NULL FILTER MOBILE RADAR (NFMRAD): CONCEPT VERIFICATION,

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The Null Filter Mobile Radar (NFMRAD) concept is described, and the need for narrow-band digital filters that can distinguish between positive and negative frequencies is explained. Two different procedures for realizing complex bandpass and bandstop filters of the infinite impulse response type are described, and their properties evaluated. A procedure for generating complex bandpass and bandstop filters of the finite impulse response type is also described, and the particular applicability of these filters to NFMRAD is discussed. The experimental NFMRAD system is described, and the field
Abstract (Continued)

installation and test program is explained. A ground based experiment using a truck traveling at about 30 mph (13.4 m/s) as the radar platform is described. Under these experimental conditions, the performance of NFMRAD is computed theoretically and compared to expected performance of systems using Chebyshev and uniform-fed antenna arrays with and without filters. Real-time processing tasks such as antenna pattern synthesis, Doppler filtering, and detection processing are described. Hardware performance and a functional description of the radar hardware is presented. NFMRAD transmitter, receiver, and digital processing specification are reviewed, and experimental system restrictions are made clear. The NFMRAD system is found to have its greatest superiority over other systems for cases in which the differences between the radial velocities of the target and the local clutter is not large.
Preface

The authors are indebted to the following individuals for their contributions during the course of this work:

Dr. William B. Goggins for original system concept and system design work. 1

Dr. August Golden for system design, software development, consultation, and continued project guidance.

Mr. Radames Gonzalez, Jr. for system fabrication and support during field testing.

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Mr. Richard Taylor for development of efficient real-time radar data processing software which made system realization possible, and for his continued support and interest.

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Concept Verification

1. INTRODUCTION

Pulse-Doppler radars may be used in a side-looking airborne configuration for the detection of slow-moving stand-off targets embedded in clutter (that is, foliage). Such a radar experiences interference from this background clutter. Clutter signals received through the mainbeam may be distinguished from a target of interest by appropriate Doppler filtering. This is possible, since the mainbeam clutter returns possess near-zero Doppler and can be effectively separated from a moving target signal possessing a Doppler frequency shift. However, clutter received through receive antenna sidelobes possesses motion relative to the moving radar platform. This sidelobe clutter can easily possess a Doppler equal to that of a moving target in the receive antenna mainbeam. Doppler filtering of this sidelobe clutter signal will not separate it from the target Doppler signal received through the mainbeam. Hence, with sufficient power, the sidelobe clutter signal can obscure the mainbeam target; thus, moving target detection is rendered difficult or impossible. The detection of a moving target will be made where no such target exists in the mainbeam.

This phenomenon results from the mechanism which generates a Doppler frequency and may be referenced to the forward motion of the radar platform by the following relation:

(Received for publication 3 October 1980)
\[ f_d = \frac{2V_o}{\lambda} \cos \theta \]

where \( \lambda \) is the wavelength of the radar frequency; \( V_o \) is the forward velocity of the airborne radar platform; \( \theta \) is measured from the line of forward motion of the radar platform, and \( f_d \) is the resultant Doppler frequency. Targets with motion parallel to the radar platform and with position in the receiver mainbeam would be at an angle of 90° and produce a zero Doppler frequency in their radar returns.

Targets illuminated by the mainbeam and producing nonzero Doppler frequencies are following courses nonparallel to that of the radar platform; hence they possess radial velocities with respect to the radar platform. A problem arises when ground clutter possesses a radial velocity with respect to the moving radar platform. This is true of ground clutter received through the receive antenna sidelobes.

To illustrate this matter, assume that the radar platform possesses a forward velocity of 180 miles per hour (mph) (80 meters per second in m s). A signal is being received in an antenna sidelobe from a large patch of ground clutter at a bearing of roughly 95° from the radar platform course heading. The return signal power would be large and detectable above the radar system's minimum discernible signal level. In addition, the sidelobe signal Doppler would be 14 m/s \( \lambda \). For a radar wavelength of 1.8 ft (0.6 m or 500 MHz), the Doppler frequency of the return would be 23.3 Hz, indicating a target moving away from the radar platform with a radial speed of 31.3 mph (14 m s).

The returned signal Doppler frequency may also be referenced to radial velocity from the moving radar platform with the following relation:

\[ f_d = \frac{2V_r}{\lambda} \]

where \( V_r \) is the apparent target radial velocity from the moving radar platform. A moving target in the mainbeam may possess the same radial velocity as side-lobe clutter:

\[ V_r = V_o \cos 80^\circ \]

Such a target, embedded in foliage, could be traveling away from the radar platform at 31.3 mph (14 m s). The course of the mainbeam target would be 80° from that of the radar platform, with a velocity of approximately 32 mph (14.2 m s) on its own course.
Examining the two cases illustrated above highlights the problem from the radar reception standpoint. With reception of clutter alone, an erroneous detection of a target may be made. With the reception of the discrete clutter at a bearing from the radar line of travel of 95°, and the simultaneous presence of the moving target in the receive antenna mainbeam, detection of the mainbeam target is rendered a difficult if not impossible task. Figure 1 illustrates the situation of clutter at 95° and the above target on a course 80° from that of the radar platform.

One possible answer to this sidelobe clutter problem is the use of a narrow-beam, low-sidelobe high-gain antenna. Unfortunately, this approach suffers from physical unmanageability. The radar operating frequencies that afford good foliage penetration performance lie well below 900 MHz. The antenna apertures required to realize a narrow mainbeam become prohibitively large at these optimum foliage penetration frequencies, and would result in an antenna size physically unmanageable in a side-looking airborne configuration.

One approach to the solution of the Doppler clutter problem is the use of antenna receive-pattern nulls placed or stirred such that the interfering clutter is attenuated. This technique could be used in conjunction with mainbeam target Doppler filtering to reveal the presence or absence of a tactical target in the mainbeam. It is for the investigation of this antenna null filtering technique that the Null Filter Mobile Radar (NFMRAD) project has been conducted.

NFMRAD is a proof-of-concept investigation. In any proof-of-concept program, economy of performance in the attainment of new information is important. To this end NFMRAD has been designed. Hardware design in both the areas of radio frequency and digital signal processing technology has been directed toward implementation of one antenna null and one Doppler filter with a similar design of processing software.

Rather than implementing NFMRAD in a costly airframe design, a truck-based moving platform has been used. This NFMRAD configuration is similar to that of a side-looking Airborne Moving Target Indicator (AMTI) on a moving platform. The moving system may be field tested against another moving target (truck), greatly simplifying field testing while addressing the objective of the most economic experimental investigation possible. Additionally, facility exists in the design for comparison of NFMRAD performance to that of conventional standoff, side-looking AMTI without the null filtering capability.

NFMRAD, as an experimental system, has been implemented in X-band. In keeping with constraints of simplicity and economy, the X-band implementation

---

Figure 1. Example of Doppler Clutter Interference
affords the use of readily available components within a compact experimental configuration. This again not only simplifies the design of the experimental rake platform but also eases field testing requirements.

Section 2 of this report covers in detail the theoretical aspects of the NFMRAWD concept. Sections 3 and 4 describe the hardware design, both radio-frequency and digital circuitry. Section 5 covers the design of supporting system software, while Section 6 deals with the experimental system field installation and testing. Results and conclusions are presented in Section 7 of the report.

2. THE NFMRAWD CONCEPT AND EXPERIMENTAL SYSTEM

2.1 The Null Filter Mobile Radar (NFMRAWD)

2.1.1 CONCEPT DESCRIPTION AND FILTER REQUIREMENTS

The NFMRAWD system operates from a moving platform and uses Doppler bandpass and bandstop filters together with null control in the receive antenna array pattern to reduce clutter return when detecting a target that is in motion relative to the ground. For a radar moving with velocity relative to the ground, there is a simple relationship between the clutter Doppler frequency shift \( \Delta f_d \) and the azimuth angle \( \theta \) given by the formula

\[
\Delta f_d = \frac{2v \cos \theta}{\lambda}
\]

where \( \lambda \) is the wavelength of the transmitted frequency and \( \theta \) is measured clockwise relative to the direction of motion of the NFMRAWD platform. By means of this relationship, all Doppler filter plots between the maximum and minimum clutter Doppler frequency shifts (\( 2v/\lambda \) and \(-2v/\lambda \)) can be exhibited as functions of azimuth angle \( \theta \), rather than clutter Doppler frequency shift. Throughout this report, we will assume that the clutter distribution is uniform in \( \theta \). The Doppler filter patterns in \( \theta \) thus represent the clutter distribution as modified by the filters.

The NFMRAWD system converts to baseband, using inphase and quadrature mixing. The received signal thus processed will contain only the Doppler frequency shifts, these will be positive or negative according to whether the objects producing the reflected signals are approaching or receding relative to the radar. Discrimination between these two cases is possible because both inphase and quadrature information is present. To quantitatively measure the frequency shifts and hence the relative velocities, bandpass filters must be designed that can measure both positive and negative frequencies and distinguish between them. At the same time, to discriminate against signals due to ground clutter, it is
In the operation of NEMRAD, many different filters and various different antenna array patterns need to be implemented, and the only practical way this can be accomplished is by digital processing. Two general types of these digital filters have been considered, one type has an infinite impulse response (IIR) and is realized by means of a recursive algorithm. The second type has a finite impulse response (FIR) and is realized by means of a non-recursive algorithm.

2.1.2 PRIMARY RECEIVED ANTENNA PATTERN

In the NEMRAD system concept, the spatial antenna sectors are the result of a set of fixed transmitting beam positions, and the Doppler frequency shift is determined by a set of bandpass filters. Figure 2(a) shows a typical beam sector pattern and Figure 2(b) shows a typical Doppler bandpass filter. Figure 2(c) shows a frequency transfer function both plotted with the same axis. For each transmitting beam position, there is also an associated bandstop-Doppler filter (filter than blanks the clutter spectrum from the beam of the transmitted pattern). Figure 2(d) shows the bandstop-Doppler frequency transfer function corresponding to the transmitted antenna pattern of Figure 2(a).

Assume there is a target located at the peak of the null by a certain antenna pattern of Figure 2(a) with a doppler velocity relative to the NEMRAD platform such that its return signal has a Doppler frequency shift that lies in the passband of the Doppler filter of Figure 2(d). Then for the transmit antenna pattern of Figure 2(a) and the clutter-reject frequency transfer functions of Figures 2(b) and 2(c), there exists a unique receive antenna array pattern that is optimum in that sense that it maximizes the ratio of the power of the received target signal to the total clutter power.

All optimum received patterns that appear in this report were computed from:


Figure 2. NFMRAD Antenna Patterns, Filter Power Transfer Functions, and Clutter Distributions
Figure 2. NEMRAD Antenna Patterns, Filter Power Transfer Functions, and Clutter Distributions (Cont.)
Figure 2. NEMRAD Antenna Patterns, Filter Power Transfer Functions, and Clutter Distributions (Cont.)

(c) Composite Doppler Filter Characteristic

CBR-6980, using modifications of a FORTRAN program devised by Goggin and Schnell,\textsuperscript{5} Figure 2(d) shows a typical optimum receive antenna array pattern for the present case. It is characteristic of this optimum pattern that it has a null region at the angular sector whose clutter Doppler return lies in the passband of the Doppler bandpass filter.

The power filter transfer function for the composite filter obtained by cascading the bandpass filter (Figure 2(h)) with the bandstop filter (Figure 2(e)), as shown in Figure 2(e). A plot of clutter power vs angle, as observed after illumination of the assumed uniformly distributed clutter by the transmit antenna pattern followed by filtering but before receive antenna processing, is shown in Figure 2(f). The product of this distribution of clutter power vs angle with the NEMRAD receive antenna pattern represents the distribution of clutter power that is ultimately present in the NEMRAD system, and is shown in Figure 2(g). The integral of the clutter distribution shown in Figure 2(g) is a measure of the total clutter power in the NEMRAD system.

Figure 2(h) shows the composite antenna pattern (product of the antenna transmit pattern with the NFMHAD receive pattern). Note that the composite filter transfer function, Figure 2(d), and the composite antenna pattern, Figure 2(h), are complementary in that the composite filter transfer function has a reject region in the angular range where the composite antenna pattern has a mainlobe, and vice versa.
(g) Clutter Power vs Azimuth After Filtering and NFMRAD Processing

(h) Two Way NFMRAD Antenna Pattern

Figure 2. NFMRAD Antenna Patterns, Filter Power Transfer Functions, and Clutter Distributions (Cont.)
Experimentla Radar System

A simplified block diagram of the experimental NFMIRAD system is shown in
Figure 3. A low-power X-band transmitter is shown in connection with an eight-
channel coherent receiver. An eight-element linear receive array feeds the RF
front end of the eight-channel coherent receiver. Imhase and quadrant signals
for each channel are generated at the output of the receiver. The receiver output
signals are converted to a digital format and transferred into an array processor.
The array processor forms the receive antenna pattern and filters the radar data.
Filtered data is transferred to the host minicomputer where it is stored and dis-
played. Detection processing takes place in the minicomputer, and detection out-
puts are also available for display on a CRT. Filtered data is stored on magnetic
tape for more sophisticated off-line detection processing, to be performed on a
CDC 6600.

The experimental radar system uses conventional AMTI processing as a basis
for performance comparison. Figure 4 depicts a block diagram of the total
NFMIRAD AMTI experimental radar system. The experimental system implemen-
ted incorporates two forms of radar processing, NFMIRAD and AMTI. The two radar
processing algorithms use common data as input; however, the NFMIRAD algorithm
differs from the AMTI algorithm in that the antenna patterns synthesized in pro-
cessing are different. AMTI processing forms an antenna pattern by uniformly
weighting each receive channel, while NFMIRAD antenna pattern synthesis weights
each channel so as to form a clutter notch in the receive pattern. Sidelobe clutter
falling outside the NFMIRAD antenna pattern clutter notch is not in the bandpass
of the NFMIRAD Doppler filter. Antenna beam-forming coefficients are a function
of the Doppler filter bandpass center frequency, therefore, to detect a wide range of
target velocities, several Doppler velocity filters are needed, with each filter
requiring a separate antenna pattern and clutter notch. The NFMIRAD and AMTI
Doppler velocity filters that follow the antenna pattern processing are identical.

The block diagram of the NFMIRAD system shown in Figure 3 depicts a low-
power monopulse transmitter operating at 9,410 GHz. The experimental transmit-
array pattern shown in Figure 16 indicates an azimuth beamwidth of 18°. The experi-
mental transmit elevation pattern of Figure 17 indicates a 3-dB elevation beamwidth
of 12°. The transmit horn shown in Figures 3(a) and 5(b) was placed in an anechoic
chamber for the pattern measurements.

The eight-element receive array shown in Figures 5(c) and 5(d) feeds the RF
front end of the eight-channel coherent receiver. A theoretical and experimental
single-element pattern is shown in Figure 20. The measurements were made in an
anechoic chamber, and the pattern is indicative of a waveguide opening into free
space.

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Figure 3. Simplified NFMRAD System

Figure 4. NFMRAD AMTI Experimental Radar System
(a) Transmit and Receive Array (Front View)

(b) Transmit Antenna (Three Views)

Figure 5. NFMRAD AMTI Transmit and Receive Antennas
(c) Eight-Element Receive Array (Side View)

(d) Eight-Element Receive Array (Top View)

Figure 1. NUMRAD/AMTI Transmit and Receive Antennas (Cont.)
I

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The experimental NMRAD and AMTI system implementation was discussed in Section 5.1 and 5.2. The Doppler filter characteristics are shown in Figure 6 and are discussed in Section 7.2. Filter design is discussed in Appendix A.

The function of the Doppler filter is to generate a frequency shift that is directly related to the NMRAD motion. The experimental radar system implemented a single NMRAD clutter filter for each corresponding Doppler filter. In addition, the NMRAD pattern and filter, a uniformly-weighted AMTI pattern is synthesized, and an identical Doppler filter is also implemented in the radar array processor. Following the coherent filtering intervals, the NMRAD and AMTI Doppler filters provide voltage magnitude per range cell.

Filtered NMRAD and AMTI data is transferred into a high-speed microcomputer for further processing, recording, and display. Detection processing is implemented in real-time for both NMRAD and AMTI data. Detection inputs are data or filtered voltage magnitude data may be displayed on the CRT. NMRAD or AMTI detections or voltage amplitude may be displayed as a function of range cell on the CRT. Filtered NMRAD and AMTI voltage magnitudes are recorded at magnetic tape in the format shown in Table 12 for off-line detection processing on a CDC 6600. Detection processing is also performed by the host mainframe computer. This processing compares each filtered voltage amplitude with a programmable threshold voltage. Detection processing results are not recorded.

Table 1 summarizes system characteristics, some of which were discussed above.

Table 1. NMRAD AMTI System Characteristics Summary

<table>
<thead>
<tr>
<th>Experimental Operation</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Limited broadside search (selectable search range)</td>
<td></td>
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<tr>
<td>Real-time antenna pattern synthesis, Doppler filtering, and detection processing achieved in radar van</td>
<td></td>
</tr>
<tr>
<td>Filtered data recorded for off-line additional detection processing on a CDC 6600</td>
<td></td>
</tr>
</tbody>
</table>

Operational Frequency

9.410 GHz
<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Antenna</td>
<td>25.4 x 25.4 in.</td>
</tr>
<tr>
<td>Power</td>
<td>18 W</td>
</tr>
<tr>
<td>Peak Power</td>
<td>22 W</td>
</tr>
<tr>
<td>Average Power</td>
<td>11 W</td>
</tr>
<tr>
<td>Receive Antenna (Figure 5)</td>
<td></td>
</tr>
<tr>
<td>Aperture Size</td>
<td></td>
</tr>
<tr>
<td>composite</td>
<td>3.875 x 7.375 in. (973 x 187.3 cm)</td>
</tr>
<tr>
<td>single element</td>
<td>0.375 x 7.375 in. (9.5 x 187.3 cm)</td>
</tr>
<tr>
<td>Polarization</td>
<td>Horizontal</td>
</tr>
<tr>
<td>Azimuth Beamwidth</td>
<td>25°</td>
</tr>
<tr>
<td>Elevation Beamwidth</td>
<td>12°</td>
</tr>
<tr>
<td>Receiver</td>
<td></td>
</tr>
<tr>
<td>Minimum Detectable Signal</td>
<td>-95 dB</td>
</tr>
<tr>
<td>Noise Figure</td>
<td>11 dB</td>
</tr>
<tr>
<td>RF Bandwidth</td>
<td>4 GHz</td>
</tr>
<tr>
<td>IF Bandwidth</td>
<td>18 MHz</td>
</tr>
<tr>
<td>Full Word Dynamic Range</td>
<td></td>
</tr>
<tr>
<td>Theoretical (no RF noise)</td>
<td>90 dB</td>
</tr>
<tr>
<td>Experimental</td>
<td>≈ 60 dB</td>
</tr>
</tbody>
</table>

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**Table 1. NFMRAD/AMTI System Characteristics Summary (Cont.)**

### A/D Conversion

<table>
<thead>
<tr>
<th>Number of Range Cells</th>
<th>16</th>
</tr>
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<tbody>
<tr>
<td>Sampling Rate</td>
<td>4 MHz (266 ns per range cell)</td>
</tr>
<tr>
<td>Interval per Range Cell</td>
<td>131 ft (40 m)</td>
</tr>
<tr>
<td>8-bit Conversion Rate</td>
<td>10 MHz</td>
</tr>
</tbody>
</table>

### RF Noise Level as a Function of Receiver Sensitivity

<table>
<thead>
<tr>
<th>IF Attenuation (dB)</th>
<th>Number of Mantissa Bits Consumed by RF Noise</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Average Value</td>
</tr>
<tr>
<td>0</td>
<td>4</td>
</tr>
<tr>
<td>6</td>
<td>3</td>
</tr>
<tr>
<td>12</td>
<td>2</td>
</tr>
<tr>
<td>18</td>
<td>1</td>
</tr>
<tr>
<td>24</td>
<td>0</td>
</tr>
<tr>
<td>30</td>
<td>0</td>
</tr>
<tr>
<td>36</td>
<td>0</td>
</tr>
<tr>
<td>42</td>
<td>0</td>
</tr>
</tbody>
</table>

Switching noise on the A/D boards consumes one mantissa bit, independent of IF attenuation. A combination of RF and switching noise yields the following:

<table>
<thead>
<tr>
<th>IF Attenuation (dB)</th>
<th>Number of Mantissa Bits Consumed by RF and Switching Noise</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Average Value</td>
</tr>
<tr>
<td>0</td>
<td>4</td>
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<tr>
<td>6</td>
<td>3</td>
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<td>12</td>
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<td>30</td>
<td>1</td>
</tr>
<tr>
<td>36</td>
<td>1</td>
</tr>
<tr>
<td>42</td>
<td>1</td>
</tr>
</tbody>
</table>

### Array Processor

- 6 asynchronous microprocessors
- 32-bit word
- 3 memory buses

Performs real-time beam formation and Doppler filtering for NFMRAD and AMTI processing

- 1 Doppler filter implemented
- 3 dB Doppler filter passband: -238.6 Hz to -401.5 Hz (diverging case)
- Angular sector of clutter in Doppler filter passband: 105° to 116° (diverging case)
Table 1. NFMRAD/AMTI System Characteristics Summary (Cont.)

<table>
<thead>
<tr>
<th>Minicomputer</th>
<th>Recorded Data Display (Table 12)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Memory</td>
<td>32 K interleaved core 4 K bipolar</td>
</tr>
<tr>
<td>Cycle Time</td>
<td>100 ns</td>
</tr>
<tr>
<td>Word Length</td>
<td>16 bits</td>
</tr>
<tr>
<td>Performs real-time detection processing, data storage and display functions</td>
<td>Filtered NFMRAD and AMTI data recorded voltage amplitude or detection output as a function of range cell</td>
</tr>
</tbody>
</table>

2.3 Theoretical Investigations Relative to the NFMRAD Experiment

2.3.1 THE NFMRAD EXPERIMENT

The following parameters and specifications are pertinent to the NFMRAD truck experiment:

Radar Van and Target Parameters

- Truck velocity = 32.7 mph (14.6 m/s)
- Target velocity = 34.4 mph (15.4 m/s)
- Target velocity relative to truck = 11.2 mph diverging (5.02 m/s)
- Target bearing relative to truck = 90° (constantly broadside)
- Target Doppler Frequency = -315 Hz (diverging case)
- Angle between truck track and target track = 19°
- Angle at which clutter return has target Doppler frequency = 110.12° (diverging case)

The layout of the tracks for the radar van and the target is shown in Figure 6. This is a constant-bearing diverging case and ideally results in constant relative velocity and therefore constant target Doppler. However, experimental errors in truck and target velocities and in relative bearing will be present. As the experiments are performed using just one Doppler filter, it is necessary that the filter passband be wide enough to accommodate the anticipated experimental errors.

A constant-bearing converging experiment is generated by reversing the directions of the target velocity and truck velocity. For this case, the target Doppler frequency changes sign to +315 Hz, and the Doppler filter is designed with its passband centered on this frequency.
2.3.2. OPTIMIZATION OF FIR FILTERS

In order to determine the range of FIR filter response for optimum system performance, various widths and magnitudes of passband and stopbands were considered. For each cascaded filter (that is, bandpass filter combined with a bandstop filter), an optimum receive pattern was determined and a figure of merit obtained for the performance of the system. For each system, the figure of merit was defined related to the ratio of the received power from the target to the total received clutter power. In order to compare systems using different transmit and receive patterns, as well as different filters, the figure of merit was scaled to be obtained from a system transmitting and receiving on a much lower band with no filtering. Thus the figure of merit reflects the loss in the maximum performance of any system over that of the optimum case and system described above.

The investigation showed that no filter characteristics have any impact on the figure of merit, only the ratio of the power present in the target to the sharpness of roll-off from bandpass to bandstop. A cascaded system with the filter passband the higher the more difficult to meet, the wider the required stopband system, reducing the width of the individual filters must also be increased above the filters required, and thus increases cost and complexity. For this reason, there are other considerations. There is no need to filter clutter if the system is to process through more than one model. Therefore the range of the filter must be large enough to encompass the entire sector of interest in the dynamics of the experiment as presented earlier. In the particular case chosen for the track experiment, the relative target velocity is small, and thus results in the clutter angular sector associated with the passband of the bandstop filter being close to the angular sector associated with the antenna mainbeam.
the cascaded filter does not have a rapid rolloff in the region between those sectors, the figure of merit will be reduced.

