1

CANADIAN THESES ON MICROFICHE

I.S.B.N.

THESES CANADIENNES SUR MICROFICHE

National Library of Canada
Collections Development Branch

Canadian Theses on
Microfiche Service

Ottawa, Canada
K1A 0N4

Bibliothèque nationale du Canada
Direction du développement des collections

Service des thèses canadiennes
sur microfiche

NOTICE

The quality of this microfiche is heavily dependent upon the quality of the original thesis submitted for microfilming. Every effort has been made to ensure the highest quality of reproduction possible.

If pages are missing, contact the university which granted the degree.

Some pages may have indistinct print especially if the original pages were typed with a poor typewriter ribbon or if the university sent us a poor photocopy.

Previously copyrighted materials (journal articles, published tests, etc.) are not filmed.

Reproduction in full or in part of this film is governed by the Canadian Copyright Act, R.S.C. 1970, c. C-30. Please read the authorization forms which accompany this thesis.

**THIS DISSERTATION
HAS BEEN MICROFILMED
EXACTLY AS RECEIVED**

AVIS

La qualité de cette microfiche dépend grandement de la qualité de la thèse soumise au microfilmage. Nous avons tout fait pour assurer une qualité supérieure de reproduction.

S'il manque des pages, veuillez communiquer avec l'université qui a conféré le grade.

La qualité d'impression de certaines pages peut laisser à désirer, surtout si les pages originales ont été dactylographiées à l'aide d'un ruban usé ou si l'université nous a fait parvenir une photocopie de mauvaise qualité.

Les documents qui font déjà l'objet d'un droit d'auteur (articles de revue, examens publiés, etc.) ne sont pas microfilmés.

La reproduction, même partielle, de ce microfilm est soumise à la Loi canadienne sur le droit d'auteur, SRC 1970, c. C-30. Veuillez prendre connaissance des formules d'autorisation qui accompagnent cette thèse.

**LA THÈSE A ÉTÉ
MICROFILMÉE TELLE QUE
NOUS L'AVONS REÇUE**

National Library
of CanadaBibliothèque nationale
du Canada

Canadian Theses Division Division des thèses canadiennes

Ottawa, Canada
K1A 0N4

59934

PERMISSION TO MICROFILM — AUTORISATION DE MICROFILMER

- Please print or type — Écrire en lettres moulées ou dactylographier

Full Name of Author — Nom complet de l'auteur

CHABOT, REMY REYNALD

Date of Birth — Date de naissance

07 OCTOBER 1952

Country of Birth — Lieu de naissance

CANADA

Permanent Address — Résidence fixe

23 DES JARDINS
BUCKINGHAM, QUEBEC
J8L 3G5

Title of Thesis — Titre de la thèse

IMPLEMENTATION OF THREE MONOPULSE RECEIVER
CONFIGURATIONS AND EXAMINATION OF THEIR
PERFORMANCE

University — Université

CARLETON UNIVERSITY

Degree for which thesis was presented — Grade pour lequel cette thèse fut présentée

MASTER OF ENGINEERING

Year this degree conferred — Année d'obtention de ce grade

1982

Name of Supervisor — Nom du directeur de thèse

B.A. SYRETT

Permission is hereby granted to the NATIONAL LIBRARY OF
CANADA to microfilm this thesis and to lend or sell copies of
the film

The author reserves other publication rights, and neither the
thesis nor extensive extracts from it may be printed or other-
wise reproduced without the author's written permission.

L'autorisation, est, par la présente, accordée à la BIBLIOTHÈ-
QUE NATIONALE DU CANADA de microfilmer cette thèse et de
prêter ou de vendre des exemplaires du film

L'auteur se réserve les autres droits de publication, ni la thèse
ni de longs extraits de celle-ci ne doivent être imprimés ou
autrement reproduits sans l'autorisation écrite de l'auteur

Date

21 Sep 82

Signature

Remy Chabot

IMPLEMENTATION OF THREE MONOPULSE
RECEIVER CONFIGURATIONS AND
EXAMINATION OF THEIR PERFORMANCE

by

© Remy Chabot, BAsC.

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

Master of Engineering

Department of Electronics
Faculty of Engineering
Carleton University
Ottawa, Canada
July 1982

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

"Implementation of three monopulse receiver configurations
and examination of their performance."

submitted by Remy Chabot in partial fulfillment of the requirements
for the degree of Master of Engineering.

Associate Professor B.A. Syrett
Thesis Supervisor

Prof. A.R. Boothroyd
Chairman
Department of Electronics

August, 1982

ABSTRACT

The conventional method of designing a monopulse radar receiver requires three channels of RF to IF conversion. This method allows the designer to extract complete angle information on each received pulse but has limitations imposed by channel imbalances. The number of receiver channels can be reduced to two or one in order to remove these imbalances.

The theory and design of one- and two-channel monopulse receivers are presented in this thesis. Two dual-channel monopulse receivers and one single-channel monopulse receiver are examined. From experimental results it is found that the angle-tracking error for all three systems is inversely proportional to the square root of the signal-to-noise ratio as predicted from theory. However the minimum angle-tracking error of the one-channel receiver is greater than the minimum angle-tracking error of both dual-channel receivers.

ii

ACKNOWLEDGEMENTS

I wish to express my gratitude to my Carleton supervisor, Professor B.A. Syrett and to my external supervisor, Dr. W.K. McRitchie for their guidance and their valuable support.

The many fruitful discussions with members of the Radar Countermeasures Section at the Defence Research Establishment Ottawa and in particular with Mr. Geoffrey Wardle are highly appreciated.

I must especially thank my wife Johanne for her understanding, encouragement and patience.

TABLE OF CONTENTS

	<u>Page</u>
CHAPTER 1. INTRODUCTION	1
1.1 Introduction	1
1.2 Thesis objectives	1
1.3 Thesis organization	2
CHAPTER 2. THEORY OF ANGLE-TRACKING RADARS	4
2.1 Introduction	4
2.2 Principles of monopulse radar	4
2.3 Time-shared monopulse concept	15
2.4 Thermal noise analysis	22
2.5 Phase and amplitude sensing with sum-and-difference comparison	25
2.6 Summary	28
CHAPTER 3. IMPLEMENTATION OF THREE RECEIVER CONFIGURATIONS	30
3.1 Introduction	30
3.2 Circuitry common to the three receiver configurations	30
3.3 Description of the first receiver configuration	33
3.4 Description of the second receiver configuration	41
3.5 Description of the third receiver configuration	48
3.6 Summary	51
CHAPTER 4. PRACTICAL CONSIDERATIONS	53
4.1 Introduction	53
4.2 Receiver noise limits	53
4.3 Timing considerations	56
4.4 Sample-and-hold considerations	67
4.4.1 Introduction	67
4.4.2 Design of a slow sample-and-hold device	68

4.5	Video pulse consideration	72
4.6	Error voltage output filter	73
4.7	Determination of servo bandwidth	74
4.8	Design of an experimental set-up	75
4.9	Logarithmic amplifier reading of signal-to-noise ratio	78
4.10	Summary	79
CHAPTER 5. EXAMINATION OF THE PERFORMANCE		80
5.1	Introduction	80
5.2	Experimental results	80
5.2.1	Experimental determination of k_m	80
5.2.2	Angle jitter measurements	85
5.2.3	Boresight shift measurements	92
5.3	Interpretation of results	92
5.3.1	Angle jitter measurements	92
5.3.2	Boresight shift measurements	95
5.4	Experimental comparison	98
5.5	Summary	100
CHAPTER 6. CONCLUSION		101
6.1	Summary and conclusions	101
6.2	Future work areas	103
REFERENCES		104

LIST OF FIGURES

<u>FIGURE NO.</u>		<u>PAGE</u>
1.1	Block diagram of tracking radar simulator at Defence Research Establishment Ottawa	3
2.1	Antenna patterns of three angle-tracking techniques: (a) monopulse, (b) sequential lobing, and (c) conical scan.	6
2.2(a)	Sum and difference power patterns as a function of angle off-boresight.	10
(b)	Error voltage as a function of angle off-boresight.	10
2.3	Block diagram of a conventional three-channel monopulse receiver.	12
2.4	Block diagram of Noblitt's(1) two-channel monopulse radar receiver.	13
2.5	Block diagram of a two-channel time-shared monopulse radar receiver.	16
2.6	Block diagram of a two-channel time-shared monopulse radar receiver with delay.	18
2.7	Block diagram of a one-channel time-shared monopulse radar receiver.	19
2.8	Block diagram of the monopulse comparator used in the three receiver configurations.	20
3.1	Block diagram of circuitry common to the three TSMP receiver configurations.	31
3.2	RF to video conversion section of the two-channel TSMP radar receiver.	34
3.3	Video interface to the range tracker for the two-channel TSMP radar receiver.	36

LIST OF FIGURES (cont'd)

<u>FIGURE NO.</u>		<u>PAGE</u>
3.4	Angle demultiplexer for the two-channel TSMP radar receiver.	38
3.5	Drive amplifier of two-channel TSMP radar receiver.	39
3.6	Timing diagram for the two-channel TSMP radar receiver.	40
3.7	RF to video conversion section of the two-channel TSMP receiver with delay.	42
3.8	Video interface to the range tracker for the two-channel TSMP receiver with delay and for the one-channel TSMP receiver.	43
3.9	Functional diagram of the angle demultiplexer for the two-channel TSMP receiver with delay and for the one-channel TSMP receiver.	44
3.10	Block diagram of the angle demultiplexer for the two-channel TSMP receiver with delay and for the one-channel TSMP receiver.	46
3.11	Timing diagram for the two-channel TSMP receiver with delay.	47
3.12	RF to video conversion section of the one-channel TSMP receiver.	49
3.13	Timing diagram for the one-channel TSMP receiver.	50
4.1	Block diagram of the timing interface for the one-channel TSMP receiver.	58
4.2	Circuit diagram of the timing interface for the one-channel TSMP radar receiver.	59

LIST OF FIGURES (cont'd)

<u>FIGURE NO.</u>		<u>PAGE</u>
4.3	Timing diagram for the timing interface of the one-channel TSMP radar receiver.	60
4.4	Block diagram of the timing interface for the two-channel TSMP radar receiver with delay. ..	61
4.5	Circuit diagram of the timing interface for the two-channel TSMP receiver with delay. ..	63
4.6	Timing diagram for the timing interface of the two-channel TSMP receiver with delay.	64
4.7	Circuit diagram of the timing interface between the range tracker and the PIN switch of the one-channel TSMP receiver.	65
4.8	Timing diagram for the timing interface between the range tracker and the PIN switch on the one-channel TSMP receiver.	66
4.9 (a)	Analog switch with driver.	69
(b)	JFET switch.	69
(c)	JFET switch with TTL compatible driver.	69
4.10	Circuit diagram of slow sample-and-hold device	71
4.11	Block diagram of the experimental set-up. ..	76
5.1 (a)	Plots of the measured sum power patterns (at the output of the monopulse comparator) for the azimuth and elevation planes.	82
(b)	Plots of the normalized sum voltage (as a function of the normalized off-axis angle Δ_n) for the azimuth and elevation planes.	82
5.2 (a)	Plots of the measured difference power patterns (at the output of the monopulse comparator) for the azimuth and elevation planes.	83

