GaAs MESFET MODELING AND ITS APPLICATIONS

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Chapter 1

INTRODUCTION

1.1 The Importance of GaAs FET Modeling

Beginning in the early nineteen seventies, due to the earnest demand for high-speed solid-state devices working in microwave frequencies, as well as devices equipped with more sophisticated techniques and technology in circuit design and semiconductor materials for the realization of circuits, Gallium Arsenide metal-semiconductor field-effect transistor (GaAs MESFET) has been being developed ever since because of its high mobility of electrons. GaAs MESFETs find many applications at high frequencies and in microwave monolithic integrated circuits (MMICs). They can be used as a basic function of amplification and switching as required for low-noise amplifiers, broadband amplifiers, power amplifiers and microwave switches[1]. More attention has been given to the GaAs MMIC as it becomes mature and ready for commercial applications. Another factor which makes GaAs MMIC more attractive and cost-effective is that GaAs is the only substrate material available for fabricating both discrete transistors and monolithic ICs capable of operating in microwave frequencies above 4 GHz. Meanwhile, developing CAD tools for circuit analysis, simulations and designs turns out to be a parallel process with this new technology, since in
any integrated circuit design one usually begins with a computer simulation of the circuit concerned. As a result, the question -how to set up a simple, yet precise mathematical model for GaAs MESFET- becomes a prominent issue. A circuit model for GaAs MESFET is needed in almost all software and simulators to effectively reproduce the physical property of the device in any new design and realization. A better model subsequently determines how viable and accurate the designed circuit is going to be. In other words, the better the model of GaAs FET used in simulation, the smaller the gap would be between the circuit-on-paper and the real product. Furthermore, a good model would shorten the design and fabricating process, increasing efficiency. In the last several years, modeling GaAs MESFET proved to be an uneasy job, mostly due to its complexity of device geometry, nonlinear characteristics of the device and the tedious, long process of nonlinear model parameters extraction. Hereunder several of the most popular models will be discussed and some modifications will be made in due course throughout the thesis.

1.2 Advantages and disadvantages of extensively used Models

There are three major circuit models which have been developed and used extensively in many time-domain or frequency-domain circuit simulation programs, such as PSPICE,
MWSPICE, LIBRA, etc. These are the Curtice quadratic model, the Statz model (Raytheon), and the Triquint model[2,5-7].

The Curtice model stresses the square-law relationship between the drain-source current $I_{ds}$ and the gate-source voltage $V_{gs}$.

$$I_{ds} = \beta (V_{gs} - V_t)^2 (1 + \lambda V_{ds}) \tanh(\alpha V_{ds})$$  \hspace{1cm} (1.2.1)

Note $V_t$ is threshold voltage($V_{to}$). The mathematical model uses the hyperbolic tangent ($\tanh$) function to predict the advent of saturated current and $(1+\lambda V_{ds})$ to describe the current variation when the transistor is operating in the saturated region. This model has been successfully implemented in most CAD programs because of its simplicity. Yet it is not very accurate when large-signal microwave circuit simulation is needed, especially for devices with large pinch-off voltage[2].

The Statz model, also known as the Raytheon model, in general, has retained the framework of the Curtice model, but enhanced it with several significant modifications.

$$I_{ds} = \frac{\beta (V_{gs} - V_t)^2}{1 + b(V_{gs} - V_t)} (1 + \lambda V_{ds}) (1 - (1 - \frac{\alpha}{3} V_{ds})^3)$$  \hspace{1cm} (1.2.2)

For the drain-source current $I_{ds}$, the Raytheon model replaces $\beta(V_{gs}-V_t)^2$ by $(\beta(V_{gs}-V_t))/(1+ b(V_{gs}-V_t))$ to smooth the connection between the small value of $(V_{gs}-V_t)$ and the large
value of \((V_{gs} - V_t)\). Another change is to replace the tanh function with an approximation of \((1-(1-(\alpha V_{ds}/3)^3))\) to reduce the computing time[6]. The Raytheon model provides an improved representation of the capacitor behavior. It has a symmetric expression for \(C_{gs}\) and \(C_{gd}\) capacitors, which is used for normal and inverse operation of the device[1]. The expression \((1 + \lambda V_{ds})\) still fails to predict the drain conductance at the wide range of the bias points. In addition, the Raytheon model is unable to track \(R_{ds}\) at very low current where \(V_{gs}\) is near cutoff and at a high current level.

The Triquint model is proposed to remedy these pitfalls. Toward the end, the Triquint model generalizes the square law relationship between the drain current and \(V_{gs}\) introducing parameter \(Q\). It presents \(V_t\) as a function of \(V_{ds}\) to get a better fit at near pinch-off value of \(V_{gs}\). Finally it adds a power feedback to the drain current, namely

\[
I_{ds} = \frac{I_{dso}}{1 + \delta V_{ds} I_{dso}}
\]

(1.2.3)

where \(I_{dso}\) is similar to \(I_{ds}\) in equation (1.2.2) (detail will be discussed in chapter 2). Since the model has a better match to the DC I-V characteristics over a wider bias range, it would provide a better small-signal model for AC characteristics as well[7]. The most salient part is that the improved model enhances the accuracy without sacrificing the simplicity of
1.3 Optimization of Model Parameters for Nonlinear Analysis

The nonlinear model of GaAs MESFET is made of a variety of parameters, some of them are closely related to device physics, while others are more like curve-fitting parameters. These parameters represent the physical device of FET in the simulation of any circuit design and optimization. A good model, therefore, results in a collection of very accurately extracted model parameters. In many applications, FET must be engineered with a very high degree of precision and accuracy if the chip is to be working within specifications[1]. This makes the FET model even more important in analyzing the effect of changes of materials and device structure. Since fabrication of new MMIC requires extremely precise design specifications, this pushes the parameters obtained to be as accurate as possible. That is the reason why optimization technique became necessary and essential.

Electrical engineers have been successfully taking advantage of the optimization theory in practice. There are a few very powerful optimizing techniques available to be chosen, depending on the nature of the problems and functions to be optimized. The problem becomes more complex and difficult when the defined function turns from linear to nonlinear function, or the number of variables increases. The
problem we will encounter in this research is optimization of an error function, which is defined as the accumulated sum of errors between measured and calculated values of AC or DC characteristics of the device. In order to minimize the error and increase the accuracy of the model by changing the model parameters, a stable, reliable and fast optimizing routine is needed. The Fletcher-Powell optimization technique[3], though it is not the most popular method used contemporarily, has been chosen and has proved to be an effective and sufficient tool to tackle the problem involved. The Fletcher-Powell method, in fact, is a modification of the steepest-decent optimization technique[3]. It is worth mentioning that the Fletcher-Powell method works best with a set of reasonable initial model parameters in this application, which means the methods we utilize to evaluate starting parameter values, and how accurate these values are before optimization, can be determinant. Initial value estimation has been a major effort in our research.

1.4 Research Objective

With all these mathematical models in hand, one could implement them in software packages and apply them in circuit simulations. As mentioned above, a nonlinear model depends on a number of parameters. But the procedure of extraction might be long and tedious, and extracted parameters might be
inaccurate and impractical. In this research, a complete procedure for extracting model parameters of GaAs FETs will be discussed. DC and AC model parameters will be extracted based on the device's measured data, both I-V characteristics and S-parameters. S-parameter is usually used to describe AC characteristics of microwave devices at small-signal analysis. The algorithm for optimizing DC and AC parameters will be defined in a way that the model will match as close as possible with the device. For DC parameters, the optimized parameters come out from the appropriate match of measured and modeled I-V characteristics. For AC parameters the parameters are optimized based on the physics property that: the intrinsic element of the equivalent circuit of GaAs FET is independent of frequency[8]. In our modeling, emphasis is placed on physically consistent models which should closely match physics based parameter values. Three devices and their measurements have been used in our research to extract model parameters: one is a half-micron gate length single ion-implanted FET fabricated by Rockwell; another is a one micron gate length ion-implanted device made by Wright-Patterson AFB Electronic Technology Laboratory in-house facilities, and the final device is 0.25μ*2*50μ HEMT also from Wright-Patterson AFB. Fortran programs have been written to carry out these optimization tasks. One is for DC optimization, the other is for AC optimization. An optimizing routine has been written based on the Fletcher-Powell method[3], with additions of
constraints, normalization of variables, and proper weighing of parameters facilities in the program. With the model parameters in hand, the accuracy of these parameters can be tested using different software accommodating these models.

PSPICE[22], a powerful simulation program developed out from the original SPICE program, is generally used to check our extracted models, circuits and designs. With additional input files, it may be used to analyze microwave circuits as well, such as S-parameter analysis. MWSPICE[29], similar to PSPICE, but with more microwave analysis technique in it, is mainly used to simulate microwave circuits. LIBRA[23], another useful CAD tool for network matching and small or large signal analysis and design of microwave circuits, is used in our research as well. The results of optimizations and simulations, which are satisfactory and close to the measured data, will be shown in the following chapters. The extracted parameters indicate that one model may be better than another in describing certain electrical properties of FETs. In using the Triquint model for optimization threshold voltage $V_{to}$ (pinch-off voltage) has been added as a variable inside the program. The result from this model will be shown as well. Finally the application of GaAs FET used as operational transconductance amplifier (OTA) [18,19,20,21] will be discussed and explored. We have designed an OTA circuit based on SPICE nonlinear models. The OTA block that could be used as an integrator or band-pass filter are simulated using the
Raytheon model in LIBRA. We use Triquint device data for this design[25]. Both schematic and layout circuits have been simulated to show characteristics of the circuits. The layout of the circuits could also be obtained using ACADEMY.

In the second chapter, the procedure of extraction and optimization of DC parameters will be presented. The third chapter is to introduce the procedure to extract AC parameters. The fourth chapter combines both DC and AC parameters as a whole to obtain nonlinear models used in popular simulators, such as PSPICE, MWSPICE and LIBRA. The results from different models will be presented and compared in this chapter. In chapter five, the application of GaAs FET will be discussed. An OTA circuit designed using GaAs FETs and its applications as integrators and band-pass filters, will be analyzed and presented. The last chapter is the conclusions and suggestions for future work.
Chapter 2

DC Nonlinear Modeling

This chapter will start with an introduction of numerical methods in applications of modeling of electronic devices. It would emphasize its usage on modeling GaAs FETs. An algorithm to extract and optimize DC model parameters of MESFETs will be presented. The measurements of DC curves of 0.5*180 micron FET device from Rockwell are used to extract and verify the optimized DC parameters.

2.1 Optimization based on Numerical Methods

The optimization theory has been used in many applications in electrical engineering. It has become part of circuit designs, circuit simulations and modeling of a new device. To improve the output of a system, the relationship of the output function and intrinsic or extrinsic variables should be carefully investigated. Changing one of the variables in the system would more or less affect the performance of the system. Every variable has a unique characteristic within the function. A function or a system usually has more than one variable. In order to have the system work within the specifications, a set of parameter values should be chosen. Sometimes, to achieve a desirable output of the system, or under more stringent conditions, one
needs to optimize the result by selecting a new set of parameters with the help of the numerical optimization technique. There are many optimizing methods available: steepest-descent, Fletcher-Powell and least pth, just to mention a few. Fletcher-Powell is a modified version of the steepest-descent method.

