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**AN INVESTIGATION OF TRANSCEIVER DESIGN
FOR HIGH BANDWIDTH, INDOOR WIRELESS
LOCAL AREA NETWORKS OPERATING
IN THE 20 - 60 GHz RANGE**

by

Erik Boch, BSEE

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

Master of Engineering

Department of Electronics

Carleton University
Ottawa, Ontario
Canada

28 January 1995

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the degree of Master of Engineering

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28 January 1995

Abstract

A variety of research thrusts into various aspects of millimetre-wave indoor wireless local area network (WLAN) systems have been recently undertaken. In order to better understand the general system implementation problem, some of these thrusts have been directed towards networking, propagation issues, error detection/correction, modulation/demodulation, and antenna/array techniques.

The aim of the work outlined in this thesis is to combine these elements with practical short/medium term MMIC transceiver implementation trade-offs. The resulting trade-off analysis provides insight into the current limitations in system functionality caused by transceiver hardware limitations. In order to fully understand implementation issues, a prototype MMIC-based transceiver operating in the 38 GHz band has been designed, built, and tested. Aspects of the transceiver performance are compared against predictions and shortcomings are investigated.

Acknowledgements

The author wishes to thank Professors Robert Harrison and David Falconer for their support and guidance throughout this thesis effort.

The author also wishes to thank the Canadian Institute of Telecommunication Research (CITR), in particular the Millimetre-Wave WLAN Project Group (under Prof. D. Falconer) for numerous technical inputs and ideas.

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1.0 INTRODUCTION AND OVERVIEW

This thesis documents an investigation of the feasibility of a radio transceiver for broad-band, indoor wireless local area network (WLAN) applications in the 20 - 60 GHz range.

The area of millimetre-wave (20 - 60 GHz) in wireless local area network (WLANs) applications is the subject of a Canadian research thrust led by Professor David Falconer at Carleton University. Under sponsorship of the Canadian Institute of Telecommunications Research (CITR), Professor Falconer leads a multi-university research team which is focusing on the feasibility of millimetre-wave technology to the general application of low-cost, broad-band radio for networking services and data in the indoor microcell environment. The project currently addresses the following key research areas:

- Measurement and characterization of indoor wireless channels,
- Microcellular architectures in wireless indoor communications,
- Advanced transceiver structures,
- Spread spectrum networks, and
- Infrared techniques.

Against the backdrop of this major research project, some fundamental understanding of the practical transceiver implementations and the accompanying limitations on systems design were deemed to be of value. A small project focused on the feasibility of the implementation of a millimetre-wave transceiver using techniques/technologies which could potentially be low cost in nature was formed outside of the major CITR program. The value of this effort lies in the need to accommodate constraints dictated by the transceiver implementations within other aspects of research being undertaken at the system level. This transceiver investigation forms the basis for the thesis effort documented herein.

The thesis begins with a statement of goals (Chapter 2) and a statement of general requirements for the transceiver (Chapter 3). These requirements provide the general direction for the remaining body of work. Following this, a literature survey is conducted which is intended to benchmark the current state of technology (Chapter 4). The broad nature of this

thesis requires a systems understanding of a large range of technical issues and these are investigated to a level suitable for the extraction of key information. The architecture of the transceiver is then addressed and first-order estimations of performance attributes are derived (Chapter 5). Using the top-level architecture, a detailed investigation of the design and attributes of the transceiver is undertaken which concludes with a comparison of expected performance versus requirements (Chapter 6). Using the detailed system architecture, a proof-of-concept is undertaken followed by a performance measurement and evaluation exercise (Chapter 7). The final chapter of the thesis addresses the summary of achievements and sets the direction for possible follow-on research efforts (Chapter 8).

2.0 STATEMENT OF THESIS GOALS

The use of wireless indoor communications is beginning to see an increased interest as a result of the need to transport more and more information in the "information office". Office buildings of the future will be constructed with the notion of employing the wireless local area network (WLAN) in each region of the structure.

The idea of using the WLAN to implement parts of the network is based on the desire not to have to rework the entire building installation when the LAN changes in architecture or is physically moved.

The key problems faced by this change in the work place is the need to handle the data rates required for Synchronous Optical Network (SONET) and/or High Performance Parallel Interface (HIPPI) data transfer between users and microcells within the office environment. Generally this dictates data rates of several hundred Mbit/s (with an absolute minimum of 25 Mbit/s) on any one user.

The use of the spectrum allocations for this type of communications has been limited to frequency ranges in which the achievement of this data rate is not feasible and the amount of interference present (in the form of other users) would make low-distortion communications nearly impossible.

The use of portions of the millimetre-wave spectrum beyond 18 GHz therefore provides a suitable location in which to design and implement a system of this nature.

The key goal of this thesis is to investigate and design a proof-of-concept millimetre-wave transceiver which operates above 18 GHz and is able to transmit and receive at these data rates with acceptable distortion.

As part of the overall millimetre-wave WLAN research program, this thesis will be among the first attempts to design and build transceivers for this application.

3.0 TOP-LEVEL REQUIREMENTS

In order to address the general goals of the proposed thesis, the top-level requirements for the WLAN transceivers are proposed and will serve as guide points to the subsequent detailed thesis effort.

Essentially, the transceivers will be configured in an inter-office LAN as illustrated in Figure 3.1 [1,2]. The master (base) station will be configured as an omni-directional unit which is ceiling mounted to allow for obstruction-free transmission/reception at the user terminal transceivers.

The choice of operating frequency is based on a number of factors. The bandwidth necessary for high-speed data transmission and the accommodation of frequency-reuse schemes dictates a higher operating frequency. Moving towards V (50-75 GHz) or W (75-110 GHz) band does not appear to be advantageous since the application does not require electrically large aperture antennas and would not benefit from added propagation loss factors (for example at 60 GHz) due to the ranges involved. Operating anywhere above 18 GHz provides a benefit in terms of almost non-existent man-made interference when compared to the low microwave region. In addition, current and near-term foundry capabilities appear to be limited to 40 GHz and are not really market driven to higher operating frequencies. Therefore the availability of components and the bandwidth needed lead to a selection of a high Ka-band (~ 38 GHz) operating frequency. This band is also used for other experimental communications systems and is therefore more likely to be approved for this application than the 34-37 GHz range which is allocated for radar use [for example].

The user terminal transceivers will be mounted in elevated positions above portable office walls and the general level of personnel traffic moving about the office space. The user terminals will be aligned to the central master on installation and will subsequently remain generally stationary. The user terminals will employ high gain antennas in order to reduce multipath effects and to allow a robust measure for the reduction of intersymbol interference (ISI) [3,4]. Circular polarization is employed to accommodate the effect of multi-path induced depolarization. A diagram of the mounting of the user and master terminals is shown in Figure 3.2.

FIGURE 3.1 INTRA-OFFICE WLAN CONCEPT [1]

FIGURE 3.2 MOUNTING OF MASTER AND SLAVE TRANSCEIVERS

The master and user terminals will be employed in a time-division multiplexed access (TDMA) scheme. A TDMA scheme which might be implemented is shown in Figure 3.3 [5]. At the beginning of the cycle, the master transmits a synchronization scheme which aligns the various users and then bursts its data frame addressed to a specific user. Following this, the individual users burst returning data to the master in their allocated time slots. The microcell user terminals therefore must retain timing synchronization for the duration of the frame cycle. On power up, the user terminals will operate in a listen-only mode until the synchronization pattern is detected and the clock is recovered from it.

The targeted data rate of the WLAN is 150 Mbit/s with a minimum acceptable rate of 25 Mbit/s (within a microcell). Investigations into the complexity and cost of operating at the various data rates will be carried out to determine a cost-optimal solution. Preliminary investigations show that multipath-induced ISI will dictate specific modulation schemes (such as M-level FSK or M-level frequency-hopped OOK for example).

The operation of the transceivers (master and users) must allow for robust operation in the presence of ISI and must be designed to achieve a usable bit-error-rate (BER) on each link. In general the separation of the users and the masters is 30 m or less and is a clear line-of-sight (LOS) path. Preliminary investigations indicate that propagation losses through building/construction materials will generally prevent useful operation of the link in non-LOS applications. These losses actually will provide suitable isolations which will allow for the minimization of the complexity of frequency re-use schemes.

In order to allow WLANs of similar design to operate synergistically within a given office environment, it is necessary to be able to select the operating frequency of a particular WLAN such that it does not operate at the same centre frequency as any of its neighbouring cells. Generally this will require the ability to select any one of a small number of centre frequencies provided that the WLAN cells are oriented as shown in Figure 3.4. The operating centre frequency of any cell will be selected on set-up and all transceiver connected on this WLAN will operate on this selected frequency.

FIGURE 3.3 WLAN MICROCELL TDMA SCHEME [5]

FIGURE 3.4 MICROCELLS OPERATING WITHIN AN OFFICE ENVIRONMENT

In summary, the preliminary performance requirements for the transceiver investigations which will form the focal point of the thesis are described below in Tables 3.1 A, B. An additional qualitative attribute of the transceiver is the use of a common design for the circuitry (less the antennas) in the user and master transceivers. This minimizes the current prototyping efforts as well as subsequent efforts to package the design as a minimized chip set.

TABLE 3.1 A
WLAN USER TERMINAL TRANSCEIVER REQUIREMENTS

FACTOR	UNITS	REQUIREMENT	NOTES
1. Operating frequency	GHz	~38	
2. Selectable operating frequencies	--	5	
3. Transmitter power	mW	>20	
4. Receiver noise figure	dB	<12	
5. Transmitter/receiver antenna gain	dBi	>20	
6. Polarization	--	circular	
7. Operating range	m	50 target, 30 min.	Calculated assuming LOS indoor environment.
8. Minimum operating SNR	dB	20 target, 10 min.	
9. Data rate	Mbit/s	155 target, 25 min.	Data bandwidth
10. Modulation scheme	--	TBD	Robust operation in multipath.
11. Power consumption	w	minimize	
12. Physical size	--	minimize	
13. Technological implementation	--	MMIC, SMT (see NOTE 1)	Readily adapted to lowest cost realization.
14. Long-term cost target	\$	200	

NOTE 1: MMIC = Monolithic Microwave Integrated Circuit
SMT = Surface Mount Technology

TABLE 3.1 B

WLAN MASTER TERMINAL TRANSCEIVER REQUIREMENTS

FACTOR	UNITS	REQUIREMENT	NOTES
1. Operating frequency	GHz	~38	
2. Selectable operating frequencies	--	5	
3. Transmitter power	mW	>20	
4. Receiver noise figure	dB	<12	
5. Transmitter/receiver antenna gain	dBi	-0	
6. Polarization	--	circular	
7. Operating range	m	50 target, 30 min.	Calculated assuming LOS indoor environment.
8. Minimum operating SNR	dB	20 target, 10 min.	
9. Data rate	Mbit/s	155 target, 25 min.	
10. Modulation scheme	--	TBD	Robust operation in multipath.
11. Power consumption	w	minimize	
12. Physical size	--	minimize	
13. Technological implementation	--	MMIC, SMT	Readily adapted to lowest cost realization.
14. Long-term cost target	\$	1,000	

3.1 References for Chapter 3

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- [4] T. Rappaport, 'Wireless Personal Communications: Trends and Challenges', IEEE Antennas and Propagation Magazine, Vol. 23, No. 5, Oct. 1991, pgs. 19 - 29.
- [5] Feasibility of Covert Millimetre-Wave Communications - Final Report, DSS contract 21ST.36001-4-2659.

4.0 LITERATURE REVIEW

This section of the thesis outlines the findings of a literature review conducted to determine the background technical data, theoretical and practical approaches which have been employed by other practitioners in the field of millimetre-wave microcell communications technology and other related fields of study.

This literature study is not intended to address all aspects of applicable work in detail, but rather to establish general boundaries and limitations which were observed during the work carried out in this thesis. Literature surveys were conducted in the following areas:

- Terrestrial millimetre-wave links,
- Indoor millimetre-wave links,
- Anti-multipath techniques applicable to the indoor propagation environment, and
- Practical millimetre-wave communication systems.

4.1 Terrestrial Millimetre-Wave Links

In the past the area of terrestrial communications, primarily for covert military voice and data communications, was the key development area using the millimetre-wave portion of the spectrum (above 30 GHz).

Driven by the need for short-range secure communications links, a variety of propagation investigations have been undertaken [1-4] to better understand the propagation channel and propagation through various (primarily natural outdoor) mediums.

The portion of the spectrum being exploited for this application lies in the 60 GHz range. As shown in Figure 4.1, in this portion of the spectrum, there is a high attenuation factor, on the order of 15 dB/km at sea level. Several advantages result from this absorption phenomenon:

- Minimization of the "overshoot" of point-to-point communications links,
- Reduced chance of ground-based interception of air-borne or space-borne signals,
- Technical infeasibility of building suitable intercept receivers,
- Frequency re-use within networks.

FIGURE 4.1 GENERAL SEA LEVEL PROPAGATION CHARACTERISTICS [1]

In addition, operation at these frequencies offers the potential for physically small but electrically large antennas providing a further potential advantage [for covert communications] in terms spatial selectivity.

Due to the covert nature of the links involved, and the limited amount of effective radiated power (ERP) which can readily be generated, the link ranges being considered are generally in the 1 to 10 km range. Within this range a variety of investigations have been carried out including fade margin requirements [3,5], bit error rates (BERs) [1], attenuation through airborne particulants (smoke, dust) and rain attenuation.

As shown in Figure 4.2a and 4.2b, observations of fade margins [3] show that margins of ~20 dB are required for short-haul (~100 m) terrestrial links at 60 GHz.

As a result of the large amount of propagation data, it has also been possible to establish propagation models for this type of application [6]. Modelled data agrees reasonably well with CCIR-predicted data [7,8] demonstrating the understanding of the propagation medium.

At lower frequencies (<40 GHz), multipath-related problems in the terrestrial application have been investigated through both fading measurements and observation of inter-symbol interference (ISI) [9,10]. Generally, fading is overcome through brute force by the use of high gain antennas, sufficient ERP margins, diversity, and coding. A typical fading profile for an outdoor 37-GHz link operating in a town environment is shown in Figure 4.3. As in the 60 GHz case, fades of 20 dB are observed, as the receiver orientation is varied through angle θ .

In addition, building materials have been investigated to determine their RF characteristics at 37 GHz [9]. Although not done indoors, these measurements are useful for a variety of applications. Figure 4.4 shows the attenuations of a number of building material types. It is interesting to note that the Radio Local Area Network (RLAN) [9] operates using simple high-speed frequency shift keying (FSK). The authors note that there is a significant cost increase when a millimetre-wave network has mobile users, because of the added beam-pointing/steering complexities.

FIGURE 4.2A FADING OBSERVED OVER A 100 m 60 GHz TERRESTRIAL LINK (OPEN FIELD) [3]

FIGURE 4.2B FADING OBSERVED OVER A 100 m 60 GHz TERRESTRIAL LINK (TOWN ENVIRONMENT) [3]

FIGURE 4.3 FADING PROFILE FOR A SHORT HAUL LINK AT 37 and 37.02 GHz IN A TOWN ENVIRONMENT [10]

FIGURE 4.4 ATTENUATIONS OF A VARIETY OF BUILDING MATERIALS AT VARIOUS FREQUENCIES [9]

Although not directly useful in very short range indoor local area networks, several useful observations can be made about the terrestrial application of millimetre-wave communications links:

- Aside from attenuations due to physical obstruction, propagation within buildings has not received a large amount of attention,
- The use of longer ranges (relatively speaking) leads to a very high degree of concern over link attenuation phenomenon,
- Typical fading margins are 20 dB or less,
- There appears to be a high degree of commonality in the link behaviours at 37 GHz and 60 GHz, and
- Multipath related ISI does not seem a central concern, whereas fading does. This is principally due to the data rates being used and the nature of the links.

4.2 Indoor Millimetre-Wave Radio Links

Indoor millimetre-wave propagation has been investigated in recent years, although not to the same degree as its terrestrial outdoor counterpart.

A variety of measurements [11-14] of fading and attenuation phenomena have been made to establish the general characteristics of millimetre-wave links operating within buildings. A variety of building materials have been measured to determine their propagation characteristics; some of these are summarized in Table 4.1¹. It is interesting to note that plasterboard provides ~3 dB attenuation while wood provides ~13 dB and concrete provides very high attenuation.

Typically, observed fading margins are on the order of 20-30 dB [11,12,14] as shown in Figure 4.5. Generally, observations indicate that Rayleigh spreading occurs as a result of reflected paths and Rician spreading occurs as a result of the line of sight (LOS) paths. These phenomena occur in measurements taken at 21.6 GHz, 37.2 GHz and 60 GHz.

¹ This table shows typical thicknesses as might be employed for construction purposes.

TABLE 4.1

PROPAGATION CHARACTERISTICS OF SOME POPULAR
CONSTRUCTION MATERIALS AT 60 GHz [11,14]

Material tested	Transmissivity, %	Reflectivity, %
Plasterboard $s = 1$ cm	42.0	2.0
Chipboard $s = 1.9$ cm	4.5	20.0
Autoclaved aerated concrete blocks $s = 10$ cm	0.0001	16.0

FIGURE 4.5 TYPICAL OBSERVED FADING PHENOMENA USING AN INDOOR PROPAGATION LINKS [11]

The 'free space' propagation losses observed in LOS links deviate somewhat from the D^2 factor normally associated with terrestrial links. Measurement of a variety of indoor links suggests that a closer approximation is $D^{1.1}$ to $D^{1.7}$ for "reflective" wall materials [11-13] but can increase to $D^{2.17}$ for "absorptive" materials [11]. The suggested reason for this is the waveguiding effects of walled rooms.

In addition, impulse response measurements have been made at a variety of frequencies in the indoor environment in order to characterize the delay spreads of the links. Low frequency measurements in the 800 MHz to 1 GHz range [15-17] show mean delay spreads on the order of 200 ns (rms) using non-directional antennas and jitter about the mean of 100 - 250 ns as shown in Figure 4.6.

A variety of modulation schemes are used in these low frequency links, primarily to avoid the spreading-related problems, particularly when using non-directional antennas. Often these schemes use spread-spectrum techniques which are feasible (from a signal processing perspective) at the low data rates being employed in this frequency range (typically 1 Mb/s or less).

At somewhat higher frequencies (1.3 GHz) [18], simulations of indoor direct and obstructed links shows delay profiles which are consistent with those observed in [15-17]. As expected, in the direct, unobstructed case, the spread is reduced as shown in Figure 4.7.

Results at 60 GHz [19] show that building materials generally reduce the spread because they inherently attenuate reflections originating outside the room in which the LOS path exists. This feature of millimetre-wave propagation is beneficial and also provides a basis for frequency-reuse schemes. The use of a combination directional (user) and omni-directional (base/hub) antennas in the link provides a further spread reduction while providing realistic antenna sizes. Figure 4.8A, B and C show the delay spreads measured in various similar room types (noted as areas A, B and C). Nominally the delay spread is 3 - 35 ns rms at 60 GHz. It is noted in [19] that a typical link symbol period should be equal to or greater than ~ 5 times the normalized rms delay spread in order to overcome ISI-related problems.

FIGURE 4.6 DELAY SPREAD JITTER OBSERVED IN LOW FREQUENCY (800 MHz) LINKS [15]

(a)

(b)

FIGURE 4.7 DELAY SPREAD SIMULATIONS OF (a) DIRECT AND (b) OBSTRUCTED INDOOR LINKS AT 1.3 GHz [18]

FIGURE 4.8A MEASURED DELAY SPREAD AT 60 GHz [19]
 (Area A, 20 m x 10 m 'L' Shaped Lab)

FIGURE 4.8B MEASURED DELAY SPREAD AT 60 GHz [19]
(Area B, 3 m x 30 m Corridor)

FIGURE 4.8C MEASURED DELAY SPREAD AT 60 GHz [19]
(Area C, 15 m x 30 m Teaching Lab)

Other delay spread measurements [20] for various indoor environments at 42 and 58 GHz show slightly longer delay spreads of 15 - 70 ns in the LOS case and up to 100 ns in the obstructed-path case. These are probably due to the omnidirectional antennas employed in the testing.

Although extensive research into the area of indoor millimetre-wave propagation has not been carried out, several general points can be made:

- Fading margins of 20 - 30 dB can be expected in the 30 - 60 GHz operating range,
- Free space loss factors of $\sim D^{1.7}$ can be used to approximate the link behaviour;
- Millimetre-wave signals penetrate glass and plasterboard with very little attenuation (3 dB or less when considering typical thicknesses employed in construction),
- Wood and concrete provide such high attenuations of millimetre-wave signals that they essentially provide complete blockage,
- For a communication link with non-directional antennas used for both transmitter and receiver, delay spreads on the order of 100 - 200 ns can be expected,
- For a communication link with directional (20 dBi) and non-directional (~ 0 dBi) antennas, delay spreads on the order of ~ 20 ns can be expected, and
- Since delay spreads are reduced through the use of spatial filtering, further delay spread reduction would probably occur in a link using two directional antennas or antennas of very high gain.