LIST OF FIGURES (cont'd)

<u>FIGURE NO.</u>	<u>PAGE</u>
5.2(b)	Plots of the normalized difference voltage (as a function of the normalized off-axis angle Δ_n) for the azimuth and elevation planes. 83
5.3	Plot of the standard deviation of azimuth angle jitter for the two-channel TSMP receiver. 86
5.4	Plot of the standard deviation of azimuth angle jitter for the two-channel TSMP receiver with delay. 87
5.5	Plot of the standard deviation of azimuth angle jitter for the one-channel TSMP receiver. 88
5.6	Plot of the standard deviation of elevation angle jitter for the two-channel TSMP receiver. 89
5.7	Plot of the standard deviation of elevation angle jitter for the two-channel TSMP receiver with delay. 90
5.8	Plot of the standard deviation of elevation angle jitter for the one-channel TSMP receiver. 91
5.9(a)	Plot of azimuth boresight shift (as a function of IF SNR) for a constant PRF of 2 KHz. 93
(b)	Plot of elevation boresight shift (as a function of IF SNR) for a constant PRF of 2 KHz. 93

LIST OF TABLESTABLE NO.PAGE

3.1	Differences between the three TSMP receivers described in chapter 3.	52
5.1	Azimuth $\sigma_{\min}$ and K obtained by the least-square method for the three systems.	96
5.2	Elevation $\sigma_{\min}$ and K obtained by the least-square method for the three systems.	97

When the effect of the antenna positioner servo is considered, the noise output power will be reduced by the factor $n = f_r / 2\beta_n$ where f_r is the pulse repetition frequency and β_n is the noise bandwidth of the servo system. Therefore n is the number of pulses integrated in the bandpass of the servo system. The rms thermal noise error σ_t then becomes

$$\sigma_t = \frac{\theta}{k_m \sqrt{B \tau (S/N) (f_r / \beta_n)}} \quad (2-16)$$

If the system is matched, i.e. if $B\tau=1$, then σ_t , the rms thermal noise error in a three channel monopulse radar, is

$$\sigma_{t_s} = \frac{\theta}{k_m \sqrt{(S/N) (f_r / \beta_n)}} \quad (2-17)$$

In the previous section, the concept of time-shared monopulse was introduced. One can compare the difference in the thermal noise error between multi-channel or single-channel monopulse radars as follows. By switching once, one effectively reduces n , the number of pulses integrated, by a factor of two. By switching twice, n is therefore

k	constant with value in the order of unity
k	Boltzmann's constant (1.38×10^{-23} joule/ $^{\circ}$ K)
k_m	normalized monopulse error slope
K	real constant
LORO	Lobing On Receive Only
m	slope function of squinted beams
M	radar measurement
n	number of integrated pulses in the servo loop
N	noise power level
N_{in}	available input noise power
N_{out}	available output noise power
Q	multivibrator output
R_{clamp}	clamp resistor
S	signal power level
s	distance between antenna phase centers
S_{in}	available input signal power
S_{out}	available output signal power
SNR (or S/N)	signal-to-noise power ratio
S&H	sample-and-hold device
t_{on}	turn-on time of analog switch
TSMP	time-shared monopulse radar
V_g	gate voltage of FET
V_{gs}	gate-to-source voltage of FET

V_{sig}	voltage of signal
X	value of resistor
α	angle off-boresight
β_n	noise bandwidth of servo loop
ϵ_m	error in radar measurement
Δ	difference signal
Δ_{az}	azimuth difference signal
Δ_{el}	elevation difference signal
Δ_n	normalized off-axis angle
θ	half-power beamwidth
λ	wavelength
π	pi 3.1415927
Σ	sum signal
σ	standard deviation of angle jitter
σ_1	standard deviation of angle jitter on one-pulse basis
σ_{az}	standard deviation of angle jitter in azimuth
σ_{el}	standard deviation of angle jitter in elevation
σ_{min}	minimum angle jitter level
σ_t	standard deviation of angle jitter due to thermal noise (or thermal noise error)
σ_{t_1}	thermal noise error of one-channel monopulse

σ_{t_2}	thermal noise error of two-channel monopulse
σ_{t_3}	thermal noise error of three-channel monopulse
τ	signal duration
τ_1	time constant of low-pass filter
ϕ	phase difference
ω_s	angular shaft speed of the microwave resolver rotating probe

CHAPTER 1

INTRODUCTION

1.1 Introduction

One of the many requirements in electronic warfare is to acquire and track targets at ranges from a few kilometres to hundreds of kilometres. A radar which must acquire and track a target over a period of time needs the capability of tracking the target in range, in elevation angle and in azimuth angle. Range tracking of the target allows target acquisition and multiple target resolution in range. Angle tracking of the target permits immediate correction of the antenna position in order to maintain antenna alignment with the target.

Many angle-tracking techniques are available to the radar designer. The most widely used technique is called monopulse. Usually a monopulse radar has a three-channel receiver in order to extract full angle information from each pulse.

1.2 Thesis objectives

Angle-tracking is possible with a one-channel or two-channel monopulse receiver as reported by Noblitt (1). Time-sharing of receiver channels has been briefly described by Barton (2,3) and other authors (4).

One of the objectives of this thesis is to implement three receiver configurations which do not have three receiver channels. Two dual-channel (dual-mixer) receivers and one single-channel (single-mixer) receiver will be discussed in the thesis. One of the two dual-channel receivers is part of a radar simulation

facility at Defence Research Establishment Ottawa. Other objectives of this thesis are to measure the angle-tracking error of the three monopulse receivers, to compare the angle-tracking performance of the three receivers, and to determine whether the angle-tracking error is inversely proportional to the square root of the signal-to-noise ratio as theory predicts.

A block diagram showing the different elements of the radar simulation facility is given in Fig. 1.1. A thirty foot long anechoic chamber separates the existing angle-tracking radar (i.e. one of the two dual channel receivers which will be implemented) from the target which is simulated with an array of half-wavelength dipole antennas. Two other arrays are available in order to simulate clutter and jamming noise. Equipment for computer control of the array and for radar signal monitoring is also part of the facility.

1.3 Thesis organization

Chapter 2 of this thesis develops the fundamental theory necessary to understand the angle-tracking process which takes place in a monopulse radar. A description in detail for all three radar configurations is given in chapter 3. The practical considerations which were addressed during the implementation phase are explained in chapter 4. Results of the overall experimentation are reported and compared with theoretical values in chapter 5. Finally, the conclusions and recommendations for future work are presented in chapter 6.

Fig. 1.1: Block diagram of tracking radar simulator at Defence Research Establishment Ottawa.

CHAPTER 2

THEORY OF ANGLE-TRACKING RADARS2.1 Introduction

This chapter is concerned with the concepts and theories which are relevant to the design of the proposed systems. A review of the basic principles of monopulse radar and thermal noise analysis is presented. Concepts of time-shared monopulse radars are introduced. This technique permits a reduction in the number of channels necessary in a monopulse radar. Also explained is how phase information can be contained in the amplitude of a difference signal. When phase sensing is used, angle information is in the relative phase of the signal and when amplitude sensing is used, angle information is in the relative amplitude of the signal since both the relative amplitude and phase of the signal vary across the antenna aperture.

2.2 Principles of monopulse radar

Monopulse is one of three methods employed by continuous angle-tracking radars (5,6). The other two methods are sequential lobing and conical scan. A continuous tracking radar supplies continuous tracking data on a particular target. A track-while-scan radar is not continuous because it only supplies sampled data on one or

more targets.

The basic difference between monopulse-, sequential lobing-, and conical scan radars is the method by which angle information is extracted from the target return. A monopulse radar has two stationary receiving beams in any tracking plane as shown in Fig. 2.1(a). The two squinted beams are used for amplitude sensing and the two parallel beams are employed for phase sensing. A sequential lobing radar has one transmitting and receiving beam switching mechanically between two positions as in Fig. 2.1(b). A conical scan radar has one transmitting and receiving beam moving continuously around the target as pictured in Fig. 2.1(c). A monopulse radar can extract complete angle information on each received pulse in the two tracking planes. Conical scan and sequential lobing radars, on the contrary, cannot extract complete angle information on each received pulse because they do not extract angle information in the same manner as monopulse radars do. In the case of sequential lobing and conical scan radars, the radar video output (6) contains the angle-tracking-error information in the envelope of pulses. The percentage modulation is proportional to the angle-tracking error, and the phase of the envelope function relative to the beam-scanning position contains direction information. Angle-tracking-error detection (error demodulation) is accomplished by a pair of phase detectors using a reference input from the scan motor.

Fig. 2.1: Antenna patterns of three angle-tracking techniques

- (a) Monopulse: squinted beams (top) and parallel beams (bottom)
- (b) Sequential lobing
- (c) Conical scan

Due to mechanical constraints, the scan rate or lobing rate cannot be as great as the PRF (pulse repetition frequency) order of magnitude. Therefore the latter two angle-tracking radars need a higher data rate or more time to process angle information signals to approach monopulse performance.

A key element of a monopulse radar is the antenna. A monopulse antenna has two receiving beams in azimuth and two receiving beams in elevation. In the present system, the beams are formed with a four-horn feed with a parabolic reflector. Then to extract angle information one needs to compare either the amplitudes or the phases of the target return. Amplitude comparison monopulse consists of comparing the amplitudes of the target return for two receiving beams in any plane. Phase comparison monopulse compares the phases of the target return of two receiving beams in any plane. The antenna used in this research employs amplitude sensing in the elevation plane and phase sensing in the azimuth plane. The obtained signals are passed through an RF sum-and-difference network which gives at its output a Σ (sum) signal, a Δ_{az} (azimuth difference) signal, and a Δ_{el} (elevation difference) signal. A sum-and-difference network is a monopulse comparator since it does amplitude comparison for the elevation plane and phase comparison for the azimuth plane.

Rhodes (7) has developed a theory of monopulse where he states that a true monopulse radar satisfies three postulates as follows:

Postulate I: Monopulse information appears in the form of a ratio.

