GEORGIA INSTITUTE OF TECHNOLOGY
Engineering Experiment Station

PROJECT INITIATION

Date: March 8, 1973

Project Title: Broadband Antenna Measurement Techniques

Project No.: A-1517

Project Director: Mr. F. L. Cain

Sponsor: Rome Air Development Center; Griffiss AFB, N. Y.

Effective: 2/16/73 Estimated to run until: 2/15/74

Type Agreement: Contract No. F30002-73-C-0194

Amount: $69,798

Partially funded at $29,798 through 6-30-73.


Sponsor Contact Person: Technical Matters
(Individual Not Named)
Rome Air Development Center
Attn: RADC/OCTG
Griffiss AFB, N. Y. 13441

Contractual Matters
(thru GTRI)
Rome Air Development Center
Procurement Division (PMA)
Griffiss AFB, N. Y. 13441


Assigned to: Radar Division

COPIES TO: Project Director
Assistant Director
GTRI
Division Chief(s)
Service Groups
Patent Coordinator
Photographic Laboratory
Security; Property, Reports Coordinator
EES Accounting
EES Supply Services
Library
Rich Electronics Computer Center
Project File
Other
GEORGIA INSTITUTE OF TECHNOLOGY
Engineering Experiment Station

PROJECT TERMINATION

Date 7-17-74

PROJECT TITLE: Broadband Antenna Measurement Techniques

PROJECT NO: A-1517

PROJECT DIRECTOR: Mr. F. L. Cain

SPONSOR: Rome Air Development Center; Griffiss AFB, N. Y.

TERMINATION EFFECTIVE: 6-19-74 (Approved Final Report Submitted)

CHARGES SHOULD CLEAR ACCOUNTING BY: 6-30-74

Contract Closeout Items Remaining: Final Invoice & Closing Documents
Gov't Property Inventory & Certificate

Radar Division

COPIES TO:
Project Director
Director
Associate Director
Assistant Directors
Division Chief
Branch Head
Accounting
Engineering Design Services

General Office Services
Photographic Laboratory
Purchasing
Report Section
Library
Security
Rich Electronic Computer Center
Rome Air Development Center
Griffiss Air Force Base,
New York 13441

Attention: Mr. Martin Jaeger, OCTS

Reference: Contract No. F30602-73-C-0194
Project No. 4506

Title: Broadband Antenna Measurement Techniques

Subject: Monthly Status Report No. 1

Dear Sir:

A summary of progress on the referenced contract for the period 16 February through 30 March 1973 is contained herein. The effective starting date of the contract is 16 February 1973. The research program has been designated by Georgia Tech as Engineering Experiment Station Project A-1517. The project is under the general supervision of Dr. H. A. Ecker, Chief of the Radar Division, and under the direct supervision of Mr. F. L. Cain, Project Director.

The technical efforts to be included in this research program involve two basic concepts for exploring new techniques to measure, record, and analyze broadband antenna gain performance of short-pulse radars. One concept involves broadband leveled noise as a test signal to provide average antenna gain characteristics, and the second concept, which can be used as a diagnostic tool, involves sweep frequency techniques in which the pattern data are rapidly sampled over a selected range of frequencies. From these efforts the feasibility of developing the necessary instrumentation will be determined. Recommendations of the optimum techniques for measuring the performance of wide signal bandwidth antenna systems will be made.

On 15 March, Mr. F. L. Cain of Georgia Tech visited Mr. Martin Jaeger of RADC to discuss various aspects of the referenced contract. Included in the discussions were the objectives of the program, the approach to be taken, and the outputs required. Both measurements concepts were discussed in detail, and the potential problem areas were outlined. It was mutually agreed that about the same rate of effort initially would be spent on each concept; however, this rate will be subject to change as work progresses.

