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**EMPIRICAL MODELING OF A GaAs/AlGaAs
HETEROJUNCTION BIPOLAR TRANSISTOR FOR
MICROWAVE CIRCUIT APPLICATIONS**

Tony Laneve

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

Master of Electrical Engineering

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

January 1995

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submitted by
Tony Laneve, B.Eng.

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

Thesis Supervisor

Thesis Co-Supervisor

Chair, Department of Electronics

Carleton University

March 1995

Abstract

The ever increasing demands of today's electronic circuits has challenged transistor designers to continually improve the performance of active devices. Advancements in epitaxial growth methods over the past twenty years has resulted in a mature GaAs/AlGaAs heterojunction bipolar transistor (HBT) technology. GaAs/AlGaAs HBTs combine the strengths of Si BJTs such as high power density and good low frequency noise characteristics with the speed and MMIC compatibility of GaAs FETs. Consequently GaAs/AlGaAs HBTs have demonstrated excellent performance in digital, high performance analog, and microwave circuit applications. In order to harness the performance potential of these devices, circuits must be built around them. This in turn requires a nonlinear model of the transistor. However, since GaAs/AlGaAs HBTs have only become commercially available recently, most commercial microwave CAD tools do not contain built-in HBT models.

This thesis presents the development and implementation of an empirical GaAs/AlGaAs HBT model based on the Gummel-Poon equivalent circuit. The model includes unique HBT characteristics such as self-heating and transit-time effects as suggested in the literature. Various parameter extraction techniques based on DC, AC, and thermal measurements are also examined. The model is capable of accurately predicting the following DC characteristics: I_C vs. V_{CE} , V_{BE} vs. V_{CE} , and the forward Gummel plot. Good agreement was also obtained between simulated and measured S-parameters from 0.4-40GHz in the forward active region of operation. The model is implemented in the commercial CAD program OSA90/HopeTM V2.5. To further assess the model's validity, load-pull measurements were performed and a microstrip oscillator was designed at 14 GHz.

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List of Symbols

γ	emitter injection efficiency
β	common-emitter current gain
$\Delta e_g, \Delta E_g$	increase in emitter bandgap as compared to a homojunction
β_{max}	maximum current gain when recombination currents are negligible
I_n	current due to injected electrons from emitter to base
I_p	current due to holes injected from base to emitter
I_r	current due to base bulk recombination
N_e	emitter doping level
P_b	base doping level
v_{nb}	average velocity of electrons at the emitter-end of the base
v_{pe}	average velocity of holes at the base-end of the emitter
k	Boltzmann's constant
T, T_j	junction temperature
μ_n	electron mobility
f_t	unity short-circuit current gain
τ_{ec}	transit time of electrons from emitter to collector
f_{max}	power gain cutoff frequency
g_o	common emitter output conductance
n_i	intrinsic carrier concentration
g_m	transconductance
I_C	collector current
I_B	base current
I_E	emitter current
V_{CE}	collector-emitter voltage
V_{BE}	base-emitter voltage
V_{BC}	base-collector voltage
K	stability factor for two-ports
Δ	$S_{11}S_{22} - S_{12}S_{21}$
V_{port}	voltage at output of OSCPORT element
V_s	estimate of voltage at output of OSCPORT element
I_D	current flowing through diode
q	electron charge
N	ideality factor in ideal diode equation

I_S	saturation current in ideal diode equation or collector saturation current in HBT model
τ	delay in forward current source
T_o	room temperature
I_{So}	collector saturation current at $T=T_o$
I_{SEo}	base leakage saturation current at $T=T_o$
I_{SCo}	collector leakage saturation current at $T=T_o$
N_F	forward collector ideality factor
N_R	reverse collector ideality factor
N_E	base leakage ideality factor
N_C	collector leakage ideality factor
B_F	ideal forward current gain
B_{Fo}	ideal forward current gain at $T=T_o$
B_R	ideal reverse current gain
R_B	base resistance
R_C	collector resistance
R_E	emitter resistance
C_{JEO}	zero bias base-emitter junction capacitance
V_{JE}	base-emitter built-in potential
M_{JE}	base-emitter grading factor
C_{JCO}	zero bias base-collector junction capacitance
V_{JC}	base-collector built-in potential
M_{JC}	base-collector grading factor
$C_{BE,diff}$	base-emitter diffusion capacitance
T_{FO}	coefficient characterizing bias dependence of base-emitter diffusion capacitance
X_{TF}	coefficient characterizing bias dependence of base-emitter diffusion capacitance
V_{TF}	coefficient characterizing bias dependence of base-emitter diffusion capacitance
X_{TI}	third temperature parameter characterizing I_S temperature dependence
E_g	base bandgap energy
E_{inf}	parameter characterizing B_F temperature dependence
p	parameter characterizing B_F temperature dependence
L_B	parasitic base inductance
L_C	parasitic collector inductance

L_E	parasitic emitter inductance
C_{BE}	parasitic base-emitter capacitance
C_{BC}	parasitic base-emitter capacitance
C_{CE}	parasitic base-emitter capacitance
R_{TH}	thermal resistance
C_{TH}	thermal capacitance
P_{diss}	DC power dissipated
T_{base}	base plate temperature
R_{je}, r_e	small-signal active emitter resistance
δ	coefficient of V_{BE} given by q/nkT used in emitter current diode equation
α	common-base current gain
ω	angular frequency
f	frequency
h_{feo}	DC short-circuit current gain
f_l	3-dB cutoff frequency of h_{21}
x	$-\text{Im}(h_{21})/\text{Re}(h_{21})$
r_{be}	small-signal active base-emitter resistance
P_{in}	available input power
P_{out}	available output power
Γ_{load}	load reflection coefficient
R_{load}	output load resistance
X_{load}	output load reactance
R_{out}	oscillator output resistance
X_{out}	oscillator output reactance
λ	wavelength
Z_o	characteristic impedance
$\epsilon(t)$	fluctuating amplitude component of real sinusoidal signal
$\Delta\phi(t)$	fluctuating phase component of real sinusoidal signal
t	time
$\mathcal{L}(f_m)$	single sideband phase noise to carrier ratio in 1Hz. bandwidth at offset frequency f_m from the carrier.

CHAPTER 1: INTRODUCTION

1.1 Overview

The ever increasing performance requirements of analog, digital and microwave systems have pushed present day semiconductor transistor technology to the limit. Conventional transistors such as Si bipolar junction transistors (BJTs) and GaAs Field Effect Transistors (FETs) are becoming less and less capable of meeting the performance needs of today. In order to meet these ever increasing demands and build the next generation of transistors, researchers have turned to an idea which is as old as the transistor itself. That idea is to build a transistor in which the active junctions are made of different materials hence the term *heterojunction*. Shockley first proposed this in his patent for the transistor in 1947 and later Kroemer performed much of the initial theoretical analysis [1]. Although the advantages of such a transistor were known at the time, the idea lay dormant because the technology required for its fabrication was not available. One type of transistor whose operation depends on the heterojunction is the Heterojunction Bipolar Transistor (HBT). HBTs can be fabricated in many material systems. These include Silicon-Germanium (SiGe), Indium-Phosphide (InP), and Gallium Arsenide/Aluminum Gallium Arsenide (GaAs/AlGaAs). Of the three systems mentioned, GaAs/AlGaAs is the most mature. Significant progress was made in GaAs HBT technology in the late 1970s due to the availability of high-precision growth methods for III-IV compounds such as molecular beam epitaxy (MBE) and metal-organic chemical vapour deposition (MOCVD). These epitaxial growth techniques provide accurate control of the composition (i.e aluminum content) thickness and doping of the various layers required for the fabrication of GaAs/AlGaAs HBTs [2], [3].

With conventional transistors, whose active junctions were made of like materials, the performance of the device was optimized by controlling mainly the doping levels and types. The heterojunction provides a new degree of freedom in device design.

The bandgap and hence the electric fields can now be controlled as well as the doping levels. By exploiting this bandgap engineering, higher performance transistors are realizable.

The HBT behaves in a similar way to its homojunction counterpart and it can replace the BJT in most of its circuit applications. The main strength of the HBT is that the heterojunction enhances the injection properties of the device and hence its current gain. This, coupled with a GaAs based material system produces a high speed bipolar transistor suitable for operation in the microwave and millimetre wave frequency ranges.

One of the areas where the GaAs HBT has made a large impact is in microwave circuits. This area of application is unique not only due to the high frequencies at which the devices must operate, but also the wide variety and diversity of the functions of modern microwave circuits. For example power amplifiers require devices with high power density, high breakdown voltage and high current handling capability. Low noise amplifiers require low noise figure, linear gain and large dynamic range. Microwave oscillators need devices with good low frequency noise performance. Such a wide criterion is difficult to fulfill with one type of technology. The GaAs FET was the first generic microwave transistor which was suitable for many applications. However, the varied needs of microwave systems could not fully be met with this device. This is why in modern systems p-i-n diodes and travelling-wave tubes are still in use. The GaAs HBT can help to better fulfill the needs of many microwave applications.

The two most promising areas of application of the GaAs based HBT are microwave power amplifiers and low phase noise oscillators. Impressive power density and power added-efficiency have been achieved. In the area of low phase noise oscillators, phase noise 10dB better than FETs has been demonstrated.

GaAs HBTs have a lot of performance potential. Improvements in their processing technology have come quickly since the late 70s. However to harness this potential, circuits must be accurately designed. To do so requires accurate transistor models. Since most applications of HBTs involve nonlinear circuits, the model must be non-

linear in nature. Furthermore it should be of an equivalent circuit type and be simple enough to implement in existing computer-aided design (CAD) programs. Transistor modeling is often the most difficult and challenging step in the circuit design process. Most microwave CAD tools contain built-in models for FETs and BJTs but not specifically for HBTs. Furthermore, it is not easy to implement user-defined models in many of such programs. A more open CAD environment is desirable in which designers can integrate the resources and specialized capabilities of many different CAD tools together.

1.2 Thesis Organization

The objectives of this thesis are to develop an empirical model of a GaAs/AlGaAs HBT. The model will be based on the standard BJT model but changes will be made in order to better describe characteristics unique to HBTs as suggested in the literature. Several parameter extraction techniques are investigated and evaluated. Computer optimization techniques are used to fit simulated to measured DC and small signal RF responses. Large signal measurements are used to determine the validity of the model under nonlinear operating conditions. An oscillator circuit is designed in order to further assess the usefulness of the model and CAD software OSA90/HopeTM V2.5.

The thesis is organized in the following manner. Chapter two briefly describes the operation of the HBT and in particular its heterojunction. The intrinsic characteristics of GaAs HBTs are then outlined. In order to better understand some of their advantages and disadvantages, HBTs are compared to conventional transistors such as Si BJTs and GaAs FETs. Some areas of application are discussed and typical performance results are quoted. Basic microwave oscillator theory is then reviewed. Linear and nonlinear design methods using microwave CAD are also examined.

Chapter three encompasses the main body of work for this thesis. It describes the empirical modeling approach in detail. The transistor used in this work is briefly described and then the model topology is presented along with all the relevant

model equations. The various transistor measurements performed are then discussed. Several parameter extraction techniques are outlined and the results shown. The optimization procedure is explained and final model parameter values are given. The simulated DC, S-parameter and nonlinear load-pull characteristics are then compared with measurement in order to assess the model's validity.

Chapter four deals with the design, fabrication and testing of an HBT microstrip oscillator using the previously developed model. The linear and nonlinear design procedure is explained in detail and the simulated results are presented. The testing of the oscillator circuit is described and various difficulties encountered with oscillator measurement are highlighted.

The conclusion and description of future work are explained in chapter five. The thesis is summarized and modeling results are discussed. Possible avenues of exploration in order to improve the model are suggested. Methods of improving the oscillator design are also suggested so that a second iteration can be attempted.

CHAPTER 2: BACKGROUND THEORY

2.1 Introduction

In this chapter the basic theory and principle of operation of the Heterojunction Bipolar Transistor (HBT) is explored. The treatment is qualitative in nature and is used only to provide some basic insight into the behaviour and properties of HBTs. The intrinsic characteristics of HBTs are outlined and how they relate to measured responses is stated. To this end, various figures of merit for measuring a transistor's performance capabilities are used. In order to further assess the strengths and limitations of GaAs/AlGaAs HBTs, a comparison with other traditional transistor technologies is made. These include GaAs Field Effect Transistors (FETs) and Silicon Bipolar Transistors (BJTs). The second part of this chapter deals with one of the main applications of GaAs/AlGaAs HBTs which is microwave oscillators. Basic microwave oscillator theory is reviewed and relevant design equations are given. In particular, nonlinear oscillator design methods using commercial microwave CAD packages are described.

2.2 Principle of HBT Operation

In bipolar transistors, unlike FETs, both electrons and holes carry the current. The major contributors to current flow in a simplified npn transistor are shown diagrammatically in Figure 2.1. In order for the transistor to provide current gain, it is necessary that the number of electrons injected from the emitter into the base be much greater than the number of holes back-injected from base to emitter. The parameter which expresses this ratio is the emitter injection efficiency given by (2.1):

$$\gamma = \frac{I_n}{I_n + I_p} \quad (2.1)$$

The hole current back injected into the emitter, I_p , is proportional to the equilibrium concentration of holes in the emitter. Therefore in order to achieve high values of γ (approaching unity) with conventional homojunction bipolar transistors, the emitter has to be much

Figure 2.1 Major components of current flow in a bipolar transistor in the forward active region of operation.

more heavily doped than the base. This makes $I_n \gg I_p$. Typically the emitter is 1 to 2 orders of magnitude more heavily doped than the base in BJTs. This places a practical upper limit on the base doping.

With HBTs however, bandgap engineering can be used to provide high injection efficiency and hence gain. A wide gap emitter is used which makes the emitter-base junction a heterojunction. The valence band barrier for holes is made larger than the conduction band barrier for electrons. This reduces the current due to holes injected from base to emitter and thus increases γ without the doping constraints required by homojunction transistors. This is illustrated in Figure 2.2 which shows the band diagrams for an npn homojunction and heterojunction transistor. Note the increase in the bandgap of the emitter given by ΔE_g .

In order to see how the increased bandgap in the emitter affects the injection properties and current gain more quantitatively, a figure of merit for the transistor, β_{max} can be considered. β_{max} represents the highest value of current gain obtainable when recombination currents are negligible. This is indicated in (2.2):

$$\beta = \frac{I_c}{I_b} = \frac{I_n - I_r}{I_p + I_r + I_s} < \frac{I_n}{I_p} \equiv \beta_{max} \quad (2.2)$$

Figure 2.2 Energy band diagrams of a heterojunction bipolar transistor (top) and a homojunction bipolar transistor (bottom) biased in the forward active region of operation.

With reference to Figure 2.2, I_n is the current due to electrons injected from the emitter into the base, I_r is the current due to bulk recombination, I_p is the current due to holes injected from the base into the emitter, and I_s the current due to recombination in the base-emitter depletion region. Kroemer gives an estimate of β_{max} for an HBT which is stated here in (2.3) [2]:

$$\beta_{max} = \frac{N_e v_{nb}}{P_b v_{pe}} \exp(\Delta e_g / kT) \quad (2.3)$$

where N_e , P_b are the doping levels in the emitter and base respectively, v_{nb} is the mean velocity of electrons at the emitter-end of the base, v_{pe} is the mean velocity of holes at the base-end of the emitter, Δe_g is the bandgap difference between the emitter and the base, k is Boltzmann's constant, and T is the temperature. Equation (2.3) differs from that for a homojunction transistor by the exponential factor. To obtain large values for β_{max} it is necessary then to make $N_e \gg P_b$ (as is done for homojunction transistors) or to make Δe_g a few times larger than kT . For example a bandgap difference of $8kT$ can offset a doping ratio of 3000 [3]. Thus in principle in an HBT, the bandgap can be chosen so that the current gain is independent of the base and emitter doping levels. The doping levels can thus be altered in order to improve the transistor performance without sacrificing current gain.

For a GaAs/AlGaAs HBT, in order to produce the desired wide gap emitter already discussed, an AlGaAs layer with typical thickness of $0.1\mu\text{m}$ is used in the emitter. The basic layer structure of a GaAs/AlGaAs HBT is shown in Figure 2.3. The bandgap

Figure 2.3 Cross Section of a GaAs/AlGaAs HBT showing layer structure (after Kim et al. [4]).

variation as a function of aluminum composition is shown in Figure 2.4. It varies by approximately 12.5 meV per % aluminum until the composition of aluminum reaches

45% [3]. In order to prevent the formation of deep donor levels (DX centers) the composition of aluminum has been limited to 30%. One of the reasons that aluminum is used to create wide gap emitters is due to the fact that at room temperature the lattice mismatch between GaAs and AlAs is only 0.14% [3]. This avoids the problem of dislocations forming.

Figure 2.4 Conduction and valence band-edge energies (E_c and E_v) respectively, measured relative to the valence band edge of GaAs vs. aluminum composition in the GaAs-AlAs system (after Asbeck et al. [3]).

Thus, in summary, the wide gap emitter employed in HBTs enhances the injection properties of the device such that current gain is obtainable without the doping constraints imposed on homojunction transistors. The advantages that this provides will be examined in the next section.

2.3 HBT Device Characteristics

As has been discussed the primary advantage of the HBT is the enhanced injection properties it possesses due to the employment of the heterojunction. This section

overviews the main characteristics of the HBT by comparing its operation to that of the Si bipolar transistor and the GaAs FET. The discussion will be concentrated on GaAs/AlGaAs HBTs and their use in microwave circuit applications.

2.3.1 GaAs/AlGaAs HBT vs. Si BJT

The GaAs HBT and Si BJT, both being bipolar transistors, operate by essentially the same principle. Consequently they both behave in a similar fashion. The use of the GaAs/AlGaAs material system instead of Si allows the HBT to retain the advantages of the BJT and simultaneously overcome many of its limitations and weaknesses. One of the main advantages of GaAs based transistors is improved speed and hence higher frequencies of operation. This is due to the fact that the steady state velocity of electrons as a function of the electric field is much higher in GaAs than Si. This point is illustrated in Figure 2.5. The electron mobility, μ_n , in undoped GaAs ($8000 \text{ cm}^2/\text{Vs}$) is approxi-

Figure 2.5 Steady-state velocity-field characteristics for electrons in GaAs and Si (after Asbeck et al. [3]).

mately 7 times larger than for silicon. This leads to a lower base transit time, and diffusion capacitance [3]. A figure of merit used to compare intrinsic device speed is the unity short circuit current gain, f_t . This is defined as the frequency at which the short circuit current gain is equal to unity and is given in (2.4):

$$f_t = \frac{1}{2\pi\tau_{ec}} \quad (2.4)$$

where τ_{ec} is the transit time for electrons from emitter to collector. Thus it is seen that the reduced base transit time leads to a higher f_t . The highest reported value of f_t for a GaAs/AlGaAs HBT is 170 GHz [3]. In comparison, the Si bipolar transistor using super self-aligned transistor (SST) technology has been able to obtain f_t values only above 20 GHz but at the expense of more complex processing [4].

In HBTs, since the heterojunction serves to enhance the injection efficiency, the constraint that the emitter doping be much larger than that of the base no longer applies. As such, the base doping in HBTs can be significantly increased which results in many performance benefits. The base doping in HBTs is about 100 times larger than that of Si BJTs. This leads to base sheet resistance values as low as 100-150 Ω /square [3], [5]. The resulting low base resistance is beneficial for microwave power amplification. The microwave power gain in an HBT depends strongly on the base resistance. This is indicated in (2.5) which is an approximate expression for the power gain cutoff frequency, f_{max} , also called the maximum frequency of oscillation. It is defined as the frequency at which the power gain of the transistor in common emitter (CE) configuration drops to unity [4], [6].

$$f_{max} = \left(\frac{f_t}{8\pi R_b C_{bc}} \right)^{1/2} \quad (2.5)$$

In (2.5), f_t is the unity short circuit current gain, R_b is the base resistance and C_{bc} is the

base-collector capacitance. Thus a decrease in the base resistance leads to higher values for f_{max} . The highest reported value of f_{max} for GaAs/AlGaAs HBTs is 218 GHz [5]. Furthermore, the high base doping also leads to lower values of common emitter output conductance, g_o (2.6) due to the minimization of base width modulation [3], [4], [5], [7].

$$g_o = \frac{dI_C}{dV_{CE}} = \frac{I_C}{V_A} \quad (2.6)$$

The Early Voltage, V_A , for HBTs is usually larger than that for Si BJTs by 10-20 times [4]. These characteristics lead to higher device linearity due to the small variation of collector current with collector-emitter voltage. The high base doping and high electron velocity of the HBT also delays the onset of high injection and Kirk effects until higher current densities [4]. Current crowding at the periphery of the emitter in HBTs is also greatly reduced compared to the Si BJT due to the low base sheet resistance. As a result, HBTs with emitters as wide as $2\mu\text{m}$ can operate at Ku-band frequencies [6]. In comparison, Si BJTs require sub-micron emitter widths to operate at these frequencies (see Figure 2.6) [5]. The larger emitter dimensions possible for GaAs/AlGaAs HBTs, simplifies the fabrication process which leads to more uniformity in the device characteristics.

GaAs substrates have high resistivities ($>10^8 \Omega\text{cm}$). This is due to the large bandgap of GaAs, 1.42eV, and the low intrinsic carrier concentration, $n_i = 2 \times 10^6 \text{ cm}^{-3}$ [3]. This semi-insulating property of GaAs provides many advantages. These include greater device isolation and simpler processing. The parasitic substrate capacitance is also decreased over that in Si which results in faster circuit speeds and low loss matching networks for Monolithic Microwave Integrated Circuits (MMICS) [7].

Advanced Si BJTs however still retain some advantages over GaAs HBTs. For example the integration complexity of SST for Si BJTs is at the large scale integration (LSI) to very large scale integration (VLSI) level while that of GaAs HBTs is currently limited to medium scale integration (MSI) to LSI [7]. Furthermore, Si BJTs are still supe-

rior to GaAs HBTs in their low frequency noise characteristics. A figure of merit used to compare the low frequency noise of devices is the $1/f$ noise corner. This is defined as the baseband frequency at which the excess noise intersects the background white noise [8]. This value is typically on the order of 100kHz for GaAs/AlGaAs HBTs which is much lower than that for FETs (see next section) but is still 10 to 100 times higher than the noise corner of Si BJTs [4], [7].

2.3.2 GaAs/AlGaAs HBT vs. GaAs FET

FETs are still the basic building blocks in microwave integrated circuits. Recent developments in GaAs HBTs have shown that these devices are a viable alternative to the traditional FETs for many high performance analog and microwave applications. This section compares the intrinsic device characteristics of the HBT and FET and points out the strengths and weaknesses of each device for different applications.

One major difference between the HBT and FET is in the physical device structure. In an FET, the current flows laterally through a thin gate region. The HBT, in contrast, is a vertical structure. The speed of this device is mainly determined by the transit time of electrons through the thin (thousands of Angstroms) vertical layers. These layers can be easily and accurately controlled by epitaxial growth. Furthermore, the minimum geometry feature of the HBT is the emitter width which is approximately 1.5-2.0 μm for microwave devices [4], [5], [7]. Thus the HBT can be fabricated using conventional optical lithography. In contrast, to operate at the same microwave frequencies, FETs require gate dimensions of 0.2-0.5 μm which necessitate the use of more complicated and costly submicron lithography techniques such as electron-beam lithography. This point is illustrated in Figure 2.6. The vertical structure of the HBT also enables it to possess a higher current per effective transistor area. With FETs, the current flows through a thin gate channel. In HBTs on the other hand, the entire emitter area contributes to current flow, thus higher current and power handling per transistor area result. These

Figure 2.6 Lithography requirements of HBTs, FETs and Si BJTs (after Bayraktaroglu [5]).

properties lead directly to the application of microwave power amplifiers. This is one of the main areas where the HBT is well suited. For example, in CW operation HBTs on average can produce power densities of 4W/mm of emitter length or greater at Ku-band [5], [7]. Under short pulse duration a power density of 18.7W/mm at X-band has been reported [5]. In order to obtain higher efficiency in power amplifiers nonlinear modes of operation such as Class B and C are used. The drain leakage currents (during the off cycle) and lower voltage gain in FETs however, reduces drain efficiency and limits their use in these modes [9]. HBTs however are ideally suited for these modes of operation because of their higher voltage gain and negligible leakage currents. For example 67.8% PAE, 11.6dB gain with corresponding 5.6W/mm power density at 10GHz has been reported for an HBT in Class B operation [9]. The base-collector breakdown voltage is an important parameter in power amplifiers. This parameter can be easily adjusted in HBTs by control of the collector region's doping level and thickness [4], [5].

