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Design and Evaluation of a Ka-Band Colpitts VCO for Wireless Communications

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

Master of Electrical Engineering

Ottawa-Carleton Institute of Electrical and Computer Engineering
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1998

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

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Abstract

Inspired by the success of the Colpitts oscillator at low frequencies, the design and performance of a Ka-Band Colpitts VCO is evaluated. The Colpitts VCO is a $300\text{ }\mu\text{m} \times 600\text{ }\mu\text{m}$ MMIC and uses a Nortel HBT. Lumped elements are used, in fact the HBT is de-stabilized by exploiting the parasitics of a spiral inductor. Operating within the HBT thermal constraints, the Colpitts VCO delivers -0.92 dBm at 33.83 GHz , and is tuned 82 MHz using the collector bias. The phase noise is -32.4 dBc/Hz at a carrier offset of 100 KHz . The measured performance variation from both circuit and electromagnetic simulations is evaluated, and the discrepancies explained. Indeed, individual sections of the Colpitts VCO are measured and evaluated, yielding further insight. Moreover, the large signal design procedure is evaluated and compared to other possible design approaches. Potential application of the Colpitts VCO to wireless communications, such as LMCS, is assessed.

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For Jen, Mom, Dad, Char and my Grandparents

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Chapter 1

Introduction

1.1 Motivation

A 1992 Hollywood movie, *Sneakers*, perhaps said it best: “The world isn’t run by weapons anymore, or energy, or money. It’s run by little ones and zeroes, little bits of data. It’s all just electrons.” Just a few short years ago, a war was fought in the Persian Gulf over that ever so precious natural resource - oil. Not much further in the past, a country’s wealth was primarily determined by the extent of its natural resources. But recently the world has undergone a quiet revolution. The power of information, or more specifically the power to communicate information, is now an immeasurable resource in the new borderless world of the late twentieth century. From the radio to satellite TV, the Internet to cellular phones, people are communicating more now than ever before. And it is precisely this communication that is causing exponential growth in science, technology, standard of living and economic power. This fact alone is what separates the first world from the developing.

Much of this communications growth is in the wireless segment. Signs of this growth are everywhere. For example,

- Today, 20 cents out of every dollar in revenue for the telecommunications providers

comes from wireless communications and that number is expected to increase to 80 cents in just over 10 years [1].

- There are more than 70 million cellular phone subscribers worldwide and that number could easily quadruple to 280 million in just 3 years [1].
- In 3 years nearly 10 million people are expected to transmit data by wireless networks [1].

It is quite possible that the revenue generated from wireless communications could exceed \$ 100 billion by the year 2000 [1]. Considering the number of possible applications of wireless communications, these forecasts are very attainable. Some of these applications are briefly outlined in the paragraphs to follow.

Mobile Communications: From Cellular Phones to PCS Cellular service worldwide is growing at a rate of 50% per year [1]. There are several reasons for this tremendous growth, one of which is decreased user costs. As wireless technology matures the user costs are lowered, since the initial infrastructure investment is financed and an increased number of competitors enter the market. With lower user costs user accessibility, and hence user demand, is increased. The popularity of cellular phones is also driving the development of personal communications services (PCS), which will provide superior coverage and digital quality to the user. While cellular phones are considered to be strictly mobile communicating devices, mobile PCS phones compete directly with existing fixed telephone service. Eventually, PCS telephone numbers may be associated with people as opposed to places as fixed telephone numbers are now. A phone call will be placed to a person, no matter where they are, rather than to a place such as a home or business.

Data Communications: From Pagers to the Internet About 75 million people worldwide use pagers and paging continues to grow at a steady pace [1]. Although the primary advantage

to pagers is low cost, modern digital display models offer messaging services as well as email delivery directly to the pager unit. Currently, most paging services are offered on a local basis, such as within a city, but soon service will be available on a much wider scale, perhaps even nationally. Wireless access to other forms of data communications, such as faxes and statistics, is also growing as the business need grows, but certainly the largest growth in data communications is the Internet. Wireless access to the Internet has the potential to offer instant access to vast amounts information with unrestricted mobility.

In-Building Communications: From the Cordless Phone to Wireless LAN At present, in-building cordless phones are so popular that they are probably not considered “high-tech” anymore. There are several hundred million cordless phones around the world and about half the homes in the United States have at least one [1]. Despite the large number of existing cordless phones, the market is growing. This growth is mostly due to replacements and to the better performance of newer models. While cordless phones are successful, a potentially larger market exists for in-building wireless local area networks (LAN). Wireless LANs offer much easier configuration and maintenance than their wired counterparts, and can even span multiple buildings that are in close proximity.

Broadband Communications: From Satellite to LMCS After the explosion of cellular phones and the Internet, the next big communications technology likely is broadband communications. Loosely speaking, broadband communication means a large amount of bandwidth, which translates to a large amount of information. With the capability of transmitting large amounts of information comes the ability to merge all forms of communication, such as telephone, television, data and the Internet, on one wireless link. Satellite communication is traditionally seen as the means to deliver broadband services. Satellites offer a large amount of bandwidth over a wide area. However, recently satellite services are being challenged by Local Multipoint Communications Systems (LMCS). Operating like a cellular network,

LMCS is a terrestrial wireless network offering two way services such as video on demand, teleconferencing, digital television, high-speed Internet service, multiple phone lines and fax lines.

ESP and the Future The development of new wireless technologies is causing the convergence of computers, communications and entertainment. But how far can this new multi-mode electronics be pushed? The question is probably best answered by the historians of the future, because we are not likely to have any idea at all. Science fiction is quickly becoming science fact and ideas that were once scoffed at are now being seriously investigated. For example, Sony is actively researching extrasensory perception (ESP)¹, trying to scientifically explain the phenomena and how the “wireless” information travels.

Enter the technology to power these new wireless systems. The voltage controlled oscillator (VCO) described and evaluated in this thesis represents one piece in the realization of these next generation networks. Perhaps once again Hollywood said it best in the 1989 movie *Field of Dreams*, “If you build it, they will come.” In the chapters and sections to follow the VCO, a small part of the overall picture, is presented and the limitless uses are left up to the people of the future.

1.2 Objective

Inspired by the success of the Colpitts oscillator at low frequencies, the thesis objective is to evaluate the design and performance of a state-of-the-art Ka-Band Colpitts voltage controlled oscillator (VCO). Thus design, manufacture and testing of a Ka-Band Monolithic Microwave Integrated Circuit (MMIC) is achieved. This MMIC is as compact as possible to allow the circuit to be integrated with other components, called higher order integration, and hence lower the cost. The classic Colpitts configuration is used because of its considerable

¹The ESPER Program, Sony Corporation

popularity at lower frequencies and its good performance with small size. The use of the Colpitts topology requires lumped elements, which at Ka-Band increases the design risk due to the lack of accurate models. The active element used is a Northern Telecom (Nortel) Heterojunction Bipolar Transistor (HBT), and Nortel fabricated the MMIC. The Nortel HBT was originally intended for digital applications, hence it has restrictive thermal limitations which are adhered to. A noise model is not available for the Nortel HBT and therefore noise was not a design consideration. Thus, the Colpitts oscillator was designed for maximum output power. The Colpitts design at Ka-Band is compared to design techniques and performance at low frequencies. Specifically, the following aspects are investigated:

- Ka-Band VCO design techniques are investigated compared to low frequency VCO design techniques,
- the applicability of the Colpitts topology at Ka-Band,
- the accuracy of the simulators (circuit and electromagnetic),
- the use of lumped elements, including parasitics, at Ka-Band,
- device de-stabilization using the parasitics of a spiral inductor,
- VCO tuning using HBT bias,
- the performance impact of an extremely compact circuit design, such as components coupling with each other,
- the impact of the digital HBT with low thermal tolerance on high power designs,
- performance comparison with other technologies, and,
- the cost effectiveness, reliability and manufacturability of the design.

1.3 History

Oscillators have been, and continue to be, essential to communications systems. Oscillators generate the signal which is modified to carry information from one point to another. Over the last century many different types of oscillators have been developed, however, the purpose remains the same. One of the first successful oscillators in communications is known as the Colpitts Oscillator, named after its inventor Edwin Colpitts, and remains very popular today.

Edwin H. Colpitts The year 1997 marks the 125th anniversary of the birth of Edwin H. Colpitts. Colpitts was a communications pioneer and is remembered primarily for the invention of the oscillator that bears his name. Additionally, Colpitts's team achieved significant advances in many types of vacuum tube circuits and, in 1915, successfully demonstrated the first transatlantic radio telephone.

Born in 1872 in New Brunswick, Canada, Colpitts attended Mount Allison College in 1893. After two years as a teacher and a principal in Newfoundland, he enrolled at Harvard University, where he studied physics and mathematics, and received a Master's degree in 1897 [2]. Colpitts remained at Harvard for two more years as a laboratory assistant, then left for the American Bell Telephone Company in 1899. In 1907 he joined the Western Electric Company, where in 1915 he proposed an oscillator circuit based on capacitive coupling as an alternative to the inductive coupling oscillator invented by his colleague, Ralph Hartley. The Colpitts oscillator, as it became known, was used in many systems over the next 4 years and was first reported in a 1919 journal paper with Edward Craft [3].

Colpitts patented his oscillator topology under the name "Oscillation Generator" in 1920 [4]. The oscillator schematic, scanned from the actual patent, is shown in Figure 1.1.

Existing Microwave Oscillators Although a Colpitts oscillator has not been implemented in HBT technology at microwave frequencies, many other oscillator types have. Table 1.1

enumerates some existing HBT microwave oscillators. Listed only from the early 1990s to the present, this list represents the current state-of-the-art in HBT microwave oscillators. The Ka-Band Colpitts oscillator performance is compared to these oscillators in Chapter 6.

1.4 Organization

The thesis contains a total of seven chapters. In this first chapter the thesis objective is outlined, along with a brief history to put the thesis into context. The motivation for the research is also included.

The second chapter outlines the essential oscillator fundamentals. Models for oscillators are also described to aid in the analysis of the circuits. The chapter concludes with a description of the traditional design techniques used for both low frequency and Ka-Band oscillators.

The third chapter describes the current state-of-the-art of Ka-Band technology. This includes such issues as MMIC fabrication, HBT theory and modeling, and computer aided design (CAD) tools.

Chapter four presents the Colpitts VCO design, starting with the evaluation criteria. A section on design strategy outlines the design approach used. Finally, the VCO design is described in detail.

Chapter five reports the Colpitts VCO performance. Circuit and electromagnetic simulations are presented, as well as the measurement procedures and results.

The sixth chapter includes a critical evaluation of the technology, design process and performance of the Ka-Band Colpitts VCO.

The seventh and final chapter concludes the thesis with a summary and recommendations for future study.

Figure 1.2 is a pictorial representation of the work flow.

Figure 1.2: Research work flow chart.

Chapter 2

Oscillator Fundamentals

2.1 Introduction

In this chapter the fundamentals of oscillator theory are presented. The details of the Colpitts topology and the models available for analysis are essential to understand the design and evaluation in later chapters. Derivations are included of only the most important concepts. An uninterested reader can easily skip the derivations as the concepts are more important than the details.

2.2 Oscillation

An oscillator produces an alternating current (AC) output signal without any AC input signal. When direct current (DC) bias is applied to the input terminals of the oscillator, AC energy is produced at the output terminals. Given that AC signal and radio frequency (RF) are synonymous, in the simplest terms, an oscillator is a DC to RF energy converter.

At its most basic level an oscillator consists of an active element and a passive network. The active element is a solid-state device and performs the energy conversion from DC

power to RF power. The passive network performs three main functions. First, it provides the proper DC bias for the device to operate. Second, the passive network presents the proper electromagnetic environment to the device terminals to initiate oscillation, shape its frequency spectrum and determine the oscillator output power. Lastly, the passive network ensures a stable thermal environment. The thermal requirements are an important consideration in oscillator design because the DC to RF energy conversion is typically not very efficient and the excess energy is dissipated in the form of heat.

Initially, an oscillator functions as a noise amplifier. When DC bias power is connected to the circuit the energy is transferred by the device to the noise present in the circuit. The device operates as an amplifier with the existing noise as the input, hence the term noise amplifier. Since the oscillator is a closed loop feedback system, the noise is amplified continuously. As the noise signal grows it is filtered by the passive network to contain the desired frequency components and becomes a useful RF signal. At some point the amplitude of the RF signal circulating in the loop is large enough to cause the device to operate in a nonlinear mode and the device nonlinearities cause the performance to degrade and the RF signal amplification decreases. This process is called device saturation. Eventually the device becomes completely saturated and the gain is unity. At this point the energy transferred from DC to RF exactly balances the loss in RF signal energy as it circulates the loop. At this point the oscillator is in equilibrium and is said to be operating at its steady-state condition.

At steady-state an oscillator is a large signal device while during start-up an oscillator operates under small signal conditions. It is therefore common to see both large signal and small signal descriptions of oscillator behavior. By small signal it is meant that the circuit elements are dependent on frequency only, and the variation of power does not affect their performance. Passive structures such as transmission lines and lumped elements are small signal elements, as are transistors operated at low power levels. Small signal behavior is a linear behavior. Circuit elements that are dependent on power as well as frequency dependent

are termed large signal elements. A transistor operated at moderate to high power levels is an example of a large signal element. Large signal behavior is nonlinear, hence nonlinear behavior is often used interchangeably with large signal behavior.

At steady-state the oscillator is a large signal circuit because the active element is a transistor. At start-up, however, when the power level is low, the transistor performance does not vary with power, hence the oscillator is a small signal circuit. Figure 2.1 shows this relationship between large and small signal oscillator behavior. As the power increases after start-up the oscillator operating point moves along the starting trajectory in Figure 2.1. The transition from small signal start-up to large signal steady-state causes the oscillator impedance, the frequency of oscillation, and output power to be different from the initial values. This is evident from Figure 2.1, however, it should be noted that the impedance variation shown is conceptual only since no distinction is made between magnitude and phase. Describing oscillator behavior in this manner is essential in most design procedures.

Figure 2.1: Oscillator small and large signal conceptual behavior [15].

Once the oscillator is operating in steady-state, the frequency of the output signal can be changed. This tuning is possible since changes in the passive network mainly determine the frequency of oscillation. Usually a passive component value is varied by a tuning voltage, hence the term Voltage Controlled Oscillator or VCO.

2.3 The Colpitts Oscillator Topology

The Colpitts oscillator topology is shown in Figure 2.2. The Colpitts oscillator, while physically very simple, is analytically fairly complex. The Colpitts oscillator is part of a large family of oscillators, called L-C oscillators, which share the same mathematical model but differ in physical realization. The Colpitts topology is distinguished from the other types primarily by:

- the type of reactive elements used in the feedback path,
- the manner in which DC power is fed to the oscillator, and,
- the method of delivering RF power to the load.

The Colpitts oscillator is roughly decomposed into the following elements:

- **The Transistor** The transistor is not specifically required by the Colpitts topology, in fact any gain producing element can be used.
- **The Inductor** The inductor primarily determines the frequency of oscillation, that is the inductor-transistor interaction determines the steady-state frequency of oscillation.
- **The Two Parallel Capacitors** The two parallel capacitors primarily destabilize the transistor to enable oscillations to occur. Loosely speaking, the capacitors use voltage division to feed back part of the output signal to the input to sustain oscillations.

Figure 2.2: The Colpitts oscillator topology.

- The Load Resistor-Transistor Interaction** The load resistor-transistor interaction determines the steady-state output power of the oscillator by controlling the saturation point of the transistor. Indeed, the load resistance controls the power level at which the loop gain falls to unity, causing the oscillation growth to cease and steady-state operation to start.

The physical simplicity of the Colpitts topology lowers its cost. The performance of the Colpitts oscillator is generally considered to be average; other oscillator types are more suited to specific applications, however the Colpitts topology provides average performance for most applications.

2.4 Oscillator Models

It is convenient to introduce models when analyzing oscillator circuits. Two models, namely the feedback and negative impedance models, are the most prevalent in the literature. Closer

examination of the two models, however, reveals their equivalence. Although equivalent, different mathematical and intuitive models of oscillator behavior is useful since certain oscillator analyses are easier with different models.

2.4.1 Feedback model

The feedback model is shown in Figure 2.3. The oscillator is divided into two parts: a gain element and a feedback network. The gain element is often a solid-state device, however, in general any amplifier circuit can be used. The feedback network is any type of frequency selective circuit.

Figure 2.3: The feedback oscillator model.

Steady-State Oscillatory Conditions

The steady-state behavior of the feedback oscillator model is best explained by referring to Figure 2.3. This topology is the classic feedback arrangement with $A(j\omega)$ as the forward gain and $B(j\omega)$ as the reverse gain. Note that the forward gain has been assumed independent of

power, which simplifies the discussion, however is not strictly speaking accurate. If the loop is broken and an input voltage, V_i , is applied at one terminal, the network transfer function is,

$$\frac{V_o}{V_i} = \frac{A(j\omega)}{1 - A(j\omega)B(j\omega)} \quad (2.1)$$

For the network to be oscillating, then $V_i = 0$. Hence for $V_o \neq 0$ when $V_i = 0$ it is necessary that

$$A(j\omega) = \infty \quad (2.2)$$

or

$$1 - A(j\omega)B(j\omega) = 0 \quad (2.3)$$

Equation 2.2 is unrealistic, hence the condition for oscillation is given by equation 2.3 and is known as the Barkhausen Criterion when it is rewritten as

$$A(j\omega)B(j\omega) = 1 \quad (2.4)$$

More insight into the oscillatory condition can be achieved if the Barkhausen Criterion is rewritten as

$$|A(j\omega_o)B(j\omega_o)| = 1 \quad (2.5)$$

and

$$\angle(A(j\omega_o)B(j\omega_o)) = 0 \quad (2.6)$$

where ω_o is the frequency of oscillation. Equation 2.5 is known as the open loop gain, or often just loop gain. Equation 2.6 is known as the loop phase.

In words, the Barkhausen Criterion states that if in a negative feedback network the open loop gain has a total phase shift of 0° (or 360°) at some frequency ω_o , the network will oscillate at ω_o provided that the loop gain is unity.

Note that sometimes the oscillatory condition is derived using a positive feedback network with a transfer function of $\frac{V_o}{V_i} = \frac{A(j\omega)}{1 + A(j\omega)B(j\omega)}$. The stability of the system is then described by

the Nyquist Criterion. Mathematically the two criteria differ only by a negative sign, however they both describe the same phenomenon for oscillations.

Start-up Condition

The start-up condition is,

$$|A(j\omega_o)B(j\omega_o)| > 1 \quad (2.7)$$

The loop gain must be larger than unity so that the noise present in the loop initially is amplified and the power can grow in the loop.

