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**Integrated Amplifier/Antenna Elements
for GaAs Monolithic Phased Arrays**

Langis J.P. Roy

**A thesis submitted in partial fulfilment of the requirements for
the degree of Doctor of Philosophy.**

**Ottawa-Carleton Institute for Electrical Engineering
Department of Electronics
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INTEGRATED AMPLIFIER/ANTENNA ELEMENTS
FOR GaAs MONOLITHIC PHASED ARRAYS

submitted by
Langis J.P. Roy, B.A.Sc., M.Eng.

in partial fulfilment of the requirements
for the degree of Doctor of Philosophy.

Chair, Department of Electronics

Thesis Supervisor

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June 1993

ABSTRACT

Future communications and radar systems operating at frequencies above 20 GHz are expected to make use of monolithic phased arrays based on GaAs Monolithic Microwave Integrated Circuit (MMIC) technology. This thesis investigates the integration of amplifiers and antennas on GaAs to form active amplifier/antenna elements suitable for use in such arrays.

The evolution of phased arrays is first considered, and it is seen that recent technological progress has led to the possibility of realizing phased arrays in monolithic form not only with microwave circuit techniques, but also with the use of emerging digital and optical technologies. Then, an overview of active antenna realizations is given, in which various transmitting, receiving and transmitting/receiving configurations are assessed in terms of the applicability to monolithic phased arrays. It is found that the desired active antenna element consists of an amplifier/antenna.

After studying the characteristics of resonant type antenna structures, the most efficient radiating element on GaAs MMIC substrates is determined to be the slot radiator. As a result, an analytical tool is developed to obtain the slot input impedance in terms of its geometry for subsequent integration with active devices. The analysis is based on an integral equation formulation for a magnetically coated dipole, this being the complementary/dual structure of the slot on a dielectric substrate, and a moment method solution is employed to obtain the input impedance numerically. Results of the computations compare favourably with experimental data. In addition, newly developed on-wafer techniques are described for the measurement of integrated-circuit slot antenna parameters such as input impedance and radiation patterns.

To demonstrate the feasibility of MMIC phased array realization, an optimum monolithic array configuration is proposed and the form of the constituent active amplifier/antenna elements is defined. The amplifier/antenna is a new MMIC structure, consisting of a P-HEMT device connected without matching circuitry to a slot radiator on the opposite side of the substrate by means of metallized via holes. A method is presented for the analysis of the amplifier/antenna using the slot input impedance calculation previously developed, and a commercial microwave circuit CAD package to represent the device and antenna feed characteristics.

The practical MMIC implementation of a novel transmitting amplifier/antenna is then described along with the results of its performance. At the 30 GHz design frequency, a gain of 7.4 dB is obtained from the amplifier portion of the active antenna, while the radiation pattern exhibits a symmetrical, single lobe characteristic, as predicted. This is the first demonstration of an MMIC which incorporates a P-HEMT and a slot radiator, and the first time via holes have been used to connect between circuit components on both sides of the wafer. It is concluded that the principle of the amplifier/antenna, as well as its analysis method, are validated. A more exhaustive study of slot antenna input impedances is recommended as future work to allow the realization of increasingly compact, enhanced performance transmit- and receive-mode amplifier/antennas.

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CHAPTER 1. INTRODUCTION

1.1 Motivation

In less than 10 years, a veritable revolution has taken place in the communications industry. Facsimile machines, cellular phones and electronic mail are but a few examples of systems that have recently sprung into existence and are now transforming the way in which people and their machines communicate with each other. Present forecasts are that the 21st century will be the Information Age, characterized by ever-increasing needs for communications [1]. Indeed, voice, data, image and video communications (together, multimedia communications) are expected to function easily, any time, anywhere, in a timely, cost-effective manner [2]. Much of this vision, referred to as Global Communications, is already fulfilled with today's "wired" networks found in metropolitan areas, whether these be based on optical fibers or traditional copper wire transmission lines [3].

A challenging aspect of Global Communications is the ability to provide multimedia communications to an individual person wherever he or she may be, a service known as Personal Communications. Implicit in this concept are the ideas of mobility and portability, which place severe constraints on the nature of the communicating terminal. It must be light weight, small in size and consume relatively little power so as to be supplied by a battery. The personal communicators of today that most closely fit this description are terrestrial-based mobile radios (i.e. cellular phones) operating in the UHF and L bands (0.3 - 2 GHz). They are currently limited to major population areas, although future L-band mobile satellite systems such as INMARSAT and MSAT will extend their reach [4]. Due to the severe crowding of spectrum in this band and the increasing requirements for high rate data communications, these mobile radios can only

satisfy demand for a short time. Tomorrow's Personal Communications terminals will therefore have to operate at higher frequencies, such as the proposed 20/30 GHz and 37 - 40 GHz bands where bandwidth is plentiful [4,5].

The precise form that future higher frequency Personal Communications networks will have is not evident at the present time: proposals for Personal Communications by satellite or via a terrestrial base station abound [4-7]. However, in all cases a handheld cellular phone style of personal terminal is called for to provide a wireless link over a distance of a few metres (in-building communications) to several thousands of kilometres (via satellite). Such a mass-produced terminal would share the same characteristics as its lower frequency counterpart with regard to weight, size and power consumption, but would have the additional feature of a directive radiation beam to ensure minimal use of RF power. At the envisaged millimetre-wave frequencies of operation, it is important to maintain the lowest possible levels of power due to inherent inefficiencies in RF generation and signal processing [8].

The technology currently in use to produce the RF electronics portion of this type of terminal is either hybrid or monolithic microwave integrated circuitry. As the required number of identical circuits grows, the GaAs monolithic microwave integrated circuit (MMIC) approach is favored, as it combines on a semiconductor substrate both the active and passive components in a single, precise, repeatable, fabrication step. Of the many types of antennas capable of producing directive antennas, only planar-style configurations are suitable for insertion into handheld terminals. Furthermore, to compensate for the motion of the terminal, the beam must be steerable. For compatibility with the personal terminal concept, mechanical steering is not a viable solution and hence an electronically steered phased array must be employed [4,6,7]. With the planar antenna consisting of an array of individual elements, the realization of a phased array involves providing an amplitude and phase control of the signal at each element.

The additional signal processing of a phased array necessitates, for the same reasons stated above, the use of MMIC technology. Since the radiating elements are in close proximity, it has been suggested that full integration of the active, passive and radiating components on the same GaAs wafer to obtain "monolithic phased arrays" is a promising technique for ultimately achieving Personal Communications [6,7]. At millimetre wave frequencies, the size of the radiators is compatible with the monolithic approach, where GaAs wafers may be up to 4" in diameter.

The need for monolithic phased arrays has also been expressed in the increasingly important field of millimetre-wave radar. With the predicted ability to provide rapid, agile electronically controlled beam steering and shaping, these arrays are ideal candidates for incorporation into an aircraft's skin to produce the so-called "Smart Skins" for airborne ECM applications [9]. Also, at the sub-millimetre wavelengths used in space-based radar the monolithic phased array approach is the only construction technique compatible with the small grid spacing of the array elements, and offers the additional benefit of being light weight [10]. In terms of civilian applications, a potentially high volume area is in automotive and traffic systems, where MMIC-based radars are being developed in the 60 to 77 GHz range for speed sensing and collision avoidance [11,12].

At present, communications satellites providing the 20/30 GHz transponder service do not make use of phased arrays. The need to further increase capacity and improve performance by generating a number of beams or reconfiguring the beam shape on command, have led to the realization that conventional reflector systems may have reached their limits. MMIC phased arrays are thought to hold the potential for achieving these latest performance requirements [13].

Phased arrays on a wafer have been studied by several researchers over the past five years but, so far, no working examples have been demonstrated for reasons related

primarily to the radiation efficiency of antenna elements on GaAs and the antenna/circuit interface, as will be seen in the remainder of this thesis. The need for monolithic phased arrays in future communications and radar applications constitutes the primary motivation of this work.

1.2 Objectives

Research towards the realization of monolithic phased arrays involves a multitude of issues which cannot possibly be studied in their entirety in a single thesis. The scope of this work is therefore limited to the problem of integrating, in MMIC form, a single radiating element of the array with its associated active circuitry. In the most general of phased array configurations, the chain of active circuits begins with an amplifier which not only amplifies the transmitted or received signal, but also provides isolation from the remainder of the signal processing components. Thus, the problem stated above is properly addressed by considering only the monolithic integration of a radiator with an amplifier. Herein lies the essence, it is believed, of MMIC phased array realization.

Such an undertaking must be pursued in the context of a specific MMIC array architecture. Since earlier attempts at producing monolithic phased arrays have been unsuccessful, no proven configuration presently exists. Therefore, a first objective of this thesis is to define a phased array configuration which is suitable for MMIC implementation and which determines the form of the amplifier/antenna interface.

The proposed integrated amplifier/antenna is a new MMIC structure, extending the capability of this technology beyond the integration of passive and active circuit elements, to also include radiating elements. As a result, an analysis method is required which

adequately describes the behavior of this new class of element. The second objective of this thesis is the development of an analytical tool allowing the design and optimization of MMIC amplifier/antenna elements, and the prediction of their performance.

A GaAs MMIC wafer containing radiating elements presents unique measurement problems at EHF frequencies due to its fragility and special feeding requirements. Thus, a third objective of this work is to develop suitable techniques for the accurate measurement of MMIC antenna parameters, such as input impedance and radiation patterns.

To validate the principle of the new amplifier/antenna device, experimental verification must take place. A fourth and final objective of this thesis is therefore the fabrication of the proposed amplifier/antenna using GaAs MMIC technology, and the practical demonstration of its operating characteristics so that comparisons may be made with the predicted performance.

1.3 Thesis Organization

The material of this thesis is presented in a logical manner throughout the remaining chapters. Chapter 2 gives an historical account of the evolution of phased array systems, highlighting how GaAs MMIC technology and emerging complementary technologies can be used to meet the needs of Personal Communications. With the system-level background in place, active antennas are reviewed in Chapter 3, as they embody the full range of current techniques from which compact phased arrays can be built. After assessing their suitability for monolithic designs, it is seen that no current structures are satisfactory. Thus, Chapter 4 is devoted to the determination of an efficient radiating element on GaAs, the slot antenna, and the development of an analytical method

to obtain its input impedance for subsequent integration with active devices. Having studied MMIC slot antennas, Chapter 5 describes novel techniques for performing measurements of their impedance and radiation patterns. The results of the previous chapters combine in Chapter 6, which deals with active antennas for GaAs monolithic phased arrays. An optimum phased array configuration is defined and a method for the analysis of its elements is given. A practical implementation of an amplifier/antenna and the results of its performance are then described. Finally, Chapter 7 concludes with a summary of the thesis work and suggestions for future studies.

CHAPTER 2. EVOLUTION OF PHASED ARRAY SYSTEMS

2.1 Introduction

Phased array systems have evolved tremendously over the last two decades as the microwave, digital, photonic and manufacturing technologies to implement them have progressed. This chapter presents an overview of past and present phased array developments with the primary aim of showing how, from a system point of view, continued advances in key technological areas may lead to the realization of monolithic phased arrays. The discussion emphasizes personal communications related applications, omitting therefore several aspects of phased array systems. A more general treatment may be found in several books and review papers on the subject [14-16].

2.2 Basic Configurations

The concept of a phased array antenna originates from the simultaneous needs to radiate microwave energy in a very narrow beam and to control the direction of this radiation. Enlarging the dimensions of a single-element antenna can lead to more narrow radiation characteristics, but another way to increase an antenna's dimensions is to form an assembly of small radiating elements in an electrical and geometrical configuration called an array. In most cases, the array elements (wires, apertures, etc.) are identical for convenience and simplicity, and are spaced up to a half free space wavelength apart [17]. The total beam is formed by the vector summation of the fields radiated by the individual elements and is a function of the relative position, amplitude excitation, phase excitation, and radiation pattern of the elements. By electronically imparting a relative phase difference between each element, the array beam may be steered in the direction desired.

Thus, phased arrays eliminate the need for mechanical scanning of the the antenna to direct its beam, a significant advantage over large single-element antennas when rapid steering and compact construction are required [18].

Depending on the application, phased arrays can require up to 10 000 elements and produce effective radiated powers reaching 10 kW or more. Their operating frequencies range from 800 MHz to 60 GHz, while their radiation capabilities include beam scanning (rapid, wide angle, two dimensional), and beam shaping (dynamic, multiple beam, adaptive nulling), among other features [10,14,19-22].

Phased array technology has followed a trend of increasing component integration to improve performance and lower costs. A conventional phased array is shown in block diagram form in Figure 1.1, which is representative of the technology in use 25 years ago.

Figure 1.1 Block diagram of a conventional phased array.

A single high power source and a receiver are connected to a power distribution network via a circulator which isolates the received and transmitted channels. Each radiating element is fed from the power divider through an individual phase shifter. The power source is a tube-type transmitter which can deliver up to 1 kW of power, but with

low reliability. The power combiner/divider is of a metal waveguide or coaxial line construction, causing RF losses ranging from 1 to 4 dB. The phase shifters are generally ferrite devices which require considerable drive power and add to the total insertion loss, while dipoles or waveguide antennas are used as radiators [24]. The capability of these arrays is limited to beam scanning, as there is no control of the individual element's amplitude excitation to allow dynamic pattern changes. Radars based on this technology have typical mean time between failure (MTBF) values of 100 hours, with the transmitter characteristics representing the principle failure mechanism of the system [18]. Examples of such radars are given by Oliner [14].

The block diagram of a more advanced phased array, which uses the transmit-receive (T/R) module approach, is shown in Figure 1.2. In this case, a lower power source is used along with the feed network to provide power to all the elements via the T/R module placed behind each radiator. This module consists of a variable attenuator, phase shifter, separate transmit and receive amplifiers isolated by RF switches, and the necessary DC bias and control lines. The exact type of individual circuits used in the T/R module varies greatly depending on the application. In communications systems, for example, the transmit and receive paths would not combine as shown in Figure 1.2. Over the past 15 years, the technology of the modules has undergone a complete transformation, starting from thick-film hybrid circuits containing silicon active devices, and progressing up to today with parts or all of the module realized in GaAs MMIC form [18,23]. The use of MMICs is expected to reduce the costs per module to the \$500 - \$1000 range [24]. At the same time, the radiating elements and feed network have evolved towards printed or microstrip arrays and feedlines, although these have not completely overtaken the more traditional structures [25,26].

Figure 1.2 Block diagram of an advanced phased array.

Compared to the conventional phased array approach, many advantages are derived from the use of T/R modules. A low power solid state source may be employed, and losses through the power divider network are much smaller in absolute terms, resulting in a highly reliable and efficient system. Also, operation of the phase shifters at lower power levels can lead to faster switching and simpler implementation [20]. An important advantage in the receiving case is that the noise figure is largely set by the module's low noise receive amplifier, instead of by the losses incurred in the signal path. A recent study by Mailloux has shown that a 20 dB increase in link margin is possible by using the T/R module configuration [16]. Typical MTBF figures for all-solid state T/R modules are of the order of 10^5 hours [18]. More importantly, the failure of a single module has very little impact on overall system performance, thereby ensuring "graceful degradation" of the phased array. Examples of T/R module based systems abound, as this has been the preferred approach over the past 15 years [27].

T/R modules and radiating elements have increasingly used microwave integrated

circuit technology and photolithographic techniques in their development. It is well recognized that higher levels of integration result in fewer individual subcircuits, and hence less manipulation to assemble them into complete modules [28]. This has a direct impact not only on cost but also on performance, since assembly operations can introduce variations from one unit to another (bond wire lengths, positioning, etc.) and transitions normally entail losses. Even if complete T/R modules are made on a MMIC chip, they must still be interfaced to the radiating elements on one side and the feed network on the other. At C-band, where a module and associated radiating element might each be 4 cm wide, the interconnection step does not pose a difficult manufacturing or performance problem [24]; however, the same is no longer true for a millimetre wave frequency module and antenna which could be only 1/4 cm in size [10]. Thus, at the EHF frequencies envisaged for Personal Communications, the interfacing step is a difficult problem.

2.3 Current Trends

Within the general framework of the T/R module-based phased array, the current techniques being followed to integrate the T/R active circuitry with the radiating elements and the feed networks (both DC and RF) fall into the categories of either hybrid or monolithic construction.

Hybrid integration, where the various components or functions are fabricated individually and combined in some sort of package, has been widely adopted in recent phased array developments [15,16,29]. In one hybrid approach, multiple MMIC chips are each optimized for a specific purpose such as low noise amplification, phase shifting or switching, and mounted on a thick film printed carrier containing the radiators and the

RF power combiner. The end result is a complete subarray formed on a single substrate, or motherboard. When the requirements of the phased array demand that several such subarrays be assembled, a "brick" style of array construction may be used [16], as shown in Figure 2.3. The parallel substrates depicted are column subarrays, which are oriented perpendicularly to the array radiating aperture.

Figure 2.3. "Brick" construction for subarray assembly.

Brick assembly is not compatible with the low profile requirement imposed by handheld terminals and other conformally mounted phased arrays. Thus, an alternative assembly technique, "tile" construction as illustrated in Figure 2.4, is preferred. In this case, area subarrays are placed side by side to form the complete array and a different hybrid approach, the layered technique, is taken for the subarray design. Each layer performs only a given function or group of functions such as RF power distribution, or phase shifting and amplification. One possible subarray implementation is shown in Figure 2.5, in which 7 layers are employed to accommodate all the necessary components,

Figure 2.4. "Tile" construction for subarray assembly.

Figure 2.5. Layout of a layered subarray.

including ferrite-based circulators and thermal conductive paths [15].

The layered structure requires numerous vertical feedthroughs to electrically connect one layer to another. Such connections have conventionally been provided through miniature coaxial cables or vertical pins, which represent prohibitive costs for large EHF arrays. A new multilayer ceramic packaging technology shows promise of alleviating the vertical interconnection problem [15]. Via holes can be stamped and filled with conductive ink where needed on multiple layers of alumina and metal, and then stacked together using the co-fired ceramic process. Thus, all circuit interconnects, either patterned horizontally or fed through vertically, can potentially be realized without wiring in a single, rigid package. The performance, yield and reliability of such a technique at EHF, however, are still unknown. In fact, the ultimate success of the hybrid approach in general will depend on the quality and cost, on a mass production scale, of the various assembly techniques it employs.

Full monolithic integration, in which an entire array or subarray is fabricated in a single processing sequence on a large GaAs chip, is also being pursued at frequencies above 20 GHz where the size of the radiating element is relatively small in terms of the available wafer area [30]. An example of a MMIC subarray on GaAs, designed for 44 GHz operation and suitable for tile assembly into an array, is given in Figure 2.6.

Figure 2.6. Layout of a GaAs monolithic subarray.

Since the chip must incorporate all of the functions on a single layer and within the half wavelength interelement spacing dictated by the array, a severe, performance degrading, crowding problem arises with this approach. A brick configuration would lead to a low density subarray, although at the expense of a nonplanar resultant array. Currently, much effort is being directed towards reducing the area requirements of MMIC components so as to make the tile configuration more feasible [31]. Other aspects of the monolithic approach which must be addressed to achieve this goal relate to the device characteristics and antenna element radiation properties of GaAs MMICs. These issues are dealt with thoroughly in the following chapters and only a brief summary is given here.

