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An Airborne Direction Finder for GPS Interference Sources

by

M. Roger Edwards, B.Sc.

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

Ottawa-Carleton Institute for Electrical Engineering

Department of Electronics

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Ottawa, Ontario, Canada

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acceptance of the thesis

An Airborne Direction Finder for GPS Interference Sources

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Abstract

The Global Positioning System (GPS) is being introduced into the air navigation system as a replacement for existing ground-based navigation aids. While GPS offers excellent three dimensional position accuracy world wide, its low signal levels make it susceptible to intentional and unintentional interference. Because of this, civil aviation agencies and Nav Canada in particular are interested in having the capability of identifying and locating interference sources as quickly as possible. One means of fulfilling this requirement is the airborne direction finder described in this thesis.

The main restrictions on the design were that the aircraft installation be simple and easily certifiable and that the signal levels to be tracked are expected to be at a very low level and hence would have a low signal to noise ratio. The proposed design measures the carrier phase differences from the outputs of three GPS antennas arranged in an equilateral triangular array. To improve the signal to noise ratio to a point at which accurate phase measurements can be made, very narrow bandwidth tracking filters implemented as phase locked loops are used. A means of assisting the filters to lock on to the desired signal using a human operator is provided.

Two revisions of the phase measurement board were built and tested with good results. A test antenna array was constructed, measured using GPS techniques and tested independently. The results showed that the basic premise is valid. The tracking filter design was simulated to determine its ability to improve signal to noise ratio for phase measurement. This was done successfully and prototypes were built and tested. These test were inconclusive due to unexpected behaviour of the phase detector used.

The design is considered to be sufficiently promising to warrant further development of a model suitable for operational testing.

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Chapter 1

Introduction

1.0 Problem Description

The navigation aids which support the world's present air navigation system require ground installations which are expensive, require periodic maintenance and calibration and limit the location of air routes. The Global Positioning System (GPS) is a satellite-based system for providing highly accurate position information to users at any point on or near the earth's surface in any weather conditions. As such, it is attractive as an air navigation system since it does not require the installation of ground-based facilities on or near the airways and it permits arbitrary routing of aircraft. For these reasons, the International Civil Aviation Organization (ICAO) and national civil aviation authorities are introducing GPS as a replacement for the present ground-based navigation aids in their air navigation systems [1][2][3].

One of the main disadvantages of GPS is that, due to the high satellite orbits, the signal strength near the earth's surface is low and, despite the processing gain resulting from the spread spectrum signal format, receivers are susceptible to intentional and unintentional jamming at relatively low power levels.

1.1 Motivation

Because any disruption of a sole means of aircraft navigation is not acceptable, civil aviation authorities, and in particular the Air Navigation Directorate of Transport Canada (now Nav Canada), need a means of identifying and locating sources of GPS interference. It will be necessary to reach a reported interference area quickly and to search large amounts of territory in a short period of time. This will most likely require a solution which includes an airborne direction finder capable of determining the relative bearing of an interference source. This relative bearing, when combined with information from the aircraft's naviga-

tion system, can pinpoint the interference source. Nav Canada has a fleet of high speed flight calibration aircraft which are typically distributed across the country at any given time and which can respond quickly to emergency situations. These aircraft are considered to be the most appropriate platforms for this function.

This thesis describes a design for such a direction finder and the development and testing of critical elements of this design.

1.2 Thesis Outline

This thesis is divided into 6 chapters. Chapters 1 and 2 provide a general discussion of GPS and its susceptibility to interference. Chapter 3 is a description of direction finding techniques and Chapter 4 describes the proposed system design. Chapters 5 and 6 include the design of the system components, simulation results and test results. Conclusions and recommendations for future design development are given in Chapter 7.

Chapter 2

Background

2.0 Summary

This first part of this chapter provides a basic description of the Global Positioning System (GPS) including some history and a description of some of the institutional issues which surround its use as an aeronautical navigation system. Also included is a brief section on differential GPS. The second part gives some background and a description of some recent work in the most important area of interest at the present time; intentional and unintentional jamming of the system in the vicinity of airports.

2.1 GPS System Description and History

The Global Positioning System (GPS) is a satellite-based navigation system developed and operated by the US Department of Defense (DOD). It permits the determination of position anywhere in the world to accuracies varying from 100m (2 distance root mean square) to less than 1m depending on the type of receiver being used. Two levels of service are provided: the Standard Positioning Service (SPS) which uses a Coarse/Acquisition (C/A) code and the Precision Positioning Service (PPS) which uses a Precise (P) code.

Early in the development of the system, the surveying community recognized its potential for high accuracy surveying with many advantages over traditional techniques. This was probably due to their experience with the US Navy Transit satellite position reference system which had been used by surveyors for more than 10 years. By installing receivers at two locations it was possible to record satellite data over long periods of time (several hours). By averaging out the random errors in the individual measurements and subtracting common errors, relative position accuracies of 1 part in 10^6 are readily achievable. The fact that, in the early stages of development, there were few satellites in orbit was not a great disadvantage because, in many circumstances, it was easier to set up a receiver, or pair of receivers, and wait until sufficient data had been recorded than it was, for example, to cut several kilometres of sight lines through dense bush. The economic benefits to the survey-

ors of even the abbreviated prototype GPS system led to a high level of activity in the development of improved receiver design and signal processing capability. One of the results of this development was the improvement of the accuracy of the C/A code receivers until it was close to that of P code receivers. However, DOD considered that the general availability of such an accurate positioning system would greatly diminish the original advantage provided by the GPS code. To restore this advantage they decided to introduce deliberate degradation into the C/A code performance of the operational constellation. This was done by introducing clock and , optionally, orbital parameter errors into the GPS signal in a process called selective availability (S/A). With S/A the unaided horizontal positional error is up to 100 metres 95% of the time. This is slightly more than twice the error without S/A. [5]

As a result of the base established by these early developments, the civilian user segment is now much larger than the military one and new applications for GPS are being found almost daily. One of the most important and fastest growing of these is air navigation and there is a great deal of effort being expended in order to adapt the system to meet the special needs of the aviation community.

The four major requirements for an all weather air navigation system are accuracy, integrity, availability and continuity. Accuracy must be sufficient for the phase of flight being considered. In the enroute phase 0.5 nautical miles (NM) might be sufficient whereas 3m might be required for the landing phase. GPS can meet the former easily and has been shown to be able to meet the latter using the differential techniques described later.

Integrity is defined as the system's ability to warn the pilot when the positioning accuracy has been degraded beyond the requirements for the particular phase of flight. For GPS this would probably result from a given satellite becoming unserviceable. This condition could be detected by a ground station and the information could be transmitted to the aircraft via a data link. Another approach would be to have the aircraft's receiver itself detect an unserviceable satellite by comparison of position solutions using redundant satellites. This latter is called receiver autonomous integrity monitoring (RAIM). Finally the system must be available for use a high percentage of the time. Continuity is the probability that a sys-

tem will be available for the duration of a phase of operation if it was available at the beginning of the phase.

One of the problems with the use of GPS for world-wide civil aviation is that it is, first of all, a military system and is controlled by the US DOD. A second problem, from the point of view of civil aviation authorities of non-North American countries, is that it is a US system and could be withdrawn at their whim. A third and potentially disastrous problem is that it is not optimized for civil use. In particular, it has been shown recently to be susceptible to jamming and unintentional interference as will be discussed later in this chapter. If this latter issue cannot be resolved satisfactorily then the whole issue of the use of GPS for civil aviation will be brought into question. This would have political consequences in the US since, in June 1994, the Administrator of the Federal Aviation Administration (FAA) announced that the FAA was terminating its development of the Microwave Landing System (MLS) in favour of GPS as a replacement for the current Instrument Landing Systems used for final approach and landing guidance in instrument weather conditions[4].

2.2 System Description

2.2.1 Basic Principles

If one can measure the distance from three satellites and if one knows the positions of those satellites at a given time then the position of the receiver can be determined by solving the following set of equations:

$$\begin{aligned}(x_1 - x_r)^2 + (y_1 - y_r)^2 + (z_1 - z_r)^2 &= r_1^2 \\(x_2 - x_r)^2 + (y_2 - y_r)^2 + (z_2 - z_r)^2 &= r_2^2 \\(x_3 - x_r)^2 + (y_3 - y_r)^2 + (z_3 - z_r)^2 &= r_3^2\end{aligned}\tag{1}$$

where the x_i , y_i and z_i are the coordinates of the 3 satellites, x_r , y_r and z_r are the coordinates of the receiver and the r_i are the measured ranges from the receiver to the satellites.

2.2.2 Range Measurement

In the GPS, the distances from the satellites are determined by measuring the difference between the time of transmission and the time of reception of a coded signal and multiplying the result by the speed of light. To achieve position accuracies in the range of metres the time difference must be known to an accuracy in the neighbourhood of 10ns. This places a tight requirement on the synchronization of satellite clocks. For this purpose each satellite carries two Cesium beam and two Rubidium frequency standards (stability on the order of 10^{-13})[19]. The performance of these standards is monitored and corrections are included in each satellite's navigation data message which is described later.

2.2.3 Satellite Position Determination

The position of each satellite is derived from the set of orbital parameters, commonly referred to as the ephemeris. This information is also included in the navigation data message transmitted at regular intervals by each satellite. This information includes:

- a) semi major axis
- b) eccentricity
- c) inclination
- d) longitude of ascending node (longitude at which the satellite crosses the equator from south to north)
- e) argument (angle in the orbital plane) of perifocal (point of closest approach to earth) passage
- f) mean anomaly at reference time
- g) mean motion difference from computed value

Note that the determination of the satellite position is also dependent on the knowledge of system time.

Range calculation demands that the receiver also have an accurate estimate of system time. However, in order to reduce the demands on the receiver clock design, the receiver clock

error is included in the position fix calculation at the cost of requiring an additional satellite for the computation. As is shown in Figure 2.1, an illustration of the problem in two dimensions, the measured range from each satellite is in error by an amount proportional to the receiver clock bias b_r (since the measured range differs from the true range, it is called the pseudorange in GPS terminology). Since this bias is common to all measurements, it may be estimated using a least squares error method except when only four satellites are used. This requires one more measurement than the number of dimensions of the position fix. Thus, a three dimensional GPS fix requires measurements from at least four satellites. With the current complete constellation of 24 satellites there are usually more than seven available at any given time so that it is not usually a problem.

Figure 2.1
An Illustration of the Effect of Clock Bias
and its Removal

2.2.4 Geometric Dilution of Precision (GDOP)

Although the accuracy of the ranging measurements is important to the overall accuracy of the system, the relative positions (geometry) of the satellites being used has a significant bearing on the final results. A measure of the “goodness” of the geometry is given by the GDOP which relates the accuracy of the position fix to the accuracies of the pseudorange measurements. The most used components of GDOP are PDOP (position dilution of precision), HDOP (horizontal dilution of position), VDOP (vertical dilution of precision) See [6] pp 220-224 .

$$GDOP = \sqrt{PDOP^2 + TDOP^2}$$

$$PDOP = \sqrt{HDOP^2 + VDOP^2}$$

Where TDOP is the time dilution of precision.

Position accuracy is therefore derived by multiplying the appropriate DOP by the pseudorange accuracy. e.g. Vertical accuracy = VDOP x pseudorange accuracy.

With the current constellation, HDOP is typically 1.0 to 1.5 and VDOP is typically 2.0 to 2.5.

2.2.5 Signal Format and Acquisition

GPS is an application of direct sequence spread spectrum Code Division Multiple Access (CDMA) since all satellites transmit on the same carrier frequencies but with different spreading codes. The two carrier frequencies used are 1575.42 MHz (designated L_1) and 1227.6 MHz (designated L_2). A 50 bit per second navigation message is sent on both frequencies. It contains the information shown in Figure 2.2. TLM is the telemetry word which contains an 8 bit Barker code for synchronization and HOW is the handover word which contains the information for determining the epoch of the P code (i.e. it identifies

which part of the P code is being transmitted at that time). This enables a P code receiver to acquire the P code in a reasonable amount of time.

