimum power (164mW). Second, the transfer characteristic is not linear over the bias range of interest ($\pm 2.24\text{V}$). The transfer characteristic shows a 5% deviation from linearity at -2.2V and at 1.7V. This nonlinearity tends to decrease the effective modulation index even further. Therefore, due to non-symmetry and nonlinearity, this 7% difference in modulation index is justifiable.

This oscilloscope measurement indicates that RF compression is limited by the bias voltage. Indeed, decreasing the external bias voltage to 1V reduces the photocurrent at compression to 10mA,

which is half of the compression current at 2V bias. An oscilloscope measurement at 1V bias and the same optical modulation conditions as the prior measurement is shown in figure 7.16:

Figure 7.16: RF compression shown for C2D2 with 1V reverse bias and 10mA of photocurrent.

At 1V bias, the peak and trough occur at 0.32V and -0.287V, respectively, with 10mA of photocurrent. These values are nearly double the values obtained for the peak and trough at 2V bias, which are 0.64V and -0.578V, respectively. At 3V bias and 31mA photocurrent, the peak and trough occur at 0.938V and -0.783V, respectively. The peak and trough voltages at 3V bias are nearly 3 times the values at 1V bias. Clearly, a linear relationship exists between the bias voltage and the RF compression current.

7.3 High Power RF Compression Measurements at 15MHz

The oscilloscope measurement in section 7.2 described the voltage-limited electric RF compression of a 100% amplitude modulated optical signal, as detected by the Intel-UCSB photodiode.

This section explores this RF compression further by testing the device with biases up to 6V and with photocurrents as high as 65mA. To achieve RF compression at biases higher than 3V, an EDFA with higher output power than the one constructed must be used. For this purpose, the Bowers Group at the University of California Santa Barbara (UCSB) has generously offered the use of their lab for these experiments.

The measurement setup is as follows. A 1550nm laser is modulated using a Mach-Zehnder Interferometer. This interferometer was characterized in figures 6.25 and 6.26. The modulator has separate AC and DC bias ports. The DC bias port is connected to a DC power supply, and the AC port is connected to an 8656B HP Signal Generator. This signal generator operates at 15MHz with +17dBm of RF power. The amplitude modulated optical signal is amplified to high optical power levels using an EDFA. From there, the optical signal is routed to a lensed fiber, which focuses the light down to a 3.3 μ m spot size on the silicon waveguide facet. Once inside the silicon waveguide, the light propagates to the photodetector and is nearly completely absorbed. This absorption produces an electrical current in the photodetector, which is proportional to the responsivity and the intensity of the incident optical signal. A 15MHz electrical signal is captured by a ground-signal (G-S) probe, which connects with the pad metallization of the photodetector. Copper-Beryllium is used for the probe material, which does not oxidize in air and provides good electrical contact with the Aluminum pads. A coaxial cable connects the probe to the measurement instrumentation, which is a spectrum analyzer. A bias-T is used to provide DC bias to the photodetector while isolating the voltage source from the AC signal. This arrangement is shown in figure 7.17:

Figure 7.17: Experimental setup for RF compression measurements under high optical power conditions

Figure 7.18 shows a photo of the measurement setup described by figure 7.17:

Figure 7.18: RF compression measurement setup showing the lensed fiber, photodiode, probe, attenuator load, coaxial RF output cable, and photodiode DC bias cable

The diode bias voltage is set to a fixed value. Then, amplitude modulated 1550nm laser light is applied to the silicon waveguide through the lensed fiber, which is subsequently absorbed by the photodetector. The optical power is varied, and the output RF power is measured. This output RF power is plotted with respect to the time-average photocurrent on a log-log plot, which graphically illustrates the compression behavior as the deviation from a straight line. These measurements are conducted for biases ranging from 1V to 6V in 1V increments. A composite plot of the RF measurements for C2D2 is shown in figure 7.19 on the following page. A modulation frequency of 15MHz is chosen because it is very low frequency and should be compatible with the test loads of chapter 8. This measurement uses the 50Ω spectrum analyzer input terminal as a load for the photodetector, and the optical modulation frequency is 15MHz:

Figure 7.19: RF compression composite plot for C2D2 with biases ranging from 1V to 6V (15MHz)

The 1dB compression point is found by projecting the linear slope of the log-log plot past compression and determining the point where the RF power drops from this line by 1dB. This is conducted using an automated MATLAB program, which graphically shows the 1dB compression point by drawing a line that is 1dB long and placing it between the projection line and the measurement response. This is shown in figure 7.20 for C2D2 at 1V bias using the 50 Ω input terminal of the spectrum analyzer as a load for the photodetector:

Figure 7.20: Electrical RF power compression of C2D2 at 1V bias and 15MHz

The optical power is obtained by measuring its value from an optical attenuator (81577A Attenuator Power Control). A characterization of this attenuator reveals that the actual output power is offset by 2dB from the actual display value. For example, if the attenuator display reads 26dBm, then the actual optical power is approximately 24dBm. An optical loss of 6dB is assumed between the lensed fiber and the photodetector. This estimate of the optical power is shown for reference only and is not used for any measurement result in this report.

The bias level tested does not exceed 6V for two reasons. First, the EDFA does not produce enough optical power to reach compression at 7V bias. Second, operation at 7V bias dissipates too much power, which may result in thermal failure. RF compression of C2D2 at 6V bias is shown in figure 7.18:

Figure 7.18: Electrical RF power compression of C2D2 at 6V bias and 15MHz

The RF power compression shown by figure 7.18 illustrates the largest RF power obtained using a 50Ω load for any device tested (15.73dBm). This is confirmed by the data shown in Table 7.1, which catalogs the RF power and photocurrent at compression obtained for four devices on chip #2 at bias levels ranging from 1V to 6V:

	RF Power at Compression (dBm)			
Bias (V)	C2D1	C2D2	C2D5	C2D6
1	-3.96	-3.62	-5.12	-4.43
2	4.1	4.41	3.33	3.5
3	8.77	8.95	7.66	8.61
4	12.02	11.98	10.57	11.64
5	13.91	14.09	12.46	13.61
6	15.57	15.73	14.01	15.09
	Compression Current (mA)			
Bias (V)	C2D1	C2D2	C2D5	C2D6
1	6.4	7.07	5.89	6.35
2	16.17	17.36	16.07	15.58
3	27.78	29.29	25.01	28.27
4	40.42	41.65	35.20	40.20
5	50.32	53.31	43.24	50.26
6	62.14	65.38	51.74	60.70

Table 7.1: RF Compression Measurements using the input terminal of a spectrum analyzer as a 50Ω load

The compression measurement discussed in chapter 8 endeavors to increase the maximum RF power through the use of higher load impedance. Indeed, the maximum RF power obtained using a 50Ω load (15.73dBm) is increased to 17.83dBm using a 100Ω load. This is discussed in further detail in chapter 8.

8. Compression Measurements Using Higher, Non-Standard Impedances

Operation using a 50Ω load is not necessarily required. Although many RF systems use 50Ω as a de-facto standard, circuit designers are not necessarily required to use this impedance. Many loads types can have impedances that are different from 50Ω. An antenna is one such example. Half-wave dipole antennas are usually fashioned with slightly shorter leads than a half wavelength in order to create a real input impedance of 70Ω. Transmission line quarterwave transformers are then used to transform a 50Ω impedance to a higher value.

Photodiodes are modeled as a constant current source with an output current proportional to the incident optical power through the optical responsivity coefficient. Power output is proportional to the load resistance multiplied by the square of the current. Therefore, as the load impedance increases, the output power should also increase:

$$P = I^2 R$$

According to this equation, if the load impedance is doubled then the output RF power should also double, which is an increase of 3dB. This hypothesis is directly observed and validated by our RF power measurements. We do not see any benefit in operating these devices with a load resistance value less than 50Ω.

The load impedance cannot be increased indefinitely. An optimum value exists that is limited by the voltage drive of the diode. Our photodiodes experience avalanche breakdown at a voltage slightly higher than 12V. If these diodes are biased at 6V, this allows a maximum ideal voltage swing of 12V peak-to-peak centered on 6V. Increasing the load impedance will result in power gains up until this condition. Any further increase in load impedance will effectively saturate the photodiode and cause clipping in the waveform.

Compression measurements are conducted at 15MHz to provide proof of principle evidence for this hypothesis. The highest RF power obtained at 1dB compression and 15MHz was +17.83dBm with a compression current of 54.57mA using a 100Ω load. In addition, an increase of 3dB and 5.5dB in RF

power were obtained by switching from a 50Ω load to 100Ω and 177Ω loads, respectively, at similar photocurrents. Using load impedances greater than 50Ω has the effect of increasing the RF output power without the need for additional photocurrent and more optical power. Operation at lower photocurrent, by using load impedances greater than 50Ω , improves ohmic power dissipation in the diode, which lowers heat generation and allows operation at higher reverse bias before thermal failure.