Postulate II: The sensing ratio for a positive angle of arrival is the inverse of the ratio for an equal negative angle.

Postulate III: The angle-output function is an odd, real function of the function of the angle of arrival.

With regard to Postulate I, a monopulse ratio is formed on each received pulse in order to obtain the angle of arrival of the target return independently of its absolute amplitude levels. This means that the absolute amplitudes and absolute phases of the target return may vary with the propagation medium but their relative values are only functions of the angle of arrival. A monopulse ratio, for example, is formed with automatic gain control (AGC) which is required in order to maintain a stable closed loop servo system for the antenna positioner (see Fig. 2.3). The AGC action is obtained by having a voltage proportional to the sum channel intermediate frequency output to control the gain of the receiver channels. With AGC, the output of the angle-error detector is proportional to the difference signal normalized or divided by the

sum signal. A monopulse ratio can also be formed through logarithmic normalization as follows:

$$\log(\Sigma + \Delta) - \log(\Sigma - \Delta) = \log \frac{(1 + \Delta/\Sigma)}{(1 - \Delta/\Sigma)} \quad (2-1)$$

$$= 2 \left(\frac{\Delta}{\Sigma} + \frac{1}{3} \left(\frac{\Delta}{\Sigma} \right)^3 + \dots \right) \quad (2-2)$$

$$= 2 \frac{\Delta}{\Sigma} \text{ for } \frac{\Delta}{\Sigma} < 1 \quad (2-3)$$

where Σ is the sum signal and Δ is a difference signal. So the above mathematics show that a difference of logarithms forms a monopulse ratio.

Postulate II above means that the antenna has to be designed in a manner such that the monopulse ratio is positive when the antenna is pointing to the right or above the target while it is negative when the antenna is directed to the left or below the target. This can be done by having radiation patterns for the Σ and Δ channels as shown in Fig. 2.2(a). The difference signals Δ_{az} and Δ_{el} change sign when the antenna axis is moved past the target. So $\log(\Sigma + \Delta)$ becomes $\log(\Sigma - \Delta)$ and $\log(\Sigma - \Delta)$ becomes $\log(\Sigma + \Delta)$, and the ratio $2\Delta/\Sigma$ changes sign.

Fig. 2.2: (a) Sum and difference power patterns as a function of angle off-boresight.

(b) Error voltage as a function of angle off-boresight.

Rhodes' postulate III means that the angle output function or error signal as a function of the angle of arrival is pictured as in Fig. 2.2(b). The error voltage with odd symmetry shown in Fig. 2.2(b) is obtained when the monopulse ratio $2\Delta/\Sigma$ is used as the antenna positioner drive signal. In summary, Postulates II and III require that a monopulse radar have odd symmetry of the angle output about the boresight axis.

An example of a three-channel monopulse radar is given in Fig. 2.3. The monopulse comparator gives three RF signals, i.e. Σ , Δ_{el} , and Δ_{az} which have to be downconverted to an intermediate frequency for amplitude or phase detection. The Σ signal is used for video purposes and for AGC of the three receiver channels. The Δ signals are used to drive the antenna positioner.

Fig. 2.4 shows a block diagram of a two-channel monopulse radar. As for the three-channel monopulse radar, there are the Σ , Δ_{az} , and Δ_{el} signals. In order to reduce the number of channels, the Δ_{az} and Δ_{el} signals are combined through a microwave resolver which gives at its output:

$$\Delta = \Delta_{az} \cos \omega_s t + \Delta_{el} \sin \omega_s t \quad (2-4)$$

Fig. 2.3: Block diagram of a conventional three-channel monopulse radar receiver.

Fig. 2.4: Block diagram of Noblitt's (1) two-channel monopulse radar receiver.

where ω_s is the angular frequency of the microwave resolver rotating probe. Then the Σ and Δ signals are combined through a 90° hybrid to obtain a $\Sigma+\Delta$ signal and a $\Sigma-\Delta$ signal. The $\Sigma+\Delta$ and $\Sigma-\Delta$ signals are then downconverted to IF for logarithmic detection after which they are passed through a difference network to obtain the $2\Delta/\Sigma$ monopulse ratio and through a sum network to have a video signal. The sum of $\log(\Sigma+\Delta)$ and $\log(\Sigma-\Delta)$ gives

$$\log(\Sigma+\Delta) + \log(\Sigma-\Delta) = \log(\Sigma^2 - \Delta^2) \quad (2-5)$$

$$= 2\log\Sigma + \log(1 - \Delta^2/\Sigma^2) \quad (2-6)$$

$$\approx 2\log\Sigma \quad (2-7)$$

when Δ/Σ is much smaller than one. The main advantage of summing logarithms is that the video pulse is independent of the Δ signal.

There is a definite interest in reducing the number of channels in a monopulse radar receiver so as to reduce the effects of gain and phase imbalances between channels. Errors in angle-tracking accuracy are induced by imbalances between channels. A reduction in the number of channels can also result in monetary savings.

2.3 Time-shared monopulse radar concept

A three-channel monopulse radar can be downgraded to a two-channel or to a one-channel monopulse radar for different reasons such as cost reduction or channel imbalance removal. For example, Noblitt (1) describes a two-channel monopulse which he names conical-scan-on-receive-only (COSRO). A block diagram of a COSRO receiver is given in Fig. 2.4. It is called COSRO because, like conical scan, the receiving beam is moving around the target when tracking but, unlike conical scan, the transmitting beam is stationary.

If one replaces the microwave resolver in Noblitt's (1) circuit by a PIN diode switch, one has another two-channel monopulse radar which involves time-sharing the elevation and azimuth difference channels. A circuit with this option is shown in Fig. 2.5. The time-shared monopulse (TSMP) concept is explained briefly in Barton (2,3). The acronym TSMP is used in order to be consistent with Wilcox (4). The operation of the two-channel TSMP radar of Fig. 2.5 is essentially similar to the COSRO radar. In the former, the PIN switch multiplexes the Δ_{el} and Δ_{az} signals in a discrete manner while in the latter the microwave resolver does so in a continuous manner. The discrete time-multiplexing creates a lobing-on-receive-only (LORO) effect which will be explained later in this section. The PIN switch

Fig. 2.5: Block diagram of a two-channel time-shared monopulse radar receiver.

also removes all information from a tracking plane for one pulse repetition interval (PRI) since it switches to Δ_{el} for one PRI and to Δ_{az} for the other PRI. The angle demultiplexers or demodulators are different because of the nature of the difference signals Δ_{az} and Δ_{el} .

One can also time-share, in addition to the difference channels, the $\Sigma+\Delta$ and $\Sigma-\Delta$ signals as in the circuits of Fig. 2.6 and Fig. 2.7. The $\Sigma+\Delta$ and $\Sigma-\Delta$ signals can be multiplexed by another RF PIN switch (Fig. 2.7) or by a delay line (Fig. 2.6). The circuit of Fig. 2.7 is a one-channel TSMP radar because it has only one mixer and the circuit of Fig. 2.6 is a two-channel TSMP since it has two mixers. These two radar circuits operate in a manner similar to the two-channel TSMP of Fig. 2.5 except that they require a different angle demultiplexer. A more detailed description of the circuit operation of the three TSMP radars which have been introduced in this section will be given later in the third chapter.

As mentioned previously in this section, the PIN switch is causing a LORO action. A block diagram of the monopulse comparator is given in Fig. 2.8. The signals from the antenna feed horns A, B, C, and D are combined through hybrids to give the Σ , Δ_{el} and Δ_{az} signals.

Fig. 2.6: Block diagram of a two-channel time-shared monopulse radar receiver with delay.

Fig. 2.7: Block diagram of a one-channel time-shared monopulse radar receiver.

Fig. 2.8: Block diagram of the monopulse comparator used in all three receiver configurations.

At the 90° hybrid output of the three TSMP radar receivers, the following signals are present:

$$E+\Delta_{az} = 1/2\{(A+B)+(C+D)\} + 1/2\{(B+C)-(A+D)\} = B+C \quad (2-8)$$

$$E+\Delta_{el} = 1/2\{(A+B)+(C+D)\} + 1/2\{(A+B)-(C+D)\} = A+B \quad (2-9)$$

$$E-\Delta_{az} = A+D \quad (2-10)$$

$$E-\Delta_{el} = C+D \quad (2-11)$$

In the case of a two-channel TSMP radar, there is paired lobing of a (B+C) beam and a (A+D) beam followed at the next pulse repetition interval by beams (A+B) and (C+D). In the case of the one-channel TSMP, there is sequential lobing of beams (B+C), (A+D), (A+B) and (C+D), in this respective order. It has to be noted that the lobing action is not a physical movement of the beam in space but a simulated lobing action through the receiver electronics.

The main advantage of TSMP is that fewer IF signal processing channels are needed. The basic advantages of monopulse over conical scan and sequential lobing are preserved. On the other hand, TSMP does not have as good a performance as a standard three-channel monopulse because it has a lower data rate.

TSMP is considered to partly satisfy Rhodes' three postulates for true monopulse. Postulate I is partly satisfied * with logarithmic normalization as used in the circuits of Fig. 2.5, Fig. 2.6 and Fig. 2.7. Postulate II is obeyed if a monopulse antenna is used. Postulate III is obeyed because the error signal is an odd function of the off-boresight angle.

2.4 Thermal noise analysis

Any radar measurement (range, angle, or frequency) is limited by the receiver sensitivity or thermal noise. With regard to angle, the limits in angle measurement (radar tracking accuracy) due to thermal noise are determined by the angle jitter. These effects of the receiver sensitivity become negligible at high signal-to-noise ratios.

The rms error δM of a radar measurement M is expressed by Skolnik (5) as

$$\delta M = \frac{kM}{\sqrt{2E/N_0}} \quad (2-12)$$

* True monopulse needs complete angle information on each received pulse. TSMP does not provide it.

where E is the received signal energy, N_o is the noise power per unit bandwidth, k is a constant whose value is of the order of unity, and M is the beamwidth (for angle measurement). The constant k depends on the shape of the aperture illumination $A(x)$.