Noise and Stability

Amplitude stability is often not a significant issue compared to the phase stability of the oscillator. The amplitude is controlled by the limiting of the device, and the deviations from the saturation of the device are slight.

There are two distinct forms of phase (or frequency, since frequency is related to phase) stability. The first type is long term stability where the frequency of oscillation changes over a period of minutes to years. Long term phase stability is due to temperature and aging effects of the oscillator components. Conversely, short term stability is measured in seconds. Short term stability affects the output frequency of the oscillator by changing the phase of the system, hence phase and frequency stability are used interchangeably. Indeed, short term stability refers to how the frequency of oscillation responds to phase changes in the system. A highly stable system in the short term processes the largest rate of change of frequency with respect to phase, $\frac{df}{d\phi}$. To visualize the effect of this last statement, see Figure 2.4, which shows the loop phase of two feedback model oscillators. According to equation 2.6, at the frequency of oscillation f_o the loop phase is 0° . If some external influence causes a shift in loop phase of 10° , then the frequency of oscillation for the first oscillator will shift by Δf_1 . The same external influence causes a frequency of oscillation shift of Δf_2 for the second

oscillator. From Figure 2.4 it is clear that $\Delta f_1 < \Delta f_2$ due to the higher $\frac{d\phi}{df}$. Consequently the first oscillator possess better phase stability than the second oscillator because $\frac{d\phi}{df}$ is higher.

Figure 2.4: Phase stability comparison of two feedback model oscillators [16].

Further, the short term frequency stability of the feedback model is [16],

$$S_F = 2Q \quad (2.8)$$

where Q is the quality factor of the feedback network. Hence, the higher the Q in the oscillator feedback network, the more stable and less noisy the oscillator will be.

Feedback Model of the Colpitts Oscillator

The Colpitts oscillator is shown as a feedback model in Figure 2.5. The Colpitts oscillator is well suited to the feedback model because the Colpitts topology is based on a parallel-resonant circuit structure.

Figure 2.5: The Colpitts oscillator using the feedback model.

2.4.2 Negative Impedance Model

The negative impedance oscillator model is shown in Figure 2.6. Kurokawa [17] is credited with much of the development of the negative impedance oscillator model. As in the feedback model, the negative impedance model describes the oscillator in two parts: the device and the resonator circuit. The noise initially present in the oscillator is modeled by the voltage source $e(t)$.

Steady-State Oscillatory Conditions

The impedance oscillatory condition is $Z_{Device} + Z_{Circuit} = 0$. The derivation follows the approach outlined by Kurokawa in a noted paper [17] and concludes with the impedance oscillatory condition in equation 2.26. A disinterested reader may skip the derivation by proceeding immediately to page 23.

Figure 2.6: The negative impedance model for an oscillator.

The oscillator is a closed loop system and hence any analysis must begin by selecting an appropriate point to break the loop. Since the oscillatory condition is defined in terms of device and circuit impedances, it is appropriate to define a reference plane at the junction of the resonator circuit and the device as shown in Figure 2.6. Note that the device is shown as a one port network, which is the case for negative resistance diodes. Two port transistors are not excluded from this analysis because they are easily configured as one port devices by de-stabilizing elements.

At start-up, noise in the circuit induces the required oscillations, therefore an equivalent voltage source, $e(t)$, is included in Figure 2.6. The oscillator is then completely described by two equations. The first equation is the RF current,

$$i(t) = \Re\{I\} = A(t) \cos[\omega t + \phi(t)] \quad (2.9)$$

where

$$I(t) = A(t)e^{j[\omega t + \phi(t)]} \quad (2.10)$$

$\Re$ denotes real part. A and ϕ are unknown amplitude and phase variables and are assumed to vary slowly with time as compared to $i(t)$. ω is the fundamental oscillation frequency. Higher order harmonics have been ignored at this point for simplicity. They are eliminated

in an integration step later in the derivation.

The second equation is the voltage loop relation,

$$v_d(t, A) + \Re[Z_c(\omega)I(t)] = e(t) \quad (2.11)$$

Solving equations 2.9 and 2.11 simultaneously will yield solutions for A and ϕ . The oscillation equation then follows from these expressions when some simplifications are made.

Equation 2.9 is already in a useful form however it is beneficial to expand the terms in equation 2.11. Hence using Ohms law we obtain the voltage drop across the device,

$$v_d(t) = i(t)(R_d + jX_d) = R_d A \cos(\omega t + \phi) - X_d A \sin(\omega t + \phi) \quad (2.12)$$

where $R_d < 0$.

The second term in equation 2.11 is more difficult to expand due to the frequency dependence of Z_c . Kurokawa used perturbation analysis to obtain a first order approximation of this frequency dependence and obtained a more explicit expression for the voltage drop across $Z_c(\omega)$. The frequency perturbation $\delta\omega$ to ω gives

$$\omega' = \omega + \delta\omega \quad (2.13)$$

Kurokawa obtained an expression for $\delta\omega$ by examining the derivative of the current

$$\frac{dI(t)}{dt} = j\omega' I(t) = \left[j\left(\omega + \frac{d\phi}{dt}\right) + \frac{1}{A} \frac{dA}{dt} \right] I(t) \quad (2.14)$$

If the second and all other higher order derivatives of A and ϕ with respect to t are ignored then a first order approximation for the m th derivative is

$$\frac{d^m I}{dt^m} \approx (j\omega')^m I(t) \quad (2.15)$$

These results are consistent with AC Laplace transform circuit theory, so that

$$\omega' = \omega + \delta\omega = \omega + \frac{d\phi}{dt} - j\frac{1}{A} \frac{dA}{dt} \quad (2.16)$$

Finally, if we assume that $|\delta\omega| \ll \omega$ then, expanding $Z_c(\omega')$ in a Taylor series about $\omega' = \omega$ gives

$$\begin{aligned} Z_c(\omega') &\approx Z_c(\omega) + \frac{dZ_c(\omega)}{d\omega} \delta\omega \\ &= R_c(\omega) + jX_c(\omega) + \left[\frac{dR_c(\omega)}{d\omega} + j \frac{dX_c(\omega)}{d\omega} \right] \left[\frac{d\phi}{dt} - j \frac{1}{A} \frac{dA}{dt} \right] \end{aligned} \quad (2.17)$$

Now the perturbed voltage drop across Z_c can be expressed as

$$\begin{aligned} \Re[Z_c I] &= \left[R_c(\omega) + \frac{dR_c(\omega)}{d\omega} \frac{d\phi}{dt} + \frac{dX_c(\omega)}{d\omega} \frac{1}{A} \frac{dA}{dt} \right] A \cos(\omega t + \phi) \\ &\quad - \left[X_c(\omega) + \frac{dX_c(\omega)}{d\omega} \frac{d\phi}{dt} + \frac{dR_c(\omega)}{d\omega} \frac{1}{A} \frac{dA}{dt} \right] A \sin(\omega t + \phi) \end{aligned} \quad (2.18)$$

Substituting equations 2.12 and 2.18 into the loop equation 2.11 yields

$$\begin{aligned} &\left[R_c(\omega) + R_d(\omega) + \frac{dR_c(\omega)}{d\omega} \frac{d\phi}{dt} + \frac{dX_c(\omega)}{d\omega} \frac{1}{A} \frac{dA}{dt} \right] A \cos(\omega t + \phi) \\ &- \left[X_c(\omega) + X_d(\omega) + \frac{dX_c(\omega)}{d\omega} \frac{d\phi}{dt} + \frac{dR_c(\omega)}{d\omega} \frac{1}{A} \frac{dA}{dt} \right] A \sin(\omega t + \phi) = e(t) \end{aligned} \quad (2.19)$$

Equation 2.19 is multiplied by $\cos(\omega t + \phi)$ and then by $\sin(\omega t + \phi)$, and in each case integrated over one oscillation period, T_o , to give

$$R_c(\omega) + R_d(\omega) + \frac{dR_c(\omega)}{d\omega} \frac{d\phi}{dt} + \frac{dX_c(\omega)}{d\omega} \frac{1}{A} \frac{dA}{dt} = \frac{1}{A} e_c(t) \quad (2.20)$$

and

$$-X_c(\omega) - X_d(\omega) - \frac{dX_c(\omega)}{d\omega} \frac{d\phi}{dt} + \frac{dR_c(\omega)}{d\omega} \frac{1}{A} \frac{dA}{dt} = \frac{1}{A} e_s(t) \quad (2.21)$$

where

$$e_c(t) = \frac{2}{T_o} \int_{t-T_o}^t e(t) \cos(\omega t + \phi) dt \quad (2.22)$$

$$e_s(t) = \frac{2}{T_o} \int_{t-T_o}^t e(t) \sin(\omega t + \phi) dt \quad (2.23)$$

Equations 2.20 and 2.21 are easily manipulated into two equations where one contains only $\frac{dA}{dt}$ and the other contains only $\frac{d\phi}{dt}$. Hence,

$$[R_c(\omega) + R_d(\omega)] \frac{dX_c(\omega)}{d\omega} - [X_c(\omega) + X_d(\omega)] \frac{dR_c(\omega)}{d\omega} + \left| \frac{dZ_c(\omega)}{d\omega} \right|^2 \frac{1}{A} \frac{dA}{dt}$$

$$= \frac{1}{A} \left[\frac{dX_c(\omega)}{d\omega} e_c(t) + \frac{dR_c(\omega)}{d\omega} e_s(t) \right] \quad (2.24)$$

and,

$$\begin{aligned} [R_c(\omega) + R_d(\omega)] \frac{dR_c(\omega)}{d\omega} - [X_c(\omega) + X_d(\omega)] \frac{dX_c(\omega)}{d\omega} + \left| \frac{dZ_c(\omega)}{d\omega} \right|^2 \frac{d\phi}{dt} \\ = \frac{1}{A} \left[\frac{dR_c(\omega)}{d\omega} e_c(t) - \frac{dX_c(\omega)}{d\omega} e_s(t) \right] \end{aligned} \quad (2.25)$$

Equations 2.24 and 2.25 are called the general dynamic oscillation conditions and they determine the oscillating current amplitude and phase at the fundamental frequency.

During steady-state, free-running conditions, the noise source is negligible compared to the oscillation signal, hence $e(t) = 0$. Additionally, the amplitude and phase are constant, hence $\frac{dA}{dt} = 0$ and $\frac{d\phi}{dt} = 0$. Using these values we obtain the expected result,

$$R_c(\omega) + R_d(\omega) + j[X_c(\omega) + X_d(\omega)] = 0 \quad (2.26)$$

Finally, the oscillator equation gives the necessary impedance conditions for stable, steady-state oscillations. Since $R_d < 0$ the oscillator must be designed to satisfy

$$R_c = -R_d \quad (2.27)$$

and

$$X_c = X_d \quad (2.28)$$

Start-up Conditions

For start-up, a rule of thumb is to set [18],

$$R_c = \frac{-R_d}{3} \quad (2.29)$$

This heuristic approximation is well known in the art, and it ensures that the negative impedance device produces enough power to overcome the losses in most resonator circuits.

If this is the case, the oscillations will grow to steady-state.

Noise and Stability

A non-ideal oscillator produces random phase and amplitude distortions around the carrier. This phenomena is known as noise. Noise and stability expressions are presented here without derivation, and an interested reader is referred to the references.

The amplitude modulation noise in an oscillator output in terms of power spectral density is [18]

$$|\delta A(f)|^2 = \frac{2 \left| \frac{dZ_c(\omega_o)}{d\omega} \right|^2 |e|^2}{\omega^2 \left| \frac{dZ_c(\omega_o)}{d\omega} \right|^4 + A_o^2 S^2} \quad (2.30)$$

The phase noise in an oscillator output in terms of power spectral density is [18]

$$|\phi(f)|^2 = \frac{2|e|^2}{\omega^2 A_o^2} \frac{\omega^2 \left| \frac{dZ_c(\omega_o)}{d\omega} \right|^2 + A_o^2 \left[\left(\frac{\partial R_d}{\partial A} \right)^2 + \left(\frac{\partial X_d}{\partial A} \right)^2 \right]}{\omega^2 \left| \frac{dZ_c(\omega_o)}{d\omega} \right|^4 + A_o^2 S^2} \quad (2.31)$$

Equation 2.26 is sufficient to guarantee that oscillations are possible however small electrical perturbations in the circuit could cause this condition to be violated. Therefore an important design parameter is oscillator stability. Stability is the ability of an oscillator, given some perturbation, to converge back to the stable operating point.

Oscillator stability is derived by perturbing the amplitude by an amount $\delta A(t)$ from the steady-state value A_o . If $\frac{\delta A}{dt} < 0$ then the oscillator will return to the stable operating point.

The stability, S , is [18]

$$S = \frac{\partial R_d}{\partial A} \frac{dX_c(\omega_o)}{d\omega} - \frac{\partial X_d}{\partial A} \frac{dR_c(\omega_o)}{d\omega} = \frac{dR_c(\omega_o)}{d\omega} \frac{\partial R_d}{\partial A} [\tan(\theta_c) - \tan(\theta_d)] > 0 \quad (2.32)$$

where

$$\tan(\theta_c) = \frac{\frac{dX_c(\omega_o)}{d\omega}}{\frac{dR_c(\omega_o)}{d\omega}} \quad (2.33)$$

$$\tan(\theta_d) = \frac{\frac{\partial X_d}{\partial A}}{\frac{\partial R_d}{\partial A}} \quad (2.34)$$

The stability expression 2.32 is useful in the oscillator design process. It should be noted, however, that it is approximate because the derivation assumes that the device impedance

is a function of RF current amplitude only. Bates *et al.* [19] have derived a more general stability criterion where the device impedance is a function of both amplitude and frequency and that an arbitrary number of frequency components may be present in the circuit. While more accurate, Bates *et al.* more general criterion is much more difficult to evaluate hence the single frequency criterion is used here.

Colpitts Oscillator as a Negative Impedance Model

The Colpitts oscillator is shown as a negative impedance model in Figure 2.7. The Colpitts topology is not as easily analyzed using the negative impedance model as compared to the feedback model, nevertheless it is equally valid to do so.

Figure 2.7: The Colpitts oscillator using the negative impedance model.

2.4.3 Model Equivalence

Although seemingly different, the feedback and negative impedance models are equivalent. Consider the feedback model as shown in Figure 2.3. Instead of using gains $A(j\omega_o)$ and $B(j\omega_o)$, represent the gain element and feedback networks as impedances, Z_A , and Z_B , respectively. If the load is included in Z_A , then the feedback model is as seen in Figure 2.8, which is identical to the negative impedance model in Figure 2.6, where $Z_A = Z_d$ and $Z_B = Z_c$. Thus the models are equivalent and the choice depends on which is more suitable for the calculations of the particular situation.

Figure 2.8: Feedback model in terms of impedances.

2.5 Traditional Design Approaches

The traditional design approaches vary depending on the frequency of oscillation. Hence, traditional design methodology is presented for low-frequency oscillators and for Ka-Band oscillators.

2.5.1 Low Frequency Oscillators

The traditional low frequency design approach uses design equations in terms of transistor parameters and feedback elements. This approach uses the feedback model to derive the design equations. For a detailed derivation the reader is referred to [20].

Steady-state Oscillatory Condition

From the loop phase oscillation condition in equation 2.6, the first steady-state design equation follows as,

$$X_1 + X_2 + X_3 = 0 \quad (2.35)$$

where X_1 , X_2 , and X_3 are the reactances of the feedback elements. Recall that in a Colpitts oscillator, the feedback network contains only reactive elements. Note also that the load is ignored for simplicity in this analysis.

Using the loop gain oscillation condition, the second steady-state design equation is found to be,

$$\frac{-A_v X_1}{X_1 + X_3} = 1 \quad (2.36)$$

where A_v is the gain of the transistor gain element.

Rearranging equation 2.35 to $X_2 = -(X_1 + X_3)$ and substituting into equation 2.36 yields,

$$\frac{A_v X_1}{X_2} = 1 \quad (2.37)$$

Therefore from equation 2.37 a usable form of the second steady-state design equation is,

$$A_v = \frac{X_2}{X_1} \quad (2.38)$$

The feedback elements are chosen using equations 2.35 and 2.38. Further, if we consider that for a Colpitts oscillator $X_1 = \frac{1}{\omega_o C_1}$, $X_2 = \frac{1}{\omega_o C_2}$ and $X_3 = \omega_o L$, then using equations 2.35 and 2.38 the Colpitts oscillator is designed for steady-state using,

$$\omega_o = \frac{1}{\sqrt{L \frac{C_1 C_2}{C_1 + C_2}}} \quad (2.39)$$

and

$$A_v = \frac{C_2}{C_1} \quad (2.40)$$

where the gain from base to collector, A_v can be expressed in terms of the linear transistor model parameters as $A_v = g_m R$.

Start-up Condition

For oscillation start-up the magnitude of the loop gain must be greater than one, as in equation 2.7. It follows from equation 2.40, in order to design for oscillation start-up, the gain from the base to collector must be set to,

$$A_v > \frac{C_2}{C_1} \quad (2.41)$$

Noise and Stability

The output signal of an ideal oscillator is described by a pure sinusoid of constant amplitude and phase. In reality the amplitude and phase are random, time-varying processes. Hence the output signal of a real oscillator is expressed as,

$$v(t) = [1 + n(t)] \cos[\omega_o(t) + o(t)] \quad (2.42)$$

with random processes $n(t)$ = amplitude noise modulation and $o(t)$ = phase noise modulation. In general the amplitude noise modulation is less significant than the phase noise modulation, and hence is not considered further.

In the time domain, the phase noise modulation causes instantaneous shifts in the oscillator operating frequency. This shifting effect is analogous to phase modulation communications theory, and is called phase noise.

In the frequency domain the phase noise effect is characterized by continuous spectral sidebands which generally decrease with increasing frequency offset, f_m , from the carrier. When

analyzing this so-called single sideband phase noise (SSB) phase noise $L(f_m)$, a frequency domain approach is most often used.

Using the feedback oscillator model, Leeson described a model for SSB phase noise [21]. Leeson postulated that at baseband frequencies within the resonator half bandwidth, phase perturbations at the oscillator input due to noise or some other parameter result in frequency shifts at the output. It followed that the SSB phase noise would have a slope of -30 dB/decade or $\frac{1}{f^3}$ below the flicker corner and a slope of -20 dB/decade or $\frac{1}{f^2}$ above the flicker corner but below the resonator half bandwidth, $\frac{f_o}{2Q}$. At frequency offsets above the half bandwidth of the resonator the SSB noise floor is flat. Combining these effects gives Leeson's equation for SSB phase noise.

$$L(f_m) = 10 \log \left[\frac{1}{2} \left(\left(\frac{f_o}{2Q_l f_m} \right)^2 + 1 \right) \left(\frac{f_c}{f_m} + 1 \right) \left(\frac{FkT}{P_s} \right) \right] \quad (2.43)$$

and this is shown pictorially in Figure 2.9.

