**AN EFFICIENT KA-BAND MMIC UPCONVERTER USING A
HETEROJUNCTION BIPOLAR TRANSISTOR**

by

Eric R. Guétre

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

Ottawa-Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Electronics
Carleton University
January 1997

©Copyright E. Guétre, 1997

National Library
of Canada

Acquisitions and
Bibliographic Services Branch

395 Wellington Street
Ottawa, Ontario
K1A 0N4

Bibliothèque nationale
du Canada

Direction des acquisitions et
des services bibliographiques

395, rue Wellington
Ottawa (Ontario)
K1A 0N4

Your file: Votre référence

Our file: Notre référence

The author has granted an irrevocable non-exclusive licence allowing the National Library of Canada to reproduce, loan, distribute or sell copies of his/her thesis by any means and in any form or format, making this thesis available to interested persons.

L'auteur a accordé une licence irrévocable et non exclusive permettant à la Bibliothèque nationale du Canada de reproduire, prêter, distribuer ou vendre des copies de sa thèse de quelque manière et sous quelque forme que ce soit pour mettre des exemplaires de cette thèse à la disposition des personnes intéressées.

The author retains ownership of the copyright in his/her thesis. Neither the thesis nor substantial extracts from it may be printed or otherwise reproduced without his/her permission.

L'auteur conserve la propriété du droit d'auteur qui protège sa thèse. Ni la thèse ni des extraits substantiels de celle-ci ne doivent être imprimés ou autrement reproduits sans son autorisation.

ISBN 0-612-19516-3

Name AN EFFICIENT KATANA FOR CONSTRUCTING A PROPOSITIONAL NETWORK
 Dissertation Abstracts International is arranged by broad, general subject categories. Please select the one subject which most nearly describes the content of your dissertation. Enter the corresponding four-digit code in the spaces provided.

ENGINEERING ELECTRONICS

0579 U-M-I

SUBJECT TERM

SUBJECT CODE

Subject Categories

THE HUMANITIES AND SOCIAL SCIENCES

COMMUNICATIONS AND THE ARTS

Architecture 0229
 Art History 0377
 Cinema 0700
 Dance 0378
 Fine Arts 0357
 Information Science 0223
 Journalism 0391
 Library Science 0399
 Mass Communications 0708
 Music 0413
 Speech Communication 0459
 Theater 0445

EDUCATION

General 0515
 Administration 0514
 Adult and Continuing 0516
 Agricultural 0517
 Art 0273
 Bilingual and Multicultural 0282
 Business 0688
 Community College 0275
 Curriculum and Instruction 0727
 Early Childhood 0518
 Elementary 0524
 Finance 0277
 Guidance and Counseling 0519
 Health 0680
 Higher 0745
 History of 0520
 Home Economics 0278
 Industrial 0521
 Language and Literature 0279
 Mathematics 0280
 Music 0522
 Philosophy of 0998
 Physical 0523

Psychology 0525
 Reading 0535
 Religious 0527
 Sciences 0714
 Secondary 0533
 Social Science 0534
 Sociology of 0340
 Special 0529
 Teacher Training 0530
 Technology 0710
 Tests and Measurements 0288
 Vocational 0747

LANGUAGE, LITERATURE AND LINGUISTICS

Language 0679
 General 0679
 Ancient 0289
 Linguistics 0290
 Modern 0291
 Literature 0401
 Classical 0294
 Comparative 0295
 Medieval 0297
 Modern 0298
 African 0316
 American 0591
 Asian 0325
 Canadian (English) 0352
 Canadian (French) 0355
 English 0593
 Germanic 0311
 Latin American 0312
 Middle Eastern 0315
 Romance 0313
 Slavic and East European 0314

PHILOSOPHY, RELIGION AND THEOLOGY

Philosophy 0422
 Religion 0318
 General 0321
 Biblical Studies 0319
 Clergy 0320
 History of 0322
 Philosophy of 0469
 Theology 0469

SOCIAL SCIENCES

American Studies 0323
 Anthropology 0324
 Archaeology 0326
 Cultural 0327
 Physical 0327
 Business Administration 0310
 General 0272
 Accounting 0770
 Banking 0454
 Management 0338
 Marketing 0385
 Canadian Studies 0501
 Economics 0503
 General 0505
 Agricultural 0508
 Commerce-Business 0509
 Finance 0510
 History 0511
 Labor 0358
 Theory 0366
 Folklore 0351
 Geography 0351
 Gerontology 0578
 History 0578
 General 0578

Ancient 0579
 Medieval 0581
 Modern 0582
 Black 0328
 African 0331
 Asia, Australia and Oceania 0332
 Canadian 0334
 European 0335
 Latin American 0336
 Middle Eastern 0337
 United States 0585
 History of Science 0398
 Law 0615
 Political Science 0616
 General 0617
 International Law and Relations 0814
 Public Administration 0452
 Recreation 0626
 Social Work 0938
 Sociology 0631
 General 0628
 Criminology and Penology 0629
 Demography 0630
 Ethnic and Racial Studies 0628
 Individual and Family Studies 0629
 Industrial and Labor Relations 0630
 Public and Social Welfare 0700
 Social Structure and Development 0344
 Theory and Methods 0709
 Transportation 0999
 Urban and Regional Planning 0453
 Women's Studies 0453

THE SCIENCES AND ENGINEERING

BIOLOGICAL SCIENCES

Agriculture 0473
 General 0285
 Agronomy 0475
 Animal Culture and Nutrition 0476
 Animal Pathology 0359
 Food Science and Technology 0478
 Forestry and Wildlife 0479
 Plant Culture 0480
 Plant Pathology 0817
 Plant Physiology 0777
 Range Management 0746
 Wood Technology 0306
 Biology 0287
 Anatomy 0308
 Botany 0309
 Cell 0379
 Ecology 0329
 Entomology 0353
 Genetics 0369
 Limnology 0793
 Microbiology 0410
 Molecular 0307
 Parasitology 0317
 Oceanography 0416
 Physiology 0433
 Radiation 0821
 Veterinary Science 0778
 Zoology 0472
 Biophysics 0786
 General 0760
 Medical 0760

EARTH SCIENCES

Biogeochemistry 0425
 Geochemistry 0996

Geodesy 0370
 Geology 0372
 Geography 0373
 Hydrology 0388
 Mineralogy 0411
 Paleobotany 0345
 Paleogeography 0426
 Paleontology 0418
 Paleozoology 0985
 Paleobotany 0427
 Physical Geography 0368
 Physical Oceanography 0415

HEALTH AND ENVIRONMENTAL SCIENCES

Environmental Sciences 0768
 Health Sciences 0566
 General 0300
 Audiology 0992
 Chemotherapy 0567
 Dentistry 0550
 Education 0769
 Hospital Management 0738
 Human Development 0982
 Immunology 0564
 Medicine and Surgery 0347
 Mental Health 0569
 Nursing 0570
 Nutrition 0380
 Occupational Health and Safety 0354
 Therapy 0381
 Ophthalmology 0419
 Pathology 0571
 Pharmacology 0572
 Physical Therapy 0573
 Public Health 0574
 Radiology 0575
 Recreation 0575

Speech Pathology 0460
 Toxicology 0383
 Home Economics 0386

PHYSICAL SCIENCES

Pure Sciences 0485
 Chemistry 0749
 General 0486
 Agricultural 0487
 Analytical 0488
 Biochemistry 0738
 Inorganic 0490
 Nuclear 0491
 Organic 0494
 Pharmaceutical 0495
 Physical 0754
 Polymer 0405
 Mathematics 0605
 Physics 0796
 General 0606
 Acoustics 0608
 Astronomy and Astrophysics 0748
 Atmospheric Science 0607
 Atomic 0798
 Electronics and Electricity 0759
 Elementary Particles and High Energy 0609
 Fluid and Plasma 0610
 Molecular 0752
 Nuclear 0756
 Optics 0611
 Radiation 0463
 Solid State 0346
 Statistics 0984
 Applied Sciences 0346
 Applied Mechanics 0984
 Computer Science 0984

Engineering 0537
 General 0538
 Aerospace 0539
 Agricultural 0540
 Automotive 0541
 Biomedical 0542
 Chemical 0543
 Civil 0544
 Electronics and Electrical 0348
 Heat and Thermodynamics 0545
 Hydraulic 0546
 Industrial 0547
 Marine 0794
 Materials Science 0548
 Mechanical 0743
 Metallurgy 0551
 Mining 0552
 Nuclear 0549
 Packaging 0765
 Petroleum 0554
 Sanitary and Municipal 0790
 System Science 0428
 Geotechnology 0796
 Operations Research 0795
 Plastics Technology 0924
 Textile Technology 0924

Psychology 0621
 General 0384
 Behavioral 0622
 Clinical 0620
 Developmental 0623
 Experimental 0624
 Industrial 0625
 Personality 0989
 Physiological 0349
 Psychobiology 0632
 Social 0451

The undersigned hereby recommend to
the Faculty of Graduate Studies and Research
acceptance of the thesis

"A KA-BAND UPCONVERTER USING A HETEROJUNCTION BIPOLAR
TRANSISTOR"

submitted by

Eric R. Guétre, B.Eng., P.Eng.

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

Dr. J.S. Wight,
Thesis Co-Supervisor

Dr. M.G. Stubbs,
Thesis Co-Supervisor

Chair,
Department of Electronics

Carleton University

January 1997

ABSTRACT

New broadband systems such as WLAN, LMCS, and broadband SATCOM are currently being proposed or implemented. Frequency crowding at the lower end of the microwave spectrum and the large bandwidth requirements of these new systems means that wireless transmission frequencies at Ka band and above are being considered. As an alternative to the GaAs MESFET, HEMT, and PHEMT, the GaAs HBT is being explored for these wireless applications. There is little information, however, on the performance of the GaAs HBT for use as a mixer at microwave and millimeter-wave frequencies.

The objective of this thesis is to investigate the potential use of the HBT as an upconverter and to design, build, and test a MMIC upconverter destined for an indoor Ka band WLAN application. The performance of the single-ended HBT upconverter is compared to a single-ended PHEMT upconverter with similar specifications. Test results of the fabricated MMIC upconverter show that the GaAs HBT is well-suited for use in an efficient Ka band upconverter design.

ACKNOWLEDGEMENTS

I would like to thank my supervisor, Dr. Jim Wight, for his support throughout my Master's program. I am grateful to him for giving me the chance to carry out my research at CRC and for exposing me to the microwave research community.

I would like to thank my co-supervisor, Dr. Malcolm Stubbs, for his invaluable guidance, assistance, and patience. I would also like to extend my appreciation to all the staff at CRC, especially those from Mr. René Douville's directorate, who helped me throughout my stay at CRC and patiently answered all my questions. Special thanks is extended to Mr. Kees Verver of CRC, who thoroughly reviewed a large portion of my thesis and who never hesitated to help me out.

I would also like to thank Mr. Herman Willemsen of Nortel for allowing Carleton University graduate students access to the Nortel GaAs HBT fabrication runs, and Mr. Alan Harrison of Nortel for providing us support with respect to the Nortel GaAs MMIC fabrication process.

Finally, I would like to acknowledge the support that I received from Ms. Lori Campbell, who also reviewed my thesis. Her encouragement and help were very valuable to me.

CONTENTS

ABSTRACT	iii
ACKNOWLEDGEMENTS	iv
LIST OF TABLES	viii
LIST OF FIGURES	ix
LIST OF ABBREVIATIONS AND ACRONYMS	xii
LIST OF SYMBOLS	xiii
CHAPTER 1: INTRODUCTION	1
1.1 Overview	1
1.2 Objective	3
1.3 Thesis motivation	3
1.4 Thesis organization	5
1.5 Thesis contribution	5
CHAPTER 2: BACKGROUND	7
2.1 Introduction	7
2.2 Mixers	7
2.2.1 Fundamentals of mixers	8
2.2.2 Performance parameters	12
2.2.3 Mixer types and configurations	16
2.2.4 Active microwave mixers	20
2.3 The GaAs heterojunction bipolar transistor (HBT)	24
2.3.1 Operation and characteristics of the GaAs HBT	24
2.3.2 HBT large-signal model	31
2.4 Non-linear analysis	35
2.5 MMIC technology	37
2.5.1 GaAs MMIC fabrication issues	38
2.5.2 HBT performance and fabrication	39
CHAPTER 3: ANALYSIS OF THE HBT AS AN UPCONVERTER	42
3.1 Introduction	42
3.2 Model confidence	42
3.3 Investigation of the HBT as an upconverter	43
3.3.1 Bias point	44

3.3.1.1	The HBT's non-linear regions	44
3.3.1.2	Upconverter configuration	48
3.3.1.3	Load-line analysis.....	50
3.3.1.4	Chosen bias point	51
3.3.2	Estimation of the conversion gain	52
CHAPTER 4: DESIGN, LAYOUT AND MEASUREMENTS.....		57
4.1	Introduction	57
4.2	Design specifications and constraints.....	57
4.3	Upconverter design.....	60
4.3.1	LO matching network.....	63
4.3.2	Conversion gain estimate	64
4.3.3	Collector bias network	66
4.3.4	IF matching network	67
4.3.5	Base bias network.....	71
4.3.6	RF matching network	72
4.3.7	MMIC layout.....	74
4.3.8	EM simulations	75
4.3.9	Some simulation results	77
4.4	Measured and simulated results of the MMIC upconverter	78
4.4.1	Return loss.....	82
4.4.2	Conversion loss	85
4.4.3	Output spectrum	87
4.4.4	The 1 dB compression point and the IP3	88
4.4.5	Conversion gain versus LO power	91
4.4.6	LO-IF isolation	93
4.5	Differences between the measured and simulated results IF and RF return losses.....	94
4.6	Summary of the results and comparison with a PHEMT single-ended upconverter.....	98
CHAPTER 5: CONCLUSION.....		101
5.1	Summary	101
5.2	Future work	103
5.3	Conclusion.....	104
REFERENCES.....		105
APPENDIX A: ANALYSIS OF A SINGLE-BALANCED MIXER.....		110
APPENDIX B: LOAD-LINE ANALYSIS TEST BENCH AND CIRCUITS.....		114
APPENDIX C: UPCONVERTER CIRCUIT AND TEST BENCHES.....		117

APPENDIX D: SIMULATION TEST CIRCUIT FOR OBTAINING THE S-PARAMETERS OF SECTION 4.3.2.....	121
APPENDIX E: DIAGRAMS OF THE MEASUREMENT SET-UPS.....	123

LIST OF TABLES

Table 2-1. Summary of FET-based and HBT microwave mixers.....	23
Table 2-2. Device merit parameter comparison	30
Table 2-3. Advantages and disadvantages of MMICs and MHMICs	38
Table 2-4. Key electrical parameters of the HBT.....	40
Table 2-5. Maximum ratings of the HBT.....	40
Table 4-1. Mixer specifications and constraints.....	58
Table 4-2. Some simulation results.....	78
Table 4-3. Test equipment used	79
Table 4-4. Vce/Vbe/PLO combinations used in the design and measurements.....	82
Table 4-5. LO return loss	84
Table 4-6. Simulated and measured output power levels of certain spurious signals.....	88
Table 4-7. Comparison between the HBT upconverter and CRC's PHEMT upconverter.....	99

LIST OF FIGURES

Figure 1-1. Separately integrated wireless transceiver.....	2
Figure 2-1. Types of non-linearity.....	10
Figure 2-2. Effect of bias point and LO power on non-linearity	11
Figure 2-3. One dB compression point and 3rd order intercept point (IP3).....	15
Figure 2-4. Single-ended active mixer	17
Figure 2-5. A passive single-balanced mixer with 180 deg hybrid couplers	19
Figure 2-6. Steady-state velocity-field characteristics for electrons in GaAs and Si	26
Figure 2-7. Current components for an HBT operating in the active region.....	27
Figure 2-8. Energy band diagram for the HBT of Figure 2-7.....	27
Figure 2-9. Nortel GaAs HBT fabrication layers	31
Figure 2-10. Standard BJT electrical model (based on the Gummel-Poon model).....	32
Figure 2-11. DC I-V curves.....	33
Figure 2-12. Electrical model of the HBT	34
Figure 2-13. Harmonic balance	36
Figure 3-1. Comparison between HBT large-signal simulation results and measurements.....	43
Figure 3-2. Non-linear regions of the HBT	45
Figure 3-3. Second (local) derivatives	46
Figure 3-4. Top view of a GaAs MESFET showing the source divided into two ports	49
Figure 3-5. Emitter-injected HBT single-ended upconverter.....	50
Figure 3-6. Load-line analysis.....	51
Figure 3-7. Upconverter with input and output matching networks	52
Figure 4-1. Upconverter block diagram	62

Figure 4-2. LO matching network.....	63
Figure 4-3. Magnitude of S21 when the HBT is biased in the active and cut-off regions	65
Figure 4-4. Collector bias network.....	67
Figure 4-5. IF matching network.....	70
Figure 4-6. Effect of the IF matching network.....	71
Figure 4-7. Simulated IF matching from 3.35 GHz to 3.85 GHz.....	71
Figure 4-8. Base bias network.....	72
Figure 4-9. Effect of the RF matching network on the unmatched RF impedance.....	73
Figure 4-10. RF matching network	74
Figure 4-11. Upconverter MMIC layout	75
Figure 4-12. EM and linear simulations of the IF side's LO trap.....	77
Figure 4-13. Simulated and measured IF return loss.....	82
Figure 4-14. Measured and simulated RF return loss	83
Figure 4-15. Measured RF output spectrum.....	85
Figure 4-16. Simulated (c0, c1, and c2) and measured (c1 and c2) conversion gain	86
Figure 4-17. Output spectrum for the c1 Vce/Vbe/PLO combination	88
Figure 4-18. Simulated conversion gain Gc (at c1 and c2) against IF input power	89
Figure 4-19. Measured IP3 (c1 Vce/Vbe/PLO combination).....	90
Figure 4-20. Measured IP3 (c2 Vce/Vbe/PLO combination).....	90
Figure 4-21. Measured conversion gain versus LO power.....	92
Figure 4-22. Simulated conversion gain, Gc, versus LO power.....	92
Figure 4-23. LO power level at the IF port.....	93
Figure 4-24. Results of the EM simulation against the linear simulation for the LO trap on the IF side.....	96

Figure 4-25. Simulated RF return loss using the linear and EM response of the LO trap on the IF side.....	97
Figure A-1. Single-balanced upconverter.....	111
Figure B-1. Load-line test bench.....	115
Figure B-2. Load-line test circuit	116
Figure C-1. Test circuit of the MMIC upconverter.....	118
Figure C-2. Test bench for all simulations except the RF return loss	119
Figure C-3. Test bench for the RF return loss only	120
Figure D-1. Test circuit for section 4.3.2	122
Figure E-1. Test set-up for the IF and RF return losses	124
Figure E-2. Test set-up for the LO return loss.....	125
Figure E-3. Test set-up for the output spectrum and the conversion loss measurements.....	126
Figure E-4. Test set-up for the two-tone third order intermodulation measurement.....	127
Figure E-5. Test set-up for the output spectrum at the IF port	128

LIST OF ABBREVIATIONS AND ACRONYMS

AC	alternating current
BJT	bipolar junction transistor
BPF	band pass filter
CAD	computer-aided design
CITR	Canadian Institute for Telecommunications Research
CRC	Communications Research Centre
dBm	milliwatt expressed in decibel where one milliwatt equals zero decibel
DC	direct current
DRC	design rule check
EM	electromagnetic
FET	field effect transistor
GHz	gigahertz
GP	Gummel-Poon
GPS	Global Positioning System
HBT	heterojunction bipolar transistor
HEMT	high electron mobility transistor
HPA	high power amplifier
I-V	current-voltage
IF	intermediate frequency
InP	indium phosphate
IP3	third order intercept point
IVHS	intelligent vehicle highway system
LNA	low noise amplifier
LO	local oscillator
LSB	lower side band
MESFET	metal semiconductor field effect transistor
MHMIC	miniature hybrid microwave integrated circuit
MHz	megahertz
MIC	microwave integrated circuit
MIM	metal-insulation-metal
MMIC	monolithic microwave integrated circuit
MN	matching network
MOCVD	metalorganic chemical vapor deposition
mS	millisiemens
NLE	non-linear element
PCM	process control monitor
PDA	personal digital assistant
PHEMT	pseudo-morphic high electron mobility transistor
RF	radio frequency
USB	upper-side band
VSWR	voltage standing wave ratio
WLAN	wireless local area network

LIST OF SYMBOLS

β	DC current gain for a bipolar transistor
γ	emitter efficiency
Γ	reflection coefficient
$\beta_{\max}$	maximum attainable DC current gain
α_T	forward emitter-collector current transfer coefficient
A	coefficient of the Fourier expansion of a square wave
c0	$V_{ce}=2$ V; $V_{be}=1.1$ V; $P_{LO}=-3$ dBm
c1	$V_{ce}=2$ V; $V_{be}=1.1$ V; $P_{LO}=-2$ dBm
c1	$V_{ce}=1.25$ V; $V_{be}=1$ V; $P_{LO}=-2$ dBm
C	capacitance
D_{nb}	diffusion constant for the electrons in the base
E_g or ΔE_g	energy height difference (or gap) between qV_n and qV_p
F	flux
$f_{\max}$	maximum frequency of oscillation
f_T	cut-off frequency
G_c	conversion gain
g_m	transconductance
GN_b	Gummel number for the base
I	current
k	Boltzmann's constant = 1.380662×10^{-23} J/K
L	inductance
M	impedance mismatch factor
n_{ib}	intrinsic electron concentration in the base layer
N_n	emitter doping density
N_p	base doping density
P	power
q,Q	charge in coulomb
qV_c	collector bias energy (relative to the base energy)
qV_e	emitter bias energy (relative to the base energy)
qV_n	energy height for holes at the emitter side of the emitter-base junction
qV_p	energy height for electrons at the base side of the emitter-base junction
R	resistance
S_{ij}	S parameter
T	temperature in Kelvin
v, V	voltage
v_{nb}	mean speed of electrons at the emitter end of the base
v_{pe}	mean speed of holes at the base end of the electrons
Y	admittance
Z	impedance

CHAPTER 1: INTRODUCTION

1.1 Overview

Wireless communication began long ago with Marconi's 1894 radiotelegraph experiment. Now, wireless communication seems poised to become ubiquitous and affordable. The promise of wireless communication is instantaneous, portable and affordable voice, video and data communications from anywhere in the world to anyone in the world. Increased bandwidth requirements of these future communication systems, coupled with frequency band crowding at the lower end of the microwave spectrum, means that the Ku, K, Ka and millimeter-wave bands¹ are being considered as frequencies of operation. An example of a wireless system being implemented is satellite personal communications, spearheaded by Motorola's Iridium program. Other wireless systems being proposed or already implemented are the Local Multipoint Communication System (LMCS), which is to provide broadband multimedia services to residential areas, and the wireless local area network (WLAN), which promises to eliminate costly wiring associated with computer networks. Monolithic microwave integrated circuits (MMICs) are promising for these microwave and millimeter-wave communications systems, as the reduced costs and high-volume capability of transmitter-receiver MMICs makes use of new frequency bands economical, and, as such, the development of advanced MMIC mixers is crucial to this [Haigh89].