One inherent difference between bipolars in general and FETs is in the transconductance, g_m . This is defined as the derivative of the output current with respect to the input voltage. This parameter is 10 to 100 times larger in HBTs than FETs. The higher transconductance results from the exponential input voltage to output current rela-

tionship for bipolars compared to that of FETs which is linear or quadratic [4]. This provides higher amplifier gain for a given DC power level. The high transconductance along with the low output conductance leads to a large value for intrinsic gain, (g_m/g_o) important in analog applications.

Another important difference between HBTs and FETs is in their noise characteristics. Advanced FETs show superior noise figure performance compared to GaAs/AlGaAs HBTs at microwave frequencies. The minimum noise figure of a GaAs HBT is 2-6 dB and increases with frequency [5], [6]. The current noise behaviour of HBTs is similar to that of $1\mu\text{m}$ FETs of about 15 years ago [5]. The low frequency noise performance of the HBT, however, is far superior to that of FET. For GaAs/AlGaAs HBTs, typical values for the $1/f$ noise corner is 100kHz while that of FETs is 10-100 times higher. Recently, low frequency noise measurements were performed on some GaAs/AlGaAs HBTs from Rockwell International which showed noise corners on the order of 100kHz [8]. This excellent noise performance has been attributed to the fact that the current flow in HBTs is through active junctions which are well shielded from surface and substrate interfaces. With FETs, the carriers travel between surface and gate channel-substrate interfaces which have greater trapping effects and produce larger $1/f$ noise [4]. In GaAs HBTs one of the major causes of $1/f$ noise is the base surface recombination current. Several methods have been used to reduce this current. The use of self-alignment techniques for the base-emitter reduces the extrinsic base surface region. Another technique is to deposit a passivation layer on the GaAs surface which reduces the surface recombination velocity [5], [6].

The low $1/f$ noise of the GaAs HBT makes it attractive for low phase noise microwave oscillators and mixers. This low frequency noise is upconverted by the nonlinear nature of the oscillator (or mixer) and appears as phase noise around the fundamental frequency of operation. Low phase noise oscillators are therefore an important application for HBTs. For example an HBT dielectric resonator oscillator (DRO) at 11.06GHz exhib-

ited a phase noise of -76 dBc/Hz and -102dBc/Hz at 1 kHz and 10 kHz from the carrier respectively [10]. This type of performance is approximately 10dB better than a FET based oscillator with similar resonator Q-values.

A major concern with GaAs/AlGaAs HBTs lies in their thermal properties. The HBT has a much higher thermal resistance than the GaAs MESFET. This has resulted in device self-heating as indicated by a negative differential resistance in the static DC-IV curves. In a comparison done by Long it was shown that the thermal resistance of the HBT was approximately six times that of the MESFET for the same specifications of a power amplifier [11]. Part of the reason is that the thermal conductivity of GaAs is low ($0.46 \text{ W/cm}^{\circ}\text{C}$, three times lower than Si). This is an important issue since although the HBT has much higher power density capability (4 W/mm of emitter length) its output power in CW operation is thermally limited as opposed to electrically limited. This could make the device less reliable in high power CW applications if careful attention is not paid to thermal management issues.

2.4 HBTs, Si BJTs and FETs: A Summary

The above sections have touched upon the similarities and differences of the different transistors. Only some of the basic properties of the GaAs/AlGaAs HBT were discussed. Many other techniques are used to optimize the performance of the device. For example carrier velocity overshoot in the base, hot electron injection, and the use of a graded base can be used to increase the speed of the device [3], [6], [7]. Use of a wide band collector, called a Double Heterojunction Bipolar Transistor (DHBT), can increase the base-collector breakdown voltage and reduce charge storage during transistor saturation [2], [3]. Complementary devices are also realizable (PNP) which have shown fair performance at microwave frequencies [5]. Other material systems for HBTs are receiving much research interest. For example InGaAs/InP HBTs are being developed which have advantages over the GaAs/AlGaAs system. These include higher electron mobility,

lower surface recombination velocity, and semi-insulating substrates with higher thermal conductivity than GaAs [3].

The strength of the GaAs HBT is that it combines the speed, high frequency performance, and MMIC compatibility of the GaAs FET while at the same time providing the advantages of bipolar technology, which includes high power density, low $1/f$ noise, and high linearity. Figure 2.7 summarizes the intrinsic characteristics of each device.

Intrinsic Device Characteristics	GaAs HBT	Advanced Si Bipolar	GaAs HEMT/MESFET
• Device Speed (f_T , f_{max})	~ 50-200 GHz, ~ 1.3 μm Emitter	~ 20-40 GHz, $\leq 0.5 \mu\text{m}$	~ 0.2-0.5 μm Gate
• Current/Effective Chip Area	~ 50 kA/cm^2	~ 2.5X Lower	~ 5-10X Lower
• Device Matching	$V_{be} \sim 1.2 \text{ mV}$	Comparable	~ 3.5X Worse
• Transconductance (g_m)	~ 5-10 $\text{K mS}/\text{mm}$	Comparable	~ 20-100X Lower
• Output Conductance (g_o)	~ 0.2 mS/mm	~ 10-20X Higher	~ 25-100X Higher
• Trapping Effects	$1/f$ Corner $< 100 \text{ kHz}$	~ 10-100X Lower	~ 10-100X Higher
• Input Impedance	Base Input Current	Same	Voltage Controlled Gate
• White Noise (10-20 GHz)	~ 4-6 dB	Comparable	~ 2-7X Lower
• Radiation Hardness	$> 500 \text{ Mrad}$, $> 1 \times 10^{14} \text{ Neutrons}/\text{cm}^2$	~ 20-50X Lower	Comparable
• Substrate Capacitance	No	Yes	No
• Power Consumption	High	High	Medium
• Integration Level	MSI LSI	LSI VLSI	LSI

Figure 2.7 Comparison of intrinsic device characteristics for GaAs HBTs, Si BJTs, and GaAs FETs (after Kim et al. [4]).

2.5 Microwave Oscillator Theory

A microwave oscillator is a circuit which converts DC power and noise into RF power. Microwave oscillators are usually divided into two parts, the active device and the passive circuit. It is the interaction between these two parts which initiates and sustains oscillations. The active device can be a two terminal device such as a Gunn or IMPATT or a three terminal device such as a transistor. The transistor of choice for microwave oscillators was traditionally the FET since Si bipolars only possess gain at the lower microwave frequencies. Today, however, the GaAs/AlGaAs HBT has shown good performance for microwave oscillator applications [4], [5], [7], [10], [12]. The passive circuit part of the oscillator consists of some type of resonator and a low loss match. The resona-

tor is used to determine the frequency of oscillation and affects the overall circuit performance. The matching circuit enables the output power to be transferred to a load.

There are many different forms of oscillators. Fixed frequency oscillators are designed to produce a steady state oscillation at one frequency only. These types of oscillators are usually named for the resonator used in the circuit. Some examples of resonators include lumped element, distributed transmission line (such as microstrip), and dielectric resonator. Some tuning can be applied (either electrical or mechanical) to adjust the oscillation frequency by a small percentage. In voltage controlled oscillators (VCOs) varactors (voltage controlled capacitors) or YIG spheres are commonly used to adjust the frequency of oscillation over a wider bandwidth.

The following sections outline the basic design techniques for microwave oscillators. This includes small-signal design procedures as well as nonlinear methods used in commercial CAD packages.

2.5.1 Microwave Network Characterization Using S-Parameters

Scattering parameters (S-parameters) are valuable tools for the design of microwave circuits such as amplifiers and oscillators. These parameters characterize a general N-port linear circuit in terms of incident and reflected waves, a and b respectively. The parameters are measured with the ports terminated in matched loads (usually 50Ω). Figure 2.8 illustrates how the scattering parameters of a two-port network are defined [13].

An important property of a two-port network for oscillator and amplifier design is its stability. The stability is a measure of the two port's resistance to oscillate. A two port is said to be *potentially* unstable if there exists some passive port terminations that when connected to one of the ports will cause the reflection coefficient looking into the other port to be greater than 1. A two port is *unconditionally* stable if no passive load can be connected to either port to make it unstable. The stability properties of a two-port

Figure 2.8 Characterization of a two-port network by its scattering parameters.

can be characterized by two quantities: the stability factor, K , and Δ [14].

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} \quad (2.7)$$

$$\Delta = S_{11}S_{22} - S_{12}S_{21} \quad (2.8)$$

Assuming $|S_{11}|$ and $|S_{22}|$ of a two-port are less than 1, then it is unconditionally stable if $K > 1$ and $|\Delta| < 1$ [14], [15]. This is the goal of the amplifier designer. If $K < 1$, then the two-port is potentially unstable and can be used as an oscillator. Therefore one of the necessary conditions for oscillations is that K must be less than unity at the frequency of oscillation.

2.5.2 N-Port Oscillation Condition

The analysis of the oscillation conditions for an N-port device connected to an N-port circuit in terms of scattering parameters has been done by Khanna [16]. The situation is illustrated in Figure 2.9. Writing an equation for the incident and reflected waves in terms of the circuit and device S-parameters results in [15]:

Figure 2.9 Scattering parameter characterization of an N-port device connected to an N-port circuit [15], [16].

$$(S_d S_c - U) a_c = 0 \quad (2.9)$$

where U is the identity matrix. Since a_c is not equal to 0 then:

$$|S_d S_c - U| = 0 \quad (2.10)$$

For the most common case, the two-port, (2.10) reduces to

$$\begin{vmatrix} S_{11}\Gamma_1 - 1 & S_{12}\Gamma_2 \\ S_{21}\Gamma_1 & S_{22}\Gamma_2 - 1 \end{vmatrix} = 0 \quad (2.11)$$

This leads to the well known 2-port oscillation conditions given by (2.12) [13], [15]:

$$\begin{aligned} S'_{11} &= S_{11} + \frac{S_{12}S_{21}\Gamma_2}{1 - S_{22}\Gamma_2} = \frac{1}{\Gamma_1} \\ S'_{22} &= S_{22} + \frac{S_{12}S_{21}\Gamma_1}{1 - S_{11}\Gamma_1} = \frac{1}{\Gamma_2} \end{aligned} \quad (2.12)$$

This oscillation condition may also be expressed in terms of impedances. Assuming that port 1 of the 2-port is the output of the oscillator and port 1 of the passive device is the

load, the following equations relating reflection coefficient and impedance can be used: $S'_{11} = (z_{out} - 1) / (z_{out} + 1)$ and $\Gamma_1 = (z_L - 1) / (z_L + 1)$. Where S'_{11} and Γ_1 are the oscillator output and load reflection coefficients respectively. Substituting these relations into (2.12) results in:

$$\begin{aligned} R_{out} + R_L &= 0 \\ X_{out} + X_L &= 0 \end{aligned} \quad (2.13)$$

This equation shows clearly that since R_L is usually a positive quantity, the output resistance, R_{out} , must be negative in order for the device to oscillate.

2.5.3 Small-Signal Design Procedure

The small-signal or linear design procedure is straightforward. A device is chosen and biased such that it is potentially unstable at the frequency of oscillation (i.e. $K < 1$). If this is not the case then the topology of the circuit should be changed (for example by changing the common terminal of the transistor) or feedback must be added. The two common forms of feedback are series and shunt as indicated in Figure 2.10.

Figure 2.10 Shunt and series feedback configurations in a microwave oscillator [15].

The feedback elements are usually reactive and can be lumped or distributed. When considering the type of feedback in the oscillator circuit, it is important that the DC bias is not disturbed. For example, shunt inductive feedback could cause problems because a DC block is needed while capacitive series feedback is also difficult due to grounding requirements.

The two-port is then reduced to a modified one-port by the feedback and resonator circuits. The values of these elements are chosen such that the one-port exhibits a large negative resistance at the frequency of oscillation. In theory, the output load matching circuit is then designed so that (2.13) is satisfied. In practice however, this is not the case. As oscillations build up in the circuit, the negative output resistance decreases in magnitude until steady state operation is reached. If (2.13) is satisfied at start-up then in steady state the loop resistance (i.e. $R_{out} + R_L$) would be positive and oscillations would cease before they could be detected. Therefore to ensure oscillations build up, designers often use the following rule of thumb [13], [15], [17]:

$$\begin{aligned} R_L &= \frac{-R_{out}}{3} \\ X_L &= -X_{out} \end{aligned} \quad (2.14)$$

After start-up, R_{out} decreases in magnitude due to the device nonlinearities, until steady state is reached where $R_L = -R_{out}$.

The above procedure usually produces good results when designing microwave oscillators. The shortcomings of this technique is that as oscillations build up in the circuit the output resistance and reactance of the device change. As a result the frequency of oscillation is usually shifted somewhat from the designed value [14]. Tuning is usually required in order to adjust the frequency of oscillation.

2.5.4 Large-Signal Oscillator Design

The inherent limitation of applying linear design procedures (such as those based on S-parameters) to oscillator applications is that *all* oscillators are in reality nonlinear circuits. Since the oscillations grow from noise present in the circuit, the small-signal characterization is only valid at start-up. It is the nonlinearity in the circuit which causes the oscillations to stop growing and reach steady state. Nevertheless, small-signal design techniques are very valuable as a starting point for oscillator design. Using small-signal linear design methods, however, there is no way to determine or optimize important oscillator parameters such as the output power level, spectral purity, phase noise, efficiency, and load and frequency pulling characteristics. For these reasons quasilinear and nonlinear techniques are often employed.

Quasilinear design techniques are an extension of the linear technique. They attempt to model the power dependence of the active device's S-parameters as a function of power level at the fundamental frequency of oscillation only. Several methods of designing oscillators using these methods have been proposed. These include the experimental device-line method proposed by Wagner [18], [19] and several others based on the use of large-signal, or power dependent S-parameters [13], [20].

Quasilinear techniques take into account only the fundamental frequency component, however in reality oscillators generate harmonics. Nonlinear techniques take the harmonics into account, and therefore have the potential to be the most accurate. The drawback however is the added complexity and computing time required. Despite the aforementioned advantages, the success of nonlinear techniques depends on the accuracy of the nonlinear model used in the design. The accurate large-signal modeling of the transistor is often the most challenging step in the design of nonlinear circuits.

The most widespread technique for nonlinear analysis and design of oscillator circuits is based on time-domain methods. One of the most popular software tools is SPICE. Using the transient analysis capability, the oscillation can be seen from start-up to

steady-state [17], [21]. Oscillations are initiated by introducing a current or voltage impulse in the circuit. This impulse is used to simulate broadband noise present in the actual circuit. With high Q circuits it can take a long time just to simulate the complete oscillator start-up. This could make optimization of the circuit (i.e for maximum output power, spectral purity etc.) impractical.

The harmonic balance method is the other major nonlinear simulation technique. It is the most common in design of microwave circuits and it is employed in most of the industry standard microwave CAD packages such as LIBRATM, MDSTM, and HarmonicaTM. This method combines frequency domain analysis of the linear portions of the circuit with time domain analysis for the nonlinear elements such as transistors. Harmonic balance methods assume that for a sinusoidal stimulus to a circuit, there exists a solution which can be approximated accurately by a finite trigonometric series. This method has been shown to be faster than pure time domain techniques such as those used in SPICE due to the fact that the simulation frequency is known beforehand and complex time-sampling algorithms are avoided [22].

The harmonic balance technique is well suited for the design of amplifiers but applying it to oscillator design is not obvious. This is because harmonic balance requires a sinusoidal excitation to be applied to the circuit which does not exist in oscillators. Furthermore, unlike time-domain techniques, the harmonic balance method only provides steady state solutions.

One method used in harmonic balance simulators for oscillator design is to introduce a special directional coupler as is done in LIBRATM [25]. This four port coupler called OSCTEST, has zero electrical length and is invisible to normal circuit operation. It is connected in the feedback path of the circuit. The coupler allows the injection of a signal at the fundamental frequency in order to stimulate the oscillations, blocks the fundamental frequency flow in the feedback path and monitors the signal returned in the feedback path. The scattering matrix of the coupler depends on whether the frequency is

the fundamental or any of its harmonics, including DC [25]. This is indicated in (2.15).

$$S_{fo} = \begin{bmatrix} 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \end{bmatrix} \quad S_{nfo} = \begin{bmatrix} 0 & 1 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \end{bmatrix} \quad n \neq 1 \quad (2.15)$$

Only the fundamental frequency is transmitted from node 3 to node 2. Transmission from node 2 to node 1 and from node 1 to node 4 is broadband for all frequencies. All frequencies except the fundamental are transmitted from node 1 to node 2.

The oscillation condition is met when the ratio of injected and returned signals have unity magnitude and zero phase as shown in (2.16).

$$\frac{|V_4|}{|V_3|} = 1 \quad (2.16)$$

$$\angle V_4 - \angle V_3 = 0^\circ$$

The oscillation conditions can be satisfied by varying of the stimulus signal and at least one circuit element value. Figure 2.11 illustrates how the directional coupler is used in a circuit for oscillator design as taken from the LIBRA 3.5TM Applications Manual [25].

Another software package, OSA90/HopeTM V2.5 handles oscillator analysis and design in a similar fashion. The user has the flexibility of defining a four-port coupler similar to the one in LIBRA to perform oscillator design. The user can, instead, use a built-in element dedicated for oscillator design. This is a specialized port termination called OSCPORT [26]. It is placed at the output of the oscillator as shown in Figure 2.12. The voltage source V_s , is an estimate of the voltage at the port termination, V_{port} , at the fundamental frequency. The controlled source stimulates the oscillations. The circuit is in a state of free oscillation if the simulation results in $V_{port} = V_s$. The voltage transfer coefficient k , is equal to unity at the fundamental frequency and 0 otherwise. In this manner

Figure 2.11 OSCTEST element placed in feedback path for oscillator design using the harmonic balance technique [25].

Figure 2.12 OSCPORT element connected to oscillator output.

the stimulus is applied at the fundamental frequency only [26]. Two built-in labels are used in conjunction with the OSCPORT element. These are OSC_GAIN and OSC_PHASE and are defined by:

$$\begin{aligned}\text{OSC_GAIN} &= 20\log_{10}\left(\left|\frac{V_{port}}{V_s}\right|\right) \\ \text{OSC_PHASE} &= \angle(V_{port} - V_s)\end{aligned}\tag{2.17}$$

In terms of these labels, the oscillation condition is satisfied if both OSC_GAIN and OSC_PHASE are equal to zero.

CHAPTER 3: DEVICE MODELING

3.1 Introduction

GaAs based HBTs possess high values f_t and f_{max} as well as other inherent device advantages over other transistor technologies. Some of these have been discussed in chapter 2. As a result they have been found to be well suited for many high performance digital, analog, and microwave applications. Furthermore monolithically combined digital and microwave functions have also been realized [7]. In particular, the high power handling capability and low $1/f$ noise properties of GaAs/AlGaAs HBTs make them attractive candidates for microwave power amplifiers and low phase noise oscillators. In order to take advantage of this technology's performance capabilities, circuits must be successfully produced. To this end, an accurate model of the nonlinear device is essential. This is particularly true for MMICs where extensive tuning of the final circuit is not possible.

Traditionally, most microwave circuits have been based on FET technology. Furthermore, GaAs HBTs have not become commercially available until recently. As a result, current microwave circuit CAD tools do not contain built-in models directly applicable for HBTs. An extensive amount of research has been performed in the area of HBT device physics. Many of these physical models are computationally intensive numerical simulators which are quite accurate, but require enormously long times to arrive at solutions. Furthermore, they are difficult to employ with conventional microwave CAD tools. At the other extreme, simpler models such as the Gummel-Poon model, which is the standard in many simulators for Si BJTs is also used to describe HBTs in circuits [27]. These models however, do not take into account some characteristics present in GaAs HBTs such as self-heating and transit time effects.

This chapter describes the approach used to model a GaAs/AlGaAs HBT for microwave circuit applications. The model topology and equations are based on those

of the Gummel-Poon model. Modifications to this model as have been proposed in the literature have been implemented in order to better represent HBT behaviour. Various commercial modeling and microwave design tools are also discussed and compared. The equivalent circuit of the model along with relevant equations is then presented. The measurement setups used to obtain data for the modeling process and several parameter extraction methods are then outlined. The optimization procedure is explained and finally DC, linear, and nonlinear measurements are compared to the model's responses in order to assess its validity.

3.2 Modeling/Simulation Tools

There are numerous simulation tools available to a microwave engineer. The CAD tools available for this work included LIBRATM V3.5, HSPICE, and OSA90/HopeTM V2.5. When modeling a transistor for the purpose of circuit design it is of course required that the model can be implemented in a CAD tool so that the overall objective of circuit design can be accomplished. For microwave transistor modeling purposes, most CAD programs fall short. SPICE based simulators, in particular HSPICETM, provide excellent analog behavioral modeling capabilities [28]. Models can be created with arbitrary topologies. Furthermore, nonlinear voltage and current sources, which are arbitrary functions of voltages and currents at other nodes or branches in the circuit, can also be defined. One limitation with SPICE, however, is that all the built-in temperature dependent elements must remain at a constant temperature during simulation. This is not the case for HBTs due to their large thermal resistance. Another difficulty is that modeling often requires optimization of model parameters in order to fit some measured characteristics of the device. Most SPICE simulators do not contain built-in optimization. HSPICETM however, does provide optimization capabilities but they are not very flexible. The most severe limitation with SPICE for microwave modeling and simulation however, is that it does not contain an extensive library of microstrip, stripline and other transmis-

sion line elements essential for microwave design. Although a user-defined transistor model can be made with HSPICE it would not be very practical to use as the primary microwave CAD tool.

LIBRATM V3.5 is one of the industry standard microwave circuit design tools. As a modeling tool however, it is limited in many ways. As is the case with SPICE, the circuit temperature for built-in elements must remain constant. Furthermore, user-defined voltage or current sources which are arbitrary functions of voltages or currents elsewhere in the circuit cannot be entered directly. In order to do so requires the use of another package called LIBRA Sr.TM. To create a user-defined model, one must encode, in C, all model equations and the derivatives of all voltage and current sources with respect to all node voltages in the model and then compile and link this with LIBRATM. This could be a very time consuming procedure which would be worthwhile only after the model has been verified to be satisfactory. The other limitation is in the area of computer optimization in order to perform parameter extraction. LIBRATM V3.5 does contain many different optimizers such as random, gradient, minimax, etc. These are mainly used in circuit design by optimizing circuit parameters in order to meet some specification. LIBRATM can fit measured to modeled S-parameters at one bias for *linear* models only. It cannot however be used to fit measured data with simulated DC or small-signal responses at several bias points for a nonlinear model. These shortcomings do not make it a very practical tool for microwave transistor modeling.

Another software tool, OSA90/HopeTM V2.5 combines the microwave circuit simulation strengths of LIBRATM with the modeling capability of SPICE. Voltage, current, and charge sources can be entered directly into the netlist to facilitate the implementation of user-defined models. These sources can be arbitrary functions of any voltage(s) or current(s) in the circuit. The program also contains state of the art L1, L2, Random, Simplex and other optimizers. These optimizers can be used to fit measured and simulated responses for practically any type of measured data supplied by the user. This

includes DC, small-signal RF, as well as large-signal data. The software also contains many built-in nonlinear models as well as transmission line and microstrip elements required for microwave circuit design. OSA90/HopeTM is based on the harmonic balance engine (as is LIBRA) which is most familiar to microwave engineers. It is also unique in that one model is used for all types of simulation. This means the user can perform DC, small-signal and harmonic balance optimization *simultaneously* on one model. This is in contrast to LIBRA in which in order to obtain the S-parameters from a nonlinear model requires an extra procedure which is tedious. In addition, OSA90/HopeTM V2.5 contains dedicated elements to facilitate the design and analysis of oscillators using the harmonic balance algorithm.

As a result of the aforementioned advantages and versatility of this software package, OSA90/HopeTM V2.5 was chosen as the main modeling and circuit design tool for this work.

3.3 Modeling Procedure

There exists numerous methods used to model active devices, each of which has their own advantages and disadvantages. A good rule of thumb is to use the simplest model which can satisfactorily represent the device for the intended application. Much analytical device modeling based on the device physics of GaAs HBTs has been carried out [29], [30], [31], [32]. Physics based models give detailed information about the different collector and base current components and their origins. These types of models can be invaluable to the engineer who is trying to monitor and refine the fabrication process and optimize the design of the transistor for different applications. For example, device designers can use these models to examine the trade-offs in device performance caused by such things as different emitter and base grading schemes, changes in doping levels, or emitter sizes etc. The model equations can become quite complex and not easily implemented in commercial CAD packages. Furthermore these analytical, 1d, and 2d

numerical models require a complete knowledge of the processing parameters as well as the physical layout and dimensions of the device.

For this work, a more empirical approach to device modeling was employed in which the device physics were not directly taken into account. This is due to the fact that knowledge of the processing parameters used by the foundry for the HBTs was not available. Furthermore, the equations resulting from most of the physics-based models are too complicated to implement into a commercial CAD package.