To conduct these measurements, test loads must be constructed for 15MHz operation at the following impedances: 50Ω , 100Ω , 177Ω . This is described in section 8.1.

8.1 Low Frequency Load Construction

We seek to determine the RF compression characteristics of these photodiodes with loads that are different from 50Ω . To do this, attenuator loads with input impedances of 50Ω , 100Ω , and 177Ω are constructed using surface-mount resistors and capacitors soldered onto a copper-clad PC board.

A resistor pi-network attenuator design is used to construct these loads. SMA connectors are soldered to the edge of the PC board to allow input and output to instrumentation. A radio frequency choke (RFC) is constructed between the diode terminal and the bias connector. The bias voltage is supplied using an SSMB connector. A schematic diagram of this circuit is shown in figure 8.1:

The resistor network raises the input impedance of the measurement instrumentation, as seen from the photodiode.

Figure 8.1: Pi- network attenuator used to construct the 50Ω, 100Ω, and 177Ω low frequency loads

Before values for R can be chosen, the input impedance of the RFC must be found. Several RFC choke combinations are constructed using ferrite beads (Lead EMIFIL Inductor Type BL01) and low-Q, ferrite core coils, which are analyzed using the VNA for frequencies up to 3GHz. The RFC is terminated by an AC short using two or three 0.1μF capacitors arranged in parallel. These RFCs are shown in figure 8.2:

Figure 8.2: Various RFC designs containing (a) all ferrite bead chokes and (b) combinations of two or three low-Q ferrite core inductors in series with three ferrite bead chokes

RFC designs are rated by impedance magnitude and rolloff of impedance with increasing frequency. S11 data is collected for these chokes up to 3GHz and the impedance is plotted. Figures 8.3 through 8.11 show the S11 data and vector impedance for each RFC design:

Figure 8.3: (a) Impedance and (b) S11 Smith Chart for the 9 ferrite bead RFC

(a)

(b)

Figure 8.4: (a) Impedance and (b) S11 Smith Chart for the 6 ferrite bead RFC

(a)

(b)

Figure 8.5: (a) Impedance and (b) S11 Smith Chart for the 3 ferrite bead RFC

(a)

(b)

Figure 8.6: (a) Impedance and (b) S11 Smith Chart for the 1 ferrite bead RFC

(a)

(b)

Figure 8.7: (a) Impedance and (b) S11 Smith Chart for the 1 ferrite coil RFC

(a)

(b)

Figure 8.8: (a) Impedance and (b) S11 Smith Chart for the 2 ferrite coil RFC

(a)

(b)

Figure 8.9: (a) Impedance and (b) S11 Smith Chart for the 1 ferrite coil and 3 ferrite bead RFC

Figure 8.10: (a) Impedance and (b) S11 Smith Chart for the RFC design with 2 ferrite coils and 3 ferrite beads. This design is chosen as the RFC for the circuit illustrated in figure 7.15.

From these tests, it appears that the RFC design with 2 ferrite coils and 3 ferrite beads shows the greatest impedance value at 15MHz (906Ω). The ferrite beads appear to lower the frequency of the impedance peak, which is desired in this application. The impedance of this design peaks at 50MHz (1.6kΩ). However, the input impedance into this RFC drops sharply at frequencies greater than 100MHz. At 100MHz, the impedance is 900Ω, and at 300MHz it drops to 330Ω. Since the addition of the pi-network attenuator will likely spread out the high impedance response of the RFC to lower frequencies, the peak input impedance (1.6kΩ) is chosen as the design impedance.

An adjustment to the 2 ferrite coil and 3 ferrite bead design involves placing a small wire loop placed just before the inductive components and immediately after the SMA connector. This wire loop is directed normal to and away from the copper board, which effectively acts like a transmission line with increased characteristic impedance. The S11 data and impedance are shown in figure 8.11:

Figure 8.11: (a) Impedance and (b) S11 Smith Chart for the RFC design with 2 ferrite coils and 3 ferrite beads with an input wire loop. Shows the highest low frequency impedance, but has poor impedance at moderately higher frequencies above 100MHz

Although the design in figure 8.11 shows the highest impedance at 15MHz (1.1k Ω), it shows very poor impedance at moderately higher frequencies. At 100MHz, the impedance drops to 500 Ω . At 300MHz, this impedance drops to 200 Ω . Even though these loads are designed for a fundamental frequency of 15MHz, good high frequency impedance is desired to route more of the harmonic content toward the measurement instrumentation instead of dissipating it in the RFC. For this reason, the RFC design shown in figure 8.10 is chosen as the final RFC design, which contains 2 ferrite coils, 3 ferrite beads, and no loop. Table 8.1 summarizes the low frequency input impedance for each RFC design:

RFC Design	Impedance (Ω)	RFC Design	Impedance (Ω)
9 Ferrite Beads	400	1 Ferrite Coil	1000
6 Ferrite Beads	300	2 Ferrite Coils	1600
3 Ferrite Beads	175	1FC + 3FB	900
1 Ferrite Bead	50	2FC + 3FB	1600
2FC + 3FB + Loop	1600		

Table 8.1: RFC impedances obtained for each design

The design resistance (R), shown in figure 7.15, is selected for each attenuator by considering three equivalent impedances in parallel: the design resistor (R), the parallel combination of the 50 Ω

instrumentation impedance and the 50Ω resistor all arranged in series with a 300Ω resistance (325Ω), and the low frequency impedance of the RFC ($1.6k\Omega$). This is illustrated mathematically in the equation shown below:

$$Z_{eq} = [(50\Omega || 50\Omega) + 300\Omega] || 1600\Omega || R$$

Table 8.2 shows the value of the resistor (R) selected for each of the design impedances (177Ω , 100Ω , and 50Ω) and the equivalent impedance calculator for that resistor value:

Design Impedance (Ω)	Resistance Selected for R (Ω)	Equivalent Impedance (Ω)
177	511	176.7
100	158	99.7
50	62	50.4

Table 8.2: Resistor selection for each desired input impedance of the pi-network attenuator

Before connecting the RFC choke, the input impedance of the pi-network attenuator is verified for the 177Ω design. The input impedance results from 325Ω placed in parallel with 511Ω , which yields an equivalent resistance of 198.7Ω . An S11 measurement reveals that the impedance of this arrangement (200Ω) is very close to the predicted impedance (198.7Ω). The S11 input impedance seen from the device port and the S11 Smith Chart is shown in figure 8.12 below:

Figure 8.12: (a) Input impedance and (b) S11 Smith Chart of the pi-network attenuator seen from the device port without the RFC choke connected

A shunt 50Ω resistor is designed into the pi-network attenuator to minimize reflections at the instrumentation port. Figure 8.13 shows the S22 Smith Chart and input impedance of the pi-network attenuator instrumentation port without the RFC attached:

Figure 8.13: (a) Input impedance and (b) S22 Smith Chart of the pi-network attenuator seen from the device port without the RFC choke connected

With the RFC choke attached, the final 177Ω attenuator design shows a measured equivalent impedance of 178Ω at 15MHz. This is shown by the S11 Smith Chart and input impedance plot shown in figure 8.14:

Compression Measurements Using Higher, Non-Standard Impedances

Figure 8.14: (a) Input impedance and (b) S11 Smith Chart of the final 177Ω pi-network attenuator design with the RFC attached

Similarly, the input impedance and S11 Smith Chart are shown for the 100Ω and 50Ω loads in figures 8.15 and 8.16:

Figure 8.15: (a) Input impedance and (b) S11 Smith Chart of the final 100Ω pi-network attenuator design

Compression Measurements Using Higher, Non-Standard Impedances

Figure 8.16: (a) Input impedance and (b) S11 Smith Chart of the final 50Ω pi-network attenuator design

Table 8.3 shows the calculated and measured input impedance for each of the attenuator designs:

Design Impedance (Ω)	Calculated Impedance (Ω)	Measured Impedance (Ω)
177	176.7	177.9
100	99.7	100.2
50	50.4	50.86

Table 8.3: Calculated and measured input impedance for each attenuator design

Figure 8.17 shows a photo of the 100Ω attenuator load. The bias SSMB connector is shown attached to the RFC choke, and two SMA connectors serve the 100Ω diode port and the 50Ω instrumentation port.