To try to satisfy the evolving requirements of microwave and millimeter-wave band communication, new microwave circuit technologies are explored and assessed for viability when existing technologies are inadequate. Because of a comparatively low maximum frequency of oscillation, the silicon bipolar transistor is not deemed suitable

¹ The Ku band covers from 12.4 to 18 GHz; the K band covers 18 to 26.5 GHz; the Ka band covers 26.5 to 40 GHz. The millimeter-wave band is from 30 GHz to 300 GHz.

for high microwave frequency operation. Instead, the gallium arsenide (GaAs) metal semiconductor field effect transistor (MESFET) and the GaAs high electron mobility transistor (HEMT) have been, to date, the transistors of choice for microwave integrated circuits (MICs) and MMICs. The GaAs heterojunction bipolar transistor (HBT), however, is one of the alternative technologies being explored. It has attracted considerable interest in the past few years for applications such as power amplifiers and low phase noise oscillators. Its popularity at microwave and millimeter-wave frequencies stems from a high cutoff frequency, large current gain, high current handling capability and use of epitaxial growth for critical dimensions. There is little information, however, on the performance of the GaAs HBT as a mixer at microwave frequencies.

Used in the above-mentioned applications (power amplifier, oscillator, and mixer), HBTs may allow the integration of a large portion of the transmit side of a wireless transceiver onto a single MMIC. Figure 1-1 shows a simplified representation of a wireless transceiver, where the receive path is integrated on a single MMIC using HEMTs and the transmit path is separately integrated using HBTs.

Figure 1-1. Separately integrated wireless transceiver.

1.2 Objective

The objective of this thesis is to investigate the potential use of the HBT as an upconverter and to design, build and test a single-ended HBT MMIC upconverter destined for an indoor Ka band WLAN application.

1.3 Thesis motivation

This thesis was originally associated with a research effort devoted to the development of a prototype Ka band indoor WLAN transceiver. A brief overview of WLANs² is given here to provide some background as to the origin of this research, and to describe an application where HBT mixers might play a role in the future.

The obvious difference between a wireline LAN and a wireless LAN is that in a wireless LAN application data would be transferred between users via microwaves, millimeter-waves, or infrared. Furthermore, a WLAN can accommodate mobile users, within its operational limits and range, and is potentially easier to install and take apart because it requires less cable. The hope is that WLANs may replace wireline LANs when wiring cost are prohibitive and when using wires as the coupling medium is impractical. For instance, a WLAN becomes an affordable and desirable alternative when mobile workers must have quick access to up-to-date information and must operate within a large perimeter. Examples of such environments are office floors, hospitals, manufacturing plants, and schools. Another type of WLAN application is when the activity site is temporary and the LAN must be set up expediently. Examples of temporary WLANs environments are sporting events, political campaigns, seasonal events, etc. Typically, a WLAN would be composed of a number of client applications (laptops, computers, PDAs, manufacturing robots, facsimile machines, etc.) either coupled to each other, or to

² The underlying project for this thesis is a Ka band WLAN research project involving many universities, industry and the CRC, and is sponsored by the Canadian Institute for Telecommunications Research (CITR). More detailed information on the CITR WLAN project can be found in [Falco96] and [Wight96].

one or many fixed base stations. The base station(s), in turn, might be connected via wireline to a server or to the outside world. Depending on the application, communication between clients and base stations could be two-way or one-way, and each base station could serve clients within a given geographical area ranging from a few meters to hundreds of meters.

In any of the above-mentioned applications, the high data rates required of the latest data and video communication equipment translate into a need for large bandwidths. Given the frequency crowding at the lower end of the microwave spectrum and these large bandwidth requirements, there has been considerable interest in operating wireless systems in the Ku band frequency range and beyond. One such commercially available indoor WLAN system, marketed for an office environment, operates at 15 Mbits/sec for a frequency of 18 GHz. This particular WLAN covers an unobstructed range of 130 feet, with a radiated power of 25 mW [Gard95]. Typically, for an indoor environment, a cell would be delimited by the walls, ceiling and floor of a building's story. Note that communications cells, with diameters of no more than a few hundred metres, are often termed pico-cells. The fact that frequencies in the high microwave or millimeter-wave range experience high atmospheric loss has made them unattractive for long-range wireless applications. It is exactly this property, however, that makes them attractive for indoor or short-range applications. High attenuation and reflection of millimeter-wave signals incident on concrete walls also helps in delimiting a cell within a building's walls. Talbi and Delisle [Talbi96] report on obstruction losses, at 37.2 GHz, of 43.0 dB for a 75.1-cm wide concrete wall. A Ka band WLAN system installed on one story of a concrete building would be likely to have its signals confined within the walls, ceiling and floor, thus forming a well-isolated pico-cell. This would allow frequency re-use from pico-cell to pico-cell.

1.4 Thesis organization

This thesis is organized into five chapters. The second chapter presents the background knowledge applicable to the thesis. Specifically, it concentrates on four areas: mixers, the GaAs HBT and its large-signal model, non-linear circuit analysis, and MMIC technology. The third chapter discusses the use of the HBT as a mixer and includes a derivation of a formula that estimates the maximum conversion gain of a single-ended upconverter. The fourth chapter outlines the design, simulation, monolithic layout, and measured results of the upconverter. It also compares the HBT measured results to a CRC-built PHEMT upconverter with similar specifications. Finally, the fifth chapter summarizes the thesis and makes several conclusions based on the outcome of the research as well as recommendations for worthwhile future research.

1.5 Thesis contribution

This thesis investigates the potential use of the HBT as a mixer. To carry out this investigation, a single-ended MMIC upconverter is designed, built, and tested. The upconverter specifications originate from the CTR WLAN project. They are intermediate frequency (IF) 3.6 GHz, local oscillator (LO) frequency 26.5 GHz and radio frequency (RF) 30.1 GHz. The input IF power is set at -10 dBm while the LO power level is specified at 5 dBm or less. The HBT upconverter measurements are compared with a PHEMT single-ended upconverter having similar specifications. The thesis contributions can be summarized as follows:

- 1) A design of a single-ended upconverter is presented. The design is done using a Nortel large-signal model already implemented in Libra™ by the CRC. The single-ended upconverter design utilizes all three ports of the HBT because the LO is emitter-injected, thereby eliminating the need for a diplexer.

- 2) A formula that estimates the upper-bound of the conversion gain is derived for the emitter-injected, single-ended upconverter.
- 3) The upconverter gives 3.5 dB of conversion loss in the 29.8 GHz to 30.4 GHz frequency range with only -2 dBm of LO power and as little as 2 mW of DC power. This compares well with a PHEMT single-ended upconverter having the same specifications and yielding 1.5 dB of conversion loss for 5 dBm of LO power and 3 mW of DC power.

CHAPTER 2: BACKGROUND

2.1 Introduction

There are four topics presented in this chapter relating the background information for this thesis. The topics are arranged sequentially so as to answer four basic and logical questions. What are mixers and how do they work? How does the HBT work? What is the best non-linear analysis method for this upconverter circuit design? What is a monolithic microwave integrated circuit (MMIC)?

The first part of chapter 2 is a brief introduction on mixer theory, and covers the "mixing" concept, the performance parameters characterizing mixers, the different mixer configurations, and a review of microwave mixer literature. The next topic is the non-linear device used for this upconverter: the HBT. Its bipolar operation is described and its characteristics are compared to the family of GaAs FET microwave devices. Next, a description of the non-linear HBT model used in this thesis is presented. Then, non-linear circuit analysis is briefly discussed, with harmonic balance being recognized as the best method for large-signal circuit analysis. Finally, the last part of the chapter describes MMICs.

2.2 Mixers

Mixers are found in almost all communication circuits. Their function, in any transceiver, is to convert a signal from one frequency to another, often from the intermediate frequency (IF) to the radio frequency (RF) or vice-versa. The IF is then mixed down to baseband. Except for the self-oscillating type, mixers are three-port circuits because of the need for a local oscillator (LO) signal to drive the mixer into non-

linear operation and to mix with the input signal to produce the desired output. Figure 1-1 shows mixers as three-port circuits.

2.2.1 Fundamentals of mixers

Mixers require one or more non-linear components to perform the mixing operation. An example of a non-linear component is a diode. These non-linear components, under large-signal excitation by the LO and fed by a small-signal modulated input signal (RF or IF), generate signals at other frequencies called harmonics and intermodulation products [Combes87]. The LO signal acts like a pump that drives the mixer circuit into non-linear operation. The output signal (IF or RF) is the desired intermodulation product.

To illustrate how intermodulation products are generated in a mixer, let us assume that a circuit being driven into non-linear operation has a current-voltage response as in equation 2-1 [Maas88].

$$I = aV + bV^2 + cV^3, \quad (2-1)$$

where a , b , and c are real coefficients.

Substituting a two-tone excitation (equation 2-2) into equation 2-1 and applying well-known trigonometric operations, we get the linear, quadratic and cubic terms. These are given by equations 2-3, 2-4 and 2-5 respectively [Maas88].

$$V = v_s(t) = V_1 \cos(\omega_1 t) + V_2 \cos(\omega_2 t) \quad (2-2)$$

$$i_a(t) = av_s(t) = aV_1 \cos(\omega_1 t) + aV_2 \cos(\omega_2 t) \quad (2-3)$$

$$i_b(t) = bv_s^2(t) = \frac{1}{2}b\{V_1^2 + V_2^2 + V_1^2 \cos(2\omega_1 t) + V_2^2 \cos(2\omega_2 t) + 2V_1 V_2 [\cos((\omega_1 + \omega_2)t) + \cos((\omega_1 - \omega_2)t)]\} \quad (2-4)$$

$$\begin{aligned}
i_c(t) = c v_s^3(t) = & \frac{1}{2} c \left\{ \frac{1}{2} V_1^3 \cos(3\omega_1 t) + \frac{1}{2} V_2^3 \cos(3\omega_2 t) \right. \\
& + 3V_1^2 V_2 [\cos((2\omega_1 + \omega_2)t) + \cos((2\omega_1 - \omega_2)t)] \\
& + 3V_1 V_2^2 [\cos((2\omega_2 + \omega_1)t) + \cos((2\omega_2 - \omega_1)t)] \\
& + 3(\frac{1}{2} V_1^3 + 2V_1 V_2^2) \cos(\omega_1 t) \\
& \left. + 3(\frac{1}{2} V_2^3 + 2V_1^2 V_2) \cos(\omega_2 t) \right\}
\end{aligned} \tag{2-5}$$

For an upconverter, the output RF signal is usually the signal with frequency $f_{IF} + f_{LO}$. Unfortunately, non-linear analysis is usually not as straightforward as equation 2-1 because the non-linear relationship is not usually known and may vary with LO power and DC bias.

Maas presents several concepts relevant to non-linear circuits [Maas88]. In general, non-linear circuits are categorized as being weakly non-linear or strongly non-linear. Weakly non-linear circuits are defined as having non-linear characteristics (current-voltage, charge-voltage or flux-current) which are describable, with reasonable accuracy, by a power series expansion of a few terms. In weakly non-linear circuits, it is assumed that the excitation levels are low enough so as not to alter the DC operating point noticeably. A diode or transistor, under small-signal excitation, are examples of weakly non-linear circuits. Conversely, strongly non-linear components and circuits are simply those that cannot be qualified as weakly non-linear.

Two other useful definitions, enunciated by Maas [Maas88] and shown in Figure 2-1, are the two-terminal non-linearity and the transfer non-linearity. A two-terminal non-linearity is simply a non-linear component, such as a non-linear resistor, whose value is a function of one independent variable. On the other hand, for a transfer non-linearity, the control voltage or current is not necessarily at the non-linear element's terminals but can be somewhere else in the circuit. An example is the bipolar junction transistor (BJT), where the collector current is a function of the base-emitter voltage. Analytically, two-

terminal transfer non-linearities are more complicated because the intermodulated and harmonic signals generated at the non-linear component can circulate back to that same non-linear component and produce other intermodulation signals. Clearly, a circuit containing many two-terminal and transfer non-linearities, usually dependent on bias voltage and LO power, will generate a multitude of intermodulation products.

Figure 2-1. Types of non-linearity: (a) two-terminal; (b) transfer.

To show the effect of the bias point and of the LO power level on the non-linear relationship associated with a component, an example involving a transfer non-linearity is presented. The example is illustrated in Figure 2-2a-c. For each case, the solid line represents the non-linear y - x relationship for the domain $x = -4$ to $x = 10$, where $y = ax + bx^2/2 + cx^3/6 - 1$ and where $a = 1$, $b = \frac{1}{2}$, and $c = \frac{1}{6}$. In Figure 2-2a-c, the dotted line represents the location of the bias point and the diamonds represent the LO swing about the bias point due to the excitation tone acting on x .

If we were to use a Taylor polynomial to expand, at the bias point, the y - x relationship described by each case of Figure 2-2a-c, then case 1 and 2 would certainly require only a few terms to be described accurately, while case 3 would undoubtedly require more terms to be described accurately. From equation 2-4, we know that it is the second term that yields the desired IF or RF intermodulation product.

Figure 2-2. Effect of bias point and LO power on non-linearity: (a) weakly non-linear; (b) weakly non-linear; (c) strongly non-linear.

Case 1 (Figure 2-2a) is not desirable because the non-linear component has been biased in a weakly non-linear region and, consequently, a large LO power is necessary to drive into non-linear operation and generate intermodulation products. Case 2 (Figure 2-2b), where the component is biased in a highly non-linear region, but with insufficient LO power, behaves in a weakly non-linear fashion and therefore does not produce strong

intermodulation products. For case 3 (Figure 2-2c), where the component is biased as in case 2 but with more LO power, the circuit is strongly non-linear and detectable intermodulation products will assuredly be generated. For an actual transistor, further increasing the LO power, although transferring more power to the desired RF (or IF) product, also transfers more power to unwanted intermodulation products. What Figure 2-2 illustrates is that there seems to be an optimum solution for DC bias and LO power. The bias point must be chosen so as to reside in a sufficiently non-linear region and the LO power must be strong enough to produce the desired intermodulation product but not so strong as to unduly increase the importance of higher order intermodulation products.

2.2.2 Performance parameters

The following is a list of the principal mixer performance parameters [Combes87], [Maas93], [Maesel92], and [Hender81]. These parameters are normally measured in a 50-ohm system, and are given with respect to a specific LO power level and ambient temperature.

a) *Conversion gain (or loss)*: The conversion gain, G_c , is defined as the ratio between the output power (RF or IF) transferred to the load and the input power (IF or RF) available from the source. The conversion loss, L_c , is the ratio between the available input power to the output power transferred to the load.

$$\begin{aligned} G_c \text{ (dB)} &= 10 \log \left(\frac{P_{\text{out/load}}}{P_{\text{ava/source}}} \right) \\ L_c \text{ (dB)} &= 10 \log \left(\frac{P_{\text{ava/source}}}{P_{\text{out/load}}} \right) \end{aligned} \quad (2-6)$$

b) *Return loss*: The return loss is a measure of the mismatch at each of the ports of a network with respect to a 50-ohm system. It is usually specified in decibels. Acceptable levels of return loss vary depending on the system requirements but most circuits specify

at least -10 dB or lower. Another measure of the mismatch is the voltage standing wave ratio, or VSWR, but for this thesis, the quality of the matching between ports will be specified in terms of return loss only.

$$\text{Return loss} = 20 \log\left(\frac{1}{|\Gamma|}\right), \quad (2-7)$$

where Γ is the reflection coefficient looking towards one port of a circuit.

c) *Compression point*: The 1 dB compression point is defined as the power level of the input signal that causes a 1 dB decrease in the mixer's conversion gain over its small-signal range. Conversion gain saturation occurs when gradually more power from the input signal and the LO tone are transferred to higher order intermodulation products and, gradually, less power goes to the desired intermodulation product. The 1 dB compression point is usually a few decibels lower than the LO power level. Figure 2-3 shows an example of the 1 dB compression point.

d) *Dynamic range*: The dynamic range is the input power range over which the mixer's input and output power have a linear relationship. In other words, it is the range over which the output power of a mixer increases by 1 dB for every 1 dB increase of the input power level. At the low end, the dynamic range is bounded by the noise floor of the mixer and the system, while at the high end, it is limited by the 1 dB compression point.

e) *Spurious signals*: Spurious signals are undesirable mixer-generated intermodulation products. Single tone spurious signals have frequencies of the form $m f_{LO} + n f_{input}$, where $m=0,1,2,\dots$ and $n=0,1,2,\dots$. Multiple tone spurious signals can have more than one input signal mixing with a harmonic of the LO. Out-of-band spurious signals can be reduced by using a filter, while in-band spurious signals can be avoided by a judicious selection of the LO, IF and RF frequencies. An example of poor selection of frequencies would be a downconverter with an IF of 4 GHz and an LO of 8 GHz. In this example, the

third harmonic signal of frequency $3x_{f_{IF}}$ and the spurious signal of frequency $2x_{f_{LO}} - f_{IF}$ coincide with the desired RF signal $f_{LO} + f_{IF}$. Another way to eliminate some spurious signals is to use a balanced mixer configuration.

f) *Suppression*: The suppression at a given port is the difference between the port's intended signal and a spurious signal. Poor suppression can distort the signal at the output of a mixer and introduce unwanted signals at the input of a mixer. For this thesis, RF to LO suppression means $P_{RF} - P_{LO/RF \text{ port}}$. In other words, the RF to LO suppression is the RF power level coming out of the RF port minus the LO power level coming out of the RF port.

g) *Isolation*: The isolation between two ports of a mixer is the attenuation of a signal at a given port with respect to the power level of the signal at its intended port. For instance, an LO-RF isolation of 20 dB means that the LO power level coming out of the RF port is 20 dB down from the LO power level going into the LO port. As a minimum, manufacturers will usually quote the LO-IF and LO-RF isolations. Isolation is the measure of the port-to-port insertion loss and is reserved for the IF, RF and LO signals while suppression is used to quantify the relative strength of spurious signals at a given port.

h) *Third order intercept point*: The 3rd order intercept point is a measure of the intermodulation distortion of two closely spaced input tones, f_1 and f_2 . In a mixer, these two tones can produce a large combination of intermodulation products. The two-tone third order spurious signals of interest have frequencies $f_{LO} + (2f_2 - f_1)$ and $f_{LO} + (2f_1 - f_2)$ and fall within passband.

Figure 2-3. One dB compression point and 3rd order intercept point (IP3).

As shown in Figure 2-3, the output power for a mixer (a single input tone mixing with an LO tone) is expected to increase by 1 dB for every 1 dB increase in the input power level. For the case when the mixer has two input tones, the power level of the two-tone third-order spurious signals rise 3 dB for every 1 dB increase in the input power. The third order intercept point (IP3) is simply the intersection of the extrapolation of these two curves past their respective saturation points.

i) *Power consumption*: Low power consumption is always desirable, and often critical in portable wireless applications, because it allows for longer usage time between battery changes or smaller batteries. Moreover, lower power consumption by the active devices of a circuit often leads to higher reliability rates for the circuit. This is because relatively low power consumption means that devices are less likely to experience failure due to thermal heating.

j) *Noise figure*: The noise figure is defined as the ratio of the input signal-to-noise ratio to the output signal-to-noise ratio, and is an important performance parameter for downconverters. However, it is not a critical performance parameter for upconverters.

This is because, presumably, the noise acquired by the signal during atmospheric propagation, from the transmitting antenna to the receiving antenna, is greater than the noise acquired in the transmitter chain and launched at the transmitting antenna. On the other hand, the noise in an upconverter is relevant because it affects the noise floor of the upconverter, and consequently, the lower bound of a mixer's dynamic range. In this thesis, the noise figure of the HBT upconverter will not be measured.

Besides the noise generated by the random motion of charged carriers in devices and materials, a mixer's noise figure can be considerably exacerbated by the additive noise from the LO source. This additive noise from the LO source is referred to as AM noise and phase noise. Phase noise is always transferred to the output signal of a mixer because the mixer "mixes" the LO tone with the input signal, including their respective phases, thus transferring every radian of the phase noise from the LO source to the output signal [Maas93]. Maas [Maas93] states that there exists no mixer design technique that effectively reduces phase noise and relegates the task of phase noise (and AM noise) reduction to the LO oscillator design. However, reduction of AM noise is possible in mixer design if a balanced mixer configuration is chosen. The reader is referred to the text by Faber [Faber95] for a mathematical treatment of AM and phase noise.

2.2.3 Mixer types and configurations

Mixers can be grouped into two categories: passive and active. Active mixers are defined as requiring DC bias while passive mixers are defined as not requiring DC bias. Active mixers usually use transistors with some sort of DC bias but can also employ diodes, as long as DC bias is applied. Passive mixers usually employ diodes and can also use transistors if no DC bias is applied. Each type has its own advantages and disadvantages. Traditionally, passive mixers can operate well into the millimeter-wave range while active mixers are limited by the transistor's maximum frequency of operation.

For example, a passive mixer using Schottky diodes, with a conversion loss of 8 dB and a noise figure of 6 dB, is used at 94 GHz [Adel90]. In this case, the Schottky diodes' cut-off frequency is reported to be 2.3 THz, well beyond the cut-off frequencies of microwave transistors. Moreover, diodes are easier to model than transistors because they have less parasitics [Maas88]. On the other hand, diode mixers are inherently lossy. For example, LORAL's K-band upconverter (model 4827) has a conversion loss of 9 dB [Loral]. Conversely, active mixers have the advantage of yielding positive conversion gain. Also, well-designed FET mixers consistently have lower noise figures than diode mixers [Maas93]. A major disadvantage of active mixers, though, is the difficulty in modelling the transistors at microwave and millimeter-wave frequencies, in part because of the greater number of parasitics. An additional disadvantage of active mixers is that they are potentially unstable.

Figure 2-4. Single-ended active mixer with diplexer.

The simplest type of mixer has a single-ended configuration. It usually requires a matching network at each port. Since the transistor is usually grounded at the source (for a field effect transistor) or at the emitter (for a bipolar transistor), one of the transistor's ports is used by two of the mixer's signals. For example, Figure 2-4 shows a single-ended upconverter, using a bipolar transistor, with the LO fed at the collector. In this case, the LO is driving the transistor at the collector. Note that a reasonably broad RF conjugate matching network at the collector will also match the LO port. A diplexer (which can be as simple as two adjoining bandpass filters) is required to prevent the LO from going into

the RF port and vice-versa. A passive diplexer is typically much larger than the rest of the circuit. For the example illustrated by Figure 2-4, we see that the greater the frequency separation between the RF and LO signal, the more difficult it becomes to simultaneously match both the RF and LO port using one output matching network (but it is easier to design the two filters that make up the diplexer, because the ample frequency separation means that the roll-off of each filter's response does not have to be as sharp). Alternatively, distinct RF and LO matching networks could be introduced in each branch of the diplexer, but this has the disadvantage of increasing the design difficulty, as each matching network is sensitive to the impedance presented by the other.

Another popular mixer configuration is the single-balanced mixer. It has the advantages of increasing the operating bandwidth, improving isolation, rejecting certain spurious signals, decreasing AM noise from the LO source, and enhancing the dynamic range. If matched transistors are used as the non-linear elements, balanced mixers can also improve the conversion gain. The improved operating bandwidth and isolation in balanced mixers comes from the use of broadband hybrid couplers or baluns. An ideal hybrid is a four-port structure that is matched at all ports. The input is equally divided between the two output ports and the remaining port is isolated. The only two types of hybrid couplers are the 180° and the quadrature (or 90°) hybrids. This fact is a consequence of the unitary property of a four-port scattering matrix [Helsz78]. Figure 2-5 shows a passive single-balanced downconverter with a 180° hybrid on the output side. There are many possible configurations for single-balanced mixers. However, according to Maas [Maas93], a single-balanced mixer can be so called only if, in addition to rejecting most spurious signals, it rejects the LO tone and its accompanying AM noise.

Baluns can be used instead of hybrid couplers. A balun (*balanced - unbalanced*) is a structure that couples the signal from an unbalanced transmission line to a balanced transmission line, or vice-versa. For signals below 18 GHz or so, transmission line

hybrids or passive baluns might be too large for monolithic circuits. Therefore, lumped-element hybrids and active baluns³ are increasingly used.