The above equation can be related to other well-known radar performance expressions. It is known that

$$E = S\tau \quad (2-13)$$

$$N_o = N/B \quad (2-14)$$

where S is the signal power level, τ is the signal duration, N is the noise power level, and B is the noise bandwidth of the receiver. Substituting θ for M , where θ is the one way half-power beamwidth, and replacing k with $1/k_m$, where k_m is Barton's normalized monopulse error slope, one obtains Barton's (2) rms error on a one-pulse basis σ_1 , which would be the error if the servo could react to each received pulse:

$$\sigma_1 = \frac{\theta}{k_m \sqrt{2B\tau(S/N)}} \quad (2-15)$$

When the effect of the antenna positioner servo is considered, the noise output power will be reduced by the factor $n = f_r / 2\beta_n$ where f_r is the pulse repetition frequency and β_n is the noise bandwidth of the servo system. Therefore n is the number of pulses integrated in the bandpass of the servo system. The rms thermal noise error σ_t then becomes

$$\sigma_t = \frac{\theta}{k_m \sqrt{B \tau (S/N) (f_r / \beta_n)}} \quad (2-16)$$

If the system is matched, i.e. if $B\tau=1$, then σ_t , the rms thermal noise error in a three channel monopulse radar, is

$$\sigma_{t,} = \frac{\theta}{k_m \sqrt{(S/N) (f_r / \beta_n)}} \quad (2-17)$$

In the previous section, the concept of time-shared monopulse was introduced. One can compare the difference in the thermal noise error between multi-channel or single-channel monopulse radars as follows. By switching once, one effectively reduces n , the number of pulses integrated, by a factor of two. By switching twice, n is therefore

reduced by a factor of four. So σ_{t_2} , the thermal noise error for a two-channel TSMP radar, is

$$\sigma_{t_2} = \frac{\theta}{k_m \sqrt{(S/N) (f_r/2\beta_n)}} \quad (2-18)$$

and σ_{t_1} , the thermal noise error for a one channel TSMP, is

$$\sigma_{t_1} = \frac{\theta}{k_m \sqrt{(S/N) (f_r/4\beta_n)}} \quad (2-19)$$

Furthermore one can relate the performance of the different systems since

$$\sigma_{t_3} = \sqrt{2} \sigma_{t_2} = \sqrt{4} \sigma_{t_1} \quad (2-20)$$

Therefore, the thermal noise error for a three-channel monopulse is improved by a factor of two over a one-channel TSMP radar.

2.5 Phase and amplitude sensing with sum-and-difference comparison

The antenna used in this work employs amplitude sensing in the elevation plane and phase sensing in the azimuth plane followed by a sum-and-difference comparator. The elevation plane uses squinted beams and therefore the signals $\hat{A}$ and $\hat{B}$, received respectively by each beam (when tracking), are in phase but have different amplitudes.

Then the difference error signal is

$$\Delta_{el} = \vec{A} - \vec{B} \quad (2-21)$$

$$= (|A| - |B|) / 0^\circ \text{ or } (|A| - |B|) / 180^\circ \quad (2-22)$$

The azimuth plane uses parallel beams and therefore the signals $\vec{A}$ and $\vec{B}$, received respectively by each beam (when tracking) are out of phase but have similar amplitudes. Then the difference error signal is

$$\Delta_{az} = \vec{A} - \vec{B} \quad (2-23)$$

$$= |A|e^{j\frac{\phi}{2}} - |A|e^{-j\frac{\phi}{2}} \quad (2-24)$$

$$= 2j|A|\sin(\frac{\phi}{2}) \quad (2-25)$$

Since in the tracking mode, the phase difference ϕ is usually very small, Δ_{az} reduces to

$$\Delta_{az} = 2j|A|(\phi/2) \quad (2-26)$$

$$= j|A|\phi \quad (2-27)$$

The phase difference ϕ is related to the angle of arrival or the off-boresight angle α by the relation

$$\phi = \frac{2\pi s}{\lambda} \sin \alpha \quad (2-28)$$

where s is the distance between antenna phase centers and λ is the wavelength. Substitution of Eqn (2-28) in Eqn (2-27) gives

$$\Delta_{az} = j|A| \frac{2\pi s}{\lambda} \sin \alpha \quad (2-29)$$

$$= jK \sin \alpha \quad (2-30)$$

where K is a constant. When Δ_{az} is normalized to the sum channel Σ and if one assumes that Σ is constant when the radar is tracking then

$$\frac{\Delta_{az}}{\Sigma} = \frac{jK}{\Sigma} \sin \alpha \quad (2-31)$$

The sine function provides odd symmetry of the angle output about the boresight axis as required by the monopulse postulates. Also it is shown that angle (through phase sensing) information is contained in the amplitude of the Δ_{az} signal.

In the case of Δ_{el} , we have a linear relationship between Δ_{el} and α , for small α ,

$$\Delta_{el} = m\alpha \quad (2-32)$$

where the slope m is a function of the angle between the squinted beams but since this angle is a constant, m is also a constant. When Δ_{el} is normalized to the sum channel Σ

$$\frac{\Delta_{el}}{\Sigma} = \frac{m}{\Sigma} \alpha \quad (2-33)$$

and if Σ is assumed constant when the radar is tracking the linear function in the above equation provides odd symmetry of the angle output about the boresight axis as demanded by the monopulse postulates. Also it is shown that angle (through amplitude sensing) information can be contained in the amplitude of the Δ_{el} signal.

2.5 Summary

In this chapter, elementary concepts of angle-tracking radar techniques have been introduced. Particular emphasis has been placed on time-shared monopulse radars since they are the main subject of this thesis. Thermal noise analysis of tracking errors has been presented to show the effects of lower data rate on angle jitter performance.

Finally, the monopulse comparator outputs Δ_{az} and Δ_{el} have been related to the return signal angle of arrival to show that phase information can be contained in the amplitude of the Δ_{az} signal and amplitude information can be contained in the Δ_{el} signal.

CHAPTER 3

IMPLEMENTATION OF THREE RECEIVERCONFIGURATIONS3.1. Introduction

This chapter presents the design of three TSMP receiver configurations. Firstly the circuitry common to the three receiver configurations is discussed. In subsequent sections, the circuitry which is special to each configuration is described.

3.2 Circuitry common to the three receiver configurations

The three receiver configurations to be realized make use of an APG 502 (Westinghouse) four-horn antenna, antenna pedestal, servo motors, an electronic control amplifier (ECA), and a waveguide sum and difference network as shown in Fig. 3.1. To this equipment, an RF PIN diode switch (General Microwave F9120) is added to time-multiplex the difference channels Δ_{az} and Δ_{el} . The switch is followed by a 90° hybrid (Anaren 1H0568-3) to form the $\Sigma+\Delta$ and $\Sigma-\Delta$ signals. In summary, the inputs to this common block are the azimuth- and elevation drive error voltages and the outputs are the $\Sigma+\Delta$ and $\Sigma-\Delta$ signals. The circuitry is common to all three systems.

Fig. 3.1: Block diagram of circuitry common to the three TSMP receiver configurations.

The APG 502 antenna used is a standard paraboloid reflector with four-horn feed. It has phase sensing in the horizontal plane and amplitude sensing in the vertical plane. Each horn is identified by the letters A, B, C and D. The signals gathered by these four horns are then combined through waveguide networks to give a sum channel Σ and two difference channels Δ_{az} and Δ_{el} . The signals Σ , Δ_{el} and Δ_{az} obtained from the waveguide hybrid network are not in phase. They are either 90 degrees or 180 degrees out of phase. The Σ signal goes through a phase shifter to adjust its phase to be 90 degrees out of phase with the Δ signals while the Δ_{az} signal uses a phase shifter to have the same phase as the Δ_{el} signal. An RF PIN diode switch time-multiplexes the Δ signals followed by two low-noise amplifiers (LNA WJ-K310-170 by Watkins Johnson) which are used for amplification and low noise figure. Then the 90° hybrid is used as an adder or subtractor depending on whether Σ is leading or lagging (at the 90° hybrid inputs) the Δ signal by a 90° phase difference. At the 90° hybrid outputs, $\Sigma + \Delta$ is obtained when Σ and Δ are in phase and $\Sigma - \Delta$ is obtained when Σ and Δ are 180° out of phase.

The main goal of all three systems is to subtract $\Sigma - \Delta_{az}$ from $\Sigma + \Delta_{az}$ and also to subtract $\Sigma - \Delta_{el}$ from $\Sigma + \Delta_{el}$. These subtractions supply the error correction signals which are fed, after amplification, to the electronic control amplifier (ECA). The ECA has been designed specifically by the manufacturer to control magnetic amplifiers.

The latter are used to drive the azimuth and elevation servo motors of the antenna pedestal and hence track the radar target.

3.3 Description of the first receiver configuration

The first receiver configuration implements the two-channel TSMP shown in Fig. 2.5. To realize this receiver, one has to design a special angle demultiplexer.

Let us first consider the RF-to-video conversion as depicted in Fig. 3.2. PIN diode attenuators (General Microwave D1958) are needed to control or regulate the amount of power received. Mixers (RHG DM8-12) downconvert the 9GHz RF signal to the 60MHz IF signal by using a local oscillator which supplies a 9.060 GHz continuous signal. Preamplifiers (RHG ICFT6020) are used to equalize and amplify the IF signals. Logarithmic amplifiers (RHG ICLX6010) provide instantaneous gain control. DC blocking capacitors bring the average value of logarithmic amplifier output noise to zero volts for good performance of the range tracker. A 91 ohm resistor must be the output termination for each logarithmic amplifier as specified by the manufacturer.

The $\log(1+\Delta)$ and $\log(1-\Delta)$ signals are used by the range tracker and also by the angle demultiplexer. First let us look at the video interface to the range tracker. The schematic

Fig. 3.2: RF-to-Video conversion section of the two-channel TSMP radar receiver.

diagram appearing in Fig. 3.3, consists basically of a summing amplifier. One has to add $\log(\Sigma+\Delta)$ and $\log(\Sigma-\Delta)$ to have a signal used by the range tracking subsystem which is independent of the difference signal Δ . This sum gives a series of pulses all with the same amplitude, allowing the range tracker circuit to range track on each video pulse. The constant amplitude of the pulses can be obtained from the summation

$$\log(\Sigma+\Delta) + \log(\Sigma-\Delta) = \log(\Sigma^2 - \Delta^2) \quad (3-1)$$

$$= 2\log\Sigma + \log\left(1 - \frac{\Delta^2}{\Sigma^2}\right) \quad (3-2)$$

$$= 2\log\Sigma \text{ for } \frac{\Delta}{\Sigma} \ll 1 \quad (3-3)$$

In practice Δ is much smaller than Σ so that each pulse has an amplitude $2\log\Sigma$ independent of Δ . The range tracker makes very extensive use of digital circuitry and is not described here. This subsystem employs a video detection threshold which is adjusted to reject the noise but accept the target return signal. When the range tracker range tracks the target return, it supplies the angle demultiplexer with two sampling control signals. These two signals have two functions: first they are a range gate in time, and second they demultiplex the azimuth signals from the elevation signals. Another function of the range tracker is to supply two switching control signals to the RF PIN switch in the common cir-

cuitry of Fig. 3.1.

The timer, like the range tracker, involves complex digital circuitry and is not described here. It provides synchronism amongst all the units of the radar system.