A function to be optimized may be defined as

\[ F(x) = E(x) \quad (2.1.1) \]

where \( x \) is a vector of components \( x_1, x_2, \ldots, x_n \). By Taylor’s expansion, a differentiable function \( E(x) \) can be written as\(^3\)

\[ E(x + \delta x) = E(x) + \sum_{j=1}^{n} \frac{\partial E}{\partial x_j} \delta x_j + \frac{1}{2} \sum_{j=1}^{n} \sum_{j'=1}^{n} \frac{\partial^2 E}{\partial x_j \partial x_{j'}} \delta x_j \delta x_{j'} + \ldots \quad (2.1.2) \]

which can be approximated as a quadratic function

\[ E(x + \delta x) = E(x) + \sum_{j=1}^{n} \frac{\partial E}{\partial x_j} \delta x_j + \frac{1}{2} \sum_{j=1}^{n} \sum_{j'=1}^{n} \frac{\partial^2 E}{\partial x_j \partial x_{j'}} \delta x_j \delta x_{j'} \quad (2.1.3) \]

or

\[ E(x + \delta x) = E(x) + \nabla E^T \delta x + \frac{1}{2} \delta x^T H \delta x \quad (2.1.4) \]

where \( \nabla E \) is the gradient. The symbol \( H \) stands for the Hessian matrix, where

\[ H = \left[ \frac{\partial^2 E}{\partial x_j \partial x_j} \right] \quad (2.1.5) \]

Note that the right hand side of (2.1.4) is the approximation
of the original function. It contains not only first derivative, but also second derivative of the function. As we have indicated before, an optimization routine usually tries to minimize an error function. In using the Steepest-descent optimization technique, the very first step is to find the direction of each parameter in the function to which the parameter should approach. That is, to find the partial derivative of the function with respect to that parameter in the function. To the end that subroutine chooses the slope-following method, in which to search for a minimum in the direction of the negative gradient\(^3\). The technique is based on the fact that the negative gradient points in the direction of the fastest set of decrease. That is,

\[
S = -\nabla E
\]  

(2.1.6)

with a second derivative in equation (2.1.4). The Newton method is used to find the direction. The new search direction has been modified as\(^3\)

\[
S = -H^{-1}\nabla E
\]  

(2.1.7)

The third term in equation (2.1.4) obviously makes the approximation of the error function more accurate, yet introduces second derivative of \(E(x)\), which involves heavy computation. Therefore, an approximation is employed in the Fletcher-Powell method. It is based on the evaluation of the gradient and simulation of the inverse of Hessian matrix\(^3,4\).
The approximation is approached by substituting the inverse of the Hessian matrix by a matrix G, because inverse of the Hessian is the one needed in the computation. G starts with an identity matrix I. Then it would iteratively approaches $H^{-1}$. Let $G = I$, I is

$$I = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ \vdots & \ddots & \ddots \\ 0 & 0 & 0 & 1 \end{bmatrix}$$  \hspace{1cm} (2.1.8)

Then use $S = -G \cdot \text{GRAD}(x)$ for a distance $dx$. For the parameter change $dx$, we have the gradient change

$$y = \text{GRAD}(x') - \text{GRAD}(x)$$  \hspace{1cm} (2.1.9)

From the change of the gradient at $x' = x + dx$, new G can be obtained as follows [3]:

$$G = G + \frac{\Delta x \Delta x^T}{d_1} - \frac{(Gy)(Gy)^T}{d_2}$$  \hspace{1cm} (2.1.10)

where $d_1$ and $d_2$ are scale factors as follows:

$$d_1 = y^T \Delta x \quad d_2 = y^T Gy$$  \hspace{1cm} (2.1.11)

Since G is initially an identity matrix, the first Fletcher-Powell iteration is equivalent to the steepest descent, but the following iterations apply a better approximation for $H^{-1}$ to improve direction-search.
There are two important points to be kept in mind when utilizing an optimization program. One is that the minimum of a function found might be a local minimum instead of a global minimum. The other, which happens more often to practicing engineers, is that the outcome after using an optimization program might not have physical significance. That means it produces unrealistic parameter values which are meaningless to the one who uses them. The first problem could be dealt with by using a different set of initial parameters and/or optimizing the problem with a combination of several optimizing techniques. Some recent techniques may use a "disturbing" facility when the error can no longer be reduced. The subroutine would probe randomly in some other directions to minimize the chance of falling into a local valley. For the latter, one can install constrains, lower and upper limits, in the program to ensure the variations of parameters be within a reasonable range. The negative side-effect of added constrains may deteriorate the efficiency of the algorithm and thus result in obtaining a greater error, which could be averted if no constraint is imposed.

To avoid falling in a local minimum and using a more complex optimization routine, choosing good starting parameter values is necessary and important. Actually the success of optimization depends on the choice of proper initial parameters.
2.2 Procedures to Obtain Initial Fitting Parameters

For nonlinear analysis of circuits including GaAs FETs, a model, consisting of all model parameters, is required to characterize the FETs. In PSPICE, a large-signal model (nonlinear model) for n-channel JFET is modified and used for GaAs FET as well. As shown in Fig. 2.1, three nonlinear elements are $C_{gd}$, $C_{gs}$ and $I_{ds}$. Among them drain conductance is the main electrical property characterizing the operation of GaAs MESFETs, which is usually emulated by a mathematical model. Usually SPICE models are loosely divided into two categories: DC parameters and AC parameters. There are some other parameters related to physical structure of FETs. More about the SPICE model will be discussed in chapter 4. Here we concentrate on how DC parameters can be extracted based on device measurements. Since the Raytheon model is an improved model for the Curtice model, more attention will be drawn onto this model. A complete procedure to extract an initial DC parameter by using the Raytheon model will be presented. A similar process may also apply to the Triquint model and the modification of the Triquint model.

For Raytheon model, alpha, beta, lambda and $b$ are fitting parameters. The drain current of MESFET may be written as

$$I_{ds} = \frac{\beta (V_{gs} - V_{t})^2}{1 + \beta (V_{gs} - V_{t}) (1 + \lambda V_{ds}) (1 - (1 - \frac{3}{\alpha} V_{ds})^{3})}$$
V\textsubscript{t} is equal to the threshold voltage (pinch-off voltage) \( V_{to} \) for the Raytheon model. It should be noted that internal voltage \( V_{gs} \) and \( V_{ds} \) used in (2.2.1) are related to the external voltages with the following relationships

\[
V_{gs} = V_{ds} - I_{ds} R_s, \quad V_{ds} = V_{ds} - (R_s + R_d) I_{ds}
\]  

(2.2.2)

\( R_s \) and \( R_d \) are parasitic source and drain resistances. They are caused by the ohmic contact of interface between metal and semiconductor. The values of \( R_s \) and \( R_d \) utilized in Eqn(2.2.2) are extracted from the ACMOD program (details discussed in Chapter 3). Extraction and optimization of alpha, beta, lambda and \( b \) is then a necessary step to establish a good model. First, one needs to estimate initial parameters values from a device's measured I-V characteristics. Beta may be approximated by using[8]

\[
\beta = \frac{I_{ds}}{(V_{gs} - V_{t})^2}
\]  

(2.2.3)

at the lowest drain current data point. The value of \( b \) is obtained by applying
at the highest drain current point. The initial alpha and lambda are estimated by inspecting the shape of I-V curves with the estimated initial beta and b. The inspection can be easily done on a PC having MATHCAD[30] installed. This would give plots and data of calculated I_{ds} compared to the measured I_{ds} in order to find suitable values for alpha and lambda. The parameter values extracted this way can achieve a certain accuracy as our optimization program would justify it in the following section.

2.3 Optimization and Extraction of DC Parameters Using Curve Fitting

With estimated initial DC parameters, the next step is to optimize these parameters to minimize the error function. First, an error function has to be defined. DC error function is a sum of the difference of modeled and measured I-V characteristics for V_{gs}, usually in the range between V_{to} and 0.5V, and gate voltage is above pinch-off (cut-off). Three ways can be used to formulate the error function: absolute difference \(|I(i)_{msd} - I(i)_{cal}|\), percentage of absolute difference \(|(I(i)_{msd} - I(i)_{cal})/I(i)_{msd}|\), and square of percent difference \(((I(i)_{msd} - I(i)_{cal})/I(i)_{msd})^2\). Different error functions would
produce different error values. In return, they affect the numerical optimization algorithm. The sum of absolute difference gives mere error and would lead the algorithm to iterate stably to reduce the error. But it is usually a large number, not a good indication of error at a bias point, especially for the final result. This method treats error due to each bias equally. The percentage of total error gives a good indication of the final error for each bias point. Note that the highest error for each bias is unity; however, sometimes we may encounter error which is more than one for a certain bias condition, because measured value is less than unity. We may need to scale the measured data, or use more accurate initial parameters. Otherwise error function may mislead iterations to a local minimum and produce bizarre results. The square of total error exponentially boosts the error for some bias points. As a return it would lead the algorithm to work on biases over which the model shows poor prediction. The error function is thus chosen as follows:

\[
\text{Err} = \frac{1}{mn} \left( \frac{1}{m} \sum_{i=1}^{m} \sum_{j=1}^{n} \left( \frac{I(ij)_{\text{meas}} - I(ij)_{\text{cal}}}{I(ij)_{\text{meas}}} \right)^2 \right) \quad (2.3.1a)
\]

\[
\text{Err} = \frac{1}{mn} \left( \frac{1}{m} \sum_{i=1}^{m} \sum_{j=1}^{n} \left( \frac{I(ij)_{\text{meas}} - I(ij)_{\text{cal}}}{I(ij)_{\text{meas}}} \right)^2 \right)^{\frac{1}{2}} \quad (2.3.1b)
\]

\[m \] is the number of different \( V_{gs} \), and \( n \) is the number of points adopted on x-axis \( V_{ds} \). A Fortran program, named DCMOD, has been
written to carry out the task, which is to minimize error function by adjusting DC parameters. Beta, b, lambda and alpha are variables for the Raytheon model in the program.

From equation (2.3.1), one can observe that $I_{ds}$ is linearly proportional to beta, the transconductance parameter, which is a major factor in the equation. Parameter b, a doping tail extending parameter[11], is also important because it is inversely proportional to the current. Alpha, the drain-source current saturation knee potential, and lambda, the channel-length modulation which indicates the slope of the I-V curve in the saturation region, are less significant in the equation in terms of how effectively they would affect the current value. After having employed the optimization program, we are able to justify the fact that the normalized gradient of error function with respect to beta is much greater than those of others at early iterations. It means that beta would tend to go to extremes and or bonds (if bonds are defined in the optimizing routine), while others would tend to stay more or less the same. In addition, an inaccurate beta would contribute the most error in the error function. To avoid extreme change of beta, one could normalize or scale the entry variables before optimization and weigh them accordingly as well. The success of applying an optimizing technique to extract model parameters depends heavily on good starting parameter values. In some cases, poor starting values could cause parameters to fluctuate and produce unreasonable
results. Local minimum and failure to converge may be also encountered if initial values are picked randomly. In general, the initial approximated parameter can be obtained from the aforementioned procedure, and the optimization program would improve the accuracy of the parameters and refine the transistor model further.

2.4 Discussion on the Optimized Results and Triquint Model

The proposed method has been applied to extract DC parameters for half-micron gate length, single ion-implanted FET. An I-V curve created using the Raytheon model comparing the measured and calculated $I_{ds}$ is illustrated in Fig. 2.1. Table 2.1, which lists initial and final optimized parameters, is presented as well. From Fig. 2.1, one can observe that the Raytheon model does not perform well at large current levels. Neither is the onset of saturation of current predicted by the model. The Triquint model, which has been newly added to the PSPICE, is proposed by McCamant et al[6]. This model improves the poor tracking of the drain-source resistor of the Raytheon model. To improve the drain conductance fit at low drain current, $V_t$ is used as a function of drain voltage

$$V_t = V_{to} - \gamma V_{ds} \tag{2.4.1}$$

The second modification intends to reduce $I_{ds}$ at high level current and voltage by using the feedback effect of delta:
The new model replaces alpha and b in the Raytheon model with gamma, delta, and Q. Gamma is in the range of 0.03 to 0.05 [6]. Initially Q is set equal to 2. We have used an optimization program to improve the model parameters for the Triquint model. After optimization, the results show a better fit. Fig.2.2 illustrates the measured Ids and computed Ids for the Triquint model. A table of initial and final parameter values is also presented in Table 2.2.

The Triquint model improves the fitting for the drain current, yet a closer I-V curve fitting is desired. In applying the Triquint model a change in algorithm is proposed by introducing Vt0 as a variable in the optimization program DCMOD. Originally Vt0 is estimated from the measured I-V characteristics; it is hard to give an exact threshold voltage by inspection. An error may be involved. The advantage of making Vt0 as a variable in the program is to reduce the error. The results show a better fit at the saturation region at a high current level. The comparison is depicted in Fig.2.3 with the parameter values given in Table 2.3.1.
Table 2.1 The initial parameter and final values after optimization by using the Raytheon model.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Initial Value</th>
<th>Final Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>alpha</td>
<td>2.5</td>
<td>1.0869</td>
</tr>
<tr>
<td>beta</td>
<td>0.016</td>
<td>0.0088</td>
</tr>
<tr>
<td>lambda</td>
<td>0.001</td>
<td>0.001</td>
</tr>
<tr>
<td>( b )</td>
<td>0.009</td>
<td>0.009</td>
</tr>
<tr>
<td>( V_0 )</td>
<td>-2.0</td>
<td>-2.0</td>
</tr>
<tr>
<td>Error</td>
<td>9.75%</td>
<td>2.81%</td>
</tr>
</tbody>
</table>

Table 2.2 The initial parameters and final values after optimization by using the Triquint model.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Initial Value</th>
<th>Final Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>alpha</td>
<td>1.8</td>
<td>1.6304</td>
</tr>
<tr>
<td>beta</td>
<td>0.009</td>
<td>0.0068</td>
</tr>
<tr>
<td>gamma</td>
<td>0.04</td>
<td>0.0306</td>
</tr>
<tr>
<td>delta</td>
<td>1.0</td>
<td>0.0006</td>
</tr>
<tr>
<td>( Q )</td>
<td>2.0</td>
<td>2.1015</td>
</tr>
<tr>
<td>( V_0 )</td>
<td>-2.0</td>
<td>-2.0</td>
</tr>
<tr>
<td>Error</td>
<td>4.91%</td>
<td>2.65%</td>
</tr>
</tbody>
</table>

Table 2.3 The initial parameters and final values after optimization by using a modified model.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Initial Value</th>
<th>Final Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>alpha</td>
<td>1.8</td>
<td>1.4035</td>
</tr>
<tr>
<td>beta</td>
<td>0.009</td>
<td>0.0071</td>
</tr>
<tr>
<td>gamma</td>
<td>0.04</td>
<td>0.0277</td>
</tr>
<tr>
<td>delta</td>
<td>1.0</td>
<td>0.0005</td>
</tr>
<tr>
<td>( Q )</td>
<td>2.0</td>
<td>1.9233</td>
</tr>
<tr>
<td>( V_0 )</td>
<td>-2.0</td>
<td>-2.0215</td>
</tr>
<tr>
<td>Error</td>
<td>4.91%</td>
<td>1.53%</td>
</tr>
</tbody>
</table>
Figure 2.1 GaAs MESFET large-signal model.
I-V Curves

Figure 2.2 Plot of the drain current for the Raytheon model compared to the measured drain current.