Generally, it was found that discriminations greater than about 40 dB for bandstop of bandpass filters did not improve the figure of merit. This is due in part to the noise produced on reception by the analog-to-digital converter because of truncation. Analysis of the experimental system to be used in the truck projects that if theoretical nulls of the order of -55 dB were desired in the optimum receive pattern, truncation in the A D converters would result in nulls only of magnitude -48 to -50 dB. The effect of the truncation is therefore equivalent to a noise source in the receiver system at a level 48 to 50 dB below the peak of the optimum (NFMRAD) pattern.

Taking into account the effect of the A D truncation noise on the optimum receive pattern, the figures of merit for many different cascaded pairs of FIR filters were computed and compared. Bandwidth considerations, based primarily on expected errors in relative velocity in the truck experiment, indicated that cascading a BP33-2-1 bandpass filter with a BS33-14-1 bandstop filter was the best compromise solution (that is, high figure of merit together with adequate bandwidth).

2.3.3 COMPARISON WITH OTHER SYSTEMS

NFMRAD was compared with systems using Chebyshev and uniform array transmit and receive patterns with and without filters. In all cases the antenna apertures used were equivalent to those as specified for the truck experiment. Comparisons were made for targets located broadside to the truck and moving at various radial velocities in the interval -11,2 mph to -32,7 mph (-5,6 m/s to -14,6 m/s). Figure 7 shows the improvement of NFMRAD over systems using uniform array transmit and receive patterns with and without filters. The cascaded FIR filters used for this analysis were composed of a BP33-3-2 bandpass filter and a BS33-14-1 bandstop filter. Figure 8 shows the improvement of NFMRAD over systems using -40 dB Chebyshev patterns for both transmit and receive with and without filters. The filters used here were a BP33-2-1 bandpass and a BS33-14-1 bandstop.

Without filters, the performance of both the uniform array system and the Chebyshev array system are far inferior to NFMRAD. Even with filters, the uniform array system performance is 35 dB or more below NFMRAD over most of the clutter frequency range that was considered. The Chebyshev array system performance with filters is at least 8 dB below NFMRAD except for clutter frequencies in a range corresponding to low radial velocities relative to the ground where NFMRAD improvement was 20 to 35 dB. For a given aperture size, the composite Chebyshev pattern (the product of the transmit and receive patterns) has a broader mainlobe than the composite NFMRAD pattern, and for targets with
Figure 7. Improvement of NFMRAD System Over a System Using Uniform Patterns on Both Transmit and Receive. (a) Without Filters, (b) With Filters. Filters were BP-33-3-2 and BS-33-14-1.

Figure 8. Improvement of NFMRAD System Over a System Using Chebyshev Array Patterns on Both Transmit and Receive. (a) Without Filters, (b) With Filters. Filters were BP-33-2-1 and BS-33-13-1.
radial velocities close to the radial velocity of the local ground. Difference in the
broadening deteriorates the performance considerably. For the Coherent echo
system with filters, the performance improves as the difference between the radial
velocity of the target and that of the local ground clutter decreases.

3. NFMRAD RF HARDWARE

NFMRAD is a "proof-of-concept" system, and the objective was the demonstra-
tion construction of a radar capable of performing the postulated function with the
least cost and the greatest economy. This economy involved simplicity in design at minimum 
cost in the addressing of the program objective. A unique and sophisticated 
eperimental tool is the product of this "economy" consideration.

NFMRAD was created in order to experimentally investigate and evaluate
the use of antenna pattern null filtering to reduce Doppler clutter. As a result,
NFMRAD performance capabilities are limited in extent. NFMRAD, as a proof-
of-concept design, is not a fully operational side-looking, stand-off, astatic
moving target indicator radar design based on the antenna null-filtering concept.

The principle performance characteristics of NFMRAD are highlighted here.
As a side-looking Airborne Moving Target Indicator (AMTI) radar, NFMRAD does not possess the performance capability of angular scanning (azimuth or elevation). NFMRAD responds to targets in its main beam and is capable of nulling Doppler
clutter received through one antenna side-lobe. This is possible since the NFMRAD possesses a multichannel receiver design. However, the antenna null is developed through digital signal processing of information from the eight receiver channels. The NFMRAD also performs conventional Doppler filter processing. However, here again only one Doppler filter of experimental interest is implemented, and, as with the antenna null formation, it is created through the digital processing of the eight-channel receiver information. These performance capabilities and limitations are a direct result of the intended purpose of the NFMRAD hardware design.

The NFMRAD may be subdivided into three major functional subsystems: the 
RF system, containing the transmitter and the receiver mixing preamplification stage; the Signal Conditioning section, consisting of the intermediate frequency amplifiers, the gain control, detectors, sample and hold circuits, analog-to-digital converters, output, timing, and control logic; and the Signal Processor, consisting of a high-speed array processor slaved to a host CYP-30 main computer. Figure 9 illustrates this simple functional subdivision of NFMRAD.

Discussion of the NFMRAD RF hardware design and operation will encompass
the RF section and portions of the Signal Conditioning section critical to the proper
functioning of the NFMRAD RF design. Remaining topics dealing with the digital
hardware will be developed in Section 4.
NAIRAD is a pulse-Doppler radar that uses a single frequency with a single transmit/receive array. The primary application for NAIRAD is short-range air defense. The transmitted signal is developed by a X-band oscillator and is amplified by a high-gain amplifier. The output of this amplifier is fed to the receiver array, which is designed to receive signals reflected from targets. The received signals are then processed by the receiver, which is used to develop a radar signal that is then fed to the processor to determine the magnitude and phase of the target signal vector. The processor then determines the rate of change of the phase of the target signal vector, which is then used by the processor.

The above receiver design approach makes use of signal processing techniques that allow for parallel processing of the received signals. This receiver configuration, typically used in a coherent receiver, allows for parallel processing of the signals from each channel. Each receive channel contains an active signal processor, which is used to process the signals from each channel. The signals are then combined to provide the processor with a broad bandwidth coverage.

Referring to Figure 10, the pulse-Doppler configuration is illustrated through the use of the receive pulse-shape filters. These filters are inserted before the transmitter/receiver array and after the transmitted signal has passed through the amplifier stage of the transmitter. The pulse-shape filter produces a pulse with a high-gain, rectangular shape. Figure 10 also illustrates the use of a high-gain, sectoral array. Note that these techniques help to optimize the performance of the radar system.
1. Introduction
Frequency Stability: 10,000 percent
Temperature Range: 32 to 140°F (0 to 60°C)
Harmonics: -50 dBc (minimum)
Sub-Harmonics: -52 dBc (-48 dBc minimum)
Sparrows: -50 dBc minimum
Output Impedance: 50 Ω

The secondary source (a 50-MHz crystal-controlled oscillator, Greenway, model number X-5111-60) specifications are as follows:

Frequency Stability: 10,000 percent
Temperature Range: 32 to 140°F (0 to 60°C)
Output Power: 300 mW (maximum)
Harmonics and Sub-Harmonics: -50 dBc (minimum)
Output Impedance: 50 Ω

The 5-GHz source passes through a 5-Watt power amplifer (No. 1, model number 301-60). Power from the 5-Watt port of the frequency multiplier is fed into the RF port of the eight-channel switch, to output a 5-GHz source. The RF input source is then selected, and is tuned to correct the output frequency. Power is passed through the eight-channel switch, to output a 5-GHz source. The RF input source is then selected, and is tuned to correct the output frequency.
The mixing products of the 9.35 GHz and 60 MHz are free to propagate from both waveguide ports of this device. The 60-MHz residual is attenuated by the cutoff characteristics of the waveguide. However, the sum and difference signals, as well as the original 9.35 GHz, are within the passband of the waveguide.

Power consisting of reflected 9.35-GHz signal and the various mixing product propagate down the waveguide in the direction of the -6 dB directional coupler output port. The directivity of this directional coupler is specified at 15 dB. This enables the coupling of mixing product power into the LO signal path to the receiver. This spurious signal power consists of the sum, difference, original 9.35 GHz, and the intermodulation products. The presence of these mixing products in the LO path to the receiver mixing preamplification stages would desensitize the receiver by raising the mixing preamplification noise floor; that is, effectively "jamming" the receiver front end. In order to avoid this difficulty, an isolator and the first pin diode modulator have been placed between the -6 dB directional coupler and the orthomode mixer. An understanding of the mechanism involved may be obtained through examination of two distinct cases: diode switch "on" condition and diode switch "off" condition.

First consider the diode switch "off" condition. During this condition the pin diode switch provides -45 dB of nondirectional attenuation. This effectively attenuates the 9.35 GHz power to maximum of 7.50 mW to a level of -16 dBm available at the orthomode mixer to a level below that required to effect mixing (2-mW minimum required signal power). Hence, these mixing products cannot be generated by the "turn-on" of the mixer by the 9.35-GHz signal power during this condition.

Given the possibility of sufficient signal current available for mixer turnover from the 60-MHz source, sufficient reverse signal attenuation exists to prevent receiver desensitization. The mixed signal would be at a level below the maximum power content measured in this case at 20 dBm. For sake of argument, choose the level at 20 dBm. This signal would traverse -45 dB of attenuation in the diode switch. -7.5 dB directional attenuation in an isolator preceding the switch, a total of -62.5 dB effectively available in the -6 dB directional coupler, as well as additional spurious power division (-2 dB channel). This would give original power with a signal attenuation of approximately -125 dB. If one refers to the receiver signal attenuation at the output of the diode switch, then a maximum signal level of 20 dBm was measured at the first mixing stage of each receiver.

The diode switch will be the remaining weakness of the receiver.

Power from the reflected 9.35-GHz signal at transmit antenna during first mixing stage of mixer is significant concern. This is true since little or no spurious power from adjacent channels actually exists at the output of the switch.
New constitute the circumstances during the first duration of 900 sec. During this period, the system operation, including input, has not yet started. The received signals are of no interest at this time. Hence, a very low signal level was received from end degradation, during which, initial biasing was made. It is important to note here that the +5 dBm level is selected for this test as an example, the power in the reflection path passes through an attenuation of 42 dB. It reduces any feedback signal to a maximum of +1.5 dBm signal in the output, prior to the first mixing stage of each receiver channel. Given the size of these power, this is signal level is insufficient to affect signal quality.

The received 9.3 GHz CW signal is mixed with the oto-6 MHz CW reference signal, difference, and 9.3-6 MHz pulsed signals. A second switch in the 9.3 GHz output protects any power reflected from the mixer from entering the receiver's front end, reflecting back to the adjacent mixer and receiver, which is a problem. The five-cycle butterworth response is a good filter to stop the lower harmonic signal at 41 GHz. The 6-MHz separation between these harmonics and the other intermediate channel frequencies is 2 MHz. The ceramic filter at 41 GHz removes the 9.3-6 MHz severely perturbs any mixing or oscillator content of the mixer output.

5.1.3 TRANSFER PULSE FORMATION

The desired transfer pulse characteristic is achieved using the two diode switches. One of these is excited before the other is excited, the other after the mixer's waveform switch, following the interstage power amplifier.

The first diode switch, the primary switch, biasing is selected for a low-level pulse, and also provides the feedback suppression control for triggering. The second switch serves two functions as well: it blocks the feedback suppression power generated by the TWTA during the interstage output, and also generates the RF pulse.

The first diode switch is a Hewlett-Packard model 3342A, and the second is a model 33222A. The model 3342A possesses SMA type RF ports and SMA type port, while the model 33222A possesses N-type RF ports with a BNC-type connector. Both switches are of compact physical design.

The diode switch is an absorption type and employs RF diodes sandwiched into the RF path. A large negative going pulse applied to the center control port is the control signal for switch turn on. In the absence of control signal bias, the switching devices provide a low-impedance path to ground. During this condition (shunting diode biased on by the RF signal), no more than +4.5 dB of incident RF power is available at the RF output port. In this state the switch is in the "off" condition. With the
negative going control pulse applied, the shunting diodes are biased off, and the resident RF power is available at the RF input port for the duration of the control pulse. Insertion loss at the control during the 'off' condition is specified to be less than 1 dB. Figure 11 illustrates the basic electrical structure of the pin diode switch.

![Diagram of a pin diode switch](image)

**Figure 11. Typical Microwave PIN Diode Switch**
(Adapted from Hewlett-Packard Application Note, Fast Microwave Switch-SpST Series 33140 and 33140A)

An RF pulse produced from the action of Model 33222A will possess rise and fall times no greater than 10 ns. Model 33142A will typically produce an RF pulse with a rise time of 3 ns and a fall time of 7 ns. Switch turn 'on' or 'off' is a function of control voltage threshold. Actual turn 'on' or 'off' times are a function of the switching characteristics of the diodes shunting the RF path. Once the negative going control pulse amplitude in fact falls below or rises above the negative control voltage threshold, the switch will turn on or off, according to the characteristic switching time of the switch diodes. Switch model 33142A is capable of absorbing 2 W of CW power in the off mode. Switch model 33222A, at the output of the TWTA, is capable of absorbing 2 W of CW power and sustaining a 75 W peak power at a maximum pulse width of one ms and 0.001 duty cycle.

Both switches are driven by a TTL compatible switch driver (device number DH0035G). The main band trigger is provided to integrated logic circuitry (ITL) which drives the switch driver of the first microwave diode switch. From the same main band trigger source, the main band trigger to the second diode switch is first supplied to an SN74123 (TTL retriggerable monostable multivibrator). This device delays and reshapes the main band trigger. The output of the SN74123 TTL device is also supplied to integrated logic circuitry (ITL) which drives the switch driver of the second microwave pin diode switch.
Referring to Figure 12, the first diode switch precedes the orthomode mixer and the second diode switch follows the TWT amplifier. As a result, a finite and significant delay exists in the propagation of the RF pulse from the first diode switch to the second diode switch. To compensate for this propagation delay, the firing of the second switch driver is delayed through the action of the SN74123 TTL device. The rising edge of the main bang trigger latches the first stage of the SN74123 on. External RC tuning circuitry is adjusted to yield a first-stage pulse time of 300 ns, equal to the propagation delay from first to the second switch. At the down clock of the first SN74123 stage, the second stage is clocked on for the predetermined period of 256 ns. The output of this stage is used to drive the second microwave pin diode switch driver. The microwave switch is turned on precisely the time the RF pulse generated by the first switch has risen to its maximum amplitude, and remains on for 206 ns. This action permits opening of the second switch when a stable RF pulse is incident at its input port and reshapes the wide RF pulse to desired pulse width.

The RF pulse generated by the first diode switch is wider than that desired for transmission. This is a result of the filtering action of the waveguide filter that eliminates the higher frequency components of the RF pulse. To correct this problem, the main bang trigger, as it originates from the Signal Processing Section—card nest board number 1 (System Timing 1), is 420 ns wide. As a result, the RF pulse generated by the first microwave diode switch is 320 ns wide with rise and fall times less than 10 ns. The RF signal bandwidth is theoretically 400 MHz wide at this stage of transmission pulse formation. However, the waveguide filter bandwidth is only 64 MHz wide. Consequently, the RF pulse incident at the input to the TWT amplifier is 320 ns wide, with rise and fall times of approximately 100 ns each. The 320 ns wide main bang trigger is delayed 50 ns by the 74123 device first stage, which in turn fires the 74123 device second stage for the desired pulse period of 256 ns. The firing of the second stage of this TTL device enables the capturing of the degraded RF pulse (when it is present at the second switch), and reshapes the RF pulse to the desired transmit pulse characteristics. The RF pulse, incident at the transmit antenna input, has a slightly rounded peak and is 206 ns wide, with a rise time of approximately 6 ns and fall time of approximately 8 ns.

This reshaping process theoretically produces a final transmit pulse with leading and trailing edge "porches." However, these porches are of duration less than 30 ns and are no greater than 0.6 mW in intensity. These porches were not detectable during system bench testing using a linear detector of 30 mV MW sensitivity. Figure 12 below illustrates the pulse firing sequence discussed above.

The rise and fall time of the switch driver output pulse is between 4 to 6 ns. This switching time is characteristic of the switch driver and independent of input pulse rise and fall time. In addition, the negative amplitude of the control pulse is
far larger than the voltage necessary for diode switch turn-on. This insures precise and reliable operation of the diode switches. Figure 13 is a schematic of the microwave diode switch firing circuitry. The quad-Nand gate and inverter (TTL devices) driving the switch driver add minor delays in the firing of the switch drivers. These minor delays have been compensated for in the design of the firing circuitry.

The second switch also attenuates noise generated by the TWTA during the interpulse period. The characteristics of the Hughes X1277H TWTA are such that 1 mW of white noise is continuously generated across its bandwidth, in the absence of input signal. With the second switch turned off during the interpulse period, white noise is reduced to a level of at least -45 dBm. This, combined with the -65 dB coupling figure from transmit to receive antennas, prevents subsequent jamming of the receiver during the interpulse period.
5.1.4 TRAVELING WAVE TUBE AMPLIFICATION

The filtered pulsed 9.41-GHZ signal must be amplified before transmission. Insertion losses through components between the orthomode mixer and the amplification stage yield a final peak signal power of 5 mW maximum incident at the amplifier input. The maximum safe permissible driving power for this TWTA amplifier (Hughes, model X1277H) is 0.6 mW. This may be achieved by adjusting the variable attenuator (adjustment located on the transmitter rack front panel) to a setting of -9.2 dB. With 0.6 mW of drive power, the TWTA amplifier output power is 22.5 W.

Continuously variable transmitter power is available in two power ranges. In the low range (0-5 mW), the TWTA is switched out of the circuit by two manual waveguide switches, one preceding and one following the TWTA. During low-range operation, the variable attenuator, preceding the TWTA, affords continuous adjustment of transmit power passing through the parallel waveguide transmission path. With the TWTA into the circuit, the same variable attenuator is used to continuously vary transmit output power from 1 mW to 22.5 W through variation of TWTA input power. Figure 14 is the measured gain characteristic of this particular TWTA. In order to protect the TWTA from reflected power, a -16 dB isolator has been installed between the amplifier output and the waveguide switch following the amplifier. This protects the TWTA during manual waveguide switch operation and during the second microwave pin diode switch off time.

![Figure 14. TWTA Gain Characteristic](image)

3.1.5 TRANSMIT ANTENNA

The transmit radiator is a simple sectorial horn situated directly above the receive horn array and polarized in the horizontal direction (H-plane vertical). The horn aperture has an E-plane dimension of 3.3 in. (8.3 cm) and an H-plane
dimension of 8.3 in. (21.2 cm). The H-plane taper length is 18.7 in. (47.5 cm), with an E-plane taper length of 19.5 in. (49.5 cm). These dimensions yield a theoretical gain of approximately 20.3 dB. Figure 15 illustrates the transmit horn.

![Transmitter Horn Diagram](image)

**Figure 15. Transmitter Horn**

Given the above horn aperture dimensions and uniform aperture excitation in the E-plane and cosine aperture excitation in the H-plane (TE$_{01}$ mode excitation of waveguide feeding a flared horn), the theoretical azimuth 3-dB beamwidth is approximately 19.3°, and the theoretical elevation 3-dB beamwidth is 10.3°. The theoretical azimuth mainlobe beamwidth is approximately 44°, and the theoretical elevation mainlobe beamwidth is approximately 25.8°. It is the azimuth directivity pattern which is of interest here, although the elevation directivity pattern determines the multipath characteristics of the system. Figure 16 compares the theoretical and experimental transmit directivity patterns, while Figure 17 illustrates the experimental elevation directivity pattern.

It is important to note that coupling between the transmit horn aperture and the aperture of any given receive antenna element aperture (located directly below the transmit horn) is approximately -65 dB. This coupling of transmitted power into the receive antenna element apertures results in the characteristic main band video at the start of range timing.
Figure 16. Theoretical and Experimental Transmit Horn Azimuth Power Patterns

Figure 17. Experimental Transmit Horn Elevation Power Pattern
3.1.3 Diagnostic Facilities

Two continuously variable output power ranges may be selected: 0-1 mW, or 1 mW to 22.5 W. In addition, facilities have been provided for the synthesis of a Doppler offset frequency. The injection of this synthetic Doppler signal into the transmitter mixing process enables efficient diagnostic testing of the receiver hardware and software. The Doppler frequency is that of a target moving away from the radar platform at a speed of approximately 11.5 mph (1.14 m/s) (or Doppler frequency of 325 Hz). The signal is created by the single sideband suppressed carrier mixing of 325 Hz with 60 MHz prior to the orthomode mixing of 60 MHz with 9.35 GHz. The 60 MHz and 325 Hz undergo phase cancellation mixing and the subsequent selection of the first lower sideband of this single sideband mixing process. Phase cancellation mixing is the only effective single sideband suppressed carrier mixing technique available. The frequency separation of the first upper and lower sidebands from the carrier (325 Hz) renders the design of the necessary sideband filter very difficult (if not impossible) for this application.

Figure 18 illustrates the Doppler target simulator. Note that the schematic illustrates the selection of the first upper sideband. Selection of the first lower sideband is accomplished through the interchange of the Gl. ports of the Hewlett-Packard Balanced mixers shown in the offset signal generator section of the Doppler generator schematic. A simple Wein Bridge audio oscillator and nullifier comprise the 325-Hz simulated Doppler signal source. The power supplied to the orthomode mixer is the same (0.3 dBm) whether it is a pure 60-MHz sinusoid or a sinusoidal 60-MHz plus Doppler sinusoid; thus, the signal level at the receiver is invariant to the selection of either mode.

Selection of the injected Doppler is accomplished by a switch provided on the front panel of the RF assembly (Transmitter Receiver Rack). With the radar platform stationary, the operator may select the injection of Doppler into the transmitter signal; thus, a Doppler target moving at the velocity of interest (-11.5 mph, -1.14 m/s) is synthesized. The signal may be directed at a convenient stationary radar target for operational check of the receiver hardware and receiver operating software, rather than attempting the simulation of this Doppler target through appropriate motion of either the radar platform and/or radar target. The operation of the receiver, with respect to range, may be checked through the appropriate selection of either the low or high-power ranges of the transmitter. This may be done with or without the synthesized Doppler target and the use of an appropriate stationary radar target.
3.1.7 SUMMARY OF TRANSMITTER SPECIFICATIONS

a. Frequency Sources

1. Primary Frequency Source (crystal controlled)
   Frequency: 9.35 GHz ± 0.0025 percent
   Power: 250 mW min to 1 W max, (typically 350 mW)
   Harmonics: -45 dBc (min), (typically -65 dBc)
   Spurious: -60 dBc (min)
   Supply: +28 V dc at 1A
   Unit is convection cooled

2. Secondary Frequency Source (crystal controlled)
   Frequency: 60 MHz ± 0.003 percent
   Power: 800 mW (min) (typically 950 mW)
   Harmonics: -20 dBc (min)

Figure 18. Doppler Target Simulator
Supply: ±24 V ±3 percent Regulation.
Unit as convection cooled.