It is worth noting that the design shown in Figure 2-5 might be impractical if the frequency separation between the LO and input signal is too large for even the most broadband of couplers. For example, an upconverter based on the specifications supplied for this thesis would need a hybrid coupler with an operating frequency band encompassing the IF of 3.6 GHz and the LO of 26.5 GHz. See Appendix A for an analysis and a discussion highlighting the balancing mechanism of a particular type of single-balanced upconverter and how it rejects spurious signals and the LO tone.

Figure 2-5. A passive single-balanced mixer with 180 degree hybrid couplers.

Doubly-balanced microwave mixers using diodes are either star or ring-shaped [Combes87]. Their advantages, over other mixer types, are inherent isolation at all ports, rejection of LO noise, greater rejection of spurious signals, and extremely broadband operation [Maas93]. Note that doubly-balanced mixers are often too complex to incorporate a 50-ohm match for all the devices, so they may have less conversion gain than either single-ended or single-balanced mixers [Maas93]. Substituting the four diodes of a diode-ring mixer by four bipolar transistors forms the basis of a commonly

³ Strictly speaking, a balanced transmission line has odd mode propagation, i.e. equal magnitude but opposite phase between two conductors. Active baluns referred to in the literature, such as in Freunhofer [Freun96] or Maas [Maas93], concern an output of two signals of equal magnitude but with opposite phases, each travelling on a distinct transmission line and where the input signal originates from a single transmission line. For the input or output, the transmission line of choice for MMICs is usually the microstrip line, which is an unbalanced transmission line. The term balun is nevertheless used to describe the equal magnitude and opposite phase output.

used mixer called the Gilbert mixer. A distinct advantage of the Gilbert cell mixer is a circuit topology that can be readily integrated in a compact monolithic form, because it has no crossover connections [Gilbert95]. Doubly-balanced active mixers are also realizable using FETs in a layout similar to the Gilbert cell [Maas93]. An in depth analysis of doubly-balanced mixers is beyond the scope of this thesis and is not done. What is important to remember for doubly-balanced mixers is that they offer very good performance in terms of isolation, spurious rejection, broadband operation, AM noise rejection, and dynamic range. The trade-offs are greater LO power consumption (and greater DC power consumption for active doubly-balanced mixers), increased complexity, and, sometimes, less conversion gain. The double doubly-balanced mixer is another type of mixer configuration and has eight non-linear elements. Note that increasing the number of non-linear elements in a balanced mixer increases the dynamic range, in part because of high order spurious cancellation, but primarily because the LO power is distributed to more non-linear elements.

The mixer configuration of interest in this thesis is the single-ended active mixer, and uses one HBT. Using one non-linear element allows a more exclusive look at the behavior of the HBT when operating the device in a non-linear mode and, most importantly, permits a performance comparison with a single-ended PHEMT upconverter already built and tested by the Communications Research Centre (CRC).

2.2.4 Active microwave mixers

Traditionally, microwave mixers have been passive mixers. In fact, Maas [Maas93] states that the non-linear device most often used in mixers is the Schottky-barrier diode. Increasingly, though, microwave mixers are active, using transistors as non-linear elements. At microwave frequencies, active mixers frequently use GaAs FET-based transistors, such as MESFETs, HEMTs or PHEMTs. They are said to have

superior gain and lower noise figure at microwave frequencies than diodes [Maas93]. In 1976, Pucel et al. [Pucel76] first showed the use of GaAs MESFETs in microwave mixers and concluded that the MESFET was promising in terms of conversion gain and noise. In 1981, Bura [Bura81] demonstrated the feasibility of using a FET in a microwave single-ended upconverter, demonstrating a conversion gain of 3.5 dB at an RF of 6 GHz. He concluded that a FET mixer could replace a diode mixer and the first stage of a power amplifier in the transmit path of a transceiver.

An important reason for the technological lag of active microwave mixers, with respect to passive mixers, is complexity. For one thing, transistors are more difficult to model and analyze, at microwave frequencies, than diodes. In addition, active microwave transistors usually require one or two bias networks, which only increase network complexity. Nevertheless, the advances in MMIC manufacturing techniques in the last ten years have created a demand for functional active microwave mixers. Also, advances in CAD techniques and computer technology are levelling the field between passive and active mixers circuit design.

As already mentioned, the microwave transistor of choice has been the MESFET, and increasingly, its HEMT and PHEMT derivatives. For example, Maas, in the 1993 edition of his "Microwave Mixers" book [Maas93], disregards bipolar transistors for use in microwave mixers and deals exclusively with MESFET or HEMT mixers. In the past few years, though, the HBT has been touted as a capable and potentially superior alternative to the FET-based family of transistors. Some researchers have, predictably, replaced the BJT with the HBT in the Gilbert cell configurations. Kobayashi et al. [Kobay94] have demonstrated a Gilbert cell InGaAs-HBT MMIC upconverter (3 GHz upconverted to 5-12 GHz) that achieved a conversion loss between 0.8 dB and 2.6 dB and with LO-RF and IF-RF isolations in excess of 25 dB. More recently, Freundhofer and Falt [Freun96] have reported on a doubly-balanced GaAs-HBT monolithic upconverter,

which produced a 1 dB conversion gain for an input power of -10 dBm, at an RF of 27 to 30 GHz, with an LO of 26 GHz and a LO-RF isolation of 20 dB. Besides higher DC power consumption and increased circuit complexity, doubly-balanced mixers can suffer from a high noise figure, because of the thermal noise originating from the use of six active devices. For example, Osafune and Yamauchi [Osaf94] report on a single-side band noise figure of 16.5 to 17.8 dB for a GaAs-HBT Gilbert cell multiplier.

Some HBT-based single-balanced mixers have already been reported in the literature. For an RF of 1.08 GHz, Koizumi et al. [Koizu95] have tested a single-balanced dual-gate FET upconverter with a conversion gain of up to 16.7 dB, a noise figure of 6.5 dB and a LO-RF isolation of greater than 40 dB. At an RF of 11 GHz, Bura and Vassilakis [Bura89] reported on a HEMT single-balanced upconverter with a conversion gain of around 3 dB. Comparatively, a single-balanced GaAs HBT upconverter by Wang et al. [Wang90] has a conversion gain of 4 dB and an IP₃ of 4 dBm, at an RF of 1.4 GHz.

No single-ended HBT upconverter results have been found in the literature. One explanation is the fact that the performance of a single-ended HBT mixer heavily depends on the matching network at all ports and, consequently, on the HBT models that enables the design of accurate tuning circuits. Unfortunately, up until a few years ago, most HBT models were either too computationally intensive to solve or were oversimplified and inaccurate [Teeter93].

Single-ended FET-based mixers have been reported. They include a PHEMT single-ended gate mixer, by Matthes et al. [Matt95], that operates in the output frequency range of 32-38 GHz and exhibits a conversion gain of 5 dB for a matched IF port of 100 MHz and a LO power of 7 dBm. The LO-RF and LO-IF isolations are quoted as being greater than 15 and 27 dB respectively.

Table 2-1. Summary of FET-based and HBT microwave upconverters.

Name	Upconverter type	RF	performance
Kobayashi et al. [Kobay94]	HBT Gilbert cell	5-12 GHz	-2.6 dB < G_c < -0.8 dB IF-RF isolation > 25 dB LO-RF isolation > 25 dB for $P_{LO} = 5$ dBm
Freundhofer and Falt [Freun96]	HBT Gilbert cell	27-30 GHz	$G_c = +1$ dB LO-RF isolation = 20 dB for $P_{LO} = 10$ dBm
Osafune and Yamauchi [Osaf94]	HBT Gilbert cell	DC-20 GHz	$G_c = +5$ dB NF = 16.5-17.8 dB for $P_{LO} = 0$ dBm
Koizumi et al. [Koizu95]	FET dual-gate single-balanced	1.08 GHz	$G_c = 16.7$ dB LO-RF isolation = 40 dB NF = 6.5 dB for $P_{LO} = 0$ dBm
Bura [Bura89]	HEMT single-balanced	10.7-11.7 GHz	$G_c = +2.8$ to $+3.5$ dB 1 dB comp. point = 7 dBm IP3 = 15 dBm for $P_{LO} = 17$ dBm
Wang et al. [Wang90]	HBT single-balanced	1.4 GHz	$G_c = +4$ dB IP3 = 4 dBm for $P_{LO} = 0$ dBm
Bura [Bura81]	FET single-ended	6 GHz	$G_c = +3.5$ dB for $P_{LO} = 14$ dBm
Matthes et al. [Matt95]	PHEMT single-ended gate	32-38 GHz	$G_c = +5$ dB LO-IF isolation > 27 dB LO-RF isolation > 15 dB for $P_{LO} = 7$ dBm
Dieudonné et al. [Dieud95]	PHEMT single-ended 1. single-gate 2. dual-gate	32-40 GHz	$G_{c1} = +5$ dB $G_{c2} = +4$ dB NF _{1,2} > 10 dB for $P_{LO} = 8$ dBm
Madihan et al. [Madi95]	HEMT single-ended source-injected	21-22.5 GHz	$G_c = -10$ dB LO-IF isolation > 15 dB LO-RF isolation > 5 dB for $P_{LO} = 3.3$ dBm

Dieudonné et al. [Dieud95] report on two active single-ended upconverters: a single-gate mixer with a conversion gain of 5 dB for a frequency range of 32-40 GHz and a dual gate mixer with a conversion gain of 4 dB for the same frequency range. Finally, Madihan et al. [Madi95] have built a source-injected HEMT upconverter yielding a conversion gain of around -10 dB and a LO-RF and LO-IF isolations better than about 5

dB and 15 dB respectively, at 23.5 GHz. Table 2-1 is a summary of some of the active microwave mixers found in the literature.

2.3 The GaAs heterojunction bipolar transistor (HBT)

Prior to the advent of gallium arsenide (GaAs) HBT technology, the transistors of choice for microwave and millimeter-wave operations were the GaAs FET-based family of transistors: MESFET, HEMT, and PHEMT. At microwave frequencies below the Ku band, high-performance silicon (Si) bipolar junction transistor (BJT) are sometimes used. Advances in manufacturing techniques in the 1980's have led to the realization of HBTs based on GaAs and other III-IV-V compounds [Asbeck87]. Indeed, by 1987, the first analog GaAs HBT integrated circuits were being demonstrated [Oyama93]. HBTs are now also being used for high performance digital circuits [Asbeck93]. Other bipolar heterostructure transistors, based on indium phosphate (InP) and silicon germanium (SiGe), are being investigated by researchers around the world, but we will limit our scope to GaAs-based HBTs. The HBT's potential as a high performance microwave transistor is presently being investigated by researchers and companies. For microwave circuits, the HBT has significant advantages when used in power amplifiers and oscillators because of high power handling capability and low $1/f$ noise respectively [Asbeck93]. Nevertheless, the GaAs HBT is a relative newcomer in the field of microwave circuits, with respect to the GaAs MESFET. As such, the HBT is still being compared against the high-performance heterojunction FETs: the HEMT and PHEMT.

2.3.1 Operation and characteristics of the GaAs HBT

In this section, the characteristics of the GaAs HBT are compared to the GaAs MESFET and HEMT (see Table 2-2). However, before doing so, the HBT's operation

and characteristics are described in comparison with the Si BJT. The HBT's behavior is easily grasped when compared to the behavior of the homojunction Si BJT.

There are two main advantages of GaAs HBT, compared to the Si BJT, and they can be summarized as follows:

- 1) GaAs has greater electron mobility, for a given electrical field strength, which results in a higher cutoff frequency f_T .
- 2) The heterojunction at the emitter-base junction enhances the emitter efficiency, γ , resulting in a decrease in the base resistance, R_b . Consequently, this means a greater maximum frequency of oscillation, f_{max} (equation 2-8 [Asbeck89]).

$$f_{max} = \left(\frac{f_T}{8\pi R_b C_{bc}} \right)^{\frac{1}{2}},$$

where R_b is the base resistance and C_{bc} is the
base - collector depletion capacitance. (2-8)

The first statement is due to the use of GaAs as the transistor substrate. Figure 2-6 [Asbeck93] shows electron velocity, in gallium arsenide and silicon, versus the steady state electric field in the material. Electron mobility, μ , is defined as the slope of the velocity-electric field curve, for a given magnitude of the electric field. From Figure 2-6, we see that the low-field mobility of GaAs is several times larger than for Si. Of course, care must be taken to operate the HBT in the high positive mobility portion of the curve, occurring at electric field values of less than 4×10^3 V/cm, and not in the negative mobility portion of the curve and beyond.

The higher electron mobility leads to a shorter transit time, τ_{EC} , across the emitter-base depletion region, the base region and the base-collector depletion region. For example, Oyama and Wong [Oyama93] compare transit times from commercially

available GaAs HBTs and Si BJTs where the transit time for the former is 4 picoseconds and the latter 12 picoseconds, corresponding to cutoff frequencies of 40 GHz and 13 GHz respectively (where $f_T = (2\pi\tau_{EC})^{-1}$). A figure of merit that is also improved in HBTs is the maximum frequency of oscillation, f_{max} . It is used for analog circuits and is the frequency at which the unilateral gain of the amplifier falls to 0 dB. For common emitter amplifiers, it is given by equation 2-8.

Figure 2-6. Steady-state velocity-field characteristics for electrons in GaAs and Si.

The second statement stems from using a heterojunction at the emitter-base junction of a bipolar transistor. Take the npn transistor of Figure 2-7 [Sze85], biased in the active region, as an example. Electrons from the emitter are injected across the emitter-base depletion region and into the base, giving rise to a current I_{en} . Most of the electrons in the base then make it to the collector-base depletion region and into the collector, forming a current I_{cn} . In the p-doped base, some electrons will recombine with holes resulting in a recombination current I_r . Electron-hole recombination does also occur in the depletion regions but this phenomenon is neglected for the purpose of simplifying the analysis. Figure 2-8 [Kroemer82] illustrates the HBT's energy-band

diagram, when the device is biased in the active region. For both Figures 2-7 and 2-8, the symbol * represents recombination between holes and electrons in the base of the HBT.

Also from Figure 2-8, note that under forward bias conditions at the emitter-base junction ($V_{eb} < 0$), some holes from the base do leak into the emitter, forming the current I_{ep} . Collector holes will also leak from the collector to the base but they have a negligible effect on the current I_c . Note that in Figure 2-7 and Figure 2-8, the arrows are in the direction of current flow and not in the direction of the flow of charged particles.

Figure 2-7. Current components for an HBT operating in the active region.

Figure 2-8. Energy band diagram for the HBT of Figure 2-7.

Kroemer [Kroemer82] uses two figures of merit for a bipolar transistor, β and $\beta_{\max}$, where β is the common emitter current gain and $\beta_{\max}$ is the highest possible value of β . Note that the recombination current, I_r , is very small compared to I_{en} or I_{ep} .

$$\beta = \frac{I_c}{I_b} = \frac{I_{cn}}{I_{ep} + I_{cn} - I_{cn}} = \frac{I_{cn}}{I_{ep} + I_r} = \frac{I_{cn} - I_r}{I_{ep} + I_r} < \frac{I_{cn}}{I_{ep}} = \beta_{\max} \quad (2-9)$$

$$\gamma = \frac{I_{en}}{I_{ep} + I_{en}} \quad (2-10)$$

From equation 2-9, an increase in $\beta_{\max}$ (and β) is achievable by decreasing the injected hole current I_{ep} , or in other words, by increasing the emitter efficiency, γ (equation 2-10). Physically, the emitter's bandgap offset in the valence band, ΔE_g , reduces hole injection from the base to the emitter, thus increasing emitter efficiency, γ . Equations 2-11 to 2-13 [Kroemer82] show that if the energy gap of the emitter is larger than that of the base by an amount ΔE_g , this enhances $\beta_{\max}$ and γ , which leads to increased current gain, β . That extra current gain, though, can be sacrificed by increasing the base doping concentration, N_p , and/or decreasing the emitter doping concentration, N_n . For example, a ΔE_g of 8 kT (200 mV) can offset an increase in the doping ratio, N_p/N_n , of approximately 3000 [Asbeck93]. The increase in base doping concentration means a decrease in the base resistance, R_b . This translates into an increase in $f_{\max}$, as seen from equation 2-8. Meanwhile, a decrease in N_n is also allowed by the increase in the doping ratio. This means less minority carrier storage in the emitter thus reducing the emitter-base capacitance, which positively affects f_T [Asbeck87].

$$J_{en} = N_n v_{nb} e^{\frac{-qV_b}{KT}} \quad (2-11)$$

$$J_{ep} = N_p v_{pe} e^{\frac{-qV_e}{KT}} \quad (2-12)$$

$$\beta_{\max} = \frac{N_n v_{nb}}{N_p v_{pc}} e^{\frac{q}{KT}(V_r - V_s)} = \frac{N_n v_{nb}}{N_p v_{pc}} e^{\frac{\Delta E_g}{KT}}, \quad (2-13)$$

where N_n = emitter doping density

N_p = base doping density

v_{nb} = mean speed of the electrons at the emitter end of the base

v_{pc} = mean speed of the holes at the base end of the emitter

qV_p, qV_n = barrier height for holes and electrons respectively

K = Boltzmann's constant

T = temperature in Kelvin

ΔE_g = amount by which the energy gap at emitter is larger than at the base

Given that the $f_{\max}$ for a Si BJT at most resides in the Ku band, then a heterojunction at the emitter-base junction is essential for the effective operation of a bipolar transistor in the K band, Ka band and beyond. For GaAs HBTs, the heterojunction is formed by employing, at the emitter, a wide energy bandgap material that preferably has a lattice constant similar to the GaAs used at the base. The lattice constant, a_0 , of a material is the equilibrium spacing between two atoms in a crystal structure where the attractive gravitational force equals the repulsive Coulomb force. In actuality, the interatomic spacing varies around a_0 because of thermal atomic vibrations. Similar lattice constants permit chemical bonding between semiconductor layer with minimal strain between the crystalline structures. The Nortel GaAs HBT uses a lattice-matched $\text{Ga}_{51}\text{In}_{49}\text{P}$ semiconductor alloy on GaAs with a bandgap of 1.43 eV.

For all the advantages of the HBT over its silicon homojunction counterpart, the workhorse transistor at microwave frequencies today still remains the Metal Semiconductor FET (MESFET) and its heterojunction improvements, the High Electron Mobility Transistor (HEMT) and the Pseudo-morphic HEMT (PHEMT). Ali and Gupta [Ali91] have presented, in Table 2-2, some merit parameters that illustrate the relative advantages of the MESFET, HEMT and HBT.

Table 2-2. Device merit parameter comparison.

Merit Parameter	MESFET	HEMT	HBT	Comparison
f_t	L	M	H	HBT has the highest speed
f_{max}	M	H	L	HEMT has the highest freq. of operation
Gain * bandwidth	M	H	H	HEMT is best for wideband at high frequency
Noise figure	M	L	H	HEMT is best for LNAs
Phase noise	M	H	L	HBT is best for VCOs
g_m/g_o	L	M	H	HBT is best for highly linear amplifiers
IP3/P _{dc}	H	M	H	MESFET is best at high frequency; HBT is best at low frequency
Collector efficiency	M	H	H	HBT potentially is best power device
Power density	M	M	H	HBT potentially is best power device
L = low	M = medium		H = large	

Compared to the MESFET or HEMT, the HBT seems well suited for microwave power applications such as power amplifiers and VCOs. The HBT has a high current capability per unit area because, being a vertical structure, the HBT's entire emitter area conducts as opposed to the thin horizontal surface channel geometry obtained in FET technologies [Asbeck93]. On the other hand, the HBT does not seem optimum for low-noise applications, such as low-noise amplifiers and downconverters, because of its higher noise figure (although Bayraktaroglu reports that InP-based HBTs have achieved noise figures 1.5 dB to 2 dB lower than GaAs-based HBTs [Bayrak93]). Another distinct advantage of the HBT is the avoidance of more expensive sub-micrometer lithography in favor of epitaxial growth to control the base thickness [Bayrak93]. The base thickness is a vertically-defined structure of the HBT and is a critical transistor dimension because it greatly impacts the transit time, τ_{EC} , across the device. Conversely, in microwave FETs, the critical transistor dimension is the gate length, which is a horizontal structure, and is defined by lithography.

A definite disadvantage of the GaAs HBT, compared to the Si BJT, is its low thermal conductivity. Usually, the HBT is a thermally limited device, where the thermal conductivity of GaAs is three times smaller than the thermal conductivity of silicon (0.54 W/cm-°C for GaAs versus 1.57 W/cm-°C for Si [Sze90]). Consequently, thermal over-heating of the HBT is a design concern.

2.3.2 HBT large-signal model

The transistor used in this thesis is a Nortel Technologies InGaP on GaAs substrate heterojunction bipolar transistor, with an emitter area of 3.5 by 6 square microns. Figure 2-9 [Nortel96] shows the cross-section of an HBT, with all the fabrication layers identified.

Figure 2-9. Nortel GaAs HBT fabrication layers.

The standard HBT model is based on the Gummel-Poon (GP) model. The starting point of the model is the intrinsic BJT model, shown in Figure 2-10 [Nortel96]. Large leakage resistances R_{be} , R_{ce} and R_{bc} , set at 10^9 ohms, are added to provide a DC path when the collector-base and emitter-base junctions are reverse-biased. A diode D_{bcx} and capacitance C_{bcx} is added to simulate the extrinsic base-collector junction.

To include the effect of the junction temperature, the model incorporates a thermal sub-circuit that approximates the temperature rise, $T - T_o$, at the emitter-base junction by multiplying I_c by V_{ce} (the thermal sub-circuit is located in the bottom right corner of Figure 2-12). The R_{th} used in the thermal sub-circuit is the equivalent thermal resistance corresponding to the junction temperature rise [Zhang96b].

The HBT thermal effect physically manifests itself in two ways: by a decrease in the emitter-base junction built-in potential and by an increase in the reverse hole injection from the base to the emitter [Nortel96]. For the base current source configuration of Figure 2-11a, Zhang et al. [Zhang96a] affirm that the dominant effect of the junction's temperature rise is an increase in the base-to-emitter hole injection current, I_{ep} . An increase in I_{ep} results in a decrease in emitter efficiency, γ , which leads to a reduction in the I_c . Since I_b is held constant, that means that β decreases as V_{ce} is increased. The increase in reverse hole injection is modeled by a temperature-controlled voltage source, V_{pe} , that drives a diode, D_{ipe} .

Figure 2-11. DC I-V curves: (a) current source at the base; (b) base-emitter voltage source.

For the base-emitter voltage configuration of Figure 2-11b, where the HBT is biased under constant V_{be} conditions, the dominant thermal effect is the reduction in base-emitter built-in potential [Nortel96]. As junction temperature increases and the base-emitter built-in potential subsequently decreases, more electrons will be injected into the base and hence more electrons will reach the collector. This inflation of the collector current further increases the junction temperature, and so on. An intrinsically low emitter resistance means that the maximum rate at which I_c increases, with respect to V_{ce} , is

high. This sudden and dramatic increase in I_c means that HBT will eventually over-heat. As shown in Figure 2-12, the reduction in the base-emitter built-in potential is modeled by the temperature-controlled voltage source, V_{JT} [Zhang96b]

Figure 2-12. Electrical model of the HBT, including the thermal sub-circuit (lower right).