4.3 Anti-Multipath Techniques

As a result of the propagation mediums encountered in the indoor radio environment, a variety of anti-multipath techniques have been implemented in an effort to reduce the effects of interference due to multipath.

Spread spectrum techniques [21-23] have become useful and provide considerable benefit in mobile, non-directive communication applications. Due to the processing overhead, these links generally operate at relatively low bit rates. Spread spectrum has also shown useful performance attributes when operating in very dense electromagnetic environments or where access interference is a prevalent condition [22]. These techniques, despite their cost, have

been proposed for millimetre-wave applications in data transmission systems able to operate at several Mbit/s [24]. An alternative suggested in [24] is to employ multi-carrier operation to achieve higher data rates with simpler transceiver designs. Presumably, the simpler design is amenable to reduced cost.

In cases where directive antennas are used, other techniques are available, including antenna sampling and equalization. In the case of antenna sampling, the incident signals arriving at the receiver are sampled to determine which one provides the optimal link and then this link is used in a timeframe which is shorter than that in which the multipath changes nominally occur. Equalization is useful when the bit period is much less than the channel delay spread but requires more complex, stable receiver/transmitters. The basic concept is to transmit redundant information (using differential phase shifting techniques) over a wide bandwidth and to use transversal filters to constructively combine all of the received signals. A common example of a spread spectrum receiver employing equalization is the "RAKE" receiver as shown in Figure 4.9. Such a receiver has several drawbacks, namely:

- Receiver/transmitter stability requirements,
- Complex equalization circuitry,
- Loss of available data bandwidth due to redundancy, and
- Operation of the equalization/spread spectrum circuitry at very high data rates.

Hybrid approaches are also possible. These use spread spectrum and antenna selection/diversity to improve data rates while retaining multipath rejection performance [23]. The performance of such a receiver is illustrated by the BER plots of Figure 4.10.

Another technique, which exploits spatial rejection of multipath, is the use of adaptive antenna arrays [25]. In this technique the receiver determines the source of the interference and adjusts the complex weighting coefficients of a phased-array antenna to place a rejection notch in the spatial pattern centred on the multipath source. This technique means a large increase in the complexity of the antenna and processing.

FIGURE 4.9 A TYPICAL "RAKE" EQUALIZING RECEIVER [21]

FIGURE 4.10 A HYBRID ARCHITECTURE WHICH EMPLOYS BOTH SPREAD SPECTRUM AND DIVERSITY TO OVERCOME MULTIPATH [23]

Practical limitations in the realization of phased array front-ends tend to reduce the usefulness of this technique at millimetre wavelengths.

A simpler technique is to employ a limiter-discriminator receiver combined with slow, cyclical frequency hopping [26] to avoid the Rayleigh fading phenomenon. Since the receiver/transmitter must retain a high level of synchronization, only low and moderate data rates can be readily achieved. A similar approach is to commute among several signal sets in such a way as to avoid the reuse of a given frequency until the multipath "ringing" has died out [27].

In a practical realization [28], the various general techniques for rejection of local scattering (Rayleigh fading) and intersymbol interference (Rician ISI) have been investigated and compared. Figure 4.11 shows the effects of the origins of these phenomena in the in-building WLAN environment. It was found [28] that adaptive equalization was too complex to implement even at 10 Mb/s and that spread spectrum required a very high spread ratio to provide practical benefits (therefore a much larger bandwidth than the data itself requires). The preferred, and simplest, approach was to use antenna diversity and selectivity for data rates of 10 Mb/s or higher.

Spatial rejection of multipath is a robust and well understood technique which provides benefit in fading and ISI conditions [28]. The use of antenna selectivity augments this technique particularly for mobile or obstructed links.

In summary:

- Most of the processing-based anti-multipath techniques limit the data bandwidth or are not compatible with implementation at millimetre-wave frequencies,
- Techniques requiring stable receiver/transmitters can provide benefit but place cost constraints on the WLAN transceiver,
- Concepts which use coding (M-ary) to reduce the symbol rate such that it is lower than the delay spread of the channel are useful provided that the channel is well understood and relatively static in nature,

LOCAL SCATTERING

Key:

UM: User Module
CM: Control Module

REFLECTIONS

FIGURE 4.11 SOURCES OF MULTIPATH IN THE IN-BUILDING WLAN ENVIRONMENT [28]

- Antenna directivity is a low cost, robust anti-multipath technique. Antenna gain is a readily available commodity at high (millimetric) frequencies, and
- Antenna diversity is useful for mobile links or links which are not generally static in nature.

4.4 Practical Millimetre-Wave Communications Links

The majority of millimetre-wave communication systems which have been implemented have been for military applications in the 60 GHz region. The reason for this was the exploitation of the atmospheric attenuation phenomenon for covert link realizations.

In 1988, the Radio Communications Agency (RA) produced specifications to allow the use of 38, 55, 58 GHz for personal communications networks (PCN) [29]. Recent commercial implementations [30] of systems operating at these frequencies have exploited the RA allocations for wireless microcell applications.

In order to assess the practicality of millimetre-wave communication links, it is useful to investigate the systems which have been developed for military 60 GHz applications. Although these do not meet the WLAN requirements, some similarities in the limitations of implementation technology exist.

First-generation voice-communication systems have been pursued for applications such as binocular mounted person-to-person links [31,32], hand-held ship-to-ship [33] and air-to-air links for fighter aircraft [34]. Within the ground-based (army) application, a large requirement for covert low/moderate data rate links has been pursued through a variety of developments [35] at Ka and V band (~ 38 GHz and 60 GHz) as shown in Table 4.2.

Higher data rates have also been sought [36] in the 60 GHz range. The preferred approach to realizing the transmitter is to use FSK-modulated fundamental Gunn oscillators. The reason for this is the low cost and availability of these sources. A critical design limitation on this approach is the stability and temperature drift of the Gunn oscillator (Figure 4.12). If uncompensated, it can be seen that drifts of several tens of megahertz are possible.

TABLE 4.2

**A SAMPLING OF MILITARY MILLIMETRE WAVE
RADIO DEVELOPMENTS [35]**

<u>MMW Radio Systems (Developmental)</u>					
<u>Frequency</u>	<u>Bandwidth</u>	<u>Format</u>	<u>Configuration</u>	<u>Contractor</u>	<u>Year</u>
60 GHz	20 MB/s	Simplex	Tripod-Mount	Lab Built	76
36 GHz	50 MB/s	Duplex	Rack-Mount	HAC	77
37 GHz	5 MHz	Simplex	Hand-Held	Norden	78
	1.5 MB/s	Duplex	Tripod	Norden	78
60/62 GHz	20 MB/s	Simplex	Rack-Mount	AIL	78
38 GHz	20 MB/s (TV)	Duplex	Mast-Mount	Norden	78
37 GHz	1.15 MB/s	Duplex	Tripod	Norden	78-79
54 GHz	Voice, IFF	Net	Omni	Norden	80
54 GHz	Voice, IFF	Simplex	Hand-Held	Norden	80
38 GHz	20 MB/S	Duplex	Mast-Tripod	Norden	80
36-38.6 GHz	20 MB/s	Tri-Tac	Militarized	Norden	80-82
		Compatible	6.3A		
54-58 GHz	5 MB/s	Duplex	Electronic	ITT	80-82
			Tunable	HAC	
				Norden	
55-57 GHz	576 KB/s	TDMA	Distribution System	Motorola	82-84

FIGURE 4.12 TYPICAL GUNN OSCILLATOR DRIFT OVER TEMPERATURE [36]

A technique useful for operating the link in full duplex is to exploit dual polarization operation and to rely on the polarization isolations (nominally > 20 dB) as shown in Figure 4.13. This technique could also be useful for simultaneous transmissions, where it could reduce the symbol rate by a factor of 2. Other examples of practical links [37,38] at moderate data rates have also exploited FSK, due to the ease and reliability of carrying out the modulations.

Time diversion multiple access (TDMA) schemes have also been exploited in military microcell networks using offset frequencies for reception and transmission [39]. Using this scheme it is possible to transmit a "beacon" from the master station continuously while the transmitters cycle through their individual frames. Although not explicitly suggested in [39], this technique could be used to maintain transmitter/receiver alignment or synchronization. In this application the TDMA scheme relies on directional slaves and an omni-directional master station, and exploits an FSK-based modulation scheme.

In commercial applications, there has been a variety of developments of millimetre-wave communication systems. One example is a railway-mounted system operating at 35 and 59 GHz [40,41]. Such systems use amplitude-shift keying (ASK) modulation(s) because of cost advantages.

In the 27 - 29 GHz range, commercial radio systems have been developed [42] as a means of overcoming the radio congestion in urban environments. Some of these exploit E-plane integrated circuit techniques [43,44] in order to achieve cost reduction. The use of E-plane techniques has also been demonstrated in the 54 - 58 GHz range for personal communications networks (PCN) [45]. Although the need for stable local oscillators is pointed out in [45], no indication is given as to how the selected ideas are realized in E-plane technology. One problem with E-plane techniques is its general incompatibility with mounting monolithic devices.

Other realizations of systems operating at 22, 31 and 50 GHz [46-48] exploit printed microwave integrated circuits (MICs) in an effort to achieve cost reduction. ASK is the preferred modulation technique due to its ease of MIC realization. Data rates of as high as 20 Mb/s are reported.

FIGURE 4.13 USE OF DUAL POLARIZATION TECHNIQUES TO ACHIEVE FULL DUPLEX TRANSMISSIONS [36]

The use of waveguide integrated circuitry has also received developmental attention [49-50]. Both FSK and ASKL schemes have been successfully employed at rates of up to 133 Mb/s. The use of waveguide is deemed to be economically feasible through the exploitation of electron discharge milling (EDM) fabrication techniques. The use of EDM allows the critical dimensions of waveguide integrated circuits to be realized with useful repeatability.

A much lower-cost technique which has recently evolved as a result of Defense Advanced Research Program (DARPA)-funded research and development [in the US] is high-yield monolithic microwave integrated circuits (MMICs). Although the DARPA thrust is toward low-cost MMIC building blocks for military applications, a variety of the developments are dual purpose. Through the influence of the DARPA program, several foundries [51] have developed the capability to provide building blocks able to operate in the 30 - 100 GHz range. Table 4.3 shows a cross section of these foundries and the maximum operating frequency of their processes. The availability of building blocks in this technological form allows the implementation of small low-cost hardware items. These can be produced using automated processes free from the traditional "tuning" required by the traditional MIC and waveguide processes. This "tuning" of production circuitry often incurred a significant cost because of the high skill required in the manual tuning process.

The availability of foundries able to produce millimetre-wave MMIC building blocks has recently led to the development of multifunction MMICs [52,53] for a variety of applications, including missile seekers. These multifunction MMICs further reduce cost by avoiding the need to combine several discrete MMICs to realize a particular function. For high-volume applications, this approach yields the minimum production cost but requires substantial design and development effort.

Currently, low and moderate quantity production relies extensively on hybrids realized through the use of several interconnected single-function MMICs. These hybrids have found their way into the commercial PCN marketplace. One example is a hybrid MMIC transmitter for PCN applications at 38 GHz [54]. The transmitter relies on "chip and wire" techniques to interconnect several MMIC devices into a single multifunction module. Currently a

TABLE 4.3
MILLIMETRE WAVE MMIC FOUNDRY CAPABILITIES [51]

TABLE III PROCESS CAPABILITY								
Years	Min. Line Width (μ)				Max. Frequency (GHz)			
	85	88	91	94	85	88	91	94
Alpha	0.50	0.25	0.25	0.25	20	35	60	80
Anadigics			0.30	0.25			30	30
Avantek	0.50	0.25	0.15	0.10	20	40	60	100
Gen'l Elec	0.25	0.20	0.15	0.10	60	60	100	100
Harris	0.50	0.50	0.25	0.10	21	21	44	60
Hughes	0.50	0.40	0.25	0.10	30	35	60	94
ITT Defense	0.80	0.40	0.40	0.25	20	20	20	40
Litton	0.50	0.20	0.10	0.08	26	50	100	> 100
M/A-COM	1.00	0.50	0.50	0.25	12	20	30	60
Raytheon		0.50	0.25	0.15		20	35	44
Texas Inst	0.50	0.50	0.25	0.15				
TriQuint	1.00	0.50	0.50	0.35	12	18	18	30
TRW			0.20	0.10			70	100
Westinghouse	0.70	0.50	0.30	0.10	12	16	20	94
Average	0.61	0.39	0.28	0.17	23	30	50	69

growing number of building blocks are commercially available [55,56], including power amplifiers, low noise amplifiers, mixers and oscillators.

In more widespread use are hybrids which exploit currently available MMIC and conventional MIC circuitry. MIC circuitry is used to interconnect MMICs and other circuitry not conventionally realized in MMIC building blocks [57]. Some examples of this are hybrid beam-steered transmitters [58] operating at 32 GHz, the ALTAIR WLAN transceiver [59] operating at 18 GHz (Figure 4.14), and tactical radio transceivers for MILSTAR applications (Figure 4.15) at 44/20 GHz [60].

Although the millimetre-wave integrated circuitry drives cost to a large extent, intricate machined housings are generally required to house these circuits. In many cases the housing also forms part of the actual circuitry. The cost of these housings often prevents truly low-cost realizations. A possible very low-cost solution is to use injection moulded, plated-plastic housings [61]. This technique has been exploited to house E-plane automotive radars operating at 80 GHz.

Several generalizations can be made concerning the technology involved in realizing practical millimetre-wave transceiver hardware:

- Cost has limited the designs to very simple modulation schemes,
- Practical limitations in the receiver/transmitter circuitry must be accounted for in the modulation scheme selected,
- A variety of modulation techniques have been employed, mainly FSK and ASK because of their simplicity,
- Very little work has been reported on very-high-data-rate realizations (i.e. > 150 Mb/s),
- The most cost-effective transceiver realization must exploit MMIC technology with a goal to employ a single-chip multifunctional RF front end, and
- Very little effort has gone into overcoming multipath-related link problems in operating systems:

FIGURE 4.14 THE ALTAIR WLAN TRANSCEIVER MMIC/MIC HYBRID [59]

FIGURE 4.15 MILSTAR TACTICAL RADIO TRANSCEIVER MMIC/MIC HYBRID [60]

4.5 References for Chapter 4

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5.0 MILLIMETRE-WAVE WLAN TRANSCEIVER ARCHITECTURE INVESTIGATION

5.1 Factors

Based on the literature review (Chapter 4) several key conclusions can be stated which [partially] dictate specific transceiver architectures:

- The operating frequency is not dictated by propagation phenomena. This is because of the commonality of characteristics (across the 20 - 60GHz range) and short ranges involved.
- Fading margins of ~ 20 dB must be accommodated.
- Aside from minor obstructions, the links must be LOS. Propagation loss factors of $\sim D^{1.7}$ are applied.
- Highly directional antennas are required as a low-cost means of reducing both Rayleigh and Rician fading effects. This may preclude mobile-user station realizations.
- Delay spreads of ~ 20 ns rms must be accommodated by the modulation scheme employed. This may be reduced through increased spatial rejection.
- The costs of complex anti-multipath techniques have largely precluded their application in millimetre-wave WLANs. This is compounded by practical limitations in transceivers.
- Practical transceivers must rely on MMIC and SMT technologies to reduce cost. Use of plated plastic is likely required to cost-effectively realize electrically critical housings.
- Existing systems operating in the millimetre-wave regime rely on very simple signalling schemes due to the high cost of millimetre-wave RF hardware.

5.2 Top-Level Design Trade-offs

The key focal area of this thesis is the millimetre-wave transceiver. As an embedded module within the WLAN terminal it is necessary to investigate the terminal design sufficiently to establish the top-level design of the transceiver module. A number of factors originating with the WLAN terminal design and operation directly influence the transceiver architecture and detailed design. When considering the top-level architecture of the WLAN terminal, two key trade-offs must be addressed, namely:

- Modulation scheme, and
- IF (indirect) or RF (direct) modulation.

5.2.1 Modulation Scheme Selection

A variety of modulation schemes is available. The selection of which is more appropriate for a specific application is dependant on a number of factors. In this case, the dominant factors are:

- Simplicity,
- Multipath resistance (ISI particularly),
- Sensitivity to transmitter and receiver instabilities, and
- Provision of adequate data rates (155 Mb/s expandable to 660 Mb/s).

Two factors allow for reduction in ISI (other than spatial rejection): the reduction of the symbol rate such that

$$T_{\text{symbol}} > 5 \times T_{\text{rms_delay_spread}}$$

or the use of adaptive equalization.

Reduction in the symbol rate can only be accommodated by exploiting a multi-carrier transmission or by increasing the information in each transmitted symbol (multi-level transmission).

Multilevel transmissions using phase modulation require increased transmitter/receiver stability while use of amplitude keying requires increased transmitter power and more complex AGC-based receiver circuitry. Multilevel-frequency keying is possible but since only one symbol is transmitted at a time a suitable number of frequency "keys" is not readily attainable. When considering delay spreads of ~ 20 ns rms, the symbol rate needs to be on the order of 10 Msymbols/s (or $T_{\text{symbol}} = 100$ ns). This requires an alphabet of 2^{16} (since the 155 Mb/s rate desired is a factor of ~ 16 faster than the allowable symbol rate of 10 Msymbols/s) which is orders of magnitude higher than what is practically achievable when considering the receiver/transmitter drifts associated with low cost, free-running oscillators at millimetre-wave (temperature stabilized).

More practical would be the use of multiple subcarriers in order to transmit more bits in parallel. A multifrequency scheme could be employed using simple ASK (OOK), BFSK, or BPSK. Table 5-1 shows a simple tradeoff of top-level characteristics of each of these cases.

Of these, PSK offers the optimal SNR performance as shown in Figure 5.1, while OOK offers the simplest circuitry, but suffers from ~ 6 dB worse SNR requirements. Nominally, to achieve a probability of error of 10^{-6} requires a minimum SNR of ~ 15 dB. FSK allows the use of limiter-based receiver circuitry and a compromise in required minimum SNR.

Effective exploitation of subcarriers requires that the simultaneous symbol transmission rate be ~ 10 Mb/s (in order to avoid ISI effects). This requires that 16 subcarriers be employed. The result is that 16 separate receive/transmit channels must exist. This presents a complex solution which is susceptible to spurious effects and crosstalk. Figures 5.2 and 5.3 show examples of an M-subcarrier BFSK modulator and demodulator to illustrate the complexity.

Adaptive equalization is commonly applied to improve multipath performance at low data rates (i.e. < 5 or 10 Mbit/s). At higher data rates however, this becomes a complex and costly solution which is applied at the receiver IF. The cost of this approach is generally associated with the need to exploit surface acoustic wave (SAW) tapped delay lines in the transversal filter circuitry [2-4]. The fabrication of high-performance SAW devices on LiNbO_3 is not amenable to high-volume, low-cost production. A more recent approach to this problem is to employ GaAs-based technology to realize acoustic charge transport (ACT) devices useful

TABLE 5-1
TRADE-OFFS BETWEEN MODULATION SCHEMES [1]

TRADE-OFF FACTOR	NON-COHERENT PHASE SHIFT KEYING (PSK)	AMPLITUDE SHIFT KEYING (OOK)	NON-COHERENT FREQUENCY SHIFT KEYING (FSK)	COHERENT FSK
1. Relative Complexity for Binary Operation	Low	Low	MOD	MOD
2. Relative Complexity for M-ARY Operation	High	High (A)	High	MOD
3. Relative Minimum Operating SNR	Low	High	MOD	MOD
4. Inherent Suppression of ISI (B)	MOD	MOD	MOD	High
5. Sensitivity to Rx, Tx, Temp. Drifts	High	Low	Low	MOD
6. Operating Bit Rate	High	High	High	MOD/Low

NOTES: (A) M-ARY refers to multi-subcarrier OOK
 (B) Assuming M-ARY operation

FIGURE 5.1 ERROR PROBABILITIES FOR VARIOUS BINARY DIGITAL MODULATION SCHEMES [1]

FIG93/002

FIGURE 5.2 A TYPICAL MULTI-CARRIER BFSK MODULATOR

FIG93/003

FIGURE 5.3 A TYPICAL MULTI-CARRIER BFSK DEMODULATOR

for programmable transversal filter (PTF) applications. Using this technique, the cost benefits of GaAs processing are exploited while simultaneously allowing the fabrication of the tapped delay line structure and the support circuitry (i.e. I/O RAM, tap weight setting circuitry, etc.) on a single device [5,6] as shown in Figure 5.4. A currently available device is shown in Figure 5.5. Such devices are able to operate with non destructive sensor (NDS) tap delay spacings on the order of 5.6 ns (and 128 taps) and can achieve 5-bit tap weighting. This device could theoretically equalize a 640 ns window at rates of ~ 180 MHz. Taps are adjusted serially in the device and require a few hundred nanoseconds per tap for the execution of a weighting modification. Devices currently under development include an ACT-based PTF which employs 256 taps with ~ 3 ns spacing and 8-bit tap weighting.