In terms of MMIC devices, a number of limitations must be overcome. Of primary importance is the low efficiency of GaAs devices which produces thermal problems in a high density monolithic array. However, new, more efficient device technology such as pseudomorphic HEMTs has the potential to substantially reduce the thermal loading. A second issue is the use of a single semiconductor process for all active circuits. Compromises in monolithic circuit performance must be made unless an optimum process is used for the realization of low noise amplifiers, another for high power amplifiers, switches, etc. Finally, the yield of wafer-scale monolithic systems is presently low, as processing variations across a GaAs wafer limit the uniformity of identical circuits at different locations. Nonetheless, recent reported results have shown substantial progress in improving the two last shortcomings [32].

The efficiency of radiating elements on GaAs is often considered an impediment to the realization of monolithic arrays, since the high dielectric constant of the material causes the fields to remain tightly bound to the structure in many cases. Certain classes of antennas, such as slot radiators, can however be used with good efficiency for this purpose. Thus, full monolithic integration of a subarray on GaAs represents a difficult

challenge due to the improvements needed in MMIC components, but holds the possibility of providing size, weight and performance benefits (as well as cost in large volume production levels) over the hybrid approach.

Before ending this section, it is instructive to compare the radiated RF power levels required from a phased array for three potential 30 GHz personal communications applications: a geosynchronous satellite link (GEO), a low earth orbit satellite link (LEO), and a wireless in-building system (WIB). The comparisons are for a 4.8 kbps data transmission channel having a 10^{-5} bit error rate, which corresponds to a carrier-to-noise ratio (C/N) of 55 dB Hz [6].

For the GEO system, the distance between transmitter and receiver is taken to be 36 000 km while the satellite G/T is assumed to be 23 dB/K, comparable to that of the ACTS satellite [6]. Under these conditions, a straightforward link budget analysis (with 10 dB link margin) yields a required EIRP from the phased array of 27 dBW. If the array contains 50 elements and the elemental gain pattern is 2 dB, the total power transmitted by the array must be 6 W, or 120 mW per element. At 20% efficiency per active element, 30 Watts would therefore be dissipated as heat. Such a thermal load could cause serious performance problems in a small MMIC array.

In the case of a LEO system such as Motorola's proposed IRIDIUM [3], the distance becomes 755 km. Repeating the analysis with the same array parameters and satellite G/T but with the link margin increased to 20 dB, a total transmitted power of 27 mW is obtained. This translates into 135 mW lost as heat, a much more manageable thermal scenario than for the GEO example.

Finally, for the WIB link one need only consider a single element possessing a broad radiation pattern. Assuming for the receiver (low cost) a G/T of -26 dB/K and taking the distance to be 20 m, the power budget analysis with a 30 dB link margin leads to approximately 1 mW of transmitted power. This represents a negligible thermal problem.

2.4 New Technologies

Digital and optical technologies have recently been applied to phased array systems, and in this sense may be considered new. Where conventional microwave technology appears to be reaching its limits in implementing certain array functions, these complementary technologies are emerging to improve the performance of future phased arrays.

In the case of RF power distribution to the array elements, the traditional network of transmission lines and analog power dividers (combiners as well in multi-beam applications) are often bulky, lossy and bandwidth limited [15]. These drawbacks can be overcome with the use of digital beamforming, a concept which completely eliminates the RF feed network [33]. A typical block diagram of a transmitting phased array using this principle is shown in Figure 2.7. The amplitude and phase of the signal to be transmitted at each element are synthesized by a computer, whose digital outputs are then converted to a low frequency analog waveform and finally introduced into the T/R

Figure 2.7. Block diagram of a phased array using digital beamforming.

module. An enhanced T/R module, which includes an upconverter and filtering, is required, as it must accept the output of a D/A converter in the MHz range. The number of array elements that can be fed this way is only limited by the processing time of the digital circuitry, although the additional complexity of the T/R module and the restricted signal dynamic range are also limiting factors.

Photonic technology has advanced to the point where it is now used for signal distribution in phased arrays. Two types of distribution are possible: control signals for the settings of each element's phase shifter and variable attenuator, or the actual RF signals that are applied to the T/R module. An example of a phased array employing both types of optical distribution is illustrated in Figure 2.8.

In terms of T/R module control lines, the signals carried in optically may be converted at the module to DC electrical inputs [34], or instead used directly to provide optical control of the semiconductor devices [35]. For the case of RF signal distribution, an appropriately modulated laser source places the RF signal on an optical carrier, which can then be linked to the individual T/R modules using fiber optics or other optical waveguides. A high-speed photodetector is required to convert the signal back to the microwave regime, resulting in a loss of typically 20 dB [35].

Figure 2.8. Block diagram of a phased array using optical signal distribution.

Another optical application to phased arrays is the replacement of conventional phase shifters with true time delay units using optical delay lines [36,37]. This is particularly well suited to wideband systems where a constant time delay, rather than a constant phase shift, is required to steer the beam without dispersion. If such a time delay scheme is used along with the optical components shown in Figure 2.8, the photodetector would be placed after the phase shifter, leading to a larger and larger portion of the phased array system being implemented optically.

The use of photonic techniques is advantageous from several points of view. It is inherently wideband, thereby allowing several control signals, or even RF signals, to be multiplexed onto a single waveguide. Furthermore, optical waveguides are a compact, light weight and low loss transmission medium, providing insensitivity to coupling and crosstalk [38]. Based on these qualities alone, the insertion of photonics into phased array systems has begun; however, to enjoy widespread use a number of drawbacks must be overcome. Optical active components (lasers, detectors, etc.) are costly, limited in dynamic range, and exhibit large RF to lightwave conversion losses which also entail poor noise figures. With the rapid pace of research and development taking place in this field, these shortcomings are expected to diminish in the next few years. Already, the problem of realizing optical components in an integrated circuit format is alleviated by the

advent of planar waveguide structures and opto-electronic integrated circuits (OEICs) [38,39]. The possibility of performing monolithic microwave/digital/optical integration for phased array applications is now being actively pursued [40].

2.5 Conclusion

The last 25 years have seen enormous changes in the way phased array systems can be built, allowing increased performance to be obtained along with the additional benefits of smaller size and weight, lower power consumption, and potentially reduced cost. This progress has brought about the possibility of achieving EHF personal communications, among other systems, through the use of monolithic phased arrays. The realization of such an array, when based on the tile approach, requires that the MMIC devices and radiators have not only good efficiencies, but also small area requirements in order to reduce the crowding problem on the wafer. The application of new technologies to phased array development has increased the number of tools by which these goals may be attained. One approach may employ digital techniques as an efficient and compact way to realize the beamforming function, leaving a more complicated T/R module to be implemented with microwave techniques. Another may reduce the microwave portion to a minimum (e.g. the output amplifier stage only), in favour of integrated optical techniques for RF distribution and phase shifting. The optimum choice, however, cannot be made without considering the efficiency and size implications of the radiating element and its integration with the active circuitry.

CHAPTER 3. ACTIVE ANTENNAS: OVERVIEW

3.1 Introduction

The term active antenna is used for any type of radiating structure where devices are either embedded in, or located close to, the radiating element. The devices are for the most part solid-state active components, configured in a wide variety of ways (i.e. oscillator, amplifier, detector) to enhance the operation of the antenna and the overall system in which it is placed. The following features of active antennas illustrate some of the enhancements that are possible.

1) Due to an obscuring of the dividing line between the active circuitry and the antenna, active antennas provide a compact way of integrating the radiator with the remainder of the system. For this reason, active antennas are also called integrated antennas.

2) Size reduction of the antenna may be achieved with the use of active devices. In a receive mode of operation, there are normally losses associated with the physical separation of the antenna and the low noise amplifier (LNA). Elimination of these losses by placing the LNA at the radiator results in an increased system figure of merit (G/T) and hence possibly smaller antenna size. On transmit, the use of a smaller active antenna that incorporates an amplifier can provide the same gain as a larger "passive" antenna.

3) Active antennas offer the possibility of increased operating bandwidth. The device employed may have an impedance characteristic that compensates the impedance variation of the radiator, thereby broadening the power transfer bandwidth of the combined unit.

4) The presence of integrated devices extends the antenna's functionality. Among the additional capabilities that may be obtained in this manner are the variation of the radiator's gain, beam direction, power pattern and polarization; the tuning of its operating frequency; and the switching of its function (transmit/receive, on/off, etc.).

The study of active antennas is important in this work, as they necessarily constitute part of a monolithic phased array. The choice of a specific type of active antenna for this application depends on such factors as the array's technology usage and configuration (Ch. 2), the degree of compactness provided by the active antenna, and ultimately its performance. In this chapter, an overview is given of the reported active antenna realizations that are most relevant to the design of monolithic phased arrays. Hybrid as well as monolithic configurations are considered, since many hybrid structures lend themselves to monolithic implementation. A familiarization of the many types of active antennas will then allow their suitability to be assessed in terms of phased array building blocks and, if need be, will also indicate the context in which new structures are to be developed. The field of active antennas is an extremely dynamic one, with an abundance of papers published (and continuing to appear) in the literature [41-60]. Thus, only the most representative examples are presented here, grouped according to their operation as transmitters, receivers, or as both transmitters and receivers.

3.2 Transmitting Antennas

Several active patch elements have been reported over the past four years, in which Gunn diodes, IMPATT diodes, or FETs are integrated directly with the radiators [41-44]. An example of a transmitting Gunn diode patch is shown in Figure 3.1, where L and W are approximately one half and one third of the guided wavelength, respectively [43]. The

Gunn device is shunt mounted between the patch and the ground plane at a distance $x' = 4$ mm so as to match the device's -8 ohm input impedance to that of the patch at resonance. For optimum radiation efficiency the substrate's relative dielectric constant and thickness are chosen to be 2.33 and 1.5 mm, respectively. DC biasing is provided by means of a high impedance microstrip line connected to the patch edge.

In terms of performance, a maximum output power of 14.7 mW is obtained at the 10.2 GHz operating frequency, which may be tuned over a 8 % range by varying the bias voltage. Injection locking to a stable source is also possible to reduce noise and frequency drift. Radiation patterns similar to those of an ideal patch are obtained, with almost equal 90 degree E- and H-plane beamwidths [45]. A variation of the same concept developed by Perkins uses a circular patch and a high power device to produce a radiated power of 14 W at 6.8 GHz [44].

Figure 3.1. Gunn diode active patch

Although the device and substrate employed have nothing in common with MMICs, the Gunn diode patch is noteworthy for two reasons. First, it demonstrates that a useful active antenna can be made without matching circuitry by properly locating the device within the radiator, with the additional benefit x' at a very compact structure results. Second, incorporation of this element in a phased array is possible since injection-locking can ensure that the individual oscillators operate at the same frequency and have appropriate relative phasing.

Two types of active transmitting patches using FET oscillators have been reported separately [41,42]. In the first circuit, shown in Figure 3.2, a patch antenna is placed in the feedback loop of a common source FET amplifier [42]. In this configuration the patch also acts along with the matching components as a frequency selective resonator, such that oscillation takes place at the operating frequency due to positive feedback. The FET-patch is built on a low ϵ_r substrate with a packaged device, and radiates 17 mW of power at 5.7 GHz. Smooth radiation patterns are obtained in the E- and H-planes with respective 3 dB beamwidths of 56 and 32 degrees.

Figure 3.2. FET-based active patch.

The second design, illustrated in Figure 3.3, directly integrates the FET into the patch antenna on a low ϵ_r substrate [41]. A narrow slot splits the patch into two sections to allow the FET's gate and drain leads to contact the opposite sides while remaining DC isolated. Bias connections are made by quarter wave RF-shorted stubs on the edges of the patch. The FET operates at zero gate-to-source bias, with capacitive feedback provided by the slot so as to oscillate at 8 GHz. In this manner, variation of the bias voltage allows frequency tuning over a range of 250 MHz, while the radiated power is estimated to be about 6 mW. The H-plane pattern agrees well with theoretical predictions for a rectangular patch, unlike the E-plane pattern which has 2 dB ripples thought to be caused by spurious radiation from the slot [41]. Recent extensions of the active patch

concept, which use two devices inside a single patch to improve the radiation characteristics and increase the power output of the active antenna, have been reported [45].

Figure 3.3. Active patch with integrated FET.

The advantages of using FETs in active antennas are mainly performance-related. FETs have higher DC-to-RF conversion efficiencies and lower noise figures than either IMPATT or Gunn diodes, and are compatible with monolithic fabrication techniques [41]. This last point, however, brings out the fundamental problem associated with all of the active patch antennas examined: monolithic integration requires that they be placed on a high ϵ_r substrate, thereby limiting their radiation efficiency to less than 55 %, as will be seen in Chapter 4.

One solution to the low radiation efficiency problem is to employ two substrates in the active antenna design. Thus, the patch can be printed on a low ϵ_r substrate and the active devices fabricated on high permittivity semiconductor material, with the interconnection being performed electromagnetically through a coupling slot. Figure 3.4 shows an example of an oscillator antenna using this approach [46]. A GaAs FET is configured as an oscillator on the lower Duroid substrate (not a semiconductor in this

Figure 3.5. Monolithic millimetre wave active antenna.

RF efficiency when the GaAs substrate thickness is chosen to be 125 μm . If the thickness is reduced to 75 μm the efficiency is found to decrease to 3 %, although it is not clear whether this is due to the IMPATT diodes operating under a less optimal loading condition, or to the antenna radiation efficiency decreasing with substrate thickness [47]. However, from the results of Chapter 4, it may be stated that the thicknesses employed are below the $0.06\lambda_0$ or 410 μm required for optimum radiation efficiency, and hence a larger radiated power would be possible from this active antenna design.

A linear array of the IMPATT active elements was also constructed by Camilleri and is shown in Figure 3.6. The problem of injection locking to ensure in-phase operation of all oscillators is solved by the use of transmission lines connecting the elements together. The coupling strength can be adjusted by varying the characteristic impedance of the line and the radiating elements may be placed less than λ_0 apart while

still locking the oscillators in phase, since the wavelength in a GaAs microstrip is about one third the free space wavelength [47]. Thus, the ability to fabricate a monolithic active transmitting array with all elements locked in phase has been demonstrated in this example.

Figure 3.6. Radiation patterns of a monolithic active antenna.

Figure 3.7. Linear array of active IMPATT elements.

The last type of active oscillator antenna to be reviewed here consists of a single substrate with a slot radiator formed on one metallized side, and the active circuitry placed on the other, as shown in Figure 3.8 [48]. This hybrid design uses an appropriately matched FET as the active device, which is electromagnetically coupled to the 1.0λ long

Figure 3.8. Active slot antenna.

slot using a quarter wave microstrip to slotline transition. Approximately 13 dBm of effective radiated power is produced by the active antenna at 9.3 GHz, in a broad, single-lobe pattern. The frequency may be tuned over a 1 % range by varying the drain bias voltage or, in a modified version, over a 0.5 % range by varying the optical illumination of the device [49]

3.3 Receiving Antennas

Integrated active receiving antennas suitable for arrays have been developed primarily around slot or dipole elements [50-53], while patches have been used in traveling wave style active antennas [54-55]. These compact units, also known as quasi-optical receivers, combine the functions of antenna, local oscillator, mixer, and low pass filter to produce an IF signal at the output. One such receiver circuit employing a modified coupled slot antenna is shown in Figure 3.9, where the dotted lines indicate components situated on the opposite side of the substrate [50].

Figure 3.9. Quasi-optical receiver using coupled slot antennas.

For a RF signal propagating in the even mode, the coupled slotline structure behaves as a pair of resonant slot antennas, whereas for a LO signal propagating in the odd coupled slotline mode, the structure becomes a coplanar waveguide. Thus, the receiver operation is as follows: the radiated RF signal is received across the slots and is downconverted at the balanced mixer (Schottky diodes located in the centre) with the LO signal, which is generated by the FET oscillator (located at the extreme right) and fed to the mixer along the coplanar waveguide line. The IF signal resulting from the mixing operation passes through the low pass microstrip filter on the left side of the circuit. Similarly to the active transmitting antenna case, the connected coupled slot is used simultaneously as a half wave resonant antenna and as an embedding circuit for the local oscillator [50].

It is found that the antenna patterns are similar to those of a single slot antenna since the spacing between the two slots is small enough that they both receive the RF signal at approximately the same phase. The efficiency of the receiver as an antenna device is characterized using the isotropic conversion loss parameter, defined as the ratio of IF power to the RF power that would be received if an isotropic antenna had been used

under the same conditions [50]. For this circuit, fabricated on a substrate with a dielectric constant of 2.5 and operating over a 9.6 to 10.5 GHz RF band, the minimum conversion loss is 4 dB at 10.1 GHz.

A variation of the previous receiver which provides a higher degree of integration and greater compatibility with FET-based MMIC technology by using no diodes, is illustrated in Figure 3.10 [51]. The mixer diodes and separate oscillator FET are replaced by two self-oscillating FET or HEMT mixers located at the centre of the coupled slots, with the circuit being fabricated on a low ϵ_r substrate and operating at 10 GHz as before. This leads to an isotropic conversion gain of 4.5 dB instead of a loss, and a noise figure of 7 dB. Although the devices are mounted on the side of the ground plane, they could be placed on the opposite side in a monolithic implementation and appropriately connected to the ground side with via holes. It should be noted that the receiver noise figure, optimized in this case with the use of low noise HEMT devices, is well above the 2 - 3 dB range of values considered acceptable for communications applications.

Figure 3.10. Quasi-optical coupled slot receiver using only FET devices.

Receivers with identical operating principles have also been reported, where twin dipoles are used instead of slots [52]. Low ϵ_r substrates are still employed but are

covered with fused quartz superstrates to prevent losses to surface modes. Conversion losses of 6 dB were obtained for a 63 GHz RF frequency. Nearly equal H- and E-plane radiation patterns result from the half wavelength spacing of the two dipoles.

Nesic recently developed a uniplanar equivalent to the quasi-optical receivers described so far [53]. The circuit, shown in Figure 3.11, uses coplanar waveguide techniques to integrate the FET local oscillator, mixer, two slot antennas, and low pass filter all on the same side of a low dielectric constant substrate. By placing the slots half a wavelength apart, an array gain of 3 dB is obtained over the coupled slot configuration in [50]. Also, the slots are operated at resonance so as to provide a broader frequency range over which the devices may be matched [53]. These improvements allow a 1 dB isotropic conversion gain to be obtained at 10.3 GHz, with approximately the same beamwidth (60 degrees) in the E- and H-planes. Since all circuitry is placed on the same side of the substrate the need for via holes is eliminated and circuit construction is simplified.

Figure 3.11. Uniplanar quasi-optical receiver.