The CRT or cathode ray tube is used to display the video return signal in range and intensity.

As shown in Fig. 2.5, the difference of the two log detector outputs is used by the angle demultiplexer. A difference amplifier and the angle demultiplexer are drawn in Fig. 3.4 while the drive amplifier is pictured in Fig. 3.5. The angle demultiplexer consists mainly of two fast sample-and-hold (Datel SHM-UH) devices which sample every second PRI but not simultaneously. The necessary timing diagram is illustrated in Fig. 3.6 to give an idea of the demultiplexing action. As shown in Fig. 3.6, the sampling event $2\Delta/\epsilon$ is the difference between the sampling events $\log(\epsilon+\Delta)$ and $\log(\epsilon-\Delta)$. The azimuth sample-and-hold device samples the azimuth $2\Delta/\epsilon$ signal and the elevation sample-and-hold device samples the elevation $2\Delta/\epsilon$ signal. Furthermore there is an RC filter between the sample-and-hold devices (S&H) and their respective drive amplifiers. Its function is to smooth out the S&H outputs since they have droop problems as shown in Fig. 3.6. The drive amplifiers are high gain amplifiers which provide the voltage levels required by the antenna positioner.

Fig. 3.4: Angle demultiplexer for the two-channel TSMP radar receiver.

Fig. 3.5: Drive amplifier of two-channel TSMP radar receiver.

Fig. 3.6: Timing diagram for the two-channel TSMP radar receiver.

3.4 Description of the second receiver configuration

The second configuration implements the two-channel TSMP radar receiver with delay shown in Fig. 2.6. As in the first two-channel TSMP receiver, this system has PIN attenuators, mixers and preamplifiers. Mixer-preamplifiers (RHG MDM8-12/12A) are used instead of separate mixers and preamps as before. A diagram of the RF-to-video conversion circuit is given in Fig. 3.7. The delay line is a 1.6 μ sec coaxial delay line. The $(I-A)$ signals are delayed with respect to the $(I+A)$ signals and then both signals are combined through a power divider. The preamplifiers are used as gain equalizers. The logarithmic amplifier provides instantaneous gain control and video conversion.

The $\log(I \pm A)$ signals are used by the range tracker and also by the angle demultiplexer. First let us look at the simple interface required between them and the range tracker. As shown in Fig. 3.8, the interface is basically an amplifier with a gain of 10. It fulfills the same function as the summing amplifier of the previous configuration.

Now consider the block diagram, Fig. 3.9, of the angle demultiplexer to understand how the demultiplexing action is done. The demultiplexing problem is to sample four different pulses instead of two as previously. This is done by first sampling all

Fig. 3.7: RF-to-Video conversion section of the two-channel TSMP receiver with delay.

Fig. 3.8: Video interface to the range tracker for the two-channel TSMP receiver with delay and for the one-channel TSMP receiver.

Fig. 3.9: Functional diagram of the angle demultiplexer for the two-channel TSMP receiver with delay and for the one-channel TSMP receiver.

$\log(\Sigma+\Delta)$ and $\log(\Sigma-\Delta)$ signals and, then selecting the azimuth or the elevation difference from the difference amplifier. So the $\log(\Sigma+\Delta)$ and $\log(\Sigma-\Delta)$ signals have to be sampled at one PRI each but the $2\Delta_{az}/\Sigma$ and $2\Delta_{el}/\Sigma$ signals have to be sampled every two PRI each.

A more detailed version of the angle demultiplexer appears in Fig. 3.10. A fast S&H device (Datel SHM-UH3) is required in order to have acquisition time in the order of tens of nanoseconds. However its held output was found to droop approximately one volt per millisecond. Such a droop prevents good tracking performance and therefore a slow S&H device was built to sample the signal 2 μ sec after the fast S&H device had sampled it. The design of this slow S&H is presented later in the thesis. A slow S&H has less droop than a fast S&H because, due to the size of its output capacitor, it acquires the signal more slowly and can hold it longer. A timing interface is necessary between the range tracker and the different sample-and-hold devices. It will be described later. The drive amplifiers are the same as those used in the first configuration except that no RC filter is used in this configuration since the servo loop itself acts as an RC filter.

A timing diagram for this system is supplied in Fig. 3.11. It explains what happens to each video pulse as it goes through the demultiplexer. This is done by showing the outputs of the four slow sample-and-hold circuits as a function of time and with

Fig. 3.10: Block diagram of the angle demultiplexer for the two-channel TSMP receiver with delay and for the one-channel TSMP receiver.

Fig. 3.11: Timing diagram for the two-channel TSMP receiver with delay.

correspondence to the $\log(\Sigma\Delta)$ sequence. No droop is shown on these diagrams because slow sample-and-hold circuits have much less droop than fast sample-and-hold devices. The droop of a slow S&H was observed to be negligible over one PRI.

3.5 Description of the third receiver configuration

The third receiver configuration implements the one-channel TSMP receiver shown in Fig. 2.6. A block diagram of the RF-to-video conversion section is given in Fig. 3.12. Contrary to the two previous systems, it has only one mixer-preamp (RHG MDM8-12/12A). Two PIN attenuators are shown but only one is essential. It is more convenient in the module interchange between systems to have two attenuators. A special timing interface is necessary between the range tracker and the RF PIN switch of the common circuitry described in section 3.2. The latter switch opens and closes every two PRI while the switch in the RF-to-video conversion part opens and closes every PRI.

As in the second configuration, this system has only one logarithmic amplifier and has four pulses to demultiplex. Since the four pulses arrive at the angle demultiplexer in the same order, the same demultiplexer shown in Figs 3.9 and 3.10 is used with a minor change in the timing interface between the range tracker and demultiplexer. It will be shown later how the timing interface is designed. A timing diagram is shown in Fig. 3.13.

Fig. 3.12: RF-to-Video conversion section of the one-channel TSMP receiver.

Fig. 3.13: Timing diagram for the one-channel TSMP receiver.

The $\log(L-\Delta)$ pulse always follows the $\log(L+\Delta)$ pulse by design..

Fig. 3.13 illustrates the sampling events as they occur in time.

The $\log(L+\Delta)$ sampling event shows that $\log(L+\Delta_{az})$ is sampled two PRI after (or before) $\log(L+\Delta_{el})$ is sampled. The $\log(L-\Delta)$ sampling event demonstrates that $\log(L-\Delta_{az})$ is sampled two PRI after (or before) $\log(L-\Delta_{el})$ is sampled. The azimuth and elevation S&H devices sample the azimuth and elevation differences respectively.

3.6 Summary

Table 3.1 summarizes the differences between the three receiver configurations which have been described in this chapter.

TABLE 3.1

DIFFERENCES BETWEEN THE THREE TSMP RECEIVERS
DESCRIBED IN CHAPTER 3

CIRCUIT DIFFERENCES	2-CHANNEL TSMP	2-CHANNEL TSMP WITH DELAY	1-CHANNEL TSMP
Mixer	2	2	1
Mixer preamp	2	2	1
Log amp	2	1	1
PIN switch	1	1	2
Delay line	0	1	0
Fast S&H	2	2	2
Slow S&H	0	4	4
Video interface	0	1	1
Timing interface	0	1	2

CHAPTER 4

PRACTICAL CONSIDERATIONS4.1 Introduction

The practical considerations that were not evident during the initial development of the three receiver configurations are now discussed. These range from the limits of the receiver sensitivity to the servo bandwidth. The experimental set-up which is used to test the three systems is also described.

4.2 Receiver noise limits

The receiver noise limits the dynamic range of the IF signal-to-noise ratio. Noise in the receiver is observed at the outputs of the logarithmic amplifiers as a noise floor with power level at about -60 dBm. This value can be predicted by the receiver noise figure. The noise figure (5) of a linear network may be defined as

$$F_n = \frac{S_{in}/N_{in}}{S_{out}/N_{out}} = \frac{N_{out}}{kT_o B G} \quad (4-1)$$

where S_{in} is the available input signal power, N_{in} is the available input noise power (equal to $kT_o B_n$), S_{out} is the available output signal power, and N_{out} is the available output noise power. The available gain G is equal to S_{out}/S_{in} , k is Boltzmann's constant i.e., $1.38 \times 10^{-23} \text{ J/}^\circ\text{K}$, T_o is the standard temperature of 290°K and B_n is the noise bandwidth

$$B_n = \frac{\int_{-\infty}^{\infty} |H(f)|^2 df}{|H(f_o)|^2} \quad (4-2)$$

where $H(f)$ is the frequency response characteristic of the IF amplifier or filter, and f_o is the frequency of maximum response which usually occurs at midband. The 3-dB bandwidth is not identical to the noise bandwidth but can be used as an approximation (5) when the 3-dB and the noise bandwidth do not differ appreciably (the IF bandwidth is 20 MHz). So the available thermal noise power at the receiver output is

$$N_{out} = F_n (kT_o B_n) G \quad (4-3)$$

The overall noise figure F_n for a cascaded amplifier is given by (5)

$$F_n = F_1 + \frac{F_2 - 1}{G_1} \quad (4-4)$$

where F_1 is the noise figure for the first stage, G_1 is the gain of the first stage, and F_2 is the noise figure of the second stage.

Since the low-noise amplifier (LNA) has a gain of 20 dB and a noise figure of 4 dB while the mixer/IF preamplifier combination has a noise figure of 9.2 dB, Eqn. (4-4) yields

$$F_n = 2.5 + \frac{8.3 - 1}{100} \quad (4-5)$$

$$= 2.573 \text{ or } 4.1 \text{ dB} \quad (4-6)$$

Now assuming B_n to be the IF bandwidth, 20 MHz, we can obtain N_{out} from Eqn. (4-3) for various gains. For example, for a gain of $G=10^4$, N_{out} is -56.86 dBm, for a gain of $G=10^3$, N_{out} is -66.86 dBm, and for a gain of $G=10^2$, N_{out} is -76.86 dBm.

The typical values in the above paragraph clearly show that the receiver noise can reduce the SNR dynamic range of the logarithmic amplifiers because, for all three available gains, the receiver noise output is higher than -80 dBm which is the logarithmic amplifier threshold. The receiver noise shows up at the output of the log amp as a noise floor on which is superimposed the target return or video pulse. Therefore the dynamic range of available video pulse

signal levels is limited to the difference between the noise floor and the saturation levels of the log amp.

4.3 Timing considerations

The range tracker is so designed as to give two sampling control signals at two pulse repetition intervals. These sampling signals are synchronized with the RF PIN switch operation. For instance, when the range tracker issues a control signal to sample the azimuth error signal it also causes the PIN switch to turn on the Δ_{az} channel. Of course a similar situation occurs for the elevation error signal and the Δ_{el} channel. In other words, the range tracker has been designed for the first configuration, i.e. the two-channel TSMP receiver.