3. Offset Doppler Test Source
Wein Bridge Audio Oscillator
Frequency: 32 kHz, less than 1 percent distortion
Amplitude: typically 0.8 V peak-to-peak
Supply: ±15 V dc

b. Pulse Characteristics (Pulse Generation, TTL Controlled)
1. Pulse Generated by First Microwave PIN Diode Switch
   Frequency: 9.35 GHz
   Prf: 1831 Hz
   Pulse Width: 320 ns
   Pulse Rise Time: 5 ns
   Pulse Fall Time: 7 ns

2. Pulse Characteristics after Filtering
   Frequency: 9.41 GHz
   Prf: 1831 Hz
   Pulse Width: 320 ns
   Pulse Rise Time: 100 ns
   Pulse Fall Time: 100 ns

3. Pulse Characteristics after Amplification and Action of Second
   Microwave PIN Diode Switch
   Frequency: 9.41 GHz
   Prf: 1831 Hz
   Pulse Width: 266 ns
   Pulse Rise Time: 6 ns
   Pulse Fall Time: 8 ns
   Leading and Trailing edge porches no greater than 30 ns in duration
   and at least -3 dB below pulse peak power.

c. Transmitter Emission Characteristics
1. Signal (Pulsed CW)
   Frequency: 9.41 GHz
   Prf: 1831 Hz
   Pulse Width: 266 ns
   Pulse Rise Time: 6 ns
   Pulse Fall Time: 8 ns
   Occupied Bandwidth: 60 MHz (approx.)
   Maximum Peak Power: 22.5 W
   Duty Cycle: 0.04 percent
   Spurious: -40 dBc minimum
1. Source Power-Spektrometer Recorder and Power Meter
   Source Power Spectrometer Recorder and Power Meter
   - Frequency: 0.4 GHz
   - Power: +3 dB below 40 MHz, source meter

2. Source Power-athedral Reference for L and Q Mixers
   Frequency: 40 MHz
   - Power: +3 dB below 40 MHz, source meter

3. Synthesized Doppler Target
   - Lower sideband emphasized for simulation, transmit power at -41 dBm, 1.044 m above the radar platform
   - Upper subband: -20.33 dB
   - Carrier: -20.4 dB

4. Schematic of XEMRAD Transmitter Subsystem - Figure 19

![Figure 19. XEMRAD Transmitter Subsystem](image-url)
Figure 20: Theoretical and Experimental Helical Duct Flow Pattern
Assuming uniform aperture excitation, the receive array produces an azimuth 3-dB beamwidth of approximately 19°. The calculated eight-element azimuth directivity pattern and the experimental pattern are illustrated in Figure 22. The eight-element directivity pattern was calculated using the principle of antenna pattern multiplication. The relationship for the single-element pattern was multiplied, point for point, with the equivalent value of a pattern produced by eight equally spaced point source radiators. The normalized relationship or array factor for eight equally spaced point sources is as follows:

\[ F = \frac{\sin(n \cdot 2\pi)}{\sin(2\pi)} \]

where

- \( n \) is the number of elements,
- \( d \) is the spacing (center to center of the receive elements),
- \( \delta \) is 0 for uniformly excited point sources,
- \( \theta \) is measured from the normal to the array face and

\[ d = \frac{2n \cdot d}{\lambda}, \quad \lambda = 3.188 \text{ cm} \]

with \( d = 4.27 \text{ cm} \), the element spacing (center to center of the receive elements). Each value of \( F \) achieved for a given \( n \) is multiplied for the same value of \( F \) achieved for the single-element pattern to yield \( F' \) or the composite amplitude pattern. The \( F' \) is squared and the log to base 10 taken for development of the eight-element azimuth directivity pattern. Note that all pattern calculations are normalized and referenced to 0 dB.

The above calculations yield a close approximation to the nature of the eight-element pattern. The correction of phase and amplitude errors, element by element, is referenced to the first channel receive element. This calibration was accomplished in the signal processing of the raw radar information with the target and radar platform stationarity. The calibration is performed on a daily basis, before and after field testing sessions (refer to Pre-Installation testing, software and field testing sections of this report).

Positioned in the horizontal direction (E-plane parallel to ground), the receive array is situated directly below the transmit horn. As a result of this configuration, energy from the transmit horn couples into each receive element at a level 6.0 dB below the peak transmit power. This results in the detection of main beam values at the beginning of same tuning.
3.2.2 RECEIVER FRONT END

Referring to the schematic in Figure 23, the received 9.41-GHz signal is mixed with the 9.35-GHz LO signal to produce a 60-MHz intermediate frequency (IF). The IF signal is then amplified. These two steps are accomplished in one unit (RHIG Model MDM 8-12 12A mixer preamplifier). The LO power is derived from the primary (9.35 GHz) transmitter signal source. Power from this primary source is coupled at a level 6 dB below its output power. This coupled power is once again divided, eight ways (through a Merrinax eight-way power divider, Model Number PDM 82-10C), and one each of the eight power divider outputs is supplied to the mixing stage of each of the eight receiver mixer preamplifiers. Typically, the LO power available to each of the receiver front end mixers is 11 mW or 10.4 dBm. The mixer preamp requires 3 to 18 dBm of LO power to effect dependable mixing, and an LO injection level in excess of 23 dBm will cause damage to the double balanced mixer. Intermodulation products of signal and LO mixing are suppressed a minimum of 20 dB below that of the 60-MHz IF signal.

The RF bandwidth of the mixer preamp is 4 GHz, centered at 10 GHz. The IF bandwidth is typically 16 MHz, centered at 60 MHz. The noise figure of the mixer preamp is typically 11 dB. Mixer preamp RF to IF gain is specified to be 20 dB minimum. With these specifications, the front-end signal sensitivity is
approximately -20 dBm per channel. This results in an approximately $1.4 \cdot 10^{-9}$ W/m^2 minimum incident signal power requirement per receiver channel for signal detection.

### 2.2.3 RECEIVER GAIN CONTROL

With the exception of channel 1, each IF signal passes through a variable attenuator (10-14 dB), Mercell Model ARM-10 that is mounted in the path of each channel. The signal in channel 1 is divided into two paths. One is sent to a log amplifier whose output is ultimately used to amplitude control the power of all channels to ensure linear operation over the complete range of expected linear signals. The other forms the input for a linear receiver channel that is channel 1. The variable attenuators in channels 2 through 8 enable balancing of these channels with respect to channel 1. Compensation of the element signals for dissimilar array elements is accomplished in the signal processing. This calibration procedure is accomplished daily to establish an antenna pattern measurement baseline, as well as to compensate for changes in receiver component operating characteristics.

The division of power in channel 1 enables the realization of an unorthodox gain control system. Unlike conventional AGC, the XMRAD stepped gain control provides a broad but linear dynamic range. This is achieved through a feed-forward control approach. The signal derived from the logarithmic amplifier in channel 1 is amplitude sampled and converted to a 4-bit digital code. The digital amplitude...
information is transformed in combinatorial logic into a 3-bit toggling code to control programmable attenuators in each of the eight linear receive channels.

Sufficient time for switching and settling of the attenuators is provided by delaying the IF signal 0.09 ms with a long RG-58 coaxial cable. With a delay of 1.6 as ft (5.27 m), 360 ft (109.7 m) of RG-58 provides the necessary 0.59 ms delay. However, a considerable insertion loss is associated with this 360 ft (109.7 m) coaxial delay line. To compensate for the loss, the signal is preamplified prior to insertion in the delay line.

In addition to toggling the switchable attenuators, the 3-bit switching code is processed and provided to the system's logic interface. This information carries real-time values of received signal level. In this fashion, the order of magnitude of the received signal amplitudes are preserved for later adjustment of the digitized linear channel outputs.

The amplification of the received signal for gain control is accomplished through a logarithmic amplifier. The signal is amplified and the logarithmic value of the signal level is sampled and converted to the 4-bit digital amplitude code. This enables the use of a 4-bit analog-to-digital converter for the first step in generation of the 3-bit attenuator switching commands. The use of linear amplification would require the use of an analog-to-digital converter of much greater than 4 bits of dynamic range. Unfortunately, the required conversion speeds constrain this portion of the design to A/D converters of 8 bits capacity or less. The choice of a 4-bit A/D converter also simplifies the design of conversion logic for the generation of attenuator switching commands as well as the digital amplitude information for later digital processing of the received signals.

Amplifiers following and preceding the switchable attenuators in each channel serve as isolation amplifiers. The attenuators, during switching, are mismatched to the 50-Ω delay lines. In order to isolate reflected spurious power during attenuator switching, a stage of IF amplification has been inserted prior to the attenuators. In addition, a stage of IF amplification following the attenuator isolates the mismatch from the input of the I and Q mixers in each channel. These amplifiers also compensate for component insertion losses throughout the receiver system, as well as provide the required system gain.

Without the influence of this gain control system, the dynamic range of the receiver (per channel basis) would be only 48 db. This is derived from the 8 bits of A/D conversion capacity for either the I or Q sides of the eight channels. Each toggling of a bit in the A/D converters corresponds to a doubling in signal level. Hence, 8 bits of conversion multiplied by 6 db of signal dynamic range per bit yields 48 db of dynamic range.

The amplifiers extend this dynamic range by 42 db for a total linear dynamic range of 90 db. The attenuation occurs at the rate of 6 db per step or 6 db per bit change on the switching command code. This operation is illustrated in Table 2.
Table 2. IF Attenuation as a Function of Base 2 Exponent

<table>
<thead>
<tr>
<th>Total Attenuation (dB)</th>
<th>Steps Set (dB)</th>
<th>Base 2 Exponent (IF attenuator switching code)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>none</td>
<td>000</td>
</tr>
<tr>
<td>6</td>
<td>6</td>
<td>001</td>
</tr>
<tr>
<td>12</td>
<td>12</td>
<td>010</td>
</tr>
<tr>
<td>18</td>
<td>12 &amp; 6</td>
<td>011</td>
</tr>
<tr>
<td>24</td>
<td>24</td>
<td>100</td>
</tr>
<tr>
<td>30</td>
<td>24 &amp; 6</td>
<td>101</td>
</tr>
<tr>
<td>36</td>
<td>24 &amp; 12</td>
<td>110</td>
</tr>
<tr>
<td>42</td>
<td>24 &amp; 12 &amp; 6</td>
<td>111</td>
</tr>
</tbody>
</table>

When the signal power present at the input to the mixer preamps approaches -54 dBm, the operation of the switchable attenuators is automatically initiated. At a level of -54 dBm of power incident on the input to the mixer preamps, the receiver will begin to saturate. It is important to note that the base-2 dynamic range assumes a noise level such that no ADC conversion bits are set by noise rather than the signal of interest. The total dynamic range is limited as opposed to logarithmic ( encountered with feedback control, AGC). This enables the signal processing that forms the XEMRAD and Doppler filters to be implemented.

3.2.4 IF BANDPASS CHARACTERISTICS AND DETECTION OF MOVING TARGETS

The three IF amplifiers serve various functions with regard to the receiver gain control system. In addition, their cascaded bandpass characteristics determine the overall IF bandpass characteristic. The composite IF bandpass is centered at 50 MHz with a 3-dB bandwidth of 5 MHz and a rolloff of 40 dB per decade.

The receiver IF bandwidth has been chosen to approximate the reciprocal of the pulse width (that is, 2 X 1.256 ns = 7.8 MHz, which in turn approximates the performance of a matched filter). This filter characteristic only crudely approximates a matched filter characteristic (that is, the IF bandwidth characteristic passes only those signal components within 4 MHz of the 50-MHz IF). Many of the signal frequency components are attenuated, resulting in pulse spreading. This would cause two consecutive pulses that are very close in time to overlap or become indistinguishable as two unique pulses. This does not pose a problem with respect to range determination, since the range of any given target is a function of the associated target pulse position within a given range bin. This dictates the range-
accuracy of the radar. This is opposed to continuous range determination referred to the leading edge of a received pulse. Hence, a need exists to preserve exact pulse characteristics to enable separation of pulses, since pulse position is time of importance only with respect to their occurrence in a given range bin. Consequently, the IF bandpass characteristic is sufficiently broad to preserve the gross nature of a received pulse and yet sufficiently narrow to band limit noise that is, the maximization of S/NR is tantamount to the preservation of exact pulse shape and the avoidance of interpulse overlap or interference.

Information carried by the received pulses (per channel) encompasses not only position information (with respect to any given range bin) but return signal instantaneous phase information as well. The division of radar range into range bins, and the timing of returned pulses with respect to these range bins, enables the extraction of target range information. However, phase discrimination of the returned signal, with respect to the transmit signal, is accomplished for the extraction of instantaneous target signal phase information. This latter phase information when compiled to form a time history of phase differences for a target of interest, constitutes information concerning the Doppler frequency of the target of interest and hence the target's radial velocity with respect to the XLRAD platform.

The IF signal undergoes quadrature mixing with the 60-MHz reference from the transmitter. This quadrature mixing yields signal magnitudes proportional to the rectangular components of the signal vector. One product of quadrature mixing is proportional to the magnitude of the real component of the signal vector, and the other product is proportional to the magnitude of the imaginary component of the same signal vector. In addition, the positive or negative nature of the real and imaginary parts of the signal are determined. Hence, the quadrature mixing of the IF signal and 60-MHz reference results in information regarding not only the magnitude of the received signal but knowledge of its complex component magnitude as well. Later processing of the component magnitudes, on a per pulse basis, yields the instantaneous phase difference with respect to the transmitted signal. Compilation of a time history of these phase differences enables measurement of the time rate of change of received signal phase. This is the definition of frequency and is the Doppler frequency of the target. The positiveness or negativeness of complex component magnitude yields information on the positiveness or negativeness of the phase difference. With the time history of positive or negative phase differences, the sign of the target Doppler may be determined with successively increasing negative phase differences indicating a target with relative velocity diverging from the radar, and a time history of successively increasing positive phase differences indicating an approaching target.

The data processing of the target signal vector components not only yields target Doppler information but also enables the appropriate processing of each channel.
output, carrying information on the same target, in order to form the receive antenna pattern null. This is covered in greater detail in Section 3.4.2 of this report.

The IF signal, after the attenuator post amplification stage, is power divided and inserted at the G-N ports of two Hewlett-Packard Double Balanced Mixers (Model 10334B). A 60-MHz reference signal from the transmitter secondary signal source is power divided through a Meritmic Quadrature Hybrid (Model QHS-20). This 60-MHz reference signal is typically of 500-mW power level, yielding approximately 50 mW, after eight-way power division, of reference signal power into each of the eight quadrature hybrids. In the quadrature hybrid, the 60-MHz reference signal is power divided. One portion is inserted directly into the R-G port of one of the model 10334B double balanced mixers. The other portion of the 60-MHz reference is phase delayed by 90° and provided to the R-G port of the remaining model 10334B double balanced mixer.

The portion of the IF signal mixed with the unshifted 60-MHz reference produces a signal in phase with the 60-MHz reference and with magnitude proportional to the magnitude of the inphase component of the signal vector. This is designated as the I component of the signal vector and is represented as the real component of the signal vector in phase with the reference. The portion of the IF signal mixed with the phase-delayed 60-MHz reference signal produces an output in quadrature with the 60-MHz reference and with magnitude proportional to the quadrature component of the signal vector. This signal is designated as the Q component of the signal vector and is represented as the imaginary component of the signal vector in quadrature phase relationship with respect to the reference signal. Figure 24 illustrates this I and Q signal relationship with respect to the 60-MHz reference vector.

Referring to Figure 24, a time history of consecutive I and Q values, indicating a change in \( \Phi \) such that each new \( \Phi \) value is successively greater than the last and changing in the positive direction, would indicate a positive target Doppler or target motion on a radial path away from the NFMRAD platform. The converse is true for \( \Phi \) changing in the negative or clockwise direction.

There is conversion loss associated with the mixing process. In addition, the only mixing products of interest are the sum and difference frequencies. The original input mixing frequencies and the intermodulation mixing products are of no interest. To overcome these difficulties, both I and Q sides of each channel possess amplification and filtering of the quadrature detection output. Filtering is accomplished by virtue of the frequency response characteristics of components following the quadrature mixers. The wideband amplifier (Hartel Model AM-1241B) following the mixers compensates for conversion loss in the mixing process. The gain of this amplifier is set to equal the conversion loss of the power division and
mixing process (typically 3.4 dB). The 3-dB bandwidth of this device is also established by the chosen gain setting. Gain and bandwidth characteristics of the device are coupled. By the selection of the wideband amplifier, but at 3.4 dB, the amplifiers 3-dB bandwidth is set at roughly 10 MHz. This not only compensates for conversion losses but also affords attenuation of all frequency mixing products above 10 MHz.

Further attenuation of unwanted high-frequency mixing products below 10 MHz is accomplished during the amplitude sampling of the combined I and Q signals. The response time of the sample and hold device (DARTEC High Speed Sample and Hold, Model SHM-12D) is such that the highest sinusoidal signal frequency that can be accurately quantized is 5 MHz (that is, a maximum sampling rate of 10 MHz; maximum required Nyquist rate to accurately sample a 5-MHz sinusoid). As a result, the 3-dB bandwidth of this device is 5 MHz; thus, additional attenuation of mixing products above 10 MHz is afforded, as well as attenuation of mixing products.
between 6 and 10 MHz. Any remaining products of mixing above 1 MHz are present as noise riding the sampled signal.

The 10-MHz sampling rate is sufficiently fast in order to sample a signal of 2-MHz bandwidth. The signal bandwidth results from a pulse train of roughly 2 KHz per pulse with each pulse carrying information of approximately 1-KHz bandwidth (1 KHz equals one-third the unfolded Doppler spectral bandwidth). Thus the bandwidth is sufficiently broad to preserve the approximate character of each pulse (e.g., 2.0 to 3.8 MHz). This later characteristic minimizes amplitude error generated from off-peak sampling of individual pulses.

The sample signal for both I and Q channels of each of the eight receive channels is converted to an 8-bit amplitude code on a pulse-to-pulse basis (Data A/D Converter Model VH88-12). The analog-to-digital converter possesses a maximum conversion rate of 10 MHz (5-MHz conversion bandwidth), again providing sufficient conversion speed for a signal of 2-MHz bandwidth. This digital amplitude code for the eight I and Q channels is sent in parallel to the Radar Data Buffer for eventual transfer into the array processor.

3.2.3 SUMMARY OF RECEIVER SPECIFICATIONS
a. Frequency: 9.41 GHz
b. Antenna
   1. Eight sectoral horns flared in the H-plane only, horizontally polarized, forming an eight-element receive array
   2. Array 3-dB Beamwidth
      Elevation (H-plane): 8°
      Azimuth (E-plane): 16°
   3. Gain
      Array: 20.0 dB (approx.)
      Element: 11.4 dB (approx.)
c. Front End (Per Channel Basis)
   1. LO Source: 9.34 GHz supplied from transmitter, typically 11-mW power level
   2. RF Bandwidth: 4 GHz (centered at 10 GHz)
   3. Maximum IF Signal Power: -23 dBm
   4. Minimum Discernible Signal: -90 dBm
   5. IF Bandwidth: 16 MHz
   6. RF to IF Gain: 20 dB
d. IF Section (Per Channel Basis)
   1. IF Bandwidth: 8 MHz centered at IF frequency with 40-dB per decade rolloff
   2. IF Frequency: 50 MHz
   3. IF Gain: 60 dB (approx.)
4. Receiver Digital Hardware

Receiver digital hardware is tasked with the analog-to-digital conversion of Land Q detector outputs. Digitally formatted data is temporarily stored in mantissa and exponent buffers, but ultimately is transferred to an array processor in which base formation and Doppler filtering takes place. Digital hardware utilized to convert the outputs of the Land Q receiver to a format appropriate for input to an array processor will be discussed in the following sections.

A functional description of receiver digital hardware will be presented to yield insight into the tasks required of each digital system. An input output approach will be used for explanation throughout the digital receiver hardware discussion, and a review of hardware design techniques is beyond the scope of this report.

Reference to Figure 26 shows that the receiver automatic gain control hardware establishes a base 2 exponent which is ultimately transferred to the array processor. Typically, the logarithm of channel 1 Hi signal establishes the signal attenuation.
Figure 26. Receiver Realization. The receiver consists of eight I and Q channels. Channel 1 is shown with the log amplifier voltage scaling circuitry. Each receiver channel uses common scaling circuitry; that is, Channel 1 IF is used to determine the base 2 exponent for all eight channels. Seven of the I and Q receiver channels are not shown.
The voltage scaling mechanism installed in the HI signal path in Figure 26, the HI voltage would be halved, and therefore the scaling factor must be \(2^4\). The case of 42 dB HI attenuation, the scaling factor is \(2^4\). With three control lines, the maximum base 2 exponent obtainable is 7, corresponding to 42 dB HI attenuation. Minimum HI attenuation of 0 dB corresponding to \(2^0\) or a scale factor of unity. Scaled \(2^0\) analog components are input to wide-band amplifiers seen in Figure 26, then sampled for analog-to-digital (A-D) conversion. Timing of sequential events and its analog sample and pulse transmission are under control of system time processor. A-D conversion is initiated upon sample and hold output stabilization. A D-convertor digital output plus 8-bit base 2 exponent are temporarily stored in quantizer and base 2 exponent buffers. When data buffers are filled, the Main Arithmetic Processor (MAP) array processor is informed of data availability via interface control logic. Under array processor control, the data is transferred from memory to temporary storage through data conditioning logic into MAP processor for beam formation and Doppler processing.

The functional hardware discussion in the next section reviews system timing, sample and hold circuits, analog-to-digital conversion, radar data buffer, digital conversion logic, and radar processor interface.

4.4 NAFRAD AMH System Timing

NAFRAD AMH system timing circuitry, shown in Figure 26, sequences periodic transmitter and receiver functions (for example, pulse transmission and analog receiver signal sampling). The system timing circuitry consists of clock oscillator, synchronous binary counters, and sense logic for detecting counter outputs.

The transient pulse command (MAIN RANG, SAMPLE, commands for the 4- and 8-bit analog-to-digital converters (A-D), the radar buffer memory, shift A-G (MAIN) commands, the AMH MAIL and DELETE command, and the radar buffer load (MAIN FIELD) command are derived from outputs of the binary counter string. A 1-MHz clock input is divided by \(2^{11}\), producing binary counter outputs \(Q_0, Q_1, \ldots, Q_{11}\) and \(Q_{12}, Q_{13}, \ldots, Q_{23}\); in addition, the counter \(Q_0\) 7-ms clock pulses 649 times. as in Figure 27.
The timing signals are derived from and locked to the radar 10-MHz clock. The 10-MHz clock is not synchronized with the 60-MHz clock. The 10-MHz clock is also run asynchronously with the radar 10-MHz clock to eliminate phase jitter. The 1-MHz clock is associated primarily with system timing and sampling events.

Typically, a timing command sequence would appear similar to Table 5. Timing signals referenced in Table 5 are defined in Table 4.
Table 4: Timing Control Sequence

Sixteen range cells are implemented in hardware; however, only three cells are shown in Table 4.
Table 1: Mode Button Settings

<table>
<thead>
<tr>
<th>Control Name</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>MAIN RANGE</td>
<td>Fine tune Analog</td>
</tr>
<tr>
<td>4-bit SAMPLE</td>
<td>Sample by pushing - Sample 4-bit button</td>
</tr>
<tr>
<td>2-bit SAMPLE</td>
<td>Sample was selected by pressing PQ buttons</td>
</tr>
<tr>
<td>CONVERT</td>
<td>Begin and end digital conversion</td>
</tr>
<tr>
<td>END OF CONVERT</td>
<td>End of analog/digital conversion</td>
</tr>
<tr>
<td>ATTENUATOR SET</td>
<td>Set IF signal attenuator</td>
</tr>
<tr>
<td>SHL F</td>
<td>Shift transmitted signal</td>
</tr>
<tr>
<td>SHL F</td>
<td>Shift transmitted and attenuated transmitted signals</td>
</tr>
</tbody>
</table>

1.2 Ultra High Speed Sample and Hold

Ultra high-speed sample and hold of the A/D system is accomplished by the SAR module which is capable of less than 20ns with proper signal conditioning. The SAR module utilizes a charge transfer scheme to achieve less than 50 nanoseconds. The 'SAMPLE' button initiates the mode and signals the SAR to start its operation. The analog signal is captured in this mode and the 'CONVERT' button initiates the A/D conversion process. The 'SAMPLE' button is pressed again to start a new conversion and the process repeats. Significant increased signals used in the A/D conversion are displayed on the reading diagram in Figure 28.

1.3 Ultra High Speed Analog to Digital Converter

Ultra high-speed 12-bit series 8-bit ADC's were utilized for conversion. The resultant output signal is a 12-bit digital signal of high fidelity. The A/D series 12-bit ADC's were capable of 12-bit conversion at 10 MHz. The SAR module CURRENT conditions sampled by the SAR module for the 12-bit A/D series is a sample rate of approximately 7 MHz. The 'SAMPLE' button initiates the A/D conversion, 'CONVERT' end conversion, and display of readings. An 'END OF CONVERT' button is presented for final display.
control. The ADC command is a TTL compatible 40 ns positive going pulse. 40 ns rising and falling edges are used to latch ADC output data in the buffer registers until SHIFT IN strobes output data into the Mantissa Buffer described in Section 4.4.1. Eight channels, one I and one Q component per channel, require 16 8-bit ABC's. Eight-bit SAMPLE command rising edge initiates synchronous NMI and ADC cycles for the A-D conversion, hardware shown in Figure 29.