Figure 2.3 Plot of the drain current for the Triquint model compared to the measured drain current.
Fig. 2.4 Plot of the drain current for the modified model in which $V_{to}$ is a parameter, compared to the measured current.

---------- measured
.......... modeled
Chapter 3
AC Linear Modeling

This chapter is intended to present AC modeling for GaAs MESFETs. AC parameters extraction and optimization algorithm will be discussed in detail with its application on a research FET device from Wright-Patterson AFB Electronic Technology Laboratory. S-parameters for small-signal analysis will be generated using a final optimized model to compare to the measured value to verify the model and to seek possibilities for improvement of accuracy.

3.1 AC Equivalent Circuit Topology for GaAs FETs

In contrast with DC nonlinear modeling, AC modeling is considered linear because a circuit is operating at a fixed bias point with small input and output signals. For a study of AC characteristics of GaAs FETs, an explicit equivalent circuit model is necessary. An accurate, sophisticated equivalent circuit which is able to precisely describe the behavior of transistors under AC analysis has been searched and developed for years to meet the accuracy requirement [8,14,15,16]. The one which has been widely accepted as an equivalent circuit for GaAs FET, shown in Fig. 3.1, has some limitations when the frequency is too high(above 60 GHz), and it is not applicable at some DC operating points. A more
complicated, extended small-signal equivalent circuit of FET proposed by M. Benoth and R. Rosch[6] is shown in Fig. 3.2. The new circuit has a similar topology to the old one, but with some additional facilities within the intrinsic device. The new model takes into account the gate currents and resistive feedback. According to the device physics and structure, FET's equivalent circuit has been categorized as two sections: the intrinsic device and the extrinsic circuit. Therefore, components in the intrinsic device are intrinsic elements. The major distinction between intrinsic and extrinsic elements is that extrinsic elements are bias invariant, that means they are constant with bias condition, while intrinsic elements are not. That makes intrinsic elements more unpredictable. External elements include parasitic components such as $L_g$, $L_d$, $L_s$, $R_g$, $R_d$, $R_s$, $C_{pg}$, $C_{pd}$, etc. Intrinsic components contain 10 elements altogether in our case. These elements make up part of the nonlinear GaAs FET model.

To extract AC model parameters, one needs first to determine initial values of these element values according to FET physics and geometry[1,14], especially for external elements to be used in the optimization program. Generally they may be computed as follows:

$$L_g = \frac{\mu_0 qZ_G}{m^2 L_g}$$

(3.1.1)
\[ L_g = \frac{L_g}{3} = \frac{\mu_0 dZ_G}{3mc^2 L_G} \]  \hspace{1cm} (3.1.2)

\[ L_d = \frac{L_g}{2} = \frac{\mu_0 dZ_G}{2mc^2 L_G} \] \hspace{1cm} (3.1.3)

\[ R_g = \frac{\rho Z_G}{3mc^2 \hbar L_G} \] \hspace{1cm} (3.1.4)

\[ R_s = \frac{\rho C}{Z_G} + \frac{L_{SG}}{\eta N \hbar \omega Z_G} \] \hspace{1cm} (3.1.5)

\[ R_d = \frac{(L_{GD} - \lambda)}{[1 - \frac{2eV_{sat} (W-d)}{\mu_0 \omega q^2 N_d^2}]} \] \hspace{1cm} (3.1.6)

\[ C_{p1} = \frac{eZ_G h_s}{L_{SG}} \] \hspace{1cm} (3.1.7)

\[ C_{p2} = \frac{eZ_G h_d}{L_{GD} + L_{SG} + L_G} \] \hspace{1cm} (3.1.8)
where the parameters on the right hand side of eqns (3.1.1-3.1.8) stand for the electrical characteristics of semiconductor materials, physical structure, dimensions and doping of MESFETs.

\( \mu_0 \): permeability of free space

\( d \): depletion depth

\( m \): number of parallel strips into which the total gate width is divided

\( Z_0 \): the gate width

\( L_G \): the metallurgical gate length

\( \rho \): resistivity

\( h \): the height of gate strip

\( q \): electron charge

\( c \): capacitance

\( N \): the doping density of in the N-channel layer

\( W \): the thickness of the N-layer under the gate

\( W_r \): the depth of the gate recess

\( W_s \): surface depletion depth

\( \varepsilon \): the permitivity

\( v_{sat} \): scatter-limited velocity

\( L_{GD} \): the gate-drain separation

\( L_{SG} \): the source-gate separation

\( h_s \): the height of the metal at source and gate

\( h_d \): the height of the metal at drain and source

\( x \): extension of space-charge layer into the gate-drain space
3.2 **AC Parameters Extraction and Optimization**

With equivalent circuit available, plus the initial estimation of extrinsic elements, an optimizing procedure for refining these values and extracting intrinsic elements' values has been developed and tested on Sun Workstations. A Fortran program, named ACMOD, which employs the Fletcher-Powell method for optimization with added facilities, is used to carry out the task. The program consists of three parts: a main program, an error function and the optimizing routine.

There are two objectives in applying this program: first is to optimize external elements, second is to extract internal elements and then optimize them. All parameters will be calculated based on given measured S-parameters for a FET operating at a given bias point. An algorithm has been set up in which steps are taken to the isolate intrinsic device[14]. With the given measured S-parameters, a procedure that extracts the intrinsic parameter elements for the circuit of Fig. 3.2 is as follows:

(1) Convert S-parameter to Z-parameter
(2) Remove effect of $L_g$ and $L_d$ from the Z-parameter
(3) Convert the Z-parameter to the Y-parameter
(4) Remove effect of $C_{ps}$ and $C_{pd}$ from the Y-parameter
(5) Convert the Y-parameter to the Z-parameter
(6) Remove $R_g$, $R_d$, $R_s$ and $L_s$ from the Z-parameter
(7) Convert X-parameter to Y-parameter

The final Y-parameters are in the form of the 2x2 Y-matrix.
From the Y-parameters, one is able to compute the intrinsic elements by applying the following relations [15]:

\[
C_{gd} = \frac{Im(Y_{12})}{\omega} \left( 1 + \left( \frac{Re(Y_{12}) + g_{fd}}{Im(Y_{12})} \right)^2 \right)
\]  

(3.2.1)

\[
R_{gd} = \frac{Re(Y_{12}) + g_{fd}}{\omega C_{gd} Im(Y_{12})}
\]  

(3.2.2)

\[
C_{gs} = \frac{Im(Y_{11}) + Im(Y_{12})}{\omega} \left( 1 + \left( \frac{Re(Y_{11}) + Re(Y_{12}) - g_{fs}}{Im(Y_{11}) + Im(Y_{12})} \right)^2 \right)
\]  

(3.2.3)

\[
R_{i} = \frac{Re(Y_{11}) + Re(Y_{12}) - g_{fs}}{\omega C_{gs} (Im(Y_{11}) + Im(Y_{12}))}
\]  

(3.2.4)

\[
g_{m} = \sqrt{((Re(Y_{21}) - Re(Y_{22}))^2 + (Im(Y_{21}) - Im(Y_{22}))^2)} D1
\]  

(3.2.5)

\[
r = \frac{1}{\omega} \arcsin\left( \frac{Im(Y_{12}) - Im(Y_{21}) - \omega C_{gs} R_{i} (Re(Y_{21}) - Re(Y_{12}))}{g_{m}} \right)
\]  

(3.2.6)
where

\[ C_{ds} = \frac{\text{Im}(Y_{22}) + \text{Im}(Y_{12})}{\omega} \quad (3.2.7) \]

\[ g_{ds} = \text{Re}(Y_{22}) + \text{Re}(Y_{12}) \quad (3.2.8) \]

\[ D_I = 1 + (\omega C_{gs} R_1)^2 \quad (3.2.9) \]

and

\[ g_{fd} = -\text{Re}(Y_{12}) \quad (3.1.10) \]

\[ g_{fs} = \text{Re}(Y_{11}) - g_{fd} \quad (3.1.11) \]

g_{fd} and g_{fs} are approximated at low frequency, e.g. 50 MHz.

The way to define the error function can be crucial sometimes because it has an interactive relationship with the optimizing routine. A typical error function may affect the way the variables change to minimize the error from the error function. There are two types of errors that are most common: absolute error and relative error. In our study the relative
error function is chosen as:

\[ e(x) = \sum_{k=1}^{m} \sum_{i=1}^{n} \left| \frac{f_k(i) - f_{k,\text{ave}}}{f_{k,\text{ave}}} \right| \]  

(3.2.20)

The total error is the summation of the relative errors of all of the intrinsic parameters calculated at different frequencies. It intuitively shows how much error has been reduced after optimization. To avoid the spread of parameters and impractical values from the optimizing routine, parameter scaling and bounds are employed in the program. Utilization of different weights is another important factor similar to the normalization of parameter, and can be used to improve optimization for better fitting of the model. Weighing the parameters in error function is one of the ways we have used to improve the efficiency of optimization.

3.3 Examples and Results

S-parameters measured at zero-bias voltage and some other points for 1.0\(\mu\text{m}\)*200\(\mu\text{m}\) GaAs device, provided by Wright-Patterson AFB Electronic Technology Laboratory, have been used in the program to extract and optimize both intrinsic and extrinsic element values. The extracting algorithm has an emphasis on the flexibility of changing the weights of
variables, depending on the degree of significance in terms of how each variable would affect the error function. So often the relationship between the error function and the variables is not self-evident, neither is it explicitly expressed in equations before applying the optimization algorithm. That means the post-adjusting weights and the step size could be as important as the starting parameter values. Sometimes one intrinsic element or more may be reduced to zero and should be ignored by the program. But these decisions will not be made and put into practice until a few iterations have been executed. The subroutine will calculate explicit gradients of error function and show how variables vary to diminish the error. Based on the information from pre-processing, one could add weights to reduce error, or to produce reasonable element values within a range.

From the initial values of the intrinsic element values, versus frequency for this particular case, we observed that $C_{gs}$ and $G_m$ contribute more error and $R_i$ is unusually large before optimization. Therefore a factor of 2 has been added to $C_{gs}$ and $G_m$ respectively and the weight of error due to $R_i$ is reduced to 0.1. Since $R_{gd}$ is zero before optimization, it is weighted zero in the error function. A scaling factor of 5 has been added to the extrinsic inductors which would enhance the physical significance of the parameters of the model. A reasonable choice of initial conditions is also important in optimization to avert unrealistic results. Since the initial
set of parameters have an influence on the process of optimization and the final result, we also applied the tuning technique in using the program. One variable may be adjusted as others are kept constant. The final error resulting from this change can be used as feedback for further iterations. In this case we estimated starting values of the external elements based on the eqns(3.1.1-3.1.8)[1].

