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**DESIGN OF AN OPTOELECTRONIC MIXER USING A
METAL-SEMICONDUCTOR-METAL PHOTODETECTOR FOR A
5.3 GHz TO 3.3 GHz APPLICATION**

Vanessa E. Behrens Zaror

**A Thesis submitted to
the Faculty of Graduate Studies and Research
in partial fulfillment of
the requirements for the degree of**

Master of Electrical Engineering

**Ottawa-Carleton Institute of Electrical Engineering
Department of Electronics
Carleton University
Ottawa, Canada**

May 1996

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**DESIGN OF AN OPTOELECTRONIC MIXER USING A
METAL-SEMICONDUCTOR-METAL PHOTODETECTOR FOR A
5.3 GHz TO 3.3 GHz APPLICATION**

submitted by

Vanessa E. Behrens Zaror, B.Eng.

in partial fulfillment of the requirements for the degree of
Master of Electrical Engineering

Thesis Supervisor

Thesis Co-Supervisor

Chair, Department of Electronics

Carleton University

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Abstract

Optoelectronic mixers are required in transmit or receive systems which employ an optical signal distribution architecture. The next generation optically-controlled phased array antennas are a candidate for this type of mixer. An optoelectronic mixer (OEM) using a metal-semiconductor-metal photodetector (MSM-PD) has been designed, fabricated and tested. The novel aspects of the OEM are that it is optically-pumped, and operates in the L-C frequency band. The OEM was developed to downconvert a 5.3 GHz RF input signal to 3.3 GHz, using a local oscillator signal provided by a 2.0 GHz intensity modulated optical source. The mixer design was based on a small-signal S-parameter approach. It is shown that the small-signal conductance of the MSM-PD must change as a function of optical power in order to obtain optoelectronic mixing. A minimum conversion loss of 44 ± 1 dB was achieved for the mixer. The OEM exhibits an LO-to-IF interport isolation of 71 dB and an RF-to-IF isolation of 49 dB under optically-pumped conditions. In conclusion, the developmental work on a lumped element equivalent circuit model of the MSM-PD necessary for the large signal design of the mixer is presented. The simulated response for the model under small-signal conditions is shown to follow a similar trend as the measured data.

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Contents

List of Tables	viii
List of Figures	ix
List of Symbols	xiv
List of Acronyms	xvii

Chapter 1: An Overview of the Optoelectronic Mixer.....1

1.1	Introduction	1
1.2	An Overview of Mixers	2
1.3	Device Selection: Metal-Semiconductor-Metal Photodetector	5
1.4	Literature Review: Optoelectronic Mixers (OEMs)	5
1.5	Objective of the Thesis	9

Chapter 2: Theory behind the Metal-Semiconductor-Metal Photodetector (MSM-PD) and its Application as an Optoelectronic Mixer.....11

2.1	Introduction	11
2.2	Photodetectors	11
2.2.1	General Operation of Photodetectors	11
2.2.2	Figures of Merit of Photodetectors	14
2.2.3	Metal-Semiconductor-Metal Device	19
2.2.4	Metal-Semiconductor-Metal as a Photodetector	22
2.2.5	Figures of Merit of the MSM-PD	27
2.2.6	MSM-PD Circuit Configurations	28
2.3	Mixers	29
2.3.1	Diode Mixer Theory	31
2.3.2	RF Mixer Figures of Merit	34
2.3.3	The Optoelectronic Mixer Process	38
2.3.4	Operating Point of the OEM	39
2.4	Design Procedure	44
2.4.1	Introduction	44
2.4.2	Small Signal (S-Parameter) Design	46
2.4.3	Large Signal Design	48

Chapter 3: Equivalent Circuit of the MSM-PD.....50

3.1	Introduction	50
3.2	Modeling Procedure	51

3.3	Choice and Equivalent Circuit Topology of MSM-PD	53
3.4	DC Equivalent Circuit for the MSM-PD	57
3.4.1	Introduction	57
3.4.2	Experimental Set-Up	57
3.4.3	DC Equivalent Circuit	61
3.5	AC Equivalent Circuit Model for the MSM-PD	67
3.5.1	Introduction	67
3.5.2	Measurement Technique and MSM-PD Mounting Structure	69
3.5.3	AC Element Starting Values and Optimization Range	72
3.5.4	AC Equivalent Circuit	75
3.5.5	Conclusion	87

Chapter 4: Design of an Optically-Pumped, MSM-PD, 5.3 GHz to 3.3 GHz Downconverter88

4.1	Introduction	88
4.2	Optoelectronic Mixer Design Based on S-Parameters	91
4.2.1	Optoelectronic Mixer Circuit	91
4.2.2	Lowpass Filter (LPF)	92
4.2.3	Bias Network and IF Return	96
4.2.4	Input and Output Matching Networks	101
4.2.5	MSM-PD	102
4.2.6	Final Mixer Circuit	102
4.3	Optoelectronic Mixer Experimental Results	105
4.3.1	Experimental Set-Up	105
4.3.2	Input Frequency Match	109
4.3.3	Output Frequency Match	110
4.3.4	Output Power Measurements	112
4.3.5	Repeatability Measurements	116
4.3.6	Comparing Results of OEM with and without Matching Networks	117
4.4	Figures of Merit of the Optoelectronic Mixer	122
4.4.1	Return Loss	122
4.4.2	Conversion Loss	125
4.4.3	1dB Compression Point	127
4.4.4	Interport Isolation	128
4.4.5	3rd Order Intercept Point	129
4.4.6	Signal-to Noise Ratio (S/N)	129
4.5	Conclusion	130

Chapter 5: Potential Areas for Further Research..... 131

5.1	Brief Review of Thesis Objectives	131
5.2	Improvements in the MSM-PD Measurement Techniques	132
5.3	Accounting for MSM-PD Contact Pads	132
5.4	Physical Modeling of the Device	133
5.5	Accounting for the Effect of Deep Traps in the MSM-PD	133
5.6	MSM-PD Parameter Dependence on Bias and Optical Power	135
5.7	Improvement in the Optical Power Measuring Technique	136
5.8	Technique to Increase the Modulation Index of the Laser	137
5.9	Mixer Improvement Through Changes in the Device Doping Profile	138
5.10	Conclusion	138

Appendices:

I.	HP-EEsof/Libra Schematic Files for the Optoelectronic mixer	140
II.	SHARP LT015MD Specification Data	144

References 147

List of Tables

Table 1.1	Review of Optoelectronic Mixers	7
Table 3.1	DC Model Element Values	67
Table 3.2	Element Value Range for Mounting Structure #2	73
Table 3.3	Final Element Values for the MSM-PD Equivalent Circuit	86
Table 4.1	RT/Duroid Substrate Parameters	92
Table 4.2	Power and Current Values	108
Table 4.3	Optoelectronic Mixing Results using the OEM	113
Table 4.4	Optoelectronic Mixing Results using the OEM without Matching Networks	119
Table 4.5	Power Difference between the OEM with and without Matching Networks	121
Table 4.6	Minimum Input and Output Return Loss Parameters of the OEM with Matching Networks	123
Table 4.7	Input Return Loss Parameters of the OEM without Matching Networks	125
Table 4.8	Conversion Loss Difference between Matched and Unmatched OEM	127
Table 4.9	Interport Isolation for the Optoelectronic Mixer	129
Table 4.10	Signal-to-Noise Measurements	130

List of Figures

Figure 1.1a)	Conventional microwave mixers - downconverter	2
Figure 1.1b)	Conventional microwave mixers - upconverter	2
Figure 1.2	Detector and downconverter system	3
Figure 1.3	Optoelectronic mixer as a downconverter	4
Figure 2.1	Physical structure of a pn diode	12
Figure 2.2	Energy band diagram for a reverse bias pn diode	13
Figure 2.3	Structure and energy band diagram for a reverse-biased pn photodiode	14
Figure 2.4	Physical representation of the MSM	19
Figure 2.5	Schematic diagram of the MSM	19
Figure 2.6	Energy band diagram and current densities of the MSM-PD at thermal equilibrium	20
Figure 2.7	Energy band diagram of an MSM under a positive bias applied at contact 2	21
Figure 2.8	Charge distribution and electric field of the MSM	22
Figure 2.9	Schematic of MSM-PD under external stimulus	23
Figure 2.10	MSM-PD physical representation	23
Figure 2.11	Cross-sectional view of MSM-PD under an applied bias and optical illumination	24
Figure 2.12	MSM-PD under applied bias and optical illumination	25
Figure 2.13	DC I-V relationship for different optical powers	26
Figure 2.14	MSM-PD connected in a 1-port and 2-port configuration	28
Figure 2.15	Schematic diagram of a downconverter	29
Figure 2.16	I-V characteristic of a diode	33

Figure 2.17	Input and output characteristic of a mixer	36
Figure 2.18	DC I/V curves for different optical powers for a device in which no optoelectronic mixing occurs	39
Figure 2.19	a-axis intercept as a function of optical power	40
Figure 2.20	DC I/V curves for different optical powers for a device in which optoelectronic mixing does occur	41
Figure 2.21	Slope of DC I/V curves	42
Figure 2.22	a-intercept of DC I/V curves	42
Figure 2.23	Impedance transformations for maximum power transfer	46
Figure 3.1	Physical topology of MSM-PD	54
Figure 3.2	Equivalent circuit components related to the physical topology	54
Figure 3.3	Optoelectronic mixer MSM-PD equivalent circuit	56
Figure 3.4	DC equivalent circuit of MSM-PD	57
Figure 3.5	DC and AC measurement set-up	58
Figure 3.6	MSM-PD	58
Figure 3.7	Photocurrent as a function of bias for different optical powers	60
Figure 3.8	Responsivity versus bias	62
Figure 3.9	Responsivity versus optical power	62
Figure 3.10	Conductance as a function of optical power for different bias levels	64
Figure 3.11	Simulated and measured DC I/V curves for different optical powers	66
Figure 3.12	DC equivalent circuit of the MSM-PD	66
Figure 3.13	DC and AC model of the MSM-PD	68
Figure 3.14	Mounting structure #1	70

Figure 3.15	Mounting structure #2	71
Figure 3.16	Magnitude of S_{11} for different bias levels	72
Figure 3.17	Comparing measured and simulated (before and after optimization) magnitude and phase (in degrees) of S_{11}	76
Figure 3.18	Comparing measured and simulated (before and after optimization) magnitude and phase (in degrees) of S_{22}	77
Figure 3.19	Comparing measured and simulated (before and after optimization) magnitude and phase (in degrees) of S_{21}	78
Figure 3.20	Comparing measured and simulated (before and after optimization) magnitude and phase (in degrees) of S_{12}	79
Figure 3.21	Mag(S_{11}) [dB] of MSM-PD taken on four different days	81
Figure 3.22	Ang(S_{11}) [deg] of MSM-PD taken on four different days	81
Figure 3.23	Mag(S_{21}) [dB] of MSM-PD taken on four different days	82
Figure 3.24	Ang(S_{21}) [deg] of MSM-PD taken on four different days	82
Figure 3.25	Mag(S_{12}) [dB] of MSM-PD taken on four different days	83
Figure 3.26	Ang(S_{12}) [deg] of MSM-PD taken on four different days	83
Figure 3.27	Mag(S_{22}) [dB] of MSM-PD taken on four different days	84
Figure 3.28	Ang(S_{22}) [deg] of MSM-PD taken on four different days	84
Figure 3.29	Optimized equivalent circuit of the MSM-PD	86
Figure 3.30	Simulated and measured Mag(S_{11}) for a range of biases and $P_{opt} = 1.36$ mW	87
Figure 4.1	One element of a conventional phased-array receiver	88
Figure 4.2	One element of a phased-array receiver using fiber optics as the distribution system	89
Figure 4.3	One element of a phased-array receiver using an OEM	89

Figure 4.4	Schematic of OEM application	90
Figure 4.5	Optoelectronic mixer module	91
Figure 4.6	Physical layout of the lowpass filter	94
Figure 4.7	Magnitude of S_{11} [dB] of the lowpass filter	94
Figure 4.8	Magnitude of S_{21} [dB] of the lowpass filter	95
Figure 4.9	Simulated S_{11} for the lowpass filter	95
Figure 4.10	Physical layout of bias network and IF return	97
Figure 4.11	Simulated S_{22} for the bias network with IF return	98
Figure 4.12	Simulated S_{22} for the bias network without IF return	99
Figure 4.13	Magnitude of the S_{11} [dB] for the bias network	100
Figure 4.14	Magnitude of the S_{21} [dB] for the bias network	100
Figure 4.15	Photograph of the optoelectronic mixer	103
Figure 4.16	Simulated input return loss of the OEM	104
Figure 4.17	Simulated output return loss of the OEM	104
Figure 4.18	Optoelectronic mixer experimental set-up	105
Figure 4.19	Laser module and bias-T	106
Figure 4.20	Input frequency match measurement set-up	109
Figure 4.21	IF power versus RF power for different LO powers	114
Figure 4.22	IF power versus LO power for different RF powers	115
Figure 4.23	Output spectrum of the OEM	116
Figure 4.24	Repeatability measurements	117
Figure 4.25	Optoelectronic mixer experiment set-up using the unmatched OEM	118

Figure 4.26	IF power of OEM with and without matching networks	120
Figure 4.27	Measurement set-up for the return loss of the OEM	122
Figure 4.28	Comparison of simulated and measured input return loss	124
Figure 4.29	Comparison of simulated and measured output return loss	124
Figure 4.30	Conversion loss of the OEM with matching networks	126
Figure 4.31	Conversion loss of the OEM without matching networks	126
Figure 4.32	Calculation of 1dB compression point for matched OEM	128

List of Symbols

A	amperes
<i>c</i>	velocity of light in vacuum
C	coulombs
C_1, C_2	junction capacitance
C_{metal}	capacitance of the MSM-PD metal fingers
C_{msm}	depletion junction capacitance of the MSM-PD
C_{pad1}, C_{pad2}	capacitance of MSM-PD contact pads
<i>d</i>	diameter of bond wire
$d_{e,h}$	distance traveled by carrier
<i>E</i>	electric field intensity
E_C	conduction band energy at thermal equilibrium
E'_C	conduction band energy under reverse bias
E_F	Fermi energy level
E_g	semiconductor band gap energy
E_{mi}	maximum electric field
E_V	valence band energy at thermal equilibrium
E'_V	valence band energy under reverse bias
<i>f</i>	frequency
f_{IF}	IF signal frequency
f_{R1}, f_{R2}	intermodulation product frequencies
f_{rf1}, f_{rf2}	RF frequencies
f_{RF}	RF signal frequency
f_{LO}	LO signal frequency
g_D	small-signal conductance
<i>h</i>	Planck's constant
I	IF mixer port
I_D	diode current
I_{dark}	dark current of the MSM-PD
I_{PD}	photodetector current
I_{ph}	photocurrent
I_S	saturation current
J	joules
J_n	electron current density
J_p	hole current density
J_{SMO}, J_{MSO}	current densities
J_{SMF}, J_{MSR}	current densities
<i>k</i>	Boltzmann's constant
<i>l</i>	length of bond wire

L	semiconductor length
L	LO mixer port
L	inductance of a bond wire
l_{abs}	absorption length
L_{bond1}, L_{bond2}	bond wire inductance
L_M	mutual inductance between two parallel bond wires
L_S	inductance of the MSM-PD metal fingers
L_{total}	total inductance of two parallel bond wires
n	ideality factor
n	donor-doped material
N_D	total number of donor atoms or sites/ cm^3
p	acceptor-doped material
P_0, P_1, P_2, P_3	coefficients of the responsivity equation
$P1, P2$	port designation for the Vector Network Analyzer
$P_{electrical}$	electrical power
P_{IF}	IF power
P_{in}	input power
P_{LO}	electrical power level of the LO signal
P_{opt}	optical power
$P_{opt,n}$	n th optical power level
P_{opt_rms}	modulated optical power
P_{RF}	RF power
$P_{sa}, P_{inc}, P_{ref}, P_2$	RF power
P_{out}	output power
q	electron charge
QE_{ext}	external quantum efficiency
R	RF mixer port
R	responsivity
R_{F1}, R_{F2}	resistance of MSM-PD metal fingers
$R_{Ground1}, R_{Ground2}$	shunt resistance associated with the DC measurement set-up
R_{msm}	resistance representing the leakage current of the MSM-PD
R_{ph}	responsivity of the MSM-PD
R_S	resistance of the MSM-PD metal fingers
R_{S1}, R_{S2}	series resistance associated with the DC measurement set-up
s	separation length between two bond wires
S_y	S-parameters
t	time
T	junction temperature
$v_{e,h}$	velocity of the carrier
V	volts

V_1, V_2	bias potentials across diode junctions
V_{bi}	built-in potential of a pn junction diode
V_{bias}	bias applied across MSM-PD
V_D	applied voltage across diode
V_{D1}, V_{D2}	built-in potentials
V_{DUT}	applied voltage across the MSM-PD
V_p	DC operating voltage
V_{RT}	reach-through voltage
W	watts
W_1, W_2	depletion region widths
x_{abs}	photodetector material thickness
a_n, b_n, c_n	equation coefficients
λ	wavelength
α	absorption coefficient
η_q	internal quantum efficiency
ρ	reflection coefficient
Φ_{n1}, Φ_{n2}	electron barrier heights
ϵ_0	permittivity of free space
ϵ_r	relative permittivity
α_1	gradient of DC I/V curve
β	gradient
η	gradient
ϕ	gradient
τ_{RC}	RC time constant
τ_{tr}	hole and electron transit time
Γ	reflection coefficient

List of Acronyms

APD	avalanche photodiode
BW	bandwidth
DUT	device under test
EMI	electromagnetic interference
FET	field effect transistor
GND	ground
IF	intermediate frequency
LO	local oscillator
MIM	metal-insulator-metal
MMIC	microwave monolithic integrated circuit
MSM-PD	metal-semiconductor-metal photodetector
OEM	optoelectronic mixer
PD	photodetector
PR1, PR2	planes of reference
RF	radio frequency

Chapter 1: An Overview of the Optoelectronic Mixer

1.1 Introduction

Optoelectronic technologies offer several advantages over the existing electronic technologies. Fiber optics are replacing the conventional waveguides or coaxial cables for signal distribution in microwave systems. The use of fiber optics reduces the weight and size of the system, provides immunity to EMI (electromagnetic interference), yields a higher bandwidth, and offers lower propagation losses. These are important requirements in airborne and satellite applications [1]. Another advantage of optoelectronic technology is its potential to optically control microwave devices. Optical technology also has the potential of being integrated with MMIC (microwave monolithic integrated circuit) components.

Microwave signals can be carried conveniently via fiber optic cables on optical carriers. Replacing the traditional transmission lines with fiber optics entails a modification to the conventional receiver configuration because optical signals are now present in addition to electrical signals. The optical signal must be converted into a microwave signal for subsequent signal processing using existing microwave devices. However, there are some microwave devices that directly use the optical signal, and these devices are referred to as optically-controlled devices. An optoelectronic mixer (OEM) is one such device that does not require the optical-to-electrical conversion. The OEM simultaneously detects the optical signal and performs frequency conversion as would a conventional mixer. An OEM differs from the conventional microwave mixer in that one of the microwave inputs is carried on an optical carrier. Optoelectronic mixers have been

investigated using various photodetectors with different characteristics [13]-[34]. An optoelectronic mixer using a metal-semiconductor-metal photodetector (MSM-PD) as an optically controlled microwave device is described in this thesis.