An innovative receiving array of integrated horn antennas has been developed by Rebeiz et al. for millimetre-wave imaging applications [56]. As shown in Figure 3.12(a), each antenna consists of a dipole suspended on a 1 μm dielectric membrane in a

pyramidal horn. The horn is formed by anisotropically etching a stack of two silicon wafers, as depicted in Figure 3.12(b), while the dipole is printed along with an integrated microbolometer (detector) on the membrane. A mixer circuit may also be obtained by replacing the bolometer with a Schottky diode epoxied at the dipole apex. In terms of operation, the energy collected by the horn is focused onto the dipole and delivered to the circuitry on the membrane, or on the semiconductor wafer between the horns (Fig. 3.12(b)), for further processing.

Figure 3.12. Integrated receiving horn array. (a) Perspective view; (b) Side view.

Several versions of the array have been developed for the 94 GHz band and other frequencies in the upper millimetre-wave range. With gold plating of the horn surfaces, the maximum attainable aperture efficiency is around 90 %, which indicates the absence of substrate mode and dielectric losses in the structure. Furthermore, the mismatch losses between the device and the dipole are virtually eliminated by conjugate matching of

their impedances. Thus, a very efficient receiver is obtained, with a single element conversion loss of 5.4 dB (SSB) reported at 92 GHz, using an LO frequency of 90 GHz.

The size of the horn apertures, which is also the period of the array, is typically 1.0 to 1.5 λ , producing a resultant multibeam pattern suitable for imaging applications. In order to be considered for phased array applications, the flare angle of the horns must be reduced from the 70.6° naturally obtained with the <111> crystal planes of the silicon (or GaAs) wafer. Smaller angles could be achieved with the use of more expensive fabrication techniques such as ion-beam milling or reactive-ion etching.

3.4 Transmitting/Receiving Antennas

The remainder of this chapter deals with active antennas which employ the same basic configuration to perform the transmitting or receiving functions. These are generally characterized by a similar device/radiator interface.

A number of monolithic active antennas and subarrays have been produced by the Rome Air Development Center specifically for phased array applications at EHF [57]. Figure 3.13 shows a 44 GHz transmitting active array with four circular patch radiators and associated FET amplifier, phase shifter and RF distribution circuitry. The patches, providing circularly polarized radiation, are separated by a half free space wavelength from centre to centre. Unlike the oscillator-radiators discussed in section 3.2, an RF signal must be supplied to the array.

Figure 3.13. Monolithic transmitting active circular patch array.

The circuit was fabricated monolithically on 125 μm thick GaAs as a success-oriented end run attempt; however, acceptable radiation performance was not obtained [57]. It is clear from the properties of microstrip patches that low radiation efficiency is at least partly responsible for this result, although other factors are also cited such as poor impedance matching and processing tolerances. A more recent redesign of the subarray achieved high efficiency amplification and low loss phase shifting, but the radiation characteristics were not reported [58].

A "hybrid monolithic" receiving active element is illustrated in Figure 3.14, in which the radiator is mounted on a low dielectric constant quartz substrate that is bonded to the GaAs layer. The active circuitry is the same as for the transmitting case, while the antenna is a dual-fed proximity coupled microstrip circular patch used to achieve circular polarization. No performance results for this circuit are given other than the claim that a passive version of the design performed well enough to warrant development of the active antenna. Four-element subarrays of these antennas were fabricated for operation at 20 GHz and, without further reports of this work since 1988, it may be deduced that satisfactory performance was not achieved [57].

Figure 3.14. Monolithic receiving active circular patch.

The two active arrays described above appear to be the only examples of monolithic phased arrays ever developed. In spite of their performance shortfalls, these circuits indicate that sufficient space is available within the half wavelength element spacing to accommodate the active circuitry required for phased arrays.

A generalization of the aperture-coupled oscillator patch which applies to both transmit and receive operation is the active antenna shown in Figure 3.15, comprising a circular patch and an electromagnetically-coupled amplifier and mixer chip [59]. The patch is fabricated on a quartz substrate ($\epsilon_r = 3.8$) while the active circuitry is realized in GaAs MMIC form. Matching of the patch to the 50 Ohm impedance of the amplifier is achieved at 38 GHz by appropriately choosing the dimensions of the coupling slot and the length of the tuning stub; however, this feeding mechanism has a total loss of 1.5 dB. The measured patterns of the transmit active antenna show single-lobe radiation in both planes with better than 16 dB front-to-back radiation ratio.

Figure 3.15. Aperture-coupled active patch antenna.

An entirely different approach to the design of active transmit or receive antennas is that of two-dimensional arrays where active devices are embedded in a single-function grid [60-62]. Power is coupled quasi-optically to and from the grid, which can be designed to perform phase shifting, frequency multiplication, mixing, amplification, or RF generation (oscillator).

A typical grid device is the 100-element Schottky diode grid mixer [60], shown in Figure 3.16(a) along with its layout in Figure 3.16(b). Diodes are placed at the center of a bow-tie shaped unit cell which is defined by the grid symmetry. From an equivalent circuit point of view, the diodes in each column are connected in series. The incident RF and LO voltages are given vertical polarization, such that the IF voltages add along each column and are collected off the top and bottom edges of the grid. If the devices are assumed to be identical, then for a grid infinite in extent, each diode lies in an equivalent waveguide unit cell with electric walls on the top and bottom, and magnetic walls on each side. Thus, circuit power scales with total area, but circuit impedances are determined by the unit cell.

Analysis of the impedance presented to the diode terminals is facilitated by the

small size of the unit cell compared to a wavelength. The grid period is on the order of $\lambda/10$, allowing the induced EMF method to be used to produce a simple transmission line model of the structure that includes the equivalent circuit of the device. In this manner, the reflection coefficient of the grid for an incident plane wave can be determined and controlled so that it matches to free space.

The grid mixer of Figure 3.16 is fabricated on a 30 x 30 mm Duroid substrate, and operates at 10 GHz with a flat metal mirror behind the grid. With an LO frequency of 10.2 GHz, the measured conversion loss and noise figure are 10 dB and 8 dB, respectively, which is comparable to the performance of a single-diode mixer. Other types of active antenna grids have also demonstrated excellent performance as amplifiers, oscillators and phase shifters. This leads them to be considered as building blocks for complete receivers or transmitters, which could be formed by stacking together appropriately functioning arrays. Current obstacles to the realization of such systems are the required ground plane backing of all grid types except the amplifier, and the need to supply to the mixer grid the LO power together with the RF input power.

Figure 3.16.(a) Schottky diode grid mixer; (b) Grid layout.

Parfitt and Griffin proposed an active antenna concept consisting of a dipole supported on a dielectric slab that is extended from the MMIC wafer, as shown in Figure 3.12(a) [63]. In principle the structure may be designed such that the dipole radiates mostly in the endfire direction, away from the MMIC containing the active components. To form an array, several of these active elements would be combined in a brick configuration, as described in Chapter 2. This allows the depth behind each radiator to be used to any extent as a way of alleviating the crowding of the active circuitry experienced in the previous tile configurations.

Figure 3.17. Proposed MMIC active dipole antenna.

A departure from conventional microwave approaches to active antennas has been suggested by Baca and McInerney in a concept paper [64]. An optical fiber carries the signal to the chip, at which point it is converted to an electrical signal in a monolithically integrated optoelectronic device on GaAs. Further amplification and radiation would then take place in the integrated amplifier and slot antenna, as shown in Figure 3.18. Results were obtained only for the flared antenna structure, which radiated in an endfire direction, indicating its likely utilization in a brick type of configuration.

A proposed active antenna, consisting of a HEMT device monolithically integrated with a slot antenna on the same side of a GaAs substrate, has been reported [76]. However, no results of its performance were given.

Figure 3.18. Proposed monolithically integrated optoelectronic active antenna.

3.5 Critical Assessment and Conclusion

Extremely diversified in concept, the active antennas studied in this chapter illustrate numerous techniques for integrating active devices with radiating elements to form compact units. Although in many cases they are intended to be used directly as integrated receivers or transmitters, all may be considered also as building blocks for phased array antennas.

The active oscillator antennas of Sec. 3.2 are attractive configurations in terms of generating the signal to be transmitted at each element, and coupling it efficiently to the radiator. In this manner, sophisticated RF distribution networks are no longer needed, replaced by simpler injection locking schemes which are capable, theoretically, of performing not only the information modulation of the oscillator, but also the setting of its phase. However, in general, phased arrays also require amplitude control of each element's excitation, and this function cannot be provided with injection locking.

Attempting to set the amplitude of the output signal by varying the oscillator's DC bias voltage results in a shift of its frequency, thereby destroying the phase locking. Thus, with so many parameters to be controlled by a single device, the oscillator antenna is not suitable for phased arrays.

With regard to the receiving active antennas of Sec. 3.3 where the down-conversion takes place at the radiating element, subsequent processing may be carried out at a lower IF, again allowing the removal of RF combining networks from the array. However, since a mixer constitutes the first stage of such receivers rather than a low noise amplifier (LNA), higher than necessary noise figures are obtained. For the 20 - 30 GHz band, the lowest noise figure of a mixer is in the 7 - 10 dB range, whereas that of a LNA is 2 to 3 dB [65]. Integrated mixer antennas are therefore inappropriate for use in low noise phased array receivers such as those required in communications applications.

Finally, among the transmit/receive active antennas presented in Sec. 3.4, the grid format was found to be ill-suited to compact phased array construction. On the other hand, the remaining designs, although not completely successful in terms of operation, are deemed acceptable for phased arrays as they integrate the device and radiator in an amplifier/antenna configuration. This way, a transmitting array may have an output amplifier as the last stage of the signal processing chain, whose bias can be varied to achieve amplitude control (along with amplification). Likewise, for a receiving array the first stage may be a LNA which ensures optimum noise performance of the system.

In conclusion, the desired active antenna for the phased array application considered in this work consists of an amplifier/antenna element. All of the techniques employed in the other active antenna configurations may still be useful for realizing such an element. However, in the case of a GaAs MMIC implementation, the performance of the radiator itself can be quite different from that of a hybrid equivalent constructed on

low dielectric constant substrates. Further study of the radiating properties of antennas on GaAs is therefore required.

CHAPTER 4. RADIATION FROM MONOLITHIC HIGH DIELECTRIC CONSTANT PLANAR STRUCTURES

4.1 Introduction

At small enough wavelengths, antennas and arrays can be made as part of a thin-film integrated circuit, fabricated on a single high dielectric constant substrate such as silicon or GaAs. These integrated-circuit (IC) antennas become feasible at frequencies above approximately 15 GHz when the element area requirements are small compared to the size of a wafer, typically 3 inches in diameter [66]. Since the material's dielectric constant is high, the electromagnetic fields tend to remain bound to the structure, leading to poor performance in some cases. The important performance considerations for IC antennas are radiation patterns, impedance and efficiency. A single beam is the only pattern requirement as the beam narrowness, or gain, can be controlled by the array characteristics. The antenna impedance must be known accurately for matching to the transmitting or receiving circuitry, and the radiation-to-air efficiency is desired to be maximized since RF power is limited on an IC. As for other parameters, a wide operating bandwidth is only of secondary importance here. At 30 GHz even a relatively small value of 3 % still provides almost 1 GHz of useable bandwidth, which is considered broad for many communications purposes. Also, circular polarization being obtainable from linearly polarized elements, antenna polarization is not an important performance parameter and the remainder of this work is limited to the linear case.

Only resonant-type metal antennas such as patches, dipoles and slots may be considered as candidate elements for monolithic phased arrays with a T/R module at each

element, as they are small enough to fit in a half-wavelength grid spacing. Excluded from this discussion therefore is the wide variety of printed circuit traveling-wave antennas (bow-ties, spirals, flared slots, etc.) as well as dielectric antennas, which are either tapered dielectric rods or leaky-wave structures (dielectric guide, image guide, and variations thereof). Both types are several wavelengths long, with the flared slot and rod providing end-fire radiation, and the leaky-wave antenna radiating from the normal to its interface [67,68]. Although their fabrication is compatible with IC technology, traveling-wave antennas have dimensions which far exceed those available for phased-array elements.

In this chapter patch, dipole and slot antennas are described with emphasis on the effects of the dielectric substrate, and some of the available techniques for improving their performance are reviewed. Then, with the slot radiator emerging as the most suitable element for integration into GaAs MMICs, a method for the analysis of the slot input impedance is presented. Finally, results of the analysis are given and compared with experimental slot input impedance measurements.

4.2 Characteristics of Resonant-Type Structures

4.2.1 Performance of Patches, Dipoles and Slots

A microstrip patch antenna consists of a conducting patch of any shape (square, circular, etc.) on one side of a dielectric substrate backed by a ground plane on the opposite side. It may be directly fed by a microstrip line as shown in Figure 4.1, or fed through the substrate by a coaxial line. In an integrated circuit the only practical means of feeding the patch is with a microstrip line which can be inset from the edge or located at

a point along one of its edges to obtain the desired input impedance. A broad enough range of impedances is available to satisfy most requirements [69].

Figure 4.1. Microstrip patch antenna.

The radiation efficiency of patch antennas is limited by the excitation of surface waves, which propagate into the substrate and give rise to end-fire radiation. These waves are TE and TM modes resulting from the propagation of a quasi-TEM wave along the patch at a phase velocity close to that of a surface wave. The lowest order TM_0 mode is always present to some degree, but its effects are found to be negligible when $h/\lambda_0 < 0.03$ for $\epsilon_r \simeq 10$, which corresponds to $h < 300 \mu\text{m}$ at 30 GHz. The lowest frequency at which higher order modes may occur is a function of the substrate ϵ_r and thickness. For a $500 \mu\text{m}$ thick GaAs wafer ($\epsilon_r \simeq 13$) the TE_1 or TM_1 mode is excited above 43 GHz(approximately).

A popular patch antenna is the rectangular patch depicted in Figure 4.1. Typical dimensions are $\lambda/2$ for the patch length and $\sim \lambda/3$ for its width, where λ is the wavelength in the microstrip. From a radiation point of view, the rectangular patch can be viewed (with some assumptions) as being equivalent to two slots spaced $\lambda/2$ apart, excited in phase and radiating in the half plane above the ground plane [69]. Each slot radiates the same field as a magnetic dipole with magnetic current proportional to the

open-circuited voltage at the patch edges. The resulting radiation patterns are given in Figure 4.2 for the case of a patch $\lambda/2$ by $\lambda/2$ in size. The E-plane pattern is dependent on L whereas that of the H plane is a function of W .

Figure 4.2. Radiation patterns of a rectangular patch antenna.

The patch efficiency, defined in terms of radiated power and losses due to the dielectric material and conductor only, increases with substrate thickness and frequency, but decreases with ϵ_r . The maximum efficiency that can be obtained on a 500 μm GaAs substrate at 30 GHz is approximately 70 % [69]. However, this value does not take into account the additional losses due to surface waves which, from the above discussion, are known to exist. If one therefore chooses a 250 μm GaAs substrate (no surface waves) the largest achievable efficiency would be no greater than 55 %. A good review article on microstrip patches has recently been published by Pozar [70].

An IC dipole is a pair of thin collinear strips of metal printed on a grounded or ungrounded dielectric substrate. Although the conventional feed mechanism is a pair of lines as shown in Figure 4.3(a), a single line connection as illustrated in Figure 4.3(b) has recently been demonstrated to simplify the feed structure [71]. Electromagnetic sidecoupling (Figure 4.3 (c)) is one possible feeding technique for the grounded substrate case, which is also referred to as a microstrip dipole [72].

Figure 4.3. Dipole antennas: (a) conventional feed; (b) single-wire feed (c) electromagnetic sidecoupling.

For the dipole on a grounded substrate, determining the power carried in the TE and TM surface wave modes requires a complicated evaluation of Sommerfield type integrals, resulting from the integral equation formulation for the antenna current distribution [73]. In the ungrounded dipole case, the surface wave power is found by applying the Lorentz reciprocity theorem. Consideration of the structure of the propagating modes and the effective substrate thickness leads to simple formulas for the power in a given TE or TM mode. Further details of the surface wave calculations are given by Alexopoulos et al. [73]. The discussion of dipoles will be limited to the microstrip configuration, as it has been shown that the maximum obtainable efficiency on GaAs for an ungrounded dipole is 20 % ($h/\lambda_0 < 0.2$) due to dominating surface waves

[74].

Dipoles are typically $\lambda/2$ long so as to allow operation at resonance, which corresponds to approximately $0.2\lambda_0$ for GaAs [73]. The strip thickness is normally less than a tenth of its length and hence the overall antenna size is much smaller than a patch. It is not necessary for the dipole to be exactly of resonant length, since a single beam pattern will still result as long as the length is kept below about λ ; however, matching to the input impedance may become more difficult as it increases to several hundred ohms at second resonance from close to 50 ohms at resonance with the correct substrate thickness [74].

The resonant dipole's input resistance R_r is given in Figure 4.4 as a function of substrate thickness h [73]. Superimposed on this figure are the input resistances due to radiation into air (R_{rr}) and the excitation of surface waves (R_{rs}), where $R_r = R_{rr} + R_{rs}$. It is seen that the surface wave resistance is always larger than the radiation resistance. Defining the radiation efficiency as

$$\eta = R_{rr}/(R_{rr} + R_{rs})$$

in which the dielectric substrate and antenna conductor losses are omitted, a maximum efficiency of 30 % is obtained at a GaAs thickness of $0.06\lambda_0$, or 600 μm at 30 GHz. Thus the microstrip dipole is also plagued by a low efficiency, as was found for the patch antenna.

Figure 4.4. Resonant dipole radiation and surface wave resistances

The E-plane radiation pattern of a $0.3\lambda_0$ long dipole on a $0.2\lambda_0$ thick substrate is shown in Figure 4.5, from which a 3 dB beamwidth of 62 degrees is measured [74]. The beamwidth of the H-plane pattern is easily made comparable to that of the E-plane with a suitable array.

Figure 4.5. Radiation pattern of a dipole antenna.

An IC slot antenna consists of a thin rectangular opening in the ground plane of a dielectric substrate. Without the substrate this would be the complementary structure of the dipole, in which case the radiation patterns are identical with orthogonal polarization, and the input impedances are related through the simple Booker relationship [75]:

$$Z_{\text{slot}} = (120\pi)^2 / (4Z_{\text{dipole}})$$

With the presence of the substrate these relationships still hold qualitatively, thereby allowing much of the well-developed dipole theory to provide useful guidance for slots [77].

Direct feeding can be achieved with a coplanar waveguide line as shown in Figure 4.6(a), or by a microstrip line which connects to the far edge of the slot through the substrate (Figure 4.6(b)). Electromagnetic coupling can also be used to feed the slot with

a microstrip line by terminating it in an open circuit a quarter wavelength beyond its far edge, as is done in Figure 4.6(c).

Similarly to the dipole, slot antennas can be operated at lengths up to approximately one wavelength in the slotline for a single beam pattern to occur. Again, the question of matching determines what the optimum length should be and it is found that operation at second resonance provides an input impedance close to 50 ohms for slots on thin, high ϵ_r substrates [30].

Figure 4.6. Slot antennas fed by (a) coplanar waveguide, (b) microstrip, (c) electromagnetically coupled microstrip line.