To exploit ACT-based equalization at IF, the receiver/transmitter drift must be maintained within the RF input bandwidth of the equalizer device and variations must be slow with respect to the equalization resetting intervals. In the case of ACT-based PTF devices, the input bandwidths can easily be several hundred megahertz. A limiter-based FSK demodulator, simple AGC-based OOK or differential phase demodulator would operate well with this device in conjunction with the appropriate weighting scheme (i.e. zero forcing) and clock-recovery circuitry. Of these possibilities differential BPSK and noncoherent BFSK have the advantages of constant signal presence and reduced SNR requirements.

Constant signal presence has benefits when considering the implementation of any required AFC or AGC loops. Stable symbol rate communication is a pre-requisite for the application of fixed tap spacing PTF-based equalization schemes. For the purposes of extracting the needed equalization coefficients, both the in-phase and quadrature components of the incoming signal must be available and relatively stable for the equalized time interval. A simple I/Q IF stage can be readily implemented to accommodate this.

The use of high speed equalization is an elegant solution and results in modulation/demodulation circuits of considerably lower complexity than their multi-carrier or multi-level counterparts.

FIGURE 5.4 ACT GaAs PROGRAMMABLE TRANSVERSAL FILTER STRUCTURE [7]

ACT Programmable Filter Module Organization

FIGURE 5.5 ACT GaAs HYBRID DEVICE [7]

5.2.2 Modulation at IF or RF

This trade-off factor is largely dictated by the type and rate of the modulation scheme being employed. Direct (or RF) modulation and demodulation is desirable due to the reduced number of RF components required. However, the use of more exotic modulation/demodulation and ISI suppression schemes, such as multicarrier transmissions or equalization are generally not viable at millimetre-wave RF.

Selection of modulation/demodulation at IF would provide an increase in flexibility at the expense of minor increases in the complexity of the RF hardware.

5.3 WLAN Terminal Architecture Definition

The WLAN terminal forms a portion of a microcell in which several slave terminals may operate in conjunction with a single master terminal. The slave terminal would employ a highly directive antenna while the master terminal would employ an omni-directional antenna. Another possible approach is to employ omni-directional user terminals (to allow mobility) and antenna switching in the master terminal (continually selecting optimal paths to a particular user). This approach is not pursued because it is expected to incur several drawbacks such as increased complexity/cost in the master terminal, increase in cost of the user terminals due to the differential antenna, and reduced available ERP (and hence path length) when considering limitations in transistor technology. For the microcell network to operate with a given hierarchy, system level configuration of the WLAN terminals is required. This is deemed to be outside the scope of the current work.

The top-level WLAN terminal architecture is shown in Figure 5.6. The terminal is essentially a converter which transforms the hardware interface (i.e. SONET, ETHERNET, HIPPI) on to a radio link which operates seamlessly and transparently from the user's point of view. The use of simple high-speed differential PSK or non-coherent BFSK modulation in combination with high-speed ACT-based equalization at IF is employed so as to achieve the required data rates in the presence of intersymbol interference (ISI).

FIG93/001

FIGURE 5.6 TOP-LEVEL WLAN TERMINAL ARCHITECTURE

The terminal operates in half-duplex time-division-multiplexed mode transmitting bursts during its predetermined time slot. RF is received/transmitted by the RF transceiver and converted to/from a low frequency IF. At IF, high-speed modulator/demodulators are employed to provide data to/from the interfacing module. The interfacing module would provide suitable interfaces to the system network (in the case of a master terminal) or to the user's computer (in the case of a slave station).

The selection of half duplex (rather than full duplex) operation is based on the minimization of the required RF hardware and the difficulties in implementing RF circulators in low-cost MMIC form. The circulators are needed to retain a single transmit/ receive antenna (for cost reasons) while transmitting and receiving simultaneously at different frequencies.

5.3.1 Terminal Interface Module

This module allows for interface to the external data link and formats data information for use by the modulation circuitry or from the demodulation circuitry. Included in this module is the control and timing of data and networking. In addition, the interface module controls the equalization of incoming data. This could be achieved through the use of equalization codes which are transmitted for the purpose of setting the equalization to a preset condition and assuming this condition will be present during the subsequent reception of data. Alternatively, adaptive equalization can be carried out through the minimization of mean squared errors of the data itself.

Being outside the focal area of the current study, no further effort is expended on this module.

5.3.2 Power Conditioner Module

This module may take on several forms, depending on its intended application:

- Operation from standard 120 VAC, and
- Battery operation.

In an office environment, it is not likely that the latter will be the dominant requirement. Focusing on 120 VAC operation, issues such as "power saver" modes are unlikely to be required and this module reduces to a DC rectification and regulation circuit which can be developed in low-cost hybrid form using conventionally available technology.

As a "conventional" module, no further effort is focused on it.

5.3.3 Modulator Module

The modulator module architecture is shown in Figure 5.7. Data is formatted for modulation by the system interface module and sent to the modulator module where it modulates the IF carrier. Differential BPSK is selected as a simple technique which consumes less bandwidth than non-coherent BFSK. Retaining the signal within the receiver passband is less problematic with differential BPSK since the transmitter/receiver drift issue is the only factor requiring consideration (aside from data bandwidth).

Limitations in the state switching speed of the modulator dictate the rate of data transmission achievable. State switching speeds of 3 ns are currently available in low-cost SMT devices. The use of FET-based switching in conjunction with delay-line phase shifting networks could theoretically provide switching times of less than 1 ns. The current technological roadblock is in the area of the voltage driving circuitry required to provide the FET gate input voltage swings.

FIG93/009

FIGURE 5.7 MODULATOR MODULE ARCHITECTURE

5.3.4 Demodulator Module

The demodulator module architecture is shown in Figure 5.8. The IF received by the transceiver module is sent to the demodulator module where IF equalization is applied and the modulation envelopes subsequently detected. Differential BPSK or noncoherent BFSK modulation selection results in very little difference in the circuitry. The resulting data stream(s) are sent to the interface module for subsequent formatting, error checking/correcting and transmission.

The demodulator architecture is based on ACT-based transversal filtering to equalize the data against the impulse-response effects of the propagation channel. In the short term these devices will dictate the achievable data bandwidth because the minimum tap spacings need to be matched to the symbol duration. Currently tap spacings of ~ 5 ns are achievable. Interleaved tap structures in development are expected to yield tap spacings of ~ 3 ns or less.

Currently-available GaAs-based frequency/phase demodulator SMT devices able to operate at maximum rates of 1 GHz are currently available.

5.3.5 RF Transceiver Module

The RF transceiver module provides the conversion of the IF to/from millimetre-wave. The module is based on GaAs monolithic microwave integrated circuitry (MMIC) to minimize the eventual cost of the millimetre-wave WLAN terminal. In addition, the transceiver provides a tunable operating frequency which facilitates frequency re-use.

FIG93/010

FIGURE 5.8 DEMODULATOR MODULE ARCHITECTURE

NOTE TO USERS

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UMI

FIGURE 5.9 MILLIMETRE WAVE RF TRANSCEIVER MODULE ARCHITECTURE

not considered. Printed circuit antennas [10] offer a low-cost solution and can be configured as arrays. Some complexity is incurred when considering gains on the order of 25 - 30 dBi.

The gain of such an array can be approximated by [10]

$$G = 10\log(4\pi L^2) - \alpha L - C$$

where L = electrical length of a side of the aperture in wavelengths

α = loss per unit wavelength

C = mismatch losses (~ 1.7 dB)

A number of printed arrays have been reported which exploit this technique to achieve gain using large numbers (e.g. 256) of the low gain printed patch or endfire slot elements [10]. Considerable complexity is involved when considering large planar arrays and feed networks able to achieve gains on the order of 25 - 30 dBi [11]. When considering cost issues, the laminating and mounting of this type of structure needs to be considered in addition to the cost of the "raw" printed circuitry (often misperceived as the only associated "cost" of this type of antenna).

Generally, simple horn antenna elements appear to provide a suitable choice in this application since conformality is not required. A variety of polarizations can be accommodated, simple linear polarization requires the least number of components (e.g. polarizing grids or transducers).

In the case of the master terminal, an omni-directional antenna is required. The simplest millimetre-wave omni-directional antenna is the biconical horn. In theory printed arrays could be applied in this application. Because a biconical antenna is not amenable to low-cost fabrication, the use of a printed omnidirectional antenna may be viable. Such an antenna array might take the form of a circularly formed linear array. The linear array provides a simple implementation and feed network [10].

The use of a biconical omnidirectional horn antenna or printed circular array on the master station tends to dictate that the system polarization be linear or slant linear, because of the inherent radiating characteristics of these designs. This is readily accommodated by the user station directional horn antenna. No data exists on the preferential characteristics of various polarizations in the intra-office radio environment (as is found in the over-sea environment for example).

Frequency reuse is accommodated through the tuning (or switching) of the local oscillator (driving both up- and down-conversion) to offset the radiating frequency of the transceiver. Typically, MMIC-based VCOs allow tuning bandwidths of 1 - 5% in this frequency range resulting in the availability of approximately 3 operating slots. Componentry upstream of the RF mixers (and including them) must accommodate the entire accessible bandwidth.

Transmit/receive (T/R) switching is required to realize the half-duplex operation of the module. T/R switching is accommodated at RF by a high speed-switch which is driven to the appropriate state through a synchronization of the terminal with the TDMA network in which it is operating.

This transceiver module forms the kernel of more detailed theoretical and practical investigations to be undertaken in a subsequent section of this thesis.

5.4 Qualitative WLAN Terminal Performance Estimation

Based on the top-level architecture outlined in previous sections, it is possible to qualitatively assess various performance attributes and compare these with the general requirements for the terminal.

Of particular interest is the operating range of a given user-master link within a microcell. This can be estimated using a standard formulation for one-way transmission [13]. Modifying the D^2 factor to $D^{1.7}$ to become [13, modified]

$$P_R = \frac{(P_T G_T G_R \lambda^2)}{(4\pi)^2 D_{TR}^{1.7} L_R L_T L_P}$$

- where P_R = received power, W
 P_T = transmitted power, 80 mW in their case
 G_T, G_R = transmitter, receiver antenna gains respectively, nominally 25 dBi and +3 dBi
 λ = transmission wavelength, 7.9 mm AT 38 GHz
 D_{TR} = distance between transmitter and receiver, 50 m (max)
 L_P = propagation loss, assumed to be ~ 0 dB in this application due to the very short ranges involved
 L_T = circuit losses between the transmitter and the antenna, ~ 2.5 dB assumed
 L_R = circuit losses between the antenna and the receiver, ~ 2.5 dB assumed. This is accounted for in the calculation of the receiver noise figure.

Using the above parameters, the received signal power is ~ -51 dBm at the input to the receiver. When considering the processing capabilities of the receiver it is necessary to evaluate several other factors. Employing OOK modulation requires a transmission bandwidth roughly equivalent to the digital transmission rate [1] (i.e. 155 MHz). When considering the receiver and transmitter, an additional 50 MHz is required to accommodate drift. This results in an IF bandwidth of approximately 200 MHz and a video bandwidth of 155 MHz. The effective noise bandwidth can be approximated by [14]

$$B_{eff} \approx [(2 B_R B_V - B_V^2)]^{1/2}$$

where B_R and B_V are the RF and video bandwidths respectively.

The resulting effective bandwidth is approximately 200 MHz. The noise floor in this bandwidth is given by [1]

$$N_{power} = \kappa T B_{eff}$$

where κ = Boltzmann's constant

T = temperature, Kelvins.

The noise power is approximately -90 dBm. In addition, the receiver noise contribution of 5 dB (due to the receiver noise figure at room temperature) will increase this level to -85 dBm.

The difference between the approximate received signal power and the noise power represents the operating SNR and must accommodate any margins required and the minimum SNR needed for detection at the appropriate false alarm rate. The fading margins necessary to ensure some probability of service are difficult to determine due to the limited amount of data on which to base this. From the literature survey it was determined that margins on the order of 20 dB are required in the indoor millimetre-wave environment. It should be noted, however, that this data does not really reflect the propagation conditions when considering a highly directional LOS propagation path above the level of reflecting bodies (i.e. along the ceiling within an office, as noted in Section 2). The resulting minimum SNR is 14 dB. Figure 5.1 shows that at this SNR, detection and false alarm performance is acceptable. Only a minor margin exists for errors in estimating the transceiver characteristics, fading margins, and distance dependencies.

Although the operating range of the link is of primary importance, a variety of operating characteristics are dictated by fundamental design constraints, a number of which have been addressed throughout this architecture investigation. Tables 5.2 and 5.3 summarize these attributes and compare them with the required levels.

TABLE 5.2

**WLAN USER TERMINAL: REQUIRED vs ESTIMATED
PERFORMANCE ATTRIBUTES**

FACTOR	UNITS	REQUIRED PERFORMANCE	ESTIMATED PERFORMANCE	NOTES
1) Operating Frequency	GHz	38	38	- somewhat dictated by limitations in GaAs MMIC technology
2) Selectable Operating Frequency	--	5	2	- ~600 MHz bandwidth usage
3) Transmitter Power	mw	>20	-50	- at antenna input
4) Receiver Noise Figure	dB	<12	-8	- at antenna input
5) Transmitter/Receiver Antenna Gain	dBi	>20	25	
6) Polarization	--	Circular	Linear	- V, H or slant orientation
7) Operating Range	m	50 target, 30 min.	50	- LOS - 20 dB margin assumed - 14 dB min. SNR
8) Minimum Operating SNR	dB	20 target, 10 min.	14	- at 7
9) Data Rate	Mbit/s	155 target, 25 min.	155 Mbit/s min., 330 Mbit/s target	- half duplex, TDMA
10) Modulation	--	TBD	Differential PSK or FSK	
11) Power Consumption	w	Minimize	TBD	
12) Physical Size	--	Minimize	TBD	- exploitation of SMT, MMIC circuitry
13) Technological Implementation	--	MMIC, SMT	MMIC, SMT	
14) Long Term Production Cost Target	\$	200	TBD	

TABLE 5.3

**WLAN MASTER TERMINAL: REQUIRED vs ESTIMATED
PERFORMANCE ATTRIBUTES**

FACTOR	UNITS	REQUIRED PERFORMANCE	ESTIMATED PERFORMANCE	NOTES
1) Operating Frequency	GHz	38	38	- somewhat dictated by limitations in GaAs MMIC technology
2) Selectable Operating Frequency	--	5	3	- -600 MHz bandwidth usage
3) Transmitter Power	mw	>20	-50	- at antenna input
4) Receiver Noise Figure	dB	<12	-8	- at antenna input
5) Transmitter/Receiver Antenna Gain	dBi	-0	+2	
6) Polarization	--	Circular	Linear	- V, H or slant orientation
7) Operating Range	m	50 target, 30 min.	50	- LOS - 20 dB margin assumed - 14 dB min. SNR
8) Minimum Operating SNR	dB	20 target, 10 min.	14	- at 7
9) Data Rate	Mbit/s	155 target, 25 min.	155 Mbit/s min., 330 Mbit/s target	- half duplex, TDMA
10) Modulation	--	TBD	Differential PSK or FSK	
11) Power Consumption	w	Minimize	TBD	
12) Physical Size	--	Minimize	TBD	- exploitation of SMT, MMIC circuitry
13) Technological Implementation	--	MMIC, SMT	MMIC, SMT	
14) Long Term Production Cost Target	\$	1,000	TBD	

5.5 References for Chapter 5

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6.0 MILLIMETRE-WAVE TRANSCEIVER DESIGN

The WLAN RF transceiver needed to implement both the user and master terminal within the microcell forms the focal point of this section of the thesis.

6.1 Transceiver Requirements

Based on efforts carried out in the terminal architecture investigation of Chapter 5 it is possible to formulate the requirements for the transceiver. Generally, the transceiver can be viewed as a three port device with defined interfaces as shown in Table 6.1.

6.2 Transceiver Architecture Design Drivers

The architecture of the transceiver is dictated by the following trade-off factors:

- Transmit/receive frequency,
- Single or double conversion,
- Simultaneous receive/transmit functionality, and
- Frequency reuse.

The operating frequency of the transceiver is largely determined by the requirements for the WLAN terminal. It has been determined (Chapter 5) that a suitable operating frequency for the terminal is ~ 38 GHz; this becomes a driving factor in the design of the transceiver. The terminal is operated in half duplex (i.e. time sharing the receive and transmit functions) to allow the use of a single carrier frequency for both the transmit and receive functions, simplifying the overall design.

TABLE 6.1

MILLIMETRE-WAVE TRANSCEIVER PERFORMANCE REQUIREMENTS

FACTOR	UNITS	NOMINAL REQUIREMENT	NOTES
1. IF input frequency	GHz	0.5 to 1	
2. IF input power	dBm	0 to -15	Manual gain control
3. RF output frequency	GHz	38 - 40	
4. RF output power	dBm	> +15	
5. Rx/Tx switching speed	ns	<10	
6. Rx noise figure	dB	8	At antenna/ transceiver interface
7. Operating freq. tunability	MHz	>500	
8. IF output frequency	GHz	0.5 to 1	
9. IF output spurious, LO leakage, image signals levels	dBc	>40	RF in at -25 dBm
10. Linear IF output dynamic range	dBm	-70 to -25	Function of range to transmitter
11. Limited IF output	dBm	-15	
12. Manual gain control	V	TBD	External
13. Tx/Rx switching control	V	TBD	External
14. Operating frequency control	V	TBD	External
15. IF in/out, RF port VSWR	Ratio	2 : 1	

The use of single stage conversion (up and down) with a suitably low IF will result in the following:

- Increased potential for spurious outputs from the transmitter,
- Decreased image suppression in the transmitted output,
- Decreased power in the desired output carrier transmitted due to spurious and LO feedthrough into the final amplification stages,
- Decrease in the available reuse bandwidth,
- Increase in the complexity of the RF preselection filter required to isolate the local oscillator from the transmitted output, and
- Increase in the spurious levels in the received IF passband.

Although there is an incremental cost associated with two stage (or double) conversion in the transmitter and receiver, the necessity to optimize the transmitted RF power in the desired carrier drives the design to this architecture configuration. In addition, the need to implement reuse schemes and the necessary suppression of spurious products in the receiver IF also dictate the two stage conversion approach.

6.3 Transceiver Detailed Design

The transceiver top-level design is shown in Figure 6.1. The 0.5 - 1 GHz IF input is upconverted to a 4 - 4.5 GHz IF by the 5 GHz local oscillator (fixed) and then to RF using a 35 GHz LO which allows for reuse tuning. The use of a narrow band power amplifier ensures that only the desired sideband is amplified and passed to the RF bandpass filter via the T/R switch. The output of the filter is then radiated from a directional (user terminal case) or omni directional (master terminal case) antenna. The RF bandpass filter ensures full suppression of spurious transmitter signals.

Incoming RF is received by the antenna and routed through to the RF bandpass filter and T/R switch to the low noise RF amplifier. The desired RF is amplified and downconverted through a double downconverting chain to the 0.5 - 1 GHz IF.

FIG93/025

FIGURE 6.1 TOP-LEVEL TRANSCEIVER BLOCK DIAGRAM

6.3.1 Frequency Plan

An important element of the transceiver design is the design of the frequency plan.

The frequency plan must consider the following:

- Spurious product generation;
- LO leakage,
- Image suppression,
- Component performance, and
- Filtering requirements vs feasibility.

The frequency plan employed in this transceiver is shown graphically in Figure 6.2. In the high frequency stage, the LO is placed at 35 GHz which allows the LO and image band to be suppressed by the respective input/output filters. The rejection requirements on the filters are not severe and thus should not drive cost. The 35 GHz LO is tunable over a 500-MHz band allowing for reuse. The tuned band is readily isolated from the IF and RF passbands due to the spectral offset.

The low frequency conversion stage frequency plan also allows for suppression of the LO and image bands from the second IF passband (0.5 - 1 GHz).