The final results of each intrinsic parameter versus frequency have been compared to its initial values. Fig. 3.3 depicts the initial and final intrinsic elements against the frequency for the 1.0μm*200μm GaAs device. We expect that the intrinsic elements be theoretically independent of frequency. Using this property we are able to obtain more accurate values of external and internal elements. The AC model parameters for the device are shown in Table 3.1. Note that the average of the intrinsic element values for the sweep of frequencies at this bias is adopted as the objective element value. The final average of all intrinsic elements is given in the table. The bias conditions for this device was $V_{ds}=2.5\text{V}$, $I_{ds}=33.09\text{mA}$, and the gate voltage, $V_{gs} = -1.314\text{V}$. We have also applied this program to model the same device at $V_{gs} = 0.5\text{V}$. The extrinsic parameters obtained at the first bias point ($V_{gs} = -1.314\text{V}$) were used as the initial values. The intrinsic parameters obtained were reasonably flat with frequency and optimization hardly reduced the error any further. This implies that using this method, the external elements can be extracted using
measured S-parameters at any reasonable bias point. The model parameter extracted at $V_{gs}=0.5\,\text{V}$ and $V_{ds}=3.0\,\text{V}$ is given in Table 3.2. LIBRA has been used to check the accuracy of the equivalent circuit and the extracted parameters. The S-parameter generated from LIBRA would be compared to that from measurements. The measured versus calculated S-parameter for $V_{gs}=-1.314\,\text{V}$ and $V_{gs}=0.5\,\text{V}$ will be presented in Smith charts and illustrated in Fig. 3.4-3.9.
Table 3.1 The extrinsic parameter and error values for 1.0*200um device (at $V_g=-1.314V$, $V_d=2.5v$)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Initial</th>
<th>Final (optimized)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_s$ (ohm)</td>
<td>2.50000</td>
<td>2.35000</td>
</tr>
<tr>
<td>$R_d$ (ohm)</td>
<td>1.76433</td>
<td>1.72264</td>
</tr>
<tr>
<td>$R_o$ (ohm)</td>
<td>4.10092</td>
<td>2.98850</td>
</tr>
<tr>
<td>$L_s$ (pF)</td>
<td>32.19800</td>
<td>72.56142</td>
</tr>
<tr>
<td>$L_d$ (pF)</td>
<td>16.09900</td>
<td>12.32269</td>
</tr>
<tr>
<td>$L_o$ (pF)</td>
<td>10.7330</td>
<td>122.15124</td>
</tr>
<tr>
<td>$C_m$ (fF)</td>
<td>45.58000</td>
<td>40.99659</td>
</tr>
<tr>
<td>$C_m$ (fF)</td>
<td>11.39500</td>
<td>2.12676</td>
</tr>
<tr>
<td>Error</td>
<td>12.4435%</td>
<td>2.1554%</td>
</tr>
</tbody>
</table>

Table 3.2 The intrinsic model parameter average values

<table>
<thead>
<tr>
<th>Parameter</th>
<th>$g_m$(mmho)</th>
<th>$G_m$(mmho)</th>
<th>$C_p$(pF)</th>
<th>$C_m$(fF)</th>
<th>$C_d$(fF)</th>
<th>Tau(ps)</th>
<th>$R_i$(ohm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>30.464</td>
<td>2.5846</td>
<td>0.2554</td>
<td>26.788</td>
<td>20.6608</td>
<td>3.269</td>
<td>20.764</td>
</tr>
<tr>
<td>Final</td>
<td>23.257</td>
<td>2.2991</td>
<td>0.2005</td>
<td>28.86</td>
<td>37.6058</td>
<td>3.27</td>
<td>8.7293</td>
</tr>
</tbody>
</table>

Table 3.3 The intrinsic model parameter average values (at $V_g=0.5v$, $V_d=3v$)
(The final extrinsic values obtained in Table 1.1 have been used)
(Initial error=4.89%, Final error=2.65%)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>$g_m$(mmho)</th>
<th>$G_m$(mmho)</th>
<th>$C_p$(pF)</th>
<th>$C_m$(fF)</th>
<th>$C_d$(fF)</th>
<th>Tau(ps)</th>
<th>$R_i$(ohm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial</td>
<td>33.0173</td>
<td>1.3053</td>
<td>0.4089</td>
<td>19.8442</td>
<td>36.752</td>
<td>3.8865</td>
<td>5.5925</td>
</tr>
<tr>
<td>Final</td>
<td>31.4672</td>
<td>1.4608</td>
<td>0.3831</td>
<td>18.3506</td>
<td>36.426</td>
<td>3.9123</td>
<td>6.86</td>
</tr>
</tbody>
</table>
Figure 3.1 Small-signal equivalent circuit of a field effect transistor.

Figure 3.2 Extended small-signal equivalent circuit of a field effect transistor with gate current and resistive feedback.
Figure 3.3 Intrinsic Elements Vs Frequency
Figure 3.3 (Continued)
Figure 3.3 (Continued)
Fig. 3.4  The measured $S_{11}$ and $S_{22}$ compared to that of AC model at bias: $V_{gs} = -1.314V$, $V_{ds} = 2.5V$. 
Fig. 3.5 The measured S12 compared to that of AC model at bias: $V_{gs}=-1.314\text{V}$, $V_{ds}=2.5\text{V}$. 
Fig. 3.6 The measured S21 compared to that of AC model at bias: $V_{gs} = -1.314V$, $V_{ds} = 2.5v$. 
Fig. 3.7 The measured S11, S22 compared to that of AC model at bias: $V_{gs}=0.5V$, $V_{ds}=3.0V$. 
Fig. 3.8 The measured S12 compared to that of AC model at bias: $V_{gs}=0.5\,V$, $V_{ds}=3.0\,V$. 
Fig. 3.9 The measured $S_{21}$ compared to that of AC model at bias: $V_{gs}=0.5V$, $V_{ds}=3.0V$. 
Chapter 4
Analyzing MESFET and HEMT Using the Complete Nonlinear Model in Simulation Packages

Chapter 4 is intended to combine the AC and DC modeling as a complete model for either linear or nonlinear circuit analysis and simulation when utilizing commercial time-domain and frequency-domain simulation programs. Some aspects of the AC or DC part of the model have been justified with the optimization programs DCMOD and ACMOD in the previous two chapters. The robustness of the algorithm, the reliability and efficiency of the optimization program, and the accuracy of the extracted GaAs FET model discussed before will be tested again. This time they will be tested by comparing the S-parameters generated based on the combined nonlinear model and the measurements on a 1*200 µm GaAs field-effect transistor. The algorithm will be extended to accommodate the model extraction for HEMT. The developmental transistor measurements are provided by Wright-Patterson AFB Electronic Technology Laboratory. Small or large signal analysis of field-effect transistors is crucial for circuit design in MMIC and thus will be emphasized in this chapter. Given simulators, one should be able to apply the appropriate model available to analyze, simulate and optimize new designs. Without any doubt, applications of the extracted GaAs FET models in circuit simulation and the accuracy of the model is our ultimate concern.
4.1 Nonlinear Device Models in SPICE

The combination of DC and AC model parameters leads to the SPICE nonlinear model. The SPICE model was originally defined in the SPICE program, which includes all sorts of different models to be used to describe nonlinear electrical properties of electronic elements, such as: capacitors, bipolar junction transistors, field-effect transistors, etc. The nonlinear SPICE model for GaAs MESFET is needed as well in PSPICE, MWSPICE, and LIBRA to accurately simulate the device and circuit behavior. Since no single empirically derived model of GaAs MESFET could fully cover all features of the device to date, there are several different models available for users within SPICE for modeling GaAs MESFET. For instance in PSPICE, there are three models available, represented by levels 1, 2 and 3. Some simulators even provide a user-defined model if needed.

The nonlinear GaAs FET model in time-domain simulation programs like PSPICE, or in CAD tools using an harmonic balance technique like LIBRA, is the same. The GaAs FET is modeled as an intrinsic FET, and represented by a list of parameters:

Vto: pinch-off voltage (volt)
Alpha: saturation voltage parameter (volt\(^{-1}\))
Beta: transconductance coefficient (amp/volt\(^2\))
b: doping tail extending parameter (volt\(^{-1}\))
Lambda: channel-length modulation (volt$^{-1}$)

Gamma: static feedback parameter

Delta: output feedback parameter (amp*volt)$^{-1}$

Q: power-law parameter

Tau: conduction current delay time (sec)

Rg: gate ohmic resistance (ohm)

Rd: drain ohmic resistance (ohm)

Rs: source ohmic resistance (ohm)

Is: gate p-n saturation current (amp)

N: gate p-n emission coefficient

M: gate p-n grading coefficient

Vbi: gate p-n potential (volt)

Cgd: zero-bias gate-drain capacitance (farad)

Cgs: zero-bias gate-source p-n capacitance (farad)

Cds: drain-source capacitance (farad)

Fc: forward-bias depletion capacitance coefficient

Vdelta: capacitance transition voltage (volt)

Vmax: capacitance limiting voltage (volt)

Eg: bandgap voltage (barrier height) (eV)

Xti: IS temperature exponent

Vtotc: Vto temperature coefficient (volt/°C)

Betatce: beta exponential temperature coefficient (%/°C)

Trgl: Rg temperature coefficient (°C$^{-1}$)

Trdl: Rd temperature coefficient (°C$^{-1}$)

Trsl: Rs temperature coefficient (°C$^{-1}$)

Kf: flicker noise coefficient
Af: flicker noise exponent

The parameters above may be used in the Curtice quadratic, Raytheon or Triquint models.

Usually for small-signal analysis such as S-parameter computation, the model can be linearized at a fixed bias. The small-signal equivalent circuit in the SPICE model is shown in Fig. 4.1.1 The conductances $G_{gs}$ and $G_{gd}$ can be obtained from

\[ G_{gs} = \text{Area} \cdot \frac{\partial I_{gs}}{\partial V_{gs}} \quad (4.1.1) \]

\[ G_{gd} = \text{Area} \cdot \frac{\partial I_{gd}}{\partial V_{gd}} \quad (4.1.2) \]

The transconductance at this bias may be calculated from [35]

\[ g_m' = g_m e^{-j\omega \tau_m} \quad (4.1.3) \]

and $g_m$ is obtained from

\[ g_m = \text{Area} \cdot \frac{\partial I_{\text{drain}}}{\partial V_{gs}} \quad (4.1.4) \]

With this information and known circuit topology, node equations can be written and Y or S-parameters can be obtained for the circuit at this bias.
For large-signal analysis nonlinearity of GaAs FET results from the nonlinear relationship of drain current and gate voltage and voltage-controlling capacitors. The mathematical model for $I_{ds}$ should be able to illustrate this nature. Eqn. (2.2.1) shows $I_{ds}$ as a nonlinear function of $V_{gs}$. Since intrinsic capacitances are dependent on bias voltages, and because of transistor gate geometry and the uneven spread of drain-source and gate voltages, the capacitor model is the function of $V_{gs}$ and $V_{ds}$ as well. In chapter 3, assuming intrinsic capacitances are not changed significantly, we considered the equivalent circuit as a linear circuit (suitable for small signal analysis), because the input signal is small and the bias is fixed. In this chapter, we will extend our study to the nonlinear modeling of GaAs field-effect transistors and high electron mobility transistors (HEMT) when a large signal is used as input, or bias conditions vary a considerable amount. Bias dependence of the equivalent circuit will also be investigated. Nonlinear model of MESFET is crucial for simulations of power amplifiers.

As an extension and conclusion of the last two chapters, both DC and AC parameters used in a model block for the SPICE model are extracted and optimized by separate Fortran programs. The Fletcher-Powell numerical method is applied to optimize model parameters and to match calculated characteristics to the measured data. The AC parameters on which we are working include $C_{gs}$ and $C_{gd}$ at zero-bias point,
$C_{ds}$, and $T_{au}$; whereas, DC parameters include alpha, beta, lambda, and $b$ for the Raytheon model. $V_{to}$ is known and can be obtained from the measured drain current of the device. Same AC parameters will be used for the Triquint model because of the similar internal capacitance model, but DC fitting parameters will be different. The first device under investigation is the $1*200 \mu M$ GaAs field-effect transistor.

4.2 Application of the Raytheon Model in MWSPICE

The Raytheon model has existed in most time-domain simulation programs since 1986. It is fairly easy to use, convenient to access, and more importantly, it is more accurate than other models currently in use. The SPICE model for GaAs FET in MWSPICE is identical to that in PSPICE[22]. The Raytheon model is labeled as MODEL 2.5 in the circuit file in MWSPICE. In general it describes variations of the drain current $I_{ds}$ similarly to the Curtice model, but with two prominent improvements. One is an enhanced drain current formulation which includes a feedback factor $b$ to improve the relationship between $I_{ds}$ and $V_{gs}$. The other is a new capacitance model which has symmetrical patterns that make it work for both normal and inverse modes of operation of the transistor. Theoretically, capacitance near the drain end should increases abruptly at the point of saturation of the electron velocity, because the voltage potential is much
higher at the drain than at the source. In practice, velocity saturation occurs more gradually, so there is no sudden change in the intrinsic capacitors. The Raytheon capacitance model calculates channel charge and then capacitance, followed by splitting it into gate-source and gate-drain capacitance values[24].