1.4 Radar Data Buffer

Two 8-bit mantissas (I and Q components), in conjunction with a 5-bit exponent, define the amplitude and phase of one channel range sample. Each channel is divided into 16 range samples separated in space by 181 ft (54.9 m). The mantissa storage buffer is loaded under control of radar system timing circuitry discussed in Section 4.1. Following the lapse of a selectable range count timing interval (referenced to t0 of Section 4.1), system timing circuitry dictates the region where range samples will be taken. Sixteen sample and hold modules synchronously respond to the master timing 8-bit SAMPLE command. Following each 2-sampling period, a CONVERT command is issued to the A-D unit(s). At the conclusion of A-D conversion, eight data channels consisting of two 8-bit words per channel are ready for loading into the radar mantissa buffer. The 3-bit positive base 2 exponent is stored per range sample in base 2 exponent storage buffer. Data representing one range sample for each channel consists of 16 8-bit words stored in inverted offset binary format (defined in Section 4.4.1), plus one positive 3-bit base 2 exponent. Eight digital I components, plus eight digital Q components, are
synchronously loaded in parallel into the mantissa buffer under the control of the master timing circuitry. The base 2 exponent is loaded in parallel prior to the loading of the mantissa buffer (one 3-bit exponent per range bin). The mantissa and exponent buffers are implemented with TIL shift registers; therefore the first range bin loaded per channel will be the first range bin shifted out when unloaded into the array processor.

A diagram depicting the parallel loading of the radar data buffer is shown in Figure 29. The data load commands, hereafter referred to as SHIFT IN commands, are generated by the radar system timing. The buffers are completely filled following the 18th SHIFT IN command.

Data is multiplexed out of the mantissa buffer in a serial string of 2,048 3-bit words. The unload buffer commands or SHIFT OUT commands are under software control of the array processor (MAP-300) input output scroll (IOS) board. The IOS board is the input output interface used to pass radar data to the array processor memory. Data contained in the mantissa buffer is read out by component, channel, and finally range bin. The pecking order of the data multiplexer is depicted in Table 5. The multiplexer strobes 16 words from one range bin and recycles with a SHIFT OUT command to advance shift register data forward toward the multiplexer in Figure 29.

Table 5. Mantissa Buffer Data Multiplexer Pecking Order

<table>
<thead>
<tr>
<th>Position in Serial Word Train</th>
<th>Signal Component Description (8-Bits Wide at Output of Mantissa Buffer)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Channel 1, I Component, Range Bin 1</td>
</tr>
<tr>
<td>2</td>
<td>Channel 1, Q Component, Range Bin 1</td>
</tr>
<tr>
<td>3</td>
<td>Channel 2, I Component, Range Bin 1</td>
</tr>
<tr>
<td>4</td>
<td>Channel 2, Q Component, Range Bin 1</td>
</tr>
<tr>
<td></td>
<td></td>
</tr>
<tr>
<td>14</td>
<td>Channel 7, Q Component, Range Bin 1</td>
</tr>
<tr>
<td>15</td>
<td>Channel 8, I Component, Range Bin 1</td>
</tr>
<tr>
<td>16</td>
<td>Channel 8, Q Component, Range Bin 1</td>
</tr>
<tr>
<td>17</td>
<td>Channel 1, I Component, Range Bin 2</td>
</tr>
<tr>
<td></td>
<td></td>
</tr>
<tr>
<td>25</td>
<td>Channel 8, Q Component, Range Bin 16</td>
</tr>
</tbody>
</table>

66
Figure 24: Matrix and Exponent Shift Register Buffer with 8-Bit Wide Data Multiplexer. SHIFT IN commands originate from the radar master timing. SHIFT OUT commands originate from the array processor I/O interface.
The output data train is combined with a 3-bit base 2 exponent and directed through format conversion hardware to insure radar data format compatibility with that of the array processor. Details of the format conversion hardware are discussed in Section 4.5.

4.4.1 MANTISSA BUFFER

Radar digital data is stored in shift register memory during the interpulse period until the radar initiates an array processor input cycle. The shift register memory may be visualized as a data matrix built of 8-bit word elements. The eight-channel receiver outputs an inphase and quadrature analog component per channel. Sixteen analog range samples are taken at 267-ns intervals; therefore, the data matrix is organized by row with regard to range cells and by column with respect to channels. A data matrix is loaded into shift register memory following every transmitted pulse. An I and Q 8-bit word is loaded for 8 channels and 16 range cells, generating a matrix of 16 range cell rows and 8 complex channel columns (256 8-bit words). The mantissa buffer shown in Figure 30 obtains input data from Datel 8-bit A/D converters. The output from the data buffer is directed into an array processor via Format Conversion Logic.

<table>
<thead>
<tr>
<th>Mantissa Buffer</th>
<th>Column 1</th>
<th>Column 2</th>
<th>....</th>
<th>Column 8</th>
</tr>
</thead>
<tbody>
<tr>
<td>Column 1</td>
<td>Channel 1</td>
<td>Channel 2</td>
<td>....</td>
<td>Channel 8</td>
</tr>
<tr>
<td>Inphase</td>
<td>Inphase</td>
<td>Inphase</td>
<td>....</td>
<td>Inphase</td>
</tr>
<tr>
<td>Quadrature</td>
<td>Quadrature</td>
<td>Quadrature</td>
<td>....</td>
<td>Quadrature</td>
</tr>
<tr>
<td>ROW 1 RANGE BIN 1</td>
<td>8 bit</td>
<td>8 bit</td>
<td>8 bit</td>
<td>8 bit</td>
</tr>
<tr>
<td>ROW 2 RANGE BIN 2</td>
<td>8 bit</td>
<td>8 bit</td>
<td>8 bit</td>
<td>8 bit</td>
</tr>
<tr>
<td>ROW 16 RANGE BIN 16</td>
<td>8 bit</td>
<td>8 bit</td>
<td>8 bit</td>
<td>8 bit</td>
</tr>
</tbody>
</table>

Figure 30. Radar Receiver Mantissa Buffer (256 8-Bit Words)
4.4.2 BASE 2 EXPONENT BUFFER

Each of 16 range bins is associated with a positive 11-bit exponent that characterizes the effective gain of a channel. The 11-bit exponent may vary from range bin to range bin if the steerable attenuators are in the automatic mode of operation. If the step attenuators are operated in the manual mode, the 11-bit exponent will stay constant for 16 range bins. The exponent is stored in a shift register array 2 bits wide and 16 range bins deep. The array input comes from conversion logic following the 4-bit A/D converter sampling the log of channel 1 IF when in automatic operation. The logarithm of the channel 1 IF signal is provided by an analog logarithmic amplifier. Channel 1 IF signal is split using an isolator or power splitter. Half of channel 1 IF signal power is used as input to the logarithmic amplifier. The log amplifier output ultimately determines the IF signal attenuation required for eight channels. The remainder of channel 1 IF signal power is time delayed by utilizing coax cable. Channels 2 through 8 IF signal power is attenuated by 3 dB and delayed in a fashion identical to channel 1. The log amplifier output is sampled and converted to a 4-bit word. The 4-bit A/D output is converted to a 5-bit IF attenuator control word, which establishes the IF signal attenuation required in all eight channels and the base 2 exponent as a function of range bin. Establishing the necessary IF attenuation requires time for A/D conversion, 3-bit conversion, and attenuator setting and stabilization. Channels 1 through 8 are delayed for the time required to establish the appropriate IF signal attenuation. While in manual operation, the 4-bit A/D output is bypassed, and the 3-bit exponent is hardwired in and loaded following each 4-bit SAMPLE command.

4.5 Data Format Conversion Logic

Data format conversion hardware transforms the 8-bit A/D output and 11-bit base 2 exponent into a normalized 15-bit floating point data word. The 15-bit floating point format is compatible with the input requirements of the CASMAP array processor. Format conversion logic follows the output of the analog IF buffer multiplexer in Figure 26. Radar data is transferred from the unattenuated exponent buffers through data conditioning (format conversion) logic and into the array processor. Data format conversion could have been implemented in array processor software; however, due to the process speed requirements of our real-time operating environment, format conversion was necessarily implemented in hardware.
4. 4.1 ANALOG-TO-DIGITAL CONVERTER INPUT AND
   OUTPUT SPECIFICATIONS

a. Input voltage range is ±10 V for the input analog-to-digital converter

   -1.25 V to 1.25 V.

   Digital output coding for Datel model ADC-8H
   series is referenced as inverted offset binary, as defined in Table 8.

b. From Table 8, A-D input sensitivity is 30 mV per bit.

c. The A-D converters used perform an 8-bit conversion in 0.1 ms (10-MHz
   conversion rate).

Table 8. Datel ADC Output Format

<table>
<thead>
<tr>
<th>Analog Input Voltage</th>
<th>Output Bit Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>1111111111111111</td>
</tr>
<tr>
<td>-0.001</td>
<td>1100000000000000</td>
</tr>
<tr>
<td>0.25</td>
<td>0000000000000000</td>
</tr>
<tr>
<td>-1.25</td>
<td>1111110000000000</td>
</tr>
<tr>
<td>-1.25</td>
<td>1111100000000000</td>
</tr>
</tbody>
</table>

4. 4.2 8-BIT INVERTED OFFSET BINARY FORMAT CONVERSION
   TO 8-BIT SIGN MAGNITUDE FORMAT

The first stage of format conversion following the ADC output is dedicated to
transferring the ADC output of sign magnitude format. If bit 7 is sensed to be
set, the 8-bit data word is transferred directly through the first section of digital
logic without modification. However, if bit 7 is not set, the conversion logic per-
forms a 2's complement operation on bits 0 through 6, with one exception. If
bits 4 through 7 are not set, bit 7 remains unchanged while bits 0 through 6 are
set.

Table 7 defines logical input and output states for the first stage of format
conversion circuits. Equation (1) relates the A-D input voltage to sign magnitude
binary components m₀, m₁, ..., m₆ before combination with Base 2 exponent.
### Table 7. Sign Magnitude Format

<table>
<thead>
<tr>
<th>Analog Input Voltage (Volts)</th>
<th>Format Conversion Input</th>
<th>Format Conversion First Stage Output</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Inverted Offset Binary bit position</td>
<td>Sign Magnitude bit position</td>
</tr>
<tr>
<td></td>
<td>7 6 5 4 3 2 1 0</td>
<td>7 6 5 4 3 2 1 0</td>
</tr>
<tr>
<td>M</td>
<td>1</td>
<td>S M</td>
</tr>
<tr>
<td>S</td>
<td>8</td>
<td>I S</td>
</tr>
<tr>
<td>B</td>
<td>B</td>
<td>G R</td>
</tr>
<tr>
<td></td>
<td>d_{7} d_{6} d_{5} d_{4} d_{3} d_{2} d_{1} d_{0}</td>
<td>m_{7} m_{6} m_{5} m_{4} m_{3} m_{2} m_{1} m_{0}</td>
</tr>
<tr>
<td>-1.270</td>
<td>1 1 1 1 1 1 1 1</td>
<td>1 1 1 1 1 1 1 1</td>
</tr>
<tr>
<td>-0.040</td>
<td>1 1 0 0 0 0 0 0</td>
<td>1 1 0 0 0 0 0 0</td>
</tr>
<tr>
<td>-0.010</td>
<td>1 0 0 0 0 0 0 1</td>
<td>1 0 0 0 0 0 0 1</td>
</tr>
<tr>
<td>0.000</td>
<td>1 0 0 0 0 0 0 0</td>
<td>1 0 0 0 0 0 0 0</td>
</tr>
<tr>
<td>+0.010</td>
<td>0 1 1 1 1 1 1 1</td>
<td>0 0 0 0 0 0 0 1</td>
</tr>
<tr>
<td>+0.040</td>
<td>0 1 0 0 0 0 0 0</td>
<td>0 1 0 0 0 0 0 0</td>
</tr>
<tr>
<td>+1.270</td>
<td>0 0 0 0 0 0 0 1</td>
<td>0 1 1 1 1 1 1 1</td>
</tr>
<tr>
<td>+1.280</td>
<td>0 0 0 0 0 0 0 0</td>
<td>0 1 1 1 1 1 1 1</td>
</tr>
</tbody>
</table>

\[ V_{\text{smag}} = (-1)^{m_7} \left(10 \text{ mV}\right) \sum_{i=0}^{6} 2^i m_i \]  

(1)

where

- \( V_{\text{smag}} \) is the input voltage to 8-bit A/D
- \( m_6, m_5, \ldots, m_0 \) are binary data bits
- \( m_7 \) is the sign bit.

#### 4.5.3 8-BIT SIGN/MAGNITUDE MANTISSA WITH POSITIVE 3-BIT BASE 2 EXPONENT CONVERSION TO 16-BIT FLOATING POINT FORMAT WITH BASE 16 EXPONENT

The 12-bit radar data word is defined as shown in Table 8 following the combination of sign/magnitude format and positive base 2 exponent. The A/D input voltage expressed (as seen at output of sign/magnitude format conversion logic after combination with base 2 exponent) in terms defined by Table 3 is seen by Eq. (2).
The data word of Table 3, following conversion to a 10-bit floating-point word, is defined as shown in Table 3 and Eq. (6). Equation (6) yields the 10-bit ADC input voltage as seen by the MAP.

\[ V_{\text{MAP}} = (10)^{N_p} \sum_{k=1}^{N_p} \frac{V_k}{10^k} \]  

\[ V_{\text{MAP}} = \sum_{i=1}^{N} S_i 2^{i-1} \]  

\[ V_{\text{MAP}} = \sum_{i=1}^{N} (S_i 2^{i-1} + S_{i+1} 2^i) \]  

\[ V_{\text{MAP}} = \sum_{i=1}^{N} (S_i 2^{i-1} + S_{i+1} 2^i) \]  

where:

- \( S_i \): Sign bit of mantissa
- \( S_p \): Sign bit at base 2 exponent follows as positive or \( S_p = 0 \)
- \( b_0 \) to \( b_{N-1} \): Binary mantissa bits
- \( v_2, v_1, v_0 \): exponent bits

Bit positions are defined in Table 3.
where
\[ \exp \text{ is derived from } S_e, e_2, e_1, e_0. \]
\[ S_e \text{ is the sign of the base 16 exponent.} \]

If \( \exp = 0, (S_e = 0) \) then
\[ \exp = 4e_2 - 2e_1 + e_0. \]

If \( \exp = 0, (S_e = 1) \) then \( e_2, e_1, \) and \( e_0 \) are in 2's complement form. Perform 2's complement operation on \( e_2, e_1, \) and \( e_0 \). Yielding \( e'_2, e'_1, \) and \( e'_0 \) then
\[ \exp = -4(e'_2 - 2e'_1 + e'_0). \]

H1, H2: Hex digits converted to base 16,
H3: Hex digit with LSB deleted converted to base 16,

Bit position for arguments of Eq. (3) are defined in Table 9.

### Table 9. MAP Base 16 Floating Point Half-Word Format

<table>
<thead>
<tr>
<th>Bit Position</th>
<th>15</th>
<th>14</th>
<th>13</th>
<th>12</th>
<th>11</th>
<th>10</th>
<th>9</th>
<th>8</th>
<th>7</th>
<th>6</th>
<th>5</th>
<th>4</th>
<th>3</th>
<th>2</th>
<th>1</th>
<th>0</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>M</td>
<td>S</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>N</td>
<td>B</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
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<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Bit Position</th>
<th>( h_{11} )</th>
<th>( h_{10} )</th>
<th>( h_9 )</th>
<th>( h_8 )</th>
<th>( h_7 )</th>
<th>( h_6 )</th>
<th>( h_5 )</th>
<th>( h_4 )</th>
<th>( h_3 )</th>
<th>( h_2 )</th>
<th>( h_1 )</th>
<th>( h_0 )</th>
<th>( c_3 )</th>
<th>( c_2 )</th>
<th>( c_1 )</th>
<th>( c_0 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>S_e</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
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<td></td>
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<td></td>
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<td></td>
</tr>
<tr>
<td>H1</td>
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<td></td>
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<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
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<td></td>
<td></td>
</tr>
<tr>
<td>H2</td>
<td></td>
<td></td>
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<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>H3</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>S_e</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
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<td></td>
<td></td>
</tr>
<tr>
<td>Exponent</td>
<td></td>
<td></td>
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<td></td>
<td></td>
<td></td>
<td></td>
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<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Hex Mantissa | Base 16 Exponent

Equation (3) relates the voltage observed by the MAP to the key components identified by Table 9.

Equation (2) is not equivalent to Eq. (3). The MAP-defined voltage, \( V_{MAP} \), differs from \( V_{MNMN} \) by a constant multiplier.

\[
V_{MNMN} = (1024/2^7)V_{MAP}
\]
\[ v_{\text{smag}} = (10 \text{ mV})(2^7)(-1)S_h 10^\text{exp} \sum_{k=1}^{3} \frac{H_k}{10^k}. \]

Equation (4) allows one to predict the A/D output utilizing Table 9, MAP hex voltage representation, and the IF attenuator settings \( x_2, \) \( S_1, \) and \( x_0. \) Equation (4) allows the conversion of MAP base 16 data to A/D input voltage.

\[ v_{\text{smag}} = (10 \text{ mV})(2^7) \frac{(10 \text{ mV})(10^\text{exp})}{4x_2^2x_1x_0} \sum_{k=1}^{3} \frac{H_k}{10^k}. \]  

Equation (4) is useful when relating A/D voltage input observations to MAP voltage observations used in processing. Equation (4) may also be used to ensure A/D linear operation by insuring that the input voltage specifications are never exceeded.

4.5.4 NORMALIZATION OF 16-BIT FLOATING POINT WORD

The 16-bit data word is normalized in hardware to insure \( H_1 \neq 0 \) (unless \( H_1, H_2, \) and \( H_3 = 0). \) If hex shifts are required for normalization, the base 16 exponent is decremented accordingly.

4.6 INPUT/OUTPUT SCROLL (IOS) INTERFACE

Discussion of the IOS interface will be developed primarily around data and control interface hardware. An event timing diagram is presented with references to RADAR IOS handshake control signals. Interface test hardware is presented in addition to detailed discussion regarding real-time data status indication. Data transfer rates realized are presented, and the data double buffering input scheme implemented is discussed. Radar data is transferred from the radar data storage buffers through format conversion logic and into an array processor under control of the array processor interface board (IOS).

4.6.1 16-BIT NORMALIZED FLOATING POINT WORD TRANSFER FROM RADAR TO IOS INTERFACE

Real-time radar operational constraints require all signal processing and data transfers be completed within one interpulse interval. Data transfer rates must be high to insure maximum signal processing rates. The radar IOS interface and conversion logic realize a 16-bit MAP word transfer in approximately 1.30 ms; 120 ns is needed for 32-bit MAP full word transfer (8 ms to transfer 16 MAP full words, \( \approx \) 546 ns per interpulse period).
4.6.2 RADAR MAP-300 LOS INTERFACE CONTROL AND TIMING SIGNALS (Figure 31, Table 10)

The array processor (LOS interface) informs the radar that the LOS processor has started via signal LOSIN. Following activation of LOSIN, the LOS interface informs the radar that processing is complete and the LOS is waiting for data via PAUSE control signal. The radar responds to the processor data request following the filling of the radar mantissa and exponent buffer discussed in Section 4.4.

When the mantissa and exponent buffers are filled with new data, the radar informs the processor that data is ready by issuing the interface control signal CLRPAUSE. The PAUSE signal and CLRPAUSE signals insure that the data transfers are synchronized, and that data is transferred only when data is available, that is, the array processor must complete all processing and data transfers within one interpulse period, and must wait until radar data is ready. The LOS then requests data, utilizing the leading edge of control signal LOSIN. Data input request is acknowledged by radar, utilizing control signal LOSACK, and the data is transferred (in handshake fashion) on the trailing edge of LOSIN. A total of 128 complex word transfers takes place in 144 usec, then processing is initiated. Three control lines were used to monitor radar data integrity and insure that the MAP did not drop a pulse by missing a data matrix, or transfer data as the radar data buffer was being updated.

![Figure 31. Radar LOS Interface Control and Timing Signals](image)

*Figure 31. Radar LOS Interface Control and Timing Signals. Only the most significant control and timing signals are shown. Active signal levels shown here do not accurately reflect interface. Positive logic was used for discussion purposes.*
Table 10. Radar IOS Interface Control and Timing Signals
Active interface logic levels are not necessarily as shown

<table>
<thead>
<tr>
<th>Edge Number</th>
<th>Name</th>
<th>Classification</th>
<th>Message</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>IOSRUN</td>
<td>MAP IOS interface control</td>
<td>Scroll is running</td>
</tr>
<tr>
<td>2</td>
<td>PAUSE</td>
<td>MAP IOS interface control</td>
<td>MAP waiting for data</td>
</tr>
<tr>
<td>3</td>
<td>RABUFFED</td>
<td>Radar tuning</td>
<td>Radar mantissa and exponent buffers start load cycle.</td>
</tr>
<tr>
<td>4</td>
<td>CLRPAUSE</td>
<td>Radar interface control</td>
<td>Falling edge of RABUFFED indicates buffer load complete. This event triggers CLRPAUSE implying new data available.</td>
</tr>
<tr>
<td>5</td>
<td>IOSIN</td>
<td>MAP IOS interface control</td>
<td>Leading edge of IOSIN requests data transfer from radar.</td>
</tr>
<tr>
<td>6</td>
<td>IOSACK</td>
<td>Radar interface control</td>
<td>Leading edge of IOSACK acknowledges data request causing data transfer on trailing edge of IOSIN.</td>
</tr>
<tr>
<td>7</td>
<td>IOSRUN</td>
<td>MAP IOS interface control</td>
<td>Radar to MAP data transfer complete, scroll is halted.</td>
</tr>
</tbody>
</table>

4.6.3 DATA STATUS INDICATOR

Three data status lines are provided under radar control to indicate data integrity. XEMRAD Doppler processing requires a constant time lapse from data group to data group; therefore, the loss of one pulse data group during a coherent processing interval is intolerable. If the array processor requires more than one interpulse period for beam formation and Doppler filter processing, invalid data will be transferred into the MAP. The radar is tasked with monitoring data integrity and reporting data status to MAP. Radar interface logic monitors transfer control signals and internal timing commands to determine data status. If the IOS requires more than 540 μs between data requests, the radar hardware writes over data stored in the radar buffer.

New radar data may become available at the ADC output as the MAP transfers data under IOS control. In this situation, new data would be shifted in the mantissa.
1.7 Receiver Simulator Hardware

Receiver simulator hardware was developed for the purposes of unit testing, interface verification, and software diagnostics. Unprocessed electrical and mechanical data are removed from the test data mount. Stimulated faults, such as simultaneous recording of data at the interface port, are introduced via the interface board 22 with the receiver simulator card (Figure 3A). Another view of Figure 32.

The receiver simulator card and associated circuitry mount to the interface board. Typically, the data matrix may be characterized by the two ranges, the two ranges of cell voltage, and the range of cell categories. Every range cell voltage magnitude is sampled, and the range of voltage magnitude is calculated. Each cell voltage magnitude is determined by the cell voltage magnitude limits. The cell voltage magnitude limits are compared, and the voltage magnitude limits are set for each cell range. The voltage magnitude limits are set for each cell range in the range of cell voltage magnitude limits.
The table below illustrates the single, relatively simple, circuit-by-circuit, and channel-by-channel approach of the II. The columns 1-12 represent the different channels of the II. The rows represent different single-channel circuit descriptions. Each row lists the attributes of a specific channel, including the channel number, and the attributes associated with that channel.