The HBTs for this work have been provided by Rockwell International. The device modeled consists of 8 emitter fingers connected in parallel. The approximate dimensions of each emitter finger is $2 \times 12 \mu\text{m}^2$. The HBTs were fabricated on a $75 \mu\text{m}$ thick GaAs substrate. The transistor was in common emitter configuration and emitter grounding was provided through via holes to the back plane metallization. A photograph of the HBT is shown in Figure 3.1.

Figure 3.1 Rockwell common emitter 8 finger HBT used in this work.

The model in this work is based on the Gummel-Poon transistor model for Si homojunction BJTs [28]. This is the most widely used model formulation for BJTs and is built-in to most CAD packages. Since BJTs and HBTs behave in a qualitatively similar manner, it seems a reasonable starting point for the modeling work. Some modifications to the Gummel-Poon model are added in order to better describe some characteristics unique to HBTs. First, DC and small-signal S-parameter measurements were taken of the device in common emitter configuration. Parameter extraction methods were then used in order to obtain initial values for some of the model parameters. DC optimization was then performed to fit the measured and simulated DC characteristics. Small-signal optimization was carried out to extract the bias independent extrinsic parameters and to determine the variation of base-emitter, and base-collector capacitances as functions of junction voltages. Finally, equations were fitted to the capacitance vs. bias characteristics in order to complete the model.

3.4 Model Description

The following describes the modified Gummel-Poon model used in this work. The circuit topology is shown in Figure 3.2. The four diodes are used to represent the base-emitter and base-collector junctions of the transistor. These diodes obey the ideal diode equation as given in (3.1):

$$I_D = I_S \left(e^{\frac{qV}{NkT}} - 1 \right) \quad (3.1)$$

where I_D is the current flowing through the diode, I_S is the saturation current, q is the electron charge, which is $1.602 \times 10^{-19} \text{C}$, V is the voltage across the diode terminals, N is the ideality factor, k is Boltzmann's constant equal to $1.3806 \times 10^{-23} \text{J}$, and T is the temperature in Kelvin. Diodes D_{BF} and D_{BR} represent the base current components which are directly proportional to the current source between the emitter and collector. Consequently, these

Figure 3.2 Modified Gummel-Poon model including parasitics.

diodes simulate an ideal constant current gain. The current through diode D_{BF} is equal to I_{CC}/B_F and similarly the current through diode D_{BR} is equal to I_{EC}/B_R . The forward and reverse ideal current gains are represented by B_F and B_R respectively. Diodes D_E and D_C represent the so called non-ideal components of base current. This component of base current in GaAs/AlGaAs HBTs is mainly attributed to recombination in the space charge region [33], [34]. The saturation currents for D_E and D_C are I_{SE} and I_{SC} respectively while the ideality factors are N_E and N_C respectively. The current source between emitter and collector has the following form (3.2):

$$I_{CT} = I_{CC} - I_{EC} = I_S \cdot \exp\left(\frac{qV_{BE}(t - \tau)}{N_F kT}\right) - I_S \cdot \exp\left(\frac{qV_{BC}(t)}{N_R kT}\right) \quad (3.2)$$

where I_S is the collector saturation current, N_F and N_R are the collector current ideality factors for the forward and reverse modes of operation respectively. This equation

assumes that the Early Voltage of the device is infinite. This is a valid assumption for GaAs HBTs because the high base doping significantly reduces the effect of base width modulation as mentioned in Section 2.3.1.

One difference between (3.2) and the corresponding SPICE equation is the inclusion of the delay factor, τ . Grossman has indicated the need for this delay term [35]. The Gummel-Poon model in SPICE assumes that the base charge changes instantaneously with the base-emitter voltage, V_{BE} . Even if all passive parasitics are accounted for, Grossman found that SPICE simulations do not match the measured data well when frequency components approach the f_t of the device. Teeter et al. have compared different models for simulation of HBTs at millimetre wave frequencies [29], [36]. An accurate physics based fully numerical model was used to compare the Gummel-Poon model with a modified Ebers-Moll model proposed by the Teeter et al. As seen in Figure 3.3, the Gummel-Poon model was not able to predict the delay in the collector current waveform. The modified Ebers-Moll model which included the delay term however gave much better agreement with the fully numerical model.

Figure 3.3 Collector current waveform computed using three different models (after Teeter et al. [36]).

The three resistors R_B , R_C , and R_E model the ohmic potential drops between the external device contacts and the internal active junctions of the transistor. In this model these resistances are assumed to be constant. The parasitic elements are modeled by ideal inductors and capacitors and are also assumed to be bias independent. The inductors L_B and L_C model short lengths of transmission lines while L_E represents the via inductance. Similarly the three capacitances, C'_{BE} , C'_{CE} and C'_{BC} represent parasitic coupling between the terminals of the transistor.

GaAs/AlGaAs HBTs show a negative slope in the DC I-V curves which has not been observed in Si BJTs. This effect has been attributed to the large thermal resistance of the transistor which causes device self-heating [35], [37], [38], [39]. The large thermal resistance is a result of the low thermal conductivity of GaAs and the geometry of the device. As such, the junction temperature, T , is not constant but instead is a function of the power applied to the device as indicated by (3.3):

$$T = T_o + R_{TH}(I_C \cdot V_{CE} + I_B V_{BE}) \quad (3.3)$$

where T is the device temperature, T_o is the room temperature, R_{TH} is the thermal resistance, I_C is the collector current, V_{CE} is the collector-emitter voltage, I_B is the base current and V_{BE} is the base-emitter voltage. This equation can be realized in an equivalent circuit form as illustrated in Figure 3.4 [35], [40]. In this electrical analog of a thermal system the ambient temperature is represented by a DC voltage source, T_o . The power applied is characterized by a current source. The thermal resistance, R_{TH} , models the proportional increase in temperature due to power while the thermal capacitance, C_{TH} , represents the time delay associated with the device junction temperature. The thermal time constant of HBTs is in the order of μ s which is much larger than the signal period at microwave frequencies [35]. Therefore an accurate value for C_{TH} was not required in this model. C_{TH} is an important parameter in pulsed applications and also in lower frequency analog circuits where thermal transients can affect gain, linearity and settling time [41]. The output volt-

Figure 3.4 Thermal sub-circuit used in HBT model. The output voltage, T represents the junction temperature.

age, T represents the junction temperature. Therefore as the power dissipated changes, the junction temperature changes accordingly and dynamically modifies the temperature dependent model parameters.

The temperature dependence of the DC currents is modeled after the work of Wang et al. [42], [43]. The model is applicable to modern HBTs with high base doping and processed by a mature technology which minimizes surface recombination. The following equations proposed by Wang et al. are based on the built-in bipolar temperature compensation equations found in SPICE. The model parameters affected by temperature are the collector saturation current, the recombination saturation currents and the current gain. The ideal current gain, represented by B_F , has the temperature dependence described by (3.4):

$$B_F = B_{F0} \left(\frac{T}{T_0} \right)^p \exp \left(\frac{E_{inf}}{kT} - \frac{E_{inf}}{kT_0} \right) \quad (3.4)$$

where T is the junction temperature, T_0 is the room temperature, B_{F0} is the ideal gain at $T=T_0$, k is Boltzmann's constant, p is a fitted parameter, and E_{inf} is a parameter characterizing the temperature behaviour of the current gain. The temperature variation of the collector saturation current, I_S , is the same as in SPICE [28] and is given by (3.5):

$$I_S = I_{So} \left(\frac{T}{T_o} \right)^{XTI} \exp \left(-\frac{E_g}{kT} + \frac{E_g}{kT_o} \right) \quad (3.5)$$

where E_g is the bandgap energy in the base and XTI is the third temperature parameter. The temperature dependence of the non-ideal forward base saturation current, I_{SE} , is represented by (3.6):

$$I_{SE} = I_{SEo} \left(\frac{I_S}{I_{So}} \right)^{1/N_E} \cdot \frac{B_{Fo}}{B_F} \quad (3.6)$$

where N_E is the base leakage ideality factor.

The difference between this model's temperature equations and those of SPICE is that the term $\left(\frac{T}{T_o} \right)^{XTB}$ is replaced by $\left(\frac{T}{T_o} \right)^p \exp \left(\frac{E_{inf}}{kT} - \frac{E_{inf}}{kT_o} \right)$ [42].

The base-collector capacitance, C_{BC} , is modeled with a depletion component only (C_{JC}) using the standard SPICE equation [28] given by (3.7):

$$C_{BC} = C_{JC} = C_{JCo} \left(1 - \frac{V_{BC}}{V_{JC}} \right)^{-M_{JC}} \quad (3.7)$$

where C_{JCo} is the zero bias capacitance, V_{BC} is the base-collector voltage, V_{JC} is the base-collector built-in potential, and M_{JC} is the grading factor. This simple model is valid only for values of V_{BC} less than V_{JC} . Most SPICE programs handle this limitation by using a different capacitance formula for values of V_{BC} greater than V_{JC} . Usually a linear approximation is employed whose slope is equal to the slope of (3.7) at a voltage close to V_{JC} in order to model the depletion capacitance in this region. In the model for this work however, it is assumed that the device will be operated in a region where V_{BC} is always less than V_{JC} so that no change in (3.7) is required. The diffusion component of C_{BC} has not been modeled in this work. Again, this is because the device will be biased in the forward

active region of operation where the base-collector junction is reverse-biased and hence the diffusion capacitance is insignificant. The charge equation corresponding to (3.7) is obtained by integrating (3.7) with respect to V_{BC} and is given in (3.8) [44].

$$Q_{JC} = \frac{V_{JC}C_{JCO}}{(M_{JC}-1)} \left[\left(1 - \frac{V_{BC}}{V_{JC}} \right)^{M_{JC}} - 1 \right] \quad (3.8)$$

The base-emitter capacitance is comprised of a depletion and diffusion component as given in (3.9).

$$C_{BE} = C_{JE} + C_{BE, diff} \quad (3.9)$$

The depletion capacitance equation is similar to (3.7) with C_{JCO} , V_{JC} , M_{JC} and V_{BC} replaced with C_{JEO} , V_{JE} , M_{JE} and V_{BE} .

$$C_{JE} = C_{JEO} \left(1 - \frac{V_{BE}}{V_{JE}} \right)^{M_{JE}} \quad (3.10)$$

The diffusion capacitance is a variation of the SPICE diffusion capacitance formula. It is related to the derivative of the stored forward charge, Q_F , as given by the following: [44].

$$Q_F = T_F I_S \left(\exp \frac{qV_{BE}}{N_F k T} - 1 \right) \quad (3.11)$$

$$C_{BE, diff} = \frac{\partial Q_F}{\partial V_{BE}} = \frac{T_F I_S q}{N_F k T} \cdot \exp \left(\frac{qV_{BE}}{N_F k T} \right) \quad (3.12)$$

The ideal forward transit time, T_F , in its simplest form is considered constant. For many Si BJTs and in GaAs/AlGaAs HBTs it has been found that the diffusion capacitance is a

function of both the base-emitter voltage and the base-collector voltage [45], [58]. An extension of (3.12) is also used in the SPICE Gummel-Poon model in which T_F is not considered a constant but instead is also a function of V_{BC} [28], [44].

$$T_F = T_{FO} \left[1 + X_{TF} \left(\frac{I_C}{I_C + I_{TF}} \right)^2 \cdot \exp \left(\frac{V_{BC}}{1.44 V_{TF}} \right) \right] \quad (3.13)$$

This empirical equation was developed for Si Bipolar transistors and represents the dependence of the base transit time on bias. In HBTs with heavily doped base, the base-width is not a strong function of bias. Furthermore, the transit time increases as the base-collector voltage is made more negative due to the increase in the base-collector depletion layer transit time [45]. Corraera et al. have proposed a simplification of (3.13) for HBTs [45] given by (3.14) which is used in this model.

$$T_F = T_{FO} \left[1 - X_{TF} \cdot \exp \left(\frac{V_{BC}}{1.44 V_{TF}} \right) \right] \quad (3.14)$$

where T_{FO} , X_{TF} and V_{TF} are treated as fitting parameters. A summary of all the HBT model parameters is given in Table 3.1.

Table 3.1 HBT Model Parameters

Parameter	Description
ISO	Collector Saturation Current
ISEO	Base Leakage Saturation Current
ISCO	Collector Leakage Saturation Current
NF	Forward Collector Ideality Factor
NR	Reverse Collector Ideality Factor
NE	Base Leakage Ideality Factor
NC	Collector Leakage Ideality Factor
BFO	Ideal Forward Current Gain

Table 3.1 HBT Model Parameters (Continued)

Parameter	Description
BR	Ideal Reverse Current Gain
RB	Base Resistance
RC	Collector Resistance
RE	Emitter Resistance
CJEO	Zero Bias Base-Emitter Junction Capacitance
VJE	Base-Emitter Built-In Potential
MJE	Base Emitter Grading Factor
CJCO	Zero Bias Base-Collector Junction Capacitance
VJC	Base-Collector Built-In Potential
MJC	Base-Collector Grading Factor
TFO	Diffusion Capacitance Fitting Factor
XTF	Diffusion Capacitance Fitting Factor
VTF	Diffusion Capacitance Fitting Factor
XTI	Third Temperature Parameter
Eg	Base Bandgap Energy
Einf	Parameter Characterizing BF Temperature Dependence
p	Parameter Characterizing BF Temperature Dependence
LB	Base Parasitic Inductance
LC	Collector Parasitic Inductance
LE	Emitter Parasitic Inductance
CBE'	Base-Emitter Parasitic Capacitance
CBC'	Base-Collector Parasitic Capacitance
CCE'	Collector-Emitter Parasitic Capacitance
RTH	Thermal Resistance
CTH	Thermal Capacitance

3.5 Measurement Setup

The modeling procedure described in this work depends heavily on device characterization through electrical measurement. The device measurements thus form the foundation for the remainder of the modeling effort. To perform the necessary curve fitting, DC, AC, and thermal measurements were performed. The following sections describe the setup for each of these measurement including the various types of equipment used. Difficulties encountered during measurement are also discussed.

3.5.1 DC and Small-Signal Measurements

The basic DC and small-signal S-parameter measurement setup is illustrated in Figure 3.5. The HP 4142B modular DC source and monitor was used to apply the necessary bias to the DUT. The output from the 4142B was quadraxial. Quadraxial cables contain four conductors and are often used to perform very accurate DC measurements. The bias input to the Network Analyzer (NA) however, was coaxial. A quadrax to coax adapter was therefore necessary to connect the DC bias supply to the NA. The RF measurements were performed with an HP 8510C NA and a 40GHz S-parameter test set. For DC measurements the computer used was an HP series 900 with the HP IMATM (Interactive Measurement and Analysis) software. For the S-parameter measurements a 486 PC was used with the software WAFER MAP. The DUT was placed on a CK1032 automatic wafer probing station with a Cascade Microtech model 154 top plate. Textronix 0-40GHz ground-signal-ground (GSG) probe heads were used to perform the DC and RF measurements. The computer, bias supply, and network analyzer were all connected using the GPIB bus allowing the data to be automatically retrieved and stored on computer.

The DC measurements performed included standard DC-IV curves with stepped base current drive and forward Gummel plots. The bias supply configurations for these two types of measurements are shown in Figure 3.6.

Some difficulties were encountered when performing these measurements.

Figure 3.5 DC and small-signal S-parameter measurement setup.

It was found that the devices were very susceptible to oscillations. Initially, to perform the DC measurements the output of the quadrax to coax adapter was connected directly to the probe heads bypassing the Network Analyzer, (NA). The DC curves however were very unstable due to low-frequency oscillations. The setup shown in Figure 3.5 was then used with a Wiltron 360 NA instead of the HP 8510C. Similar instabilities were observed. Finally, it was found that applying the bias through the HP 8510C provided much more stable results. The internal HP bias tees seemed to be better at suppressing the low frequency oscillations than those of Wiltron. A similar problem occurred with the Gummel

Figure 3.6 Bias Supply configurations for DC measurements performed on HBT.

plot measurement. Applying the bias through the NA resulted in smooth, stable curves but a problem occurred at low voltage (current) levels. It was observed that the base and collector currents would roll-off and become almost flat at low values of V_{BE} . This was due to a parallel conductance effect. It was found that the HP 8510C NA contains a $1\text{ M}\Omega$ resistor shunted to ground at the output of each port of the S-parameter test set. At low values of V_{BE} the device is practically turned off and presents a large impedance in parallel with the $1\text{ M}\Omega$ resistors. As a result, the current which was measured was that passing through the resistors instead of the device. By bypassing the NA and applying the bias directly to the device, the current at low values of V_{BE} could be measured, however instabilities occurred at higher values of V_{BE} . These two effects can be seen in Figure 3.7. Therefore to cover the desired voltage range, the measurement was made twice. Once with the bias applied directly to the device and once with the bias provided through the NA. The results were then superimposed.

The scattering parameters of the HBT were measured from 0.4-40GHz for I_B from 0 to 1.4mA in 0.2mA steps and V_{CE} from 1.5-4.5V in 0.5V steps using the setup of Figure 3.5. The calibration method used for the S-parameter measurements was the Short-Open-Load-Thru (SOLT) technique [46]. The calibration was performed with a Cascade Microtech Impedance Standard Substrate. The substrate was 0.025" thick and

Figure 3.7 Effect on forward Gummel plot by biasing the DUT directly as opposed to through the NA.

made of alumina. It contained the necessary standards and verification structures realized in coplanar waveguide applicable for the ground-signal-ground probe head configuration.

3.5.2 Thermal Resistance Measurement Setup

The setup for the measurement of the thermal resistance is the same as shown in Figure 3.5 except for the placement of the DUT on the chuck of the probing station. This is shown in Figure 3.8. The device was placed on a copper block which had a hole drilled into the side. K-type thermocouple wire was dipped in silicone thermal grease and inserted into the hole in order to measure the base plate temperature. A Cole-Parmer Digi-Sense thermocouple meter was used to take the temperature readings. The backside of the copper was coated with the thermal grease and placed on top of a Peltier cooler/heater. A Peltier cooler is a semiconductor device whose temperature can be changed by varying the DC bias applied to it. It is normally used as a heat sink in circuit applications. However, by reversing the polarity on the bias supply, it also acts as a voltage controlled heater. By applying the necessary bias voltage to the Peltier cooler, the base plate temper-

Figure 3.8 Setup used for measurement of HBT thermal resistance.

ature could be altered and monitored with the thermocouple meter.

3.6 Parameter Extraction

This section addresses the topic of parameter extraction based on electrical measurement. In this work computer optimization was employed in order to fit simulated to measured transistor responses. However, it is important to determine initial values for as many model parameters as possible before resorting to optimization. Reliance solely on optimization techniques leads to difficulties especially when the model contains a large number of unknowns. This has often been the case where S-parameters are fitted to small-signal models. In these instances, the parameter values validity is not guaranteed since they depend on the starting point and may vary in a nonphysical manner [47]. Furthermore, many solutions may exist which give satisfactory agreement between measurement and simulation. It is very difficult however to extract many of the HBT model parameters accurately from electrical measurement without the use of more elaborate measurement facilities and special test structures. The extraction methods outlined in the following sec-

tions are approximate and are used to guide the optimization procedure by providing initial guesses to *some* of the model parameters. This was attempted in order to reduce the optimization parameter space as much as possible. Many of the parameters however could not be extracted prior to computer optimization and thus had to be obtained solely by optimization methods.

3.6.1 Extraction of Saturation Currents and Ideality Factors

A common method to extract base and collector saturation currents and ideality factors for BJTs is to use the Gummel plot characteristic. This is a plot of base and collector currents as a function of V_{BE} for some constant value of V_{BC} (usually $V_{BC} = 0V$) on a semilog scale. A typical forward Gummel plot for a homojunction BJT is given in Figure 3.9. It can be seen from (3.1) that plotting the ideal diode equation on a

Figure 3.9 Typical forward Gummel plot for a Si homojunction bipolar transistor (after Hafizi et al. [33]).

semilog scale results in the y-intercept being equal to the saturation current and the slope can be used to determine the ideality factor according to (3.15).

$$N = \frac{q}{[\ln(10) \cdot \text{slope} \cdot k \cdot T]} \quad (3.15)$$

Referring to Figure 3.9, one notices a region on the graph where the collector and base currents are parallel. In this region the transistor is exhibiting constant current gain. The existence of this constant current gain region simplifies the extraction procedure for the ideality factors, saturation currents and ideal forward current gain. The two components of base current can be separately identified due to the noticeable change in slope. The component due to space-charge recombination is dominant at lower V_{BE} and the component proportional to I_C is dominant at higher V_{BE} . The parameters for the two base-emitter diodes and ideal current gain can thus be obtained graphically as indicated in Figure 3.9 [33].

A forward Gummel plot for the HBT used in this work is shown in Figure 3.10. These curves are typical for GaAs/AlGaAs HBTs. In this plot there is one very

Figure 3.10 Forward Gummel plot of HBT used in this work.

noticeable difference compared to Figure 3.9. There is no region in which the collector and base currents are parallel, indicating the absence of a constant current gain region.

This complicates the extraction procedure. The collector current is similar to that of the BJT since it is quite linear and exhibits an ideality factor close to unity. Therefore an estimate of the collector saturation current and ideality factor can still be obtained from the y-intercept and slope respectively of the I_C vs. V_{BE} in Figure 3.10. The base current is drastically different than that of a BJT. Usually the dominant component of base current for GaAs/AlGaAs HBTs is the space-charge recombination (SCR) current which is usually characterized by an ideality factor closer to 2 [33]. However this component cannot be separated from the component proportional to collector current since an abrupt change in slope of the base current cannot be observed (see Figure 3.10). As a result the model parameters NE and ISEO must be obtained through computer optimization. A rough initial guess can be made by assuming that at lower V_{BE} the SCR recombination is dominant and then use the graphical technique to obtain an approximate value for ISEO and NE. The parameter BFO (maximum forward current gain) must also be optimized. For BJTs it is a relatively simple parameter to extract since the maximum gain occurs in the constant current gain region of the curve as in Figure 3.9. For the HBT, the current gain increases with V_{BE} but then rolls off due to series resistance and temperature effects. Therefore only a lower limit for BFO, equal to the maximum gain observed on the Gummel plot (see Figure 3.10) can be obtained. A summary of the initial values used for optimization for the parameters BFO, ISO, NF, ISEO and NE is given in Table 3.2.

Table 3.2 Initial Values for Parameters Extracted from Gummel Plot

Parameter	Initial Value
ISO	1.2×10^{-22}
ISEO	5.0×10^{-19}
NF	1.14
NE	1.55
BFO	>32

3.6.2 Extraction of Thermal Resistance

The thermal resistance of a device is the proportionality constant relating the increase in the junction temperature to the power dissipated in the transistor. GaAs/AlGaAs HBTs have been shown to have large values of thermal resistance when compared to Si BJTs and GaAs FETs [5], [11], [41]. This has caused the well known negative slope in GaAs HBT DC-IV curves attributed to self-heating as illustrated in Figure 3.11.

Figure 3.11 DC IV curve of HBT used in this work. The base current is stepped from 0.2mA to 1.4mA in 0.2mA steps.

In power applications it has been found that it is in fact the thermal properties of the HBT instead of its electrical properties which limit its performance [5], [11]. The thermal resistance is thus an important model parameter to extract. There are several methods used to measure thermal resistance. In general however, it can be troublesome to measure the junction temperature of small devices. Optical infra-red (IR) thermometry has limited spatial resolution and liquid crystal techniques such as liquid crystal transition are tedious [48]. In both methods, the observed surface temperature must be related to the internal junction temperature [48]. One of the preferred electrical measurement techniques for transistors is to switch from a high current level with known power dissipation to a low

current level already calibrated [39]. This method is employed in order to remove the feedback effects due to base-width modulation. The disadvantage is that a high enough switching speed must be used so that the device does not cool down significantly [39].

GaAs/AlGaAs HBTs have been shown to possess negligible base-width modulation (i.e. high Early Voltage) due to the high base doping ($> 10^{19} \text{ cm}^{-3}$) [3], [4] [5], [6]. As a result, I_C is independent of V_{CE} in the forward active region for low DC power dissipation. This can be seen in Figure 3.11 which shows a flat, nearly ideal I_C vs. V_{CE} characteristic at low values of I_B for the Rockwell eight finger HBT. At higher current levels, the slope of the I_C curve becomes progressively more negative. This increasingly negative slope is therefore directly the result of device self-heating *without* base-width modulation causing an offsetting factor to the observed slope [39]. This observation leads to simple CW methods to measure the thermal resistance. Dawson et al. [39] and Waltrop [48] both describe a CW method of measuring the thermal resistance. The method of Dawson was used in this work due to its simplicity and is described below.

The measurement setup used for this extraction procedure is shown in Figure 3.8. The method to be described uses the linearity of the device's DC current gain, β , as a function of temperature in order to calculate the thermal resistance. To confirm this assumption, the collector current, I_C , was plotted as a function of base-plate temperature, T_{base} , with constant base current, I_B (Figure 3.12). A linear approximation and a quadratic fit are also shown. It is seen that the linear approximation is quite accurate over the measured temperature range.