1.2 An Overview of Mixers

A mixer is a device capable of frequency conversion. A more detailed description of mixers is given in Chapter 2. The schematic of a conventional microwave mixer is illustrated in Figure 1.1 where a) and b) respectively illustrate a downconverter and an upconverter. It is important to note that the RF, LO, and IF signals of the conventional mixer lie within the radio frequency bands as opposed to the optical regime. The LO signal refers to the pump signal and the IF or RF, depending whether the mixer is a down or up converter, represent the input information or the output signal. The discussion on mixers will be limited to a downconverter, and hence the input information signal will be termed the RF signal and the output will be the IF as shown in Figure 1.1a)

Figure 1.1 Conventional microwave mixers

The advent of fiber optical cables for signal distribution requires the intensity modulated optical signal to first be converted into an electrical signal prior to subsequent processing using existing microwave devices. The optical signal must be first demodulated using a photodetector and the resultant radio frequency signal fed into the conventional microwave mixer as shown in Figure 1.2. Though the figure illustrates the optical signal as carrying the information, the optical signal could also be the local oscillator.

Figure 1.2 Detector and downconverter system

One should note that the RF, LO, and the IF signals lie again in the radio frequency regime. Depending whether the detector is equipped with gain, an amplifier might be needed between the detector and the mixer to increase the RF signal applied to the mixer.

A drawback of the system illustrated in Figure 1.2 is the need for the photodetector to convert the optical signal into an RF signal prior to processing. The requirement for a photodetector introduces an additional component compared to the conventional microwave mixer of Figure 1.1.

The complexity of the optical receiver system can be reduced if the detector and mixer are replaced with an optoelectronic mixer as shown in Figure 1.3. An optoelectronic mixer uses an intensity modulated optical signal directly to either pump the device or as the information signal to be demodulated. The LO and RF signals can either be in the optical or radio frequency (electrical) spectrum. The IF output signal of the downconverted mixer is in the radio frequency spectrum because the mixer generates an electrical current. It should be noted that an IF amplifier is usually required at the output of an optoelectronic mixer. However, the amplifier between the photodetector and mixer required in the system of Figure 1.2 is no longer needed, since mixing and detection of the optical signal is achieved simultaneously by the optoelectronic mixing device.

Figure 1.3 Optoelectronic mixer as a downconverter

Another advantage of the optoelectronic mixer over the conventional mixer is that no electrical matching or parasitic component problems are present at the optical port. Optical matching is however required to maximize the amount of light illuminating the photosensitive device. A further advantage of the optoelectronic mixer is the almost infinite interport isolation between the electrical and the optical ports, i.e., the incoming electrical signal does not leak into the optical port.

1.3 Device Selection: Metal-Semiconductor-Metal Photodetector

The device used for the OEM must satisfy two criteria: it must be an efficient photodetector and it must possess a mixing mechanism. The device used as an OEM was the MSM-PD because it was readily available at the research establishment¹. The MSM-PD offers high speed [4], [9], [10]; wide bandwidth [8], [9], [10], [11]; low power consumption (low bias) [8]; easy integration with conventional GaAs MESFET circuit technology [3], [4], [5], [6]; and better noise performance. The efficiency of the MSM-PD as the mixing element will be determined throughout the development of the optoelectronic mixer.

1.4 Literature Review: Optoelectronic Mixers (OEMs)

Optoelectronic mixers have been researched in the past using different devices in order to meet the requirements in a variety of applications.

There are two types of OEM: optically-pumped and electrically-pumped. The former refers to an optical local oscillator and the latter refers to an electrical local oscillator. The electrically-pumped mixer has been reported using MESFETs [2,12], photoconductors [18], point-contact diode structures [35,36], Schottky diodes [37,38], avalanche photodiodes (APD) [13,21], metal-insulator-metal (MIM) diodes [39], and most recently metal-semiconductor-metal devices [30]. Optically-pumped mixers have been reported using photoconductive devices [15], and heavily doped metal-semiconductor contacts [20].

¹ Communications Research Center (CRC)

Table 1.1 gives a brief overview of the optoelectronic mixers found in the literature. The table is spread across two pages due to the number of columns. The letter 'o' or 'e' in the third column on page 7 respectively indicates whether the mixer is optically or electrically pumped. The letter 'd' in the seventh column on page 7 refers to the downconverted mixer product. The letter 'p' in the last column on page 8 indicates that the conversion loss is a predicted value, all other quoted values being derived from measured data. The symbols in the last column on page 8 indicate papers by the same authors. The given conversion losses are the optimum values quoted in the papers, and the cited LO, RF, and IF frequencies correspond to these given values.

Table 1.1 Review of Optoelectronic Mixers

Ref #	Device	pump (o/e)	conversion loss	LO Frequency	RF Frequency	IF Frequency
13	APD	e	2 dB	100 MHz	70 MHz	30 MHz (d)
14	APD	e	2 dB	100 MHz	70 MHz	30 MHz (d)
15	InP switch	O				
16	APD	e		100 MHz	50 MHz	50 MHz (d)
	GaAs photo-conductor					
17	GaInAs PIN PD	e	(27-40) dB	(10-20) GHz	(10-20) GHz	0.5 & 2 GHz
18	GaAs photo-conductor	O	N/A	4.5 GHz		500 MHz (d)
19	APD	e	5 dB	100 MHz		10 MHz (d)
20	Schottky-N GaAs tunnelling junction	O	43 dB	130 MHz	140 MHz	10 MHz (d)
21	APD	e	(8-28)dB	100 MHz		10 MHz
22	GaAs FET	e	N/A		89 GHz	6.5 GHz
23	Mott diode	O	6.5 dB	2.2 GHz		60 MHz (d)
			9 dB	9.9 GHz		270 MHz (d)
24	Schottky-N GaAs tunnelling junction	O	20 dB	100-110 MHz (d)		1-10 MHz (d)
25	GaAs MESFET	O	48 dB	400 MHz	1 GHz	600 MHz
26	AlGaAs/GaAs HEMT	O	N/A	6 GHz	9 GHz	3 GHz
27	GaAs interdigital PD	e	N/A	101 MHz	100 MHz	1 MHz
28	InGaAs PIN PD	e	8.1 dB	8 GHz	0.9 GHz	8.9 GHz
	InGaAs PIN PD	O	47.7 dB	8 GHz	0.9 GHz	8.9 GHz
29	InGaAs PIN PD	e	8 dB	10 GHz	0.9 GHz	10.9 GHz
30	GaAs MSM PD	e	14 dB	100 MHz	60 MHz	40 MHz
31	S.I. GaAs PD	e	33 dB	3.596 GHz	3.467 GHz	0.129 GHz (d)
32	S.I. GaAs PD	e	41 dB	3.598 GHz	3.467GHz	0.131 GHz (d)
33	HEMT	e	32 dB	(1-20) GHz	(1-20) GHz	0.4 GHz
34	MESFET	e	(7-11 dB)	(60 MHz-3GHz)		50 MHz (d)

Table 1.1 (continued)

Ref #	Device	laser λ	Date	Type of mixers	
13	APD	He-Ne laser	1970	diode	••
14	APD	He-Ne laser	1972	diode	••
15	InP switch		1981	photoconductive	
16	APD	750 nm	1982	diode	
	GaAs				
	photo-conductor				
17	GaInAs PIN PD	1300 nm	1983	diode	
18	GaAs photo-conductor		1984	photoconductive	
19	APD	780 nm	1985	diode	(a)
20	Schottky-N	780 nm	1985	diode	(a)
	GaAs tunnelling junction				
21	APD	780 nm	1986		
22	GaAs FET	830 nm	1986	three-terminal	
23	Mott diode		1987	diode	p #
					p #
24	Schottky-N	797 nm	1989	diode	#
	GaAs tunnelling junction				
25	GaAs MESFET		1989	three-terminal	
26	AlGaAs/GaAs		1987	three-terminal	
	HEMT				
27	GaAs interdigital PD	780 nm	1990	MSM	
28	InGaAs PIN PD	InGaAsP laser	1991	diode	•
	InGaAs PIN PD	InGaAsP laser	1991	diode	•
29	InGaAs PIN PD	InGaAsP laser	1991	diode	•
30	GaAs MSM PD	780 nm	1993	MSM	
31	S.I. GaAs PD	840 nm	1991		&
32	S.I. GaAs PD	840 nm	1991		&
33	HEMT	850 nm	1993	three-terminal	
34	MESFET	850 nm	1995	three-terminal	

1.5 Objective of the Thesis

The research documented in this thesis is focused on the development of an optoelectronic mixer used as a downconverter in a optoelectronic receiver system for a phased-array application. The incoming signal is an electromagnetic wave which is converted to electrical energy through a receiving antenna and must be downconverted using a mixer. The mixer is traditionally pumped using an electrical local oscillator signal. However, the next generation phased-array antennas will have an optical distribution network to transport the signal to and from the payload and the front end of the receiver. Therefore the local oscillator signal would be in the optical regime.

Therefore the objective of the research is to design, fabricate and test an optically-pumped optoelectronic mixer for a phased array application. The mixer will downconvert a 5.3 GHz input electrical signal to a 3.3 GHz electrical signal using a 2.0 GHz intensity modulated optical signal. A Metal-Semiconductor-Metal Photodetector (MSM-PD) is used as the mixing device. The approach used in this thesis differs from previous research in that the MSM-PD will be optically-pumped as opposed to the more conventional electrically-pumped approach [30]. Furthermore fabrication of the circuit will be carried out on RT/Duroid as opposed to the more commonly used alumina substrate. Duroid was selected to reduce fabrication difficulties and costs. A large signal model of the MSM-PD will also be developed using a lumped element approach. To the author's knowledge, this is the first attempt to provide a large signal model of the MSM-PD. The existing models of the MSM-PD are only valid under small-signal conditions [45] - [53].

The thesis is divided into five sections. A brief description on the general operation and physical structure of photodetectors, concentrating on the MSM-PD, is given in Chapter 2, along with a description of conventional and optoelectronic mixer theory. The development of a large-signal equivalent circuit of the MSM-PD is described in Chapter 3. The design, fabrication, and testing of an optoelectronic mixer using a small-signal (S-parameter) approach is provided in Chapter 4. Results and calculated figures of merit are also presented in Chapter 4. In conclusion, a summary of the results and suggested future areas of work are presented in Chapter 5.

Chapter 2: Theory Behind the Metal-Semiconductor-Metal Photodetector (MSM-PD) and its Application as an Optoelectronic Mixer

2.1 Introduction

The background theory on photodetectors and mixers, including their operating mechanism and associated figures of merit is introduced in this chapter. More specifically MSM-PD and optoelectronic mixers will be described. Optoelectronic mixer theory will be paralleled to the general microwave mixer theory to highlight deviations from the standard mixer theory. In conclusion, an outline is presented of the design steps needed in the fabrication of an optoelectronic mixer.

2.2 Photodetectors

A photodetector is used to convert incident optical energy (photons) into electrical energy. The photodetector material absorbs incident photons having energy greater than the bandgap energy level of the material to create an electron-hole pair within the semiconductor material. The optically generated carriers that reach the external contacts will form a continual carrier movement called the photocurrent.

2.2.1 General Operation of Photodetectors

The general operation of a photodetector will be explained using a pn photodiode. A pn photodiode consists of a p-type semiconductor material placed in intimate contact with an n-type semiconductor material as shown in Figure 2.1. At equilibrium, i.e., no external stimulus is applied to the diode and more specifically no bias

across the device, a depletion region is formed near the junction of the two materials. This region is devoid of majority carrier concentration. The diffusion and drift components of the current are equal and opposite, thus resulting in a zero net flow of carriers. Neudeck [40] gives a more detailed explanation on the operation of the pn diode.

Figure 2.1 Physical structure of a pn diode

Applying a reverse bias across the diode increases the barrier potential encountered by the holes in the p-region wanting to diffuse to the n-region. The barrier potential increases as well for the electrons in the n-region wanting to diffuse to the p-region. Consequently, the electron and hole diffusion components of the current are reduced to less than their respective thermal equilibrium values. The electron and hole drift components remain at their equilibrium values since their values are independent of the barrier potential. The drift component is small since it is limited to the thermally generated carriers. The overall current for a reverse-biased pn diode, termed the leakage current, is primarily due to the drift component and is negatively directed. The energy band diagram for the reverse bias case as compared to the thermal equilibrium case is shown in Figure 2.2. E_v, E_c respectively describe the valence band and conductance band energy levels at

thermal equilibrium. E'_v, E'_c also describe the same energy levels when the diode is reverse biased. V_{bi} refers to the built-in potential and V refers to the applied voltage.

Figure 2.2 Energy band diagram for a reverse bias pn diode

Applying sufficient optical energy to the depletion region of a reverse-biased diode creates electron-hole pairs. The optical energy must be greater or equal to the band gap energy of the semiconductor material to create carriers. The electric field created by the static ion sites sweeps the optically generated carriers across the depletion region towards their respective majority sides as shown in Figure 2.3. The created electrons are swept towards the n-type material while the holes are swept towards the p-type material. This carrier movement is termed the photocurrent of the device and is in addition to the leakage

current. The leakage current is much smaller than the photocurrent. Therefore the overall current is primarily due to the optically generated carriers. A forward bias diode will have a diffusion current component in the opposite direction to the drift and optical generated current components. Hence the overall current won't be only caused by the optically generated carriers. A forward-biased diode will also have a small depletion region thus limiting the sensitive region of the device, and making the device an inefficient photodetector. Hence the diode is operated in the reverse bias mode to maximize its responsivity as a photodetector, where the responsivity is defined as a measure of the generated photocurrent to the incident optical power.

Figure 2.3 Structure and energy band diagram for reverse-biased pn photodiode

2.2.2 Figures of Merit of Photodetectors

Photodetectors are very important in optical-fiber communications systems since they must convert the optical signal into an electrical signal without loss of

information. They must therefore satisfy stringent requirements such as high sensitivity at the operating frequency, high frequency response, minimum noise, and they should also be compact in size, use low biasing voltages, and be reliable under operating conditions. Some figures of merit associated with photodetectors are absorption coefficient, quantum efficiency, responsivity, dark current, and bandwidth.

Absorption coefficient (α)

The absorption of photons in a photodetector generates carriers, and the collected carriers form a photocurrent. The amount of photons absorbed is dependent on the absorption coefficient α of the material. The absorption coefficient is strongly dependent on the wavelength of the optical signal. The absorption length l_{abs} is defined as the thickness of the material required to absorb $(1-1/e)$ of the incident photons. The absorption length and coefficient have the relationship given by (1).

$$\alpha = \frac{1}{l_{abs}} \quad (1)$$

If the thickness of the photodetector material is x_{abs} , then the quotient x_{abs}/l_{abs} represents the number of absorption lengths of the device. A material should have a few absorption lengths to absorb 100% of the incident photons. For example, the absorption length for GaAs at 830 nm is approximately $(1-2) \mu m$, hence the thickness of GaAs must be greater than $(1-2) \mu m$ to absorb all of the incident light.

Quantum Efficiency

Quantum efficiency describes the ability of the material to convert absorbed photons into usable carriers. This parameter represents the fraction of absorbed incident photons that generate electron-hole pairs which are subsequently collected by the contacts. The internal quantum efficiency η_q depends on the absorption coefficient as shown in (2) and is therefore dependent on the photon wavelength. Two factors reducing η_q are bulk recombination and surface recombination of the created carriers. A percentage, *refl*, of the incident power is reflected at the interface between air and semiconductor. The external quantum efficiency *QE* takes this effect into account as shown in (3). The reflection may be due to contact reflection or material reflection.

$$\eta_q = 1 - e^{(-\alpha \cdot x_{\text{abs}})} = 1 - e^{(-x_{\text{abs}}/l_{\text{abs}})} \quad (2)$$

$$QE = (1 - \text{refl}) \cdot \eta_q \quad (3)$$

Responsivity (R)

Responsivity is a measure of the output photocurrent to the incident optical power as shown in (4). Responsivity is often used to characterize the performance of the photodetector instead of the quantum efficiency since it relates to the incident power.

$$R = \frac{I_{ph}}{P_{opt}} \quad [A/W] \quad (4)$$

The responsivity is defined as the electric charge produced from one photon ($q \cdot QE$) per energy of photon ($h \cdot f$) as shown in (5) and has the unit of amps per watt.

$$R = \frac{q \cdot QE}{h \cdot f} \quad \frac{[C / \text{photon}]}{[J / \text{photon}]} = \frac{[A] \cdot [s]}{[W] \cdot [s]} = \frac{[A]}{[W]} \quad (5)$$

The responsivity of a photodiode for an 830 nm wavelength optical signal is 0.667 A/W assuming 100% quantum efficiency and no reflected incident power.

Dark current

The dark current is the current that flows in the absence of light and is ideally small so as not to significantly contribute to the overall current. The dark current is comprised of drift, generation-recombination (G-R) and tunneling currents. The drift current was previously discussed in the “General Operation of Photodetectors” section. The G-R current is due to the generation and recombination of electron-hole pairs in the diode depletion region. Tunneling current refers to the band-to-band tunneling in the presence of high fields which reduce the effective barrier potential, allowing the carriers to cross the bandgap. The dark current relates to the detector’s sensitivity and ability to detect weak optical signals.

Bandwidth

The frequency response of photodetector is limited by its RC time constant and carrier transit times. The bandwidth is limited due to the capacitive transient charging effect which arise from the capacitance C associated with the depletion region and the combined resistance of the load and the device R. The RC time constant is given by (6).

$$\tau_{RC} = RC \quad (6)$$

Another factor that limits the bandwidth of a photodetector is the time it takes the carriers to cross the depletion region. The electron and hole transit times are given by (7) where the carrier velocities are assumed to be independent of position.

$$\tau_{tr} = \frac{d_{e,h}}{v_{e,h}} \quad \text{where } d_{e,h} = \text{distance traveled by the carrier} \quad (7)$$

$$v_{e,h} = \text{velocity of the carrier}$$

If the depletion region is wide, then the capacitance associated with the junction decreases and the RC time constant also decreases. However, the transit time increases due to the larger distance the carriers are required to travel. There is therefore a trade-off between both bandwidth limiting factors associated with the depletion width. The overall bandwidth of the photodetector is given by (8).

$$BW = \frac{1}{2 \cdot \pi \cdot (\tau_{RC} + \tau_{tr})} \quad (8)$$

There exist two types of bandwidths associated with an optical detector: electrical and optical bandwidths. The 3-dB electrical bandwidth refers to the frequency at which the detector output microwave power drops 3 dB down from the low-frequency reference level. The 3-dB optical bandwidth of the detector is defined as the modulation frequency at which the photocurrent drops to half its low-frequency level. The optical bandwidth is larger than the electrical bandwidth because as shown in (9) the optical power is proportional to the photocurrent, whereas the electrical power is proportional to the square of the photocurrent. Hence when the photocurrent falls to half its low-frequency value, the optical power is 3 dB down on the low-frequency reference level, but the electrical power is already 6 dB down.

I_{ph} = photogenerated current

P_{opt} = incident optical power

$$I_{ph} \propto P_{opt}$$

$$P_{electrical} \propto I_{ph}^2$$

$$\therefore P_{electrical} \propto P_{opt}^2 \quad (9)$$

2.2.3 Metal-Semiconductor-Metal Device

The Metal-Semiconductor-Metal (MSM) structure is a two terminal device consisting of a semiconductor material sandwiched between two metal contacts as shown in Figure 2.4.