For AC simulation in the SPICE model we chose $R_g$, $R_d$, $R_s$, $C_{gd}$, $C_{gs}$, and $T_{au}$ for optimization. $C_{gd}$ and $C_{gs}$ are at zero-bias capacitance when used in the SPICE model. One could obtain $C_{gd}$ and $C_{gs}$ by applying the program ACMOD with measured S-parameters at bias conditions other than zero bias point. Then the equations below could be used to get $C_{gso}$ and $C_{gdo}$ for the SPICE model.

$$C_{gso} = \frac{(C_{gs} - K3(C_{gs} + C_{gd})(K2 + K3))(1 - (V_u/V_{bd}))^{1/2}}{(area \cdot K1K2 - (K1K2 + K1K3))} \tag{4.2.1}$$

$$C_{gdo} = \frac{(C_{gs} K1 - (K1K3(C_{gs} + C_{gd})(K2 + K3)))}{(area \cdot K1K2 - (K1K2 + K1K3))} \tag{4.2.2}$$

where $K1$, $K2$, $K3$, $V_u$, and $V_n$ are computed using following equations

$$K1 = (1 + (V_n - V_{bp})/((V_n - V_{bp})^2 + \delta^2)^{1/2})/2 \tag{4.2.3}$$
\[ K_2 = (1 + (V_{gs} - V_{gd})/((V_{gs} - V_{gd})^2 + (1/\alpha)^2)^{1/2})/2 \]  
(4.2.4)

\[ K_3 = (1 - (V_{gs} - V_{gd})/((V_{gs} - V_{gd})^2 + (1/\alpha)^2)^{1/2})/2 \]  
(4.2.5)

\[ V_e = (V_{gs} + V_{gd} + ((V_{gs} - V_{gd})^2 + (1/\alpha)^2)^{1/2})/2 \]  
(4.2.6)

\[ V_n = \begin{cases} X, & X < V_{\text{max}} \\ V_{\text{max}}, & \text{otherwise} \end{cases} \]  
(4.2.7)

where

\[ X = (V_e + V_{to} + ((V_e - V_{to})^2 + \text{delta}^2)^{1/2})/2 \]  
(4.2.8)

In MWSPICE \( V_{\text{max}} = 0.5 \), delta=0.2. Area is the relative device area and defaults to 1. An alternative is to extract \( C_{gs0} \) and \( C_{gd0} \) directly by using ACMOD with S-parameters measured at zero-bias condition as an input file. The output from the program would be the optimized gate-source and gate-drain capacitances. The latter has been chosen in our study. Table 4.1 shows the initial and final AC small signal equivalent circuit parameter values.
R_g, R_d, and R_s are external parasitic resistances. As mentioned before, they are variables in ACMOD. These resistances may impose considerable impact on input gate-sources and gate-drain voltages and cause an erroneous outcome when using DCMOD, especially when the drain-source current is at high levels. Thus, internal V_{gs} and V_{ds} should be calculated when using DCMOD. Provided that gate current I_g is small and can be neglected (as it is in normal operation), the voltage across the gate resistance may be dropped. Similar to the procedures discussed in chapter 2, the Raytheon mathematical model for drain current in DCMOD would calculate I_{ds} and then evaluate error function. Four DC fitting parameters in the SPICE model need to be optimized. These are alpha, beta, lambda and b. Starting with the estimated initial values, we have optimized the parameters to reduce error between the calculated and the measured I-V characteristics. Table 4.2 shows the initial and final parameter values using the Raytheon model for the 1*200 μM device in this study.

4.3 Application of the Triquint Model in PSPICE

Triquint is considered to be a better model, not only for a good match of I-V characteristics, but also for small-signal analysis, especially for S-parameters analysis of circuits. It offers more accurate prediction of AC characteristics of GaAs FET over a wider bias range[7]. We have used the Triquint
model to improve S11 and S22 behavior of circuits with two ports. There is a limitation in using this model because currently no Triquint model has been installed in MWSPICE. Therefore the one in PSPICE has been chosen to generate S-parameters. The Triquint model newly added to PSPICE has better tracking of drain resistance $R_{ds}$ over the Raytheon model and provides a better model for AC characteristics, such as S-parameters[7]. A comparison has been made on S-parameters generated by Triquint and that from the Raytheon model. Triquint will show its due advantage as we move on.

We use the same equivalent circuit to isolate the intrinsic circuit and to extract all model parameters (see Fig.3.2). The same gate-source capacitor, $C_{gs}$, and gate-drain capacitor, $C_{ds}$, at zero bias, as well as the parasitic resistances $R_g$, $R_d$, $R_s$, and time delay $T_{au}$ are used for Raytheon as they are for Triquint.

Since the Triquint model has a different way of calculating drain current from Raytheon, it includes somewhat different sets of DC parameters. The fitting parameters are alpha, beta, gamma, delta and Q. Gamma is used to make $V_t$ as a function of $V_{ds}$, which in turn would improve drain conductance near the cut-off region. The adjustment of gamma, on the other hand, could make up the discrepancy of the drain resistance $R_{ds}$ between low frequency and high frequency. In our study, we may also include $V_{to}$ as a parameter in our program in order to achieve a better fit of calculated $I_{ds}$ and measured
I_{ds}. But the procedure to optimizing them remains the same. The Triquint model uses the same gate-diode and capacitance model as Raytheon. Table 4.3 lists starting and final optimized values for the DC model in the model block. To avoid local minimum in applying the optimizing program DCMOD, initial values should be reasonably close to the optimized in order for the optimization procedure to converge to an acceptable final model.

Based on the output of DC and AC programs, table 4.4 and 4.5 presents the complete sets of parameter values we have obtained for the Raytheon and Triquint models.

4.4 Tuning

Tuning of the intrinsic elements in the SPICE model block is another technique to achieve the desirable result. Analytically, S-parameters of the GaAs FET may be expressed as follows:

\[
S_{11} = \frac{(1-y_{11})(1+y_{22})+y_{12}y_{21}}{\Delta_2}
\]  
(4.4.1)

\[
S_{21} = \frac{-2y_{21}}{\Delta_2}
\]  
(4.4.2)
where $y_{ij}'$ are $Y$-parameters of the 2-port equivalent circuit of GaAs FET. $Y_{ij}' = Y_{ij}Z_0$, $Z_0$ is the characteristic impedance, and

$$\Delta_2 = (1 + y_{11}') (1 + y_{22}') - y_{12}' y_{21}'$$  \hspace{1cm} (4.4.5)$$

To simplify the relationship of $S$-parameters with intrinsic elements, for instance, for $S_{22}$, we assume $y_{12}'$ is small and negligible. Thus, $S_{22}$ can be approximated as below:

$$S_{22} = \frac{(R_{ds} - 1) + j\omega R_{ds} (C_{ds} + C_{gd})}{(R_{ds} + 1) + j\omega R_{ds} (C_{ds} + C_{gd})}$$  \hspace{1cm} (4.4.6)$$

Obviously, $S_{22}$ depends on the values of $R_{ds}$, $C_{ds}$ and $C_{gd}$. While DC parameters were kept constant, AC parameters were tuned to adjust the $S$-parameters so as to achieve the best fit of calculated and measured values. $C_{gs}$ and $C_{gd}$ are contributing quite a bit to tuning in approaching a better match of the
calculated and measured S-parameters.

4.5 Use of Microwave Simulation Program (LIBRA)

LIBRA is a CAD program designed to analyze and simulate microwave devices[23]. It provides harmonic balance measurements, which include power, voltage and current in both time and frequency domains. We used LIBRA to generate S-parameters for AC equivalent circuits and to compare measured and modeled S-parameters on Smith charts. Fig. 4.4.1-4.4.3 presents the comparisons of measured S-parameters and computed S-parameters using the Raytheon model in MWSPICE. Fig.4.4.4-4.4.6 compares the measured S-parameters to the computed ones by using the Triquint model in PSPICE. From the curves on the Smith charts, we observe that the Raytheon model provides a good match for $S_{11}$ and $S_{21}$, but not for $S_{22}$ and $S_{12}$. Fig. 4.4.4-4.4.6 indicates that Triquint would give a better model in calculating $S_{22}$ and $S_{12}$ without sacrificing much accuracy on $S_{11}$ and $S_{21}$. This means that, as a whole, the Triquint model is more suitable for simulating GaAs MESFET devices and circuits in compromising the accuracy on ($S_{11}, S_{21}$) versus ($S_{22}, S_{12}$). Note that the device models presented here can be used to extract S-parameters at any bias point in the normal operating region.

Harmonic balance analysis in LIBRA and transient analysis in MWSPICE and PSPICE need consistent and non-linear models
for each device for the simulation to be reliable. Both models discussed here satisfy this requirements, yet the Triquint model seems slightly more accurate.

4.6 Modeling of HEMT Devices

4.6.1 Introduction

The high electron mobility transistor (HEMT) is another device used to take advantage of the superior transport properties of electrons, as its name implies. HEMT is a heterostructure field effect transistor. In contrast to MESFET, HEMT improves significantly in noise figure and can work at higher frequencies, that is associated with more complex structure, difficult fabrication, extra cost and low yields[31]. In order to understand HEMT, we would compare the similarities and differences in structure and materials between MESFET and HEMT. Shown in Fig. 4.6.1 is the cross-sectional view of a conventional HEMT structure. Compared to Fig. 4.6.2, the cross-section of MESFET, we can see immediately some obvious changes in conduction channel HEMT. Both AlGaAs and GaAs are used in HEMTs. One could find that there is an obvious difference between HEMT and MESFET in terms of semiconductor layers. In HEMT, the thickness of n-type AlGaAs and the undoped AlGaAs spacer layer are critical in determining device behavior. The most distinctive part of
HEMT occurs at the AlGaAs/GaAs boundary. A high carrier concentration resides in this narrow area along the side of the heterojunction. The high free-electron concentration over the area is termed as a two-dimensional electron gas (2-DEG). Since the GaAs is undoped, electrons traveling in this region do not collide with ionized donors. Due to the fact that electron mobility is highest for lightly doped material, the structure and transport properties in the 2-DEG are favorable for best response time and high-frequency operation. Geometric dimensions of HEMT are crucial factors in determining electrical properties of HEMT. The layout of dimensions of MESFET can be applied to HEMT. As is shown in Fig.4.6.3, these dimensions are $L$, $Z$, $L_{gs}$, $L_{gd}$, $L_{g}$, and $L_{d}$, illustrated in the diagram. Another important dimension characterizing the HEMT physical structure is the gate length, which is $L$ in Fig.4.6.3. This dimension is a major factor of governing maximum frequency limits for HEMT.

As its counterparts of MESFET, HEMT has ohmic contacts at the source and drain. The gate to semiconductor material is a Schottky barrier.

4.6.2 Parameter Extraction for 0.25μ*2*50μ HEMT

From the measurements on the 0.25μ*2*50μ device supplied by Wright-Patterson AFB, we extracted the model parameters using the procedures discussed in previous chapters. Since
there is no complete model especially designed for HEMT and installed in popular simulators such as PSPICE, we can take advantage of the current MESFET model available in the software to simulate the HEMT device for the purpose of analysis and simulations if necessary. As we have examined the simplified physical structures of HEMT and MESFET, we have noticed that they have many aspects in common, which may be easily justified by the MESFET empirical model. Meanwhile, a great deal of effort has been done to make HEMT a faster, more effective device than MESFET, especially with new structures in the conduction channel. To extract model parameters, more attention will be drawn onto those exceptions of HEMT when we are applying our algorithm. The AC equivalent circuit for MESFET, shown in Fig.4.6.3, seems adequate in describing HEMT. Therefore, program ACMOD is used to extract and optimize the AC model parameters for HEMT. Extrinsic elements are optimized to make intrinsic elements independent of changes of frequency, a physical property also shared by HEMT. Table 4.6.1 & 2 presents the initial and final element values of the equivalent circuit of HEMT. The new device shows a smaller time delay than MESFET. As expected, the drain conductance is higher than that of MESFET. So is beta, the transconductance factor of HEMT. Fig. 4.6.4 illustrates the I-V characteristics of the 0.25μ*2*50μ device. Observing the curves, one can see that the drain current is evidently increased for HEMT. The device shows an increase in conduction even after saturation
of current, which is different from MESFET. Beyond $V_{ds}=3.0\text{V}$, it is the break-down region. For DC fitting parameters, the program DCMOD is applied for matching the models to the physical device. We found that beta and alpha are much larger than those of MESFET, as shown in Table 4.6.3, which presents the initial and final parameter values optimized with the Raytheon model in DCMOD. Table 4.6.4 presents the initial and final parameter values for HEMT using Triquint in DCMOD.