The next step was to measure I_C and device power dissipation at different base plate temperatures for V_{CE} from 2-4.5V in 0.5V steps. These curves fall on a series of straight lines as shown in Figure 3.13. If the temperature dependence of β is assumed linear, I_C can be expanded as in (3.16) [39]:

$$I_C(T_j) = \left[\beta_1 + \frac{\Delta\beta}{\Delta T}(T_j - T_1) \right] I_B \quad (3.16)$$

Figure 3.12 I_C vs. base-plate temperature at constant I_B .

Figure 3.13 Plot of I_C versus power at a constant I_B of 1.0mA . The collector current shows a linear dependence on power dissipation which is used to extract the thermal resistance.

where T_j is the average junction temperature, T_1 is the temperature at which the linear expansion is made, β_1 is the current gain at T_1 , and $\Delta\beta/\Delta T$ is the slope of the current gain versus temperature curve at T_1 . I_B is held constant. The junction temperature and thermal resistance are related by (3.17):

$$T_j = T_{base} + R_{TH} \cdot P_{diss} \quad (3.17)$$

where T_{base} is the base plate temperature and P_{diss} is the power dissipated. Substituting (3.17) into (3.16) leaves [39]:

$$I_C(T_{base}, P_{diss}) = \left[\beta_1 + \frac{\Delta\beta}{\Delta T}(T_{base} - T_1) + \frac{\Delta\beta}{\Delta T}R_{TH}P_{diss} \right] \cdot I_B \quad (3.18)$$

From (3.18) it is seen that I_C is a linear function of base plate temperature and internal power dissipation. On a plot of I_C versus P_{diss} at constant T_{base} and constant I_B as in Figure 3.13, the slope is equal to $(\Delta\beta/\Delta T)R_{TH}I_B$. Similarly, a plot of I_C versus T_{base} at a constant P_{diss} and I_B given in Figure 3.14 yields a slope equal to $(\Delta\beta/\Delta T)I_B$.

Therefore R_{TH} can be approximated by dividing the average slope of Figure 3.13 by the slope of Figure 3.14. This calculation gives an R_{TH} of approximately 280°C/W for the Rockwell 8 finger device.

3.6.3 Parameter Extraction from Measured S-Parameters

In this section several equivalent circuit extraction procedures based on measured S-parameters found in the literature are examined. Although some of these methods are used to extract most of the small-signal model parameters it was found that only good initial values for the series resistances R_B and R_E and the parasitic inductance, L_B , could be obtained. As a result the treatment will be focused on the base and emitter series resistances. The applicability to GaAs HBTs of some DC extraction methods used for Si BJTs is also discussed.

Figure 3.14 I_C as a function of base plate temperature at different power levels. I_B is held constant at 1.0mA. The data for this graph is obtained from Figure 3.13.

One common technique used to extract R_B and R_E for homojunction BJTs is to use the Gummel plot characteristic as shown in Figure 3.9 [49]. The base current roll-off at high values of V_{BE} is due only to base and emitter series resistance because of the insignificant emitter conductivity modulation [49]. Therefore the deviation of I_B from the ideal $\exp(qV_{BE}/kT)$ relationship is directly related to the series resistance and can be used to extract R_B and R_E directly. This method however cannot be used for GaAs/AlGaAs HBTs. This is because the I_B characteristic does not display the ideal behaviour that it does for Si BJTs. This is obvious from Figure 3.10. There is no region on this curve in which the base current exhibits the ideal $\exp(qV_{BE}/kT)$ behaviour required as a reference by this method. Furthermore, the deviation of the I_B at high values of V_{BE} is not only caused by series resistance effects for GaAs/AlGaAs HBTs. It is also due to the device temperature increasing due to the self-heating effect. Therefore, the temperature which appears in the exponential function of the ideal diode equation cannot be considered a

constant which complicates the extraction procedure. These two factors prevent the method from being used to accurately extract the base and emitter series resistances for GaAs/AlGaAs HBTs.

Another popular DC technique used to extract the emitter resistance is the open collector method. This method is simple and requires only the measurement of the collector-emitter saturation voltage, V_{CES} , as a function of I_B with the collector open-circuited [49], [50]. The emitter resistance is related to I_B and V_{CES} according to (3.19).

$$V_{CES} = \left(\frac{kT}{q} \right) \ln(\alpha_I) + R_E I_B \quad (3.19)$$

The emitter resistance can thus be obtained from the slope of a V_{CES} vs. I_B curve. It has been typically found for bipolars that the R_E value obtained in this manner is a function of I_B . This contradicts the device physics because R_E is expected to be independent of current due to the high emitter doping. For this reason the accuracy of this method is in question [33], [49], [50].

AC methods are also used extensively to extract model parameters. These methods involve measuring the scattering parameters of the transistor and fitting to an equivalent small-signal model. Optimizing model parameters to fit measured results often leads to difficulties because many solutions can exist which give adequate results. The final values of the parameters are very sensitive to the initial values and hence unrealistic solutions can be found. What is desirable is a simple method which can give good starting values for some parameters before optimization is attempted. Costa et al. give a robust method which yields all of the small-signal model parameters directly from the scattering parameters without the need for optimization [47]. The difficulty with this method is that special test structures are required. These test structures employ the same physical layout as the transistor however the device is replaced with an open and a short circuit. This facilitates the de-embedding of the parasitics from the intrinsic device. Although this

method is very systematic and leads to good results, the special test structures required were not available and hence was not attempted.

The strategy proposed by Maas is straightforward and leads to accurate values for the emitter resistance [51]. The linear model used for this method is the common T-circuit illustrated in Figure 3.15. As observed by Maas, the inverse trans-imped-

Figure 3.15 Small-signal equivalent circuit of HBT (after Maas [51]).

ance parameter, Z_{12} , is predominantly real for HBTs at low frequencies. For example the real and imaginary parts of the measured Z_{12} for the Rockwell 8 finger device used in this work is shown in Figure 3.16. Even as high as 5 GHz, Z_{12} is predominantly real.

With reference to Figure 3.16, because Z_{12} is real it can be related to the emitter resistance by the following simple expression:

$$Z_{12} = R_E + R_{je} \quad (3.20)$$

Furthermore, the emitter current can be approximated by (3.21) [51]:

Figure 3.16 Measured real and imaginary parts of Z_{12} versus frequency. Note that Z_{12} is predominantly real.

$$I_E = I_S \left(e^{\delta V_{BE}} - 1 \right) \quad (3.21)$$

where δ is given by q/nkT . R_{je} is found by taking the derivative of (3.21) with respect to V_{BE} [51].

$$R_{je} = \left(\frac{\partial I_E}{\partial V_{BE}} \right)^{-1} \approx \frac{1}{\delta I_E} \quad (3.22)$$

From (3.20) and (3.22) it is seen that a plot of $\text{Re}[Z_{12}]$ versus $1/I_E$ will be linear with the y-intercept equal to R_E which is represented by the model parameter RE.

The Z-parameters and I_E were determined from the measured S-parameters. A simple program was written in MathematicaTM to accomplish this and is shown in Appendix I. Z_{12} versus $1/I_E$ was plotted for several biases in the active region at a frequency of 0.4GHz. The results are given in Figure 3.17. The curve is linear over the

Figure 3.17 Measured $\text{Re}[Z_{12}]$ vs. $1/I_E$ at 0.4GHz along with linear fit. The y-intercept is equal to the emitter resistance, R_E .

range of emitter and base current indicating that it is independent of bias and agrees well with this simple theory. The y-intercept and hence emitter resistance, R_E , as calculated from Figure 3.17 is approximately equal to 1.37Ω .

Another extraction routine proposed by Pehlke et al. provides most small-signal equivalent circuit parameters directly from S-parameter measurement [52], [53]. The T-model used for this procedure is given in Figure 3.18. It is a simplified model because the parasitic capacitances are not included. In the work by Pehlke, the h-parameters were calculated in terms of the impedance blocks outlined in Figure 3.18. The formulation is simpler however in terms of Z-parameters. The common emitter Z-parameters for the circuit of Figure 3.18 can be written as follows:

$$Z_{11} = z_B + z_{BE} + z_E \quad (3.23)$$

$$Z_{12} = z_{BE} + z_E \quad (3.24)$$

Figure 3.18 Small-signal T equivalent circuit of HBT (after Pehlke [53]).

$$Z_{21} = z_{BE} + z_E - \alpha z_{BC} \quad (3.25)$$

$$Z_{22} = z_{BE} + z_E + z_C + (1 - \alpha) z_{BC} \quad (3.26)$$

The solution for the impedance blocks can be written by rearranging the above equations to get:

$$z_B = Z_{11} - Z_{12} \quad (3.27)$$

$$z_{BC} + z_C = Z_{22} - Z_{21} \quad (3.28)$$

$$z_{BE} + z_E = Z_{12} \quad (3.29)$$

$$\alpha z_{BC} = Z_{12} - Z_{21} \quad (3.30)$$

Since $z_B = R_B + j\omega L_B$, the base resistance R_B , can be extracted from the real part of (3.27) and the base inductance, L_B , from the imaginary part of (3.27). A MathematicaTM program was used for this calculation (see Appendix I). Figure 3.19 shows the

Figure 3.19 Real part of z_B as a function of frequency and base current injection level calculated using the method described in [53].

real part of z_B versus frequency using this method. The imaginary part of z_B versus frequency is shown in Figure 3.20 from whose slope L_B can be calculated. This curve is very linear and shows no significant bias dependence which agrees well with the theory. The slope and hence base inductance was calculated to be approximately 0.1047nH. The accuracy of the extraction of R_B using this method however is in question. The results show a decrease in R_B with increasing injection level which makes sense however the variation from 1.3 Ω to 4.7 Ω seems too large. The order of magnitude of R_B although does seem reasonable. This device was expected to have a small value of base resistance because the contacts are self-aligned. Furthermore the device consists of 8 emitter fingers in parallel (see Figure 3.1) which also reduces the value of R_B . The accuracy of the variation of R_B

Figure 3.20 Imaginary part of z_B versus frequency for different base current levels using the method of [53].

with bias and frequency is doubtful because the parasitic capacitances of the model were not taken into account. It is expected that these parasitics would be fairly large due to the device geometry which contains a large number of interconnects.

The validity of this method is even more in doubt when the extraction of the collector resistance, R_C is performed. In the work by Pehlke, the collector resistance was obtained from the real part of (3.28) at higher frequencies. This is because z_{BC} in series with z_C can be written mathematically as [53]:

$$z_{BC} + z_C = \left[R_C + \frac{r_{BC}}{1 + (\omega r_{BC} C_{BC})^2} \right] + j \left[\omega L_C - \frac{\omega r_{BC}^2 C_{BC}}{1 + (\omega r_{BC} C_{BC})^2} \right] \quad (3.31)$$

The real part of equation (3.31) should simplify to R_C , when $\omega^2 \gg 1/r_{BC} C_{BC}^2$. This however resulted in a negative value for R_C which is of course unrealistic. Other researchers using this extraction procedure have had similar results [54]. This again is most likely

due to large capacitive parasitics not taken into account in the equivalent circuit. Therefore it seems this method cannot be directly applied for devices with large parasitics without some method of separating the intrinsic device from the extrinsic parasitics. This could theoretically be accomplished with special test structures as proposed by Costa [55].

A final method of direct parameter extraction from S-parameters was attempted following the work of Prasad [56], [57]. This technique uses a simplified conventional hybrid- π model shown in Figure 3.21. If the device is operating in the forward

Figure 3.21 Low frequency simplified hybrid- π model of HBT (after Prasad [56]).

active mode with C_{BE} much larger than C_{BC} and neglecting the affect of r_o due to the high Early Voltage of HBTs, the following equations are given in [56]:

$$h_{21} = \frac{h_{feo}}{1 + j\left(\frac{f}{f_1}\right)} \quad (3.32)$$

where h_{feo} is the DC short-circuit current gain and f_1 the 3-dB cutoff frequency. The ratio of the real and imaginary parts of (3.32) may be written as [56]:

$$x \equiv \frac{f}{f_1} = -\frac{\text{Im}(h_{21})}{\text{Re}(h_{21})} \quad (3.33)$$

and solving for h_{feo} ,

$$h_{feo} = \text{Re}(h_{21}) [1 + x^2] \quad (3.34)$$

Furthermore, the real and imaginary parts of h_{11} yield the following two expressions [56].

$$\text{Re}(h_{11}) = R_B + R_E + \frac{r_{be} + h_{feo}R_E}{1 + x^2} \quad (3.35)$$

$$\text{Im}(h_{11}) = -\frac{x(r_{be} + h_{feo}R_E)}{1 + x^2} \quad (3.36)$$

where r_{be} is given by:

$$r_{be} = \frac{h_{feo}kT}{qI_C} \quad (3.37)$$

Solving (3.37) and (3.36) for R_E and R_B respectively yields:

$$R_E = -\frac{\text{Im}(h_{21}) \frac{[1 + x^2]}{x} + r_{be}}{h_{feo}} \quad (3.38)$$

$$R_B = \text{Re}(h_{11}) + \frac{\text{Im}(h_{11})}{x} - R_E \quad (3.39)$$

Therefore the procedure to calculate R_B and R_E is as follows: The measured S-parameters are converted to h parameters. Equations (3.33), (3.34), and (3.37) are used to compute x , h_{feo} , and r_{be} . Equation (3.38) is used to calculate R_E which is substituted

into (3.39) to calculate R_B . It is mentioned in [56] that this calculation should be performed at a frequency where the real and imaginary parts of h_{21} are comparable, that is where x is approximately equal to 1. Prasad also states that this method works best at moderate bias levels. Note also that in the calculation of r_{be} , the temperature was assumed to remain at 300K in [56]. Here, the temperature was assumed to vary according to power level as in (3.3), using the calculated value of R_{TH} from Section 3.6.2. The program used to calculate R_B and R_E is shown in Appendix I.

The values of R_B and R_E calculated using this technique are indicated in Table 3.3 and Table 3.4 respectively at medium levels of base current drive and well in the active region...

Table 3.3 Calculated Values of R_B

$V_{CE} \backslash I_B$	0.6	0.8	1.0
2.5	4.30	5.40	4.35
3.0	5.16	4.17	4.10
3.5	4.15	4.30	4.27
4.0	4.71	4.79	4.17
4.5	4.30	4.33	4.39

Table 3.4 Calculated Values of R_E

$V_{CE} \backslash I_B$	0.6	0.8	1.0
2.5	1.61	1.54	1.45
3.0	1.63	1.50	1.44
3.5	1.61	1.50	1.43
4.0	1.62	1.50	1.44
4.5	1.61	1.50	1.44

The calculated R_E is approximately 1.4-1.6 and agrees well with the value obtained using the method proposed by Maas outlined previously. The calculated R_B is 4-5 Ω and doesn't show the strong bias dependence that was found earlier using Pehlke's method.

The extraction routines discussed in which parameters are extracted solely from S-parameter data were capable of providing only rough approximations for the base and emitter resistances and other model parameters could not be extracted. This is due to a number of factors. The device being modeled is composed of 8 single emitter finger devices connected in parallel. The large number of interconnects present increases the value of, and the distributed nature of the parasitics. In these methods the equivalent circuits are simplified in order to be able to obtain analytical solutions. In doing so, the parasitics are ignored which leads to inaccuracies for the extracted parameters. The above extraction procedures however yield good initial values which can be used in the optimization procedure to follow.

3.7 Computer Optimization

The model contains a large number of parameters, most of which are difficult to extract only from electrical measurement. As a result, optimization techniques were used to fit the simulated to the measured responses. Initially, DC optimization was performed until good agreement was reached between simulation and measurement. Small-signal AC optimization was then performed at multiple biases in the active region of operation of the transistor. This was used to determine the bias independent parasitics, the base-emitter capacitance, the base-collector capacitance, and the forward delay. The capacitances values obtained were then fit to the appropriate equations to complete the model. The complete optimization procedure is explained in detail below.

3.7.1 DC Optimization

The DC parameters to be optimized included ISO, ISEO, ISCO, NF, NC, NE, NR, BFO, BR, Einf, XTI, RTH, RB, RE, and RC (see Table 3.1). Initial values were entered for the parameters previously extracted and their optimization range was tightly constrained. Other parameters whose values were not known were allowed to optimize over a wide range. To further reduce the number of unknowns, DC optimization was first applied to the forward active region of the device. In this region the only parameters affecting simulation were ISO, ISEO, NF, NE, BFO, RE, RB, XTI, RTH, and Einf. The other parameters were held constant. Optimization was then performed in the saturation region in which ISCO, BR, NR, NC, and RC were allowed to vary. The other parameters were fixed at the values previously obtained during the forward active region optimization.

The measured data used in the optimization was the I_C versus V_{CE} at constant base current drive characteristic and the corresponding V_{BE} versus V_{CE} characteristic. The optimization objective was to match the measured and simulated I_C and V_{BE} over the measured range. A combination of optimizers was used including Random, L1, and Simplex. All optimization was performed with the software package OSA90/HopeTM V2.5. An example of the netlist file used for DC optimization is shown in Appendix II.

The results of the DC optimization are indicated in the following diagrams. Figure 3.22 shows the simulated versus measured I_C - V_{CE} curve. The necessary V_{BE} corresponding to each point in Figure 3.22 is given in Figure 3.23. Note that in order to represent the device characteristics more accurately both the I_C vs. V_{CE} and V_{BE} vs. V_{CE} characteristics need to be fitted. Finally, the simulated versus measured forward Gummel plot is shown in Figure 3.24. These results show that the DC characteristics of the HBT are represented very well by the model.

Figure 3.22 Simulated and measured I_C vs. V_{CE} curve (TOP). Zoomed in view of reverse active region (BOTTOM).

3.7.2 Small-Signal Optimization

S-parameter characterization of the device was performed in order to obtain the data for small-signal optimization. The S-parameters were measured from 0.4GHz to 40GHz at a large number of bias points in the active region of operation of the

Figure 3.23 Simulated and measured V_{BE} vs. V_{CE} characteristic.

Figure 3.24 Simulated vs. measured forward Gummel plot of HBT.

device. The bias points used in optimization included V_{CE} from 1.5 to 4.5V in 0.5V steps and I_B from 0.2mA to 1.4mA in 0.2mA steps. This covered the active region of the device. The measurement setup is explained in detail in Section 3.5.

Optimization used to fit simulated to measured S-parameters has the potential to cause problems when the number of unknowns in the circuit is large [55]. As has been discussed, the solutions can be sensitive to starting points and the parameters may vary in an unrealistic manner. The optimization method used attempts to reduce as much as possible, the number of unknowns optimized simultaneously. Since the DC model has already been completed, the only elements remaining to be optimized are the base-emitter capacitance, base-collector capacitance, forward transit delay and the parasitics. In order to determine the variation of these parameters with bias, each parameter was defined as a matrix in the OSA90/Hope netlist (see Appendix II). These matrices were designated CBC[I,J], CBE[I,J], and TAU[I,J]. Each column of the matrix represents a change in I_B for a constant V_{CE} and similarly each row represents the variation of V_{CE} for a constant I_B . Initially three bias points in the active region were optimized *simultaneously*. These points were chosen at low, medium, and high values of base current drive. The parasitics and three elements in each of the matrices CBC, CBE, and TAU were allowed to vary. After good agreement was obtained at these three points, the parasitics were fixed. Each bias point was then optimized independently and successively with only one corresponding element in each matrix allowed to vary. This was done for all bias points in the active region which were measured. At the completion of the optimization process, the bias variation of these parameters was determined.

The variation of C_{BC} with bias is indicated in Figure 3.25. The C_{BC} is composed of a depletion component only since the base-collector junction is reverse biased. The curve shows the expected variation of C_{BC} with V_{BC} for a depletion capacitance. There is also a variation with base current drive although it is less substantial. This variation could be due to a temperature dependence of the depletion capacitance as sug-

gested in [59].

For simplicity this capacitance was fit to the standard SPICE equation given by (3.7). This was done by creating a new file in OSA90/HopeTM and optimizing the parameters CJCO, MJC, and VJC of (3.7). The optimization objective was to minimize the error between the capacitance calculated from (3.7) and the optimized values from Figure 3.25 over the bias range. Since in (3.7) C_{BC} is only a function of V_{BC} , the variation of C_{BC} with I_B in not included in the curve fit shown in Figure 3.25.

Figure 3.25 Optimized variation of base-collector capacitance, C_{BC} , with bias.

The base-emitter capacitance variation with bias is shown in Figure 3.26. This capacitance is composed of a diffusion and depletion component, according to (3.10) and (3.12). From these two equations it is seen that C_{BE} is a function of V_{BE} , V_{BC} , and temperature, T . In order to perform the curve fitting, V_{BC} , V_{BE} , and T were determined at each of the bias points from a DC simulation in OSA90/HopeTM. These values were entered into a matrix in a new netlist file. The parameters TFO, XTF, VTF, CJEO, VJE,

MJE (see Table 3.1) were optimized until good agreement was obtained. This curve fit, which is the sum of equations (3.10) and (3.12) is shown in Figure 3.26 and agrees well with the optimized values over the entire bias range.

Figure 3.26 Optimized base-emitter capacitance, C_{BE} , as a function of bias along with curve fit.

Finally, the variation of the delay term, τ_{AU} , with bias is shown in Figure 3.27. The delay increases with increasing V_{CE} . This trend makes sense physically because as V_{CE} is increased, the base collector junction is more strongly reverse biased. This results in a widening of the base-collector space charge region which in turn increases the time taken for carriers to traverse it [3]. In all simulations, the delay term is considered constant at the value obtained through small-signal optimization.

The results of the small-signal optimization are shown in Figure 3.28, Figure 3.29, and Figure 3.30. These graphs show the simulated vs. measured S-parameters at three bias points in the active region. The results show good agreement from 0.4-40GHz, and are representative of the fit between simulation and measurement over the entire

Figure 3.27 Variation of forward delay with bias.

active region. One can observe on the Smith charts however, a phase error in S22. The reason for this is explained as follows. When fitting the base-emitter capacitance variation with bias as shown in Figure 3.26, the capacitance equations (3.10) and (3.12) were used. However the simulated results shown in Figures 3.28-3.30 use the corresponding charge equations. The charge formulation is required for harmonic balance simulation which is used in OSA90/HopcTM. Therefore in order to use *one* model for DC, small-signal, and nonlinear design and analysis, the charge formulation instead of nonlinear capacitance must be employed. This presents no problem for the base-emitter and base-collector depletion capacitances since they depend on only one variable. The problem occurs with (3.12) where the charge is a function of two voltages, V_{BC} and V_{BE} . It therefore cannot be represented exactly by a single diffusion capacitance as is done in SPICE and was done here for small-signal fitting purposes. To reduce the error in the phase of S22 the model should be reoptimized using the charge equations. This was not done because it would

Figure 3.28 Simulated vs. measured S-parameters from 0.4GHz to 40GHz. The bias point is $I_B = 0.2\text{mA}$ $V_{CE} = 1.5\text{V}$.

Figure 3.29 Simulated vs. measured S-parameters from 0.4GHz to 40GHz. The bias point is $I_B = 0.8\text{mA}$ $V_{CE} = 3.0\text{V}$.

Figure 3.30 Simulated vs. measured S-parameters from 0.4GHz to 40GHz. The bias point is $I_B = 1.4\text{mA}$ $V_{CE} = 4.5\text{V}$.

require that *all* the AC parameters be optimized simultaneously which leads to convergence difficulties, long optimization times, and inability of the optimizers to find the minimum of the error function.

The final values for all model parameters are given in Table 3.5.

Table 3.5 Final Values of HBT Model Parameters

ISO = 2.207×10^{-22} A	RE = 1.35 Ω	p = -1
ISEO = 5.442×10^{-19} A	CJE0 = 0.31385 pF	Eg = 1.43 eV
ISCO = 5.605×10^{-13} A	MJE = 0.33	Einf = 0.0764 eV
NF = 1.1685	VJE = 1.7136 V	LB = 0.1051 nH
NR = 1.1052	CJC0 = 0.11352 pF	LC = 0.0829 nH
NE = 1.5779	MJC = 0.5	LE = 0.002 nH
NC = 2.0	VJC = 1.2638 V	CBE' = 0.0629 pF
BFO = 63.728	TF0 = 8.364 PS	CBC' = 0.1048 pF
BR = 0.10935	XTF = 0.6422	CCE' = 0.0698 pF
RB = 4.4047 Ω	VTF = 3.8263	RTH = 282 $^{\circ}\text{C/W}$
RC = 1.0973 Ω	XTI = 4.398	CTH = ∞

3.8 Large-Signal Measurement and Results

With the DC and S-parameter optimization completed, the model needed to be verified under large-signal excitation. To this end, two types of measurements were performed. The first was a measurement of P_{out} vs. P_{in} at two bias points corresponding to Class A ($I_B=0.8\text{mA}$, $V_{CE}=3.0\text{V}$) and Class AB ($I_B=0.2\text{mA}$, $V_{CE}=4.0\text{V}$) operation at 14GHz. The source and load were both terminated into 50 Ω . The measurement setup is shown in Figure 3.31. An RF source, the HP8350B, was used to supply power to the DUT. The output power was read on an HP437B digital power meter with a HP8487A power sensor. The bias was applied through bias tees using the 4142 DC modular source. The input and output losses were de-embedded by separate measurement of the insertion

Figure 3.31 Setup used to measure the P_{out} vs. P_{in} characteristic of the HBT.

loss of the cables, adapters, bias tees, and microwave probe heads.