Surface wave modes for slots have also been studied by Alexopoulos et al., who calculated the power contained in them as a function of GaAs substrate thickness in a similar way to the dipole case [73]. From their results plotted in Figure 4.7, the surface wave power is practically zero for thicknesses less than about $0.05\lambda_0$, with a normalized radiated power of 1 on the same scale (not shown). Thus of all the IC radiating structures considered so far, the slot antenna is the one which can exhibit the highest efficiency on appropriately thin substrates; however, this result is somewhat mitigated by the fact that the slot radiates in both directions, with a larger proportion going into the substrate [74]. In some situations unidirectional radiation is desired and hence the effective efficiency will be less than 50 % unless corrective measures such as those described in the next section are employed.

Figure 4.7. Normalized surface wave power versus substrate thickness.

A typical radiation pattern of an elementary slot antenna (similar to that of a resonant slot) on $0.04\lambda_0$ thick GaAs is plotted in Figure 4.8 [74]. With an array of slot elements the very broad E-plane pattern, which has no null on the air side, can be made narrower and the degradation of the pattern smoothness, due to the finite surrounding ground plane, almost completely corrected [78].

Figure 4.8. Radiation pattern of a slot on $0.04\lambda_0$ thick GaAs.

4.2.2 Hybrid Techniques for Performance Enhancement

The integrated-circuit radiating structures considered previously to produce truly monolithic phased arrays all suffer to some degree from poor efficiency due to surface waves or bidirectional radiation. If one accepts to move away from the monolithic concept, the efficiency and several other performance parameters may be improved through the use of lenses or multilayered substrates, as described in this section.

As was seen earlier, surface waves limit the antenna efficiency and can even dominate the radiated power. From the ray point of view illustrated in Figure 4.9, surface waves may be understood as being rays which are incident at an angle larger than the critical angle θ_c and are therefore completely reflected, or trapped [74].

Figure 4.9. Trapped rays in a thin dielectric slab.

The simplest way to eliminate the surface waves and also to increase the antenna sensitivity to radiation from one side only, is to use a substrate lens as shown in Figure 4.10. The rays are now incident normal to the surface and hence do not undergo total internal reflection. A hemispherical lens (Figure 4.10(a)) allows the transmitted rays to pass through unaffected, whereas the hyperhemisphere of Figure 4.10(b) refracts the rays, effectively decreasing the antenna's beamwidth. In designing dielectric substrate lenses, material absorption and reflection losses (at the interface) must be considered as they may be significant at millimetre wave frequencies [74].

Figure 4.10. Substrate lens: (a) hemispherical, (b) hyperhemispherical.

Another approach to the problem involves the use of multi-layered substrates, each less than a wavelength in thickness, as shown in Figure 4.11. In the case of two layers ($\epsilon_3 = 1$), a radiation efficiency of 100 % for dipoles can be obtained through the elimination of surface waves only if the superstrate is a much higher dielectric constant material than the substrate [79]. This would correspond, for example, to an antenna

fabricated on a grounded Teflon substrate ($\epsilon_1 \simeq 2.2$) and covered with a layer of GaAs ($\epsilon_2 \simeq 13$), the opposite configuration to what is needed for IC antennas.

Figure 4.11. Multilayered substrates for dipole and slot antennas.

Roger and Neikirk have considered three and five layer dielectric stacks, with each layer being a quarter of a dielectric wavelength thick, or resonant [80]. When the contrast between alternating layers is made large, i.e. $\epsilon_r = 13$, $\epsilon_r = 2.2$, $\epsilon_r = 13$, surface wave modes are largely suppressed and the radiation efficiency can be more than doubled. Figure 4.12 shows the results of the above choice of dielectrics for slot and dipole elements. In the case of a single slot, the efficiency increases from 18 % to 40 % whereas that of broadside-spaced slots, separated by $0.244\lambda_0$ to obtain identical E and H-plane patterns, increases from 34% to about 80%. Similar results apply to dipoles, with further improvements obtained in a five layer structure. The effect of increasing the number of layers and increasing the contrast between them is to narrow the beam patterns. This comes at the expense of added ground plane and dielectric losses, which increase with frequency [80].

A double layer configuration also exists to improve the performance of patch antennas. As illustrated in Figure 4.13, the radiating patch is electromagnetically coupled through an aperture in the ground plane to a microstrip line feed (direct via connections

Figure 4.12. Radiation efficiency of slots and dipoles on a 3-layer substrate.

may also be employed). This configuration provides several advantages over the single substrate patch, due to the presence of two distinct media. A low ϵ_r for the antenna substrate and a large substrate thickness can be chosen to optimize the radiation efficiency independently of the feed substrate. Only the size of the aperture need be varied to control the input impedance over a wide range of values. Also, peak radiation from the feed side of the ground plane has been found to be at least 20 dB below the peak of the patch radiation, which means that pattern degradation resulting from spurious feed radiation is minimal [25].

Techniques have been investigated to make slots radiate unidirectionally. The simplest way, employed by Yoshimura, is to use a ground plane reflector located a certain distance behind the antenna [78]. This method works well for a single slot but has been found to excite a waveguide mode in the parallel plate region when an array of slots is used [28]. Other approaches exist, such as the cavity-backed slot where a metal box surrounds the aperture, but are incompatible with IC technology [28,81].

Figure 4.13. Double layer patch antenna.

In order to avoid the complexities of achieving unidirectional radiation and to focus on the concept of a monolithic active antenna, only slot antennas radiating bidirectionnally into free space will be considered in the remainder of this work.

4.3 Analysis of Slots with Arbitrary Feedpoint Locations

The slot radiator is seen to be the most efficient type of radiator on thin IC substrates, and is therefore appropriate for use in a monolithic amplifier/antenna element. In this section, a method is described for analyzing the input impedance of slots to allow their subsequent matching to MMIC devices. The method is applicable to broadside radiating slots (i.e. non-travelling wave type) which are fed at an arbitrary point along the slot length. Arbitrary feedpoint location, as opposed to conventional centre feeding, offers greater flexibility in obtaining a desired input impedance.

The geometry of the slot element under consideration is shown in Figure 4.14, in which the dielectric substrate and ground plane are assumed to extend to infinity in both the x and y directions. An ideal delta-gap generator represents the feed mechanism such that the driving point impedance of the slot is obtained. In this form, the slot impedance may easily be modified to account for a different feeding arrangement such as the type employed in Chapter 6.

Figure 4.14. Geometry of the slot antenna to be analyzed.

A number of analytical methods exists to determine the input impedance of a slot on a dielectric layer. Due to the finite width of the slot and the presence of the substrate, an exact solution cannot be obtained in closed form. Thus, other approximate numerical methods must be pursued which are suited to radiating structures whose dimensions are electrically small. These may be broadly classified as time domain or frequency domain techniques.

In the time domain, an increasingly popular approach is the finite-difference time-domain (FDTD) algorithm, where Maxwell's time-dependent curl equations, subject to the boundary conditions, are solved by converting them into finite-difference equations across a spatial grid. The application of this method (and the closely related TLM method) to antenna problems generally results in higher CPU time and memory requirements than does frequency domain analysis [82].

The integral equation (IE) technique is the most widely applied method to the analysis of small antenna structures and may be formulated in the frequency domain as well as in the time domain [83,84]. Again, the procedure is more complicated in the time domain and hence only the frequency domain will be considered further. This method represents the antenna problem in the form of an integral in which the unknown quantity, usually the current density on the structure, is part of the integrand. Numerical techniques such as the Moment Method can then be used to solve the integral. Derivation of the IE appropriate to a given problem is based on boundary-condition matching for the electric or magnetic fields.

Many types of planar slot antenna structures have been analyzed in recent years using the basic IE technique. Hadidi and Hamid studied slot apertures in an infinite conducting screen (with no dielectric present) and obtained a numerically efficient solution to the resulting Pocklington IE by using the Fourier domain [85]. Only the

voltage distribution along the slots was given, although the input impedance is easily obtained from these results. Kominami et al. investigated dipole and slot elements (and arrays) printed on infinitely thick dielectric substrates [86]. They also employed a moment method solution in the Fourier Transform domain to derive the input impedance, resonant length and resistance, and radiation patterns for these elements.

In the previous two studies, the same gap generator as the one depicted in Figure 4.14 is assumed for the slot feed. However, other people have considered a non-contacting microstripline feed configuration, where a slot is printed in the ground plane of the substrate beneath the microstrip [87,88]. The slot can be treated as a discontinuity in the propagating quasi-TEM fields of the line, in a similar way to the analysis of waveguide slot elements. By solving for the reflection and transmission coefficients in the equivalent transmission line circuit, the slot impedance may be found. Pozar developed IEs in the Fourier Domain for this case by applying the reciprocity theorem along with a boundary condition on the aperture, and used a moment method solution to obtain the series impedance presented by the slot to the line [87]. An extension of the method to include multi-layered substrates has also been reported [89].

None of the analyses described so far is directly applicable to the slot configuration of Figure 4.14, either because the substrate geometry is different or a specific propagating mode is assumed for the fields exciting the slot. However, important work by Singh and Wilton has recently led to an IE formulation which allows the desired slot driving point impedance to be calculated for the single dielectric layer case, as well as many other cases such as multiple dielectric layers and cavity backing [90]. The key to their formulation is the determination of appropriate Green's functions for each geometry, which in turn require (for the dielectric layer case) that magnetic and electric vector potentials be obtained in terms of Floquet harmonics. Although this rigorous treatment of the problem produces highly accurate results, uncovering even erroneous data reported in the literature, a simpler and more intuitive method proposed by Nesic and described in the next section may be considered as a very good alternative.

4.3.1 Method of Complementary/Dual Structures for Analyzing Slot Antennas on a Dielectric Substrate

The analysis presented in this section follows the procedure of Nesic and Popovic, which basically consists of formulating the problem in terms of an equivalent dipole antenna and solving for the input impedance using numerical techniques. In their paper [91], the transformation of the slot antenna to a magnetically covered cylindrical dipole is given. The numerical results for input impedance provided at the end of the paper are obtained using the moment method solution outlined in the book by Popovic et al. [93].

As stated in Section 4.2, the slot aperture is the complementary structure to a dipole strip formed by the metal removed in making the slot. When a dielectric is present, Booker's principle cannot be directly applied to obtain the slot input impedance from well-known dipole theory, since the symmetry is no longer preserved [1c]. This problem has been overcome by Nesic and Popovic, who found that Booker's relationship can be extended to approximately complementary/dual structures arrived at through a series of transformations on the original antenna [91]. The basic idea, therefore, is to replace the slot and its dielectric medium by an equivalent dipole of circular cross-section and with magnetic coating, then solve for the dipole impedance with relatively simple methods, and finally convert the result back to the impedance of the original slot using Booker's relationship.

The cross-section of the slot antenna on a dielectric substrate is shown in Figure 4.15(a). For the first transformation, a coplanar waveguide (CPW) is formed by considering a pair of infinitely long slots separated by a distance d (to be defined later), as depicted in Figure 4.15(b). Then, by placing a symmetric dielectric layer above the conductor plane, as is done in part (c) of the figure, an equivalent CPW in the TEM-mode sense is constructed, where the inductances and capacitances of the guides are equal. The

system is finally transformed to that of Figure 4.15(d), consisting of complementary conductors in a dual magnetic medium ($\epsilon_r = \mu_r$), which is also equivalent to the two-wire

(a) Slot on a dielectric substrate.

(b) Coplanar waveguide made of two slots spaced a distance d apart.

(c) Equivalent coplanar waveguide in a dielectric sandwich.

(d) Complementary/dual structure of (c).

(e) Equivalent pair of magnetically coated conductors.

Figure 4.15. Transformation of slot antenna on a dielectric substrate to a magnetically coated dipole.

structure in Figure 4.15(e) under the following conditions:

- (i) The radius of the wire conductor is taken to be

$$a = \frac{W_a}{4} \quad (4.1)$$

which is the equivalent radius of an infinitely thin flat strip of width W_a (obtainable by conformal mapping).

- (ii) The outer radius of the coaxial magnetic cover is chosen as

$$b = \frac{W_a}{4} \exp \left\{ \left(\frac{\epsilon_r}{\epsilon_r - 1} \right) \left(\ln \left(\frac{4d}{w_a} \right) - \frac{\pi \epsilon_0}{C} \right) \right\} \quad (4.2)$$

where d is the distance between slots and C is the capacitance per unit length of the system in Figure 4.15(c) [92].

It should be noted that although the distance between slots is not unique in the transformation process, the variation of C as a function of d ensures that a single value of outer radius b is obtained for the final structure. Good results are reported for $d/a \sim 10$.

The equivalent dipole problem (*i.e.*, one of the magnetically coated wires) can be analyzed with the use of the two-potential IE extended to include the effect of a cover, as outlined in the next section. From the resulting current distribution on the wire structure, the input impedance Z_d of the dipole is easily found. Booker's principle may finally be employed to obtain the slot input impedance Z_s as

$$Z_s = \frac{\eta^2}{4Z_d} \quad (4.3)$$

with η the characteristic impedance of free space.

4.3.2 Numerical Evaluation of the Two-Potential Integral Equation for Dielectrically or Magnetically Coated Wire Antennas

Integral equations governing the current distribution along cylindrical antennas can be written in several forms depending on the expressions chosen for the underlying magnetic vector-potential and electric scalar-potential equations, and subsequent mathematical manipulations. Although from a numerical point of view the commonly found Hallen and Pocklington integral equations are quite simple to solve, the "two-potential" equation will be used here instead, as it can be more easily extended to represent a wire antenna with magnetic or dielectric coating [93]. The basic two-potential integral equation for a bare cylindrical conductor (Fig. 4.16), arrived at by enforcing the boundary condition for the total electric field on the wire surface, is

$$\int_0^l \left[I(z') + \frac{1}{k^2} \frac{dI(z')}{dz'} \frac{d}{dz} \right] g(r_a) dz' = \frac{E_{iz}(z)}{j\omega\mu_0}, \quad (4.4)$$

where $I(z')$ is the unknown current flowing along the wire axis,

$$g(r_a) = \frac{e^{-jkr_a}}{4\pi r_a}$$

is the Green's function for unbounded homogeneous medium, $r_a = \sqrt{a^2 + (z - z')^2}$

is the distance between the observation point and the conductor surface, $E_{iz}(z)$

is the z-directed incident E-field on the structure and $k = \omega\sqrt{\epsilon_0\mu_0}$

is the phase coefficient.

Figure 4.16 Wire antenna represented by the two-potential equation.

The case of a wire antenna covered with a thin dielectric or ferrite coating is depicted in Figure 4.17. Based on a quasi-static approximation of the electric and magnetic fields in the covering material, the effect of the coating may be reduced to that of polarization charges and currents, and of magnetization currents, located along two tubular layers of radius a and b [93]. This results in an extended form of the two-potential equation for wire antennas as follows:

$$\int_0^l \left\{ \left[\mu I(z') + \frac{1}{\epsilon_r k^2} \frac{dI(z')}{dz'} \frac{d}{dz'} \right] g(r_a) + \left[-(\mu_r - 1)I(z') + \left(1 - \frac{1}{\epsilon_r} \right) \frac{1}{k^2} \frac{dI(z')}{dz'} \frac{d}{dz'} \right] g(r_b) \right\} dz' = \frac{E_k(z)}{j\omega\mu_0} \quad (4.5)$$

in which

$$r_b = \sqrt{b^2 + (z - z')^2} \quad ,$$

ϵ_r , μ_r are the material parameters of the coating layer and all remaining terms are defined in equation (4.4). Since a slot on a dielectric substrate transforms into a dipole with magnetic coating only ($\mu_r > 1$, $\epsilon_r = 1$), equation 4.5 simplifies to

$$\int_0^l \left\{ \left[\mu_r I(z') + \frac{1}{k^2} \frac{dI(z')}{dz'} \frac{d}{dz'} \right] g(r_a) - (\mu_r - 1) I(z') g(r_b) \right\} dz' = \frac{E_{iz}(z)}{j\omega\mu_0} \quad (4.6)$$

Figure 4.17 Coated wire antenna represented by the extended two-potential equation.

To obtain the current distribution $I(z)$ along the equivalent dipole, a moment method solution may be followed. This general procedure for solving differential, integral and integro-differential equations is well described in several books [Harrington], thus only its application to equation (4.6) will be considered here. Three basic steps are involved in the procedure. First, the current distribution along segments of the antenna is expressed in terms of appropriate basis (or expansion) functions with unknown coefficients. Second, a linear system of equations is formed by "sampling" the defining integral equation with some weighting (or testing) functions. Finally, the system of equations is solved using various matrix techniques.

For the first step, the wire is divided into three equal length segments and a five term power series is chosen as the basis function to represent the current distribution along each segment. Thus,

$$I(z) = \sum i_j(z) \quad , j = 1, 2, 3 \quad (4.7)$$

where

$$i_j(z) = \begin{cases} a_j + b_j z + c_j z^2 + d_j z^3 + e_j z^4 & , (j-1)\Delta \leq z \leq j\Delta \\ 0 & , \text{elsewhere} \end{cases}$$

with $\Delta = \frac{\ell}{3}$ being the segment length.

Upon substituting equation (4.7) into equation (4.6) and performing the derivative of $I(z)$ one obtains

$$\sum_{j=1}^3 \int_{(j-1)\Delta}^{j\Delta} \left\{ \mu_r (a_j + b_j z + \dots + e_j z^4) g(r_d) + \frac{1}{k^2} (b_j + c_j z + \dots + 4e_j z^3) g'(r_d) - (\mu_r - 1) (a_j + b_j z + \dots + e_j z^4) g(r_b) \right\} dz' = \frac{E_z(z)}{j\omega\mu_0} \quad (4.8)$$

which has a total of 15 unknown $a_j, b_j, \dots, e_j$ coefficients. To solve for these, a system of 15 linear equations is therefore required. By stipulating that equation 4.8 be satisfied at three points on each segment (*i.e.*, by evaluating it at discrete matching points $z = z_m$ for $m = 1, 2, 3$), nine of the 15 equations will be obtained. This is the sampling process referred to above, where the weighting functions are Dirac delta-functions. Such a choice has the advantage that no double integrals need be evaluated, but can lead to numerically unstable behavior for certain locations of the matching points. Popovic et al. have found that placing the matching points at

$z_m = \frac{\Delta}{6}, \frac{\Delta}{2}$ and $\frac{5\Delta}{6}$ along each segment generally produces stable solutions. The set

of equations which results from sampling can be expressed in matrix form as

$$[x_{mn}][I_n] = [V_m] \quad , \quad \begin{matrix} m = 1, 2, \dots, 9 \\ n = 1, 2, \dots, 15 \end{matrix} \quad (4.9)$$

Where:

$$x_{mn} = \begin{cases} \int_0^{\Delta} \{\mu_r g(a, m, z')\} dz', & n=1 \\ \int_0^{\Delta} \left\{ \mu_r z^{(n-1)} g(a, z_m, z') + \frac{(n-1)}{k^2} z^{(n-2)} g'(a, z_m, z') \right. \\ \quad \left. - (\mu_r - 1) z^{(n-1)} g(b, z_m, z') \right\} dz', & n=2, 3, 4, 5 \end{cases} \quad (4.10)$$

are the impedance matrix elements,

$$I_n = \begin{bmatrix} a_1 \\ \cdot \\ e_1 \\ \cdot \\ a_2 \\ \cdot \\ e_2 \\ \cdot \\ a_3 \\ \cdot \\ e_3 \end{bmatrix} \quad \begin{array}{l} \text{is the current vector containing the unknown coefficients of the} \\ \text{power series in equation (4.7).} \end{array} \quad (4.11)$$

$$V_m = \begin{bmatrix} E_k(z_1) \\ \cdot \\ \cdot \\ \cdot \\ E_k(z_a) \end{bmatrix} \quad \begin{array}{l} \text{is the excitation vector describing the electric field} \\ \text{impinging at each matching point } z_m \text{ due to the} \\ \text{generator voltage } v_i, \end{array} \quad (4.12)$$

and

$$g'(a, z_m, z') = \frac{\partial}{\partial z} g(r_a) \big|_{z=z_m} = \frac{-(z-z')}{4\pi r_a^2} \left(jk + \frac{1}{r_a} \exp(-jkr_a) \right) \big|_{z=z_m} \quad (4.13)$$

Functions $g(r_a)$ and $g(r_b)$ are now written in terms of their z_m and z' dependencies. The expressions given for $n = 1$ to 5 in equation 4.10 are identical for $n = 6$ to 10 and $n = 11$ to 15 except that the integration extends from Δ to 2Δ , and 2Δ to 3Δ , respectively.