Figure 2.4 Physical representation of the MSM

Each metal-semiconductor junction can be described as a Schottky barrier assuming certain conditions are met [40]. Since the MSM has two of these barriers, it is described as two Schottky diodes back-to-back, and hence the reason for its schematic representation as illustrated in Figure 2.5.

Figure 2.5 Schematic diagram of the MSM

At thermal equilibrium, the energy band diagram of the MSM is as given in Figure 2.6 where Φ_{n1} and Φ_{n2} represent the electron barrier heights and V_{D1} and V_{D2} represent the built-in potentials for contact 1 and 2 respectively. At thermal equilibrium the Fermi level inside a group of materials in intimate contact is independent of position [42], i.e. E_F is at the same energy level for all x . For a symmetrical MSM, the barrier potentials of the metals are the same ($\Phi_{n1} = \Phi_{n2}$) as well as for the built-in potentials ($V_{D1} = V_{D2}$). The current density J_{SM0} that flows from the semiconductor to the metal is opposite and equal to the current density J_{MS0} that flows from the metal to the semiconductor at each junction. This results in a zero net current density crossing the junction.

Figure 2.6 Energy band diagram and current densities of the MSM-PD at thermal equilibrium

However, when a positive bias is applied at contact 2, the junction at contact 2 becomes forward biased whilst the junction at contact 1 becomes reverse biased. The energy band diagram for the MSM structure now resembles that of a forward and reverse Schottky barrier diode as shown in Figure 2.7. A depletion region is formed at each

junction, and for small applied voltages across the MSM, the sum of the depletion widths is smaller than the thickness L of the semiconductor.

Figure 2.7 Energy band diagram of an MSM under a positive bias applied at contact 2

The applied bias V_{msm} is shared between the two contacts in such a way that $V_{msm} = V_1 + V_2$. Current continuity requirements demand that $J_{MSR} = J_{SMF}$. The current I_{msm} as shown in the insert of Figure 2.7 is limited by the reverse-biased contact and is therefore not very large. Under small bias conditions, the total current is predominantly due to the electron contribution. The hole current contribution becomes significant at higher biases. Sze et al. [41] give a detailed analysis of the MSM currents under different bias conditions.

The charge distribution and the electric field of the MSM at a small bias level is illustrated in Figure 2.8.

Figure 2.8 Charge distribution and electric field of the MSM

As the voltage increases, the depletion region of contact 1 becomes larger and the depletion region of contact 2 becomes smaller. Eventually the depletion regions touch each other resulting in the depletion width being equal to the thickness L of the semiconductor material. The voltage at which this happens is termed the reach-through voltage V_{RT} .

2.2.4 Metal-Semiconductor-Metal as a Photodetector

The MSM device can be used to detect optical energy and is more appropriately referred to as a metal-semiconductor-metal photodetector (MSM-PD). The MSM-PD operates on the same principles as the pn photodetector described in section

2.2.1 where the optically generated carriers are swept across the depletion region by the electric field. The carriers collected at the terminals form the photocurrent of the device.

Therefore the MSM-PD has two sources of stimuli: voltage applied across the junction and light illuminating the active area as shown in the schematic diagram of Figure 2.9.

Figure 2.9 Schematic of MSM-PD under external stimulus

The MSM-PD is an interdigitated structure consisting of metal fingers on top of a semiconductor material as shown in Figure 2.10. The behavior of the MSM-PD depends on the finger width, the finger spacing, the optically sensitive area as defined by the fingers, the doping concentration, and the type of semiconductor material.

Figure 2.10 MSM-PD physical representation

When a potential difference is applied across the MSM-PD, one set of fingers will be at a higher potential than the other set as shown in the cross-sectional view of the MSM-PD in Figure 2.11. The back-to-back Schottky diodes are located between the fingers and the semiconductor material.

Figure 2.11 Cross-sectional view of MSM-PD under an applied bias and optical illumination

The bias applied across the device causes the formation of a depletion region between the fingers. As the bias increases, the depletion region associated with the reverse-biased junction approaches that of the forward-biased junction while the depletion region of the forward-biased junction decreases. At V_{RT} the depletion regions touch each other and the semiconductor is completely depleted. In the absence of light, the current generated is very small, being of the order of 40 nA for an MSM-PD with a $2\ \mu\text{m}$ finger width, a $2\ \mu\text{m}$ finger spacing and an area of $100 \times 100\ \mu\text{m}^2$ [53].

When the device is completely depleted, optical illumination can be applied because the generated carriers will be swept across the depletion region and collected to

form a current. Illuminating the MSM-PD with photons having energy greater or equal to the bandgap energy of the material will generate carriers in the depletion region. The optical beam is placed close to the MSM-PD so as to illuminate the active area which is defined by the fingers. Some of the optical power will be reflected from the metal finger and some will be reflected due to the reflective properties of the material.

The MSM-PD can be used as a photodetector when the depletion region extends throughout the semiconductor. The depletion region creates an electric field which sweeps the optically generated carriers across the depletion region to form a continual carrier movement as shown in Figure 2.12. The measured current is composed of two components: photocurrent I_{ph} and dark current I_{dark} . The contribution of the dark current to the overall current is not significant for the detection of high optical powers. The dark current becomes an important issue if the optical power to be detected is comparable to the dark current.

Figure 2.12 MSM-PD under applied bias and optical illumination

The photocurrent depends on the bias across the device and the light illuminating it. Typical DC I/V curves of the MSM-PD for positive applied bias levels are shown in Figure 2.13 where $P_{opt1} < P_{opt2} < P_{opt3} < P_{opt4} < P_{opt5}$. Since the device is symmetrical, a negative voltage applied across the device will yield equal and opposite current through the device compared to that obtained when a positive voltage is applied.

Figure 2.13 DC I-V relationship for different optical powers ²

An increase in voltage signifies an increase in the electric field across the depletion region resulting in an increase in carriers which are swept and collected at the terminals to create the photocurrent. The same figure also shows that the photocurrent increases as the optical power increases for any given voltage. An increase in the optical power increases the number of generated carriers in the depletion region, hence increasing the photocurrent.

² Measurements performed on the MSM-PD used in this thesis

2.2.5 Figures of Merit of the MSM-PD

The MSM-PD is characterized by its dark current, quantum efficiency, sensitivity, capacitance and transit time. Different finger spacing yield different capacitance and transit times. The finger spacing and width determine the responsivity of the device. The dark current depends on the doping profile of the MSM-PD. Quantum efficiency depends on the semiconductor material used (absorption coefficient). The sensitivity of the material also depends on the material bandgap. These parameters are the same as for the photodetector described in Section 2.2.2.

The capacitance of the MSM-PD is normally specified for two voltage ranges: voltages less than or greater than the reach-through voltage V_{RT} .

$$\underline{V < V_{RT}}$$

Since the depletion regions are not touching in this case, the capacitance per unit area can be treated as two capacitances in series as shown in (10).

$$C = \frac{C_1 \cdot C_2}{C_1 + C_2} \tag{10}$$

$$\text{where } C_1 = \sqrt{\frac{q \cdot \epsilon_0 \cdot \epsilon_r \cdot N_D}{2 \cdot (V_1 + V_{D1})}} \quad C_2 = \sqrt{\frac{q \cdot \epsilon_0 \cdot \epsilon_r \cdot N_D}{2 \cdot (V_{D2} - V_2)}}$$

V_1, V_2 = voltages across contacts 1 and 2 respectively

V_{D1}, V_{D2} = built-in potentials for contacts 1 and 2 respectively

$$\underline{V > V_{RT}}$$

The semiconductor is now completely depleted, and the capacitance per unit area is the same as that for a parallel plate capacitor [41] as shown in (11).

$$C = \frac{\epsilon_0 \cdot \epsilon_r}{L} \quad (11)$$

2.2.6 MSM-PD Circuit Configurations

The MSM-PD can be used in a 1- or 2-port configuration as shown in Figure 2.14. The MSM-PD is typically DC biased through one port at the desired potential whilst the other is grounded regardless of the port configuration. The port configuration becomes important when considering the AC behavior of the MSM-PD. The circuit topology of the MSM-PD is typically a 1-port configuration when used to detect a constant or intensity modulated optical signal. The generated photocurrent is collected at the load. The MSM-PD can be connected in a 2-port configuration when used as an optoelectronic mixer. The AC voltage signal is superimposed onto the DC bias across the device. The generated AC current appears at the load of the second port.

Figure 2.14 MSM-PD connected in a 1-port and 2-port configuration

2.3 Mixers

A mixer is a device capable of frequency conversion. The mixer is respectively termed an upconverter or a downconverter if the frequency of a signal is translated to a higher or lower frequency. Mixing is caused by devices that have a current-voltage relationship that is nonlinear, time-variant or both. A mixer will produce an output composed of many different frequencies which are not present at the input. The desired frequency component is selected using an appropriate filter.

A mixer is a three-port device with two input ports and one output port. A downconverter is illustrated in Figure 2.15. The signal at the R-port is termed the RF signal and represents the information signal into the mixer. The signal at the L-port is referred to as the LO (local oscillator) signal or the pump signal of the mixer. The I-port contains all of the output products generated by the mixer. The IF (intermediate frequency) signal refers to the downconverted signal generated by the mixer. A similar description is applicable to an upconverter except that the IF signal is the input and the RF signal is the upconverted output signal.

Figure 2.15 Schematic diagram of a downconverter

The LO signal is usually of larger amplitude than the RF signal, and is referred to as a pump signal. The function of the LO signal is to modulate the conductance of the device. This modulation causes the mixing phenomena which will be described later. A higher LO results in better intermodulation suppression compared to the case where the LO and RF signals have similar power levels. The LO signal should be at least 20 dB higher in power than the RF signal [43].

The output frequencies generated by the mixer are given by (12) which include a DC term, fundamental, harmonics, as well as the sum and difference frequency components of the input signals. The products most desired are the sum and difference of the fundamental RF and LO signals where $m = n = 1$ giving $f = \pm f_{LO} \pm f_{RF}$. The other products will be referred to as higher order intermodulation products.

$$f = \pm m \cdot f_{RF} \pm n \cdot f_{LO}$$

(12)

where $m = 0, 1, 2, 3, \text{etc.}$
 $n = 0, 1, 2, 3, \text{etc.}$

Any nonlinear device can be used as a mixer such as diodes, BJTs, FETs, etc. The choice of nonlinear device depends upon considerations of gain (or loss), noise figure, stability, dynamic range and generation of undesirable frequencies. The degree of the nonlinearity will determine the number of frequencies generated. Different devices will yield different weights on the various frequency components. A brief description of diode mixer theory is given in the next section. The diode theory is discussed since it is similar to the MSM-PD in that they are both two-terminal devices.

2.3.1 Diode Mixer Theory

A diode can be used as a mixer because it has a nonlinear current-voltage relationship as shown in equation (13).

$$I_D = I_s \left(e^{V_D / n \cdot k \cdot T} - 1 \right) \quad (13)$$

where I_s = saturation current

q = electron charge = $1.602 \cdot 10^{-19}$ [C]

n = emission coefficient or ideality factor

k = Boltzman constant = $1.38 \cdot 10^{-23}$ [J / K]

T = junction Kelvin temperature [K]

V_D = applied voltage

I_D = diode current

The diode's exponential relationship can be expanded into a power series as shown in equation (14).

$$\begin{aligned} I_D &\cong I_s e^{(V_D / V_T)} \\ &= I_s \sum_{n=0}^{\infty} \frac{V_D^n}{V_T^n \cdot n!} \\ &= I_s + \frac{I_s}{V_T} \cdot V_D + I_s \cdot \frac{V_D^2}{V_T^2 \cdot 2} + I_s \cdot \frac{V_D^3}{V_T^3 \cdot 6} + \dots \end{aligned} \quad (14)$$

where $V_T = [q / (n \cdot k \cdot T)]^{-1}$

When two or more sinusoidal signals are applied simultaneously across the diode, the signals will mix and produce mixing components. As an example, assume that the applied voltage V_D is the superposition of two sinusoidal voltage signals each having different frequencies and an operating DC voltage V_p . The applied voltage is given in (15), and the square term of the applied voltage found in the power series is given in (16).

$$V_D = V_p + V_1 \cos(\omega_1 t) + V_2 \cos(\omega_2 t) \quad (15)$$

$$V_D^2 = V_p^2 + \frac{V_1^2}{2} + \frac{V_2^2}{2} + 2V_p V_1 \cos(\omega_1 t) - \frac{V_1^2}{2} \cos(2\omega_1 t) + 2V_p V_2 \cos(\omega_2 t) - \frac{V_2^2}{2} \cos(2\omega_2 t) + V_1 V_2 \cos(\omega_1 - \omega_2)t + V_1 V_2 \cos(\omega_1 + \omega_2)t \quad (16)$$

Equation (16) shows that a device whose current depends on the square of the applied voltage has the following current components: DC term, fundamentals and second harmonics of both input signals, as well as the sum and difference between the two signals. The cubic term of the applied voltage yields the same terms as the square of the applied voltage, as well as the third harmonics of the input signals, and the $2f_1 \pm f_2$ and $2f_2 \pm f_1$ intermodulation products. Higher order terms of the applied voltage in (14) will yield higher order harmonics and intermodulation products.

The above mathematical development shows that mixing is possible if the output current depends on the square of the voltage. Equation (16) also shows that the amplitude of the sum and difference frequency components are dependent on the product $V_1 V_2$ of the input signal amplitudes. If both input amplitudes are small then the amplitude of the sum and difference frequencies will be insignificant compared to that of the other generated frequencies. Consequently, the sum and difference frequency terms are neglected and mixing does not occur. However, if one of the input signal amplitudes is large, then the sum and difference components cannot be neglected.

A more conceptual way to think about mixing, is to work in terms of device conductance modulation. Mixing occurs when the small-signal conductance of the device

varies as a function of bias. A graphical representation of the I-V relationship is shown in Figure 2.16.

Figure 2.16 I-V characteristic of a diode

An AC signal is superimposed onto the operating voltage. Assume that the signal variation is small enough that the conductance does not vary. Therefore the I-V relationship can be linearized about the operating point. The derivative of the I-V relationship at the operating point defines the small-signal conductance g_D of the device (17) and can be seen in Figure 2.16. Hereafter, the term conductance will be used to represent the small-signal conductance.

$$g_D = \left. \frac{dI_D}{dV_D} \right|_{V_p} = \left(\frac{q \cdot I_s}{n \cdot k \cdot T} \right) \cdot e^{V_p/q/nkT} \quad (17)$$

In the case of a mixer, if the input signals are small, the value of the conductance will not change as the voltage varies across the junction in a sinusoidal fashion. The diode can therefore be represented as a resistor. Therefore mixing is not possible. For mixing to

occur, the conductance must be modulated. Hence a large signal is needed to achieve the conductance modulation. So long as the conductance is varied, then mixing will occur. Only one of the input voltage signals needs to be large. The other signal can be small which is usually the case. If both signals are large, then more mixing products will be generated. This is seen in the generated current equation where large signals relate to more weight being applied to the other components which are otherwise neglected under small-signal conditions.

2.3.2 RF Mixer Figures of Merit

The performance of a mixer is crucial to the overall operation of the system since it is usually the first or second component from the RF input. The performance of the mixer is judged according to figures of merit such as: conversion loss, VSWR, isolation, dynamic range, intermodulation products, 3rd-order intercept point, and noise figure. These figures of merit will now briefly be reviewed.

1) Conversion Loss

The conversion loss (CL) is one of the most important figures of merit of a mixer since it represents the efficiency of the frequency conversion. The CL is defined as the difference in dB between the received signal power entering the R-port and the output power of the desired IF sideband exiting the I-port (18), (refer to Figure 2.15 for a graphical description of the ports).

$$CL = P_{RF}|_{dB} - P_{IF}|_{dB} \quad (18)$$

Normally only one of the up or down IF products is desired and the other product is filtered out. This results in the loss of half the power and leaves an automatic minimum 3 dB conversion loss. Further losses occur because some of the power is lost to unwanted higher-order mixing products. Other power losses are in form of heat loss from the mixing device and from the mismatch at the mixer ports. Typical range for single sideband conversion loss is from 6-9 dB [43] for diode type mixers. Conversion loss has a strong dependence on the LO power since changing the LO signal (pump) will affect the impedance, and hence the match between the system and the mixer.

2) Voltage Standing Wave Ratio (VSWR) & Return Loss (RL)

VSWR and RL are measures of mismatch between the system and the mixer. These parameters are given for the RF and IF ports, and they are usually specified over a given bandwidth and are a function of the LO power and temperature. A change in the LO power results in a change in the operating point of the device, hence changing the impedances at all of the mixer ports. An impedance change affects the mismatch between the system and mixer. The relationship between the VSWR and the reflection coefficient is given by (19).

$$VSWR = \frac{1 + |\rho|}{1 - |\rho|} \quad (19)$$

3) Isolation

Interport isolation specifies the insertion loss between any two mixer ports. It is usually specified over a given bandwidth and is a function of the LO power and temperature. Maximum interport isolation is important to achieve because unwanted

signals can obscure the desired signals at a port. Normally the isolation between the L and R port is given as well as the isolation between the L and the I ports. The RF-to-IF isolation is not as important as the other isolation parameters because the RF signal, being weak, will not have a significant effect at the other port. The interport isolation of the RF, LO or IF signals is calculated using equation (20). P_{in} represents the power level of the desired signal at any port, and P_{out} represents the power level of the same signal at another port.

$$\text{Interport isolation} = P_{in}|_{dB} - P_{out}|_{dB} \quad (20)$$

4) 1-dB compression point

This parameter, expressed in dBm, describes the RF power at which the conversion loss increases by 1 dB. The 1-dB compression point is the RF power at which the IF to RF relationship deviates from linearity by 1 dB as shown in Figure 2.17. For a diode type mixer, the 1-dB compression point is generally 5 to 10 dB lower than the LO input power [43].

Figure 2.17 Input and output characteristic of a mixer

5) Noise Figure

Noise figure is defined as the difference in dB between the input RF signal-to-noise ratio and the output IF signal-to-noise ratio. It is a single sideband measurement that is taken for either the up- or down-converted IF product.

6) 3rd-Order Intermodulation Distortion

The mixer will generate two-tone intermodulation products when two input RF signals are incident at the mixer's R-port. The intermodulation products have the following frequencies: $(2 \cdot f_{R2} - f_{R1}) \pm f_{LO}$.

The two RF signals whose frequency separation is small are combined and fed into the RF input to the mixer. The intermodulation product signal strength is measured for a range of RF signal strength and is also shown in Figure 2.17. A 1 dB increase in the RF signals will lead to a 3 dB increase in the intermodulation product signal strength. The power at which this relationship intercepts the fundamental relationship is termed the 3rd-order intercept point. This parameter defines the upper limit of the dynamic range, which is the input power range over which a mixer is usable. The lower limit of the dynamic range is defined by the noise figure.

2.3.3 The Optoelectronic Mixer Process

In a conventional mixer, the conductance of the nonlinear device must be modulated in order to observe the mixing phenomena. Different DC operating points will yield different conductance modulations.

Conductance modulation is achieved by applying a large sinusoidal voltage across the nonlinear device. This signal is superimposed onto the DC operating point as well as a small sinusoidal voltage. A signal is "small" if the nonlinear voltage-to-current relationship of the device can be linearized about the operating point where the small signal is superimposed. The conductance is assumed to be constant as the signal goes through its cycle at the DC operating point. A change in bias point will change the linearization, hence the conductance. The application of a large voltage signal implies a continual change in the conductance since the voltage has a nonlinear relationship with the current. If the relationship were linear, then conductance would remain constant and no mixing would occur.

