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**A Hybrid High-Tc Superconductor Filter/Low-Noise Amplifier
Microwave Integrated Circuit**

Daniel J. Drolet

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

Master of Electrical Engineering

**Ottawa-Carleton Institute of Electrical Engineering
Department of Electronics
Faculty of Engineering
Carleton University
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March 1994

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Microwave Integrated Circuit**

submitted by
Daniel J. Drolet, B. Eng.

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

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ABSTRACT

The increasing demands for higher performance electronic circuits in both digital and analog applications is leading to the development and implementation of alternative technologies. One area of interest is the operation of circuits and subsystems at cryogenic temperatures to improve their speed, reliability, and power handling requirements. Semiconductor devices for microwave applications such as the high electron mobility transistor (HEMT) exhibit superior performance at lower temperatures due to improved transmission gain, noise figure, and frequency of operation.

Superconductivity is a promising alternative for improving circuit performance. The recent discovery of high temperature superconductors (HTSCs) which operate above the temperature of liquid nitrogen (77K) has made its application in analog and digital electronics much more feasible. Previous low temperature superconductors, requiring expensive cooling systems, have demonstrated superior performance to conventional circuits in terms of speed, sensitivity, and switching capability. The extremely low loss properties of HTSCs, along with the simpler refrigeration systems necessary for operation, allow high Q resonators to be used in microwave components such as narrowband filters and low phase-noise oscillators.

This thesis presents the design and performance of a hybrid semiconductor/superconductor K-band filter/amplifier microwave integrated circuit. The circuit combines the high frequency, high performance advantages of a cryogenically cooled GaAs HEMT with the low loss, high Q properties of a narrowband HTSC filter and matching networks.

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CHAPTER 1: INTRODUCTION

1.1 Overview

Cryogenic temperatures have been used to improve the speed, reliability, and performance of both digital and analog electronic systems. Low temperatures improve the speed of digital circuits by enhancing the mobility of solid state devices, and improves reliability by reducing the effects of heating caused by higher power densities. In high-speed analog electronics, cryogenics have also been implemented to minimize the thermal noise in low noise microwave receiver systems. Low temperatures provide both an improvement in conventional semiconductor device operation, and permit the implementation of superconductor devices. Moreover, with recent advances in superconductor technology, integration of semiconductors and superconductors can produce hybrid circuits with higher performance than conventional circuits at 77K.

The performance demands on conventional semiconductor electronics are being pushed to physical limits that cannot be exceeded due to fundamental laws of nature and the intrinsic properties of materials.^[1] Semiconductor technology offers a trade-off of high speed capability versus high power dissipation in digital circuitry. The speed is limited by the switching ability of the transistor, as well as the lengths of the internal wiring and interconnections, while the power dissipation limits the size of digital logic circuits.

Semiconductor components operated at low temperatures exhibit a number of improvements, such as higher Q (i.e. quality factor) circuits due to lower resistivity and reduced substrate loss. Increases in operating speeds by a factor of two have been reported in CMOS circuits due to the enhanced carrier mobility, at lower supply voltages.^[2] The ETA-10 supercomputer is one example

of an application which uses liquid nitrogen to exploit the performance advantages of CMOS devices at 77K.^[3] As circuit topology reduces in size and power densities increase, cooling is required to maintain circuit reliability by eliminating heating problems. These heating effects contribute to higher failure rates and shorter lifetimes in solid state devices. A controlled operating temperature environment which cryogenics provides will reduce any degradation in the performance of temperature-sensitive components.

In microwave electronics, active semiconductor devices such as FETs and high electron mobility transistors (HEMTs) have demonstrated significant performance improvements at cryogenic temperatures. These devices have yielded increased carrier mobilities and transconductance values at 77K, which provides a higher frequency of operation well into the millimeter-wave frequencies. The higher transconductance and lower resistance resulting from low temperature operation also results in higher gain and lower noise figure. Extremely low noise levels are required for radio astronomy and deep space communications at microwave and millimeter-wave frequencies, where the incoming signal is mixed with cosmic and atmospheric noise components. The thermal noise, which is directly dependent on temperature, is inherently reduced by operating in a cryogenic environment. Recently, cooled HEMT devices have been capable of meeting these noise requirements as a less expensive alternative to traditional components such as MASERS and parametric amplifiers.^[4,5]

Another significant advantage of using cryogenic technology in electronics is the possible application of superconductors. Since the discovery of superconductivity in 1911, continuous research has been undertaken into developing new superconductors with higher transition, or critical, temperatures (T_c). The early superconductors were metallic elements such as niobium and lead which required liquid helium (4K boiling point) as a cryogen in order to

superconduct. Because of their extremely low transition temperatures, these materials are referred to as low temperature superconductors (LTSCs). Apart from exhibiting extremely low losses, another significant contribution of LTSCs has been the Josephson junction, which produces high-speed switching capabilities in digital electronics faster than any semiconductor device. The high-speed switching also occurs with very low power dissipation, hence circuit density can be increased and chip size can be reduced. Superconducting wiring and interconnections in digital systems provide another significant advantage over conventional technology, in that dispersion is negligible. Pulses are therefore able to propagate along a transmission line with minimal distortion.

The Josephson junction has also found application in high performance analog electronics to produce voltage-controlled oscillators (VCOs) at millimeter-wave frequencies.^[6] Other applications include a Josephson parametric amplifier^[7] which yields extremely low noise levels, and Superconducting Quantum Interference Devices (SQUIDs)^[8] which act as highly sensitive magnetometers. Superconducting-insulating-superconducting (SIS) mixers are replacing conventional technology in high performance microwave systems such as for radiometry and radar applications.^[9]

In 1987, high temperature superconductors (HTSCs) were discovered which operate at liquid nitrogen (77K boiling point) temperatures. These materials are ceramics that typically exhibit extremely poor conductivity above their transition temperature. Liquid nitrogen refrigeration systems are more feasible than the bulky, hard-to-handle liquid helium coolers due to a 60:1 improvement in the latent heat of vaporization. Also, a HTSC device can be cooled using liquid nitrogen at an annual cost of \$35, as compared to liquid helium systems, which cost upwards of \$50,000 annually.^[10]

One suitable application for HTSCs is in low power passive microwave devices such as high Q resonators and filters. These high Q components have been implemented in planar transmission line circuits to yield comparable performance to conventional waveguide and dielectric resonators, at a significant reduction in size.^[11,12,13] While the cost of the refrigeration system must be considered, the implementation of planar HTSC filters into miniature filter banks required for applications such as radar and electronic warfare can offset the cost and inconvenience of cooling the entire system.

The implementation of a cryogenic refrigeration system must justify the cost of cooling the components. Refrigeration systems can either utilize a short duration, inexpensive open-cycle method such as a liquid nitrogen dewar, or a long duration closed-cycle method such as the Stirling cryocooler and Gifford-McMahon techniques which require electrical power.^[14] Sharing refrigeration systems among several analog and digital systems can reduce the cost of achieving low temperatures. The use of superconductivity in deep space applications is especially promising due to the inherently cooled environment in which only passive radiators are required.

It is preferable to use liquid nitrogen refrigeration systems over liquid helium to cool semiconductor circuits not only because of the enormous cost difference, but also because little advantage is gained at temperatures lower than 77K. Dopant freeze-out occurs in GaAs FETs at temperatures approaching the liquid helium boiling point, which severely degrades the device performance. Also, resistivities in conventional metals such as copper and gold reach anomalous limits for thin-film transmission lines at temperatures slightly below 77K. An exception is in extremely low noise applications, such as radio astronomy, where temperatures of 4K are necessary to considerably lower the thermal noise in the receiver front end.

Superconductors exhibit certain limitations which need to be overcome before they are able to achieve the wide-spread use of semiconductor devices. One of the main disadvantages of superconductor technology is the lack of a suitable three-terminal device to produce gain. While the two-terminal Josephson junction is a mature device for LTSCs, HTSC devices are still in the developmental stages. Therefore, a promising alternative is the integration of both semiconductor and superconductor components into high performance hybrid microwave circuits. These hybrid circuits can exploit the advantages of gain-producing solid-state devices and high Q, low loss HTSCs which both operate better at liquid nitrogen temperatures. Thin-film HTSC processing techniques have not yet reached the maturity of semiconductor processing technology, and therefore superconductor circuits are quite expensive to fabricate.

Several examples of these hybrid semiconductor/superconductor microwave integrated circuits (MICs) have been reported since the discovery of HTSCs. A millimeter-wave receiver system has been proposed which utilizes superconducting antennas and filters to achieve low losses and high Qs, and semiconductor devices for the amplifiers and analog signal processors.^[15] An analog signal processing system was developed which uses superconducting tapped delay lines to achieve an attenuation competitive with surface acoustic wave (SAW) devices.^[16] A hybrid semiconductor/superconductor Ku-band oscillator was designed at the Communications Research Centre (CRC) using a high-Q HTSC linear resonator and monolithic MIC (MMIC) amplifiers.^[17] A similar oscillator hybrid circuit was developed at Dupont for C-band application.^[18] A Ku-band low noise amplifier (LNA) was designed for operation at 77K using HTSC matching networks and a GaAs HEMT.^[17] Also, an

integrated HTSC filter/GaAs LNA has been developed for possible application in high sensitivity microwave receivers.^[19]

While these examples illustrate the ability to interconnect superconductor and semiconductor devices into a circuit, recent papers have reported three-terminal devices using high- T_c superconducting material. Examples of these hybrid devices include the superconducting FET (SuFET)^[10], and the Superconductor-Base Semiconductor-Isolated Transistor (SUBSIT).^[20]

1.2 Thesis Objectives and Organization

The objective of this work is to investigate the potential for combining semiconductor and superconductor technologies into a hybrid microwave integrated circuit. The benefits of integrating the two technologies to produce higher performance components is emphasized. Specifically, a narrowband HTSC filter is integrated at the input of a HTSC/GaAs amplifier cooled to 77K for operation at K-band. Such filter/amplifier combinations are useful at the front end of microwave receivers to lower the noise figure while removing any intermodulation products due to spurious signals outside the frequency band. This particular circuit configuration was chosen to demonstrate the improved performance of an active semiconductor device at cryogenic temperatures in a low-noise amplifier, and also to illustrate the advantage of HTSCs for low loss, narrowband filters. A further objective of this work is to investigate the reliability of various commercial CAD software packages in analyzing HTSC passive structures.

Chapter 2 provides background information on various forms of conventional MIC technology and the advantages of operating MICs at cryogenic temperatures. An overview of the theory of superconductivity for LTSCs and HTSCs is presented, including the Two-Fluid model, London and BCS theories.

HTSC thin-film processing techniques and circuit fabrication are examined. Finally, the cryogenic measurement technique used in this work is described, including an overview of the cryocooler apparatus and the calibration procedure used to perform S-parameter and noise figure measurements.

Chapter 3 presents the design of a K-band HTSC/GaAs amplifier MIC at 77K. Firstly, the characterization of a HEMT device at 77K is described. Based on this characterization, the amplifier was designed using HTSC matching networks. Measurements and results of this amplifier, as well as a modified version, are then presented. The final section of the chapter discusses CAD software used in simulating the HTSC matching networks.

The development of narrowband HTSC microstrip filters is presented in Chapter 4. A description of the filter design procedure is followed by measurements and results, and comparisons with CAD simulations. The filter was then integrated with the amplifiers described in Chapter 3 to produce the hybrid semiconductor/superconductor filter/amplifier MIC. S-parameter measurements and noise figures are presented for two variations of the hybrid circuit.

The final chapter summarizes the performance benefits of combining semiconductor and high temperature superconductor technology. A review of the accomplishments of this work is presented, along with possible modifications and improvements to the circuits. The application of HTSCs in future hybrid circuits is explored, as well as possible improvements in the design and measurement of these components at cryogenic temperatures.

CHAPTER 2: SEMICONDUCTOR AND SUPERCONDUCTOR TECHNOLOGIES: BACKGROUND

2.1 Introduction

The integration of semiconductors and superconductors in electronics applications has permitted circuit designers to exploit the benefits of each technology. Combining these two technologies may be carried out on three different levels.^[15] The first level involves the combination of semiconductor chips with superconductor chips or interconnections. The next level involves integrating semiconductor and superconductor devices on the same chip. At the highest level of integration, the two technologies are combined within the same active device. This thesis focuses on the first level, as this is the simplest to fabricate, in which semiconductor active devices will be integrated with HTSC passive structures on separate chips.

In this chapter, an overview of semiconductor microwave integrated circuit (MIC) technology will be presented, including the areas of performance improvement by operating circuitry at cryogenic temperatures. A comparison of the advantages and disadvantages of MICs versus other circuit fabrication technologies is presented. Following this discussion, an overview of the theory of superconductors for LTSCs and HTSCs is described. The various processing techniques for HTSC thin-films will be discussed, as well as the microwave characterization of these films. The final section will present the cryogenic apparatus used for S-parameter and noise figure measurements, along with the calibration methods used.

2.2 Microwave Integrated Circuits (MICs)

2.2.1 MIC Technology

Microwave integrated circuits have increasingly replaced conventional microwave circuits using coaxial and waveguide components due to the more stringent demands on size and weight for applications such as satellite communications, electronic warfare, and radar systems. MICs can also yield higher reliability, better performance, and lower cost along with the significant reduction in size as compared to these conventional technologies. A basic MIC consists of planar transmission lines on a single low loss dielectric substrate which form distributed matching elements and interconnect the discrete devices. These discrete devices include both passive components such as resistors, capacitors, and inductors, as well as active semiconductor devices such as diodes and transistors. These components are usually attached to the substrate in chip form and are connected into the circuit using wire bonds. This type of circuit is referred to as a hybrid MIC.

An extension of the hybrid MIC is the miniature hybrid MIC (MHMIC), wherein all passive components are fabricated onto the MIC substrate using thin-film processing. Only the active devices are added separately to the circuit. MHMICs provide a smaller circuit topology allowing for broader bandwidths, as well as lower assembly costs and improved circuit reliability compared to conventional circuits. The disadvantages include more complex processing and a longer turnaround time for circuit fabrication.

Hybrid MIC technology offers the availability of a wide selection of devices, including high-precision laser-trimmed resistors and stacked capacitors which allow for a wide range of capacitances while maintaining a small base area. Active components include Gunn and varactor diodes, as well as various transistors such as FETs, HEMTs and HBTs. A significant problem with these

discrete active devices is the wide variety of electrical characteristics between batches caused by slight variations in processing conditions. Also, since these devices are usually wire bonded into the circuit, the variation in the placement and length of the bond wires can affect the circuit performance and reliability.

Another advantage of hybrid MICs is the variety of substrates which are available. The substrate must have a uniform dielectric constant, with a relatively high value to reduce the circuit size. It also must provide mechanical stability from shock and vibration, and high thermal conductivity and temperature stability for producing the various circuit elements and structures.^[21] Alumina (Al_2O_3) is one of the most suitable substrate materials in thin-film technology on the basis of the above criteria. Its substrate properties are listed in Table 2.1, along with some other common substrate materials.

	Relative Permittivity ϵ_r	Loss Tangent (at 10 GHz), $\tan\delta$	Specific Thermal Conductivity, $\text{W/cm}\cdot\text{K}$
Alumina (99.5% pure)	9.8	0.0001	0.37
Sapphire	9.4	0.0001	0.42
Quartz glass	3.8	0.0001	0.017
GaAs (highly resistive)	12.9	0.002	0.46
Silicon ($\rho = 10^3 \Omega\text{-cm}$)	11.9	0.015	1.45
PTFE	2.1	0.0003	0.002

Table 2.1 Properties of various substrate materials for MICs^[21]

The relatively large distributed matching networks employed in MICs are useful for narrow bandwidth applications such as low noise amplifiers. Also, the availability of high Q components such as dielectric pucks and YIG

sphere resonators in MICs can produce stable oscillators and narrowband filters with Q s greater than one thousand.