In the excitation vector above, $E_{iz}(z)$ needs to be specified. Since a delta-gap generator is assumed for the excitation of the dipole, where the gap width d tends to zero, an infinite E field (V_g/d) is created between the generator terminals and cannot be treated numerically. This problem is circumvented by using the magnetic-current ring representation of the delta generator, as shown in Figure 4.18.

Figure 4.18 Magnetic-current ring representation of the dipole excitation.

The generator in the gap is now replaced by equivalent magnetic currents on the surface of the conductor according to the general equivalence theorem [93], such that the voltage across the ring is equal to v_i . In this manner, the impressed field along the z -axis can be written as

$$E_z(z) = \frac{v_i a^2}{2} \left(\frac{1 + jkr}{r^3} \right) \exp(-jkr) \quad (4.14)$$

where $r = \sqrt{(z - z_f)^2 + a^2}$ is the distance between the generator (feed point) located

at z_f and an observation point z on the z -axis.

Equation 4.14 shows that the impressed field decays rapidly along the wire due to its

$\frac{1}{r^3}$ dependency.

To present, nine equations have been obtained in terms of the fifteen current-distribution coefficients by evaluating the two-potential equation at each of the nine matching points. To complete the matrix formulation, six additional equations which take into account the junction effects at the end of each segment remain to be defined. These are referred to as the junction-field constraints [93], and must be included in any solution of the current distribution based on the 2-potential equation in order to obtain accurate results. A first set of constraints stipulates that Kirchhoff's first current law be satisfied at each segment end (or node). Thus,

$$I_{in}(node\ j) = I_{out}(node\ j) \quad , \quad j = 1, 2, 3, 4 \quad (4.15)$$

which is the condition that forces the current to be zero at each of the dipole ends ($z_1 = 0$ and $z_2 = \ell$) and continuous across adjacent segments ($z_3 = \Delta$ and $z_4 = z\Delta$). Equation 4.15 is easily incorporated into equation (93), adding four rows to the impedance matrix ($m = 10$ to 13) and the same number of zero entries to the excitation vector.

The second set of constraints applies to the fields in the vicinity of a junction of two segments. Since the solution of the current distribution is only approximate, thereby giving rise to possibly large axial total electric fields (instead of zero theoretically) near the junctions, a nonzero complex power is generated which reduces the accuracy of the result. From consideration of the voltage along the wire axis in the junction region, it can be shown that this power will be zero if

$$\int_{z_{k1}}^{z_{k2}} (E_z(z) + E_{iz}(z)) dz = 0 \quad , \quad k = 1, 2 \quad (4.16)$$

where the integration extends a distance of four wire radii along each segment (i.e., for a junction located at z_k , $z_{k1} = z_k - 4a$ and $z_{k2} = z_k + 4a$) and the sum of $E_z(z)$ and $E_{iz}(z)$ forms the total field. $E_z(z)$ results from the current distribution along the antenna and is obtained from equation (4.5), while $E_{iz}(z)$ is given by equation (4.14).

The integral in equation (4.16) can be evaluated numerically by summing the integrand at four equally spaced points within the (z_{k1}, z_{k2}) interval. Through straight-forward manipulations, equation (4.16) can be seen to consist of the sum of four point matching equations given by equation (4.8), for $k = 1$ and 2. Thus, the second junction constraint adds two final rows to the impedance matrix with elements defined as

$$x(m, n) = \sum_{\ell=1}^4 x(\ell, n) \quad , \quad m = 14, 15 \quad (4.17)$$

with $x(\ell, n)$ given by equation (4.10), while the corresponding entries to the excitation vector are

$$v(m) = \sum_{\ell=1}^4 E_z(z_\ell) \quad , \quad m = 14, 15 \quad (4.18)$$

where $E_z(z)$ is written in equation (4.14) and z_ℓ are the four numerical integration points located in the junction region.

The system of fifteen equations in fifteen unknowns representing the equivalent dipole current is now completely defined, and may be solved using various numerical techniques. To this end, a FORTRAN computer program running on a Sun SPARC2 workstation has been written. It was found that the accuracy (stability) of the computed results is greatly enhanced by using built-in library programs, the IMSL routines, to perform the integration and matrix reduction functions. The results are also affected by the location of the matching points, which is to be expected in the point matching technique. The points are placed according to equation (4.8), itself a function of the segment length Δ . After conducting several numerical experiments, the numerical stability was determined to be greatest for the point locations still given by the rule above, but with the length of the segment

containing the generator reduced from $\frac{\ell}{3}$ to approximately $\frac{\ell}{30}$ (with the

two remaining segments made proportionally longer). This modification is justified by the need to sample the impressed field with a higher density of points in the excitation zone.

4.3.3 Validation of the Computed Results

Presented in the last section was the formulation for a moment method solution of the current distribution along a magnetically coated dipole. From the computed current distribution the dipole input impedance and that of the equivalent (complementary/dual) slot antenna are easily obtained. A computer program has been developed to perform the calculations; in this section, the code will be validated by comparing some computed results with experimental values obtained from measurements or reported in the literature.

As a first example, a cylindrical dipole with no coating is considered. The antenna's input admittance is computed for a radius a of 0.007022λ as a function of normalized length, L/λ . In the calculations, no special treatment is given to the effect of the wire ends, which behave as flat caps with small radial currents going to zero at their centers. A certain correction is achieved, however, by taking the antenna to be $a/2$ longer at each end. The admittance results obtained this way are plotted in Figure 4.19 along with

Figure 4.19. Input admittance of a bare cylindrical dipole with radius $a=0.007022\lambda$.

reported experimental data. Since a delta-function generator is used in the program, for proper comparisons with the measured results it is important that a similar feeding configuration be employed experimentally. This is the case with the measured values in Figure 4.19, where the contributions to the antenna susceptance have been removed [93]. The agreement is seen to be very good, with less than 10% deviation between the two sets of data.

To verify the accuracy of the computations for the case of a coated wire antenna, a dipole of radius $a = 3.92$ mm, covered with magnetic material of outer radius $b = 5$ mm, is analyzed. The relative permeability of the coating is $\mu_r = 2.1$, while the dipole length is $L = 252.2$ mm. Computed and measured admittance results are compared in Figure 4.20, where the measured values are adapted from an equivalent monopole antenna (by dividing the admittance by two) [92]. Since no correction is made for the coaxial feedline of the experimental antenna, some discrepancy is expected. Nonetheless, Figure 4.20 shows excellent agreement up to the second resonance frequency of 0.8 GHz.

Figure 4.20. Input admittance of a 252.2 mm magnetically coated dipole with $a = 3.92$ mm, $b = 5$ mm and $\mu_r = 2.1$.

By far the most important test of the program, as it also validates the overall analytical approach, is the analysis of a slot antenna on a thin dielectric substrate. The antenna chosen for this example is a slot fed by a coplanar waveguide (CPW) feedline, as shown in Figure 4.21. The advantage of the CPW feed is that its dual/complementary structure is a two wire feedline, also with magnetic coating, which exhibits minimal discontinuity in the excitation zone of the equivalent dipole. Thus, the delta-function generator model can be expected to adequately represent the CPW feed mechanism in the calculated results. Referring to Figure 4.21, the slot is of length $L = 2H = 44.5$ mm and width $w = 3.81$ mm, realized on a Duroid substrate of thickness $h = 0.635$ mm and relative dielectric constant $\epsilon_r = 10.2$. The surrounding ground plane is sufficiently large in extent (200 x 200 mm) that no coupling effects exist between the slot and the outside edges which may affect the input impedance. Both the fabrication and measurement of this slot antenna were done as part of this work.

Figure 4.21. Slot on a dielectric substrate fed by a CPW feedline.

To perform the analysis, parameters μ_r , a and b of the equivalent dipole must first be determined. From section 4.3.1, these are obtained as $\mu_r = 10.2$, $a = w/4 = 0.9525$ mm and, with some calculation, $b = 0.160$ mm [92]. The computed input impedance results are shown in Figure 4.22, together with measured data corresponding to the impedance at the plane of the slot (i.e. de-embedded data). Very good agreement is

Figure 4.22. Input impedance of a slot on a dielectric substrate ($L=44.5$ mm, $w=3.81$ mm, $h=0.635$ mm, $\epsilon_r = 10.2$).

obtained over the entire 1.6 to 3.6 GHz range, where the experimental and theoretical curves are found to be within 10% of each other. The program can therefore be said to give reliable input impedance predictions for slot antennas on thin dielectric substrates. Other examples of the program's results for integrated circuit slot antennas on GaAs are given in Chapter 6.

Before proceeding further, the limitations of the method presented in this chapter should be briefly considered. Since the sequence of transformations on the original slot structure to an equivalent dipole is based on quasistatic theory, it is not possible to account for possible surface wave trapping in the substrate. Thus, the analytical method is limited to slots on thin substrates where surface waves are not present (described in section 4.2).

Limits on the range of slot dimensions (w , L) that can be treated are determined by the conditions under which the two-potential equation is valid. Thus, the slot width must be electrically thin, i.e. less than approximately 0.05λ , while the length has no upper limit provided the basis function polynomial is of sufficiently high order to

represent the current variation along each antenna segment. Finally, the principle limitation to the overall accuracy of the method stems from an approximation made in the transformation process of the slot antenna. Only an approximately complementary/dual structure was obtained from the original one-sided geometry. For Booker's principle to be strictly applicable, an exact complementary/dual structure is required, consisting of two dielectric layers placed symmetrically about the ground plane.

4.4 Conclusion

In this chapter, resonant-type antenna structures suitable for integration into monolithic circuits were reviewed, and some hybrid techniques for their performance enhancement were presented. Having found the slot to be the most efficient radiator for thin GaAs IC substrates, an analysis was undertaken to determine its input impedance. The method was based on the principle of complementary/dual structures whereby the slot is transformed into an equivalent magnetically coated dipole and the input impedances of the two antennas are related through Booker's rule. A solution of the current distribution along the dipole using the moment method was described and examples given to compare the input impedance results against experimental data. From the very good agreement that was obtained, the method is validated and hence an analytical tool to predict the input impedance of slot antennas on thin dielectric substrates has been successfully developed. This capability will be of use in the design of monolithic active antenna elements on GaAs, as described in Chapter 6.

CHAPTER 5. MEASUREMENT TECHNIQUES FOR INTEGRATED-CIRCUIT ANTENNAS

Physically small integrated-circuit (IC) antennas, operating at millimetre wave frequencies, present unique problems regarding the measurement of certain performance parameters such as input impedance and radiation patterns. Generally speaking, at frequencies above a few GHz, connections to the measuring equipment can introduce significant discontinuities, feedline losses tend to increase, and the small size and sometimes fragile nature of the antenna may render its manipulation difficult. This chapter describes two techniques and associated apparatuses developed specifically for the measurement of 20- 30 GHz IC slot antennas used in this work. The size of the antennas is on the order of 10 to 20 mm a side, with the GaAs substrate typically being of 100 μm thickness. Experimental results obtained using the newly developed setups are given in Chapter 6.

5.1 Input Impedance Measurements of IC Slot Antennas

In this section, a method is presented for measuring the input impedances of IC slot antennas fed by a microstripline or a CPW line. The CPW fed slot, shown in Figure 5.1(a), is realized on only one side of the substrate, while the microstrip configuration (Figure 5.1(b)) has the feedline and the slot on opposite sides of the substrate. In both cases, however, the input impedance Z_{in} is desired at the "plane of the slot" or, more

precisely, at the transverse reference plane which coincides with the first slot edge, as illustrated in Figure 5.1. This reference plane is chosen for its usefulness in subsequent integration of the slot with active circuit components.

Figure 5.1. Slot antennas fed by (a) a CPW line or (b) a microstrip line.

A conventional way of performing such measurements would be to mount the antenna in an appropriate test fixture having a connector attached to the feedline, measure the input impedance at the plane of the connector, and process the measured data analytically such that the de-embedded Z_{in} is obtained at the plane of the slot. These steps can be quite tedious, given the size and fragility of the device, and introduce additional uncertainties into the results.

Since many of the measurement problems described are those encountered in non-radiating MMIC measurements, it is appropriate to make use of equipment already developed for such circuits, i.e. a wafer prober measurement system. Shown in Figure 5.2 is the Cascade Microtech Model 54 probing station, which is used extensively for MMIC testing. The primary advantage of an on-wafer measurement facility is that no custom-made test fixtures and related connectors are required, as the wafer is mounted

on the chuck located at the center of the Figure, and the signals are fed to and from the device under test (DUT) by means of precision miniaturized probes. Furthermore, when properly combined with a network analyzer, any discontinuities resulting from the probe contact with the DUT can be completely compensated for during calibration.

Figure 5.2. Cascade Microtech probing station for MMIC measurements.

In order to obtain accurate slot antenna input impedance measurements, three issues must be addressed. First, the antenna must appear to be radiating into free space, such that repeatable input impedance values are obtained which do not depend on the antenna surroundings. Second, the finite sized ground plane in which the slot is cut must be large enough in extent to simulate the condition of a slot in an infinite ground plane, again for reasons of measurement repeatability. Finally, an appropriate calibration must be performed to set the measurement reference plane at the slot.

To meet the free space radiation condition in the presence of the wafer probing station's numerous metallic structures, a few modifications to the setup are required. Since the slot radiates both upward and downward, reflections from the four probe arms

above the wafer, and the metal base plate (chuck) beneath the wafer, can be of sufficient magnitude to affect the input impedance. The remaining metal plates and microscope attachment (not shown) above the wafer are more than a few wavelengths away so as to have negligible effect on the slot impedance [32]. From experiments performed at 4 GHz on a crude frequency scaled model of the probing station, it was found that the free space input impedance of slot antennas is obtained above 20 GHz when the three unused probe arms are placed at least 5 cm laterally from the slot, and the base plate setup shown in Figure 5.3 is employed.

Figure 5.3. Wafer probe base plate setup for slot impedance measurements.

A layer of dielectric RF absorbing material is inserted below the wafer such that the downward radiation is effectively dissipated before reaching the metal surface (the 1 cm thickness chosen provides 30 dB attenuation). To prevent dielectric loading of the antenna, an additional 1.3 cm styrofoam layer ($\epsilon_r \simeq 1$) is placed between the absorber and the wafer, as depicted in the figure. The two layers are machined as 10 cm diameter discs which exactly cover the metal base plate.

In order that the input impedance of a slot in a finite ground sheet be substantially the same as that in an infinite ground plane, the edge of the sheet must be one

wavelength or more away from the slot [94]. Thus, the IC slot antennas designed for 30 GHz operation in Chapter 6 are fabricated with at least a 10mm (1λ) ground plane surrounding each side of the aperture. This is a conservative guideline, however, as the dielectric substrate of the antenna reduces the wavelength considerably. It has been experimentally verified, by means of slot input impedance measurements on 90 μm thick GaAs substrates, that identical values in the 20 -30 GHz band are obtained with only 5 mm of surrounding ground plane.

Setting the measurement reference plane to the input of the slot is easily done using a network analyzer and the Through-Reflect-Line (TRL) calibration technique. This two port method requires a set of three transmission line standards appropriate to the length and type of the antenna's feedline, as illustrated in Figure 5.4. For a slot fed by a

Figure 5.4. TRL calibration standards for an antenna feedline of length F .

2 of/de 2

MICROCOPY RESOLUTION TEST CHART
NATIONAL BUREAU OF STANDARDS
STANDARD REFERENCE MATERIAL 1010a
(ANSI and ISO TEST CHART No. 2)

microstrip line of length F , the Through standard (Figure 5.4(a)) is a simple transmission line of length $2F$ connecting the two ports (in this case probe tips) of the analyzer. The Reflect standard consists of an open circuited transmission line of length F connected to each of the two ports of the analyzer, as depicted in Figure 5.4(b). Lastly, the Line standard (Figure 5.4(c)) is identical to the Through except that its length is $2F + D$, where D may be arbitrarily chosen in the range of 0.1 to 0.25 guided wavelengths at the highest measurement frequency [95]. The resulting reference plane location is at the desired distance F away from the probe tips, and the measurement may be performed using either one of the two ports of the analyzer. It should be noted that any probe-to-feedline discontinuities are automatically calibrated out with the TRL method.

5.2 Radiation Pattern Measurements of IC Slot Antennas

The measurement of IC slot antenna radiation patterns is now considered. Following the same reasoning as in the previous section, it is desired to make use of a wafer probing system to allow connector-less feeding of the antenna in the 20 - 30 GHz range. This approach, however, is complicated by the needs to mount the antenna vertically in an anechoic test chamber and that it radiate into free space without obstruction from the mounting fixture in at least one half-space.

In Figure 5.5 is photographed a portable wafer prober and mounting fixture which has been developed for the antenna pattern measurements. Essentially unimpeded radiation of the antenna occurs from the front face of the apparatus, shown in Figure

Figure 5.5. Portable mounting fixture for IC slot antenna pattern measurements.

(a) Front view.

Figure 5.5 (continued). Portable mounting fixture for IC slot antenna pattern measurements. (b) Back view.

5.5 (a), while the probes contacting the antenna feedline are located on the reverse side, shown in Figure 5.5 (b). Although only one probe is required to feed the input signal to a passive radiator, two probes are provided in this design to also allow for the DC biasing of an active antenna. With the use of threaded-screw positioner mechanisms, each probe can be moved along two axes to make contact with the antenna feedline.