The ephemeris is the set of orbital parameters for the transmitting satellite. The almanac contains a low resolution ephemeris for each of the other satellites in the constellation. This allows the receiver to determine which satellites are in view and to calculate an estimate of their carrier doppler shifts. This reduces acquisition time.

Figure 2.2
Summary of the Format of the GPS Navigation Message

The L_1 signal consists of two carrier components in quadrature. One is modulated with the spread by the P code at 10.23 MHz and the other is spread by the C/A code at 1.023 MHz. The L_2 carrier is BPSK modulated and spread by the P code only. Both the C/A and the P codes are unique to a given satellite.

The high chip rate of the P code was designed to provide the highest level of accuracy and was thus intended only for military users. In order to increase the difficulty for unauthorized users to exploit this capability, the P code is very long. The P code is the product of two PN codes $X_1(t)$ and $X_2(t+n_i T)$ where X_1 has a period of 15,345,000 chips and X_2 has a period of 15,345,037 chips. This results in a set of 37 codes

$$X_{p_i} = X_1(t) X_2(t+n_i T) \text{ where } T \text{ is the chip period } (0 \leq n_i \leq 36)$$

The period of this code is slightly more than 38 weeks. Thus, by resetting the codes on a weekly basis, each satellite in effect can have a different non overlapping segment of the code. This code is reset at the beginning of each GPS week (midnight Saturday/Sunday, GPS time; GPS time is shifted an integral number of seconds from Coordinated Universal Time (UTC)).

To permit fast acquisition of the P code a short, easily acquired code (the C/A code) is added as described above.

The C/A code is a Gold Code of length 1023 generated by the two polynomials [6]:

$$G_1(X) = 1 + X^3 + X^{10}$$

$$G_2(X) = 1 + X^2 + X^3 + X^6 + X^8 + X^{10}$$

At the chipping rate of 1.023 MHz the period of the C/A code is 1 ms.

For a “cold” start-up, with no knowledge of the almanac, the receiver does not know which satellites are in view (i.e. which codes are available), nor can it estimate the signal Doppler shift which can be up to 4 kHz. Thus, for each code and for an IF bandwidth of 1 kHz it must search approximately 9,000 combinations of frequency and code offsets. Theoretically, 1023 codes would have to be searched but since the number of codes in use at a given time is about 24, those codes can be embedded in the receiver software and thus the scope of the initial search can be limited. Normally, the time to first lock is approximately 90 seconds [6]. Once the first lock has been achieved, the navigation message can be decoded and the subsequent searches can be reduced significantly since the satellites in view (and hence their codes) can be determined and their Doppler shifts can be estimated from their positions.

Since the period of the C/A code is 1 ms, there is an initial ambiguity of the satellite range as the lock occurs for integral multiples of $1 \times 10^{-3} \times 3 \times 10^8 = 3 \times 10^5$ metres. Once four satellite codes have been acquired it is possible to resolve this ambiguity through the solu-

tion of the position fix calculations using the assumption that the receiver is near the earth's surface.

2.3 Implementation

From a functional point of view the GPS is divided into the Space Segment, the Control Segment and the User Segment

2.3.1 Space Segment

The nominal space segment consists of 24 satellites (21 primary + 3 active spares) arranged in six orbital planes equally spaced around the earth with each plane being oriented at 55° to the equator. The orbit radius is approximately 26560 km which results in a sidereal period of about 2 minutes short of 12 hours. This departure from synchronicity is designed to avoid strong orbital resonances which would decrease the usable life of the satellites.

The constellation configuration was designed to maximize a criterion called the "constellation value" (CV) which is the fraction of the earth's surface and time that the PDOP is less than 10 and at least 4 satellites have an elevation angle greater than 5° . It was also designed to provide minimum degradation of availability in the event of satellite failures. (see Table 2.2).

Table 2.2 Constellation Values with Satellite Failures

Number of Failed Satellites	Best CV	Average CV	Worst CV	Median HDOP	Median PDOP
0	1.0000	1.0000	1.0000	1.1	2.0
1	1.0000	0.9999	0.9997	1.2	2.1
2	1.0000	0.9993	0.9961	1.2	2.2
3	0.9998	0.9969	0.9769	1.3	2.3
4	0.9997	0.9905	0.9475	-	-

2.3.2 Control Segment

The Control Segment is made up of a master control station in Colorado Springs and a series of 5 tracking stations distributed around the equator. The Tracking Stations are located at precisely known locations and are used to determine the precise orbital parameters and the clock errors for each GPS satellite. The tracking information is sent to the master control station where it is correlated and used to generate the navigation data messages for each satellite. These messages are uplinked to the satellites. The other function of the master control station is to maintain the satellites close to their nominal orbits.

2.3.3 User Segment

The user segment is simply the collection of receivers in use at any given time.

2.4 Error Sources and Compensation

2.4.1 Ionospheric Delay

The GPS range calculation is based on a time difference and a nominal signal speed. However, the GPS signal must travel through the ionosphere, which reduces the speed of the wave modulation by an amount which depends on the local electron density. The observed effect is a delay in the arrival of the signal. This delay is also affected by the thickness of the ionosphere over the signal path. These delays produce an error in the pseudorange of about 10m.

This error is estimated by the receiver in one of two ways depending on whether P code is available or not. If P code is available then the difference in delay at both L_1 and L_2 frequencies may be calculated. The ionospheric delay is dependent on frequency and the observed ranges:

$$R_1 = R_{TRUE} + \frac{k}{f_1^2}$$

$$R_2 = R_{TRUE} + \frac{k}{f_2^2}$$

where R_1 and R_2 are the measured ranges at f_1 and f_2 respectively.

By combining and rearranging, the true range may be determined:

$$R_{TRUE} = \frac{R_1 - \left[\frac{f_2}{f_1} \right]^2 R_2}{1 - \left[\frac{f_2}{f_1} \right]^2}$$

It is not strictly true that one must have access to the P code in order to take advantage of the two frequency method of ionospheric delay correction. Certain high performance (and high cost) survey quality receivers use a cross-correlation technique for achieving access to the difference in delays at L_1 and L_2 . By knowing that both frequencies are spread by the P code and because the time offset between them is relatively small, it is possible to compare one with the other to permit decoding. [7]

In the case where the P code is not available, it is possible to mitigate the effects of ionospheric delay by using a model built into the GPS signal.

Extensive studies summarized in [8] have shown that the ionospheric delay can be modelled by:

$$\Delta t = F \left[5 \times 10^{-9} + \sum_{i=0}^3 \alpha_i \phi_M^i \left(1 - \frac{x^2}{2} + \frac{x^4}{4} \right) \right]$$

Where

$$F = 1 + 16(0.53 - E)^3$$

$$x = \frac{2 \cdot \pi \cdot (t - 50400))}{\sum_{i=0}^3 \beta_i \phi_M^i}$$

E is the satellite elevation angle (normalized to 180°)

α_i and β_i are parameters for the i th satellite and ϕ_M^i is the geomagnetic latitude of the i th satellite (normalized to 180°)

The parameters of the equation are included in the GPS navigation message.

This technique reduces the ionospheric delay error by about 50%.

2.4.2 Selective Availability (S/A)

As was mentioned in the Introduction, developments in the capabilities of C/A receivers have led the US Department of Defense to introduce artificial errors into the GPS signal. These errors consist of clock offsets and are designed to produce errors of 100m or less 95% of the time. This represents an increase in error magnitude of about 2.5/1.

2.4.3 Tropospheric Delay

Due to the fact that the index of refraction of the lower atmosphere is not quite 1 (typically 1.0003) a delay of up to 0.5 m may be introduced for low elevation satellites. This however is primarily a function of elevation angle and may be reduced by 90% by a relatively simple model.

2.4.4 Ephemeris Error

Even without S/A the broadcast positions of the satellites are in error by up to 4m. This is due to errors in tracking and random perturbations of the orbit between ephemeris updates.

2.5 Differential GPS (DGPS)

In an effort to compensate for the errors detailed above, a technique known as differential GPS is used. The basic principle is simple: place a reference GPS receiver at an accurately known location and measure the errors in the GPS signal, then transfer those error estimates to other receivers which are making position measurements at unknown locations.

The reference station normally measures and transmits errors in the pseudoranges themselves rather than simply measuring and transmitting a position error. This is because there is no way of knowing which satellites the user receiver is using, and, if the two are not using the same satellites, their position errors will not be the same. With individual pseudorange

range errors the user may select only those which are of use. For off-line measurements such as surveying applications there does not have to be any real time communication between receivers. Since all GPS measurements are very closely synchronized, the receivers may simply record the data which may then be combined at a later date using special computer software packages provided by the receiver manufacturers. For real time use, however, some sort of data link is needed to transmit the corrections from the reference to the measurement receiver.

There are two basic configurations of real time DGPS: Local Area and Wide Area. These differ in the means of error measurement and the means of transmitting the error information.

2.5.1 Local Area DGPS

In the Local DGPS, only one reference station is used and corrections are transmitted via any of several means. For aviation use VHF Comm, MF beacons and Mode S radar have been suggested. Mode S radar is an extension of secondary surveillance radar which permits two way communication of data between ground and aircraft. In Europe and North America there are also commercial DGPS services which use subcarriers on local FM broadcast stations. The improved accuracy provided by these services has not been opposed by the US DOD presumably because of their local nature and because the links could be jammed relatively easily if necessary.

The accuracy of local area DGPS at short range has been shown to be about 0.2 m both horizontally and vertically [10]. Such accuracy would allow the use of DGPS for both vertical and horizontal guidance of aircraft for all-weather landings. The accuracy depends on the distance of the user from the reference station due mainly to the decorrelation of ionospheric errors with distance. The deterioration is about 1 part in 10^6 of the base line length for single frequency and 1 part in 10^7 for dual frequency receivers. The accuracy also depends on the rate of the correction signal and degrades at about $2 \times 10^{-3} T^2$ metres where T is the period of the updating message in seconds[9].

2.5.2 Wide Area DGPS

One of the drawbacks of local area DGPS from the aviation point of view is that, to provide the coverage necessary to allow three dimensional guidance to given airfields, a local area system would have to be installed at each airfield. Also a separate receiver is required to handle the data link part. In order to avoid this, the American Federal Aviation Administration (FAA) is in the process of implementing a system which provides a much larger area of coverage per reference station and which uses the GPS L_1 frequency to transmit corrections.

The ground system consists of a network of reference stations roughly 500 nautical miles apart. As in the local DGPS, each station measures pseudorange errors, but in the wide area system these errors are transmitted to a central location (at present at the FAA Technical Center in Atlantic City NJ). Here the data are used to develop a two dimensional model of the error distribution. A grid with spacing 5° Lat x 5° Long has been defined and the GPS correction for each grid intersection is computed. The resulting array is transmitted to an INMARSAT earth station from where it is transmitted to an INMARSAT satellite for re-transmission on the GPS L_1 frequency to the area covered by the reference stations.

Preliminary flight tests performed by Transport Canada and the FAA showed that the accuracy was better than 3 meters horizontally and 6 meters vertically [11].

An additional function is the provision of integrity service which is the detection of out-of-tolerance conditions and the transmission of a warning message to users. The time permitted between detection of a failure and reception of the warning by the user is 6 seconds.

At present the test bed system consists of reference stations at Gander Newfoundland, Ottawa Ontario, Winnipeg Manitoba, Dayton Ohio, FAA Technical Center Atlantic City NJ, Anderson South Carolina, Denver Colorado, Seattle Washington, Arcata Santa Paula and Riverside California, Prescott Arizona, Elko and Columbus Nevada, Great Falls Montana, Grand Forks North Dakota, Green Bay Wisconsin, Oklahoma City Oklahoma, Green-

wood Mississippi, Miami Florida, Bangor Maine, Cold Bay, Bethel, Notzebue and Fairbanks Alaska and Honolulu and Mauna Loa Hawaii.