A second type of MIC is the monolithic MIC (MMIC), in which all active and passive circuit elements and interconnections are fabricated on a semiconductor substrate in a complete technological process. MMIC technology has gained prominence due to several factors, including: the production of large diameter semi-insulating Si and GaAs wafers; the development of the MESFET; and the increased demand for large quantities of MICs. The use of highly resistive GaAs substrates allows for a higher frequency of operation by reducing the leakage effects. The small size of MMICs makes it possible to reproduce large numbers of identical circuits on a single wafer, and also allows for broad bandwidth applications. Another advantage offered with MMICs is the potential for using customized circuit elements to achieve the desired performance. This includes fabricating passive components with specific values, as well as diodes and FETs with various dimensions to suit the required application. Also, as all elements are interconnected by patterning thin-film metal layers rather than using non-repeatable bond wires, the reliability of the circuit is enhanced.

Due to the common substrate that MMIC devices share, the selection of active devices is limited to those which comply with the doping profile, hence certain devices such as PIN diodes are unavailable in many MMIC applications. Another disadvantage of MMICs is the limited power handling capability of the GaAs substrates due to their poor thermal conductivity. Also, the reduced size of MMICs results in smaller spacings between the circuit elements, thus creating circuit parasitics which become significant at higher frequencies.

The most common forms of planar transmission lines in MICs are microstrip, stripline, and coplanar waveguide, which are displayed in Figure 2.1.

The microstrip line is the most commonly used for its ease of fabrication, compact size, low ground plane and radiation losses. The non-uniform dielectric, consisting of air and the substrate, causes a quasi-TEM mode of propagation in the microstrip. Values for the effective permittivity and the characteristic impedance of microstrip lines are generally computed using CAD software due to the complexity of the expressions. Another transmission line is coplanar waveguide, which has the advantage that all metallization is on one side of the substrate, allowing for low inductance grounding which is beneficial at millimeter-wave frequencies. However, circuit topologies should be kept relatively simple due to space limitations and difficulty in realizing complex circuit geometries.

Figure 2.1 Various planar transmission line configurations^[21,22]

A stripline configuration reduces the radiation losses by confining the electromagnetic fields between two ground planes. Stripline also supports TEM propagation due to the uniform dielectric surrounding the conductor. These

advantages are offset by the difficulties in fabricating and tuning stripline circuits, since the elements are sandwiched between the ground planes.

The transmission lines are formed by depositing the metal layer, usually gold or copper, onto the substrate using thin- or thick-film technology. Thin-films, which will be used extensively in this work, are deposited onto the substrate by various methods such as electron-beam evaporation and sputtering. The films are then patterned using etching or lift-off processing techniques. Hybrid MICs usually require a single layer of metallization, and are less expensive as compared to MMICs. A typical MIC can be fabricated and assembled within one week. A MMIC usually requires at least ten masks and complex processing to pattern all layers onto the substrate. The cost of fabricating a few 3" wafers at an MMIC foundry using the same mask set can be upwards of \$100K, with a typical turnaround time of sixteen weeks.

The decision on whether to use hybrid MICs or MMICs depends upon the application, performance requirements, cost and time considerations. MMICs are more suitable for applications requiring large quantities of small, lightweight, and broadband circuits. Conversely, MICs are more applicable to small numbers of low complexity circuits requiring low noise figure, narrow bandwidth or high power, as well as a short turnaround time. For these reasons, and the fact that superconductor-based MMICs do not presently exist, all circuits designed for this work were fabricated using hybrid MIC technology.

2.2.2 MICs at Cryogenic Temperatures

Operating MICs at cryogenic temperatures offers a number of benefits to the performance of the circuit. A cooled environment will increase the Q factor of a circuit by reducing conduction and substrate losses. Active semiconductor devices exhibit improved gain and higher frequency of operation, as well as extremely low noise figures under low temperature conditions. Other

factors such as temperature stability, circuit reliability, and power density benefit from a cooled operating environment.

Transmission line structures have three main loss mechanisms: resistive conductor loss, dielectric loss, and radiation loss which are related to Q_c , Q_d , and Q_r respectively. Each mechanism degrades the overall Q of a circuit according to equation 2.1. Recall that the quality factor describes the performance of a resonant circuit, and is defined as the ratio of the time average stored energy to the energy lost per second.

$$\frac{1}{Q} = \frac{1}{Q_c} + \frac{1}{Q_d} + \frac{1}{Q_r} \quad (2.1)$$

The resistive loss of normal conducting metals decreases significantly when cooled to liquid nitrogen temperatures (77K). The resistivity of bulk gold, for instance, drops to one-sixth of its room temperature value, while similar decreases to one-eighth and one-tenth the room temperature resistivity occur for copper and aluminum, respectively.^[23] The graph in Figure 2.2 displays the decrease in the DC resistivity of bulk gold over temperature.

In thin-films, the reduction in resistivity over temperature is not as significant due to a "size effect" which occurs when the conductor dimension approaches the mean free path of the carriers.^[24] As the thin-film is cooled, the electron mean free path increases, causing an increase in surface scattering. This results in a higher resistivity as compared to bulk material. Figure 2.3 displays the effect of temperature on the resistivity of thin-film copper relative to its room temperature value. At 77K, the difference between the bulk resistivity and that of a 1 μ m thick film is less significant as compared to lower temperatures.

Figure 2.2 Resistivity of bulk gold versus temperature^[23]

Figure 2.3 Resistivity of thin-film copper versus temperature^[24]

Resistive losses become more severe at higher frequencies due to the "skin effect" in which the fields and associated currents are confined to a very thin layer of the conductor surface. This effect is measured by the penetration depth, δ , as defined in equation 2.2 for good conductors. Based on this parameter, the surface resistance, R_s , is derived in equation 2.2 displaying a square root dependence on frequency.

$$\delta = \frac{1}{\sqrt{\pi f \mu_0 \sigma}} \quad R_s = \frac{1}{\sigma \delta} = \sqrt{\frac{\pi f \mu_0}{\sigma}} \quad (2.2)$$

At low temperatures, the surface resistance reaches anomalous limits as shown in Figure 2.4 for bulk copper. This phenomenon is similar to the size effect, in which the higher mean free path at cold temperatures and reduction in δ at higher frequencies causes surface scattering, thus limiting R_s to a minimum value. The AC resistances of both copper and gold are approximately reduced by a factor of four at 77K.

Figure 2.4 R_s of copper vs. temperature at various frequencies^[24]

A second consideration affecting the Q -factor is the contribution due to the dielectric losses of the substrate. The loss tangent, $\tan\delta$, is inversely proportional to Q_d , and is defined as the ratio of the displacement current to the conduction current. The loss tangent of various substrates has been shown to decrease with temperature, thus yielding a higher Q_d .^[23] This trend was also observed in more recent experiments of substrates such as LaAlO_3 , which are suitable for supporting HTSC structures.^[25]

The final loss mechanism, radiation loss, becomes a factor at higher microwave frequencies. Also, since radiation loss is unaffected by temperature, it becomes significant at low temperatures. Aside from using a stripline configuration, radiation losses can be reduced by removing transmission line discontinuities, using thinner substrates, providing shielded enclosures, and minimizing the circuit area.

Power density of MICs and MMICs improves at low temperatures due to the enhanced thermal conductivity of pure materials, as displayed in Table 2.2. This is not true for composites and materials with impurities such as silver epoxy, however. Circuit reliability also improves at low temperatures by extending the lifetime of GaAs devices, as found by Roesch.^[26] FETs have been the main failure mechanism of MICs in which interdiffusion of the junction metals into the GaAs substrates has been caused by heating. Cryogenic temperatures implicitly provide insensitivity to changes in external ambient temperature, thus minimizing heating problems.

Another important advantage of cooling MICs is the dramatic improvement in performance of active devices such as FETs and HEMTs. At low temperatures, the transconductance and frequency of operation increase significantly, along with a corresponding reduction of the device noise figure. These effects will be examined in more detail in Chapter 3.

GaAs	10x
Gold	8x
Alumina	4x
Ag epoxy	0.5x

Table 2.2 Thermal conductivity improvement from 300K to 77K^[27]

2.3 Theory of Superconductors

Superconductivity was discovered in 1911 by Kammerlingh Onnes, who found that a sample of mercury cooled to 4K using liquid helium yielded zero resistance. For many years following, discoveries of various niobium-based alloy superconductors with higher critical temperatures (T_c) were made. A number of theories have been developed for superconductivity such as the Two-Fluid model and London theory^[28] based on the Meissner effect, discovered in the 1930's. The Bardeen-Cooper-Schrieffer (BCS) theory^[29] in 1957 introduced the concept of Cooper pairs, while Josephson's theory^[30] in 1961 explained the Josephson junction. An overview of the evolution of superconductors and their development is displayed in Figure 2.5.

The second major breakthrough occurred in 1986 when Bednorz and Müller discovered a ceramic-based superconductor $\text{La}_{2-x}\text{Ba}_x\text{CuO}_4$ (LaBaCuO) which attained a T_c above 30K.^[31] Within one year, Chu at the University of Houston broke the liquid nitrogen (77K) barrier with the discovery of $\text{YBa}_2\text{Cu}_3\text{O}_7$ (YBCO) with a critical temperature of 93K.^[32] The discovery of these high temperature superconductors (HTSCs) has sparked enormous research into finding new HTSCs such as $\text{Bi}_2\text{SrCa}_2\text{Cu}_2\text{O}_8$ (BSCCO) with $T_c=115\text{K}$, and $\text{Tl}_2\text{Ba}_2\text{Ca}_2\text{Cu}_3\text{O}_{10}$ (TBCCO) with a T_c of 125K. These superconductors could find application in numerous areas including high Q resonators for filters and oscillators, RF antennas to exploit low loss, and high speed interconnects in

analog signal processors to remove loss and dispersion effects. This section describes the basic theory of low temperature superconductors, and how these concepts have been extended to explain high temperature superconductors.

Figure 2.5 Chronological history of superconductivity^[33]

2.3.1 Low Temperature Superconductors

A material enters the superconducting state when the temperature falls below the material's critical (or transition) temperature, T_c . A superconductor is characterized by zero resistance, (i.e. the resistivity drops to immeasurably small quantities within a very small temperature range), and perfect diamagnetism. Diamagnetism, referred to as the Meissner effect in superconductors, prevents magnetic fields from penetrating inside a superconductor as illustrated in Figure 2.6, and is demonstrated by magnets

levitating above superconducting disks. The Meissner effect distinguishes perfect conductors from superconductors since, although both are able to prevent a magnetic field from penetrating inside their interior, only superconductors can expel a magnetic field already existing within the material.

Figure 2.6 Meissner effect in superconductors^[34]

A superconductor must not only operate below its critical temperature, T_c , but it is also limited by a critical magnetic field intensity, H_c , and a critical current density, J_c . The three parameters are interdependent according to the three-dimensional surface shown in Figure 2.7. For instance, the farther below its critical temperature a superconductor is operated, the greater the current density and magnetic field it can withstand before reverting to a normal resistivity state.

Superconducting materials can be classified as either Type I or Type II superconductors. Type I superconductors have only one critical magnetic field, below which surface currents cancel any magnetic field inside the material. Type II superconductors, such as all ceramic HTSCs, have a second, higher critical magnetic field due to the flux quantization within the material. Flux

quantization refers to the penetration of the magnetic field into the superconductor in discrete domains.

Figure 2.7 Relationship between T_c , J_c , and H_c in superconductors^[35]

Below T_c , the resistance declines to immeasurably small values due to the formation of Cooper pairs, in which two electrons interact with a phonon (lattice vibration) to lower the effective energy of the pair. Assuming an infinite momentum relaxation time (i.e. no electron scattering), classical mechanics and electromagnetics can be used to define the London equations. The relationships for the force, F , acting on a charge in the presence of an electric field, E , and the current density, J , of a number of superconducting electrons, n , are defined in equation 2.3 as

$$F = m \frac{dv}{dt} = qE \qquad J = nqv \qquad (2.3)$$

where m , q , and v represent the mass, charge, and velocity of the electrons, respectively. Combining the two equations above yields

$$\frac{dJ}{dt} = \frac{nq^2}{m} E. \quad (2.4)$$

The Maxwell equations presented below show the connection between the electric field, current density, and magnetic field.

$$\nabla \times E = -\mu_0 \frac{\partial H}{\partial t} \quad \nabla \times H = J + \frac{\partial D}{\partial t} \quad (2.5)$$

Applying the "curl" operator to both sides of equation 2.4 and substituting the first Maxwell's equation of 2.5 into the right side produces,

$$\frac{\partial}{\partial t} (\nabla \times J + \frac{1}{\lambda_L^2} H) = 0 \quad \lambda_L^2 = \frac{m}{\mu_0 n q^2}. \quad (2.6)$$

λ_L , referred to as the London penetration depth, defines the extent to which a magnetic field is able to penetrate the superconductor.

Due to the Meissner effect, the magnetic field is not only time-invariant, but zero within the superconductor. London's first equation is derived from equation 2.6 using the Meissner effect to describe the field-current interaction inside a superconductor. The second London equation follows directly from equation 2.4 by substituting the London penetration depth. The two London equations are displayed in equations 2.7.

$$\nabla \times J + \frac{1}{\lambda_L^2} H = 0 \quad \frac{dJ}{dt} = \frac{1}{\mu_0 \lambda_L^2} E \quad (2.7)$$

London's first equation can be rewritten in terms of the magnetic field only. Taking the curl of the second Maxwell's equation in 2.5 and neglecting the displacement current at relatively low frequencies, the current density term of London's first equation can be replaced. Noting that the divergence of a magnetic field is always zero, the $\mathbf{H}$ -field can be solved as a Laplace equation, expressed below.

$$\nabla^2 \mathbf{H} = \frac{\mathbf{H}}{\lambda_L^2} \quad (2.8)$$

While superconductors exhibit negligible losses at DC, losses increase in proportion to the square of frequency. This phenomenon is explained by the Two-Fluid model, as shown in Figure 2.8, which assumes that the total number of electrons, n , is divided into superconducting electrons, n_s , and normal electrons, n_n .^[34] The ratio n_s/n_n varies with temperature below T_c according to equation 2.9. The superconducting electrons flow through the inductor, while the normal electrons flow in the resistor and are responsible for the AC losses.

Figure 2.8 Two-Fluid model for superconductors

$$\frac{n_n}{n} = \left(\frac{T}{T_c}\right)^4 \quad \frac{n_s}{n} = 1 - \left(\frac{T}{T_c}\right)^4 \quad (2.9)$$

The conductivity can be described in terms of a complex quantity, $\sigma = \sigma_1 - j\sigma_2$, in which the resistive part arises from the normal electron conduction, and σ_2 represents the inertia of the superconducting electrons and depends on the London penetration depth. The condition $\sigma_2 \gg \sigma_1$ is valid for all superconductors below T_c , hence the surface impedance $Z_s = R_s + jX_s$ can be represented as,

$$R_s = \frac{1}{2} \omega^2 \mu_0^2 \lambda_L^3 \sigma_1 \quad X_s = \omega \mu_0 \lambda_L. \quad (2.10)$$

Due to the temperature dependence of the superconducting electrons, the London penetration depth is also dependent on temperature. Experiments have predicted that λ_L varies with temperature according to equation 2.11, and becomes infinitely large as the temperature approaches T_c . This reflects the kinetic inductance effect, which becomes a factor when operating superconductors close to their T_c . This will be demonstrated in the HTSC filters of Chapter 4.

$$\lambda_L(T) = \frac{\lambda_L(0)}{\sqrt{1 - (T/T_c)^4}} \quad (2.11)$$

Superconducting thin-film transmission lines exhibit very low dispersion up to and beyond microwave frequencies, due to the extremely low surface resistance. Neglecting R_s , the phase velocity in superconductors is only dependent upon λ_L and film thickness for temperatures well below T_c (i.e. $n_n/n \ll 1$). Since λ_L is independent of frequency, the dispersion can be considered negligible. This phenomenon makes superconductors attractive for use as interconnections in digital systems, delay lines, and filtering applications.