In both cases, since the LO is spectrally separated from the respective RF passbands, an improved spurious suppression performance will be achieved. The production of spurious signals is particularly important when considering the receiver chain since undesired spurious signals tend to compete with desired tones effectively reducing the S/I ratio. In digital communications systems this factor manifests itself in reduced BER performance.

In order to assess the spurious production of this frequency plan, an analysis of the low and high frequency conversion circuit structures has been undertaken using the prediction method outlined in [1]. The following spurious factors can be estimated using this approach:

- High-order mixing products (so called MxN products),
- LO leakage,
- IM products, and
- Undesired 1x1 products (i.e. out of band signals due to insufficient suppression).

FIG93/024

FIGURE 6.2 TRANSCEIVER FREQUENCY PLAN

The model employed for the analysis is shown in Figure 6.3. The following input parameters were employed for the high frequency down conversion stage:

- 38-40 GHz (3 dB points) RF filter with 30 dB/octave selectivity,
- 4-4.5 GHz (3 dB points) IF filter with 40 dB/octave selectivity,
- LO at 35 GHz, and
- A single-balanced mixer with 7 dB conversion loss and +13 dBm input third-order intermodulation intercept point (P_{3IM}).

An analysis of mixing products up to 4x4 terms was carried out using an RF range of 36 - 41 GHz while observing any outputs in the 3.5 -5 GHz range. A fixed input RF power of -20 dBm is employed, being the maximum anticipated worst case level. When considering the range performance analysis carried out in Chapter 5, this would allow operation down to < 5 m. The results of this analysis are shown in tabular format in Table 6.2. This table shows all outputs which are spurious in the 3.5 - 5 GHz range and which are suppressed by 60 dB or less (with respect to the output carrier). The results shown in the table indicate that the only undesired products are 1x1 out-of-band terms whose suppression is limited by the selectivity of the filters employed.

A second analysis using the same approach was carried out on the low frequency converter stage. In this case a DC - 1 GHz low pass filter is employed as the final filter due to the difficulty in building a 0.5 - 1 GHz bandpass filter (this is a very large percentage bandwidth).

The input/output filters employ 40 dB/octave roll-offs. In this case the input signal is swept over 3.5 - 5 GHz and the output is observed over the DC - 1.5 GHz range using a fixed LO at 5 GHz. The output spurious is shown in tabular form in Table 6.3.

The outputs shown in Table 6.3 indicate that the spurious generation is limited to 1x1 products and some 2x2 products. The 2x2 products are partially suppressed by the balancing techniques employed [2]. As in the high frequency converter stage, a single balanced design is employed for analysis purposes although in this frequency range double and even triple balancing is readily attainable for larger suppression levels. Undesired products are suppressed by ~ 60 dB in this case.

FIG93/023

FIGURE 6.3 CONVERTER MODEL USED IN SPURIOUS ANALYSIS [1]

TABLE 6.2
HIGH FREQUENCY CONVERSION STAGE SPURIOUS ANALYSIS RESULTS

SPUR FREQ (MHz)	SIGNAL FREQ (MHz)	SIG HAR	LO OSC FREQ (MHz)	LO HAR	ATTEN LEVEL (dB)
3500	38500	1	35000	1	63.3985
3550	38550	1	35000	1	59.4171
3600	38600	1	35000	1	55.1405
3650	38650	1	35000	1	50.5214
3700	38700	1	35000	1	45.5001
3750	38750	1	35000	1	40
3800	38800	1	35000	1	33.9199
3850	38850	1	35000	1	27.1229
3900	38900	1	35000	1	19.4171
3950	38950	1	35000	1	10.5214
4550	39550	1	35000	1	10.5214
4600	39600	1	35000	1	19.4171
4650	39650	1	35000	1	27.1229
4700	39700	1	35000	1	33.9199
4750	39750	1	35000	1	40
4800	39800	1	35000	1	45.5001
4850	39850	1	35000	1	50.5214
4900	39900	1	35000	1	55.1405
4950	39950	1	35000	1	59.4171
5000	40000	1	35000	1	63.3985

NOTES: Frequencies in MHz
 Levels in dB

TABLE 6.3
LOW FREQUENCY CONVERSION STAGE SPURIOUS ANALYSIS RESULTS

SPUR FREQ (MHz)	SIGNAL FREQ (MHz)	SIG HAR	LO OSC FREQ (MHz)	LO HAR	ATTEN LEVEL (dB)
1300	3700	1	5000	1	72.623
1250	3750	1	5000	1	63.3985
1200	3800	1	5000	1	53.3369
1150	3850	1	5000	1	42.2633
1100	3900	1	5000	1	29.9384
1050	3950	1	5000	1	16.0215
450	4550	1	5000	1	10.5214
400	4600	1	5000	1	19.4171
350	4650	1	5000	1	27.1229
300	4700	1	5000	1	33.9199
250	4750	1	5000	1	40
200	4800	1	5000	1	45.5001
150	4850	1	5000	1	50.5214
100	4900	1	5000	1	55.1405
50	4950	1	5000	1	59.4171
0	5000	1	5000	1	63.3985
50	5050	1	5000	1	67.1229
100	5100	1	5000	1	70.6214
150	5150	1	5000	1	73.9199
200	5200	1	5000	1	77.04
250	5250	1	5000	1	80
1200	4400	2	5000	2	77.4171
1150	4425	2	5000	2	73.1405
1100	4450	2	5000	2	68.5214
1050	4475	2	5000	2	63.5001
1000	4500	2	5000	2	58
950	4525	2	5000	2	63.5001
900	4550	2	5000	2	68.5214
850	4575	2	5000	2	73.1405
800	4600	2	5000	2	77.4171

NOTES: Frequencies in MHz
 Levels in dB

6.3.2 Transmitter Chain Detailed Design

The detailed design of the transmitter chain is based on the use of surface mount technology (SMT) in the low frequency stage and available MMIC chip and wire technology in the high frequency (millimetre-wave) stage. A block diagram of the specific devices implemented in the transmitter chain is shown in Figure 6.4. The use of SMT technology at the low frequencies allows for exploitation of low-cost devices. As frequency increases, the limitations on packaging performance do not allow for the use of SMT devices and therefore MMIC chip-and-wire technology is exploited. Both of these implementations allow for the use of conventional printed microstrip as a convenient interconnect mechanism. A high-Q filtering function is, however, not readily attained on microstrip. One region in the transmitter chain where this is an issue is down stream of the millimetre-wave mixer. It would be beneficial to place an RF filter before the transmitter amplifiers in order to optimize the RF power gain applied to the carrier of interest. Although it would be feasible to do this, added stages of RF gain would be required to overcome the losses associated (2-4 dB) with such a filter. The approach which is taken is to select a final power amplifier device with a bandpass gain characteristic which inherently suppresses the 35 GHz LO and RF image (~ 31 GHz) signals, thus avoiding the need for this filter.

The components applied to the realization of the transmitter chain are summarized in Annex A. With the exception of the transition from microstrip to waveguide and the RF bandpass filter, these components are commercially available devices or are custom devices manufactured using commercially available foundry processes¹.

The transition to waveguide¹ is required in order to convert the transmission line medium to one which allows a high-Q filter function to be realized. In this application a tapered E-plane transition from microstrip to wave guide is employed and is readily designed using a variety of approaches [3-8] depending on the geometry required. The design of this transition is detailed further in Annex B.

Note 1: The SPDT switch, the microstrip-to-waveguide transition and the RF BPF are components shared with the receiver chain.

FIG93/021

FIGURE 6.4 TRANSMITTER CHAIN COMPONENTS

The RF bandpass filter¹ is readily realized in printed form using E-plane circuitry in a unilateral finline configuration. This approach provides a high Q and good selectivity while retaining low in-band loss [9-11]. The technique described in [12] readily allows filters of this configuration and moderate bandwidths (up to ~ 15%) to be realized. The design of this bandpass filter is detailed further in Annex C.

The key attributes of the transmitter chain which must be investigated to ensure that the design will operate properly when realized are:

- In-band signal gain,
- In-band compression, and
- In-band output power.

Other issues related to operating frequency have been addressed through the frequency plan and spurious analysis and will not be reviewed here.

The above characteristics, combined with the general attributes of the selected/design components are addressed as shown in the spread sheet in Figure 6.5. The first-order analysis indicates that the signal gain is ~ 15.5 dB and that the only stage operating at saturation is the final power amplifier stage. This is an important factor since it is this stage that will dictate the overall efficiency of the transmitter. In addition the millimetre-wave mixer operates near compression in an effort to increase the RF power output level thus reducing the number of required RF gain stages. This is important due to the cost of the RF amplification.

Nominally, a first order analysis can be improved upon, however, the "typical" nature of the SMT and MMIC componentry manifests itself in characteristics which are not exact but are extremely repeatable once known. In addition higher order analysis requires detailed component data not generally available due to the variability of different applications.

The transmitter output power, which is the key attribute of this chain, is 45 mW which compares with the level employed in the user terminal performance assessment of Section 5.4 (+19 dBm and 2.5 dB losses provides ~ 45 mW of signal power).

Note 1: The SPDT switch, the microstrip-to-waveguide transition and the RF BPF are components shared with the receiver chain.

Notes

- 1: This is saturated power.
- 2: P1dB Out is the 1 dB compression point referenced to the component output.
- 3: P3IM Out is the third order intermodulation intercept point referenced to the component output.

FIG93/022

FIGURE 6.5 TRANSMITTER CHAIN IN-BAND SIGNAL FLOW SPREAD SHEET

6.3.3 Receiver Chain Detailed Design

A block diagram of the receiver chain components is shown in Figure 6.6. Incoming RF is bandpassed through the 38-40 GHz preselection filter and routed through the T/R switch to the receiver low noise RF amplifier (the RF filter and T/R switch are shared with the transmitter chain). The amplified RF is double downconverted to the .5 - 1 GHz IF through a two stage downconverter operating exactly opposite to the chain described in the transmitter chain.

In addition to the spurious analysis carried out in a previous section, the receiver chain is analyzed to determine the following primary characteristics:

- In-band gain,
- In-band noise figure and noise bandwidth reduction factors,
- In-band minimum signal power levels, and
- In-band thermal noise power levels.

This first order analysis has been carried out and is summarized in the spreadsheet shown in Figure 6.7. The components employed in the spreadsheet are summarized in Annex D. Signal power levels employed to establish the minimum and maximum received RF power are dictated by the 50 m link with and without 20 dB fade margin. As shown in the right column of the figure, an SNR of ~ 5 dB is possible during 20 dB fading. This is somewhat marginal when considering the effects which this SNR may have on BER. However, this is predetection SNR and does not include the effects of post-detection signal processing. In addition, the 20-dB fade margin is not a value based on exact measurement.

Verification of these SNR levels has been carried out using a commercial link analysis software tool [13] with similar results, as shown in Figure 6.8 and 6.9.

From Figures 6.8 and 6.9 one can readily derive the maximum RF signal power of concern. At 5 m range, the SNR is ~ 45 dB (no fading), and assuming a thermal noise power of ~ -81 dBm one can readily calculate that a maximum received signal power of -36 dBm might be expected. A suitable safety margin of 16 dB is considered such that the absolute maximum level would be on the order of -20 dBm.

FIG83/044

FIGURE 6.6 RECEIVER CHAIN COMPONENTS

FIG93/046

FIGURE 6.7 RECEIVER CHAIN IN-BAND SIGNAL FLOW SPREADSHEET

TRANSMITTER POWER = 45 mW
 TRANSMIT ANTENNA GAIN = 2 dB
 RECEIVE ANTENNA GAIN = 25 dB
 RECEIVE ANTENNA OHMIC EFFICIENCY = 95 %
 PHYSICAL RECEIVE ANTENNA TEMPERATURE = 25 DEG C
 FREQUENCY = 39 GHz
 EXCESS (ABSORPTIVE) PATH LOSS = 0 dB
 BACKGROUND TEMPERATURE = 300 DEGREES K
 ANTENNA-TO-RECEIVER FEEDLINE LOSS = 0 dB
 ANTENNA-TO-RECEIVER FEEDLINE TEMPERATURE = 25 DEG C
 RECEIVER NOISE FIGURE = 8 dB
 SYSTEM NOISE BANDWIDTH = 500 MHz

RANGE (METERS)	S/N ₀ (dBc-Hz)	S/N (dB)
5.000	131.242	44.252
10.000	125.221	38.232
15.000	121.700	34.710
20.000	119.201	32.211
25.000	117.253	30.273
30.000	115.679	28.689
35.000	114.340	27.350
40.000	113.180	26.190
45.000	112.157	25.167
50.000	111.242	24.252
55.000	110.414	23.424
60.000	109.658	22.689
65.000	108.963	21.973
70.000	108.319	21.330
75.000	107.720	20.730
80.000	107.160	20.170
85.000	106.633	19.643
90.000	106.137	19.147
95.000	105.667	18.677
100.000	105.221	18.232

**FIGURE 6.8 ESTIMATED SNR FOR AN OMNI-TO-DIRECTIONAL (USER)
 LINK - NO FADE MARGIN CASE**

TRANSMITTER POWER = 45 mW
 TRANSMIT ANTENNA GAIN = 2 dB
 RECEIVE ANTENNA GAIN = 25 dB
 RECEIVE ANTENNA OHMIC EFFICIENCY = 95 %
 PHYSICAL RECEIVE ANTENNA TEMPERATURE = 25 DEG C
 FREQUENCY = 39 GHz
 EXCESS (ABSORPTIVE) PATH LOSS = 20 dB
 BACKGROUND TEMPERATURE = 300 DEGREES K
 ANTENNA-TO-RECEIVER FEEDLINE LOSS = 0 dB
 ANTENNA-TO-RECEIVER FEEDLINE TEMPERATURE = 25 DEG C
 RECEIVER NOISE FIGURE = 8 dB
 SYSTEM NOISE BANDWIDTH = 500 MHz

RANGE (METERS)	S/N ₀ (dBc-Hz)	S/N (dB)
5.000	111.242	24.252
10.000	105.221	18.232
15.000	101.700	14.710
20.000	99.201	12.211
25.000	97.263	10.273
30.000	95.679	8.689
35.000	94.340	7.350
40.000	93.180	6.190
45.000	92.157	5.167
50.000	91.242	4.252
55.000	90.414	3.424
60.000	89.658	2.669
65.000	88.963	1.973
70.000	88.319	1.330
75.000	87.720	0.730
80.000	87.160	0.170
85.000	86.633	-0.357
90.000	86.137	-0.853
95.000	85.667	-1.323
100.000	85.221	-1.768

FIGURE 6.9 ESTIMATED SNR FOR AN OMNI-TO-DIRECTIONAL (USER)
 LINK - 20 dB FADE MARGIN CASE

This power level is employed to carry out a simple investigation to ensure that none of the sensitive low noise amplifier stages is compressing prematurely².

The spreadsheet summary of maximum received signal powers (referenced to the component outputs) is shown in Figure 6.10. In conjunction with the signal power levels, the component output compression points are also included in the spreadsheet. No compression-related limitations are envisioned.

A secondary factor in the design is protection of the receiver from damaging RF power emitted by external sources or from the transceiver transmitter leakage. Operating within the 39 GHz portion of the spectrum readily provides isolation from other signals and thus interference will tend to take the form of other 39 GHz microcell transmitters operating with the same transmitter powers. The maximum safe RF power (CW) which can be sustained by the receiver is ~ 17.5 dBm which is higher than the designed transmitter power. Hence it is essentially impossible to couple sufficient power into the receiver to cause damage. "Own-transmitter" leakage into the receiver chain will also not cause receiver damage due to the high (> 30 dB) isolation in the T/R switch.

6.3.4 High Frequency LO Detailed Design

The high frequency LO source design is shown in Figure 6.11. The LO is realized in chip-and-wire MMIC form using a microstrip interconnect. Components employed in this chain are summarized in Annex E.

In order to implement re-use, an externally-controlled voltage is employed on the VCO to facilitate changes in the transceiver centre frequency of ~ 500 MHz (typically the available tuning bandwidth is ~ 1 GHz). This tuning bandwidth must be accommodated by all components between the RF mixer and the antenna. Generally this is not a problem except in the case of the RF band pass filter which must be opened up in pass band width to allow for the variation in operating frequency. In addition, the VCO is provided with a second centre

NOTE 2: This maximum power level was also employed in previous calculations of spurious signal generation within the mixing stages employed.

FIG93/045

FIGURE 6.10 SPREADSHEET SUMMARY OF MAXIMUM IN-BAND SIGNAL AMPLITUDE

FIG93/041

FIGURE 6.11 HIGH FREQUENCY LO COMPONENT CHAIN

frequency adjustment point to facilitate thermal stabilization through the use of thermistor-based temperature tracking circuitry (not pursued in the scope of this thesis effort) needed to ensure that the operating centre frequency remain stable.

The output power of the oscillator source is $\sim +10$ dBm nominal across the tuning bandwidth. This is usually sufficient power for driving the unbiased single-balanced mixers employed in receivers but is usually insufficient to drive upconverting (transmitter) mixers. A driver amplifier is employed to amplify the LO signal to $\sim +15$ dBm in order to improve upconversion efficiency while not providing serious noise increases in the downconverting mixer. The driver amplifier also provides isolation of the local oscillator from load pulling effects [2] associated with mixer LO port loading and transient loading associated with switching the LO between the receiver and transmitter mixers.

It is expected that the quality factor (Q) of this oscillator will be relatively poor due to the achievable quality factor of the on-chip resonant circuitry. This poor quality factor may partially dictate the usefulness of various modulation and multipath suppression schemes which might be employed in a wireless WLAN architecture. The quality of the high frequency LO signal will be the subject of further investigative efforts based on measured quality indicators such as phase noise performance.

6.3.5 Low Frequency LO Detailed Design

The low frequency LO is a fixed frequency design feeding a transmit/receive select switch. The block diagram of this LO chain is shown in Figure 6.12. The LO is realized using low cost SMT components employing a microstrip interconnect. The selected components are summarized in Annex F.

The LO source employs a varactor-controlled thermal compensation facility (not exploited in this thesis effort) and provides sufficient RF output power to drive either the up or downconverting mixers. The LO source employs an internal buffer amplifier to reduce load-pulling effects on the oscillator, therefore no external amplifiers are required (as in the case of the millimetre-wave LO).

FIG93/042

FIGURE 6.12 LOW FREQUENCY LO COMPONENT CHAIN

Overall stability of this oscillator is expected to provide single-sideband (SSB) phase noise levels of ~ -100 dBc/Hz @ 100 KHz offset. This moderate performance level is deemed to be acceptable when considering the cost of high-quality oscillators (this is intended to be a low-cost GaAs MMIC unit) and the fact that the lower performance of the millimetre-wave LO will dominate the amount of signal distortion caused by phase-noise effects.

6.4 Transceiver Expected Performance

The expected performance attributes of the transceiver are summarized and compared with the required performance in Table 6.4.

6.5 Implementation of the Transceiver

The preliminary implementation design of the transceiver module focuses on the following factors:

- Control circuitry,
- Power supply design, and
- Mechanical design.

6.5.1 Control Circuitry

Control circuitry for this type of module generally takes the form of the one of the following:

- Control interface for RF switches,
- Control interface for local oscillators, and
- Internal thermal compensation circuitry.

These factors are deemed to be outside the scope of this thesis research and are hence not pursued any further. The transceiver implementation will employ a "pin" connection for these control functions to be accessed where required. No special attention will be paid to the standardization of voltages, connectors etc.

TABLE 6.4

**EXPECTED vs REQUIRED PERFORMANCE LEVELS
OF THE TRANSCEIVER DESIGN**

FACTOR	UNITS	NOMINAL REQUIREMENT	EXPECTED PERFORMANCE	NOTES
1. IF input frequency	GHz	0.5 to 1	0.5 to 1	
2. IF input power	dBm	0 to -15	-5 to +5	Manual gain control
3. RF output frequency	GHz	38 - 40	39 - 40	
4. RF output power	dBm	> +15	+16.5	
5. Rx/Tx switching speed	ns	<10	<10	
6. Rx noise figure	dB	8	6.5	At antenna/transceiver interface
7. Operating freq. tunability	MHz	>500	500	
8. IF output frequency	GHz	0.5 to 1	0.5 to 1	
9. IF output spurious, LO leakage, image signals levels	dBc	>40	>60	RF in at -25 dBm
10. Linear IF output dynamic range	dBm	-70 to -25	-70 to -20	Function of range to transmitter
11. Limited IF output	dBm	-15	N/A	
12. Manual gain output	V	TBD	0 to -5	External
13. Tx/Rx switching control	V	TBD	0 to -5	External
14. Operating frequency control	V	TBD	0 to +10	External
15. IF in/out, RF port VSWR	Ratio	2 : 1	2 : 1	

6.5.2 Power Supply Circuitry

Power supply design falls largely into the same category as that described for the control circuitry. The key limitations in the design of the power supply are:

- Voltage settings are often critical in achieving a particular module performance,
- Size is to be minimized to allow for field use,
- Voltages often must be sequenced to prevent device damage, and
- Supplies must be filtered and spike protected to prevent device damage.