One has to make the range tracker operation compatible with the demultiplexing of four pulses. Therefore a timing interface is needed as mentioned in chapter 3. The video pulses are in the following sequence: $\log(I+\Delta_{az})$, $\log(I-\Delta_{az})$, $\log(I+\Delta_{el})$, and $\log(I-\Delta_{el})$. Demultiplexing is done by first sampling $\log(I+\Delta)$ and holding it until the $\log(I-\Delta)$ pulse arrives. When $\log(I-\Delta)$ arrives, the difference is taken immediately after it is sampled in order to have an error voltage. This signal processing philosophy is realized by the configuration shown in Fig. 3.9 and in Fig. 3.10.

When the range tracker receives a train of pulses as above, it issues two sampling control signals which will sample $\log(\Sigma+\Delta)$ and $\log(\Sigma-\Delta)$. So it naturally demultiplexes $\log(\Sigma+\Delta)$ and $\log(\Sigma-\Delta)$. One has only to delay the $\log(\Sigma-\Delta)$ sampling signal in order to generate two 4PRI sampling signals for the azimuth and elevation error voltages. This is implemented as shown in Fig. 4.1. The range tracker issues 40 nanosecond sampling pulses which are not sufficient to make the slow sample-and-hold circuit work. Therefore a multivibrator (TI SN74LS221) is used to stretch the $\log(\Sigma+\Delta)$ sampling pulse and also the $\log(\Sigma-\Delta)$ sampling pulse. The latter pulse is used to generate a 4PRI control signal with a multivibrator (TI SN74LS221). The 4PRI control signal is fed to the inputs of two more multivibrators (TI SN74LS221). One works on the rising edge and the other works on the falling edge of the 4PRI control signal. This results in two sampling control signals at 4PRI: one for the elevation error slow sample-and-hold device and the other for the azimuth error slow sample-and-hold device. The circuit diagram for this timing interface is drawn in Fig. 4.2. A timing diagram is given in Fig. 4.3.

The timing interface is more complicated for the two-channel TSMP with delay than for the one-channel TSMP receiver. As shown in Fig. 4.4, one has to combine the range tracker sampling signals through an OR gate in order to sample every PRI because there is a

Fig. 4.1: Block diagram of the timing interface for the one-channel TSMP receiver.

Fig. 4.2: Circuit diagram of the timing interface for the one-channel TSMP receiver.

Fig. 4.3: Timing diagram for the timing interface of the one-channel TSMP radar receiver.

Fig. 4.4: Block diagram of the timing interface for the two-channel TSMP radar receiver with delay.

$\log(\Sigma+\Delta)$ to sample every PRI. The timing diagram of Fig. 4.6 shows this requirement. But there is a $\log(\Sigma-\Delta)$ following $\log(\Sigma+\Delta)$ by a delay of approximately two microseconds. So to sample $\log(\Sigma-\Delta)$, one has to delay the $\log(\Sigma+\Delta)$ sampling signal by two microseconds with the use of two multivibrators. The output of the second multivibrator can be used, after going through a 50 ohm line driver, as a fast sample-and-hold sampling control pulse. This output is stretched to be used as a slow sample-and-hold sampling pulse. The latter is used to generate a 2PRI control signal which is fed to the inputs of two multivibrators. One works on the transition low-high and the other works on the transition high-low. This gives two sampling control signals at 2PRI, one for the elevation error slow sample-and-hold circuit and the other for the azimuth error slow sample-and-hold circuit. The circuit diagram for this timing interface is drawn in Fig. 4.5 and timing diagram is given in Fig. 4.6.

Furthermore, a timing interface between the range tracker and the RF PIN switch is necessary in the one-channel TSMP receiver configuration. The range tracker naturally gives two switching control signals at 2PRI. As shown in Fig. 4.7, one has to take one of these signals and create two 4PRI switching control signals. This is done with a multivibrator which is fed one 2PRI switching signal and it gives two outputs at 4PRI one at the Q output and the other at the $\bar{Q}$ output. A timing diagram is given in Fig. 4.8.

Block diagram of the frequency response measurement system. The system consists of a RANGE TRACKER, a NOR GATE, and three MULTIVIBRATORs.

The RANGE TRACKER is connected to the NOR GATE. The NOR GATE has two outputs: FAST S8Hz and SLOW S8Hz.

The FAST S8Hz output is connected to the first MULTIVIBRATOR. The SLOW S8Hz output is connected to the second MULTIVIBRATOR. The third MULTIVIBRATOR is connected to the output of the second MULTIVIBRATOR.

Each MULTIVIBRATOR has a 5V input, a 27K resistor, a 120pF capacitor, and a 15V output.

The output of the first MULTIVIBRATOR is labeled FAST S8Hz $\log(z-d)$. The output of the second MULTIVIBRATOR is labeled SLOW S8Hz $\log(z-d)$. The output of the third MULTIVIBRATOR is labeled SLOW S8Hz (E1).

63

Fig. 4.6: Timing diagram for the timing interface of the two-channel TSMP receiver with delay.

X(ohms)	PRF(Hz)
6.8K	5000
15.7K	2000
30.1K	1000
40.7K	750
59.4K	500
84.3K	350
145K	200

Fig. 4.7: Circuit diagram of the timing interface between the range tracker and the PIN switch for the one-channel TSMP receiver.

Fig. 4.8: Timing diagram for the timing interface between the range tracker and the PIN switch of the one-channel TSMP receiver.

4.4 Sample-and-hold considerations

4.4.1 Introduction

A sample-and-hold is a device which samples a signal and holds it until another sample is taken. A fast sample-and-hold device is needed because the video pulse (target return) has a 0.5 μsec duration. The Datel SHM-UH and SHM-UH3 can acquire a signal within 50 nsec but have a minimum droop of 50 $\mu\text{V}/\mu\text{sec}$. This droop is unacceptable for proper demultiplexing operation since the sampled signal has lost its amplitude information by the end of one PRI. The bad droop problem is corrected with the addition of a slow sample-and-hold circuit in cascade with a fast sample-and-hold (S&H) device. A slow S&H circuit can hold the signal well enough to preserve amplitude information over one PRI.

Since the demultiplexing configuration of Fig. 3.9 was selected, one needs to add a slow S&H circuit after each fast S&H circuit as shown in Fig. 3.10. This is necessary due to the subtraction process which has to take place after $\log(I+A)$ and $\log(I-A)$ have been sampled. The amplitude information contained in $\log(I+A)$ and $\log(I-A)$ must be preserved. Otherwise the obtained subtraction can be meaningless for tracking purposes.

In the case of the first configuration, there is no need for a slow S&H because the subtraction process is done prior to sampling and also because the sampled value is of the order of a few millivolts. When tracking a target the bad droop of a fast S&H device is much less significant when the sampled value is close to zero volts because the droop tends toward zero volts.

To preserve amplitude information, an analog-to-digital converter could have been used but it was found that a slow sample-and-hold can sample a signal much faster than it takes a cheap A/D converter to do analog-to-digital conversion. Therefore analog signal processing was found more advantageous than digital signal processing for the angle demultiplexer.

4.4.2 Design of a slow sample-and-hold device

A slow sample-and-hold device can be purchased but, due to time constraints, one was built as follows. To design a sample-and-hold one usually needs (8) an analog switch and a switch driver. This is pictured in Fig. 4.9(a). If a FET is used as the analog switch, one needs to design a driver compatible with proper operation of the FET. A JFET shunt-resistor driver basically is as shown in Fig. 4.9(b). The JFET is on when $V_{gs} = 0$ and this state may be achieved by placing a resistance between the gate and the source. So long as the current through this resistance is zero, the gate-to-source voltage will also be zero.

Fig. 4.9: (a) Analog switch with driver
 (b) JFET switch
 (c) JFET switch with TTL compatible driver.

A slow sample-and-hold circuit was designed, using the TTL compatible driver shown in Fig. 4.9(c). Operation of the circuit is described with reference to the circuit diagram of Fig. 4.10. Assume the sampling control is 5 volts. Q_1 is on and I_1 turns Q_2 on. This pulls its collector to $-V_{cc}$ which is -15 volts. Since the gate of Q_3 is tied to the collector of Q_2 , the gate voltage V_g is also at $-V_{cc}$. In this condition Q_3 is off and will remain off so long as the most negative value of V_{sig} is greater than $\{V_g - V_{gs}(\text{off})\}$. If $V_{gs}(\text{off})$ is assumed to be $-3.5V$ then the most negative allowable input signal is

$$V_{sig} = -15 - (-3.5V) = -11.5V \quad (4-5)$$

If the rated V_{sig} is 10 volts, then I_s will flow when V_g is at $-15V$ (Q_3 is off). The value of I_s depends on R_{clamp} and introduces a trade-off between switching time t_{on} and leakage current I_s : the faster the switching time, the larger the leakage current. From Siliconix data, a value of $10K\Omega$ for R_{clamp} results in $t_{on} = 0.821 \mu\text{sec}$ and $I_s = 2,500 \mu\text{A}$.

From the description in the above paragraph, one can notice that the JFET driver works on negative logic. In effect, when the sampling control voltage is 5 volts, the switch is off and when it is 0 volts, the switch is turned on.

Fig. 4.10: Circuit diagram of slow sample-and-hold device.

Input and output buffers are necessary. The output buffer, a FET-input operational amplifier, is needed to present an high output impedance to the capacitor when the switch is off. An input buffer is needed for the JFET driver to provide an high input impedance to the timing interface circuitry.

4.5 Video pulse consideration

For proper operation, one needs to supply the range tracker with a good video signal. A video signal is good if it is well above the noise and does not drop below the range tracker video threshold level.

In the first configuration, the $\log(I+\Delta)$ and $\log(I-\Delta)$ video pulses are simply added to give $2\log I$. The range tracker is thus supplied with pulses which all have the same amplitude. This is very important when the IF signal-to-noise ratio is very small. In effect, the range tracker video threshold has to be adjusted just above the noise level. If pulses are irregular in amplitude and the video threshold is set to be just above the noise level, there will be a problem caused by the pulses which are below the video threshold.

In the second and third configuration, the $\log(I+\Delta)$ and $\log(I-\Delta)$ video pulses are not added, so a different video interface has to be designed. As stated previously, one needs a video signal well above the noise level for proper operation. The best way is

to use a gain amplifier as shown in Fig. 3.8. Such an interface is desired because the emphasis of this thesis is put on angle-tracking performance rather than on range-tracking performance. Also most of the experimental results will be taken at low signal-to-noise ratios.