Paralleling the conventional mixer process to the optoelectronic mixer process, the conductance of the optoelectronic mixer must also be modulated. Since the optoelectronic mixer has two stimuli, its conductance can be modulated through the use of a large optical or large voltage signal.

An optically-pumped mixer refers to an optoelectronic mixer in which the optical signal is large. The optical signal is therefore responsible for the modulation of the conductance, and hence the generation of mixing products. The electrical signal is small

and does not significantly change the conductance. Similarly, an electrically-pumped mixer is an optoelectronic mixer in which the electrical signal is large in comparison to the modulated optical signal. The electrical signal will modulate the conductance and be the cause of the mixing effect.

2.3.4 Operating Point of the OEM

The design of an efficient optoelectronic mixer demands the appropriate selection of the MSM-PD's operating point, i.e., bias and optical power. This section will outline the operating point in an optically and an electrically pumped mixer to achieve optimum mixing. It will also yield the necessary conditions needed to obtain mixing.

The investigation into the optimum optoelectronic mixing conditions starts by examining the I/V characteristics of different devices. Mixing will not occur for a device whose slope (small-signal conductance) as a function of the optical power remains constant. The $I-V$ curves for a device yielding no mixing products is shown Figure 2.18.

Figure 2.18 DC I/V curves for different optical powers for a device in which no optoelectronic mixing occurs

Different optical powers yield different DC I/V relationships where each DC I/V curve is described by a different equation as shown in (21).

$$\begin{aligned} P_0 &\rightarrow I(V) = \alpha_1 \cdot V - 0 \\ P_1 &\rightarrow I(V) = \alpha_1 \cdot V - a_1 \\ P_2 &\rightarrow I(V) = \alpha_1 \cdot V - 2 \cdot a_1 \end{aligned} \quad (21)$$

This set of equations shows that the current is dependent on both bias and optical power. The family of curves of the current can be represented as one equation (22) where the slope of the DC I/V curve does not vary as a function of the optical power (23) whereas the a-axis intercept does vary with optical power (24). The optical power dependence of the a-intercept is shown in Figure 2.19.

$$I[V, P] = \alpha(P) \cdot V + a(P) \quad (22)$$

$$\alpha(P) = \alpha_1 \quad (23)$$

$$a(P) = -\beta \cdot P + a_1 \quad (24)$$

Figure 2.19 a-axis intercept as a function of optical power

The overall current equation (25) is obtained by substituting the slope equation (23) and the a-intercept equation (24) into the current equation (22). Time dependence is

incorporated into the resultant current equation using the following mathematical descriptions for the sinusoidal voltage and optical signals superimposed onto a DC level:

$$V(t) = V_1 + v_1 \cdot \cos(\omega_{RF} \cdot t)$$

$$P(t) = P_1 + p_1 \cdot \cos(\omega_{LO} \cdot t)$$

Thus:

$$I[V(t), P(t)] = \alpha_1 \cdot [V_1 + v_1 \cdot \cos(\omega_{RF} \cdot t)] - \beta \cdot [P_1 + p_1 \cdot \cos(\omega_{LO} \cdot t)] + a_1 \quad (25)$$

There are no sum and difference frequencies ($f_{RF} \pm f_{LO}$) present in the output current as shown in equation (25). Therefore a device with a DC I/V characteristic where the small-signal conductance does not change as a function of optical power as shown in Figure 2.18 is not suitable for optoelectronic mixing.

The device with DC I/V characteristics that vary as a function of the optical power as shown in Figure 2.20 is examined as a potential optoelectronic mixing candidate.

Figure 2.20 DC I/V curves for different optical powers for a device in which optoelectronic mixing does occur

Different optical powers yield different DC I/V characteristics where each DC I/V curve is given by a separate equation as shown in (26).

$$\begin{aligned} P_0 &\rightarrow I(V) = \alpha_1 \cdot V + 0 \\ P_1 &\rightarrow I(V) = 2 \cdot \alpha_1 \cdot V - a_1 \\ P_2 &\rightarrow I(V) = 3 \cdot \alpha_1 \cdot V - 2 \cdot a_1 \end{aligned} \quad (26)$$

This family of curves is again represented by equation (22). The equation representing the slope of the DC I/V curves (27) as well as the a-intercept (28) were calculated from Figure 2.21 and Figure 2.22 respectively. As in the development of equation (25), the slope from equation (27), the a-intercept from equation (28) and the time dependence were incorporated into (22). The voltage and optical signals are represented by a sinusoidal signals superimposed onto a DC level. The resultant current is given in equation (29).

Figure 2.21 Slope of DC I/V curves

Figure 2.22 a-intercept of DC I/V curves

$$\alpha(P) = \phi \cdot P \quad (27)$$

$$a(P) = -\eta \cdot P + a_1 \quad (28)$$

$$\begin{aligned}
I[V(t), P(t)] &= \phi \cdot [P_1 + p_1 \cdot \cos(\omega_{LO} \cdot t)] \cdot [V_1 + v_1 \cdot \cos(\omega_{RF} \cdot t)] - \eta \cdot [P_1 + p_1 \cdot \cos(\omega_{LO} \cdot t)] + a_1 \\
&= \phi \cdot V_1 \cdot P_1 + \phi \cdot V_1 \cdot p_1 \cdot \cos(\omega_{LO} \cdot t) + \phi \cdot v_1 \cdot P_1 \cdot \cos(\omega_{RF} \cdot t) + \frac{\phi \cdot v_1 \cdot p_1}{2} \cdot [\cos(\omega_{RF} - \omega_{LO}) \cdot t + \\
&\quad \cos(\omega_{RF} + \omega_{LO}) \cdot t] - \eta \cdot P_1 - \eta \cdot p_1 \cdot \cos(\omega_{LO} \cdot t) + a_1
\end{aligned} \tag{29}$$

The up and down frequency components ($f_{RF} \pm f_{LO}$) are present as well as a DC component and the fundamentals. Greater strength in the up and down frequency components can be achieved by increasing the parameters ϕ , v_1 , and p_1 . An increase in these parameters will also increase the strength of the fundamental and DC signals. An increase in ϕ signifies an increase in the rate at which the slope of the DC I/V curve changes ($\alpha(P) = \phi \cdot P$). An increase in ϕ can only be obtained by modifying the physical structure of the device. An increase in v_1 and p_1 can be achieved by increasing the AC amplitude of the voltage and the optical signals.

If the DC I/V relationship as a function of voltage, see Figure 2.20, were nonlinear, mixing would still occur so long as the slope (small-signal conductance) changes as a function of the optical power. The more nonlinear the DC I/V characteristics as a function of voltage, the stronger the mixing products but more frequencies will be generated due to the increased nonlinearity. A nonlinear DC I/V characteristic with no slope change as a function of optical power will only lead to the generation of harmonics.

Therefore, optoelectronic mixing will occur if the small-signal conductance of the MSM-PD changes as a function of the optical power. Furthermore, a greater sensitivity in the device small-signal conductance to the optical power will yield increase strength in the

mixer products. It should also be noted that a linear change in the small-signal conductance as a function of optical power will produce mixing components.

In a typical DC I/V plot for the MSM-PD, as shown in Figure 2.13, a small-signal conductance change is apparent at low bias levels. At high bias levels, the slope remains fairly constant for different power levels. Therefore, the MSM-PD should be operated in the region where the gradient is the most sensitive function of the optical power, this occurs near the zero bias point. To increase the strength in the optoelectronic mixing products, the optical signal should be modulated to a high degree, thus leading to greater changes in conductance.

2.4 Design Procedure

2.4.1 Introduction

The steps involved in the design of an optoelectronic mixer (OEM) are the same as those of a conventional microwave mixer. The OEM circuit consists of a nonlinear device as well as several circuit elements such as the input and output matching network, the bias network, IF return, and filters. The design of a mixer involves the individual design of each of these circuit elements.

The design of an OEM first involves the choice and characterization of a nonlinear device. The design of the other circuit elements is based on the impedances presented by the nonlinear device. One of the differences between the OEM and a conventional diode mixer is the nonlinear device. The nonlinear device of an OEM is subjected to optical illumination and hence is dependent on optical power as well as on voltage.

Once the nonlinear device has been chosen, there are two possible design approaches. The first approach involves the design of the OEM based on the model of the nonlinear device. An accurate model will simulate the behavior of the nonlinear device under both small and large-signal conditions. The model will yield the impedance seen at the input and output of the nonlinear device when subjected to the desired power levels. The design of the matching networks, filters and IF return are subsequently based on these simulated impedances.

The second design approach is undertaken in the event of the unavailability of a nonlinear model. This approach consists of designing the circuit elements based on (small-signal) S-parameter measurements of the nonlinear device. In reality, the OEM will be subjected to large signals, hence the real impedances will differ from the simulated impedances which were obtained under small signal conditions. This discrepancy between the real and simulated impedances will increase as the device's nonlinearity increases. The small-signal design of a device with a small degree of nonlinearity will yield acceptable results assuming that the impedances under large and small signals do not significantly differ.

Both approaches involve the characterization of the nonlinear device, which is the MSM-PD in this case. The first approach utilizes the measurements to develop the nonlinear model. The second approach designs the mixer circuit elements directly based on the measurements. Characterizing the device involves the measurements of the S-parameters at different bias levels applied across the device. For an OEM, the S-

parameters are also measured for different optical power levels illuminating the device. The photocurrent is also measured for different bias levels and optical powers.

2.4.2 Small Signal (S-Parameter) Design

This design approach is based on the S-parameters measured at the desired operating conditions (bias and optical power levels). The operating conditions are chosen to yield maximum optoelectronic mixing products.

The bias network, IF return and filters are designed in isolation according to desired requirements. The purpose of the input and output matching network is to respectively allow maximum power transfer of the RF signal from the input to the MSM-PD and maximum power transfer of the IF signal from the MSM-PD to the output. The input matching network is designed to transform the input impedance of the MSM-PD into the $50\ \Omega$ of the generator. The output matching network is designed to transform the output impedance of the MSM-PD into the $50\ \Omega$ load. These impedance or reflection coefficient transformations are shown in Figure 2.23.

Figure 2.23 Impedance transformations for maximum power transfer

The bias network, IF return, filter and the nonlinear device are placed together prior to the design of the input and output matching networks because their presence will affect the impedance of the circuit. The input impedance transformation is performed by conjugately matching the input impedance of the nonlinear device with the impedance seen looking into the bias network at the desired input frequency ($\Gamma_1 = \Gamma_{in}^*$). The output impedance transformation conjugately matches the impedance seen looking into the output port of the MSM-PD to the load impedance at the desired output frequency ($\Gamma_2 = \Gamma_{out}^*$).

The input matching network is designed prior to the output matching network. The output match is subsequently designed with the input matching network present in the circuit since the presence of the input matching network will affect the impedance of the circuit. The designed output matching network is placed in the circuit, and it will consequently load the circuit, i.e. the impedance seen looking into the nonlinear device is affected by the output matching network. The input match must be redesigned using the new impedance seen looking into the MSM-PD. However, the new input matching network will change the impedance needed to design the output matching network. Therefore, the design of the input and output matching network is an iterative process.

However, if the forward and reverse transmission coefficients, S_{12} and S_{21} , of the nonlinear device are small, then loading of the circuit does not affect the impedances looking into the input or output port of the nonlinear device. For small S_{12} and S_{21} , the reflection coefficients at the input and output of the nonlinear device are equal to the reflection coefficient of the nonlinear device as shown in (30). Therefore, the design of

the input matching networks is not affected by the presence of the output matching network and vice versa.

$$\Gamma_{in} = S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_2}{1 - S_{22} \cdot \Gamma_2} \quad \Gamma_{out} = S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_1}{1 - S_{11} \cdot \Gamma_1} \quad (30)$$

$$\Gamma_{in} = S_{11} \quad \Gamma_{out} = S_{22} \quad \text{if } S_{12}, S_{21} \text{ are small}$$

The small-signal approach suffers from the fact that the design is based on small-signal impedances whereas in reality large signals are present in the circuit. Therefore the signals are subjected to impedances different from the impedances used to design the matching networks. A small-signal approach is feasible if the device nonlinearity is small or if the applied signal is such that the conductance variation is small, hence leading to the linearization of the DC I-V characteristic equation in both cases.

2.4.3 Large Signal Design

A large-signal design is based on the impedances obtained from a nonlinear model of the device as opposed to the measured impedances obtained under small-signal conditions. This approach necessitates the development of a model which represents the behavior of the device under both small- and large-signal conditions. A model topology is chosen and the simulated S-parameter response is compared to the measured S-parameters for different bias levels. The development of a model used in an OEM would also involve the comparison of simulated and measured S-parameters at different optical power levels. To verify the accuracy of the nonlinear model, the model's simulated and measured S-parameters are compared at the different bias and optical powers levels. A model that

accurately represents the device under small-signal conditions for different stimuli levels is still not an absolute guarantee that the model accurately represents the device under large-signal conditions.

A nonlinear design approach will yield more accurate results compared to the small-signal approach. Although there are a few disadvantages in using the large-signal design approach. Modeling a device can be very time consuming and can lead to highly complex models having long simulation times under large-signal conditions. Convergence problems can also arise depending on the degree of complexity of the model. An accurate model will require very good experimental measurements, and the device must be measured exactly how it will be incorporated into the circuit.

Chapter 3: Equivalent Circuit of the MSM-PD

3.1 Introduction

A model provides a representation of the device's performance. The designer can use the model in a circuit and predict the circuit performance under different operating conditions, hence making circuit redesigns possible. Models are divided into three categories: physical models, empirical models and data-based models [44].

Physical models are based on obtaining solutions to the semiconductor carrier transport equations for a particular device. Physical models are capable of providing DC, transient and steady-state AC results. Physical models also provide insight to the physical aspects of the device operation aiding in the design and fabrication of an improved device. An advantage of this type of model is the ability to predict the performance of new devices which have not yet been experimentally characterized. Another advantage is that there is no need to change the model to obtain both small and large signal results. Physical models, however, lead to long computation time due to the complex nature of the charge transport equations involved. Accuracy must often be traded against computation time in obtaining a workable model of the device.

Empirical models are represented by an equivalent circuit whose components are derived both from device physics and empirical equations. Device nonlinear behavior is modeled with parameterized empirical equations that are functions of control voltages, currents or other variable parameters present in the circuit. The values are extracted by curve-fitting simulated results to measured data. The accuracy of this type of model

depends on the accuracy of the measured data, the curve-fitting process, and the complexity of the empirical equations. Some of the equation are simplified to facilitate their use in simulators. This simplification means setting limits on voltages, currents or any other control variable in the circuit.

Data-based models are based on measured data (S-parameters vs. bias) and are subsequently independent of technology and process. This type of model provides a higher degree of accuracy in circuit simulation, but does not provide insight to the design and fabrication of better devices.

The choice of model depends on the purpose of the simulation exercise. If further device improvement is desired then a physical model is advantageous. However, physical dependence of the components needs precise fabrication data which is not always readily available. An empirical model was chosen because it does not require knowledge of the detailed physical parameters of the device. The empirical model is also sufficient to determine the impedance of the device necessary for the OEM design.

3.2 Modeling procedure

A model must represent the device under static and dynamic conditions. A static condition exists when the stimulus remains constant after having reached steady-state. The analysis of a model under static conditions is termed the DC analysis. The AC analysis represents the device under a continually changing stimulus.

The empirical based lumped-element approach is chosen to model the MSM-PD. The purpose of the MSM-PD model as an optoelectronic mixer is to design matching networks for maximum power transfer. Therefore, the physical dependencies of the MSM-PD are not necessary and only the device's performance at the operating conditions is needed. An empirical type model suffices.

The next step in the development a model involves the selection of a topology that will represent the device under DC and AC conditions. The model should accurately represent the device with the least number of elements. A model can have many possible combinations of elements all yielding the same behavior. However, a combination of elements having no physical meaning diverts from the understanding of the device and does not aid in the improvement of the device. Therefore, each element in the model should represent a real physical behavior.

The modeling is divided into two stages. The DC performance of the device is modeled prior the AC performance since the AC response of the model depends on the DC components. The DC portion of the model is however not affected by the AC components under small-signal conditions. Device parasitics can dramatically affect the performance of the device, and hence must be taken into account in the model. External parasitics such as bond wire inductance must also be considered, and it would be ideal to eliminate their effect by performing measurements directly on the device. If this is not possible, the parasitics must be represented in the model by other circuit elements.

Once the topology has been chosen, a starting value is assigned to each element. A starting value is an estimation of the real value and can be obtained from equations or existing data. The element values are subsequently optimized within a range of the starting value to yield the closest representation of the model to the data. A range of values restricts the optimizer to a physically possible solution.

Once the optimizer has been exercised, the simulated results are compared to the measured data. If they are similar, the model accurately represent the device in a certain region. The model can hence be used to predict the device's performance under both static and dynamic conditions. However if the simulated data does not compare to the measured data, then the model is inaccurate. The next step involves the investigation of possible error sources, such as a physical effect that has not been taken into account leading to the omission of a representative element. Another source of error could be the inaccuracy of the measured data. Repeatability studies can be used to quantify measurement errors. Another possibility could be that certain physical effects cannot be modeled using a lumped-element approach.

The following sections will document how the empirical modeling procedure is applied to the MSM-PD.

3.3 Choice and Equivalent Circuit Topology of the MSM-PD

The metal-semiconductor-metal photodetector (MSM-PD) has been modeled in the past as a lumped-element equivalent circuit [45]-[52]. These classical models represent the MSM-PD as a photodetector and are inadequate in representing the MSM-

PD as an optoelectronic mixer. The OEM operates over a broad range of biases and light powers. The existing models are inadequate since they are valid only for a particular bias and optical power. The existing models can represent the MSM-PD as detectors of either constant or intensity modulated optical signals. If an intensity modulated optical signal is to be detected, the unmodulated optical carrier power is used in the model.

The physical topology of the MSM-PD is shown in Figure 3.1, and its equivalent circuit components as they relate to the physical structure are given in Figure 3.2.

Figure 3.1 Physical topology of MSM-PD

Figure 3.2 Equivalent circuit components related to the physical topology

The photogenerated current is represented by a current source of value I_{ph} . The inductance and resistance associated with the fingers are respectively represented by an inductance L_s , and resistance R_s . The depletion capacitance of the MSM-PD is represented by a capacitance C_{msm} . The capacitance of the metal fingers is represented by a shunt capacitance C_{metal} across the current source and junction capacitance. The shunt resistance R_{msm} is used to model the leakage current (dark current). This resistance is

usually very large since the dark current is very small, in the order of nanoamperes. These elements represent the extrinsic and intrinsic characteristics of the device. Parasitics such as pads and bond wires are not included, although they must be included if they are present in the measurements. The values of the elements will depend on the MSM-PD structure, material and measurement set-up.

The classical models and the model used for an optoelectronic mixer have the same elements as shown in Figure 3.2. The classical models however represent the photocurrent by a constant current source, whereas the OEM model represents the photocurrent by a bias and optical power dependent current source. The OEM model represents the MSM-PD over a range of voltages and light powers. To the author's knowledge, a lumped-element-based model representing the MSM-PD over a range of voltage and optical power has not been developed. The MSM-PD mixer model is an extension of the photodetector model, and the existing model is modified to cover a broader range of biases and light intensities. Figure 3.3 illustrates the MSM-PD model used as an optoelectronic mixer. This model represents the extrinsic and intrinsic parts of the MSM-PD and does not include the parasitic effects due to the measurement set-up.