These factors are deemed to be outside the scope of this thesis research and are hence not pursued any further.

6.5.3 Proof-of-concept Transceiver Test Housing Mechanical Design

The mechanical design of the transceiver test circuit housing requires a great deal of attention to the sizes of cavities, exact device placements, routing of power supply lines, etc. These considerations form the basis for the effort needed to design suitable test housings for the proposed circuitry. The detailed design of the test housings for the transceiver forms one of the steps in the build and test activity outlined in a subsequent section of this thesis.

A critical factor in the design of the transceiver is the preliminary layout of the module. The layout must account for the following factors:

- Testability of built-up circuit sections,
- Device orientations,
- Access to required power supply, and
- Signal routing.

The testability of the transceiver blocks is facilitated through the use of MMIC and MIC test fixtures which allow testing of carrier-mounted devices without the need for a test housing. These are currently commercially available operating to 60 GHz.

Device orientations are of concern since access to the device ports must be accomplished in a uni-planar realization. This limitation is particularly important when considering high frequency operation (into millimetre-wave) where multi-layering or trace-jumping is not possible.

Access to the required power supplies is based on the desire to avoid DC cabling inside the RF cavities of the various MIC and MMIC circuits. The cabling can provide circuit problems such as cavity resonances or can be the source of circuit damage when inadvertently pushed onto small circuit bond wires or ribbons.

Signal routing issues can create problems similar in nature to those outlined for device orientation. Complex routing designs are generally not feasible at millimetre-wave unless uni-planar circuit structures are pursued.

A preliminary mechanical layout for the proof-of-concept RF transceiver test housing/circuit is shown in Figure 6-13.

The layout employs a WR-19 waveguide section which interfaces to the terminal antenna. The RF bandpass filter and waveguide-to-microstrip transition are realized using split block printed E-plane techniques. The MMIC devices are grouped together on three carrier assemblies which are suitable for individual testing on a universal test fixture. The IF (4 - 4.5 GHz) is interfaced on to a microstrip SMT circuit which houses the low-frequency components. The 0.5 - 1 GHz input and output are accessed via standard SMA connectors; these also provide a simple transition to microstrip through the housing wall.

The design of the RF MMIC circuitry readily allows testing of built-up assemblies in order to verify their individual operation. The SMT circuitry, however, does not allow this feature. Suitable test circuitry may be required to assess the operation of these components in combination to ensure proper operation.

DC and control signals are fed into the transceiver through a multipin connector. The power supply provides the various voltages required by the circuit components. Provision of wiring interconnects needed for power/control connection to the specific components is facilitated by threading wiring along the backside of the assembly to the respective circuit elements.

FIGURE 6.13 TRANSCEIVER TEST CIRCUIT/HOUSING PRELIMINARY MECHANICAL LAYOUT

6.6 References for Chapter 6

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7.0 PROOF-OF-CONCEPT TRANSCEIVER EVALUATION

The transceiver concept design (Chapter 6) is taken as far as a proof-of-concept model implementation which is employed to derive laboratory performance results which can be compared with theoretical predictions. Differences between predicted and measured performance can then be investigated to establish their respective causes. At the completion of this exercise, the transceiver evaluation will provide insight into the limitations of the technological approach taken to realize cost-effective millimetre-wave WLAN transceivers. Some of these limitations can be discussed with respect to impacts on WLAN systems operating at millimetre-wave frequencies.

In order to construct the evaluation transceiver, the transceiver circuitry (outlined in Chapter 6) must be broken down and allocated to mechanical carriers which can be independently built up, tested and then integrated together into the final transceiver. The "carrier" serves as a physical partition of the transceiver. Figure 7.1 shows the partitioning of the circuitry into the respective carriers. Care must be taken to construct manageable (in complexity) carriers which balance the reduction in interconnection losses with an acceptable risk in fabrication damage/uncertainty. This is particularly true when considering a first prototype. Later versions of the same design may indeed migrate to full-complexity circuit realizations, using chip-and-wire interconnection on a single carrier. This would eliminate losses associated with circuit interconnection (alumina 50-ohm lines in this case).

As shown in the breakdown of Figure 7.1, the following carrier assemblies were designed, constructed and tested:

- Filter/transition carrier,
- T/R switch carrier,
- Amplifier carrier,
- Converter carrier,
- LO carrier, and
- IF carrier.

FIG078.CDR

FIGURE 7.1 PARTITIONING OF THE TRANSCEIVER INTO PHYSICAL "MODULE" BUILDING BLOCKS (simplified block diagram shown)

The following sections detail the design, evaluation plan, test set-ups, testing limitations, measured data and measured versus expected performance for each carrier.

Once all of these carrier building blocks have been completed they are assembled into the final transceiver test housing and the performance of the integrated assembly is investigated.

7.1 Filter/Transition Module

This module consists of the finline bandpass filter and the antipodal finline transition from the waveguide E-plane on to microstrip. This transition is necessary in order to employ the required planar circuitry needed for MMIC mounting and interconnection. The preselection bandpass filter is realized in finline due to the lower loss potentially available. The filter is required to reject image and local oscillator outputs from the transmitter and to prevent the reception of undesired out-of-band signals (including the RF image) which will distort the transceiver's response to the desired signal (in the receive mode).

The test circuit used to evaluate the performance of these circuits is shown in Figure 7.2. The test housing is exactly the same as the final transceiver housing configuration. The finline transition and filter circuits are constructed from .005" thick soft Duroid substrate such that effective compression can be achieved in the clamping area. The waveguide split-block housing (upper half removed in Figure 7.2) is machined such that the clamping area is electrically very small and the depth is such that the substrate is squeezed slightly to ensure a good RF ground connection to the waveguide, to keep the circuit physically in place, and to prevent any undesired propagation effects in the clamping region.

The tested attributes of this circuit module are:

- Scalar insertion loss,
- Scalar return loss, and
- Frequency selectivity characteristics.

FIG024.CDR

FIGURE 7.2 FILTER/TRANSITION MODULE TEST CIRCUIT CONFIGURATION

The measurement of these characteristics is accomplished with the scalar network analyzer set-up shown in Figure 7.3.

The test circuit employs a waveguide input and a planar output. This combination prevents truly accurate calibration of the test set-up. Accurate measurement requires the calibration of the through path using "like" transmission structures. In the case of the test circuitry employed, this is not possible. A calibrated coaxial transition (loss of 2.5 dB at 39 GHz) is employed after the WR-28 waveguide calibration process.

Initially the filter center frequency was ~ 0.5 GHz higher than expected. Tuning this type of filter structure is relatively straightforward and is accomplished by cutting the resonant sections apart and changing their separation.

The results of the measurement using the final filter configuration are shown in Figure 7.4. The in-band insertion loss is ~ 3.5 dB and the return loss is better than 10 dB in the band of interest. Rejection of the LO (at 35 GHz) is 37 dB and image rejection is 35 dB. Due to the fact that this was somewhat higher than expected, the finline bandpass filter and antipodal finline-to-microstrip transition were independently investigated.

In order to investigate the characteristics of the filter, it was tested in isolation in a split-block E-plane fixture as shown in Figure 7.5. The filter was tested using the Ka-band test set-up and the 40 - 60 GHz test set-up as shown in Figure 7.6. The insertion loss of the filter in the 30 - 50 GHz range is shown in Figure 7.7. The measured high-side rejection is 25 - 30 dB. This level of rejection is somewhat less than expected. An investigation into the test set-up revealed that the tripler source is limited to ~ 0 dBm and that the detectors provide ~ -40 dB sensitivity. When testing using 10 dB couplers, the available dynamic range is ~ 30 dB and therefore ultimately limits the measured rejection. Additionally, the tripler source produces a significant second-harmonic component which propagates through the WR-19 waveguide if it is above cut-off. The filter rejects this signal (in addition to its suppression with respect to the tripled output) and is not expected to be a contributing factor under the circumstances. Generally, the filter provides insertion loss and rejection performance which is comparable to what was expected.

FIG019.CDR

FIGURE 7.3 - SCALAR Ka-BAND TEST SET-UP

FIGURE 7.4 MEASURED INSERTION AND RETURN LOSS OF THE
TRANSITION/FILTER MODULE

FIG026.CDR

FIGURE 7.5 FILTER IN SPLIT BLOCK TEST FIXTURE

FIG020.CDR

FIGURE 7.6 40 - 60 GHz SCALAR NETWORK ANALYZER TEST SET-UP

FIGURE 7.7A RF FILTER CHARACTERISTICS IN THE 30 - 40 GHz RANGE

FIGURE 7.7B RF FILTER CHARACTERISTICS IN THE 40 - 50 GHz RANGE

Back-to-back tapers were constructed and mounted in a split-block test jig (as shown in Figure 7.8) to determine whether the insertion loss was higher than expected. The tapers were tested in the Ka-band scalar network analyzer test set-up; the results are shown in Figure 7.9. These results show that the single-taper loss is ~ 1.5 dB which is 1 dB higher than expected. When considering the longer 50-ohm line employed in the transition/filter test circuit, it is evident that use of this transition in the transceiver would incur ~ 2 dB loss.

The added insertion loss of the transition/filter module will result in higher receiver noise figure and reduced available transmitter output power. Both of these commodities are rare in this frequency range. Further work would be required to minimize the loss of the transition.

7.2 T/R Module

The T/R module is a high speed switch which is switched between the transmit and receive paths. This is done in synchronism with switching the first and second local oscillators as well as the RF/data streams. The implementation of this module is shown in Figure 7.10. The MMIC single-pole, double-throw (SPDT) chip is mounted using standard hybrid assembly techniques and is tested in a universal test-figure. The module carrier measures 0.200" x 0.800". The key attributes which were investigated are:

- Insertion loss,
- Return loss, and
- Isolation.

G025.CDR

FIGURE 7.8 BACK-TO-BACK ANTIPODAL FINLINE TAPERS MOUNTED IN A SPLIT-BLOCK TEST HOUSING

**FIGURE 7.9 LOSS CHARACTERISTIC OF THE BACK-TO-BACK ANTI-
PODAL FINLINE-TO-MICROSTRIP TRANSITIONS**

FIG012.CDR

FIGURE 7.10 T/R MODULE IMPLEMENTATION

The switches do not cause output power compression until the RF input power reaches $\sim +23$ dBm. As a result, testing under small-signal conditions was deemed adequate. The test set-up of Figure 7.3 was used to accomplish this. Figure 7.11 shows the insertion loss and isolation of the module and Figure 7.12 shows the measured return loss while switched to both states. The measurements agree with predictions with the exception of insertion loss which was expected to be ~ 1 dB. The measured loss of 3 dB will further increase the receiver noise figure and reduce the available transmitter power. No apparent reason for this could be found from inspection of the module assembly or from careful examination of the diode I/V characteristics. Discussions with the device foundry also proved futile. A second switch was constructed to investigate the possibility of an error in handling/or assembly. Virtually identical results were obtained. A suggestion from the foundry was to employ chip-and-wire techniques rather than microstrip interconnects so as to achieve performance closer to their probed on-wafer measurements. In follow-on developments, as the circuitry becomes better understood, it will be possible to migrate to the chip-and-wire technique (as is normally the case). The ultimate cost of this approach would warrant such measures or even the design of a complete multi-function ASIC.

7.3 LNA/Power Amplifier Module

The purpose of this module is to provide the RF low-noise amplification (LNA) function (receive mode) as well as the power amplifier function (transmit mode). The constructed module is shown in Figure 7.13. The LNA strip is constructed from two amplifier stages and their associated bias circuitry. The power amplifier consists of a driver amplifier and power amplifier, together with associated support circuitry. The module carrier measures .800" x .800" and is assembled with standard hybrid assembly techniques.

Measurements on the LNA and power amplifier are treated separately in the subsequent sections.

FIGURE 7.11 T/R MODULE INSERTION LOSS AND ISOLATION PERFORMANCE

FIGURE 7.12 T/R MODULE RETURN LOSS PERFORMANCE IN BOTH SWITCHING STATES

FIG011.CDR

FIGURE 7.13 LNA/POWER AMPLIFIER MODULE LAYOUT

7.3.1 LNA

The LNA strip on this module was tested for the following attributes:

- Gain,
- Return loss,
- Compression characteristic, and
- Noise figure.

Noise figure testing was not possible due the failure of the noise source required for this testing. The noise figure is evaluated in a subsequent section that addresses the completed transceiver module (see receive mode of operation).

The scalar network analyzer set-up shown in Figure 7.3 was employed to evaluate the small-signal gain and return loss. The results of this measurement are shown in Figure 7.14. The measured gain is ~ 4 dB lower than predicted and the return loss indicates that some kind of resonance has been inadvertently excited. Isolation of 35 - 40 dB is achieved. The compression characteristics of the LNA were determined by measuring the gain under various input powers, as shown in Figure 7.15. It is seen that 1 dB gain compression occurs at ~ +5 dBm (assuming that the gain at -20 dBm is in the linear region).

Due to the low gain and the input return loss characteristics the LNA circuit was modified to correct what appeared to be a damaged on-chip inductor. This resulted in better overall LNA performance, as shown in Figure 7.16. The achieved gain is very close to the 14 dB expected for this module.

7.3.2 Power Amplifier

The power amplifier portion of this module was tested for the following attributes:

- Scalar gain,
- Scalar input return loss,
- Gain compression, and
- Mid-band output power.

FIGURE 7.14 - MEASURED LNA GAIN AND RETURN LOSS (RF AT -10 dBm)

FIGURE 7.15 MEASURED LNA GAIN UNDER VARIOUS INPUT POWER CONDITIONS

FIGURE 7.16 MEASURED LNA GAIN AND RETURN LOSS AFTER PREPARATION OF ON-CHIP BIAS CIRCUITRY

The gain, return loss, and compression of the power amplifier were measured using the test set-up shown in Figure 7.3. The measured gain and return loss of the power amplifier are shown in Figure 7.17. The measured gain of $\sim +19$ dB at mid band is 4 dB lower than expected. Since the transmitter path through the transceiver is approximately linear, added input signal power can be provided to offset this effect (provided that no other devices in the chain compress). The amplifier provides better than 10 dB suppression at the RF image in the 31 GHz range and 8 dB of suppression at the 35 GHz LO mid-band range.

The gain of the amplifier was measured under various input drive levels in order to determine the compression characteristic. As shown in Figure 7.18, the 1 dB compression point is ~ -2 dBm (interpolating between the 0 and -5 dBm responses). At a drive level of +5 dBm, the amplifier is well into compression. Using a calibrated downconverter and spectrum analyzer (in place of the final detector in the test set-up of Figure 7.3) it was possible to measure the mid-band maximum output power available from the power amplifier. Figure 7.19 shows a maximum available signal power of $\sim +16$ dBm. Close-in sidebands were identified as originating in the test set-up. The output power level of the amplifier is 3 dB lower than expected. No explanation for this was available, so a second unit was constructed. It also produced +16 dBm. Discussions with the foundry indicated that their data was "typical" data only and is measured using on-wafer probing techniques.

7.4 Up/Downconverter Module

The up/downconverter module consists of a pair of single-balanced Schottky diode mixers, see Figure 7.20. The mixers are identical and are simply operated in reverse. As in the case of the other MMIC-based modules, standard hybrid assembly techniques are employed in the module construction. The module carrier measures .600" x .800". The upconverter and downconverter portions of this module are treated separately in the subsequent sections.

FIGURE 7.17 MEASURED POWER AMPLIFIER GAIN AND RETURN LOSS

FIGURE 7.18 MEASURED POWER AMPLIFIER COMPRESSION CHARACTERISTIC

FIGURE 7.19 MEASURED POWER AMPLIFIER MAXIMUM OUTPUT SIGNAL

FIG013.CDR

FIGURE 7.20 UP/DOWNCONVERTER MODULE LAYOUT

7.4.1 Upconverter

The upconverter was tested to determine the following attributes:

- Conversion loss,
- LO-to-RF isolation,
- LO and IF port return losses, and
- Output spurious and image suppression.

These attributes of the module were measured using the test set-up shown in Figure 7.21. Mixer testing generally involves avoiding the effects of spurious signals. In this case, a spectrum analyzer is employed to ensure that the wideband detectors are in fact measuring the signal of interest.

The conversion loss (power input - power output) of the mixer at three LO frequencies (34.5, 35, and 35.5 GHz) is derived from data shown in Figure 7.22. This data is taken through a 10 dB coupler (measured to be 9 dB) employing a calibrated coaxial IF launcher with 2.5 dB loss. The resulting upconversion loss is ~ 10 dB and remains very constant across the LO tuning band. This loss is high compared to the expected 7 dB level. Investigation of this difference revealed that the 7 dB level was not a probed measurement taken on-wafer by the foundry but a predicted value based on mixer performance with a 500 MHz IF input. Additionally (as noted in Figure 7.26) image signals produced by the mixer add to the measured output power as a result of the use of wideband detectors (resulting error = 1.8 dB max., based on -3 dB spurious levels).

The increased loss incurred in the mixer can be readily overcome through added gain in the IF drive circuitry provided that the mixer does not compress. In this circuit, the driver and power amplifiers saturate at ~ -2 dBm input power. The required IF input to the mixer is therefore +7 dBm.

The LO-to-IF isolation is shown in Figure 7.23. The LO leakage must be suitably suppressed so that it does not compete with the desired RF signal in the power amplifier causing premature saturation and added output spurious. The measured LO attenuation is ~ 37 dB. The LO will typically be 20 dB below the desired RF signal.

FIG022.CDR

FIGURE 7.21 UPCONVERTER TEST SET-UP

OUTPUT POWER

(Pin = 0 dBm, measured through an 11.5 dB coupling loss)

CH1:	A	-	21.49	dBm	
10.0	dB/	REF	+	.00	dBm

FIGURE 7.22 UPCONVERSION LOSS (LO at +10 dBm, IF input at 0 dBm)

FIGURE 7.23 LO LEAKAGE AT THE RF PORT OF THE UPCONVERTER
(LO at +10 dBm)

INPUT RETURN LOSS

CH1:	B	-M	-	7.73 dB
10.0 dB/	REF	-	-	.00 dB

FIGURE 7.24 UPCONVERTER IF PORT RETURN LOSS
 (35 GHz LO at +10 dBm, IF at +5 dBm)

FIGURE 7.25 REFLECTED LO SIGNAL (INCIDENT LO at 35 GHz, +10 dBm)

The IF return loss is shown in Figure 7.24 and the reflected LO signal, as measured through a high directivity (>30 dB) coupler is shown in Figure 7.25. The reflected LO signal is ~ 24 dB down from the incident signal indicating a good LO power transfer to the mixer diodes.

The spurious output signals produced by the upconverter are difficult to measure since the spectrum analyzer is not preselected and is a downconversion unit yielding a large number of spurious signals in itself. For particular (known) tones, one can employ the analyzer to provide reasonably accurate power measurements on a frequency selective basis. Figure 7.26 shows the output spectrum of the upconverter. In the centre of this figure one can see the LO leakage at 35 GHz and the tone on the far left is the image signal at 30.750 GHz. The tone second from the right is the upconverted RF signal at 39.250 GHz produced from a mid-band IF at 4.250 GHz. The remaining tones are spurious signals produced by the spectrum analyzer itself (partially identified using the "SIGNAL IDENT" function within the analyzer).

From the figure it can be seen that both the desired RF signal and image are produced as would be expected from a single-balanced mixer. The image is suppressed by ~ 3 dB due only to the fact that mixer die has been optimized for 39 GHz operation (rat race construction). This degree of image suppression requires the power amplifier to have a suppression characteristic in order to prevent the image from being amplified.

7.4.2 Downconverter

The downconverter was tested to determine the following attributes:

- Conversion loss,
- RF port return losses (LO return loss was tested in the upconverter portion of this module), and
- Output spurious and image suppression.

The test set-up employed for downconverter testing is shown in Figure 7.27.

The measured conversion loss (power input - power output) is derived from Figure 7.28. The loss (2.5 dB) of the calibrated coaxial probe is not accounted for in these measurements. The resultant conversion loss is 11 - 12 dB. As in the case of the upconverter, this loss is higher than predicted, for similar reasons.