Amplitude noise is not a dominant characteristic in oscillators and is therefore not addressed.

2.5.2 Ka-Band Oscillators

The traditional design approach for Ka-Band oscillators uses reflection coefficients in the design equations. Reflection coefficients are easily determined from s-parameters, and s-parameters are easily measured at Ka-Band. This approach uses the negative impedance model to derive the design equations.

Steady-state Oscillatory Condition

Kurokawa's oscillation conditions can also be represented in terms of reflection coefficients.

Consider the circuit in Figure 2.10. The reflection coefficient seen from the device looking into the circuit is Γ_c . Γ_d is the reflection coefficient looking into the circuit from the device.

Figure 2.9: Oscillator SSB phase noise according to Leeson's model [21].

Figure 2.10: Oscillator schematic in terms of reflection coefficients.

The goal of this section is to develop the steady-state oscillation condition, $Z_d + Z_c = 0$, in terms of Γ_c and Γ_d .

Using the relationship between impedance, Z , characteristic impedance, Z_o , and reflection coefficient, Γ , we can rewrite the oscillator condition. Specifically, using

$$\Gamma = \frac{Z - Z_o}{Z + Z_o} \quad (2.44)$$

the oscillation condition becomes

$$\frac{1 + \Gamma_d}{1 - \Gamma_d} + \frac{1 + \Gamma_c}{1 - \Gamma_c} = 0 \quad (2.45)$$

where the characteristic impedance of the oscillator, Z_o , was divided out of the equation. With simple algebraic manipulation equation 2.45 reduces to

$$\Gamma_c = \frac{1}{\Gamma_d} \quad (2.46)$$

Since most resonator circuits are passive, $0 \leq |\Gamma_c| \leq 1$. Therefore in order for oscillations to take place equation 2.46 requires that $|\Gamma_d| > 1$. Having a device reflection coefficient greater than one means that the device returns more energy than it receives. Thus the device generates power, which is indicated by the negative device resistance.

Start-up Condition

For start-up,

$$\Gamma_c > \frac{1}{\Gamma_d} \quad (2.47)$$

Noise and Stability

Expressions for the noise and stability, in terms of reflection coefficients, are as follows. For derivations, the reader is referred to the references.

The amplitude modulation noise spectrum at modulation frequency ω_m is [18]

$$|\delta A(\omega_m)|^2 = \frac{\frac{1}{|\Gamma_c|^2} \left| \frac{d\Gamma_c}{d\omega} \right|^2 2|b_n|^2}{|A_o|^2 \frac{1}{|\Gamma_c|^2} S^2 + \frac{1}{|\Gamma_c|^4} \left| \frac{d\Gamma_c}{d\omega} \right|^4 \omega_m^2} \quad (2.48)$$

and phase noise spectrum at modulation frequency ω_m is [18]

$$|\delta \phi(\omega_m)|^2 = \frac{2|b_n|^2 \frac{1}{|\Gamma_c|^2} \left| \frac{d\Gamma_c^{-1}}{dA} \right|^2 |A_o|^2 + \frac{1}{|\Gamma_c|^2} \left| \frac{d\Gamma_c}{d\omega} \right| \omega_m^2}{\omega_m^2 |A_o|^2 \frac{1}{|\Gamma_c|^2} S^2 + \frac{1}{|\Gamma_c|^4} \left| \frac{d\Gamma_c}{d\omega} \right|^4 \omega_m^2} \quad (2.49)$$

where

$$b_n = \frac{|e|^2 |1 - \Gamma_d|^2}{4Z_o |\Gamma_d|^2} \quad (2.50)$$

Therefore, good noise performance is achieved if the circuit and device reflection coefficient loci intersect orthogonally, $\left| \frac{1}{|\Gamma_c|} \frac{d\Gamma_c}{d\omega} \right|$ is a maximum and $\left| \frac{1}{|\Gamma_c|} \frac{d\Gamma_c^{-1}}{dA} \right|$ is a minimum.

The stability is given by [18]

$$S = \left| \frac{d\Gamma_c}{d\omega} \right| \left| \frac{d\Gamma_d^{-1}}{dA} \right| \sin \theta > 0 \quad (2.51)$$

where θ is the angle between the device and the circuit impedance versus frequency loci, plotted on the reflection coefficient plane. When stability is expressed in the form of equation 2.51, the oscillator is stable for $0 \leq \theta \leq \pi$.

Chapter 3

Advanced Ka-Band Technology

3.1 Introduction

In this chapter the current state-of-the-art in Ka-Band technology is presented. This includes an explanation of the theory of operation and model for the active device, namely the Nortel HBT, used in the Colpitts VCO. Monolithic microwave integrated circuits are introduced as are the computer aided design tools used in the current research.

3.2 Gallium Arsenide Heterojunction Bipolar Transistors

Metal Semiconductor Field Effect Transistor (MESFET) technology was the first gallium arsenide (GaAs) transistor to be used for microwave circuits. Indeed, it is still popular today for applications from 1 to 20 GHz, primarily due to mature performance models. In order to extend the frequency performance, MESFET technology was modified to produce High Electron Mobility Transistors (HEMTs) and Pseudomorphic High Electron Mobility Transistors (PHEMTs). HEMTs and PHEMTs are quite good millimetrewave transistors in many performance areas.

While GaAs transistors were being researched for microwave circuits, silicon (Si) designers were working to improve the high frequency performance of Silicon Bipolar Junction Transistors (BJTs). Si BJTs are much cheaper than GaAs devices, and today Si BJTs can be used successfully up to 15 GHz. As with the GaAs technology, recently BJTs have been modified to produce Silicon Germanium (SiGe) Heterojunction Bipolar Transistors (HBTs) which perform very well at high frequencies.

Gallium Arsenide Heterojunction Bipolar Transistor technology has been under development since the early 1980's. Today, the technology has matured to provide high performance, low cost components for Ka-Band applications. At present, research continues to exploit HBT technology since HBTs have not been conclusively proven to be superior to other millimetrewave technologies, such as HEMTs and PHEMTs. In the sections which follow, the theory of operation of the GaAs HBT is presented in the context of its superior performance to other technologies.

3.2.1 Theory of Operation

The HBT was first patented by William Shockley in 1948. In 1957, Kroemer [22] outlined the theory of the use of a wide-band-gap emitter to improve the performance of a n-p-n transistor. To achieve the wide-band-gap emitter, materials such as InGaP or AlGaAs are used. Since the base is GaAs, the emitter-base junction is no longer homogeneous but instead heterogeneous, hence the name heterojunction bipolar transistor.

The theory of operation of an HBT is presented in comparison with Si BJTs. Sedra *et al.* [20] offers an excellent review of Si BJT operation and hence it is not covered here. The operation of a GaAs HBT is most easily understood when compared to a Si BJT.

There are two main advantages of the GaAs HBT compared to the Si BJT [23]:

- The higher cutoff frequency, f_T , of GaAs, due to the higher electron mobility.

- The higher maximum frequency of oscillation, f_{max} , due to the emitter-base heterojunction.

These two points are further illustrated as follows.

Higher Cutoff Frequency Electron mobility, μ , is defined as the slope of the velocity-electric field curve, at a given steady-state electric field. A GaAs transistor substrate has greater low-field electron mobility than a Si transistor substrate, as shown in Figure 3.1. This higher electron mobility leads to a shorter transit time, τ_{EC} , across the emitter-base depletion region, the base region and the base-collector depletion region. The cutoff frequency, f_T , can be expressed as

$$f_T = \frac{1}{2\pi\tau_{EC}} \quad (3.1)$$

and since the cutoff frequency is inversely proportional to transit time, the cutoff frequency for GaAs is higher than for Si.

Figure 3.1: Steady-state velocity-electric field characteristics for electrons in GaAs and Si [24].

Maximum Frequency Oscillation In order to understand why the heterojunction increases the maximum frequency of oscillation, f_{max} for the HBT, we must start by defining f_{max} as

$$f_{max} = \sqrt{\frac{f_T}{8\pi R_b C_{bc}}} \quad (3.2)$$

In equation 3.2 R_b is the base resistance and C_{bc} is the base-collector depletion capacitance. Therefore to improve f_{max} , R_b and C_{bc} must be reduced and this is accomplished by the heterojunction.

Specifically, the improvement is as follows. The base resistance is decreased by an increased base doping concentration, N_p . The base-collector depletion capacitance is decreased by less minority carrier storage in the emitter, or a decreased emitter doping concentration, N_n . Hence, if the doping ratio, $\frac{N_p}{N_n}$, is increased, the maximum frequency of oscillation is improved. The doping ratio is related to maximum current gain, β_{max} , by [25]

$$\beta_{max} = \frac{N_n}{N_p} \frac{v_{nb}}{v_{pe}} e^{\frac{\Delta E_g}{kT}} \quad (3.3)$$

Therefore from equation 3.3 an increase in doping ratio will decrease the maximum current gain. In order to offset the decrease in β_{max} the heterojunction is used to increase the emitter band gap offset in the valence band, ΔE_g .

In summary, by engineering the bandgap of the emitter-base junction, the maximum frequency of oscillation can be increased. This so-called bandgap engineering is accomplished by using a wide energy bandgap material at the emitter to create a heterojunction, as opposed to the homojunction found in conventional Si BJTs. It is desirable for the material used at the emitter to have a lattice constant similar to that of the GaAs in the base. The lattice constant, a_o , of a material is related to the spacing between two atoms at equilibrium, where equilibrium is defined as when the gravitational force is balanced by the Coulomb force. In actuality the interatomic spacing varies around a_o due to thermal atomic vibrations. Similar lattice constants allow chemical bonding with minimal strain between crystalline structures. The Nortel GaAs HBT uses a lattice-matched $\text{Ga}_{51}\text{In}_{49}\text{P}$ semiconductor alloy with a bandgap of 1.43 eV [23].

3.2.2 Model

A device model that is valid for a wide range of bias and signal power is required for accurate microwave circuit simulation. The large signal model of the Nortel HBT was developed by Zhang et al. [26] and this section is intended to provide an overview only. The Nortel model accounts for several key HBT effects, namely,

- thermal effects,
- dependence of the junction depletion capacitance and transit time on collector current, and,
- the variation of the effective electron barrier in an emitter heterojunction.

The model is shown in Figure 3.2 and is modified from the BJT Gummel-Poon Model. The thermal equivalent circuit, to calculate the temperature rise at the emitter junction, is shown in Figure 3.3.

The Nortel model is developed for normal HBT operation only, hence a breakdown mechanism is not included. The Kirk effect and substrate capacitance, which are believed to be unimportant for GaAs-based HBTs under normal operation conditions [26], are also ignored in the model.

Model Parameters The model parameters from Figures 3.2 and 3.3 are listed in Table 3.1, and further described as follows.

- The bias independent elements of the model account for those parasitics of the HBT that do not depend on the bias point of the device.
- Five diodes, D_E , D_{CI} , D_{CX} , D_L , and D_R , are used to account for the currents that flow in the two p-n junctions and those currents caused by parasitic effects. It is assumed

Figure 3.2: Large signal model for the Nortel HBT [26].

Figure 3.3: Thermal equivalent circuit for temperature rise at the Nortel HBT emitter junction [26].

Table 3.1: Nortel HBT model parameters.

Bias-independent elements	R_E	Emitter Resistance
	R_B	Base resistance
	R_C	Collector resistance
	L_E	Emitter lead inductance
	L_B	Base lead inductance
	L_C	Collector lead inductance
Diodes	D_E	Emitter junction
	D_{CI}	Intrinsic collector junction
	D_{CX}	Extrinsic collector junction
	D_L	Base-to-emitter leakage
	D_R	Base-to-emitter recombination
Current-controlled current source	I_{CE}	From Gummel-Poon model
Thermal effects	R_{th}	Junction equivalent thermal resistance
	C_{th}	Junction equivalent thermal capacitance
	$V_{J,T}$	Thermal reduction of emitter potential
	$I_{PE,T}$	Backward hole injection current
Voltage-controlled voltage source	V_B	Effective electron barrier variation
Capacitors	C_I	Base-emitter capacitance
	C_{CI}	Intrinsic base-collector capacitance
	C_{CX}	Extrinsic base-collector capacitance
	C_{Tr}	Transit time

that the currents represented by the diodes can be expressed as,

$$I_d = I_s \left(e^{\frac{V}{N V_T}} - 1 \right) \quad (3.4)$$

where I_s is the saturation current of a diode, N is the ideality factor of the diode current, and $V_T = \frac{kT}{q}$ is the thermal voltage at junction temperature T .

- The current controlled current source, I_{CE} from the Gummel-Poon model, and is defined as,

$$I_{CE} = \beta_F I_{EE} - \beta_R I_{CC} \quad (3.5)$$

where β_F and β_R are the forward and reverse current gains, and I_{EE} and I_{CC} are defined in Figure 3.2.

- GaAs substrates have a large thermal resistivity, therefore thermal effects such as self-heating are important considerations at large signal. The most significant effects from a rise in device temperature occur at the emitter junction, and hence using the equivalent circuit in Figure 3.3, the temperature rise at the emitter junction is calculated. Two major consequences from a rise in HBT temperature are the reduction of emitter junction built in potential, $V_{J,T}$, and the increase in the reverse hole injection current from the base to the emitter, $I_{PE,T}$. The former is responsible for the increase in the collector current under constant V_{BE} conditions, and the latter causes the decrease in the collector current under constant I_B conditions as V_{CE} increases [26].
- It is assumed that the electron tunneling through the abrupt emitter junction conduction band spike is bias dependant. This effect is modeled using a voltage controlled voltage source, V_B , defined as,

$$V_B(V_{BE}) = V_{BE} N_o \left(\frac{1}{N_o} - \frac{1}{N(V_{BE})} \right) \quad (3.6)$$

where N_o is the ideality factor without the barrier effect, and $N(V_{BE})$ is the ideality factor that varies with bias because of barrier variation.

- The capacitors are used to model the transit time (C_{Tr}) and charge storage mechanisms (C_I, C_{CI} and C_{CX}) at large signal.

3.3 Monolithic Microwave Integrated Circuits

By “Monolithic,” an integrated circuit approach is meant, where all active and passive circuit elements and interconnections are formed into the bulk, or onto the surface, of a semi-insulating substrate by some deposition scheme. Possible deposition schemes are epitaxy, ion implantation, sputtering, evaporation, diffusion, or a combination of these processes and others. The Monolithic Microwave Integrated Circuit, or MMIC, concept is not new. In fact it was attempted as early as 1964 [27], however it has not until the late 70’s that the aforementioned deposition schemes were advanced enough to fabricate high performance MMICs [25]. Additionally, advanced hybrid circuits of that time period, that is circuits where some elements are integrated and some are connected separately using bond wires, were becoming too complicated to be practical. Hence the MMIC approach quickly became very favourable.

Specifically, the advantages of MMIC technology are [27]:

- low cost when manufactured in large quantities,
- improved reliability and reproducibility,
- small size and weight,
- broadband performance, and,
- multiple function chips.

These advantages are countered by a number of disadvantages of MMIC technology, such as [27]:

- unfavourable device-to-chip area ratio,
- circuit tuning impractical,
- trouble-shooting difficult,
- potential difficulty of suppressing undesired RF coupling, and,
- high power limitations.

3.3.1 Foundry Fabrication Process

The concept of the HBT has been known for many years, however, it is only recently that fabrication technologies have become available to manufacture these devices. Northern Telecom Ltd. fabricated the Colpitts VCO in Ottawa, Canada, and complete details of the fabrication process are described in the *Nortel HBT Design Handbook* [28]. The Nortel HBT process was initially designed for digital applications, and Nortel is currently commercially fabricating HBTs for digital circuits. The use of HBT in analog circuits, such as the Colpitts VCO in this thesis, is still in the experimental stages.

The cross section the Nortel HBT is shown in Figure 3.4. Some Nortel HBT key electrical parameter values are given in Table 3.2. Some maximum ratings for the Nortel HBT are shown in Table 3.3.

The *Nortel HBT Design Handbook* [28] outlines eight thermal design rules to avoid performance degradation due to HBT self-heating. Some rules specify that the rating included in Table 3.3 not be exceeded. Others deal with mutual heating of multiple HBT devices, which are not used in the Colpitts VCO. The most important rule for the Colpitts VCO is to operate the HBT within its safe area of operation, as shown in Figure 3.5. The operating limits are defined by the maximum breakdown voltage, the maximum current density and the maximum power dissipation. The maximum current density and voltage are found in Table 3.3 and the

Figure 3.4: Nortel HBT cross-section [28].

Table 3.2: Nortel HBT electrical parameters [28].

HBT Parameter	Minimum	Mean	Maximum
F_t	50 GHz	60 GHz	n/a
F_{max}	60 GHz	75 GHz	n/a
g_m	20 mS	30 mS	40 mS
β	50	100	200
V_{be} turn on voltage	1.34 V	1.4 V	1.45 V
BV_{ebo}	3 V	6 V	n/a
BV_{ceo}	6 V	8 V	n/a
BV_{cbo}	10 V	16 V	n/a
MIM Capacitance	4.5 pF/35000 μm^2	4.9 pF/35000 μm^2	5.5 pF/35000 μm^2
NiCr sheet resistance	45 $\Omega/\square$	50 $\Omega/\square$	55 $\Omega/\square$

Table 3.3: Nortel HBT maximum ratings [28].

Parameter	Rating
Storage Temperature	-65 to 100°C
Junction Temperature	-55 to 125°C
Emitter current density	$5 \times 10^4 \text{ A} \cdot \text{cm}^{-2}$
Metal 1 (me1) current density	$3 \text{ mA}/\mu\text{m}$ width
Metal 2 (me2) current density	$5 \text{ mA}/\mu\text{m}$ width
NiCr current density	$0.5 \text{ mA}/\mu\text{m}$ width
MIM capacitor voltage	20 V

Figure 3.5: Nortel $6.5 \times 3 \mu\text{m}$ HBT thermal safe area of operation with 80°C maximum self-heating.

maximum power dissipation is defined as,

$$P_{max} = \frac{\Delta T_{self-heating}}{\theta_{Q,R}} \quad (3.7)$$

where $\Delta T_{self-heating}$ is the maximum allocated self-heating temperature rise and $\theta_{Q,R}$ is the device thermal impedance. The maximum allocated self-heating temperature is taken to be $80^{\circ}C$, consistent with an ambient environment of $35^{\circ}C$ and the device thermal impedance is $2.3^{\circ}C/mW$ from the *Nortel HBT Design Handbook* [28]. Therefore, the maximum power dissipation is $34.8 mW$.

3.4 Computer Aided Design

Computer aided design (CAD) is essential at Ka-Band frequencies. The accurate performance of circuit elements require the use of complex representations, and to trade-off the multitude of circuit performance parameters requires the use of powerful computers.