The results of this measurement are shown in Figure 3.32 for Class A bias and Figure 3.33 for Class AB bias. These curves show the measured response and the simulated response using OSA90/HopeTM V2.5. The measured small-signal transducer gain in a 50Ω system, given by $20\log_{10}(|S_{21}|)$ is also indicated on the graph. This was obtained from previous small-signal S-parameter measurements. The S_{21} was extrapolated from small-signal conditions and appears as a straight line (constant gain) in Figure 3.32 and Figure 3.33. In both cases, the simulated gain in the linear region (low input power) is lower than the measured value by approximately 1dB. This is most likely due to inaccuracies in de-embedding of the losses of the cables and connectors. The reason the measurement accuracy is suspected and not the model is because the *measured* small-signal S_{21} , shown in the figures as a dotted line agrees very well with the simulated value at both biases. At higher power levels, the simulated gain rolls off at a much slower rate than the measured value. This indicates that the nonlinearities of the device have not been repre-

Figure 3.32 P_{out} vs. P_{in} characteristic of Rockwell HBT in Class A operation. The plot shows the measured and simulated response as well as the small-signal S_{21} linearly extrapolated to higher power levels.

Figure 3.33 P_{out} vs. P_{in} characteristic of Rockwell HBT in Class AB operation. The plot shows the measured and simulated response as well as the small-signal S_{21} linearly extrapolated to higher power levels.

sented accurately by the model and requires further work in order to improve the response. This is discussed in more detail in chapter 5.

The other measurement obtained was a load-pull characterization of the device. This measurement is used to characterize the performance of a device when the load and/or source is terminated in impedances other than 50Ω . This technique is often used in the design of power amplifiers. This type of measurement involves some manner of changing the load presented to the DUT and measuring the output power and load impedance. The two main setups include passive load-pull and active load-pull systems. Passive systems use some form of mechanical tuner which is manually or automatically tuned in order to vary the load impedance. Active load-pull systems inject a signal back into the output of the DUT of varying amplitude and phase in order to simulate different load reflection coefficients to the DUT [60]. Active systems have the advantage of being able to realize load reflection coefficients, Γ_{load} , greater than or equal to unity. Passive systems cannot produce a Γ_{load} close to unity due to losses in the connectors and the tuner itself. Thus active systems can cover a greater range of impedances on the Smith chart than passive systems. Active systems are also more versatile. For example multi-harmonic load-pull setups can be realized fairly easily with active systems [61]. This type of system injects signals of varying amplitude and phase at the fundamental and a number of harmonics. In this way the effect of different load terminations at harmonic frequencies can be determined.

The load-pull measurement of the Rockwell HBT was performed at Ecole Polytechnique in Montreal, Canada. To prepare the device for this measurement a square hole was drilled into a 10mil Alumina substrate which contained input and output 50Ω microstrip lines. The HBT was placed in the hole and then the substrate and device were epoxied to a gold plated brass carrier. Bond wires were used to connect the device to the input and output microstrip lines. The measurement setup for the load-pull characterization is shown in Figure 3.34. In order to ensure the device was operating under nonlinear

2 of/de 2

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

PRECISIONSM RESOLUTION TARGETS

Figure 3.34 Ecole Polytechnique's active load-pull setup used to measure the Rockwell 8 finger HBT.

conditions, the 1-dB compression point with 50Ω terminations was first determined. This was found to occur at an input power of 8.0dBm by Ecole Polytechnique. The RF power absorbed by the transistor was then maintained at 8.0dBm for the duration of the load-pull measurement. The frequency of operation was 14.0GHz. The results are shown in Figure 3.35 which shows the contours of constant output power on the Γ_{load} plane. The crosses correspond to actual measurement points. The power contours are then determined by a computer algorithm which interpolates between measurement points in order to draw the contours.

An attempt was made to simulate the load-pull characteristic using OSA90/HopeTM V2.5. This software allows the load and source terminations to be altered by the user. Some difficulties were encountered in this area. The first was that the RF power absorbed had to be kept constant at 8.0dBm. In order to do so, the available

Figure 3.35 Measured constant output power contours of Rockwell 8 finger HBT at 14GHz. The transistor was biased at $I_B = 0.8\text{mA}$ $V_{CE} = 3.0\text{V}$.

power from the source was increased until the RF power absorbed by the device was equal to 8.0dBm. This had to be done individually for each of the load terminations that were measured which was time consuming and inefficient. An uncertainty in the simulation regarding harmonic terminations was also encountered. The load-pull measurement system uses narrow band 6-port networks to measure input and output power. At other than the measurement frequency of 14.0GHz these 6-ports present a reflective termination to the device which was unknown. Consequently, during simulation the same impedance was presented to the transistor for all frequencies. Finally, problems with convergence occurred with the load-pull simulations. For some of the terminations the harmonic balance simulation was not able to converge to a solution. Furthermore, "false convergence" sometimes occurred in which the solution was not realistic or repeatable. The convergence was also found to be sensitive to the input power level. The reason for these con-

vergence difficulties is probably due to the fact that the model topology and equations are user-defined. As a result, the program must calculate the necessary derivatives of the non-linear elements numerically. With built-in models the equations are known *a priori* and thus derivatives can be calculated analytically which eases the convergence process.

The simulated and measured output power for different loads is given in Table 3.6. The R_{load} and X_{load} values shown in Table 3.6 are normalized to 50Ω . As can

Table 3.6 Measured vs. Simulated P_{out} for Various Load Terminations

R_{load}	X_{load}	Pout Simulated [dBm]	Pout Measured [dBm]	Error [dB]
0.384	-0.167	13.37	11.72	1.65
0.596	0.209	15.75	13.90	1.85
0.771	0.496	15.37	13.95	1.42
1.063	0.922	13.81	12.93	0.88
1.926	2.079	10.11	9.73	0.38
9.19	6.873	4.21	3.88	0.33
2.668	0.072	10.88	10.25	0.63
2.146	-1.647	9.11	8.43	0.69
1.344	-1.646	8.75	7.87	0.88
0.805	-1.315	8.65	7.82	0.83
0.575	-0.912	9.41	8.52	0.89
0.548	..612	10.85	9.86	0.99
0.582	-0.395	12.30	11.19	1.11
0.640	-0.184	13.63	12.31	1.31
0.743	0.040	14.63	13.28	1.36
1.205	0.418	14.02	13.11	0.91
0.887	0.207	14.90	13.52	1.38
0.333	0.137	15.62	13.38	2.24

Table 3.6 Measured vs. Simulated P_{out} for Various Load Terminations (*Continued*)

R_{load}	X_{load}	Pout Simulated [dBm]	Pout Measured [dBm]	Error [dB]
0.226	-0.038	13.21	11.50	1.71
0.143	-0.133	10.76	9.28	1.48
0.061	-0.227	6.58	5.05	1.53
0.178	0.057	13.56	11.80	1.76
0.303	-0.252	11.80	10.65	1.15
0.179	-0.442	8.53	7.49	1.04
0.097	-0.576	4.99	4.12	0.87

be seen in the table, the model consistently overpredicts the output power. The error ranges from approximately 0.3 to 2.25dB. The terminations which did not converge are not shown. The fact that the model overestimated the output power makes sense since it was seen in Figure 3.32 that the simulated P_{out} vs. P_{in} curve was more linear and rolled off at a slower rate than the measured curve. Although most of the discrepancy is probably due to the model's inaccuracy under nonlinear excitation other sources of error can also be considered. For example the device which was sent to Ecole Polytechnique for measurement was not the same one which was modeled. Some error could be due to the nonuniformity of different devices. Some uncertainty also arose in de-embedding the effect of one of the connectors. Any phase error in de-embedding this connector would result in a rotation of the curves shown in Figure 3.35 which may also be responsible for some error. Finally as already mentioned, the terminations during simulation were kept constant for all frequencies. This was not true during measurement because the 6-ports were narrow band. However, the harmonic terminations in the measurement were unknown and thus could not be reproduced during simulation. This could also account for some of the discrepancy between simulation and measurement.

3.9 Summary

This chapter has presented the empirical modeling procedure for a GaAs/AlGaAs HBT [62], [63]. The model was based on the SPICE Gummel-Poon BJT model. Several modifications were made to the SPICE model as suggested in the literature in order to better represent GaAs/AlGaAs HBT characteristics. These included self-heating, transit time effects, and variation of diffusion capacitance with V_{BC} . The model was implemented in the commercial microwave CAD package OSA90/HopeTM V2.5 and can be used to design microwave circuits using the harmonic balance method.

Several parameter extraction techniques were performed in order to obtain initial values for the model parameters. Parameters were extracted from DC, small-signal, and thermal measurements. Due to the large number of model parameters, computer optimization was also used to curve fit simulated and measured responses. The optimization was first performed on DC data and then S-parameter data at multiple bias points. This procedure was adopted in order to reduce the number of unknowns being optimized simultaneously. To complete the model the optimized variation of the base-emitter and base-collector capacitances was fit to the appropriate equations.

The simulated DC responses including I_C vs. V_{CE} , V_{BE} vs. V_{CE} , and forward Gummel plot agreed well with measurement. The simulated and measured S-parameters also showed good agreement from 0.4-40GHz over most of the active region of operation of the device. The simulated load-pull characteristics however, overestimated the measured results by 0.3-2.2dB which requires further investigation.

CHAPTER 4: OSCILLATOR DESIGN

4.1 Introduction

One of the main applications of GaAs/AlGaAs HBTs is low-phase noise microwave oscillators. This is due to their high speed operation (f_t and f_{max} 50-200GHz [3], [5]) and excellent low frequency noise characteristics. This low frequency noise is upconverted by the nonlinearities of the oscillator and appears as phase noise at the fundamental frequency of oscillation. The noise corner is an important quantity when comparing the low frequency noise properties of transistors. It is defined as the baseband frequency at which the excess noise intersects the background white noise. Advanced Si BJTs still possess the lowest noise corners with typical values of 10kHz. The noise corner of GaAs/AlGaAs HBTs is lower than 1MHz while that of FETs is 10 to 100 times higher [4]. Presently GaAs/AlGaAs HBTs have shown approximately 10dB improvement in phase noise over FET type oscillators with similar resonators [10]. When low phase noise was required at lower microwave frequencies, the silicon homojunction bipolar transistor was the device of choice. The GaAs/AlGaAs HBT now gives the oscillator designer the low phase noise characteristics of bipolar technology at the microwave and millimetre wave frequencies.

In order to further evaluate the model it was used in a practical application. This chapter deals with the first iteration design, fabrication and measurement of a simple HBT oscillator. As the primary purpose of the circuit is to further evaluate the validity of the model, no attempt was made to design for any specifications. The small-signal, large-signal, and bias circuit design are described. The complete design was performed using OSA90/HopcTM V2.5. The measurement setup used to determine the oscillation frequency and output power is then outlined. The chapter then concludes with a discussion of the measurement results.

4.2 Design Procedure

The oscillation frequency for this application was 14.0GHz. The device bias point was chosen for class A operation at $I_B=0.8\text{mA}$ and $V_{CE}=3.0\text{V}$. The first step in the design was to choose the topology of the circuit. For oscillator applications all three possible transistor orientations; common emitter, common base, and common collector can be used. The common base configuration is usually able to produce a broad band negative output resistance and is popular for VCOs. The common collector configuration is often used because it is generally unstable without the need for external feedback which simplifies the circuit layout. Since the modeling procedure was performed with the transistor in common emitter mode, the circuit was designed with this transistor configuration.

The next step was to determine the stability factor, K , at the designated bias point. This was calculated from an AC simulation in OSA90/HopeTM as stated in Section 2.5 and shown in Figure 4.1. In order to be used in an oscillator application, K should be less than unity at the oscillation frequency. As can be seen from Figure 4.1 the device is potentially unstable at low frequencies and is stable at frequencies above 12 GHz. Feedback must then be added to the device in order to make it potentially unstable at 14.0GHz. There are several methods of doing this as explained in Section 2.5.3. A simple series feedback arrangement was chosen because the circuit was fabricated using MIC technology due to time constraints. The feedback was implemented with a short-circuited microstrip transmission line from emitter to ground. A short-circuited line rather than an open-circuited line is required in order not to disturb the DC operating point of the device. Next, an open-circuit microstrip line is added to the base of the transistor. This is used to produce a large negative resistance at the output, or collector port of the transistor. For simplicity, the output load circuit for the oscillator was kept at 50Ω so that no additional output match circuitry was required.

Figure 4.1 Stability factor, K , as a function of frequency. The bias point is $I_B = 0.8\text{mA}$ $V_{CE} = 3.0\text{V}$.

The two lengths of microstrip lines were optimized such that the following conditions were met at 14.0GHz.

$$\begin{aligned} R_{out} &\leq -100\Omega \\ X_{out} &= 0\Omega \end{aligned} \tag{4.1}$$

The microstrip lines were modeled using the built-in microstrip elements found in the OSA90/HopeTM V2.5 library [26]. The bond wires were modeled as ideal inductors of 0.15nH. This inductance was calculated based on the rule of thumb that the inductance per unit length of the bond wire is approximately 0.8nH/mm. Since two bond wires are usually connected to each pad this value is reduced by approximately one half to 0.4nH/mm. The length of the wire was calculated to be 0.38mm based on an estimate of the physical layout of the circuit. Since there is no built-in model for via holes in the OSA90/HopeTM V2.5, the via hole was also modeled as an ideal inductor of approximately 0.07nH. The basic oscillator topology is shown in Figure 4.2. The optimized out-

Figure 4.2 Basic oscillator topology showing optimized line lengths in order to produce a large negative resistance at the output.

put resistance and reactance at the output (collector) terminal of Figure 4.2 are shown in Figure 4.3. As shown in the Figure 4.3, the output resistance is equal to -105Ω and the

Figure 4.3 Optimized output resistance and reactance. The design frequency is 14.0GHz.

reactance is equal to 0 at 14.0GHz. The optimized emitter line length and base open-circuited line length were 3203 μm and 2052 μm respectively. These lines both have a characteristic impedance of 50 Ω . This forms the initial RF design of the oscillator.

The design of the DC bias circuit was then performed. The bias circuit consisted of a high characteristic impedance quarter-wave length transmission line followed by a low impedance quarter-wave length open-circuited line. These line lengths were optimized in order to provide an effective open-circuit at plane A-A' and a short-circuit at plane B-B' at the oscillation frequency as shown in Figure 4.4. By doing so, the

Figure 4.4 Oscillator bias circuit.

bias circuit will be theoretically transparent to the RF operation of the circuit at the design frequency. The DC bias is fed in at the interface between these two lines with a high impedance line connected to a bonding pad where the DC bias was actually applied. Also at this interface a 50 Ω resistor and 100pF capacitor were connected in series to ground through a via. This R-C network is used in order to suppress low frequency oscillations [15]. The bias circuitry described is connected in shunt at plane A-A' to the base stub. A

similar bias circuit it connected to the output 50Ω line in the circuit.

The *complete* oscillator circuit topology was then entered into the OSA90/Hope™ V2.5 netlist shown in Appendix III. A 10pF DC blocking capacitor was added to the output. The oscillator topology is illustrated in Figure 4.5. To perform the large-signal design, the OSCPORT element was connected to the output of the circuit as described in Section 2.5.4 and shown in Figure 4.5. The following specification was used in the optimization at the design frequency of 14.0GHz:

$$\begin{aligned} \text{OSC_GAIN} &= 0 \\ \text{OSC_PHASE} &= 0 \end{aligned} \quad (4.2)$$

OSC_GAIN and OSC_PHASE are defined in (2.17). In order for this criterion to be met, one or more parameters in an oscillator must be allowed to vary. In this case, the length of the base open-circuited stub was optimized as well as the estimate of the output voltage of the circuit, V_s . The optimization was carried out with a combination of the LI, random, and simplex optimizers. Two problems which were encountered during the course of optimization were convergence difficulties and the inability of the optimizer to meet the oscillation conditions of (4.2). A number of “false convergences” also occurred. What is meant by this is that many times the program would optimize the circuit without any convergence error messages. The error function would be reduced to practically zero indicating the oscillation conditions had been met. However, when subsequent simulations were performed with all the circuit element values remaining constant as determined by the optimizer, it was found that OSC_GAIN and OSC_PHASE were not equal to zero! Furthermore if any further optimization was performed from that point, the error function, instead of being equal to zero as had been previously obtained, would restart at a large value. When these types of problems occurred, the optimization was restarted with a different initial value for V_s and the length of the base stub was also reset to the value obtained from small-signal optimization. The large-signal optimization was repeated until

Figure 4.5 Complete HBT oscillator topology as entered into OSA90/Hopc™ V2.5 for oscillator design using the OSCPORT element.

the conditions of (4.2) were met and no other convergence difficulties were encountered. After the optimization was completed, the following values were obtained:

$$\begin{aligned} \text{OSC_GAIN} &= -0.0041\text{dB} & \text{OSC_PHASE} &= 0.0011^\circ \\ V_s &= 1.268\text{V} & \text{Base Stub} &= 502.7\mu\text{m} \end{aligned}$$

The simulated output power spectrum of the circuit is shown in Figure 4.6. The fundamental frequency output power was approximately 12.0dBm and the second harmonic is below the fundamental by more than 22dB. The time domain output voltage

Figure 4.6 Simulated output power spectrum of HBT oscillator.

waveform is given in Figure 4.7 which was obtained using the built-in inverse FFT function in OSA90/HopeTM V2.5.

The oscillator circuit was laid out using the software package DW2000 on a Macintosh Quadra computer. The glass plate masks were obtained from Shaw Photogrammetric and the fabrication was carried out using the in-house facilities available at

Figure 4.7 Time domain output voltage waveform of HBT oscillator.

CRC. The circuit was fabricated on a 10mil alumina substrate with 2 μ m thick gold microstrip lines. The transistor, chip capacitors, chip resistors, and DC bias wires were connected to the circuit using silver epoxy. The substrate was also epoxied to a gold plated brass carrier. A photograph of the complete oscillator circuit is shown in Figure 4.8.

4.3 Measurement And Results

The circuit was mounted in the Wiltron Universal Test Fixture as shown in Figure 4.9. The setup used to measure the oscillation frequency and output power is illustrated in Figure 4.10. The DC bias was applied using the 4142 modular source and controlled via an HP computer with the IMA software package.

The transistor mounted in the first circuit was the same device which was used in the modeling procedure described in chapter 3. When the circuit was powered up, it was found that the bias point was approximately half of what was expected. The bond

Figure 4.8 Photograph of MIC HBT oscillator.

Figure 4.9 Photograph of oscillator mounted in Wiltron Universal Test Fixture.

Figure 4.10 Test setup for oscillator measurement.

wires on the circuit were removed and the transistor was probed in order to perform a DC test. It was found that the current gain was much lower than in previous measurements which indicated some damage to the transistor had occurred during the fabrication process. One possible reason for this may have been the exposure of the transistor to a high temperature when it was epoxied to the substrate. Another circuit was then fabricated with a similar device. For this circuit, the transistor and other components were epoxied at a temperature of 50°C and cured for 8 hours instead of 150°C for 5 minutes as was done previously. After the circuit was built, the transistor was DC tested before the bond wires were connected. It was found that this device showed a normal DC response.

Through the course of device measurement it was found that these transistors were very susceptible to instabilities due to low frequency oscillations. This was noticed when DC measurements were performed as was explained in Section 3.5. Whenever DC bias was applied to the transistor it had to be done through a bias tee in order to obtain meaningful results. The bias tees provided a good low frequency match which

removes the bias fluctuations. This behaviour was also noted when attempting to measure this circuit. Various methods were tried in an attempt to stabilize the DC operating point. Shunt bias decoupling capacitors of various sizes were connected to the base and collector lines but no improvement was noticed. Filter-cons were also connected to the bias lines but these also did not improve stability. Ferrite beads of different sizes were also inserted around the bias wires. Such beads are often used to remove instabilities by absorbing low frequency RF energy. This also had little affect on the stability. Although not practical, the DC bias had to be applied through external bias tees for testing purposes.

Testing the circuit proved to be quite difficult. The circuit was extremely sensitive to the loading provided to it. The DC operating point often became unstable and results were not always reproducible. It was found that the spectrum analyzer had to be connected to the -13dB arm of the coupler in order to reduce its loading effect on the circuit. If the spectrum analyzer was connected to the through arm of the coupler, the DC bias would fluctuate. It was also found that a wide band coupler was required in order to get a more stable DC operating point. A slight movement of the setup or a change in the amount of torque applied to a 50Ω load connected to the output would affect the stability of the Q point of the circuit.

Using the setup shown in Figure 4.10, the output spectrum of the oscillator is illustrated in Figure 4.11. The oscillation frequency was approximately 13.0GHz. The output power was determined using the power meter as shown in Figure 4.10 and de-embedding the loss of the cables, adapters and coupler. This loss was calculated by separate S-parameter measurement. The power was calculated to be 11.1dBm. When the scale was expanded however, it can be seen in Figure 4.12 that the spectrum was irregular. This was due to the fact that the DC operating point was fluctuating. It was found that adjusting the torque on the connector and terminations improved the stability of the DC operating point. The output spectrum then changed to that shown in Figure 4.13. As can be seen, the oscillation frequency was lowered to 12.7GHz. The power also increased to

Figure 4.11 Output spectrum of oscillator.

Figure 4.12 Expanded view of oscillator output spectrum.

12.2dBm. The output spectrum shown in Figure 4.13, is much improved over that shown Figure 4.12. This result however was impossible to reproduce at a later date.

In an attempt to stabilize the operating point in a reproducible fashion, a triple stub tuner was connected to the output of the oscillator circuit at plane X-X' as shown in Figure 4.10. The arms of the tuner were adjusted until a stable operating point and

Figure 4.13 Output spectrum of oscillator with more stable DC operating point.

clean output spectrum resulted. It was found that adjusting the tuner over the full range only produced approximately three different oscillation frequencies. These occurred at 12.2GHz, 12.5GHz, and 13.0GHz. The tuning had more of an effect on changing the output power of the oscillator. The cleanest spectrum was produced at 12.5GHz. Only a small change in the length of the middle arm of the tuner was sufficient to produce a clean output spectrum. Increasing the length of the middle arm increased the output power. The oscillator was tuned for the maximum output power at approximately 12.5GHz. The output spectrum is shown in Figure 4.14. Unfortunately, during testing the transistor was damaged and had to be replaced. The circuit was then retested with a new transistor using the same tuning arrangement. The frequency response was very similar to that shown in Figure 4.14. A wide band output and close up of the tuned circuit is given in Figure 4.15 and Figure 4.16 respectively. The remaining measurements were performed with this configuration.

The frequency of oscillation was 12.502GHz with an output power of 14.8dBm at the fundamental frequency. This power was calculated by measuring the

Figure 4.14 Tuned output spectrum of oscillator.

Figure 4.15 Wide band output spectrum of tuned circuit.

insertion loss of the K-cable, coupler, and connectors at the oscillation frequency and adding this value to the measured output power read on the power meter.

The pushing characteristic was then measured. This describes the change in oscillation frequency due to a change in DC bias. Figure 4.17 shows this result. The collector-emitter voltage, V_{CE} was varied from 1.5V to 4.5V in steps of 0.25V for base

Figure 4.16 Narrow band output spectrum of tuned circuit.

Figure 4.17 Percentage change in oscillation frequency as a function of applied DC bias.

current levels of 0.7mA, 0.8mA, and 0.9mA. It was found that for values of V_{CE} outside of the range plotted in Figure 4.17, the DC operating point would again become unstable. These points were not plotted. The maximum change in oscillation frequency was only

0.046% over the measured range. The corresponding output power variation with bias is illustrated in Figure 4.18.

Figure 4.18 Output power vs. applied DC bias of HBT oscillator.

The phase noise of a microwave oscillator is an important quantity related to the phase instability of the output waveform. There are two types of phase instability, deterministic, and random. Deterministic instabilities show up as distinct components on a spectral density plots and can be related to known phenomena [64]. Random phase fluctuations are described by the term phase noise and have a continuous spectrum. For example, a real sinusoidal signal can be modeled by an ideal sine wave with a fluctuating amplitude and phase component as described by (4.3) [64]:

$$V(t) = [V_o + \epsilon(t)] \cdot \sin [2\pi f_o t + \Delta\phi(t)] \quad (4.3)$$

where $\epsilon(t)$ = amplitude fluctuations and $\Delta\phi(t)$ = randomly fluctuating phase term known as phase noise. An indirect measure of the noise energy is $\mathcal{L}(f_m)$ which is the single sideband phase noise to carrier ratio per Hz. It is usually specified in dB relative to the carrier level

in units of dBc/Hz at a specified offset frequency.