The emitter barrier effect is a corollary of the conduction band spike that appears in abrupt heterojunctions [Bluy79]. It can be mitigated, but cannot be completely eliminated, by compositional grading [Nortel96]. The conduction band spike impedes electron injection from the emitter to the base. Although the spike height, which depends on the junction compositional grading, is independent of bias, the electron tunneling probability through the conduction spike is not [Sze69]. Thus, it effectively acts as a voltage-dependent emitter barrier [Zhang96b]. Therefore, as shown in Figure 2-12, the model incorporates a voltage source, V_{eb} . The voltage source V_{eb} is controlled by the base-emitter voltage and effectively simulates the emitter barrier effect. The source V_{eb} has a non-linear dependence with V_{be} that is empirically determined [Nortel96].

2.4 Non-linear analysis

Small-signal circuit analysis, in microwave engineering, is synonymous with linear circuit analysis. However, when at least one component of a circuit cannot be linearized without significant loss of accuracy, the circuit is termed non-linear.

Time domain analysis of microwave circuits is impractical because circuits may contain large time constants, with respect to the inverse of the fundamental frequency, which might necessitate thousands of numerical integrations of the time domain differential equations in order to ignore the transient part of the response and be left with the steady state part only [Nahkla96]. Also, at microwave frequencies, time domain analysis of distributed elements is more computationally intensive (again requiring many integration cycles) than frequency domain techniques. This is because a transmission line can be simply and accurately represented by a phasor rotation in the complex plane [Gilmore86]. Moreover, a time domain package such as SPICE does not take into account the frequency dispersion of distributed transmission lines.

Weakly non-linear circuits can be solved by the Volterra series method. The Volterra series method is based on successive approximation where, during the first approximation, all circuit elements are linearized and a solution is found. This linearized solution then serves as a basis for a non-linear approximation of second degree, and so on [Maas88]. Linearization of strongly non-linear circuits can introduce large errors in the initial approximate solution, leading to the subsequent approximation being grossly inaccurate.

The preferred non-linear analysis method for strongly non-linear circuits is the harmonic balance method. The first step of the harmonic balance method is to partition the circuit into a linear sub-circuit and a non-linear sub-circuit. The second step of the harmonic balance is to iteratively match the frequency domain solution of the linear sub-

circuit to the time domain solution of the non-linear sub-circuit. As illustrated in Figure 2-13 [Maas88], a non-linear circuit with 2 sources is divided into an N-port sub-circuit consisting of the circuit's non-linear elements and a (N+2)-port sub-circuit consisting of the remaining linear elements [Gilmore86]. The computation time and the accuracy of the harmonic balance method are a function of the number of harmonics initially specified. Too few harmonics can lead to convergence problems and accuracy errors; too many unduly increases the computational time [Maas88]. Harmonic balance is a hybrid method because the linear sub-circuit is solved in the frequency domain while the non-linear sub-circuit is solved in the time domain and is the preferred method when only the steady-state solution of the circuit is required.

Figure 2-13. Harmonic balance.

The model shown in Figure 2-12 contains linear elements such as R_{ce} or C_{bcx} but also has many non-linear elements. These include the diodes of the intrinsic Gummel-Poon model as well as the diodes D_{cbx} and D_{ipe} , all of which can be termed two-terminal non-linearities. In addition to V_{eb} , which has a non-linear dependence with V_{be} , we have the sources V_{JT} and V_{pe} , both of which are dependent on the junction temperature and hence on I_c and V_{ce} . Such a combination of strongly non-linear

components and dependent sources means that manipulating the HBT model, under large-signal excitation, is considerably easier with Libra's harmonic balance simulation test bench.

2.5 MMIC technology

The use of monolithic microwave integrated circuits (MMIC) is increasingly the norm at microwave and millimeter-wave frequencies. The MMIC differs from its counterpart, the miniature hybrid monolithic microwave integrated circuit (MHMIC), in that the active and passive components of the circuit share a common semi-insulating substrate, thereby eliminating the need for attachments and interconnects [Goyal89]. The impetus behind the emergence of MMICs is the lower cost of production when circuits are manufactured in large quantities.

So far, MMIC technology mostly uses GaAs because the electron mobility in GaAs is substantially higher than in Si. MMIC applications are numerous and diversified: phased-array radar, electronic warfare, communications satellites, remote sensing, global positioning system (GPS), vehicle anti-collision radar, wireless communications, etc. Table 2-3 presents the principal advantages and disadvantages of MMIC compared to hybrid microwave integrated circuits [Rober95].

The principal advantages of MMICs are lower cost and increased reliability, when large volumes are required, like in the burgeoning wireless communications industry. On the other hand, the circuit designer is limited to using the foundry's devices and components. Because of the integrated nature of a MMIC, the designer must also rely on CAD techniques and on the accuracy of the foundry's electrical models for its active and passive components. Another disadvantage, from the circuit designer's point of view, is that the very small size of MMIC makes it harder for the designer to tune the circuit

during the testing phase. Also, the long manufacturing time of MMICs makes building several iterations of a circuit design inconvenient in most cases. Consequently, the simulated performance of a MMIC needs to be very accurate, yet, the circuit design must be forgiving enough to deliver the specified performance despite variations in the manufacturing process.

Table 2-3. Advantages and disadvantages of MMICs and MHMICs.

MMICs	MHMICs
Cheap in large quantities (>1000).	Simple circuits can be cheaper; and automatic assembly is possible. Cheaper in smaller quantities (<1000).
Very good reproducibility.	Poor reproducibility due to device placement and bond wires.
Small and light.	Relatively larger than MMICs.
Reliable.	MHMICs must use epoxy and bond wires to attach the active devices to the rest of the integrated circuit; reliability suffers.
Less parasitics - more bandwidth and higher frequencies of operation.	The best transistors are always available for LNAs and HPAs.
Chip space is at a premium; the circuit must be made as small as possible.	Substrate is cheap, which allows microstrip to be used abundantly.
Limited choice of components.	A vast selection of devices is available.
Long turn around time (2-3 month).	Turn around time can be fast (2-3 weeks), making redesigns easy.
Very expensive to start up a MMIC manufacturing facility	Less capital equipment is required to start-up a MHMIC assembly facility.

2.5.1 GaAs MMIC fabrication issues

A GaAs MMIC is basically a planar structure and is usually fabricated on 3 or 4 inch wafers, where one wafer can contain hundreds of MMICs. The metal that forms the transmission lines on the substrate is defined by lithography. Active components, that have layers that are doped differently, are fabricated by a combination of material growth (epitaxy or ion implantation) and lithography. As seen from Figure 2-9, the HBT's critical structure that governs speed is the p-doped base and is vertically defined. It is

fabricated by epitaxy and ion implantation. For vertical structures, epitaxy is used and is considered more reliable for critical dimension growth (emitter and base layers) [Goyal89]. As already mentioned, high performance MESFETs, HEMTs and PHEMTs, with horizontal gate lengths less than 1 micrometer, must use lithography. Optical lithography technology is stretched to its limit in the sub-micrometer range so that at these dimensions some manufacturers are constrained to use very expensive electron lithography [Rober95].

2.5.2 HBT performance and fabrication

The HBT of concern in this thesis uses GaInP at the emitter, in part because there is good conduction-band line up between the emitter's GaInP alloy and the GaAs at the base [Nortel96]. The HBT layers are grown by metal-organic chemical vapor deposition (MOCVD) epitaxy, which is described by Goyal [Goyal89] as an epitaxial process well suited for high rate, low-cost production. The p type base is doped using carbon while the dopant in the emitter and collector is silicon [Nortel96].

Although reproducibility is an advantage of MMICs over MHMICs, process variations do affect device performance within a wafer and from wafer to wafer. The HBT model used in this thesis is based on the average performance of Nortel's GaAs HBT so far. Table 2-4 shows some key HBT electrical performance parameters as well as the minimum and maximum values of these parameters, recorded from previous fabrication runs [Nortel96].

Table 2-5 gives some of the maximum ratings for the process. The HBT process utilizes two metal layers, which are used to fabricate the metal-insulation-metal (MIM) capacitors. Metal 2 is one of two metals used in this process and, having the lowest resistivity, is normally the conductor of choice for the microstrip transmission lines. The

maximum emitter current density for either the constant I_b or the constant V_{be} biasing condition is provided. For an emitter area of $3.5 \mu\text{m} \times 6 \mu\text{m} = 21 \mu\text{m}^2$, the maximum emitter current is 10.5 mA. Above that current, this particular Nortel GaAs HBT is more likely to malfunction because of thermal difficulties. In Table 2-5, the alloy NiCr is used to fabricate lumped resistors.

Table 2-4. Key electrical parameters of the HBT.

HBT parameter	Min	Mean	Max
f_T (GHz)	50	60	N/A
f_{max} (GHz)	60	75	N/A
g_m (mS)	20	30	40
β	50	100	200
V_{be} turn on voltage	1.34	1.4	1.45

Table 2-5. Maximum ratings of the HBT.

Parameter	Rating
storage temperature	-65 to 100°C
junction temperature	-55 to 125°C
emitter current density	$5 \times 10^4 \text{ A cm}^{-1}$
metal 1 current density	3 mA/ μm width
metal 2 current density	5 mA/ μm width
NiCr current density	0.5 mA/ μm width
MIM capacitor voltage	20 volts

Finally, the MMIC fabrication process used in this thesis includes via holes. They are basically metallized holes within a circuit layout that establish a localized ground. Vias are always slightly inductive (typically less than a nanohenry). Without via holes, grounded devices and components would have to be placed near the edge of the MMIC, with either wrap-around metallization or bond wires providing the ground contact to the back plane [Rober95]. At low microwave frequencies, it is possible to share vias by using a ground bar, but at millimeter-wave frequency, the higher inductance of the ground bar can create an undesirable coupling path in the circuit and should be avoided.

CHAPTER 3: ANALYSIS OF THE HBT AS AN UPCONVERTER

3.1 Introduction

In this chapter, the HBT is investigated for use as an upconverter. First, confidence in the model is established. Then, a suitable bias point is found, based on the non-linear characteristic of the HBT and on the chosen mixer configuration. Finally, a formula for estimating the upper-bound of the conversion gain is derived.

3.2 Model confidence

This model, originally developed for SPICE, was implemented in Libra by Laneve et al. [Laneve96]. They found good agreement between simulated and laboratory measurements for the DC curves, small-signal S-parameters, and a 14-GHz power amplifier output power. Figure 3-1 [Laneve96] shows a graph of simulated large signal measurements for an HBT power amplifier at 14 GHz. The simulation results of the output power and of the gain of the power amplifier are in good agreement with the measured results of that power amplifier.

For the model implemented in SPICE, Nortel has found good agreement between measured and simulated data when comparing the forward h-parameter, h_{21} , and the maximum gain of an HBT under high power conditions [Nortel96]. For the Nortel comparison, the error between the measured and modeled data is within 10% over a wide bias range [Nortel96]. The Nortel document also reports very good agreement between measured and modeled DC I-V measurements.

The above-mentioned agreement between simulated and measured data is taken as proof that the model is valid under certain small-signal and large-signal conditions. This does not mean that the model is valid for all conditions at all frequencies but, for this

thesis, it is taken as evidence that the model depicts the HBT realistically. The Nortel GaAs HBT is still an experimental process and the model will not predict all the process variations from wafer to wafer, or even from HBT to HBT on the same wafer. Nevertheless, the model remains the best tool to extensively investigate the HBT's behavior as an upconverter and to design an upconverter based on that investigation.

Figure 3-1. Comparison between HBT large-signal simulation results and measurements.

3.3 Investigation of the HBT as an upconverter

The aim of section 3.3 is to investigate the non-linear potential of the HBT and to identify a suitable bias point for an upconverter design. The HBT is not considered for use as a downconverter in this investigation because the HBT model used does not include noise parameters. As is evident from Table 2-2, we see that the HBT has worse

noise performance than either MESFETs or HEMTs. This makes the HBT comparatively less attractive as a downconverter if low noise figure is, indeed, a critical specification.

3.3.1 Bias point

The bias point for a single-ended upconverter must be such that the mixer can be driven into a non-linear mode of operation. The HBT has more than one non-linear region that can be exploited when using the transistor in a mixer design. The best region of operation should be chosen so as to maximize upconverter performance. As a minimum, however, the non-linear region must noticeably generate the desired mixer product, $f_{RF} = f_{IF} + f_{LO}$. The remainder of section 3.3.1 explores several criteria that help to identify the best possible non-linear region.

3.3.1.1 The HBT's non-linear regions

From Figures 3-2a and 3-2b, we see that the HBT has two potential biasing regions that can be exploited for a mixer application: the non-linear transition between the saturation region and the active region (henceforth called the knee region) of Figure 3-2a and Figure 3-2b, and the non-linear transition between cut-off and the active region (henceforth called the cut-off region) of Figure 3-2a and Figure 3-2b. Both Figure 3-2a and Figure 3-2b differ only by their choice of bias control mechanism (constant base current for Figure 3-2a and constant base-emitter voltage for Figure 3-2b) and thermal effects.

The knee region represents the same non-linear transition between the saturation and active modes of the HBT; in the saturation region, both junctions are forward biased, and in the active region, the emitter-base junction is forward biased and the collector-base junction is reverse biased. Thus, the knee region occurs when the collector-base junction goes from being forward-biased to being reverse-biased.

The cut-off region represents the turn-on region of the base-emitter junction. Note that in Figure 3-2a, there does not seem to be a non-linear relationship between I_b and I_c . This is because the ratio $\beta = I_c/I_b$, for a given V_{ce} , is relatively constant and therefore linear. For either Figure 3-2a or Figure 3-2b, however, there is a non-linear dependence between V_{be} and I_c , as the base-emitter diode-like junction of the HBT turns on.

Figure 3-2. Non-linear regions of the HBT: (a) current source at the base; (b) base-emitter voltage source.

In the knee region, the device can be driven into a non-linear region of operation by varying V_{ce} . For the cut-off region, non-linear operation can be achieved by the variation of V_{be} . In each case, the output is I_c at the collector port of the HBT. To evaluate how strong each non-linear region is, a simple graphical technique is used: the second local derivative of the dependent variable, I_c , is taken with respect to the independent variable, where the independent variable is varied by the LO tone (V_{ce} for the knee region, and V_{be} for the cut-off region). Figure 3-3a-c is a graphical representation the second (local) derivative, for the knee and cut-off regions respectively, with respect to the independent variable. The analysis is carried out for the cut-off region of Figure 3-2b only because it demonstrates the non-linear I_c - V_{be} dependence.

knee region of Figure 3-2a: $\frac{\partial^2 I_c}{\partial (V_{ce})^2}$ for $I_b = 0.08$ mA;

knee region of Figure 3-2b: $\frac{\partial^2 I_c}{\partial (V_{ce})^2}$ for $V_{be} = 1.36$ volts; and

cut-off region of Figure 3-2b: $\frac{\partial^2 I_c}{\partial (V_{be})^2}$ for $V_{ce} = 2$ volts.

Figure 3-3. Second (local) derivatives for the three non-linear bias regions of the HBT.

The greater the value of the second (local) derivative, the larger the non-linear relationship. By visual inspection of Figure 3-3a-c, it is apparent that the strongest non-linear relationship occurs in the cut-off region, where the second derivative follows an exponential relationship. Since the cut-off region corresponds to the base-emitter turn-on

region, an exponential relationship is expected. The second derivative carried out in the cut-off region corresponds to the first derivative of the transconductance for a bipolar transistor, where the transconductance is defined as $g_m = \partial i_c / \partial v_{be}$ [Sedra87]. Anholt [Anholt95] gives a mathematical description for the forward operation of a bipolar device, equation 3-1, where we see that I_c is, indeed, exponentially dependent on V_{be} . Additionally, Oyama and Wong state that the HBT's transconductance is fifty times greater than for HEMTs [Oyama93]. This implies that the HBT, biased in the cut-off region, has greater conversion gain potential when compared to the HEMT.

$$J_c = q \left\{ \exp \left[\frac{q V_{be}}{kT} \right] - 1 \right\} \alpha_T \frac{D_{nb} n_{ib}^2}{GN_b}, \quad (3-1)$$

where J_c = collector current density

GN_b = Gummel number for the the base

D_{nb} = diffusion constant for the electrons in the base

V_{be} = applied base - emitter voltage

α_T = forward emitter - collector transfer coefficient

n_{ib} = intrinsic electron concentration in the base layer

K = Boltzmann's constant

T = temperature in Kelvin

The cut-off region is the bias region with the strongest non-linear potential generating the desired intermodulation product with a minimum amount of LO power. As an added benefit, DC power consumption is not very large in the cut-off region, as the device is cut-off for half of the LO cycle.

Compared with the constant I_b biasing of Figure 3-2a, it is easier to bias in the cut-off region when biasing with a constant V_{be} because it requires a constant voltage source at the base. For a portable application, the constant voltage can be directly supplied by the batteries, as opposed to a more complex constant current source circuit. Note that in an actual application, a scheme to limit the collector current below the thermally safe threshold would have to be implemented.

3.3.1.2 Upconverter configuration

The choice of configuration is dependent on the requirements imposed on the mixer. For a downconverter, where a low noise figure and a large dynamic range are usually critical specifications, a balanced mixer configuration is often essential. For upconverters, where low conversion loss (or high conversion gain) is desired and where the input IF power is habitually held constant over time, a single-ended configuration is an adequate configuration. In this case, it is desirable to minimize the amount of LO power required of the oscillator (while minimizing the conversion loss) because the oscillator is to operate at Ka band. A further advantage of using a single-ended configuration for a microwave upconverter, when low LO power is a requirement, is that single-ended mixers traditionally consume the least amount of LO power. The obvious disadvantage of single-ended mixers is that they do not inherently suppress spurious signals and must often rely on an output filter for that task. Nevertheless, the upconverter configuration chosen in this thesis is the single-ended configuration. First, a single-ended upconverter will permit a more exclusive look at the HBT when it is operating in a non-linear mode. Second, it has the potential of perhaps yielding high conversion gain while minimizing the required amount of Ka band LO power.

Traditionally, a microwave single-ended upconverter that uses a MESFET (or HEMT or PHEMT) needs a diplexer. Diplexers are passive circuit elements that are typically much larger than the rest of the circuit, and typically, the diplexer consists of two adjoining bandpass filters. Since the MESFET's source is usually grounded by a via hole, this leaves only two transistor ports (the gate and drain) for handling the RF, IF and LO signals. The source is grounded because MESFETs are usually laid out with the source divided into two ports and located on either side of the gate-drain axis. From Figure 3-4 [Ladbr89], we see that using the source as an input port for the LO means using a cross-over line over the drain line or the gate line. Alternatively, letting the LO in

through only one of the source's port (with the other source port grounded) prevents using a cross-over line but complicates the modelling of the mixer because of the asymmetrical usage of the MESFET's source. On the other hand, the Nortel HBT used in this thesis has only one emitter port. Thus, the LO can be fed at the HBT's emitter port without requiring a cross-over line and without complicating the HBT model. In other words, the IF signal is fed at the base of the HBT, the LO signal is injected at the emitter, and the RF signal exits from the collector (see Figure 3-5).

Figure 3-4. Top view of a GaAs MESFET showing the source divided into two ports.

Using a transistor as a three-port network has been reported by Madihan et al. [Madi95] for a K band HEMT single-ended upconverter. At an operating frequency below 1 GHz, Yip et al. [Yip92a] report on a FET single-ended upconverter and a BJT single-ended upconverter. A similar emitter-injected concept, involving a single-ended HBT mixer, has not been found in the literature. Even though using a diplexer is avoided, even an emitter-injected single-ended mixer will probably require an RF band pass filter (BPF) in order to effectively suppress the spurious signals. However, with the LO-injected single-ended configuration, the mixer can be tested without the BPF. The resulting mixer will either be an MHMIC, if the BPF is fabricated separately and must be

bonded to the "mixing" circuit, or can be completely integrated, if the "mixing" circuit and the BPF are put on a single MMIC.

Figure 3-5. Emitter-injected HBT single-ended upconverter.

3.3.1.3 Load-line analysis

A load-line analysis (Figure 3-6) using Libra provides a useful insight into the large-signal behavior of an HBT emitter-injected mixer at the two non-linear bias regions for the constant V_{be} biased HBT. The Libra test circuit and a typical load-line analysis test bench are shown in appendix B. For the load-line analysis, the LO power level is set at 5 dBm.

Note that the test circuits, as shown in appendix B, have LO quarter-wave open-circuited stubs at the IF (input) and RF (output) ports because, in an actual mixer, it will be necessary to prevent the emitter-injected LO from leaking at either the IF or RF ports. Of course, an LO quarter-wave stub located at the IF and RF ports forces an LO short-circuit at both ports, implying an instantaneous DC bias point variance for V_e only. In other words, for this configuration, both $V_{be} = V_b - V_e$ and $V_{ce} = V_c - V_e$ will vary with the LO pump signal ($P_{LO} = 5$ dBm).

From Figure 3-6a, we see that when the HBT is biased in the knee region, the load-lines do not quite follow the non-linear I_c - V_{ce} relationship. On the other hand, Figure 3-6b shows that the load-line biased in the cut-off region is still switching the HBT between the cut-off and active regions, and thus is more faithful to the (sought after) exponential relationship of the cut-off region.

Figure 3-6. Load-line analysis for: (a) the knee region; (b) the cut-off region.

3.3.1.4 Chosen bias point

Taking into account the degree of non-linearity of each non-linear region, the chosen upconverter configuration, and the above load-line comparison of each non-linear region, the bias point for the single-ended upconverter is chosen to be in the cut-off region. The cut-off region takes advantage of a large non-linearity that behaves in an exponential fashion and is well suited for an emitter-injected concept. Biasing the HBT at the edge of the cut-off region, and subjecting it to the action of the LO pumping signal, drives the HBT into the active region. On the positive half of cycle of the LO, the device

is cut-off, while on the negative half, the device is active. The base-emitter bias voltage (or operating voltage), shown in equation 3-2, is the instantaneous base-emitter bias voltage that varies with the LO tone.

$$V_{be/instantaneous} = V_{DC/be} + V_{LO} \cos(\omega_{LO} t) \quad (3-2)$$

3.3.2 Estimation of the conversion gain

Figure 3-7. Upconverter with input and output matching networks.

From Figure 3-7, we see that the IF signal is transferred to the HBT, where it is mixed with the LO signal, resulting in an RF signal. In Figure 3-7, MN_{IF} represents the IF matching network, MN_{RF} represents the RF matching network, $P_{avs/IF}$ is the available IF power from the source, $P_{in/IF}$ is the IF power transferred to the HBT, $P_{avt/RF}$ is the available power from the HBT, and $P_{l/RF}$ is the RF power transferred to the load. Also in Figure 3-7, note that the impedance of the IF source, Z_S , and the impedance of the RF load, Z_L , are equal to $Z_0 = 50$ ohms. The symbols V_{IF} , V'_{IF} , V_{RF} , V'_{RF} , I_{IF} , and I_{RF} are phasor representations of the corresponding time-varying IF and RF voltages and currents.

The conversion gain of an emitter-injected HBT can be estimated by separating the mixer of Figure 3-7 into three events: the IF power transfer to the HBT, the mixing process, and the RF power transfer to the output load. To simplify the analysis, it is assumed that the HBT is unilateral, i.e., $S_{12} = 0$. For the first event, the IF source power transferred to the HBT is calculated as per equation 3-3 [Collin92].

$$\begin{aligned}
 P_{in/IF} &= \frac{1}{2} \operatorname{Re} \{ V_{IF}' (I_{IF})^* \} = \frac{1}{2} \operatorname{Re} \left\{ V_{IF} \frac{Z_{IF}}{Z_{IF} + Z_s} \frac{V_{IF}^*}{Z_{IF}^* + Z_s^*} \right\} \\
 &= \frac{|V_{IF}|^2}{8Z_s} \frac{4 R_{IF} Z_s}{|Z_{IF} + Z_s|^2} = \frac{|V_{IF}|^2}{8Z_s} M_{IF} = P_{avs/IF} M_{IF} \\
 &= \frac{|V_{IF} \sqrt{M_{IF}}|^2}{8Z_s} = \frac{|V_{IF}'|^2}{8Z_s},
 \end{aligned} \tag{3-3}$$

where $0 < M_{IF} < 1$ is the input impedance mismatch factor.