Circuit Simulation The circuit simulator used to design the Colpitts VCO is HP EEsof Libra version 6.1. By circuit simulator it is meant that the circuit elements are represented in Libra by ideal lumped element, frequency domain models.

For linear analysis, Libra generates frequency domain s-parameters for the circuit elements and then combines them to get an overall set of s-parameters for the entire circuit. Once the s-parameters for each element and those for the overall circuit are known the circuit performance parameters are calculated.

For non-linear analysis, Libra uses the harmonic balance technique. Harmonic balance is an iterative solution that assumes that sinusoidal excitation exists and that a suitably accurate solution can be represented by a finite number of Fourier harmonics. To that end, the circuit is divided into linear and non-linear elements, as seen in Figure 3.6. The linear and non-linear networks are connected by nodes. These node voltages take on a set of amplitudes

and phases for all harmonic frequencies, hence the solution is postulated in the form of a finite Fourier series. The linear sub-network is solved using the straightforward frequency domain s-parameter analysis. The non-linear sub-network is solved in the time domain, and is transformed to the frequency domain using Fourier techniques. Therefore, two frequency domain solutions for the node currents are known, one from the linear sub-network and one from the non-linear sub-network. According to Kirchhoff's Current Law (KCL) these currents must sum to zero at all nodes. The probability of achieving this balance on the first iteration is extremely small, hence an error function is formed. The error function is a measure of how much KCL is violated, and hence the voltage amplitudes and phases are adjusted for the next iteration. When the analysis converges, that is the error function is reduced to some small predefined value, then the resulting voltage amplitudes and phases approximate the steady-state solution.

Figure 3.6: Harmonic Balance technique setup.

Electromagnetic Simulation The electromagnetic simulator used is Sonnet EM. Sonnet EM is a planar 3-D analysis using the Method of Moments technique. Sonnet EM sub-sections the circuit into conductor squares and then evaluates the electric field everywhere due to the current in the sub-section square. Sonnet EM then repeats the same analysis for every

sub-section in the circuit. With the appropriate boundary conditions, Sonnet EM can solve geometries for s-parameters taking into account all possible coupling mechanisms, including dispersion, stray coupling, discontinuities, surface waves, moding, metalization loss, dielectric loss and radiation analysis.

A more scientific description of the analysis technique Sonnet EM employs is [29]:

The analytical full-wave numerical approach invokes a Galerkin method of moments techniques which develops the dyadic Green's function operator as a bi-dimensional infinite vectorial summation of homogeneous rectangular waveguide eigenfunctions resulting in a subsectional technique using roof-top expansion and testing functions and closely related to spectral domain techniques.

Chapter 4

Colpitts VCO Design

4.1 Introduction

The design of the Colpitts VCO is presented as follows. After the evaluation criteria and design constraints are outlined the design strategy is described. The Colpitts VCO is treated as a gain element and a feedback network, which is finally combined to form the Colpitts VCO.

4.2 Design Criteria

The design criteria are sub-divided into three categories: design constraints, design criteria evaluated and design criteria not evaluated. Each category is further outlined below.

4.2.1 Design Constraints

The design is constrained by fabrication technology, topology, and total MMIC area available. These design constraints are described in detail in the paragraphs to follow. For a summary of the design constraints, refer to Table 4.1.

Table 4.1: Design Constraints.

Criteria	Constraint
Fabrication technology	GaAs MMIC using a Nortel $6.5 \times 3 \mu m$ HBT
Topology	Colpitts type oscillator
Total MMIC area available	Maximum of $2500 \times 1150 \mu m$
Output Impedance	50Ω

Fabrication Technology The fabrication technology is a GaAs MMIC using a Northern Telecom $6.5 \times 3 \mu m$ HBT. No other foundries are possible because Northern Telecom is a participant in this research project.

Topology The topology is a Colpitts type oscillator. The topology is constrained to be a classic low frequency configuration because the focus of the thesis is to apply classical techniques to modern MMIC technology. Other classic topologies, such as Hartley, Clapp and Pierce, could have been equally chosen, but the Colpitts was chosen due to its high profile and considerable popularity.

Total MMIC Area Available The maximum allowable MMIC area for all circuits is $2500 \times 1150 \mu m$. All circuits and test structures to be built must fit within this area. This requirement is driven by cost, hence this area was the largest possible given budgetary constraints.

Output Impedance The output impedance of the Colpitts VCO must be 50Ω to allow for testing and integration into standard equipment and systems.

4.2.2 Criteria Evaluated

The VCO is evaluated based on numerous criteria, and these criteria are design goals. Each evaluation criteria is categorized as one of either physical, electrical or design robustness. Table 4.2 lists the evaluation criteria, and the related categories.

Table 4.2: Criteria evaluated.

Criteria		Evaluation
Physical	Size	VCO size
	Layout	Inter-element coupling
Electrical	Bias	Bias point
	Frequency	Oscillation frequency
	Power	Tuning bandwidth and sensitivity Power at fundamental and harmonics Power variation over tuning bandwidth
Design Robustness	Thermal	Foundry thermal limits
	Noise	Amplitude and phase noise

Physical

The evaluated physical aspects of the design are categorized by size and layout.

Size Size is a critical factor to evaluate. The Colpitts VCO is intended for higher order integration hence it must be as compact as possible. A clear benefit of a compact design is lower cost.

Layout The layout is examined to determine the effects, if any, on the VCO performance of inter-element coupling. The effect of some element mutual coupling is accounted for in the electromagnetic simulations, however the analysis is approximate. Therefore the measured performance deviates from the electromagnetic simulations, and part of this deviation is due to inter-element coupling.

Electrical

The evaluated electrical characteristics of the design are grouped by bias, frequency, and power.

Bias The bias point is critical to the VCO performance and hence the design bias point is evaluated. Deviation from the design bias point will affect the output power, frequency and thermal performance of the VCO.

Frequency The oscillation frequency is designed for 28 GHz which is consistent with current prototype wireless communications systems, such as LMCS. The measured oscillation frequency deviates from 28 GHz , and these deviations are examined. Furthermore, the VCO tuning bandwidth and sensitivity is evaluated. It is desirable that the tuning bandwidth be as large as possible. Finally, the linearity of the VCO tuning sensitivity is of interest for some possible VCO applications.

Power The fundamental output power must be as large as possible over the entire tuning bandwidth. A large output power is required since many wireless systems use the VCO as a local oscillator to drive a phase detector, an upconverter or a downconverter, or some combination of the three. If the output power is large enough, a buffer amplifier is not required, thereby reducing the VCO size and cost. Further, the harmonic output power must be low enough such that it does not interfere with the fundamental. Moreover, for both the fundamental and the harmonics, the output power must also not vary substantially over the tuning bandwidth.

Design Robustness

The design robustness is evaluated based on thermal issues and noise.

Thermal Issues The thermal limits as imposed by the foundry design rules [28] are adhered to at all times. Northern Telecom defines the thermal limitations by specifying the maximum currents and voltages on the device and the elements, and by stating the adequate spacing for heat dissipation.

Amplitude and Phase Noise A noise model is not available for the Northern Telecom HBT, hence noise is not a design requirement. Nevertheless, amplitude and phase noise performance is measured and evaluated. Low noise performance is highly desirable in wireless communications systems.

4.2.3 Criteria Not Evaluated

Numerous performance criteria are not evaluated completely. Outlined in Table 4.3, these items are not evaluated, primarily due to the inability to test or simulate them with sufficient accuracy.

Table 4.3: Criteria Not Evaluated.

Criteria		Evaluation
Electrical	Bias	Long and short term bias fluctuations
	Power	DC to RF efficiency
	Matching	Mismatch
	Transient	Load pull
Design Robustness		Start-up and settling time
		Long and short term post tuning drift
	Stability	Spurious signals
		Low frequency oscillations
		Long term frequency and power stability
	Yield	Tolerance of element variations
	Reliability	Operate within foundry constraints

Electrical

The electrical characteristics of the design that are not evaluated are: bias, power, matching, transient.

Bias Deviation from the design bias point affects the output power, noise and frequency of the VCO. The effects of long and short term bias fluctuations are not investigated, they are

instead left for future study.

Power The DC to RF efficiency is not evaluated. In most cases, the operating point for maximum power is not the same as for maximum efficiency. Thus, since the Colpitts VCO is designed for maximum power, the maximum DC to RF efficiency is not a design requirement. Further, the design techniques for maximizing efficiency involve manipulating the output harmonics. At Ka-Band, the harmonics are too high in frequency to be simulated or measured, and the ability of the device to produce predictable harmonic signals above its cutoff frequency is questionable. Efficiency is important if the Colpitts VCO is to be used in a portable unit of a wireless system because battery life is finite, however there are many applications where the Colpitts VCO would not be in a portable system and hence where efficiency is not as critical.

Matching The VCO is designed to be matched to $50\ \Omega$, however the variation of the actual impedance match will affect the frequency and output power. The variation of frequency and power with load is called load pull. The load pull is not investigated because of the difficulty of measuring the effects of a variable load. Additionally, the load mismatch must not reflect a large amount of power back to the device, and this effect is known as load tolerance. The load tolerance is not evaluated for the same reasons as the load pull.

Transient Time Analysis A transient, or time domain, parameter of interest is the amount of time the VCO takes to start-up and settle to within some prescribed value of the operating frequency. This parameter is known as the VCO settling time. Another transient parameter is the deviation from a new operating frequency after the VCO has been tuned to that new frequency. In the case of this post tuning drift, both short term, on the order $1\ \mu s$ to $1\ s$, and long term, on the order of $1\ s$ to $1\ hr$, are important considerations [52]. The settling time and post tuning drift are not evaluated due to the complexity of the required measurements.

It is common to analyze the start-up and steady-state performance of oscillators in the time domain. Frequency domain analysis is used exclusively in this thesis for a number of reasons. First, the transient time domain analysis in the Libra circuit simulator ignores all frequency dependent effects, clearly an unacceptable assumption. The convolution analysis, that is frequency domain analysis converted into the time domain using a FFT, will not converge for the Colpitts VCO. This convergence problem is most likely due to a bug in the software. The problem was raised in August, 1997, with technical support personnel with HP EEsof in California, USA, however, to date they have not been able to solve the problem. Hence, convolution analysis is not used to evaluate the transient performance of the Colpitts VCO.

Design Robustness

The design robustness is not evaluated in the areas of stability, yield and reliability.

Stability Spurious signals and low frequency oscillations must be suppressed in a well designed VCO. These effects are difficult to measure however, since they are often due to parasitics and other apparently random physical phenomena. Therefore, spurious signals and low frequency oscillations are not evaluated.

Additionally, long term stability, such as days or even weeks, of the VCO operating frequency and output power is another key VCO issue. This parameter is not evaluated and is left for future study.

Yield The yield, or number of VCOs that pass a given performance criteria, is a critical issue for a commercial VCO. The VCO must be designed such that the performance can tolerate variations in the element values that occur due to manufacturing or aging. Yield is not possible to evaluate because the VCO is a prototype design with a small number of

circuits manufactured. Should the Colpitts VCO ever be manufactured in large numbers, the yield can be evaluated.

Reliability The VCO must operate within foundry constraints at all times. The specific foundry issues have been presented in Section 3.3.1, and are designed for, but not measured. Further, the reliability of the topology is not considered a design constraint because the Colpitts topology is fixed.

4.3 Colpitts VCO Design

The section includes a literature review followed by a the design approach used to design the Colpitts VCO.

4.3.1 Literature Review

In addition to the approach used, many other design approaches exist in the literature, as seen in Table 4.4. These approaches are classified as small signal and large signal, or alternatively linear and non-linear device operation. Clearly, since oscillators operate at large signal at steady-state, the small signal design approaches are rough approximations. Nevertheless, small signal approaches are much easier than the large signal approaches, hence sometimes the small signal inaccuracy is compensated by ease of use.

All of the procedures in this section are presented for single frequency oscillators for simplicity. All of the procedures are easily applied to VCOs by applying the procedures at multiple frequencies, and iterating where required.

The design procedures included in Table 4.4 are evaluated in the following paragraphs.

Table 4.4: Existing design procedures.

Design Type	Author(s)	Year	Reference
Small signal	Wilson <i>et al.</i>	1989	[30]
	Boyles	1986	[31]
	Esdale <i>et al.</i>	1981	[32]
	Golio <i>et al.</i>	1978	[33]
	Khanna <i>et al.</i>	1981	[34]
	Martinez <i>et al.</i>	1992	[35]
	Krowne	1977	[36]
	Dougherty	1989	[37]
Large signal	Pucel <i>et al.</i>	1975	[27]
	Fukumoto <i>et al.</i>	1983	[38]
	Abe	1986	[39]
	Johnson	1979	[40]
	Wagner	1979	[41]
	Berini <i>et al.</i>	1994	[42]
	Hu <i>et al.</i>	1989	[43]
	Cheng <i>et al.</i>	1990	[44]
	Mitsui <i>et al.</i>	1977	[45]
	Gilmore <i>et al.</i>	1983	[46]
	Vehovec <i>et al.</i>	1968	[47]
	Rauscher	1980	[48]
	Kotzebue	1984	[49]
	Xuan <i>et al.</i>	1987	[53]
	Xuan <i>et al.</i>	1989	[50]

Existing Small Signal Design Procedures

The small signal design procedures cannot predict steady-state operation, hence only start-up is considered. Although all the small signal design procedures are essentially the same, Wilson *et al.* [30] presents perhaps the clearest small signal design procedure. Using the negative impedance model and the oscillatory conditions in equations 2.27 and 2.28 the oscillator is designed satisfied in eight clear steps. These steps not only ensure start-up at the correct frequency, but also address frequency stability and low noise. Wilson's work follows closely that of Boyles [31], which lays the foundation for Wilson's procedure by considering the oscillator as a reflection amplifier. With the negative impedance model, Boyles uses the

reflection coefficient expressions in Section 2.5.2. Furthermore, the reflection coefficient versions of Kurokawa's impedance oscillatory conditions are considered a design procedure in its own right, developed by Esdale *et al.*

Inherent in these small signal design procedures that use the negative impedance model, is the destabilization of the device. Golio *et al.* [33] presents some simple theory to maximize the device instability by way of maximizing the small signal device reflection coefficient. The expressions derived, however, assume a particular device configuration and this limits the generality, and hence usefulness, of the procedure.

All of the small signal procedures use three port transistors, de-stabilized by some sort of feedback to produce a two port for the negative impedance model. Khanna *et al.* [34] is credited with extending the two port diode oscillatory conditions to n -ports such as three port transistor and four port multiple gate input devices. The general n -port oscillatory condition is a useful design method if the device has more than three ports.

In 1992, Martinez *et al.* [35] proposed a small signal design procedure based on the so-called "circle function". The circle function is a flow graph representation of the loop gain. The procedure is limited to small signals since the flow graph is in terms of small signal s -parameters. Indeed, the difficulty to accurately measure large signal s -parameters is the only limitation to the circle function procedure. Perhaps the most useful feature of Martinez's analysis is that the circle function can be used on both the negative impedance and feedback oscillator models.

Many other authors have reported on the small signal design technique [36, 37], and while not adding much new, sometimes the different wording is useful. None of the aforementioned design procedures were used for the Colpitts VCO because of the limited performance prediction attained with small signal techniques. The Colpitts VCO is designed for output power at steady-state, thus requiring large signal techniques. Nevertheless, review of the small signal techniques is essential as most large signal design procedures incorporate the small signal

aspect to ensure start-up.

Existing Large Signal Design Procedures

The number of approaches to large signal design is quite varied as compared to small signal. This variation is due to the number of different simplifying assumptions in the large signal analyses.

Output Power vs. Input Power In 1975, Pucel *et al.* [27] proposed a method to predict the output power of a FET oscillator. They plotted P_{out} vs. P_{in} for the FET as an amplifier extended well into the non-linear region. Then, based on the feedback oscillator model, $P_{osc} = P_{out} - P_{in}$ is the power available from the device when operated as an oscillator, that is the net output power after some is fed back to sustain the oscillations. The maximum oscillator power is predicted by evaluating $\frac{\partial P_{osc}}{\partial P_{in}}$.

Reike Diagram Fukumoto *et al.* [38] develops an oscillator large signal model based on a polynomial expression for oscillator admittance using measured Reike diagrams. This model is a more accurate version of Van der Pol's landmark oscillator model [51].

Modeling Abe [39] assumes small signal s-parameters to generate an equivalent model for the transistor. The model is used to give greater insight into the non-linear device operation at large signal than the small signal s-parameters can. The maximum power is expressed as a function of these device parameters and the resulting voltages and currents are used to find the feedback network. This procedure is difficult to apply in general because a model must be derived. Further, gaining insight into nonlinear operation from the device model may be problematic, depending on the experience with the device.

Gain Saturation Approximation The so-called “Gain Saturation Approximation,” outlined by Johnson [40], is a more general non-linear device operation method than that of Abe. The Gain Saturation Approximation is simple approach for modeling the large signal power saturation and gain roll off phenomena observed in transistor oscillators. It is general because a model is not required, instead the device S_{21} is used. It is assumed that the device S_{21} is the only device parameter to change under large signal conditions, and hence an approximate S_{21} at large signal is chosen. This approximate large signal S_{21} is used in conjunction with the other small signal s-parameters to design the oscillator.

As with Abe’s modeling method, the difficulty with the Gain Saturation approach is approximating the device large signal behavior. Johnson provides some guidance in this regard by assuming an exponential device power saturation characteristic. Using this device saturation characteristic and the feedback oscillator model, an approximate large signal power gain, S_{21} is calculated.

Device Line Wagner’s [41] Device Line technique is also based on the large signal saturation of the device. Wagner considers the oscillator using the negative impedance model, and plots the device impedance as a function of power as the device saturates. The device impedance that corresponds to the maximum output power is used to satisfy the oscillatory conditions in equation 2.46. When the optimal device impedance is measured this technique yields accurate steady-state results. The device line approach is problematic, however, in that it does not address start-up, and often the device-circuit impedances for maximum power at steady-state do not guarantee start-up.

Research by Berini *et al.* [42] at the Ecole Polytechnique in Montreal has recently produced an extension to Wagner’s Device Line technique. Berini *et al.* have developed a technique for measuring large signal device lines at multiple harmonics, and can synthesize oscillators with specific harmonic output powers to a high degree of accuracy.

Volterra Series Volterra series [64] is a method that closely approximates the device impedance in the negative impedance model as a function of nonlinear currents. Cheng *et al.* [44] and Hu *et al.* [43] are examples of procedures that use Volterra series to predict the power output of an oscillator. Although the Volterra Series approach produces accurate results, it can quickly become mathematically complex and require a computer if a higher than third order solution is required. Additionally, Volterra Series is most suited to weakly nonlinear circuits, and for some oscillators this assumption may not be valid.