The London theory is a simple phenomenological theory which gives an accurate qualitative description of superconductors, but is less reliable in providing precise quantitative results. The BCS theory provides a more detailed, quantum mechanical explanation for superconducting phenomena. Electrons normally experience Coulomb repulsion forces when they approach each other. In a superconductor, BCS theory assumes the existence of additional attractive forces which compensate for the repulsive forces. These attractive forces arise when two electrons exchange a phonon, which is a quantum of thermal energy in the lattice vibrations, to form a Cooper pair. If all the conduction electrons form these Cooper pairs, in which the electrons have equal and opposite momentum, the energy of the system can be lowered. The average distance at which the attractive and repulsive forces balance is known as the coherence length, ξ . The coherence length can span across many other electron pairs to form Cooper pairs with these other electrons, since they all have the same quantum mechanical state.

As predicted with the Two-Fluid model, normally conducting electrons exist along with the superconducting electrons but at different energy levels. The energy gap parameter, Δ , corresponds directly to the binding energy of the Cooper pairs. It refers to the amount of energy that must be added to each superconducting electron to break up a pair and create individual normal electrons. The density of electron states in superconductors is zero inside the energy gap, peaks at the edge of the gap, and decays to the normal carrier density as the energy increases.^[34] The energy gap decreases as the operating temperature rises, and becomes zero at T_c .

The peak in the density of states gives rise to the tunneling effect which takes place in a superconductor-insulator-superconductor (SIS) junction. No current flows in the junction until a voltage exceeding the energy gap is

applied. A large surge in current is produced by the normal electrons, as opposed to the Cooper-paired superconducting electrons, as shown in the current-voltage relationship of Figure 2.9a). SIS junctions are suitable for detectors and mixers in microwave and millimeter-wave applications, such as radio astronomy in which high sensitivities are required.

Figure 2.9 Current-voltage relationship for a) SIS and b) Josephson junctions^[1]

A special type of SIS junction is the Josephson junction, in which the Cooper pair superconducting electrons are the tunneling mechanism. These junctions consist of a thinner insulating barrier to enable the Cooper pairs to flow with no applied voltage, resulting in the I-V characteristic of Figure 2.9b). The flow of current at $V=0$ is referred to as the DC Josephson effect. When a finite voltage is applied, the junction behaves as a regular SIS junction. However, a second phenomenon known as the AC Josephson effect occurs at very small voltages, whereby an alternating current is produced by the electrons to radiate this excess energy. Voltages on the order of a few microvolts will produce AC currents in the microwave frequencies, where $f = 2qV/h$. The Josephson effect is possible due to the preservation of long-range order between the two superconductors, which occurs when each superconductor maintains a fixed phase relationship between their phase functions, known as the phase coherence, δ_0 . The current flowing through a Josephson junction can be expressed as,

$$J = J_0 \sin(\delta_0 - \frac{2qV}{\hbar}t) \quad (2.12)$$

The Josephson junction equivalent circuit model is displayed in Figure 2.10, in which the total current comprises three components. The $J_0 \sin \delta_0$ component represents the current due to the Cooper pairs at zero voltage, the resistance is related to the flow of the normal electrons as in the SIS junction, and the capacitance is determined by the insulator. Josephson junctions are used in high-speed digital switching applications, and also provide the basis for Superconducting Quantum Interference Devices (SQUIDS) which are extremely sensitive magnetometers.

Figure 2.10 Equivalent circuit model for a Josephson junction^[34]

2.3.2 High Temperature Superconductors

There is a continuing debate as to whether the same theories which apply to LTSCs are also applicable to HTSCs. HTSCs exhibit the same properties as traditional superconductors such as zero resistance, the Meissner effect, flux quantization, and the Josephson effect. The BCS theory still applies in some cases, such as the formation of Cooper pairs, but cannot explain whether it is due to the electron-phonon exchange process or some other phenomenon.

Type II superconductors, including the ceramic-oxides, consist of three phases that depend on the magnetic field intensity. At sufficiently small fields, the magnetic flux is completely expelled from the material according to the Meissner effect. Above a critical field, H_{c1} , the material remains superconducting but the field partially penetrates the material in quantized flux regions. Above a second critical field, H_{c2} , the superconducting state is completely destroyed. Type II superconductors have been described using the Ginzburg-Landau theory^[36] which uses quantum mechanics to explain how magnetic fields penetrate into bulk and film superconductors, and how the magnetic structure is developed. This theory characterizes the superconductor by a wave function to describe the superconducting electrons, and a magnetic vector potential. The focus of this theory is to determine solutions for these functions which minimize the total free energy of a sample. The London penetration depth, $\lambda_L(T)$, and the coherence length, $\xi(T)$, are shown to be temperature dependent, varying proportionately to $(T_c - T)^{-1/2}$. The Ginzburg-Landau parameter, defined as $\kappa = \lambda_L(T)/\xi(T)$, is independent of temperature and determines the extent to which the long-range order of the superconductor exists. Type II superconductors are defined for $\kappa > 1/\sqrt{2}$. Ceramic superconductors such as YBCO display values of κ greater than 200 at $T=11K$, thus demonstrating their Type II tendencies.^[14]

The key to superconductivity in the copper oxides lies in the bonds between the oxygen and copper atoms. The oxygen forms a pyramidal polyhedron around the copper atoms, and it is the Cu-O planes which allow electrons to move freely through the ceramic.^[10] The YBCO has a perovskite crystal structure, as does all HTSCs, in which consecutive planes of barium are separated by alternating planes of copper oxide and held together by planes of yttrium, as shown in Figure 2.11. YBCO displays anisotropic properties in which

the coherence length is considerably larger in the a,b plane than in the c-direction of the crystal

Figure 2.11 Crystal structure for YBCO superconductors^[35]

A popular contemporary theory for explaining the superconducting phenomena in HTSCs is known as the "resonating valence bond" theory.^[37] This theory utilizes the fact that the conduction mechanism lies in the copper-oxygen planes. CuO_2 is an insulator by itself, but when mixed with the other elements (Y, Ba), the valence state of copper changes from +2 to an average of +2.33. The copper and oxygen atoms exchange electrons by means of a sea of electrons, or conduction band. When in the superconducting state, the conduction band is filled with loose electrons, leaving positively charged holes in the lattice. The holes are the primary charge carriers in ceramic superconductors.

Several other recent phenomenological theories to explain the behaviour of HTSCs have been proposed. One of these, the bipolaron model, provides the basis for software models of HTSC microstrip and coplanar lines.^[38]

This model predicts the surface impedance of the HTSC film based on the material's London penetration depth, which is temperature dependent. It also relates how the effective dielectric constant of a microstrip or coplanar line is affected by kinetic inductance. Another HTSC theory, the enhanced three-fluid model, can more accurately describe the characteristics at temperatures far below T_c .^[39] This theory incorporates a temperature dependent momentum relaxation time, τ , to predict a phenomenological "coherence peak" observed in the real part of the complex conductivity for YBCO thin-films and bulk material.

Apart from a higher T_c , HTSCs experience a higher energy gap parameter than LTSCs, which should allow for higher operating frequencies in superconductive electronics. One disadvantage of HTSCs, however, is a larger London penetration depth, resulting in an increased kinetic inductance. This is evident from equation 2.11, where λ_L becomes large as the temperature rises towards T_c . Other disadvantages of HTSCs as compared to LTSCs include lower coherence lengths and energy state densities. A comparison of the superconducting properties of niobium and YBCO is presented in Table 2.3.

	T_c	$\lambda_L(0), \mu\text{m}$	Δ (meV)	ξ (nm)	n (/m ³)
Nb	9.2K	0.07	1.5	38	14.7e27
YBCO	90K	0.22	~15	~2	2.6e27

Table 2.3 Comparison of properties for a LTSC and HTSC^[14,40]

2.4 Fabrication and Processing of HTSCs

Thin-films must exhibit several properties to ensure their suitability for practical application of microwave devices, including: availability on both sides of large area substrates; low surface resistance, R_s ; high power handling

capability; and a surface topology compatible with standard photolithographic techniques.^[41] Thin-films are deposited using various techniques, as shown in Figure 2.12, such as electron beam co-evaporation, RF sputtering, pulsed laser ablation, and molecular beam epitaxy (MBE). Films may be applied using an in-situ process in which the material is deposited onto a substrate maintained above the material's crystallization temperature. For an ex-situ process, the material is deposited onto the substrate at a temperature below the crystallization temperature. The film is crystallized in a subsequent annealing procedure.

Presently, the TBCCO thin-films have the highest reported T_c of any practical superconductors, and hence were the material of choice for circuit fabrication in this work. Thallium-based superconductors exist in several phases, such as $\text{Tl}_2\text{Ba}_2\text{CaCu}_2\text{O}_8$ (2212), $\text{Tl}_2\text{Ba}_2\text{Ca}_2\text{Cu}_3\text{O}_{10}$, and $\text{TlBa}_2\text{Ca}_2\text{Cu}_3\text{O}_9$. Thin-films for this work were fabricated at Superconductor Technologies Inc. (STI) using the $\text{Tl}_2\text{Ba}_2\text{CaCu}_2\text{O}_8$ compound, which permits for easier preparation at the expense of a slightly lower T_c . These STI thin-films are fabricated in two stages, involving the deposition of a precursor using laser ablation, and sintering under an appropriate thallium-oxygen environment. The precursor films contain stoichiometric amounts of calcium, barium, and copper, which are deposited onto the substrate. The amorphous film is then annealed at a temperature between 830°C and 900°C in the presence of a controlled Tl_2O vapor to form the TBCCO film with a 2212 phase. Special deposition facilities are required since thallium is a highly toxic element, although the TBCCO compound is believed to be more benign than its constituent elements. The sintering can be applied using a lidded or sealed tube anneal, or a flow-through furnace annealing process.^[42]

Electron Beam Evaporation

Off-Axis Sputtering

Laser Ablation

Molecular Beam Epitaxy

Figure 2.12 HTSC thin-film deposition techniques^[43]

Patterning of the thin-film HTSCs requires a process that does not degrade the film properties, is environmentally benign, and can produce the desired feature size. The most common forms of patterning include the application of stencil masks, wet etching or ion milling. STI currently uses an ion milling process which allows for line widths of $25\mu\text{m}$ and tolerances of $\pm 0.5\mu\text{m}$ to be achieved. Film thickness is specified to be $0.8\mu\text{m} \pm 0.1\mu\text{m}$ which is sufficiently greater than the London penetration depth of $0.3\mu\text{m}$ at 77K .^[44]

A key property in determining film quality is the surface resistance. R_s is proportional to the square root of frequency for conventional metals (equation 2.2), but it increases in proportion to the square of frequency for superconductors (equation 2.10). Even though R_s increases more rapidly with frequency in superconductors, it can outperform normal conductors over frequencies extending to hundreds of GigaHertz. The surface resistance of STI's TBCCO thin-films decreases rapidly at a transition temperature of typically 105K, as shown in Figure 2.13a). At 77K and a frequency of 10 GHz, the R_s of these TBCCO films is approximately 0.25m Ω , which is an improvement of almost forty times as compared to oxygen free high conductivity (OFHC) copper under the same conditions (Figure 2.13b)). The best surface resistances reported to date are 130 $\mu\Omega$ and 181 $\mu\Omega$ at 77K and 10 GHz for $Tl_2Ba_2CaCu_2O_8$ and $Tl_2Ba_2Ca_2Cu_3O_{10}$, respectively.^[41,45] Patterned films generally result in lower T_c and higher R_s compared to unpatterned films, since the films are slightly degraded in the etching process. This is especially true at the line edges, where the current density is highest.

Another factor which determines the quality and usefulness of superconductor films is the critical current, J_c . The TBCCO films from STI possess a typical J_c of 5×10^5 A/cm² at 77K. The best J_c s reported to date for TBCCO thin films at 77K is 2×10^6 A/cm².^[42] Yttrium based YBCO films have reported J_c s of 5×10^6 A/cm² and higher,^[46] hence YBCO is more suitable for high-power performance than thallium compounds, at the expense of a lower T_c and higher R_s . Higher power levels cause current levels and magnetic fields to increase towards their critical limits, thus increasing the surface resistance and lowering the Q. STI specifies typical Q values of between 12000 and 15000 for each wafer, which are tested using a 5.6 GHz microstrip resonator. A 14 GHz TBCCO thin-film linear resonator measured at CRC had an unloaded Q of 10500 at 77K,

compared to a similar gold resonator providing a Q of 470.^[47] A 20 GHz TBCCO linear resonator produced an unloaded Q of 2800, which could be improved by adjusting the cover height to further reduce radiation loss.

Figure 2.13 Performance of R_s for STI's TBCCO thin-film^[44]

Single crystal films must be grown on substrates which share the same perovskite structure and thermal expansion coefficient as the HTSCs. The TBCCO thin-films at STI are deposited onto a lanthanum aluminate (LaAlO_3)

substrate. The substrates are presently available as 2" or 3" diameter wafers, with thicknesses of 0.010" or 0.020". The dielectric constant of LaAlO_3 is typically 24, while the loss tangent, $\tan\delta$, is specified to be 3×10^{-5} .^[44] Both these substrate parameters have been shown to decrease slightly at lower temperatures.^[25] The dielectric constant and loss tangent of various substrates compatible with HTSCs are listed in Table 2.4. Due to the low loss tangent which is inversely proportional to the Q-factor, Q values of a HTSC thin-film microstrip resonator can approach 30000, and become limited mainly by radiation loss. A disadvantage of the LaAlO_3 substrate is the "twinning" problem, whereby the a,b crystal orientation is randomly interchanged. This problem occurs when the processing temperature exceeds the twinning temperature of 450°C , thus causing changes in the dielectric constant and surface roughness.

Material	Relative permittivity, ϵ_r	Dielectric loss factor, $\tan\delta$
lanthanum aluminate (LaAlO_3)	24.0	3×10^{-5}
lanthanum gallate (LaGaO_3)	23.1	10^{-4}
magnesium oxide (MgO)	9.6	4×10^{-5}
R-plane sapphire ($\alpha\text{-Al}_2\text{O}_3$)	10	1.5×10^{-8}
strontium lanthanum aluminate (SrLaAlO_4)	16.8	4×10^{-5}

Table 2.4 Properties of HTSC substrates at 77K and 10 GHz^[25,48]

Other substrates suitable for HTSC thin-films include lanthanum gallate (LaGaO_3) which possesses similar properties to LaAlO_3 , but also experiences twinning. Magnesium oxide (MgO) offers a low loss tangent and a

dielectric constant which is close to that of conventional MIC substrates, however dielectric losses are susceptible to increases due to water absorption. Sapphire substrates, which are used in MIC technology, would be ideal for HTSC applications with their extremely low loss tangent. Thermal mismatch with the HTSC thin-films which causes cracking, and the fact that the aluminum lowers the T_c during the annealing process limit its present usefulness. However, the addition of a buffer layer between the film and the sapphire substrate can reduce these effects.