Table I. II Channel Description

<table>
<thead>
<tr>
<th>Board</th>
<th>Channel Description</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
</tr>
<tr>
<td>1</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td></td>
</tr>
<tr>
<td>4</td>
<td></td>
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<td>5</td>
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<td></td>
</tr>
<tr>
<td>16</td>
<td></td>
</tr>
<tr>
<td>17</td>
<td></td>
</tr>
</tbody>
</table>
5. **SYSTEM SOFTWARE**

System processing functions are implemented in the C -signal processor, MAP-300 array processor, CSP-30 and MAP-300 processor as their respective system maintenance tools. Maintenance and validation of these software are presented throughout the following discussion.

The Computer Signal Processor (CSP) was built in the 1980s and subsequently updated in 1990. It is a multifunction processor that can be configured with the software of the CSP-30, 50 or 90A, and also includes software for the maintenance of the MAP-30, MAP-300, and MAP-300. The Computer Signal Processor (CSP) is configured with the Corresponding Software for the program. These are custom software packages that are specific to the system and any arithmetic operations. The computer signal processor is a software tool designed for both the computer and human interface to ensure a real-time online operating environment. The algorithm processor is independent of the MAP-30 series processors and is designed to be a standalone software package.

5.1 **CSP-30 MAP-300 System Initialization Software**

The computer signal processor (CSP) is designed to control system initialization and phase difference compensation for all channels. A system error signal is placed in parallel to the 300-channel error array in order to identify any errors. A phase error signal is also placed in parallel to the receiver array in order to identify any phase errors. The computer signal processor is designed to be a standalone software package that is capable of initializing complex multipliers. The software is designed to be a standalone software package that is capable of initializing complex multipliers. The computer signal processor is designed to be a standalone software package that is capable of initializing complex multipliers. The computer signal processor is designed to be a standalone software package that is capable of initializing complex multipliers. The computer signal processor is designed to be a standalone software package that is capable of initializing complex multipliers.
During the course of system initialization, the MAP processor is tasked with averaging 16 pulse returns. After 16 pulses are averaged by the MAP, the largest voltage magnitude from channel 1 is chosen from the group of 16 range bins. The range cell containing the largest voltage magnitude is the range cell containing the stationary reflector used for channel 1 through 3 normalization. The range cell containing the reflector is first determined, and then the eight complex average voltages chosen from the range cell containing the reflector are transferred to the host minicomputer.

These averaged voltages represent a plane wave parallel to the line of the eight-element receive array. This voltage data is used by CSP-30 software to calculate an amplitude and phase correction factor. The software correction factors are normalized to channel 1 (that is, the remaining seven channels are normalized in phase and amplitude to channel 1). Once the CSP-30 receives the eight-element voltage array, the CSP-30 software calculates antenna weight correction factors normalized to channel 1. The eight-element receive array scattering matrix and the average voltage array transferred to the host by the MAP are used to determine the antenna weight correction factors.

CSP-30 software modifies AMTI and NEMRAD beam-forming coefficients to reflect system channel gain and phase variations, and then transfers the modified beam-forming coefficients and the Doppler filter coefficients to the MAP processor. The CSP-30 function during initialization is to supply the MAP with corrected beam-forming coefficients and Doppler filter coefficients (Doppler coefficients are not altered during initialization).

5.2 CSP-30/ MAP 300 Real-Time Operation

In real-time operation, the MAP-300 APIUs are responsible for AMTI and NEMRAD beam-forming calculations. Typically, a beam-forming calculation for a single range cell would involve eight complex multiplications and a summation of complex products over eight channels. The APIUs are also tasked with Doppler filter implementation. Two identical Doppler filters are implemented within the MAP. One filter is for NEMRAD and one is for AMTI processing. To implement a single filter over one range cell requires 32 pulse returns, 32 complex multiplications and a cumulative sum for a single range cell.

The I/OU is responsible for transfer of receiver data from the radar into MAP memory. The I/OU is tasked with the transfer of processed radar data from MAP memory to CSP-30 memory.

A survey of MAP real-time functions indicates this group of six MAP processors is tasked with NEMRAD/AMTI beam forming, NEMRAD/CSP-30 Doppler processing, and data transfer from the radar and into the CSP-30.
CSP-30 real-time processing functions consist of detection processing, pro-
cessed data recording, and driving a CRT display. Under real-time operating
conditions, the CSP-30 signal processor is almost exclusively utilized to record
processed XEBRAD and ATRI voltage magnitudes, and data batch status. Pro-
cessed data is recorded to cartridge tape for more sophisticated detection process-
ing at a later time. Detection processing is discussed in detail in Section 6.2.4.

3.3 Maintenance Software

Maintenance software consists of a group of programs designed to aid in
receiver hardware troubleshooting and diagnosis. Using the array processor in
combination with the host minicomputer and line printer, hardware diagnosis soft-
ware is used to ensure that the eight-channel receiver is operational. If a problem
is detected, maintenance software is used as a troubleshooting tool.

MAP maintenance software consists of a straightforward transfer of receiver
data from the receiver into the host minicomputer. MAP-100 maintenance soft-
ware transfers one 128 complex word array of radar data into a block of MAP
memory, then the HIM transfers the entire block of unprocessed data into CSP-30
memory. During system maintenance, the MAP serves only as a receiver data
channel. No numerical processing responsibilities are placed on the MAP during
the maintenance mode of operation.

CSP-30 maintenance software consists of several programs. One program
entitled MDS (Minimum Detectable Signal) is written to run with the receiver in a
maintenance configuration. The function of this software is the generation of an
RF transfer characteristic for each receiver channel. Typically, low-power RF
CW is coupled into each channel of the RF front end. The RF energy phase and
power level is under control of external test hardware (phase shifter and wave
guide attenuator). Data is collected over a range of input power levels from approxi-
ately -100 dBm to +10 dBm. At the conclusion of the test, eight RF curves are
produced, relating receiver input power to output power or voltage.

A second maintenance routine, entitled RITCHCK, was written to gain ins-
ight into receiver operation with constant power input. Generally, the receiver
input may be RF CW at constant phase and power, or noise. Independent samples
are taken and compiled over an extended period of time, and a histogram is produced
showing the number of observations as a function of the output voltage magni-
tude. An example of this routine operation is as follows: With no signal present,
noise is present at the output of the receiver. By noting how much ADC dynamic
range is consumed by the system noise, the dynamic range lost due to noise may be
determined.
5.4 MAP Software

Repeating the MAP architecture for use in Section 5.4 discussion, there are six processors within the MAP-300, all of which operate synchronously: the host interface module (HIM) for communication between the CSP-30 host machine and the MAP-300, the I/O control (IOC) for communication between the radar and the MAP-300, the central signal processing unit (CSPU) for processing the returned and controlling the other processors, the arithmetic processing unit (APU), which run in parallel, and the addresser processing control (APCS) which generates a stream of input and output addresses for the API. The various processors are loaded with the necessary programs, started by the CSPU, and their status is reported to the CSPU by the setting on digital output.

5.4.1 MAP Initializations

The system initialization function and purpose is discussed in Section 5.4 which reviews MAP processor activity when the system initialization program is executed.

Initially, the host signals initialization routine to the MAP system, which contains a 1. The CSPU reads a 128-word load program from memory which will be used for numerical summarization. The CSPU then loads the 128-word radar-to-MAP data transfer program, loads the API with a data averaging program, and loads the APS with the corresponding addressing program.

For 64 times, the CSPU turns on the ICS, waits for the radar-to-transmit, then the ICS program turns off the ICS at the end of the transmit, turns on the ICS which in turn starts the API, and waits for an API data flag to be set. This operation divides each data element by 64 and sums them for 64 samples. This complex voltage averaging process takes place with the radar and reflector stationary. The reflector signal is large when compared to the signals in the remaining 16 range cells. The phase and amplitude of the stationary target return signal are averaged for 64 pulses. The receiver channels are to be used as in gain and phase to channel 1. This average phase and amplitude data are sent to the host for each receiver channel normalization.

Next, the CSPU loads the APS with a vector-magnitude program, loads the APS with the proper addressing program. The APS is turned on at this time and resets.

The CSPU waits for the APS to signal that it is ready, which triggers a set of 32-bit vector magnitudes in Memory Bank 2 from memory location 000001. The 16 vector magnitudes are calculated from the average complex data loaded over the range cells of channel 1. The collect MAP (TCP) decrements the range cell with the nearest voltage magnitude, and declares this range cell to contain the stationary initialization target.
The CSPU loads the MPU with a program to determine which range bin of the radar data contains the greatest magnitude, and next loads the APS with the addressing program. The APS is started and the CSPU waits for the APS door flag to be set (it was necessary to introduce additional timing delay to allow time for the result to be placed in memory).

The CSPU loads the HIM with a program to transfer: eight I and Q voltages to the CSP-30 host. The host uses this average voltage data to normalize the gain and phase of channels 2 through 5 to channel 1. A complex normalization factor is calculated in the host by using the eight-element receiver array scattering matrix and the average voltage array transferred from the MAP. The normalization factors are used to modify the AMTI and NEMRAD beam-forming coefficients resident in CSP-30 core memory.

The CSPU loads the HIM with a program to transfer the modified NEMRAD beam-forming coefficients, or antenna weights (eight complex factors), the modified AMTI antenna weights (eight complex factors), the NEMRAD filter coefficients (65 complex factors), and the AMTI filter coefficients (65 complex factors) from the host to the MAP. The CSPU waits for a host ready flag to be set to allow the host machine to compute these quantities and place them in the correct host memory area. When the CSPU senses that the host is ready, it starts the HIM processor, and the antenna weights are placed in identical memory areas on MAP Memory Bus-2 and Bus-3. The filter coefficients are placed on MAP Memory Bus-1. After the data has been transferred to the MAP, the CSPU loads the processors for real-time processing and then waits for another host ready signal before proceeding. This completes system initialization.

4.4.2 MAP REAL-TIME PROCESSING

Assuming that the MAP has performed system initialization and is waiting for the host, all that is necessary to begin real-time processing is to set the host ready flag. Once real-time processing has begun, the host can only reset the MAP with an I/O reset, otherwise the MAP runs indefinitely, as determined by the availability of radar data.

If the MAP has been halted by an I/O reset from the host, but is not ready to go, its memory is not disturbed and real-time processing can be continued without having to go through system initialization by sending the MAP an I/O reset signal for a number greater than 16.

The APF program multiplies eight complex receiver voltages, as determined by the NEMRAD antenna weights, by complex conjugates of the complex result of the integrator. The complex result of the integrator is the complex conjugate of the complex result of the integrator. Location of the complex conjugates is computed by the real-time processor, and the complex conjugates are transferred to NEMRAD core memory.
Beam formation and Doppler filtering is repeated for AMI processing, using identical radar data, identical Doppler filter coefficients, and AMI antenna weights. The API and APS run continuously for a block of 16 radar pulses (or they should finish calculating before more data is available, then set towritten loops).

\[
\begin{align*}
\mathbf{R}_{k, p} &= \sum_{n=1}^{N} \mathbf{N}_n \cdot \mathbf{A}_{k, n, p} \\
\mathbf{R}_{k, p} &= \sum_{n=1}^{N} \mathbf{R}_{k, n, p} \\
N &= \text{Beam forming coefficients weight complex vector}\\
A &= \text{Complex vector complex voltage from radar receive complex voltages} \\
B &= \text{System static and pattern complex voltage per pulse} \\
\mathbf{N}_n &= \text{Doppler Filter coefficients for complex weight} \\
\mathbf{A}_{k, n, p} &= \text{Antenna element weight between receive array} \\
\mathbf{N}_{R_{k, n, p}} &= \text{Range bin for the range bin} \\
\mathbf{R}_{k, p} &= \text{4th order pulse to pulse Coherent Processing Interval at PRI} \\
C &= \text{Doppler filter output (one complex voltage per range bin per CPD)}
\end{align*}
\]

After 16 radar pulses, the API and APS are turned off, switched to the complex vector magnitude program, and turned back on. Following each coherent processing interval, the magnitude of the 16 output filter voltages is calculated and transferred to the host for storage and display.

The radar data enter the AMI's 16-bit floating point numbers. All calculations are performed using 32 bits, and the final vector magnitudes are converted to 16-bit floating point format on AMI memory Bank 2. The MIN program processes the 16-bit floating point magnitudes, assigns an operating range, and an I/O status word to each of 16-bit stored to the host via a serial interface using the data pulses. The CPU checks to see if the data machine takes the data, then proceeds to set up for the next block of data.

The I/O status word sent to the host contains information necessary to set transfers. The status word indicates data validity, error condition, and data count words. There are three separate invalid data codes available. They are:
a. Radar data became available during IOS read (data the IOS read was invalid because as the IOS read data from the radar data buffer, the radar data buffer was being written over with new data).

b. The IOS was running a few microseconds behind the radar (in this case one pulse of radar data was dropped).

c. Radar data becomes available while the MAP was processing and before the IOS started running (at least one pulse of radar data was dropped).

If no data status error or combination of errors occurred during the CPU, the CPU is said to be valid. For the purpose of definition, the CPU output combined with the status word is referred to as a batch.

In order for MAP processing to keep up with the radar data, it was necessary to employ a double buffering input scheme and to have a very efficient API to support the specific features which were necessary. Hence:

a. The IOS is turned on and the radar data, when ready, is transferred to MAP Memory Bus-3.

b. The APU-APS is ready to process data on Bus-3.

As soon as the data has been transferred, the IOS turns itself off and the USPI sets a flag to begin the APU out of an idle state to process that data.

c. The IOS is switched to Memory Bus-2 and enabled, awaiting the next pulse of data which will occur before the APU has finished processing Bus-3 data.

d. After the data has been transferred to Memory Bus-2 and the IOS turns off, the USPI sets a flag to switch the APU to process Bus-2 data. If the APU has finished processing Bus-2 data, it continues to process Bus-3 data until it is ready, at which time it immediately begins to process Bus-2 data. If the APU has finished Bus-2 data before Bus-3 data is ready, it sets an enabling of the flag to switch it to Bus-2 data. The IOS is ready after setting the APU Bus-2 flag to transfer radar data to Bus-3. The USPI ensures an APU flag that ensures that the APU has finished with the data of a particular memory bus before allowing the IOS to transfer the data with another pulse of radar data.

e. This routine continues until the entire set of data has been processed. The data status word is then set to a value indicating the status of all the memory buses, indicating each data ready to be processed (e.g., if Bus-3 is ready the status word is 3). The next cycle commences until the entire set of data has been processed.

f. At the end of the processing, the APU and APS are turned off. The USPI sets the EEPROM vector locations, writes on the APS and APU, then sets the APU to perform the calculations. As soon as the APU is ready, the USPI enables the IOS to transfer the data to Memory Bus-1 and turns off the IOS. The data transfer results in a processor, if the processor is not empty, memory scanning locations and turns on the APS and the APU to perform the calculations.
flag was cleared by the APU before turning off to process another 65 pulses of radar data.

5.4.3 MAP SOFTWARE TEST MODE

If the host sends the MAP a mode word less than zero, the CSU executes a test program that does everything the real-time program does except for the radar-to-MAP data transfers. The MAP Bus-2 and Bus-3 memory locations used for radar data are filled instead with simulated test data from the host machine. The MAP performs all calculations on this test data. The MAP is now synchronized by the data coming from the host machine and will run at a much slower rate.

The host machine is tasked with simulating the radar data matrix and transferring the test data to the MAP for beam formation and filter processing. In the Software Test Operating Mode, the MAP looks to the host for radar data, independent of the operational status of the radar. The simulated test data is generated based on simple target and clutter models. The host calculates a new test matrix on a simulated pulse-to-pulse basis. The host first calculates a test matrix, then transfers the simulated data into the MAP. As the MAP is processing simulated test data, the host also processes the simulated test data in a fashion identical to the MAP. When the host has completed processing of the test data, the MAP transfers processed results back to the host. The host then compares both sets of results and reports any significant errors to the operator. The next sequential simulated data pulse is calculated by the host and transferred to the MAP, and the test process repeats until halted by the operator.

The Software Test Mode of Operation allows a check of the digital processing software and hardware independent of the operational status of the radar.

5.5 MAP-300 Double Buffered Input

In a rigorous real-time radar operation, the time required to transfer the radar data plus the time required to process the data must be less than one interpulse period. This constraint may be loosened somewhat by implementing a double buffer input scheme.

Radar data is double buffered at the input of the MAP-300. Double buffering allows additional time for processing during the radar interpulse period. As will be shown, double buffering allows the processing time to extend through the data transfer interval and extend beyond the interpulse period. Real-time software requires a few microseconds more than the interpulse period to complete the processing of one data pulse; therefore, the additional processing time is acceptable.

Radar data is loaded into two separate buffers located on two data buses, specifically 2 and 3. Each data load pulse is televised into memory on buses 2 and 3, and every memory location is loaded at 150 μs. Data loads begin at 150 μs.
...and data from memory. But as data is actively processed, therefore, the essence of double buffering is the data transfers and data processing proceed in cycles. The data transfer interval appears transparent to the MAP arbitration. However, this double buffering scheme does have serious limitations if the processing interval extends beyond one inter pulse period. There is a processing set-up time for the digitizer's pulse to reach the processing...
input scheme provides the additional time necessary to complete interpulse processing for 65 pulses, while not dropping pulses due to CPU execution speed limitations.

To quantitatively determine the additional available processing time due to double buffering the input data, refer to Figure 34. The time available for real-time processing (p) without double buffering equals the interpulse period (IPP) minus the time required to transfer one pulse of radar data (T), or

\[ p = IPP - T \]

The additional processing time per pulse due to the double buffering input scheme may be found by the use of Figure 34. From the double-buffered input time diagram in Figure 34

\[ 2IPP = T + W + P + B \]

Also note:

\[ W = w-10 \]

\[ P = IPP - T \]

To ensure that MAP memory is not destroyed, constrain B > 0,

\[ IPP = T + (n-1) + 1 > 0 \]

Solving for the lag time (L) beyond the interpulse period (IPP), the processing may extend:

\[ L = IPP - T \]

Letting \( n \) equal the number of pulses contained within one CPI, we get:

\[ L = IPP - T \]

Therefore, the time allowed for processing with double buffering:

\[ P = IPP - \frac{IPP - 1}{CP} \]
Subtracting the available processing time without double buffering from the processing time available with double buffering yields (Eq. 67 minus Eq. 66) Eq. 68.

\[ \Delta t = P - p = \frac{1}{CPI} (\text{PP} - T) \]

\[ \Delta t = T \times \frac{1}{CPI} \]

The processing time gained by double buffering made real-time possible using the MAP-300. Double buffering allowed the processing time to be expanded by the data transfer interval plus a fraction of the processing time available without double buffering \((p)\). The fraction of \(p\) decreases as the CPI increases. The limit of the processing improvement as the CPI approaches infinity is \(T\).

For the NEMRAD/AAI/A processing application:

- CPI = 45
- IPP = 160 μs
- T = 154 μs

Therefore:

\[ \Delta t = 154 \times (45 - 154) = 154 \text{ μs} \]

With 160 μs between pulses, up to 62 μs may be used for data processing with a 68-pulse CPI.

6. FIELD INSTALLATION AND TESTS

Section 5.1 described the components used in experimental testing and the techniques employed to test the 알아 and test systems and the data collection.

Section 6.1 reviews the field installation and testing. The system uses data collection, target detection, and performance evaluation techniques described in Section 5.2. The testing is designed to test the system's target detection accuracy. A weighted target alignment technique is employed to align the target to the system screen. The target was a metallic plate reflector, used as the water target.
Section 5.2 reviews the four phases of field testing. Section 5.3 describes the method used to measure torsional and training parameters. Section 6.1 provides the experimental data. Section 6.2 and 6.3 review the M1-11A1 experiments and external test conditions. Section 6.4 describes the experimental setup.

6.1 Experimental Setup
Figure 35. Radar Van Installation with Generator
NULL FILTER MOBILE RADAR (NFMRAD): CONCEPT VERIFICATION, (U)

OCT 80  W BUCHANAN, H GODLEWSKI, F S HOLT
Figure 37. Radar Equipment Layout (End View of Van)

Figure 38. Floating Radar Platform with Shocks and Sway Braces (Top View)
6.1.3 ENHANCED RADAR TARGETS

6.1.3.1 Pickup/Corner Reflector Combination Target

The enhanced radar target utilized in moving target experiments described in Section 6.2.4 consisted of a large corner reflector mounted on the passenger side of a six-passenger pickup (see Figure 42). The corner reflector is depicted in Figure 43. A technician riding in the radar target vehicle insured that the target vehicle was positioned broadside to the radar van as the two vehicles traverse taxiways "Raytheon" and "Whiskey" shown in Figure 48. This was accomplished by the use of a boresight mounted adjacent to the large target corner reflector. A spotlight mounted above the radar van receive array served as an optical alignment aid for use with the boresight.

6.1.3.2 Flat Plate Reflector

The 12-in. (30.5 cm) square flat metal plate shown in Figure 44 was tripod mounted and used for system initialization. An optical alignment procedure insured broadside flat plate positioning. A light source held adjacent to the rifle scope seen in Figure 44 was used to insure broadside flat plate alignment relative to the receive array face. The experimental layout for system initialization is similar to that shown in Figure 45. Approximately 293 ft (89.3 m) from the radar van, a series of metal plates canted skyward were dispersed to break up the specular multipath component. Figure 45 shows the flat plate 450 ft (137 m) away from the radar.
Figure 41. Remote Test Facility

Figure 42. Enhanced Radar Target
van used as a target for two-way antenna pattern measurements or for system initialization. Typically, the flat plate or corner reflector (Section 6.1.3.1) would be used as a stationary target for antenna pattern measurements or system initialization. Due to the fact that the corner reflector provided a substantially larger radar cross section over a broader beamwidth, the corner reflector was used as a target most often.

6.2 Field Tests

NFMRAD [AMTI] field testing was divided into four separate phases. Phase 1 of field testing yielded two-way antenna patterns for the experimental radar system. Phase 2 consisted of experimentally verifying Doppler filter performance. A Doppler filter transfer characteristic was obtained with the radar position held constant measuring radar performance for converging and diverging targets traversing the radar main beam. In phase 3, target detection threshold voltages were determined. After target detection thresholds were established, phase 4 involved
Figure 44. Flat Plate Reflector

Figure 45. Field Site Used for Antenna Pattern Measurements (Top View)
moving target data collection in an effort to establish the probability of detection
\( P_d \) for NEMInAD processing and conventional AMTI processing.

The discussion that follows will clarify the experimental details involved with
generation of the data described above. Conditions in which the experimental
antenna patterns, filter characteristics, target thresholds, and \( P_d \) curves were
obtained will be discussed.

2.2.1 MEASUREMENT OF TWO-WAY ANTENNA PATTERNS

The discussion above deals with the generation of the antenna patterns
shortly...

Several antenna pattern measurement techniques were tried before a final
scheme was adapted for antenna pattern generation. Initially, with the radar van
stationary, a flat plate reflector was moved in azimuth at constant range. Pattern
data was accumulated and stored on tape. Typically, the target was moved 2\( ^\circ \) in
azimuth, realigned optically with respect to the radar van receive array, and
pattern data was recorded on tape. The cycle would repeat from 0\( ^\circ \) to 180\( ^\circ \) bearing
angle. An antenna pattern measurement consumed 4 hours; therefore, the measure-
ment scheme was modified. The modified measurement scheme involved moving
the receive array with respect to a stationary passive flat plate or corner reflector.

The modified scheme involved displacing the radar van in azimuth and holding
the reflector stationary. Figures 45 and 46 depict the experimental layout utilized
when transmit/receive patterns were measured. The van and receive array would
advance 2\( ^\circ \) in azimuth and step, pattern data was recorded on magnetic tape, the
truck would then be positioned forward 2\( ^\circ \), and the cycle repeated. This modified
antenna pattern measurement scheme was not without drawbacks, however.