Large Signal S-Parameters Many oscillator design procedures involve small signal s-parameters. Indeed, a good number of large signal design procedures use small signal s-parameters to approximate large signal behavior. Hence, why not use large signal s-parameters to design oscillators?

Large signal s-parameters are not used often because they are difficult to measure accurately. Nevertheless, if large signal s-parameters can be accurately measured, they can be used to design the oscillator. Mitsui *et al.* [45] has demonstrated a successful oscillator design using large signal s-parameters. Power and efficiency are estimated by using drain and gate currents which are a function of the large signal s-parameters.

Gilmore *et al.* [46] have also used large signal s-parameters to design oscillators. Given the large signal s-parameters, the inverse s-parameters are used to synthesize the feedback network to deliver a given power to the load.

Voltage and Current Although a voltage or a current in the traditional sense cannot be measured at microwave frequencies, voltage and currents are invaluable mathematical tools to design large signal oscillators. If the device is characterized at large signal by voltages and currents, then traditional circuit analysis can be used to design the oscillator. This concept is further illustrated by the procedures in the following paragraphs.

As early as 1968, Vehovec [47] published an oscillator design procedure that used the

large signal device port voltage ratio to get equations for the feedback network. Further, he shows how to optimize for output maximum power and he addressed the uniqueness of the feedback network for a given maximum power output.

Further, Rauscher [48], in attempting to eliminate the error introduced by large signal s-parameters in the procedures of Johnson [40] and that of Mitsui [45], de-embeds the device dominant nonlinearities. The voltages and currents resulting from these dominant nonlinearities are then used to form the feedback network for the optimum oscillation conditions. To design the oscillator for optimum performance it is necessary to synthesize feedback network to:

1. Simultaneously provide optimum terminating impedance for the device and optimum feedback to maximize output power transfer to the load.
2. Yield a single, stable state of oscillation while suppressing spurious oscillations.

Rauscher uses Johnson's approximate saturation exponential relationship as a base to derive similar expressions for device voltages in terms of device parameters, and hence the optimum device voltage for maximum output power. With the device voltages known, the load line is known. It immediately follows that the device currents are known and everything that is needed to synthesize the feedback network is known.

As with Rauscher, Kotzebue [49] also modifies Johnson's approach by using device voltages and currents. Kotzebue uses a transistor configured as an amplifier to find the port voltages and currents that maximize added power. It is then assumed that these values are the same for an oscillator, and Kotzebue uses the values to modify the design equations used by Johnson. Kotzebue's new design equations are supposed to give a maximum output power, but no derivations are included.

In 1987, Xuan *et al.* [53] published the Double Device Line technique used to design the Ka-Band Colpitts VCO. In 1989, Xuan *et al.* [50] generalized their procedure to any

feedback topology and any number of harmonics. This procedure is used to design the Colpitts VCO, and is described in detail in Section 4.3, and is evaluated in Section 6.3.2.

4.3.2 Design Strategy

The design strategy used is summarized as follows, is known as the Double Device Line approach. The Colpitts topology is used and a heterojunction bipolar transistor is used as the gain element. The VCO is designed for maximum output power at the desired frequency of oscillation. Therefore, it is necessary to determine the steady-state RF operating conditions for maximum power output. This is accomplished by varying the currents and voltages on the input and output ports of the HBT until maximum net power is generated by the device. At steady-state, these voltages and currents are produced by the interaction of the device and feedback elements. Hence, these voltages and currents are used to determine the feedback network, such that the oscillator will oscillate at the desired frequency and generate maximum output power. The output power and the frequency of oscillation are the design goals; all other criteria are evaluated.

Initially, it is assumed that the frequency of oscillation depends only on the small signal loop phase, and that the steady-state output power depends only on the large signal steady-state loop gain. These assumptions are removed later in the design process. Hence, given the Colpitts topology and the HBT as the gain element, the design strategy is summarized as follows:

1. Select a quiescent operating point and design the DC bias network.
2. Find the steady-state RF operating conditions for maximum power output from the device at the desired frequency of oscillation.
3. Use these RF operating conditions to approximate the ideal feedback elements for stable steady-state oscillations.

4. Use the approximate ideal feedback elements to realize physical elements with interconnections.
5. At small signal, alter the feedback elements to give a 28 GHz frequency of oscillation and ensure oscillation start-up.
6. At large signal, ensure that steady-state oscillation is achieved.
7. At large signal, ensure that thermal restrictions are met. Alter the load as seen from the device to reduce the output power and hence the thermal stress on the device.
8. Design a matching network to match $50\ \Omega$ to the new device termination.
9. Examine the tuning of the VCO.
10. Layout the oscillator.
11. Simulate performance. Use the harmonic balance technique and electromagnetic simulations to analyze the start-up and the steady-state oscillator performance.
12. Build and test the oscillator.

The first two steps are used to analyze the gain element. The feedback network is understood using steps three through six. Steps seven, eight, nine and ten pertain to the oscillator as a whole. The last two steps are covered in Chapter 5.

4.3.3 Gain Element

Step 1: DC Analysis

The DC bias point for the HBT is the operating state which the HBT operates at in the absence of an input signal. The DC bias point, or quiescent point as it is sometimes referred

to, is visualized as the average operating state. The input signal determines the instantaneous operating state about this average.

The HBT used in the gain block is in common emitter (CE) configuration. Common emitter, where the potential of the emitter of the HBT is equal to RF ground, is used to exploit the large current and voltage gain of the CE configuration. Large gain is desirable to ensure that oscillation start-up occurs. The other possible transistor configurations, specifically the common base (CB) and the common collector (CC), could equally have been used. In fact, CB and CC configurations are more unstable than the common emitter. This higher degree of instability is because the internal device feedback capacitances between the input and output of the device are larger for CB and CC configurations. Rough designs using the CB and CC configurations as the gain element for the Colpitts VCO indicate that the feedback element values are not as convenient as the CE configuration, hence the CE configuration is chosen.

For the Colpitts VCO a Class A bias point is used. The bias class is where the quiescent point is located on the bias curves, and depending on the location, the device will act differently when stimulated by an RF signal. Class A is the simplest bias point, and is chosen in the centre of the possible bias curves. This bias point ensures that no clipping occurs when the output RF current and voltage vary around the quiescent point. Thus the device conducts for the entire RF signal period and maximum power is achieved. Other bias classes, such as B, C, and F, are used primarily for high efficiency applications, which is not the focus of the Colpitts VCO.

Specifically, the bias point is 3.65 mA collector current with 3 V on the collector at a base current of 0.04 μ A. The simulated DC bias curves for the $6.5 \times 3 \mu\text{m}$ Nortel HBT, with the quiescent point, is included as Figure 4.1. Simulations indicate that this bias point is the best compromise between output power and device operating constraints such as thermal restrictions.

The most direct way to bias the HBT at the desired quiescent point is accomplished

Figure 4.1: Simulated DC bias curves.

using a voltage and a current source connected to the HBT collector and base, respectively. The current source provides the most control when testing the Colpitts VCO, and is suitable should the VCO be integrated with other functions since current mirrors are standard in many integrated circuits. Should the VCO ever be used in a hybrid integrated circuit or some other system on its own however, the current source becomes impractical and more expensive. Additionally, the current source bias method does not provide good stability of the quiescent point over variations in HBT device parameters and temperature, as compared to a resistive bias network. The presence of an emitter resistor provides negative feedback, which helps to stabilize the transistor from temperature and device parameter variations [20].

The resistor bias network is shown in Figure 4.2. The simulated DC conditions for the Colpitts VCO are also shown in Figure 4.2.

Although the purpose of the bias network is to provide the correct DC conditions, the

Figure 4.2: Resistive bias network and DC conditions.

RF characteristics of the network must also be considered since RF signals are present in the VCO as well the DC conditions. The circuitry that provides the base and collector bias must allow DC to pass from the external source, while blocking the RF signal produced by the circuit. This is referred to as a bias choke.

The base choke is provided by the resistor network. The impedances of R_{b1} and R_{b2} are large enough (200Ω and 1000Ω , respectively) that the RF is effectively blocked from the base bias supply, and this is confirmed using simulations. This type of choke was dictated by the choice of a resistive bias network, and using the resistive network to set the DC bias point as well as to isolate the base bias supply reduced the number of elements in the VCO, thereby reducing the VCO size.

The collector choke is a lumped element spiral inductor. The inductor provides a large RF impedance and low DC impedance simultaneously, and hence the collector bias supply is isolated from the RF while passing the correct collector voltage. This performance is verified by simulation.

Both chokes used lumped instead of distributed elements. The electrical performance of the two methods is adequate, however, the lumped elements are far more compact. Indeed, the size of the Colpitts VCO is greatly reduced by using lumped elements wherever possible.

The quiescent point must be selected such that the thermal restrictions are adhered to. The thermal aspects of the bias network are covered, as with other thermal aspects, in Section 4.3.5.

Gain Element

The transistor, bias network and chokes are the necessary components of the gain element. Interconnecting transmission lines are included and are assumed to not affect the DC analysis. The assumption is valid because the loss of the short interconnecting transmission lines is low.

The final gain element is shown in Figure 4.3. Note that all simulation values given in this and other sections are based on this configuration.

Step 2: Determine Steady-State RF Conditions for Maximum Output Power

The biased device is embedded within feedback elements to interact such that particular RF conditions are imposed on the device for maximum power generation, at the desired frequency of oscillation. Note again that power, not noise, was the design driver because there is no noise model for the Nortel HBT.

To determine the RF conditions for maximum power, the following three-step algorithm was used [53].

Figure 4.3: Final gain element configuration for the Colpitts VCO.

1. Vary the input power, and hence the input voltage, at each gain element port and measure the resulting current.
2. From the input voltage and measured currents at each port, calculate the net output power.
3. Choose port voltages and currents for maximum net output power.

These steps are further outlined as follows.

In the first step, the biased device is stimulated by two power sources, one on the base and one on the collector, as seen in Figure 4.4. The emitter is grounded due to the common emitter configuration. The power into the device at each port is varied. By varying the power, the voltage seen by the device at the base and the collector is also varied. The current produced by the device is measured. In the second step, using the proper sign convention,

the voltage and current are used to calculate the net power generated by the device.

Figure 4.4: Setup to determine the RF operating conditions for maximum device power generation.

Referring to Figure 4.4, the net power generated by the device is given by

$$P_{net} = V_1 I_1^* + V_2 I_2^* \quad (4.1)$$

In the setup of Figure 4.4 the device is stimulated by sources. In the Colpitts oscillator the device is stimulated by voltages and currents generated by the interaction between the device and the feedback network. From the device point of view, however, these two situations are equivalent, that is the sources can be used to simulate the voltages and currents that the device-feedback network interaction produces in the Colpitts oscillator. Hence, equation 4.1 is used to calculate the net output power generated by the device at any harmonic, given that the appropriate voltage and current harmonics are used.

In the third and final step, the power supplies are varied, hence varying the stimulating voltage. The currents are measured for each case and the net output power generated is calculated. This procedure is repeated for a number of different supply powers until the RF conditions that produce the greatest net device output power are known. The voltages and currents at this optimum power point are those which the device-feedback network interaction

must produce to stimulate the device such that it produces maximum power output. Using these voltages and currents the feedback network is designed.

Before the feedback network is synthesized, the bias point is reconsidered since the algorithm to find the optimal RF conditions for maximum output power is performed at one bias point only. Previously, in Section 4.3.3 the bias point was selected primarily based on DC and thermal conditions. The large signal power generation capability of the bias point was only approximated by the selection of a Class A bias point. Hence, to determine if the bias point previously selected is indeed optimum for maximum output power, the aforementioned algorithm is performed at multiple bias points. As previously mentioned in Section 4.3.3, the bias point in that section is optimum, considering all relevant requirements and restrictions.

As with the bias point, the thermal restrictions of the device are also considered when the optimum RF conditions are found. The thermal aspects of the design are considered in Section 4.3.5.

The final steady-state RF operating conditions for maximum output power are shown in Table 4.5.

Table 4.5: Optimum RF steady-state conditions for maximum output power.

Parameter	Value
V_1	$0.2461 + j0.1244 \text{ V}$
V_2	$-1.3241 + j1.0178 \text{ V}$
I_1	$4.3692 + j10.5996 \text{ mA}$
I_2	$-1.1328 - j8.5223 \text{ mA}$

4.3.4 Feedback Network

Once the gain element DC and RF steady-state operating conditions for maximum output power are determined, the feedback network is synthesized. The feedback network is designed using the concept of loop gain, as outlined in Section 2.4.1.

Step 3: Feedback Element Synthesis

The voltages and currents that generate the maximum output power from the device are used to synthesize the feedback network elements. The device-feedback network interaction must produce the same RF conditions at the device ports as the power supplies did. Using Kirchoff's Laws the feedback elements are derived from the optimized voltages and currents.

The algorithm to synthesize the feedback network elements is as follows [53].

1. Consider the Colpitts oscillator to be arranged as a feedback model oscillator as shown in Figure 2.5.
2. Given the optimum RF operating voltages and currents are known, Kirchoff's Laws are used to synthesize the feedback network, as seen in Figure 4.5. This amounts to solving the system of linear equations 4.2,

$$\begin{pmatrix} 0 & -\Im(V_1) & 0 & -(\Im(V_1) - \Im(V_2)) \\ 0 & \Re(V_1) & 0 & \Re(V_1) - \Re(V_2) \\ \Re(V_2) & 0 & -\Im(V_2) & -(\Im(V_2) - \Im(V_1)) \\ \Im(V_2) & 0 & \Re(V_2) & \Re(V_2) - \Re(V_1) \end{pmatrix} \begin{pmatrix} G_2 \\ B_1 \\ B_2 \\ B_3 \end{pmatrix} = - \begin{pmatrix} \Re(I_1) + \Re(V_1)G_1 + (\Re(V_1) - \Re(V_2))G_3 \\ \Im(I_1) + \Im(V_1)G_1 + (\Im(V_1) - \Im(V_2))G_3 \\ \Re(I_2) + (\Re(V_2) - \Re(V_1))G_3 \\ \Im(I_2) + (\Im(V_2) - \Im(V_1))G_3 \end{pmatrix} \quad (4.2)$$

which is derived below. Note that in equation 4.2 $\Re$ denotes real part and $\Im$ denotes imaginary part and there are six unknowns but only four equations. This issue is easily resolved by the Colpitts topology, since two of the feedback elements are purely reactive. Hence, $G_1 = 0$ and $G_3 = 0$, and the system is now readily solved with four equations and four unknowns. The values of G_1 and G_3 represent the conductor losses in the reactive elements, and hence they are assumed to be lossless to simplify the algorithm.

Figure 4.5: KCL nodes in the feedback network.

Equation 4.2 is now derived, and may be skipped by proceeding directly to page 73. Kirchoff's Current Law (KCL) is solved at the two nodes in the feedback network, I and II, as shown in Figure 4.5. The resulting two equations can be considered as four, two real and two imaginary, since all the parameters are complex. Hence, there are four equations with six variables, G_1 , B_1 , G_2 , B_2 , G_3 , and B_3 . To correspond to the Colpitts feedback network, the conductances of elements 1 and 3 are chosen, since they represent the conductor loss and are easily estimated. Therefore, G_1 and G_3 are assumed known.

KCL at node I gives,

$$I_1 + V_1 Y_1 + (V_1 - V_2) Y_3 = 0 \quad (4.3)$$

which, when written explicitly in real and imaginary parts, is equivalent to,

$$\Re\{I_1\} + \Re\{V_1\}G_1 - \Im\{V_1\}B_1 + (\Re\{V_1\} - \Re\{V_2\})G_3 - (\Im\{V_1\} - \Im\{V_2\})B_3 = 0 \quad (4.4)$$

and

$$\Im\{I_1\} + \Re\{V_1\}B_1 + \Im\{V_1\}G_1 + (\Re\{V_1\} - \Re\{V_2\})B_3 + (\Im\{V_1\} - \Im\{V_2\})G_3 = 0 \quad (4.5)$$

In equations 4.4 and 4.5, $\Re$ denotes the real part and $\Im$ denotes the imaginary part.

Similarly, KCL at node II gives,

$$I_2 + V_2 Y_2 + (V_2 - V - 1) Y_3 = 0 \quad (4.6)$$

which, when written explicitly in real and imaginary parts, is equivalent to.

$$\Re\{I_2\} + \Re\{V_2\}G_2 - \Im\{V_2\}B_2 + (\Re\{V_2\} - \Re\{V_1\})G_3 - (\Im\{V_2\} - \Im\{V_1\})B_3 = 0 \quad (4.7)$$

and

$$\Im\{I_2\} + \Re\{V_2\}B_2 + \Im\{V_2\}G_2 + (\Re\{V_2\} - \Re\{V_1\})B_3 + (\Im\{V_2\} - \Im\{V_1\})G_3 = 0 \quad (4.8)$$

Also in equations 4.7 and 4.8, $\Re$ denotes the real part and $\Im$ denotes the imaginary part.

Combining and rearranging equations 4.4, 4.5, 4.7, and 4.8 gives the matrix in equation 4.2 which is used to solve the system of equations.

The algorithm to solve equation 4.2 is implemented using Matlab, a mathematical software package. The algorithm interface is defined as,

- Input $\rightarrow$ optimized complex RF voltages and currents at gain element ports, given in Table 4.5
- Output $\rightarrow$ optimum ideal feedback elements for Colpitts topology, and these results are given in Table 4.6

The Matlab code used is as follows.

```
clear;

v1=0.2461+j*0.1244;
v2=-1.3241+j*1.0178;
i1=4.3692e-3+j*10.5996e-3;
i2=-1.1328e-3-j*8.5223e-3;

g1=0;
g3=0;
```

```

freq=28e9;
omega=2*pi*freq;

A(1,1)=0;
A(1,2)=-1*imag(v1);
A(1,3)=0;
A(1,4)=-1*(imag(v1)-imag(v2));
A(2,1)=0;
A(2,2)=real(v1);
A(2,3)=0;
A(2,4)=real(v1)-real(v2);
A(3,1)=real(v2);
A(3,2)=0;
A(3,3)=-1*imag(v2);
A(3,4)=-1*(imag(v2)-imag(v1));
A(4,1)=imag(v2);
A(4,2)=0;
A(4,3)=real(v2);
A(4,4)=real(v2)-real(v1);

Y(1,1)=real(i1)+real(v1)*g1+(real(v1)-real(v2))*g3;
Y(2,1)=imag(i1)+imag(v1)*g1+(imag(v1)-imag(v2))*g3;
Y(3,1)=real(i2)+(real(v2)-real(v1))*g3;
Y(4,1)=imag(i2)+(imag(v2)-imag(v1))*g3;

B=-1*Y;
X=inv(A)*B

g2=X(1,1);
b1=X(2,1);
b2=X(3,1);
b3=X(4,1);

R2=1/g2
C2=b2/omega
C1=b1/omega
L=1/(b3*omega)

```

Step 4: Physical Realization of Feedback Elements and Interconnects

The ideal values in Table 4.6 are realized as physical lumped elements. There are no available dimensions for realizable lumped elements at 28 GHz using the Nortel fabrication process.