Note that the magnitude of the voltage waveform transferred to the HBT can be written as $|V_{IF}'| = |V_{IF}| \sqrt{M_{IF}}$. For the third event, we treat the HBT as the RF "source", where the RF signal is generated by the "mixing" of the IF signal with the LO tone. The RF power transfer to the 50-ohm load is written in equation 3-4, where the impedance of the RF source is the resistance looking into the RF port, $R_{RF} = \operatorname{Re}\{Z_{RF}\}$.

$$\begin{aligned}
 P_{I/RF} &= \frac{1}{2} \operatorname{Re} \{ V_{RF}' (I_{RF})^* \} = \frac{1}{2} \operatorname{Re} \left\{ V_{RF} \frac{Z_l}{Z_{RF} + Z_l} \frac{V_{RF}^*}{Z_{RF}^* + Z_l^*} \right\} \\
 &= \frac{|V_{RF}|^2}{8R_{RF}} \frac{4 R_{RF} Z_l}{|Z_{RF} + Z_l|^2} = \frac{|V_{RF}|^2}{8R_{RF}} M_{RF} = P_{avI/RF} M_{RF},
 \end{aligned} \tag{3-4}$$

where $0 < M_{RF} < 1$ is the output impedance mismatch factor.

The second event is the actual frequency conversion occurring in the HBT. To estimate the voltage conversion gain (or loss) of the transistor, Yip [Yip92b] proposes a

method that takes into account the time variation of the forward S-parameter, S_{21} . Since the HBT is biased at cut-off, one half-cycle of a sufficiently strong LO tone will drive it into the active region, while the other half-cycle will maintain the HBT at cut-off as seen in equation 3-2. This is analogous to turning the HBT on and off, at the LO frequency. Assuming an ideal transistor (rise-time, fall-time, delay time and storage time are set to zero), the variation of the magnitude of S_{21} , as seen by the IF input signal, can be approximated by a square wave with frequency, f_{LO} . This is shown in Figure 3-8 [Yip92b]. The Fourier series of the corresponding square wave is represented by:

$$S_{21}(t) = \sum_{n=0}^{\infty} A_n \cos(n\omega_{LO}(t)) = \frac{|S_{21\max}| + |S_{21\min}|}{2} + \frac{2(|S_{21\max}| - |S_{21\min}|)}{\pi} [\cos(\omega_{LO}t) - \frac{1}{3}\cos(3\omega_{LO}t) + \frac{1}{5}\cos(5\omega_{LO}t) - \dots], \quad (3-5)$$

$$\text{where } A_0 = \frac{|S_{21\max}| + |S_{21\min}|}{2} \text{ and } A_{1,3,5,\dots} = \frac{2(|S_{21\max}| - |S_{21\min}|)}{n\pi}$$

In Figure 3-8, $|S_{21\max}|$ is the magnitude of S_{21} while the HBT is in the active region and $|S_{21\min}|$ is the magnitude while the HBT is cutoff.

Figure 3-8. Variation of S_{21} (at 3.6 GHz) with respect to time.

The output voltage is a square wave that can be written as in equation 3-6. Note that S_{21} is a parameter for a two-port device in a 50-ohm system. The on and off S_{21} parameters are measured or simulated before the IF and RF matching networks are connected, and

might not be identical to the S_{21} parameter once the matching networks are connected. Strictly speaking, equation 3-6 only provides an indication of the output voltage waveform without the IF or RF matching networks.

$$\begin{aligned}
 v_{\text{out}}(t) &= S_{21}(t) * v_{\text{IF}}(t) = \sum_{n=0}^{\infty} A_n \cos(n\omega_{\text{LO}}t) * |V_{\text{IF}}'| \cos(\omega_{\text{IF}}t) \\
 &= \sum_{n=0}^{\infty} \frac{A_n |V_{\text{IF}}'|}{2} [\cos((n\omega_{\text{LO}} - \omega_{\text{IF}})t) + \cos((n\omega_{\text{LO}} + \omega_{\text{IF}})t)] \\
 &= A_0 |V_{\text{IF}}'| \cos(\omega_{\text{IF}}t) + \frac{A_1 |V_{\text{IF}}'|}{2} \cos((\omega_{\text{LO}} \pm \omega_{\text{IF}})t) \\
 &\quad + \frac{A_2 |V_{\text{IF}}'|}{2} \cos((2\omega_{\text{LO}} \pm \omega_{\text{IF}})t) + \dots
 \end{aligned} \tag{3-6}$$

The RF signal is the upper side band signal corresponding to the frequency term $\omega_{\text{LO}} + \omega_{\text{IF}}$ in the series expansion of equation 3-6. By substituting the corresponding value for A_1 of equation 3-5 into the RF term of equation 3-6, we get:

$$\begin{aligned}
 v_{\text{RF}}(t) &= \frac{A_1 |V_{\text{IF}}'|}{2} \cos((\omega_{\text{IF}} + \omega_{\text{LO}})t) \\
 &= |V_{\text{IF}}'| \frac{|S_{21\text{max}}| - |S_{21\text{min}}|}{\pi} \cos((\omega_{\text{IF}} + \omega_{\text{LO}})t)
 \end{aligned} \tag{3-7}$$

The voltage conversion gain of the HBT, $G_{\text{cv/HBT}}$, is the magnitude of the ratio of the RF voltage waveform originating from the mixing process of the HBT, $v_{\text{RF}}(t)$, over the IF voltage waveform transferred to the base of the HBT, $v_{\text{IF}}(t)$:

$$G_{\text{cv/HBT}} = \frac{|v_{\text{RF}}(t)|}{|v_{\text{IF}}(t)|} = \frac{|V_{\text{RF}}|}{|V_{\text{IF}}'|} = \frac{|S_{21\text{max}}| - |S_{21\text{min}}|}{\pi} \tag{3-8}$$

Finally, the mixer conversion gain, G_c , is obtained by taking the ratio of the RF power transferred to the load, $P_{I/RF}$, over the IF power available from the source, $P_{avs/IF}$. G_c is obtained by combining equations 3-3, 3-4, 3-8, and recalling that $|V_{IF}| = |V_{IF}'|/\sqrt{M_{IF}}$:

$$\begin{aligned}
 G_c &= \frac{P_{I/RF}}{P_{avs/IF}} = \frac{|V_{RF}|^2 M_{RF}}{8R_{RF}} \frac{8Z_s}{|V_{IF}|^2} = \frac{|V_{RF}|^2}{|V_{IF}'|^2} \frac{M_{RF}Z_s}{R_{RF}} \\
 &= \frac{|V_{RF}|^2}{|V_{IF}'|^2} \frac{M_{RF}M_{IF}Z_s}{R_{RF}} = \left[\frac{|S_{21max}| - |S_{21min}|}{\pi} \right]^2 M_{IF}M_{RF} \frac{Z_s}{R_{RF}} \quad (3-9)
 \end{aligned}$$

Of course, the expression for G_c is an upper limit of the mixer's conversion gain because it assumes a unilateral ideal transistor, lossless matching networks, and a narrowband input signal. However, it does provide the designer with an appreciation of the expected conversion gain (or loss) of a reactively matched emitter-injected mixer. Armed with that estimate, the circuit designer can decide a priori if performance trade-offs are possible (or necessary) between the conversion gain and other performance parameters such as stability, return loss, and useful bandwidth. In chapter 4, an estimate for G_c is calculated for that purpose.

CHAPTER 4: DESIGN, LAYOUT AND MEASUREMENTS

4.1 Introduction

This chapter forms the core of the thesis. It elaborates on the various phases leading to a GaAs HBT MMIC upconverter: design, layout, measurements, and performance comparison. Section 4.2 is a prelude to the MMIC design, where the specifications and constraints applicable to the design are reviewed. Section 4.3 outlines the design itself. Also presented in section 4.3 are the actual MMIC layout and EM simulation on selected parts of the design. Section 4.4 outlines the testing done on the upconverter. It shows the results obtained from the MMIC upconverter measurements, compares the measurements with the simulation results, and offers possible explanations to account for the discrepancies between the measured and simulated results. Section 4.4 also provides a brief account of the difficulties encountered during the testing phase. In section 4.5, the design is re-examined in order to find out why the measured IF and RF return losses do not quite agree with the simulated return losses. Finally, in section 4.6, the HBT upconverter is compared to a single-ended PHEMT upconverter with similar specifications.

4.2 Design specifications and constraints

Table 4-1 summarizes CRC's specifications and constraints for the design of an upconverter for a wireless LAN. Each port must be conjugately matched to 50 ohms, within the specified bandwidth. This facilitates the insertion of the mixer within a 50-ohm system, and allows maximum power transfer at each port. The desired return loss at each port is -10 dB or less. Matching the IF and RF ports to 50 ohms, by means of conjugate matching, also ensures that the maximum conversion gain is realized, as maximum power is transferred from the IF source to the RF load. Within the RF

operating frequency band (29.85 GHz to 30.35 GHz), the conversion gain should vary by no more than a few decibels.

Table 4-1. Mixer specifications and constraints.

SYSTEM SPECIFICATIONS	
IF, RF, LO	3.6 GHz, 30.1 GHz, 26.5 GHz
Operating bandwidth	500 MHz
P_{IF}	-10 dBm
P_{LO}	+5 dBm or less
DESIGN SPECIFICATIONS	
IF port return loss	-10 dB or better from 3.35 GHz to 3.85 GHz
RF port return loss	-10 dB or better from 29.85 GHz to 30.35 GHz
LO port return loss	-10 dB or better at 26.5 GHz
LO-RF isolation	as high as possible
LO-IF isolation	as high as possible
Conversion gain	as high as possible
DESIGN CONSTRAINTS	
maximum MMIC size	2.5 x 3.7 mm ²
minimum distance between adjacent lines	one substrate height, or 100 microns
use of lumped-element components	below 18 GHz

As stated earlier, single-ended mixers do not inherently suppress the LO tone. Consequently, LO suppression at the IF and RF ports should be included in the upconverter design. As will be seen in section 4.3, quarter-wave open-circuit LO microstrip stubs are inserted on the IF and RF side in order to meet the specified LO-IF isolation and LO-RF isolation.

Due to the allotted space on the wafer, the maximum size of the MMIC is 2.5 mm by 3.7 mm. Because the mixer is to be matched to 50 ohms at each port, the circuit must incorporate a matching circuit at all three ports of the upconverter. In addition, the upconverter requires two bias networks: one for the base and one for the collector. These five networks, plus the HBT, form the upconverter.

As will be seen later in this chapter, to fit all the networks into the prescribed area, it is necessary to use lumped-element spiral inductors in the IF matching network and the base bias network. Above 18 GHz, lumped element inductors are not recommended, not only because they may have self-resonant frequencies close to or below the intended frequency of operation, but also because they are usually not well-characterized [Comsat89]. Hence, above 18 GHz, it is advisable to use distributed elements, such as open-circuited and short-circuited stubs. At all microwave frequencies, distributed elements are well-characterized, and, above 18 GHz or so, are short enough in length to fit in MMICs. Another constraint is to keep the distance between transmission lines, or any components on the circuit layout, at least one substrate height apart or 100 microns [Comsat89]. This rule-of-thumb distance is enough to prevent unanticipated parasitic coupling between adjacent components.

For the Nortel MMIC process, the lumped-element spiral inductors include, but are not restricted to, a library of pre-tested spiral inductors. The advantage of using inductors from the library is that they are already laid-out in MMIC form and are ready to be inserted in the circuit layout. Note that these inductors are not measured above 20 GHz, and most of them have a self-resonant frequency, f_{res} , of less than 20 GHz. For Libra simulation, the Nortel spiral inductors are taken into account by using the supplied S-parameters data files, originating from laboratory measurements done at Nortel.

Different types of capacitors include metal-insulation-metal (MIM), interdigitated, and end-coupled. MIM capacitors have the advantage of being approximately modeled by a lumped-element capacitor and a series transmission line of length and width equal to the MIM capacitor actual length and width [Nortel96]. A simple scaling formula for MIM capacitors is given by equation 4-1.

$$\text{Area of MIM capacitor} = 1000^2 \frac{C}{142}, \quad (4-1)$$

where C is in pF.

As mentioned in section 2.5.2, the Nortel MMIC process uses via holes. For this process, a via hole is modeled as a series inductance and resistance of about 0.05 nH and 0.053 ohms respectively.

For this MMIC process, the microstrip line has the following parameters: relative permittivity of GaAs $\epsilon_r=12.9$, substrate height $h=100 \mu\text{m}$, metal thickness $t=2 \mu\text{m}$, resistivity relative to gold $\rho=1.5$, and loss tangent $\delta=0.001$ [Nortel96]. Therefore, for a 50-ohm microstrip line, the effective relative permittivity⁴ at 3.6 GHz is 8.167. At 26.5 GHz, it is 8.317, and at 30.1 GHz, it is 8.347.

4.3 Upconverter design

The bias point of the single-ended, emitter-injected upconverter is set in the transistor's cut-off region (section 3.3.1). For a mixer, if the HBT is to be biased in region 3, then V_{ce} should be kept relatively low. This is a preventive measure to ensure that the instantaneous bias point of the device does not swing into the thermally unstable region. If the quiescent V_{ce} is 3 volts as opposed to 2 volts, Figure 3-2 shows that the HBT is more susceptible to over-heat because it takes less V_{be} to force the HBT into the

⁴ The relative permittivities are calculated using HP-EEsof's Linecalc™.

thermally unstable region. On the other hand, an emitter-injected HBT, with a quiescent point that has a low V_{ce} , is more likely to have its instantaneous bias point enter the saturation region. It is important that the instantaneous bias point of the HBT swing into the active region, as opposed to the saturation region, in order to maximize the conversion gain of the mixer (or minimize its conversion loss). As a compromise between avoidance of device over-heating and of the saturation region, the collector-emitter voltage, V_{ce} , is set at 2 volts.

The quiescent base-emitter voltage, V_{be} , and the LO power level determine the quality of the non-linear relationship (recall section 2.2.1). The simulated values of V_{be} and P_{LO} were varied and the combination that gave the maximum simulated conversion gain (without the collector current exceeding 10.5 mA) was chosen to be 1.1 volts and -3 dBm respectively. Why is V_{be} set at 1.1 and not higher? The reason is that increasing V_{be} past 1.1 volts or so made the HBT unstable (according to Libra simulations).

At $V_{be} = 1.1$ volts and $P_{LO} = -3$ dBm, the HBT's base-emitter junction turns on and off, yet I_c stays below the maximum allowable collector current of 10.5 mA. This is verified by a simulated load-line analysis. Note that the small value of LO power means that less AM noise is transferred to the mixer, thereby keeping the noise floor as low as possible. Low LO power also helps the RF to LO suppression and the IF to LO suppression. Recall, however, that the upper bound of the dynamic range (the 1 dB compression point) will be lower than the P_{LO} value.

Figure 4-1 shows a block diagram of the upconverter that includes the five networks: the IF matching network, the RF matching network, the LO matching network, the base bias network, and the collector bias network. Figure 4-1 serves as a starting point for the design, and is referred to throughout section 4.3 in order to clarify the design procedure. As will be apparent in section 4.3.5, the collector bias network could actually

be placed after the RF matching network without significantly disturbing the behavior of the circuit. However, as will be seen in section 4.3.4, placing the base bias network after the IF matching network would have a more significant effect.

Figure 4-1. Upconverter block diagram.

From Figure 4-1, notice the quarter-wave LO open-circuit stub next to the base of the HBT. It serves as an LO trap that blocks the LO signal from leaking out of the IF port. Nevertheless, the LO trap is used to guarantee suppression of the LO tone at the IF port. As will be seen in section 4.3.5, a distributed LO trap is also inserted in the RF matching network. Finally, there are the capacitors located near the IF and RF ports. Both are MIM capacitors of approximately 5 pF each and serve as DC blocks. At 5 pF, the MIM capacitor has a negligible effect on the impedance seen at the RF port. At the IF port, the capacitor has a small but manageable effect on the IF matching network.

Appendix C shows the Libra circuit of the completed upconverter, and the two test benches necessary for simulation the upconverter. The first test bench is used to simulate the conversion gain, the spurious suppression at all ports, and the return losses at

the LO and IF ports. The second test bench in appendix C is used to calculate the return loss at the RF port.

4.3.1 LO matching network

One of the goals of the LO matching network is to match the LO impedance looking into the emitter port of the HBT to 50 ohms. The other goal is to provide a DC ground for the emitter port of the HBT. With only the LO trap on the IF side and looking towards the emitter, the LO impedance is $Z_{LO} = 42.8 - j23.9$ ohms. This translates into a reflection coefficient of $\Gamma_{in/LO} = 0.261 \angle -92.2^\circ$. This value of the LO signal's reflection coefficient, $\Gamma_{in/LO}$, yields a value of return loss of -11.7 dB, which is better than the desired minimum return loss for the LO port.

Figure 4-2. LO matching network.

Because the LO matching is already acceptable with only the LO trap, the principal aim of the LO matching network is to provide a DC ground return path, without disturbing the inherent LO impedance. To do so, the LO matching network incorporates a quarter-wave short-circuit stub. At DC, the via hole at the end of the quarter-wave LO stub provides a low-resistance path to ground. At the LO frequency, the stub is seen as an open circuit, which does not affect the LO impedance, Z_{LO} , at that point. Both objectives

of the LO matching network are thereby met. Figure 4-2 shows an updated version of the upconverter block diagram.

The inclusion of an LO matching network at the emitter port does have a detrimental effect on the gain of the HBT. In section 4.3.2, the simulated small-signal S21 parameter for the HBT, when the HBT is attached to the LO matching network and to the LO trap on the side, is less than the S21 parameter for the case of a stand-alone common emitter HBT. The LO matching network is seen by the emitter, at 3.6 GHz, to be an inductive impedance. Such an inductive load between the emitter and ground adversely affects the conversion gain of the bipolar transistor [Yip92a].

4.3.2 Conversion gain estimate

The goal of section 4.3.2 is to get an estimate of the upper bound of the conversion gain (equation 3-9). As stated in chapter 3, the estimate provided by equation 3-9 is an upper bound for two reasons. First, equation 3-9 does not take into account the losses in the IF and RF matching networks. Second, equation 3-9 is based on the assumption that the HBT's instantaneous bias point switches back and forth in time, between the cut-off and active regions, like an ideal square wave. It is worthwhile to use equation 3-9, however, to determine whether the emitter-injected single-ended HBT mixer configuration has enough potential conversion gain to be considered as a viable upconverter configuration at Ka band.

The parameters in equation 3-9 are derived in the following manner. The small-signal S-parameters are simulated on Libra for the case when the HBT is connected to the LO matching network at the emitter port, the LO trap connected at the base port, and the LO port is terminated by a 50-ohm load. See appendix D for a view of the Libra test circuit and test benches necessary to obtain the simulated S-parameters. Figure 4-3 shows

the magnitude of $S_{21\text{max}}$ and $S_{21\text{min}}$ from 1 to 40 GHz, corresponding to the active region ($V_{ce} = 2$ volts; $V_{be} = 1.385$ volts) and the cut-off region ($V_{ce} = 2$ volts; $V_{be} = 0$ volts). $V_{be} = 1.385$ volts is the maximum V_{be} seen during a load-line simulation of the HBT when it is connected to the LO matching network, the IF port LO trap, and 50-ohm loads at the IF and RF ports (this load-line simulation uses a test bench similar to that of appendix B). As stated earlier, the LO power for the load-line simulation is set at -3 dBm. Finally, we obtain the simulated RF impedance looking towards the collector, $Z_{RF} = 10.63 - j12.3$, ohms, with $|\Gamma_{RF}| = 0.667$ (with the LO trap on the IF side and LO matching network attached to the HBT). At the IF port, looking towards the base of the HBT (and again including the LO trap on the IF side and the LO matching network), we see a large IF mismatch, where $|\Gamma_{IF}| = 0.962$.

Figure 4-3. Magnitude of S_{21} when the HBT is biased in the active region (line with marker M1) and cutoff region (line with marker M2).

Given the relatively good match on the RF side, $|\Gamma_{RF}| = 0.667$, we assume that it is possible to effectively match the RF match impedance to 50 ohms over the entire 500-MHz bandwidth. On the other hand, the IF side is severely mismatched, with $|\Gamma_{IF}| = 0.962$. Strictly speaking, the conversion gain estimate should take into account Fano's limit, where Fano's limit is the best possible reflection coefficient that is

achievable, taking into account the impedance mismatch and the desired bandwidth [Gonza84]. Nevertheless, what is calculated here is only an upper bound estimate of the conversion gain that provides an insight into the conversion gain potential of the emitter-injected single-ended HBT upconverter. Hence, it is reasonable to set M_{IF} (and M_{RF}) to 1 and find out what that potential is.

The parameters needed to calculate an upper-bound estimate of the conversion gain, G_c , are: $S_{21max} = 4.61$, $S_{21min} = 0.05$, $R_{RF} = 10.63$ ohms, $M_{RF} = 1$, and $M_{IF} = 1$. Substituting all these values into equation 3-9 yields $G_c = 9.9$ dB. This estimate for G_c seems unrealistically high. However, this is because it does not take into account losses in the IF and RF matching networks, is based on the assumption that both the IF and RF ports are perfectly matched, and assumes an HBT that ideally switches between the cut-off and active regions. Nevertheless, this estimate for G_c lets the designer know that the emitter-injected single-ended HBT mixer configuration has enough potential conversion gain to be considered as a viable upconverter configuration at Ka band.

4.3.3 Collector bias network

The goal of the collector bias network is to apply a bias voltage at the collector port of the HBT. In doing so, however, RF leakage into the collector bias network must be prevented so as not to adversely affect the conversion gain. Also, the collector bias network must attenuate low-frequency signals as much as possible. This prevents the low-frequency signals from leaking into the collector bias network and feeding back into the base bias network, perhaps causing oscillations (the low-frequency signals are either generated by the noise present in the mixer or can be spurious signals such as f_{IF}).

To achieve an RF block at the collector bias network, an RF quarter-wave short-circuit stub is connected to the transmission line coming out of the collector port of the

HBT, as seen in Figure 4-4. A 0.26-picofarad MIM capacitor is connected in series with the via hole of the quarter-wave stub in order to block the DC current. The capacitor value is chosen so as to tune out, at 30.1 GHz, the via hole's inherent inductance. This RF short-circuit becomes an RF open-circuit a quarter-wave later, at the location where the stub joins the transmission line.

An 8-pF MIM capacitor is also connected at the point where the DC line attaches. At low frequencies (less than a few gigahertz), this larger capacitor acts as a short-circuit by creating an alternative path for low-frequency signals, thereby preventing substantial leakage into the DC bias network.

Figure 4-4. Collector bias network.

4.3.4 IF matching network

It is important to design the IF matching network before the RF matching network. The IF signal, with a wavelength just over eight times larger than the RF wavelength, will tend to "see" the stubs of the RF matching network as open-circuit stubs, having small electrical lengths. For example, an ideal LO trap at the RF matching network with a stub length of $0.25\lambda_{LO}$ equals a stub length of only $0.034\lambda_{IF}$ at IF. This is essentially an open-circuit stub of small electrical length connected in shunt to the RF

output transmission line and, from the IF point of view, has a negligible effect on Z_{IF} . Conversely, the RF will be very sensitive to small variation in the IF matching network. Hence, it is preferable to design the IF matching network before the RF matching network.