Figure 3.3 Optoelectronic mixer MSM-PD equivalent circuit

An optoelectronic MSM-PD mixer model differs from the classical MSM-PD model in that the current source is bias and optical power dependent as opposed to an ideal independent source. The DC modeling section will determine the equation used to represent the photogenerated current of the MSM-PD. All other parameters are similar to the ones found in the classical model. The determination of the reactive elements is explained in the AC modeling section.

3.4 DC Equivalent Circuit for the MSM-PD

3.4.1 Introduction

The equivalent circuit representing the DC behavior of the MSM-PD is shown in Figure 3.4. The DC model includes finger resistance, R_{F1} and R_{F2} and the photogenerated current, I_{ph} . Evaluation of these element values involved measuring the MSM-PD under DC conditions and optimizing the element values until the model's DC response is equal to the measured DC response.

Figure 3.4 DC equivalent circuit of MSM-PD

3.4.2 Experimental Set-Up

The DC measurement involves applying a constant bias across the MSM-PD and illuminating it with a constant optical power. The current generated under these applied stimuli is termed the photocurrent of the MSM-PD. The photocurrent is measured for a range of biases and optical powers, hence producing a family of curves characterizing the device performance. The set-up used for the DC measurements, as well as for the AC measurements is shown in Figure 3.5.

Figure 3.5 DC and AC measurement set-up

The MSM-PD consists of Ti/Pt/Au contact pads and interdigitated fingers on top of an undoped semi-insulating GaAs substrate as shown in Figure 3.6. The photodetector has a $3\text{ }\mu\text{m}$ finger width, a $100\text{ }\mu\text{m}$ finger length, and an active area covering $100\times 100\text{ }\mu\text{m}^2$ which is defined by the fingers.

Figure 3.6 MSM-PD

An MSM-PD with a $5\text{ }\mu\text{m}$ finger spacing was chosen as the device in the OEM experiment. The finger spacing affects the responsivity and bandwidth of the

photodetector. Therefore selection of the finger spacing must consider the effect on these two parameters. The $5\text{ }\mu\text{m}$ finger spacing provided the MSM-PD with sufficient responsivity and bandwidth for the intended application.

The MSM-PD array was placed in a universal test fixture. This fixture allows device measurements to be performed without the need of connectors. The MSM-PD was illuminated with an 830 nm wavelength optical signal generated by a Sharp LT015MD semiconductor laser diode. The light from the laser diode was focused into a multimode optical fiber cable via a lens system. The laser module includes the semiconductor laser package, a lens system, and a $47\text{ }\Omega$ series resistor. The resistor was included in the laser module for optical power modulation purposes as will later be explained. A photodetector, included in the laser diode package, monitors the back reflection of the laser and is used to control the optical signal generated from the laser. The optical fiber was brought into close proximity of the active region of the MSM-PD using a micropositioner. The positioner was adjusted to obtain a maximum change in the measured magnitude of S_{11} compared to the no-optical-illumination situation. The change in $|S_{11}|$ was observed on the VNA (Vector Network Analyzer) screen. Since the fiber end was located very close to the active area of the MSM-PD, the entire area of the laser beam was assumed to fall within the sensitive area of the device. Approximately half of the optical power illuminating the active region is reflected due to the Ti/Pt/Au fingers, and another 30% of the light is reflected due to the reflection properties of the GaAs substrate. The laser output power and temperature were controlled by an ILX laser diode

controller. The temperature of the laser diode was kept at approximately 20° C. The laser was operated in constant output power mode.

The HP4142 was programmed via the Wafer Map software package³ to apply a range of bias values across the MSM-PD and to measure the photocurrent for each applied bias. The applied bias is not the bias applied across the MSM-PD because there are voltage drops across resistances associated with the measurement set-up. These must be taken into account to properly model the MSM-PD. The photocurrent measurement is repeated for different optical powers up to a few milli-watts. The DC measurement yielded a family of curves as shown in Figure 3.7.

Figure 3.7 Photocurrent as a function of bias for different optical powers

Figure 3.7 shows that the photocurrent increases as the voltage across the MSM-PD increases for various optical powers illuminating the device. Only positive voltages

were applied across the MSM-PD since it is a symmetrical device in theory. Hence, negative voltages were assumed to yield equal but negative photocurrent values.

3.4.3 DC Equivalent Circuit

The generated photocurrent is represented by a current source. The equation associated with the current source is dependent on the bias v_{ds} across the MSM-PD and the optical power P_{opt} . The photocurrent I_{ph} can be modeled by a 3rd-order polynomial expression as shown in (31).

$$I_{ph}(v_{ds}, P_{opt}) = a_0 + a_1(P_{opt})v_{ds} + a_2(P_{opt})v_{ds}^2 + a_3(P_{opt})v_{ds}^3 \quad (31)$$

$$a_1(P_{opt}) = b_0 + b_1 P_{opt} + b_2 P_{opt}^2$$

$$a_2(P_{opt}) = c_0 + c_1 P_{opt} + c_2 P_{opt}^2$$

$$a_3(P_{opt}) = d_0 + d_1 P_{opt} + d_2 P_{opt}^2$$

The a_i coefficients in the current equation depend on the optical power P_{opt} and are also given by polynomial expressions. A higher order polynomial in either the current or coefficient equations will yield a closer representation to the data but at the expense of computation time and convergence efficiency.

The current equation (31) can be simplified by expressing it as a function of the MSM-PD responsivity. The responsivity is defined as the ratio of photocurrent to optical power and is calculated from the DC I/V measurements of Figure 3.7. The responsivity as a function of bias of the MSM-PD for different optical power levels is given in Figure 3.8. For a fixed bias, the responsivity should remain independent of the optical power. The

³ Data acquisition software by Meta Graphics Software Corporation

responsivity of the MSM-PD has however a dependence on the optical power as shown in Figure 3.9. This behavior indicates from [59] that the GaAs MSM-PD exhibits a large trap-induced gain due to charge trapping close to the metal fingers.

Figure 3.8 Responsivity versus bias

Figure 3.9 Responsivity versus optical power

Assuming that the responsivity of the MSM-PD does not vary as a function of optical power (except at low powers), then the current source equation can now be simplified by expressing it as the responsivity multiplied by the optical power. The equation used to represent the responsivity can take any form as long as it accurately models the photocurrent. The responsivity R_{ph} is represented by a hyperbolic tangential equation as shown in (32). The hyperbolic function was chosen because it resembles the DC I/V curves in the low-bias range.

$$\begin{aligned}
 I_{ph} &= P_{opt} \cdot R_{ph} & [A] \\
 R_{ph} &= p_0 \cdot \tanh(p_1 \cdot v_{dsr}) \cdot (1 + p_3 \cdot |p_2 \cdot v_{dsr}|) & [A/W]
 \end{aligned} \tag{32}$$

where p_0, p_1, p_2 , and p_3 are real

A family of curves as shown in Figure 3.7 is obtained when equation (32) is plotted. Representing the current source by this tangential equation greatly reduces the complexity (computation time, convergence) but accurately models the current only over a narrow range of bias values. Depending on the region of operation, the coefficients of the hyperbolic equation will take on different values. The model will have to be re-optimized for the different bias regions of application to yield suitable coefficients.

The model is valid over a range of optical power and bias levels. The region of operation depends on the application. As shown in Chapter 2, optoelectronic mixing improves as the rate of change in small-signal conductance as a function of optical power increases. The conductance of the MSM-PD as a function of the optical power for different bias levels is shown in Figure 3.10. The conductance was obtained from the

DC I/V data of Figure 3.7 by taking the slope of the photocurrent as a function of voltage for different optical powers and bias levels.

Figure 3.10 Conductance as a function of optical power for different bias levels

The rate of change of the conductance as a function of the optical power is greater at low bias levels. Hence for an optoelectronic mixer, the MSM-PD will be biased near 0 volt. The current source coefficients are optimized so that the model generated current equals the measured photocurrent in the bias range of 0-2 volts. The coefficients are also optimized for an optical power level. The chosen optical power value was based on an optoelectronic mixing experiment previously performed on the MSM-PD where the optical power was varied until maximum mixing products were obtained. An optical power of 1.36 mW yielded the best mixing results

The current coefficients, p_0, p_1, p_2 , and p_3 in (32) were optimized using the simulation software package OSA90/HopeTM V2.5.⁴ The software optimized the coefficients until the current drawn by the model equaled to the measured current over a 0-2 volt bias range and for an optical power of 1.36 mW.

The resistances associated with the DC path of the Wiltron VNA, $R_{S1}, R_{S2}, R_{Ground1}$, and $R_{Ground2}$, were taken into account since they affect the DC behavior. These resistance values were measured and included in the model as shown in Figure 3.12. The MSM-PD finger resistance, R_{F1} and R_{F2} , were also included in the model. A 1.5Ω resistance (obtained from [53]) is used to represent the fingers on each side of the junction. The coefficients were then re-optimized to yield the best representation of the measured data.

The simulated and measured DC I/V curves for the MSM-PD over the 0-2 volt bias range and over the 0.1-1.79 mW optical power range are displayed in Figure 3.11. The measured and simulated photocurrent in this bias range yields a good fit for the optical power of 1.36 mW. A discrepancy exists between the measured and simulated photocurrent at the other optical powers because the model responsivity is not dependent on optical power whereas in reality it is, as was shown in Figure 3.9. A good fit should be possible for optical power levels in the vicinity of 1.36 mW, the value required for good optoelectronic mixing.

⁴ OSA90/HopeTM V2.5 is a general CAD software package capable of simulating, modeling, statistical analysis, nominal and yield optimization, and data visualization of linear and nonlinear analog circuits.

Figure 3.11 Simulated and measured DC I/V curves for different optical powers

The DC model of the MSM-PD including the resistance associated with the measurement set-up is illustrated in Figure 3.12. The values of the current coefficients and the resistance are given in Table 3.1.

Figure 3.12 DC equivalent circuit of the MSM-PD and associated resistance of the VNA

Table 3.1 DC Model Element Values

Parameter	Value
p_0	0.0494671
p_1	0.162863 [V^{-1}]
p_2	8.95531 [V^{-1}]
p_3	4.31077
R_{F1}	1.5 Ω
R_{F2}	1.5 Ω
R_{S1}	3.4 Ω
R_{S2}	1.6 Ω
$R_{Ground1}$	$3.49 \cdot 10^6 \Omega$
$R_{Ground2}$	$3.44 \cdot 10^6 \Omega$

3.5 AC Equivalent Circuit Model of the MSM-PD

3.5.1 Introduction

AC modeling consists of adding reactive elements to the completed DC model such that the AC behavior is now represented without modifying the DC behavior of the model. The equivalent circuit that represents both the DC and AC behavior of the MSM-PD is shown in Figure 3.13. The current source I_{ph} and the finger resistances R_{F1} and R_{F2} were determined in the DC modeling of the MSM-PD. All of the reactive elements were added to the DC model. The parasitic elements C_{pad1} , C_{pad2} , L_{bond1} , and L_{bond2} are associated with the mounting structure.

$$I_{ph} = P_{opt} \cdot R_{ph} \quad [A]$$

$$R_{ph} = p_0 \cdot \tanh(p_1 \cdot v_{dut}) \cdot (1 + p_3 \cdot |p_2 \cdot v_{dut}|) \quad [A/W]$$

where p_0, p_1, p_2 , and p_3 are real

Figure 3.13 DC and AC model of the MSM-PD

The values of the model elements are obtained by varying them over a predefined range. The element values are varied until the error between the simulated and measured AC response of the model becomes small. Optimization of the model requires AC data (S-parameter measurements). The optimizer also requires a starting value for each element as well as a range of values from which the final value will be chosen. The following sections describe each of these requirements.

3.5.2 Measurement Technique and MSM-PD Mounting Structure

S-parameter measurements were initially performed directly on the MSM-PD chip using microwave probes to characterize the device alone. GSG (ground-signal-ground), GS (ground-signal) by Cascade⁵ and 8-pin Picoprobes⁶ were used for S-parameter measurements. Each of these types of probes introduced a degree of uncertainty. The GSG probes might add an extra capacitance to ground due to the floating pin. The picoprobes did not yield this floating pin problem, however, a good calibration was not achieved. The GS probes didn't have this floating ground either, but the probe pitch was comparable to the MSM-PD pad separation. The measurements were jittery as a consequence of poor probe-to-pad contact. Another problem was that the ground contact was possibly not at the same potential as the negative pin on the probes. Therefore due to the above mentioned shortcomings, the probe measurements were replaced by standard coaxial-type measurements. This method yielded no contact or grounding problems.

This coaxial-type measurement required the MSM-PD chip to be connected to a transmission medium for subsequent testing. The MSM-PD chip was connected to transmission lines by bond wires. The effects of the transmission lines and bond wires were eliminated using a TRL (Thru-Reflect-Line) calibration which provides very accurate coaxial measurements [63]. This calibration determines the full 12-term error model and subsequently de-embeds the effects of the connectors, lines, etc... up to the plane of the device which, in this case, includes the bond wires.

⁵ Trade name

⁶ Trade name

Two mounting configurations of the MSM-PD were used to perform the TRL calibration technique. Mounting structure #1 is shown in Figure 3.14. The MSM-PD devices were available in an array block which consisted of 20 MSM-PD devices (10 rows by 2 columns) each having different finger spacing for the same active area. The array block was simply epoxied onto an etched portion of Duroid between two gold-plated 50 Ω transmission lines. The selected MSM-PD (5 μm finger spacing) was connected to the transmission lines using bond wires.

Figure 3.14 Mounting structure #1

The transmission lines are not considered part of the model since the TRL calibration eliminates all effects up to the bond wires. A starting value was assigned to each element of the MSM-PD model using mounting structure #1. The simulated response of the model was compared to the measured data. There was not a good correlation between both sets. A potential problem was the mounting configuration. The bond wires were very long due to the elevation of the device, and the difference in substrate height could introduce effects not modeled by lumped elements. The first mounting structure was replaced by a second configuration to eliminate these potential problems as shown in Figure 3.15. The second structure reduces the bond wire length, and maintains a uniform substrate height.

Figure 3.15 Mounting structure #2

The AC response of the model was simulated and compared to the AC data obtained from mounting structure #2. The experimental set-up for the AC measurements is the same as the DC measurements set-up shown in section 3.4.2. The mounting structure #2 or the device under test (DUT) was placed in the universal test fixture which allows the device to be measured without having to use connectors. As in the DC measurements, the HP4142 was programmed to apply a range of DC biases for different light powers illuminating the MSM-PD. S-parameter and DC data were measured for a range of bias and optical power levels. A sample of S-parameters at a certain optical power level for different bias is given in Figure 3.16.

Figure 3.16 Magnitude of S_{11} for different bias levels

3.5.3 AC Element Starting Values and Optimization Range

A starting value is assigned to each element of the model. The optimizer varies the values within a predetermined range of the starting values to obtain good agreement between the simulated AC response of the model and the AC experimental data. Since the optimizer can yield many solutions each having a similar response, the range of optimization must be limited to a physically realizable set of values. This reduces the possible combinations of values associated with each element of the model.

The determination of the starting value for each element will be described in this section. Each element is individually varied over a predetermined range to observe its effect on the circuit's AC behavior. These observations together with theoretical

limitations (tolerances, etc.) determine the optimization range associated with each element. The range of values for each element is given in Table 3.2. There is no minimum or maximum value for the finger series resistance, R_{F1} and R_{F2} , since they were determined in the DC modeling of the device.

Table 3.2 Element Value Range for Mounting Structure #2

Element	minimum value	starting value	maximum value
R_{F1}, R_{F2}		1.5 Ω	
L_{F1}, L_{F2}	5 pH	10 pH	50 pH
L_{bond1}	0.38 nH	0.47 nH	0.56 nH
L_{bond2}	0.33 nH	0.41 nH	0.49 nH
C_{metal}	0.01 pF	0.05 pF	0.1 pF
C_{mm}	0.08 pF	0.2 pF	0.3 pF
C_{pad1}, C_{pad2}	0.01 pF	0.015 pF	0.05 pF

Bond wires: L_{bond1}, L_{bond2}

The MSM-PD has two bond wires connecting each of its pads to a 50 Ω transmission lines. The inductance equation for a bond wire in free space and the mutual inductance between two parallel bond wires are respectively given in (33) and (34) [54]. The total inductance of two parallel bond wires is shown in (35).

$$L = 2 \cdot l \cdot \left[\ln\left(\frac{l}{d}\right) + 0.5 + 0.22 \cdot \left(\frac{2 \cdot d}{l}\right) \right] \quad [\text{nH}] \quad (33)$$

$$L_M = 2 \cdot l \cdot \left[\ln\left(\frac{l}{s}\right) - 1 + \left(\frac{s}{l}\right) - 0.25 \cdot \left(\frac{s}{l}\right) \right] \quad [\text{nH}] \quad (34)$$

$$L_{total} = \frac{L}{2} + L_M \quad [\text{nH}] \quad (35)$$

where l = length of bond wire (cm), d = diameter of bond wire (cm)
 s = separation between two bond wires (cm)

A tolerance of 20% is given to the inductance value to take into account errors due to the tolerance associated with the bond wire lengths and other non ideal conditions. The inductance for a $17.8\ \mu\text{m}$ diameter and gold bond wire was also calculated using the 0.8 nH/mm rule of thumb. The inductance values using both methods yielded approximately the same results.

Finger inductance: L_{F1}, L_{F2}

The inductance of each finger on each side of the device are lumped together as one equivalent inductance. The inductance of a finger is calculated assuming that the end of each finger is not open-ended. Equation (33) is used as an approximate equation to calculate the inductance of one finger. The inductance is divided by the total number of fingers since they are connected in parallel. The mutual inductance between the metal fingers is not taken into account, resulting in a possible discrepancy of the total inductance value. Again a tolerance of 20% is given to the inductance value.

Parameter optimizations of this element have been done in the past [53] using the same MSM-PD device with same active area but having different finger widths and spacings. The paper presents inductance values for a device having a smaller gap between fingers, subsequently resulting in more fingers. Fewer fingers increase the inductance, hence the values used in this simulation will be higher than the ones presented in [53].

Junction capacitance: C_{mm}

The value for the junction capacitance is determined from [45] where capacitive measurements were performed on MSM-PDs having the same finger spacing and

approximately the same finger width. The measurements yielded different values for different light power conditions and showed that the capacitance had a slight dependence on the voltage across the device.

Metal finger capacitance: C_{metal} and contact pad capacitance: C_{pad1}, C_{pad2}

The capacitance of a pad-to-ground and the capacitance between two sets of fingers was calculated using equations found in [55] and [56].

3.5.4 AC Equivalent Circuit

The next step involved in modeling is the optimization of the element values within the predetermined range. The optimizer compared the simulated AC response to the measured response. The stimulus applied to the circuit in the simulation must be the same as in the measurement for comparison purpose. The measurements used in the optimization were when the MSM-PD was biased at 0.1 volt and illuminated with 1.36 mW of optical power. The optical stimulus was chosen because it was found to yield the optimum optoelectronic mixing components when the experiment was previously performed on the MSM-PD alone (no matching networks). Figure 3.17-3.20 depicts the simulated AC response before and after optimization. The simulation of the model before optimization used the starting values for the element values. The optimization analysis resulted in the element values tending towards either their minimum or maximum values, as assigned for the optimization procedure. An improvement in the correlation between the measured and simulated responses is observed from the graphical data before and after optimization. The optimized model and parameters are presented at the end of the chapter.