![Figure 45. Field Site Used for Antenna Pattern Measurements (Side View)]

There was concern the van would move outside the beamwidth of the stationary
reflector. A second concern was multipath interference due to the low reflector
and receiver array height. The terrain over which pattern data was collected was
not flat, which lead to concern over vertical beam deflection as two-way patterns
were measured. When advancing the van over unlevel concrete, the reflector
height would stay constant; however, as the truck advanced through arc, the relative height and tilt of the antenna array face with respect to the stationary reflector surface would vary slightly. The reflector illumination would vary due to multipath lobes formed in the beam; therefore, the pattern measured would typically be a composite of an antenna pattern generated primarily by radar azimuthal position but modulated by radar vertical displacement. The passive reflector was located at a range of approximately 450 ft (137 m). The van was driven in a tight turning arc. The radar van was displaced radially approximately 10 ft (3 m) from the reflector as data was collected over azimuth bearing angles of 44° through 136°. The receive array displacement parallel to the reflector was approximately ±17 ft (±5.2 m) referenced to broadside of the reflector. Every 8° of arc, the flat plate was optically aligned and the multipath reflector wall repositioned. Flat plate height relative to the ground was 4.5 ft (1.37 m). Height of the receive/transmit antenna array was 8.5 ft (2.57 m). (When using the corner reflector shown in Figure 43 for making pattern measurements, it was not necessary to adjust reflector position as the radar van advanced through arc. The reflected power variation due to the van displacement was less than 1 dB.) The illuminated multipath patch of earth was masked as seen in Figure 45. The earth surface between the radar van and target was not perfectly flat, which presented a problem when trying to accurately ascertain the ground patch significantly contributing to multipath interference. A shotgun approach was adopted and a large ground patch was masked, using long metal reflectors canted skyward and X-band absorber material.

Single-element patterns were measured for each of the eight-receiver channels. Two-way patterns, measured as described above, were obtained by uniformly weighting each receive channel and summing over the eight-element array to form one composite pattern (hereafter referred to as the AMTI pattern). Finally, each receive channel was weighted to form the XFMRAD pattern. Theoretical and experimental patterns are presented in Sections 7.1 and 7.2. Patterns previously discussed in Section 6.2.1 are referenced in Chapter 7 as AMTI and XFMRAD.

Typically, antenna patterns were measured using common real-time operating software with modified antenna weights and Doppler filter coefficients. For example, to measure an antenna pattern for channel 1, the following software modifications are required. Channel 1 antenna weight is set to 1+j0; weights for channels 2 through 8 are set to 0+j0. Doppler filter coefficients are modified to form a de pass filter, or all 65 filter coefficients are uniformly weighted to 1.65+j0. The resulting software zeros channels 2 through 8 receiver data and sums the weighted receiver data over all eight channels. After 65 pulses, an average channel 1 receiver output is generated for use as data in channel 1 single-element antenna pattern.
Similarly, to generate an AMTI antenna pattern, antenna weights in each channel are set to 1 \( \times \) 0. Doppler filter coefficients are uniformly weighted to form a de pass filter just as in the single-element pattern.

One additional concern is the requirement for the eight-receiver channel characteristics to appear uniform in phase and amplitude from channel to channel; all phase path lengths and gains must be uniform channel to channel. In practice, this constraint is not realized; therefore, the receiver channels are normalized in to ensure channel uniformity before each series of measurements. The channel weight correction is a software normalization of channels 2 through 8 to channel 1. Before data is collected for processing, a channel normalization takes place within the software. Specifically, the gain and phase of all receiver channels are normalized to channel 1. A correction factor for each channel is calculated based on receiver response to an incident plane wave parallel to the eight-element array aperture. The correction factor assures channel to channel uniformity by modifying each channel weight approximately, and assumes linear time invariant receiver operation. Before radar data is collected for processing, the channel normalization procedure is performed.

A flat plate or corner reflector is used as the signal source in the normalization procedure. A reflector is set normal to the receiver array and aligned optically. A multipath wall is constructed over an appropriate ground patch. Data is accumulated to determine gain and phase differences of each channel, and then the channels are normalized to channel 1. Antenna weights are modified by the appropriate correction factors, and the normalization procedure is complete.

To calculate an antenna pattern, generally several (5 to 10) records of data were obtained per azimuth bearing angle. For the purpose of system term definition, one record consists of 60 data batches. One data batch is output following the processing of 65 received pulses over which the coherent processing interval extends. Every data batch contains processed data resulting from two radar processing techniques operating on a common radar data base. In the case of normal real-time radar operation, NFMRAD and AMTI processing would be implemented; consequently, the batch output would consist of NFMRAD and AMTI processed data. In addition to processed output data, the data batch also contains a batch status work indicator. This status word is an indication of data validity that is, did the MAP processing stay ahead of the radar or was data lost over the 65-pulse coherent processing interval? If a data pulse was dropped or written over, or if the MAP ran "barely behind" the radar during the coherent processing interval, the status word indicates the integrity of each processing interval. Every data batch is written to cartridge tape in groups of 50. Sixty batches make up a record. In normal radar operation, one data batch consists of 16 NFMRAD range bin words (one voltage magnitude per range bin, no phase information) followed by 16 AMTI
range bin magnitudes. The final word in the data batch serial string is the status word. Table 12 depicts the contents of one tape record during normal real-time radar operation. Generally, several records were averaged per bearing angle or velocity to obtain an antenna pattern or Doppler filter characteristic. A record is constructed of 1880 words or 60 batches, and is defined for normal radar operation as shown in Table 12.

Table 12. Tape Record
Obtained during normal
real-time radar operation

| NFMIRAD voltage magnitude, range bin 1, batch 1 |
| NFMIRAD voltage magnitude, range bin 2, batch 1 |
|                                |
| NFMIRAD voltage magnitude, range bin 16, batch 1 |
| AMTI voltage magnitude, range bin 1, batch 1 |
| AMTI voltage magnitude, range bin 2, batch 1 |
| AMTI voltage magnitude, range bin 16, batch 1 |
| STATIS word indicator 65 pulse interval, batch 1 |
| NFMIRAD voltage magnitude, range bin 1, batch 2 |
| AMTI voltage magnitude, range bin 16, batch 2 |
| STATIS word indicator 65 pulse interval, batch 2 |
| NFMIRAD voltage magnitude, range bin 1, batch 3 |
| AMTI voltage magnitude, range bin 16, batch 3 |
| AMTI voltage magnitude, range bin 16, batch 60 |
| STATIS word indicator 65 pulse interval, batch 60 |

6.2.2 MEASUREMENT OF DOPPLER FILTER CHARACTERISTICS

Measurements were made in an effort to verify digital Doppler filter performance. Theoretical filter characteristic is shown in Section 7.4, Figure 32. For a radar moving platform speed of 32 mph (14.4 m/s), the 3 dB bandwidth velocities extend from 7.75 mph (3.47 m/s) diverging through 14 mph (6.21 m/s) diverging, with a peak centered around 11.25 mph (5.03 m/s). Band reject velocities extend from approximately 4 mph (1.79 m/s) converging to 4 mph (1.79 m/s).
Experimental Doppler filter characteristics were obtained by measuring the delay in a single pulse in response to moving target signal returns. The target velocities varied from 0 to approximately 40 mph (64 km/h) for two-way, opposing-direction flights. Typically, several records of data were taken with the moving target centered in the radar main beam. Figure 47 indicates the experimental layout used when measuring the radar Doppler filter characteristic.

![Figure 47. Field Site Used for Measurement of Doppler Filter Characteristic (Top View)](Image)

One record extends over 40 coherent processing intervals (65 pulses per interval). An average of 1 pulse is lost when a batch output is transferred from the MAP-300 to CIES-30 minicomputer for tape storage. Approximately 3960 pulses are transmitted during one record at a rate of 1831; therefore, approximately 2.1 sec expire during the accumulation of one record. The problem of multipath was not addressed when measuring Doppler filter data. The change in target height relative to the radar was also assumed to be negligible due to the short distance traveled over one record and the apparent flatness of the surface traversed.
Each experimental data point in Figure 42 was averaged over many test runs to obtain the characteristic shown. Each point on the experimental curve represents an average over many coherent processing intervals with the number of CPI's per data point a function of the target velocity. Figure 47 indicates the approximate range and sampling interval over which data was accumulated to the interferometric data. The stationary laser recorded a wing target located on the tower traversed through the rear of the tower.

The target was face was situated on the roof of the tower facing the interferometer. The target was illuminated to a level of the RADAR beam by the use of a light source. The device was 10 meters high and 10 meters wide, and the beam was incident at a level of 1000 megawatts. The maximum power available was 7.2 kwhr of 1000 megawatts.

TARGET DETECTION THRESHOLD VOLTAGE DETERMINED

The purpose of Section 4.2 was to determine the threshold voltage of the target by the use of the interferometer. A target was placed in the interferometer and the voltages of the RADAR (and the radar) were measured. The threshold voltage was found to be 50 volts.

Typically, data was collected by moving the radar along a target to the extent of the radar's coverage at a constant speed of 32 mph (14 ft/s) while the target was being measured. The target's position relative to the radar was recorded at the beginning of a series of measurements of 1000 megawatts. The data was then analyzed with the use of a computer to determine the target's position relative to the radar.

Channel normalization was carried out in a controlled moveable target within the extent of the testing. Data collected was assumed to be spatially related but leveled from the mean at a rate of 1 clove taxiway Whiskey through the final pass over Whiskey taken days later. Threshold data collection on taxiway Whiskey was then calculated.
only if the radar van had reached $v_o$ before passing the transmitter tower shown in Figure 48. Typically, 10 records were accumulated per pass down taxiway Whiskey. The pass length extended for 1015 ft (309.4 m) and took approximately 22 s. Due to the roughness of the taxiway–runway intersection, data collection was terminated before the intersection. The effort to obtain data starting position repeatability was driven by the need to have data from pass to pass related on a batch-to-batch, record-to-record basis (that is, batch 1, record 1, of pass 1 was measured over-looking the same terrain as batch 1, record 1 of pass 2). This constraint was imposed by the analysis method employed when calculating target thresholds and $P_d$ curves.

Samples were gathered in an effort to obtain a data base with sufficient information to insure false alarm rates as low as 1 in 24 K. Approximately $2.4 \times 10^3$ samples were collected for the threshold data base.
Three separate analyses for determining target thresholds were implemented. Each scheme utilized a common data base for threshold calculation. The receiver detection threshold voltage is chosen to achieve the desired false alarm probability, $P_f^*$. The false alarm probability of method 1 and method 2 may vary as the receiver gain and noise level drift; however, for analysis, the system was assumed time invariant. Method 3 utilizes a constant false alarm rate (CFAR) average; therefore, uniform drift in receiver gain and noise levels are of little consequence to the detection process (drift does affect system performance by causing channels to not be uniform).

Method 1 involved compiling, sorting, and tabulating range bin voltage data, and forming a histogram. Histograms were necessary for AMTI and NFMRAD data; therefore, two separate histograms were formed and separate thresholds were obtained. All range bins were lumped into one common data group, yielding a histogram containing $16 \cdot R \cdot B \cdot P$ samples where $R$ total number of records per pass, $B$ total number of batches per record, and $P$ total number of data collection passes. There are $960$ NFMRAD [AMTI] voltage magnitudes contained within one record after lumping every NFMRAD [AMTI] range sample together. A single pass down taxiway Whiskey yields $9600$ NFMRAD [AMTI] voltage samples, assuming 10 valid records. To achieve false alarm rates of $1 \times 10^5$, 1050 data collection passes are required. Clearly, so many Whiskey data runs are outside practical limits. A total of 105 passes are required to obtain sufficient data to insure false alarm rates of $1 \times 10^5$. This figure was determined to be beyond tolerable hardware limits; therefore, in the interest of all concerned, 25 passes were run to develop the voltage threshold data base. 240 K samples were collected, yielding a false alarm rate of $1$ in $24 \, K$. The data contained within this data base is filtered radar data; each batch represents one coherent processing interval. One might argue that since each voltage magnitude represents a weighted summation of 65 voltages, the false alarm rate should be $1$ in $1.56 \times 10^5$ instead of $1$ in $24 \, K$.

Target thresholds are determined for NFMRAD [AMTI] by selecting voltages with magnitudes sufficiently large to realize the desired false alarm rate. Since no target is present, any voltage above the threshold voltage chosen should be considered a false alarm. Typically, NFMRAD [AMTI] voltage thresholds, NFMRAD \text{ AMTI} $V_T \left( P_{fa}^* \right) \left[ V_T \left( P_{fa}^* \right) \right]$, are specified given a probability of false alarm $P_{fa}^*$; for example, a NFMRAD false alarm rate of $1 \times 10^5$ yields a threshold voltage written NFMRAD as $V_T \left( 10^{-05} \right)$.

The second method implemented for threshold selection is a modified version of the first. Data is grouped on a range bin basis, forming 16 histograms for NFMRAD and 16 histograms for AMTI. Sixteen threshold voltages are chosen for
NEMRAD (AMTI) detection processing, resulting in one threshold voltage per range bin. The threshold selection scheme is similar to that described earlier. The significant difference in this scheme is the reduction in relative size of available threshold data base. The gain of method 2 is a spatially localized threshold level. Thresholds derived from method 2 are referenced using a k index, for example, range bin 2. NEMRAD probability of false alarm, \( P_f \), as written as

\[
\text{NEMRAD} \quad V_{(2)} \quad (10^{-3})
\]

Method 3 employs a form of CFAR average. For a given NEMRAD (AMTI) data batch, 16 samples are averaged and multiplied by a constant \( \sigma_{\text{NEMRAD}} \) to \( \text{AMTI} \), resulting in a batch-dependent threshold

\[
\text{NEMRAD} \quad \text{AMTI} \quad V_{(8,10)} \quad (V_{(8,10)})
\]

Separate NEMRAD (AMTI) \( \sigma \) values are required. Following CFAR Averages, NEMRAD (AMTI) detection processing is performed. If the threshold value, \( V_{(8,10)} \), is less than any range bin voltage magnitude, \( V_{(RR)}(k) \), \( V_{(RR)}(8) \), a false alarm is counted in the corresponding \( k \) th range bin. This procedure is repeated every R:R:P batches. The number of false alarms is calculated following the conclusion of R:R:P batches. If the number of false alarms is larger than the number of false alarms specified, \( \sigma_{\text{NEMRAD}} \) and \( \sigma_{\text{AMTI}} \) are incremented, and the averaging continues until an \( a \) is established that produces the desired false alarm rate (given the constraint of data base size limitation). The calculated \( \sigma \) values are used to determine the probability of detection when operating on the moving target data base. The CFAR averaging and threshold decisions are performed utilizing data that contains both background terrain and target data. After calculation \( \sigma \) for a given \( P_f \), a \( P_d \) curve is calculated utilizing a separate data base containing the enhanced moving target.

6.2.4 NEMRAD (AMTI) \( P_d \) DETERMINATION

The map of Figure 48 indicates taxiway Whiskey utilized in Section 6.2.3, Target Detection Threshold Voltage Determination, is also used in Section 6.2.4. Data collected for determining detection probability was gathered by using two moving vehicles. In addition to the radar van traversing taxiway Whiskey, an electromagnetically enhanced target vehicle travels Raytheon taxiway in the direction shown. Radar moving platform velocity, \( v_p \), equals 32 mph (14.3 m/s), as in Section 6.2.3 testing; however, Section 6.2.4 differs from Section 6.2.3 in that an
enhanced target vehicle travels broadside to the moving radar platform at \( v_p = 31.3 \text{ mph} \) (14.2 m/s) (see Figure 48 for velocity directions). Raytheon taxiway 16° diverging angle is such that target radial velocity \( v_r \) is approximately \( 11.23 \text{ mph} \) (5.03 m/s), the center frequency of Doppler filter passband.

Data collection begins after the radar van passes taxiway Whiskey transmitter tower if the moving platform velocity \( v_o \) and target radial velocity \( v_r \) are achieved before passing transmitter tower and the target vehicle position is broadside to eight-element receive array mounted on moving platform. The target position was aligned optically with respect to the moving platform eight-element receive array.

Typically, \( v_o \) was achieved before reaching the transmitter tower. Shortly thereafter, the moving target position was aligned followed by stabilization of \( v_r \). If \( v_r \) was realized before passing the transmitter tower, data collection was initiated and continued until moving platform approached runway 23 of Figure 48.

Radio communications were established between the moving radar platform and target vehicle to insure experimental coordination. An optical bore sight alignment technique was implemented i.e., target vehicle to insure proper broadside target positioning. A bore sight mounted in the target vehicle was used for broadside target alignment in conjunction with a narrow-beam light source mounted on the moving radar platform. An observer riding adjacent to the large corner reflector mounted on the target vehicle monitored broadside target position. Verbal feedback from passenger position monitor to target vehicle driver was utilized in conjunction with radio communications to the moving platform. In summary, verbal and optical feedback coupled with nighttime testing made for a somewhat oscillatory (overshoot/undershoot) target position and velocity control. These experimental errors were due to accuracy limitations of vehicle speedometer readings, coupled with optical and verbal velocity and position control. However, velocity variations of \( v_r \) \( (v_r \text{ deviations} \) related to fluctuations in \( v_o, v_r, \) and bearing angle \( \theta \) fall within the 3-dB Doppler passband characteristic of Figure 52. The theoretical NFMRAD and AMTI patterns of Figures 51 and 50 indicate NFMRAD performance is more sensitive to target position fluctuation at broadside than is AMTI, due to the forward center of the NFMRAD main beam.

Real-time software used in Section 6.2.4 measurements was identical to software used in Section 6.2.3; however, the significant difference is the presence of an enhanced moving target traversing Raytheon taxiway broadside to radar moving platform.

Data accumulated pass to pass is again considered to be related batch to batch, record to record (as in Section 6.2.3). Approximately 25 passes were made down taxiway Whiskey, collecting data in Section 6.2.4 testing. Analysis of moving target data base toward the calculation of \( P_d \) values for NFMRAD and AMTI was implemented as described in the following discussion.
Section 6.2.4 data collected traversing taxiway Whiskey was compiled by grouping data batches as a function of taxiway Whiskey position in $x$, as shown in Figure 49. For purposes of this data analysis discussion, let

- $P :=$ data collection passes = 25
- $B :=$ data batches per tape record = 60
- $R :=$ records accumulated per data collection pass = 10
- $K :=$ range bins involved in detection processing
- $SF :=$ distance scale factor
- $p :=$ integer data collection passes index where $1 \leq p < P$
- $b :=$ integer batch index where $1 \leq b < B$
- $r :=$ integer record index where $1 \leq r < R$
- $\bar{x} :=$ normalized Whiskey taxiway position where $0 \leq \bar{x} \leq 1$
- $x :=$ Whiskey taxiway position relative to transmit tower where $0 \leq x \leq S$1
- $k :=$ integer range bin index

\[ V_{\text{FMRAD}}(P_{fa}) \{ V_{\text{FMRAD}}(P_{fa}) \} = \text{FMRAD} \{ \text{ATM} \} \text{ threshold voltage as function of } P_{fa} \]

\[ P_{fa} := \text{ probability of false alarm} \]

\[ V_{\text{FMRAD}(p)} \{ V_{\text{FMRAD}(p)} \} = \text{FMRAD} \{ \text{ATM} \} \text{ range bin voltage magnitude} \]

\[ V_{\text{RB}(k)}(x) \{ V_{\text{RB}(k)}(x) \} = \text{FMRAD} \{ \text{ATM} \} \text{ range bin voltage magnitude} \]

- $V_{\text{FMRAD}}(P_d) \{ P_d \} (x, p) := \text{FMRAD} \{ \text{ATM} \} \text{ probability of detection as function of Whiskey taxiway position and probability of false alarm}$

\[ \text{FMRAD} \{ \rho \} \{ \text{ATM} \} = \text{CFAR real coefficients} \]

\[ V_{\text{FMRAD}}(x, p) \{ V_{\text{FMRAD}}(x, p) \} = \text{CFAR batch dependent voltage threshold} \]

112
We set 

where \( \tilde{\chi} \) is a function of \( l \) and \( u \). The variable \( \tilde{\chi} \) may be expressed as

\[
\tilde{\chi} = \frac{\text{data}}{\text{model}}
\]

where \( \text{data} \) represents the data that \( \mathbf{E} \mathbf{= E} \text{ and } \mathbf{I} \mathbf{=} \mathbf{I} \),

\[
y = \tilde{\chi}
\]

where \( y \) is a total, from having established \( \tilde{\chi} \), for example, total 1, total 2, etc. and it fixes moving platforms position at a Whiskey immediately to the right of transmitter tower. In other words,

\[
\tilde{\chi} \approx \tilde{\chi}(1, 1) \quad \frac{1}{\text{mm}}
\]
or at \( x(1, 1) \), 100 total taxiway pass has been completed. At \( b = B \) and \( r = R \),
\[
\tilde{x} \quad \tilde{x}(b, r) \quad \tilde{x}(B, R) \quad \frac{(10)}{(10)} \quad 1
\]
or a \( \tilde{x}(R, B) \) total taxiway pass has been completed. For any Whiskey position, \( \tilde{x}(b, r) \), \( P \) data batches are used to calculate \( P_d(\tilde{x}) \). The \( p \)th batch specified at \( \tilde{x} \)
contains 16 NFMRAD voltage magnitudes, \( V_{\text{RR}(k)} \), 16 AMTI voltage magnitudes, \( V_{\text{RR}(k)} \), and 1 batch status word, STATWD. The status word is used only to ensure data integrity; STATWD is not used in detection processing. If STATWD indicates batch data is invalid, batch data is not processed. If the \( p \)th data batch at \( \tilde{x}(b, r) \) is valid, detection processing is performed over \( K \) NFMRAD range bins and \( K \) AMTI range bins. Three separate detection procedures are implemented in an effort to determine \( P_d(x) \) sensitivity to the detection procedure utilized. Target threshold

NFMRAD AMTI
\[
V_{\text{RR}(k)} \quad \text{and} \quad \text{1 batch status word, STATWD.}
\]

Three separate detection methods were implemented as follows.

Method 1, described in Section 6.2.3, yields \( V_{\text{RR}} \) \( (P_{fa}) \) and \( V_{10} \) \( (P_{fa}') \). Given probability of false alarm, \( P_{fa} \), one NFMRAD threshold and one AMTI threshold voltage are produced following method 1 detection threshold analysis.

Method 1 detection processing utilized \( V_{\text{RR}} \) \( (P_{fa}) \) \( \{V_{10} \} \) to ultimately calculate \( P_d \) \( \{P_{fa}\} \) as a function of moving platform taxiway Whiskey position \( x(b, r) \). NFMRAD \( \{P_{fa}\} \) AMTI probability of detection \( P_d \) \( \{P_{fa}\} \) for a specified probability of false alarm \( P_{fa} \) at taxiway position \( x(b, r) \) may be written as

\[
\begin{align*}
\text{NFMRAD} & \quad \text{AMTI} \\
\text{P}\_d(x(b, r), P_{fa}) & \quad \text{P}\_d(x, P_{fa}) \quad \{P_d(x(b, r), P_{fa}) \} \quad \text{P}\_d(x, P_{fa}) \quad \text{P}\_d(x, P_{fa}) \\
\end{align*}
\]

A typical analysis to evaluate \( P_d(x, P_{fa}) \) at \( x \) \( x(b', r') \) and \( P_{fa} \) \( P_{fa}' \) follows:

Given:

\( b', r' \) (establishes taxiway position on Whiskey as \( x \))
\( P_{fa}' \) (establishes probability of false alarm)
\( x(b', r') \) is equivalent to \( P \) data batches collected at \( x \).
If $V_{RB}(k) \sim V_1(P_{fa})$ for any $k$ such that $1 \leq k \leq K$ and $p = 1$, then a single target hit is counted.

If $V_{RB}(k) \sim V_1(P_{fa})$ for all $k$ such that $1 \leq k \leq K$ and $p = 1$, then no hit is counted.

Pass counter $p$ is incremented and the detection procedure is repeated for $p = 2, 3, 4, \ldots, P-1$. Following detection processing conclusion at $x_i(k)$, the total number of hits over $P$ passes are totaled. A target is assigned present at $x_i$ if $P_{fa}$. Total hit count at $x_i$ is divided by $P$ passes to obtain probability of:

if $V_{RB}(k) \sim V_1(P_{fa})$, tests $P_{fa}$ from $x_i$ through $x_i$. The constant $x_i$ is varied over the entire Wiskok taxiway $x_i \times x_i$. A threshold is computed to calculate $P_{fa}(x, P_{fa})$.