During circuit fabrication at STI, ohmic contacts are deposited onto the thallium films using e-beam gold deposition, creating a bondable contact with a resistivity of $10^{-7} \Omega\text{-cm}^2$, and an insertion loss of 0.01dB at 10 GHz.^[44] A $3\mu\text{m}$ polyimide passivation layer is applied to protect the HTSC film and to allow gold crossovers. A gold film is deposited onto the ground plane to allow for substrate attachment to gold-plated kovar carriers using conductive silver epoxy, or a low melting point 150°C indium/silver solder. The backside gold is applied in a "window frame" around the perimeter of the substrate in order not to reduce the R_s and hence degrade the Q . Dicing of the circuits requires caution due to the brittleness of the LaAlO_3 substrates. The mask process thus requires various layers for circuit fabrication: an ohmic contact layer, a superconductor layer, polyimide passivation layer, gold on polyimide, and a substrate layer to define the dicing.

Computer files containing AutoCAD layouts are sent to STI via modem. The mask production, fabrication of films and dicing is performed at STI, with a normal turnaround time of six weeks. The cost of a double-sided 2" diced wafer with gold contacts is approximately \$15K(US). Upon receipt, the circuits are assembled into the kovar carriers which are attached onto a centre section of the test fixture.

2.5 Cryogenic Measurement Technique

With the improvement of low temperature refrigeration systems, it has become more feasible to cool semiconductor devices in order to achieve better performance. The potential application of HTSCs in low loss microwave circuits has provided further incentive to utilize cryogenic technology. However, a significant problem has been the accurate measurement of circuits inside cooling systems. This requires proper calibration of the network analyzer at low temperatures in order to remove the effects of the device test fixture and external cables. This section describes the Through/Reflect/Line (TRL) calibration technique, as well as its application for low temperature measurements. The cryogenic measurement apparatus used in this work is also described in detail.

2.5.1 TRL Calibration Procedure^[49]

Calibration involves the cancellation of systematic errors in a measurement system in order to obtain accurate measurements. These systematic effects such as leakage, test port mismatch, and frequency response at microwave frequencies will affect measured data. The effects are repeatable, and therefore can be determined for a network analyzer by measuring a series of known standards. The systematic effects are contained in a twelve-term error model which performs an "error-correction" to subsequent measurements. In a conventional two-port calibration, three known impedances and one transmission standard are required. Imperfections in the calibration standards cause residual effects which are the portion of the uncorrected systematic error that remains after calibration is performed.

TRL is a two-port calibration technique that relies on transmission lines rather than a set of impedance standards. Transmission lines offer a number of advantages in that they are the simplest elements to realize in non-coaxial media such as microstrip. Also, the impedance of transmission lines can be

accurately determined from the physical dimensions and materials. The basic TRL calibration process is displayed in Figure 2.14. A total of sixteen measurements are required to determine the twelve unknowns. In a perfectly balanced test system which ignores the different terminating impedances as the RF switch changes between ports 1 and 2, only eight error terms are required.

Figure 2.14 TRL procedure and ideal S-parameters for each standard^[49]

The TRL procedure initially assumes a balanced system in which a through connection is made between ports 1 and 2 with a short length of transmission line. The transmission frequency response and port match are measured in both directions to produce four measurements. Next, an open or short reflection standard is connected to each port and the two reflection coefficient measurements are performed. Lastly, a known length of transmission line relative to the through is connected between port 1 and 2 such that the four measurements of frequency response and port match in each direction can be made. These ten measurements result in ten equations, but since only eight

unknowns exist in the balanced system, the Γ of the reflect standard and the propagation constant, γ , of the line standard are determined as constants. To correct for the different terminating impedances at the RF switch, the ratio of incident signals are computed during the through and line standard measurements. All twelve terms in the error model have thus been accounted for.

2.5.2 Cryogenic Apparatus

Figure 2.15 Cold head of the cryocooler apparatus^[47]

The cryogenic apparatus consists of a customized cold head on an RMC Cryosystems LTS-22-IR helium refrigerator. The refrigerator, or cryocooler, is capable of cooling a 0.25 Watt thermal load to temperatures below 10K. The cryocooler is integrated with the HP8510C or Wiltron 360 network analyzer to

perform accurate cryogenic measurements up to 30 GHz. Semi-rigid 0.118" coaxial K-cables connect a split-block test fixture to coaxial feedthroughs which are connected to the test ports of the network analyzer. The cryocooler can achieve a minimum temperature of 45K within one hour when the two halves of the test fixture are directly connected together. A diagram of the cryocooler based apparatus is illustrated in Figure 2.15.

Figure 2.16 Refrigeration system for cryocooler setup^[47]

The cooling system operates by circulating high pressure helium gas into and out of the cold head according to the Gifford-McMahon cycle, as shown in the schematic diagram of Figure 2.16. As the gas enters through valve A, it expands and absorbs heat. The resulting low pressure gas returns through valve B to the water cooled compressor where it gives up its heat. In order to achieve extremely low temperatures, the cold head is evacuated using a vacuum pump to reduce conduction and convection heat loss. When performing a measurement, the temperature is stabilized using a silicon diode thermometer

and a resistance heater mounted on the second stage cold station. When returning to room temperature, dry nitrogen is supplied into the cold head to increase the heat transfer and to prevent condensation from forming on the test fixture substrates.

2.5.3 Cryogenic Measurement Technique

The TRL calibration technique allows the effects of the test fixture's microstrip lines, microstrip-to-coaxial launchers, and interconnecting cables to be de-embedded at cryogenic temperatures.^[50] Only the effects of the bond wires connecting the device under test are not mathematically removed. The gold microstrip lines, having a characteristic impedance of 50Ω , are patterned onto 0.010" thick alumina substrates to form the standards as well as the input and output lines of the measurement test fixtures. The standards include a zero-length through, open circuits on the input and output ports, and a 0.050" long line standard. The line standard provides an 11ps delay and is designed for calibration of frequencies up to 40.5 GHz. The alumina substrates and centre pins of the K-connector launchers are attached using an indium/lead solder to ensure a solid connection at 77K. Test fixtures are fabricated using Thermkon 70M (85% Mo, 15% Cu) which possesses approximately the same thermal expansion coefficient as alumina.^[51] This alloy is easier to machine than kovar and has better thermal conductivity, but requires a longer gold plating procedure. A photograph of the split-block test fixture is displayed in Figure 2.17.

Figure 2.17 Photograph of Thermkon split-block test fixture

To perform a TRL calibration on the network analyzer, the through standard is formed by connecting the two halves of the test fixture directly together. Next, the line standard is mounted onto a shim and sandwiched between the halves. The through and line standards are connected using gold ribbon bonds to ensure solid connections at 77K. Finally, the open reflection standards are attained by separating the two halves of the test fixture. The cooling cycle requires approximately fifty minutes for the test fixture to reach 77K, and thirty minutes to reheat back to room temperature. To complete a full two-port calibration, roughly five hours is needed for the three cooling cycles plus wire bonding time. Thereafter, each circuit requires up to two hours for the assembly, cooling cycle, and S-parameter measurements.

Results of a calibration over a 1 GHz to 26 GHz frequency range are displayed in Figure 2.18, in which the through standard was remeasured after

the TRL calibration. The transmission response shows a deviation of within $\pm 0.05\text{dB}$ and one degree in the angle, while the return losses yield better than 40dB over the entire range. Ideal conditions for the through should produce a transmission response of $S_{21} = 1\angle 0^\circ$ and input and output reflections of $S_{11}=S_{22}=0$. Deviations from the ideal conditions are mostly due to variations in the bond wire connections as well as the drift of the network analyzer during the long calibration and measurement cycles.

Figure 2.18 Remeasurement of "through" standard after calibration

Measuring the same standard used to calibrate does not reflect the true accuracy of this method since they have been defined by the calibration to be ideal. Therefore, a 0.101" long microstrip line which is twice the length of the line standard was measured. The results in Figure 2.19 show that the insertion loss is better than 0.2dB up to 20 GHz, while the return losses are below 26dB for the input and output. Sources of error are due to bond wires and network analyzer drift, as mentioned above, along with the longer transmission line.

Figure 2.19 Measurement of a 0.101" line to verify calibration

This TRL calibration technique provides accurate, broadband, de-embedded S-parameters of microwave devices operated at cryogenic temperatures.^[52] To reduce the length of calibration time, as well as to eliminate the need for bond wire connections, a cryogenic wafer prober is currently being

developed in-house at the Communications Research Centre. The probing station will allow on-wafer measurement of HTSC and semiconductor circuits up to the millimeter-wave frequencies, at liquid nitrogen temperatures (77K).

CHAPTER 3: CRYOGENIC AMPLIFIER DESIGN AND EXPERIMENTAL RESULTS

3.1 Introduction

Cryogenic cooling is used in microwave receiver systems to achieve better performance in low noise amplifiers and mixers. In radio astronomy, for instance, cryogenic temperatures are required to minimize system noise to levels comparable with the natural limits of galactic, cosmic, and atmospheric noise. Parametric amplifiers and MASERS have been used to achieve superior noise performance to traditional MESFETs. With the advent of the High Electron Mobility Transistor (HEMT), performance comparable to paramps can be achieved at a reduction in cost. The HEMT, as well as the Pseudomorphic HEMT (PHEMT), provides much better noise performance and higher operating frequency than the MESFET. These advantages are even more pronounced at cryogenic temperatures where the modulation doping of HEMTs can be further exploited.

In this chapter, the development of a hybrid high- T_c superconductor/semiconductor K-band amplifier is described. A brief description of the principle of operation of HEMTs and PHEMTs is presented, along with a small-signal device model for a HEMT at room temperature and 77K. Based on this model, an amplifier using HTSC passive structures for matching has been designed. The final section presents measured results of the amplifier and compares these to simulations. An analysis of the ability of various CAD software packages to predict the performance of HTSC passive structures is also presented.

3.2 Device Characterization at 77K

The HEMT device was invented in 1980 by researchers at Thomson and Fujitsu simultaneously.^[53] The structure is physically identical to the MESFET except for the active layer doping profile. The active layer is modulation doped in the vertical dimension to create a "super lattice" to give the HEMT its superior performance. The active layer contains a combination of doped AlGaAs, a thin layer of undoped AlGaAs, and pure GaAs. A heterojunction forms at the AlGaAs/GaAs interface as shown in Figure 3.1a). Electrons in the doped, higher bandgap AlGaAs migrate through the thin layer of pure AlGaAs into the pure GaAs. These carriers form a two-dimensional electron gas free of coulomb scattering effects which are not obstructed by collisions with donor ions and hence produce a higher mobility. The pseudomorphic HEMT or PHEMT is an extension of the HEMT, wherein the undoped GaAs layer is replaced with the lower bandgap material InGaAs as shown in Figure 3.1b). This results in a further increase of the electron density and mobility of the two-dimensional electron gas.

Figure 3.1 Cross-section of the structure for a) HEMT^[22] and b) PHEMT^[54]

Cooling FETs and HEMTs improves performance due to higher carrier mobility, larger resistance in the semi-insulating substrate, and higher conductivity in the active region. With higher mobility, the electron velocity

increases, resulting in a higher transconductance, g_m , for the device. At the cutoff frequency, the input current is equal to the output current, and is expressed in terms of the device model elements as,

$$f_T = \frac{g_m}{2\pi C_{gs}} \quad (3.1)$$

Figure 3.2 Mobility vs temperature for MESFET and HEMT channel structures^[27]

From equation 3.1, it is evident that a higher f_T will result from a higher transconductance. The electron mobility improves at cold temperatures due to the reduction of lattice vibrations. The mobility versus temperature is displayed in Figure 3.2 for both a modulation doped GaAs HEMT and uniform doped GaAs FET. This illustrates the benefit of the undoped GaAs layer in HEMTs for

electron mobility at lower temperatures. Another advantage of cooling the device is a higher conductivity, which decreases the drain-to-source resistance, thereby increasing the frequency of operation and reducing the noise. At temperatures below ~50K, however, extensive dopant freeze-out occurs which severely degrades device performance. The higher resistance of the semi-insulating GaAs at low temperatures helps reduce circuit parasitics such as substrate leakage currents.

Perhaps the most important advantage of cooling devices is seen in the noise performance. Noise figure, NF, is defined as the input signal-to-noise ratio divided by the output signal-to-noise ratio,

$$NF = \frac{SNR_{in}}{SNR_{out}}. \quad (3.2)$$

The noise characteristics for a FET or HEMT depend on four noise parameters: the minimum noise figure, NF_{min} , optimum noise admittance, $(g_o + jb_o)$, and the equivalent noise resistance, r_n . The noise figure can be expressed in terms of these parameters as

$$NF = NF_{min} + \frac{r_n}{g_o} ((g_s - g_o)^2 + (b_s - b_o)^2). \quad (3.3)$$

The minimum noise figure occurs when the source admittance $(g_s + jb_s)$ equals the optimum noise admittance, and is shown as,

$$NF_{min} = 1 + 2\pi K_f f C_{gs} \sqrt{\frac{R_s + R_g}{g_m}} \quad (3.4)$$

where K_f is a fitting factor to represent the quality of the channel material. From this equation, one can see that the higher transconductance and lower gate resistance of HEMTs at lower temperatures results in a lower NF_{min} .

The basic MESFET small-signal equivalent circuit is displayed in Figure 3.3. The main intrinsic elements of this model are the gate-to-source capacitance, C_{gs} , the drain-to-source and channel resistances, R_{ds} and R_i , and the transconductance, $g_m = \partial I_{ds} / \partial V_{gs}$. The extrinsic (parasitic) elements include the source resistance, R_s , the drain resistance, R_d , the gate resistance, R_g , and the drain-to-source capacitance, C_{ds} .

Figure 3.3 Small-signal equivalent circuit model for a MESFET^[55]

A small-signal model for the $0.25\mu\text{m} \times 100\mu\text{m}$ Fujitsu FHR10X HEMT has been derived by Smuk et al at both room temperature (300K) and 77K.^[52] This model uses FETlink software which incorporates the standard FET model, as shown in Figure 3.3, with additional parasitics. The model is based on previous S-parameter measurements from 1 GHz to 20 GHz at bias conditions of 2V drain-to-source and 5mA drain current, in a source ground configuration. The device model was found to provide an accurate agreement to measured data.^[52]

De-embedding was performed to the edge of the test fixture but did not include wire bonds.

	297 K	77 K
gm (mS)	26.9	42.5
Cgs (pf)	0.116	0.113
Ggs (mS)	0	0.077
Rds (Ω)	334	232
Ri (Ω)	2.2	0
τ (ps)	1.9	0.8
Cgd (pf)	0.015	0.021
Cds (pf)	0.058	0.063
Lg (nH)	0.135	0.161
Rg (Ω)	1.0	0.5
Ld (nH)	0.089	0.148
Rd (Ω)	2.8	5.6
Ls (nH)	0.021	0.024
Rs (Ω)	0.8	1.0
f _T (GHz)	32.7	50.5

Table 3.1 Element values for small-signal FHR10X HEMT model^[52]

Figure 3.4 S-parameters derived from a HEMT model at 300K and 77K

Element values are displayed in Table 3.1 for both 77K and room temperatures. Note that by cooling the device, the transconductance, g_m , showed a 58% increase, while the drain-to-source resistance, R_{ds} , dropped by 30%. The cutoff frequency for the device increased more than 50% to 50 GHz. A plot of the S-parameters is shown in Figure 3.4. At 20 GHz, the transmission gain increased by 2.5dB after cooling, while the output reflection coefficient dropped by 1.8dB relative to room temperature.