Figure 3.17 Comparing measured and simulated (before and after optimization) magnitude and phase (in degrees) of S_{11}

Figure 3.18 Comparing measured and simulated (before and after optimization) magnitude and phase (in degrees) of S_{22}

Figure 3.19 Comparing measured and simulated (before and after optimization) magnitude and phase (in degrees) of S_{21}

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PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

PRECISIONSM RESOLUTION TARGETS

Figure 3.20 Comparing measured and simulated (before and after optimization) magnitude and phase (in degrees) of S_{12}

These graphs show that there still exists a discrepancy between the behavior of the MSM-PD model and the measurements. Since a discrepancy still existed between the simulated and measured S-parameters, measurements were taken on four different days (Day 1, Day 2, Day 3, Day 4) to check the repeatability of the measured data. Figures 3.21 - 3.28 each present the magnitude or phase (in degrees) of the S-parameters taken on the four different days. The MSM-PD was illuminated with 1.36 mW of optical power and biased near 0 volt. The magnitude and phase of the S-parameters (S_{11} , S_{21} , S_{12} , S_{22}) for all of the different days followed the same trend except for data collected on Day 4. Day 4 can be identified by the upside-down triangle data markers. A possible reason for the difference observed in the Day 4 data could be a poor calibration. It was concluded that the measurement errors were not sufficient to cause the discrepancy between the model's simulated response and the experimental data. Therefore any day (except Day 4) could be chosen as the data used to design the matching networks and model the MSM-PD. Day 3 (empty circle data points) was chosen since data was collected for a particular optical power needed for designing the matching networks.

Figure 3.21 $\text{Mag}(S_{11})$ [dB] of MSM-PD taken on four different days

Figure 3.22 $\text{Ang}(S_{11})$ [deg] of MSM-PD taken on four different days

Figure 3.23 Mag(S_{21}) [dB] of MSM-PD taken on four different days

Figure 3.24 Ang(S_{21}) [deg] of MSM-PD taken on four different days

Figure 3.25 $\text{Mag}(S_{12})$ [dB] of MSM-PD taken on four different days

Figure 3.26 $\text{Ang}(S_{12})$ [deg] of MSM-PD taken on four different days

Figure 3.27 $\text{Mag}(S_{22})$ [dB] of MSM-PD taken on four different days

Figure 3.28 $\text{Ang}(S_{22})$ [deg] of MSM-PD taken on four different days

To further investigate the discrepancy between the simulated and measured AC response, an RF signal was superimposed onto the DC bias. Since the RF signal used to measure S-parameters is a small signal, the generated DC current should be the same as when no RF signal is applied. This was experimentally proven to alleviate any possible source of error. The MSM-PD was illuminated with a constant optical power. There was no significant change in the photocurrent when a modulated electrical power of less than -20 dBm (small signal) was superimposed on the DC bias. The frequency of the modulated signal was swept from 10 MHz to 20 GHz. However, for powers greater than approximately -5 dBm, the photocurrent was not the same as under small signal conditions. The difference was significant at 0 dBm. Under large signal conditions, the difference in photocurrent was noticeable at low frequencies but gradually disappeared above 10 GHz.

This effect is important if the MSM-PD is to be modeled as an electrically-pumped mixer where the input microwave signal is large. If the model is fitted over a wide range of voltages, then the model could handle large RF signals. The model developed in this thesis was only designed for a limited range, and is therefore only applicable to an optically-pumped mixer. For an electrically-pumped mixer, the current source representing the device's photocurrent would have to be modified to represent the device over a large range of biases.

The MSM-PD model is re-illustrated in Figure 3.29 for convenience, and the optimized component values are given in Table 3.3. This model was designed to represent the MSM-PD over the bias range of 0-2 volt and at an optical power of 1.36 mW.

$$I_{ph} = P_{opt} \cdot R_{ph} \quad [A]$$

$$R_{ph} = p_0 \cdot \tanh(p_1 \cdot v_{ds}) \cdot (1 + p_3 \cdot |p_2 \cdot v_{ds}|) \quad [A/W]$$

Figure 3.29 Optimized equivalent circuit of the MSM-PD

Table 3.3 Final Element Values for the MSM-PD Equivalent Circuit

Element	final value
p_0	0.0494671
p_1	8.95531 [V^{-1}]
p_2	0.162863 [V^{-1}]
p_3	4.31077
R_{F1}, R_{F2}	1.5 Ω
L_{F1}, L_{F2}	50 pH
L_{bond1}	0.56 nH
L_{bond2}	0.49 nH
C_{metal}	0.01 pF
C_{mem}	0.08 pF
C_{pad1}, C_{pad2}	0.005 pF

The simulated and measured S-parameters for the bias range of 0-0.5 volt and an optical power of 1.36 mW is shown in Figure 3.30.

Figure 3.30 Simulated and measured $\text{Mag}(S_{11})$ for a range of biases and $P_{\text{opt}} = 1.36 \text{ mW}$

3.5.5 Conclusion

There is a discrepancy between the simulated and measured S-parameters. Improvements to the model and measuring techniques are discussed in Chapter 5. The validity of the model can be determined by comparing the simulated response from a large-signal analysis of the model with large signal measurements performed on the MSM-PD.

Chapter 4: Design of an Optically-Pumped, MSM-PD, 5.3-to-3.3 GHz Downconverter

4.1 Introduction

The OEM was designed for integration into a conventional phased-array receiver such as illustrated in Figure 4.1. The receiver downconverts the incoming signal for subsequent signal processing. The incoming signal is weak and must be pumped using a local oscillator (LO) signal provided by the signal processing unit. The LO signal is brought to the front end via conventional waveguide or coaxial transmission medium.

Figure 4.1 One element of a conventional phased-array receiver

In an optical receiver, the fiber optical cables replace the conventional waveguide or coaxial transmission medium. In such a system, the optical signal must be converted into an electrical signal prior to entering the conventional mixer as shown in Figure 4.2.

Figure 4.2 One element of a phased-array receiver using fiber optics as the distribution system

An optoelectronic mixer is advantageous in that it replaces the photodetector, amplifier and mixer as illustrated in Figure 4.3. An OEM simultaneously detects the optical signal and performs frequency conversion of the input signal, thus providing a marked reduction in the receiver complexity, fabrication cost, and design time.

Figure 4.3 One element of a phased-array receiver using an OEM

One of the requirements for the phased-array receiver is to receive a 5.3 GHz signal and downconvert it into a 3.3 GHz signal. The schematic of the OEM with these frequency requirements is shown in Figure 4.4.

Figure 4.4 Schematic of OEM application

Conventional mixer terminology represents the weak signal as the RF and the strong (pump) signal as the LO. A strong LO signal is required to modulate the conductance of the device and subsequently convert the RF frequency to an IF frequency. The optoelectronic mixer under study uses the optical signal as the pump signal.

A mixer is composed of passive circuits as well as a nonlinear device. Hereafter the term mixer will refer to the entire mixer circuit including the nonlinear device and the other passive circuits. This chapter describes a design procedure for the optoelectronic mixer based on small-signal (S-parameter) measurements. Results and figures of merit are presented for the optoelectronic mixing measurements.

4.2 Optoelectronic Mixer Design Based on S-Parameters

4.2.1 Optoelectronic Mixer Circuit

The optoelectronic mixer can be broken down into five sub-circuits as shown in Figure 4.5: input matching network, bias network/IF return, nonlinear device, output matching network and lowpass filter.

Figure 4.5 Optoelectronic mixer module

The input matching network provides the maximum RF power transfer from the input port to the mixing device. The bias network, in conjunction with the DC return, provides the desired bias across the MSM-PD. The IF return impedes any generated IF to travel back to the input port. The metal-semiconductor-metal photodetector (MSM-PD) is the nonlinear element that will produce the mixing components. The output matching network provides the maximum power transfer of the IF to the output port. Finally, the lowpass filter passes the desired downconverted IF frequency and blocks other higher generated frequencies.

Each circuit was individually designed to achieve certain requirements which will be later explained. All of the designs were done using HP-EEsof/Libra V6.0 simulation

software package unless otherwise mentioned. The entire mixer module was simulated once all of the individual circuit designs were complete. Optimization was performed on the matching networks to improve the input and output match of the overall circuit.

Each circuit was designed based on a distributed (transmission lines) approach rather than a lumped element approach. Lumped elements are difficult to model at high frequencies due to their associated parasitics. At these frequencies, transmission lines are accurately modeled, and the size of the lines is not a physical problem. The circuits were fabricated using microstrip technology on a RT/Duroid dielectric material which is 25 mil thick, has a relative dielectric constant of 10.2 and is plated on both sides with 0.5 oz rolled copper. One side of the material was used for the design and the other side was used as the ground plane. The material parameters are given in Table 4.1

Table 4.1 RT/Duroid Substrate Parameters

Parameter	Value
dielectric constant	10.2
metal thickness	0.7 mil
dielectric height	25 mil
effective dielectric constant	6.77
50 Ω line width	23 mil

4.2.2 Lowpass Filter (LPF)

The lowpass filter used in this application needs to pass the IF signal of 3.3 GHz. The filter also needs to reject higher frequencies such as the 5.3 GHz RF signal as well as the second harmonics of the RF (10.6 GHz) and LO (4 GHz). The LPF needs to provide a short for the RF signal and a through for the IF signal. The provided short will

reflect the RF signal with a 180° phase shift so as to not let any RF power through to the output. A further requirement of the LPF is that it must present the minimum attenuation to the IF signal. The filter was designed using a circuit simulator software package called Eagleware⁷. The design was then imported to HP-EEsof/Libra V6.0 where the values of the line lengths and widths were optimized to approach the desired specifications as closely as possible.

The chosen filter topology was an elliptical filter which consists of series inductors and a combination of inductors and capacitors in shunt. The narrow transmission lines represent the inductors and open-circuit stubs represent the capacitors to ground. The elliptical filter differs from a conventional lowpass filter in that its shunt elements consist of the combination of inductors and capacitors instead of only capacitors. Both types of LPFs have similar responses in the passband, but differ in the rejection band. The elliptical filter produces a series of peaks and valleys once the rejection frequency is reached. A conventional lowpass response attenuates in a continuous fashion. The elliptical filter was chosen because it provides a sharper transition from the passband to the rejection band compared to the conventional lowpass filter. The sharp transition is needed because the RF and IF frequencies are close together. Figure 4.6 illustrates the physical layout of the filter along with the physical dimensions.

⁷ Simulator package

Figure 4.6 Physical layout of the lowpass filter

The magnitudes of the reflected and transmission coefficients at port 1 are respectively illustrated in Figure 4.7 and Figure 4.8. The filter is a passive and bi-directional device ($|S_{11}| = |S_{22}|$ and $|S_{21}| = |S_{12}|$), and hence the simulated magnitudes of the reflection and transmission coefficient at port 1 are only given.

Figure 4.7 Magnitude of S_{11} [dB] of the lowpass filter

Figure 4.8 Magnitude of S_{21} [dB] of the lowpass filter

The phase of the reflection coefficient is different at the two ports since the filter is not a symmetrical device. The phase information of the reflection coefficient at port 1 is given in Smith chart form as shown in Figure 4.9. The 5.3 GHz RF signal sees a short and the 3.3 GHz IF sees a through at the input to the filter as seen in the Smith chart.

Figure 4.9 Simulated S_{11} for the lowpass filter

The reflection coefficient of Figure 4.7 shows that the filter is a good match at the desired 3.3 GHz IF frequency. The bandwidth of the IF signal at 3.3 GHz is approximately 92 MHz (2.8 %). This bandwidth is the frequency range over which the input reflection coefficient about 3.3 GHz is -20 dB or less. The cutoff frequency and insertion loss of the filter are calculated using the transmission coefficient given in Figure 4.8. The cutoff frequency is the frequency at which the response falls 3 dB below the insertion loss of the filter. The insertion loss of the filter is defined as the average internal loss through the filter in the passband. This filter has an insertion loss of 0.3 dB, a cutoff frequency of 3.725 GHz, and a passband ripple of ± 0.3 dB. The magnitudes of the reflection and transmission coefficients for the frequencies of interest are respectively given below the graphs in Figure 4.7 and Figure 4.8. The filter provides a return loss of 0.08 dB and an insertion loss of 70 dB at 5.3 GHz.

4.2.3 Bias Network and IF Return

The bias network will provide a potential difference across the nonlinear device. This network needs to provide an open circuit at the RF, i.e., it needs to appear invisible to the RF frequency. The bias network must however provide a short at the IF, i.e., it should reflect all of the power at the IF frequency with a phase shift of 180° . The bias network must ensure that the IF power generated from the mixer is returned to the mixer and that none reaches the input port.

The physical structure of the bias network is illustrated in Figure 4.10. The bias circuit consists of two $\lambda/4$ transmission lines at the RF frequency, an open-circuited stub

called the IF return, a DC bias pad and a decoupling capacitor.

Figure 4.10 Physical layout of bias network and IF return

The two $\lambda/4$ transmission lines convert the open-ended transmission line into an open at the design frequency. The open for the RF frequency does not mean that the IF signal will see a short. An open at RF and a short at IF are difficult to achieve simultaneously since both frequencies are close in value. Thus another open-circuited stub called the IF return was provided apart from the two $\lambda/4$ lines. The open-circuited stub corresponds to approximately $\lambda/2$ at the RF frequency and hence appears like an open and does not affect the RF signal. The open-circuited stub represents a finite impedance value for the IF signal. This impedance in parallel with the impedance of the bias network results in a reflection coefficient which is a closer approximation to a short at the IF frequency. The reflection coefficient at port 2 with and without the IF return are respectively shown in Figure 4.11 and Figure 4.12. There is a significant improvement in

the impedance presented to the IF signal. The capacitor provides a path to ground for the frequencies generated by the mixer other than the RF or IF frequencies.

Frequency 1.0 to 6.0 GHz

M1 Frequency=3.3 mag=0.9572 ang=160.4737

M2 Frequency=5.3 mag=0.0093 ang=109.4082

Figure 4.11 Simulated S_{22} of bias network with IF return

Frequency 1.0 to 6.0 GHz
 M1 Frequency=3.0 mag=0.2275 ang=102.2462
 M2 Frequency=5.3 mag=0.0320 ang=91.5903

Figure 4.12 Simulated S_{22} of bias network without IF return

Figure 4.13 and Figure 4.14 respectively illustrates the magnitude of the reflection and transmission coefficient at port 1 of the bias network including the IF return. The bandwidth of the RF signal at 5.3 GHz is approximately 249 MHz (4.7 %). This bandwidth is the frequency range over which the input reflection coefficient about 5.3 GHz is -20 dB or less. The graphs show that most of the RF energy is passed and that most of the IF energy is reflected.

Figure 4.13 Magnitude of the S_{11} [dB] for the bias network

Figure 4.14 Magnitude of the S_{21} [dB] for the bias network

4.2.4 Input and Output Matching Networks

The input matching network was designed to provide maximum power transfer of the RF signal under small-signal conditions from the input $50\ \Omega$ system to the mixer circuit. The output matching network was designed to provide maximum power transfer of the IF signal from the mixer circuit to the output $50\ \Omega$ system once again under small-signal conditions. The input matching network is designed first and is present in addition to the other components when designing the output matching network. The output match is based on the small-signal impedance seen looking into the MSM-PD from port 2 as shown in Figure 2.23. The impedance seen is not only from the MSM-PD, but from the bias network, IF return and input matching network. Once the output matching network is designed, it is placed into the circuit. The output matching network does not affect the input circuit since the transmission coefficient S_{21} of the MSM-PD is very small, in the order of 0.28 at the RF frequency of 5.3 GHz. Hence the input reflection coefficient looking into the MSM-PD is independent of the applied load as shown in equation (30). Similarly, the input matching network will not load the output circuit. It was thus possible to independently design the input and output matching networks.

The matching networks were initially designed using a single-stub matching technique. The resulting matching network yielded a narrow bandwidth response. Double-stub matching was attempted to increase the bandwidth. Single-stub matching was used in the final circuit since the insignificant bandwidth increase of the double-stub matching technique did not warrant the added stub. The input matching network used an open-circuited stub as opposed to a short-circuited stub since an open circuit can be

modeled more accurately than a short which requires a via model that accounts for the uncertainties of the via location and structure. A balanced single stub was used to provide design symmetry. A short-circuited stub was used for the output matching network because it provides a DC path to ground for the applied bias across the MSM-PD.

4.2.5 MSM-PD

The 10x2 MSM-PD array was placed in mount structure #2 because the structure has fewer parasitic elements than mounting structure #1. The input and output contact pads of the MSM-PD were each connected to a 50 Ω transmission line by two gold bond wires. A 5 μm finger spacing MSM-PD was chosen from the array as the mixer element. The MSM-PD used in the optoelectronic mixer circuit was not the same photodetector used for DC and S-parameter measurements.

4.2.6 Final Mixer Circuit

The completed optoelectronic mixer consisting of the matching networks, bias circuit, IF return and filter is illustrated in Figure 4.15. All of the copper lines of the circuit were gold plated to enable the MSM-PD to be wire bonded to the transmission lines. The via holes were plated through so that contact could be made with the ground plane located on the back side of the RT/Duroid substrate. The final dimensions of the microstrip lines used in the circuit are given in the HP-EEsof/Libra V6.0 schematic files found in Appendix I.

Figure 4.15 Photograph of the optoelectronic mixer

The optoelectronic mixer was simulated under small-signal conditions using the HP-EEsof/Libra V6.0 circuit simulator. The OEM was biased at 0.1 volt and illuminated with 1.36 mW of optical power. The simulated input and output return losses are respectively given in Figure 4.16 and Figure 4.17.

Figure 4.16. Simulated input return loss of the OEM

Figure 4.17 Simulated output return loss of the OEM

4.3 Optoelectronic Mixer Experimental Results

4.3.1 Experimental Set-Up

This section describes the set-up and experimental results of the optoelectronic mixer based on the small-signal (S-parameter) design. The set-up used for the optoelectronic mixer experiment is illustrated in Figure 4.18. The OEM has three input ports and one output port. The MSM-PD within the module is biased using an external DC power supply. The bias level across the MSM-PD, V_{bias} , and its generated photocurrent, I_{ph} , are monitored. The electrical RF signal is applied to port 1 of the OEM. The RF signal has a frequency f_{RF} , and a power level of P_{RF} .

Figure 4.18 Optoelectronic mixer experimental set-up

The MSM-PD is illuminated using the same technique used in the device characterization measurements of Chapter 3 with the difference that the optical signal is

intensity modulated with a microwave signal of frequency f_{LO} and power P_{in} , instead of being held constant. The power P_{in} refers to the electrical power applied to the laser module. This microwave signal is superimposed onto a DC signal using a bias-T network. The laser diode controller provided the laser diode with a laser drive current I_{laser} needed to generate the desired unmodulated (DC) optical power P_{opt} . Finally, the modulated optical power incident on the MSM-PD is denoted by P_{opt_rms} .

The laser module consists of a $47\ \Omega$ resistor in series with a SHARP LT015MD laser diode as shown in Figure 4.19. The output optical signal from the laser diode is focused into a multimode fiber using a lens system. The laser diode has three pins. Pin 1 provides the drive current needed to bias the laser at the desired operating point as well as the modulated signal. Pin 2 provides the ground. Pin 3 provides a way of regulating the output optical power by monitoring the current generated by the photodetector (included in the laser diode package) due to laser back reflection.