12.6 Interference

The potential benefits of GPS for agencies that provide navigation services to the aviation industry are very great. At present, for example, Nav Canada, the successor to Transport Canada maintains about 150 ground-based en route navigation aids many of which are in remote locations. There are also about 110 instrument landing systems. All of these systems must be maintained and calibrated on a regular basis. GPS has the potential to replace all of these especially the en route aids.

However, there are four major requirements which a navigation aid must meet before being certified for general use in instrument weather conditions. These are: accuracy, integrity availability and continuity. Accuracy is self evident and GPS more than meets the requirements especially for enroute use. Integrity was defined above and can be achieved either through an integrity channel such as the Wide Area DGPS, or by an autonomous technique like Receiver Autonomous Integrity Monitoring (RAIM) in which the receiver constantly compares position solutions using different sets of satellites to detect erroneous range data. Continuity is the probability that the system will function properly throughout the duration of a given flight procedure.

Initially it was thought that the availability would be determined by the constellation configuration. However as more and more experience is gained with the airborne use of GPS, especially with C/A code receivers, it appears that there is a potential problem with both intentional and unintentional interference. It is rather ironic that the usefulness of a system, which was designed for the military to be resistant to jamming, is being questioned for precisely that reason. T

The normal criterion for performance of communications links, bit error rate, does not re-

ally apply to GPS once initial lock has been achieved since the data (navigation message) is relatively static and the loss of bits does not have a great effect on performance. The major result of interference is loss of lock of the code tracking loop or the carrier phase locked loop. This is independent of the type of code being tracked.

Concern for the susceptibility of GPS was raised in late 1994 when a “hole” was discovered in the GPS coverage in the area around St. Louis, Missouri [12]. The hole was characterized by loss of GPS data in a funnel shaped volume with the area of outage being smaller at lower altitudes. The phenomenon lasted for several days but no definite cause was established.

In addition, there is an international political dimension to the problem in that the European countries, especially, have been reluctant to embrace GPS as the answer to all air navigation problems. In particular they are resisting the US initiative to have GPS recognized by the International Civil Aviation Organization (ICAO) as the replacement for the current VHF Instrument Landing System (ILS) as the world standard system for guidance for approach and landing in low ceiling and low visibility conditions. This is mainly because they feel that the system will not have been tested sufficiently by the time ILS replacement is needed and because MLS could be made available even though, as was mentioned earlier, MLS development in the US has been terminated. It is therefore not surprising that two of the three available papers on this subject originate in the UK. These papers are listed as references [13], [14] and [15].

References [13] and [15] attempt to derive analytical expressions for the level of interference required to force the receiver's carrier tracking loop to lose lock. The argument used in reference [15] is that the carrier tracking loop is used to aid the code tracking Delay Locked Loop which has a low bandwidth ($\sim 1\text{Hz}$) and hence can not follow aircraft manoeuvres by itself. Thus the critical point is maintaining carrier loop lock. It is then stated without proof or reference that the carrier loop requires a 16dB SNR to maintain lock. Also the processing gain is defined as the ratio of the input signal bandwidth to the carrier loop tracking bandwidth (assumed to be 20Hz and thus the “Processing Gain” is $2\text{MHz}/20\text{Hz} = 50\text{dB}$). Note that normal processing gain for the C/A code is $2\text{MHz}/50\text{Hz} = 43\text{ dB}$. Finally,

it is stated the antenna received power of the desired signal is -157 dBW for a 0 dBi antenna. The expected level at which an interfering signal would cause loss of lock is:

$$-157\text{dBW} + 50\text{ dB} - 16\text{dB} = -123\text{ dBW}$$

However Owen then adjusts this to -130dBW and justifies this by saying “Because of their short length and 1ms repetition interval, the C/A codes produce a line spectrum that leads to peaks in the cross-correlation product and false lock detections, effectively reducing the signal processing gain by several dB.”

Flight tests described in reference [15] consisted of radiating a jamming signal at an aircraft as it flew a racetrack pattern as shown in Figure 2.3. The jamming signal was an L_1 carrier modulated with FM noise at a 2 MHz bandwidth.

Figure 2.3
Geometry of Jamming Flight Trials [15]

Figure 2.4
Results of Airborne Jamming Trials [15]

Results are shown in Figure 2.4. As can be seen, it is difficult to determine the power level at which lock was lost but it appears to be between -125dBW and -130dBW.

Ground tests were performed using CW, FM modulated carrier and AM and FM noise but it is reported that no significant differences in jamming threshold were reported.

Unfortunately the receiver tested was not identified either by make or by type.

Reference [13] evaluated two commercially available navigation receivers, the Garmin GPS 75 and the Magellan 5000D. The Garmin is a single channel, time multiplexed receiver while the Magellan is a 5 parallel channel type. The theoretical analysis follows that of Owen except that it uses a processing gain of 43 dB (by using the data rate of 50 Hz). This results in a jamming threshold of -130dBW or -100dBm.

The laboratory tests consisted of subjecting the receivers to a jamming signal radiated from

an antenna 45" directly above the receiver antenna. Tests were done with the following types of modulation:

- a) CW, pulsed at prfs of 10kHz, 500kHz and 2 MHz at various pulse widths
- b) FSK at 1.023 MHz rate with deviations of 2.0 MHz and 4.0 MHz and 2 MHz rate with deviations of 4.0 MHz and 8.0 MHz.

The results of the FSK tests are shown in Table 2.3. Note that, for direct comparison with the previous paper's results, the LAS (lost all satellites) column should be used and 30 dB should be subtracted to convert to dBW.

Results of the pulsed jammer tests are shown in Figure 2.5a and 2.5b.

While the results are in general agreement with the levels obtained in [15] (-130dBW), there is significant difference in their assessment of the effects of different types of jamming modulation.

Table 2.3 Results of FM Jamming Tests

FM Rate	FM Deviation	Garmin GPS 75		Magellan 5000D	
(MHz)	(MHz)	Threshold (dBm)	LAS (dBm)	Threshold (dBm)	LAS (dBm)
1.023	2.0	-110	-101	-107	-99
1.023	4.0	-105	-94	-102	-93
2.0	4.0	-107	-99	-107	-99
2.0	8.0	-97	-86	-90	-84

LAS: Lost All Satellites

Figure 2.5a
Pulsed Jamming Results for Garmin GPS 75
Receiver

Figure 2.5b
Pulsed Jamming Results for Magellan
5000D Receiver

Reference [14] describes comprehensive tests done on five representative types of receiver. No theoretical analysis is offered but the results of laboratory and airborne tests are described. The receivers tested are described in Table 2.4.

Table 2.4 Types of GPS Receivers Tested in Reference [14]

System	GPS Type
GPS No.1	standard correlator, carrier phase tracking, single level A/D, coherent system, dual frequency
GPS No.2	standard correlator, single level A/D, non-coherent system, aviation unit
GPS No.3	standard correlator, single level A/D, non-coherent system, DGPS station
GPS No.4	narrow correlator, multi-level A/D, non-coherent system, interference detector
GPS No.5	standard correlator, multi-level A/D, coherent system

Unfortunately the type of modulation used is not defined. It is simply stated that “The RF field was modulated using various type of modulation to obtain the worst case susceptibility profile”.

Laboratory tests were done by radiating the jamming signal from an antenna at an elevation angle of 60° and a distance of 1m from the GPS antenna. The field strength was measured by an isotropic field sensor. Since the results of these tests are given in field strength terms (dB referred to 1μV/m) it is necessary to convert these to power output from the antenna to provide a direct comparison with the previous papers’ results. The power density for a given field strength is given by

$$P = \frac{s_f^2}{120\pi} 10^{-12}$$

where P is the power density (W/m²)

s_f is the field strength (μV/m) and

120π is the impedance of free space

To obtain the antenna power output for a given incident power density it is also necessary to multiply by the effective antenna area. In the absence of any specific detail from the paper a value of -22dB is used. This is the value provided in reference [13].

Thus the conversion factor is $-120\text{dB} - 10 \log_{10}(120\pi)\text{dB} - 22\text{dB} = -168\text{dB}$

The results of the laboratory tests are given in Table 2.5

Table 2.5 Results of Laboratory Tests

GPS Under Test	Field Strength for Loss of Lock (dB μ V/m)	Power Density for Loss of Lock (dBW/m ²)	Antenna Power Output for Loss of lock (dBW)
GPS No.1	30	-116	-138
GPS No.2	40	-106	-128
GPS No.3	49	-97	-119
GPS No.4	53	-93	-115
GPS No. 5	55	-91	-113

Thus the receiver architecture appears to have a great influence on the level of susceptibility.

Susceptibility tests were also performed on a NovAtel Model 3951 at Carleton University as part of a 4th year engineering project [16]. The RF power densities for loss of all satellites were found to be -84 dBW/m² for CW, -89 dBW/m² for FM with 10 kHz deviation and -89 dBW/m² for 100% AM sine modulation at 10 kHz. This compares reasonably well with the above results for GPS receiver #4 which is obviously a NovAtel (being the only one with a narrow correlator).

Chapter 20 of reference [17] includes a comprehensive review of interference sources and mechanisms. It also covers techniques for mitigating the effects of interference.

At the 1998 ICAO conference in Rio de Janeiro, the FAA announced that it could not support the use of GPS as a “sole source” (no other system required) navigation system due to the potential for interference. This prompted a study of the problem by Johns Hopkins University. Their report [18] concludes that with the implementation of regulatory procedures

and receiver improvements, the risk could be reduced to an acceptable level. The methods and conclusions are being questioned, however, [19] and the situation remains unresolved at this time.

2.7 Conclusions

While GPS is a system with a large and growing number of applications, its use in aeronautical navigation is still open to many questions especially in the area of approach and landing guidance. This may cause problems in the near future due to the wholehearted commitment of the US and the FAA to this system and the reluctance of other countries to follow this path. Since there is growing pressure to replace the current ILS and since airlines will have to equip (at great cost) their aircraft with receivers compatible with the replacement system, there will undoubtedly be interesting times ahead. Thus it is expected that there will be great activity in efforts to increase the jam proofing of the GPS and, looking farther down the road, to developing its successor.

Chapter 3

Direction Finding

3.0 Introduction

Direction finding of radio frequency sources is as old as radio itself. It has been used by the military for determining the position of enemy communications and radar transmitters and has been used in peacetime for navigation and the control of illegal broadcasting stations. Because of the many applications, a large number of techniques and devices have been developed [20]. Generally these techniques fall into one of 3 groups: Polar Diagram DF, DF by phase measurement and Doppler DF.

3.1 Polar Diagram DF

The polar diagram technique exploits antenna directivity to determine the angle of arrival (AOA). Either an antenna pattern maximum (peak) or a null may be used. The peak method has the advantage of permitting the measuring receiver to operate at maximum signal strength but it requires an antenna with an unambiguous main lobe and the accuracy is proportional to the width of that lobe. Since the width of the lobe is inversely proportional to the antenna aperture, high accuracy systems require large antennas (or antenna arrays).

Figure 3.1
The Effect of Signal to Noise Ratio on Null
Direction Finding Resolution

The null technique may be implemented using small antennas (e.g. airborne Automatic Direction Finder (ADF) which uses a rotatable loop) but since it uses the antenna minimum the receiver operates on a low signal level and the accuracy is limited by the receiver sensitivity and the signal to noise ratio as shown in Figure 3.1.

In both of these cases the antenna has to be rotated to place the emitter at either the maximum or the minimum.

Excessive size and mechanical complexity has limited the use of polar diagram DF in airborne applications. One notable exception has been the ADF, a medium frequency band (200 - 500 kHz) navigation system widely used in World War II. Until recently the antenna used in this system was a rotatable loop which provided accuracies in the range of 3° to 5° . Lately this antenna has been replaced by a small commutated doppler array.