Figure 8.17: 100Ω pi-network attenuator load and circuit diagram

Figure 8.18 shows the 177Ω load:

Figure 8.18: 177Ω pi-network attenuator load

The purpose of the attenuator networks is to provide controlled input impedance to the diode port while allowing measurement of the RF power at the diode through the instrumentation port. The electrical distance between the photodetector and the resistor pi-network is minimized to avoid transmission line effects. Power at the diode port is proportional to the power at the instrumentation port through several current divider equations.

The current flowing through the instrumentation port is determined by the measured RF power:

$$P_{meas} = I_{50}^2 50$$

$$I_{50} = \sqrt{\frac{P_{meas}}{50}}$$

Since this current is effectively doubled by the 50Ω resistor placed in parallel, the current flowing through the 300Ω resistor is twice that flowing through the instrumentation port:

$$I_{300} = 2I_{50} = 2\sqrt{\frac{P_{meas}}{50}} = \sqrt{\frac{P_{meas}}{25}}$$

The voltage across the parallel combination of two 50Ω resistors in series with the 300Ω resistor (V_{325}) is found using Ohm's law:

$$V_{325} = I_{300} 325 = 325 \sqrt{\frac{P_{meas}}{25}}$$

The power seen by the diode is found using this voltage and the equivalent parallel resistance (R_{eq}) of the attenuator and RFC:

$$P_{diode} = \frac{V_{325}^2}{R_{eq}} = \frac{325 \sqrt{\frac{P_{meas}}{25}}}{R_{eq}}$$

The equivalent resistance is represented by the following equation:

$$R_{eq} = [(50\Omega || 50\Omega) + 300\Omega] || 1600\Omega || R$$

$$R_{eq} = 270.1\Omega || R$$

The equivalent resistance obtained for each attenuator was described in table 8.3. Using these equivalent resistances, an equation relating the power transmitted at the diode port to the power received at the instrumentation port can be shown. This power relation for the 177Ω load is the following:

$$P_{diode,177\Omega} (mW) = 47.7 P_{meas,177\Omega} (mW)$$

$$P_{diode,177\Omega} (dBm) = 16.79dB + P_{meas,177\Omega} (dBm)$$

Similarly, the power relation for the 100Ω load is as follows:

$$P_{diode,100\Omega} (mW) = 84.5 P_{meas,100\Omega} (mW)$$

$$P_{diode,100\Omega} (dBm) = 19.27dB + P_{meas,100\Omega} (dBm)$$

Finally, the power relation for the 50Ω load is calculated:

$$P_{diode,50\Omega} (mW) = 169 P_{meas,50\Omega} (mW)$$

$$P_{diode,50\Omega} (dBm) = 22.28dB + P_{meas,50\Omega} (dBm)$$

These attenuator loads are used for low frequency RF compression testing, which is described in the following subsection.

8.2: High Power RF Compression Measurements at 15MHz

The pi-network attenuator loads are used to test the RF compression outputs at different load resistances than 50Ω. A composite plot of the RF power versus log-magnitude current for C2D1 is shown in figure 8.19, which uses the 177Ω pi-network attenuator as a load.

Figure 8.19: RF compression composite plot for C2D1 using a 177Ω pi-network attenuator load

RF compression plots are generated for each of the pi-network attenuator loads (50Ω, 100Ω, and 177Ω). This is repeated for each of the five devices on chip 2. The best data collected was for C2D2, which showed a 3dB increase in RF power when transitioning from the 50Ω to 100Ω loads and a 2.5dB increase between measurements with the 100Ω and 177Ω loads. This decibel ratio can be found using the following equation:

$$dB = 10 * \log\left(\frac{R_2}{R_1}\right)$$

$$dB_{50\Omega \rightarrow 100\Omega} = 10 * \log\left(\frac{100}{50}\right) = 3dB$$

$$dB_{50\Omega \rightarrow 177\Omega} = 10 * \log\left(\frac{177}{50}\right) = 5.5dB$$

RF compression data for C2D2 is shown in figure 8.20:

Figure 8.20: RF Power versus the photocurrent at compression for several loads (C2D2)

This result shows that additional RF power can be extracted from these photodetectors, without increasing the optical power and photocurrent level, by using larger impedance loads. This gives the advantage of increasing RF power extraction without raising the level of DC power dissipated by the time-average photocurrent. Although more power is extracted at similar photocurrent levels, this is not true of similar reverse biases.

The photodiode is modeled by a constant current source with a photocurrent that is proportional to the input optical signal magnitude. Therefore, doubling the load resistance should also double the RF power, assuming that the photocurrent remains constant. Figure 8.20 confirms this result with an increase in RF power of 3dB between the 50Ω and 100Ω loads. Furthermore, a 5.5dB increase in RF power is obtained by switching from a 50Ω load to a 177Ω load. These results are consistent with theory.

Table 8.4 shows the RF compression data illustrated in figure 8.20:

<u>Spectrum Analyzer Load (50ohm)</u>		
Bias (V)	Compression Current (mA)	RF Power at Compression (dBm)
1	7.07	-3.62
2	17.36	4.41
3	29.29	8.95
4	41.65	11.98
5	53.31	14.09
6	65.38	15.73
<u>177Ω Load</u>		
Bias (V)	Compression Current (mA)	RF Power at Compression (dBm)
1	5.59	0.23
2	11.92	7.06
3	19.78	11.4
4	27.94	14.19
5	33.82	15.94
6	42.81	17.67
<u>100Ω Load</u>		
Bias (V)	Compression Current (mA)	RF Power at Compression (dBm)
1	6.65	-0.64
2	14.62	6.61
3	24.4	11.03
4	34.78	14.14
5	44.28	16.18
6	54.57	17.83
<u>50Ω Load</u>		
Bias (V)	Compression Current (mA)	RF Power at Compression (dBm)
1	7.56	-2.33
2	17.55	5.18
3	30.95	10.02
4	42.29	12.85
5	56.37	15.19
6	70.39	17.07

Table 8.4: RF compression data for C2D2

A 0.75dB offset between the measurements using the 50Ω pi-network attenuator load and the 50Ω spectrum analyzer terminal load is observed, which is likely due to changes in the measurement setup between the two load types. A bias-T is used when the spectrum analyzer measurement terminal is used as a load. This bias-T is not needed during the pi-network attenuator measurements. The extra

impedance may be due to a DC blocking capacitor placed in series with the spectrum analyzer. This capacitor is not present during measurements with the spectrum analyzer load because the bias-T already contains a DC block.

This result occurs for other devices as well. RF compression data is illustrated for C2D4 in figure 8.21 below:

Figure 8.21: RF Power versus the photocurrent at compression for several loads (C2D4)

Like the compression data for C2D2, the result for C2D4 shows a 3dB increase in RF power when switching from the 50Ω to 100Ω load and a 2.5dB increase in RF power when switching from the 100Ω to 177Ω load.

C2D1 shows a similar result. This is shown in figure 8.22:

Figure 8.22: RF Power versus the photocurrent at compression for several loads (C2D1)

C2D1 was destroyed during compression measurements with the 50Ω pi-attenuator load. The device failed after 1dB compression was obtained. A photocurrent of 90mA triggered the current compliance on the Keithley voltage source, which was sustained for several seconds. Nevertheless, the data shown in figure 8.21 for the 100Ω and 177Ω loads shows a 2.5dB offset, which is consistent with the expected results.

C2D5 and C2D6 also show data that is consistent with the expected results. The RF compression data for C2D5 is shown in figure 8.23 and the data for C2D6 is shown in figure 8.24:

Figure 8.23: RF Power versus the photocurrent at compression for several loads (C2D5)

Figure 8.24: RF Power versus the photocurrent at compression for several loads (C2D6)

A severe offset is seen with the 50 Ω pi-network attenuator load in figures 8.23 and 8.24. This offset is likely due to a cold solder joint, which opened partially during testing with C2D5 and continued to exhibit this behavior during testing with C2D6. Cold solder joints result from a poor solder connection to one or more pads, which give poor mechanical strength to the solder joint. Even with this poor mechanical performance, the solder joint may still make intermittent electrical contact. This intermittent behavior tends to vary the resistance of the solder joint depending on the quality of electrical contact at any point in time. The measurements in figures 8.23 and 8.24 indicate increased load impedance in the 50 Ω pi-network attenuator, which is best explained by a cold solder joint. Nevertheless, the 100 Ω and 177 Ω loads show a 2.5dB offset, which is consistent with theory and prior measurements.