Figure 4.19 Laser module and bias-T

The $47\ \Omega$ resistor was chosen for matching purposes assuming the laser diode had an approximate impedance of $3\ \Omega$ at the DC operating point. The $47\ \Omega$ resistor together with $3\ \Omega$ of the laser diode presents a matched $50\ \Omega$ termination to the microwave source used to modulate the laser diode. Most of the incident microwave power is dissipated in the $47\ \Omega$ resistor which protects the source by minimizing back reflections. It was later found that the laser diode had a complex impedance of $5.68 + j62.87\ \Omega$ at the LO frequency of 2.275 GHz. This impedance was obtained from measurements made available on a laser diode of the same part number as the one used in the OEM measurements. The calculated powers in the system listed in Table 4 2 were based on the impedance value for the measured laser diode.

The measured laser diode impedance differed from the initial impedance estimate. This discrepancy led to an impedance mismatch between the source and the laser diode. Hence not all of the available power P_{in} was delivered to the laser diode. The power P_{LO} delivered to the laser diode, as well as the current I_{laser} through the laser diode were calculated using the HP-EEsof/Libra V6.0 simulation package. It should be noted that the power available at the laser diode is electrical, and the power available at the surface of the MSM-PD is optical. The responsivity in W/A of the laser diode was measured and then used to calculate the optical power incident on the MSM-PD. The measured laser responsivity of 0.41 W/A was 50% less than the manufacturer's specification. Specification data for the laser diode is included in Appendix II. The discrepancy in the responsivity values was due to the electrical-to-optical conversion, coupling losses of the lens, and coupling losses of the fiber.

Table 4.2 Power and Current Values

P_{in} (dBm)	P_{LO} (dBm)	$I_{laser(rms)}$ (mA)	P_{opt_rms} (mW)	P_{opt_rms} (dBm)
+10	-1	11.8	4.79	6.8
+5	-6	6.6	2.7	4.3
+3	-8	5.3	2.18	3.4
0	-11	3.7	1.51	1.8
-5	-16	2.05	0.84	-0.76
-10	-21	1.2	0.49	-3.1
-15	-26	0.64	0.26	-5.9
-20	-31	0.35	0.15	-8.2
-25	-36	0.21	0.087	-10.6

The table 4.2 shows that there is an 11dB reduction in the electrical power from the modulated source (P_{in}) to that incident on the laser diode (P_{LO}). Therefore very little microwave power was available to modulate the laser diode. Most of the microwave power incident on the laser module was dissipated in the $47\ \Omega$ series resistance.

The signals generated from port 2 of the OEM module are observed on the spectrum analyzer. The power levels of the generated signals were taken directly from the display screen of the spectrum analyzer. Ideally, a power meter should be used for accurate power measurements. The power meter available at the time of the experiment was not sensitive enough to measure the relatively weak powers of less than -40 dBm. Thus the spectrum analyzer was chosen as the power measuring tool. An average difference between both instruments was obtained by comparing their power readings for different power levels and frequencies. The spectrum analyzer readings were approximately 0.5 dB lower than the power meter readings. The power readings of the spectrum analyzer also fluctuated ± 2 dB about the average signal power value.

4.3.2 Input Frequency Match

The input match frequency was determined prior to performing optoelectronic mixing. At this frequency there is maximum power transfer into port 1 of the OEM module. The device was designed to have an input match at 5.3 GHz. Figure 4.20 illustrates the set-up used to determine the input match frequency.

Figure 4.20 Input frequency match measurement set-up

The electrical RF frequency is fed into port 1 of a directional coupler. The signal is attenuated by 0.74 dB as it passes through the coupler. The reflected power at the OEM travels back through the coupler, and an approximate 10% is coupled into port 2 of the directional coupler and is displayed on the spectrum analyzer. A 50 Ω load is placed at port 2 of the OEM module to emulate the spectrum analyzer that would be present in the OEM experiment.

The reflected power from the OEM module varies depending on the operating conditions, i.e., bias, optical power, RF power and LO power. The input frequency match measurement was performed under the design conditions required for OEM operation. Consequently, the laser output DC power was adjusted to the design value of 1.36 mW

and the optical signal intensity modulated with a 2.0 GHz signal having a -1 dBm power level. The bias applied across the MSM-PD was adjusted to +0.1 volt. The frequency of the input RF signal into the OEM module was swept, and the reflected power was observed on the spectrum analyzer. A low RF power level of -20 dBm was chosen to insure that the OEM operated in the optically-pumped mode. The input match which occurs when the reflected RF power from port 1 falls to a minimum value was found to be located at approximately 5.4 GHz. At the design bias level of +0.1 volt, the reflected power decreased to yield a calculated return loss of 17.5 dB at a frequency of 5.429 GHz. A description of the return loss calculation will be given later in the chapter.

Maintaining the design bias level, the optical power was swept until a minimum reflected power was obtained. An DC optical power of 2.38 mW was found to yield the optimum match. The match was unaffected by DC optical powers greater than 2.38 mW. Under this condition, the frequency of optimum match was found to be 5.418 GHz with a calculated return loss of 31.3 dB.

The bias levels were also changed to observe their effect on the match. The match became worse as the bias increased beyond near-zero bias. The response remained constant after +5 volt.

4.3.3 Output Frequency Match

The output match was determined using the same set-up as for the input match measurement except that port 2 of the OEM module was connected to the coupler and port 1 was terminated in 50 Ω . Since a different range of frequencies was to be swept

for the determination of the output match, a different directional coupler was used having an insertion loss of 0.56 dB and a coupling factor of -10.1 dB. Repeating the same measurement procedure as for the input frequency match with a laser DC optical power of 1.36 mW and a bias level of +0.1 volt, the output frequency match was found to occur at 3.154 GHz with a calculated return loss of 21.5 dB.

Maintaining the design bias of +0.1 volt, a DC optical power of 1.80 mW was found to yield the optimum output match. The reflected power reached a minimum, and DC optical powers greater than 1.80 mW were found to not affect the match. Under this condition, the output match was found to be 3.167 GHz with a return loss of 35.0 dB.

Therefore under the operating condition of a +0.1 volt bias and 1.36 mW of unmodulated optical power, the maximum power transfer to and from the OEM occurred at an input RF signal frequency of 5.429 GHz and at an output IF signal frequency of 3.154 GHz. The LO signal will have a 2.275 GHz frequency which was calculated from the difference between the RF and IF frequencies. The design frequencies for the RF, LO and IF were 5.3 GHz, 2.0 GHz, and 3.3 GHz respectively. Comparing the design and measured frequencies, the percentage difference relative to the design frequencies were calculated to be 2.4 %, 13.75%, and 4.4 % for the RF, LO, and IF respectively. The OEM measurements were performed using the design bias and optical power levels to check the validity of the design procedure. It should be noted however that DC optical power levels different from the design level yielded improved matching.

4.3.4 Output Power Measurements

Optoelectronic mixing was performed for a range of RF and LO powers. The RF power values were swept from -25 dBm to +10 dBm, and the LO power values were swept from -36 dBm to -1 dBm. As previously stated, the LO power refers to the electrical power available at the laser diode. The optical power P_{opt} available at the surface of the MSM-PD varied from -10.6 dBm to +6.8 dBm. A greater power level was not possible due to the power dissipated in the 47 Ω resistor, and the signal generator used in the optoelectronic mixing experiment was not equipped to provide more than +10 dBm of output power. Despite the low modulated-optical power available at the surface of the MSM-PD, optically-pumped mixing was still observed with -20 dBm of electrical RF power and -1 dBm of LO power. The RF power was measured at the plane of reference PR1, and the IF power was measured at the plane of reference PR2 as defined in Figure 4.18.

The measured output power levels P_{out} of the RF, 2RF and RF-LO (or IF) signals are given in Table 4.3. The values contained within the shaded cells represent the characteristics of the signals applied to the OEM. Empty entries in tables indicate that the power levels in the 40 MHz frequency span of the spectrum analyzer were too weak to be measured. The bias level across the MSM-PD and the DC optical power illuminating the device were kept constant. The operating conditions for the OEM under which the table values were recorded are as follows:

$$\begin{aligned} V_{bias} &= 0.1 \text{ V} \quad (I_{msm} = 300 \text{ } \mu\text{A}) \\ P_{opt} &= 1.36 \text{ mW} \quad T = 20.2 \text{ } ^\circ\text{C} \end{aligned}$$

Table 4.3 Optoelectronic Mixing Results using the OEM

Prf (dBm)	Pout (dBm)	Plo (dBm)								
		-36	-31	-26	-21	-16	-11	-8	-6	-1
-25	RF = 5.429 GHz	-73	-73	-74	-73	-74	-72	-72	-72	-73
	2RF = 10.858 GHz									
	LO = 2.275 GHz									-73
	RF-LO = 3.154 GHz									
	RF+LO = 7.704 GHz								-72	-70
-20	RF	-70	-69	-69	-69	-69	-69	-69	-69	-69
	2RF									
	LO									-72
	RF-LO						-71	-69	-68	-63
	RF+LO									
-15	RF	-66	-66	-66	-66	-66	-66	-66	-66	-66
	2RF					-70	-70	-69	-69	-70
	LO									
	RF-LO					-72	-69	-66	-63	-58
	RF+LO									
-10	RF	-61	-61	-61	-61	-61	-61	-61	-61	-61
	2RF	-63	-63	-63	-63	-63	-63	-63	-63	-63
	LO									
	RF-LO				-73	-68	-63	-61	-60	-55
	RF+LO									
-5	RF	-57	-57	-57	-57	-57	-57	-57	-57	-57
	2RF	-60	-60	-60	-60	-60	-60	-60	-60	-60
	LO									
	RF-LO				-69	-64	-60	-57	-54	-50
	RF+LO								-69	-65
0	RF	-52	-52	-52	-52	-52	-52	-52	-52	-52
	2RF	-52	-52	-52	-52	-52	-52	-52	-52	-52
	LO									
	RF-LO		-71	-67	-62	-58	-53	-50	-47	-42
	RF+LO							-65	-63	-59
3	RF	-49	-49	-49	-49	-49	-49	-49	-49	-49
	2RF	-54	-54	-54	-54	-54	-54	-54	-54	-54
	LO									
	RF-LO		-69	-64	-60	-56	-50	-48	-45	-41
	RF+LO								-62	-58
5	RF	-47	-47	-47	-47	-47	-47	-47	-47	-47
	2RF	-55	-55	-55	-55	-55	-55	-55	-55	-54
	LO									
	RF-LO	-72	-67	-63	-58	-53	-48	-45	-43	-38
	RF+LO						-65	-63	-61	-56
10	RF	-42	-42	-42	-42	-42	-42	-42	-42	-42
	2RF	-53	-54	-54	-54	-54	-54	-54	-54	-54
	LO									-66
	RF-LO	-70	-65	-61	-56	-51	-46	-43	-41	-36
	RF+LO						-65	-62	-62	-56

Figure 4.21 illustrates the power of the IF signal versus RF power for different LO powers. Figure 4.22 illustrates the power of the IF frequency signal vs LO power for different RF powers. Both graphs depict an increase in the IF power as the RF or LO power increases for a respective constant LO or RF power. Figure 4.21 shows a deviation in the linear trend when the RF power reaches approximately +5 dBm. This power value is known as the 1-dB compression point indicating that a change in RF power beyond +5 dBm will not yield a proportional increase in the generated IF output power. Not all of the input RF energy will be converted into the desired output power, but some will be dissipated as heat and unwanted IM products. Figure 4.22 shows no significant deviation from the linear trend since the optical power range is never very large due to the difficulty in achieving a high modulation index of the optical signal. This difficulty arises because a small amount of power is available to modulate the laser diode.

Figure 4.21 IF power versus RF power for different LO powers

Figure 4.22 IF power versus LO power for different RF powers

It was also observed that as the RF power was swept over the given power range for high RF power values (> -5 dBm), the measured DC photocurrent decreased compared to its value when no RF power was applied. The importance of this effect was mentioned in Chapter 3 with reference to the large-signal modeling of an electrically-pumped OEM.

An example of the output spectrum observed on the spectrum analyzer is illustrated in Figure 4.23. The spectrum is centered at the IF frequency of 3.154 GHz and covers a 40 MHz frequency span.

Figure 4.23 Output spectrum of the OEM

4.3.5 Repeatability Measurements

Optoelectronic mixing was performed again on another day to ensure repeatability of the measurements. The set-up was disassembled and reassembled so as to ensure equipment handling did not significantly affect the results. The bias was set at +0.1 volt and the DC optical power illuminating the MSM-PD was set to 1.4 mW. The results from both experiments are displayed in Figure 4.24. Taking an average for both sets of power measurements, an approximate ± 1 dB power difference can be expected.

Figure 4.24 Repeatability measurements

4.3.6 Comparing Results of the OEM with and without Matching Networks

Optoelectronic mixing was also performed without matching networks. In other words, the MSM-PD alone was bonded to mounting structure #2 in place of the OEM module (with matching networks). This measurement was to determine whether the OEM with matching networks (small-signal design) yielded an improvement over the OEM without matching networks. The set-up used of this mixer experiment is shown in Figure 4.25.

Figure 4.25 Optoelectronic mixer experiment set-up using the unmatched OEM

The optoelectronic mixer using the MSM-PD required the matched OEM to be replaced with the MSM-PD, and the inclusion of external bias-Ts to provide the necessary potential difference across the MSM-PD. This experiment includes no matching networks, hence the output signals are directly fed into the $50\ \Omega$ system of the spectrum analyzer. There is also no lowpass filter present and consequently signals with frequencies that would have been otherwise eliminated or attenuated using the OEM module appear at the output. Hereafter the MSM-PD alone will be referred to as the unmatched OEM. Table 4.4 gives the output signal powers generated by the mixer using the unmatched OEM. The RF input power values are measured at plane of reference PR1, and the output

power values are measured at plane of reference PR2 as shown in Figure 4.25. Therefore the power into the MSM-PD must take into account the insertion loss (IL) of the bias-T #1 and the output power must take into account the insertion loss of the bias-T #2. The power levels given in Table 4.4 includes the losses associated with the bias-Ts. The same power terminology applies as in Table 4.3.

Table 4.4 Optoelectronic Mixing Results using the OEM without Matching Networks

Prf (dBm)	Pout (dBm)	Plo (dBm)				
		-21	-16	-11	-6	-1
-11.5	RF = 5.429 GHz	-22.3	-22.3	-22.3	-22.3	-22.3
	2RF = 10.858 GHz					
	LO = 2.275 GHz					-67.3
	RF-LO = 3.154 GHz				-68.3	-65.3
	RF+LO = 7.704 GHz					
-6.5	RF	-17.3	-17.3	-17.3	-17.3	-16.3
	2RF					-71.3
	LO					-68.3
	RF-LO		-71.3	-67.3	-64.3	-60.3
	RF+LO				-64.3	-60.3
-1.5	RF	-12.3	-12.3	-12.3	-12.3	-12.3
	2RF	-62.3	-62.3	-62.3	-62.3	-61.3
	LO				-70.3	-66.3
	RF-LO			-64.3	-60.3	-56.3
	RF+LO			-64.3	-60.3	-56.3
3.5	RF	-7.3	-7.3	-7.3	-7.3	-7.3
	2RF	-59.3	-59.3	-59.3	-59.3	-59.3
	LO					-63.3
	RF-LO	-69.3	-64.3	-59.3	-55.3	-50.3
	RF+LO	-69.3	-63.3	-59.3	-55.3	-50.3
8.5	RF	-2.3	-2.3	-2.3	-2.3	-2.3
	2RF	-55.3	-55.3	-55.3	-55.3	-55.3
	LO				-68.3	-62.3
	RF-LO	-64.3	-59.3	-54.3	-51.3	-45.3
	RF+LO	-64.3	-59.3	-54.3	-50.3	-45.3

Figure 4.26 compares the IF power as a function of the RF power for different LO powers of the optoelectronic mixer experiment obtained using the matched and unmatched OEMs.

Figure 4.26 IF power of OEM with and without matching networks

There is an approximate 10 dB improvement in IF output power using the matched OEM module as opposed to the unmatched OEM. Table 4.5 shows the difference in powers between the measurements for the OEM with and without matching networks.

Table 4.5 Power Difference between the OEM with and without Matching Networks

Prf (dBm)	Pout (dBm)	Plo (dBm)				
		-21	-16	-11	-6	-1
-10	RF = 5.429 GHz	-39.7	-39.7	-39.7	-39.7	-39.7
	2RF = 10.858 GHz					
	LO = 2.275 GHz					
	RF-LO = 3.154 GHz				9.3	10.3
	RF+LO = 7.704 GHz					
-5	RF	-39.7	-39.7	-39.7	-39.7	-40.7
	2RF					
	LO					
	RF-LO		7.3	8.3	10.3	11.3
	RF+LO					
0	RF	-39.7	-39.7	-39.7	-39.7	-39.7
	2RF	8.3	8.3	8.3	8.3	7.3
	LO					
	RF-LO			11.3	11.3	12.3
	RF+LO				-3.7	-3.7
5	RF	-39.7	-39.7	-39.7	-39.7	-39.7
	2RF	9.3	9.3	9.3	9.3	9.3
	LO					0.3
	RF-LO	11.3	11.3	11.3	12.3	12.3
	RF+LO			-4.7	-5.7	-5.7
10	RF	-39.7	-39.7	-39.7	-39.7	-39.7
	2RF	2.3	2.3	2.3	2.3	2.3
	LO					
	RF-LO	9.3	8.3	9.3	10.3	9.3
	RF+LO			-9.7	-9.7	-11.7

A positive difference indicates that there is more power available at the specified frequency when optoelectronic mixing is performed using the matched OEM, and a negative difference indicates the opposite. The matched OEM shows that there is an approximate 40 dB decrease in the RF power signifying that the lowpass filter provides significant RF power attenuation at the mixer output in comparison with the unfiltered (unmatched) case. The negative RF+LO power difference indicates that in the matched

OEM, the RF+LO signal is also being attenuated by the filter. The smaller negative values of the RF+LO power difference in comparison with those for the RF indicates that the filter rejection is not as effective for the RF+LO signal as for the RF signal. The reason for this is that the elliptical filter exhibits a low pass response until approximately 5.3 GHz beyond which it passes higher frequencies with progressively less attenuation, as was shown in Figure 4.8. For this same reason, there is an increase in the 2RF power level. Finally the 10 dB increase in the RF-IF power means that the matched OEM is more efficient than the unmatched OEM.

4.4 Figures of Merit of the Optoelectronic Mixer

The figures of merit of the OEM experiment with and without matching networks are given in this section. The figures to be discussed are: return loss, conversion loss, 1-dB compression point, 3rd-order intercept point, noise figure and interport isolation.

4.4.1 Return Loss

The measurement set-up shown in Figure 4.27 was used to determine the return loss (RL) of the OEM with matching networks.

Figure 4.27 Measurement set-up for the return loss of the OEM

Equation (36) is used to calculate the return loss. P_{in} and P_{inc} respectively represent the input power into the directional coupler and into the OEM. P_{ref} refers to the power reflected from the OEM, P_{sa} is the power delivered to the spectrum analyzer and P_2 is the power returned to the input. IL and CL describe the directional coupler and respectively denote the insertion loss and the coupling factor.

$$RL = P_{sa}|_{dB} - CF|_{dB} - (P_{in}|_{dB} - IL) \quad (36)$$

The minimum input and output return loss values calculated under the specified operating conditions are given in Table 4.6.