3.2 Phase Measurement DF

Direction Finding by phase measurement, as the name implies uses the difference in signal phase of the outputs from two or more antennas. The phase measurement is unambiguous as long as the antenna separation l is greater than the wavelength λ .

Figure 3.2
Phase Measurement

In Figure 3.2 the phase between the antenna outputs will be $2\pi \cdot \frac{l}{\lambda}$ radians.

but $l = d \cdot \cos\theta$ where θ is the angle of arrival to be measured.

thus $\varphi = \frac{2\pi \cdot d \cdot \cos\theta}{\lambda}$ where φ is the phase between the antenna outputs.

and $\cos\theta = \frac{\varphi \cdot \lambda}{2\pi \cdot d}$

The sensitivity varies with the angle of arrival, with $\frac{d\varphi}{d\theta}$ being proportional to $\sin\theta$.

If the spacing between array elements is small, errors in bearing measurement may result from wave distortion due to reflections as shown in Figure 3.2

Figure 3.3
Errors Due to Wave Distortion

3.3. Doppler DF

A modification of the phase measurement technique uses a circular array of antenna elements. A commutator connects opposite pairs of elements to the receiver in a sequential manner. The output signal from this system is almost the same as that from an arrangement in which two antennas are mounted at either end of a rod and the rod is rotated in a horizontal plane about its centre. The resulting signal at the receiver input is frequency modulated sinusoidally, due to the Doppler shift resulting from the relative motion of the antennas and the signal source. In the case of the commutated array, the rate of rotation is equal to the angular rate of the commutation. The frequency deviation is proportional to the product of the angular rate of commutation and the radius of the array. The direction of the source relative to the primary axis of the array is determined by the phase angle between the received frequency modulation and the commutation.

3.4 Position Fixing with DF Systems

The ultimate objective of this system is to determine the position of an interfering source. However the output of this device is only a relative bearing to the source. To determine the location of the source more information is required. Information from systems on board the aircraft includes the aircraft heading, the position and the velocity. This information, combined with the relative bearing can be used to determine position as follows.

As shown in Figure 3.4 the position of the source relative to the aircraft can be computed from the true bearing of the source (θ_T) and its range (R). θ_T is the sum of the aircraft heading (ϕ) provided by the aircraft's navigation system and the relative bearing (θ_R) as determined by the direction finder.

The range R can be calculated from the relationship between the rate of change of bearing and the component of the aircraft velocity perpendicular to the line joining the aircraft and the target. In Figure 3.5, the change in bearing $d\theta$ is related to the change in position along the perpendicular to the bearing ds by the equation $d\theta = \frac{ds}{R}$ or $R = \frac{ds}{d\theta}$.

Thus, using rates of change with respect to time $R = \frac{\left(\frac{ds}{dt}\right)}{\left(\frac{d\theta}{dt}\right)}$.

$\frac{ds}{dt}$ is $V \cos \alpha$ the component of aircraft velocity perpendicular to R where α is true bearing (θ) minus true heading (ϕ).

Once the position of the source relative to the aircraft has been established, the absolute position can be determined since the absolute position of the aircraft is known. Note that the rate of change of bearing is maximum when the relative bearing is either 90° or 270° .

The direction finder described in this thesis may thus be used, in conjunction with other equipment on board the investigating aircraft, to determine the position of an interferer.

Figure 3.4
Geometry of Position Fix

Figure 3.5
Determination of Range from Rate of Change of Bearing

Chapter 4

Proposed System

4.0 System Requirements

The main constraint in the design of the direction finder stemmed from the fact that the system was to be installed on an aircraft. This severely limited the amount of area available for the sensor and also imposed restrictions on the sensor itself in that it had to be certifiable as an airworthy installation.

The format of the system was originally specified as a PC interface board but this was found to be impractical. Subsequently, the objective became minimum weight and volume.

The DF accuracy objective was to achieve the minimum possible error less than 10° since the more accurate the bearing the less flying time would be required to locate the radiating source.

Successful detection and bearing measurement were to be achieved at the minimum interfering level of -105dBm received from the GPS antenna and the minimum output data rate was to be 1 Hz.

4.1 System Design

4.1.1 Measurement Technique

A review of the literature on direction finding [20] showed that there are two main techniques in general use which do not require rotating antennas. These are

Phase Measurement
Doppler

. Doppler was thought to be impracticable because it usually employs large arrays, but Doppler using a minimal array is still being pursued as a possible alternative

The most promising was Phase Measurement with either a four or three element array and thus the system architecture shown in Figure 4.1. was adopted.

Figure 4.1.
System Architecture

4.1.2 Antenna Array

Antenna arrays of 4 and 3 elements in the configurations shown in figure 4.2. were considered.

Figure 4.2
Candidate Antenna Arrays

4.1.3 Analysis for the 4 Antenna Array Using Phase Measurements

In the 2 dimensional case, i.e if the source is in the same plane as the array

$$\Delta t_{13} = \frac{L \cos \theta}{c} \quad \Delta t_{24} = \frac{L \cos(\theta - 90)}{c}$$

$$\Delta t_{12} = \frac{0.707L \cdot \cos(\theta + 45)}{c}$$

$$\Delta t_{23} = \frac{0.707L \cdot \cos(\theta - 45)}{c}$$

where Δt_{ij} is the time for the interfering signal to travel from antenna i to antenna j.

L is the antenna spacing as shown in Figure 4.2

θ is the angle of arrival of the interfering signal and

c is the speed of light

converting these to phase differences where φ_{ij} is the phase difference between the outputs of antennas i and j:

$$\varphi_{13} = \frac{L \cos \theta c}{\lambda} \quad \text{or if L is made equal to } \lambda, \frac{L}{\lambda} = 1$$

$$\text{then } \varphi_{13} = \cos(\theta)$$

similarly

$$\varphi_{24} = \cos(\theta - 90) = \sin \theta$$

$$\varphi_{12} = 0.707 \cdot \cos(\theta + 45)$$

$$\varphi_{23} = 0.707 \cdot \cos(\theta - 45) = 0.707 \cdot \sin(\theta + 45)$$

In the 3 dimensional case, when the source is not in the plane of the array, the effective spacing of the array and hence the measured phase angle is reduced as a result of the depression angle α

Figure 4.3
Illustration of Effective Length

Therefore

$$\varphi_{13} = \frac{L \cos \alpha \cos \theta}{c} \frac{c}{\lambda} = \frac{L \cos \alpha \cos \theta}{\lambda} = \cos \alpha \cos \theta \quad (4.1)$$

similarly

$$\varphi_{24} = \cos \alpha \sin \theta \quad (4.2)$$

$$\varphi_{12} = 0.707 \cdot \cos \alpha \cos (\theta + 45) \quad (4.3)$$

$$\varphi_{23} = 0.707 \cdot \cos \alpha \sin (\theta + 45) \quad (4.4)$$

Dividing (4.2) by (4.1)

$$\frac{\varphi_{24}}{\varphi_{13}} = \frac{\sin \theta}{\cos \theta} = \tan \theta$$

and dividing (4.4) by (4.3)

$$\frac{\varphi_{23}}{\varphi_{12}} = \frac{\sin (\theta + 45)}{\cos (\theta + 45)} = \tan (\theta + 45)$$

Finally

$$\theta = \text{atan}\left(\frac{\varphi_{24}}{\varphi_{13}}\right) \text{ and } \theta = \text{atan}\left(\frac{\varphi_{23}}{\varphi_{12}}\right) - 45$$

Note that antenna pairs 4-1 and 3-2 give the same result as the latter equation providing a third redundant measurement.

Due to the fact that their spacing is $0.707L$, the sensitivity of pairs 1-2, 2-3, 3-4 and 4-1 is reduced by a factor of 0.707 compared to 1-3 and 2-4 and the standard deviation of their errors is increased by a factor of 1.404. This is compensated for by the fact that they form redundant pairs and thus the mean of each pair of measurements has a standard deviation which is 0.707 of the standard deviation of each.

4.1.4 Analysis for the 3 Antenna Array using Phase Measurements

In the two dimensional case where the source is in the same plane as the antenna array

$$\Delta t_{21} = \frac{L \cos(\theta - 30)}{c} \quad \Delta t_{13} = \frac{L \cos(\theta + 90)}{c} \quad \Delta t_{32} = \frac{L \cos(\theta - 150)}{c}$$

where Δt_{ij} , θ and c are as defined in the 4 antenna analysis.

converting to phase differences

$$\varphi_{21} = \frac{L \cos(\theta - 30) c}{\lambda} \text{ or } \varphi_{21} = L \frac{\cos(\theta - 30)}{\lambda}$$

Similarly

$$\varphi_{13} = L \frac{\cos(\theta + 90)}{\lambda}$$

and

$$\varphi_{32} = L \frac{\cos(\theta - 150)}{\lambda}$$

now since $\cos(a + b) = \cos(a) \cdot \cos(b) - \sin(a) \cdot \sin(b)$

$$\varphi_{21} = \frac{L}{\lambda}(A \cos(\theta) - B \sin(\theta)) \text{ where } A = \cos(-30)^\circ \text{ and } B = \sin(-30)^\circ$$

$$\varphi_{13} = -\left(\frac{L}{\lambda} \sin(\theta)\right) \quad \text{and}$$

$$\varphi_{32} = \frac{L}{\lambda}(C \cos(\theta) - D \sin(\theta))$$

$$\text{but } C = \cos(-150)^\circ = -A \quad \text{and} \quad D = \sin(-150)^\circ = B$$

Thus

$$\varphi_{32} = \frac{L}{\lambda}(-A \cos(\theta) - B \sin(\theta)) \quad (3)$$

Therefore

$$(\varphi_{32} - \varphi_{21}) = \frac{L}{\lambda}(-2A \cos(\theta)) \text{ and}$$

$$\frac{\varphi_{13}}{\varphi_{32} - \varphi_{21}} = \frac{-\sin(\theta)}{-2A \cos(\theta)} \left(\frac{L}{\lambda} \cdot \frac{\lambda}{L} \right) = \frac{\tan(\theta)}{2A}$$

or

$$\theta = \text{atan}\left(\frac{2A\varphi_{13}}{\varphi_{32} - \varphi_{21}}\right)$$

As in the 4 antenna case, when the source is not in the plane of the array, the effective spacing of the array and hence the measured phase angle is reduced as a result of the depression angle α . However, since the wavelength does not appear in the final equation the depression angle does not affect the calculation of bearing. It does, however, affect the sensitivity of the phase angle measurement and hence the final accuracy.

4.1.5 Comparison of Accuracy of 3 and 4 Antenna Configurations

In order to obtain an estimate of the phase measurement accuracy required, a Monte Carlo analysis was performed using the 3 and 4 antenna equations. The results for a phase error standard deviation of 10 degrees are shown in Figure 4.4

Figure 4.4
Comparison Between the Error Characteristics of the 4 Antenna and
3 Antenna Arrays

Since there was little difference between the two sets of results, especially at 90 and 270 degrees relative bearing where a running position fix is most effective, the 3 element array was chosen

4.1.5 Amplification

To obtain reasonable sensitivity from the tracking filter phase locked loop mixer the input signal level should be a minimum of -20 dBm. Approximately 60dB of amplification (in addition to antenna preamplification) is required to provide this level to the next stage. This is provided by a 60 dB low noise gain stage. Signal levels are shown in Figure 4.5.

Figure 4.5
Requirements for Amplification

4.1.6 Input Filter and Signal Acquisition

In order to improve the signal to noise ratio, provide a stable signal for phase measurement and permit the operator to select the signal that he wishes to locate, an input filter and acquisition block is provided. This consists of one phase locked loop (PLL) for each antenna input plus some switching circuitry. The phase locked loops act as tracking filters which have a very narrow bandwidth to improve the signal to noise ratio but also are capable of following a drifting signal, or providing the average of a frequency modulated signal.

In this mode the output of the master channel is connected to the inputs of the other channels. Thus all three channels are tuned to the same frequency and the VCO outputs are in phase.