In this work, the addition of source inductance to improve the device stability was investigated. This source inductance provides series feedback to reduce the input and output VSWR. By adding 0.1nH source inductance to the FHR10X HEMT model, the stability factor, K , of the device increased from 0.54 to 1.1 at 20 GHz, while the gain dropped by 0.6dB. (The stability factor is defined in terms of the scattering parameters, and must be greater than one for a transistor to be unconditionally stable).

Two sets of measurements at 77K were also performed on the device to confirm the effects of different bond wire source inductance. The increase in inductance was achieved by lengthening, and reducing the number of bond wires from the source to ground. As a rule of thumb, the 0.0007" diameter gold bond wires have an approximate inductance of 0.8 nH/mm. The measurements were taken from 1 GHz to 26 GHz, and the S-parameters and stability factor are displayed in Figure 3.5, along with those of the model (FHRMODL2 and FHRMEAS2 contain 0.1nH added source inductance). The discrepancy between the measurements and model occur due to the fact a different device was used in each case. Slight variations in the processing conditions for different batches of devices can alter their characteristics slightly. However, the general trend of increasing stability, and larger reduction in gain over frequency is apparent in both cases.

Figure 3.5 Effect of source inductance on HEMT performance

Certain cooled HEMT devices are known to be sensitive to light. These HEMTs at cryogenic temperatures exhibit superior performance only under illuminated conditions in which they are time-invariant, memoryless devices.^[56] In darkness, the threshold voltage shifts and distortion of the I-V

characteristics occurs. Illuminating, or heating the device up to approximately 130K restores the initial device characteristics. For a Fujitsu FHR01X HEMT, it was reported that illumination affects the I-V characteristics at relatively low drain voltages, while for $V_{DS} > 2V$ the effect is minimal.^[57] The collapse of the I-V characteristic in darkness is attributed to the existence of DX centres in the AlGaAs layer giving rise to trapping effects. A DX centre is a deep donor level in which electrons freeze out, or are "trapped".

Measurements of a Fujitsu FHR10X HEMT were taken in both a partially illuminated (i.e. cryocooler partially shields device from fluorescent and natural light) and dark environment. When the device was cooled to 77K under partial illumination, a gate voltage of $V_{GS} = -0.564V$ was required to achieve 5mA drain current at $V_{DS} = 2V$. When the illumination was removed, the drain current increased to 7mA at the same V_{GS} . A second cooling cycle was performed in which the HEMT was in complete darkness. To achieve $I_{DS} = 5mA$ at $V_{DS} = 2V$, a slightly lower gate voltage of $-0.584V$ was required. Subsequently exposing the device to illumination caused no immediate appreciable changes in the bias conditions. For the amplifier designs, S-parameter measurements were performed under partial illumination to remove the possibility of electron trapping, and to maintain a consistent operating environment.

3.3 Amplifier Design Procedure

Low noise amplifiers are an integral component of microwave receiver systems in minimizing the system noise figure. In very low noise receivers, cryogenic amplifiers are implemented to achieve further noise figure reduction. Not only does cooling the LNA improve the noise temperature, it also increases the transmission gain and provides a controlled temperature operating environment. Cryogenic temperatures are also ideal for combining

semiconductor devices with superconducting passive structures. In an amplifier, for instance, the input and output matching circuits could be fabricated using superconductors. This would reduce the losses of the matching networks, and therefore lower the overall amplifier noise figure.

A single-stage K-band amplifier was designed for maximum gain using microstrip HTSC matching networks. Although the design does not utilize noise parameters to achieve optimum noise performance, the amplifier is inherently low noise due to operation at cryogenic temperatures. The procedure for obtaining noise parameters at low temperatures requires a tuning mechanism which would be difficult and lengthy to implement in the present cryogenic system. The overall circuit complexity was kept to a minimum.

A narrowband design configuration was used in which the input and output matching networks consisted of single open-circuited shunt stubs. The microstrip line widths were determined using LineCalcTM, and are dependent on the characteristic impedance, the effective dielectric constant, and the substrate thickness. To achieve optimum gain at the centre frequency, the stubs were designed to conjugate match the device using S-parameter data from the FHR10X HEMT model at 77K. Standard impedance matching techniques for single-stubs were used as follows:^[58]

- 1) The complex conjugate of the input reflection coefficient, S_{11}^* , at 20 GHz was located on the Smith Chart.
- 2) A rotation along the constant VSWR circle was made in the direction of the load until it intersected the $\tilde{G}=1$ circle.
- 3) The distance traversed in wavelengths around the Smith Chart determined the placement of the stub away from the device.

- 4) From the $\tilde{B}=0$ point, a rotation was completed around the outer perimeter of the Smith Chart until the admittance matched that from step 2). This rotation in wavelengths corresponded to the length of the stub.

The same procedure was repeated for the output reflection coefficient, S_{22}^* , in the output matching network.

The initial transmission lines and stub lengths were optimized using Touchstone™ to achieve minimum input and output reflections at 20 GHz. The final optimized values of the input matching network deviated by only 2% of the initial predictions. The output matching stub varied over 50% from its original value, indicating a high degree of insensitivity to changes in length.

Figure 3.6 Photograph of assembled HTSC/HEMT amplifier

The HTSC matching networks were fabricated at Superconductor Technologies Incorporated (STI) using TBCCO thin films grown on both sides of a 254 μm thick lanthanum aluminate substrate, as described in Chapter 2. The

TBCCO film becomes superconducting at around 105K, and measurements should be carried out well below this transition temperature (77K in this case) in order to avoid kinetic inductance effects. Circuit assembly must therefore provide a very good thermal contact and thermal expansion match between the circuit and carrier. Carriers were made of gold-plated kovar, with the HTSC circuits attached using either a low temperature silver epoxy or indium solder. The silver epoxy compound cures in five minutes at 150°C, which should not degrade the TBCCO thin films. The input and output matching circuits were epoxied onto the carrier, on either side of a centre pedestal protruding from the carrier. The semiconductor device was then mounted onto this pedestal using silver epoxy. The gate and drain of the device were connected to the gold contact pads on the matching networks using 0.0007" diameter gold wire bonds. The source pads of the HEMT were grounded to the carrier via wire bonds. A photograph of the assembled amplifier is shown in Figure 3.6.

The bias circuitry consists of high impedance/low impedance quarter-wave transmission line networks on the gate and drain side. These networks present an effective open circuit at the design frequency to prevent the signal from propagating into the bias circuit. The high impedance line has a width of 25 μ m, which is the minimum allowable feature size according to STI's specifications.^[44] To simplify the analysis and assist in minimizing reflections, the bias feed line was placed opposite the shunt stub. In the middle of each of these networks, a 50 Ω resistor and a 100pf capacitor were placed in series and grounded to the carrier to help stabilize the amplifier at the lower frequencies. A 2.2k Ω resistor was added in the bias circuit on the gate side to protect against excessive gate currents. The laser trimmed chip resistors and multilayer chip capacitors were attached to the substrate using the low temperature silver epoxy. The resistor and capacitor in the stabilizing configuration were connected with

wire bonds, and the resistor was grounded to the carrier using epoxy. DC biasing was applied to the gate and drain via ultrasonic bonded 0.0005" x 0.003" gold ribbons. These ribbons connect to two filtercons attached to the centre section of the test fixture, which in turn connect to the power supplies through phosphor-bronze bias lines in the cryogenic test setup. Due to the emphasis on design simplicity for these amplifiers, DC blocking capacitors on the input and output 50 Ω transmission lines were omitted. Hence, it was also possible to apply the bias directly through the network analyzer.

Two slightly different variations of the amplifier were implemented, designated as amplifiers 'A' and 'B'. Amplifier 'A' was designed and fabricated according to the preceding procedure. A schematic diagram of this amplifier is presented in Appendix I. However, due to a shortage of Fujitsu FHR10X HEMTs, and with the availability of PHEMT devices fabricated at Comsat, it was decided to substitute a PHEMT into this amplifier. While the design was based on a small-signal model for a HEMT, the PHEMT used had identical gate dimensions and also offered the possibility of better performance. At 20 GHz, the gain and stability of a measured PHEMT device, as compared to the HEMT model, is displayed in Table 3.2. Amplifier 'B' contained alterations to the matching networks, based on empirical modifications of a previous HTSC/HEMT LNA.^[17] This involved adjusting the lengths of the open-circuited stubs as well as the transmission line between the stubs and the transistor. Results of both amplifiers are presented in the following section of this chapter.

	$ S_{21} $ dB	K
PHEMT	6.5	1.13
HEMT	6.0	1.08

Table 3.2 Comparison of HEMT and PHEMT devices at 20 GHz

A second iteration amplifier design (amplifier C) undertaken at a later date attempted to improve on the stability. This new design incorporated a resistor on the drain side to stabilize the device at the expense of transmission gain. The matching networks were designed in the same manner as the previous amplifiers, with the exception that S-parameter data from the measured HEMT in Figure 3.5 with higher source inductance was used. Another feature incorporated into this design was the addition of DC blocking capacitors. The results of this amplifier are presented in the next section. A brief description of each of the three amplifiers is provided in Table 3.3.

Amplifier	Device	Stabilizing Resistor
A	PHEMT	No
B	HEMT	No
C	HEMT	Yes

Table 3.3 Comparison of design characteristics for three amplifiers

3.4 Measurements and Results

This section presents S-parameter measurements at 77K of the amplifiers described in the previous section and compares the results to simulations using TouchstoneTM. The original set of amplifiers (A and B) were designed specifically for maximum gain at 20 GHz. Noise figure results are also presented for amplifier 'B'. The later version amplifier 'C' was designed to address the problem of stability in greater depth. An analysis of the individual matching networks using various CAD software packages for amplifier 'C', as well as a previous amplifier,^[17] were conducted.

S-parameter measurements were performed using the cryogenic technique described in section 2.5. The amplifier was inserted between the two halves of the test fixture and connected with wire bonds, as shown in Figure 3.7. A significant step in line width occurs between the 50Ω transmission lines on the alumina substrates and the input and output lines on the high dielectric, superconductor substrates. To minimize any reflections due to the step junction, tapered contact pads were used at the amplifier input and output. A Through-Reflect-Line (TRL) calibration was performed as described in section 2.5 over a 1 GHz to 26 GHz frequency range. Measurements were performed using either the HP8510C or the Wiltron 360 Automatic Network Analyzer.

Figure 3.7 Photograph of HTSC/HEMT amplifier 'B' in test fixture

3.4.1 HTSC/PHEMT Amplifier 'A' Results

The amplifier based on a $0.25\mu\text{m} \times 100\mu\text{m}$ Comsat PHEMT device was measured at a bias condition of 2V drain-to-source voltage and 6mA drain current. These are the specified room temperature bias values for optimum noise performance. The measured amplifier gain $|S_{21}|$ in dB is displayed in the graphs of Figure 3.8, along with predicted results. The gain reaches a peak value of 13dB at 19 GHz with corresponding input and output reflections of -14dB and -7dB, respectively. The discrepancy in centre frequency between the original design and this measurement is attributed to the use of a different type of device (PHEMT vs. HEMT). The Touchstone simulation predicts a slightly lower gain peak which is shifted down in frequency by 500 MHz. This simulation incorporates device measurements of a quarter micron gate length PHEMT at 77K along with simulated HTSC matching and bias networks.

The amplifier response was found to be extremely sensitive to the value of source inductance. By varying the source inductance slightly in the simulation to approximate the inductance caused by the wire bonds, a closer fit between simulation and measurement may be found. Lowering the source inductance will cause a downward shift in frequency with a corresponding increase in gain and decrease in stability factor, K . The discrepancy between simulation and measurement can be further attributed to the use of a different device in each case.

Figure 3.8 a) and b) Measured vs. predicted matching for amplifier 'A'

Figure 3.8c) Measured vs. predicted gain of amplifier 'A'

A second gain peak occurs at around 8 GHz in both the measurement and simulation. This secondary peak indicates a potentially unstable device, as confirmed by the input reflection coefficient, $|S_{11}|$ which shows a tendency towards oscillations. This was likely caused by extraneous effects in the bias circuit. An attempt was made in the second iteration amplifier 'C' design to eliminate this instability problem.

3.4.2 HTSC/HEMT Amplifier 'B' Results

Amplifier 'B' was designed and fabricated using the original Fujitsu HEMT device with slight variations in the open-circuited stub lengths. These stubs were lengthened and placed closer to the transistor to shift the gain peak upwards in frequency. Results of this amplifier over a 1 GHz to 25 GHz frequency range are displayed in Figure 3.9. The transistor was biased at its room temperature optimal noise bias condition of 2V drain-to-source voltage and 5mA drain current. The gain reaches a peak value of 7.8dB at 20.5 GHz.

Corresponding return losses are -7.5dB on the input and -9dB on the output. Touchstone predicted a maximum gain of 9.7dB at 20 GHz with input and output matching below -15dB. The lower gain peak and higher frequency of the measured amplifier is attributed to a higher source inductance for increased stability. Touchstone predicted that an increase of 0.05nH source inductance for the simulated amplifier would more closely match the measurements. A secondary peak in gain emerges at the lower frequencies (10 GHz) which is again attributed to effects in the bias circuit.

Figure 3.9a) Measured vs. predicted gain of amplifier 'B'

Figure 3.9b) and c) Measured vs. predicted matching of amplifier 'B'

Noise figure was measured for this amplifier using the setup shown in Figure 3.10. An HP346C noise source was connected to the input vacuum feedthrough port of the cryogenic chamber. The output vacuum

feedthrough was connected to the HP8970 noise figure meter via an isolator and external mixer. An LO signal is generated using the 83570A RF plug-in connected to a sweep oscillator. The losses of the external mixer and cables are removed by directly connecting the noise source to the isolator/mixer on the output and performing a through calibration. The noise diode was then connected to the device under test (DUT) so the noise figure, NF_{meas} , could be measured. This value includes the noise contribution from the K-cables and test fixture inside the chamber as well as the DUT.

Figure 3.10 Block diagram for noise figure measurement setup

Using the standard expression for cascaded noise figures, the measured noise figure can be written as,

$$NF_{meas} = NF_i + (NF_{DUT} - 1)L_i + \frac{(NF_o - 1)L_i}{G_{DUT}} \quad (3.5)$$

where $NF_i = L_i$ corresponds to the noise figure of the input losses and NF_o is the noise figure of the output losses. The DUT has a gain, G_{DUT} , and noise figure, NF_{DUT} . The losses due to the input and output K-cables and split-block test fixture were characterized by connecting the two halves of the test fixture together to form a through, and measuring the transmission response using the

network analyzer. As the K-cables are of equal length, half of this loss, $L=L_1=L_2$, is attributed to the input and thus directly contributes to the overall noise figure. The noise contribution of the output side is reduced by the DUT gain. The corrected DUT noise figure is obtained by rearranging equation 3.5 to give,

$$NF_{DUT} = \frac{NF_{meas}}{L} - \frac{(L-1)}{G_{DUT}} \quad (3.6)$$

The noise figure of the HTSC/HEMT amplifier 'B' was found indirectly by determining the noise figure of the hybrid filter/amplifier circuit, described in Chapter 4. The noise figure of the stand-alone amplifier was determined by subtracting the noise figure of the filter (i.e. insertion loss in dB) from the total filter/amplifier noise figure (in dB). At 19.8 GHz, and biasing of 5mA drain current at 2V, the amplifier noise figure was found to be 1.0dB. By lowering the drain current to 3mA, a minimum noise figure of 0.8dB was extracted for this amplifier. Note that these values were measured at frequencies slightly offset from the amplifier peak, and therefore may not be the true minimum values.