Table 8.5 shows the maximum RF power obtained for each device:

Device	Width (μm)	Length (μm)	Load (Ω)	RF Power (dBm)	Compression Current (mA)
C2D1	5.8	100	100	16.48	47.04
C2D2	5	250	100	17.83	54.57
C2D4	4.4	250	100	16.41	47.63
C2D5	4.4	500	100	16.66	48.46
C2D6	4.4	250	100	17.71	55.75

Table 8.5: Maximum 1dB compressed RF power at 6V reverse bias and 15MHz modulation

From table 8.5, C2D2 appears to have the highest output RF power (+17.83dBm) with a photocurrent of 54.57mA at 6V bias using a 100 Ω load. C2D6 yields a similar result (+17.71dBm) with a photocurrent of 55.75mA at 6V bias using a 100 Ω load. In comparison, Ramaswamy⁵¹ was able to extract a maximum RF power of 14.35dBm and a compression current of approximately 60mA at 1GHz modulation and 8V reverse bias with a 50 Ω load. This result (17.83dBm) is over 3dB greater than the result obtained by Ramaswamy⁵¹ (14.35dBm) using similar photodiodes. This is expected since the load impedance in our experiment is doubled, and the modulation frequency (15MHz) is much lower than that used by Ramaswamy⁵¹ (1GHz).

Chapter 9 conducts similar measurements using a quarterwave transformer to provide a 100 Ω impedance at microwave frequency. The results of this measurement are expected to be similar.

9. Microwave-Frequency Power Compression

Measurements at 15MHz provided proof of principle evidence for increased RF power extraction by raising the input impedance at low frequencies. This chapter outlines the method that could be used to perform the same RF compression measurements as chapter 8 except at microwave frequencies. Thin film quarterwave transformers are used to provide a 100Ω load at the reference plane of the photodiode for two microwave frequencies: 3GHz and 7GHz.

9.1 Microwave Frequency Load Design

Quarterwave transformers are used to create a set of 100Ω loads for testing the 1dB power output compression at 3GHz and 7GHz, which are in the microwave frequency range. To match a 50Ω load to a 100Ω input impedance, a quarter-wavelength long transmission line section with characteristic impedance equal to the geometric mean of these impedances is used. This is shown algebraically below:

$$Z_0 = \sqrt{R_{in}R_L} = \sqrt{(100)(50)} = 70.71\Omega$$

Similar to the 50Ω coplanar waveguide used for C3D1, the 100Ω quarter-wavelength transformers are constructed from a gold thin film coplanar waveguide that is electrically connected to the aluminum pads of the diode by a gold wirebond. An SMA connector is electrically connected to the other side of the thin film chip with a bead of silver filled epoxy. Chip capacitors are secured on top of the group plane on both sides of the center conductor and then wirebonded to the 100Ω terminal. The purpose of these chip capacitors is to provide an AC short to the ground terminal, which serves as a connection point for a gold wire. This wire has an inductance that is approximately 0.7nH/mm, which parallel resonates with the junction capacitance of the diode. Recall from figure 6.35 that the junction capacitance varies from 0.3pF to 1.1pF. The parallel resonance frequency should be close to the design frequency. A precise resonance frequency can be found during measurements by tuning to the modulation frequency with the highest power output.

A diagram showing the thin film coplanar chip, diode chip, SMA block, chip capacitors, and wirebond sites is shown in figure 9.1 below:

Figure 9.1: Microwave measurement setup showing the positions of the 100Ω quarter-wavelength thin film coplanar waveguide, photodetector, and SMA connector

Quarterwave transformers are designed using a thin film coplanar waveguide for operation at 3GHz and 7GHz. Table 9.1 outlines the parameters used to simulate these coplanar waveguides using ADS LineCalc's CPWG module¹:

Input Parameters		3GHz	7GHz
Parameter	Description	Value	Value
L	Waveguide length	407 mil	174 mil
W	Center conductor width	9 mil	9 mil
G	Separation between the center conductor and ground plane	23.9 mil	23.9 mil
H	Substrate thickness	25 mil	25 mil
T	Metallization thickness	0.125 mil	0.125 mil
ϵ_r	Relative permittivity	9.8	9.8
μ_r	Relative permeability	1	1
Cond	Conductor conductivity	4.52E7	4.52E7
TanD	Loss Tangent	0.0002	0.0002
f	Frequency of operation	3GHz	7GHz
Simulation Output		3GHz	7GHz
Parameter	Description	Value	Value
Z_0	Characteristic impedance	70.63 Ω	70.63 Ω
E_Eff	Effective electrical distance (degrees)	90.03	89.80
K_Eff	Effective dielectric constant	5.844	5.843
A_dB	Insertion loss	0.056 dB	0.033 dB
SkinDepth	Center conductor skin depth	0.053 mil	0.035 mil

Table 9.1: Coplanar waveguide simulation parameters for Agilent's ADS LineCalc CPWG module¹

The design methodology is as follows. A metallization-free padding of 8 mil is given to the outer rim of the waveguide chip to electrical contact with the silver filled epoxy joining the sidewalls of the waveguide chip to the brass block. 10 mil of metallization fills in the padding to either side of the SMA terminal to connect the shield of the coaxial cable to the ground planes of the coplanar waveguide. Silver filled epoxy is used to electrically connect the ground places and the center conductor to the SMA port. Separation between the ground planes and the center conductor is widened at the 50 Ω port to fit the dimensions of the center pin of the SMA connector. 60 mil of 50 Ω impedance waveguide follow this section to provide good load impedance for the quarterwave transformer. Tapers with 4 mil width are used to reduce electrical reflections at the impedance step sites. A 70.71 Ω section follows the 50 Ω section to provide an effective impedance of 100 Ω at the diode port.

Diagrams of the 3GHz and 7GHz thin film coplanar quarterwave transformers are shown in figures 9.2 and 9.3, respectively:

Figure 9.2: 3GHz quarter wavelength transformer matching 50Ω to 100Ω

Figure 9.3: 7GHz quarter wavelength transformer matching 50Ω to 100Ω

9.2 High Power Microwave Compression Measurements at 3GHz and 7GHz

The quarterwave transformers illustrated in figures 9.2 and 9.3 could be used as higher-impedance loads to measure the 1dB compression output power of the photodiode at 3GHz and 7GHz, respectively. Future research in this area may reveal whether an improvement in RF power can be obtained at microwave frequencies by adjusting the load impedance.

The test fixtures are prepared using the following procedure:

1. Prepare brass blocks with dimensions that fit the thin film quarterwave transformer and waveguide photodetector chips.
2. Position the chips on top of the brass blocks and apply silver filled epoxy to secure the chips in place. Bolt the brass blocks to a brass SMA block and use silver filled epoxy to connect the center and outer conductors to the thin film coplanar waveguide, as shown in figure 9.1.
3. Gold wirebond the signal and ground terminals of a single waveguide photodetector to the corresponding terminals of the thin film coplanar waveguide.

The operating frequency of these test fixtures can be found by tuning the modulation frequency until the highest RF output power at 1dB compression is obtained. At the resonant frequency the quarterwave transformer offers its peak input impedance to the photodetector, which results in an output RF power maximum at that frequency. A simulation of the 3GHz and 7GHz quarterwave transformers from figures 9.2 and 9.3 are shown in figure 9.4 on the following page:

Figure 9.4: Quarterwave transformer simulation showing a real input impedance of 100Ω at (a) 3GHz and (b) 7GHz using a 50Ω load and a 70.71Ω transformer characteristic impedance

After testing the 1dB RF compression power and compression current with this setup, we will attempt to eliminate the junction capacitance by placing a shunt inductor in parallel with the diode. Figure 9.1 shows this inductor as an elliptical wirebond from the signal plane to the ground plane at the diode terminal. A chip capacitor is placed at the ground plane wirebond site for DC blocking. This inductor has a value that is tuned to the frequency of operation (3GHz or 7GHz) through the LC resonance formula:

$$f_o = \frac{1}{2\pi\sqrt{LC}}$$

C5D1 is chosen for the 3GHz compression measurements. It has a junction capacitance of 0.7855pF, which corresponds to a resonant inductance of 3.58nH. At 3GHz, the impedance of this inductor is the following:

$$Z_L = j\omega L = j * 2 * 3.14159 * 3GHz * 3.58nH = j67.5\Omega$$

Likewise, C4D3 is chosen for the 7GHz compression measurements. Its junction capacitance is 0.6457pF, which corresponds to a resonant inductance of 0.8nH. The inductor impedance is the following:

$$Z_L = j\omega L = j * 2 * 3.14159 * 7GHz * 0.8nH = j35.2\Omega$$

A 7-segment transmission line arranged in tandem is modeled in MATLAB to select an appropriate wirebond length. The shape of the wirebond is assumed to be an ellipse that is twice as long as it is tall. The height of each segment is the average height of each of the segment extremes. This is shown in figure 9.5 below:

Figure 9.5: 7-Segment transmission line approximation of an elliptical wirebond

Hall²⁵ describes the characteristic impedance (Z_0) of a microstrip transmission line of arbitrary height above a ground plane, which is shown below. These formulas are accurate so long as the conductor width (W), height (H), and effective relative permittivity (ϵ_r) obey the following conditions: $0.25 \leq W/H \leq 6$ and $1 < \epsilon_r < 16$. Since the dielectric surrounding the wirebond is air, the effective relative permittivity is very close to unit and can be ignored. Assuming that $\epsilon_r = 1$ and $W/H \leq 1$, the equations described by Hall²⁵ reduce to the following:

$$Z_0 = \sqrt{\mu_0 \epsilon_0} \frac{1}{C_a}$$

$$C_a = \frac{2\pi\epsilon_0}{\ln(8H/W + W/4H)}$$

The 7-segment transmission line is modeled by cascading the load impedance through the chain of transmission lines. Considering a two transmission line example, the input impedance of the transmission line toward the load serves as the load impedance of the transmission line arranged toward

the generator. In our simulation, the chip capacitor is assumed to be an AC short circuit and will serve as the initial load to the transmission line. The load impedance (Z_L) is transformed through each transmission line segment using the following equation:

$$Z(-l) = Z_0 \frac{Z_L \cos(\beta l) + jZ_0 \sin(\beta l)}{Z_0 \cos(\beta l) + jZ_L \sin(\beta l)}$$

The input impedance versus wirebond length of an elliptical wirebond loop is shown in figure 9.6:

Figure 9.6: 7-segment transmission line model showing the wirebond length needed for C5D1

This simulation yields a wirebond length of 142mil for an input impedance of $j67.32\Omega$, which is very close to the value needed to parallel resonate the junction capacitance in C5D1. Likewise, the simulation for C4D3 yields a wirebond length of 84 mil with an input impedance of $j35.11\Omega$.

If two parallel inductors are desired then the input impedance for each should be doubled. The result is a wirebond length of 246mil for C5D1 and a wirebond length of 147mil for C4D3. These results are summarized in table 9.2:

Device	Wirebond Length of a Single Inductor (mil)	Wirebond Lengths of two Parallel Inductors (mil)
C4D3	84	147
C5D1	142	246

Table 9.2: Wirebond length selection summary

As seen in figure 9.4, the operating frequency can be tuned several hundred MHz. This makes it possible to tune the modulation frequency to simultaneously achieve high RF power while operating at the LC resonant frequency. Microwave-frequency RF compression measurements can then be made without the influence of the junction capacitance, which tends to sink a large portion of the RF signal to ground.

These Microwave RF compression measurements are an example of future research that can be done with these devices.

10. Summary

As VLSI architectures become smaller with limited real estate for copper interconnects, optical interconnects will serve as an important alternative data transmission medium in future telecommunications technology. Optical interconnects do not suffer from crosstalk and can have adjacent channels that are very closely packed, with a physical separation on the order of 1-3.5 μm . Recent advances in monolithic integration of photodetectors on silicon have widened the possibility for achieving scalable, low-cost, on-chip optical interconnects.

A new Si-Ge PIN photodiode has been developed by Yin et al.⁷³ that is fabricated using only standard CMOS processes. This photodetector is capable of high power output, as much as 17.07dBm with 70.39mA photocurrent into a 50 Ω load. By increasing the load resistance from 50 Ω to 177 Ω , we are able to improve RF power extraction by 5.5dB for similar photocurrent levels. Using a 100 Ω load impedance, we were able to extract up to 17.83dBm of RF output power. This testing is performed at 15MHz modulation to provide a proof-of-concept for improved RF power extraction using load impedance optimization. High electrical output power is an important characteristic for a low-cost, low-noise photodetector because it allows for elimination of the post amplifier that is usually present to drive logic circuitry. Eliminating this amplifier significantly improves the noise figure for optical receivers.

In addition to high power output, these photodetectors also demonstrate large 3dB bandwidth. Testing by Yin⁷³ revealed a bandwidth of 31GHz for devices with very small cross sectional area. The bandwidth of these detectors is essentially limited by the RC time constant at high bias, and is limited by the transit time of photocarriers in the active region under low bias. Vector network analyzer measurements revealed bandwidths between 3GHz and 10GHz.

Input impedance measurements provide information about the junction capacitance, which is around 0.6pF-1pF at 6V bias. This shunt capacitance is an important consideration for microwave-frequency measurements. A method has been described in chapter 9 to eliminate the effect of this shunt capacitance through parallel resonance with an inductor. This inductor is realized by a wirebond

Summary

connection to a chip capacitor, which is connected to the ground plane of the thin film coplanar waveguide using silver filled epoxy. Quarter-wavelength coplanar waveguides are designed to transform a 50Ω impedance into 100Ω for microwave-frequency testing of the RF power extraction improvement using higher load impedance. Future research goals have been summarized to perform these RF power extraction measurements at the microwave frequencies of 3GHz and 7GHz.

Bibliography

- [1] Agilent Corporation. "CPWG (Coplanar Waveguide with Lower Ground Plane)." 2008. 21 May 2010. <[<http://edocs.soco.agilent.com/display/ads2008U2/CPWG+\(Coplanar+Waveguide+with+Lower+Ground+Plane\)>](http://edocs.soco.agilent.com/display/ads2008U2/CPWG+(Coplanar+Waveguide+with+Lower+Ground+Plane))>
- [2] Alferov, Zh. I. "The history and future of semiconductor heterostructures." American Institute of Physics. Semiconductors 32.1 (1998): 1-14.
- [3] Alles, M., et al. "Distributed Velocity-Matched 1.55 μ m InP Travelling-Wave Photodetector for Generation of High Millimeterwave Signal Power." Microwave Symposium Digest. 3 (1998): 1233-1236.
- [4] Altera Corporation. "Using 10-Gbps Transceivers in 40G/100G Applications." White paper. 5 May 2010. Feb 2010 < <http://www.altera.com/literature/wp/wp-01080-stratix-iv-gt-40g-100g.pdf>>.
- [5] Ang, Kah-Wee, et al. "Low Thermal Budget Monolithic Integration of Evanescent-Coupled Ge-on-SOI Photodetector on Si CMOS Platform." Selected Topics in Quantum Electronics. 16.1 (2010): 106-113.
- [6] Beling, A., et al. Impact of Voltage-Dependent Responsivity on Photodiode Non-Linearity. Proc. of IEEE Lasers and Electro-Optics Society (LEOS) Conf., 9 Nov 2008 21 May 2010. Acapulco: IEEE, 2008.
- [7] Beling, A., et al. "High-Speed Miniaturized Photodiode and Parallel-Fed Traveling-Wave Photodetectors Based on InP." Selected Topics in Quantum Electronics. 13.1 (2007): 15-21.
- [8] Beling, A., et al. "High-Power Monolithically Integrated Traveling Wave Photodiode Array." IEEE Photonics Technology Letters. 21.24 (2009): 1813-1815.
- [9] Celtek, O., S. Ozer, and C. Besikci. "High Responsivity InP-InGaAs Quantum-Well Infrared Photodetectors: Characteristics and Focal Plane Array Performance." IEEE Journal of Quantum Electronics. 41.7 (2005): 980-985.
- [10] Chang, Kai. Microwave Solid-State Circuits and Applications. Wiley: New York, 1994.
- [11] Chau, T., et al. Long-Wavelength Velocity-Matched Distributed Photodetectors. Proc. of IEEE Lasers and Electro-Optics Society (LEOS) Conf., 3 May 1998 21 May 2010. San Francisco: IEEE, 2002.
- [12] Chiu, W. Y., et al. "Leaky-Wave Photodiodes With a Partially p-Doped Absorption Layer and a Distributed Bragg Reflector (DBR) for High-Power and High-Bandwidth-Responsivity Product Performance." IEEE Photonics Technology Letters. 18.11 (2006): 1267-1269.
- [13] Choe, Joong-Seon, Yong-Hwan Kwon, and Kisoo Kim. "Characteristics of Waveguide Photodetectors at High-Power Optical Input." IEEE Photonics Technology Letters. 17.4 (2005): 881-883.
- [14] Chtioui, Mourad, et al. "Thick Absorption Layer Uni-Traveling-Carrier Photodiodes With High Responsivity, High Speed, and High Saturation Power." IEEE Photonics Technology Letters. 21.7 (2009): 429-431.
- [15] Davidson, Andrew and Robert Marsland. "Demand for High-Speed Detectors Drives Research." Laser Focus World. 34.4 (1998): 101-106.
- [16] DeCorby, R. G., A. J.P. Hnatiw, and J. N. McMullin. Techniques to Improve Speed and Efficiency of Photodetectors. Proc. of IEEE Canadian Conference on Electrical and Computer Engineering. Conf., 9 May 1999 21 May 2010. Alberta: IEEE, 1999.
- [17] Derickson, Dennis, ed. Fiber Optic Test and Measurement. Upper Saddle River: Prentice Hall, 1998.
- [18] Demiguel, Stephane, et al. High-Responsivity High-Speed and High-Power Partially Depleted Absorber Waveguide Photodiodes with Relaxed Coupling Tolerances. Proc. of Optical