A key element of the mounting fixture is the styrofoam disc (Figure 5.5(a)), 10 cm in diameter, used as the wafer carrier. Similarly to the setup for input impedance measurements, the styrofoam offers mechanical support while behaving practically as air in terms of microwave propagation. Thus, the wafer, attached first to the styrofoam carrier, is held vertically on the front of the mounting frame and accessed from behind by the probes.

To avoid creating a metal cavity effect on the feed side of the antenna, the complete assembly, except for the probe heads and the styrofoam wafer carrier, is made of low permittivity Delrin dielectric material ($\epsilon_r = 2.1$) rather than metal. This decreases reflections in general and reduces the possibility of forming a metallic enclosure with the clustered structures on the back side, which may alter the characteristics of the antenna.

Two small holding brackets at the periphery of the styrofoam disc are the only protrusions into the forward half-space of the antenna. Since these are fabricated of dielectric material and are located a few wavelengths away from the radiating aperture (placed at the center of the disc), their effect on the radiation pattern is expected to be minimal. The mounting fixture can therefore be used for measurements of IC slot antenna radiation patterns over a -90 to +90 degree scan.

CHAPTER 6. ACTIVE AMPLIFIER/ANTENNAS FOR GaAs MONOLITHIC PHASED ARRAYS

6.1 Introduction

In the preceeding chapters it was seen that the desired active antenna element for communication phased arrays is an amplifier/antenna, which can be realized efficiently in a GaAs MMIC format using slot radiators. To illustrate the system insertion of such an active element, the block diagram of a possible monolithic phased array is drawn in Figure 6.1. A single wafer is used to integrate the radiator, amplifier, phase shifter, variable attenuator and RF power combiner circuitry found in a receiving array (a

Figure 6.1. Monolithic receiving phased array.

similar diagram can be drawn for a transmitting array). The phased array operates, therefore, in either a transmit or receive mode only, such that diplexers separating the two channels and dual frequency radiating elements, among other complexities, are not required. Referring to the figure, the first stage of the receiver chain (or last stage of the transmitter chain) is an amplifier, to be integrated with the radiator so as to become an active amplifier/antenna.

In this chapter, a suitable phased array architecture for monolithic implementation is proposed and the precise form of its amplifier/antenna elements defined. The issues of MMIC device characteristics, slot radiators and antenna-device coupling which pertain to the design of active antenna elements, are then discussed. Finally, the practical MMIC implementation of a novel transmit-mode amplifier/antenna is described in detail, along with the results of its performance.

6.2 Proposed Monolithic Phased Array Architecture and Required Active Antenna Elements

The design of a phased array requires that many electrical, mechanical and architectural concerns be taken into account. In this section, the relevant aspects of these issues will be addressed to define an "optimum" monolithic phased array configuration and the precise form of the constituent active antenna elements.

In terms of electrical performance, the type of substrate, bandwidth, maximum scan range, type of polarization and spurious radiation are all issues to be considered [96].

The first two are already fixed, as the object of this work is to use a GaAs substrate for both the circuitry and the radiators, even though it is known that lower ϵ_r substrates are preferable for broad antenna bandwidths [69]. The scan range of a phased array is limited by the generation of grating lobes and "scan blindness", an effect which results from the array elements having unity reflection coefficients at wide scan angles [20]. Grating lobes are avoided by keeping the spacing of the elements less than $0.5\lambda_0$ apart, whereas scan blindness, a function of the element type, element spacing and substrate parameters, is generally unavoidable at some critical angle far off boresight. As mentioned earlier, linear polarization is chosen in this work for simplicity. Spurious feed- or active circuitry coupling with the antenna elements may degrade the radiation characteristics (sidelobe level, polarization, gain) and therefore must be minimized [20]. Also, the design should attempt to minimize the generation of surface waves which reduce efficiency.

Mechanical issues to consider are the number of elements, substrate area and heat transfer [96]. The number of elements has a bearing on the complexity of the feed and control circuitry, but will not be addressed further here due to its dependence on the specific application of the phased array. Sufficient substrate space must be available for the radiating elements, feed network, active circuitry (amplifiers, phase shifters, etc.) and bias lines, which must all fit in a $0.5\lambda_0 \times 0.5\lambda_0$ area. At 30 GHz, this amounts to 25 mm^2 , while at 20 GHz the available area becomes 56 mm^2 . Finally, the density of active devices employed in the array is sufficiently high that heat removal from the wafer is required. Since a single-wafer approach has already been chosen, heat removal depends mainly on factors external to the phased array.

The architecture of a phased array on a wafer falls under the class of either a single-sided geometry or a two-sided geometry [96]. The single-sided case is shown in Figure

6.2 for patch radiating elements. Since all the active circuitry and feed lines are crowded together on the same side as the patch, this design is susceptible to strong coupling between these components, which may also affect the radiation characteristics of the array. Scan blindness, however, is close to endfire for very thin substrates [28].

Figure 6.2. Single-sided phased array geometry.

A truly monolithic two-sided geometry does not appear to exist. However, the two-substrate aperture-coupled patch array shown in Figure 6.3 can still be used as an example to illustrate the benefits of this approach. As was seen in chapter 4, the radiation properties of the patch are enhanced by placing it on a low ϵ_r substrate. With the radiator removed from the front side, there is much more space available for the remaining circuitry. Also, the ground plane effectively isolates this circuitry from the radiating elements, resulting in reduced spurious coupling and radiation [96].

Figure 6.3. Two-substrate phased array configuration.

The monolithic phased array configuration proposed in this work is shown in Figure 6.4. Such a two-sided geometry is similar to what would be obtained by removing the second substrate and patches from Figure 6.3. Thus, the radiating elements are slots, although in this case they are approximately of resonant length, and are connected to the front side with through-substrate metallized via holes rather than electromagnetic coupling. For a purely monolithic design, this represents an optimum configuration in several ways: the benefits of double-sided construction are achieved (increased area for active circuitry, reduced spurious coupling), and slots have the highest radiation efficiency of all IC radiators, approaching 100 % under appropriate substrate conditions unlike patches or dipoles. Further, a very compact construction results from the novel use of direct via connections through the substrate, instead of electromagnetically coupled feedlines which may extend up to a quarter wavelength beyond the location of the slot.

Figure 6.4. Proposed double-sided monolithic phased array configuration using slot radiators.

It is now possible to focus on the nature of the amplifier/antenna elements used in the MMIC phased array. The basic structure consists of a transistor (FET) operating as an amplifier and connected with substrate vias to a slot in the underlying ground plane. Since the object is to feed the voltage across the slot to the FET terminals, the vias are located at opposite edges of the slot so as to connect the FET source to one edge and either the gate or drain to the other edge. To a first order approximation, each

metallized via hole in a 90 μm thick GaAs substrate behaves as a 25 pH inductor. The MMIC implementation for receiver operation is illustrated in Figure 6.5 along with its equivalent circuit representation. The transmitting case is identical except that the gate and drain bias pads are interchanged.

Figure 6.5. Integrated FET/slot antenna structure.

In both cases, DC biasing may be applied without affecting the RF performance or the geometry of the structure. For example, to bias the gate of the device negatively with respect to the source, a capacitor may be inserted between the gate and the via. This ensures that the gate is DC isolated from the vias which are all at ground potential. Appropriate DC voltages can then be applied through RF chokes to the gate and drain in the normal way.

It should be noted that to reduce area consumption, no matching circuitry is inserted between the device and the radiator. Low noise amplification is therefore achieved in the receive mode when the slot impedance presented to the FET is such that the minimum noise figure match is obtained. Similarly, in the transmit mode, maximum power is radiated from the slot when its input impedance is conjugately matched to the

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Figure 6.5. Integrated P-HEMT/slot antenna structure.

FET's optimum load impedance. In effect, the issue becomes one of achieving appropriate impedance matching for the two cases described.

Considering the active antenna structures of Chapter 3, it may be stated that along with the novelty of monolithically integrating a FET and slot radiator on GaAs, the proposed configuration contains a novel feed mechanism for the slot. Up to now, substrate via holes have been used for the sole purpose of grounding MMIC circuit elements (mainly FETs); this design, to be more fully described in Section 6.4, would be the first known attempt to use via holes for RF signal transmission to a circuit component on the opposite side of the wafer.

6.3 GaAs Monolithic Amplifier/Antenna Design Considerations

In order to successfully implement the amplifier/antenna in GaAs MMIC form, a number of important design issues must be considered. These are the MMIC device characteristics, the effect of via hole feeding on the slot's input impedance, and the coupling between the device and the radiator located in close proximity.

6.3.1 MMIC Device Characteristics

Until recently, the dominant device technology used in MMICs has been the GaAs MESFET. Today, High Electron Mobility Transistor (HEMT) and Heterojunction

Bipolar Transistor (HBT) heterostructure devices are also widely available at commercial foundries, increasing the number of choices the designer faces for realizing a specific MMIC. In this section, some of the basic characteristics of these three key devices are presented so as to allow selection of the most appropriate device for the proposed amplifier application. More complete information may be obtained in the referenced material.

The device structures of the GaAs MESFET, AlGaAs/GaAs HEMT and AlGaAs/GaAs HBT are drawn in Figure 6.6 [97]. Many variations exist of the basic structures shown in the figure. The MESFET is said to be a homojunction device as it consists of junctions of like materials (n^+ , n and undoped GaAs layers). In the n -doped channel where the drain to source current flow takes place, carrier mobility is limited due to coulomb scattering caused by collisions with ionized donors (impurities). The HEMT, on the other hand, consists of epitaxially grown layers of compound semiconductors having different bandgap energies, so that the interface between them forms a heterojunction [98]. Such a device is a technologically advanced FET with current carriers densely confined in a thin layer referred to as a two-dimensional electron gas (2-DEG). Electrons in this region, located between the undoped AlGaAs spacer and GaAs layers, undergo little scattering due to the absence of ionized donors, and hence exhibit a very high electron mobility. A further refinement to the structure results from growing an additional InGaAs layer "pseudomorphically" (due to a mismatch of lattice constants) onto the GaAs layer to increase the electron mobility in the channel to an even greater level. Pseudomorphic HEMTs, or P-HEMTs, are now the fastest three-terminal devices in the world [98].

The HBT derives its advantage over conventional bipolar devices from the use of different bandgap energy materials for the emitter and base regions of the device. This produces a high emitter injection efficiency of the carriers into the base, enabling the

Figure 6.6. MESFET, HEMT and HBT device structures.

HBT to have a very thin base and hence reduced base carrier transit time. Compositional grading of the AlGaAs base layer (by varying the mole fraction along the vertical direction between the collector and emitter) further reduces the transit time through the base [99].

In terms of high frequency operation, the three devices have unity gain cutoff frequencies (f_T) in excess of 100 GHz and may be used well into the millimetric frequency range. However, important distinctions emerge with respect to their suitability for specific applications.

Microwave and millimetre wave low noise amplification is best achieved with a MESFET or P-HEMT, as the HBT's noise performance rapidly deteriorates at these frequencies. From a physics based analysis [100], the minimum noise figure, F_{min} , can be written in terms of the device equivalent circuit parameters as

$$F_{\min} = 1 + 2\omega(C_{gs}/g_m)((R_s + R_g)/R_i)^{0.5}$$

Since the 2-DEG allows the $g_m R_i^{0.5}$ product to be higher for a HEMT than in a MESFET, the lowest noise will be obtained with a HEMT based amplifier [100]. State-of-the-art noise figures at 30 GHz for 0.15 μm gate length MESFET and P-HEMT devices are 1.8 dB and 0.7 dB, respectively [98]. At low frequencies, the HBT has lower 1/f (flicker) noise compared to FET devices, with a 1/f corner frequency 10 - 100 times smaller than for P-HEMTs or MESFETs [99]. This makes the HBT ideal for use in oscillator circuits for which the best phase noise performance is desired. Oscillator phase noise is dependent on the Q-factor of the resonator and the upconverted flicker noise of the active device. Frequency conversion circuits such as mixers and dividers are also optimally implemented with HBTs.

The power performance of a transistor can be characterized in terms of its power-added-efficiency (PAE) and its maximum output power. At X-band, the three devices currently have a comparable PAE of approximately 65%, while the HBT produces a four times larger output power (4W/mm of emitter length) than the two FET devices [101]. Although unproven at millimetric frequencies, the HBT is thought to hold promise as an efficient power device for this band due to the higher current densities per effective transistor area that can be achieved from its vertical structure. However, as the PAE comparison of the three devices in Figure 6.7 shows, the power HEMT presently provides the highest PAE above 20 GHz [97]. Also, the output power of the device is maintained at 1 W/mm of gate width up to 60 GHz. Such good performance is made possible by using a double heterojunction structure, where a heavily doped

Figure 6.7. Power-added-efficiencies of MESFET, HEMT and HBT devices.

AlGaAs layer is added beneath the InGaAs channel of the P-HEMT to allow high current density for increased power handling.

It has been shown that pseudomorphic devices built for optimum power handling also exhibit excellent noise performance, and are therefore suitable to be used in a wide range of millimetric applications. This is a markedly different characteristic from the MESFET, whose gate length and layer parameters can only be optimized for large signal capability or low noise, but not both together [100].

The question of device yield must still be addressed. Compared with the relative processing simplicity and maturity of the MESFET, and the relaxed fabrication dimensions of the HBT (only a 1 - 3 μm minimum lithographic dimension is required), reproducibility of the multi-layer HEMT is somewhat more difficult to achieve. However, heterostructure device technology has progressed rapidly in the past decade and with this trend expected to continue, new heights in P-HEMT performance and yield will be reached [98].

From the above discussion, the pseudomorphic HEMT emerges as the most appropriate device for implementing the critical low noise and power amplification functions in a monolithic phased array. This type of device has been chosen for the amplifier/antenna implementation described in Section 6.4.

To complete the description of MMIC devices, it should be noted that the device gate width is a variable in foundry processes. Specification of this parameter allows the designer a certain degree of control over the device characteristics. For example, variation of a P-HEMT's gate width from 75 μm to 300 μm produces the input and output reflection coefficient characteristics (S_{11} and S_{22}) plotted in Figure 6.8. Scattering parameters S_{12} and S_{21} also change with gate width, but are not shown in the figure. Referring to the small signal equivalent circuit of the device given in Figure 6.9, the effect of changing the gate width can be represented by appropriate scaling of the circuit element values. Neglecting the parasitic elements associated with the terminal pads and electrode cross-overs in parallel-integrated device structures, the equivalent circuit capacitor and transconductance values scale proportionally to gate width, while the resistor and inductor values are inversely proportional to the width of the gate.

Figure 6.8. 20 -30 GHz input and output reflection coefficient characteristics for 75 μm , 100 μm and 300 μm gate width P-HEMT devices.

Figure 6.9. Small signal equivalent circuit of a FET device.

Thus, use of the equivalent circuit model in a microwave CAD tool would allow the device characteristics to be analyzed over a range of gate widths based on information from a single device.

6.3.2. Slot Radiator Feeding Configurations and Antenna-Device Coupling

Although the amplifier/antenna configuration selected in Section 6.2 employs through-the-substrate feeding and places the device in close proximity to the radiator, quite accurate analysis of the structure is possible by considering the slot, feed and device components separately, as will be shown in this section.

In Chapter 4, the analytical computation of the IC slot's driving point impedance was seen to produce results corresponding to CPW feeding of the antenna. For a microstrip feed mechanism, shown in Figure 6.10, the slot input impedance is assumed

to differ little from the CPW case if the via hole inductance is small and the feedpoint location is close to the second slot edge. In this manner, the only significant difference in the geometry of the feed structure is that the center strip is no longer coplanar to the slot, but raised a distance of one substrate thickness above it. For a thin substrate, the strip to ground separation at the first slot edge is of the same order as the spacing that already exists between the CPW center strip and its adjacent ground (which becomes the first slot edge). Thus, under appropriate conditions the input impedances obtained by the two feeding configurations are expected to be essentially equal, and approximated by the driving point impedance.

Figure 6.10. Geometries of microstrip and CPW fed slots.

The above hypothesis has been verified experimentally and analytically using two identical slots fed by one or the other of the configurations in question. The slot antennas were 5.3×0.45 mm in size and realized on $90 \mu\text{m}$ GaAs substrates. First, the CPW fed slot was measured and compared with the calculated input impedance values from the computer program. The measurement conditions were the same as those described in Chapter 5, except for a shift of the reference plane to the probe tips.

Correction of the measured data was achieved by removing the presence of the 13 mm CPW feedline using a microwave CAD package. The resulting values, de-embedded to the first slot edge, are plotted in Figure 6.11 along with the computed curve. Agreement is very good, as expected from the validation process of Chapter 4, and the results confirm that the first slot edge is the correct reference point for comparing measured and computed values.

A similar procedure was then followed for the microstrip fed slot. In this case, the strip is connected through a 25 pH metallized via hole to a point 200 μm beyond the second slot edge, the closest contact to the edge that the fabrication process allows. The

Figure 6.11. Computed and measured input impedance for a CPW fed slot antenna.

inductive reactance introduced by the metallized hole is only $j4.7$ Ohms at 30 GHz, while the effect of feeding the slot $0.02\lambda_0$ away from the edge is negligible in terms of lateral voltage decrement, according to a rough slotline analysis [102]. Thus, the input impedance should be approximately the same as for the CPW configuration and the computed results. After de-embedding to the plane of the slot, the input impedance for

microstrip feeding is as shown in Figure 6.12, where the original calculated curve is also plotted. Again, good agreement is obtained, thereby demonstrating the near equivalence of the two antennas.

For the comparison made in Figure 6.12, no account was taken of the via hole inductance in the computed results. However, closer inspection of the two curves reveals a shift of a few degrees with respect to each other. It has been found that a simple series addition of the via hole reactance to the input impedance improves the accuracy of the calculated result. The new calculated curve obtained by adding the reactance of the 25 pH inductor in this manner, as well as the microstrip fed slot measurement, are plotted in Figure 6.13. This simple correction is seen to bring the two curves even closer together, and hence can be used as a conversion rule between the two feeding mechanisms.

Figure 6.12. Computed and measured input impedance for a microstrip fed slot antenna.

Figure 6.13. Computed and measured input impedance for a microstrip fed slot antenna. Via hole inductance correction applied to computed values.

It should be stressed that the addition of a lumped inductor in series with the antenna input impedance is not strictly correct in the physical sense. In this case the inductance is added externally to the antenna (at the first slot edge), while in reality it is situated inside the excitation zone. Since the computed input impedance is referenced to the first slot edge, there is no way (in the program's current state) of inserting the inductance at the second slot edge where it actually exists. However, the error is slight by modelling the inductance at the input to the antenna, as the difference in distance amounts to the width of the slot. This is confirmed by the results shown in Figure 6.11.

The remaining issue to be addressed in the amplifier/antenna design is the close proximity of the device and the radiator, which may lead to unaccounted for coupling. This effect must be considered in all active antenna realizations, as undesired coupling may alter the antenna's radiation pattern or affect the impedance matching condition of the two components.