The goal of the IF matching network, then, is to conjugately match the IF impedance, looking towards the base of the HBT (before the LO trap), to 50 ohms, or similarly, match the IF admittance to 0.02 mho. As seen in section 4.3.2, there is a large capacitive IF mismatch when looking towards the base port of the HBT, where $|\Gamma_{IF}| = 0.962$. The reason the IF matching network is placed before the base bias network, relative to the HBT, is because the IF impedance is relatively large, compared to 50 ohms. Indeed, for the base bias network to be effective (i.e., to block the IF signals from going in the bias network), it has to present an impedance considerably larger than the IF impedance looking towards the HBT. Another reason to have a large impedance looking into the base bias network is that changes of the base bias network impedance, attributable to process variations or modelling discrepancies, will not have a significant effect on the IF impedance looking towards the base of the HBT. Hence, since the unmatched IF impedance is relatively large ($Z_{IF} = 14 - j169.41$), it is easier to insert the base bias network after the IF matching network; once the IF impedance is matched 50 ohms or so. In other words, the impedance (at IF), looking into the base bias network, has to be many times higher than the impedance looking towards the base of the HBT, Z_{IF} , in order to block the IF signal without affecting Z_{IF} . Also, recall that because of the maximum allowable MMIC size, lumped-element components must be used on the IF side of the upconverter. Since the goal of the base bias network is to present an impedance many times larger than the IF impedance (looking towards the base), then less inductors are required to do so when the IF impedance (looking towards the base) is 50 ohms than when it is about 122 ohms. A consequence of matching Z_{IF} to 50 ohms and

then connecting the base bias network is that the IF matching cannot contain a series capacitor or a shunt branch containing only spiral inductors or thin film resistors. Otherwise, the DC bias current will not reach the base of the HBT.

The large IF mismatch, looking towards the base of the HBT, can be explained by the fact that the HBT's bias point is located in the cut-off region, when the device is basically an open-circuit. The IF impedance, $Z_{IF} = 170\angle-85.3^\circ$, is capacitive because of the inherent depletion region capacitance at cut-off. As seen in section 4.3.2, Fano's limit means that the best possible return loss achievable at the IF port, from 3.35 GHz to 3.85 GHz, is -11 dB. Fano's limit is unattainable because it assumes a return loss of -11 dB within the frequency band and a return loss of 0 dB outside. Of course, such a sharp distinction at the band edge is unrealizable. In reality, to get a return loss of at least -10 dB from 3.35 GHz to 3.85 GHz for such a mismatch, with reactive lumped-element components only, would invariably necessitate an elaborate matching network design. Because of the severe IF mismatch, a purely reactive lumped-element matching network would require too many stages and would be too large for the amount of MMIC area available. A solution, to get a return loss of at least -10 dB for the entire IF band, is to use a resistive element in the IF matching network. Recall from section 4.3.2 that the upconverter has a relatively high potential conversion gain (9.9 dB). Trading some conversion gain for more input bandwidth is a reasonable compromise when the device has considerable conversion gain potential.

The IF matching network is shown in Figure 4-5. It is composed of two inductors in series and a shunt capacitor-resistor branch. The inductors have a combined inductance of about 8 nH and a total resistance of 11 ohms. Two spiral inductors are used because their respective self-resonant frequencies are greater than 10 GHz, while a single 8-nH spiral inductor has a self-resonant frequency dangerously close to 3.6 GHz. The shunt capacitive-resistive branch is made up of a 0.6 pF MIM capacitor and a 30-ohm

NiCr thin film resistor. The 30-ohm resistor, chosen empirically from simulations, sufficiently broadens the IF match to the entire 500-MHz bandwidth.

Figure 4-5. IF matching network.

Figure 4-6 is a snapshot showing the approximate effect of each component (starting with the lossy spiral inductors and ending with the shunt resistor) of the IF matching network on the unmatched IF impedance (at 3.6 GHz), $Z_{IF} = 14 - j169$ ohms. The unmatched IF impedance is located near the edge, in the capacitive hemisphere of the Smith Chart (point A). From there, Z_{IF} is brought to the inductive hemisphere of the Smith Chart by the lossy series inductors (point B). The shunt capacitor-resistor brings the IF impedance close to 50 ohms (points C and D). The IF impedance, at 3.6 GHz, is not brought to exactly 50 ohms because it is critical to match the entire 500 MHz bandwidth. Hence, as shown in Figure 4-7, the simulated matched IF impedance, for the entire 500 MHz bandwidth, resides in the vicinity of the center of the Smith Chart so as to provide the best possible match over the entire frequency band.

The IF matching network, with its shunt thin film resistor and lossy spiral inductors, introduces approximately 3 dB of loss (this value is obtained by looking at the difference between the simulated conversion gain of the mixer that has the IF matching network described above and the simulated conversion gain of the mixer that has an IF

matching network, yielding the same return loss, but composed of lossless reactive lumped-elements).

Figure 4-6. Effect of the IF matching network.

Figure 4-7. Simulated IF matching from 3.35 GHz to 3.85 GHz.

4.3.5 Base bias network

The base bias network must block the IF signal in order to prevent the IF from leaking into the base bias network. Therefore, it must present a large enough impedance (at 3.6 GHz) at the junction so as to achieve that and, at the same time, not improperly load the IF impedance looking towards the HBT. Figure 4-8 shows the base bias

network. It is composed of two identical lumped-element inductors of value 5.3 nH, with a self-resonating frequencies, f_{res} , equal to 9.0 GHz. They are connected in series by a 10- μm wide microstrip line, having a characteristic impedance of 91 ohms.

Figure 4-8. Base bias network.

The resulting simulated impedance (at 3.6 GHz), looking into the base bias network, is approximately 7 times larger than the matched IF impedance. Therefore, variations in the base bias impedance, due to process variations or modelling inaccuracies, will not significantly affect the matched IF impedance. In addition to the series inductors, a 10-pF MIM capacitor is connected in parallel to the base bias line. As with the collector bias network, this large MIM capacitor attenuates low frequency tones generated by noise processes in the mixer, thereby minimizing the chance that low frequency oscillations will occur.

4.3.6 RF matching network

The first goal of the RF matching network is to match the RF impedance, looking towards the collector of the HBT, to 50 ohms. The second goal of the RF matching network is to block the LO signal from exiting out from the RF port, thus realizing the upconverter's specified LO-RF isolation of at least 25 dB.

The Smith Chart of Figure 4-9 shows, approximately, how the RF matching network brings the unmatched RF load, Y_{RF} , to 50 ohms (note that it is easier to show the effect of stubs in admittance coordinates). Figure 4-10 shows the RF matching network.

The Smith Chart of Figure 4-9 shows, approximately, how the RF matching network brings the unmatched RF load, Y_{RF} , to 50 ohms (note that it is easier to show the effect of stubs in admittance coordinates). Figure 4-10 shows the RF matching network.

Figure 4-10. RF matching network.

4.3.7 MMIC layout

Following the upconverter design phase is the actual MMIC layout. Figure 4-11 shows the MMIC layout. On the IF side, a via hole is shared between the IF input matching network and the bottom input pad. Similarly, the base bias network's shunt capacitor shares a via hole with the upper input pad. This is done to reduce the number of via holes used in the design. Reducing the number of via holes is recommended, as vias are one of the lowest yield components on the MMIC. However, this is only done at low microwave frequencies, such as on the IF side. At higher microwave frequencies, such as at LO or RF, the via's inductance becomes significant enough to couple signals between the circuit branches that share the via.

The final circuit design occupies most of the allocated wafer space (2.5 mm x 3.7 mm). In the available space to the left of the LO matching network, a spiral inductor of value $L = 3.5$ nH and $R = 4.8$ ohms is laid out. This is one of the inductors used in the IF matching network. It is placed there for on-wafer measurement, in order to compare the spiral inductor's measured response with Nortel's S-parameter EM files. As the IF matching network is a critical part of the MMIC upconverter, it is useful to do an on-

wafer measurement of an inductor that is part of the IF matching network, in order to check for reliability and accuracy.

Figure 4-11. Upconverter MMIC layout.

4.3.8 EM simulations

An electromagnetic (EM) simulation is an important tool for MMIC applications at Ka band, especially for the sections of the circuit involving transmission line discontinuities because Libra uses lumped-element models for these discontinuities. These lumped-element models are relatively accurate at low microwave frequencies, but, at high microwave frequencies, simulations using the lumped element models start to diverge from reality. For example, the open-end of a stub is described by Edwards [Edwar81] as producing radiation and surface-wave propagation. A shunt admittance at the end of the open-circuit microstrip stub can model that effect. However, at K band and

above, it is advisable to do an EM simulation because lumped-element models are increasingly inaccurate.

S-parameter files (from measurements done at Nortel) are already used for the spiral inductors, while MIM capacitors, resistors and vias are considered to be accurately modeled by their lumped-element models [Nortel96]. EM simulations are carried out for the LO trap on the IF side and for the RF matching network. EM simulations are performed on these parts of the circuit because they directly affect critical specifications (LO-IF isolation, LO-RF isolation and the RF match) that operate at high microwave frequencies. The EM simulations were done using HP-EEsof's E-Sync EM simulator and HP-EEsof's Momentum EM simulator.

From the layout of Figure 4-11, we see that the IF side's LO trap has three discontinuities: a mitered corner, an open end, and the junction with the transmission line. The LO trap, connected to a piece of transmission line, is run through Libra's EM simulator. It has to be shortened by 13 microns for the EM results to agree with the Libra's linear test bench simulation. Figure 4-12 shows the magnitude of S_{11} and S_{22} , electromagnetically simulated from 24 to 29 GHz, superimposed on the linear simulation results, once the modification is done. For Figure 4-12, the EM simulation is carried out when the LO trap is connected in the middle of a 200- μm piece of transmission line. This is done in order to take into account the junction between the stub and the transmission line. As a result, the simulation results are for a 2-port network.

Similarly, for the RF matching network, the LO trap had to be trimmed by 25 μm and the distance between both stubs had to be shortened by 6 μm for the EM simulator's S-parameters to approximately correspond to the S-parameters of the linear simulation.

Figure 4-12. EM and linear simulations of the LO trap (located on the IF side).

Once the EM simulation of the critical distributed elements is complete, the MMIC layout is put through a design rule check (DRC), which ensures that the circuit does not violate any fabrication rules. In addition, the DRC verifies that the layout of components obeys certain guidelines. These guidelines ensure that the electrical characteristics of the MMIC's components are within the foundry's stated specifications. If the layout contravenes any of the rules or guidelines, it has to be modified accordingly and re-checked until there are no more DRC errors.

4.3.9 Some simulation results

Table 4-2 presents some simulation results in order to show the reader the *expected* results from the fabricated MMIC GaAs HBT upconverter. In the next section, the measured results are compared with the simulation results in more detail.

The most interesting result in Table 4-2 is that 2.7 dB of conversion gain is expected with only -3 dBm of LO power. Recall from section 4.3.2 that the estimated upper-bound of the conversion gain was 9.9 dB and that the IF matching network

described in section 4.3.4 introduces approximately 3 dB of IF loss. This brings the upper-bound estimate for G_c down to 6.9 dB, but still 4.2 dB above the simulated G_c . This 4.2 dB difference between the estimated and simulated conversion gain can be attributed to the fact that HBT does not ideally switch between cut-off and the active region.

Table 4-2. Some simulation results.

IF return loss peak IF return loss	< -11 dB from 3.35 GHz to 3.85 GHz -19 dB at 3.6 GHz
RF return loss peak RF return loss	< -19 dB from 29.85 GHz to 30.35 GHz -25 dB at 30.3 GHz
LO return loss (with LO and IF signals) LO return loss (with LO signal only)	-10.5 dB at 26.5 GHz -7.7 dB at 26.5 GHz
Conversion gain	2.7 dB at 30.1 GHz
LO-RF isolation	35 dB
LO-IF isolation	65 dB
LO power required V_{cc} V_{be}	-3 dBm 2 volts 1.1 volts

4.4 Measured and simulated results of the MMIC upconverter

The goal of this section is to characterize the HBT upconverter in terms of conversion loss, return loss, IP3, isolation, and suppression. In this section, the measured data is presented in conjunction with the simulated measurements.

The upconverter was tested on wafer using a probing station. On-wafer probing of the MMIC upconverter eliminates the need to mount the MMIC on a jig with coaxial cable inputs (a jig with coaxial cable inputs introduces uncertainties during the measurements because of the added bond wires, coaxial-to-microstrip connectors, and

extra line lengths of the microstrip carriers). On-wafer testing of the upconverter is possible because the LO is emitter-injected, thereby avoiding the need to connect the upconverter to an off-chip diplexer. The probing station is configured so that a ground-signal-ground probe can touch down on each of the MMIC's IF, LO, and RF signal launch-pads (left, bottom and right of Figure 4-11). On the top side of the MMIC shown in Figure 4-11, DC probes touch down on the corresponding V_b and V_c pads.

All the testing was carried out at the CRC. Table 4-3 is a list of the equipment used during the measurements. Diagrams depicting the test set-up for each measurement are shown in appendix E. The frequency counters, the digital multimeter, and the power meter listed in Table 4-3 are omitted from the test set-up diagrams of appendix E in order not to unnecessarily clutter the diagrams. Frequency counters were used to verify the frequency of the sweep oscillators throughout the measurements. The power meter was sometimes used to validate the output power level from the sweep oscillators. The digital multimeters were continuously used during the measurements in order to adjust the V_{be} and V_{ce} to the proper levels and to monitor I_c and I_b .

Table 4-3. Test equipment used.

Cascade Microtek Probing Station model 1032 with Bausch&Lomb microscope
HP8510 Network Analyzer (DC to 40 GHz)
HP8350B Sweep Oscillator with HP83570A RF plug-in (18 GHz to 26.5 GHz)
HP8350A Sweep Oscillator with HP83594A RF plug-in (2 GHz to 26.5 GHz)
HP8620A Sweep Oscillator with HP8612A RF section (DC to 12 GHz)
HP8564E Spectrum Analyzer (30 Hz to 40 GHz)
HP8563A Spectrum Analyzer (DC to 26.5 GHz)
EIP 578 Source Locking Microwave Counter
HP5343A Microwave Frequency Counter
Anatek DC source
Fluke 8010A Digital Multimeters (4)
HP437B Power Meter
HP7550A Graphics Plotter

At the start of the testing, two noteworthy problems were encountered. The first problem was that the device seemed to oscillate, even though during the design phase, the small-signal stability of the upconverter was verified using Libra's two-port small-signal stability simulation capability. The small signal stability Libra simulation showed that the completed upconverter is unconditionally stable from DC to 100 GHz. This is a predictable result, considering that the IF matching network does contain a thin film resistor and lossy spiral inductors (see [Gonza84] for a discussion on how resistors can increase the stability of a circuit). Nevertheless, oscillations were observed. Symptoms of oscillations are jittery read-outs of the DC digital multimeters monitoring I_c , I_b , V_{ce} and V_{be} , and short-lived spikes appearing at low frequencies on the display of the spectrum analyzer connected at the RF port of the upconverter.

The oscillations were eliminated by changing the standard DC probes to probes that have a set of de-coupling capacitors (120 pF) very close to the probe tips and another set of de-coupling capacitors (0.1 μ F) halfway between the probe tips and the DC probe's input connector. This discrepancy between the unconditional stability verdict of the small signal stability simulation of the upconverter and the oscillations observed at the start of the testing phase is probably due to the fact that the simulations do not take into account the inductive coupling that occurs in the off-chip bias network, especially at low frequencies. The 10-pF and 8-pF shunt MIM capacitors, placed in the MMIC's base bias network and collector bias network respectively, proved insufficient to fully reflect the low frequency signals. Note that when mounting different MMICs together to construct a transceiver circuit, it is common practice to use large de-coupling capacitors between the bias and the MMIC. Hence, using DC probes with large de-coupling capacitors for the on-wafer measurements is realistic.

The second problem is, presumably, over-heating. More than one mixer was lost before it was realized that the $V_{ce}/V_{be}/P_{LO}$ combination used during the simulation

caused the circuit to malfunction. The malfunctions became apparent when I_b suddenly climbed to a value above 2 mA and I_c abruptly rose to a value greater than 20 mA. After each of these occurrences, the mixer never worked again. This symptom concords with the thermal runaway afflicting the HBT when it is biased with a constant V_{be} as described in section 2.3. Load-line simulations showed that for the $V_{ce}/V_{be}/P_{LO}$ combination used in the simulation, the DC I_c would be approximately 2 mA, and that the time-varying I_c would stay under the thermal limit. During the initial phase of the measurements, it was found that keeping I_c below 1.7 mA DC ensured that the upconverter did not malfunction. Note that the HBT is biased in the cut-off region, yet a current of 1.7 mA is measured by the digital multimeter. This is because the IF and LO signals mixing in the HBT produce DC components that result in measurable base and collector currents, I_c and I_b . The magnitude of I_c and I_b depend on the amount of LO and IF power injected in the mixer and on the magnitude of V_{be} (typically, I_b was thirty to forty times less than I_c). Monitoring I_c to keep its magnitude under the prescribed safe limit of 1.7 mA DC did prevent further malfunctions of the MMIC upconverter.

After some experimentation, two alternative $V_{ce}/V_{be}/P_{LO}$ combinations that keep I_c below 1.7 mA were selected. The first combination is $V_{ce} = 2$ volts, $V_{be} = 0.989$ volts, and $P_{LO} = -2$ dBm. This $V_{ce}/V_{be}/P_{LO}$ combination was chosen because it uses the same V_{ce} as the $V_{ce}/V_{be}/P_{LO}$ combination chosen in the design section with V_{be} and P_{LO} are only marginally different. The second combination had $V_{ce} = 1.25$ volts, $V_{be} = 1$ volt, and $P_{LO} = -2$ dBm and was chosen because by lowering V_{ce} to 1.25 volts, V_{be} can be increased slightly to a value closer to the V_{be} of 1.1 volts initially specified in the design phase. Note that the lower the V_{ce} , the lower the DC power flowing through the HBT.

Table 4-4 summarizes the three $V_{ce}/V_{be}/P_{LO}$ combinations used for the upconverter: one for the design phase and two for the testing phase. Throughout section

4.4, the design $V_{ce}/V_{be}/P_{LO}$ combination is referred to as c0 and the two testing $V_{ce}/V_{be}/P_{LO}$ combinations are referred to as c1 and c2 respectively. Combinations c1 and c2 were used in nearly all the measurements.

Table 4-4. $V_{ce}/V_{be}/P_{LO}$ combinations used in the design and measurements.

	V_{ce} (volt)	V_{be} (volt)	P_{LO} (dBm)
Design $V_{ce}/V_{be}/P_{LO}$ combination: c0	2	1.1	-3
First testing $V_{ce}/V_{be}/P_{LO}$ combination: c1	2	0.989	-2
Second testing $V_{ce}/V_{be}/P_{LO}$ combination: c2	1.25	1	-2

4.4.1 Return loss

Figure 4-13. IF return loss: simulated and measured.

The return loss at the IF port was measured using the setup shown in Figure E-1 of appendix E. The measured and simulated IF return loss at each of the $V_{ce}/V_{be}/P_{LO}$ combinations are compared in Figure 4-13. The three curves that show data from 3.0

GHz to 4.2 GHz are the simulated IF return loss, while the curves spanning from 2 GHz to 5 GHz are the measured data. Note that the three simulated curves have minimum return losses at 3.6 GHz while the measured curves have minimum return losses at approximately 3.75 GHz. We see that the measured IF return loss is below -10 dB from 3.5 GHz to 4.2 GHz, while the simulated IF return loss covers the frequency band from 3.35 GHz to 3.9 GHz. The measured IF return loss is shifted by about 175 MHz, or 4.9 %, compared to the simulated IF return loss. Note that the return loss is relatively invariant with respect to changes in c0, c1 or c2.

Figure 4-14. Measured and simulated RF return loss.

The return loss at the RF port was measured using the set-up shown in Figure E-1. Figure 4-14 shows the measured RF return loss and the simulated RF return loss with the c0, c1 and c2 combinations. Note that the measured RF return loss at c0 and c1 is unsteady compared to the measured RF return loss at c2. The RF return loss at c0 and c1 was measured using a different calibration of the HP8510 network analyzer than the RF return loss at c2. It is assumed that the measured RF return losses at c0 and c1, showing

wobbliness, is a consequence of an imperfect calibration but these measurements still show the general trend of the RF return loss. The measured RF return loss peaks at around 31 GHz and provides a return loss of approximately -10 dB or better from about 30.75 GHz to 31.7 GHz. This is in contrast to the simulated RF return loss that provides a -10 dB return loss or better from 29.5 GHz to 30.6 GHz. The measured RF return loss has its minimum return loss shifted upwards by about 0.9 GHz, or +3 %, with respect to the specified center RF frequency of 30.1 GHz. Unfortunately, given the 1.7% bandwidth, this +3% shift considerably degrades the RF match within the RF frequency band. Moreover, as is apparent in Figure 4-14, the quality of the RF return loss is degraded. The minimum simulated RF return loss is around -25 dB but the minimum measured RF return loss is greater than -18 dB. The discrepancies between the simulated and measured RF and IF return losses are further discussed in section 4.5.

The LO return loss was measured using the set-up shown in Figure E-2 of appendix E. Note that the LO return loss was not measured with the network analyzer because the network analyzer can only provide -10 dBm of power at 26.5 GHz whereas the LO return loss measurement requires a large-signal LO tone to properly emulate the design conditions. The LO return loss is the difference between the incident input power at the LO port of the MMIC and the reflected LO power. Table 4-5 shows the LO return loss from 26.2 GHz to 26.67 GHz, where we see that the LO return loss at the c1 $V_{ce}/V_{be}/P_{LO}$ combination is -11 dB and the LO return loss for the c2 combination is -11.8 dB. The corresponding simulated LO return loss for either combinations is -10.5 dB at 26.5 GHz.

Table 4-5. LO return loss.

frequency (GHz)	26.2	26.3	26.4	26.5	26.6	26.67
measured return loss (dB) at c1	-8.8	-8.2	-9.5	-11	-12	-11.8
measured return loss (dB) at c2	-9	-8.5	-10.2	-11.8	N/A	-12.7

4.4.2 Conversion loss

The set-up used in this section is shown in Figure E-3 of appendix E. For this test, the IF input source is swept from 3.1 GHz to 4.3 GHz and the spectrum analyzer's scan rate is slowed down in order to view the entire output band of the upconverter. Figure 4-15 shows the measured RF output band from 29.6 GHz to 30.8 GHz for the c2 $V_{ce}/V_{be}/P_{LO}$ combination.

Figure 4-16 shows the simulated conversion gain of the upconverter at the c0, c1, and c2 $V_{ce}/V_{be}/P_{LO}$ combinations, versus the measured (negative) conversion gain, at the c1 and c2 $V_{ce}/V_{be}/P_{LO}$ combinations. The measured conversion gain curves of Figure 4-16 are adjusted data points taken from Figure 4-15 and another similar measurement. This is done by adding the cable loss and probe loss, between the RF port of the MMIC and the input of the spectrum analyzer, to the power levels recorded by the spectrum analyzer and shown in Figure 4-15.

Figure 4-15. Measured RF output spectrum (IF: 3.1 GHz to 4.3 GHz).