- Fiber Communication Conference. Conf., 6 Mar 2005 21 May 2010. Anaheim: IEEE, 2005.
- [19] Duan, Ning, et al. "Thermal Analysis of High-Power InGaAs-InP Photodiodes." IEEE Journal of Quantum Electronics. 42.12 (2006): 1255-1258.
- [20] Fang, Alexander W., et al. "A Continuous-Wave Hybrid AlGaInAs-Silicon Evanescent Laser." IEEE Photonics Technology Letters. 18.10 (2006): 1143-1145.
- [21] Franco, Sergio. Design with Operational Amplifiers and Analog Integrated Circuits. 3rd ed. McGraw-Hill: New York, 2002.
- [22] Ghioni, Massimo, et al. "A VLSI-Compatible High-Speed Silicon Photodetector for Optical Data Link Applications." IEEE Transactions on Electron Devices. 43.7 (1996): 1054-1060.
- [23] Giboney, Kirk S. "Travelling-Wave Photodetectors." Diss. U. of California Santa Barbara, 1995.
- [24] Giboney, Kirk S., et al. "Travelling-Wave Photodetectors with 172-GHz Bandwidth and 76-GHz Bandwidth-Efficiency Product." IEEE Photonics Technology Letters. 7.4 (1995): 412-414.
- [25] Hall, Stephen H., Garrett W. Hall, and James A. McCall. High-Speed Digital System Design: A Handbook of Interconnect Theory and Design Practices. Wiley: New York, 2000.
- [26] Harari, Joseph, et al. "Modeling of Waveguide PIN Photodetectors Under Very High Optical Power." IEEE Transactions on Microwave Theory and Techniques. 43.9 (1995): 2304-2310.
- [27] Harari, Joseph, et al. "Theoretical Study of p-i-n Photodetectors' Power Limitations from 2.5 to 60 GHz." IEEE Transactions on Microwave Theory and Techniques. 45.8 (1997): 1332-1336.
- [28] Hietala, Vincent M., et al. "Traveling-Wave Photodetectors for High-Power, Large-Bandwidth Applications." IEEE Transactions on Microwave Theory and Techniques. 43.9 (1995): 2291-2298.
- [29] Huang, Beiju, et al. Monolithic Integration of Light Emitting Diodes, Photodetector and Receiver Circuit in Standard CMOS Technology. Proc. of the 9th International Conference on Solid-State and Integrated Circuit Technology. Conf., 20 Oct 2008 21 May 2010. Beijing: ICSICT, 2008.
- [30] International Business Machines (IBM) Corporation. "IBM's Strained Silicon Breakthrough Image Page." 8 June 2001. 21 May 2010.
<<http://www.research.ibm.com/resources/press/strainedsilicon/>>
- [31] Intel Corporation. "Hybrid Silicon Laser." 2006. 10 Jan 2007
<<http://www.intel.com/research/platform/sp/hybridlaser.htm>>.
- [32] Islam, M. S., et al. Correlation Between the Failure Mechanism and Dark Currents of High Power Photodetectors. Proc. of Lasers and Electro-Optics Society 2000 Annual Meeting. Conf., 13 Nov 2008 21 May 2010. Rio Grande: IEEE, 2000.
- [33] Islam, M. Saif, et al. "High Power and Highly Linear Monolithically Integrated Distributed Balanced Photodetectors." Journal of Lightwave Technology. 20.2 (2002): 285-295.
- [34] Islam, M. Saif, et al. "Recent Advances and Future Prospects in High-Speed and High Saturation-Current Photodetectors." Proceedings of SPIE. 5246 (2003): 448-457.
- [35] Ito, H., et al. "60GHz High Output Power Uni-Travelling-Carrier Photodiodes with Integrated Bias Circuit." Electronics Letters. 36.8 (2000): 747-748.
- [36] Ito, H., et al. "InP/InGaAs Uni-Travelling-Carrier Photodiode with 310GHz Bandwidth." Electronics Letters. 36.21 (2000): 1809-1810.
- [37] Jalali, B., L. Naval, and A. F. J. Levi. "Si-Based Receivers for Optical Data Links." Journal of Lightwave Technology. 12.6 (1994): 930-935.
- [38] Jasmin, Serge, et al. "Diluted- and Distributed-Absorption Microwave Waveguide Photodiodes for High Efficiency and High Power." IEEE Transactions on Microwave Theory and

Bibliography

- Techniques. 45.8 (1997): 1337-1341.
- [39] Johnson, Howard and Martin Graham. High-Speed Signal Propagation: Advanced Black Magic. 2nd ed. Upper Saddle River: Prentice Hall, 2003.
 - [40] Knier, Gil. "NASA scientists work to improve optical computing technology." 23 April 2000. 10 Jan. 2007 <http://science.nasa.gov/headlines/y2000/ast28apr_1m.htm>.
 - [41] Lasaosa, Daniel, et al. "Traveling-Wave Photodetectors With High Power-Bandwidth and Gain-Bandwidth Product Performance." IEEE Journal of Selected Topics in Quantum Electronics. 10.4 (2004): 728-740.
 - [42] Li, Zhi, et al. "High-Power Integrated Balanced Photodetector." IEEE Photonics Technology Letters. 21.24 (2009): 1858-1860.
 - [43] Lin, L. Y., et al. "High-Power High-Speed Photodetectors—Design, Analysis, and Experimental Demonstration." IEEE Transactions on Microwave Theory and Techniques. 45.8 (1997): 1320-1331.
 - [44] Liu, Fei, et al. "Photoconductive Gain of SiGe/Si Quantum Well Photodetectors." Optical Materials. 27.5 (2005): 864-867.
 - [45] Liu, J. F., et al. Waveguide-Integrated Ge p-i-n Photodetectors on SOI Platform. Proc. of 3rd IEEE International Conference on Group IV Photonics. Conf., 13 Sept 2006 21 May 2010. Ottawa: IEEE, 2006.
 - [46] Muramoto, Y., et al. 46.5-GHz-Bandwidth Monolithic Receiver OEIC Consisting of a Waveguide PIN Photodiode and a Distributed Amplifier. Proc. of Integrated Optics and Optical Fibre Communications 11th International Conference. Conf., 22 Sept 1997 21 May 2010. Edinburgh: IET, 2006.
 - [47] Biever, Celeste. "Secret of 'strained silicon' chips revealed." New Scientist. 20 Dec 2003. 25 May 2010. <<http://www.newscientist.com/article/dn4493-secret-of-strained-silicon-chips-revealed.html>>
 - [48] Palik, Edward D., ed. Handbook of Optical Constants and Solids. Academic Press: New York, 1985.
 - [49] Park, Hyundai, Alexander W. Fang, Di Liang, et al. "Photonic Integration on the Hybrid Silicon Evanescent Device Platform." Advances in Optical Technologies. 2008 (2008): 17 pages.
 - [50] Pauchard, Alexandre, et al. "Infrared-Sensitive InGaAs-on-Si p-i-n Photodetectors Exhibiting High-Power Linearity." IEEE Photonics Technology Letters. 16.11 (2004): 2544-2546.
 - [51] Ramaswamy, A., et al. A High Power Ge n-i-p Waveguide Photodetector on Silicon-on-Insulator Substrate. Proc. of 6th IEEE International Conference on Group IV Photonics. Conf., 9 Sept 2009 21 May 2010. San Francisco: IEEE, 2009.
 - [52] Razeghi, Manijeh. Fundamentals of Solid State Engineering. 2nd ed. Springer: New York, 2007.
 - [53] Shi, J. W., et al. High-Responsivity, High-Speed, and High-Saturation-Power Performances of Evanescently Coupled Photodiodes with Partially p-Doped Photo-absorption Layer. Proc. of Electron Devices Meeting. Conf., 13 Dec 2004 21 May 2010. San Francisco: IEDM, 2004.
 - [54] Shi, Jin-Wei, et al. "Extremely High Saturation Current-Bandwidth Product Performance of a Near-Ballistic Uni-Traveling-Carrier Photodiode With a Flip-Chip Bonding Structure." IEEE Journal of Quantum Electronics. 46.1 (2010): 80-86.
 - [55] Shimizu, N., et al. "RF Saturation Mechanism of InP/InGaAs Uni-Travelling-Carrier Photodiode." Electronics Letters. 36.8 (2000): 750-751.
 - [56] Sun, Chi-Kuang, I-Hsing Tan, and John E. Bowers. "Ultrafast Transport Dynamics of p-i-n Photodetectors Under High-Power Illumination." IEEE Photonics Technology Letters. 10.1 (1998): 135-137.
 - [57] Takeuchi, T., et al. "High-Speed, High-Power and High-Efficiency Photodiodes With Evanescently Coupled Graded-Index Waveguide." Electronics Letters. 36.11 (2000): 972-973.