FIGURE 7.26 UPCONVERTER OUTPUT SPECTRUM (LO at 35 GHz, +10 dBm)

FIG023.CDR

FIGURE 7.27 DOWNCONVERTER TEST SET-UP

OUTPUT POWER

(Pin = 0 dBm, measured through a 2.5 dB coupling loss)

FIGURE 7.28A DOWNCONVERTER CONVERSION LOSS
(LO at 34.5 GHz/+10 dBm, RF at 0 dBm)

OUTPUT POWER

(Pin = 0 dBm, measured through a 2.5 dB coupling loss)

FIGURE 7.28B DOWNCONVERTER CONVERSION LOSS
(LO at 35 GHz/+10 dBm, RF at -0 dBm)

OUTPUT POWER

(Pin = 0 dBm, measured through a 2.5 dB coupling loss)

CH1: MEM	-	15.62 dBm	-
10.0 dB/	REF	+	.00 dBm

FIGURE 7.28C DOWNCONVERTER CONVERSION LOSS
(LO at 35.5 GHz/+10 dBm, RF at .0 dBm)

The RF input return loss was measured and the results are shown in Figure 7.29. The RF return loss of Schottky diode mixers is generally poor due to the impedance characteristics of the diodes and the desire to achieve optimized LO matches. The mixer exhibits a return loss of only 3 - 5 dB over the central portion of the input band. This may cause stability problems when being driven by the LNA module.

The spurious signals of interest residing near the IF output portion of the spectrum (4.0 - 4.5 GHz) are shown in Figure 7.30. This figure shows the output spectrum of the downconverter across a 2.5 GHz portion of the spectrum centred at 4.25 GHz (IF midband point). No other signals are observed within ~ 45 dB of the desired signal.

7.5 Local Oscillator Module

The VCO module is constructed as shown in the carrier layout of Figure 7.31 (.800" X .600") and was assembled using standard hybrid assembly techniques. The module includes a driver amplifier to ensure adequate LO drive and to isolate the LO from the loading effects caused by the transmit-receive switching (the LO is shared between transmit and receive paths). The single pole double throw (SPDT) switch is activated in synchronism with the T/R switch such that the LO drive signal "de-activates" the respective mixer, thereby increasing isolation. The switch is common in configuration with the T/R switch and its characteristics (speed) would be similar.

The following characteristics were measured on the VCO module:

- Output power versus tuning frequency,
- Phase noise, and
- Tuning voltage characteristic.

The test set-up employed for testing the VCO is shown in Figure 7.32.

The output power characteristic versus tuned operating frequency is shown in Figure 7.33. The measured data show a stable LO power characteristic which is well in excess of the +10 dBm nominally required for fully-driven single balanced mixers. The power flatness across frequency is ± 0.5 dB.

FIGURE 7.29 DOWNCONVERTER RF PORT RETURN
(LO at 35 GHz/+10 dBm, RF at -10 dBm)

FIGURE 7.30A DOWNCONVERTER OUTPUT SPURIOUS
 (LO at 35 GHz/+10 dBm, RF at 39 GHz/-5 dBm)

FIGURE 7.30B DOWNCONVERTER OUTPUT SPURIOUS
 (LO at 35 GHz/+10 dBm, RF at 39.5 GHz/-5 dBm)

FIG010.CDR

FIGURE 7.31 VCO MODULE LAYOUT

FIG021.CDR

FIGURE 7.32 VCO MODULE TEST SET-UP

FIGURE 7.33 VCO MODULE POWER OUTPUT VERSUS FREQUENCY

The phase noise characteristic was measured using spectrum analyzer techniques (assuming that the phase noise characteristic of the spectrum analyzer is good compared to that of the VCO) and is shown in Figure 7.34. At a 1 MHz offset, the phase noise is ~ 90 dBc in a 1 Hz bandwidth. This performance level is suitable for simple transmission/modulation applications.

The VCO tuning voltage characteristic was measured using a frequency counter and is shown in Figure 7.35. The curve shows a very linear response and tuning range well in excess of the 1 GHz minimum desired. This tuning bandwidth is employed to effect the frequency re-use tuning of the transceiver.

7.6 IF Module

The IF module consists of IF circuits which both upconvert and downconvert signals to and from the respective operating bands. A simplified block diagram outlining the IF module is shown in Figure 7.36. The IF converter is assembled using standard quasi-surface mount techniques onto a .010" thick soft-substrate printed interconnect circuit.

The design of the IF converter follows the frequency plan derived in an earlier section.

7.6.1 IF Downconverter Strip

The testing of the downconverter portion of the IF converter focused on the following attributes:

- Swept gain and input return loss, and
- Output spurious.

In order to test these attributes, the test set-up shown in Figure 7.36 was employed. In all tests, LO power was +12 dBm.

Measurement of swept gain is shown in Figure 7.38. The input RF is swept (at -10 dBm) across the 3 - 5.5 GHz range, clearly showing the design passband from 4 - 4.5 GHz. The band edges are primarily defined by the IF bandpass filter. The forward gain of the module is close to the expected (+2 dB) level and flatness is approximately 1 dB.

FIGURE 7.34

VCO PHASE NOISE CHARACTERISTIC
 (LO tuned to 35 GHz)

FIGURE 7.35 VCO MODULE: TUNING VOLTAGE vs FREQUENCY CHARACTERISTIC

FIG073.CDR

FIGURE 7.36 IF MODULE BLOCK DIAGRAM

FIG072.CDR

FIGURE 7.37 IF CONVERTER MODULE TEST SET-UP

FIGURE 7.38 IF DOWNCONVERTER SWEPT SCALAR GAIN AND INPUT RETURN LOSS CHARACTERISTIC

The output spurious signals were investigated by stepping the incident RF across the 4 - 4.5 GHz passband and employing the spectrum analyzer to measure spurious outputs. The results of this measurement are shown in Figure 7.39.

The output spectrum shows two in-band spurious products of concern. The first, at ~ 500 MHz is a subharmonic of the 1 GHz tone, which is the largest signal towards the center. This component is suppressed by 50 dBc. It is not clear whether this product results from the test set-up or from the module itself. Investigation of the RF signal generator being used as a signal source indicated that a number of spurious signals were being produced. The second spurious signal located at 1 GHz and is probably a harmonic of the 500 MHz tone, which is the largest signal towards the left. In this case it is suppressed by 42 dBc.

The spurious outputs, being harmonically related to the IF signals, are not unexpected due to the octave bandwidth selected for the final IF band. A narrower operating bandwidth (i.e. for transmission of 155 Mb/s) could easily be configured to avoid/filter these spurious signals.

7.6.2 IF Upconverter Strip

The upconverter portion of the IF module was tested for similar attributes as the downconverter (with the exception that the IF input signal was a fixed-power large signal). The following characteristics were investigated:

- Scalar gain and input return loss, and
- Spurious output performance.

The test set-up used is shown in Figure 7.37. (A Fluke generator was used in this case), the LO power was +12 dBm.

Swept scalar gain and return loss are shown in Figure 7.40. As expected, the input return loss of the upconverter is somewhat worse than that of the downconverter. This is due to the mixer IF input-port characteristic. The gain agrees with predicted levels (+2 dB) and has a flatness of approximately ± 1 dB.

FIGURE 7.39 IF DOWNCONVERTER OUTPUT SPURIOUS PERFORMANCE
 (RF stepped in 100 MHz increments, -10 dBm)

FIGURE 7.40 IF UPCONVERTER SWEPT GAIN AND INPUT RETURN LOSS
(IF input at +12 dBm)

Output spurious was more difficult to measure due to the number of spectral lines being generated. Because of this, cumulative outputs (for inputs across the band of interest) could not be superimposed. Testing was carried out in a similar fashion as for the downconverter. Input signals were stepped in 100 MHz increments across the 0.5 - 1.0 GHz band and the output spectrum was recorded. Two of these output plots are shown in Figure 7.41. In Figure 7.41A, the input is a 0.5 GHz signal; the resulting spectrum shows the fundamental output at 4.5 GHz, along with the LO leakage and a $1 \times \text{LO} - 2 \times \text{IF}$ spurious signal suppressed by ~ 35 dBc. As expected, due to the octave wide input bandwidth, harmonics of the input signals are providing output spurious. An investigation into the spectral characteristics of the test signal generator revealed that the second harmonic output levels were typically ~ 35 dBc. This leads to the probability that the module is not producing these products but merely translating them up in frequency.

The LO leakage (at 5 GHz in Figure 7.41) shows that ~ 35 dB of suppression is achieved. This is somewhat lower than expected since the mixer achieves 30 dB LO isolation and the 4 - 4.5 GHz filter achieves an additional 30 dB rejection at 5 GHz. The substantial difference between the measured and expected levels can only be attributed to cavity effects in the IF circuit. Since no special attention was paid to the suppression of modes in the cavity it is reasonable to expect that coupling through the cavity is possible. This is particularly true when considering the poor input match which the leaked LO experiences at the bandpass filter (since it is rejected by this filter).

Figure 7.41B shows a similar output spectrum when the IF input was set to 1 GHz. In this case the fundamental output is located at 4 GHz. The undesired LO leakage signal is present at 5 GHz as well.

-

FIGURE 7.41A IF UPCONVERTER SPECTRAL OUTPUT
 (IF input at 500 MHz, +12 dBm)

FIGURE 7.41B IF UPCONVERTER SPECTRAL OUTPUT
 (IF input at 1 GHz, +12 dBm)

7.7 Integrated Transceiver Unit

The previous subsections have dealt with the attributes of the modules which when integrated together form the transceiver unit, as shown in Figure 7.1.

This section outlines the performance of the transceiver after the modules have been integrated into the final test housing. The assembly process employs standard MMIC carrier hybrid interconnect techniques using 0.001" x 0.010" gold ribbon jumpered between the microstrip traces of the various carrier modules. A photograph of the final assembly is shown in Figure 7.42.

The integrated transceiver was evaluated for the following characteristics:

- Receive mode conversion loss, selectivity and input return loss,
- Receive mode dynamic range compression,
- Receive mode noise figure,
- Receive mode output spurious,
- Transmit mode conversion gain and input return loss,
- Transmit mode output power,
- Transmit mode output spurious, and
- T/R transient switching characteristics.

The receive mode conversion loss and selectivity were measured under small-signal conditions using the test set-up shown in Figure 7.27. The transceiver was switched⁽¹⁾ to the receive mode and the swept conversion loss of the transceiver was measured with the VCO tuned to 34.5, 35 and 35.5 GHz as shown in Figure 7.43. Since a calibrated sweep of this characteristic is not possible due to the frequency translation, a detailed analysis of conversion loss was carried out by measuring the input vs output power every 50 MHz across the respective RF bands. This data is presented in Figure 7.44. Measured flatness is ± 2.5 dB within the defined passbands (these would ideally exhibit 3 dB roll-offs).

(1) This switching was achieved by diode jumpering due a failure of one of the on-chip switching diodes.

FIGURE 7.42A INTEGRATED TRANSCEIVER TEST UNIT
(split block in place, actual size is 3"W x 4"L x .75"H)

FIGURE 7.42B INTEGRATED TRANSCEIVER TEST UNIT
(split block removed, actual size is 3"W x 3"L x .75"H)

LOSS

CH1:	A	-M
10.0	dB/	REF + .00 dB

FIGURE 7.43A INTEGRATED TRANSCEIVER RECEIVE MODE
CONVERSION LOSS AND SELECTIVITY

FIGURE 7.43B INTEGRATED TRANSCEIVER RECEIVE MODE
CONVERSION LOSS AND SELECTIVITY

**FIGURE 7.44 INTEGRATED TRANSCEIVER CONVERSION LOSS
FLATNESS MEASURED USING CALIBRATED POWER METER**

A nominal conversion loss of 6 dB is achieved which is considerably lower than expected (+6 dB gain). The source of the difference is attributed to the following:

- Higher-than-expected losses through the RF bandpass filter and transition (1.5 dB expected, 3.4 dB measured loss),
- Higher-than-expected losses through the T/R switch (1 dB expected, 3 dB measured loss),
- Higher-than-expected losses in the downconverter mixer (7 dB expected, 12 dB measured loss),
- Lower-than-expected LNA gain (14 dB expected, 12 dB measured gain), and
- Interconnect losses due to poor assembly practice (~ 1 dB).

As a result, the nominal conversion loss difference between the expected and measured values can be accounted for. The critical contributor is the millimetre-wave downconverter. As discussed in a previous section, no data was available from the foundry on the performance of this device when using an IF as high as that employed here (4 - 4.5 GHz).

A selectivity of 40 - 50 dB is measured as shown in Figure 7.43. Limitations in the dynamic range of the test equipment ultimately limit the selectivity measurement to ~ - 40 dBm. Since the transceiver employs a waveguide section which cuts off at ~ 32 GHz, a marked increase in selectivity would be expected in this frequency range.

The input return loss of the transceiver operating in the receive mode is shown in Figure 7.45. As expected, the transceiver shows reasonably well-matched operation in the 38 - 40 GHz range, as this is the design operating frequency range of the unit.

The compression characteristic of the transceiver (receive mode) was evaluated by measuring the conversion gain at various input power levels. The results of this measurement are shown in Figure 7.46. The curves in the figure indicate that gain compression occurs just below +5 dBm. This higher-than-expected compression is largely due to the increase in conversion loss of the receive path components upstream of the receiver LNA and downconverter.

FIGURE 7.45 INTEGRATED TRANSCEIVER RECEIVE MODE INPUT RETURN LOSS (RF at -10 dBm)

FIGURE 7.46 TRANSCEIVER RECEIVE MODE RELATIVE CONVERSION LOSS (CALIBRATED APPROXIMATELY AT $P_{IN} = -10$ dBm) AT VARYING INPUT POWER LEVELS (RF input from -10 to +10 dBm, LO at 35 GHz)

The receive-mode noise figure was measured using the test set-up shown in Figure 7.47. Using the Y-factor technique [1], the following procedure was applied to calculate the transceiver receive-mode noise figure:

- Calculate the noise figure of a 0.5 - 1 GHz, high-gain amplifier using the low-frequency noise source (shown dotted in Figure 7.48), giving N_{amp} .
- Determine the gain (or loss) of the unit under test (UUT) (from a previous section), giving G_{uut} .
- Measure the total system (i.e. UUT + amplifier) noise output power ratio (hot/cold) of the UUT across the frequency range of interest, giving N_2/N_1 .
- Compute the system (i.e. UUT + amplifier) noise figure from

$$NF_{sys} = ENR / ((N_2/N_1) - 1),$$

where ENR is the excess noise ratio produced by the noise source.

- Compute the UUT noise figure (N_{uut}) from the relation for cascaded components:

$$NF_{sys} = NF_{uut} + (NF_{amp} - 1) / G_{uut}$$

This process was repeated with the LO set at 34.5, 35, and 35.5 GHz, and with measurement, stepped in 100 MHz increments across the input RF band. The results are presented graphically in Figure 7.48. The noise figure measurements are higher than expected by an amount which is nominally attributable to the increased losses in components upstream of the LNA. It should be pointed out that this measurement is very sensitive due to the fact that the UUT noise figure is not significantly lower than the ENR of the noise source used in the measurements. The result of this test condition is that the N_2/N_1 ratio measurement needs to be performed to ~ 0.1 dB accuracy. A spectrum analyzer is used as the frequency selective power measurement device (using the "max hold" facility). The measurement error which it introduces is not likely to be insignificant with respect the required overall measurement accuracy.

FIG076.CDR

FIGURE 7.47 NOISE FIGURE TEST SET-UP

FIGURE 7.48 TRANSCEIVER RECEIVE MODE NOISE FIGURE
(LO at 34.5, 35 and 35.5 GHz).

The receive mode spurious was investigated. However, due to the "noisy" nature of the multiplied test source and the impossibility of fully sorting unknown spectral components using a harmonic spectrum analyzer, the source of transceiver output spurious components could not be determined.

The transceiver transmit-mode conversion loss was evaluated using the swept scalar network analyzer test set-up shown in Figure 7.21. The conversion gain is shown in Figures 7.49 (three LO frequencies employed) and input return loss is shown in Figure 7.50.

As noted in the case of the receive-mode testing, calibrated conversion loss measurements not possible, due to the frequency translation of the UUT. A further study of the conversion loss and maximum output power was undertaken using discrete RF input tones (rather than swept testing). Testing was carried out using LOs set to 34.5, 35 and 35.5 GHz and input RF stepped across the input band in 100 MHz increments. The results of this measurement are shown in Figure 7.51. The output power produced by the transceiver (using RF input power = +8 dBm for optimal conversion gain) is nominally +12 dBm and exhibits a flatness of $\sim \pm 3$ dB. The source used in this test exhibits an approximate flatness of ± 1 dB. The nominal conversion gain is therefore +4 dB. The differences between the expected and measured conversion gain and output power are largely attributable to the added losses in the T/R switch and transition/filter as discussed in previous subsections.

Transmit-mode output spurious were investigated using the spectrum analyzer. Since harmonic conversion is employed in the analyzer, only expected products can be readily investigated. The output spurious at the LO frequencies and image bands were investigated. Since the output waveguide of the transceiver has been designed to provide a cut-off higher than the image band, the image signals are attenuated and are below the measuring capacity of the spectrum analyzer. Suppression is estimated at > 60 dBc. The LO signal leakage was measured at the low, middle and high end of its tuning range (34.5 - 35.5 GHz) with the typical results shown in Figure 7.52. Suppression of the LO is on the order of 47 dBc. The expected suppression level was a cumulation of effects from the following:

- Mixer isolation (37 dB),

FIGURE 7.49A INTEGRATED TRANSCEIVER TRANSMIT MODE
CONVERSION LOSS (LO at 34.5 GHz, RF measured
through a 20 dB coupler)

FIGURE 7.49B INTEGRATED TRANSCEIVER TRANSMIT MODE
CONVERSION LOSS (LO at 35 GHz, RF measured
through a 20 dB coupler)

FIGURE 7.49C INTEGRATED TRANSCEIVER TRANSMIT MODE CONVERSION LOSS (LO at 35.5 GHz, RF measured through a 20 coupler)

INPUT RETURN LOSS

FIGURE 7.50 INTEGRATED TRANSCEIVER TRANSMIT MODE INPUT RETURN LOSS (RF input at +10 dBm)

FIGURE 7.51A INTEGRATED TRANSCEIVER TRANSMIT MODE OUTPUT POWER/CONVERSION LOSS (LO at 34.5 GHz, RF in at +8 dBm)

FIGURE 7.51B INTEGRATED TRANSCEIVER TRANSMIT MODE OUTPUT POWER/CONVERSION LOSS (LO at 35 GHz, RF in at +8 dBm)

FIGURE 7.51C INTEGRATED TRANSCEIVER TRANSMIT MODE OUTPUT POWER/CONVERSION LOSS (LO at 35.5 GHz, RF in at +8 dBm)

FIGURE 7.52 INTEGRATED TRANSCEIVER TRANSMIT MODE LO LEAKAGE (LO at 35 GHz, typical)

- Power amplifier rejection (8 dB), and
- Bandpass filter rejection (30 dB).

This should nominally result in a rejection of near 75 dB. A potential explanation for this is the difference in matching characteristics (in the LO band) which are exhibited in-circuit versus in the discrete test set-up for the modules. This probably caused increased LO radiation into the RF cavity and reduced isolation. Since no avenue for testing this hypothesis was possible, no further investigation was pursued.

An investigation into T/R switching characteristics was not possible due the complex equipment needed to measure RF transients at millimetre-wave frequencies. A discussion with the foundry that designed the T/R switches indicated that only minimal testing of that nature had been performed. Typical test results for the switching characteristics of the device indicate that 2 ns typical switching speeds (50% control to 90% RF) are achievable. [2]

7.8 References for Chapter 7

- [1] S.F. Adam, Microwave Theory and Applications, Prentice Hall, 1969, Section 6.3.
- [2] Alpha Industries Foundry data.

8.0 SUMMARY AND CONCLUSIONS

8.1 Proof-of-Concept Transceiver Performance Summary

Based on the measured results of Chapter 7.0, the proof-of-concept transceiver can be evaluated through a review of the required and predicted performance attributes. Table 8.0 shows this summary of achieved versus required/predicted performance.

The achieved proof-of-concept performance falls short of the predicted performance in several areas. Of these, the noise, figure and transmit-mode output power provide the most severe impacts on the usefulness of the transceiver in the WLAN system application.

As discussed in Chapter 7.0, these performance factor shortcomings can be related to the higher-than-expected losses in the transition from microstrip to waveguide and the T/R switch module. The transition can be improved by more careful tuning. In the open literature, a number of transitions of this nature have been reported with losses of ~ 1 dB or less. The T/R switch losses are probably due to implementation process shortcomings. Several of these modules were built-up and evaluated to determine if this was indeed the cause of the higher than predicted losses. All of these exhibited similar losses. Discussions with the foundry did not provide insight into the problem with the exception that the test circuitry used differs in thickness from that employed here (0.005" versus 0.010" thick). It is the author's experienced opinion that the substrates employed here are suitable and are unlikely to be the source of the observed performance. However, having said this, not all things RF are completely understood. Subsequent implementations would need to employ the thinner substrate in order to assess this factor.