As seen from Figure 4-16, the simulated conversion gain, at all $V_{ce}/V_{be}/P_{LO}$ combinations, is higher than the measured conversion gain. The averaged simulated conversion gain, over the specified RF output frequency band, is approximately 2.5 dB at c0, 1.3 dB at c1, and 0.7 dB at c2. This is in contrast to the average measured conversion loss of 3.5 dB at c1 and c2 combination. The discrepancy between the simulated and measured results can be blamed on the poor match at the RF port. From Figure 4-10, we see that the RF return loss at 3.6 GHz is only 3 dB. This means that only half the RF energy coming out of the HBT is transferred to the output port. Nevertheless, as seen on Figure 4-16, the fact that the IF and RF match are shifted upwards in frequency does seem to even the conversion loss within the specified 29.85 GHz to 30.35 GHz output band, compared to the sloping conversion gain obtained during simulations. Note that an average of 3.5 dB of conversion gain, from 29.8 GHz to 30.4 GHz, is achieved with only -2 dBm of input LO power. This proves that the HBT mixes efficiently when biased in the cut-off region and when configured as emitter-injected single-ended upconverter. The frequency band from 29.8 GHz to 30.4 GHz is the frequency band where the conversion loss varies from a low of 2 dB to a high of 5 dB, thus varying by no more than 3 dB. Within that band, the average conversion loss is approximately 3.5 dB.

Figure 4-16. Simulated (c0, c1, and c2) and measured (c1 and c2) conversion gain.

4.4.3 Output spectrum

The output spectrum of the upconverter (at the RF port) was measured using the test set-up shown in Figure E-3 of appendix E. The spectrum shown in Figure 4-17 is measured at the c1 $V_{CC}/V_{be}/P_{LO}$ combination. All the output power levels of Figure 4-17 must be adjusted upwards by an amount corresponding to the frequency-dependent losses in the coaxial cable connecting the spectrum analyzer to the probe and the frequency-dependent losses in the output ground-signal-ground probe.

As is seen from Figure 4-17, most spurious signals in the DC to 40 GHz band are suppressed by at least 20 dB, except for the following spurious tones: f_{IF} , $2xf_{IF}$, $f_{LO} - f_{IF}$, f_{LO} , and $f_{LO} + 2xf_{IF}$. In Table 4-6, the simulated power levels of these spurious and RF tones (at c0 and c1) are compared to the measured power levels at c1. From Table 4-6, we see that the simulated RF to LO suppression, at c1, is 28.8 dB. This does not compare well to the measured RF to LO suppression of 5.3 dB. Given that the LO input power is -2 dBm, this means a measured LO-RF isolation of 16 dB, down by 19.5 dB from the LO-RF isolation simulated at the c1 $V_{CC}/V_{be}/P_{LO}$ combination. This large difference between the measured and simulated LO-RF isolation is likely due to the LO trap of the RF matching network being slightly offset in frequency.

The simulation results also predict better RF to spurious signal suppression. The measured output spectrum shows that the upconverter does not sufficiently suppress certain spurious tones. Suppression at the RF port is exacerbated by the fact that the measured RF output power, for an IF power of -10 dBm, was less than simulated. The upwards shift in frequency of the output match might also inadvertently transfer more power from the $f_{LO} + 2xf_{IF}$ spurious signal to the output load.

Figure 4-17. Output spectrum for the c1 $V_{ce}/V_{be}/P_{LO}$ combination.

Table 4-6. Simulated and measured output power levels of selected spurious signals at c1.

Input power $P_{IF} = -10$ dBm LO power $P_{LO} = -2$ dBm	c1 power levels (dBm) simulated	c1 adjusted power levels (dBm) measured
$f_{IF} = 3.6$ GHz	-22	-21.6
$2 \times f_{IF} = 7.2$ GHz	-22	-21.3
$f_{LO} - f_{IF} = 22.9$ GHz	-21.5	-18.9
$f_{LO} = 26.5$ GHz	-37.5	-18
$f_{RF} = 30.1$ GHz	-8.7	-12.7
$f_{LO} + 2 \times f_{IF} = 33.7$ GHz	-22.7	-17

4.4.4 The 1 dB compression point and the IP3

Two MMIC upconverters were damaged by trying to directly measure the 1 dB compression point. Most probably, the HBT over-heated as the IF input power was increased to over -8 dBm (and I_C increased to over 2 mA). Undoubtedly, the actual 1 dB compression point is between -8 dBm and -2 dBm. The simulated 1-dB compression

points, for the c1 and c2 $V_{ce}/V_{be}/P_{LO}$ combinations, are shown in Figure 4-18. The simulated 1-dB compression point for the c1 $V_{ce}/V_{be}/P_{LO}$ combination is approximately -5 dBm while the simulated 1-dB compression point for the c2 $V_{ce}/V_{be}/P_{LO}$ combination is about -7 dBm.

Figure 4-18. Simulated conversion gain G_c (at c1 and c2) versus IF input power.

Even if the 1 dB compression point cannot be determined exactly, the IP3 can. This is because the IP3 is found by the intersection of the extrapolated curve of the fundamental response of the mixer and the extrapolated curve of the two-tone third order intermodulation product response. The two-tone intermodulation test set-up is shown in Figure E-4. Figure 4-19 shows the measured IP3, for the c1 combination, to be 6 dBm. Figure 4-20 shows the measured IP3, for the c2 combination, to be -2 dBm. The IP3, for any of the $V_{ce}/V_{be}/P_{LO}$ combinations, could not be simulated because the computer platform used did not have enough memory to perform a harmonic balance simulation when two input tones and one LO tone are stipulated. For Figure 4-19, the measured two-tone third order intermodulation product is $f_{LO} + (2f_{IF2} - f_{IF1})$, where $f_{IF1} = 3.5$ GHz and $f_{IF2} = 3.7$ GHz. For Figure 4-20, the measured two-tone third order

intermodulation product is $f_{LO} + (2f_{IF2} - f_{IF1})$, where $f_{IF1} = 3.55$ GHz and $f_{IF2} = 3.65$ GHz.

Figure 4-19. Measured IP3 (c1 $V_{ce}/V_{be}/P_{LO}$ combination).

Figure 4-20. Measured IP3 (c2 $V_{ce}/V_{be}/P_{LO}$ combination).

As expected, the IP3 for Figure 4-20 is lower than the IP3 for Figure 4-19 (- 2 dBm compared to 2 dBm). The IP3 is a direct measure of the third order intermodulation distortion of two closely spaced inputs. It is also a measure of the strength of the third order response of a non-linear I-V relationship. The higher the IP3, the weaker the third order component of a mixer's inherent non-linear relationship is, and the less problematic the mixer's third order intermodulation products become. As is seen in Figure 3-3c, I_C is exponentially dependent on V_{be} . For a given P_{LO} (and a V_{ce} greater than 1 volt), we see that the I_C - V_{be} relationship gradually becomes more non-linear as V_{be} is increased from zero to a value just below the base-emitter turn-on voltage. As the I_C - V_{be} relationship becomes more non-linear, the third order component of a power series characterizing the I_C - V_{be} non-linear relationship will become more significant. For the c2 $V_{ce}/V_{be}/P_{LO}$ combination, V_{be} is set at 1 volt and that makes the I_C - V_{be} relationship more non-linear than the I_C - V_{be} non-linear relationship at the c1 $V_{ce}/V_{be}/P_{LO}$ combination, where V_{be} is set at 0.989 volts. Therefore, as expected, the IP3 for the c2 $V_{ce}/V_{be}/P_{LO}$ combination is lower than the IP3 for the c1 $V_{ce}/V_{be}/P_{LO}$ combination.

4.4.5 Conversion gain versus LO power

Another useful test to characterize a mixer is to find out at what LO power the conversion loss is minimized. For this upconverter, this is done for the specified IF input power of -10 dBm, the IF set at 3.6 GHz, and the LO at 26.5 GHz. This test was performed using the test set-up shown in Figure E-3 of appendix E. Figure 4-21 shows the simulated and measured progression of the conversion gain against LO power, for the c1 combination. Figure 4-22 shows the simulated conversion loss versus LO power for the c1 and c2 $V_{ce}/V_{be}/P_{LO}$ combinations. From the simulation results, we see that the optimum LO power, for the c1 and c2 $V_{ce}/V_{be}/P_{LO}$ combinations, is 2 dBm and -0.5 dBm respectively. The measured conversion loss curves versus LO power of Figure 4-21

extend up to $P_{LO} = -1.5$ dBm only. The LO power was not increased beyond -1.5 dBm in order to avoid increasing I_C beyond the 1.7 mA limit. Any further increase of the LO power might have pushed I_C beyond the empirically-determined I_C limit of 1.7 mA and damaged another MMIC upconverter.

Figure 4-21. Measured conversion gain versus LO power for c1 and c2.

Figure 4-22. Simulated conversion gain, G_c , (at c1 and c2) versus LO power.

The simulated curves of Figure 4-22 show that the LO power yielding the best possible conversion gain is 2 dBm for the c1 $V_{ce}/V_{be}/P_{LO}$ combination and -0.5 dBm for the c2 $V_{ce}/V_{be}/P_{LO}$ combination. This means that the LO power for the two $V_{ce}/V_{be}/P_{LO}$ combinations selected for the measurements can be considered to be optimum, because both the c1 and c2 $V_{ce}/V_{be}/P_{LO}$ combinations produce an I_c just under the prescribed thermal limit of 1.7 mA.

4.4.6 LO-IF isolation

The LO-IF isolation was measured using the test set-up shown in Figure E-5 of appendix E. The spectrum analyzer is connected to the IF port, the RF port is terminated by a 50-ohm load, and a -2 dBm LO tone at 26.6 GHz is incident at the LO port of the MMIC upconverter. Figure 4-23 shows the spectrum on the IF side where we see that the LO signal is more than 40 dB down from the LO power level of -2 dBm injected at the LO port of the mixer. Figure 4-23 shows the LO suppression for the c2 $V_{ce}/V_{be}/P_{LO}$ combination. The same result is observed for the c1 $V_{ce}/V_{be}/P_{LO}$ combination. The corresponding simulated LO-IF isolations, at the c0, c1, and c2 $V_{ce}/V_{be}/P_{LO}$ combinations, are all greater than 60 dB. The unilateral character of the HBT coupled with the LO trap on the IF side contributes to the good LO-IF isolation observed during measurements.

Figure 4-23. LO power level at the IF port.

4.5 Differences between the measured and simulated results IF and RF return losses

As seen in section 4.4, both the IF and RF return losses have their minimum return losses shifted upwards in frequency by 4.9% and 3% respectively. Recall from Figure 4-14 that the minimum RF return loss is shifted upwards in frequency by 3% but, given the 1.7% RF bandwidth, this leads to a significant mismatch in the specified RF frequency band of 29.85 GHz to 30.35 GHz. This RF mismatch downgrades the upconverter's expected conversion gain within the RF output band to a average conversion loss of about 3.5 dB. Note, however, that this seems to counteract the simulated conversion gain's downward slope in the output frequency band. From Figure 4-13, we see that the minimum IF return loss is shifted upwards in frequency by 4.9 %. Fortunately, given the 14% IF bandwidth, this shift does not unduly degrade the match within the IF frequency band.

According to Nortel [Harr96], the process control monitor or PCM (the Nortel verification test performed for each wafer fabricated) showed that HBT's base-emitter turn on voltage, transconductance, DC current gain, and DC I-V curves are within allowable specifications. The process control monitors also revealed that the 5-pF MIM test capacitors on the wafer was measured to be 4.9 pF, implying that the MMIC's MIM capacitors are fairly accurate. The only problem reported during the PCM is that the thin film resistors were found to be 10% lower than specified. This means that the 30-ohm thin film resistor inserted in the IF matching network is likely to be 27 ohms in reality.

Recall that one of the spiral inductors used in the IF matching circuit was inserted in the allocated MMIC layout area (see Figure 4-11) in order to test and compare it with the spiral inductor's S-parameters file supplied by Nortel. The spiral inductor was measured on wafer and the microstrip line and taper connecting both sides of the spiral inductor were de-embedded using Libra. At 3.6 GHz, the spiral inductor measured on the

MMIC has an inductance of 3.8 nH and resistance of 9.2 ohms while the Nortel-supplied S-parameter file has an inductance of 3.6 nH and 5.5 ohms. Although the spiral inductor is more lossy than predicted by the Nortel data, its inductance is very close to expected value. Unfortunately, inserting the measured data into the simulated upconverter shifts the minimum IF return loss downwards by about 0.1 GHz (and degrades the simulated conversion gain by about 1 dB). This is in contrast to the measured data which actually show a shift upward by 0.175 GHz, as is shown in Figure 4-13. The second spiral inductor in the IF matching could not be fitted on the allocated wafer reticule so it is not known if its inductance is above or below the expected value or if it is more lossy than expected. Decreasing the value of the thin film resistance by 10%, from 30 ohms to 27 ohms, does not significantly affect the simulated IF return loss or the simulated conversion gain. According to the PCM, the remaining major component of the IF matching network, the 0.6-pF MIM capacitor, is supposed to be accurate and cannot reasonably be considered to be the cause for the +4.9% frequency shift between the measured IF return loss and the simulated IF return loss.

From Figure 4-14, we see that the measured minimum RF return loss is approximately 10 dB greater than the simulated minimum RF return loss and, in addition, the measured minimum RF return loss appears to be shifted in frequency by +3% with respect to the simulated minimum RF return loss. To try to explain this discrepancy, the circuit design was re-checked in order to try to find a probable cause. EM simulations, using Sonnet⁵, were carried out on the RF matching network and parts of the LO matching network. The upconverter was re-simulated using those EM results but the resulting RF return loss was similar to the simulation results obtained during the design phase. A sensitivity analysis, where the value of components in the circuit were varied and the simulated effect on the RF return loss observed, did not conclusively explain why

⁵ Sonnet is a well-respected EM simulation program.

the measured RF return loss is shifted upwards in frequency by 3% or why the minimum return loss is not as good as predicted in the simulations.

During this re-examination of the circuit, however, it was found that the LO trap on the IF side of the upconverter had not been EM-simulated below and above the vicinity of the LO frequency. Figure 4-24 shows how the response of this quarter-wave open-circuit LO stub diverges in the IF and RF frequency bands.

Although the EM and linear simulations do match up nicely at LO, this is not the case at RF and IF. Figure 4-25 shows how re-simulating the upconverter using the EM response of the LO trap made the RF return loss worse compared to the original simulations that used linear elements for the LO trap. However, the result shown in Figure 4-25 still does not explain the +3% shift in frequency of the output match. Note that neither the frequency response nor the magnitude of the simulated IF return loss are significantly altered when the LO trap's linear elements are replaced by the EM response. The fact that a wideband EM simulation of the IF side's LO trap was not taken into account at the onset of the design phase is a design error.

Figure 4-24. Results of the EM simulation against the linear simulation for the LO trap on the IF side.

Furthermore, this extensive re-examination of the circuit highlighted a design flaw: an RF trap should have been inserted on the IF side of the upconverter and another RF trap should have been incorporated in the LO matching network. Adding these RF quarter-wave open-circuit stubs would create RF short-circuits in the IF and LO side of the MMIC upconverter. This would eliminate the dependence of the RF impedance, looking towards the collector, on process variations or on inaccurate modelling of elements on the IF and LO sides of the circuit.

Figure 4-25. Simulated RF return loss using linear and EM response of the LO trap on the IF side.

Unfortunately, the inaccurate modelling of the IF side's LO trap does not explain why the measured minimum IF and RF return loss are shifted upwards in frequency when compared with the simulated return losses; it only explains why the measured RF return loss is worse than the simulated RF return loss. Perhaps this discrepancy is a result of the large-signal behavior of the HBT on this wafer differing from the large-signal behavior predicted by the model. Note that a MMIC HBT oscillator, designed by the CRC and tested by staff at the CRC, was also on the same wafer as the MMIC upconverter. It was designed to oscillate at 38 GHz but ended up oscillating at 36.5 GHz. Differences

between simulated and measured results from the MMIC upconverter and oscillator on that particular wafer are not proof that there the model did not accurately predict the large-signal behavior of the HBT, but it does warrant further investigation.

4.6 Summary of the results and comparison with a PHEMT single-ended upconverter

The measured results for the GaAs HBT upconverter are summarized and presented in Table 4-7, together with the measured results of a GaAs PHEMT single-ended upconverter. The PHEMT upconverter was designed and tested by the CRC [Verver96]. It has the same system specifications as the HBT upconverter. The obvious difference between both mixers, besides the choice of transistor, is that the CRC upconverter requires a diplexer, because both the RF and LO signals use the drain port of the PHEMT. Therefore, the CRC PHEMT upconverter is a MHMIC consisting of a MMIC the size of the HBT upconverter attached to a large off-chip diplexer. The HBT upconverter does not require a diplexer but will most likely require an RF pass-band filter (BPF) in order to filter out the larger than expected undesired spurious signals. From the simulations, it was hoped that no output filter would be required, thereby making the HBT upconverter very compact compared to traditional single-ended mixers. Nevertheless, a filter is smaller than a diplexer, such that an MHMIC version of this thesis' single-ended upconverter is potentially more compact than the traditional single-ended upconverters requiring diplexers.

The principal advantages of the HBT single-ended upconverter are smaller size (even with an RF BPF), lower LO power required, and lower DC power consumption. The 1 dB compression point of the two upconverters both lie between -8 and -2 dBm. It is assumed that the IP3 of the HBT single-ended upconverter is comparable to the IP3 of the PHEMT single-ended upconverter. On the other hand, the conversion loss of the

HBT upconverter (without an RF BPF) is about 2 decibels higher than the conversion loss of CRC's PHEMT upconverter.

Table 4-7. Comparison between the HBT upconverter and CRC's PHEMT upconverter.

	GaAs PHEMT single-ended upconverter with an LO/Rf diplexer	GaAs HBT single-ended upconverter without an RF BPF
Conversion loss (dB)	1.5 dB (at $P_{IF} = -15$ dBm)	3.5 dB (at $P_{IF} = -10$ dBm)
1-dB compression point (IF power)	-4.5 dBm	between -8 dBm and -2 dBm
LO-RF isolation	40 dB	16 dB
LO-IF isolation	N/A	> 40 dB
IF-RF isolation	> 50 dB	11.6 dB
LSB suppression	> 40 dB	6.2 dB
IF port return loss	poor	< -10 dB
RF port return loss	< -10 dB	poor
LO port return loss	< -10 dB	< -10 dB
IP3	N/A	-2 dBm
LO power required	+5 dBm, at the mixer port and not the diplexer port	-2 dBm
bias	$V_g \approx 0.29$ volts; $V_d \approx 0.46$ volts; $I_d \approx 6.5$ mA $P_{DC} = V_d \cdot I_d \approx 3$ watts	$V_{ce} = 1.25$ volts; $V_{be} = 1$ volt; $I_c \approx 1.6$ mA $P_{DC} = V_{ce} \cdot I_c \approx 2$ watts

The principal disadvantage of the emitter-injected HBT single-ended upconverter is that it does not suppress spurious signals as well as the PHEMT single-ended configuration requiring a diplexer. Adding an RF BPF at the output port of the single-ended upconverter would most likely improve the RF to spurious signal suppression to a level comparable to CRC's PHEMT upconverter, but would unquestionably exacerbate

the conversion loss by an amount corresponding to the BPF's insertion loss. Another disadvantage of this particular HBT upconverter is that it seems more susceptible to detrimental self-heating effects. In fact, the thermal limitations of the HBT, biased as it is in the cut-off region, means that an HBT emitter-injected upconverter cannot be used at its optimum LO power level (where the optimum LO power level yields the lowest possible conversion loss for a given bias point).

CHAPTER 5: CONCLUSION

5.1 Summary

In recent years, we have seen a dramatic increase in the number of wireless communications systems being proposed. Future communication systems such as WLANs, LMCS, and broadband satellite communication will require large frequency bandwidths in order to support the high data rates required for data and video transmission. Due to frequency crowding at the lower end of the microwave spectrum, wireless transmission frequencies at the Ku, K, Ka band and above are being considered. The requirements of these new high-performance wireless systems, operating at high microwave frequencies and at millimeter-wave frequencies, means that new technologies are being assessed when older ones are inadequate or need improvement.

The GaAs HBT has emerged as a microwave transistor capable of rivaling the other GaAs FET-based microwave transistors: the MESFET, HEMT, and PHEMT. The HBT has proven to be especially successful when used in power amplifiers and in low phase noise oscillators. However, the HEMT and PHEMT, with lower noise figures, are better suited for the low thermal noise applications.

The goal of this thesis was investigate the HBT for use as an upconverter and to design and test a GaAs HBT MMIC upconverter. The HBT offers two distinct non-linear regions that can potentially be exploited to mix signals. The HBT is investigated using a large-signal model developed by Nortel, implemented in Libra by the CRC, and verified for accuracy by the CRC and Nortel. The model is designed to take into account the HBT's self-heating effect and emitter-barrier effect. The design of the upconverter is done using Libra, which is capable of carrying out harmonic balance circuit analysis.

Each non-linear region of the HBT is investigated by comparing the degree of non-linearity and its suitability with the single-ended emitter-injected upconverter configuration. The single-ended configuration is chosen in order to compare the HBT upconverter with a previously-built PHEMT single-ended upconverter. The LO tone is injected at the emitter port of the HBT in order to avoid using a diplexer and to make the upconverter more compact. The HBT is biased in the cut-off region to take advantage of I_C 's exponential dependence on V_{be} . A formula is derived that gives an upper-bound estimate of the conversion gain for an emitter-injected single-ended upconverter. This estimate provides the designer with an appreciation for the conversion gain potential available from the HBT and the possible trade-offs that can be done in exchange for more bandwidth and stability.

One disadvantage of this HBT is that unless monitored closely, it is susceptible to over-heat and malfunction. For this particular HBT and given that the IF input power was specified at -10 dBm, the $V_{ce}/V_{be}/P_{LO}$ combination must be judiciously chosen so as not to over-heat the HBT. The HBT's susceptibility to thermal runaway is the reason why the 1 dB compression could not be measured directly nor could the optimum LO power be determined.

At the beginning of the design phase, the upper bound of the conversion gain, for an HBT ideally switching between cut-off and the active region, was estimated at 9.9 dB. As a result of the high conversion gain potential of this emitter-injected single-ended HBT upconverter configuration, some loss is purposely added to the IF matching network in order to broaden the IF match over the entire 500 MHz. The simulation results shown in the results section predict 2.5 dB of conversion gain at 30.1 GHz, an LO-RF isolation of 35 dB, and an LO-IF isolation of more than 60 dB, with -3 dBm of incident LO power required. The actual measured results show an average conversion loss of 3.5 dB, with an LO-RF isolation of only 16 dB, and an LO-IF isolation greater than 40 dB, with an LO

power of -2 dBm. The IF return loss is below -10 dB at 3.6 GHz, but the RF return loss is poor at 30.1 GHz. The poor RF return loss is the principal reason why the HBT upconverter yielded conversion loss instead of the expected positive conversion gain. The suppression at the RF port is only 5 dB for certain spurious signals while the LO-RF isolation is 16 dB. This means that for practical use, this MMIC upconverter would have to be accompanied by a BPF centered on RF. The IP3 depends on the relationship between V_{be} and P_{LO} .

The HBT single-ended upconverter is particularly efficient, compared to the PHEMT upconverter, because it only needs -2 dBm of LO power and 2 mW of DC power for a conversion loss of 3.5 dB, compared with the PHEMT upconverter which requires 5 dBm of LO power and 3 mW of DC power for a conversion loss of 1.5 dB. Moreover, because the HBT upconverter has the LO tone injected at the emitter port of the HBT, it is more compact (even with an RF BPF) than a traditional single-ended mixer requiring a diplexer.

5.2 Future work

This GaAs HBT upconverter has the distinctness of being the first emitter-injected single-ended HBT upconverter. More work could be done on improving the design of this single-ended configuration. Specifically, an RF quarter-wave open-circuit stub should be inserted on the IF side of the circuit and another RF quarter-wave open-circuit stub should be incorporated in the LO matching network in order to desensitize the RF matching circuit to process variations and modelling inaccuracies on the IF and LO sides of the upconverter.