Bibliography

- [58] Tulchinsky, David A. and Keith J. Williams. High Power Photodetectors for Microwave Photonic Links. Proc. of Avionics Fiber-Optics and Photonics. Conf., 20 Oct 2005 21 May 2010. IEEE, 2005.
- [59] Tulchinsky, D. A. and K. J. Williams. "High Power RF Photodiodes." 2005. 21 May 2010 < [http://www.nrl.navy.mil/content_images/05Optical\(Tulchinsky\).pdf](http://www.nrl.navy.mil/content_images/05Optical(Tulchinsky).pdf)>.
- [60] Umbach, A., et al. "High-Speed, High-Power 1.55 μ m Photodetectors." Optical and Quantum Electronics. 33 (2001): 1101-1112.
- [61] Vivien, L., et al. Germanium Photodetector Integrated in a Silicon-On-Insulator Microwaveguide. Proc. of 4th IEEE International Conference on Group IV Photonics. Conf., 19 Sept 2007 21 May 2010. Tokyo: IEEE, 2007.
- [62] Vivien, L., et al. 42 GHz Waveguide Germanium-on-Silicon Vertical PIN Photodetector. Proc. of 5th IEEE International Conference on Group IV Photonics. Conf., 17 Sept 2008 21 May 2010. Sorrento: IEEE, 2008.
- [63] Wang, J., et al. "Evanescent-Coupled Ge p-i-n Photodetectors on Si-Waveguide With SEG-Ge and Comparative Study of Lateral and Vertical p-i-n Configurations." IEEE Electron Device Letters. 29.5 (2008): 445-448.
- [64] Wang, Xin, et al. "InGaAs-InP Photodiodes with High Responsivity and High Saturation Power." IEEE Photonics Technology Letters. 19.16 (2007): 1272-1274.
- [65] Williams, A. R., et al. "InGaAs/InP Waveguide Photodetector with High Saturation Intensity." Electronics Letters. 28.24 (1992): 2258-2259.
- [66] Williams, Keith J. and Ronald D. Esman. "Photodiode DC and Microwave Nonlinearity at High Currents Due to Carrier Recombination Nonlinearities." IEEE Photonics Technology Letters. 10.7 (1998): 1015-1017.
- [67] Williams, Keith J., David A. Tulchinsky, and Alex Hastings. High-Power and High-Linearity Photodiodes. Proc. of IEEE Laser and Electro-Optics Society. Conf., 9 Nov 2008 21 May 2010. Newport Beach: IEEE, 2008.
- [68] Williams, Keith J. and Ronald D. Esman. "Large-Signal Compression-Current Measurements in High-Power Microwave PIN Photodiodes." Electronics Letters. 35.1 (1999): 82-84.
- [69] Williams, Keith J. and David A. Tulchinsky. High-Power Photodiodes. Proc. of 2007 IEEE International Topical Meeting on Microwave Photonics. Conf., 3 Oct 2007 21 May 2010. Victoria: IEEE, 2007.
- [70] Wu, Y.-S. and J.-W. Shi. "Dynamic Analysis of High-Power and High-Speed Near-Ballistic Unitraveling Carrier Photodiodes at W-Band." IEEE Photonics Technology Letters. 20.13 (1998): 1160-1162.
- [71] Wu, M. C., et al. Recent Progresses in High Frequency, High Power Photodetectors. Proc. of MWP '98 International Topical Meeting on Microwave Photonics, 1998. Conf., 12 Oct 1998 21 May 2010. Princeton: IEEE, 1998.
- [72] Xue, Chunlai, et al. "1x4 Ge-on-SOI PIN Photodetector Array for Parallel Optical Interconnects." Journal of Lightwave Technology. 27.24 (2009): 5687-5689.
- [73] Yin, Tao, et al. "31 GHz Ge n-i-p Waveguide Photodetectors on Silicon-on-Insulator Substrate." Optics Express. 15.21 (2007): 13965-13971.
- [74] Itakura, S. et al. "High-Current Backside-Illuminated Photodiode Array Module for Optical Analog Links." Journal of Lightwave Technology. 28.6 (2010): 965-971.

Glossary

III-V – Semiconductor alloys consisting of elements from groups III and V on the periodic table

1dB Compression – The output and input power conditions that cause the gain to decrease by 1dB

Absorber – Active region of the photodiode

Compression Current – The photocurrent that occurs at 1dB compression

Data Link – The information exchange medium and its supporting technologies

Electrical Interconnect – A long metallization trace that transports electrical signals from one point to another on a printed circuit board

IV Characteristic – A plot of the diode current versus bias voltage

High Power Photodiode – A photodiode capable of several mA of photocurrent at high optical power

Load – resistive impedance that is connected across the photodiode terminals to dissipate energy

Load Impedance Optimization – selecting a load impedance that will extract more RF power

Monolithic Integration – Epitaxial growth of structures directly on the substrate material

Optical Communication Link – A data link that uses an optical waveguide as its communication medium

Optical Interconnect – A waveguide structure that routes optical signals from one point to another; can be monolithically integrated in certain circumstances to serve as the direct optical equivalent to the electrical interconnect

P-I-N – A photodiode consisting of a PN structure with an intrinsic layer placed between them

Post Amplifier – A transimpedance amplifier that boosts the signal strength of the photogenerated signal at the expense of the signal-to-noise ratio

Power Compression – Conditions at which the output power of the photodiode drops by 1dB

Power Extraction – Output RF power

Quantum Efficiency – Percent of the incident optical power absorbed by the active region

Responsivity – A coefficient defining the change ratio between the current and optical power

SiGe – An alloy of Silicon and Germanium, which is compatible with standard CMOS processing

Signal Integrity – A discipline of electrical engineering concerned with the elimination or minimization of reflection, noise, insertion loss, and interference in electrical signals

Si-rib – A term referring to the shape of the silicon waveguide structure

Si-waveguide – A monolithically-integrated waveguide structure consisting of doped silicon

Signal-to-Noise (SNR) ratio – The ratio between the signal power and the noise power

Silicon-on-Insulator – A wafer consisting of bulk silicon, a thin buried layer of silicon oxide, and a silicon surface

Standard CMOS Process – Epitaxy on a silicon substrate that does not involve group III-V materials