From Fig. 4.6.5 & 6, one can see that either the Triquint or the Raytheon model could predict and match fairly well the HEMT device at a low current level and a low gate bias, such as $V_{gs}=-0.5\text{V}$. The Raytheon model is more favorably able to predict the increase in drain conductance after the saturation of the drift velocity of electrons in HEMT, but still not good enough at greater $V_{gs}$ and $I_{ds}$. Triquint retains Raytheon's good model at a low current level, while using a positive delta to depress the saturation current, which might be suitable for MESFET, but not for HEMT in this case. To be more useful and practical, we would use the SPICE model in PSPICE to explore AC characteristics of the research HEMT device. The Raytheon and Triquint models for MESFETs in PSPICE may be used to generate small-signal S-parameters for the $0.25\mu*2*50\mu$ device. Table 4.6.5 & 6 lists the model parameters needed in MODEL block in PSPICE, which we had obtained earlier. The results of modeled S-parameters while applying the Raytheon model are compared to the measured S-parameters on Smith charts in
Although MESFET models could somehow be applied to simulate and model HEMTs for small-signal analysis, they fail to predict the HEMT's unique electrical properties when a large signal is necessary. Since little work has been done to produce a large-signal HEMT model that could be incorporated into large-signal circuit simulator routines, an existing MESFET model may be modified to meet the needs. Among characteristics of HEMT, transconductance is considered most critical to the accurate prediction of many important large-signal effects[31]. As shown and discussed previously, transconductance of HEMT in the linear region (controlled by alpha) or in the saturation region (controlled by lambda for Raytheon or delta for Triquint) is much larger than its counterparts of MESFET, that makes HEMT distinct from MESFET and gives the edge over MESFETs. Transconductance is a nonlinear function of $V_{gs}$ as well, and peaks at a certain value of $V_{gs}$. Besides the transconductance, another nonlinear element, gate capacitance (charge) is different from MESFET, a detail that can be found in chapter 1 of *Microwave MESFETs and HEMTs* by J. M. Golio. For more accuracy the MESFET model in LIBRA/PSPICE needs to be modified. A starting point may be the justifications presented in chapter 2 of the same book by J. M. Golio.
Table 4.1: Initial parameters' values and their optimized values for a small signal equivalent circuit at $V_{gs}=0.0V$, $V_{ds}=3.0V$

<table>
<thead>
<tr>
<th>Initial values</th>
<th>Final values</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_g=32.2 \text{ PH}$</td>
<td>$L_g=77.19 \text{ PH}$</td>
</tr>
<tr>
<td>$L_d=16.1 \text{ PH}$</td>
<td>$L_d=16.05 \text{ PH}$</td>
</tr>
<tr>
<td>$L_s=10.73 \text{ PH}$</td>
<td>$L_s=141.4 \text{ PH}$</td>
</tr>
<tr>
<td>$R_g=2.5 \text{ ohm}$</td>
<td>$R_g=2.5 \text{ ohm}$</td>
</tr>
<tr>
<td>$R_d=1.764 \text{ ohm}$</td>
<td>$R_d=1.764 \text{ ohm}$</td>
</tr>
<tr>
<td>$R_s=4.1 \text{ ohm}$</td>
<td>$R_s=4.1 \text{ ohm}$</td>
</tr>
<tr>
<td>$C_{pg}=45.58 \text{ FF}$</td>
<td>$C_{pg}=48.22 \text{ FF}$</td>
</tr>
<tr>
<td>$C_{pd}=11.4 \text{ FF}$</td>
<td>$C_{pd}=11.3 \text{ FF}$</td>
</tr>
<tr>
<td>$C_{gd0}=17.87 \text{ FF}$</td>
<td>$C_{gd0}=19.0 \text{ FF}$</td>
</tr>
<tr>
<td>$C_{gs0}=0.587 \text{ PF}$</td>
<td>$C_{gs0}=0.33 \text{ PF}$</td>
</tr>
<tr>
<td>$\tau=3.69 \text{ PS}$</td>
<td>$\tau=3.61 \text{ PS}$</td>
</tr>
<tr>
<td>$C_{ds}=20.3 \text{ FF}$</td>
<td>$C_{ds}=30.43 \text{ FF}$</td>
</tr>
<tr>
<td>Error=25.26%</td>
<td>Error=2.405%</td>
</tr>
</tbody>
</table>

Table 4.2: Initial DC parameters and their optimized values for the Raytheon model

<table>
<thead>
<tr>
<th>Initial Values</th>
<th>Final Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha=2$</td>
<td>$\alpha=1.4$</td>
</tr>
<tr>
<td>$\beta=0.003$</td>
<td>$\beta=0.0079$</td>
</tr>
<tr>
<td>$\lambda=0.2$</td>
<td>$\lambda=0.0363$</td>
</tr>
<tr>
<td>$B=0.15$</td>
<td>$B=0.2045$</td>
</tr>
<tr>
<td>$e_0=5.44%$</td>
<td>$ef=4.2%$</td>
</tr>
</tbody>
</table>

Table 4.3: Initial DC parameters and their optimized values for the Triquint model

<table>
<thead>
<tr>
<th>Initial Values</th>
<th>Final Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha=2$</td>
<td>$\alpha=1.691$</td>
</tr>
<tr>
<td>$\beta=0.003$</td>
<td>$\beta=0.0058$</td>
</tr>
<tr>
<td>$\gamma=0.04$</td>
<td>$\gamma=0.0407$</td>
</tr>
<tr>
<td>$\delta=1$</td>
<td>$\delta=-0.0019$</td>
</tr>
<tr>
<td>$Q=2$</td>
<td>$Q=1.811$</td>
</tr>
<tr>
<td>$V_{to}=-4$</td>
<td>$V_{to}=-3.9737$</td>
</tr>
<tr>
<td>Ini.err=7.68%</td>
<td>Final err=4.33%</td>
</tr>
</tbody>
</table>
Table 4.4: Parameter values for the Raytheon model in MWSPICE

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>MODEL</td>
<td>2.5</td>
</tr>
<tr>
<td>VTO</td>
<td>-4V</td>
</tr>
<tr>
<td>BETA</td>
<td>0.0079</td>
</tr>
<tr>
<td>LAMBDA</td>
<td>0.0362</td>
</tr>
<tr>
<td>ALPHA</td>
<td>1.39</td>
</tr>
<tr>
<td>TAU</td>
<td>3.61ps</td>
</tr>
<tr>
<td>RS</td>
<td>4.1</td>
</tr>
<tr>
<td>RD</td>
<td>1.764</td>
</tr>
<tr>
<td>CGS</td>
<td>0.33PF</td>
</tr>
<tr>
<td>CGD</td>
<td>0.019PF</td>
</tr>
<tr>
<td>VBI</td>
<td>0.8</td>
</tr>
<tr>
<td>RG</td>
<td>2.5</td>
</tr>
<tr>
<td>B</td>
<td>0.2038</td>
</tr>
<tr>
<td>CDS</td>
<td>0.03PF</td>
</tr>
</tbody>
</table>

Table 4.5: Parameter values for the Triquint model in PSPICE

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>LEVEL</td>
<td>3</td>
</tr>
<tr>
<td>VTO</td>
<td>-3.974</td>
</tr>
<tr>
<td>BETA</td>
<td>0.0058</td>
</tr>
<tr>
<td>ALPHA</td>
<td>1.691</td>
</tr>
<tr>
<td>GAMMA</td>
<td>0.0407</td>
</tr>
<tr>
<td>TAU</td>
<td>3.61ps</td>
</tr>
<tr>
<td>RS</td>
<td>4.1</td>
</tr>
<tr>
<td>RD</td>
<td>1.764</td>
</tr>
<tr>
<td>CGS</td>
<td>0.33PF</td>
</tr>
<tr>
<td>CGD</td>
<td>0.019PF</td>
</tr>
<tr>
<td>DELTA</td>
<td>-0.0019</td>
</tr>
<tr>
<td>Q</td>
<td>1.811</td>
</tr>
<tr>
<td>VBI</td>
<td>0.8</td>
</tr>
<tr>
<td>RG</td>
<td>2.5</td>
</tr>
<tr>
<td>CDS</td>
<td>0.03PF</td>
</tr>
</tbody>
</table>
Table 4.6.1 The initial and final extrinsic element values for 0.25μ*2*50μ HEMT.

<table>
<thead>
<tr>
<th>Element</th>
<th>Initial</th>
<th>Final (optimized)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_q$(ohm)</td>
<td>2.5</td>
<td>0.837</td>
</tr>
<tr>
<td>$R_d$(ohm)</td>
<td>1.764</td>
<td>1.097</td>
</tr>
<tr>
<td>$R_s$(ohm)</td>
<td>4.1</td>
<td>2.968</td>
</tr>
<tr>
<td>$L_q$(pH)</td>
<td>32.198</td>
<td>50.09</td>
</tr>
<tr>
<td>$L_d$(pH)</td>
<td>16.099</td>
<td>11.47</td>
</tr>
<tr>
<td>$L_s$(pH)</td>
<td>10.733</td>
<td>10.20</td>
</tr>
<tr>
<td>$C_{ps}$(fF)</td>
<td>45.58</td>
<td>9.94</td>
</tr>
<tr>
<td>$C_{pd}$(fF)</td>
<td>11.395</td>
<td>13.0</td>
</tr>
<tr>
<td>Error</td>
<td>4.581%</td>
<td>2.775%</td>
</tr>
</tbody>
</table>

Table 4.6.2 The initial and final intrinsic element values for 0.25μ*2*50μ HEMT at $V_{gs}$=-0.5V, $V_{ds}$=1.5V.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Initial</th>
<th>Final (optimized)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$g_m$(mmho)</td>
<td>20.429</td>
<td>19.575</td>
</tr>
<tr>
<td>$g_m$(mmho)</td>
<td>3.0289</td>
<td>2.979</td>
</tr>
<tr>
<td>$C_{gs}$(fF)</td>
<td>51.2</td>
<td>86.66</td>
</tr>
<tr>
<td>$C_{gd}$(fF)</td>
<td>13.218</td>
<td>13.0</td>
</tr>
<tr>
<td>$C_{ds}$(fF)</td>
<td>30.274</td>
<td>28.218</td>
</tr>
<tr>
<td>Tau(ps)</td>
<td>1.074</td>
<td>1.65</td>
</tr>
<tr>
<td>$R_d$(ohm)</td>
<td>13.7067</td>
<td>2.29</td>
</tr>
</tbody>
</table>

Table 4.6.3 Initial and final optimized fitting parameter values for the Raytheon model

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Initial</th>
<th>Final</th>
</tr>
</thead>
<tbody>
<tr>
<td>alpha</td>
<td>2.9</td>
<td>4.5278</td>
</tr>
<tr>
<td>beta</td>
<td>0.035</td>
<td>0.0167</td>
</tr>
<tr>
<td>lambda</td>
<td>0.35</td>
<td>0.3594</td>
</tr>
<tr>
<td>$b$</td>
<td>0.8</td>
<td>0.8343</td>
</tr>
<tr>
<td>Vto</td>
<td>-1.0</td>
<td>-1.1955</td>
</tr>
<tr>
<td>Error</td>
<td>11.14%</td>
<td>8.018%</td>
</tr>
</tbody>
</table>
Table 4.6.4 Initial and final optimized fitting parameter values for the Triquint model

<table>
<thead>
<tr>
<th></th>
<th>Initial</th>
<th>Final</th>
</tr>
</thead>
<tbody>
<tr>
<td>alpha</td>
<td>2.5</td>
<td>3.0087</td>
</tr>
<tr>
<td>beta</td>
<td>0.035</td>
<td>0.0211</td>
</tr>
<tr>
<td>gamma</td>
<td>0.06</td>
<td>0.0601</td>
</tr>
<tr>
<td>delta</td>
<td>0.05</td>
<td>0.03</td>
</tr>
<tr>
<td>Q</td>
<td>2.0</td>
<td>0.7593</td>
</tr>
<tr>
<td>Vto</td>
<td>-1.0</td>
<td>-0.682</td>
</tr>
<tr>
<td>Error</td>
<td>22.14%</td>
<td>8.17%</td>
</tr>
</tbody>
</table>

Table 4.6.5 Model parameters for Raytheon in PSPICE

<table>
<thead>
<tr>
<th>LEVEL=2</th>
<th>Vto=-1.1955</th>
<th>BETA=0.0167</th>
</tr>
</thead>
<tbody>
<tr>
<td>LAMBDA=0.3594</td>
<td>ALPHA=4.5278</td>
<td>TAU=1.65 PS</td>
</tr>
<tr>
<td>RS=2.968</td>
<td>RD=1.097</td>
<td>CGS=86.66fF</td>
</tr>
<tr>
<td>CGD=13.0fF</td>
<td>VBI=0.8</td>
<td>RG=0.837</td>
</tr>
<tr>
<td></td>
<td></td>
<td>B=0.8343</td>
</tr>
</tbody>
</table>

Table 4.6.6 Model parameter for Triquint in PSPICE

<table>
<thead>
<tr>
<th>LEVEL=3</th>
<th>Vto=-0.682</th>
<th>BETA=0.0211</th>
<th>ALPHA=3.0087</th>
</tr>
</thead>
<tbody>
<tr>
<td>GAMMA=0.0601</td>
<td>TAU=1.65 PS</td>
<td>R S = 2 9 6 8</td>
<td></td>
</tr>
<tr>
<td>RD=1.097</td>
<td>CGS=86.66fF</td>
<td>CGD=13.0fF</td>
<td>DELTA=0.03</td>
</tr>
<tr>
<td>VBI=0.8</td>
<td>RG=0.837</td>
<td>CDS=28.218ff</td>
<td></td>
</tr>
</tbody>
</table>
Fig. 4.1.1 Small-signal equivalent circuit of GaAs MESFET.
Fig. 4.6.1 Cross-section view of a conventional HEMT structure.

Fig. 4.6.2 Cross-section view of a MESFET.
Fig. 4.6.3 Dimensions of a MESFET.

Fig. 4.6.4 Equivalent circuit for HEMT.
I-V Characteristics (with break-down)

Fig. 4.6.5 I-V characteristics of 0.25μ*2*50μ HEMT device.