Method 2, described in Section 6.2.3. yields $V_{RB}(P_{fa}) = V_{RB}(P_{fa})$ where $1 \leq k \leq K$. Given $P_{fa}$, $K$ threshold voltages are produced $K$ NFRAD [AMTI].

Method 2 detection processing utilizes $V_{RB}(P_{fa}) = V_{RB}(P_{fa})$ and $V_{RB}(P_{fa})$ to calculate $P_{d}(x, P_{fa})$ where $1 \leq k \leq K$; generally $K - 1$. In summary, Section 6.2.3 method 2 analysis yields $K$ NFRAD thresholds and $K$ AMTI thresholds. The results from method 2 analysis yield $K$ NFRAD [AMTI] $P_d$ curves for a single $P_{fa}$ value; generally, one pair of $P_d$ curves per range bin. Method 2 processing yields $K$ $P_d$ curve pairs, or one NFRAD AMTI curve pair per range cell for specified $P_{fa}$.

Method 1 detection processing analysis yields one composite NFRAD AMTI curve pair per $P_{fa}$.

Method 2 analysis to evaluate $P_{d}(x, P_{fa})$ for range cell 1 follows. Determine NFRAD probability of detection at $x \times x_i, P_{fa}, P_{fa}^1 \leq 1$. Obtain $V_{RB}(P_{fa})$ from method 2 voltage threshold detection analysis.

If $V_{RB}(x) \sim V_{RB}(P_{fa})$ where $p = 1$, then a target is declared for range bin 1, on pass 1.

If $V_{RB}(x) \sim V_{RB}(P_{fa})$ where $p = 1$, then no target is recorded for pass 1, range bin 1, at $x_i, P_{fa}^1$. Pass counter $p$ is incremented and the detection procedure stated above is repeated for $p = 2, 3, 4, \ldots, P-1, P$. Following $P$ detection calculation decisions, the total number of range bin 1 target hits at $x_i, P_{fa}$ is summed. If target was known to be present in range bin 1, it is assumed present for $P$ taxiway passes. The total number of target hits divided by $P$ yields
NFMRAD

$P_{d}(x', P_{fa})$. If target was known to be outside range cell 1, all target hits are counted as false alarms, and a $P_{fa}$ curve is calculated.

The total number of samples taken for range cell 1 equals $65 \, R \, B / 16$. Let total samples taken for $k^{th}$ range cell equal $sample_{tot}$. To calculate false alarm probability, total false alarm count is divided by $sample_{tot}$. The procedure outlined above yields $P_{d}$ or $P_{fa}$ calculated at $x'$ only. To generate entire curve over $x$, $x$ must vary over entire taxiway, or from 0 through SF.

Method 3 described in Section 6.2.4 yields $NFMRAD$ and $AMTI$ given $P_{fa}$. These $\phi$ values are used in implementing a form of CFAR averaging. Typically,

\[ NFMRAD(\phi), \quad AMTI(\phi) \]

where $1 < k < K$ over $P$ passes. The resultant product is treated as a batch-dependent threshold voltage. To calculate NFMRAD batch-dependent threshold

\[ V_{lb}(b, P_{fa}) = \frac{NFMRAD(\phi)}{AMTI(\phi)} \sum_{k=1}^{K} V_{RB(k)(\phi)} \]

where $p = 1, 2, 3, \ldots P$. The batch-dependent threshold was treated in method 3 detection analysis as in method 1. A single composite curve pair is generated given $P_{fa}$, and 16 curve pairs are generated in a fashion similar to method 2.

7. EXPERIMENTAL RESULTS AND CONCLUSIONS

7.1 Two-Way Antenna Patterns

7.1.1 UNIFORMLY WEIGHTED ANTENNA PATTERN

The antenna pattern shown in Figure 50 was measured as described in Section 6.2.1. Figure 50 depicts the two-way experimental antenna pattern with crosshatches, and the theoretical pattern with a solid line. Experimental pattern data was collected over bearing angles ranging from $44^\circ$ to $136^\circ$. Experimental error in azimuth resolution was estimated to be less than $2^\circ$. The experimental pattern shown is averaged over many tens of data batches, or each crosshatch represents an average of several data records. The theoretical nulls at $68^\circ$ and $112^\circ$ were contributed by the transient pattern; however, the forward null at $68^\circ$ and the aft null at $112^\circ$ were not experimentally verified, as seen in Figure 50.
7.1.2 NFMRAD ANTENNA PATTERN

The antenna pattern shown in Figure 51 was measured as described in Section 6.2.1. Figure 51, like Figure 50, depicts the experimental two-way antenna pattern with crosshatches, and the theoretical pattern with a solid line.

The experimental NFMRAD pattern shows the most serious problem encountered thus far during field testing. Results indicate the experimental NFMRAD (AMTI) radar is unable to synthesize a broad antenna pattern null. Theoretically, the NFMRAD null should extend from approximately 104° to 114°; however, this broad null has thus far not been obtained. Narrow -45 dB nulls have been measured; however, they generally are not repeatable. For the NFMRAD processing scheme to realize performance improvement over AMTI processing, the 14° antenna pattern null is essential. A solution to this problem is now being implemented.

The minicomputer software is being modified to include an adaptive beam-forming algorithm. Beam-forming coefficients will be dynamically updated and the null will be formed adaptively.
7.2 Doppler Filter Characteristics

The Doppler filter characteristic, shown in Figure 32, was measured as described in Section 6.2.2. Experimental filter data was collected over a range of target velocities from 32 mph (14.3 m/s) converging to 32 mph (14.3 m/s) diverging. The experimental data outside the bandstop bandpass region of the filter characteristic is approximately 15 dB higher than the theoretical response.

7.3 Threshold Voltage Determination

Data described in Section 6.2.3 was obtained for determining appropriate target thresholds as a function of false alarm rate; however, our inability to realize a broad repeatable antenna null insured the collected data was inadequate to prove the NEMRAD concept.

7.4 Receiver I/O Characteristic

Figure 33 shows a typical receiver I/O characteristic. RF power varying from -96 dBm to -26 dBm was fed to each of the eight receiver channels. The receiver output voltage was measured, using the MAP and CSPI processors. This typical characteristic exhibits a linear receiver dynamic range of approximately 64 dB. The receiver theoretical dynamic range (assuming no noise) is 90 dB. As seen in Section 7.5, noise due to the RF front end may consume up to a bit or 30 dB of dynamic range. Generally, 24 dB of receiver dynamic range is consumed by the mean noise voltage.

Figure 31. NEMRAD Two-Way Azimuth Pattern
(Theoretical and Experimental)
Figure 52. Doppler Filter Characteristic

Figure 53. Receiver I/O Characteristic
7.5 Receiver Noise Histogram

Figure 54 is a typical histogram of receiver channel noise. The noise amplitude was calculated for 1000 complex independent voltage samples. One may approximate the number of bits consumed by receiver noise using Figure 54. For worst-case analysis, assume the noise voltage is equal to the $Q$ voltage component, and the $I$ voltage component is zero. The mean noise voltage over all eight channels was calculated to be 80 mV. ADC input sensitivity of 10 mV/change of ADC output state; therefore, eight changes of state occur at the output of the ADC. Thus, the mean noise voltage consumes 4 bits of the ADC dynamic range (worst case). From Figure 54, a noise magnitude of 200 mV occurred 10 times out of 1000 samples or 1 percent of the total number of samples. Worst-case analysis of this case yields 20 changes of ADC output state or 5 ADC bits may be consumed by receiver noise (7 mantissa bits total given a constant base 2 exponent).

![Receiver Noise Histogram](image)

A noise histogram, generated using integer $I$ and $Q$ voltage components, will contain voids around 1, 5 and 4, $3 \times 10^{-2}$ V. This apparent problem concerns the histogram generation algorithm and the quantized voltage data, not the statistical nature of the noise data. For an explanation, let $n$ be the noise voltage magnitude or

$$n = \sqrt{I^2 + Q^2}$$
where $L$ and $Q$ are integers. Then the case where $L = 2$ and $Q$ does not exist for any integer $L$ and $Q$ combination. The discrete nature of the quantized voltage data and the histogram, search and sort algorithm are the causes for the foregoing noise test data.

7.6 Preliminary Conclusions and Future Directions

Thus far, we have seen that it is extremely difficult to form images ($Q^T$) and select $Q^T$ matches in our receive antenna pattern.

The beam-forming approach allowed an accurate knowledge of the receive array scattering matrix. The explanation of our experimental results indicates that the correct antenna null may lie in the accuracy with which the scattering matrix was known. Figure 35 shows the sensitivity of null depth to large distributed errors in the scattering matrix. With relatively small errors in the scattering matrix, the null interval $101$ and $102$ rapidly nulls. The scattering matrix is expected because it was measured in an anechoic chamber using the receive array with the transmitter turned off. It seems reasonable to assume that the scattering matrix may be altered slightly due to the interactions between the receive antenna elements and the location of the turbine. One may use the results that we have obtained.

An important consideration is that the channel correction procedure on the receive side may be used to correct for the calculation of some errors between scattering factors. Scatter matrix errors that occur may partially explain the errors between antenna pattern and Doppler filter discrepancies.

The channel correction procedure seems reasonable; however, it cannot conclusively show its utility. Typically, the correction factors were repeatable within one or two decimal digits when compared over as long as 4 hours. Generally, after 4 hours of operation, the drift would be sensed at the output of each or some of the receiver channels. The problem was compounded by having to find and correct for patterns to check and calibrate. Following calibration, channel errors are considered to be under control.

Data are not available to develop software that will relatively easy to receive antenna pattern. The modified system will sacrifice the environment and may lose its advantage in terms of optimize the SNR. This system will ultimately sacrifice is relatively easy to receive antenna pattern to correct for X MiNi customer's field.

An effort to measure the scattering matrix of the receive array antenna in the field is also being pursued.
Figure 76. Effect of Uniformly Distributed Error in the Scattering Matrix on the Null Region of Receiver Pattern. Errors indicated are maximum percent error.
References


Appendix A
Complex Recursive Infinite Impulse Response
(IIR) Digital Filters

AI. FILTER SPECIFICATIONS AND NORMALIZED REAL
LOW-PASS FILTER DESIGN

It is desired to design digital filters that have a single passband or a single
stopband in the frequency interval -pi < \Omega < pi with \Omega = f_p is the pulse repetition
frequency. Except for the special cases where the passbands or the stopbands are
centered at dc, it will be necessary to use filters whose transfer functions are
complex.

For any particular design, the following specifications are needed:
1) Passband (or stopband) extend \omega_1 < \omega < \omega_2
2) Upper bound on ripple in the passband
3) Filter discrimination (ratio of maximum signal power passed to maximum
signal power in the reject region)
4) Rolloff or rate of transition from the pass regions to the reject regions of
the filter
5) Filter response time
6) Sampling period (T = \Omega_p).

Two methods were used to develop desired IIR digital bandpass and bandstop
filters. The first method proceeds as follows:
Let $0 < \omega_1 < \omega < \omega_2$ be the desired passband (or stopband). Using the design procedures described by Gold and Rader, $A_1$, frequencies $\omega_{A_1}$, $\omega_{A_2}$, $\omega_A$, and $\omega_{A_C}$ are defined by the relations

$$\begin{align*}
\omega_{A_1} &= \tan \left( \frac{\frac{1}{2}}{2} \right) \\
\omega_{A_2} &= \tan \left( \frac{\frac{1}{2}}{2} \right) \\
\omega_A &= \frac{2}{2} \\
\omega_{A_C} &= \frac{1}{2}
\end{align*}$$

Based on the specifications of ripple in the passband, discrimination, and rolloff rate, the monographs described in Christian and Eisenmann$^{A2}$ determine suitable filter types (that is, Butterworth, Chebycheff, or Cauer), and for each type the associated voltage transfer loss function, $H(s)$, $S = \Omega / \omega$ that will meet the specifications in a normalized lowpass filter. A plot of a typical power transfer loss function $A(\Omega) = 10 \log_{10} |H(\Omega)|^2$ for a normalized lowpass filter is shown in Figure A1. The quantity $A_{\text{max}}$ is the maximum attenuation in the passband and is a measure of the ripple in the passband. $A_{\text{min}}$ is the minimum attenuation in the reject region, and hence the difference $A_{\text{min}} - A_{\text{max}}$ is a measure of the filter discrimination. The frequency $\Omega_s$ is related to the rolloff rate and, in particular, the quantity $\left(F_s - 1\right)/2$ is the ratio of the width of the transition region at one edge of the passband to the full width (-1 $\leq$ $\Omega$ $\leq$ 1) of the passband. The tables in Christian and Eisenmann$^{A2}$ specify the voltage transfer loss function $A(S)$ by tabulating the zeros $\Omega_{A_1}$, $\Omega_{A_2}$, ... and poles $\Omega_{A_1}$, $\Omega_{A_2}$, ... of $H(s)$ (see Figure A1). The voltage transfer function $F(s)$ for the normalized lowpass filter is given by

$$F(s) = \frac{1}{H(s)}$$

Filters with a single passband or a single stopband can be generated from the same normalized lowpass filter by means of well known transformations described in the next section.


A2. TRANSFER FUNCTIONS FOR SINGLE PASSBAND COMPLEX FILTERS

The transformation

\[ S = \frac{\omega_{A2}^2 - (S')^2}{S'\left(S_{A2}^2 - \omega_{A1}^2\right)}, \quad S' = \omega_{A2}^2 - \omega_{A1}^2 \]

cconverts the normalized lowpass filter into a bandpass filter with voltage transfer function

\[ F_{bp}(S') = F\left(\frac{\omega_{A1}^2 - (S')^2}{S'\left(S_{A2}^2 - \omega_{A1}^2\right)}\right) \]

and symmetrical passbands

\[ 0 < \omega_{A1}^2 \leq \omega' \leq \omega_{A2}^2 \]
and

\[-\omega^*_{A2} < \omega^* < -\omega^*_{A1} < 0\]

and will be denoted by \( B \) and \(-B\) respectively.

In carrying out the present design procedure, it will be necessary eventually to identify the roots and poles of \( F_{bp}(S) \). The voltage transfer function \( F(S) \) for the normalized lowpass filter is ordinarily prescribed by means of its roots and poles, and it is advisable to transform this function factor by factor in order to be able to determine the roots and poles of \( F_{bp}(S) \). Thus each factor \( S - S_0 \) transforms into a rational fraction as follows:

\[
S - S_0 \frac{(S - r^+)(S - r^-)}{S_0}
\]

where

\[
p = \frac{(\omega^*_{A2} - -\omega^*_{A1})S_0}{2} = \sqrt{\left[(\omega^*_{A2} - -\omega^*_{A1})S_0\right]^2 - \omega^*_{A1} - \omega^*_{A2}}
\]

and

\[p = 0\,.

Since \( F_{bp}(S) \) is real on the real axis, its roots and poles will occur in conjugate pairs, and the behavior of the filter in the region \( \omega^* > 0 \) that contains the passband \( B \) will be due primarily to those roots and poles that lie in the upper half-plane. Conversely, the behavior of the filter in the region \( \omega^* < 0 \) that contains the passband \(-B\) will be due primarily to the roots and poles of \( F_{bp}(S) \) that lie in the lower half-plane.

Discarding the roots and poles of \( F_{bp}(S) \) that lie in the lower half-plane creates a complex voltage transfer function that represents a filter with only the single passband \( B \). Conversely, discarding the roots and poles in the upper half-plane results in a complex voltage transfer function that represents a filter with only the single passband \(-B\).

This method can be used to generate a single passband filter, provided that the passband does not overlap \( \omega^* = 0 \). If \( \omega^* = 0 \) is overlapped and \( \omega^*_{A2} = -\omega^*_{A1} \), then a single real lowpass filter will be sufficient. The case where \( \omega^* = 0 \) is overlapped and \( \omega^*_{A2} \neq -\omega^*_{A1} \) will be discussed later.
The single passband filter can be normalized in different ways, resulting in several different forms (see Figure A2). The form shown in Figure 2(b) has been normalized to the average value in the passband.

![Diagram of filters](image.png)

Figure A2. Power Transfer Functions of Filters Used to Realize an Asymmetrical Single-Band Lowpass Filter. Filter (a) cascaded with filter (b) produces filter (c).

### A3. TRANSFER FUNCTIONS FOR SINGLE STOPBAND COMPLEX FILTERS

The transformation

\[ S = \frac{\omega_s (\omega_A - \omega_A^2 s^2)}{\omega_A \omega_A^2 + (s^2)} \]

converts the normalized lowpass filter with voltage transfer function \( H(s) \) into a bandstop filter with voltage transfer function.
\[ F_{bs}(S) = \frac{1}{\frac{1}{M^2} \left( \frac{\omega}{M} - \omega_i \right)^2} \]

and symmetrical stopbands \( B \) and \(-B\).

As with the bandpass filters, it is desired to identify the roots and poles of \( F_{bs}(S) \), and to this end it is advisable to transform the voltage transfer function \( E(S) \) factor by factor in order to determine the roots and poles of \( F_{bs}(S) \). Thus each factor \( S - S \) transforms into a rational fraction as follows:

\[
S - S \frac{(S^2 - r^2)(S^2 - p^2)}{(S - p)(S^2 - p^2)}
\]

where

\[
r^2 = \frac{\Omega^2 (\omega_{A2} - \omega_A)}{2S} \pm \sqrt{\frac{[\Omega^2 (\omega_{A2} - \omega_A)]^2}{[2S]^2} - \omega_A^2 \omega_A}
\]

\[
p^2 = \pm j\sqrt{\omega_A \omega_A}
\]

The zeros and poles of \( F_{bs}(S) \) will as before occur in conjugate pairs, and the behavior of this filter in the regions \( \omega > 0 \) which contains the stopband \( B \) is due principally to the roots and poles that lie in the upper half-plane. Also, the behavior of the filter in the region \( \omega < 0 \) which contains the stopband \(-B\) is due principally to the roots and poles that lie in the lower half-plane.

Discarding the lower half-plane roots and poles in \( F_{bs}(S) \) creates a complex voltage transfer function with only the one stopband \( B \). Conversely, discarding the roots and poles of \( F_{bs}(S) \) that lie in the upper half-plane creates a complex voltage transfer function with only the one stopband \(-B\).

This procedure can be applied as long as the desired stopband does not include \( \omega = 0 \), if the desired stopband contains \( \omega = 0 \) and the interval is symmetric with respect to \( \omega = 0 \), then a real lowstop filter will be the solution. The design of filters for asymmetrical intervals including \( \omega = 0 \) is discussed below.
Assume \( \omega_2 = 0 \), \( \omega_3 = 0 \), and \( \omega_4 = 1 \). Then \( \omega_1^A = 0 \), \( \omega_1^M = 0 \), \( \omega_1^N = \omega_1 \), the interval 1 given by \( \omega_1^A \times \omega_1^M \times \omega_1^N \) is asymmetric and contains \( \omega_1^M = 0 \). Figure A200. The transfer function for a complex bandpass filter with bandwidth \( \omega_1^A \times \omega_1^M \times \omega_1^N \) is given by the transfer function for a complex bandstop filter with \( \omega_1^A + \omega_1^M + \omega_1^N \) (see Figure A20). The transfer function for a complex bandstop filter with bandwidth \( \omega_1^A + \omega_1^M + \omega_1^N \) is given by the transfer function for a complex bandpass filter with bandwidth \( \omega_1^A \times \omega_1^M \times \omega_1^N \) (see Figure A20). The transfer function for a complex bandpass filter with bandwidth \( \omega_1^A \times \omega_1^M \times \omega_1^N \) is given by the transfer function for a complex bandstop filter with bandwidth \( \omega_1^A + \omega_1^M + \omega_1^N \) (see Figure A20). The transfer function for a complex bandstop filter with bandwidth \( \omega_1^A + \omega_1^M + \omega_1^N \) is given by the transfer function for a complex bandpass filter with bandwidth \( \omega_1^A \times \omega_1^M \times \omega_1^N \) (see Figure A20).

**Figure A31:** Partial Transfer Functions of Filters Used to Realize an Asymmetrical Single-Bandpass-Bandstop Filter. Filter (a) cascaded with filter (b) produces filter (c).

If \( \omega_2 = 0 \), \( \omega_3 = 0 \), and \( \omega_4 = 1 \), then \( \omega_1^A = 0 \), \( \omega_1^M = 0 \), and \( \omega_1^N = \omega_1 \) and the interval 1 given by \( \omega_1^A \times \omega_1^M \times \omega_1^N \) is asymmetric and contains \( \omega_1^M = 0 \). Proceeding, similar to those described above, can be used to produce transfer functions for complex bandpass or bandstop filters with \( \omega_1^A \times \omega_1^M \times \omega_1^N \) or \( \omega_1^A + \omega_1^M + \omega_1^N \), respectively.
V3. DIGITAL FILTER REALIZATION

Whether bandpass or bandstop, each of the voltage transfer functions of the single-band filters created by the methods described above is of the form of a ratio of polynomials in \( S \), the coefficients in general, being complex. The generating function \( G(Z) \) for producing the digital realization of a given voltage transfer function \( E(S) \) is obtained by means of the transformation

\[
S = \frac{Z - 1}{Z - 1}
\]

This transformation maps the imaginary axis in the \( S \) plane onto the boundary of the unit circle in the \( Z \) plane. In particular, it maps the points \( S = \pm e^{j\alpha} \) and \( S = \pm e^{j\beta} \) into the points \( Z = e^{j\alpha} \) and \( Z = e^{j\beta} \) respectively. Thus

\[
G(Z) \equiv \left\{ \frac{Z - 1}{Z - 1} \right\}
\]

and \( G(Z) \) will be a ratio of polynomials in \( Z \). The coefficients of these polynomials determine recursive forms of digital realizations of \( E(S) \).

V6. IMPROVED DESIGN PROCEDURE

An improved design procedure for deriving digital filter generating function for single-band bandpass or bandstop filters has been developed. The design

1. Initially, the single-band bi-quad filter is selected to be used.
2. A set of the bi-quad filter is selected in step 1, for the given design

   \[
   e^{-j\theta} \left( e^{j\theta} - e^{-j\theta} \right) 
   \]

   are used, where \( \theta \) is the angle that \( \theta \) is the angle that corresponds to the required

   \[
   e^{-j\theta} \left( e^{j\theta} - e^{-j\theta} \right) 
   \]

   are used, where \( \theta \) is the angle that \( \theta \) is the angle that corresponds to the required
where

\[ \omega'_n = \tan \frac{\omega_1 - \omega_2}{4} \ T \]

This produces a voltage transfer function \( H_p(s) \) for a lowpass filter with passband

\[ \omega'_n < \omega < \omega'_N \]

The transformation

\[ S \equiv \frac{Z' - 1}{Z' + 1} \]

converts \( H(s) \) to a generating function in the \( Z' \) plane for a lowpass filter with

passband

\[ \frac{Z_1 - 1}{Z_1 + 1} < \frac{Z - 1}{Z + 1} < \frac{Z_N - 1}{Z_N + 1} \]

The transformation

\[ T \equiv \frac{1}{2} \ln \left( \frac{Z + 1}{Z_1 - 1} \right) \]

rotates the center of the lowpass filter to the center of the passband. The center is at \( \omega_1 \), denoted by \( c \). The generating function is then given by

\[ G_{b/c} = \frac{1}{2} \sum_{k=1}^{N-1} \frac{\omega_1 - \omega_k}{\omega_1 - \omega_{k+1}} \ \left( \frac{Z - 1}{Z + 1} \right)^{k-1} \]

This completes the proof and establishes the relationship between the \( H(s) \) and \( H_p(s) \) of the lowpass filter and the \( S \) and \( T \) transformations.
and

\[
\mathcal{G}(c) = \left\{ \frac{\log(\log 4)}{2} \right\}
\]

as before, shift the \( \epsilon \)-plane and rotate the stopband of the lowpass filter to the new location. The generating function \( \mathcal{G}(c) \) is then given by

\[
\mathcal{G}(c, d) = \left\{ \frac{-c, \log(4)}{1 - \frac{1}{c} + \frac{1}{c} - 1} \right\}
\]

A2 CONCLUSIONS

The use of Chebyshev's correction function and the implementation of the Chebyshev-type digital filters in the MuMAD tests, the application of the direct \( \Delta - \Delta \) filter, and the reduction of the filter's frequency response to the accuracy of the constants in the filter. It is shown that the reduced numerical results are matched by the Chebyshev-type digital filters very satisfactorily. Filter reconstruction using reduced accuracy to the accuracy of the constants,


A new MuMAD filter was implemented using the transfer function of the model of the MuMAD system. The model was defined filters for the filter design. The filter was designed using digital filters and was implemented using digital filters. The filter was tested using digital filters. The results were very promising.
Figure A4. IR Digital Filter Response Times
Appendix B

Complex Nonrecursive Finite Impulse Response
FIR Digital Filters

B. FREQUENCY SAMPLING DESIGN METHOD

Using the frequency sampling technique given in Reference 7, the filter coefficients were designed by designating one pole of the complex poles of the desired response at the desired frequency. The remaining poles are then placed at the reciprocal frequencies of the desired response. The magnitudes of the desired response are then computed using the recursive equations. The desired responses are then approximated by the impulse response of a lowpass filter with a magnitude response equal to the desired response. The filter is then designed to have the desired magnitude response, and the filter coefficients are then computed using the recursive equations. The filter coefficients are then quantized to the desired word length, and the filter is tested for the desired performance.