An attempt was made at directly measuring the noise figure of the amplifier at the gain peak of 20.5 GHz. This was rather inaccurate, however, due to relatively large variations above 20 GHz in the transmission response, $|S_{21}|$, of the through loss in the K-cables and test fixture. These measurements between 18 GHz and 21 GHz are shown in Figure 3.11. Also, the necessity for using bias T's as a DC block further decreased the measurement accuracy. For this reason, no noise figure measurements were performed for the HTSC/HHEMT amplifier 'A'.

LOG MAG.

REF--4.000dB

1.000dB/DIU

S21 FORWARD TRANSMISSION

Figure 3.11 Measurement of "through" loss in K-cables and test fixture

3.4.3 2nd Iteration Amplifier 'C'

A second iteration amplifier design was implemented to achieve an improvement in the stability of the device and remove the secondary gain peak present at lower frequencies. This involved adding a 300Ω shunt resistor on the drain side of the device to suppress the gain at frequencies outside the passband, without degrading the noise figure. The resistor was connected to the 50Ω output line via a 1.9mm transmission line, and grounded to the carrier using conductive epoxy. A schematic diagram of this amplifier is illustrated in Appendix II.

The measured S-parameters displayed in Figure 3.12 show a peak gain of 7dB at 15 GHz. This does not agree with the predicted peak response of 9dB at 20 GHz. While the input reflection at 20 GHz matched the predicted response, the measured output reflection yielded a minimum of -18dB at 15 GHz. The poor matching at the output caused the significant shift in the peak. This is

attributed to inaccuracies in modelling the resistive stabilizing network. For instance, the laser-trimmed chip resistor (0.9mm x 0.5mm) was found to display an appreciable reactive component at these frequencies. Re-simulating the amplifier in Touchstone with this added reactance provided very close agreement with the measurements. A 5% increase in the DC resistance at 77K as compared to the room temperature value had a minimal effect on the amplifier response.

Figure 3.12 Measured S-parameters for 2nd iteration amplifier 'C'

To verify the performance of the matching networks, separate test jigs were created to allow for measurement of the individual matching circuits. Measurements and the corresponding Touchstone simulations are illustrated in

the Smith Chart of Figure 3.13. While a phase shift is evident for the input matching network, the output matching circuit consisting of the stabilizing network and matching stub displays some resonance effect. This resonance is caused by the reactance of the stabilizing chip resistor, as explained above. These matching circuits contained slight deviations in their design as compared to those of the measured amplifier.

Figure 3.13 Performance of input and output matching networks

3.4.4 CAD Software Analysis of HTSC Circuits

A comparison of various CAD software packages was undertaken to simulate single-stub HTSC matching networks for a 14 GHz narrowband HTSC/HEMT amplifier designed at the Communications Research Centre (CRC).^[17] S-parameter measurements of separate input and output matching networks were compared to simulations using Touchstone, EMSimTM by EEsof, and Sonnet's EMTM. Both EMSim and EM employ a full-wave analysis technique,

while Touchstone utilizes a quasi-static approach. The full-wave analysis considers the effects of coupling and surface waves, as well as radiation loss. The results from 12 GHz to 16 GHz are displayed on the Smith Chart in Figure 3.14. For the input circuit, EMSim provides the closest match to the measurements. Touchstone offered a better prediction of the output matching network. It should be noted that the external bias circuitry was omitted for the EMSim and EM simulations for simplicity of analysis. While EMSim required only 24 patching segments to simulate the input matching, EM used 142 segments and thus required a much longer computation time. Touchstone, with its wide range of modelling elements and superior speed, proved adequate for analysis of these simple HTSC passive structures.

Figure 3.14a) Results of input HTSC matching network

Figure 3.14b) Results of output HTSC matching network

3.5 Summary

This chapter presented the design, fabrication and testing of three K-band HTSC/HEMT amplifiers. The amplifiers were based on a small-signal model of a FHR10X device, which predicted a 2.5dB increase in gain at 20 GHz and 77K, as compared to room temperature operation. The device stability was highly sensitive to the source inductance caused by source-to-ground bond wires. Increasing the source inductance by 0.1nH in the device model to approximate the longer bond wires improved the stability factor, K , from 0.54 to 1.08 at 20 GHz and 77K.

The amplifiers were designed for maximum gain by implementing high temperature superconductor matching networks at the device input and

output. Amplifier 'A' achieved a passband gain of 13dB at 19 GHz, with corresponding matching of -14dB and -7dB on the input and output, respectively. Amplifier 'B' produced 7.8dB gain at 20.5 GHz, and noise figures of 1.0dB with 5mA drain current and 0.8dB with 3mA drain current at 19.8 GHz. The noise figures for the amplifier were extracted from measurements of the HTSC filter/amplifier integrated unit. These values agree with a Ku-band HTSC/HEMT amplifier developed at CRC which reported a minimum noise figure of 0.4dB at 13.2 GHz.^[17] The results exceed the performance of a two-stage, packaged FHR10X HEMT LNA developed at CRC which produced 1.9dB noise figure at 20 GHz and 297K.^[59]

A second iteration HTSC/HEMT amplifier 'C' was designed for improved stability by including a shunt resistor at the transistor output. The measured response displayed a significant shift in the passband from 20 GHz to 15 GHz. These erroneous results are mainly attributed to inaccuracies in modelling the chip resistor at high frequencies and 77K.

CHAPTER 4: INTEGRATED SUPERCONDUCTING FILTER/AMPLIFIER PERFORMANCE

4.1 Introduction

High Temperature Superconductors (HTSCs) have produced very high Q resonators due to their extremely low loss properties.^[11,47,60] These achievements in the quality factor make HTSC materials desirable for very narrow bandwidth filters. Thin-film planar circuits using conventional metals are considered too lossy to achieve the demanding specifications on bandwidth, such as those required in satellite communications. Now, filters fabricated using HTSC planar technology such as microstrip and coplanar waveguide can achieve comparable performance to waveguide filters at a significant reduction in size. The cost of cryogenically cooling microwave components can be offset by implementing many HTSC components in a filter bank, for instance. Filter banks are frequently used in wide-band radar and communication systems to select a much narrower bandwidth signal. The current use of waveguide technology in filter banks requires a relatively large volume.

In microwave receiver systems, the low noise amplifier is the key component in reducing the noise figure and improving the sensitivity of the system. At the receiver input, the amplifier reduces the noise contributions of all following receiver components such as mixers, filters and IF amplifiers. However, to avoid intermodulation products by signals outside the desired frequency band, a bandpass filter at the receiver input is desired. The losses due to this filter will directly affect the overall noise figure. By implementing a HTSC filter, however, the insertion loss can be reduced, thus lowering the noise figure.

The first two sections of this chapter present the design, implementation, and testing of high- T_c superconducting microstrip filters. The

filter responses at 77K are presented and compared to predicted results using Touchstone and EMSim. The effect of temperature variation on filter response is investigated. Also, an analysis on the individual resonator sections of the filters is performed to determine the coupling, as well as the accuracy of these CAD software packages. Measurements and simulations are presented later in the chapter.

The final section of the chapter reports the integration of the HTSC filters with their respective amplifiers described in the previous chapter. The S-parameter and noise figure measurements are presented along with predicted results for both hybrid circuits. The problem of stability due to mismatch between filter and amplifier is addressed.

4.2 HTSC Filter Design Procedure

Two 3-pole Chebyshev filters were designed to produce a narrowband (2.3%) response at centre frequencies of 19 GHz and 20 GHz to coincide with amplifiers 'A' and 'B', respectively, described in Chapter 3. Both designs implement quarter-wave parallel-coupled microstrip resonators to achieve a passband ripple of 0.01dB. A three-pole configuration with a moderately narrow bandwidth was chosen for simplicity of design and implementation, as this was a first iteration design using HTSC thin-films. Standard design procedure using parallel-coupled microstrip resonators was followed.[61]

The first step in the filter design procedure was to determine the low pass prototype element values g_0 to g_4 . For a Chebyshev response with an odd number of sections, the first and last elements were chosen to be $g_0=g_4=1$. The remaining element values are determined using the Tables in [61] based on the desired passband ripple. To convert the low pass prototype into a bandpass

filter, a transformation of the low pass frequency, ω' , to the bandpass frequency, ω , is given as,

$$\frac{\omega'}{\omega_1} = \frac{2}{w} \left(\frac{\omega - \omega_0}{\omega_0} \right) \quad (4.1)$$

where ω_1' is the cutoff frequency of the low pass filter, w is the fractional bandwidth, and ω_0 is the centre frequency. The roll-off is determined from graphs of the attenuation, L_a , versus the frequency ratio, $\frac{\omega'}{\omega_1}$.^[61] A three-section filter with an attenuation of 15dB corresponds to a frequency ratio $\frac{\omega'}{\omega_1} = \pm 3.2$.

Substituting these two values into equation 4.1 yields the upper and lower frequency points which correspond to an attenuation of 15dB.

Once the low pass prototype elements have been determined, they can be used to calculate the impedance inverters. These inverters are used to convert the bandpass filter network into one using only series tuned circuits. For a Chebyshev filter with N sections, $N+1$ impedance inverters are required. Due to the symmetrical nature of Chebyshev filters with an odd number of poles, only the first $(N+1)/2$ inverters need to be computed. The inverter sections are expressed as,

$$\begin{aligned} \frac{Z_0}{K_{01}} &= \sqrt{\frac{\pi w}{2g_0g_1}} & \frac{Z_0}{K_{34}} &= \sqrt{\frac{\pi w}{2g_3g_4}} \\ \frac{Z_0}{K_{12}} &= \frac{\pi w}{2\omega_1} \cdot \frac{1}{\sqrt{g_1g_2}} & \frac{Z_0}{K_{23}} &= \frac{\pi w}{2\omega_1} \cdot \frac{1}{\sqrt{g_2g_3}} \end{aligned} \quad (4.2)$$

After the impedance inverters have been computed, the even and odd mode characteristic impedances, Z_{0e} and Z_{0o} , can be determined using the formulas,

$$Z_{0e} = Z_0 \left[1 + \frac{Z_0}{K_{ij+1}} + \left(\frac{Z_0}{K_{ij+1}} \right)^2 \right] \quad (4.3)$$

$$Z_{0o} = Z_0 \left[1 - \frac{Z_0}{K_{ij+1}} + \left(\frac{Z_0}{K_{ij+1}} \right)^2 \right]$$

From these impedances, the resonator lengths, widths and gap spacings were computed using LineCalc for each of the coupled line sections. A shunt capacitive loading at the ends of the open-circuited coupled lines effectively lengthens each line. To offset this fringing capacitance, the physical lengths of these resonators were reduced by an amount proportional to the capacitive loading at each end. The length reduction, Δl , can be expressed as

$$\Delta l = 0.412h \left(\frac{\epsilon_{eff}(f) + 0.3}{\epsilon_{eff}(f) - 0.258} \right) \left(\frac{\frac{w}{h} + 0.264}{\frac{w}{h} + 0.8} \right) \quad (4.4)$$

where $\epsilon_{eff}(f)$ is the adjusted effective dielectric constant at the centre frequency due to dispersion effects.^[62] Note that this value is simply an initial estimate, since dispersion in superconducting lines is practically negligible.

The resonator dimensions were implemented in Touchstone for further optimization using the microstrip coupled-line filter element, MCFIL. A summary of the coupled-line section dimensions is described in Table 4.1. All optimized dimensions deviated from the initial design calculations by less than 1.5%, except for the gap spacings of the 20 GHz filter which differed by approximately 10%. These filter elements were analyzed using a quasi-static method which is mainly valid for substrate relative dielectric constants less than

18.^[63] For this reason, a full-wave analysis package. EMSim, was used to verify the Touchstone simulations. EMSim incorporates an integral equation technique using Green's Function to account for surface waves and radiation loss from the resonator coupling sections.

	first and last sections (μm)			middle sections (μm)		
	length	width	spacing	length	width	spacing
19 GHz	973	75	237	968	83	943
20 GHz	917	79	237	914	84	953

Table 4.1 Resonator dimensions for the HTSC filters

The filters were fabricated at Superconductor Technologies Inc. using TBCCO thin-films on a 2" diameter lanthanum aluminate ($\epsilon_r=24$) substrate wafer. A layout of the two inch wafer, which includes the HTSC filters and amplifier matching networks, is shown in Figure 4.1. A polyimide passivation layer was applied over the substrate to protect the HTSC thin films. The 7.5mm x 3.3mm filters were assembled at CRC by mounting the circuits onto gold-plated kovar carriers using silver epoxy. Gold contact pads were included on the 50 Ω HTSC input and output lines to allow for wire bonding to the gold transmission lines on the test fixture. A photograph of the 20 GHz centre frequency HTSC microstrip filter is shown in Figure 4.2. Experimental and predicted results are covered in the next section.

To examine in more detail the coupling in HTSC microstrip lines, individual coupled-line sections were fabricated on the STI wafer. Both wide gap and narrow gap sections were analyzed using Touchstone and EMSim to compare with experimental results.

Figure 4.1 AutoCAD layout of 2" STI wafer containing HTSC filters

Figure 4.2 Photograph of 20 GHz HTSC microstrip filter

4.3 Filter Measurements and Results

The 3-pole HTSC microstrip filters were measured at 77K using the cryogenic measurement technique as described in section 2.5. S-parameter measurements were obtained using the Wiltron 360 Automatic Network Analyzer over the frequency range of 16 GHz to 24 GHz. The power levels at the output ports of the network analyzer were set to -30dBm to reduce the sensitivity of the unloaded Qs of the resonators. At power levels above 0dBm, the resonator Q-factor is degraded due to the current densities at the conductor edges exceeding the critical current density, J_c , of the thin-films.^[44] While the thin-film TBCCO becomes superconducting below 105K, the filter response does not stabilize until the temperature drops below 90K, due to the kinetic inductance effect. The temperature dependence of the filter response will be examined later in the section.

4.3.1 19 GHz HTSC Filter

The insertion and return losses of the 19 GHz filter are displayed in Figure 4.3, along with the predicted results using Touchstone and EMSim. The measurements show a 0.78dB insertion loss at a centre frequency of 18.65 GHz, with return losses of 17dB and 18dB at the filter input and output, respectively. The Touchstone response predicted an insertion loss of 0.56dB at a centre frequency of 18.8 GHz. EMSim predicted a significantly higher centre frequency of 19 GHz with a corresponding 0.66dB insertion loss. The S-parameters of measured and simulated results at their respective centre frequencies are displayed in Table 4.2. A further analysis of the filter was performed using a finite difference (FD) method software package developed at McMaster University.^[64] While the computation time was extremely long (more than three hours per frequency point using a SUN Workstation), this software predicted the centre frequency of the filter most accurately. Results were obtained over the

frequency range 18 GHz to 19 GHz, with the centre frequency S-parameters displayed in Table 4.2. The finite difference method assumed a perfect conductor, which explains the extremely low predicted insertion loss.

Figure 4.3 Measured vs. simulated response for 19 GHz HTSC filter

	$f_0(\text{GHz})$	S11	S21	S22
Measured	18.65	-17	-0.78	-18
Touchstone	18.8	-35	-1.56	-35
EMSim	19.0	-28	-0.66	-28
FD	18.6	-25	-0.1	-25

Table 4.2 S-parameters (in dB) at the centre frequency for the 19 GHz HTSC filter

To compare the performance of the HTSC filters to conventional metals, a gold microstrip filter on LaAlO_3 was simulated in Touchstone. An identical configuration was used for the gold filter, to produce a narrow 2.4% passband at a centre frequency of 18.8 GHz. Upon optimization in Touchstone, the filter had a minimum predicted insertion loss of 2.8dB at 297K and 1.7dB at 77K.