Fig. 4.6.6 Calculated and measured I-V curves for 0.25μ*2*50μ device.
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Fig. 4.4.1 Comparison of measured and calculated $S_{11}$, $S_{22}$ using the Raytheon model in LIBRA for a 1*200uM device at zero-bias, $V_{ds}=3.0V$. 
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Fig. 4.4.3 Comparison of measured and calculated $S_{21}$ using the Raytheon model in LIBRA for a $1*200\mu M$ GaAs MESFET device at zero-bias, $V_{ds}=3.0V$. 
Fig. 4.4.4 Comparison of measured and calculated S11, S22 using the Triquint model in PSPICE for a 1*200μM GaAs MESFET device at zero-bias, $V_{ds}=3.0V$. 
Fig. 4.4.5 Comparison of measured and calculated $S_{12}$ using the Triquint model in PSPICE for a 1*200$\mu$M GaAs MESFET device at zero-bias, $V_{ds}=3.0V$. 
Fig. 4.4.6 Comparison of measured and calculated S21 using the Triquint model in PSPICE for a 1*200μM GaAs MESFET device at zero-bias, $V_{ds}=3.0V$. 
Fig. 4.6.6 Comparison of measured and calculated \$S_{11}, S_{22}\$ using the Raytheon model in PSPICE for a \$0.25\mu m \times 2 \times 50\mu m\$ HEMT device at bias: \$V_{gs}=-0.5V, V_{ds}=1.5V\$. 
Fig. 4.6.7 Comparison of measured and calculated S12 using the Triquint model in PSPICE for a 0.25μ*2*50μ HEMT device at bias: \( V_{gs} = -0.5V, V_{ds} = 1.5V \).
Fig. 4.6.8 Comparison of measured and calculated S21 using the Triquint model in PSPICE for a 0.25μ*2*50μ HEMT device at bias: \( V_{gs} = -0.5V \), \( V_{ds} = 1.5V \).
Chapter 5

Circuit Simulations and Applications Examples

Chapter 5 is intended to explore and illustrate applications of GaAs MESFET technology in circuit designs, such as high speed building blocks in microwave or optical communication systems. For instance, an operational tranconductance amplifier (OTA) is a valuable building block in implementing linear microwave and linearly derived circuits[18,19,20]. Its versatile applications extend to voltage-controlled amplifiers, filters and impedances[34]. In this chapter an OTA utilizing commercially available GaAs MESFET devices is studied and simulated using PSPICE, LIBRA, and MWSPICE. The OTA to be used as an invaluable high frequency integrator and band-pass filter has been developed and engineered to meet certain requirements. To simulate new circuit designs, batch files, and schematic and layout files have been created and analyzed by PSPICE and LIBRA respectively. The layout circuit files basically contain the physical information of the constructing materials, which is one step closer than the schematic input file to testing an actually fabricated chip. The AC analysis of the circuits, including checking the output voltage versus the frequencies for an integrator, and the output power of a band-pass filter versus the frequencies, has been carried out in the simulations. The result is discussed in detail in the
following sections. At the final stage of this design work we have used ACADEMY, a program released by EESOF, which provides an integrated schematic capture, layout editing, and simulations for microwave design work in order to realize the layout of the OTA.

5.1 Operational Transconductance Amplifiers

A high speed analog circuit demands high frequency response transistors; therefore, GaAs MESFET technology has been developed to meet the rapidly expanding market for faster devices. Transconductance amplifiers, which take advantage of BJT and MESFET technology, are being explored and developed in an attempt to do what the traditional operational amplifier could not normally do in high frequency applications, or in improving performance. In contrast to conventional VCVS operational amplifiers, transconductors are innovative in circuit design concepts in terms of current instead of voltage. They have become useful and important building blocks in realizing active devices[1]. They are especially favorable and promising candidate for MMIC in modern microwave and optical communication engineering. In addition, transconductors can be used as a unique element, such as resistor, in building up the certain amplifiers concerned[33]. OTA is also a valuable building block for realizing continuous-time components, such as filters at lower
frequencies. Thus, the demand and superior performance of transconductors applying GaAs MESFET technology in high frequency applications justifies the higher cost and complexity in fabricating GaAs MESFET chips.

Qualitatively, an OTA is essentially a voltage controlled current source, or VCCS, functioning distinctly from conventional and more familiar voltage-controlled voltage source (VCVS) operational amplifiers. A simplified construction of OTA, similar to other operational amplifiers, is intuitively expressed in Fig. 5.1[32,34]. The output current \( i_o \), as expected, is controlled by the difference of input voltages \( V_1 = V_i - V_2 \). The output might also be written as \( i_o = g_T V_i \), where \( g_T \) is the dynamic forward transfer conductance of the OTA. As shown in Fig. 5.2, an equivalent circuit for VCCS is presented and used to assist in analyzing AC characteristics of complex OTA circuits. Some flexibility of OTA can be seen immediately in Fig. 5.1; transfer conductance could be controlled by an externally supplied amplifier bias current, \( I_{abc} \); and the VCCS can be easily converted to VCVS by using \( R_L \) as load, thus \( V_o = i_o R_L \).

5.2 GaAs MESFET Current Mirrors in Transconductance Circuits

Fig. 5.3 is the circuit block which contains two pairs of nonlinear current mirrors, one differential amplifier and one MESFET as a current source regulating and providing the
current that the OTA needs. For this analysis purpose, the circuit can be split into smaller parts since each part serves a unique function inside the integrated circuit. First, we may examine Fig. 5.4, which has a single current mirror consisting of MESFET B4, B5, B6 and B7. The circuit virtually consists of only two linearized current mirrors, which are referred to as negative current mirrors[19]. B4 and B5 in the circuit are named nonlinear, inverting, voltage-following, negative current mirrors. Note that these current mirrors are nonlinear. Usually a simplified model for the drain current of GaAs MESFET in the saturation region (ignoring channel length modulation) is chosen [19]

\[ I_d = \left( \frac{K \cdot W}{2} \right) [V_{gs} - V_t]^2 \] (5.2.1)

where K is a transconductance factor, W is the gate width, \( V_{gs} \) is the gate-source voltage and \( V_t \) is the threshold voltage. If we assume that the GaAs MESFET drain current in the saturation region can be generally expressed as

\[ I_d = W \cdot F(V_{gs}) \] (5.2.2)

where F is a nonlinear function independent of \( V_{ds} \), the input current for B4 is

\[ I_{d4} = W_4 \cdot F(V_{gs4}) \] (5.2.3)
This gate-source $V_{gs4}$ is inverted and applied to B5, then the drain current for B5

$$I_{d5}=W_5 \cdot F(-V_{gs4}) \quad (5.2.4)$$

As the drain current in B5 is equivalent to that in B6, its gate-source voltage would also be $-V_{gs4}$. Therefore, we have

$$I_{d6}=W_6 \cdot F(-V_{gs4}) \quad (5.2.5)$$

Eventually the gate-source voltage of B6 is inverted again and applied to B7, as a result, the output drain current is given by

$$I_{d7}=W_7 \cdot F(V_{gs4}) = \left( \frac{W_7}{W_4} \right) I_{d4} \quad (5.2.6)$$

Since $(W_7/W_4)$ is a constant, linear operation with a gain of $(W_7/W_4)$ is achieved. On the other hand, the frequency response of the current-mirror circuit is crucial in determining the capability and stability of the circuit for high frequency applications. Quantitatively DC and AC gain are related as [19]

$$DC(gain) = \frac{1}{1+\frac{g_0}{g_m}} \quad (5.2.7)$$
This new circuit has a single pole and a single zero, and thus is stable with a wide bandwidth[19]. The OTA operates up to 10 GHz. A differential pair of input FETs B1 and B2 forms a push-pull architecture where the current mirror linking the outputs of the differential pair cancels the controlled bias current and at the same time doubles the transconductance. Desired voltage across each device is 2.5 volts. Three diodes provide the necessary voltage shift for the output of the circuit. In this case, the ideal bias at the output port is zero volts.

5.3 GaAs FET Nonlinear Model and Simulations

Prior to applying this OTA block in practice, a thorough analysis of DC characteristics of this circuit and simulations need to be performed with the tremendous assistance of CAD and CAE programs. As constantly emphasized throughout this paper, the nonlinear model of GaAs FET is playing a major role in analyzing circuits in either frequency domain simulator (LIBRA) or time-domain simulator (PSPICE). According to the OTA circuit, since the circuit is DC coupled, a linear model
can not be used directly. External control voltage $V_{ct}$ may be varied to tune the circuit. One may also adjust other bias voltages such as $V_{ss}$ and $V_{dd}$.

The model we use in simulations is the Raytheon empirical model, provided by Triquint Semiconductor, Inc[25]. The standard structure of the gate from Triquint Semiconductor is 6 fingers, with 0.5 microns of length and a 50 micron width for each finger. Thus the gate area is $W'=300$ microns. In our design the current source FET is using a 2 fingers gate and the rest of FETs are using a 1 finger gate. As for diodes, we may choose either one finger or two according to our needs. Therefore, in our simulations we need to scale down the area sizes that are represented by 100 microns and 50 microns respectively in the model. In other words, each channel (finger) is 0.5 by 50. The parameters in the model are normalized. Note that the gate resistance $R_{gs}$ is also scaled accordingly to represent the actual value for selected devices. The parameters for the model are listed in Table 5.1 and 5.2.

DC transfer characteristics of the OTA block are simulated using PSPICE. The results of the output currents over a sweep of input voltages are illustrated in Fig. 5.5 and 5.6. Polarities of OTA with differential input amplifiers have been determined. It can be observed from Fig. 5.5 that the transfer conductance possesses a positive slope as the positive input voltage is applied to $B_1$ and negative input to
B2; whereas, when the polarities are interchanged, the transconductance is negative. We designate B1 input as a positive terminal and B2 input as a negative terminal and use this convention accordingly in the following discussions. External $I_{ABE}$ can be controlled in terms of $V_{CT}$ as designed in the circuit, which can be adjusted in simulations. Unfortunately this introduces voltage offset at output. Observing Fig. 5.5, we find that $g_m$ is increased from 5.8 mA/V to 6.2 mA/V when $V_{CT}$ varied from -7.0V to -7.5V. When $V_{CT}$ reaches 8.5 volts, the dynamic range of DC sweep is shortened and $g_m$ is reduced as well. $g_m$ is the derivative of the output current with respect to the input voltage, which are represented by nearly two constant curves in Fig. 5.5. The circuit can have a dynamic range (input swing) of -0.5V to +0.5V when operating in the range of -7.0V and -7.5V of $V_{CT}$. A load of 50 ohm is used to make the circuit more practical. In addition, frequency response of the amplifier determines its usefulness in high speed applications. A plot of output current versus a wide range of frequencies is presented in Fig. 5.7. We can use this OTA amplifier operating up to 10GHz.

Applicability, versatility and flexibility are a few important elements in designing a new device. To illustrate any potential applications of OTA, we may adjust some parameters in the circuit in order to meet certain requirements. One way is to reduce the sizes of transistors to half. The offset of the output voltage may be eliminated by
adjusting the size of the diodes. The transconductance and input dynamic range and frequency are illustrated in Fig. 5.8 and 5.9. The length of the diode (finger or gate width of diode) is chosen at 45 microns in this case. We can get 3.2 mA/V $g_m$, which is almost half of this in previous cases, and which is almost linearly related to the change of sizes of the transistors. Frequency response stays more or less the same, since no change of the gate length of the transistors has been made.

Another way to change $g_m$ without changing the sizes of transistors, besides making the amplifier more controllable, is to add two more MESFETs to the differential inputs while using an external voltage source to control the conductance of these two transistors. The circuit is shown in Fig. 5.10. As shown, $V_{CG}$ is an external controlling source instead of $V_{CT}$ in this case. The transconductance of this circuit is illustrated in Fig. 11. As $V_{CG}$ varies from -5V to -7V, $g_m$ may be increased from 5.4 mA/V to 6.4 mA/V. Since the threshold voltage of transistors is -1.8V, -7V of $V_{CG}$ cuts off the conductance of two additional transistors and makes the circuit function the same way as the previous one. The advantage of this circuit is that there is no offset voltage caused and no rearrangement of sizes of MESFETs. The limitation is that $g_m$ can go down as far as 5.4 mA/V.