The logarithmic amplifiers always provide a noise floor. It is necessary to lower this noise floor to make use of the dynamic range of the range tracker. A DC blocking capacitor $.1\mu\text{F}$ inserted between the logarithmic amplifier and the 91Ω resistor as shown in Figs. 3.2, 3.7 and 3.12 performs this function. The video interface in all configurations is connected to the 91Ω resistor, while the angle demultiplexer is connected between the logarithmic amplifier and the DC blocking capacitor. Some of the amplitude information is lost in the removal of the noise floor, and therefore, for proper angle demultiplexer operation, the connection needs to be made before the capacitor.

4.6 Error voltage output filter

In the first configuration, each sample-and-hold output is filtered by a low pass RC filter as shown in Fig. 3.5. Filtering is done to remove transients or to smooth out the error voltage output signal. The time constant of this low pass filter is

$$\tau_l = RC = (620\Omega)(25\mu F) = 15.5 \text{ msec} \quad (4-6)$$

The servo bandwidth usually is of the order of a few hertz and therefore gives a time constant much greater than the one above. This means that the overall effect of the error voltage output filter is not seen directly in the performance of the radar because the servo filter has a lower cut-off frequency than the output filter.

To prove that the effect of the output filter is negligible, the second and third configurations were tested with and without the output filter. There was no appreciable difference in the radar performance.

4.7 Determination of servo bandwidth

The servo bandwidth limits the speed of the radar response to an input change. The servo bandwidth was determined from the study of the radar response to a square-wave input. Experimentally, a square wave input is simulated by the switching of two dipoles in the anechoic chamber.

In theory, the risetime is inversely proportional to the 3-dB bandwidth. The risetime is defined as the time it takes to go from 10% to 90% of the final value. From the square-wave test, the risetime was found to be 0.5 sec. Consequently, the servo bandwidth is 2 Hz.

4.8 Design of an experimental set-up

The best way to examine performance for such a complex system is as follows. As long as the system maintains angle tracking, the main parameters which affect performance, i.e. radar return signal level and pulse repetition frequency, can be varied. The random error or jitter in boresight position with which the system tracks the target can be measured as these parameters are varied. The resulting jitter has a Gaussian probability distribution with a standard deviation which increases if the signal-to-noise ratio decreases.

The experimental set-up is shown in Fig. 4.11. For pulse generation, there is a pulse-forming chain consisting of the timer, a programmable pulse generator and a radar target simulator. This chain determines the pulsewidth and other target simulation features such as the range and the speed. An RF CW signal is fed into a microwave source modulator as the frequency reference signal. The RF output signal can be attenuated over a dynamic range of 100 dB.

The RF output signal is amplified by a TWT before being radiated through an anechoic chamber. Targets are simulated with an array of dipoles of half wavelength. One dipole, for example, represents a point source target. Transmitted power level can be

Fig. 4.11: Block diagram of the experimental set-up.

read over 40 dB dynamic range at the output of an RF logarithmic detector.

After propagating through the anechoic chamber, the signal is picked up by the four-horn antenna dish. After being processed by the radar receiver, it can be observed as a video signal at the outputs of the receiver logarithmic amplifiers. This gives an approximate value of the power levels, both signal and noise levels, received.

On the antenna structure, there is an elevation and an azimuth synchro motor. These two synchros will give a reading on the antenna position in two coordinates at the same time and for the same experimental conditions. So it is possible to have an exact knowledge of the antenna position in relative terms at all times. To have it in absolute terms would require the installation of optical position-sensing devices.

Since the main experiment is to study the effects of a random error over time, the synchro output data need to be recorded for a certain period of time before being stored for later data processing. This is done by doing an analog-to-digital conversion on the analog synchro outputs for a period of 100 seconds at a sampling frequency of 10 Hertz. The data gathered is then stored on magnetic disks.

Data processing is done on the data banks in two ways as follows. The computer, the same one as used for analog-to-digital conversion, controls a digital plotter which gives the data probability distribution. This is helpful in determining whether the data distribution is Gaussian. The computer also must calculate the mean and the standard deviation of the synchro output data. The obtained standard deviation is a measure of the angle jitter in absolute terms.

The tracking error is always the sum of a bias and a random error. The jitter or random error is actually the amount of movement around the mean which does not necessarily correspond to the true position of the target. In effect, the difference between the mean and the true position of the target is the bias error.

4.9 Logarithmic amplifier readings of signal-to-noise ratio

A logarithmic amplifier has typical specifications as follows (9)

Center frequency/bandwidth	60/20 MHz
Input dynamic range	-80 to 0 dBm
Log accuracy	1 dB
Rise time	50 μ sec
Video output	0.1 to 2.1 volts

Because a logarithmic amplifier has a linear relationship between the IF input power and the video output voltage, it can be used as an IF power measurement device. These measurements are fast and give an approximate value of the IF power level. This is particularly helpful in IF signal-to-noise ratio measurements. In effect, one can have an approximate value of the signal IF power level by looking at the output voltage reached by the signal video pulse. Also, an approximate value can be obtained for the IF noise level by reading the average value of the noise floor on an oscilloscope. Once both signal and noise power levels are measured, a difference is made between the two levels in order to obtain the IF SNR.

4.10 Summary

This chapter has presented the practical considerations which had to be dealt with during the implementation of the three receiver configurations. The main considerations were the timing of sampling control signals, the design of a slow sample-and-hold circuit and the description of the experimental set-up.

CHAPTER 5

EXAMINATION OF THE PERFORMANCE5.1 Introduction

In this chapter an experimental evaluation of the three systems is made. Experimental results are compiled and interpreted. A comparison is then made between the three configurations.

5.2 Experimental results5.2.1 Experimental determination of k_m

As per Barton (2), the normalized monopulse error slope k_m can be obtained from measurements of the sum, elevation and azimuth difference RF signal power patterns. The slope k_m is a determining factor in the thermal noise equation (refer to section 2.4). To find k_m , one has to normalize the difference voltage (obtained from the azimuth difference or elevation difference RF signal). Since the ratio of two voltages is equal to the square root of the ratio of their respective powers, the normalized difference voltage is obtained by dividing the difference signal power by the sum signal maximum power.

22
OF / DE

The sum and difference RF signal power patterns as a function of the off-boresight angle are given in Fig. 5.1(a) and Fig. 5.2(a) respectively. The sum and difference patterns are different in azimuth and in elevation because of the shape of the paraboloid antenna. The manufacturer specifies the half-power beamwidths to be 3.6° in azimuth and 6.2° in elevation.

The normalized sum and difference voltage patterns as a function of the normalized off-axis angle Δ_n are given in Fig. 5.1(b) and Fig. 5.2(b) respectively. Δ_n is the off-boresight angle normalized to the half-power beamwidths. The sum and difference voltage patterns are (within experimental measurement accuracy) identical in the azimuth and the elevation plane. Both sum and difference voltage patterns are normalized to the maximum sum voltage, i.e., elevation difference and sum are normalized to the maximum elevation while azimuth difference and sum are normalized to the maximum azimuth sum.

The observed normalized sum voltage E_r approaches the following empirical relation(2)

$$E_r = \cos^2(1.14\Delta_n) \quad (5-1)$$

(a)

(b)

Fig. 5.1: (a) Plots of the measured sum power patterns (at the output of the monopulse comparator) for the azimuth and elevation planes.
 (b) Plot of the normalized sum voltage (as a function of the normalized off-axis angle Δ_n) for the azimuth and elevation planes.

Fig. 5.2: (a) Plots of the measured difference power patterns (at the output of the monopulse comparator) for the azimuth and elevation planes.
 (b) Plot of the normalized difference voltage (as a function of the normalized off-axis angle Δ_n) for the azimuth and elevation planes.

while the observed normalized difference voltage E_e approaches the relation

$$E_e = 0.707 \sin (2.28\Delta_n) \quad (5-2)$$

By taking the derivative of E_e , one can find k_m which is the slope of the normalized difference voltage

$$\frac{dE_e}{d\Delta_n} = 1.6 \cos (2.28\Delta_n) \quad (5-3)$$

The value of k_m is found by making Δ_n equal to zero in the above equation

$$k_m = \left. \frac{dE_e}{d\Delta_n} \right|_{\Delta_n=0} = 1.6 \quad (5-4)$$

In summary, k_m is identical for the azimuth and elevation tracking plane. Therefore if one looks at the thermal noise equation (Eqn.(2-17)) one sees that the tracking planes differ (with regard to angle jitter performance) only in their respective beamwidths because the other variables in Eqn.(2-17) do not change with tracking plane.

5.2.2 Angle jitter measurements

The influence of thermal noise on angle jitter measurements can be recorded when the single-pulse SNR is less than 20 dB in the sum channel (3). The single pulse SNR at IF was set at approximately 10 dB for the three systems. The standard deviation of the angle jitter about the average antenna position was then recorded for various values of pulse repetition frequency.

Plots of the standard deviation of angle jitter versus pulse repetition frequency for all three configurations in azimuth and in elevation are given in Figs. 5.3, 5.4, 5.5, 5.6, 5.7 and 5.8. The variation of σ with PRF was modelled as $\sigma = \sigma_{\min} + K \cdot \text{PRF}^{-1/n}$ for $n = 1, 2, 3$. The best least-squares fit was found to occur for $n = 2$. The solid lines through the experimental points are obtained by curve fitting the above relation with $n = 2$ using the least-squares method; the fitted equations are shown in Figs. 5.3-5.8. Measurement accuracy is estimated to be ± 0.05 mrad. Some angle jitter measurements (in Figs. 5.4, 5.5, 5.7 and 5.8) fall more significantly away from their respective fitted curves than other angle jitter measurements (in Figs. 5.3 and 5.6) because the angle demultiplexer used in the two-channel TSMP receiver with delay and in the one-channel TSMP receiver induces an additional error due to the sampling of the $\log(I+\Delta)$ and $\log(I-\Delta)$ signals before the meaningful subtraction result $2\Delta/I$ is obtained. In Fig. 5.7 one measurement (at 5000Hz) was removed from the values used in the curve-fitting calculations because its value was much too great to be within the normal decrease (due to an increase in PRF) in angle jitter values.

Fig. 5.3: Plot of the standard deviation of azimuth angle jitter for the two-channel TSMP receiver.

Fig. 5.4: Plot of the standard deviation of azimuth angle jitter for the two-channel TSMP receiver with delay.

Fig. 5.5: Plot of the standard deviation of azimuth angle jitter for the one-channel TSMP receiver.

Fig. 5.6: Plot of the standard deviation of elevation angle jitter for the two-channel TSMP receiver.