4.1.6.1 Signal Acquisition

To ensure that the phase locked loops are all tracking the same signal and to provide the operator with a means of selecting a particular signal to be located, a manual acquisition mode is employed.

One phase locked loop is designated as the master and a portion of its output is provided to

the operator's spectrum analyzer as well as to the inputs of the other two filters. The operator can control the tuned frequency by means of a bias control voltage in the master channel loop. Once the operator is satisfied with the selection of frequency and that the master is locked on to the correct signal, the two slave channels are disconnected from the master and are connected to their respective antenna outputs. Each is then free to seek the correct phase as provided by its antenna. Note that during the acquisition process, it may be necessary to provide a higher frequency low pass filter which could be switched into the loops to allow for the higher slewing speeds. A lock indicator for all channels is to be provided as part of the internal validity checking.

Figure 4.6
Signal Acquisition Technique

4.1.7 Phase Measurement

In order to avoid the extra complexity and phase error contribution from downconversion, it was decided to implement the phase measurement at the RF frequency. This is done with a phase comparator as shown in figure 4.7.

Figure 4.7
Phase Comparator

Figure 4.8.
Phase Measurement

This can be implemented using microstrip construction techniques with the 90° phase shift being simply an extra length of line. This is possible since the span of frequencies is quite small (about 0.1% of centre frequency).

The outputs of the phase comparator are DC voltages which are proportional to the sine and

cosine of the phase difference between the two RF inputs. These are followed by analogue to digital converters which provide the input to the microprocessor as shown in Figure 4.8

4.1.8 Bearing Calculation and Interface Control

Once the sines and cosines of the phase differences are available in digital form it is relatively simple to compute the relative bearing using the on board microcomputer. The microcomputer also looks after the processing of commands from the host processor and also the transmission of the bearing and status data to the host computer.

Chapter 5

System Component Design

5.0 Introduction

The major part of the work in this thesis was devoted to the design, construction and testing of the three most critical components of the proposed system: the antenna array, the phase comparator and the tracking filter.

5.0.1 Phase Measurement

Measuring phase difference at low frequencies is relatively easy. By using zero crossing detectors the period of the two signals and the time shift between them may be determined. The ratio of the time shift to the period multiplied by 360° is then the phase shift. This technique may be applied to L band signals down converting the signal to a sufficiently low intermediate frequency using a common local oscillator.

Measuring the phase shift directly at the L band frequency involves splitting one of the signals into inphase and quadrature components and then mixing them with inphase components of the other signal. The outputs of the mixers, after low-pass filtering are DC levels proportional to the sine and cosine of the phase difference. Thus by dividing the sine output by the cosine output and computing the arctangent of the result, the phase angle may be obtained.

Figure 5.1
Phase Comparator

To minimize the complexity of the overall system and to decrease the incidence of unwanted phase shifts, the direct phase measurement technique was selected.

During the course of the development two phase comparator designs, designated initial and final, were built and tested. Although the initial design was found to be unsuitable it is included here because it has some features which might be of use in further developments.

5.0.2 Phase Locked Loop Tracking Filter

A tracking filter implemented by a phase locked loop was chosen for the following reasons:

- a. to maintain a constant input level to the phase comparator
- b. to permit the selection of a particular interfering signal
- c. to provide an external indication of the frequency to which the filter is tuned
- d. to increase the signal to noise ratio of the phase comparator inputs

The error characteristics of the direct phase measurement technique described above are dependent to a large degree on the magnitude of the two input signals. To stabilize the errors and thus to permit calibration and correction it is therefore necessary to ensure that the phase comparator input signals are maintained at a constant power level.

To accommodate the case where more than one interfering source is found it will be necessary to differentiate between them, thus a controllable means of selecting the desired signal is necessary.

The overall design philosophy requires that operator intervention and control be used wherever appropriate. The phase locked loop tracking filter provides the possibility of providing an output signal whose frequency is that to which the filter is tuned. This allows the operator, with the aid of a spectrum analyzer, to guide the direction finder to the desired signal frequency. It also allows the operator to monitor the performance of the system.

Phase measurement to the accuracy needed in this system is not possible at the signal to

noise ratio available at the output of the RF amplifiers (about 2 dB). It is therefore necessary to increase the signal to noise ratio. One way of increasing the S/N ratio is to pass the incoming signal through a filter with sufficiently narrow bandwidth. Standard bandpass filters present problems in this application including (1) sufficiently narrow bandwidths ($<0.01\%$) are difficult to achieve and (2) to cover the band of interest, the filter must be tunable and must be able to track drifting signals. Fortunately, in this application, the filter is required to preserve only the phase and not the magnitude of the input signal and thus a phase locked loop designed as a tracking filter will satisfy the requirement. Such a tracking filter has the additional advantage of being able to provide an auxiliary output at its tuned frequency even without an input signal. This provides the capability of steering the filter to a desired signal.

5.1 Antenna Array Design

The basic sensor selected for the antenna array was the NovAtel Model 511 Active GPS Antenna. This antenna is built specifically for installation on aircraft and is certified [21] for such installation by the US Federal Aviation Administration (FAA) under TSO Technical Standard Order) C 115a. The antenna includes a built-in low noise amplifier with a gain of 26 dB $\pm$ 3 dB. The amplifier power is supplied through the centre conductor of the RF coaxial cable. The nominal bandwidth is 4 MHz.

Three antennas were mounted on a 0.6m x 1 m aluminum ground plane to simulate the aircraft skin in the operational configuration. The mounting configuration was an equilateral triangle with 190mm sides. All antennas were mounted in the same orientation. The array is shown in Figure 5.2

Figure 5.2
Direction Finder Antenna Array

5.2 Phase Comparator Design

5.2.1 Initial Phase Comparator

The main components of the phase detector are the power splitters and the mixers. A survey of the RF component market initially failed to locate a 90° power splitter which was specified for operation in the 1.5 GHz range and thus it was decided to implement this function using microstrip techniques. The splitters themselves were implemented as Wilkinson dividers [22] while the 90° phase shift was to be introduced as a $\lambda/4$ delay line. This was, however, subsequently modified as discussed below.

The configuration of the Wilkinson divider is as shown in Figure 5.3.

Figure 5.3
Wilkinson Divider Design

To maintain the correct impedance match at the input, the semicircular arms should have an characteristic impedance of 70.7 ohms and the RF paths from the intersection of the output segment to the 100 ohm resistor should be a quarter wavelength. Since this path is irregular it is difficult to determine the actual electrical length by inspection. To obtain the correct dimensions of the semicircular sections, the circuit was reproduced in Libra, an RF simulation program. Libra simulation was then used to adjust the radius of the semicircular arms until the circuit s_{11} was a minimum at the frequency of interest (1.57542 GHz). This dimension was then used to lay out the microstrip design for the test circuit.

The original idea of implementing the 90° phase shift by using a quarter wavelength delay line was modified to reduce the size of the layout. The original configuration is shown in Figure 5.4

Figure 5.4
Original Phase Comparator Configuration

Primarily to reduce the size of the circuit it was decided to eliminate the redundant $\lambda/8$ sections as shown in Figure 5.5

Figure 5.5
Modified Phase Comparator Configuration

This has the effect of shifting the measured angle by 45° as follows:

Top Leg:

$$V_{COS} = \cos(\omega t) \cdot \cos\left(\omega t + \theta + \frac{\pi}{4}\right) = \frac{1}{2} \left(\cos\left(-\frac{\pi}{4} - \theta\right) + \cos\left(2\omega t + \frac{\pi}{4} + \theta\right) \right)$$

which, when low pass filtered gives

$$\frac{1}{2} \cos\left(-\frac{\pi}{4} - \theta\right) = \frac{1}{2} \cos\left(\theta + \frac{\pi}{4}\right)$$

Bottom Leg

$$V_{SIN} = \cos\left(\omega t + \frac{\pi}{4}\right) \cdot \cos(\omega t + \theta) = \frac{1}{2} \left(\cos\left(\frac{\pi}{4} - \theta\right) + \cos\left(2\omega t + \frac{\pi}{4} + \theta\right) \right)$$

which, when low pass filtered gives

$$\frac{1}{2} \left(\cos\left(\frac{\pi}{4} - \theta\right) \right) = \frac{1}{2} \sin\left(\frac{\pi}{4} - \theta + \frac{\pi}{2}\right) = \frac{1}{2} \sin\left(\theta + \frac{\pi}{4}\right)$$

Thus

$$\frac{V_{SIN}}{V_{COS}} = \tan\left(\theta + \frac{\pi}{4}\right)$$

or

$$\text{atan}\left(\frac{V_{SIN}}{V_{COS}}\right) = \frac{\pi}{4} + \theta$$

finally

$$\theta = \text{atan}\left(\frac{V_{SIN}}{V_{COS}}\right) - \frac{\pi}{4}$$

Thus the offset can be accounted for by subtracting 45° from the measured angle and thus does not pose a problem in this application. An additional benefit of this modification is that it makes the circuit pattern symmetrical which cancels out some of the errors introduced by irregular track shapes.

The selection of the device to perform the mixer function was influenced by the RF power levels available and the requirement for low distortion.

The RF power level at the mixer is determined by the power output of the tracking filter VCO and the fact that each tracking filter supplies RF to two phase comparators. Thus there is an inherent 6 dB loss plus insertion losses which are estimated to be a further 3 dB. Surface mount VCOs with supply voltages in the 10V range have output powers of about 10dBm so that the power at both the RF and Local Oscillator (LO) inputs of the mixer is about 0 dBm.

The two devices considered were a balanced mixer and a Gilbert Multiplier model IAM 81028 manufactured by Hewlett Packard.

The advantage of the balanced mixer is simplicity since the device does not need any external power. It has the following disadvantages:

- a. it requires a relatively high power level into the LO port, typically a minimum of 3dBm
- b. the RF and especially the LO ports are not matched
- c. the inherent conversion loss results in a low level output signal

The Gilbert mixer on the other hand provides gain instead of loss and its RF and LO ports

are relatively well matched with the only major disadvantages being that it requires external power and that the output is offset by a DC voltage which is dependent on the supply voltage and the RF input level.

The Gilbert Mixer was chosen for the initial design and a simple test circuit was constructed and tested to determine the suitability of this device before the first prototype DF RF board was designed.

The circuit was implemented using Rogers Duroid 5870, 32 mil substrate as shown in Figure 5.5. The 50 ohm track width was 90.2 mils and the 70.7 ohm $\lambda/4$ transformer track width was 45.2 mils. The radius of the $\lambda/4$ transformer track was 405 mils.

The results of the tests on this board showed that the phase error variation was about 5° and that the variation over time was small so that, with calibration, the error in phase measurement could be reduced to about 2° .

As a result, a full scale RF board for the Direction Finder was designed. The phase comparator part of the board is shown in Figure 5.6. As can be seen, the layout mirrors the physical arrangement of the antennas. This is due to the use of symmetry to ensure that the phase lengths of the RF path are equal and that the signals are all subject to the same irregularities.

Because of the initial restriction of board size (specifically that the width should be 4.5" or less) it was necessary to reduce the physical size of the individual phase detector and divider circuits. Since the size of the features depend to a large extent on the wavelength of the signal in the microstrip it was necessary to decrease the wave velocity by a factor of 2 which necessitated an increase in the substrate relative permittivity from 2.5 to about 10.

Rogers Duroid 3010, with a relative permittivity of 10.2 and a thickness of 25 mils was chosen as the substrate for this purpose. As before, Libra was used to determine the dimensions of the microstrip structures. The 50 ohm track width was 45.2 mils and the 70.7 ohm $\lambda/4$ transformer track width was 15.5 mils. The radius of the $\lambda/4$ transformer track was 240 mils

The length of the phase comparator was reduced from 4.25 in to 2.15 in. and the remainder of the board was reduced proportionally.