The phase noise of an oscillator has important effects on the performance of many practical systems. In Doppler radars a narrow bandwidth receiver is used to detect the frequency return of a moving target relative to the return from the ground called clutter. Since the power of the clutter is much greater than the target a very narrow bandwidth receiver which is tuned to the return frequency is required. Phase noise has the effect of spreading the target energy thus requiring a larger IF bandwidth in order to receive most of the target energy. This in turn causes more clutter energy to appear in the IF bandwidth which can partially or totally mask the target signal. Phase noise in the transmitter or receiver limits the range resolution and sensitivity of the radar [65]. The effect of phase noise on phase-modulated data systems is to degrade the bit error performance. The noise is often specified in terms of equivalent rms degrees of phase noise modulation in the channel bandwidth [65].

Several different methods exist for measuring the phase noise performance of a microwave oscillator. Common to all the methods is the use some form of frequency selective analyzer. To make accurate phase noise measurements usually requires dedicated and sometimes expensive equipment. This equipment was not available, therefore only a quick approximation of the phase noise was made. An approximate calculation of phase noise can be made directly using a spectrum analyzer. This requires that the noise of the DUT is higher than that of the analyzer itself and that correction factors be included for the non-ideal response of the analyzer. As described in [65], the correction formula is given by:

$$\begin{aligned} \text{Phase Noise (dBc/Hz)} = & \text{Spectrum Analyzer Reading} - 2.5 \\ & + 10\log(1.2 * \text{Resolution Bandwidth}) \end{aligned}$$

From Figure 4.14, at 100kHz offset from the carrier frequency, the signal is approximately 52dB below the carrier. Substituting this value in the above correction for-

mula yields:

$$\begin{aligned}\text{Phase Noise (dBc/Hz)} &\cong 52 - 2.5 + 10\log(1.2 \cdot 10000) \\ &= 90\text{dBc/Hz at } 100\text{kHz}\end{aligned}$$

Although this measurement was made with the triple stub tuner in place the result is reasonable for a free-running microstrip oscillator.

4.4 Summary

This chapter has presented the design and testing of an MIC HBT microstrip oscillator at 14GHz. The design was carried out with the CAD tool OSA90/HopeTM and included linear and nonlinear techniques. The simulated oscillation frequency and output power were 14.0GHz. and 12.0dBm respectively. Unfortunately, during testing it was found that the circuit was extremely sensitive to loading and often the DC operating point became unstable. Initial measurements showed the oscillation frequency to be between 12.7-13GHz with an output power of between 11.1 and 12.2dBm. These results however were not reproducible. In order to stabilize the circuit, tuning was required at the output. With the tuner connected, a clean stable output was obtained with a frequency of oscillation of 12.5GHz and output power of 14.8dBm. An estimate of the phase noise was obtained using a spectrum analyzer. The phase noise at 100kHz from the center frequency was approximately 90dBc/Hz.

As a result of the need for a triple stub tuner in order to stabilize the circuit a meaningful comparison between simulation and measurement cannot be made. This first iteration design however was valuable for identifying problem areas which require further investigation for future designs incorporating these transistors. A discussion of possible improvements to the design and some reasons for the discrepancy between simulation and measurement are addressed in the next chapter.

CHAPTER 5: CONCLUSION

5.1 Summary and Discussion

This work has presented the nonlinear empirical modeling of a GaAs/AlGaAs HBT from Rockwell International and a first iteration design of a microstrip oscillator at K band. The model includes self-heating and transit time effects as well as accounting for the variation of diffusion capacitance with base-collector voltage. The model can predict with good accuracy the DC characteristics of the transistor as well as the S-parameters in the forward active region of operation. Load-pull measurement results showed that the model overestimates the output power by 0.3-2.2dB. The P_{out} vs. P_{in} measurements also seem to confirm this because the simulated output power curve was more linear and saturated at a much slower rate than the measured result. The model was implemented in the microwave CAD tool OSA90/HopeTM V2.5. Using this single model the DC, small-signal RF and large-signal harmonic balance simulations and optimizations can be performed. As such it can be used to design linear and nonlinear microwave circuits, including oscillators using the harmonic balance algorithm.

Limited success was achieved with the first iteration oscillator design. The measurements showed the circuit had tendencies to oscillate fairly closely to the desired frequency. The circuit however was plagued with DC instabilities and was extremely sensitive to loading. These difficulties which could only be resolved using a triple-stub coax tuner at the output. A second iteration design would be necessary perhaps incorporating a high Q resonator in the feedback path to improve stability.

Chapter 2 described the basic characteristics of HBTs and their potential areas for application. This chapter showed how HBTs are the end result of applying band-gap engineering and the exploitation of material properties and processing technology in the quest to build a better bipolar transistor. GaAs HBTs retain the advantages of bipolar technology and extend their use to the microwave and millimetre wave frequency range.

A comparison of HBTs with other transistor technologies including Si BJTs and GaAs FETs was presented. The intrinsic characteristics and main performance advantages of each technology were outlined. The GaAs/AlGaAs HBT's high power handling capability and good low frequency noise characteristics are its biggest advantages for microwave circuits applications. Improvements to HBT transistors which include optimized physical structures geared to specific applications and use of other materials were also discussed. Finally microwave oscillator design theory was reviewed. The basic design equations were presented for small-signal and large-signal design. The method of oscillator analysis and design with some commercial harmonic balance microwave CAD tools was also outlined.

Chapter 3 dealt with the main effort of this work which was device modeling. An overview of some commercially available modeling and CAD tools was presented. OSA90/HopeTM was found to be the most flexible, powerful and appropriate tool for the type of modeling performed in this work. A description of the model was presented and the modeling procedure was outlined. The model in this work was based on the Gummel-Poon model which is the standard in most CAD tools. Modifications to this model were implemented in order to account for phenomena unique to HBTs. The most notable being the temperature dependence of model parameters and the self-heating effect.

The empirical modeling approach used in this work was heavily dependent on electrical measurement and computer optimization. The measurement setups for DC, thermal, and S-parameter measurements were explained. Difficulties arose with some measurements due to the instability of the transistor. It was found that the stability of the transistor was very sensitive to the DC biasing method used.

Some methods of parameter extraction were then evaluated. Good initial values for parameters are essential when a computer optimization routine is to be used to fit simulated to measured responses. This reduces the chance of obtaining non physical parameter values. It was found that due to the non-ideal nature of the base current in

GaAs HBTs, extraction of parameters from Gummel plots as is commonly done with bipolar transistors was not possible. This was further complicated due to the self-heating effect because the temperature variable could no longer be treated as a constant. Parameter extractions based solely on S-parameter data were attractive for this work because none of the physical and processing parameters for the transistor were available. Three different methods were tried and each met with limited success in providing initial parameter values for the resistive elements in the model. More work needs to be done in this area. Results of thermal resistance measurements were encouraging and showed the correct trend as described in the literature. The variation of current gain with temperature and power was quite linear as predicted from theory.

The next section described the optimization procedure used in the modeling effort. Due to the large number of variables, the optimization was done in stages. This reduced the number of unknowns being optimized at any one time. DC optimization was performed first with the objective of minimizing the error between the measured and simulated collector current, I_C as well as the measured and simulated base-emitter voltage, V_{BE} . When satisfactory agreement was obtained, the S-parameters were optimized. At this stage all DC parameters had been obtained so the only variable in need of optimization were the capacitances, forward delay, and the parasitics. The parasitics were obtained by optimizing three bias points covering the active region. Each bias point was then optimized individually. From these results the variation of capacitance with bias was obtained and fit to the modified SPICE equations.

The disadvantage of this procedure was that there is no guarantee that the values obtained are in any way unique. Since no physical parameters were known and no special test structures or calibration standards were provided there was no way to accurately separate the parasitics from the intrinsic device. As a result, there was heavy reliance on computer optimization. This causes many uncertainties when initial values are not close to the actual solution or when there are many unknowns. Furthermore, conver-

gence often became a problem in the course of optimization. This was due to the fact that the model was user-defined and as a result there were no built-in guards to prevent internal calculations from over or underflowing. In addition, the program must estimate the necessary derivatives of voltage and current sources with respect to the controlling node voltages numerically. With built-in models this is done analytically which produces much better convergence properties. It was found in the course of modeling that OSA90/HopeTM would inaccurately linearize the nonlinear charges in the model for small-signal simulation and optimization*. This led to errors in S-parameter simulations which were verified by comparing OSA90/HopeTM results with those of LIBRATM for a simple test circuit. Other difficulties were encountered with the optimization process. Often, the optimization had to be restarted after only a few iterations despite the fact that the error function had not reached a minimum value which rendered the optimization procedure much less automated than required.

With the model completed, some large-signal measurements were made in order to determine its validity under nonlinear operation. These included, load-pull measurements obtained from Ecole Polytechnique and P_{out} vs. P_{in} measurements. Several delays occurred in obtaining the load-pull measurement results. Several of the devices were damaged during the course of measurement and replacement devices needed to be sent to Ecole Polytechnique. Some uncertainty also arose in de-embedding the load-pull results which required further investigation. As a result of these delays, the model was not verified under nonlinear conditions until the circuit design was begun. The load-pull results were simulated with OSA90/HopeTM V2.5 on a point by point basis. It was found that the model consistently overestimated the output power with a maximum error of

* The problem was discovered to be due to the fact that the internal perturbation value used to estimate derivatives numerically was too large for this application. This was pointed out to the staff at OSA who reduced perturbation value and updated the program to allow the user to have more control over this parameter.

2.2dB. The simulations proved tedious however because of convergence difficulties with the software. Some uncertainties in comparing the simulation with measurement under nonlinear excitation also exist. These include the value of the harmonic terminations seen by the DUT, bond wire inductance uncertainties, and variations between individual devices.

The P_{out} vs. P_{in} measurement with 50Ω source and load terminations was also compared with the model. It was found that the measurement when compared to the measured S-parameters, overestimated the linear gain by approximately 1dB. This was due to inaccurate de-embedding of the losses in the measurement setup. The model also did not predict the saturation region of the curve well. In both class A and class AB operation, the simulated P_{out} vs. P_{in} curve is more linear than the measurement.

Chapter 4 dealt with the design and testing of a microstrip oscillator based on the model developed in chapter 3. A simple design procedure was employed in which the transistor was embedding in a series feedback configuration used to produce a large negative resistance at the output terminal. Both small-signal and large-signal design methods were used in order to verify the oscillation condition. Designing using this model with OSA90/HopeTM V2.5 proved to be difficult. This was due to the many convergence difficulties encountered with the harmonic balance optimization. There were circumstances where the optimizer was not able to meet the oscillation condition. Other times the error function would falsely reduce to a very small value indicating that the oscillation condition was satisfied. However, if the optimization was restarted, the error would become large again. Furthermore if a simulation was performed the values of OSCGAIN and OSCPHASE would not be zero indicating the oscillation conditions were not met.

The measurement results were then presented. The oscillator was found to be sensitive to loading and could only be stabilized with a triple stub tuner at the output. It is therefore difficult to compare simulated and measured results. Even using the tuner, it was found that the oscillation frequency was between 12.5 and 13GHz instead of the

design frequency of 14GHz. One reason for the discrepancy could be the way in which the transistor was mounted. These transistors contained vias which connected the emitter terminal to the back plane substrate metallization. These transistors were more intended for wafer probing and for two port applications such as amplifiers. For the oscillator, the transistor was epoxied to a square pad on gold on the alumina which connected the emitter to a short-circuited microstrip line. This parasitic was not taken into account in the simulation which would have resulted in the emitter short-circuited stub appearing electrically longer which would tend to reduce the oscillation frequency as was observed.

5.2 Future Work

In the area of device modeling, many areas of future work and improvements are possible. One particular area of further investigation is in parameter extraction. In this work some model parameters were extracted based on the various techniques found in the literature. Unfortunately, many parameters could not be obtained accurately and as such there was heavy reliance on computer optimization techniques. Although such techniques are useful they are not a substitute for accurate parameter extraction. Optimization of many parameters is often very time consuming and can lead to nonphysical results. Most of the difficulty with extracting parameters solely from S-parameter data was that the parasitics could not be de-embedded from the intrinsic device which prevented the accurate extraction of various parameters.

One way of circumventing this problem is to model a smaller area device. For example, the device in this work consisted of eight $2 \times 12 \mu\text{m}^2$ devices connected in parallel to form a larger device. One $2 \times 12 \mu\text{m}^2$ device should be the starting point for the modeling procedure. This device would have a much simpler parasitic structure due to the reduced number of interconnects when compared to those required for the large power devices. A more elaborate calibration process should then be used. Test structures could be provided in which the device is replaced in one case with a short-circuit and in another

with an open-circuit. These could then be used as calibration standards which would effectively remove the effects of parasitics so that the measured S-parameters will better reflect the performance of the intrinsic device. Another technique could be to use Transmit, Reflect, Line, (TRL) calibration which removes the effects of probe pads and transmission lines so that the reference plane is set closer to the intrinsic device. If these types of techniques were used, the methods of extracting model parameters from measured S-parameters would probably lead to more accurate values. Furthermore, more of the parameters could be extracted. This modified modeling approach could also be used for the larger devices. Other transistors which are built on this single emitter structure could possibly then be modeled by scaling the parameter values obtained for the $2 \times 12 \mu\text{m}^2$ device [66].

It was found that the model requires more work in improving its prediction of the device operation under large-signal excitation. The only types of measurements which were optimized were DC and S-parameters. Nonlinear measurements should also be used and optimized simultaneously with the DC and S-parameter data. This would require a readily accessible load-pull measurement setup. In this way more feedback and communication between the simulation and measurements could occur and problems and difficulties could be resolved more quickly. Currently a passive load-pull measurement system is being setup at CRC. The tuners and control software are available commercially from Focus Microwave Inc. [67]. Source and load-pull measurements can be performed as well as P_{out} vs. P_{in} measurements. This system also gives information about the load and source terminations seen by the device at the harmonic frequencies. Such a system could give valuable information for device modeling purposes as well as being an excellent design tool. The results of these measurements could be entered into OSA90/HopcTM V2.5 in order to further constrain the optimization process. This would lead to better agreement between simulation and measurement in the nonlinear region of operation of the device where it will eventually be used in most applications.

Another area of improvement is in the modeling of the reverse active region of operation. In this work, the optimization and parameter extraction were concentrated on the normal forward active region of operation. The reverse Gummel plot was not measured or used in extracting the saturation currents or ideality factors for diodes D_C or D_{BR} . This was because the devices were in common emitter configuration with the emitter connected to the back plane metallization of the substrate with via holes. In order to perform this measurement, a positive voltage needed to be applied to the emitter with the collector grounded. This meant that the entire back plane of the wafer would be at a positive potential. When this measurement was attempted however, a large emitter current was measured indicating a short-circuit somewhere else on the wafer. A transistor in common-collector configuration or without via holes would have facilitated the measurement of a reverse Gummel plot. Furthermore, the base collector capacitance in this model is composed of a depletion component only. This is because the S-parameter optimization was carried out in the active region where the base-collector junction is reverse-biased and is composed only of a depletion capacitance. However in many microwave applications such as power amplifiers, the device can enter the saturation region of operation for part of the duty cycle. As such, the base-collector junction becomes slightly forward biased. In this case, the diffusion component of the base-collector junction can become comparable or even greater than the depletion component. One method which could possibly be used to extract the diffusion component is to measure the S-parameters of the transistor in the reverse active mode of operation. In this region, the base-emitter junction is reverse biased while the base collector junction is forward biased. Since the depletion component of base-collector capacitance has already been extracted from the forward region, only the diffusion component needs extraction. The variation of diffusion capacitance with bias can be ascertained using the same procedure used in this work.

Further work can be performed regarding the thermal aspects of the model. Accurate temperature dependent measurements could be made in order to experimentally

investigate the temperature dependence of model parameters. In this work a only a crude temperature measurement system was used in order to extract the value of thermal resistance. A thermal probing station from Cascade Microtech is commercially available which facilitates accurate on wafer DC and S-parameter measurements at controlled temperatures [68]. This information could then be used to measure Gummel plots and S-parameters at multiple temperatures. Temperature variations of parameters such as ISO, ISEO, and BFO could be verified experimentally. This information could also be implemented to further constrain the optimization process.

Future work on the oscillator would begin with improvements to the non-linear model. Some of the possible improvements to the model which could be explored have been discussed above. One topic of further investigation is in the biasing of these transistors. It has been found through device and circuit measurement that these transistors were unstable unless the bias was provided through some sort of bias tee. Of course this is impractical for any circuit application. The simple bias network used for this circuit did not prove adequate and more thought should be given to this problem.

This circuit used series feedback from emitter to ground in order to make the device potentially unstable at 14GHz. As was previously explained, these transistors had vias connecting the emitter to ground which made them less applicable for oscillators. Some transistors have recently been obtained from Rockwell which do not have vias. They can thus be used as three ports which facilitates oscillator design. Other device orientations such as common base and common collector could also be examined and modeled.

It was found that this oscillator was very sensitive to the loading presented to it. One avenue of exploration could be in using a high Q resonator in the feedback path. This could further help to stabilize the oscillator and improve phase noise performance. Preliminary measurements have been performed on dielectric resonators coupled to microstrip lines. They showed a resonant frequency of approximately 13.5GHz but can

easily be lapped in order to raise the frequency to 14.0GHz. The unloaded Q's at 14 GHz have also been measured using Khanna's method [69] and were found to be greater than 4000.

Further work could also be done on optimizing the phase noise performance of the oscillator. One simple way of doing this is to employ a dielectric resonator as mentioned above. The design could also be optimized for phase noise by employing the Kurokawa's technique of having the amplitude and frequency trajectories of the output impedance intersect orthogonally [15], [70].

Appendix I:

**Mathematica™ Files for Parameter Extraction
from Measured S-Parameters**

```

(*****
* Mathematica file used to calculate the emitter resistance, RE      *
* using Maas' technique.                                           *
* =====*
*          V A R I A B L E    D E C L A R A T I O N S              *
* filebeg : refers to the first part of the S-Parameter file      *
*           which designates the base current bias. For example   *
*           0p2 means 0.2mA base current.                         *
* fileend : refers to the last part of the S-Parameter file      *
*           which designates the collector-emitter voltage bias.  *
*           For example 1p5 means 1.5V.                           *
* re       : Emitter resistance                                    *
* s        : 2-port S-parameters                                  *
* z        : 2-port Z-parameters                                  *
* magz     : two dimensional array. The first columns stores 1/IE  *
*           and the second column stores Mag[Z12].                *
* dr       : directory where files are kept.                      *
* ext      : file extension                                        *
* Freq     : Frequency                                            *
* p        : index variable                                       *
*****)

(* Input function used to convert from S-Parameters to either h or
   Z-Parameters *)

<<"D:\WNMATH22\DOCS\s2hz.m"

(* Initialize Variables *)

filebeg = {"0p2_", "0p4_", "0p6_", "0p8_", "1p0_", "1p2_", "1p4_"};
fileend = { "1p5", "2p0", "2p5", "3p0", "3p5", "4p0", "4p5" };
re = Table[ 0, {i,7},{j,7}];
s = { {0,0}, {0,0} };
magz =Table[0,{i,49},{j,2}];
dr = "b:s";
ext = ".s2p";
Freq=.;
p=1;

(* Loop over all base current and collector-emitter voltage biases *)

Do[
  Do[
    (* Open file for reading and skip first two lines *)
    sparfile = StringJoin[ dr,filebeg[[j]],fileend[[i]],ext ];
    spar = OpenRead[sparfile];
    Skip[ spar, Record,2 ];

    (* Determine value of base voltage *)
    Skip[ spar, Character, 14 ];
    VBE = Read[ spar, Word ];
    VBE = StringTake[ VBE, StringLength[VBE] - 1 ];
    VBE = ToExpression[VBE];

    (* Determine value of base current *)
    Skip[ spar,Character, 3];
    IB = Read[ spar, Word ];
    IB = StringTake[ IB, StringLength[IB] - 2 ];
    IB = ToExpression[IB];

    (* Determine value of collector-emitter voltage *)
    Skip[ spar, Character, 14 ];
    VCE = Read[ spar, Word ];
    VCE = StringTake[ VCE, StringLength[VCE] - 1 ];

```

```

VCE = ToExpression[VCE];
(* Determine value of collector current *)
Skip[ spar, Character, 3];
IC = Read[ spar, Word ];
IC = StringTake[ IC, StringLength[IC] - 2 ];
IC = ToExpression[IC];

(* Calculate the emitter current IE *)
IE = IB + IC;

(* Read frequency from data file *)
Skip[ spar, Record, 3];
Freq = Read[ spar, Real ] 10^9;

(* Read S-Parameters from data file and convert to rectangular
coordinates. It is assumed the data file contains the s-parameters
in polar coordinates in the order S11, S21, S12, S22 *)
Do[
  Do[
    mag = Read[ spar, Number ];
    ang = Read[ spar, Number ];
    s[[1,k]] = N[ mag Cos[ ang Degree ] + I mag Sin[ ang Degree ] ],
    {1,2} ];,
  {k,2} ];

(* Convert S-Parameters to Z-parameters. The second argument of the
function "s2z" refers to the characteristic impedance in which the
S-parameters were measured *)

z=s2z[s,50];

(* Store 1/IE and the magnitude of Z12 *)

magz[[p,1]]=1/IE;
magz[[p,2]]=Abs[ z[[1,2]] ];
p=p+1;

(* Close data file *)

Close[ spar ];,
(* Loop over all bias points *)
{ j,1,7}];,
{i,1,7}]

(***** E N D   O F   F I L E *****)

```



```

(*****
* Mathematica file used to calculate the small signal equivalent
* circuit parameters from measured S-Parameters using the method
* described by Pehlke.
*=====
*          V A R I A B L E   D E C L A R A T I O N S
* filebeg : refers to the first part of the S-Parameter file
*           which designates the base current bias. For example
*           0p2 means 0.2mA base current.
* fileend : refers to the last part of the S-Parameter file
*           which designates the collector-emitter voltage bias.
*           For example 1p5 means 1.5V.
* zb      : Base impedance
* zc      : Collector impedance
* s       : 2-port S-parameters
* z       : 2-port Z-parameters
* dr      : directory where files are kept.
* ext     : file extension
* Freq    : Frequency
* NumFreq : number of frequency points
* NumIB   : number of base current points
*****)

```

```
(* Input function to convert from S-parameters to h or Z parameters*)
```

```
<<"d:\wnmath22\docs\s2hz.m"
```

```
(* Initialize variables *)
```

```

filebeg = {"0p2_", "0p4_", "0p6_", "0p8_", "1p0_", "1p2_", "1p4_"};
fileend = { "1p5", "2p0", "2p5", "3p0", "3p5", "4p0", "4p5" };
NumFreq=190;
NumIB=7;
Freq = Table[0, {i, NumFreq}];
zb = Table[ 0, {i, NumFreq}, {j, NumIB}];
zc = Table[ 0, {i, NumFreq}, {j, NumIB}];
s = { {0,0}, {0,0} };
z = { {0,0}, {0,0} };
dr = "b:s";
ext = ".s2p";

```

```
(* Loop over base currents *)
```

```
Do[
```

```
(* Open file for reading and skip first nine lines *)
```

```

  sparfile = StringJoin[ dr, filebeg[[j]]!, fileend[[4]], ext ] ;
  spar = OpenRead[sparfile];
  Skip[ spar, Record, 9 ];

```

```
(* Loop over Frequency *)
```

```
Do[
```

```
(* Get Frequency *)
```

```
Freq[[i]] = Read[ spar, Number ] 10^9;
```

```
(* Read S-Parameters and convert to rectangular form *)
```

```
Do[
```

```
Do[
```

```
mag = Read[ spar, Number ];
```

```
ang = Read[ spar, Number ];
```

```
s[[1,k]] = N[ mag Cos[ ang Degree ] + I mag Sin[ ang Degree ] ],
```

```
{1,2} ];;
```

```
{k,2} ];
```

```

*****PHYSICAL CONSTANTS*****
k      = 1.3806E-23;      ! Boltzman's Constant
qe     = 1.602E-19;      ! electron charge
qbyk   = qe/k;

!*****DIODE SATURATION CURRENTS*****
ISCO   = 5.60528;
ISEO   = 25.44195?;
ISO    = 20.1 2.2071 10?;

ISCOT   = ISCO*1E-13;
ISEOT   = ISEO*1e-19;
ISOT    = ISO*1E-22;

!*****IDEALITY FACTORS*****
NE = 1.5 1.57786 2.0;
NC = 2.0;
NR = 1.10515;
NF = 1 1.16849 1.18?;

!*****DEVICE RESISTANCES*****
RB = 3.5 4.04712 6.0?;      !Base Resistanc
RE = 1.3 1.35 1.5?;        !Emitter Resistance
RC = 1.09731;              !Collector Resistance

!*****DC GAIN*****
BFO = 32 63.7275 1000?;
BR  = 0.109346;

!*****TEMPERATURE PARAMETERS*****
XTI = 4.39785?;
RTH = 282;                  ! Thermal Resistance;
To = 300;                   ! Room Temperature;

Eg = 1.43;
Egq = Eg*qe ;
Einf = 0.0764115?;
Einfq = Einf*qe;

!*****OHMIC LOSS IN S-PARAMETER TEST SET*****
RPORT1 = 1.9;
RPORT2 = 2.1;

RSWITCH = 1E7; ! Large resistor in parallel with base current source
            ! for convergence purposes

!*****EXTRINSIC PARASITICS*****
CBE' : 0.062904PF;
CCE' : 0.0698387PF;
LB   : 0.105107NH;
LC   : 0.0829073NH;
LE   : 0.00199806NH;
CBC' : 0.104809PF;

END !-----

!*****MODEL BLOCK*****
! This block contains the complete definition of the
! user-defined nonlinear model