The high receiver noise figure and low transmit power exhibited by the proof-of-concept transceiver would nominally render this design useless for WLAN applications even when considering very high gain antennas on the user terminals. If this type of transceiver were to be employed in conjunction with lossier antennas (phased arrays with complex feed networks as are being considered in the CITR millimetre-wave WLAN project, for example) further degradation could be expected. An additional system factor which also requires

TABLE 8.1
PROOF-OF-CONCEPT TRANSCEIVER PERFORMANCE
SUMMARY/COMPARISON

FACTOR	UNITS	REQUIRED PERFORMANCE	ESTIMATED PERFORMANCE	ACHIEVED PERFORMANCE	NOTES
1) Operating Frequency	GHz	38	38	38	Somewhat dictated by limitations in GaAs MMIC technology
2) Selectable Operating Frequencies	—	5	3	3	~1 GHz bandwidth usage achieved
3) Transmitter Power	mw	>20	~50	10	At antenna input
4) Receiver Noise Figure	dB	<12	~8	~12	At antenna input
5) Transmitter/Receiver Antenna Gain	dBi	>20	25	N/A	
6) Polarization	—	Circular	Linear	Linear	V, H or slant orientation
7) Operating Range	m	50 target, 30 min.	50	~ 50 m with minor circuit optimizations as noted	LOS 20 dB margin assumed 14 dB min. SNR
8) Minimum Operating SNR	dB	20 target, 10 min.	14	14	At 7)
9) Data Rate	Mbit/s	155 target, 25 min.	155 Mbit/s min., 330 Mbit/s target	500 MHz usable data bandwidth	Half duplex, TDMA
10) Modulation	—	TBD	Differential PSK or FSK	Differential PSK or FSK	This modulation was accommodated but not included in proof-of-concept work
11) Power Consumption	w	Minimize	TBD	TBD	
12) Physical Size	—	Minimize	TBD	~3" x 4" x 1"	Exploitation of SMT, MMIC circuitry
13) Technological Implementation	—	MMIC, SMT	MMIC, SMT	MMIC, SMT	
14) Long Term Production Cost Target	\$	200	TBD	TBD	

consideration, but is outside of the scope of this thesis, is the amount of transmitter 'back-off' needed for the selected implementation. Multi-carrier schemes or phase sensitive modulations often require several dB of back-off from the transmitter saturation in order to achieve acceptable system performance. It is therefore imperative to achieve these critical transceiver performance factors.

The MMIC/SMT approach investigated here would provide a 'starting point' for design intended to minimize overall cost. Ultimately, the SMT portion of the transceiver would need to be reduced to a single device as would the MMIC circuitry.

8.2 Areas for Follow-on Research

Aside from the performance shortcomings addressed in the previous section, other factors require research attention because they affect the transceiver and system designs but are only touched upon within the scope of this thesis effort. Among these are the following;

- BER performance of the fixed LOS data link (omni hub to directional user). Measurements on the temporal characteristics of the indoor WLAN propagation channel are underway and/or published. These measurements indicate that rates in excess of several 10's of Mbits/s will require some form of multipath suppression. However, reports have also been made [1] of very simple LOS links using directional user terminals which have achieved 622 Mbits/s data rates with very low BERs without any form of suppression circuitry. Further evaluation of this 'brute force' performance should be pursued in order to establish the overall need for multipath suppression and the extent (and cost) to which it should be applied.
- Complex equalization schemes potentially needed for the suppression of multipath require circuit speeds which appear to be out of the reach of current/near-term digital technology. One equalization realization discussed briefly in this thesis is the use of ACT transversal filters to achieve very high speed equalization circuits able to operate with bit spacings of only a few nanoseconds. The usefulness of

ACT technology should be investigated with a goal to determine the extent of its applicability to millimetre-wave indoor WLANs.

8.3 Conclusions

The general goal of this thesis was to investigate the feasibility of an MMIC-based transceiver for broadband wireless local area network (WLAN) applications in the 20 - 60 GHz frequency range. This goal has been addressed through the course of this thesis effort and concluded with the design, construction and test of a proof-of-concept model which generally exhibited the predicted performance attributes. Areas in which predicted attributes disagreed with experimentally derived results have been identified and their causes investigated to the point where solutions have been provided.

8.4 References for Chapter 8

- [1] P.F. Driessen, '622/Mb/S Indoor Wireless Systems with Directional Antennas', IEEE Wireless '94 Symposium Digest, pgs. 492 - 496.

APPENDIX A

TRANSMITTER COMPONENT SUMMARY

IF LPF - Avantek PLP-105

IF mixer - RHG M-186

IF amplifier - Pacific Monolithics AP-0600

RF mixer - Alpha AM035N3-00

RF amplifier - Litton VSM9599-H5 (now LMA-140)

RF power amplifier - Alpha AA84P3-00

RF SPDT T/R switch - Alpha AP640R2-00

-

PLP-105
Thin-Film Low Pass Filter

FEATURES

- DC - 1 GHz
- Lumped Element Design
- Small Size
- Hermetic Surface Mount Package
- Light Weight

APPLICATIONS

- Communications Circuits
- Cellular Telephone Systems
- Fiber Optics
- Instrumentation
- Missile Systems

DESCRIPTION

The PLP-105 is a thin-film, lumped-element, 5-pole low-pass filter packaged in a small 3/8" x 3/8" PlanarPak surface mount package. Use of thin-film lumped elements provides a circuit

much smaller and more repeatable than conventional distributed or discrete lumped element designs.

ELECTRICAL SPECIFICATIONS (Measured in a 50-ohm system @ $T_a = -55$ to $+85^\circ\text{C}$ unless otherwise noted)

Symbol	Characteristic	Typical $T_a = 25^\circ\text{C}$	Guaranteed Specifications		Unit
			$T_a = 0^\circ$ to 55°C	$T_a = -55^\circ$ to $+85^\circ\text{C}$	
FP	Passband Frequency	DC-1	DC-1	DC-1	GHz
FS	Stopband Frequency	2.3-5.5	2.3-5.5	2.3-5.5	GHz
VSWR	Passband VSWR	1.2:1	1.5:1	1.5:1	—
IL, Pass	Passband Insertion Loss (Maximum)	0.9	1.2	1.3	dB
IL, Stop	Stopband Insertion Loss (Minimum)	45	35	35	dB
AM	Amplitude Matching (Unit-to-Unit), $T_a = 25^\circ\text{C}$	± 0.1	—	—	dB
PT	Phase Tracking (Unit-to-Unit), $T_a = 25^\circ\text{C}$	± 2	—	—	Degree
P	Maximum Continuous Power	≥ 30	—	—	dBm

SCHEMATIC

TYPICAL PERFORMANCE OVER TEMPERATURE

MAXIMUM RATINGS

Continuous RF Power	-55°C to $+125^\circ\text{C}$	≥ 30 dBm
Operating Case Temperature	-55°C to $+125^\circ\text{C}$	
Storage Temperature	-62°C to $+150^\circ\text{C}$	
R Series Burn-in Temperature	$+125^\circ\text{C}$	

WEIGHT: (typical) .5 grams

Avantek, Inc. • 405 Redwood Drive, Redwood City, CA 94063 • Contact your local representative, distributor or sales office for further information. Leadtimes are to the back of 500 units.

6-4

PLP-105

Thin-Film Low Pass Filter

AUTOMATIC NETWORK ANALYZER MEASUREMENTS (Typical production unit @ $+25^\circ\text{C}$ ambient)

S-PARAMETERS

FREQ GHz	S_{11}		S_{22}		S_{21}		S_{12}		GPOEL (dB)
	Mag	Ang	dB	Ang	dB	Ang	Mag	Ang	
0.25	.07	-125.6	-63.4	-96.3	-63.7	-94.8	.07	-124.8	848
0.50	.08	-170.4	-64.6	-114.7	-64.9	-115.1	.08	-170.2	701
0.75	.08	-80.9	-65.7	-177.8	-65.9	-178.3	.08	-79.8	846
1.00	.08	-80.1	-65.1	-105.9	-65.5	-105.4	.08	-80.7	1,140
1.25	.07	-84.6	-65.1	4.7	-64.4	4.2	.06	-81.8	811
1.50	.07	-165.6	-16.0	-98.2	-16.0	-98.6	.06	-162.4	450
1.75	.07	-152.6	-27.10	-108.7	-27.10	-109.4	.02	-158.5	302
2.00	.04	-123.0	-34.85	-136.0	-34.85	-136.8	.03	-127.3	186
2.25	.05	-98.3	-46.17	-153.7	-46.17	-154.2	.04	-102.6	-246
2.50	.05	-75.9	-60.09	-131.6	-60.09	-135.6	.04	-82.4	-723
2.75	.06	-58.4	-69.57	-62.6	-69.57	-67.3	.06	-62.0	79
3.00	.06	-53.3	-63.97	-72.8	-63.97	-66.9	.04	-52.9	89
3.25	.06	-55.1	-63.97	-81.8	-63.97	-80.8	.04	-53.8	152
3.50	.06	-4.6	-63.97	-80.5	-63.97	-101.1	.04	4.8	294
3.75	.04	-24.6	-66.44	-118.4	-66.44	-120.4	.03	-14.2	294
4.00	.03	-64.8	-66.45	-136.7	-66.45	-136.5	.02	-33.9	228
4.25	.03	-65.4	-66.45	-157.1	-66.45	-161.1	.02	-63.8	236
4.50	.02	-65.8	-17.35	-178.3	-17.35	-178.2	.01	-74.0	232
4.75	.02	-105.8	-17.85	-180.7	-17.85	-180.2	.00	-84.6	243
5.00	.01	-128.1	-45.11	-138.6	-45.11	-138.0	.00	-116.5	234
5.25	.01	-148.9	-43.09	-115.9	-43.09	-114.7	.00	-138.8	285
5.50	.01	-171.8	-41.38	82.8	-41.38	96.1	.00	-161.1	285

LINEARIZATION RANGE: 25 to 5.5 GHz

(From -55°C to $+85^{\circ}\text{C}$)

Frequency Range	RF Port	4-8 GHz
	LO Port	4-11 GHz
	IF Port	DC-3.0 GHz
Conversion Loss**		7.0 dB Max
Isolation	LO to RF	(4-8 GHz) 25 dB Min
	LO to IF	(4-8 GHz) 15 dB Min
	RF to IF	(4-8 GHz) 17 dB Min

Impedance	50 Ohms Nominal
-----------	-----------------

Maximum Input Total Power	300 mW Max @ 25°C Dereated 3.2 mW/°C
------------------------------	---

RF Input	
1 dB Compression	+2.0 dBm Typical
1 dB Desensitization	0 dBm Typical

SSB Noise Figure	Within 1 dB of Conversion Loss Max
------------------	------------------------------------

3rd Order Input Intercept	
40 GHz	+15 dBm Typical
60 GHz	+12 dBm Typical
80 GHz	+11 dBm Typical

Package Type M-186 Flatpack (FP-10)
M-186C Connectorized (C-3)
(See pages 475 and 481 for physical dimensions.)

Environmental

These units are designed to meet the environmental and screening requirements of Table 1C, page 498 of the Adams-Russell catalog.

Pin Configuration	RF, P3, LO, P1, IF, P2 Case ground
-------------------	---------------------------------------

All calculations were done when operated at $\lambda = 7$ μm and the LO power was 50 dBm source and load impedance

FORM NO. 10-76 (Rev. 1-76)

Model No.	Part No.	Connectors	Unit Price (5-9 Units)
M-186	9519	Pin	\$200
M-186C	9514	SMA	287

Delivery is from stock.

Typical Performance

RHG **Make the Connection...** **Adams & Russell**
161 E. Industry Court Deer Park, NY 11729 Fax (516) 243-1222
COMPONENTS GROUP

FP-10

WEIGHT (APPROX.): 8.35 OUNCES 10 GRAMS

PACIFIC MONOLITHICS
GALLIUM ARSENIDE SYSTEMS

DATA SHEET

2-6 AND 2-8 GHz MMIC AMPLIFIERS IN DROP-IN PACKAGES

FEATURES

10 TO 30 dB GAIN BLOCKS
SMALL, SURFACE-MOUNT PACKAGES
LOW-NOISE, LOW CURRENT, AND
MEDIUM-POWER OPTIONS
GAIN CONTROL

Pacific Monolithics offers a variety of GaAs FET MMIC amplifiers in surface-mount packages for microstrip applications. These amplifier options include low noise, small signal, and medium power for cascadable 10 dB to 30 dB gain blocks, in 2-6 and 2-8 GHz versions.

These devices use fully matched and biased monolithic GaAs amplifier chips that significantly reduce the number of bond wires and MIC components required. The use of MMICs allows for fairly complex amplifiers to be built into small, surface-mount packages. Additionally, MMIC amplifiers provide higher reliability and lower life-cycle costs than comparable fully hybrid designs. Pacific Monolithics also offers temperature compensation as an option. Temperature compensation will reduce amplifier gain variation over the -54°C to $+100^{\circ}\text{C}$ temperature range.

MIL-STD 883 screening available

MODEL AM0608—10 dB GAIN BLOCK WITH GAIN CONTROL

SMALL-SIGNAL GAIN

GAIN vs. CONTROL VOLTAGE

245 Santa Ana Court, Sunnyvale, CA 94086 • TWX 510-600-3726 • (408) 732-8000

Annex A

SPECIFICATIONS (TYPICAL AND GUARANTEED)

MODEL	FREQUENCY (GHz)	GAIN (dB)	GAIN FLATNESS ($\pm$ dB)	NOISE FIGURE (dB)	POWER OUT @ 1dB COMP (dBm)	THIRD-ORDER INTERCEPT	VSWR IN/OUT	DC CURRENT @ +8 V (mA)	GAIN CONTROL $V_{GC} = 0$ to -4 V (MAX ATT) (dB)	PACKAGE
		MIN	MAX	MAX	MIN	TYP	MAX	TYP	TYP	

LOW NOISE

AM0401	3.7-4.2	10	0.25	2.75	+ 5.0	14	2:1	20	—	D,E
AM0501	5.4-5.9	10	0.25	3.00	+ 5.0	14	2:1	20	—	D,E
AM0601	5.9-6.4	10	0.25	3.00	+ 5.0	14	2:1	20	—	D,E

SMALL SIGNAL/LOW CURRENT CONSUMPTION

AM0607	2.0-6.0	9	0.75	7.50	+ 5.0	14	2:1	20	—	D,E
AM0627	2.0-6.0	18	1.50	8.00	+ 5.0	26	2.2:1	55	30	D,E
AM0637	2.0-6.0	27	1.75	8.00	+ 5.0	14	2:1	40	—	D,E
AM0608	2.0-6.0	10	0.50	8.50	+10.0	14	2:1	60	—	D,E
AM0608	2.0-6.0	19	1.25	8.00	+10.0	23	2:1	40	30	D,E
AM0628	2.0-6.0	19	1.25	8.00	+10.0	20	2:1	60	30	D,E
AM0638	2.0-6.0	30	1.50	8.00	+10.0	20	2:1	105	30	D,E
AM0609	2.0-6.0	12	0.50	9.00	+13.0	20	2:1	50 **	30	D,E
AP0610	2.0-6.0	10	1.00	6.00	+16.0	23	2:1	75 **	30	D,E
AM0629	2.0-6.0	21	1.25	9.00	+13.0	23	2:1	125 **	30	D,E
AM0639	2.0-6.0	32	1.75	9.00	+14.0	24	2:1	125 **	30	D,E

MEDIUM POWER

AP0600	2.0-6.0	12	0.75	7.00	+18.0	28	2.2:1	70 **	30	D,E
AP0621	2.0-6.0	21	1.50	6.50	+18.0	28	2.2:1	125 **	30	D,E
AP0631	2.0-6.0	30	2.00	6.50	+18.0	28	2.2:1	170 **	30	D,E
AP0612	2.0-6.0	9	1.00	10.0	+23.0	33	2:1	240	30	D,E
AP0622	2.0-6.0	20	1.75	9.50	+23.0	33	2:1	290	30	D,E
AP0632	2.0-6.0	29	2.00	9.50	+23.0	33	2:1	325	30	D,E
AP0614	2.0-6.0	9	1.75	10.0	+25.5	35	2:1	480	30	D,E
AP0624	2.0-6.0	20	2.25	10.0	+25.5	35	2:1	555	30	E
AP0634	2.0-6.0	30	2.50	9.00	+25.5	35	2:1	590	30	E

** 10

MICROPHASE CORPORATION

587 CONNECTICUT AVENUE
P O BOX 960
NEW HAVEN, CT 06526-0960
(203) 866-8000

****FAX****

05 FEBRUARY 1993

LOCKHEED CANADA INC.
STITTSVILLE, ONTARIO, CANADA

ATTENTION: ERIK BOCH

REFERENCE: YOUR FAXED REQUEST FOR QUOTATION OF 01 FEBRUARY 1993
MICROPHASE EN 93-118B

DEAR MR. BOCH:

MICROPHASE HAS COMPLETED ITS REVIEW OF THE ABOVE REFERENCE RFQ AND IS PLEASED
TO SUBMIT THE FOLLOWING QUOTATION:

BANDPASS FILTER

<u>QUANTITY</u>	<u>UNIT PRICE</u>
2	\$345
10	275

SPECIFICATIONS

MICROPHASE WILL MEET THE FOLLOWING SPECIFICATIONS:

INSERTION LOSS	- 1.5 dB (MAX.) 4.0 GHz TO 4.5 GHz
VSWR	- 2.0:1 (MAX.) 4.0 TO 4.5 GHz
REJECTION	- 30 dB (MAX.) AT 3.7 GHz AND 4.8 GHz
	50 dB (MAX.) DC TO 3.5 GHz
	AND
	5.0 GHz TO 10.0 GHz
SIZE	- 1.00 (L) X 0.50 (W) X 0.22 (DP) LESS 0.12 PINS
MOUNTING	- 4 (#1-64 TAPPED HOLES)

DELIVERY: 10 WEEKS ARO

SHOULD YOU REQUIRE FURTHER INFORMATION, PLEASE FEEL FREE TO CONTACT ME.

VERY TRULY YOURS,

LORI A. MUCCI
MICROPHASE CORPORATION

LAM/dmk/0455n

Ka-Band Monolithic Balanced Mixer

AM035N3-00

FEATURES

- Small Size 126 x 126 x 8 (mils)
- No Bias Required
- Microstrip Interface
- DC Coupled IF Port
- AC Coupled RF and LO Ports
- Low Noise Figure
- High LO/RF Isolation
- High G Survivability
- Backside via Grounds

Description

The AM035N3-00 is a Ka-band gallium arsenide monolithic balanced mixer. The circuit consists of matched Schottky diodes and a 180° hybrid in microstrip transmission medium to give excellent performance for communication, radiometer, and radar applications. Gold plated microstrip interfaces are compatible with hybrid bonding techniques.

Electrical Specifications @ 25°C

Center Frequency	30 to 45 GHz
Bandwidth	± 7%
Noise Figure, DSB	4.5 dB, Max
Conversion Loss	5.0 dB
RF/LO Isolation	24 dB, Typ.

Chip Configuration

Typical Performance

Specifications subject to change without notice.

April 1992

4-22

Alpha

Alpha Industries, Inc./GaAs MMIC Products Group/20 Sylvan Road/Woburn, MA 01801 U.S.A./ (617) 935-5150/Telex: 949436/Fax: (617) 935-4939

GaAs Monolithic Microwave Integrated Circuit VSM9599-H5

DESCRIPTION

The Litton VSM9599-H5 is a 10.0 - 40.0 GHz GaAs HEMT monolithic distributed low noise amplifier.

FEATURES

Frequency:	10.0 - 40.0	GHz
Small Signal Gain:	6.0	dB
SSG/Frequency:	± 1.5	dB
Noise Figure:	6.0	dB
P_{out} @ -1 dB:	+8.0	dBm
VSWR Input:	2.0 : 1	typical
VSWR Output:	2.0 : 1	typical
Voltage:	3.0	Vdc
Current:	50.0	mAdc
Chip Dimensions	.040 x .096 x .004	inch

DEVICE LAYOUT

Litton

Solid State

3251 Olcott Street / Santa Clara, CA 95054 / 408-988-1331 or 408-988-1845

TYPICAL MICROWAVE PERFORMANCE

GaAs Ka-Band MMIC Driver Amplifier

Three Stage

AA840P3-00

FEATURES

- 16dB Gain, 39 GHz
- 17 dBm Power Output
- Single FET Bias Design
- Size: 244 x 145 mils
- Backside via Ground
- Optional - microstrip or coplanar

Description

Alphas AA840P3-00 is a Ka-Band, three stage monolithic medium power amplifier featuring 0.25 μm MBE grown MESFET devices. The gates are E-beam written for high uniformity and reliability. Suitable for Commercial or Military communications transmitter.