While both the Touchstone and EMSim responses estimated a higher centre frequency, the quasi-static analysis was within 150 MHz of the experimental value, representing an error of less than 1%. The full-wave analysis using EMSim showed a 2% error in the centre frequency. This can be partly attributed to a discrepancy in the relative dielectric constant, ϵ_r . While EMSim did not predict the centre frequency of the filter response as well as Touchstone, the full-wave analysis did predict the asymmetric shape of the frequency response more accurately. Re-simulating in EMSim using $\epsilon_r=24.8$ provided an excellent match to the actual response. The variation in ϵ_r may be due to the twinning effect of the lanthanum aluminate substrate, and the passivation layer ($\epsilon_r \sim 3$) protecting the TBCCO thin-films. STI's most recent specifications on substrate parameters reveals an empirical value for ϵ_r of 24.7. Another factor which was not accounted for in the CAD software is the negligible dispersion in

superconductors. However, this would result in a slightly lower effective dielectric constant, ϵ_{eff} , and hence an upward shift in frequency.

A significant cause of frequency shifts in filter responses is due to the kinetic inductance effect of superconductors. This effect can result in a dramatic change in the centre frequency due to the increased magnetic penetration depth when the superconductor is operated close to its transition temperature, T_c . As the HTSC is cooled below T_c , the normal electrons are transformed into superconducting electrons, whose kinetic inductance results in the rapid frequency shift.^[11] This dependence of the response on temperature is displayed in the graph of Figure 4.4. These results were determined by measuring the 19 GHz filter response over a range of temperatures from 50K to T_c . The advantage of using thallium-based thin films over the yttrium compound (YBCO) lies in the higher T_c value of TBCCO (105K as compared to 90K for YBCO). When operating at 77K, a temperature ratio of $T/T_c = 0.73$ achieved for TBCCO is superior to the $T/T_c = 0.86$ for YBCO. The kinetic inductance effect should be minimal in this case, as is demonstrated in Figure 4.4.

A resolution test chart featuring various patterns of vertical and horizontal lines. Each pattern is accompanied by a numerical value representing its resolution. The values include 1.0, 1.1, 1.25, 1.4, 1.6, 1.8, 2.0, 2.2, 2.5, 2.8, 3.2, 3.6, 4.0, 4.5, 5.0, 5.6, 6.3, 7.1, 8.0, 9.0, and 10.0. The patterns become progressively smaller and more densely packed as the numerical value increases.

MICROCOPY RESOLUTION TEST CHART

Figure 4.4 Kinetic inductance effect of filter response over temperature

The insertion loss of the filter was slightly degraded from the predicted value due to an overestimation of the surface resistance, R_s , for the HTSC thin films. The simulators assume that R_s is proportional to the square root of frequency while superconductors exhibit a f^2 dependence. By choosing a modest value of resistivity ($\rho=0.01\rho_{\text{gold}}$), the R_s can be reasonably approximated at higher frequencies. Another factor responsible for the degradation in insertion loss across the passband is the inclusion of bond wires at the filter input and output.

To improve the filter insertion loss, the return loss should be reduced below 15dB.^[65] Insertion loss is now limited by mismatches at the input and output ports (i.e. bond wires) as conductor losses are now small enough to be of the same order as matching loss. The high return losses have also been attributed to the random twinning of the LaAlO_3 substrate.^[66]

The 3dB bandwidth for the filter was determined to be 2.5%, which is slightly broader than the 2.3% predicted by Touchstone. This can be attributed to radiation losses which are not accounted for in Touchstone but will have a significant effect in superconducting filters where conductor losses are minimal. EMSim, which considers radiation loss, accurately predicts the 3dB bandwidth. A proper cover placed over the filter would reduce radiation losses, hence increasing the overall Q and providing a narrower bandwidth response.

4.3.2 20 GHz HTSC Filter

A second filter was designed for a 2.4% bandwidth at a centre frequency of 20 GHz. This filter also incorporates a three-pole Chebyshev response using quarter-wave parallel-coupled microstrip resonators. The design procedure was identical to the 19 GHz filter, with the filter dimensions optimized in Touchstone and verified using EMSim. The resonator dimensions and gap spacings are displayed in Table 4.1. Again, the predicted results overestimated the actual centre frequency due to a higher actual ϵ_r , as shown in the graphs of Figure 4.5. The centre frequency, insertion and return losses are displayed in Table 4.3. Discrepancies between measured and predicted results are attributed to the same factors discussed in section 4.3.1. This filter will be integrated with the HTSC/HEMT amplifier 'B' which produced a peak gain at 20.5 GHz, as described in Chapter 3.

	$f_0(\text{GHz})$	S11	S21	S22
Measured	19.75	-12	-0.83	-13
Touchstone	19.8	-33	-0.53	-33
EMSim	19.95	-16	-1.09	-16

Table 4.3 S-parameters (in dB) at the centre frequency for the 20 GHz HTSC filter

Figure 4.5 Measured vs. simulated response for 20 GHz HTSC filter

4.3.3 Analysis of HTSC Resonator Coupling Sections

To further evaluate the performance of the HTSC thin films, as well as the ability of the CAD software to analyze them, single coupled-line sections identical to those used in the filters were measured and modelled. The wide gap spacing (i.e. ratio of gap/line width > 10) of the 19 GHz filter was measured over the frequency range of 16 GHz to 24 GHz. The results are shown in Figure 4.6a). While Touchstone matches the experimental results more closely around 16 GHz, it underestimates the coupling by approximately 3dB at 22 GHz. This deviation is likely attributed to the reduced accuracy of Touchstone's coupled-line model for dielectric constants above 18. EMSim produced ill-conditioned results at lower frequencies around 12.5 GHz, but yielded an excellent match at frequencies above 18 GHz. These accurate results could be attributed to the inclusion of radiation losses and surface waves in the full-wave analysis which gain significance at higher frequencies.

For the 20 GHz filter, an analysis of the small gap spacing coupled line sections (i.e. gap/line width = 3) was performed from 16 GHz to 22 GHz at 77K. Both Touchstone and EMSim predict better coupling below 18 GHz and less coupling above 20 GHz, as shown in Figure 4.6b), with deviations from the measurements within 1dB.

Figure 4.6a) Coupling for wide gap resonator sections of 19 GHz filter

Figure 4.6b) Coupling of narrow gap resonator sections for 20 GHz filter

While these 3-pole, 2.3% bandwidth HTSC filters had modest specifications, more stringent filter designs in the future will emphasize the potential superior performance of HTSC technology as compared to conventional

metals. These designs, incorporating a larger number of poles to achieve very narrowband sharp skirt responses ($<1\%$), will allow for more efficient use of microwave communication bands with more closely spaced channels.

4.4 Hybrid HTSC Filter/Amplifier MIC

This section presents the design and performance of the integrated HTSC filter/amplifier MIC, based on the results of the amplifiers and filters described in the previous sections. The 18.65 GHz centre frequency filter was integrated with a slightly modified version of the HTSC/PHEMT amplifier 'A'. Results of this hybrid circuit have been published in the 23rd European Microwave Conference.^[67] The 19.75 GHz filter was paired with the HTSC/HEMT amplifier 'B'.

4.4.1 19 GHz HTSC Filter/Amplifier MIC

Amplifier 'A' described in Chapter 3 had 13dB gain at 19 GHz, but was found to be potentially unstable at certain frequencies. Therefore, the amplifier was modified to achieve better stability by adding a larger source inductance. Open-circuited stubs were re-optimized to achieve minimum reflections at 19 GHz when matching to measured S-parameters of a $0.25\mu\text{m} \times 100\mu\text{m}$ Comsat PHEMT device at 77K. As a result of the increased stability, the amplifier had a lower predicted gain of 9dB. Also, a 10pf chip capacitor was added to the output network to provide DC blocking, and the bias circuit was altered slightly in order to remove any potential coupling with the filter. Simulations showed that a high out-of-band reflection coefficient of the filter would also create a potential stability problem at certain frequencies when connected to the amplifier, and that a section of transmission line between the filter output and amplifier input would eliminate this problem. As the filter and input matching network were combined on the same substrate, the transmission

line was readily included in the final layout. While the amplifier contained various design modifications from the initial version, a filter identical to the 19 GHz filter described in the previous section was used. A Touchstone listing of the filter/amplifier integrated unit is included in Appendix III.

Figure 4.7 Photograph of the HTSC filter/amplifier 'A' MIC

A photograph of the combination HTSC filter/amplifier MIC is shown in Figure 4.7. The overall circuit dimension is 10.5mm x 4.2mm. The lanthanum aluminate substrates were attached to a gold-plated kovar carrier, similar to those used for the amplifiers, using conductive silver epoxy. The PHEMT device was epoxied onto the elevated pedestal and connected to the input and output matching networks using wire bonds. The chip resistors and capacitors were mounted onto gold contact pads on the substrate as before. Biasing was applied through filtercons attached to the circuit test jig via the phosphor-bronze bias lines in the cryogenic test setup. To suppress any external RF signals which might propagate along the bias lines from the power supplies causing oscillations, 0.1 μ f ceramic capacitors were added in shunt to the

filtercons on both the gate and drain. Ferrite beads were also included on the bias feed lines to assist in removing extraneous RF signals.

The filter/amplifier MIC was connected into the cryogenic test setup and measured using the Wiltron 360 Automatic Network Analyzer at 77K. When the transistor was biased to its previous operating conditions of 2V drain-to-source voltage and 6mA drain current, the circuit oscillated at 20.2 GHz, which was noticeable in the output reflection coefficient. The oscillations were mainly attributed to variations between the actual PHEMT and the device used in the simulations, whereby the out-of-band filter output impedance resulted in device instability. The oscillations subsided when the drain-to-source voltage was reduced to 1.5V. Measured results are displayed in Figure 4.8 over the frequency range of 17 to 21 GHz. The gain reaches a peak value of 6.4dB at 18.9 GHz, but shows a significant ripple across the passband. At the centre frequency of 18.7 GHz, the gain is reduced below 6dB with corresponding input and output return losses of less than 5dB. This ripple is mainly caused by a mismatch between the filter passband and the peak of the amplifier, which were offset from each other by over 300 MHz.

The simulated version of the filter/amplifier circuit had a predicted passband gain of 8.4dB at a centre frequency of 18.8 GHz, with corresponding input and output matching of -22dB and -23dB, respectively. The simulation incorporated the Touchstone filter response as opposed to the EMSim version, due to its greater flexibility in simulating as well as its closer approximation to the measured response. Even though the predicted filter's passband did not directly coincide with the peak of the amplifier response, the excellent return loss of the filter provided a good match in the overall circuit.

Figure 4.8 Measured vs. simulated a) S_{11} and b) S_{22} of HTSC filter/amplifier 'A'

Figure 4.8c) Measured vs. simulated gain of HTSC filter/amplifier 'A'

When designing the filters for integration with the amplifier, the centre frequencies were deliberately set slightly lower than the peak of the amplifier response, since the gain roll-off was more gradual on the lower frequency edge. For instance, with a 19 GHz peak amplifier response, centre frequencies of 18.8 GHz and 19 GHz were achieved with Touchstone and EMSim, respectively. This would help compensate for any variation in dielectric constant, noting that ϵ_r has been found to decrease at lower temperatures,^[25] thus causing an upward shift in centre frequency. The sharper gain roll-off at the upper frequency edge of the amplifier passband would considerably distort the filter/amplifier passband response.

The discrepancy between the predicted and measured performance of the integrated unit is clearly reflected in the comparison of the measured and simulated filter responses in Figure 4.3, as their centre frequencies differ by 150 MHz. As the filter was fabricated on the same wafer as the filter/amplifier MIC,

no opportunity existed to modify the filter to more accurately match the amplifier response. Therefore, a second iteration filter design accounting for the frequency shift would be required to solve the mismatching problem.

Figure 4.9 Noise figures of the HTSC filter/amplifier MIC's

Noise figure for the HTSC filter/amplifier microwave integrated circuit was measured at 77K using the technique described in Chapter 3. By connecting the noise source to the input vacuum feedthrough port and the output port connected to the noise figure meter, as in Figure 3.10, the overall noise figure of the system can be measured. The noise figure of the DUT can be determined once the K-cable and test fixture losses have been accounted for. Noise figure for the filter/amplifier integrated unit was measured over a frequency range extending from 18.4 to 19.2 GHz in order to coincide with the narrow passband. The optimum noise bias conditions at room temperature of 2V and 6mA were applied to the drain, as no oscillations occurred in this frequency range. The results are displayed in Figure 4.9, showing a minimum noise figure

of 1.8dB at 18.9 GHz. Varying the drain current above and below 6mA produced no improvement in noise figure, confirming that this was also the optimum noise bias condition at 77K.

4.4.2 20 GHz HTSC Filter/Amplifier MIC

A second filter/amplifier integrated circuit was developed using the HTSC/HEMT amplifier 'B' described in Chapter 3 and the 20 GHz HTSC filter presented in section 4.3.2. The filter and amplifier were mounted on separate carriers for individual measurement, and then interconnected with wire bonds for the overall circuit. The combined circuit was measured at 77K over the frequency range of 16 GHz to 24 GHz. The transistor was biased with 5mA and 2V applied to the drain, and displayed no oscillation problems at these bias values.

Measured and Touchstone results are shown in Figure 4.10. The measured gain at the 19.8 GHz passband centre frequency was 7.2dB with corresponding matching of -17dB at the input and -8dB on the output. The Touchstone simulation predicted 8.8dB of gain at an identical centre frequency of 19.8 GHz. Once again, the discrepancy between the measured and predicted results is attributed to mismatching between the filter and amplifier responses. The lower frequency edge of the transmission response, $|S_{21}|$, clearly demonstrates this problem. The filter passband is 700 MHz lower than the peak amplifier response at 20.5 GHz which corresponds to an error of almost 4%. For the simulated hybrid circuit, the amplifier was to provide a peak at 20 GHz which would properly match the filter with a predicted centre frequency of 19.8 GHz in Touchstone and 20 GHz in EMSim.

Figure 4.10 Measured vs. predicted a) S_{11} and b) S_{22} for HTSC filter/amplifier 'B'

Figure 4.10c) Measured vs. predicted gain of HTSC filter/amplifier 'B'

To correct the mismatch, the amplifier response could be shifted downward in frequency by appropriately reducing the amount of source inductance. This may be achieved by shortening the bond wires which connect the source of the transistor to the gold-plated carrier (i.e. ground). A lower source inductance would also increase the peak transmission gain, hence reducing the amplifier stability. To offset any stability problem, a specific length of transmission line could be included. This would be more easily realized by integrating the filter and input matching circuit on a single substrate, as was the case for the previous filter/amplifier MIC. An alternative solution to eliminating the mismatch between the filter and amplifier is to redesign the filter to achieve the required higher centre frequency, rather than alter the existing amplifier.

The noise figure of this HTSC filter/HEMT amplifier integrated unit was measured at various bias conditions, with results for $I_{DS}=3\text{mA}$ and 5mA displayed in Figure 4.9. Minimum noise figures at 19.8 GHz of 1.7dB and 1.5dB

correspond to drain currents of 5mA and 3mA, respectively. Measurements performed at 10mA yielded significantly degraded noise figures. These results are consistent with noise figures measured for a Ku-band HEMT low noise amplifier at 77K which produced optimum noise figures at a drain current of 3.5mA.^[47] Considering that the device was matched for maximum gain rather than optimum noise performance, these initial results are very encouraging. An improvement in the matching between filter and amplifier should further improve noise figure.