Table 4.6: Synthesized feedback elements for maximum output power.

Parameter	Value
R_2	583Ω
C_2	$0.00977 fF$
C_1	$0.0357 fF$
L	$0.985 nH$

Therefore, the dimensions for the $28 GHz$ lumped elements were initially estimated based on available lumped element dimensions used by Nortel at $20 GHz$. These $20 GHz$ -based dimensions are optimized at $28 GHz$ using the circuit simulator. The physical dimensions of the lumped elements are provided as Table 4.7. Note that the parameter values in Table 4.7 are different than those in Table 4.6 because of realizable dimensions, transmission line interconnects, parasitics etc. These effects are outlined in the paragraphs to follow, however the final parameter values are introduced here. Finally, simulations indicate the resonant frequency of the realizable inductor to be higher than the frequencies of interest.

Table 4.7: Realizable feedback lumped elements.

Parameter	Value	Final Dimensions (if required)
R_2	583Ω	n/a
C_2	$0.0155 fF$	parasitic of lumped inductor
C_1	$0.0441 fF$	parasitic of lumped inductor
L	$0.617 nH$	3.25 turns at $5 \mu m$ wide with a $5 \mu m$ spacing

There are no dimensions for the resistor, R_2 , since the output 50Ω load of the VCO is matched to the value required by R_2 . The output matching is described in Step 8. There are also no dimensions for the capacitors, C_1 and C_2 , since the capacitance values indicate they can be considered as parasitics of the inductor. The use of the capacitors as parasitics is described in Step 5 and Figure 4.6.

In the final VCO design, the lumped elements are interconnected with transmission lines, which require the element dimensions to be further adjusted. These adjustments are outlined in

the sections to follow. Note that the values in Table 4.7 include the effects of interconnecting transmission lines.

Step 5: Start-up and Frequency of Oscillation

With the feedback elements defined, the small signal loop phase is used to set the frequency of oscillation. Additionally, the small signal loop gain is used to guarantee start-up of oscillations. Once again, the frequency of oscillation is assumed to be a small signal parameter only, hence not dependent on power, and this assumption is removed later in the analysis. For small signal analysis an input power level of -50 dBm is assumed. This was determined to be an appropriate level since simulations indicate that the s-parameters of the device do not vary with power until the power reaches much higher levels than -50 dBm, such as greater than -5 dBm .

The start-up frequency for the oscillator is the frequency where the small signal loop phase is zero. The feedback elements obtained from the Matlab code cause the loop phase to be zero close to, but not exactly at, the desired frequency of oscillation. This deviation is due to the effect of the transmission lines between the feedback elements. These interconnections are not accounted for by the synthesis technique given by equation 4.2. As outlined in Section 2.3, it is assumed that the inductor of the feedback network is the element that most affects the frequency of oscillation. The inductor dimensions are optimized to shift the start-up frequency to the desired frequency of oscillation.

The small signal loop gain is greater than one to ensure start-up. It is assumed that the capacitors of the feedback network are responsible for destabilizing the device, and hence causing oscillations to build up, as outlined in Section 2.3. The values of the capacitors obtained from evaluating equation 4.2 indicate that the parasitics of the inductor can be used to destabilize the device. Lumped element, spiral inductors are modeled as shown in Figure 4.6. This inductor model is identical to the feedback network topology solved

by equation 4.2, if the inductor loss modeled by R is taken to be zero and the feedback capacitance, C_3 , is neglected, which is often done for simplicity. The capacitor C_3 is not significant in the inductor model, so the assumption does not affect the model significantly.

Figure 4.6: Model for a lumped inductor [54].

Therefore, since the parasitics of the inductor can be used to destabilize the device, two capacitors and their connections to ground through via holes are eliminated. This greatly reduces the VCO complexity and size. The feedback network capacitors would ordinarily be adjusted to further destabilize the device. However, since in this Colpitts oscillator the capacitors are the parasitics of the inductor, the capacitors cannot be optimized to further destabilize the device.

The small signal loop gain and phase for the Colpitts VCO is shown in Figure 4.7.

Step 6: Large Signal Loop Gain

With the start-up of oscillation assured at the correct frequency, the large signal loop gain is used to predict the steady-state output power of the oscillator. As the power circulating in the feedback loop grows, the device's ability to amplify it decreases. Therefore, the device saturates and the loop gain falls to one. This represents the steady-state operation point of the oscillator. Hence, for the analysis, the power in the loop is increased until the loop gain falls

Figure 4.7: Small signal loop analysis for the Colpitts oscillator.

to one, and that power is the steady-state operating power required to sustain oscillations. The output at the load is found by subtracting this sustaining power from the total power generated by the device.

The loop phase is also measured at this point as it is not zero, since the assumption that the frequency is independent of power is an approximation only. Therefore, the predicted small signal start-up frequency is not the steady-state operating frequency, however, the deviation is slight. The more the loop phase deviates from zero as the device saturates, the more the steady-state operating frequency deviates from the frequency predicted by the small signal analysis. Therefore the small signal loop phase analysis is re-performed and the small signal loop phase altered slightly, such that when the large signal analysis is performed the shift in frequency results in the desired frequency of oscillation. Indeed, at a loop gain equal to one, the loop phase should be made equal to zero.

The large signal loop gain and phase for the Colpitts VCO are shown in Figure 4.8.

Figure 4.8: Large signal loop analysis for the Colpitts oscillator.

Feedback Loop

To complete the feedback loop, a DC blocking capacitor is necessary to prevent the collector bias from entering the base and the base bias from entering the collector. Often, a large capacitor that does not affect the other elements is used. To keep the VCO size to a minimum, a different approach is taken. A small capacitor is used, specifically $30\ \mu\text{m} \times 30\ \mu\text{m}$ yielding a capacitance value of $0.126\ \text{pF}$. The feedback inductor is altered to compensate for the addition of the capacitor. The inductor dimensions in Table 4.7 are those after the adjustment for the DC blocking capacitor.

The final feedback loop, with interconnecting transmission lines, used to generate the simulations and layout, is shown in Figure 4.9. Note that the load is not shown to keep with the feedback model topology in Figure 2.3.

Figure 4.9: Final feedback loop configuration for the Colpitts VCO.

4.3.5 Colpitts VCO

With the large and small signal loop analysis complete, the Colpitts VCO is analyzed as a whole.

Step 7: Thermal Analysis

The Colpitts VCO adheres to the foundry thermal restrictions since the HBT junction temperature is the limiting performance factor for the device. While the electrical performance is limited by the collector current density and the collector-base pn-junction reverse breakdown voltage, these limits occur at junction temperatures higher than permitted in the HBT. To ensure the maximum junction temperature of 125°C given in Table 3.2 is not exceeded, the quiescent point, and the instantaneous RF signal swing around the quiescent point, must remain inside the safe area of operation in Figure 3.5 at all times.

The bias point of $I_b = 0.04$ mA and $V_c = 3$ V is in the centre of the safe area of operation to allow maximum instantaneous RF signal swing. An emitter resistor is used to provide feedback to increase the temperature stability.

To analyze the instantaneous RF signal swing, a dynamic load line (DLL) analysis is used. A DLL superimposes the RF output current (collector current) as a function of the RF output

voltage (collector voltage) on the DC bias curves for the device. The RF output current as a function of RF output voltage is plotted at each point in time over one cycle and traces out an ellipse on the DC bias curves. The DLL shows the complete RF current and voltage values for a given cycle. The thermal limitations for the device are plotted on the DC bias curves giving a region of safe operation. Therefore, the RF current and voltage traces remain within the region of safe operation for the thermal requirements to be met.

A DLL is performed on the Colpitts VCO and the thermal performance examined. After the first iteration the maximum junction temperature was exceeded. To reduce the junction temperature for the device to operate within its safe area of operation, the device must saturate at a lower power level. Hence, to reduce the power in the oscillator, the load resistance as seen by the device is reduced. The loop gain saturation, and hence output power, is mostly affected by the load resistance as seen by the device, as outlined in Section 2.3. Hence, to ensure the thermal requirements are met, the load resistance is changed such that saturation occurs before the signal builds up enough to damage the device. After a small number of iterations, the load resistance was reduced from 583Ω to 100Ω . The resistance could have also been increased, however simulations indicate that a larger change would have been required. Additionally, the VCO output impedance of 50Ω is matched to this device resistance, hence a value closer to 50Ω is desirable.

The DLL analysis resulting from this reduced load resistance is shown in Figure 4.10. The thermal restrictions are met since the instantaneous RF signal swing is within the safe area of operation for the HBT at all times.

The change in load resistance has a small effect on the loop gain and phase, hence these parameters are re-analyzed for frequency of oscillation and output power. The results in Figures 4.7 and 4.8 include the reduced load resistance.

Figure 4.10: DLL for the Colpitts VCO.

Step 8: Output Matching Network

The output matching network is designed to transform the required $50 \, \Omega$ VCO output impedance to $100 \, \Omega$, the optimum load resistance as seen by the device to meet the thermal requirements. This transformation is seen on the Smith Chart in Figure 4.11. The output matching network uses a series capacitor and a shunt inductor to achieve the match.

The specific output matching topology is chosen because in this configuration the two elements, the series capacitor and shunt inductor, provide dual functions. The series capacitor also is a DC blocking capacitor, preventing the DC bias from exiting the VCO. The shunt inductor is also the collector choke. By using the output matching elements for multiple functions at least $0.062 \, \mu m^2$, or 34%, of extra space is conserved by not including two separate output matching elements.

The output matching network is shown in Figure 4.12. Circuit simulations indicate a return loss of at least $-15 \, dB$ and an insertion loss greater than $-0.32 \, dB$ over $6 \, GHz$,

Figure 4.11: Colpitts VCO output matching.

Figure 4.12: Final output matching network for the Colpitts VCO.

Figure 4.13: Colpitts VCO simulated output match.

centred at 28 GHz, as seen in Figure 4.13.

Step 9: Examine VCO Tuning

The VCO tuning is accomplished by varying the collector bias voltage. This tuning method is known as frequency pushing. As the collector bias is tuned, the VCO must meet the thermal requirements at all times. Hence, the DLL analysis must also allow for the tuning of the oscillator within the safe area of operation. The RF output trace will shift as the collector bias voltage is changed. The safe region of operation must not be violated for all tuning voltages. According to Figure 4.10 the RF swing is narrow enough that a shift in collector tuning voltage of about 1 V will not violate the thermal safe area of operation.

Output power optimization over the entire tuning bandwidth is not necessary. Simulations indicate that the power variation with bias is not large for collector bias above 2.5 V, which is another reason why a nominal collector bias of 3 V is chosen for the HBT. Similarly, the

output matching bandwidth is maximized to ensure a good match is obtained over the tuning range.

Colpitts VCO Schematic

The final circuit schematic of the Colpitts VCO is seen in Figure 4.14.

Step 10: Layout and Fabrication

The oscillator is laid out in microstrip, adhering to the foundry design rules [28], and converted to GDSII format. The total Colpitts VCO size is $600\ \mu\text{m} \times 300\ \mu\text{m}$, or $0.18\ \text{mm}^2$. Microstrip is used because the models in the circuit simulator are most accurate for microstrip. The layout is shown in Figure 4.15, and the layout rationale is as follows.

The RF probe pads for testing are coplanar with a ground-signal-ground configuration. Two via holes are required for the two ground pads. These two via holes are used as the circuit grounds, and the circuit requires a minimum of two via holes. The via holes are cap-on-vias, that is capacitors with the bottom plate as the top of the via hole. The cap-on-vias are constructed to occupy less space.

The bias pads for the base and collector are at the edges, and opposite sides, of the chip area because this is how the prober in the lab is configured. The transmission lines are just wide enough to meet the current density requirements, as are the bias resistors. The bias resistors are placed in close proximity, however mutual coupling is predicted by the electromagnetic simulations.

The feedback network, output capacitor and choke inductor are placed at right angles from each other and from the device and bias network to increase mutual isolation. The choke is placed close to the via hole to reduce the VCO size and coupling resulting from this arrangement is simulated using electromagnetic software. The output capacitor is placed between the two via holes to further reduce the VCO size. Any effects from the grounds in

Figure 4.14: Colpitts VCO final schematic.

Figure 4.15: Colpitts VCO layout and microscope exposure.

close proximity are simulated by the electromagnetic analysis.

The usual rule of thumb of one substrate thickness ($100\mu m$) element separation is violated in the Colpitts VCO. This is done to attempt to reduce the size of the VCO as much as possible.

The effects of interconnections and packaging are not included as the Colpitts VCO is intended to be used as a block in a higher order integration MMIC.

The Nortel Advanced Technology Laboratory in Ottawa, Canada, fabricated the Colpitts VCO. A picture of the circuit using a microscope exposure is shown in Figure 4.15.

Chapter 5

Colpitts VCO Performance

5.1 Introduction

The Colpitts VCO performance is evaluated in the following sections based primarily on simulated and measured performance. The simulated performance is further evaluated in terms of electrical, thermal and design robustness performance parameters, using both circuit and electromagnetic simulations.

5.2 Colpitts VCO Simulated Performance

5.2.1 Simulated Electrical Performance

Steady-state Performance

The steady-state performance of the Colpitts VCO is simulated in the frequency domain using the harmonic balance technique. The resulting power spectrum is shown in Figure 5.1. The simulated fundamental output power is -4.74 dBm at 28.04 GHz and the simulated second harmonic output is -32.04 dBm at 56.16 GHz . Simulation results indicate that the other

harmonics are below -32.04 dBm , however since all harmonics above the second are above the device cutoff frequency, the simulated results for the third and higher harmonics are highly suspect.

Figure 5.1: Colpitts VCO simulated power spectrum at the nominal operating bias point.

Using the collector bias voltage, V_{ce} , to tune the VCO, the simulated tuning range for the Colpitts VCO is 350 MHz , as seen in Figure 5.2. Over this tuning range the fundamental output power variation is $\pm 0.635\text{ dBm}$, from the value at the centre frequency. Similarly, the second harmonic output power variation is $\pm 1.86\text{ dBm}$.

Start-up Performance

The start-up is guaranteed by Figure 4.7. It is common practice for time domain analysis to be used to simulated the oscillator start-up, however for the reasons outlined in Section 4.2.3, this is not done.

Figure 5.2: Colpitts VCO simulated oscillation frequency and output power tuning range.

5.2.2 Simulated Thermal Performance

In addition to the DLL technique used to design for thermal performance, other aspects of the Colpitts VCO are analyzed for thermal performance. Specifically, the simulated steady-state collector current waveform is compared to the thermal limits. This waveform is shown in Figures 5.3, and clearly does not exceed the thermal limit of 9.75 mA , which is calculated based on the maximum emitter current density in Table 3.3.

Additionally, the junction voltage is represented within the HBT model as a voltage at node T , relative to node T_o , in Figure 3.3. This voltage is simulated and converted to a junction temperature according to,

$$T_{\text{junction}} - T_o = V_{T,T_o} \quad (5.1)$$

where $T_o = 27^\circ\text{C}$. A voltage of 10.8 V is simulated, giving a simulated junction temperature at steady-state of 37.8°C , which is less than the maximum of 125°C given in Table 3.3.

The thermal performance of the Colpitts VCO due to the component location is simulated

Figure 5.3: Colpitts VCO simulated collector current waveform for thermal analysis.

because the circuit simulator does not account for inter-element coupling.

5.2.3 Simulated Design Robustness

Noise Analysis

The noise performance of the Colpitts VCO is not simulated because a noise model is not available for the Nortel HBT.

5.2.4 Electromagnetic Analysis

The passive elements of the Colpitts VCO are simulated using electromagnetic CAD tools, that solve Maxwell's equations, and are more accurate than the circuit simulator models. It is not possible to analyze the entire circuit; the memory requirements would be of the order of tens of gigabits and the computation time would be as much as a week. Therefore, the circuit

was sectioned into five pieces, as shown in Figure 5.4. It is assumed that the electromagnetic coupling between these sections is low. The basis for this assumption is somewhat heuristic, however, the relative placement of each section is likely to produce a low amount of coupling as compared to other possible groupings. Moreover, within each section, the elements are placed close to each other to reduce the overall size of the VCO. Although this close inter-element spacing is not normally implemented in MMIC design practice, it is acceptable since the sections are analyzed electromagnetically. Each section is briefly described below.

Feedback Network The feedback network is analyzed because the parasitics are necessary to destabilize the device.

Output Capacitor The output capacitor is close to two large ground planes, hence electromagnetic simulation is required.

Choke Inductor The choke inductor is affected by the nearby via hole and is therefore analyzed.

DC Bias Network The distributed effects and mutual coupling of the bias resistors are analyzed, as is the proximity of the via hole ground plane.

Cap-on-via The cap-on-via is electromagnetically simulated since the circuit simulator model is highly inaccurate.

The substrate layers used in the electromagnetic analysis are the GaAs substrate and the major dielectric layers between the three metal layers used; base metal, metal one and metal two. The layers are shown in Figure 5.5. The number of dielectric layers used increases the computation time, however, in this case the extra accuracy obtained from five layers is important to evaluate the accuracy of the electromagnetic analysis. The layers are bounded

Figure 5.4: Electromagnetic analysis sections of the Colpitts VCO.

by ground planes on all sides, except for the top, which is defined as free space with an impedance of 377Ω . The side walls are nominally specified at least one substrate thickness away from any simulated element so as to not affect the simulation.

Figure 5.5: Sonnet EM layers.

The results of the electromagnetic analysis are used in place of the circuit simulator models and the Colpitts VCO performance is re-simulated. The electromagnetic results are more accurate than the models, hence the Colpitts VCO performance simulated with electromagnetic results is more accurate as compared to the circuit simulator. With the electromagnetic results, the simulated frequency of oscillation is 27.33 GHz , with an output power of -1.27 dBm . The simulated second harmonic occurs at 54.66 GHz , with a power level of -21.51 dBm . These results are seen in the simulated power spectrum for the Colpitts VCO using the electromagnetic results, shown in Figure 5.6.

The Colpitts VCO with electromagnetic data has a tuning bandwidth of 390 MHz , and a

Figure 5.6: Simulated power spectrum of the Colpitts VCO using electromagnetic analysis results.

fundamental and second harmonic output power tuning bandwidth variation of $\pm 0.205 \text{ dBm}$ and $\pm 0.880 \text{ dBm}$, respectively, as seen in Figure 5.7.

Using the electromagnetic data, the start-up performance is simulated by ensuring the small signal loop gain, as is shown in Figures 5.8, is greater than one.