VLSI – Very large scale integration

Index

- 1dB compression current 124, 166, 167
- 1dB compression measurements... 124, 140, 144, 146, 148, 162, 170, 176
- 1dB compression point 52, 62
- 3dB bandwidth..... 60, 69, 104, 105
 - bandwidth limitations..... 24
 - bandwidth measurements..... 97, 99, 100, 108
 - bandwidth performance regions..... 112
 - bandwidth-efficiency product *See* bandwidth-efficiency tradeoff, *See* bandwidth-efficiency tradeoff
 - bandwidth-efficiency tradeoff..... 27, 28, 59
- abrasive cleaning..... 105
- absorber..... *See* active region
- absorption coefficient..... 13, 16, 17, 56
- absorption efficiency..... 59
- active region
 - cross sectional area 25, 59, 76, 83, 104, 110
 - pure Ge absorber..... 17
 - SiGe absorber..... 17
 - surface-normal absorber..... 27
 - lateral absorber..... 27
 - germanium absorber..... 62
 - device dimensions..... 68
 - absorber thickness..... 24-28, 59, 64, 65, 108
 - absorber structures 27
 - absorber sidewalls..... 65
 - absorber length..... 91
- aluminum oxide 105
- amplitude modulation 140
- APD *See* avalanche photodiodes
- ASE measurement..... 132
- Auger recombination 37
- avalanche breakdown..... 21, 34, 124, 148
- avalanche multiplication 33, 34
- avalanche photodetectors 27, 32, 47, 52
- avalanche region 21
- balanced distributed photodetectors 51
- beam propagation modeling..... 64
- Bragg reflector 47
- break frequency..... 25
- calculated junction capacitance..... 110
- carrier mobility..... 12, 24, 27
- carrier velocity 24
- cascaded quantum wells..... 37
- CMOS compatible process 1, 8, 9, 10, 12, 17, 57
- communications links 4, 12
- communications wavelengths 16
- confinement factor 30, 42
- coplanar waveguide 49, 70, 171, 173
 - simulation 71, 172
- cost reduction..... 8-14
- crosstalk..... 5, 7
 - capacitive-mode coupling..... 5
 - inductive-mode coupling 5
- dark current..... 13, 17, 19, 34, 35, 65, 74, 76, 81
 - bulk contribution..... 63
 - dark current performance regions..... 78
 - sidewall contribution 63
- density of states 38
- dielectric
 - Alumina 71
 - RT Duroid 5880..... 7
- dislocation defects 12, 13, 16, 18, 64, 65
- distributed element model..... 48
- distributed P-I-N..... vi, 50
- doping profile v, 31, 43, 44, 56
- drift velocity 44
- dynamic junction resistance..... 22, 77
- edge of conduction..... 23
- electrical interconnects 1, 8
 - bandwidth 4
 - interconnect miniaturization 7
- electrical losses
 - dielectric losses..... 6, 7
 - performance regions 7
 - resistive losses 6
 - skin effect 6
- electron-hole pair 20, 23
- energy band diagram viii, 126
- epitaxial growth 10
- Erbium energy band diagram..... 125
- Erbium-doped Fiber Amplifier .. 15, 94, 124, 125
- Erbium-doped Fiber Amplifier simulator..... 124
 - OASiX 124, 127
- external quantum efficiency 30
- eye diagram..... 15
- fabrication..... 63
 - chemical-mechanical polishing 64
 - die attach..... 9, 12
 - direct epitaxial growth 12
 - Ge-on-Si epitaxy..... 17
 - oxygen plasma assisted bonding..... 12
 - pad metallization..... 59
 - Rapid Thermal Annealing..... 65
 - selective epitaxial process..... 64
 - wafer annealing..... 64
 - wafer bonding 10

- Fabry-Perot resonant cavity46
- FC/APC.....130
- fiberoptic communication14
- fiber-to-waveguide coupling efficiency30
- GPIO102, 108
- G-S probe.....68
- heterostructure.....45, 56
- high-power photodetectors.....14
- ideal diode equation80
- ideal responsivity26
- i-Ge buffer layer.....17
- III-V photodetectors.....11
- impact ionization.....33, 34
- indirect bandgap.....9, 13, 14, 62
- insertion loss correction100
- Intel Corporation.....8
- internal quantum efficiency46
- intersubband energy states39
- intrinsic region31
- Joule heating19, 21, 45
- junction capacitance.23, 100, 108, 109, 110, 171
- kinetic energy.....33
- laser sources.....9, 10
- laterally-illuminated photodetectors28
- lattice parameter.....12, 13, 14, 16
- lensed fiber alignment.....89
- load impedance optimization ...20, 148, 170, 173
- lumped-element model.....48
- Mach-Zehnder interferometer.....9, 97, 102
- maximum obtainable compression current66
- maximum obtainable RF power.56, 66, 124, 170
- metal-semiconductor-metal.....27, 34
- microprocessor speeds8
- monolithic integration.....8-10, 14, 16, 28
- monolithic modulator.....9
- MQW *See* multiple quantum well
- MSM *See* metal-semiconductor-metal
- multi-chip module.....8
- multiple quantum well37, 41, 43
- noise figure.....15
- noise sources15, 19, 21, 127
 - shot noise15, 19, 20, 21, 51
 - thermal noise15, 20
 - RIN.....15, 47, 51
 - 1/f noise.....21
 - avalanche noise21
 - generation-recombination noise.....20
 - Johnson noise *See* Thermal Noise
 - photon noise21
 - temperature noise21
 - amplified spontaneous emission127
- optical amplifier.....125
- optical confinement29, 30, 42
- optical coupling efficiency18, 29, 46, 85
- optical data link4
- optical filter.....134
- optical gain elements9
- optical interconnects1, 4, 8, 9, 10, 16
- optical links.....14
- optical losses
 - facet coupling loss65, 66, 91, 92
 - free carrier loss64, 65, 66, 92
 - scattering losses66
 - transmission loss.....65, 66, 91
- optical modes.....9
- oscilloscope measurements.....136
- parallel plate capacitor model48, 59, 108
- parasitic capacitance64
- partially depleted absorber.....24, 27, 43, 52
- PCB..... *See* printed circuit board
- PDA *See* partially-depleted absorber
- photodiode model22, 164
- P-I-N27, 31, 32, 43, 47, 58
- polarization sensitivity.....65
- post amplifiers14, 15, 61
- printed circuit board.....6
- propagation speed match49
- quantum cascade laser40
- quantum efficiency22, 26, 60
- quantum tunneling39
- quantum well39
- quantum well duty cycle.....30
- quantum well heterostructure24, 29, 36, 41
- quantum well intersubband photodetector.... 27, 37, 40, 52
- quarterwave transformer.....171
- QWIP *See* quantum well intersubband photodetector
- radio frequency choke149
- RC bandwidth limited 4, 19, 25, 56, 59, 108, 113
- RC time constant .4, 24, 25, 27, 48, 59, 104, 108
- RCE *See* resonant cavity enhanced
- recombination current.....80
- recombination lifetime.....19, 24, 25
- refractive index *See* index of refraction
- refractive index modulation.....9
- refractive index of SiGe.....18
- resistivity83
- resonant cavity enhanced.....27, 46, 52
- responsivity.....26
 - device responsivity91, 92, 94
 - system responsivity.....87

- responsivity correction factor.....92
- RF power extraction.....164
- rib waveguide.....18, 28, 59, 65, 66, 124
- RTA *See* rapid thermal annealing
- saturation photocurrent45, 80
- saturation velocity.....33, 34, 108
- SCH.. *See* separated confinement heterostructure
- Schottky contacts65
- Schottky-barrier junction34
- SEM cross section.....63, 64
- separated confinement heterostructure.....41, 43
- series inductance23
- series resistance.....vi, 19-24, 44, 48, 77-83, 108-109, 124
- Si-Ge PIN photodiodes61
- signal integrity4
- signal to noise ratio19, 20
- silicon waveguide.....9, 29, 62, 65
- silicon waveguide gain element9
- Silicon-Germanium Evanescent Waveguide
 - Photodetector *See* Si-Ge PIN photodiodes
- silicon-on-insulator12, 58, 59, 64
- silver filled epoxy70, 173
- SNR..... *See* Signal to Noise Ratio
- space charge effects19, 23, 43-45, 56, 124
- space charge layer.....19
- spontaneous emission *See* amplified spontaneous emission
- spontaneous emission time constant134
- spur-free dynamic range15
- strained silicon9, 12, 13, 17
- superlattice barrier36
- superlattice period.....37
- surface leakage current65
- surface recombination current.....80
- surface traps26
- surface-illuminated photodiode.....45
- TEC..... *See* thermoelectric cooler
- telecommunications1
- thermal conductivity11, 13, 57, 62, 67
- thermal dissipation.....11, 62, 66, 70, 124
- thermal expansion coefficient.....12
- thermal failure.....11, 24, 34, 62, 81, 145, 149
- thermal generation20, 24
- thermal resistivity11
- thermal runaway19, 24
- thermal simulation66
- thermoelectric cooler21
- thin film coplanar waveguide *See* coplanar waveguide
- trace density6
- transimpedance amplifier.....15
- transit time 19, 24, 27, 28, 32, 34, 35, 45, 46, 48, 56, 59, 108, 109, 112
- traveling wave photodiode.....48, 52, 56
- TWPD..... *See* traveling wave photodiode
- type-I superlattice heterostructure37
- type-II superlattice heterostructure27, 35
- uni-traveling carrier24, 27, 45, 52
- UTC *See* uni-traveling carrier
- velocity mismatch.....56
- velocity-matched distributed photodetectors ..27, 49
- velocity-mismatch bandwidth limitation49
- vertical P-I-N18
- via67
- VMDP..... *See* velocity-matched distributed photodetectors
- voltage dependent responsivity.....62, 94, 96
- wavefunction36
- waveguide facet66, 69, 85
- waveguide photodetector9, 28, 58, 60
- wavelength-division multiplexer129
- WGPD *See* waveguide photodetector
- wirebond70, 171
- XYZ positioner72