An alternative design is to use the emitter-injected concept in a single-balanced configuration based on the diagram shown in Figure A-1. The advantage of such a

balanced configuration is that no broadband hybrids, at the input or output ports, are needed. To implement this configuration in a monolithic form, it would be necessary to design a compact active IF balun at the input because a passive balun would invariably be very large for a MMIC. The LO balun could be made using a Wilkinson divider and a 180° delay in one of the two LO branches. The RF combiner could simply be a Wilkinson combiner. If the LO and RF frequencies are sufficiently high, then the LO balun and RF combiner can be made small enough to fit in a small area.

Another area of future work might be to verify, and possibly, improve the HBT model so that it is more accurate at high frequencies under large-signal conditions. This might involve performing large-signal measurements on the HBT using a simultaneous small-signal input and a large-signal LO tone.

5.3 Conclusion

In this thesis, the GaAs HBT was investigated for use as an upconverter and it was found that the HBT is well suited for use in an efficient upconverter design. The MMIC upconverter yields an average of 3.5 dB of conversion loss while requiring -2 dBm of LO power and 2 mW of DC power. Hence, by improving on the design outlined in this thesis, it is reasonable to attempt to integrate a GaAs HBT upconverter with an HBT power amplifier and oscillator onto a single chip.

REFERENCES

- [Adel90] Adelseck, B., Dieudonné, J.M., Schmiegner, K.E., "A Monolithic 94 GHz Balanced Mixer", *IEEE MTT-S International Microwave Symposium Digest*, pp. 193-196, 1990.
- [Ali91] Ali, F., Gupta, A., *HEMTs and HBTs: Devices, Fabrication, and Circuits*, Artech House, Boston, 1991.
- [Anholt95] Anholt, R., *Electrical and Thermal Characterization of MESFETs, HEMTs, and HBTs*, Artech House Inc., Norwood, Massachusetts, 1995.
- [Asbeck87] Asbeck, P.M., Chang, M.F., Wang, K-C, Miller, D.L., Sullivan, G.J., Sheng, N.H., Sovero, E., Higgins, J.A., "Heterojunction Bipolar Transistors for Microwave and Millimeter-wave Integrated Circuits", *IEEE Trans. on Electron Devices*, Vol. ED-34, No. 12, pp. 2571-2577, December 1987.
- [Asbeck89] Asbeck, P.M., Mau-Chung, F.C., Higgins, J.A., Sheng, N.H., Sullivan, G.J., Wang, K-C, "GaAlAs/GaAs Heterojunction Bipolar Transistors: Issues and Prospects for Application", *IEEE Trans. on Electron Devices*, Vol. 36, No. 10, pp. 2032-2041, October, 1989.
- [Asbeck93] Asbeck, P.M., Chang, F., Wang, K-C, Sullivan, G.J., Cheung, D.T., "GaAs-based Heterojunction Bipolar Transistors for Very High Performance Electronic Circuits", *Proceedings of the IEEE*, Vol. 81, No. 12, pp. 1709-1726, December 1993.
- [Bayrak93] Bayraktaroglu, B., "GaAs HBT's for Microwave Integrated Circuits", *Proceedings of the IEEE*, Vol. 81, No. 12, pp. 1762-1785, December 1993.
- [Bluy79] Bluysen, H.J.A., Van Ruyen, L.J., Williams, F., "Effects of Quantum Confinement and Compositional Grading on the Band Structure of Heterojunctions", *Solid-State Electronics*, Vol. 22, pp. 573-579, 1979.
- [Bura81] Bura, P., "70 MHz to 6 GHz FET Up-Converter", *Proceedings of the 11th European Microwave Conference*, pp. 215-218, 1981.
- [Bura89] Bura, P., Vassilakis, B., "A Balanced 11 GHz HEMT Up-Converter", *Proceedings of the 19th European Microwave Conference*, pp. 1299-1302, 1989.
- [Faber95] Faber, M.T., *Microwave and Millimeter-Wave Diode Frequency Multiplier*, Artech House, Norwood, Massachusetts, 1995.

- [Collin92] Collin, Robert E., *Foundations for Microwave Engineering*, McGraw-Hill, New York, 1992.
- [Combes87] Combes, P.F., Graffeuil, J., Sautereau, J.F., *Microwave components, devices, and active circuits*, John Wiley & Sons, Toronto, 1987.
- [Comsat89] COMSAT Laboratories, *COMSAT Gallium Arsenide MMIC Design Course Notes*, 29-31 March 1989.
- [Dieud95] Dieudonné, J.-M., Adelseck, B., Narozny, P., Dämbkes, H., "Advanced MMIC Components for Ka-Band Communications Systems. A survey.", *IEEE MTT-S International Microwave Symposium Digest*, pp. 409-412, 1995.
- [Edwar81] Edwards, T.C., *Foundations for Microstrip Circuit Design*, John Wiley & Sons, Chichester, United Kingdom, 1981.
- [Falco96] Falconer, D., "A System Architecture for Broadband Millimeter-Wave Access to an ATM LAN", *IEEE Personal Communications*, Vol. 3, No. 4, pp. 36-41, August 1996.
- [Freun96] Freundhofer, A.P., Jamani, Y., "A Ka-Band GaInP/GaAs HBT Double-Balanced Upconvert Mixer", *Proceedings of the 1996 Canadian Conference on Electrical and Computer Engineering*, May, 1996.
- [Gard95] Gardiner, J., West, B., *Personal Communication Systems and Technologies*, Artech House, Inc., Norwood, Massachusetts, 1995.
- [Gilbert95] Gilbert, B., Baines, R., "Fundamentals of Active Mixers", *Applied Microwave and Wireless*, pp. 10-27, Winter 1995.
- [Gilmore86] Gilmore, R., "Non-linear Circuit Design Using the Modified Harmonic Balance Algorithm", *IEEE Trans. on Microwave Theory and Techniques*, Vol. MTT-34, No. 12, pp. 1294-1307, December 1986.
- [Gonza84] Gonzalez, G., *Microwave Transistor Amplifiers, Analysis and Design*, Prentice-Hall Inc., Englewood Cliffs, New Jersey, 1984.
- [Goyal89] Goyal, R., *Monolithic Microwave Integrated Circuits*, Artech House, Norwood, Massachusetts, 1989.
- [Haigh89] Haigh, Everard J., *GaAs Technology: its impact on circuits and systems*, Peter Peregrinus Ltd., London, United Kingdom, 1989.
- [Harr96] Telephone conversation with Dr. Alan Harrison, Advanced Technology Laboratory, Advanced GaAs Devices, Nortel Technologies Inc, December 10, 1996.

- [Helsz78] Helszajn, J., *Passive and Active Microwave Circuits*, John Wiley & Sons, New York, 1978.
- [Hender81] Henderson, Bert C., "Mixers: Part I", Watkins-Johnson Company Tech Notes, Vol. 8, No. 2, pp. 2-14, March/April 1981.
- [Koizu95] Koizumi, H., Nagata, S., Tateoka, K., Kanazawa, K., Ueda, D., "A GaAs Single Balanced Mixer MMIC with Built-in Active Balun for Personal Communication Systems", *Proceedings of the IEEE Microwave and Millimeter-Wave Monolithic Circuits Symposium*, pp. 77-80, 1995.
- [Kroemer82] Kroemer, H., "Heterostructure Bipolar Transistors and Integrated Circuits", *Proceedings of the IEEE*, Vol. 70, No. 1, pp. 13-25, 1982.
- [Ladbr89] Ladbrooke, P.H., "MMIC Design: GaAs FETs and HEMTs", Artech House, Inc., Norwood, Massachusetts, 1989.
- [Laneve96] Laneve, T., Do-Ky, H., Watson, P., Stubbs, M.G., "Implementation of a Large-Signal HBT Model in Libra™", *Proceedings of the GAAS '96 conference*, session 6A2, 1996.
- [Loral] LORAL, Microwave Narda, Specification Sheet for models 4825, 4826 and 4827.
- [Maas88] Maas, Stephen A., *Non-linear Microwave Circuits*, Artech House, Inc., Norwood, Massachusetts, 1988.
- [Maas93] Maas, Stephen A., *Microwave mixers*, Artech House, Inc., Norwood, Massachusetts, 1993.
- [Madi95] Madhian, M., Desclos, L., Maruhashi, K., Onda, K., Kuruhara, M., "A Ka-Band Monolithic CPW Upconverter Utilizing a Source Mixing Concept", *Proceedings of the IEEE Microwave and Millimeter-Wave Monolithic Circuits Symposium*, pp. 163-166, 1995.
- [Maesel92] Maesel, M., "Mixers Made Simple", *Soquelec Telecommunications Newsletter*, Vol. 2, No. 3, pp. 2-3, Fall/Winter 1992.
- [Matt95] Matthes, M., Dieudonné, J.-M., Stiebler, W., Klapproth, L., "A PMHFET based MMIC Gate Mixer for Ka-Band Applications", *Proceedings of the IEEE Microwave and Millimeter-Wave Monolithic Circuits Symposium*, pp. 159-162, 1995.
- [Nakhla96] Nakhla, Micheal S., Vlach, J., "A Piecewise Harmonic Balance Technique for Determination of Periodic Response of Non-linear Systems", *IEEE Trans. on Circuits and Systems*, Vol. CAS-23, No. 2, pp. 85-91, February 1996.

- [Nortel96] HBT Design Handbook, Bell-Northern Research, Advanced Technology Laboratory, January 17, 1996.
- [Osaf94] Osafune, K., Yamauchi, Y., "20-GHz 5-dB Gain Analog Multipliers with AlGaAs/GaAs HBT's", IEEE Transactions on Microwave Theory and Techniques, Vol. 42, No. 3, pp. 518-520, March, 1994.
- [Oyama93] Oyama, Bert K., Wong, Brian P., "GaAs HBT's for Analog Circuits", Proceedings of the IEEE, Vol. 81, No. 12, pp. 1744-1761, December 1993.
- [Pucel76] Pucel, R.A., Massé, D., Bera, R., "Performance of GaAs MESFET Mixers at X Band", IEEE Transactions on Microwave Theory and Techniques, Vol. MTT-24, pp. 351-360, June, 1976.
- [Rober95] Robertson, L.D., *MMIC Design*, The Institution of Electrical Engineers, London, United Kingdom, 1995.
- [Sedra87] Sedra, A.S., Smith, K.C., *Microelectronics Circuits*, Saunders College Publishing, Fort Worth, Texas, 1987.
- [Sweet90] Sweet, A.A., *MIC & MMIC Amplifier and Oscillator Circuit Design*, Artech House, Norwood, Massachusetts, 1990.
- [Sze69] Sze, S.M., *Physics of Semiconductor Devices*, Wiley-Interscience, New York, 1969.
- [Sze85] Sze, S.M., *Semiconductor Physics*, John Wiley & Sons, New York, 1985.
- [Sze90] Sze, S.M., *High-Speed Semiconductor Devices*, John Wiley & Sons, New York, 1990.
- [Talbi96] Talbi, L., Delisle, G.Y., "Experimental Characterization of EHF Multipath Indoor Radio Channels", IEEE Journal on Selected Areas in Communications, Vol. 14, No. 3, pp. 431-439, April, 1996.
- [Teeter93] Teeter, D.A., East, J.R., Mains, R.K., Haddad, G.L., "Large-Signal Numerical and Analytical HBT Models", IEEE Transactions on Electron Devices, Vol. 40, No. 5, pp. 837-845, May, 1993.
- [Verver96] Verver, C.J., "CITR EHF In-Building Communications Project, Frequency Converters Test Report", CRC Technical Report, May 1996.
- [Wang90] Wang, H., Algani, C., Laporte, M., "HBT Oscillator and Mixer", Proceedings of the 20th European Microwave Conference, pp. 1728-1730, 1990.

- [Wight96] Wight, J.S., Stubbs, M.G., Shafai, L., "EHF System Architectures for Broadband Indoor Wireless LAN and LMCS Applications", Conference Proceedings of the 1996 Symposium on Antenna Technology and Applied Electromagnetics, pp. 729-732, 1996.
- [Yip92a] Yip, P.C.L., Leung, C.K., Lund, T., "An Emitter-injected Mixer with Improved Conversion Gain, Efficiency", Proceedings of the 1992 IEEE Region 10 International Conference, pp. 690-694, 1992.
- [Yip92b] Yip, P.C.L., Leung, C.K., Lund, T., "Upper Estimation on the Conversion Gain of a Mixer", Proceedings of the 1992 IEEE Region 10 International Conference, pp. 685-689, 1992.
- [Zhang96a] Zhang, Q.M., Hu, J., Sitch, J., Surridge, R.K., Xu, J.M., "A New Large Signal HBT Model", IEEE MTT-S International Microwave Symposium Digest, pp. 1253-1256, 1995.
- [Zhang96b] Zhang, Q.M., Hu, H., Surridge, R.K., Xu, J.M., "A New Large Signal HBT Model", IEEE Transactions on Microwave Theory and Techniques", IEEE Transactions on Microwave Theory and Techniques, Vol. 44, No. 11, pp. 2001-2009, 1996.

APPENDIX A: ANALYSIS OF A SINGLE-BALANCED MIXER

Appendix A shows a mathematical analysis applied to an active single balanced upconverter and illustrates the effectiveness of singly balanced mixers in rejecting the LO tone and some spurious signal.

The recombination of the output signals, from each non-linear element (NLE), that lead to the spurious signals and LO tone rejection is of interest in this appendix, as it is not obvious at first glance why this is so. To gain insight into the NLEs' interaction in balanced mixers, a mathematical example based on the mixer configuration shown in Figure A-1 is done.

Figure A-1. Single-balanced upconverter.

In this example, the NLE is assumed to be a three port element, where the voltage waveforms in ports 1 and 2 combine to give the output current waveform in port 3. Recall that baluns are meant to split the input signal into anti-phase components of equal magnitude. The subsequent analysis examines how the output current waveforms for each NLE, I_1 and I_2 , combine to produce the RF output signal, yet reject undesirable spurious signals and the LO tone.

Equation A-1 shows the non-linear relationship associated with each NLE. Each NLE is assumed to be identical and oriented in the same way. For simplicity, the non-linear relationship between ports 1, 2 and port 3 is assumed to be a transfer non-linearity.

$$I = aV + bV^2 + cV^3, \text{ where } V \text{ is the total AC voltage across the NLE and } I \text{ is the total AC output current.} \quad (\text{A-1})$$

The input voltages at each NLE, V_1 and V_2 , are the combination of the anti-phase input IF voltages at port 1 and the anti-phase LO voltages at port 2:

$$\begin{aligned} V_1 &= V_{LO} \cos(\omega_{LO}t) + V_{IF} \cos(\omega_{IF}t) \\ V_2 &= -V_{LO} \cos(\omega_{LO}t) - V_{IF} \cos(\omega_{IF}t), \end{aligned} \quad (\text{A-2})$$

where V_{LO} is the magnitude of LO excitation signal and V_{IF} is the magnitude of IF input signal.

Substituting the input excitation voltages of equation A-2 into equation A-1, and applying well-known trigonometric identities, yields the equation for the output currents I_1 and I_2 , for NLE₁ and NLE₂ respectively. The first terms for each NLE, I_{11} and I_{21} , are given by:

$$\begin{aligned} I_{11} &= V_1 = a\{V_{LO} \cos(\omega_{LO}t) + V_{IF} \cos(\omega_{IF}t)\} \\ I_{21} &= V_2 = -a\{V_{LO} \cos(\omega_{LO}t) + V_{IF} \cos(\omega_{IF}t)\} \end{aligned} \quad (\text{A-3})$$

The second and third terms, for each NLE, are given by equation A-4 and A-5 respectively:

$$\begin{aligned} I_{12} &= V_1^2 = \frac{1}{2}b\{V_{LO}^2 + V_{IF}^2 + V_{LO}^2 \cos(2\omega_{LO}t) + V_{IF}^2 \cos(2\omega_{IF}t) \\ &\quad + 2V_{LO}V_{IF}[\cos((\omega_{LO} + \omega_{IF})t) + \cos((\omega_{LO} - \omega_{IF})t)]\} \\ I_{22} &= V_2^2 = \frac{1}{2}b\{V_{LO}^2 + V_{IF}^2 + V_{LO}^2 \cos(2\omega_{LO}t) + V_{IF}^2 \cos(2\omega_{IF}t) \\ &\quad + 2V_{LO}V_{IF}[\cos((\omega_{LO} + \omega_{IF})t) + \cos((\omega_{LO} - \omega_{IF})t)]\} \end{aligned} \quad (\text{A-4})$$

$$\begin{aligned} I_{13} &= V_1^3 = \frac{1}{2}c\{\frac{1}{2}V_{LO}^3 \cos(3\omega_{LO}t) + \frac{1}{2}V_{IF}^3 \cos(3\omega_{IF}t) \\ &\quad + 3V_{LO}^2V_{IF}[\cos((2\omega_{LO} + \omega_{IF})t) + \cos((2\omega_{LO} - \omega_{IF})t)] \\ &\quad + 3V_{LO}V_{IF}^2[\cos((\omega_{LO} + 2\omega_{IF})t) + \cos((\omega_{LO} - 2\omega_{IF})t)] \\ &\quad + 3(\frac{1}{2}V_{LO}^3 + 2V_{LO}V_{IF}^2)\cos(\omega_{LO}t) + 3(\frac{1}{2}V_{IF}^3 + 2V_{LO}^2V_{IF})\cos(\omega_{IF}t)\} \end{aligned} \quad (\text{A-5})$$

$$\begin{aligned}
I_{23} = V_2^3 = & -\frac{1}{2}c\left\{\frac{1}{2}V_{LO}^3 \cos(3\omega_{LO}t) + \frac{1}{2}V_{IF}^3 \cos(3\omega_{IF}t) \right. \\
& + 3V_{LO}^2 V_{IF} [\cos((2\omega_{LO} + \omega_{IF})t) + \cos((2\omega_{LO} - \omega_{IF})t)] \\
& + 3V_{LO} V_{IF}^2 [\cos((\omega_{LO} + 2\omega_{IF})t) + \cos((\omega_{LO} - 2\omega_{IF})t)] \\
& \left. + 3\left(\frac{1}{2}V_{LO}^3 + 2V_{LO} V_{IF}^2\right) \cos(\omega_{LO}t) + 3\left(\frac{1}{2}V_{IF}^3 + 2V_{LO}^2 V_{IF}\right) \cos(\omega_{IF}t)\right\}
\end{aligned}$$

The output current waveform, $I_{out} = I_1 + I_2$, is obtained at the output of the RF Wilkinson combiner. Note that this particular balanced mixer configuration is particularly adapt at canceling odd order spurious products.

$$\begin{aligned}
I_{out} = b\{ & V_{LO}^2 + V_{IF}^2 + V_{LO}^2 \cos(2\omega_{LO}t) + V_{IF}^2 \cos(2\omega_{IF}t) \\
& + 2V_{LO} V_{IF} [\cos((\omega_{LO} + \omega_{IF})t) + \cos((\omega_{LO} - \omega_{IF})t)] \}
\end{aligned} \tag{A-6}$$

From equation A-6, we see that the LO tone is canceled at the RF port (thereby isolating the RF port from the LO port). The cancellation of all third order intermodulation products means an improvement in the 1 dB compression point. Recall that the 1 dB compression point occurs when the RF output power saturates by 1 dB with respect to a linear increase dependent on the input IF power. This is due to the input power gradually transferring more power to higher order intermodulation products. Less higher order spurious signals means it takes comparatively more IF input power before the onset of the 1 dB compression point and of the IP3. See Maas [Maas93] for an extensive treatment on balanced mixers.

APPENDIX B: LOAD-LINE ANALYSIS TEST BENCH AND CIRCUITS

This appendix shows the Libra load-line analysis test bench and test circuit used in section 3.3.1.3. The test bench (Figure B-1) and test circuit (Figure B-2) are used to obtain the load-line for each of the identified non-linear regions, when the LO is injected through the emitter port of the HBT. The LO power level is set at 5 dBm.

Load-line test bench

Figure B-1. Load-line test bench.

Load-line test circuit

Figure B-2. Load-line test circuit.

APPENDIX C: UPCONVERTER CIRCUIT AND TEST BENCHES

This appendix shows the Libra test circuit and the two test circuits used to design the MMIC upconverter (section 4.3) and to get the simulation results (section 4.6). The test circuit (Figure C-1) shows the upconverter. The first test bench (Figure C-2) is used to get all simulation results, except the RF return loss. The second test bench (Figure C-3) is used to obtain the RF return loss.

Mixer test bench

Figure C-2. Test bench for all simulations except the RF return loss.

Mixer test bench (for RF return loss only)

Figure C-3. Test bench for the RF return loss only.

APPENDIX D: SIMULATION TEST CIRCUIT FOR OBTAINING THE S-PARAMETERS OF SECTION 4.3.2

This appendix shows the Libra test circuit used for section 4.3.2, when an upper bound for the conversion gain is calculated. The S_{21} parameter is calculated for the case when the HBT is connected to the LO matching network and the LO trap at the LO. In this case, the LO port is terminated by a 50-ohm load.

Test circuit for S21 parameter extraction

Figure D-1. Test circuit for section 4.3.2.

APPENDIX E: DIAGRAMS OF THE MEASUREMENT SET-UPS

This appendix shows the measurements set-ups used during the measurement phase. Each set-up identifies the principal test equipment used. Omitted from each test set-up is the equipment that is used for all the tests and is not specific to any particular test: the frequency counters, the DC source, the digital multimeters, the power meter, and the graphics plotter. Two frequency counters were used, the HP5343A frequency counter were used to measure the frequency at IF while the EIP578 frequency counter were used to measure the frequency at LO and RF. The DC source provided DC power to the MMIC while the Fluke 8010A digital multimeters were used to monitor V_{ce} , V_{be} , I_c , and I_b . The HP473B power meter was used to validate the power read-out on the spectrum analyzers. Finally, the HP7550A graphics plotter was connected to the HP8564E spectrum analyzer in order to record some of the test results.

When measuring the IF return loss, one port of the HP8510 network analyzer is connected to the IF port while the RF port is terminated by a 50-ohm load. For the RF return loss, the IF port is terminated by 50-ohm load while the RF port of the MMIC is connected to the HP8510. The HP8510 is calibrated up to the probe tips by a through-short-open-load calibration.

Figure E-1. Test set-up for the IF and RF return loss.

The LO return loss cannot be measured by the HP8510 network analyzer because the analyzer cannot output more than -10 dBm at 26.5 GHz. A large-signal LO tone needs to be incident at the LO port of the MMIC in make the HBT behave in a realistic non-linear fashion. A -10 dBm signal is incident on the IF port. The LO return loss, in dB, is the difference between the reflected LO power in dBm and the incident LO power in dBm.

Figure E-2. Test set-up for the LO return loss.

For the output spectrum measurement, the full range of the HP8564E spectrum analyzer (30 kHz to 40 GHz) is used. The power levels recorded by the spectrum analyzer are referred back to the MMIC RF port by taking into account the frequency-dependent losses in the coaxial cable and probe between the HP8564E and the MMIC.

Figure E-3. Test set-up for the output spectrum and the conversion loss measurements.

For this test, the input power of each tone is measured at the IF input of the prober using a spectrum analyzer and then referred to the IF port of the MMIC. Once the initial input power levels are known, the HP8564E analyzer is used to measure the relative power increase of each input tone.

Figure E-4. Test set-up for the two-tone third order intermodulation measurement.

For the LO-IF isolation measurement, -2 dBm of LO power is injected as usual in the emitter port of the MMIC upconverter. The HP8564E spectrum analyzer measures the LO tone coming out of the IF port.

Figure E-5. Test set-up for the output spectrum at the IF port.