```

```

(*****
* Mathematica file used to calculate the base and emitter resistance *
* using the method outlined by Prasad. *
* =====*
*          V A R I A B L E   D E C L A R A T I O N S *
* filebeg : refers to the first part of the S-Parameter file *
*           which designates the base current bias. For example *
*           0p2 means 0.2mA base current. *
* fileend : refers to the last part of the S-Parameter file *
*           which designates the collector-emitter voltage bias. *
*           For example 1p5 means 1.5V. *
* rb      : Base resistance *
* re      : Emitter resistance *
* s       : 2-port S-parameters *
* h       : 2-port h-parameters *
* x       : Real part of h21 divided by imaginary part of h21. *
* dr      : directory where files are kept. *
* ext     : file extension *
* Freq    : Frequency *
* RTH     : Thermal resistance *
* T       : Junction Temperature *
* hfeo    : DC short circuit current gain *
* rpi     : small signal base-emitter resistance *
*****)

```

```
(* load in package to convert from s pars to z/h pars *)
```

```
<<"d:\wnmath22\docs\s2hz.m"
```

```
(* initialize variables *)
```

```
filebeg = {"0p2_", "0p4_", "0p6_", "0p8_", "1p0_", "1p2_", "1p4_"};
fileend = { "1p5", "2p0", "2p5", "3p0", "3p5", "4p0", "4p5" };
```

```
rb = Table[ 0, {i,7},{j,7}];
re = Table[0, {i,7},{j,7}];
x = Table[ 0, {i,7},{j,7}];
s = { {0,0}, {0,0} };
h = { {0,0}, {0,0} };
RTH = 282;
T = 300;
dr = "b:s";
ext = ".s2p";
```

```
(* loop over all bias points *)
```

```
Do[
  Do[
    (* read in S-Parameter file *)
    sparfile = StringJoin[ dr,filebeg[[j]],fileend[[i]],ext ] ;
    spar = OpenRead[sparfile];
    Skip[ spar, Record, 2 ];

    (* get value of VBE *)
    Skip[ spar, Character, 14 ];
    VBE = Read[ spar, Word ];
    VBE = StringTake[ VBE, StringLength[VBE] - 1 ];
    VBE = ToExpression[VBE];

    (* get value of IB *)
    Skip[ spar, Character, 3 ];
    IB = Read[ spar, Word ];
    IB = StringTake[ IB, StringLength[IB] - 2 ];
    IB = ToExpression[IB];
  ]
];
```

```

(* get value of VCE *)
Skip[ spar, Character, 14 ];
VCE = Read[ spar, Word ];
VCE = StringTake[ VCE, StringLength[VCE] - 1 ];
VCE = ToExpression[VCE];

(* get value of IC *)
Skip[ spar, Character, 3];
IC = Read[ spar, Word ];
IC = StringTake[ IC, StringLength[IC] - 2 ];
IC = ToExpression[IC];

(* pick desired frequency for calculation of rb and re *)
Skip[ spar, Record, 5];
Freq = Read[ spar, Number ] 10^9;

(* read in S-Parameters and convert to rectangular form *)
Do[
  Do[
    mag = Read[ spar, Number ];
    ang = Read[ spar, Number ];
    s[[1,k]] = N[ mag Cos[ ang Degree ] + I mag Sin[ ang Degree ] ],
    {1,2} ];,
  {k,2} ];

(* convert to h parameters *)
h=s2h[s,50];

(* calculate junction temperature T, *)
T = 300 + Rth (IC VCE + IB VBE)/1000;

x[[i,j]] = -Im[ h[[2,1]] ]/Re[ h[[2,1]] ];
hfeo = Re[ h[[2,1]] ] ( 1 + x[[i,j]]^2 );
rpi = hfeo 8.618 10^-2 T/IC;

(* calculate re and rb *)
re[[i,j]] = -( Im[ h[[1,1]] ] ( 1 + x[[i,j]]^2 ) /
  x[[i,j]] + rpi )/hfeo;

rb[[i,j]] = Re[ h[[1,1]] ] - re[[i,j]] -
  ( rpi + hfeo re[[i,j]] )/( 1 + x[[i,j]]^2 );

Close[ spar ];,
{j,1,7}];,
{i,1,7}];

(***** E N D   O F   F I L E *****)

```

Appendix II:

**OSA90/Hope™ V2.5 Netlist Files for DC and S-Parameter
Simulation and Optimization.**

```

!*****
!* OSA90/Hope netlist file
!* This file provides an example of DC simulation and
!* optimization in the forward active region of operation
!* for the Rockwell 8 finger HBT.
!*****

=====

! FILE:dcopt.ckt

!*****CONTROL BLOCK*****
! Various statements for controlling convergence,
! simulation and display parameters

CONTROL
! ALTERNATIVE_SOLVER;
! TWO_SIDED_PERTURBATION;
! NON_MICROWAVE_UNITS;
! NEWTON_EPS_RELATIVE=0.1;
! DISPLAY_N_DIGITS=6
END

!*****EXPRESSION BLOCK*****
! This block contains the definitions of user-defined
! variables and statements for the inclusion of measured
! data files.

EXPRESSION !-----

!*****MEASURED DC DATA IN SEPARATE ASCII FILES*****
! Each file contains the data in a matrix format.
! rw8f2dc1.dat contains the IC vs VCE data in the matrix: DCIV_MEAS
! rw8f2dc2.dat contains the VBE vs VCE data in the matrix: DCVV_MEAS
! 8f2_27p0 contains the Gummel plot data in the matrix: 8f2_27p0

#include "rw8f2dc1.dat"; ! DC IC vs. VCE data
#include "rw8f2dc2.dat"; ! DC VBE vs. VCE data

#define N_VCE_POINTS 161 !No. of collector-emitter voltage (VCE) points
#define N_IB_POINTS 9 !No. of base current (IB) points
#define N_VBE_POINTS 81 !No. of base-emitter voltage (VBE) points.

! Measured VCE extracted from first column of DC-IV data file
VCE_MEAS[N_VCE_POINTS] = COL(DCIV_MEAS,1);

! Base current array in uA
IB_MEAS[N_IB_POINTS] = [0 0 200 400 600 800 1000 1200 1400 ];

IBASE = 0.0A; !Applied Base Current
VBASE = 0.0V; !Applied Base-Emitter Voltage
VCOL = 0.0V; !Applied Collector Voltage

!***INDEX VARIABLES***

I = 1;
J = 1;
L = 1;
P = 1;
Q = 1;

```

```

*****PHYSICAL CONSTANTS*****
k      = 1.3806E-23;      ! Boltzman's Constant
qe      = 1.602E-19;      ! electron charge
qbyk    = qe/k;

!*****DIODE SATURATION CURRENTS*****
ISCO    = 5.60528;
ISEO    = ?5.44195?;
ISO     = ?0.1 2.2071 10?;

ISCOT    = ISCO*1E-13;
ISEOT    = ISEO*1e-19;
ISOT     = ISO*1E-22;

!*****IDEALITY FACTORS*****
NE = ?1.5 1.57786 2.0;
NC = 2.0;
NR = 1.10515;
NF = ?1 1.16849 1.18?;

!*****DEVICE RESISTANCES*****

RB = ?3.5 4.04712 6.0?;      !Base Resistance
RE = ?1.3 1.35 1.5?;      !Emitter Resistance
RC = 1.09731;      !Collector Resistance

!*****DC GAIN*****
BFO = ?32 63.7275 1000?;
BR = 0.109346;

!*****TEMPERATURE PARAMETERS*****
XTI = ?4.39785?;
RTH = 282;      ! Thermal Resistance;
To = 300;      ! Room Temperature;

Eg = 1.43;
Egq = Eg*qe;
Einf = ?0.0764115?;
Einfq = Einf*qe;

!*****OHMIC LOSS IN S-PARAMETER TEST SET*****
RPORT1 = 1.9;
RPORT2 = 2.1;

RSWITCH = 1E7; ! Large resistor in parallel with base current source
             ! for convergence purposes

!*****EXTRINSIC PARASITICS*****
CBE' : 0.062904PF;
CCE' : 0.0698387PF;
LB : 0.105107NH;
LC : 0.0829073NH;
LE : 0.00199806NH;
CBC' : 0.104809PF;

END !-----

!*****MODEL BLOCK*****
! This block contains the complete definition of the
! user-defined nonlinear model

```

```

MODEL !-----

!*****RESISTORS*****
RES 10 4 R=RB;
RES 40 90 R=RPORT1;
RES 50 99 R=RPORT2;
RES 6 30 R=RE;
RES 20 5 R=RC;

!*****VOLTAGE AND CURRENT LABELS*****
VLABEL 4 6 NAME = VBEDELAY TAU=0;
VLABEL 4 6 NAME = VBE ; ! Intrinsic base-emitter voltage label
VLABEL 40 0 NAME = VBE1; ! Extrinsic base-emitter voltage label
VLABEL 50 0 NAME = VCE1; ! Extrinsic collector-emitter voltage label

ILABEL 40 @IB1 NAME = IB; ! Base current label
ILABEL 50 @IC1 NAME = IC; ! Collector current label

!*****
!* DEFINITION OF THERMAL SUBCIRCUIT *
!*****
VSOURCE 80 0 NAME = TEMPER VDC=To;
RES 80 81 R = RTH;
CAP 80 81 C = CTH;
NLCCS 0 81 I = (IC*VCE1 + IB*VBE1);
VLABEL 81 0 NAME = TEST;

! Guard against floating point errors by assigning
! T=TEST only if the output of the thermal sub-circuit produces
! a realistic value for the junction temperature, T. A realistic value
! is between 300K and 400K

T = IF ( (TEST > 400) + (TEST < To) ) (400)
      ELSE ( MAX(To,TEST) );

!*****EQUATIONS FOR SATURATION CURRENTS AND IDEAL GAIN*****
ISAT = ISOT*((T/To)^XTI)*Exp(MAX(1E-99, (Egq/k)*(-1/T+1/To)));
BF = BFO*((T/To)^(-1))*Exp((Einfq/k)*(1/T-1/To));

ISBF = ISAT/BF;
ISBR = ISAT/BR;

ISE = ISEOT*((ISAT/ISOT)^(1/NE))*(BFO/BF);
ISC = ISOT*((ISAT/ISOT)^(1/NC))*(BFO/BF);

!*****ARGUMENTS OF EXPONENTIAL FUNCTION*****
xa = qbyk*VBEDELAY/(NF*T);
xb = qbyk*VBC/(NR*T);

! Limit maximum value of xa and xb to 60 to avoid floating
! point errors
xaa = EXP(MIN(60,xa));
xbb = EXP(MIN(60,xb));

!*****DIODES AND NONLINEAR CURRENT CONTROLLED CURRENT SOURCE*****
DIODE 4 5 IS=ISC N=NC TEMP=T; ! DC
DIODE 4 6 IS=ISE N=NE TEMP=T; ! DE
DIODE 4 5 IS=ISBR N=NR TEMP=T; ! DBR
DIODE 4 6 IS=ISBF N=NF TEMP=T; ! DBF
NLCCS 5 6 I=(ISAT*(xaa - xbb)); ! ICT

```



```

*****PARASITICS*****
IND @IB1 10 L=LB;
IND 30 0 L=LE;
IND @IC1 20 L=LC;
CAP 10 30 C=CBE';
CAP 20 30 C=CCE';
CAP 10 20 C=CBC';

*****BASE BIAS SUPPLY*****
ISOURCE 1 0 NAME=BASE IDC=IBASE; ! Base current source
RES 1 0 R=RSWITCH; ! Large shunt resistance included
! for convergence purposes
IND 90 1 L=1000NH; ! RF choke
CAP 41 40 C=10000PF; ! DC block

*****COLLECTOR BIAS SUPPLY*****
VSOURCE 2 0 NAME=COLLECTOR VDC=VCOL; ! Collector voltage source
IND 99 2 L=1000NH; ! RF choke
CAP 51 50 C=10000PF; ! DC block

*****INPUT AND OUTPUT PORT DEFINITION*****
PORT 41 0 NAME=INPUT;
PORT 51 0 NAME=OUTPUT;

!-----END OF MODEL DEFINITION-----

CIRCUIT HARM=5;

*****POSTPROCESSED OUTPUT LABELS*****

ICMA=ICOLLECTOR_DC*1000; ! Simulated collector current in mA
DCIV_MEAS_MA = DCIV_MEAS[J,I]*1000; ! Measured collector current

END !-----

*****SWEEP BLOCK*****
!This block contains the statements necessary for DC simulation and
!for setting up graphical views

SWEEP !-----

      DC:      I: FROM 3 TO (N_IB_POINTS) STEP = 1
                J: FROM 1 TO (N_VCE_POINTS) STEP = 1
                RSWITCH = 1E7
                IBASE = (IB_MEAS[I]*1E-6)
                VCOL = VCE_MEAS[J]

!      Output variables used in graphical views
      VCE_MEAS[J], ICMA, VBASE_DC, DCVV_MEAS[J,I], DCIV_MEAS_MA

! Plot of simulated and measured IC vs VCE characteristic

(PARAMETRIC P=J I=ALL X=VCE_MEAS[J] Y=ICMA.WHITE &
DCIV_MEAS_MA.GREEN.TRIANGLE NYTicks=4)

Plot of simulated and measured VBE vs VCE characteristic

(PARAMETRIC P=J I=ALL X=VCE_MEAS[J] Y=VBASE_DC.white &
DCVV_MEAS[J,I].GREEN.TRIANGLE YMIN=1.3 YMAX=1.5 NYTicks=4)
;
END !-----

```

```

!*****SPECIFICATION BLOCK*****
! This block contains the statements to perform DC optimization
! to match the simulated to measured DC characteristic

SPECIFICATION !-----
DC: I: FROM 2 TO (N_IB_POINTS) STEP = 1
      J: FROM 11 TO N_VCE_POINTS STEP=2
      RSWITCH=1E7
      IBASE = (IB_MEAS[I]*1E-6)
      VCOL  = VCE_MEAS[J]

!*****OPTIMIZATION CRITERIA*****
ICMA = DCIV_MEAS_MA
VBASE_DC=DCVV_MEAS[J,I] w=5;

END !-----

!*****E N D   O F   F I L E *****

```

```

!*****
! * OSA90/Hope netlist file
! * This file provides an example of S-Parameter simulation
! * and optimization at bias point IB=0.8mA VCE=3.0V.
! * This is not a complete netlist file. It only contains the
! * necessary additions and changes to the file of APPENDIX XX
! * in order to perform S-parameter simulation and optimization.*
!*****

```

```

=====
! FILE:acopt.ckt

```

```

!*****CONTROL BLOCK*****
! Various statements for controlling convergence,
! simulation and display parameters

```

```

CONTROL
! ALTERNATIVE_SOLVER;
! TWO_SIDED_PERTURBATION;
! NON_MICROWAVE_UNITS;
! NEWTON_EPS_RELATIVE=0.1;
! DISPLAY_N_DIGITS=6
END

```

```

!*****EXPRESSION BLOCK*****
! This block contains the definitions of user-defined
! variables and statements for the inclusion of measured
! data files.

```

```

EXPRESSION

```

```

!*****MEASURED DC DATA IN SEPARATE ASCII FILES*****
! Each file contains the data in a matrix format. The matrix
! has the same name as the file, i.e. s0p8_3p0.

```

```

#include "s0p8_3p0.s2p"; ! S-parameter data file.
! The bias point is IB=0.8mA VCE=3.0V

```

```

#define N_FREQ_POINTS 100 ! No. of measured frequency points

```

```

! Measured frequency points extracted from first column of data file
FREQUENCY[N_FREQ_POINTS]=COL(0p8_3p0,1);

```

```

IBASE : 0.0A; ! Applied base current
VCOL : 0.0V; ! Applied collector voltage

```

```

!***Index Variables***

```

```

I = 1;
J = 1;
P = 1;
Q = 1;

```

```

!
!
!
!
! PHYSICAL CONTSTANTS
! DIODE SATURATION CURRENTS
! IDEALITY FACTORS
! DEVICE RESISTANCES
! DC GAIN
! TEMPERATURE PARAMETERS
! OHMIC LOSS,
! EXTRINSIC PARASITICS

```

```
! All of these parameters are the same as those given in the dc file
!
!
!
END !-----
```

```
!*****MODEL BLOCK*****
```

```
! This block contains the complete definition of the
! user-defined nonlinear model
```

```
MODEL !-----
```

```
! The following shows the additions to the model block for S-parameter
! optimization. Statements which are the same as the dc netlist are
! omitted.
```

```
! Each element in the following matrices, DELAY, CBC, CBE corresponds
! to a DC bias point. Each row represents a different base current
! from 0.2mA to 0.8mA in 0.2mA steps. Each column refers to a
! collector emitter voltage from 1.5V to 4.5V in 0.5V steps.
```

```
!DELAY matrix: Forward delay in current source
DELAY[7,7]=[
```

```
1.78520PS 2.01826PS 2.28506PS 2.50753PS 2.68757PS 2.88055PS 3.05027PS
1.74212PS 2.07314PS 2.18981PS 2.32027PS 2.45325PS 2.58098PS 2.71285PS
1.79039PS 2.004PS 2.16726PS 2.15994PS 2.36458PS 2.45385PS 2.59572PS
1.66976PS 1.94127PS 1.98507PS ?2.24547PS? 2.34577PS 2.42514PS 2.57466PS
1.67002PS 1.85181PS 2.14599PS 2.27261PS 2.37582PS 2.48075PS 2.56412PS
1.59926PS 1.86417PS 2.14808PS 2.14745PS 2.4028PS 2.52955PS 2.60355PS
1.5991PS 1.88808PS 2.15317PS 2.30755PS 2.42494PS 2.56669PS 2.6687PS
];
```

```
!Base-collector capacitance
CBC[7,7]=[
```

```
0.124526PF 0.106669PF 0.0945285PF 0.0863842PF
0.080343PF 0.0756PF 0.0716828PF

0.119848PF 0.0991794PF 0.0883844PF 0.0809769PF
0.0754304PF 0.0711539PF 0.0676769PF

0.1155PF 0.0952678PF 0.083911PF 0.0785934PF
0.0721335PF 0.0685104PF 0.0651534PF

0.113267PF 0.091911PF 0.0830949PF ?0.0738483PF?
0.0696949PF 0.0665022PF 0.0631199PF

0.109153PF 0.0876128PF 0.0761833PF 0.0701888PF
0.0667276PF 0.0636105PF 0.0616452PF

0.11796PF 0.0832717PF 0.0727342PF 0.0703343PF
0.0644159PF 0.0611677PF 0.0597547PF

0.100988PF 0.0786761PF 0.0699548PF 0.0645376PF
0.0615556PF 0.0589814PF 0.0571993PF
```

```
];
```

```

!Base-emitter capacitance
CBE[7,7]= [

1.12758PF 1.20099PF 1.25529PF 1.30819PF
1.35625PF 1.39876PF 1.43711PF

1.73358PF 1.85289PF 1.98599PF 2.10023PF
2.199PF 2.29056PF 2.375PF
2.30939PF 2.52971PF 2.71084PF 2.89701PF
3.02422PF 3.16003PF 3.26745PF

2.89294PF 3.18699PF 3.46248PF ?3.63708PF?
3.82449PF 3.98752PF 4.10396PF

3.46961PF 3.86315PF 4.11597PF 4.36183PF
4.57861PF 4.7418PF 4.89023PF

4.22293PF 4.50899PF 4.79122PF 5.13536PF
5.29921PF 5.44679PF 5.59075PF

4.58975PF 5.12941PF 5.45902PF 5.73101PF
5.94824PF 6.08857PF 6.18398PF

];

! The voltage label VBEDELAY has changed so that the delay factor is
! included.

VLABEL 4 6      NAME=VBEDELAY TAU=DELAY[P,Q];

!
!
!
!
!      OTHER VOLTAGE AND CURRENT LABELS
!      THERMAL SUB-CIRCUIT
!      SATURATION CURRENTS AND GAIN EQUATIONS
!      DIODES AND NONLINEAR CURRENT SOURCE
!      BIAS SUPPLY
!      All the same as in dc netlist file
!
!
!
!

!Base-collector capacitor defined by matrix CBC[P,Q]
CAP 4 5 C=CBC[P,Q];

!Base-emitter capacitor defined by matrix CBE[P,Q]
CAP 4 6 C=CBE[P,Q];

CIRCUIT HARM=5;

END

*****SWEEP BLOCK*****
! This block contains the statements necessary for S-parameter
! simulation and for setting up graphical views.

SWEEP

!      Sweep from 0.4GHz to 40GHz
AC: I: FROM 1 TO 100 STEP = 1
      P=4 Q=7
      RSWITCH=1E7
      FREQ=(FREQUENCY[I]*1E9)

```

```

IBASE = 800UA
VCOL = 3.0
!Output variables to be plotted
FREQ,MS11,PS11,MS21,PS21,MS12,PS12,MS22,PS22,
s0p8_3p0[I,2],s0p8_3p0[I,3],s0p8_3p0[I,4],s0p8_3p0[I,5],
s0p8_3p0[I,6],s0p8_3p0[I,7],s0p8_3p0[I,8],s0p8_3p0[I,9]

! Setup Smith chart views of simulated vs. measured S11 and S22
( SMITH MP=( MS11,PS11 ).YELLOW &
  ( s0p8_3p0[I,2],s0p8_3p0[I,3] ).YELLOW.CIRCLE &
  ( MS22,PS22 ).GREEN &
  ( s0p8_3p0[I,8],s0p8_3p0[I,9] ).GREEN.CIRCLE
)

!Setup Polar plot of simulated vs. measured S21 and S12
( POLAR MP=( MS21,PS21 ).YELLOW &
  ( s0p8_3p0[I,4],s0p8_3p0[I,5] ).YELLOW.CIRCLE &
  ( MS12,PS12 ).GREEN&
  ( s0p8_3p0[I,6],s0p8_3p0[I,7] ).GREEN.CIRCLE
)

! Setup X-Y plots of simulated vs. measured S-parameters
( PARAMETRIC P=I Y=MS11.YELLOW & MS22.GREEN &
  s0p8_3p0[I,2].YELLOW.CIRCLE &
  s0p8_3p0[I,8].GREEN.CIRCLE X=FREQ)

( PARAMETRIC P=I Y=MS21.YELLOW & MS12.GREEN &
  s0p8_3p0[I,4].YELLOW.CIRCLE &
  s0p8_3p0[I,6].GREEN.CIRCLE X=FREQ )

( PARAMETRIC P=I Y=PS11.YELLOW & PS22.GREEN &
  s0p8_3p0[I,3].YELLOW.CIRCLE &
  s0p8_3p0[I,9].GREEN.CIRCLE X=FREQ)

( PARAMETRIC P=I Y=PS21.YELLOW & PS12.GREEN &
  s0p8_3p0[I,5].YELLOW.CIRCLE &
  s0p8_3p0[I,7].GREEN.CIRCLE X=FREQ)

;
END !-----

!*****SPECIFICATION BLOCK*****
! This block contains the statements to perform S-Parameter
! optimization to match the simulated to measured responses.

SPECIFICATION !-----

AC:  I: FROM 1 TO (N_FREQ_POINTS-35) STEP = 1
      RSWITCH=1E7
!      Select appropriate element in bias dependent matrices
      P=4 Q=7
      FREQ=(FREQUENCY[I]*1E9)
!      Set bias values
      IBASE = 800UA
      VCOL = 3.0

```

! Optimization Criteria

MS11=s0p8_3p0[I,2] w=20
PS11=s0p8_3p0[I,3] w=0.1
MS21=s0p8_3p0[I,4] w=1
PS21=s0p8_3p0[I,5] w=0.1
MS12=s0p8_3p0[I,6] w=1
PS12=s0p8_3p0[I,7] w=0.1
MS22=s0p8_3p0[I,8] w=5
PS22=s0p8_3p0[I,9] w=0.1;