4.5 Summary

This chapter presented the development and results of two high temperature superconducting filter/low-noise amplifier hybrid circuits. Two 3-pole narrowband (2.5%) Chebyshev microstrip filters were fabricated using TBCCO HTSC thin-films on a LaAlO_3 substrate. The filters provided 0.8dB insertion loss and return losses of 18dB and 13dB at centre frequencies of 18.65 GHz and 19.75 GHz, respectively. CAD simulations using a quasi-static approach (TouchstoneTM) and a full-wave analysis (EMSimTM) predicted a lower centre frequency for both filters. This was attributed to a higher actual dielectric constant due to the twinning effect of the LaAlO_3 substrate. As a comparison to the performance of the HTSC filters, a similar gold filter designed on LaAlO_3 predicted 2.8dB insertion loss at room temperature and 1.7dB at 77K.

The lower frequency filter was integrated with amplifier 'A' to produce the first known HTSC filter/low-noise amplifier hybrid circuit at K-band. The transmission response achieved a peak value of 6.4dB at 18.9 GHz, with corresponding noise figure of 1.8dB. The 19.75 GHz HTSC filter/amplifier 'B' integrated unit provided 7.2dB gain and 1.7dB noise figure at the nominal drain bias condition of 2V and 5mA, and 1.5dB noise figure at 3mA drain

current. These are excellent results for a first iteration design, and compare favourably to a C-band HTSC filter and GaAs LNA hybrid circuit which reported a noise figure of 1.5dB.^[19]

CHAPTER 5: CONCLUSIONS

5.1 Summary and Discussion

The potential for combining semiconductor and superconductor components to form hybrid microwave integrated circuits has been investigated. At present, the greatest benefits of these hybrid circuits can be obtained by utilizing active semiconductor devices and passive superconducting structures. The processing of three-terminal active devices using semiconductor technology is a mature area, whereas similar superconducting devices are still at the development stages. Moreover, the performance of these devices improves considerably when operating at cryogenic temperatures, making them suitable for integration with superconductors. HTSC passive components such as resonators and filters have demonstrated much lower losses and higher Q s than similar circuits fabricated using conventional technology. Their performance is comparable to dielectric resonators and waveguide filters traditionally used in narrowband applications, such as for satellite communications. HTSC planar circuits, however, are capable of meeting the narrow bandwidth requirements with a significant reduction in size.

This work has presented the design and development of a hybrid HTSC/GaAs K-band amplifier and filter at 77K. The filter and matching networks were fabricated using HTSC thin-films, while the amplification was accomplished with a HEMT device. This filter/amplifier integrated unit provides a basic component for the front end of a microwave receiver by minimizing the system noise figure while removing any spurious signals and noise outside the desired frequency band. This K-band semiconductor/superconductor hybrid circuit was the first known attempt at integrating the two technologies in this circuit configuration at such a high frequency.

Chapter 2 presented an overview of microwave integrated circuit (MIC) technology, and the performance improvements at cryogenic temperatures. Cooled MICs exhibit an increase in resonator Q-factors due to lower resistivities and substrate losses. Low temperature operation also reduces the on-chip power density by enhancing the thermal conductivity in materials, increases the circuit reliability by eliminating device failures due to heating effects, and provides a stable temperature environment. Perhaps the most significant advantages of cooling MICs is the improved performance of the active devices such as FETs and HEMTs. These devices exhibit higher gain and operating frequency due to the enhanced carrier mobility, and lower noise figures attributed to the intrinsically lower thermal noise.

The following section examined the development of low temperature superconductors, including the Two-Fluid model, and the London and BCS theories. These theories were extended to high temperature superconductors, as well as some more recent explanations into the behaviour of HTSCs. Following, the thin-film fabrication and processing techniques for HTSCs were discussed. The chapter concluded with a description of the cryocooler apparatus and TRL calibration technique developed at CRC.^[50] This test setup provides accurate S-parameter measurements at cryogenic temperatures by de-embedding the test fixture and K-cables.

Chapter 3 reported the design and development of two K-band HTSC/GaAs cryogenic amplifiers. HEMT devices operated at 77K have yielded 3dB improvements in gain, along with significantly lower output impedance. S-parameter measurements of the amplifier showed reasonable agreement to the Touchstone simulations, which revealed a high sensitivity to bond wire source inductance. Fewer bond wires connecting the source of the device to ground resulted in a higher inductance, hence reducing the input and output reflection

coefficients and providing stability for the amplifier. A secondary peak at the lower frequencies which affected the amplifier stability was caused by extraneous effects of the bias circuitry. The elimination of this potential stability problem was the objective of a second iteration amplifier design. However, due to modelling inaccuracies of the resistive stabilizing network, the amplifier yielded poor results.

Noise figures of less than 1 dB for the 20 GHz amplifier were achieved. This is a promising result considering the device was matched for optimum gain rather than minimum noise figure. Matching the device noise parameters would improve the noise figure, however the tuning mechanisms required to obtain these noise parameters at low temperatures are extremely difficult to implement. Noise models for various devices at low temperatures have been developed for application in extremely low noise cryogenic amplifiers.^[57,68,69]

While these amplifiers have demonstrated the potential for integrating superconductors and semiconductors, there is no conclusive evidence that this hybrid circuit can provide better performance than a conventional cooled GaAs amplifier. The improvements in losses of the HTSC matching networks compared to conventional metals at 77K is minimal for these relatively short transmission lines. Also, the high dielectric constant of HTSC substrates considerably reduces the line widths, creating a large discontinuity when connecting the circuit into the test fixture. This effect was partially accounted for by implementing tapered transitions at the ends of the 50 Ω input and output HTSC lines. The main advantages of HTSCs become evident in high Q resonators and narrow bandwidth filters with greater numbers of coupling sections.

The HTSC microstrip filters designed in this work had measured insertion losses of approximately 0.8dB at K-band frequencies, for 3dB

bandwidths less than 2.5%. Both Touchstone and EMSim simulations predicted a higher centre frequency, which may be attributed to a lower estimated dielectric constant. While Touchstone provided a closer estimate to the centre frequency, EMSim predicted the asymmetric shape of the response more accurately. Kinetic inductance, which causes frequency shifts when operating near the transition temperature, was minimized by using TBCCO thin-films with a T_c greater than 100K.

The ability of these commercial CAD software packages to analyze HTSC structures was further examined by measuring individual resonator coupling sections of the filters. EMSim showed excellent agreement for a wide gap section above 18 GHz, but produced erroneous results at a frequency of 12.5 GHz. Touchstone predicted the wide gap coupling accurately below 18 GHz, but underestimated the coupling at higher frequencies. The improved modelling of EMSim at higher frequencies is likely due to the consideration of radiation loss and surface waves which this full-wave analysis software is capable of handling. For the narrow gap coupling section, both Touchstone and EMSim results were within 1dB of the measurements over the 16 GHz to 24 GHz frequency range.

These software packages, along with Sonnet's EM, were also used to simulate the frequency response of HTSC matching networks for amplifiers. EMSim and Touchstone provided the closest match to measurements of an input and output matching circuit, respectively. While EM is more user friendly than EMSim, it requires a longer computation time with no apparent improvement of results. Although EM and EMSim did not consider the effects of the bias circuits in their analyses in order to shorten the computation time, Touchstone appears to be sufficient for modelling these simple HTSC passive structures due to its simplicity of modelling and superior computation speed.

The HTSC filter/amplifier hybrid MICs provided excellent results for a first iteration design. The integrated unit containing a PHEMT device consisted of the HTSC filter and input matching network on a single substrate, illustrating an advantage of utilizing HTSC matching circuits in the amplifier. A measured transmission response of 6dB at the centre frequency deviated significantly from the 8dB predicted by Touchstone, mainly caused by a downward shift in frequency of the filter response. This mismatch was evident in the passband ripple, which could be suppressed by improving the matching between the filter output and amplifier input. Due to the high out-of-band reflection coefficient of the filter, the amplifier must be unconditionally stable over the entire frequency range to prevent oscillations. A similar hybrid circuit utilizing a HEMT device provided a higher gain of 7.2dB at the centre frequency, with less ripple in the passband. This result could be improved by reducing the 700 MHz discrepancy between the filter and amplifier centre frequencies. By lowering the source inductance of the device to increase the gain and shift the amplifier response down in frequency, results closer to the 9dB predicted passband gain could be achieved. Improved matching between this filter and amplifier could be more readily realized by combining the filter and input matching network on a single substrate.

An impressive minimum noise figure of 1.5dB achieved for the HTSC/HEMT amplifier 'B' and filter illustrates the potential for hybrid semiconductor/superconductor circuits in low noise applications. Improvements in filter insertion loss and matching between the filter and amplifier could potentially lower this value close to 1dB. This type of performance would make these circuits suitable as low noise, narrow bandwidth components for channelized receivers in satellite communications.

5.2 Future Work

Improvements for the HTSC filter/amplifier MIC must focus on achieving better stability of the device at low temperatures without significant degradation in the gain and noise figure. Varying the bond wire source inductance to accomplish this task is highly sensitive to slight changes in bond wire length and numbers. Alternative forms of stability such as resistive stabilizing networks have been investigated, however these methods will require a more detailed analysis into frequency and temperature effects of chip components.

Another area of improvement lies in the design of HTSC filters. A second iteration filter design would allow the designer to predict the frequency response with greater confidence. While the filters designed in this work yielded moderately narrow bandwidths, the high-Q capabilities of HTSCs can be further exploited in extremely narrowband applications (<1%), which require higher-order filters to achieve a steeper roll-off. To attain the highest possible Qs for HTSC resonators and filters, a properly designed cover should be implemented to reduce radiation loss. Rather than matching the filter's output to a 50Ω system, the filter could be inherently matched to the input impedance of the amplifier.^[70] This would eliminate the need for an input matching network and therefore reduce the area of the overall hybrid circuit.

Due to restrictions on substrate dielectric constant in conventional software packages, and the unique properties of superconductors, customized HTSC software will be required to analyze more complex structures. Kinetic inductance, which represents the temperature dependence of the London penetration depth, must be considered when operating at temperatures close to T_c . The behaviour of the surface resistance for thin-film superconductors should

be accurately modelled, as well as the limitations on current densities and the degradation of Q-factors at high power levels.

The processing of HTSCs must continue to improve to meet the demand for higher critical current densities and lower surface resistance. Also, substrate quality must be improved to ensure the higher performance criteria for HTSCs are met. LaAlO_3 provides good thermal matching properties to the HTSC thin-films, however it poses problems due to a high dielectric constant, "twinning" planes, and brittleness which makes via holes difficult to implement. Substrates such as R-plane sapphire provide extremely low substrate loss and a lower dielectric constant, but do not provide a good thermal match to the HTSC thin-films.

The current cryogenic measurement facility at CRC is being augmented by a cryogenic wafer prober, which will allow for on-wafer S-parameter measurements at 77K. This probing station will significantly reduce the calibration time, and improve measurement accuracy by eliminating the need for wire bonds. The accuracy of noise figure measurements using the current cryogenic setup should be improved by utilizing a higher excess noise ratio (ENR) noise diode. Higher ENRs will compensate for the significant losses incurred in the K-cables, test fixture, and mixer, especially at high frequencies. For accurate measurements of extremely low noise devices, a hot-cold noise source using liquid nitrogen as the "cold" reference would provide the best results.

Future application of semiconductor/superconductor hybrid circuits will extend to MMICs, wherein superconducting layers can be deposited onto a GaAs substrate. Furthermore, integration of semiconductors and superconductors into single hybrid devices such as the Superconducting FET (SuFET), and Resonant Tunneling Transistor (RTT), are currently being

developed.^[10] The SuFET will be capable of providing large-signal gain in digital applications with very low power dissipation, while the RTT will be useful for very high frequency (THz) applications with carrier transit times on the order of a few picoseconds.

APPENDIX I HTSC/PHEMT AMPLIFIER 'A' SCHEMATIC DIAGRAM

APPENDIX II HTSC/HEMT AMPLIFIER 'C' SCHEMATIC DIAGRAM

APPENDIX III TOUCHSTONE LISTING OF HTSC FILTER/LNA

```
!FIL LNA.CKT
!SIMULATION OF HTSC FILTER/ PHEMT-AMPLIFIER MIC
!MAY '93
```

```
DIM
LNG UM
COND /KOH
```

```
VAR
L1=549
D1\602
L2=893
D2\727
WA=75
SA=237
LA=973
WB=83
SB=943
LB=968
line=1200
```

```
CKT
MSUB ER=24 H=254 T=0.8 RHO=0.01 RGH=0
TAND TAND=0.00003
```

```
! input matching network for amplifier
MLIN 1 2 W=86 L=100
MCROS 2 3 4 5 W1=86 W2=86 W3=86 W4=25
MLEF 3 W=86 L=D1
MLIN 4 6 W=86 L=L1
MLIN 5 7 W=25 L=1100
mtee 7 8 9 w1=25 w2=500 w3=25
MLEF 8 W=500 L=787
CAP 8 15 C=10
RES 15 0 R=50
mlin 9 10 w=25 l=600
mbend 10 11 w=25 ang=45 m=0
mlin 11 12 w=25 l=677
res 12 13 r=2000
mlef 13 w=350 l=350
DEF2P 1 6 IPMATCH
```

```
! output matching network for amplifier
MLIN 10 11 W=86 L=L2
MCROS 11 12 13 14 W1=86 W2=86 W3=86 W4=25
MLEF 12 W=86 L=D2
MLIN 13 15 W=86 L=1040
cap 15 26 c=10
MLIN 14 16 W=25 L=1100
mtee 16 17 18 w1=25 w2=500 w3=25
MLEF 17 W=500 L=787
CAP 17 25 C=10
RES 25 0 R=50
mlin 18 19 w=25 l=500
mbend 19 20 w=25 ang=60 m=0
mlin 20 21 w=25 l=600
mstep 21 22 w1=25 w2=350
mlef 22 w=350 l=350
DEF2P 10 26 OPMATCH
```

```
! simulated cryogenic HTSC/PHEMT amplifier 'A'
ipmatch 40 41
S2Pb 41 42 45 ../fhr10x/LNF100_2.S2P
ind 45 0 l=0.05
OPMATCH 42 43
DEF2P 40 43 AMP
```

```

! 3-pole Chebyshev bandpass microstrip filter
MLIN 1 2 W=86 L=500
MCFIL 2 3 W^WA S^SA L^LA
MCFIL 3 4 W^WB S^SB L^LB
MCFIL 4 5 W^WB S^SB L^LB
MCFIL 5 6 W^WA S^SA L^LA
MLIN 6 7 W=86 L^line
DEF2P 1 7 BANDPASS

! simulated HTSC filter/amplifier MIC
BANDPASS 1 2
MLIN 2 3 W=86 L^line
AMP 3 4
DEF2P 1 4 FIL_AMP

! measured filter/amplifier MIC
s2pd 1 2 0 ../fillna/fplna3.s2p
def2p 1 2 meas

FREQ
  sweep 16 22 0.05

OUT
FIL_AMP DB[S11] GR1
FIL_AMP DB[S21] GR2
FIL_AMP DB[S22] GR3
MEAS DB[S11] GR1
MEAS DB[S21] GR2
MEAS DB[S22] GR3

GRID
FREQ 16 22 2
GR1 -30 5 5
GR2 -20 10 5
GR3 -30 5 5

OPT
FREQ 18.9 19.1
AMP MAG[S11]=-0
AMP MAG[S22]=-0

```

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