Fig. 5.7: Plot of the standard deviation of elevation angle jitter for the two-channel TSMP receiver with delay.

Note: The value at 5000Hz is not included with the values used in the least-squares curve fitting.

Fig. 5.8: Plot of the standard deviation of elevation angle jitter for the one-channel TSMP receiver.

5.2.3 Boresight shift measurements

To measure boresight shift, it is necessary to establish the target position relative to the boresight antenna axis. Since the synchro outputs are proportional to the boresight position, the target position was taken to be the average of boresight positions measured from the synchro outputs. The pulse repetition frequency was selected to be 2 KHz and only the single pulse SNR was varied.

Boresight shift plots for the azimuth and the elevation planes are given in Fig. 5.9(a) and Fig. 5.9(b) respectively. These plots show that the antenna position changes with a variation in signal level (i.e. SNR) and that these changes are different (in their magnitudes and their directions) from one receiver configuration to the other.

5.3 Interpretation of results

5.3.1 Angle jitter measurements

The angle jitter consists of two components—random and systematic errors. The thermal noise error is part of the random errors while the minimum level of angle jitter which can be obtained is part of the systematic error. The minimum angle jitter level is independent of the signal strength, the target dynamics, the target range

Fig. 5.9: (a) Plot of azimuth boresight shift (as a function of IF SNR) for a constant PRF of 2KHz.
 (b) Plot of elevation boresight shift (as a function of IF SNR) for a constant PRF of 2KHz.

and the antenna elevation angle. The target which is simulated in this thesis is at fixed range and is not moving. Furthermore, the antenna elevation angle does not induce a random (multipath) error in an anechoic chamber due to the wall microwave absorption. Therefore, only the thermal noise error should be observed because the only parameter to be varied is the signal strength.

Thermal noise analysis has shown that angle jitter decreases when the PRF increases and that it theoretically follows the relation (3)

$$\sigma = \sigma_{\min} + \frac{K}{\sqrt{\text{PRF}}} \quad (5-5)$$

where $\sigma_{\min}$ is the the standard deviation of the minimum angle jitter and K is a constant independent of the PRF. Referring to Eqn.(2-16), one sees that K is a function of the antenna beamwidth θ , the single pulse SNR, the servo bandwidth B_n , the normalized monopulse error slope k_m , the IF bandwidth B and the pulse duration τ . The variables composing K are all constant in the measurement of angle jitter when only the PRF is varied.

The noise in the measuring system, mainly synchro output noise, will be part of $\sigma_{\min}$ because such noise is not necessarily independent of the antenna position. The magnitude of this noise was found to be significant. When the antenna was locked both in

azimuth and elevation, the measured noise was .11 mrad in azimuth and .39 mrad in elevation. This test could be done only at the zero degree position in both planes.

As stated in section 5.2.2, the least-squares method of curve fitting was used to join experimental points. Experimental values of $\sigma_{\min}$ and K were then obtained from the least-squares fit. Values for these parameters for the three configurations and both tracking planes are given in Table 5.1 and Table 5.2. These values characterise the performance of the three TSMP receivers. They show that the three TSMP receivers have different minimum angle jitter levels and different thermal noise errors for the same experimental conditions.

5.3.2 Boresight shift measurements

By inspection of Fig. 5.9, one can see the shift in apparent target position, i.e. the position where the radar determines the target is, as a function of single pulse SNR. The boresight shift magnitude is similar for all three systems but the direction of the boresight shift changes from one configuration to another. This means that: 1) the magnitude of the boresight shift is a function of components common to all three systems, such as the monopulse comparator; and 2) the direction of the boresight shift as a function of the single pulse SNR is a peculiarity of the selected receiver configuration.

TABLE 5.1

AZIMUTH $\sigma_{\min}$ AND K OBTAINED BY THE LEAST-SQUARES
METHOD FOR THE THREE SYSTEMS.

System identification	$\sigma_{\min}$ (mrad)	K
2-channel TSMP	.285	16.8
2-channel TSMP with delay	.38	10.1
1-channel TSMP	.595	16.8

TABLE 5.2

ELEVATION $\sigma_{\min}$ AND K OBTAINED BY THE LEAST-SQUARES
METHOD FOR THE THREE SYSTEMS.

System identification	$\sigma_{\min}$ (mrad)	K
2-channel TSMP	.7	17.5
2-channel TSMP with delay	.52	23.8
1-channel TSMP	.45	25

5.4 Experimental comparison of the three monopulse receivers

As just stated, the boresight shift magnitude is similar, but the direction is not, for the three systems. In order to compare angle jitter for the three receivers, a plot of the azimuth and elevation angle jitter measurements as a function of PRF is given in Fig. 5.10. It can be seen that there is no substantial difference between the two-channel TSMP receiver (curves B) and the two-channel TSMP receiver with delay (curves C) for the PRF experimental range even though the experimental $\sigma_{\min}$ and K values are different.

But there is a major difference in Fig. 5.10 between the one-channel TSMP receiver (curves A) and the other two systems. The cause of this behavior is believed to be the non-coherent subtraction process used in the one-channel TSMP receiver. A subtraction is non-coherent when the $\log(\Sigma - \Delta)$ pulse subtracted from the $\log(\Sigma + \Delta)$ pulse does not come from the same received RF pulse. The two dual-channel TSMP receivers have coherent subtraction because they subtract video pulses which come from the same RF received pulse and therefore preserve monopulse normalization.

The non-coherent subtraction error is part of the minimum error and will show its effects more in the azimuth plane than in the elevation plane because the azimuth plane has more angular accuracy than the elevation plane as indicated by the minimum error levels in Fig. 5.10. Also, as shown in Figs. 5.1(a) and 5.2(a), the Σ and Δ power patterns vary faster in angle for the azimuth plane than for the elevation plane, so that non-coherent subtraction will introduce greater angle-tracking error in the azimuth plane. The plots of Fig. 5.10 show that the non-coherent subtraction error in the elevation plane cannot be measured since all three receivers have similar minimum error levels.

Fig. 5.10: Plots of the standard deviation of angle jitter for the three TSMP receivers and for both tracking planes.

In the azimuth plane, however, a difference in minimum error levels between curves B and C and curve A can be seen. This difference gives the non-coherent subtraction error value which is approximately .3 milliradians.

A non-coherent subtraction effect is observed because one cannot say that the E and Δ signals remain the same from one pulse to the next pulse even if the pulse radiated in the anechoic chamber does not have pulse-to-pulse variations. But it would be false to say that the antenna pulse-to-pulse movement is negligible since angle tracking is done in the null of the antenna difference pattern. Any slight movement of the null due to an antenna movement changes the pulse-to-pulse values of the difference signals which are produced by the monopulse comparator and therefore the monopulse ratio $2\Delta/E$ as derived in Eqn.(2-3) does not apply when there is non-coherent subtraction.

5.5 Summary

This chapter has presented and interpreted the experimental results which were obtained from radar tests. The minimum angle jitter level was found to be the major difference between the three TSMP receivers.

CHAPTER 6

CONCLUSIONS6.1 Summary and conclusions

In this thesis three monopulse receiver configurations have been investigated as alternatives to the conventional three-channel monopulse receiver. Each configuration uses the principle of time-sharing of receiver channels. One configuration has explored the effects of a delay at IF on the overall tracking performance. Time-sharing techniques save on the number of receiver channels necessary for tracking with regard to a conventional three-channel monopulse radar.

The experimental results (Figs. 5.3-5.8) have verified the theoretical prediction of thermal-noise induced angle jitter, namely that the induced angle jitter decrease (when the SNR is increased) is inversely proportional to the square root of the SNR. The variation in SNR was simulated by a variation in PRF with a constant single pulse SNR and the fitted curves (Figs. 5.3-5.8) show that the angle jitter decrease is inversely proportional to the square root of the PRF. Also it is important to note that the thermal-noise constant K of Eqn.(5-5) was found to be greater in elevation than it was in azimuth and this can be explained by a beamwidth wider in the elevation plane than in the azimuth plane.

The differences in boresight shift for the three receiver configurations show the influence of reducing duplicate components in parallel channels which introduce channel imbalances. On the other hand, the magnitude of the boresight shift is not significantly greater in one receiver compared to the others proving that the magnitude is more a function of the antenna monopulse comparator errors than the following circuitry.

The major difference between systems is not seen in the thermal noise error or boresight shift but in the minimum angle jitter level. In other words, the major difference comes from non-coherent subtraction versus coherent subtraction. It is concluded that monopulse normalisation cannot be preserved with non-coherent subtraction and for this reason, a receiver, such as a one-channel TSMP radar, using non-coherent subtraction will be subject to possible angle error effects from target amplitude fluctuation or scintillation as are conical scan radars and sequential lobing radars. It is not foreseen that the two dual-channel TSMP receivers investigated here will be susceptible to such amplitude scintillation because they employ coherent subtraction and therefore preserve monopulse normalization.

6.2 Future work areas

It is recommended that some future efforts be devoted to using the digital subtraction principle instead of the analog subtraction principle. The digital subtraction principle means that the $\log(I \pm \Delta)$ signals are converted from an analog to a digital form, then subtracted with a digital subtractor, and finally the obtained difference is converted from a digital to an analog signal in order to drive the antenna positioner. Those efforts should be compared to the analog subtraction results in the three receiver configurations. Also recommended is inserting time delay at RF rather than at IF. This should be compared with the results obtained for IF time delay. By introducing time delay at RF, only one receiver channel would be necessary.

Some work could also be devoted to looking at the effects of amplitude scintillation on the angle jitter performance of a one-channel TSMP receiver.

REFERENCES

1. R.S. Noblitt, "Reliability without redundancy from a monopulse receiver", *Microwaves*, pp. 56-60, December 1967.
2. D.K. Barton, *Radar System Analysis*. Englewood Cliffs, N.J.: Prentice-Hall, 1964.
3. D.K. Barton and H.R. Ward, *Handbook of Radar Measurements*. Englewood Cliffs, N.J.: Prentice-Hall, 1969.
4. R.E. Wilcox, "A time-shared monopulse approach to air/surface radar ranging", *EASCON REC IEEE Electronic and Aerospace Systems Convention*, pp 249-254, Sept 1978.
5. M.I. Skolnik, *Introduction to Radar Systems*. New York: McGraw-Hill, 1980.
6. M.I. Skolnik, *Radar Handbook*. New York: McGraw-Hill, 1970.
7. D.R. Rhodes, *Introduction to Monopulse*. New York: McGraw-Hill, 1959.
8. *Analog Switches and Their Applications*, Siliconix incorporated. Dallas, 1976.
9. R.B. Hirsch, "Plain talk on log amps", *Microwaves*, pp 40-44, March 1978.

END

05109184

FIN