For the thermal aspects of the design, the electromagnetic results are used to simulate the collector current, given in Figure 5.9. Indeed, the collector current is less than 9.75 mA , which is calculated based on the maximum emitter current density in Table 3.3.

The noise is addressed using electromagnetic simulations by considering the steepness of the phase versus frequency relationship in Figure 5.8, according to the theory outlined in Section 2.4.1.

Finally, it should be noted that the Colpitts VCO circuit elements are not altered based on the discrepancy between the circuit and electromagnetic simulation because this difference is itself one of the items being investigated.

Figure 5.7: Simulated tuning bandwidth of the Colpitts VCO using electromagnetic analysis results.

Figure 5.8: Simulated start-up using electromagnetic small signal loop gain.

Figure 5.9: Simulated collector current of the Colpitts VCO using electromagnetic analysis results.

5.3 Colpitts VCO Measured Performance

Initially, the wafer was probed to measure the VCO performance. This testing was unsuccessful, however, due to short circuits in the DC bias path for both the base and the collector. This situation is shown graphically in Figure 5.10. The metal connection between the RF ground probe pad and the cap-on-via is to the top plate instead of the bottom plate. The cap-on-via must isolate the DC and the RF on the circuit side from the pad side, however, due to a connectivity error, this is not possible. Hence, the wafer could not be successfully probed.

To alleviate the problem, the VCO was cut out of the wafer and put in a test jig. Microstrip transmission lines with alumina dielectric were ribbon bonded to the signal probe pad of the VCO. In the test jig, the RF ground is the carrier and the misconnected RF ground probe pads are not used. The test jig carrier schematic is shown in Figure 5.11, the jig assembly

Figure 5.10: Short circuits in the DC bias caused by the RF probe pads.

Figure 5.11: Colpitts VCO test jig carrier schematic.

schematic in Figure 5.12, and a photograph of the completed jig is included as Figure 5.13. Note that the jig includes three test structures in between two versions of the VCO, as seen in Figure 5.12.

The jig carrier is brass with a thin layer of gold plating to enhance the electrical continuity between the VCO ground plane and the alumina ground plane. Ferrite beads and 1000 pF de-coupling capacitors are used to prevent low frequency oscillations. To be effective at Ka-Band, the HBT possesses extremely high low frequency gain and hence is prone to low frequency oscillations. During the testing it was found experimentally that adding a further $0.1\text{ }\mu\text{F}$ to each de-coupling capacitor improved the VCO stability.

The jig is placed in the Wiltron Universal Test Fixture, which uses K-connectors up to 40 GHz , and connected to the HP8565 Spectrum Analyzer using semi-rigid cable. The spectrum analyzer is connected via HP-IB to a Pentium computer running HP VEE test

Figure 5.12: Colpitts VCO test jig assembly schematic (two VCOs and three test structures).

Figure 5.13: Colpitts VCO in test jig.

software. The computer is used to control the spectrum analyzer and obtain soft copies of test data. The test set-up is shown in Figure 5.14.

Figure 5.14: The test set-up for the Colpitts VCO.

At the design bias point of $V_b = 3\text{ V}$ and $V_c = 3\text{ V}$, the Colpitts VCO jig oscillates at a frequency of 33.8257 GHz with a power of -0.92 dBm . The power is measured using a HP 437B Power Meter, and the purity of spectral content is verified using the spectrum analyzer so that the power meter result is accurate. The difference between the power measured by the

spectrum analyzer and that measured by the power meter in only 0.180 dBm at the frequency of oscillation. Comparing the spectrum analyzer power to the power meter reading, it is concluded that the spectrum analyzer, while not as accurate as the power meter, is accurate enough to be used to measure the power.

The measured spectrum of the Colpitts VCO, at the design bias point, is given in Figure 5.15. The VCO does not produce any other signals, as can be seen in Figure 5.16.

Figure 5.15: Measured spectrum of the Colpitts VCO at the design bias point of $V_b = 3\text{ V}$ and $V_c = 3\text{ V}$.

The frequency and tuning range for the VCO is also measured. The collector bias voltage is tuned from 2.7 V to 4.2 V , giving a tuning range of 82 MHz and a power variation of $\pm 2.58\text{ dBm}$. The measured tuning performance of the VCO is shown in Figure 5.17.

Finally, the phase noise performance of the Colpitts VCO is -32.4 dBc/Hz at a carrier offset of 100 kHz , provided as Figure 5.18. The phase noise is measured by the HP 85671A expansion card for the HP 8565 Spectrum Analyzer.

These measured results represent the best performance of the three jigs manufactured.

Figure 5.16: Measured full frequency span spectrum of the Colpitts VCO.

Figure 5.17: Measured frequency and power tuning range of the Colpitts VCO.

Figure 5.18: Measured phase noise performance of the Colpitts VCO.

Chapter 6

Colpitts VCO Evaluation

6.1 Introduction

There are many risk factors evaluated for the Colpitts VCO, as outlined below.

The HBT The Nortel HBT is an experimental device and experience designing with it is limited. Additionally, the Nortel HBT was designed for digital applications, hence it is not meant for high power operations. Hence, the thermal restrictions degrade and complicate the design process.

The Colpitts Topology The Colpitts topology is traditionally used at frequencies much lower than Ka-Band. The design approach used is not conventional by microwave standards. Further, the resonator is a lumped element, and hence the noise performance is poor.

Compact Size The size of the VCO is extremely compact, thus increasing the risk due to many factors. The parasitics of the lumped inductor are used to destabilize the device. The choke inductor and output DC blocking capacitor double as a matching network. The close proximity of all elements increase the likelihood of inter-element coupling altering

component values. Specifically, the choke inductance is most likely altered by the nearby via hole. Finally, the output is more coplanar than microstrip, due to the two via holes.

Fabrication Process variations change the performance of the device and the element values. Yield is an issue, especially for the cap-on-vias.

Simulator Accuracy The simulator is only as accurate as the models it uses. Lumped element models are questionable at Ka-Band. Also, the HBT model approximates the device performance.

DC Issues A resistive bias network is used, which decreases the control of the quiescent point when testing the VCO. Additionally, collector bias is used to tune the VCO, instead of the more traditional varactor.

These issues are outlined in the sections to follow.

6.2 Ka-Band Technology Evaluation

6.2.1 HBT

As previously stated no microwave transistor has emerged as best overall. Each transistor has its own unique advantages and disadvantages. Table 6.1 compares the dominant GaAs microwave transistors in use today.

Referring to Table 6.1, the HBT is most suited to power applications and VCOs. Hence the HBT is an appropriate device choice for a Colpitts power VCO. Comparatively speaking, the HBT has a high current capacity per unit area, primarily due to the vertical structure of the HBT where the entire emitter area conducts. Current capacity is limited in FET technology due to the thin horizontal surface channel geometry. The power density capability

Table 6.1: Figure of merit comparison for microwave transistors [55].

Figure of Merit	MESFET	HEMT	HBT	Comparison
f_T	Low	Medium	High	HBT has the highest speed
f_{max}	Medium	High	Low	HEMT has highest frequency of operation
Gain-Bandwidth	Medium	High	High	HEMT is best for wideband operation at high frequency
Noise Figure	Medium	Low	High	HEMT is best for LNAs
Phase Noise	Medium	High	Low	HBT is best for VCOs
$\frac{g_m}{g_o}$	Low	Medium	High	HBT is best for highly linear amplifiers
$\frac{IP3}{P_{DC}}$	High	Medium	High	MESFET is best at high frequency; HBT is best at low frequency
Collector Efficiency	Medium	High	High	HBT is potentially the best power device
Power Density	Medium	Medium	High	HBT is potentially the best power device

of HBT technology is also enhanced by the high transconductance and low parasitic device resistances [56]. For highly linear power applications, the HBT has demonstrated a high $\frac{IP3}{P_{DC}}$ figure of merit, as shown in Table 6.1. Although the HBT has a non-linear transconductance, the device non-linear emitter-base capacitance can be used to cancel intermodulation distortion products [56].

Another advantage of the HBT is avoidance of sub-micrometer lithography, which is expensive. Lithography is avoided due to the vertical HBT structure, unlike horizontal FETs that do require lithography to define the critical gate width parameter. A further advantage of HBTs is the ease of integration of high speed digital and microwave circuits on a single chip.

A key disadvantage of GaAs HBT technology is low thermal conductivity. Consequently, junction overheating is a design concern and often thermal properties limit HBT performance before the electrical properties. Additionally, the noise figure for a typical HBT is higher than for FET devices, hence HBTs are not the device of choice for low noise amplifiers (LNAs).

Overall, the Nortel HBT is adequate for the Colpitts VCO. It provides adequate small signal gain, and large signal steady-state performance prediction in terms of frequency and output power.

Thermal Aspects In terms of thermal aspects, the Nortel HBT is extremely restrictive due primarily to its small size. Designed for digital applications, the Nortel HBT was not originally intended for analog applications. Despite this, the HBT is used successfully for the Colpitts VCO, except that the output power is severely restricted due to the Nortel HBT thermal restrictions.

This success is attributed to the load of the Colpitts topology. More specifically, the saturating loop power is due primarily to the load in the Colpitts topology. Hence the load is easily used to control the loop power, and by consequence the thermal performance of the HBT.

Of course this success is measured using a DLL. There is some debate as to the accuracy of this approach. Since the safe area of operation in Figure 3.5 is defined DC parameters, that is the maximum current density, maximum breakdown voltage and maximum DC power, one view is to only ensure that the bias point is within the safe area of operation. Additionally, the safe area of operation is in fact only defined by these DC parameters, so are there any RF limitations? In 1996, Gupta *et al.* [57] reported on a study of the affects of RF stress on HBTs. The results were that no new stress mechanisms are introduced by the RF stress. Therefore, the DC boundary conditions for the safe area of operation can be used.

In the Colpitts VCO design, the entire signal swing, not just the bias point, is contained within the safe area of operation. At millimetrewave frequencies, the peak values of the device currents occur quite often, hence the peak values can degrade the device performance. Thus the entire signal swing is contained within the safe area of operation.

Tuning The tuning of the Colpitts VCO is not large, and this is due to the low HBT internal capacitance variation with bias voltage. If an external tuning element is used, then this HBT characteristic is quite useful since frequency pushing due to bias fluctuations would be low. If the HBT is used as the external tuning element, that is using one junction as a varactor, a similarly low frequency variation would be achieved.

Therefore, to achieve a large tuning variation, another type of tuning element is necessary. One possibility is an active tunable inductor [58, 59, 60], which is particularly suited to the Colpitts topology since the feedback inductor is the element that primarily determines the output frequency.

In all cases, the large amount of unaccounted parasitics associated with a Ka-Band design will alter the frequency response of the circuit and invariably limit the tuning range of the VCO [61].

Model and Yield Given the small number of VCOs evaluated it is impossible to make any conclusions concerning the accuracy of the HBT model or the yield, other than to say that some of the variation in performance from simulation to measurements is most likely due to model inaccuracy and yield issues.

6.2.2 MMICs

Lumped Elements: Inductors, Capacitors and Resistors Lumped elements are difficult to use at Ka-Band because accurate models are not available. Hence, lumped elements are suited to applications where the specific values are not critical, such as blocking capacitors, or where broadband performance is required. Additionally, lumped elements are much smaller than their distributed counterparts.

Aside from a high impedance line, all microstrip inductors depend on mutual coupling between various line segments to achieve a high inductance in a small area. The performance

of lumped inductors is often degraded by shunt capacitance to ground. In the Colpitts VCO, however, these parasitics are successfully used as part of the circuit. This is possible because the exact values of the capacitors are not critical to the Colpitts topology as it is simply the ratio of the two capacitors, C_1 and C_2 , that de-stabilizes the device and enables oscillation start-up. Indeed, by using the parasitic capacitors, two capacitors and two via holes to ground are avoided. Thus, the use of the parasitics avoids an extra 0.1 mm^2 of circuit size, or an increase of 56%. This size accounts for two average capacitors, two via holes, and at least $50 \text{ }\mu\text{m}$ spacing between elements.

Using the inductor parasitics depends on not needing accurate values for the capacitances. The Colpitts topology requires that the ratio of the two capacitances be approximated to ensure start-up according to equation 2.41. Using electromagnetic simulations to predict the parasitics of the inductor is clearly accurate according to the comparison with measured data in Figure 6.9.

The final inductor issue is the resonant frequency, that is the frequency where the reactance of the microstrip spiral inductor is capacitive instead of inductive. This effect, also due to the parasitic capacitances, is avoided by careful electromagnetic simulation. The inductors used in the Colpitts VCO have resonant frequencies a few gigahertz higher than the frequency of oscillation.

Lumped capacitors are either interdigital or metal-insulator-metal (MIM) structures. MIM capacitors are preferred for the Colpitts VCO because they offer a higher capacitance per unit area than interdigital structures. The disadvantage of MIM structures is a greater process variation than interdigital capacitors. Further, interdigital capacitors are more accurately simulated by electromagnetic simulators.

The MIM capacitor in the feedback loop is a DC blocking capacitor, and hence is not critical to the Colpitts VCO operation, except that it causes a slight output frequency shift which is compensated by the feedback inductor. The output match/DC blocking capacitor

however is a key parameter. The measured results from the output capacitor are compared to the electromagnetic results from Sonnet EM in Figure 6.10. Although the S_{21} and S_{12} are identical, the input and output reflection coefficient, S_{11} and S_{22} , vary slightly. The electromagnetic simulation accounts for coupling with the ground plane presented by the via holes, however, as mentioned in Section 6.4.2, the ground planes only approximate the via holes. The explanation for the difference in Figure 6.10 could also be that the dielectric losses are not modeled adequately in Sonnet EM. The literature indicates that often MIM capacitor measured performance is degraded as compared to simulated performance due to extremely lossy dielectric films [27].

Thin film resistors are used in the DC bias network for the Colpitts VCO. The fabricated resistors correspond to the design values since the measured DC currents from the bias supplies are comparable to simulation, as seen in Table 6.2. The resistor lengths are restricted to be no longer than a fraction of a wavelength to avoid frequency dependent distributed effects. Further, although the resistance of curved and non-rectangular structures can be estimated [62], the resistors are straight to enhance the predictability of the resistance value. Finally, a metal film, NiCr, is used to avoid the nonlinear behavior at large current densities that is common with semiconductor films.

Table 6.2: Comparison of simulation and measured results for DC parameters.

DC Parameter	Simulation	Measured
Collector Voltage, V_c	3.0 V	3.0 V
Base Supply Voltage, V_{R1}	3.0 V	3.0 V
Collector Current, I_c	3.91 mA	4.07 mA
Base Supply Current, I_{R1}	2.53 mA	2.69 mA

Bond Wires and Ribbon Bond wires and their wider cousins, ribbons, are electrically short and hence are considered as inductors with a small but finite series resistance to account for conductor loss. Expressions for the inductance of bond wires are available [63] taking into

account the ground plane, adjacent wires and curved geometry.

The effects of the ribbon on the Colpitts VCO are outlined in Section 6.4.2. These effects could be greatly reduced by using a shorter ribbon, which is clearly possible from Figures 5.11, 5.12 and 5.13. The pedestal is wider than necessary, needing a longer ribbon, because the jig was fabricated before the VCO was cut out of the wafer and the residual scribe channel size was approximated on the larger side.

Grounds Using microstrip transmission lines for the Colpitts requires a method of connecting some elements to the ground plane on the backside of the integrated circuit. Indeed, there is an amount of inductance associated with any ground connection because of the $100\text{ }\mu\text{m}$ distance between the signal conductors and the backside ground plane. The ground connections in the Colpitts VCO are made through via holes, that is through metalized cylinders etched through the circuit substrate.

The characterization of this via hole inductance is not very accurate, in fact Nortel recommends only an ideal inductor. Using Sonnet EM to electromagnetically simulate the structure is better, however, the computation time is large due to the large area of the via hole. Also, the via hole substrate discontinuities affect the fields of the components that are in close proximity. Moreover, the via hole dimensions may vary across the substrate, causing electrical variations which are difficult to predict.

In addition to the variability of the via hole inductance, the via holes in the Colpitts VCO have capacitors directly on top of the via hole. The capacitors on these so-called cap-on-vias, share the bottom plate with the top of the via, hence they are shunt capacitors to ground. The cap-on-vias are a brand new element being fabricated by Nortel, hence the characterization and the process yield is very questionable.

Transmission Line Effects Every bend and junction in the transmission lines affects the overall Colpitts VCO performance. The need for bends is minimized by using the inductor

spirals and MIM capacitors for the 90° bends. The microstrip cross at the output capacitor/choke inductor/feedback loop/device node is simulated using a Libra model as opposed to Sonnet EM. Electromagnetic simulation would provide a more accurate characterization, and should be done in any future version of the Colpitts VCO.

6.3 Colpitts VCO Design Evaluation

An important evaluation criterion is the design approach used. This type of evaluation occurs chronologically before the design commences, and accordingly the evaluation is included as literature review in the design chapter. Thus, in addition to the double device line approach, many other design approaches exist in the literature, as seen in Table 4.4.

6.3.1 High Power Design Considerations

There are many design procedures for high power oscillators that exist in the literature. Most depend on multiple cell transistors to increase the power output. Increasing the transistor area to increase the current carrying capacity of the transistor does not evaluate the Colpitts topology, which is the focus of this thesis. Hence, the procedures evaluated attempt to maximize the output power of the Colpitts topology only, and not the transistor within the Colpitts topology. It is understood that a larger transistor will increase this maximum power.

Additionally, power combining of multiple oscillator is another high power design technique. As with the transistor area issue, combining multiple Colpitts oscillators will increase the total power output, however the thesis focus is to maximize the output power from the Colpitts topology itself. While it is acknowledged that the topology would need slight modifications to operate at a combined oscillator, no additional insight would be gained into the power generating capability of the Colpitts topology itself.

6.3.2 The Double Device Line Approach

The double device line approach is a successful method for the design of the Colpitts VCO. The method achieves optimum performance by careful investigation of the device-feedback circuit interaction. This investigation is primarily concerned with two aspects of the feedback circuit, and how they interact with the device. These aspects are:

1. Forming terminating impedances for the device input and output.
2. Providing a feedback path for oscillation.

For optimum oscillator performance, the entire feedback network is determined simultaneously given the above two aspects for consideration. To accomplish this, the basic philosophy of the Double Device Line approach is to match the feedback network to the conditions for optimal device operation.

Thus, there are many benefits of this approach, such as avoiding limitations in other design approaches when only part of the feedback network is connected to the device. Further, when the device is characterized for optimum operating conditions, the current flowing from the power sources is determined exclusively by the device, without any modification or limitation due to its terminations. Given that the terminating impedances are determined by the device only, and they are equivalently realized by the feedback network, a high degree of design accuracy is achieved. Moreover, the approach is easily applied at higher harmonics by considering voltages and currents at said higher harmonics.