Table 4.6 Minimum Input and Output Return Loss Parameters of the OEM with Matching Networks

Parameters	Port 1	Port 2
IL (dB)	0.74	0.56
CF (dB)	-9.5	-10.1
P_{in} (dBm)	-20	-18
P_{sa} (dBm)	-47.7	-50.2
RL (dB)	-17.5	-21.5

OEM operating conditions for the measurements of Table 4.6

$$V_{bias} = 0.1V \quad (I_{bias} = 300 \mu A)$$

$$P_{opt} = 136 mW \quad T = 20.2^\circ C$$

$$F_{rf} = 5.429 GHz \quad F_{lo} = 2.275 GHz \quad F_{if} = 3.154 GHz$$

Figure 4.28 and Figure 4.29 respectively illustrate the comparison between the simulated and measured input and output return loss of the OEM (small-signal design)

over a range of RF input frequencies. Considering the small-signal design approach employed, the simulations compare well with the measured data.

Figure 4.28 Comparison of simulated and measured input return loss

Figure 4.29 Comparison of simulated and measured output return loss

The same set-up as Figure 4.27 was used to measure the input return loss of the OEM without matching networks. The MSM-PD was biased through a bias-T which was inserted between the coupler and the MSM-PD. Therefore the same return loss equation (36) is valid with the inclusion of twice the insertion loss of the bias-Ts. The operating conditions are the same as in the matched case. However, the input return loss minimum was not calculated, and the input return loss was only calculated for an RF input signal having a 5.429 GHz frequency. Measurement and RL values are given in Table 4.7.

Table 4.7 Input Return Loss Parameters of the OEM without Matching Networks

Parameters	Port 1
IL (dB)	0.74
CF (dB)	-9.5
P_{in} (dBm)	-15
P_{sat} (dBm)	-33
RL (dB)	-4.8

4.4.2 Conversion Loss

The conversion loss (CL) for the OEM with and without matching networks are respectively illustrated in Figures 4.30 - 4.31. A minimum of 44 ± 1 dB conversion loss is obtained when the matched OEM is fed with a local oscillator power of -1 dBm. Both figures show that the matched OEM yields better conversion loss compared to the unmatched OEM.

Figure 4.30 Conversion loss of the OEM with matching networks

Figure 4.31 Conversion loss of the OEM without matching networks

The conversion loss difference between the matched and unmatched OEM is given in Table 4.8. The positive values in the table indicate the amount by which the conversion loss is decreased through utilizing a matched OEM.

Table 4.8 Conversion Loss Difference between Matched and Unmatched OEM

P_{rf} (dBm)	$P_{lo} =$ -21 (dBm)	-16 (dBm)	-11 (dBm)	-6 (dBm)	-1 (dBm)
-10				6.8	8.8
-5		5.8	5.8	8.8	8.8
0		9.8	9.8	11.8	12.8
5	9.8	9.8	9.8	10.8	10.8
10	6.8	6.8	6.8	8.8	7.8

4.4.3 1-dB Compression Point

The 1-dB compression point is the RF input power at which the RF input versus IF output relationship deviates from linearity by 1 dB. Figure 4.32 illustrates the 1-dB compression point to be +6.7 dBm for an LO power of -1 dBm. Therefore +6.7 dBm is the maximum input power level, and beyond this point, the RF energy is dissipated into heat and unwanted intermodulation products.

The 1-dB compression point was calculated for other LO powers and was found to be approximately the same. This parameter is a function of the LO and should increase at higher LO power. A possible reason for not being able to observe this could be due to the small LO power available in the first place. The 1-dB compression point of the unmatched OEM is not given because the IF power was not found to deviate from its linear trend as a function of RF power.

Figure 4.32 Calculation of 1 dB compression point for matched OEM

4.4.4 Interport Isolation

The interport isolation for the OEM with and without matching networks is shown in Table 4.9. In both cases, the interport isolation is presented for the optically and electrically pumped modes of the mixer operation. The interport isolation is given for a certain LO and RF power. The isolation at other power levels can be calculated using the power values given in Table 4.3 and Table 4.4. The RF-to-LO interport isolation is given as infinite since no electrical power is coupled into the optical port. The OEM yields better interport isolation in the matched case. The improvement in the RF-to-IF isolation is achieved by the lowpass filter which attenuates the RF signal. The LO-to-RF interport isolation is not available due to the inaccessibility of measuring the LO power at the RF port during the interport measurements.

Table 4.9 Interport Isolation for an Optoelectronic Mixer

Interport Isolation	Optically-Pumped		Electrically-Pumped	
	Matched OEM	Unmatched OEM	Matched OEM	Unmatched OEM
P_{lo}	-1 dB	-1 dB	-21 dB	-21 dB
P_{rf}	-20 dB	-11.5 dB	+10 dB	+8.5 dB
LO-to-IF	71 dB	66.3 dB	> 59	> 56
RF-to-LO	∞	∞	∞	∞
RF-to-IF	49 dB	10.8 dB	52	10.8

4.4.5 3rd Order Intercept Point

The RF frequencies, $f_{r1} = 5.429 \text{ GHz}$ and $f_{r2} = 5.529 \text{ GHz}$, were fed into a power combiner whose output was used as the RF input to the matched OEM. The two-tone, third-order IM (intermodulation) products were not observable for LO powers below -1 dBm. The 3rd-order intercept point measurement involved measuring the power at one of the IM products as a function of the RF input power. A 1 dB change in the RF power should yield a 3 dB change in the IM power. Accurate IM power measurements were not possible due to signal fluctuations.

4.4.6 Signal-to-Noise Ratio (S/N)

The S/N ratios of the matched OEM were calculated from equation (37) using the IF signal power and noise power readings taken from the spectrum analyzer. The noise power was measured at an offset of 20 MHz from the IF frequency. Different resolution bandwidths were used to measure the noise power. The measurements were performed for an RF power of 0 dBm and an LO power of -1 dBm. The relevant data is given in Table 4.10.

$$S / N = P_{IF} \Big|_{dBm} - P_{noise} \Big|_{dBm} \quad [dB] \quad (37)$$

Table 4.10 Signal-to-Noise Measurements

Resolution BW	IF power	Noise power	S/N
	(dBm)	(dBm)	(dB)
1 Hz	-42	-131	91
1 KHz	-41	-91	50
100 KHz	-40	-83	43

4.5 Conclusion

The time and cost involved in the design of an OEM based on small-signal S-parameters provides justifiable performance improvement over the unmatched OEM. Further performance gains would be expected through the implementation of an accurate large-signal model for the MSM-PD.

Chapter 5: Potential Areas for Further Research

5.1 Brief Review of Thesis Objectives

The aim of the thesis was to design an optically-pumped mixer using a Metal-Semiconductor-Metal Photodetector (MSM-PD) to downconvert a 5.3 GHz RF signal into a 3.3 GHz IF signal using a 2.0 GHz intensity modulated optical signal as the local oscillator. Two approaches were implemented to achieve this objective. The first investigated the possibility of using a nonlinear model of the MSM-PD to facilitate the design of an optoelectronic mixer. The second approach was to design the optoelectronic mixer based on small-signal (S-parameter) measurements of the MSM-PD.

In the first design approach, a nonlinear model of the MSM-PD was developed to simulate the device behavior over a selected range of biases and optical powers. Future work would aim to analyze the developed nonlinear model under large signal conditions and subsequently compare the simulated results with the experimental results. The discrepancy between the experimental and simulation results would determine whether further refinement of the model would be necessary prior to using the model in a mixer design. This chapter will provide suggestions in the area of device modeling and device characterization felt necessary to improve the model's representation of the MSM-PD.

The second (small-signal) design approach showed that the MSM-PD can be used as an optoelectronic mixer in the aforementioned microwave range.

5.2 Improvements in the MSM-PD Measurement Techniques

Modeling of the MSM-PD requires accurate data measurement. An improvement in the device measurement would be to replace the test fixture in which the MSM-PD is mounted by an on-wafer measuring technique. On-wafer probing would eliminate all parasitic effects up to the plane of the device, and the resultant measurements would consequently be of the device itself. This technique would eliminate the need for bond wires which might dominate the behavior of the MSM-PD. The technique would also eliminate parasitic effects not modeled at the interface between the transmission lines and the bond wires.

On-chip measurements would thus eliminate parasitic effects associated with the device interconnections to the external circuit. The model then need only represent the components of the device. Elimination of the interconnect parasitics would lead to fewer unknown components in the model, hence improving the accuracy to which the device could be modeled.

5.3 Accounting for MSM-PD Contact Pads

The MSM-PD is equipped with metal pads to provide contact between the device and the outside world. Ideally the effects of the pads should also be accurately modeled to obtain a true representation of the device performance. The present model crudely represents the contact pads by capacitors to ground. An improvement of the contact representation would be to measure the pad structure in isolation and add the resultant S-parameter file to the model. An alternative approach would be to model the pads using an

electromagnetic simulator and then again incorporate the generated S-parameter file into the model.

5.4 Physical Modeling of the Device

An empirical model is currently used to model the MSM-PD. The present model is restricted to a particular physical dimension and composition. Since the device structure is predefined, an accurate empirically based model is sufficient for use in circuit design of the optoelectronic mixer.

The coefficients of the nonlinear equation representing the current source were obtained from curve fitting the equation to the measured data. Thus the coefficients have no physical meaning. Consequently, it is impossible to know how each coefficient relates to the physical parameters of the device. Hence the effect on the mixer performance due to changes in the physical properties of the device cannot be explored using the present model. Ideally a model based purely on the physical structure of the device would allow mixer performance to be examined for changes in the device structure. Such a model would allow optimization of the physical structure of the MSM-PD for optimum mixer performance. The model could simulate MSM-PDs with different finger spacing, widths, active areas, and doping profiles.

5.5 Accounting for the Effects of Deep Traps in the MSM-PD

A trap is caused by an impurity atom or a lattice defect in the semiconductor and introduces an allowed electronic level between the conduction and valence bands. A trap

can temporarily capture a carrier and release it after a period of time. Note that even though trapping will be hereafter discussed in terms of electrons, holes follow a similar explanation. Trapping levels located deep in the band gap are slower in releasing the electrons than those located closer to the conduction band. Consequently electrons held in deep traps need more energy to be released. Since more energy is required to excite an electron from a deep trap, the probability of such an event is less than for lower energy transitions. The reduced probability of electron excitation from the deep traps means that there is a delay in the release of the electrons from the traps. Thus the rate at which the electrons are captured and released depends on the energy level at which the traps are located. Trap energy level can vary substantially even for devices of different batches [46]. A study of GaAs photoconductors [62] has shown that deep traps account for carrier transits that can persist up to approximately 2000s.

Traps will affect the optical responsivity of the device. Responsivity is the ratio of the measured photogenerated current to the illuminating optical power. If the optical signal is modulated with a frequency whose period is long compared with the trapping time, the traps will have negligible effect on the responsivity. As the frequency at which the optical signal is modulated increases, the trapping and release times become long compared to the rate at which the optical signal is modulated, consequently the responsivity of the device decreases as statistically fewer carriers are available to constitute the photocurrent.

Work done by Gouin et al. [46] has shown that the responsivity depends on the modulation frequency applied to the optical signal. Results from Finite Element Modeling

of the MSM-PD structures indicate that the responsivity of the MSM-PD structures are very sensitive to the density of deep traps within the semiconductor [61].

Optoelectronic mixing requires an intensity modulated optical signal. The responsivity dependence on the optical signal modulation frequency should be incorporated into the model of the MSM-PD since different responsivities lead to different power levels of the mixing products. The responsivity dependence is not needed for characterization of the device since the device characterization measurements were performed under constant optical conditions, i.e., zero modulation frequency. The present model represents the responsivity of the MSM-PD as a function of bias. Any future model should include the dependency of the responsivity on the optical signal modulation frequency. The responsivity dependence could be incorporated into the present model by making the current source a function of the modulation frequency.

5.6 MSM-PD Parameter Dependence on Bias and Optical Power

In the present model, the only parameter that depends on both bias and optical power is the current source. Further investigation should be done to establish whether other components exhibit a dependency on bias and optical power. The present model has the junction capacitance of the MSM-PD as a constant, however this parameter is a potential candidate for the bias and optical power dependence. Gouin et al. [46] have reported experiments that show that the junction capacitance is only affected by the constant optical power once the bias across the MSM-PD falls below a volt. This observation was substantiated by Veilleux [60] who also reports a slight dependence of

the junction capacitance on the optical power. In the development of the present model, the junction capacitance dependence on optical power and bias was not felt significant enough to warrant incorporation into the model. However, due to the discrepancy between simulated and measured data, it is now felt that the dependency of the junction capacitance with applied optical power should be fully investigated and incorporated if necessary into a second iteration of the model.

5.7 Improvement in the Optical Power Measuring Technique

Accurate and repeatable measurement of the optical power incident on the MSM-PD is necessary for characterization of the device parameters which depend on the optical signal level.

The measurement technique currently employed to monitor the power output of the laser diode involves adjustment of the drive current to the diode to obtain the required optical power at the fiber tip. The optical power was monitored using an optical power meter and once the required power was achieved, the fiber was relocated over the MSM-PD. This procedure necessitates that the fiber be moved between the power meter input and the location over the MSM-PD each time a new power level is applied. Clearly this procedure results in a repeatability problem in terms of accurate repositioning of the optical fiber over the MSM-PD. To circumvent this problem, an optical power splitter should be used to allow constant monitoring of the optical signal level being applied to the device. The splitter would divide the optical signal from a laser source into two fiber optical arms. The optical power from one arm would represent a known percentage of the

applied laser power. Thus monitoring of the power in one arm would allow the optical power being applied to the device through the second arm to be determined on a continuous basis without the need to change the fiber position. Hence the repeatability of the measurement would be greatly improved. The use of a splitter however requires a higher power laser to compensate for the power lost in the monitoring arm. Since initially only low power laser sources were available, this splitter technique was not adopted. However, future work would aim to use a higher power laser to facilitate the aforementioned splitter technique for optical power monitoring.

5.8 Technique to Increase the Modulation Index of the Laser

Improved conversion loss in the optoelectronic mixer can be achieved by increasing the modulation index of the laser. At present the laser supplies a very low power modulated signal due to the $47\ \Omega$ resistance placed in series with the laser diode. Most of the RF power incident on the termination is dissipated in the $47\ \Omega$ series resistance, and consequently very little RF power is available to modulate the laser diode.

The signal generator used in the optoelectronic mixing experiment was not equipped to provide more than +10 dBm of output power. The power available to modulate the laser diode was calculated to be -1 dBm, the matching network accounting for the 11 dB of loss. An increase in the mixer's conversion efficiency would be seen if an increase in the modulation depth of the optical power could be achieved. A possible solution to increase the modulation index would be to introduce a lossless matching network at the desired frequency between the signal generator and the laser diode. A

lossless matching network could be realized in microstrip technology employing single or multiple stubs depending on the bandwidth requirements.

5.9 Mixer Improvement Through Changes in the Device Doping Profile

Optically-pumped mixing requires a change in the device's small-signal conductance (hereafter referred to conductance) as a function of the optical power. Near the zero bias region of the DC I/V characteristics of the MSM-PD, the conductance changes as the optical power changes. Further research should be done to understand how the rate of change in conductance depends on the optical power. The objective of such research would be to increase the rate of change in device conductance with applied optical power. Electrically-pumped mixing requires a change in the conductance as a function of voltage. A more nonlinear change of the conductance as a function of voltage results in stronger mixing components.

Therefore, the density and profile of dopants should be investigated to obtain a nonlinear rate of change in conductance as a function of optical power and applied bias.

5.10 Conclusion

An optoelectronic mixer using an MSM-PD was designed, fabricated and tested based on small-signal approach. The optoelectronic mixer developed in this thesis differs from an OEM using an MSM-PD found in the literature [30] in that it is optically-pumped, operates in the L-C frequency band and is fabricated on Duroid substrate.

A first iteration of a large-signal model of the MSM-PD was also developed. The model differs from the small-signal models of the MSM-PD found in the literature since the MSM-PD is typically used as a photodetector and not as a mixer.

Optoelectronic mixing using a photosensitive device is a novel approach to frequency conversion, especially in the case of an optical receiver where signal distribution is achieved via optical fibers. Further investigation in the use of different devices as optoelectronic mixers or ways of improving the performance of existing devices should be studied.

The minimum conversion loss obtained for the designed optoelectronic mixer was 44 dB. The OEM yields very good interport isolation characteristics, and will thus find application where isolation between the LO and RF signal is of critical importance. The MSM-PD, being simple to fabricate and compatible with the MESFET process, warrants further research as an optoelectronic mixer.

Appendix I:

HP-EEsof / Libra Schematic Files for the Optoelectronic Mixer

Optoelectronic Mixer Simulation Diagram

Appendix II:
SHARP LT015MD Specification Data

LT015MD/MF

Features

- High power (maximum optical power output 40 mW)
- Wavelength 830nm
- Single transverse mode

Applications

- Optical disk memories
- Medical apparatus
- Optical floppy disks
- Optical memory cards
- Information processing equipment

Absolute Maximum Ratings

(T_c = 25°C)

Parameter	Symbol	Ratings	Units
Optical power output	P _o	40	mW
Reverse voltage	Laser	2	V
	PIN	30	
Operating temperature *1	T _{opr}	-10 ~ +50	°C
Storage temperature *1	T _{stg}	-40 ~ +85	°C
Soldering temperature *2	T _{sol}	260 (less than 5 seconds)	°C

*1 Case temperature *2 At point 1.6 mm from lead base

Electro-optical Characteristics *1

(T_c = 25°C)

Parameter		Symbol	Condition	Ratings			Units
				MIN	TYP	MAX	
Threshold current		I _{th}	—	—	60	80	mA
Operating current		I _{op}	P _o = 30mW	—	95	130	mA
Operating voltage		V _{op}	P _o = 30mW	—	1.75	2.2	V
Wavelength *2		λ _p	P _o = 30mW	815	830	845	nm
Monitor current		I _m	P _o = 30mW V _a = 15V	75	250	750	μA
Radiation characteristics	Angle	Parallel to junction	P _o = 30mW	8	9.5	14	deg
		Perpendicular to junction	P _o = 30mW	20	27	38	deg
	Ripple		P _o = 30mW	—	—	± 20	%
Emission point accuracy	Angle	Δφ ₁	P _o = 30mW	—	—	± 2	deg
	Position *4	Δx, Δy, Δz	P _o = 30mW	—	—	± 3	deg
Differential efficiency		η	20mW I ₁ (30mW) - I ₁ (10mW)	0.5	0.8	1.1	mW/mA

*1 Initial value

*2 Single transverse mode

*3 Angle at 50% peak intensity (full width at half maximum)

*4 Not specified for LT015MF

Electrical Characteristics of Photodiode

(T_c = 25°C)

Parameter	Symbol	Condition	Ratings			Units
			MIN	TYP	MAX	
Sensitivity	S	V _{in} = 15V	—	8.3	—	μA/mW
Dark current	I _d	V _a = 15V	—	—	150	nA
Terminal capacitance	C _t	V _a = 15V	—	8	20	pF

LT015 Series Characteristics Diagrams

Fig. 90-1 Forward Current vs. Forward Voltage

Fig. 90-2 Optical Power Output vs. Forward Current and Monitor Current

Fig. 90-3 Threshold Current vs. Temperature

Fig. 90-4 Wavelength vs. Temperature

Fig. 90-5 Optical Power Output Dependence of Wavelength

These data are typical only and are not intended as a specification. The physical characteristics of the device may vary from device to device.

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