Figure 5.6
Initial Phase Comparator Board Layout

To obtain the high relative permittivity, the substrate incorporates a high proportion of Teflon. This factor, and the resulting narrowness of some of the microstrip tracks, in particular the semicircular arms of the Wilkinson Dividers, made production of this board very difficult and expensive. Special procedures required to produce plated through holes increased the processing cost and semicircular sections of the Wilkinson dividers regularly peeled off the board during the final cleaning procedure. Only one board was built successfully and this one was not robust.

5.2.2 Final Phase Comparator

Because of the fragility of the initial board design, it was decided to relax the board size restrictions and to proceed with a second prototype using Rogers RO4003 32 mil substrate which can be processed using normal PCB manufacturing procedures. In addition, the circuit design was changed as a result of the discovery of commercially available 0° and 90° power dividers which operate in the 1.5GHz frequency range.

This board, shown in Figure 5.7 was manufactured and was tested to determine stability of the phase measurement error characteristics over time.

Figure 5.7
Final Phase Comparator Board

5.2.3 Phase Comparator Errors

The accuracy of the phase comparator depends primarily on the accuracy of the 90° phase shift, the symmetry of the two signal paths and the distortion introduced by the mixers.

The accuracy of the 90° phase shift and the asymmetry of the signal paths produce the same effect since, if the 90° phase shift is actually 91° it produces the same effect as if the phase shift were 90° and the signal path were 1° longer.

The phase measurement involves the calculation:

$\theta = \text{atan}\left(\frac{V_{SIN}}{V_{COS}}\right)$ where V_{SIN} and V_{COS} are proportional to the sine and cosine of θ respectively.

If there is an extraneous phase shift of δ in the sine channel, for example, then the resulting measured angle is $\text{atan}\left(\frac{\sin(\theta + \delta)}{\cos(\theta)}\right)$

The resulting error for various values of δ was determined using a MATLAB program and are shown in Figure 5.8

Figure 5.8
The Effect of Path Differences in Sine and Cosine Channels

This shows that the error is approximately sinusoidal for small path differences with the maximum error being close to the path difference. Note that the error trace becomes more distorted with increasing path difference. The first maximum of the 15 degree trace is 15.06° and it occurs at 34° rather than at the 45° for the 5 degree trace.

Since this error originates in the physical layout of the board it will be a constant and it will be possible to compensate for it by calibration and thus its contribution to the overall sys-

tem error will be limited to the accuracy of the calibration procedure.

If the outputs of the mixers are distorted (i.e. depart from pure sine or cosine), then the resulting angle measurement will be in error. Such distortions are a function of the input RF power levels and the power supply voltage (V_{CC}). It was therefore necessary to perform experiments to determine the optimum levels for this device (see chapter 6)

5.3 Tracking Filter Design

The main purpose of the tracking filter is to provide to the phase comparators a signal whose phase is the same as that of the input signal, has a constant power level, and has a high signal to noise ratio.

The output power level is primarily a function of the power supply voltage and thus it was necessary to determine the sensitivity of the power output to changes in the supply voltage.

The critical design characteristics of the filter are the phase error, output phase signal to noise ratio, the ability to maintain lock under low input signal to noise conditions and the pull-in or acquisition range.

The initial design parameters were:

Acquisition range: 1MHz

Output Phase Noise Standard Deviation(σ_{ϕ_0}): $5^\circ = 0.0873$ rad

Input Signal Level: -20 dBm

Input Signal to Noise Ratio: 3 dB

The input signal to noise ratio was determined by using the minimum reported jamming signal (-105dBm) and assuming an antenna/LNA bandwidth of 4MHz.

Thus with $k = 1.381 \times 10^{-23}$ J/K

$T = 300K$

$$B=4 \times 10^6 \text{ Hz}$$

$$kTB=1.658 \times 10^{-14} \text{ W} = -137.8 \text{ dBW} = -107.8 \text{ dBm}$$

Therefore: $S/N = -105 - (-107.8) = 2.8 \text{ dB}$

From Reference [23]

$$\sigma_{\Phi_0}^2 = \frac{N_0 \omega_n}{A^2 4 \zeta} \left[1 + \left(2\zeta - \frac{\omega_n}{K} \right)^2 \right]$$

where A^2 = signal power,

K = loop gain,

ω_n = natural frequency of the loop

ζ = damping factor

and N_0 = Noise power density

Assuming $\frac{\omega_n}{K} \ll 1$ and $\zeta \approx 1$

$$\omega_n \approx \frac{A^2 4}{N_0 5} \sigma_{\Phi_0}^2$$

Now $A^2 = 3.162 \times 10^{-11} \text{ mW}$

and $N_0 = 1.381 \times 10^{-23} \times 300 = 4.166 \times 10^{-21} \text{ W/Hz} = 4.166 \times 10^{-18} \text{ mW/Hz}$

thus $\frac{A^2}{N_0} = 7.62 \times 10^6$

and $\omega_n = 0.8 \times 7.62 \times 10^6 \times (0.0873)^2 = 46468 \text{ rad/s} = 7395 \text{ Hz}$

The VCO chosen for the prototype was the Z-Communications model V618SE03

This device has the following characteristics:

Frequency range: 1570-1680 MHz

Power Output: $12 \pm 2.5 \text{ dBm}$

Tuning Sensitivity (K_1): $30 \text{ MHz/V} = 188 \text{ Mrad/V}$

Supply Voltage: 10V

The phase detector chosen for this implementation was the same one that was used in the phase comparator circuit. The sensitivity (K_3) was measured with the RF input at -20dBm and the Local Oscillator input at 0 dBm. The result was 300mV/rad

Therefore, assuming no gain in the filter, $K=K_1K_3=(0.3 \times 188 \times 10^6)=5.6 \times 10^7$.

Thus
$$\tau_1 = \frac{K}{\omega_n^2} = \frac{5.6 \cdot 10^7}{(46468)^2} = 2.612 \cdot 10^{-2} s$$

The filter used was a low pass with phase lead correction.

$$\tau_2 = \frac{2 \cdot \zeta \cdot \omega_n \cdot \tau_1 - 1}{K} = \frac{2 \cdot 1 \cdot 46468 \cdot 2.612 \cdot 10^{-2} - 1}{5.6 \cdot 10^7} = 6.08 \cdot 10^{-3} s$$

Choosing C to be 1 μ F

$$R_1 = 26.12 \text{ k}\Omega$$

$$R_2 = 60.8 \text{ }\Omega$$

Checking for phase margin:

$$\omega = \frac{\tau_2}{\tau_1} K = 132137 \text{ rad/s}$$

$$\begin{aligned} \text{Phase margin} &= 90 + \text{atan}(\omega\tau_1) - \text{atan}(\omega\tau_2) \\ &= 90+90-83=97^\circ \end{aligned}$$

which is acceptable.

Acquisition Range for a phase locked loop using a low pass filter with phase lead correction

$$\text{is } \Omega_{acq} = 2\sqrt{K\zeta\omega_n}$$

$$= 3.86 \times 10^6 \text{ rad/s} = 613 \text{ kHz}$$

This is 50% of the design objective but is considered adequate for the initial design.

If subsequent tests indicate that the acquisition range should be greater this can be accomplished by adding gain into the loop filter.

The loop filter must accomplish two functions other than actually filtering the loop signal;

it must compensate for the DC bias of the Gilbert mixer and it must bias the loop VCO to a frequency close to the frequency of interest (1.57542 GHz). This is accomplished by using an operational amplifier as a differential amplifier as shown in Figure 5.9

Figure 5.9
Loop Filter Design for Tracking Filter

In this circuit R1, R2 and C1 form the low pass filter with phase lead correction as described above. R3 controls the gain of the circuit and is set equal to R1 to provide unity gain. R4 and R6 are chosen so that the voltages at points A and B straddle the required adjustment range thus making the adjustment of R5 as sensitive as possible.

In practice R5 is adjusted so that the frequency of the VCO output is 1.575 GHz.

The maximum frequency offset is 1 MHz which requires a correction voltage of $\frac{1}{K_1}$ or

33mV which gives a maximum phase offset (error) of $\frac{180}{\pi} \frac{1}{K} = \frac{180}{\pi} \frac{33}{300} = 6.3^\circ$.

This is acceptable for the initial design. However, as in the case of the acquisition range, this error could be decreased, if necessary, by the use of gain in the loop filter.

The output of tracking filter #1 (designated as the master filter) must supply a signal to the spectrum analyzer input, to its own phase detector and, when in steering mode, to filters #2 and #3. This will be accomplished by means of microstrip directional couplers.

The resulting tracking filter was constructed and is shown in Figure 5.10.

Figure 5.10
Tracking Filter Board

Note that a microstrip directional coupler was used to provide the signal path from the VCO to the phase comparator. This serves two purposes. The first is to isolate the VCO from the input to the Gilbert Mixer and thus provide a better output match for the VCO. The second, in the case of the master tracking filter, is to permit the VCO output to be applied to the other tracking filters and to the monitoring spectrum analyzer. It was also necessary to provide a DC block in the RF path since a DC voltage is present on the Gilbert Multiplier output ports. The directional coupler was designed using Libra for a coupling of 12 dB. With a VCO output of +12dBm the input to the Gilbert Mixer was 0 dBm.

5.4 Signal Processing Design

The primary function of the signal processor is to

- a. convert the outputs of the Gilbert mixers to digital form
- b. compute the phase angles
- c. use the computed phase angles to determine the relative bearing of the source

In addition the processor must implement the communications protocol with the host computer and perform the calibration and self-test tasks.

5.4.1 Signal Conversion

As was mentioned previously, the outputs of the Gilbert mixers are biased by about 4 V and the useful signal, with a peak amplitude of about 300 mV, is superimposed on this bias. In order to keep the quantization error due to the analog to digital conversion compatible with the overall measurement error (3°), it is necessary to limit the resolution of the ADC to

2mV. Thus, if the raw signal is converted, $\frac{4.3}{2 \cdot 10^{-3}} \approx 2000$ quantization steps are required.

This implies an 11 bit (12 in practice) ADC. However, if only the useful signal is converted

$\frac{0.3}{2 \cdot 10^{-3}} \approx 150$ quantization steps are required. This implies an 8 bit ADC but requires that

the bias be removed before conversion. This can be done using a differential amplifier with a bias voltage on one input as shown in Figure 5.9.

Chapter 6

Simulation and Testing

6.0 Phase Comparator Testing

Two test procedures were used to evaluate the phase comparator circuit. The first, a dynamic test was intended to provide quick estimates of the magnitude of the phase measurement errors. The second, static test, was used to provide an absolute calibration of the phase measurement circuit.

6.0.1 Dynamic Test

The dynamic test setup consisted of two microwave signal generators, a digital oscilloscope and a Compaq computer for data acquisition as shown in Figure 6.1.

Figure 6.1
Dynamic Test Setup

The equipment used was as follows:

Signal Generators #1 and #2: Marconi 2032

Digital Oscilloscope: Tektronix 2440

Computer: Compaq Portable 385 with IEEE 488 interface

The output of one signal generator was set to a frequency of 1575.42 MHz while the output of the other was set to 1575.421 MHz thus providing a difference frequency of 1 kHz. The output power levels were both 0 dBm. Due to the fact that the outputs of the Gilbert mixers contain a DC bias, the oscilloscope inputs were AC coupled.

The sine and cosine output traces were captured on the oscilloscope screen and then transferred to the computer via the IEEE 488 data bus. The result is a 1024 x 2 array which was stored in a file on a floppy disk. A sample file plot is shown in Figure 6.2

Figure 6.2
Sample Raw Oscilloscope Traces

This file was then imported into a MATLAB program which normalizes the magnitudes of the two traces and then computes the measured angle by taking the arctangent of the ratio of the sine trace to the cosine trace.