The Double Device Line approach does however have a few numerical limitations. The coefficient matrix in the system of equations to be solved can sometimes be ill-conditioned due to the properties of the device. To overcome this difficulty, scaling of rows is sometimes necessary. Additionally, sometimes the diagonal element in the coefficient matrix may be too small, relative to the other elements, and this is a result of the phase difference between the device port voltages. Clearly, care must be taken in terms of actual numerical implementation

of the Double Device Line approach, even if the theory is sound.

It is useful to compare the Double Device Line approach to the traditional negative impedance approach in Section 2.5.2 and Wagner's approach [41]. The Double Device Line approach extends the traditional concepts of device and load loci to multi-dimensional space, and does not make any terminating impedance assumptions. Thus, the Double Device Line approach is an optimal design as compared to the traditional Ka-Band approach.

As compared to the traditional low frequency design method in Section 2.5.1, the Double Device Line approach is much more accurate. While the low frequency approach does include large signal device behavior, the parasitics associated with Ka-Band operation are too important to ignore. Hence, the low frequency method does not accurately predict the performance of the Colpitts VCO, although it can be used to provide insight into the operation of the topology.

6.4 Colpitts VCO Performance Evaluation

Overall, the performance of the Colpitts VCO can be summarized as average and predictable. There are many factors that lead to this generalization. These factors are further outlined in the sections to follow. The first two sections address the discrepancy between the simulated and measured results. Following this understanding, conclusions are made concerning the Colpitts topology performance at Ka-Band.

6.4.1 Physical Characteristics

The fabricated Colpitts VCO is $300\ \mu m \times 600\ \mu m$ in size, giving a area of $0.180\ mm^2$, and therefore the Colpitts VCO is well suited for use in higher order integration systems. This size is clearly more compact than existing VCOs listed in Table 1.1. This is an advantage of using the Colpitts topology, which is physically simple, and also using the parasitics of the

inductor to eliminate the layout of two elements. Further, using the HBT bias to tune the VCO eliminates the need for a varactor and the bias circuitry it would require.

It is assumed that as a result of the compact nature of the Colpitts topology, the inter-element coupling affects the predictability of the oscillator performance. With the use of electromagnetic simulators and test structures, however, these effects are reduced to a manageable source of error. This is evident due to the acceptable correlation between the simulated results and the measured results in this design.

6.4.2 Electrical Performance

The measured electrical performance differs from the simulated performance according to Table 6.3. As compared to the existing VCOs listed in Table 1.1, the output power of the Colpitts VCO is acceptable as it falls in the middle of the existing achievable performance. Although at first glance the Colpitts VCO tuning range appears low as compared to Table 1.1, it is acceptable since it should be compared primarily to [5] as that VCO was manufactured with same Nortel HBT as the Colpitts VCO in this thesis.

Table 6.3: Simulation and measured performance variation in key electrical parameters.

Parameter	Circuit Simulation	Electromagnetic Simulation	Measured
Frequency	28.04 GHz	27.33 GHz	33.83 GHz
Fundamental Power	-4.74 dBm	-1.27 dBm	-0.92 dBm
Tuning Range	350 MHz	390 MHz	82 MHz

Much of the variation in Table 6.3 is explained by the effects of the jig on the circuit.

Jig Effects

The Colpitts VCO is not designed to be used in a jig, rather it is intended to be integrated directly with other circuit functions. Therefore the effects of the jig cause the performance of the Colpitts VCO to differ from the circuit and electromagnetic simulations.

The major jig effect on the Colpitts VCO performance is due to the inductance introduced to the VCO from the ribbon connection from the alumina transmission line to VCO signal probe pad. The ribbon, which is $6.80 \mu m \times 946 \mu m$, is modeled using the ribbon model included in Libra, that is with an impedance of $12.03 + j231.34 \Omega$. This corresponds to a conductor loss of 11.98Ω and an inductance of $1.15 nH$ at $32 GHz$. The ribbon and load impedance are transformed through the output matching network and hence the device is presented with a complex impedance, $10.62 + j76.79 \Omega$, rather than the designed optimum real impedance of 100Ω . The net result is an extra inductor in parallel with the feedback inductor, as seen in Figure 6.1.

Figure 6.1: Colpitts VCO with ribbon impedance transformed through the matching network.

The extra inductance presented to the device by the ribbon adds to the phase shift through the feedback network. Therefore the device phase shift must also increase to attain a loop

phase shift of 0° . The phase shift through the device increases with increasing frequency. Hence the frequency of oscillation increases because the conditions for zero loop phase occur at a higher frequency. Simulations support this analysis, and the loop phase with the ribbon is shown in Figure 6.2.

Figure 6.2: Loop phase with the ribbon.

Additionally, the output power provided by the Colpitts VCO is also affected by the ribbon because the real impedance presented to the device is not the designed impedance. Hence, the transistor will saturate at a different loop power level and produce a different output power than originally designed for.

To verify this analysis, the Colpitts VCO performance is simulated with the ribbon impedance, and the electromagnetic results from Sonnet EM are also used. The resulting spectrum is shown in Figure 6.3 and the tuning range in Figure 6.4. Simulations indicate that with a ribbon connection the Colpitts VCO would oscillate at 32.03 GHz with an output power of -2.61 dBm . This new tuning range would be 710 MHz , with a fundamental output power variation of $\pm 3.17\text{ dBm}$. Clearly, these results are close to the measured results in

Figure 6.3: Simulated Colpitts VCO power spectrum with electromagnetic simulation data and jig components.

Figure 6.4: Simulated frequency and power tuning range of the Colpitts VCO with electromagnetic simulation results and jig components.

Table 6.3, with the exception of the tuning range.

Test Structures

Although the ribbon effect explains much of the measured result deviation, it does not account for the entire variation. The three circuit structures, shown in Figure 6.5, assumed most likely to cause variation in the measured results are fabricated separately and tested.

The feedback inductor, as seen in Figure 6.6, is the most crucial part of the Colpitts VCO to test separately since the inductor parasitic capacitances are used to de-stabilize the HBT. The effect of the output capacitor and associated ground planes on the output match is also critical to the Colpitts VCO output power, hence this structure is measured. This structure is shown in Figure 6.7. Finally, the choke inductor in Figure 6.8 is also part of the aforementioned output match, hence its performance with a via hole in such close proximity is also crucial. Note, however, that the choke inductor structure measured separately is not exactly as used in the Colpitts VCO. In the Colpitts VCO the choke inductor is connected to ground through the via hole. In the choke inductor test structure the choke inductor is not connected to the via hole because the inductor under test would be short circuited, and hence is not measured. Thus the choke inductor test structure is an approximation of the actual VCO section.

Certainly there are other circuit parts which affect overall measured performance, however, only the three parts mentioned are fabricated due to space restrictions by Nortel.

The test structures are de-embedded to the tapers using a Through-Reflect-Line (TRL) calibration technique. These standards were fabricated on the same wafer as the Colpitts VCO. The tapers are required due to the offset inputs of the spiral inductors, and the non- $50\ \Omega$ transmission lines used. These tapers are simulated in Sonnet EM and the resulting s-parameters are negated in Libra to de-embed the tapers from the measured results.

The measured results for the feedback inductor as compared to the equivalent structure

Figure 6.5: Parts of the Colpitts VCO tested separately.

Figure 6.6: Feedback inductor test structure.

Figure 6.7: Output match/DC blocking capacitor and via holes test structure.

Figure 6.8: Output match/choke inductor and via hole test structure.

simulated in Sonnet EM are shown in Figure 6.9. Clearly, Sonnet EM is able to successfully predict the performance of a single spiral inductor, including the associated parasitic capacitances.

The measured results for the output match/DC blocking capacitor and via holes are shown in Figure 6.10. Sonnet EM is not as successful at predicting the performance of the test structure and this deviation has a slight effect on the overall performance of the Colpitts VCO by altering the output matching. The measured deviation is due to the assumption in the Sonnet EM analysis that the via holes are perfect ground planes, which is not correct. Additionally, via holes alter the coplanar and/or microstrip modes of propagation along the transmission lines with respect to those predicted by the Sonnet Em simulations.

The output match/choke inductor test structure performance is predicted by Sonnet EM to a reasonable degree of accuracy, given the ground connection assumption previously mentioned. The comparison between measured and Sonnet EM results for the output match/choke inductor test structure is shown in Figure 6.11.

6.4.3 Thermal Performance and Design Robustness

Thermal Performance

In terms of electrical performance, the output power is successfully limited in the Colpitts VCO to achieve satisfactory thermal performance of the device. Clearly, the approach to alter the device impedance does limit the loop power, thereby reducing the thermal stress on the device. This is evident by the fact that the measured output power is predicted by simulation after the device impedance is altered, as presented in Table 6.3.

Figure 6.9: S_{11} , S_{22} (upper) and S_{12} , S_{21} (lower) comparison from 20 GHz to 40 GHz for the feedback loop inductor.

Figure 6.10: S_{11} , S_{22} (upper) and S_{12} , S_{21} (lower) comparison from 20 GHz to 40 GHz for the output match/DC blocking capacitor and via holes test structure.

0.75 0.50 0.25 0.25 0.50 0.75

Figure 6.11: S_{11} , S_{22} (upper) and S_{12} , S_{21} (lower) comparison from 20 GHz to 40 GHz for the output match/choke inductor and via hole structure.

Design Robustness

The noise is not evaluated as compared to simulation since there is no noise model available for the Nortel HBT. Leeson's model, however, does predict the noise profile, as seen in the measured spectrum in Figure 6.12 compared to the ideal profile of SSB phase noise in Figure 2.9. Leeson's model was not used to predict the actual SSB phase noise for the Colpitts VCO because the noise figure for the Nortel HBT is not available. Additionally, it has been reported that the accuracy of Leeson's model is limited to simply predicting the general shape of the Colpitts VCO phase noise [65]. This is because the Colpitts VCO it is difficult to justify the Q factor, output power and noise figure in Leeson's model.

Figure 6.12: Colpitts VCO measured spectrum close to the carrier to study the phase noise.

The Colpitts VCO has average phase noise performance. This is due to the low Q of the feedback network, that is the spiral inductor. This is evident by examination of the small signal loop phase simulations, which do have a steep frequency variation required for low noise performance, as shown in Figure 2.4.

Comparing the Colpitts VCO phase noise performance with the existing VCOs in Table 1.1, is it clear that the Colpitts VCO is not as good. Although this is due in part to the reasons mentioned in the previous paragraph, it should be noted that the VCOs in Table 1.1 were primarily designed for low noise performance, where the Colpitts VCO in this thesis was not.

As previously outlined, the yield and reliability, are not evaluated for the Colpitts VCO.

Chapter 7

Conclusions

7.1 Summary

In this thesis, Chapter 1 introduces the motivation for research into the application of the Colpitts topology for Ka-Band VCOs. This includes a brief history of the Colpitts topology and microwave oscillators in general to help place the present body of work into context. To further contextualize the thesis, the applications of a Colpitts VCO in wireless communications are also outlined.

In Chapter 2, the necessary oscillator fundamentals to understand the material in the later chapters is presented. The feedback and negative impedance oscillator models are used to illustrate conventional low frequency and microwave oscillator design aspects, including steady-state, start-up, stability and noise.

The third chapter describes the current state-of-the-art of Ka-Band technology. This includes the theory and operation for the HBT device, along with the model used. Aspects of MMICs are introduced, along with fabrication and CAD issues.

Chapter 4 outlines the design process for the Colpitts VCO. Starting with the design constraints and evaluation criteria, the chapter presents the gain element and feedback network

design issues. The chapter concludes with sections on thermal and tuning issues, and finally with the Colpitts VCO layout.

The fifth chapter presents the Colpitts VCO performance from circuit simulation, electromagnetic simulation and measurements. The measurements are performed in a jig, hence the jig design is also described.

The Colpitts VCO is evaluated in the sixth chapter. The evaluation is based on the electrical performance of the Colpitts VCO, and primarily focuses on the explanation of the deviations from simulations. Additional evaluation criteria is also presented, along with an evaluation of the Ka-Band technology. The chapter concludes with an evaluation of the design procedure used, as compared to the other possible design approaches, which are also presented.

The seventh and final chapter outlines a summary of the investigation and suggests future research that stems from the current work. The conclusions from the thesis are also included.

7.2 Conclusions

7.2.1 The Colpitts VCO

The design and performance of a $300\mu m \times 600\mu m$ MMIC Ka-Band Colpitts VCO is evaluated. The Colpitts VCO is successfully designed at Ka-Band using a Nortel $6.5 \times 3\mu m$ HBT. The Colpitts VCO delivers an output power of -0.92 dBm at 33.83 GHz . The Colpitts VCO is tuned 82 MHz using the collector bias, and the phase noise is -32.4 dBc/Hz at a carrier offset of 100 KHz . Clearly, the VCO needs to be phase-locked and the tuning range could be improved. The output power and frequency, however, were well predicted by simulations.

The Colpitts VCO uses lumped elements to achieve compactness for higher order integration. The parasitics of one of these lumped elements, namely the feedback spiral inductor, are used to de-stabilize the HBT. The load impedance is matched to a device impedance such

that the device operates within its thermal limitation.

The Colpitts topology, and simulations thereof, is further evaluated using measured test structures. It is concluded that the Sonnet EM simulator can provide accurate simulations if the assumptions for the simulations are carefully considered.

Moreover, the low frequency design technique is not suitable for the Colpitts VCO at Ka-band, primarily due to parasitics not being included. The traditional Ka-Band technique can be used, however, there are other design procedures which are better.

The Colpitts VCO is compact and suited to higher order integration with other components. The performance impact of inter-element coupling is noted, however, it is not large.

The thermal requirements for the Nortel HBT limit the overall Colpitts VCO performance. A larger HBT would dramatically improve performance by increasing output power, tuning range and decreasing the phase noise by allowing operation at a higher bias voltage.

Overall, the Colpitts VCO as manufactured is average in terms of performance, and exceptional in terms of small size. Its size makes the Colpitts VCO well suited to applications in wireless communications.

7.3 Future Research

As with most research topics, the investigation into a Ka-Band Colpitts VCO may lead to as many potential research topics as there are topics to evaluate in the present work. The possible future research is roughly grouped into further Colpitts topology research and investigations into phase-locked oscillators, which is the most likely application for the Colpitts VCO in wireless communications. The ideas outline below are but a sampling only, and by no means should they be considered as a complete listing.

7.3.1 Colpitts Topology

One of the most interesting possibilities for investigation of the Colpitts topology would be to experimentally measure the loop gain, as outlined by Middlebrook [66]. If some tunable elements are included, much insight would be achieved into the small and large signal performance of the Colpitts topology. Additionally, measured s-parameters for the feedback elements can be used to determine the Q factor, yielding further insight into the phase noise of the Colpitts VCO [67].

Various other phase noise analyses can also be applied. From general frequency domain techniques [68] to the well documented role of baseband noise upconversion in HBTs [69, 70], to impedance noise matching [71], there is much future research to be done.

Further to the phase noise, it has been reported that Nyquist analysis techniques can be applied to oscillators to address stability issues [72]. Investigation into the Colpitts topology using this technique could yield a wealth of new insight.

Finally, the effect of load pull on both frequency and power could be evaluated further, either by experiment or analysis [73].

7.3.2 Phase-Locked Oscillators

The primary use of the Colpitts VCO in wireless communications of any type is in the form of a phase-locked oscillator (PLO). A PLO in wireless communications is used as the system local oscillator (LO) to produce the signals for up or down converters. An extremely low phase noise oscillator is required for the LO, hence by using a phase-locked loop (PLL) to lock the Colpitts VCO to a stable reference, such as a crystal, a low noise signal for the LO is produced. The Colpitts VCO is well suited for use as a PLO, due primarily to its small size. By multiplying a crystal frequency up to Ka-Band, the Colpitts VCO will by far less noisy since its output frequency attempts to follow the crystal frequency, which is very stable.

PLO sources for LMCS are currently being rigorously investigated [74, 75, 76, 77, 78, 79].

There are many different configurations, for the PLO as a PLL can be implemented in a variety of ways, with different phase detectors and loop filters. Additionally, the locking can occur at a variety of different frequencies since the Colpitts VCO and crystal reference frequencies can be multiplied and divided as needed.

Perhaps the greatest challenge in designing a Ka-Band PLO is the loop filter. In order to take advantage of the compact Colpitts VCO and associated benefits of the MMIC technology, the loop filter must also be integrated. This is difficult due to the large passive structure required for frequency selective circuits. Research in active filter design promises to mitigate some of these issues [80].

7.4 The Big Picture

In an invited paper by Robert Lucky [81] in the October 1997 issue of *Proceedings of the IEEE*, the abstract states,

Historically, the deployment of new telecommunications services has taken years of effort and large amounts of investment. Thus, being able to predict the market acceptance before taking the business risk is critically important. Nevertheless, the record of success in such predictions had been relatively poor. Many new services are unable to overcome the barrier posed by Metcalfe's law, which says that there is a small value to any service shared by only a small number of users. An outstanding example of breaking this barrier was the World Wide Web, which was jump-started by the promulgation of free browser software. The failure of the industry to have foreseen the Web is indicative of the fundamental inability to predict the future. Whereas engineers sometimes invoke the "field-of-dreams" argument (if we build it, they will come), industry executives demand quantitative business projections. In spite of this fundamental uncertainty in societal acceptance of a particular service, there are common themes in human needs - both in the dreams and aspirations and in the everyday life of average citizens. Telecommunications has helped fulfill those aspirations and needs and will do so to a greater extent in the future [81].

Clearly, despite the statistics-based predictions included in Chapter 1, some people dispute the specific predictions as to the future of wireless communications, such as the *Field of Dreams* analogy. The growth is not in dispute, just the form it is going to take. Perhaps then

the big picture for the Colpitts VCO is not so much the form of the wireless communications systems, but rather the reasons for building them. The *Proceedings of the IEEE* article cites four human needs fulfilled by communications systems [81]:

1. **We want to be somewhere we are not.** Communications systems projects our presence somewhere else and brings others closer to us.
2. **We want to be in some time we are not.** With the explosion of voice mail and e-mail we can live our lives more asynchronously, and be in two places at once.
3. **We want to be someone we are not.** Vicariously sharing the experiences of others is a fundamental feature of existing and future forms of communications systems, usually through entertainment.
4. **We want to be something we are not.** With the global resources future communications systems, a new sense of community will enable personal growth not seen in the past.

The greatest advancements of the last century have not been new technologies, but the human gains from the new technologies. To think of the Colpitts VCO, and the communications systems it will be a part of, in terms of these human factors, is truly the big picture. It is hoped the research on the Colpitts VCO will in some way contribute to attaining these factors for the people of the future.

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IMAGE EVALUATION TEST TARGET (QA-3)

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