Electrical Specifications

@ 25 °C

Frequency Response	38.5 - 39.5 GHz
Gain	16 dB, typ. 14 dB, min.
P_{1dB}	17 dBm, typ. 14 dBm, min.
P_{sat}	19 dBm, typ. 17 dBm, min.
DC Bias	$I_{ds} = 500\text{mA}$ $V_{gs} = 5.5\text{V}$
I/O VSWR	Input 2.5:1 Output 2:1 max.

AA840P3-00

Typical Performance

Specifications subject to change without notice.

April 1992

4-20

Alpha

Alpha Industries, Inc./GaAs MMIC Products Group/20 Sylvan Road/Woburn, MA 01801 U.S.A. (617) 935-5150/Tel: 949-4767/Fax: (617) 935-4939

GaAs MMIC SPDT PIN Switch 26-40 GHz Reflective

AP640R2-00

FEATURES

- Broad Bandwidth
- High Isolation
- Fast Switching
- Small Size:
100 x 70 x 4 (mil)
- I/O DC Blocks
- Backside via Grounds
- Bias Network Included on Chip

Description

Alphas AP640R2-00 GaAs SPDT Monolithic PIN Switch consists of one shunt PIN per arm to obtain high isolation (35dB) and low insertion loss (1dB max.).

The design combines small size, fast speed (2 nS) and state-of-the-art performance at low production costs to offer the system engineer a real alternative to PIN hybrid switches. Switch includes complete "on-chip" bias network.

Electrical Specifications @ 25°C

Frequency Range	26 to 40GHz
Insertion Loss	1dB max. up to 35 GHz 1.5dB max. up to 40 GHz
VSWR	2:1
Isolation	30dB min.
Switching Speed	2nS
Switching Current	
On	24mA
Off	-5v

AP640R2-00

Typical Performance

Specifications subject to change without notice.

April 1982

4-23

Alpha

Alpha Industries, Inc./GaAs MMIC Products Group/20 Sylvan Road/Woburn, MA 01801 U.S.A./ (617) 935-5150/Telex: 948436/Fax: (617) 935-4939

APPENDIX B

MICROSTRIP-TO-WAVEGUIDE TRANSITION DESIGN

.

B.1 OBJECTIVES

The following design objectives were implemented:

- Low loss: <0.5 dB per transition,
- Input microstrip impedance: 50 ohm,
- Output waveguide: rectangular WR-19,
- Size: minimize, and
- No post-assembly tuning.

B.2 DESIGN APPROACH

The design approach taken to realize this transition is outlined in some detail in [1,2]. This type of transition [being a travelling-wave structure rather than a standing-wave] is able to operate with low insertion loss over full waveguide bandwidths without the need for post-assembly tuning. A reported example of this type of transition [1] indicates that a 0.25 dB insertion loss per transition is possible in the 18-26 GHz range.

Figure B-1 shows the general transition. It is based on a "thin" ridged waveguide structure. The antipodal nature of the printed finline effectively rotates the E-field which is then passed through a balun which interfaces the balanced line (shown as cross-section C-C in Figure B-1) with the unbalanced microstrip (shown as cross-section E-E in Figure B-1). This balun is designed somewhat empirically. A balun structure able to operate at 40 GHz is shown in Figure B-2.

Generally the balanced line impedance is designed to closely match the 50 ohm microstrip. Since this section of line is balanced the impedance matching process becomes somewhat experimental in nature [1,2]. The tapers employed in the transition shown in Figure B-2 are designed to be ~ a wavelength in length (or longer) at/near the lower edge of the operating frequency range. A design edge of 35 GHz is selected to avoid the need to calculate the effective dielectric constant (and hence guide wavelength) of the tapers. Since the effective dielectric constant is changing along the antipodal finline tapers, this is a difficult task. The

tapers shown in Figure B-2 are of single exponential design. It has been found through analysis and experimentation that the exponential technique provides a lower insertion loss solution [3] than other transition designs as shown in Figure B-3. This is generally associated with superior return loss performance and will be pursued in this thesis effort.

FIGURE B-1 PRINTED WAVEGUIDE-TO-MICROSTRIP TRANSITION STRUCTURE [2]

FIGURE B-2 A PRINTED TRANSITION ABLE TO OPERATE AT 40 GHZ [2]

2. Values for the reflection coefficient (a) and impedance (b) for a given finline configuration and length are easily obtained from the above design curves. These curves are based on the closed-form characteristic impedance definition.

FIGURE B-3 COMPARISON OF TAPERED FINLINE TRANSITION TYPES [3]

B.3 REFERENCES

- [1] J. Van Heuven, 'A New Integrated Waveguide-to-Microstrip Transition', IEEE Transactions on Microwave Theory and Techniques, March 1976, Pgs. 144 - 147.
- [2] L. Lavedan, 'Design of Waveguide-to-Microstrip transitions Specially Suited to Millimeter-Wave Applications', Electronic Letters, 1977 (reprinted), Pgs. 81 - 82.
- [3] P. Pramanick, P. Bhartia, 'Design Tapered FInlines Using A Calculator', Microwaves & RF, June 1987, Pgs. 111 - 114.

APPENDIX C
RF BANDPASS FILTER DESIGN

FILTER REQUIREMENTS

- i) TCHEBYSCHIEFF Implementation, .2 dB Ripple Design
- ii) Passband 3 dB points: 38.5 GHz, 40.5 GHz
- iii) Selectivity: >40 dB @ 35 GHz, 44 GHz

DETAILED DESIGN

Employing the process outlined in [1]; the following parameters are calculated:

$$\lambda_{ga} @ 35 \text{ GHz} = .337''$$

$$\lambda_{g1} @ 38.5 \text{ GHz} = .306''$$

$$\lambda_{g2} @ 40.5 \text{ GHz} = .291''$$

$$\lambda_{gb} @ 44 \text{ GHz} = .262''$$

$$\lambda_{go} = (\lambda_{g1} + \lambda_{g2})/2 = .298''$$

$$\Delta\omega = 2 \left[\frac{\lambda_{ga} - \lambda_{gb}}{\lambda_{ga} + \lambda_{gb}} \right] = .050 \text{ or } 5\%$$

In order to establish the low pass filter prototype on which to generate the bandpass filter, the method outlined in [2] is employed where:

$$L_A(\omega^1) = 10 \text{ LOG}_{10} \left[1 + \epsilon \left(\frac{\omega^1}{\omega_1^1} \right)^{2n} \right] \quad (1)$$

where; ω_1^1 = Lower Equi-ripple band edge
 ω^1 = Frequency at which rejection, $L_A(\omega^1)$, is to be attained
 $\epsilon = (10^{L_{AR}/10}) - 1$; L_{AR} = Ripple expressed in dB

Completing (1) for the filter requirements results in:

$$\left| \frac{\omega^1}{\omega_1^1} \right| = 7.03$$

$$\therefore \left| \frac{\omega^1}{\omega_1^1} \right| - 1 = 6.03$$

Using Figure C-1 it can be seen that an $n = 3$ filter design will achieve the required rejection (~ 50 dB). Applying Table C-1 gives the low pass prototype values:

$$g_0 = 1.000$$

$$g_1 = 1.2275$$

$$g_2 = 1.1525$$

$$g_3 = 1.2275$$

$$g_4 = 1.000$$

Using these values to calculate the normalized K-inverter impedance values [1] using:

$$\frac{K_{01}}{Z_0} = \frac{K_{34}}{Z_0} = \frac{\pi}{2} \cdot \left(\frac{\Delta\omega}{g_{0,3} g_{1,4}} \right)^{\frac{1}{2}}$$

$$\frac{K_{12}}{Z_0} = \frac{K_{23}}{Z_0} = \frac{\pi}{2} \cdot \Delta\omega \cdot \left(\frac{1}{g_{1,2} g_{2,3}} \right)^{\frac{1}{2}}$$

results in;

$$\frac{K_{01}}{Z_0} = \frac{K_{34}}{Z_0} = .252$$

$$\frac{K_{12}}{Z_0} = \frac{K_{23}}{Z_0} = .066$$

FIGURE C-1 CONVERSION OF FILTER PARAMETERS TO FILTER ORDER (n) [2]

TABLE C-1

PROTOTYPE FILTER TABLES [2]

VALUE OF n	ϵ_1	ϵ_2	ϵ_3	ϵ_4	ϵ_5	ϵ_6	ϵ_7	ϵ_8	ϵ_9	ϵ_{10}	ϵ_{11}
0.01 db ripple											
1	0.0960	1.0000									
2	0.4488	0.4077	1.1007								
3	0.6291	0.9702	0.6291	1.0000							
4	0.7128	1.2003	1.3212	0.6476	1.1007						
5	0.7563	1.3049	1.5773	1.3049	0.7563	1.0000					
6	0.7813	1.3600	1.6896	1.5350	1.4970	0.7098	1.1007				
7	0.7969	1.3924	1.7481	1.6331	1.7481	1.3924	0.7969	1.0000			
8	0.8072	1.4130	1.7824	1.6833	1.8529	1.6193	1.5554	0.7333	1.1007		
9	0.8144	1.4270	1.8043	1.7125	1.9057	1.7125	1.8043	1.4270	0.8144	1.0000	
10	0.8196	1.4369	1.8192	1.7311	1.9362	1.7590	1.9055	1.6527	1.5817	0.7446	1.1007
0.1 db ripple											
1	0.3052	1.0000									
2	0.8430	0.6220	1.3554								
3	1.0315	1.1474	1.0315	1.0000							
4	1.1088	1.3061	1.7703	0.8180	1.3554						
5	1.1468	1.3712	1.9750	1.3712	1.1468	1.0000					
6	1.1681	1.4039	2.0562	1.5170	1.9029	0.8618	1.3554				
7	1.1811	1.4228	2.0966	1.5733	2.0966	1.4228	1.1811	1.0000			
8	1.1897	1.4346	2.1199	1.6010	2.1699	1.5640	1.9444	0.8778	1.3554		
9	1.1956	1.4425	2.1345	1.6167	2.2053	1.6167	2.1345	1.4425	1.1956	1.0000	
10	1.1999	1.4481	2.1444	1.6265	2.2253	1.6418	2.2046	1.5821	1.9628	0.8853	1.3554
0.2 db ripple											
1	0.4342	1.0000									
2	1.0378	0.6745	1.5386								
3	1.2275	1.1525	1.2275	1.0000							
4	1.3028	1.2844	1.9761	0.8468	1.5386						
5	1.3394	1.3370	2.1660	1.3370	1.3394	1.0000					
6	1.3598	1.3632	2.2394	1.4555	2.0974	0.8838	1.5386				
7	1.3722	1.3781	2.2756	1.5001	2.2756	1.3781	1.3722	1.0000			
8	1.3804	1.3875	2.2963	1.5217	2.3413	1.4925	2.1349	0.8972	1.5386		
9	1.3860	1.3938	2.3093	1.5340	2.3728	1.5340	2.3093	1.3938	1.3860	1.0000	
10	1.3901	1.3983	2.3181	1.5417	2.3904	1.5536	2.3720	1.5066	2.1514	0.9034	1.5386
0.5 db ripple											
1	0.6986	1.0000									
2	1.4029	0.7071	1.9841								
3	1.5963	1.0967	1.5963	1.0000							
4	1.6703	1.1926	2.3661	0.8419	1.9841						
5	1.7058	1.2296	2.5408	1.2296	1.7058	1.0000					
6	1.7254	1.2479	2.6064	1.3137	2.4758	0.8696	1.9841				
7	1.7372	1.2583	2.6381	1.3444	2.6381	1.2583	1.7372	1.0000			
8	1.7451	1.2647	2.6564	1.3590	2.6964	1.3389	2.5093	0.8796	1.9841		
9	1.7504	1.2690	2.6678	1.3673	2.7239	1.3673	2.6678	1.2690	1.7504	1.0000	
10	1.7543	1.2721	2.6754	1.3725	2.7392	1.3806	2.7231	1.3485	2.5239	0.8842	1.9841

Using the curve shown in Figure C-2 results in:

$$\left. \begin{array}{lcl} W_{01} & = & W_{34} = .014" \\ W_{12} & = & W_{23} = .060" \end{array} \right\} \text{Sceptum Widths}$$

$$\theta_{01} = \theta_{34} = -.65 \text{ rads}$$

$$\theta_{12} = \theta_{23} = -.58 \text{ rads}$$

The physical separation of the sceptums can be calculated from:

$$\theta_i = \pi + \frac{1}{2} [\theta_{j, j+1} + \theta_{j+1, j+2}]$$

$$\text{and } L_R = \frac{\lambda_{g0} \cdot \theta_i}{2\pi} ; L_A = \text{Sceptum separation in inches}$$

therefore;

$$\theta_A = \theta_C = 2.52$$

$$\text{and } L_A = L_C = .119"$$

$$\theta_B = 2.56$$

$$\text{and } L_B = .121"$$

A sketch showing the final filter is shown in Figure C-3.

Theoretical values of impedance inverter parameters of bilateral finline strips in U-band. (Dielectric thickness = 0.005 in., $\epsilon_r = 2.22$, — : K/Z_0 , and - - - : ϕ)

FIGURE C-2 CONVERSION OF SEPTUM PARAMETERS [1]

FIGURE C-3 FINAL FINLINE FILTER CONFIGURATION

REFERENCES

- [1] C. Nguyen, "Graphical Design of Millimetre-wave Fineline Bandpass Filters", International Journal of Infrared and-Millimetre-waves, Vol. 8, No. 12, 1987, pgs. 1581 - 1603.
- [2] G. Matthaei, L. Young, E. Jones, Microwave Filters, Impedance-Matching Networks, and Coupling Structures, Artech House, 1980, Chapter 4.
- [3] B. Bhatt, S. Kaul, Analysis, Design and Applications of Finlines, Artech House, 1987.

APPENDIX D

RECEIVER CHAIN COMPONENT SUMMARY

IF LPF - Avantek PLP-105 (see appendix A)

IF mixer - RHG M-186 (see appendix A)

IF bandpass filter - Microphase EN 93-118B (see appendix A)

IF amplifier - Pacific Monolithics AP-0600 (see appendix A)

RF mixer - Alpha AM035N3-00 (see appendix A)

RF amplifier - Litton VSM9599-H5 (now LMA-140) (see appendix A)

RF SPDT T/R switch - Alpha AP640R2-00 (see appendix A)

-

APPENDIX E

HIGH FREQUENCY LO COMPONENT SUMMARY

LO module - Alpha AO040N8-00

LO amplifier - Litton VSM9599-H5 (see appendix A)

LO SPDT - Alpha AP640R2-00 (see appendix A)

7.

GaAs Ka-Band MMIC Voltage Controlled Oscillator

AO040N8-00

FEATURES

- Center Frequency - 39 GHz
- Integrated FET and Varactor Diodes
- 10mW Output Power
- Single FET Bias Design
- Small Size:
88 x 104 x 4 mils
- Broad Tuning Bandwidth (500 MHz)
- Backside via Ground

Description

Alphas AO040N8-00 is a gallium arsenide MMIC chip that incorporates a 0.25 micron gate length FET and two Schottky varactors in a microstrip oscillator circuit. One of the tuning varactors can be used to provide temperature compensation of frequency with an external thermistor circuit. The backside of the chip is gold plated and die attachment can be accomplished with either gold-tin solder, indium solder, or silver epoxy. The circuit metal is 3μm thick plated gold and connections to the chip are made with conventional thermocompression bonding of gold ribbon or wire.

AO040N8-00

Electrical Specifications

@ 25 °C

Frequency	Specify
Tuning Bandwidth	500 MHz
Output Power	+8mW Typ +6mW Min
Max Load VSWR	2:1 Max
Operating Voltage	+5V
Operating Current	100 mA
Tuning Voltage	0 to +10V

Typical Performance

Specifications subject to change without notice.

April 1992

4-21

Alpha Industries, Inc./GaAs MMIC Products Group/20 Sylvan Road/Woburn, MA 01801 U.S.A./ (617) 935-5150/Telex: 949436/Fax: (617) 935-4939

APPENDIX F

LOW FREQUENCY LO COMPONENT SUMMARY

LO module - Pacific Monolithics OVC0550

SPDT switch - Alpha AK006M2-01

PACIFIC MONOLITHICS
GALLIUM ARSENIDE SYSTEMS

DATA SHEET

0.8 TO 17.0 GHz VOLTAGE-CONTROLLED GaAs MMIC OSCILLATORS

FEATURES

WIDE TUNING BANDWIDTHS
UP TO 200 mW OUTPUT POWER
HIGH LOAD ISOLATION
INTERNAL VOLTAGE REGULATORS

SURFACE-MOUNT CAPABILITY
SMALL SIZE
HIGH RELIABILITY

Pacific Monolithics offers a family of small, surface-mountable, highly reliable voltage-controlled oscillators. These oscillators use a negative-resistance ("R") GaAs MMIC chip with spiral inductors and an external varactor diode. The "R" GaAs MMIC chip contains a series feedback FET, interconnection matching circuits, buffer amplifier, and all the necessary bias network on a single 48 x 48 MIL area. The external varactor diode approach provides freedom of frequency and bandwidth selection, along with high Q characteristics. The monolithic VCOs can be easily interconnected with MMIC amplifiers, switches, mixers, prescalers, power dividers, and voltage regulators to achieve multifunction capabilities in a reduced size. The small, surface-mountable GaAs MMIC oscillators are ideal for most compact and densely packaged equipment designs. They are well suited for applications in built-in test equipment (BITE), compact phase-locked sources for EW systems, and low-cost commercial equipments, in addition to conventional oscillator applications. Hermetic sealing and MIL-STD 883 screening are available.

TYPICAL PERFORMANCE AT 25°C
OVC01S1

TYPICAL PERFORMANCE AT 25°C
OVC55S1

245 Santa Ana Court, Sunnyvale, CA 94086 • TWX 510-600-3726 • (408) 732-8000

GaAs SPDT-FET MMIC Switch WITH INTEGRAL DRIVER

Non-Reflective DC-6GHz AK006M2-01, AK004M2-11, AK002M2-04

FEATURES

- Low D.C. Power Consumption
~20mW per Arm
- Non-Reflective
- Integral Driver, +5v, -4v;
CMOS & TTL* Compatible

Description

The GaAs SPDT non-reflective chip is offered in three separate packages for convenience of mounting. At the higher frequencies, 6GHz, the 7 lead flat pack (AK006M2-01) has optimum performance since all leads are RF isolated. These devices are useful as modulators as well as switches in military, instrumentation and commercial applications.

AK006M2-01
AK006M2-10⁽¹⁾

AK004M2-11⁽¹⁾

AK002M2-04

Electrical Specifications @ 25° C

Package Style		01	11	04	
Insertion Loss (2)	DC-.5GHz	0.8	0.8	0.8	dB Max
	DC-1GHz	0.9	0.9	1.0	dB Max
	DC-2GHz	1.1	1.1	1.2	dB Max
	DC-4GHz	1.4	1.4	-	dB Max
	DC-6GHz	1.8	-	-	dB Max
Isolation	DC-.5GHz	52	50	45	dB Min
	DC-1GHz	52	50	40	dB Min
	DC-2GHz	48	45	35	dB Min
	DC-4GHz	37	30	-	dB Min
	DC-6GHz	22	-	-	dB Min
VSWR	DC-1GHz	1.3:1	1.3:1	1.7:1	Max
	DC-2GHz	1.5:1	1.5:1	2.0:1	Max
	DC-4GHz	1.8:1	1.8:1	-	Max
	DC-6GHz	2.0:1	-	-	Max

Operating Characteristics @ 25° C

Impedance	50 Ohms Nominal		
Switching Characteristics			
RISE, FALL (10/90% or 90/10% RF)		10 ns	Typ
ON, OFF (50% CTL to 90/10% RF)		20 ns	Typ
Video Feedthru (3)		30 mV	Typ
Input Power for 1dB Compression			
0.5-6 GHz	+20	dBm	Typ
0.001 GHz	+12	dBm	Typ
Logic Drives (volts)*			
	Min	Max	
Low (0)	0	2	
High (1)	3.5	5	
Bias Power (4)			
	+5v	.5v @ 1mA	Typ
	-4v	.25v @ 3mA	Typ ^{PM}

Specifications subject to change without notice.

Alpha

IMAGE EVALUATION TEST TARGET (QA-3)

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