Table 5.1 Model parameters for GaAs FET device with area of 50
in LIBRA

<table>
<thead>
<tr>
<th>Mode</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>MODEL</td>
<td>2.5</td>
</tr>
<tr>
<td>VTO</td>
<td>-1.8</td>
</tr>
<tr>
<td>BETA</td>
<td>0.205E-3</td>
</tr>
<tr>
<td>LAMBDA</td>
<td>0.09</td>
</tr>
<tr>
<td>ALPHA</td>
<td>2.7</td>
</tr>
<tr>
<td>TAU</td>
<td>2.3E-12</td>
</tr>
<tr>
<td>RS</td>
<td>900</td>
</tr>
<tr>
<td>RD</td>
<td>900</td>
</tr>
<tr>
<td>CGD</td>
<td>2.9E-16</td>
</tr>
<tr>
<td>IS</td>
<td>3.0E-15</td>
</tr>
<tr>
<td>VBI</td>
<td>0.8</td>
</tr>
<tr>
<td>RG</td>
<td>12.6</td>
</tr>
<tr>
<td>B</td>
<td>1.0</td>
</tr>
<tr>
<td>CDS</td>
<td>2.3E-16</td>
</tr>
<tr>
<td>FC</td>
<td>0.5</td>
</tr>
<tr>
<td>N</td>
<td>1.15</td>
</tr>
</tbody>
</table>

Table 5.2 Model parameters of GaAs FET device with area of 100 in LIBRA

5.4 OTA used as an Integrator

The integrator is one of the most important applications of conventional operation amplifiers. It has a tremendous versatility of usage in the processing of linear analog signals, for instance, when used as a functional module, analog computation or active filter. Traditionally, OTA that acted as an integrator employed BJTs in the circuit. Without changing the configuration, we simulated the integrator using an OTA block with GaAs FETs instead. The quality of the integrator is the primary concern for an electronic circuit engineer in the process of new circuit design. The characteristics of ideal integrator may be derived from the
transfer function \([21,32]\)

\[ H_I = \frac{K}{s} \quad (5.4.1) \]

where \(K\) is a constant and, \(s = \sigma + j\omega\), and \(\omega = 2\pi f\). The magnitude and phase of the transfer function for an ideal integrator at the sinusoidal steady state \((\sigma = 0)\) may be written as

\[ |H(j\omega)|^2 = \frac{K^2}{\omega^2} \quad (5.4.2) \]

and

\[ \phi(j\omega) = -90^\circ \quad (5.4.3) \]

If high speed OTA makes a good integrator, we expect the frequency response be close to the ideal response. Toward the end, a series of circuit simulations and analysis is needed. The simplified configuration of the lossless integrator circuit is illustrated in Fig. 5.12[34]. The output capacitor is chosen \(C = 100\) PF. The capacitors \(C_{11}\) and \(C_{22}\) to the positive and negative terminals are 100 PF respectively. The load is 500 ohms to make it like an open-circuit. PSPICE and LIBRA are used to analyze and verify DC bias and AC response of this circuit. First, the magnitude and phase of the output voltage
simulated by PSPICE is shown in Fig. 5.13 and 5.14. Between 200MHz and 1 GHz, the phase of output voltage is within a 10 degree variation from the perfect -90°. The same integrator is simulated using LIBRA as well as for checking layout effects. Fig. 5.15 illustrates a somewhat deteriorated performance of the integrator with a shorter useful bandwidth between 200MHz and 650MHz, with 10 degree deviation from an ideal one. The degradation may come from parasitic inductors and capacitors of transmission lines and built-in capacitors and resistors.

5.5 OTA used as a Band-pass Filter

The configuration of a band-pass filter is shown in Fig. 5.16. Three OTA blocks are used in this circuit. The capacitors C1, C2 are chosen 0.5PF respectively. The circuit is first simulated using PSPICE to check its voltage output versus frequency. The plots of magnitude and phase are depicted in Fig. 5.17 and 5.18. It is evident that the circuit can be used as a band-pass filter between 1GHz to 3GHz. To further justify the performance of this filter, LIBRA is used to simulate a schematic circuit file which is similar to the input file used in PSPICE. However the difference lies in the different techniques of analyzing the circuits. In LIBRA harmonic balance analysis is used and the output is the power versus frequencies, instead of voltage. Fig. 5.19 presents the frequency response of the band-pass filter without layout
effects. Eventually the layout circuit file, in which all elements are represented by physical dimensions and material properties is created and analyzed by a harmonic balance technique. The output is again output power versus frequencies, illustrated in Fig. 5.20. Evidently, yet unusually, as indicated in the plot, the peak output power is not attenuated compared to that in Fig. 19, but it is enhanced by layout effects. One possibility is the parasitic inductors and capacitors added to the circuit help match the network output port that in turn would increase the output power. If the circuit is matched at the output, the highest possible gain of the circuit is above unity. Fig. 5.21 shows the phase of the output voltage of the band-pass filter including layout effects.

5.6 Layout of the OTA

Layout of the OTA circuit is a necessary step in order for the design to be fabricated and for chips to be made available for further testing. Usually designers use computer-aided designs to configure the layout for the design circuit and then simulate it with selective simulator(s) associated with the CAD programs. Since we have simulated OTA and have obtained satisfactory results, we shall concentrate the work on using ACADEMY[28] to turn the schematic circuit into a layout format.
ACADEMY is one of the powerful CAD programs among those released by EESOF for microwave design. It provides capabilities for schematic circuit editing, layout editing, and running simulations with LIBRA, TOUCHSTONE, MWSPICE, etc. Although it is equipped with an option of changing the schematic circuit for automatic layout, the conversion is not always satisfactory and the output is sometimes confusing and complex. Thus, designers need to decide which layout is the best for the circuit. At this point interactive and step-by-step development of the layout becomes important.

We will present three steps in the process of creating the layout by drawing three different circuit formats. Fig. 5.22 illustrates a pure schematic circuit with all the electrical elements needed in the design. This format is self-evident and easily applied for simulations. The annotations beside each element describe the property of that element. Ports can be connected to DC bias voltage and AC excitations. Following that is an extra step toward designing a layout of the circuit. Fig. 5.23 shows the configuration of the layout that includes many layout artworks such as transmission lines, thin film resistors and capacitors. It also contains connections elements like T-type connections, cross intersections and bends. It is, however, still in the schematic mode. The final step is to realize this configuration in the layout mode. Since we have done each previous step, the final layout is relatively easy to obtain.
and justify. The layout of OTA is depicted in Fig. 5.24. Note that some elements may not have artworks and do not show up in the layout; yet they are part of the layout and important in simulations. These elements may be substrates, nonlinear elements like GaAsFETs and diodes. During the simulations, the program will use the element identities to evaluate the electrical properties of each element.

In conclusion, an OTA amplifier using current mirrors of GaAs MESFETs was presented in this chapter. The amplifier was tested by simulations for possible usage at high frequency integrators and band-pass filters. Both DC and AC characteristics of the circuits were investigated to reveal the potential applications. Circuit layout was included for better understanding of layout effects in circuit performance.
Fig. 5.1 Symbol of VCCS.

Fig. 5.2 Equivalent circuit model of VCCS.
Fig. 5.3 Operational transconductance amplifier with linearized current-mirrors.
Fig. 5.4 Linearized current-mirrors.

Fig. 5.12 Integrator.

Fig. 5.16 Band-pass filter.
Fig. 5.5 Transfer characteristics and $g_m$ of OTA with positive input.
Fig. 5.6 Transfer characteristics of OTA with negative input.
Fig. 5.7 Frequency response of OTA.
Fig. 5.8 Transfer characteristics of OTA with MESFETs' size reduced to half.
Fig. 5.9 Frequency response of OTA with MESFET's size reduced to half.
Fig. 5.10 Operational transconductance amplifier with modified differential input that may be used to control output current.
Fig. 5.11 Transfer characteristics of OTA with tunable $V_{CO}$. 
Fig. 5.13 Magnitude of output voltage of the integrator.
Fig. 5.14 Phase of output voltage of the integrator.
Fig. 5.15 Phase of integrator with layout effects.
Fig. 5.17 Magnitude of output voltage of the band-pass filter.
Fig. 5.18 Phase of output voltage of the band-pass filter.
Fig. 5.19 Magnitude of output power of the band-pass filter without layout effects.
Fig. 5.20 Magnitude of output power of the band-pass filter with layout effects.
Fig. 5.21 Phase of output voltage of the band-pass filter with layout effects.
Fig. 5.22 Schematic circuit generated using ACADEMY.
Fig. 5.23 Semi-layout circuit.
Fig. 5.24 Layout of OTA circuit generated using ACADEMY.
Chapter 6

Conclusions

6.1 Summary of research

In our research a complete, standard procedure has been established to extract and optimize a model for a GaAs MESFET device based on the measurements of its I-V characteristics and small-signal S-parameters. These measurements are standardized and thus could be obtained from on wafer measurements of these devices. A fast, robust, easy-to-use, accurate and physically consistent model is the goal of the projected ideas, which is crucial in the modeling and simulation of electronic circuits. This goal has been mostly, if not completely, fulfilled.

The study has been approached according to the transistor's most useful properties, DC and AC characteristics. To the end, an algorithm to extract a DC nonlinear model has been defined using a conventional I-V curves fitting. A number of advanced, standard, empirical GaAs FET models have been implemented in our programs, which are also incorporated into most popular simulation programs such as PSPICE, LIBRA and MWSPICE. Although mathematical model and numerical optimization techniques are employed in the process, the final optimized model is usually geared to be as physically significant as possible. On the other hand, an
algorithm has been defined to extract an AC linear model using improved equivalent circuit topology. To enhance similarities of the empirical model and the physical model (or actual device), the property of independence of frequency of intrinsic elements has been used in our AC optimization program. Since the external parasitic resistors can be estimated or measured in the lab quite accurately, an emphasis of optimization is placed on other extrinsic and intrinsic components. Finally, a combination of DC and AC extracted models is incorporated into the SPICE nonlinear model for GaAs MESFET which is becoming standard model and extensively used in time-domain and frequency-domain simulation programs.

Model parameters for three devices are extracted using this procedure, two for GaAs FETs and one for HEMT. A series of justifications and simulations are made to compare calculated and measured data, which include I-V characteristics, transconductance and small-signal S-parameter analysis. The results are satisfactory for modeling two GaAs FETs. The combined model (SPICE model) shows a fairly good match of S-parameters on Smith charts. As for the extracted HEMT model, the result indicates that the currently existing GaAs MESFET model in PSPCIE and LIBRA is not adequate to describe the superior performance of the transconductance of HEMT and its much more complex structure. In our research we also compare the different empirical models, especially the Raytheon and Triquint models, when using optimization
programs. These two models were basically derived from the Curtice quadratic model with some major improvements in describing behavior of the drain current, gate-source and gate-drain capacitors. The result indicates that the Triquint model has a better match with respect to I-V characteristics. The Triquint model also shows a bit of improvement in calculating $S_{22}$ and $S_{12}$.

In the last stage of the research, an OTA block utilizing GaAs FET current mirror was designed to be used as a high speed integrator and microwave band-pass filter. PSPICE and LIBRA are used to simulate these circuits and execute the nonlinear analysis. The GaAs FET nonlinear models are used in both types of software, even though they have adopted different approaches in analyzing circuits. The outcome from the PSPICE simulations are justified by that from LIBRA with some layout effects. Layout components contain transmission lines, cross-shape intersections, T-shape intersections and bends. Thin film capacitors and thin film resistors are used in circuit layouts. In most cases, any layout effect is negative that would more or less impair the performance of an ideal circuit design. For instance, layout of the integrator contributes to the shortening of usable frequency bandwidths and limits applications in a higher frequency. But, in a few occasions, layout effect may enhance the output of the circuit design, such as the case of the band-pass filter. As shown in chapter 5 the layout might help match the network at the
output and thus increase the output power.

6.2 Suggestions for future study

We have noticed in DC nonlinear modeling that neither the Triquint nor the Raytheon model could accurately predict drain current at high current levels, such as $V_{gs}=0$ or above, if any. The nonlinearity of drain currents dependent on gate-source voltage has not been fully solved by existing empirical models. An innovative way for calculating drain conductance at high current levels may be needed to diminish error. Modeling the physical identification of parameters and their consistency may be as important as the accuracy of the extracted model. This is because design, analysis, and simulation of new devices and circuits, and the realization of them, should be interactive and be as closely related as possible. But usually physically derived models are not as accurate as empirical models or mathematical models in simulations. An empirical model may need some changes from the physical perspective. As for HEMT, a modification of GaAs MESFET models is needed. HEMT in our study does not increase current through the channel, but increases enormously the transfer conductance and drain conductance in the saturation region. Therefore, using the FET model to describe HEMT is not easy and accurate. Besides, the nonlinearity of the drain current with $V_{gs}$ is different from that of GaAs FET. The
voltage-dependent relationship becomes more complex. HEMT model development is suggested for future work.

One can never overestimate the importance of the modeling of GaAs MESFET or HEMT, as they are so crucial in nonlinear analysis and simulations. When it comes to practical applications, such as designing a new integrator or a band-pass filter, the accuracy of a nonlinear model directly affects the performance of the circuits. The optimization of the GaAs FET model and consideration of layout effects in simulation is one important step in reducing the gap between the design on paper and the real product. We have performed the simulations successfully, yet it would be better if the designs were verified by the processing of some of the designs presented in this thesis.
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