The approaches that can be followed in the design range from full treatment of the coupling to assuming the device and radiator are uncoupled in parasitic terms. An analytical method which includes all coupling effects associated with an active antenna structure has been developed by Gillard et al. [103]. Based on an integral equation formulation for generalized loaded microstrip structures, a representation was found for the active device such that its lumped equivalent circuit could be incorporated into the equations. Although reasonably good results were obtained, it is not always necessary to resort to such complicated techniques.

The group which developed the mixer-dipole antenna integrated on a membrane (Figure 3.12), successfully accounted for the change in input impedance of the dipole due to the mixer chip placed at its feed, by analyzing the dipole with a dielectric disc loading the feed zone.

In many situations where the device is not placed directly in the excitation region of the antenna, it is sufficient from an impedance point of view to treat the two components as having no parasitic coupling. The design may then be performed as a conventional non-radiating microwave circuit, provided the antenna's radiation does not contain appreciable surface waves which hinder isolation between circuit components. This is the approach adopted in the majority of the active antennas described in Chapt. 3 [104,105], and will also be taken in the amplifier/antenna implementation presented in the next section.

6.4. A Novel MMIC Amplifier/Slot Antenna

Having outlined in the previous sections the specific configuration for the amplifier/antenna and the issues to be considered in its design, a practical implementation may now be undertaken. The remainder of this chapter describes the design, fabrication and results of a 30 GHz transmit-mode MMIC amplifier/slot antenna. A commercial foundry, Comsat Laboratories, offering a 0.25 μm pseudomorphic HEMT process on 90 μm thick GaAs wafers, was chosen for the MMIC fabrication.

6.4.1. Amplifier/Antenna Design

Since the object of this implementation is to demonstrate successful operation of a MMIC amplifier/antenna structure, somewhat relaxed specifications were set as follows. For the amplifier portion, the maximum possible gain is desired at 30 GHz, although input matching to 50 Ohms is not required, and no target is set for the bandwidth. In terms of antenna operation, a single lobe, linearly polarized pattern must be obtained in both principle planes, but no restriction is made that the overall length be less than $\lambda_0/2$ (as would typically be the case in a phased array).

The first step of the design is to choose an appropriate device-slot combination which provides conjugate matching of each other's impedance at 30 GHz. By considering a wide variety of slot antennas and pseudomorphic HEMTs [30], this was

found to be achievable with a $150\text{ }\mu\text{m}$ P-HEMT device and a $5.3 \times 0.45\text{ mm}$ microstrip fed slot (described in Section 6.3.2) placed $200\text{ }\mu\text{m}$ away from the device to ensure impedance isolation, as shown schematically in Figure 6.14. The device's drain impedance characteristics, obtained from an equivalent model, are plotted from 25 to 35 GHz in Figure 6.15 for the bias conditions indicated. At the 30 GHz design frequency, the

Figure 6.14. MMIC amplifier/slot antenna preliminary schematic.

Figure 6.15. Drain impedance characteristics of a $150\text{ }\mu\text{m}$ P-HEMT device.

reflection coefficient is $0.45\angle-85^\circ$. Based on the computed results obtained previously for the 5.3×0.45 mm slot input impedance, the addition of a $200\ \mu\text{m}$ microstrip transmission line (simulated on SuperCompact) produces the transformed slot impedance curve shown in Figure 6.16. Thus, at 30 GHz the impedance presented to the drain of the P-HEMT is $0.48\angle+88^\circ$, approximately the conjugate of the drain impedance.

Figure 6.16. MMIC microstrip fed slot impedance with $200\ \mu\text{m}$ feedline.

Before proceeding further, it should be verified that the dimensions chosen for the slot radiator lead to a single lobe radiation pattern. As mentioned in Section 4.2.1, a centre-fed slot must be less than approximately λ_s in length for single lobe radiation to occur, where λ_s is the wavelength in the slotline. From published slotline design curves [106, the wavelength in a 0.45 mm wide slot on $90\ \mu\text{m}$ GaAs is found to be 7.1 mm at 30 GHz. This makes the 5.3 mm slot $0.76\ \lambda_s$ in length, sufficiently short to generate a single lobe pattern.

The remainder of the design consists of adding the P-HEMT DC bias circuitry in a way which is essentially transparent to the RF behavior of the circuit. On the input (gate) side of the device, the required gate-to-source voltage is applied along with the

input RF signal by means of wafer probes, so no further bias circuitry is needed. Biasing of the the P-HEMT's drain, however, cannot be done through a second set of 50 Ohm probes connected directly to the device, as this would alter the device-radiator matching condition. Instead, the drain voltage is applied with the probes through a high-impedance 550 Ohm resistor, as shown in Figure 6.17. To prevent DC shorting of the bias current to ground through the via hole feeding the slot, a 1.7 pF blocking capacitor is also added to the circuit between the via and the resistor.

Figure 6.17. MMIC amplifier/antenna schematic showing biasing circuitry.

Simulation of the complete circuit using SuperCompact allows the performance of the amplifier/antenna to be predicted. With the additional biasing components, the impedance now seen by the drain is found to be $0.47/+84^\circ$, almost exactly the desired value. Also, a verification of the K parameter confirmed the stability of the amplifier below 30 GHz (where the device is not unconditionally stable). Of greatest interest are the power gain and bandwidth performance parameters. These quantities are more difficult to calculate for an antenna structure presenting an arbitrary load impedance to the amplifier, rather than the standard $50\ \Omega$. However, by computing the difference between the power delivered to the slot and the power input (from a $50\ \Omega$ source) to the gate of the P-HEMT, the gain of the amplifier is obtained. The results of this operation are shown in Figure 6.18. At 30 GHz, the amplifier is seen to have a maximum power

Figure 6.18. Predicted power gain response of the amplifier/antenna.

gain of 7.4 dB while its 3 dB bandwidth extends from 26.2 to 31.5 GHz. Considering that the maximum available gain of the P-HEMT is 10.0 dB at the design frequency and no matching of the input is provided, the simulated gain value could be expected. On the other hand, the 5.3 GHz bandwidth appears, at 18%, rather wide for a resonant-type antenna. Such performance is one of the possible benefits of active antennas alluded to in Chapter 3, whereby the operating bandwidth of an antenna may be enhanced when interfaced with the proper device impedance characteristic. In this design, no attempt was made to achieve a broad bandwidth, but the device and slot impedances combine in a manner favouring such a result.

6.4.2. MMIC Fabrication

The layout of the amplifier/antenna is shown in Figure 6.19. Along with the active antenna, a "passive" microstrip fed slot radiator and individual P-HEMTs were included on the wafer as test structures. Fabrication of the 3 inch MMIC wafer took place at Comsat Laboratories' foundry. Since reverse side processing for the realization of circuit components in the ground plane was done as a special request, the patterning of the slot radiators could not be guaranteed. A visual inspection of the fabricated wafers revealed that the slots were indeed poorly defined (improperly etched) and that post-processing was required. Using manual methods, the unwanted chrome and gold metallization layers partially filling the slot aperture were successfully removed for a few active and passive antennas.

Figure 6.19. Active and passive MMIC slot antenna layouts.

As part of the foundry process, the ground plane of the wafer had been broken between circuits along horizontal and vertical lines to facilitate dicing into chips. Since this results in the slot's surrounding ground plane being limited to approximately 1 mm, it was also necessary to protect the properly etched slots and sputter gold on the remainder of the surface to recreate a continuous ground plane. Approximately 1 μm of gold was deposited in this fashion on a quarter wafer, bringing the total ground plane thickness surrounding the slots to 3 μm everywhere except in the original dicing channels.

6.4.3. MMIC Amplifier/Antenna Results

Measurements were performed on the active and passive MMIC slot antennas using the equipment and techniques described in Chapter 5. A photograph of the quartered wafer mounted in the portable wafer probing fixture is shown in Figure 6.20.

Figure 6.20. MMIC amplifier/antenna wafer mounted in portable test fixture.

Before testing the antenna's radiation characteristics, the matching condition was verified by measuring the passive slot radiator's input impedance and the P-HEMT device drain impedance individually. The results of the measurement are plotted in Figure 6.21. It is seen that the device and slot impedances are nearly conjugates and agree very well with the predicted values ($0.47/+84^\circ$ calculated, $0.51/+90^\circ$ measured for the slot input impedance).

Figure 6.21. Measured input impedances of the passive slot and P-HEMT device.

Extensive pattern testing of the MMIC antennas was done in the anechoic chamber. In order to obtain patterns which are representative of the antennas in a complete wafer, the ground plane was extended with copper tape to completely cover the 4 inch styrofoam disc holding the wafer. A calibration of the measurement system was performed using transmit and receive standard gain horns. Since the transmit horn was replaced in subsequent measurements by the MMIC antennas, all absolute power levels are with respect to the 23.5 dB gain of the calibration horn.

Figure 6.22 shows the measured H-plane pattern of the passive slot antenna at 30 GHz. Although ripples are present due to the finiteness of the ground, the antenna radiates in a generally symmetrical, single lobe pattern, as desired. At boresight, the

pattern gain is seen to be -21.5 dB. When accounting for the slot's theoretical 1.5 dB gain, mismatch loss, and 0.5 dB connector loss, the gain is expected to be -21.3 dB. Thus, no significant power is being lost to surface wave radiation or in the structure's metal and dielectric materials.

Figure 6.22. Measured H-plane pattern of MMIC passive slot antenna.

The E-plane pattern is plotted in Figure 6.23. Although it exhibits the general shape expected from Chapter 4, the pattern lacks symmetry unlike the H-plane case. This may be explained by the fact that the slot antenna must be turned 90° to measure the E-plane cut. By doing so the asymmetry of the mounting fixture (the vertical stem for the

foot on one side only) is more visible to the radiation, since there no longer exists a pattern null in this direction as was the case for the H-Plane measurement.

Figure 6.23. Measured E-plane pattern of MMIC passive slot antenna.

H-plane pattern results of the active antenna are plotted in Figure 6.24 along with the previously measured passive slot. The pattern is virtually identical in form, but shifted upward by approximately 5 dB. Thus, the principle of the amplifier/antenna is validated, as the signal is successfully amplified before being radiated. The predicted 7.4 dB gain should not be confused with the measurement which shows only a 5 dB gain. The latter value is the gain obtained over a passive slot radiator, not the power gain of the amplifier. In order to compare the measurement with theory, the reflected powers must be taken into account. It can be shown that the active the active slot, with its 2.3 dB return loss, accepts 2.5 dB less power than the passive slot having an input return loss of 5.7 dB. By adding the 2.5 dB penalty to the measured gain, the predicted performance is obtained.

Figure 6.24. Measured H-plane pattern of MMIC active and passive slot antennas.

Several other pattern measurements were taken to complete the characterization of the MMIC amplifier/antenna. The E-plane pattern, shown in Figure 6.25, is simply an amplified version of the passive case (Figure 6.23). The cross-polarization performance is illustrated in Figure 6.26. This result indicates that a very good linear polarization of the antenna is obtained, as the crosspolar component is 15 dB below the copolar level. To determine the frequency dependence of the radiation pattern, measurements were performed at 27 GHz and 34 GHz. These are presented in Figure 6.27 along with the 30 GHz pattern. The curves no longer exhibit the symmetry that was obtained at 30 GHz. This is likely due to the fact that they were measured during a different session and the positioning of the antenna was altered, thereby leading to a skewed cut off the H-plane. Nonetheless, the antenna exhibits single lobe radiation across this frequency range, demonstrating its pattern bandwidth to be comparable to its impedance bandwidth. Finally, the behavior of the active antenna under different bias conditions is

given in Figure 6.28. Reduction of the amplifier current from the nominal 10 mA to 0 mA shows a 20 dB variation in radiated power with practically no change to the radiation pattern. Such a wide range of control shows the active antenna's potential for use in phased arrays and other adaptive environments.

Figure 6.25. Measured E-plane pattern of MMIC active slot antenna.

Figure 6.26. Measured cross-polarization of MMIC active slot antenna.

Figure 6.27. Measured H-plane pattern of MMIC active slot antenna at 27, 30 and 34 GHz.

Figure 6.28. Measured H-plane pattern of MMIC active slot antenna for various bias currents of the amplifier.

6.5 Conclusion

The topic of active amplifier/antennas for GaAs monolithic phased arrays has been thoroughly treated in this chapter, in which the three remaining thesis objectives have been met. In fulfillment of the first objective, a new two-sided monolithic phased array architecture was defined which offered potentially better performance than previously attempted one-sided geometries. This led to the definition of a new MMIC amplifier/antenna structure consisting of a FET device connected without matching circuitry to a slot radiator through metallized via holes. By considering the various issues involved in the design, a method for analyzing this new active antenna was presented, thus achieving the second objective of the thesis. In essence, from the calculated driving point impedance of the slot, a conversion rule is used to obtain the impedance of a slot fed by a microstrip line through the via hole. Then, with the device placed a suitable distance away from the slot, the combined amplifier/antenna may be analyzed from an impedance point of view by treating the two components separately. It was also seen that the optimum device for amplifier applications at millimetric frequencies is the P-HEMT and its impedance may be controlled by varying the gate width. Thus, the complete amplifier/antenna structure can be analyzed using the analytical tool developed in Chapter 4 and a commercially available microwave circuit CAD package.

A novel MMIC amplifier/slot implementation was then described. The steps in its design and fabrication were outlined, and measured results of its performance given. From the excellent agreement obtained between the predicted and measured performance, the analysis method is considered validated. The successful operation of the amplifier/antenna, which provides 7.4 dB gain and a single lobe radiation pattern at

30 GHz, fulfills the fourth thesis objective of experimental verification. The amplifier/antenna structure is the first MMIC to integrate a P-HEMT and a slot radiator on GaAs. In addition, this is the first time that circuit components have used both sides of an MMIC with via holes providing the through-substrate connections [107].

Considering the use of the amplifier/antenna in an MMIC phased array antenna as was done in Section 2.3, a serious thermal problem would arise in a geosynchronous satellite system application due to the large required radiated powers. However, the low earth orbit satellite link and the wireless in-building system have far more modest power requirements and hence are suitable applications for the developed active element.

CHAPTER 7. THESIS SUMMARY AND SUGGESTIONS FOR FUTURE WORK

This thesis has investigated the integration of antenna/amplifier elements for GaAs monolithic phased arrays, leading to a novel MMIC active slot antenna which has been successfully demonstrated at 30 GHz.

Starting from a broad perspective, the evolution of phased array systems was presented in Chapter 2. The basic building blocks for modern phased arrays was seen to be the T/R module, whose integration with the radiating elements becomes difficult at millimetric frequencies. Current approaches for achieving this integration were shown to fall into the hybrid brick or tile assembly techniques, or the fully monolithic construction technique. Finally, the application of digital and optical technologies to phased arrays was found to hold promising prospects for improving array performance or implementation. It was concluded that recent technological progress has brought about the possibility of achieving monolithic phased arrays for personal communications and other applications.

An overview of active antennas most relevant to the design of phased arrays was given in Chapter 3. Grouped according to their operation as transmitters, receivers or as both transmitters and receivers, the active antennas studied illustrated numerous techniques for integrating active devices with radiating elements to form oscillator/antennas, mixer/antennas, grid multi-function antennas and amplifier/antennas. Following a critical assessment of the various configurations and their associated performance, it was concluded that the amplifier/antenna is the desired active antenna for the phased array application considered in this work.

The results of Chapter 3 did not indicate which type of radiating element was appropriate for the amplifier/antenna's realization on high dielectric constant GaAs. This question was addressed in Chapter 4, where the characteristics of resonant-type patch, dipole and slot antennas were described, and hybrid techniques for their performance enhancement presented. With the slot emerging as the preferred radiator, an analysis was undertaken to determine the slot's input impedance as a function of its length, width and feedpoint location for subsequent integration with MMIC active devices. The technique of complementary/dual structures was chosen to represent the problem as an equivalent magnetically coated dipole, and a moment method solution was employed to obtain the input impedance of the slot numerically. Comparisons of the computed results with experimental data confirmed the accuracy of the calculations. Hence, an analytical tool was successfully developed to predict the input impedance of slot antennas on thin substrates.

In Chapter 5, measurement techniques for integrated circuit antennas operating at frequencies above 20 GHz were presented. Two techniques and associated apparatuses developed for the measurement of MMIC slot antenna input impedances and radiation patterns were described. The key to performing both these measurements was seen to reside in the use of an on-wafer probing system whereby specialized test fixtures and connectors are avoided. The newly developed measurement setups, used to obtain the experimental results in Chapter 6, fulfill the third thesis objective of devising suitable techniques for measuring integrated circuit antenna performance parameters.

The results of the previous chapters combined in Chapter 6, which dealt specifically with amplifier/antennas for GaAs monolithic phased arrays. The first thesis objective, that of defining a MMIC phased array architecture and determining the form of the amplifier/antenna, was met with the proposed double-sided geometry for a FET/slot

antenna. The new MMIC structure was seen to represent an optimum element for monolithic phased arrays in several performance-related respects. In fulfillment of the second thesis objective, an analysis method was developed for the amplifier/antenna by considering the MMIC device characteristics, slot radiator feeding configurations and antenna-device coupling. It was found that these three factors could be treated separately in the design under appropriate conditions. Finally, the new amplifier/antenna element was implemented in MMIC form. After outlining the steps in its design and fabrication, the results of its performance were given. The successful operation of the amplifier/antenna, which provided 7.4 dB gain at 30 GHz and a suitable single lobe radiation pattern, was seen to validate the analysis method and achieve the fourth thesis objective of experimental verification. This is the first time that a P-HEMT device and a slot radiator have been integrated on the same GaAs MMIC. It is also the first time that circuit components have used both sides of an MMIC with via holes providing the through-substrate connections. It was concluded that such an active element would be suitable for low earth orbit satellite and wireless in-building personal communications.

Although the goals of this study have been reached, the work pertaining to MMIC phased array development is by no means complete. The following suggestions are made for furthering the work initiated here.

The analytical tool developed in Chapter 4 for computing slot impedances should be combined with an optimization package such that antenna synthesis may be performed. Thus, upon specifying a desired input impedance, the software would determine the various slot dimensions and feedpoint locations which produce such a result. In this manner, shorter slot lengths can be obtained for the amplifier/antenna elements to allow their realization in a half wavelength grid.

In order to achieve the noise performance benefits of integrating a low noise amplifier closely to the receiving antenna, a receive-mode version of the amplifier/antenna should be developed which employs gate to slot (rather than drain to slot) coupling.

To further pursue the development of monolithic phased arrays, several enhancements to the basic amplifier/antenna should be made. These include increased miniaturization of the active antenna by placing the device in the feeding zone of the radiator, the achievement of effectively unidirectional radiation by mounting the wafer on a high dielectric constant lens, and the construction of a monolithic array of active elements. Each of the suggested improvements requires a more powerful method of analysis than the one provided in this work, in order to account for the resulting additional coupling effects. Thus, a new analytical tool should be developed, similar in principle to the integral equation approach taken by Gillard (Section 6.3.2) but with Green's functions appropriate to slot antennas rather than microstrip patch geometries. Such an analysis capability should allow the realization of high performance wafer scale monolithic phased arrays.