If the phase measurement were perfect the resulting trace would be a sawtooth with a peak to peak amplitude of 360° , however, the actual trace departs from the sawtooth due to phase dependent errors i.e. non-bias type errors. This is shown in Figure 6.3. In order to assess the magnitude of these errors, the ideal sawtooth trace is subtracted from the actual trace with the result being plotted. The selection of the start point for the reference sawtooth is arbitrary but it influences only the overall bias of the error curve. In this case the start point was chosen as the first point after the calculated phase angle changed from $+180^\circ$ to -180° . The final plot for this example is shown in Figure 6.4 and indicates that the peak to peak error magnitude is about 16° . Thus the magnitude of the non-bias errors is expected to be about 8° .

Figure 6.3
Reference and Phase Measurement Traces

Figure 6.4
Sample Error Plot

6.0.2 Static Testing

While the dynamic test is useful for the preliminary assessment of the phase measurement circuits, determination of the circuits' absolute accuracy requires a different approach. In these tests, a known, calibrated phase shift is introduced between the inputs to the test circuit and the computed phase detector output is compared with it to determine the measurement error for that phase.

The phase measurement standard for these tests was a Hewlett-Packard Model 8720 A Network Analyzer Serial Number 2749A00258.

Figure 6.5
Static Phase Measurement

The stated error (1σ) for this instrument at 1.575 GHz is less than 1° .

6.0.2.1 Preliminary Measurements

- a. The phase shift through the 0° power splitter and its associated RF cables was determined using the Network Analyzer. The output cables were selected such that the phase shift AB and AC (shown in Figure 6.5) were equal within 0.5°
- b. The Network Analyzer was then set up to measure the phase shift CD. Using the Line Stretcher (Watkins Johnson Model 3144) adjustment, the settings for phase shifts from 0° to 350° in 10° increments were established. The results are shown in Table 6.1

Table 6.1 Delay Line Calibration

ANGLE	SETTING	ANGLE	SETTING	ANGLE	SETTING
0	29.4	120	35.73	240	42.08
10	29.92	130	36.24	250	42.60
20	30.46	140	36.78	260	43.12
30	31.00	150	37.32	270	43.65
40	31.50	160	37.84	280	44.22
50	32.02	170	38.38	290	44.70
60	32.52	180	38.90	300	45.26
70	33.06	190	39.42	310	45.76
80	33.60	200	39.96	320	46.30
90	34.14	210	40.48	330	46.84
100	34.64	220	41.00	340	47.36
110	35.20	230	41.52	350	47.86

c. In order to make the required measurements, a test jig was constructed. The circuit of this test jig is shown in Figure 6.6. The test jig provided the supply voltages to the devices and also provided access to the Gilbert mixer output voltages. As is shown in the figure, the mixer output biases were removed by the differential amplifiers so that they were close to zero under zero RF power input. This was to simulate the signal conditioning required by the system microcontroller.

d. Preliminary tests were done using this test setup to determine the error characteristics of the phase comparators under various RF power levels since there is a trade-off between the errors introduced by distortion at higher levels of RF and the low signal output magnitudes at lower RF levels.

Figure 6.6
Phase Comparator Test Jig

e. The measured phase angles were computed as follows:

For each pair of ports the outputs of the sine and cosine comparators were measured at 10° intervals thus producing two 1×36 measurement vectors. The maximum and minimum readings in each vector were found and used to determine a reference offset value $\frac{V_{MAX} + V_{MIN}}{2}$, which was then subtracted from all readings in the vector.

The two vectors were then normalized by multiplying the cosine vector by the ratio

$$\frac{V_{MAXSIN} - V_{MINSIN}}{V_{MAXCOS} - V_{MINCOS}}. \text{ Finally the measured angle was computed as the arctangent}$$

of the ratio of each element of the sine vector to the corresponding element of the cosine vector.

f. The error was evaluated as the computed value minus the nominal value.

The rationale for this process of assessing the error characteristics is that parameters such as the offset and ratio of the sine and cosine outputs of the comparator can be determined by the use of a calibration procedure and can be used to correct subsequent measurements. To use such a method, however, it is necessary to determine the amount of variation of these variables over time and with changes in RF power and supply voltage. Variation with temperature would also be a concern given the ultimate use of this equipment in an aircraft but lack of time and equipment did not permit an assessment of this factor. Note that the offset of the comparator DC outputs is affected by two factors: the drift of the Gilbert Mixer and the drift of the bias voltage on the operational amplifier. The bias drift was assessed by measuring the phase comparator outputs with no RF applied prior to each stability run. This information was not applied to the results. Typical values are shown in Table 6.2

Note that the VCC was monitored and maintained at $10V \pm 10mV$.

Initial assessment of phase measurement for the first prototype board was done using input power levels of +3, 0 and -3 dBm although the test at 0 dBm was done only for phase comparator 1-2.

The assessment of phase measurement stability for the final prototype board was done over time for power levels -3dBm, 0 dBm, +3dBm and +6dBm. Note that these were the output power levels of the signal generator and that the output of the 0° power splitter, which had an insertion loss of 0.8 dB was 3.8 dB less.

This test simulated an operational procedure in which the phase comparator would be calibrated and subsequent measurements would be corrected using the offsets, ratios and errors determined during the calibration.

Thus the calibration run measured and recorded the sine and cosine trace offsets, the sine/cosine ratio and the errors for each 10° point. This was done for each of the four input power levels

The stability measurements consisted of measuring the sine and cosine voltages, correcting them by the recorded calibration values for offset and ratio and finally subtracting the appropriate recorded error. The resulting value was then subtracted from the nominal angle to determine the error.

Five sets of measurement runs were performed over a period of 3 weeks.

6.0.2.2 Results

6.0.2.2.1 Initial Phase Comparator

The results of the initial phase comparator assessment are shown in Figures 6.7. At the input power level of 3 dBm the peak to peak outputs of the phase detectors were about 860 mV. At the input power of 0dBm the peak to peak outputs were about 370 mV and at the input power of -3 dBm the peak to peak outputs were about 140 mV.

The mean error for the 3 power levels remains relatively constant while the error spread increases with increasing power. This is probably due to increasing distortion in the Gilbert mixer. The fact that the mean error remains constant indicates that it is a function of the board construction and is unlikely to change over time. These results show that the errors are a function of input power levels and that these should be kept constant for the calibration values to remain valid.

Table 6.2 Error Statistics for Figure 6.7

Phase	Power(dBm)	Mean (°)	Peak-peak (°)
1-2	3	-14	11
	0	-14	17
	-3	-14	21
2-3	3	7.5	7.7
	-3	7.2	19.7
3-1	3	1.4	7
	-3	0.6	18

Figure 6.7
Error Plots for Initial Phase Comparator

6.0.2.2.1 Final Phase Comparator

While the initial phase comparator test results showed a maximum mean error magnitude of 14° , the tests on the final phase comparator showed mean errors between 23° and 29° . A typical error plot for input port power levels of -3, 0 +3 and +6dBm is shown in Figure 6.8. Note that the error spread is dependent on the input power.

Figure 6.8
Phase Comparator Error Plots for Four Input Power Levels

As expected, the amplitude of the sine and cosine outputs was strongly dependent on input port power. As well, it was noticed that the average of the sine and cosine inputs were also dependent on input power. Typical values of these variables are shown in Table 6.3 Note: Offset was adjusted to 0mv with no RF input. .

Table 6.3 Typical Gilbert Mixer Outputs vs Input Power

Input Power (dBm)	Peak (mV)	Offset (mV)
-3	180	10
0	350	16
3	635	20
6	1000	23

These observations indicate that a stable input power is very important to the accuracy of the phase comparator measurements. In this design, the required power stability is achieved by using the phase locked loop (PLL) implementation of the tracking filter

Figure 6.9
Phase Measurement Stability Test Results

The results of the stability tests are shown in Figure 6.9. This graph shows the mean error per angle as well as the mean error $\pm$ the standard deviation of 12 samples (4 sets of 3 measurements, one measurement for each pair of ports). The standard deviation is about 1.5 but, since the calibration measurement is effectively part of the same population, the estimated accuracy of the phase comparator is about $\sqrt{2} \times 1.5^\circ \approx 2^\circ$ which is well within the requirements of the system.

6.1 Tracking Filter Simulation and Testing

6.1.0. Introduction

The purpose of the tracking filter simulation was to verify the acquisition range and noise filtering performance of the filter design. This was done using Hewlett-Packard EEsof Omnisis simulation software.

Because both the characteristics of interest were functions of time, the Omnisis Discrete Time test bench was used. The operating principle of this test bench is to first determine

the frequency response of the circuit elements and then to perform a Fast Fourier Transform to obtain the time domain response. Two important control variables in this process and which limited the extent of these tests are the Maximum Impulse Time (MaxImpTime) and the Time Step Size (TStep). The Maximum Impulse Time is the length of time over which the impulse response is evaluated. Omnisys documentation recommends that this be at least 5 times the reciprocal of the lowest frequency of interest ($5/f_{\text{MIN}}$) which, in the case of the tracking filter is the 3dB point of the loop filter. The Time Step Size determines the maximum frequency (f_{MAX}) of the response which can be observed. According to the Nyquist criterion, the Time Step Size should be $1/2f_{\text{MAX}}$, however HP recommend using $1/5f_{\text{MAX}}$. These two parameters determine the size of the FFT required to perform the simulation, the size being $2 \times \text{MaxImpTime}/\text{TStep} = 2 \times (5/f_{\text{MIN}})(5f_{\text{MAX}})$ or approximately 50 times the ratio of the maximum to minimum frequencies of interest. Unfortunately, the maximum frequency in this case is about 1 MHz (acquisition range and noise bandwidth) and the lowest frequency is about 10 Hz (3dB point of loop filter). Thus a 5 million point FFT is required to accommodate a simulation of the actual design. In practice, due to memory limitations, a 1 million point FFT was the largest that could be computed by the available computer. Therefore, simulation was performed within the restrictions of the existing software and the results were extrapolated to the actual case.

To determine if the maximum impulse time is sufficient, Hewlett Packard recommends that the component with the lowest bandwidth be tested as follows. The frequency response is determined using the swept frequency test bench then the impulse response is measured with the Time Domain Test bench. The spectrum of the impulse response is then compared with the response measured in the first part. If the agreement is not satisfactory then the parameters of the test bench are adjusted until agreement is achieved. The comparison for the Tracking Filter loop filter is shown in Figure 6.11.

The element in the tracking filter with the lowest bandwidth is the loop filter as shown in Figure 6.10.

Figure 6.10
Tracking Filter Loop Filter

Figure 6.11
Comparison of Loop Filter Frequency Response Using Swept Frequency
and Impulse Response/FFT methods

Figure 6.12
Tracking Filter Configured for Noise Simulation

6.1.1. Simulation

Because Omnisys does not have the source modules necessary to simulate additive white Gaussian noise directly, it was found necessary to develop an equivalent stimulus. Using the idea from Blanchard [19] that the effects of additive Gaussian noise on a phase locked loop with a sinusoidal phase detector can be considered as phase modulation noise with variance $1/(S/N)$, the Omnisys test bench was set up as shown in Figure 6.12. Blanchard points out that this is only apparent phase modulation (in the case of a sinusoidal phase detector) and that its effect is that of a carrier modulated with noise but operating into a phase locked loop with a linear phase detector. Thus the tracking filter design was modified by replacing the sinusoidal phase detector with a linear one with the same sensitivity. This is shown in Figure 6.12.

An additional problem with this simulation is the frequency range which must be covered. The minimum frequency, which determines the length of the simulation is about 2 Hz while the maximum frequency is determined by the noise bandwidth to be investigated, in this case 4 MHz. The ratio of these frequencies dictates the size of the FFT that the simulation must use and the length of the simulation time. Experience has shown that a 1 million point FFT is the upper limit for the computer system available. A million point FFT simulation can take from about 36 to 72 hours.

The upper limit of frequency for the simulation of the system under test was 500kHz (TStep=1 μ s). This was limited more by time than memory. The simulation ran for 36 hours which is at the low end of run times but the run times increase by a factor of 4 for an increase of 2 in frequency. Thus, it was considered impractical to try to test at 1 MHz (TStep=0.5 μ s).