END !-----

!***** END OF FILE *****

Appendix III:

**OSA90/HopeTM V2.5 Netlist File for Nonlinear
Oscillator Design.**


```

!*****
! OSA90/Hope V2.5 netlist file used for non-linear oscillator
! design using the OSCPORT element.
!*****

!***CONTROL BLOCK***
! Various statements for controlling convergence,
! simulation and display parameters

CONTROL
! ALTERNATIVE_SOLVER;
! TWO_SIDED_PERTURBATION;
! NON_MICROWAVE_UNITS;
!NEWTON_EPS_RELATIVE = 0.05
display_n_digits=6
END

MODEL

PIN=0DBM;

!***INDEX VARIABLES***

I = 1;
J = 1;
L = 1;
P = 1;
Q = 1;

!***PHYSICAL CONSTANTS***

k      = 1.3806E-23;    ! Boltzman's Constantanst
qe     = 1.602E-19;    ! electron charge
qbyk   = qe/k;

!***SATURATION CURRENTS***

ISCO = 5.60528;
ISEO = 5.44195;
ISO  = 2.2071;
ISCOT = ISCO*1E-13;
ISEOT = ISEO*1e-19;
ISOT  = ISO*1E-22;

!***IDEALITY FACTORS***

NE =1.57786;
NC =2.0;
NR =1.10515;
NF =1.16849;

!***RESISTANCES***

RB=4.04712;  RE =1.35;  RC =1.09731;

RPORT1=0.01; RPORT2=0.01;      !OHMIC LOSS THROUGH HP8510
RSWITCH=1E7; ! LARGE RESISTOR IN PARALLEL WITH CURRENT SOURCE

!***DC CURRENT GAINS***

BFo=63.7275;    BR =0.109346;

```

!***THERMAL PARAMETERS***

XTI = 4.39785; Rth = 282; TMEAS = 300; To = 300;
Eg = 1.43; Egq = Ego*qe; Einf=0.0764115; Einfq =Einf*qe;

!***BASE AND COLLECTOR SOURCE VARIABLES***

IBASE : 0.0A;
VBASE : 0.0V;
VCOL : 0.0V;

!***EXTRINSIC PARASITICS***

CBE' : 0.062904PF;
CCE' : 0.0698387PF;
LB : 0.105107NH;
LC : 0.0829073NH;
LE : 0.00199806NH;
CBC' : 0.104809PF;

!***BOND WIRE INDUCTANCES***

LBOND=0.15NH;

!***BASE EMITTER DEPLETION CAPACITANCE PARAMETERS***

CJEO=0.313849PF;
VJE= 1.71356;
MJE=0.33;

!***BASE COLLECTOR DEPLETION CAPACITANCE PARAMETERS***

CJCO=0.113516PF;
VJC= 1.26376;
MJC=0.5;

!***BASE EMITTER DIFFUSION CAPACITANCE PARAMETERS***

TFO=8.36403PS;
XTF=0.642167;
VTF=3.82625;

IND 40 102 L=LBOND;
IND 111 50 L=LBOND;

RES 10 4 R=RB;
RES 6 30 R=RE;
RES 20 5 R=RC;

!***VOLTAGE AND CURRENT LABELS***

VLABEL 4 5 NAME = VBC ;! intrinsic base-collector voltage
VLABEL 4 6 NAME = VBE ;! intrinsic base-emitter voltage
VLABEL 4 6 NAME = VBEDELAY TAU=2.24547PS; !Delayed VBE

ILABEL 102 @IB1 NAME = IB; !Base current
ILABEL 111 @IC1 NAME = IC; !Collector current

VLABEL 10 20 NAME = VBE1;! Extrinsic base emitter voltage
VLABEL 111 0 NAME = VCE1;! Extrinsic collector-emitter voltage

```

!*****
!* DEFINITION OF THERMAL SUBCIRCUIT *
!*****
VSOURCE  80  0  NAME=TEMPER VDC=TMEAS;
RES       80 81  R=RTH;
CAP       80 81  C=1F;
NLCCS     0 81  I=(IC*VCE1 + IB*VBE1);
VLABEL    81  0  NAME=TEST;

! Guard against floating point errors by assigning
! T=TEST only if the output of the thermal sub-circuit produces
! a realistic value for the junction temperature, T. A realistic value
! is between 300K and 400K

T = IF ( (TEST > 400) + (TEST< To) ) (400)
      ELSE ( MAX(To,TEST) );

!*****EQUATIONS FOR SATURATION CURRENTS AND IDEAL GAIN*****
ISAT = ISOT*((T/To)^XTI)*Exp(MAX(1E-99, (Egq/k)*(-1/T+1/To)));
BF    = BFO*((T/To)^(-1))*Exp((Einfq/k)*(1/T-1/To));
ISBF  = ISAT/BF;
ISBR  = ISAT/BR;

ISE = ISEOT*((ISAT/ISOT)^(1/NE))*(BFO/BF);
ISC = ISCOT*((ISAT/ISOT)^(1/NC))*(BFO/BF);

!*****ARGUMENTS OF EXPONENTIAL FUNCTION*****

xa = qbyk*VBEDELAY/(NF*T);
xb = qbyk*VBC/(NR*T);

! Limit maximum value of xa and xb to 60 to avoid floating
! point errors
xaa = EXP(MIN(60,xa));
xbb = EXP(MIN(60,xb));

!***DIODES AND NCNLINEAR CURRENT CONTROLLED CURRENT SOURCE***

DIODE 4 5 IS=ISC N=NC TEMP=T;      ! DC
DIODE 4 6 IS=ISE N=NE TEMP=T;      ! DE
DIODE 4 5 IS=ISBR N=NR TEMP=T CJ0=CJCO VJ=VJC EJ=MJC FC=1; ! DBR
DIODE 4 6 IS=ISBF N=NF TEMP=T CJ0=CJEO VJ=VJE EJ=MJE FC=1; ! DBF
NLCCS 5 6 I=(ISAT*(xaa - xbb));    ! ICT

!***PARASITICS***
IND  @IB 110 L=LB;
IND  30 150 L=LE;
IND  @IC 120 L=LC;
CAP  10 30 C=CBE';
CAP  20 30 C=CCE';
CAP  10 20 C=CBC';

XBC=(VBC/1.44*VTF);
EBC=(EXP( XBC ));

TAUF = IF (VBC > -20 & vbc < 20 ) ( TFO*(1-XTF*Exp( VBC/(1.44*VTF) ) ) )
      ELSE (TFO);

!*** Nonlinear charge source between base and emitter

NLCQ 4 6 Q=( TAUF*ISAT*(xabb-1) );

!*****END OF MODEL*****

```

!***Substrate definition***

MSUB EPSR=9.96H=0.254MMT=0.001MMTAND=0.001RHS=0 ROC=2.44E-8;

!***SHORT CIRCUITED STUB IN EMITTER FOR SERIES FEEDBACK***

LESTUB = 3206.84UM;

MSL 150 151 W=0.245MM L=LESTUB;

IND 151 0 L=0.07NH;

!***OPEN CIRCUITED STUB TO RESONATE INPUT PORT OF TRANSISTOR***

LBSTUB=?502.687UM?;

MOPEN 201 W=245UM L=LBSTUB;

!*****

! * BASE BIAS CIRCUIT *

!*****

QUARTER1=2082UM;

QUARTER2=1946UM;

WTHIN = 75UM;

W50 = 245UM;

WWIDE = 300UM;

WPAD = 750UM;

MSL 40 200

W = W50 L = 1400UM;

MTEE 200 201 202

W1 = W50 W2 = W50 W3 = WTHIN;

MSL 202 203

W = WTHIN L = QUARTER1;

MCROSS 203 204 205 206

W1 = WTHIN W2 = WTHIN W3=WWIDE;

RES 206 210

R = 50;

CAP 210 0

C = 100PF;

MOPEN 205

W = WWIDE L = QUARTER2;

MSL 204 222

W = WTHIN L = 1500UM;

MSTEP 222 223

W1 = WTHIN W2 = WPAD;

MSL 223 209

W = WPAD L = WPAD;

!***BASE BIAS***

ISOURCE 209 0

NAME=BASE IDC=0.8MA;

RES 209 0

R=RSWITCH;

!*****

! * COLLECTOR BIAS CIRCUIT *

!*****

MSL 50 300

W = W50 L = 150UM;

MTEE 300 301 302

W1 = W50 W2 = W50 W3 = WTHIN;

MSL 302 303

W = WTHIN L = QUARTER1;

MCROSS 303 304 305 306

W1 = WTHIN W2 = WTHIN W3 = WWIDE;

RES 304 310

R = 50;

CAP 310

C = 100PF;

MOPEN 305

W = WWIDE L = QUARTER2;

MSL 306 307

W = WTHIN L = 1500UM;

MSTEP 307 308

W1 = WTHIN W2 = WPAD;

MSL 308 309

W = WPAD L = WPAD;

!***COLLECTOR BIAS***

VSOURCE 309 0

NAME=COLLECTOR VDC=3.0V;

!*****OUTPUT LINE AND DC BLOCK CAPACITOR*****

MSL 301 311 W = W50 L = 200UM;

CAP 311 312 C = 10PF;

MSL 312 54 W = 245UM L = 1000UM;

```

! Estimate of output voltage
OSC_VOLT =?0.6 1.26769 1.75?;

! OSCPORT element used for non-linear oscillator design
OSCPORT540NAME=OUT R=50 V=OSC_VOLT;

!*****
!* OUTPUT VARIABLES *
!*****

!Output power in each harmonic in dBm
PO_DBM[0:N_SPECTRA]=IF (PWOUT > 0) (10*LOG10( PWOUT ) + 30)
                     ELSE (NAN);
END !-----

!*****SWEEP BLOCK*****
! This block contains the statements necessary for HB simulation and
! for setting up graphical views
SWEEP !-----
  HB:
    FREQ:14
    OSC_GAIN,OSC_PHASE,PO_DBM
    MVOUT PVOUT PWOUT

! Time domain output voltage waveform
{ WAVEFORM SPECTRUM=(MVOUT,PVOUT)
  TMIN=0 TMAX=150PS NT=100}
END !-----

!*****SPECIFICATION BLOCK*****
! This block contains the statements to perform Harmonic Balance
! optimization in order to meet the oscillation conditions
! at the design frequency of 14GHz.
SPEC
  HB:FREQ=14.0
! Optimization criteria
  OSC_GAIN = 0
  OSC_PHASE = 0
END

!***** E N D   O F   F I L E *****

```

REFERENCES

- [1] H. Kroemer, "Theory of a Wide-Gap Emitter for Transistors," *Proc. IRE*, pp. 1535-1537, Nov. 1957.
- [2] H. Kroemer, "Heterostructure Bipolar transistors and Integrated Circuits," *Proc. IEEE*, vol. 70, no. 1, pp. 13-25, Jan. 1982.
- [3] P. M. Asbeck et al., "GaAs-Based Heterojunction Bipolar Transistors for Very High Performance Electronic Circuits," *Proc. IEEE*, vol. 81, no. 12, pp. 1709-1726, Dec. 1993.
- [4] M. E. Kim et al., "GaAs Heterojunction Bipolar Transistor Device and IC Technology for High-Performance Analog and Microwave Applications," *IEEE Trans. Microwave Theory Tech.*, vol. 37, no. 9, pp. 1286-1303, Sept. 1989.
- [5] B. Bayraktaroglu, "GaAs HBT's for Microwave Integrated Circuits," *Proc. IEEE*, vol. 81, no. 12, pp. 1762-1785, Dec. 1993.
- [6] P. M. Asbeck et al., "GaAlAs/GaAs Heterojunction Bipolar Transistors: Issues and Prospects for Application," *IEEE Trans. Electron Dev.*, vol. 36, no. 10, pp. 2032-2042, Oct. 1989.
- [7] F. Ali and A. Gupta, editors, *HEMTs and HBTs : Devices, Fabrication, and Circuits*, Artech House, Boston, 1991.
- [8] J. Graffeuil and R. Plana, "Low Frequency Noise Properties of Microwave Transistors and their Application to Circuit Design," *1994 European Microwave Conference*, pp. 62-75, Sept. 1994.
- [9] N. L. Wang et al., "Ultrahigh Power Efficiency Operation of Common-Emitter and Common Base HBTs at 10GHz," *IEEE Trans. Microwave Theory Tech.*, vol. 38, no. 10, pp. 1381-1390, Oct. 1990.
- [10] M. A. Khatibzadeh and B. Bayraktaroglu, "Low Phase Noise Heterojunction Bipolar Transistor Oscillator," *Electron. Lett.*, vol. 26, pp. 1246-1247, 1990.
- [11] S. I. Long, "A Comparison of the GaAs MESFET and the AlGaAs/GaAs Heterojunction Bipolar Transistor for Power Microwave Amplification," *IEEE Trans. Electron Devices*, vol. 36, no. 7, pp. 1274-1278, July 1989.
- [12] M. Madihian et al., "A Low-Noise Microwave Oscillator Employing a Self-Aligned AlGaAs/GaAs HBT," *IEEE Trans. Microwave Theory Tech.*, vol. 37, no. 11, pp. 1811-1814, Nov. 1989.
- [13] G. D. Vendelin, A. M. Pavio and U. L. Rohde, *Microwave Circuit Design Using Linear and Non-Linear Techniques*, Wiley, NY, 1990.
- [14] G. G. Gonzalez, *Microwave Transistor Amplifiers: Analysis and Design*, Prentice-Hall, N.J., 1984.
- [15] E. L. Holzman and R. S. Robertson, *Solid-State Microwave Power Oscillator*

Design, Artech House, MA, 1992.

- [16] A. P. S. Khanna and J. Obregon, "Microwave Oscillator Analysis," *IEEE Trans. Microwave Theory Tech.*, vol. 29, no. 6, pp. 606-607, June 1991.
- [17] D. A. Warren, J. M. Golio and W. L. Seely, "Large and Small Signal Oscillator Analysis," *Microwave Journal*, pp. 229-243, May 1989.
- [18] W. Wagner, "Oscillator Design by Device Line Measurement," *Microwave Journal*, vol. 22, no. 2, pp. 43-48, Feb. 1979.
- [19] Hewlett-Packard Application Note 975, "A 4.3 GHz Oscillator Using the HXTR 4101 Bipolar Transistor."
- [20] H. Umeda, "Method of Optimum Oscillator Design Using Power-Dependent S-Parameters," *IEE Proceedings-G*, vol. 138, no. 4, pp. 533-539, Aug. 1991.
- [21] Petted, B. E., "Oscillator Computer-Aided Design Process Utilizes S-Parameter and Time-Domain Techniques," *Microwave Journal*, pp. 91-98, July 1990.
- [22] O. Pitzalis and Tom Reeder, "Oscillator Design with Non-Linear CAE," *Electronic Product Design*, pp. 23-25, Feb. 1991.
- [23] LibraTM V3.5 Reference Manual, EESOF, 5601 Lindero Canyon Road, Westlake Village, CA 91362.
- [24] LibraTM V3.5 Element Catalog, EESOF, 5601 Lindero Canyon Road, Westlake Village, CA 91362.
- [25] LibraTM V3.5 Applications Manual, EESOF, 5601 Lindero Canyon Road, Westlake Village, CA 91362.
- [26] OSA90/HopeTM V2.5 User's Manual, Optimization Systems Associates Inc., P.O. box 8083 Dundas, Ont., Canada, L9H 5E7.
- [27] C. T. Matsuno, A. K. Sharma and A. K. Oki, "A Large-Signal HSPICE Model for the Heterojunction Bipolar Transistor," *IEEE Trans. Microwave Theory Tech.*, vol. 37, no. 9, pp. 1472-1474, Sept. 1989.
- [28] HSPICETM User's Manual, Meta-Software Inc., 1300 White Oaks Road, Campbell, CA 95008.
- [29] D. A. Teeter et al., "Large-Signal Numerical and Analytical HBT Models," *IEEE Trans. Electron Devices*, vol. 40, no. 5, pp. 837-845, May 1993.
- [30] C. D. Parikh and F. A. Lindholm, "A New Charge Control Model for Single and Double-Heterojunction Bipolar Transistors," *IEEE Trans. Electron Devices*, vol. 39, no. 6, pp. 1303-1311, June 1992.
- [31] J. J. Liou and J. S. Yuan, "Surface Recombination Current of AlGaAs/GaAs Heterojunction Bipolar Transistors," *Solid-State Electronics*, vol. 35, no. 6, pp. 805-813, June 1992.
- [32] I. C. Kizilyalli and K. Hess, "Simplified Device Equations and Transport Coeffi-

- cients for GaAs Device Modeling," *IEEE Trans. Electron Devices*, vol. ED-34, no. 11, pp. 2352-2354, Nov. 1987.
- [33] M. E. Hafizi, C. R. Crowell and M. E. Grupen, "The DC Characteristics of GaAs/AlGaAs Heterojunction Bipolar Transistors with Application to Device Modeling," *IEEE Trans. Electron Devices*, vol. 37, no. 10, pp. 2121-2129, Oct. 1990.
 - [34] J. J. Liou, "Calculation of the Base Current Components and Determination of their Relative Importance in AlGaAs/GaAs and InAlAs/InGaAs Heterojunction Bipolar Transistors," *J. Appl. Phys.*, pp. 3328-3334, Mar. 1991.
 - [35] P. C. Grossman and J. Choma, Jr., "Large Signal Modeling of HBT's Including Self-Heating and Transit Time Effects," *IEEE Trans. Microwave Theory Tech.*, vol. 40, no. 3, pp. 449-464, March 1992.
 - [36] D. A. Teeter, J. R. East and G. I. Haddad, "High Frequency Large Signal Characterization of Heterojunction Bipolar Transistors," *Proc. IEEE/Cornell Conf. Advanced Concepts in High Speed Semiconductor Devices and Circuits*, pp. 315-324, Aug. 1991.
 - [37] P. Baureis, D. Seiter and U. Schaper, "Modeling of Self-Heating in GaAs/AlGaAs HBT's for Accurate Circuit and Device Analysis," *1991 IEEE GaAs IC Symposium*, pp. 125-128.
 - [38] P. Baureis, W. McKinley and D. Seitzer, "A New Large Signal Model for Heterojunction Bipolar Transistors Including Temperature Effects," *IEEE 1991 Custom Integrated Circuits Conference*, pp. 23.3.1-23.3.4.
 - [39] D. E. Dawson, A. K. Gupta and M. L. Salib, "CW Measurement of HBT Thermal Resistance," *IEEE Trans. Electron Devices*, vol. 39, no. 10, pp. 2235-2239, Oct. 1992.
 - [40] R. S. Vogel song and C. Brzezinski, "Extending Spice for Electro-Thermal Simulation," *IEEE 1989 Custom Integrated Circuits Conference*, pp. 21.4.1-21.4.4.
 - [41] B. K. Oyama and B. P. Wong, "GaAs HBT's for Analog Circuits," *Proc. IEEE*, vol. 81, no. 12, pp. 1744-1761, Dec. 1993.
 - [42] H. Wang et al., "Temperature Dependence of DC Currents in HBT," *1992 IEEE MTT-S Digest*, pp.731-734.
 - [43] H. Wang et al., "GaAs/GaAlAs Power HBTs for mobile communications," *1993 IEEE MTT-S Digest*, pp.549-552.
 - [44] P. Antognetti and G. Massobrio, *Semiconductor Device Modeling with SPICE*, NY, McGraw-Hill, 1988.
 - [45] F. S. Correra and S. Tadayon, "A Modified Gummel-Poon model applied to HBTs," *1993 European Microwave Conference*, pp. 96-98.
 - [46] Cascade Microtech, *Microwave Wafer Probe Calibration Constants-Instruction Manual*.
 - [47] D. Costa, W. U. Liu and J. S. Harris, Jr., "Direct Extraction of the AlGaAs/GaAs

- Heterojunction Bipolar Transistor Small-Signal Equivalent Circuit," *IEEE Trans. on Electron Devices*, vol. 38, no. 9, pp. 2018-2024, Sept. 1991.
- [48] J. R. Waldrop, K. C. Wang, and P. M. Asbeck, "Determination of Junction Temperature in AlGaAs/GaAs Heterojunction Bipolar Transistors by Electrical Measurement," *IEEE Trans. Electron Devices*, vol. 39, no. 5, pp. 1248-1250, May 1992.
 - [49] T. H. Ning and D. D. Tang, "Method for Determining the Emitter and Base Series Resistances of Bipolar Transistors," *IEEE Trans. Electron Devices*, vol. ED-31, no. 4, pp. 409-411, April 1984.
 - [50] J. S. Park and A. Neugroschel, "Parameter Extraction for Bipolar Transistors," *IEEE Trans. Electron Devices*, vol. 36, no. 1, pp. 88-95, Jan. 1989.
 - [51] S. A. Maas and D. Tait, "Parameter-Extraction Method for Heterojunction Bipolar Transistors," *IEEE Microwave and Guided Wave Letters*, vol. 2, no. 12, pp. 502-504, Dec. 1992.
 - [52] D. R. Pehlke and D. Pavlidis, "Evaluation of the Factors Determining HBT High Frequency Performance by Direct Analysis of S-Parameter Data," *IEEE Trans. Microwave Theory Tech.*, vol. 40, no. 12, pp. 2367-2373, Dec. 1992.
 - [53] D. R. Pehlke and D. Pavlidis, "Direct Calculation of the HBT Equivalent Circuit From Measured S-Parameters," *1992 IEEE MTT-S Digest*, pp. 735-738.
 - [54] M. Cacho, A. Werthof and G. Kompa, "Broadband 40GHz Si/SiGe HBT Equivalent Circuit Using a Successive Analytical Model Parameter Extraction," *1993 European Microwave Conference*, pp. 515-517.
 - [55] D. Costa, W. Liu and J. S. Harris, Jr., "A New Direct Method for Determining the Heterojunction Bipolar Transistor Equivalent Circuit Model," *IEEE 1990 Bipolar Circuits and Technology Meeting 5.6*, pp. 118-121.
 - [56] S. J. Prasad, "A Method of Measuring Base and Emitter Resistances of AlGaAs/GaAs HBTs," *IEEE 1992 Bipolar Circuits and Technology Meeting 10.2*, pp. 204-207.
 - [57] S. J. Prasad, "Monitoring the Base and Emitter Resistances in An AlGaAs/GaAs HBT Line," *1992 IEEE GaAs IC Symposium*, pp. 271-274.
 - [58] B. A. Kramer and R. J. Weber, "Base-Emitter Diffusion Capacitance in GaAlAs/GaAs HBTs," *Electronics Lett.*, vol. 28, no. 12, pp. 1106-1107, June 1992.
 - [59] Gateway Modeling, Thermal Workshop, *IEEE 1993 MTT-S*.
 - [60] D. M. Drury, D. C. Zimmermann and W. A. Davis, "Large-signal Characterization of Dual-Gate Field Effect Transistors Using Load-Pull Measurements," *IEEE Trans. Microwave Theory Tech.*, vol. 41, no. 2, pp. 183-189, Feb. 1993.
 - [61] F. M. Ghannouchi, R. Larose and R. G. Bosisio, "A New Multiharmonic Loading Method for Large-Signal Microwave and Millimeter-Wave Transistor Characterization," *IEEE Trans. Microwave Theory Tech.*, vol. 39, no. 6, pp. 986-992, June 1991.

- [62] H. Do-Ky, T. Laneve et al., "Linear and Nonlinear Modeling Verification of Power HBTs, *1994 European Microwave Conference*, pp. 810-813.
- [63] S. Sychaleun, T. Laneve et al., "GaAs/AlGaAs Modeling for Microwave Circuit Applications," *ANTEM 1994*, pp. 538-541.
- [64] Hewlett-Packard Product Note 11729C-2, "Phase Noise Characterization of Microwave Oscillators: Frequency Discriminator Method".
- [65] Hewlett-Packard Application Note 207, "Understanding and Measuring Phase Noise in the Frequency Domain," Oct. 1976.
- [66] R. Hajji, F. M. Ghannouchi, and A. B. Kouki, "Small-Signal Modeling of High Power HBTs Using the Scaling Approach," *1994 MTT-S Digest*, pp. 1249-1252.
- [67] Focus Microwave Inc., *Computer Controlled Microwave Tuner Operation Manual*, Release 4.0, Jan. 1994.
- [68] D. d'Almeida and R. Anholt, "Device Characterization with an Integrated On-Wafer Thermal Probing System," *Microwave Journal*, pp. 94-105, Mar. 1993.
- [69] A.P.S Khanna and Y. Garault, "Determination of Loaded, Unloaded, and External Quality Factors of a Dielectric Resonator Coupled to a Microstrip Line," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, no. 3, pp. 261-264, Mar. 1983.
- [70] R. G. Rogers, *Low Phase Noise Microwave Oscillator Design*, Artech House, MA, 1991.

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