REFERENCES

1. Heilmeier, G.H., "Global Begins at Home," *IEEE Communications Magazine*, Vol. 30, N^o 10, pp. 50-56, October 1992.
2. Rosenberg, J., et al, "Multimedia Communications for Users," *IEEE Communications Magazine*, Vol. 30, N^o 5, pp. 20-24, May 1992.
3. Davis, J.H., et al, "Technologies for Global Communications," *IEEE Communications Magazine*, Vol. 39, N^o 10, pp. 35-43, October 1992.
4. Hayes, E.J., et al, "Personal Satellite Communications Utilizing the KA-Band," *Proceedings of the 1st International Conference on Universal Personal Communications*, Dallas, Texas, September 1992.
5. Raffaelli, L. and Stewart, E., "A Standard Monolithic Transmitter for 38 GHz PCN Applications," *Microwave Journal*, Vol. 35, N^o, pp. 24-30, October 1992.
6. Estabrook, P., et al, "A 20/30 GHz Personal Access Satellite System Design," *International Communications Conference*, Boston, Mass., June 1989.
7. Tsang, E.K. and Douville, R., "A Potential 21st Century Satellite Communications Application: Personal Communications," *Space Communications*, Vol. 7, N^o 3, pp. 133-138, August 1990.
8. Bhartia, P. and Bahl, I.J., "Millimeter Wave Engineering and Applications," John Wiley & Sons, New York, pp. 3-5, 1984.

9. Mailloux, R.J., "Antennas and Radar," *Microwave Journal*, Vol. 31, N° 3, pp. 28-33, March 1988.
10. Kinzel, J.A., et al, "V-Band Space-Based Phased Arrays," *Microwave Journal*, Vol. 30, N° 1, pp. 89-102, January 1987.
11. Colquhoun, A. and Schmidt, L.P., "MMICs for Automotive and Traffic Applications," *Proceedings of GaAs IC Symposium*, Miami, Fl., pp. 3-6, October 1992.
12. Raffaelli, L. and Stewart, E., "mm-Wave Monolithic Components for Automotive Applications," *Microwave Journal*, Vol. 35, N° 2, pp. 122-127, February 1992.
13. Zoghoul, A.I., et al, "Advances in Multibeam Communications Satellite Antennas," *Proceedings of IEEE*, Vol. 78, N° 7, pp. 1214-1232, July 1990.
14. Oliver, A., "Phased Array Antennas," Artech House, New York, 1972.
15. Tang, R. and Burns, R.W., "Array Technology," *Proceedings of the IEEE*, Vol. 80, N° 1, pp. 173-182, January 1992.
16. Mailloux, R.J., "Antenna Array Architecture," *Proceedings of the IEEE*, Vol. 80, N° 1, pp. 163-172, January 1992.
17. Balanis, A.B., *Antenna Theory: Analysis and Design*, Harper & Row, New York, pp. 204-206, 1982.

18. Lockerd, R.M. and Csain, G.E., "Airborne Active Element Array Radars Come of Age," *Microwave Journal*, Vol. 33, Nº 1, pp. 101-117, January 1990.
19. Psiolo, M., et al, "Multifunction Transmit and Receive Modules for EW Arrays," *Microwave Journal*, Vol. 34, Nº 2, pp. 83-92, February 1991.
20. Moynihan, R.L., "Phased Arrays for Airborne ECM," *Microwave Journal*, Vol. 30, Nº 1, pp. 34-53, January 1987.
21. Howell, J.M., "Phased Arrays for Microwave Landing Systems," *Microwave Journal*, Vol. 30, Nº 1, pp. 129-137, January 1987.
22. Mansky, L., "Broadband Phased Array Antennas," *Microwave Journal*, Vol. 27, Nº 9, pp. 159-169, September 1984.
23. Wisseman, W.R., "X-Band GaAs Single-Chip T/R Radar Module," *Microwave Journal*, Vol. 30, Nº 9, pp. 167-173, September 1987.
24. "Application of T/R Modules to Mission Specific Phased Array Radars," Technical Panel KTP-4, Subgroup K, The Technical Cooperation Program, May 1990.
25. James, J.R. and Hall, P.S., "Handbook of Microstrip Antennas," Vol. 1, Peter Peregrinus, London, pp. 693-751, 1989.
26. Campanella, S.J., et al, "Satellite Communications Systems and Technology, Circa 2000," *Proceedings of the IEEE*, Vol. 78, Nº 7, pp. 1039-1056, July 1990.

27. Schell, A.C., "Trends in Phased Array Development," *Microwave Journal*, Vol. 29, № 3, pp. 26-32, March 1986.
28. Pozar, D.M. and Schaubert, D.H., "Comparison of Architectures for Monolithic Phased Array Antennas," *Microwave Journal*, Vol. 29, № 3, pp. 93-104, March 1986.
29. Huang, Y., "Satellite Antennas," *Proceedings of the IEEE*, Vol. 80, № 1, pp. 183-193, January 1992.
30. Roy, L. et al, "Slot Antennas on GaAs for Integrated Array Applications," *Proceedings of ANTEM 92*, pp. 531-534, August 1992.
31. Bauhahn, P.E. and Geddes, J.J., "High-Density Circuit Approach for Low-Cost MMIC Circuits," *SPIE Proceedings*, Vol. 1475, pp. 122-128, April 1991.
32. Whicker, L.R. and Murphy, J.D., "RF-Wafer Scale Integration: A New Approach to Active Phased Arrays," *1992 IEEE Conference on Wafer Scale Integration*, pp. 291-299, January 1992.
33. Perkins, T.O., "Digital Beamforming Phased Arrays," *Microwave Journal*, Vol. 32, № 1, January 1989.
34. Richard, M.A. et al, "Optical Control of an 8-Element Ka-Band Phased Array using a High-Speed Optoelectronic Interconnect," *IEEE MTT-S Proceedings*, pp. 770-773, 1990.
35. Seeds, A.J., et al, "Optical Control of Microwave Semiconductor Devices,"

- IEEE Transactions on Microwave Theory Tech.*, Vol. MTT-38, pp. 577-585, May 1990.
36. Tang, R. et al, "Applications of Photonic Technology to Phased Array Antennas," *Proceedings of IEEE*, Vol. 78, N° 7, pp. 758-781, July 1990.
 37. Hartman, N.F. et al, "A New Time Time Delay Concept Using Integrated Optics Techniques," *IEEE AP-29*, pp. 1132-1135, 1991.
 38. Florence, J.M., "Monolithic Optical Time Shift Network for Phased Arrays," *IEEE MS-134*, pp. 782-785, 1990.
 39. Taylor, G.W. et al, "Electronic/Photonic Inversion Channel Technology for OEICs and Photonic Switching," *SPIE Proceedings*, Vol. 1476, pp. 2-13, April 1991.
 40. Swishun, S, et al, "Mixed Application MMIC Technologies: Progress in Combining RF, Digital and Photonic Circuits," *SPIE Proceedings*, Vol. 1475, pp. 223-230, April 1991.
 41. York, R.A. et al, "Active Patch Antenna Element for Array Applications," *Electronics Letters*, Vol. 26, N° 7, pp. 494-495, March 1990.
 42. Chang, K. et al, "Active Radiating Element Using FET Source Integration with Microstrip Patch Antenna," *Electronics Letters*, Vol. 24, N° 21, pp. 1347-1348, October 1988.
 43. Navarro, J.A., et al, "Active Integrated Antenna Elements," *Microwave Journal*,

Vol. 34, № 1, pp. 115-126, January 1991.

44. Perkins, T.O., "Active Microstrip Circular Patch Antenna," *Microwave Journal*, Vol. 30, № 3, pp. 109-117, March 1987.
45. Wu, X.D. et al, "Novel FET Active Patch Antenna," *Electronics Letters*, Vol. 28, № 20, pp. 1853-1854, September 1992.
46. Shen, Y. et al, "Active Radiating Oscillator Using a Reflection Amplifier Module," *Electronics Letters*, Vol. 28, № 11, pp. 988-991, May 1992.
47. Camilleri, N. et al, "Monolithic Millimeter Wave IMPATT Oscillator and Active Antenna," *IEEE MTT-36*, pp. 1670-1676, December 1988.
48. Kawasaki, S. and Itoh, T., "A Layered Negative Resistance Amplifier and Oscillator Using a Fet and Slot Antenna," *IEEE MTT-S*, pp. 1261-1264, 1991.
49. Kawasaki, S. and Itoh, T., "Electronically and Optically Controlled Active Integrated Antenna," *Proceedings of ISAP-92*, pp. 821-824.
50. Hwang, V.D. et al, "Quasi-Optical Integrated Antenna and Receiver Front End," *IEEE MTT-36*, pp. 80-85, January 1988.
51. Hwang, V.D and Itoh, T., "Quasi-Optical HEMT and MESFET Self-Oscillating Mixers," *IEEE MTT-36*, pp. 1701-1705, December 1988.
52. Chew, W. et al, "Printed Circuit Antennas with Integrated FET Detectors for Millimeter-Wave Quasi Optics," *IEEE MTT-37*, pp. 593-597, March 1989.

53. Nesic, A., "Integrated Uniplanar Antenna Mixer and Local Oscillator," *Electronics Letters*, Vol. 27, N° 1, pp. 56-58, January 1991.
54. Birkeland, J. and Itoh, T., "Planar FET Oscillators Using Periodic Microstrip Patch Antennas," *IEEE MTT-37*, pp. 1232-1236, August 1989.
55. Birkeland, J. and Itoh, T., "Quasi-Optical Planar FET Transceiver Modules," *1989 IEEE MTT-S Digest*, pp. 119-122.
56. Reberg, G. et al, "Integrated Horn Antennas for Millimeter-Wave Applications," *Ann. Télécommun.*, 47, N° 1-2, 1992, pp. 38-48.
57. McIlvenna, J.F., "Monolithic Phased Arrays for EHF Communications Terminals," *Microwave Journal*, Vol. 31, N° 3, pp. 113-125, March 1988.
58. Yen, H.C. et al. "A Monolithic Approach for Q-Band Integrated Active Phased Array Transmitting Antenna," *Proceedings of IEEE*, May 1992, pp. 126-129.
59. Uhmme, M. et al, "An MMIC Aperture-Coupled Microstrip Antenna in the 40 GHz Band," *Proceedings of ISAP 92*, pp. 1105-1108.
60. Hacker, J.B. et al, "A 100-Element Planar Schottky Diode Grid Mixer," *IEEE Transactions on Microwave Theory and Techniques*, pp. 1-10, June 1991.
61. Sovero. E.A. et al, "A Monolithic 35 GHz HBT Quasi-Optical Grid Oscillator," *Proceedings of IEEE GaAs Ic Symposium*, pp. 305-308, 1992.

62. Rutledge, D. et al, "Oscillator and Amplifier Grids," *IEEE MTT-S Digest*, pp. 815-818, September 1992.
63. Parfitt, A.J. and Griffin, D.W., "Monolithic Microwave and Millimeter Wave Integrated Circuit Antennas," *IREECON International Convention*, Melbourne, 1989. pp. 169-172.
64. Baca, L.J. and McInerney, J.G., "Design of a Broadband High-Frequency Antenna on Gallium Arsenide Using the Effective Index Method," *Electronics Letters*, Vol. 26, N° 14, pp. 1083-1085, July 1990.
65. Chow, P.D. et al, "InGaAs HEMT MMIC LNA and Doublers for EHF SATCOM Ground Terminals," *SPIE*, Vol. 1475, pp. 42-44, 1991.
66. Zaghloul, A.I. and Mailloux, R.J., "MMIC Phased Arrays," in AGARD Lecture Series N° 151,
67. Shiau, Y., "Dielectric Rod Antennas for Millimeterwave Integrated Circuits," *IEEE MTT-24*, pp. 869-872, November 1976.
68. Bhartia, P. and Bahl, I.J., *Millimeter Wave Engineering and Applications*, John Wiley & Sons, New York, N.Y., pp. 477-607, 1984.
69. Bahl, I.J. and Bhartia, P., *Microstrip Antennas*, Artech House, Dedham, Mass., pp. 1-80, 1980.
70. Pozar, D.M., "Microstrip Antennas," *IEEE Proceedings*, Vol. 80, N° 1, pp. 79-91, January 1992.

71. Parfitt, A.J. and Griffin, D.W., "The Single Wire Fed Dipole on Slab: An Antenna for Integration with MMICs," 1990 IEEE AP-S Symposium, pp. 795-798.
72. Oltman, H.G. et al., "Electromagnetically Coupled Microstrip Dipoles," *IEEE AP-29*, pp. 151-157, 1981.
73. Alexopoulos, N.G. et al., "Substrate Optimization for Integrated Circuit Antennas," *IEEE MTT-31*, pp. 550-557, July 1983.
74. Rutledge, D.B. et al., "Integrated Circuit Antennas," in *Infrared and Millimeter Waves*, Vol. 10, Academic Press, pp. 1-87, 1983.
75. Jordan, C.E. and Balmain, K.G., *Electromagnetic Waves and Radiating Systems*, Prentice-Hall, N.J., pp. 513-522, 1968.
76. Ogawa, H. et al., "Optical/Microwave Circuit Technologies for Microwave and Millimetre-Wave Fiber Optic Links," *Proceedings of ISAP 92*, Sapporo, Japan pp. 825-828.
77. Wesic, A. and Popovic, B.D., "Analysis of Slot Antenna on Dielectric Substrate," *IEEE Proceedings*, Vol. 132, Pt. H, N° 7, pp. 474-476, December 1985.
78. Yoshimura, Y., "A Microstripline Slot Antenna," *IEEE MTT-20*, pp. 760-762, November 1972.
79. Alexopoulos, N.G. and Jackson, D.R., "Fundamental Superstate (Cover)

Effects on Printed Circuit Antennas," *IEEE AP-32*, pp. 807-816, August 1984.

80. Rogers, R.L. and Neikirk, D.P., "Radiation Properties of Slot and Dipole Elements on Layered Substrates," *International Journal of Infrared and Millimeter Waves*, Vol. 10, N° 6, pp. 697-728, 1989.
81. Hirokawa, J. et al, "Cavity-Backed Wide Slot Antenna," *IEEE Proceedings*, Vol. 36, pt. H, N° 1, pp. 29-33, February 1989.
82. Tirkas, P.A., Balais, C.A. "FPTD Method for Antenna Radiation," *IEEE Transactions on A & P*, Vol. 40, N° 3, pp. 334-340, March 1992.
83. Balais, C.A. "Antenna Theory: A Review," *Proceedings of the IEEE*, Vol. 80, N° 1, pp. 7-22, January 1992.
84. Miller, E.K., Burke, "Low-Frequency Computational Electromagnetics for Antenna Analysis," *Proceedings of the IEEE*, Vol. 80, N° 1, pp. 24-42, January 1992.
85. Hadidi and Hamid, "Analysis of Slot Antennas in Infinite Ground Planes," *IEEE Transactions on Antennas and Propagation*, pp. 1433-1438, January 1989.
86. Kominami, M. et al, "Dipole and Slot Elements and Arrays on Semi-Infinite Substrates," *IEEE Transactions on Antennas & Propagation*, pp. 600-607, June 1985.
87. Pozar, D.M., "A Reciprocity Method of Analysis for Printed Slot and Slot-Coupled Microstrip Antennas," *IEEE Transactions on Antennas and*

Propagation, Vol. AP-34, N° 12, pp. 1139-1145, December 1986.

88. Das, B.N., "Impedance of a Radiating Slot in the Ground Plane of a Microstripline," *IEEE Transactions on Antennas and Propagation*, Vol. AP-30, N° 5, pp. 922-926, September 1992.
89. Rogers, R.L. et al, "Impedance Measurements and Calculations of a Microstrip-Fed Slot Antenna on a Multilayered Electrically Thick Dielectric Substrate," *International Journal of Infrared and Millimeter Waves*, Vol. 10, N° 6, pp. 687-729, 1989.
90. Singh, S. and Wilton, D.R., "Analysis of an Infinite Periodic Array of Slot Radiators with Dielectric Loading," *IEEE Transactions on Antennas and Propagation*, Vol. 39, N° 2, pp. 190-196, February 1991.
91. Nesic, A. and Popovic, B.D., "Analysis of Slot Antenna Dielectric Substrate," *IEE Proceedings*, Vol. 132, Pt. H, N° 7, pp. 474-476, December 1985.
92. Miller, E.K. and Burke, G.J., "Low-Frequency Computational Electromagnetics for Antenna Analysis," *IEEE Proceedings*, Vol. 80, N° 1, pp. 24-42, January 1992.
93. Popovic, B.D. et al, "Analysis and Synthesis of Wire Antennas," *Research Studies Press*, John Wiley & Sons Ltd., U.K., 1982.
94. Krams, J.D., "Antennas," *Electrical and Electronics Engineering Series*, McGraw-Hill, p. 371, 1950.

95. Hewlett-Packard, Appr. Notes on IRL.
96. James, J.R. and Hall, P.S., "Handbook of Microstrip Antennas," Vol. 1, Peter Peregrinus, London, pp. 693-751, 1989.
97. Tsang, H.Q. et al, "Advances in Power MMIC Amplifier Technology in Space Communication," *SPIE*, Vol. 1475, pp. 74-88, 1991.
98. Nguyen, T et al, "Ultra-High-Speed Modulation-Doped Field-Effect Transistors: A Tutorial Review," *Proceedings of the IEEE*, Vol. 80, N° 4, pp. 492-518, April 1992.
99. Wong, K., "HEMTs and HBTs," McGraw Hill, First Edition, New York, 1991.
100. Ladbrooke, P.H., "MMIC Design: GaAs FETs and HEMTs," Artech House, Boston, 1989.
101. Itoh, H. et al, "Current Gain Enhancement in Graded Base AlGaAs/GaAs HBTs Associated with Electron Drift Motion," *Japanese Journal of Applied Physics*, 1985.
102. Pozar, D.M., "Microstrip Antennas," *IEEE Proceedings*, Vol. 80, N° 1, pp. 79-91, January 1992.
103. Gillard, R. et al, "Une Approche Générale du Concept d'Antenne Active pour les Antennes et les Réseaux Microreban à Partir de la Théorie des Diffuseurs Chargés," *Ann. Télécommun.*, 46, N° 5-6, pp. 251-257, Mai-Juin 1991.

104. Hall, P.S., "Analysis of Radiation from Active Microstrip Patch Antennas," *Journ. Int'l'es de Nice sur les Antennes*, France, pp. 446-449, November 1990.
105. Fusco, U.F., "Series Feedback Integrated Active Microstrip Antennas Synthesis and Characterisation," *Electronics Letters*, Vol. 28, N° 1, pp. 89-91, January 1992.
106. Janaswamy, R. et al, ""Dispersion Characteristics for Wide Slotlines on Low-Permittivity Substrates," *IEEE Transactions on Microwave Theory and Techniques*, Vol. MTT-33, N° 8, pp. 723-726, August 1985.
107. Roy, L. et al, "An Integrated HEMT/Slot Active Antenna Structure Using MMIC Technology," *Proceedings of European Microwave Conference*, pp. 241-244, August 1992.

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