Figure 6.13
Test Bench for Noise Performance Simulation

Because the filter could not be tested at the desired frequency, the performance was assessed by measuring the filter output at frequencies of 62.5kHz, 125kHz, 250kHz and 500kHz and attempting to extrapolate the results to 4 MHz.

Note that in Figure 6.12 the output of the VCO is also fed to a phase detector with a sensitivity of 0.01 V/degree. In the simulation, the other input port of this detector is fed from a reference oscillator and thus the output of the detector is a DC voltage proportional to phase.

In the test bench Figure 6.13, the phase comparator output is fed to an RMS voltmeter to give the standard deviation of the noise on the phase output. This is used because Omnisys lack measurement modules for performing statistical analysis on outputs.

The results of the simulation are shown in Figure 6.14

Figure 6.14
Plot of Phase Noise versus Test Noise Bandwidth

It appears that the extrapolated value would be at or near the design value of 5°.

6.1.2. Testing

Three tracking filters were built and tested. Unfortunately, the circuits as implemented were very sensitive to external conditions and thus conclusive testing was not possible. The

use of the Gilbert Mixer as a phase detector was found to present problems in the acquisition performance. This was because the input RF affects the DC bias of the mixer so that, if the loop is in lock, and the input signal is removed, the VCO tuning voltage changes and shifts the VCO frequency, usually out of acquisition range. An additional problem was that the VCO tuning voltage was sensitive to the motion of people near the circuit, making it difficult to adjust the frequency to achieve lock with the variable resistor on the board.

Thus, the implementation of the tracking filter must be improved before reliable conclusions can be drawn.

6.2. Antenna Array Testing

6.2.0. Introduction

The purpose of the antenna array testing was to determine if the array, as constructed, could be used to measure the relative bearing of a source and to what accuracy. The antenna array was tested under close to ideal signal to noise ratio conditions. The testing was divided into two parts: a) establishing the correspondence between the physical location of the antennas and the location of the antenna phase centres and b) confirming that the phase difference measurements from the actual antennas can be used to determine relative bearing of the source of received radiation.

6.2.1 Determination of Array Dimensions

Part of the requirement for the measurement is a knowledge of the locations of the phase centres of the individual antennas. The antennas used in the array were designed for aircraft navigation systems and thus, as opposed to antennas designed for surveying use, the locations and stability of the phase centres were not specified.

To make these measurements it was decided to take advantage of the fact that GPS can be used to determine relative positions of GPS antennas extremely accurately. This is done routinely by surveyors using survey quality GPS receivers and appropriate post-processing

software. The receivers used in this case were Ashtech model Z12s which have the capability of using both the L1 and L2 carrier signals, although in this case, due to the fact that the Novatel antennas are limited to the L1 frequency, only L1 signals were used.

Two survey receivers were available and thus measurements were made for pairs of antennas (i.e. Antennas 1 and 2, antennas 2 and 3 and antennas 3 and 1). The third set is redundant but was included as an accuracy check. Also, the data reduction software has the capability of adjusting position data for a closed figure to provide a better estimate of position.

The antennas were physically installed in an equilateral triangle with 190mm sides.

The raw results of the survey are shown in Table 6.3

Table 6.4 Antenna Surveyed Locations

ANTENNA	x(m)	y(m)	z(m)
1	1108546.153	-4348215.032	4517549.623
2	1108546.330	-4348214.960	4517549.640
3	1108546.292	-4348215.105	4517549.520

Note that these results are in the Earth-Centred Earth Fixed (ECEF) coordinate system used by the GPS. This system is Cartesian with the origin at the earth's centre of mass. The z axis coincides with the earth's spin axis, and the x axis passes through the intersection of the equator and the prime (or Greenwich) meridian. The y axis completes a right handed orthogonal system.

For ease of interpretation it is preferable to convert ECEF coordinates to a locally level system with the y axis pointing True North, the x axis pointing East and the z axis vertical. This is accomplished by one rotation of $(90^\circ - \text{longitude}(\lambda))$ about the z axis and a subsequent rotation of $(90^\circ - \text{latitude}(\Phi))$. To simplify the presentation, the first 6 digits were dropped from each coordinate thus putting the origin at 110854, -434821, 451754.

The resulting coordinate transformation is:

$$\begin{bmatrix} x_{LOC} \\ y_{LOC} \\ z_{LOC} \end{bmatrix} = \begin{bmatrix} \cos(90 - \Phi) \cos(90 - \lambda) & \cos(90 - \Phi) \sin(90 - \lambda) & -\sin(90 - \Phi) \\ -\sin(90 - \lambda) & \cos(90 - \lambda) & 0 \\ \sin(90 - \Phi) \cos(90 - \lambda) & \sin(90 - \Phi) \sin(90 - \lambda) & \cos(90 - \Phi) \end{bmatrix} \begin{bmatrix} x_{ECEF} \\ y_{ECEF} \\ z_{ECEF} \end{bmatrix}$$

Where x_{LOC} , y_{LOC} and z_{LOC} are the locally level coordinates and x_{ECEF} , y_{ECEF} and z_{ECEF} are the ECEF coordinates.

The locally level coordinates of the antennas are shown on table 6.4.

Table 6.5 Locally Level Antenna Coordinates

ANTENNA	x(m)	y(m)	z(m)
1	4.719	2.206	11.342
2	4.909	2.236	11.336
3	4.836	2.059	11.343

The distances between the antennas were:

1-2: 192mm
 2-3 191mm
 3-1 188mm

The estimated error in the survey results was 0.5 mm

These results indicate that the phase centres are very close to being in the same relative physical position for all of the antennas. This permits the positioning of the antennas on an aircraft using only the physical location as a reference.

6.2.2. Measurement of the Antenna Array Characteristics

6.2.2.1 Measurement Procedures

The fundamental premise of this system is that the relative bearing of the interference source can be determined by measuring the phase differences between the outputs of the three antennas and using the equation:

$$\theta = \text{atan}\left(\frac{2A\phi_{12}}{\phi_{32} - \phi_{21}}\right)$$

as developed in chapter 4.

While the feasibility of making the necessary phase measurement has been demonstrated as part of this thesis, it was decided to prove the theory of the antenna array measurement using independent means. This was done (1) to simplify the measurements since the full data processing subsystem had not yet been implemented and (2) to permit isolation of error contributions and (3) to permit a more accurate assessment of the array performance.

The primary instrument for this was the Hewlett-Packard Network Analyzer previously described in connection with the calibration of the phase shift measurement . As was mentioned, the phase measurement accuracy of this instrument is better than 1°.

The DF array and test transmitting antenna were installed in an anechoic chamber in the configuration shown in Figure 6.15. The electrical configuration is shown in Figure 6.16.

Figure 6.15
Physical Layout for Antenna Phase Measurements

Figure 6.16
Electrical Layout for Antenna Phase Measurements

Bias tee #1 was required to supply power to the low noise preamplifier incorporated

in the GPS antenna. Bias tee#2 was used to block the 15VDC which is present on the centre conductor of the 60 dB low noise amplifier.

The antenna pedestal is equipped with a rotation speed control and an angle readout with a precision of 0.01° . The angle readout was set to read zero when the DF array antenna #2 was closest to the transmitting antenna.

The network analyzer was configured to measure the phase between the output and input ports.

Since only one measurement channel was available, each DF array antenna was tested individually.

The measurement procedure was as follows:

- a) The antenna to be tested was connected to the LNA input cable.
- b) The pedestal angle was set to 0°
- c) The network analyzer was calibrated
- d) The pedestal was set to 10°
- e) The phase shift was read from the network analyzer and the pedestal angle was incremented by 10° .
- f) step e) was repeated until the angle reached 360° .

6.2.2.2 Results

The raw results are shown in Figure 6.17. Since the reference phase angles for each set of measurements was different, it was necessary to bring all of the measurements to a common reference. This was done by assuming that the phase difference between two antennas was zero when the line joining two antennas was 90° to the direction of the transmitting antenna. This occurred at 60° for antennas 2 and 3 and 180° for antennas 1 and 3. The resulting plot is shown in Figure 6.18.

Figure 6.17
Raw Antenna Phase Shift Measurements

Figure 6.18
Corrected Antenna Phase Shift versus Relative Bearing of Array

The phase differences between each pair of antennas was then calculated for each 10° increment and was used to determine the calculated relative bearing. Finally the nominal bearing was subtracted from the calculated bearing to establish the bearing error. The results are shown in Figure 6.19

Figure 6.19
Errors in Relative Bearing Measurement

The mean error is -2.4° and the standard deviation is 3.7°

To investigate the effect of the other antennas on a given antenna phase measurement, the phase shift versus relative bearing was measured for antenna 3 with the other two antennas removed from the array. The results are shown in Figure 6.20.

Figure 6.20
Effect of Other Antennas in Array

The maximum difference is 14° at a bearing of 70° and the average is 2° . This indicates that, at least at the vertical angle at which the test was conducted (12.5°) there is little interference between antennas.

It is interesting to note that, during the 360° rotation of the antennas, 360° of phase shift is accumulated. It was verified that this was a property of the antennas as follows:

A single antenna was mounted on a ground plane and placed on the rotating pedestal with its physical centre coincident with the platform vertical axis. The phase shift versus angle of rotation was measured using the procedure detailed above. The results are shown in Figure 6.21

Figure 6.21
Accumulation of Phase Shift with Rotation of GPS Antenna

Chapter 7

Conclusions and Future Work

7.0 Conclusions

The objective of the work covered by this thesis was to determine the feasibility of using a simple array of GPS antennas to measure the relative bearing of a low power source of interfering radiation. The proposed design included three critical features which required development and investigation to establish their suitability: the antenna array itself, a baseband phase measurement technique and a tracking filter for improvement of signal to noise ratio and provision of a stable power level into the phase measurement subsystem.

This thesis has shown that the antenna array, under general conditions, can measure the relative bearing to a radiating source with a maximum error of 8° . This is acceptable in itself but can probably be improved upon by the use of additional calibration procedures.

It was shown that the phase differences between the antenna pairs can be measured to an accuracy of about 3° with stable input power conditions.

A simulation of the tracking filter indicated that the phase variance produced by a low signal to noise ratio can be reduced to an acceptable value.

These results indicate that there is a high likelihood that the original direction finder design can be developed into a system which will meet the original requirements.

7.1 Future Work

7.1.1 Antenna Array

The antenna array was tested using a vertical angle of 12.5° . This is the equivalent of a depression angle for the aircraft installation. This angle would be observed at an altitude of 6000 Ft. and a ground range of 4.5 nautical miles from a source on the ground. Thus to

achieve a reasonable range of detection the aircraft's altitude would have to be increased considerably. It is recommended that the performance of the antenna array be investigated at lower depression angles. Also, since the antenna phase response is a function of relative angle, it would be useful to investigate whether calibrating this function would result in improved performance. This implies a prior knowledge of the relative bearing but this could be handled by using a two stage procedure in which the bearing is first measured without correction and then with the correction applied.

7.1.2 Phase Comparator

The phase detector performance is acceptable and the only development which might be useful or interesting is to determine the origin of the large error biases. Repackaging might be necessary to accommodate an operational system design.

7.1.3 Tracking Filter

The main problem with the current version of the tracking filter is the use of the Gilbert mixer as a phase detector. As has been reported, this causes severe problems with acquisition. Other problems are associated with layout and packaging. These should be improved to reduce the sensitivity to external influences. Also, the source for the offset voltage should be changed to a digital to analog converter. This would remove the requirement for physical adjustments which would be unacceptable in the operational version in any case.

A loop lock detection capability is essential to provide an indication of system status. It would also be very useful in the implementation of a dual filter mode in which the bandwidth is wide prior to acquisition and narrow afterward.

The requirement for the phase noise variance could probably be relaxed since variations in the phase could be further filtered or smoothed after the phase detector considering that the data rate is expected to be in the neighbourhood of 1 Hz. This would improve the acquisition performance.

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