References


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The complex single passband filter of the type needed for NEMRAD consists in a real lowpass FIR filter of the desired bandwidth and discrimination is designed and then rotated in the Z plane to position the passband at the desired center frequency. For a single passband filter, a real lowpass FIR filter whose stopband region has the desired stopband width is designed and then rotated in the Z plane by the amount necessary to locate the stopband at the desired center frequency. The Z plane-rotation needed for the stopband filter is essentially that which corresponds to the difference frequency between half the pulse repetition frequency for 2D) and the Doppler clutter frequency at the center of the antenna transmission beam.

The action of an n-point bandpass FIR filter, together with an n-point bandstop FIR filter, can be achieved by cascading to form a single (2n-1) point FIR filter. For FIR filters with n = 14, 33, or 65, the response time is far less than for the FIR filters described earlier. For this reason, cascaded FIR filters were used for all subsequent investigations.

In the work that follows, FIR filters are designated according to their design. Figure 31 shows the design of a 35-point FIR bandpass with a points in its bandpass and 2 transition points, T1 and T2. The designation code for this filter, 1F-35-1-2, translates as follows:

1F bandpass filter (B) bandstop filter
35 total number of points in the filter
2 transition points in the bandpass 2-NEW-1
1 number of transition points.
Figure B1. Specification of Filter Response Values for a BP33-3-2 FIR Filter.

Table B1 shows some of the filters that were used in the theoretical investigations associated with NFMRAD.

Table B1. FIR Filters Used in NFMRAD Investigation

<table>
<thead>
<tr>
<th>Bandpass Filters</th>
<th>Bandstop Filters</th>
</tr>
</thead>
<tbody>
<tr>
<td>BP15-1-1</td>
<td>BS15-7-1</td>
</tr>
<tr>
<td>BP33-1-1</td>
<td>BS33-12-1</td>
</tr>
<tr>
<td>HP33-2-1</td>
<td>BS33-13-1</td>
</tr>
<tr>
<td>BP33-3-2</td>
<td>BS33-14-1</td>
</tr>
<tr>
<td>BP65-3-1</td>
<td>BS65-23-1</td>
</tr>
<tr>
<td>BP65-3-2</td>
<td></td>
</tr>
</tbody>
</table>
Appendix C

Control Panel Operational Range Equations
Range Cell Width = 131.15 ft (40 m)
Range Coverage ≥ 2100 ft (640 m)

Figure C1 depicts the three operational range modes of the radar: NEGATIVE, SHORT, and LONG. The position of switches E and F selects one of the radar range modes as defined in Table C1.

![Diagram of Control Panel](image)

Figure C1. Radar Control Panel
Table C1. Radar Operating Range Mode

<table>
<thead>
<tr>
<th>Mode</th>
<th>E</th>
<th>F</th>
</tr>
</thead>
<tbody>
<tr>
<td>LONG</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>SHORT</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>NEGATIVE</td>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>

Equations (C1), (C3), and (C5) express the radar search range as a function of range bin (RB) and control panel configuration (DCA, TRP, D7, etc.). The range mode is determined by the position of control panel switches 1, 2, and 3.

### C1. LONG DELAY RANGE EQUATION

Equation (C1) expresses radar search range (γ) as a function of 10, D1, ..., D7 control panel switch position, and the range bin (RB) of interest when in the LONG delay range mode. D_set is the control panel switch (D7, D6, D5, ..., D0) set during radar operation. To calculate γ when in the LONG delay range mode: Let D7 = 7, D6 = 6, D5 = 5, ..., D0 = 0, and RB is an integer representing the range bin of interest such that 1 ≤ RB ≤ 16.

**LONG Delay Range Equation**

\[
\gamma = 63,5763(15(D7-D_{set}) + 10 + 2 RB) \text{ ft} - 517,25 \text{ ft} \quad (C1)
\]

The time delay (Δt) between the transmitter MAIN BAND and the radar 16-bit SAMPLE command is given by Eq. (C2).

**LONG Delay Time Equation**

\[
\Delta t = \frac{2}{15 \text{ MHz}} (15(D7-D_{set}) + 10 + 2 RB) \quad (C2)
\]

### C2. SHORT DELAY RANGE EQUATION

Equation (C3) expresses radar search range (γ) as a function of A, B, C, D control panel switch positions and the range bin (RB) of interest when in the SHORT delay range mode. Control switches A, B, C, D are binary weighted when calculating γ as shown in Table C2.
Table C2. DCBA\textsubscript{10}, Equivalent for Range Calculations

<table>
<thead>
<tr>
<th>Control Panel Switch Position</th>
<th>Base 10 Equivalent</th>
</tr>
</thead>
<tbody>
<tr>
<td>D C B A</td>
<td></td>
</tr>
<tr>
<td>0 0 0 0</td>
<td>0</td>
</tr>
<tr>
<td>0 0 0 1</td>
<td>1</td>
</tr>
<tr>
<td>0 0 1 0</td>
<td>2</td>
</tr>
<tr>
<td>0 0 1 1</td>
<td>3</td>
</tr>
<tr>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
</tr>
<tr>
<td>1 1 1 1</td>
<td>15</td>
</tr>
</tbody>
</table>

SHORT Delay Range Equation

\[ \gamma = 131,1525(16,375 + RB - DCBA) \text{ ft} - 517.25 \text{ ft} \]  \hspace{1cm} (C3)

\[ \Delta t \text{ may be calculated as expressed in Eq. (C4).} \]

SHORT Delay Time Equation

\[ \Delta t = \frac{4}{15 \text{ MHz}} (16,375 + RB - DCBA) \]  \hspace{1cm} (C4)

C. NEGATIVE DELAY RANGE EQUATION

Equation (C5) expresses the radar search range (\( \gamma \)) as a function of A, B, C, D and RB when in the NEGATIVE delay range mode. A negative result from Eqs. (C5) and (C6) indicates the 16-bit SAMPLE command of interest occurs before the transmitter MAIN BANG. A positive result indicates the SAMPLE command occurs after the MAIN BANG.

NEGATIVE Delay Range Equation

\[ \gamma = 131,1525 (1/2 + RB - DCBA) \text{ ft} - 517.25 \text{ ft} \]  \hspace{1cm} (C5)

NEGATIVE Delay Time Equation

\[ \Delta t = \frac{4}{15 \text{ MHz}} (1/2 + RB - DCBA) \]  \hspace{1cm} (C6)
Appendix D
ADC Alignment Procedures

The purpose of this procedure is to ensure that the ADC assembly is calibrated and will not introduce additional errors into the receiver/processor system. All comments and adjustments will be made with reference to the I side of the assembly; however, they apply equally to the Q side.

Recommended Equipment
- Dual trace oscilloscope
- Signal generator
- 0-5 V dc variable power supply
- LED display box (or eight-channel logic analyzer)
- BNC to SMA adapters as required
- Digital multimeter (DMM)
- Two 18-wire ribbon connectors (MH PAC 362940-01 or equivalent).

D1. 8-BIT ADC ALIGNMENT

D1.1 Initial Setup

Connect an 18-wire cable from the 4-bit ADC power distribution block to the 8-bit ADC assembly under test. The cable must be connected to the 8-bit ADC board row C, pins 19-32. Ensure the black white pair of wires at the ADC board (pins 19-20) is connected to the 4-bit power distribution block (pins 1-2 or 16-21).
Reversing the supply sequence can cause serious damage to the ADC (refer to Figure D1).

Connect a coax cable to the 8-bit SAMPLE command input port from the receiver digital hardware backplane (refer to Figure D2).

Connect the dual trace scope as follows:
1. Trace "A" to the analog input
2. Trace "B" to the input of the ADC (pins 17 and 18)
(Refer to Figure D3(a) for pin locations.)

Connect the signal generator to the analog input port of the ADC under test.
Turn each power on.
D1.2 ADC Assembly Gain Adjustment (Refer to Figure D3(b))

Ensure the analog input (signal generator) is set for 100 mV peak to peak at 250 KHz as measured with the scope (trace A).

Monitor the input to the ADC (trace B) and adjust the GAIN pot of the wideband amplifier for a gain of 6.3 (630 mV) output.
DL3 ADC Assembly Offset Adjustment (Refer to Figure D3(b))

The purpose of this adjustment is to ensure the analog input has a 0 V dc offset (with respect to ground) as measured with an oscilloscope. Any dc offset to the analog signal will cause the ADC output to shift a proportional amount and invalidate any gathered data.

Monitor scope trace A and adjust the analog input (signal generator) for 0 V dc offset.

Remove the trace B probe from the ADC (pins 17 and 18) and connect it to the sample and hold unit input (pins 10 and 11). (Refer to Figure D3(a) for pin locations.)

Monitor the scope (trace B), and adjust the OFFSET pot of the side-band amplifier for 0 V dc offset.
II. 8-bit ADC Output Calibration Setup (Refer to Figures 14 and 15)

1. For the following steps, refer to the 8-bit ADC output calibration setup diagram (Figure 14) for reference.

2. Connect the BNC connector to the ADC output calibration setup diagram (Figure 14). The output feature is to be activated.

3. Connect the DMM to the output of the ADC (pins 17 and 18).

4. Connect the 0-1 V dc variable power supply to the input of the ADC (pins 17 and 18).

5. Turn rack power on. Allow 10-minute warm-up time for the ADC to stabilize.

II.5 8-bit V/D Adjustments (Refer to Figure 15(b) for Location of Pots)

1. Monitoring the DMM, adjust the variable power supply to -1, 110 V dc.

2. Adjust the -1 V DC pot until the LED display indicates the following: (red LEDs represent L side, Green LEDs represent Q side)

   - X7, X6, X5 should be on.
   - X4 should be blinking.

3. As the voltage is decreased to -1, 110 V dc., X4 should be on, and as the voltage is increased to -1, 120 V dc, X4 should be off.

4. Monitoring the DMM, adjust the variable power supply to +1, 120 V dc.

5. Adjust the +1 V DC pot until the LED display indicates the following:

   - All LEDs should be off.
D2.1 4-bit ADC Gain Adjustment

Before starting, make sure the 4-bit ADC module is on.
Connect a signal generator to each analog input. Ensure that the generator is set for 250 kHz at 100 mV peak to peak.
Monitor current trace A and adjust the signal generator for 0 V dc offset.
Monitor current trace B and adjust the signal generator for ±100 mV peak.

D2.4 4-bit ADC Output Calibration (Refer to Figures D4 and D6)

In the following procedure, the F100 tester is used to allow visual monitoring of the actual ADC output. An eight-channel logic analyzer may be used if desired (refer to Table D5 for 8-bit ADC output presentation).

1. Turn power on.
2. Disconnect the oscilloscope probe and signal generator from the 4-bit ADC output.
3. Connect an 800 Ω cable from the ADC output connector to each RIGOL 1400 scope's channel 1-120 or the F100 Tester's channel 1-120.
Table D1: 4-Bit ADC Output

<table>
<thead>
<tr>
<th>Analog Input (V DC)</th>
<th>Digital Output</th>
</tr>
</thead>
<tbody>
<tr>
<td>-2.000</td>
<td>1111</td>
</tr>
<tr>
<td>-1.526</td>
<td>1100</td>
</tr>
<tr>
<td>-1.250</td>
<td>1000</td>
</tr>
<tr>
<td>-0.750</td>
<td>0000</td>
</tr>
<tr>
<td>0.000</td>
<td>0000</td>
</tr>
</tbody>
</table>

(Note: The most negative analog input corresponds to full-scale output.)

Connect the V+ to the LED display box using a 485mV source of ±10 Vdc.
Connect the analog input to the ADC, by connecting the positive input to the voltage divider and the negative input to the ground. Connect the center conductor of the channel to connect the ADC to the display box.

To make the system functional, the four 4-bit ADC's must be operational:

Connect the 0-5 Vdc variable power supply to the input of the ADC (pins 17 and 18).

Connect the DMM to the output of the ADC (pins 15 and 16),

Turn on the power supply and press the power button.
Adjust the LED display to match the variable power supply to ±2.400 V dc, then the LED display indicates the following:

V7, V8, V9, V6 should be:
the 4-bit ADC used with these four LEDs.

Diode the DMM, adjust the variable power supply to ±2.400 V dc. The LED display should indicate the following:

V5, V11, V12, V1 should be off,

since the 4-bit ADC is not a bipolar unit, only the 1-REF selection is available.
Turn off power supply and remove all test equipment.

DI: 4-BIT ADC DC BUS ADJUSTMENT (Refer to Figure E7)

The purpose of this procedure is to ensure the output of the 4-bit ADC is and that the manual attenuators begin switching in at the correct power level. If the switching is too late, the input to the 8-bit ADCs may saturate, causing errors.
Additional Equipment:

- Waveguide attenuator (FNR 500, N = N2 - 43, 47-100 W/cm²)
- Waveguide phase shifter (HP MD 26, N = N2 - 44, 0-100 W/cm²)
- Waveguide sections as required
- Power meter and test set equipment

1. Configure the ADC deck for normal operation as follows:
   a. Remove all test equipment.
   b. Reinstall all cables and modules as per previous procedures.
   c. Ensure that all SAMPLE, procedure, and modules are correctly installed.
   d. Ensure that all ADC controls are set correctly for any DUT to be tested.
      (Refer to Table D2 for reference)
   e. Ensure that all waveguide cables are properly connected.

2. Configure the transmitter for test conditions as follows:
   a. Ensure both waveguide sections are set for 0.5 W/cm² (50 dB test position).
   b. At the rear of the transmitter, check that all the components are properly installed and that all waveguide sections are in place as shown in Figure D7.
   c. Disconnect the diode switch driver module and switch the internal switches into the transmitter (this will be a function of the CW model).
   d. Set the 0-50 dB variable attenuator to 0 dB. Set the power meter attenuator for 50 dB.

3. Turn power on for both transmitter and receiver.

   a. Adjust the transmitter front panel attenuator and receiver attenuator (0 dB) as obtained on the power meter. Note the setting. If 0 dB cannot be obtained, note the power meter and the receiver attenuation. This will be used at a later time to set the final power level.

Figure D7: 3-bit ADC Base Arrangement
Table D2: ADC to Digital Hardware Backplane Cable Test (For Figure 42 for card locations)

<table>
<thead>
<tr>
<th>From</th>
<th>To</th>
</tr>
</thead>
<tbody>
<tr>
<td>4-byte ADC</td>
<td>Receiver Hardware</td>
</tr>
<tr>
<td>8-pin B pins 1-18</td>
<td>Card 2 8-pin B pins 1-18</td>
</tr>
<tr>
<td>8-byte ADC</td>
<td></td>
</tr>
<tr>
<td>B1 1 8-pin B pins 1-18</td>
<td>Card 4 8-pin A pins 1-18</td>
</tr>
<tr>
<td>B1 2 8-pin B pins 1-18</td>
<td>Card 5 8-pin A pins 1-18</td>
</tr>
<tr>
<td>B1 3 8-pin B pins 1-18</td>
<td>Card 6 8-pin A pins 1-18</td>
</tr>
<tr>
<td>B1 4 8-pin B pins 1-18</td>
<td>Card 7 8-pin A pins 1-18</td>
</tr>
<tr>
<td>B1 5 8-pin B pins 1-18</td>
<td>Card 8 8-pin A pins 1-18</td>
</tr>
<tr>
<td>B1 6 8-pin B pins 1-18</td>
<td>Card 9 8-pin A pins 1-18</td>
</tr>
<tr>
<td>B1 7 8-pin B pins 1-18</td>
<td>Card 10 8-pin A pins 1-18</td>
</tr>
<tr>
<td>B1 8 8-pin B pins 1-18</td>
<td>Card 11 8-pin A pins 1-18</td>
</tr>
</tbody>
</table>

b. Adjust the 0-50 dB attenuator for 50 dB, and the front panel attenuator for maximum attenuation.

c. Connect the oscilloscope to the IF front end as follows (refer to Figure E1(a) for location IF front end):
1. Trace A to the 6-dB test point,
2. Trace B to the 12-dB test point,

d. Determine the front panel attenuator setting to be used (refer to the power reading/attenuator setting taken in step 3).
   1. If a 1 mW power level was obtained, add the attenuator setting to -75 dB to obtain the proper value, that is, attenuation setting
      
      \[
P = \begin{cases} 
      2.7 \text{dB} & \text{if } P = 4 \text{ dB} \\
      7.7 \text{dB} & \text{if } P = 12 \text{ dB} 
      \end{cases}
      \]
      
      will be the total system attenuation to be inserted via the 0-50 dB attenuator and the front panel attenuator 0-50 dB set to 50 dB and the front panel attenuator set to 27.5 dB.
      
      2. If a 1 mW power level was not obtained, take the power level reading of front panel attenuation and subtract the total attenuation desired, that is, power level reading at 3.2 dBm,

      \[
      \begin{align*}
      &\frac{7.8 dB}{71.8 dB} \\
      &\text{total system attenuation.}
      \end{align*}
      \]
After the proper attenuation level has been set, monitor the scope and adjust the 4-bit ADC dc bias pot (refer to Figure D8) until the 6-dB IF attenuator line just begins to switch in. Ensure the manual attenuation control on the radar control panel is in the OFF position (refer to Figures C1 and E1(a) for location.

Note: By monitoring the 6- and 12-dB IF attenuator lines, you ensure that 6 dB is being calibrated.

D4. DYNAMIC SYSTEM CHECK (Refer to Figure D8)

The dynamic test for each of the I and Q sides is the same for all eight channels. Channel IQ will be described because of its accessibility.

![Figure D8. ADC Dynamic Test Configuration](image)

Configure the transmitter for TEST, as outlined in Section D3, 2.

Remove the RF input cable from the channel 1 RF input port, and connect this cable end to an eight-way power divider (MEHRR/MAC PDM-82).

Using eight coax cables (36-in. long), connect each of the output ports of the power divider to the input ports on the RF front end.

In order to determine the saturation point for each of the ADC unit, it may be necessary to measure the total insertion loss of the eight-way power divider from the input cable to each of the eight output cables. This may be accomplished using a network analyzer. (Note: Although use of the eight-way power divider is not mandatory, it does shorten the time required to perform checks on all 16 ADCs.)
If not used, the RF input cable must be moved from channel to channel as the test progresses.

1. On the radar control panel, set the IF attenuator switches to 0 dB of manual IF attenuation.

2. Connect an oscilloscope to the sample and hold input pins (10 and 11) of the ADC under test. (Refer to Figure D3(a) for pin locations.)

3. Set the 0-50 dB attenuator to 50 dB of attenuation. Set the transmitter front panel attenuator to 20 dB of attenuation.

4. Turn rack power on (both transmitter and receiver racks). Allow 10 min for warmup. Turn the transmitter "offset oscillator" on.

5. Monitoring the scope, adjust the 0-360° phase shifter for the largest possible signal level (with respect to ground).

6. Set the transmitter front panel attenuator to 50 dB. On the radar control panel, set the auto/manual IF attenuator switch to AUTO.

7. Monitoring the oscilloscope, begin decreasing the amount of attenuation via the front panel attenuator. At the point where the 6 dB IF attenuator begins to switch in, measure the signal level (reading should be less than 1.25 V peak).

8. Continue decreasing the attenuation on the front panel attenuator. As each succeeding IF attenuator begins switching in, note the signal level.

(Note: The full range of 42 dB of IF attenuation may be achieved in the following manner:

After the front panel attenuator has been adjusted through its full range (to 0 dB), set the front panel attenuator to 50 dB and the 0-50 dB attenuator to 0 dB of attenuation. This will give the system another 50 dB of signal range.)

9. Repeat steps 1. through 7. for the remaining 15 A/D sides.

At no point should the signal level measured in the automatic mode of IF attenuation exceed 1.25 V peak. If the signal level does exceed this amount, the ADCs may saturate.

10. If the signal level for any one of the 16 ADCs exceeds 1.25 V peak, repeat the 4-bit ADC bias adjustment procedure outlined in Section D3.)
Appendix E
System Component Identification

The purpose of Appendix E is to identify the equipment located in each of the four equipment racks that make up the NFMRADI (AMTI) systems. As each level is identified within an equipment rack, reference will be made to the chapter and/or other documentation for functional descriptions.

E.1. RECEIVER (Refer to Figure E1(a))

- **Level 1**: Receiver IF
  - Ref. Section 3.2.4
- **Level 2**: Radar Control Panel
  - Ref. Appendix C
- **Level 3**: Receiver Digital Hardware
  - Ref. 4.1, 4.4 through 4.7
- **Level 4**: Analog to Digital Converter (ADC) Deck
  - Ref. 4.2, 4.3, Appendix D
- **Levels 5 and 6**: Power Supplies
  - None
level 7

    Rack Blower

    none

E2. TRANSMITTER (Refer to Figure E1(b))

level 1

    Delay Lines

    Ref. 3, 2, 3

level 2

    Receiver Front End

    Ref. 4, 2, 2

    Transmit/Receive Array

    Ref. 2, 2

level 3

    Transmitter

    Ref. 3, 4

    a. RF Sources

    Ref. 3, 4.4

    b. Pulse Generation

    Ref. 3, 4.3

    c. Pulse Generation

level 4

    Power Supplies

    none

level 5

    Traveling Wave Tube Amplifier

    (TWA)

    Ref. 3, 4.4, also Instruction and

    Maintenance Manual for Instrument-

    ation Traveling Wave Tube Amplifier

    Model No. 1277 H07 TWA, Hughes

    Aircraft Co., Electron Dynamics

    Division, 5100 W. Fanita Blvd.,

    Fountain, CA 93010

level 6

    Rack Blower

    none

E3. CSP 30 MINICOMPUTER (Refer to Figure F2(a))

level 1

    Tri-Data Cartridge Unit

    Ref. Cartridge 20 Instruction Manual,

    Tri-Data, 800 Marine Ave., Mountain

    View, CA 94040

100
level 2

Control panel

Ref, CSPI Operation and Maintenance Manual, Doc. No. JP3000-000-04, CSPI, 40 Linnet Crt., Billerica, MA 01821

level 3

Upper Mainframe Cooling Fans

Ref, none

level 4

Mainframe Card Nest

Ref, see level 2 above

level 5

Lower Mainframe Cooling Fans

Ref, none

level 6

Power Supplies

Ref, see level 2 above

54. MAP ARRAY PROCESSOR (Refer to Figure 2(h))

level 1

Display Scope

Ref, Tektronix Operation and Maintenance Manual, Type 611 Mod. 1620, Tektronix, Inc., P.O. Box 500, Beaverton, OR 97075,

level 2

Hard Copy Unit

Tektronix Mod. 4601

Ref, Tektronix Operation and Maintenance Manual, 4601.

level 3

MAP Monitor Panel

Ref, Installation and Operation Booklet No. AS7130-000-01,

level 4

Expansion Power Supply

Ref, Operation Maintenance Manual AT-0000-004-PR1-4,

level 5

MAP-300 Arithmetic Processor

MAP Programmer's Reference Manual
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