A variable bandwidth noise scheme is presently being considered in block diagram form under Concept 1. The key components of the transmit instrumentation are the noise generator, a voltage controlled oscillator, a modulator, and
a power amplifier. The key components of the receive instrumentation consist of a detector, averager, and amplifier. The general principle of operation of the complete system is as follows. At the transmit end, the signal from a synchronized wideband contiguously stepped sweeper is modulated through a double balanced mixer with narrow-band noise, and the output signal is amplified and leveled with various feedback circuits. Because the carrier (signal from the sweeper) is suppressed, the bandwidth of the transmitted signal can be adjusted by selecting the proper sweep frequency range of the synchronized wideband sweeper. At the receive end, the signal is detected by a square law device, averaged by frequency synchronized signal averaging circuits, and amplified before feeding to the Scientific-Atlanta 1520 pattern recorder.

This variable bandwidth noise approach appears to offer several advantages.

1. It is versatile and is functionally compatible with both Concepts 1 and 2, with some additions.
2. Broadband noise of precisely controlled characteristics is effectively produced via the mixing technique.
3. The broadband noise apparently can be precisely assessed using square law detection and electronic signal averaging.
4. Virtually all pattern range instrumentation variables are under control of a master clock-tuner-synchronizer that compensates for known device variations on either an open-loop or a closed loop basis.

Other approaches, as well as variations of the one described, will be examined during this program.

An investigation of the characteristics and costs of the equipment that is needed to implement the sweep frequency technique of Concept 2 has been initiated. Initially, a survey is being conducted to determine the off-the-shelf equipment that might be applicable. Most sweep frequency generators currently available operate over octave bandwidths, e.g., 1-2 GHz, 2-4 GHz, etc.; however, one recently advertised generator operates over the complete 1-15 GHz frequency range of interest. The survey indicates that TWT amplifiers will be required with all of the currently available sweep frequency generators to obtain 1 to 10 watts of output power. Conventional TWT amplifiers, like most sweep frequency generators, operate over octave bandwidths.

The network analyzer and harmonic converter were also considered as equipment that possibly could be implemented in Concept 2. In a system using these two components, both the reference signal from the sweep frequency generator and the output of the antenna under test would be fed to the harmonic converter which feeds the network analyzer. The analog phase and amplitude outputs of the network analyzer would then be digitized and fed to the computer. Of course, other
control circuits and calibration information would be necessary for the complete implementation of Concept 2. One weakness of the network analyzer is its low sensitivity, approximately -75 dBm. This low sensitivity requires that high-power transmitters be used on long antenna ranges because otherwise the space attenuation of the signal may be great enough to limit the dynamic range of the received signal to less than -40 dB.

Another potential problem which must be studied further in Concept 2 is the large amount of data which must be sampled in real time. Typical adjustable sweep times for an octave bandwidth range from a minimum of 0.01 second to over 100 seconds. The sampling rate and real time processing capability of available computers will be among the parameters which determine the total measurement time needed to record the wideband response of an antenna.

Next month, work on both Concept 1 and Concept 2 will continue. In Concept 1, a more detailed investigation will be made of presently available hardware. Also, equations for system sensitivity will be developed to determine that an adequate signal for processing is available and that proper averaging is performed to yield acceptable results. In Concept 2, the compatibility of available individual component characteristics with the overall system requirements will be studied. Also, it is anticipated that an investigation to study the sensitivity of an antenna response to phase information will be initiated.

During this reporting period, it is estimated that 136 professional man-hours were expended and that approximately 5% of the technical effort has been completed. These estimates also are cumulative because this reporting period is the first.

Respectfully submitted,

Fred L. Cain
Project Director

FLC:jf

Approved:

H. A. Ecker
Chief, Radar Division
Rome Air Development Center
Griffiss Air Force Base,
New York 13441

Attention: Mr. Martin Jaeger, OCTS

Reference: Contract No. F30602-73-C-0194
Project No. 4506

Title: Broadband Antenna Measurement Techniques

Subject: Monthly Status Report No. 2

Dear Sir:

A summary of progress on the referenced contract for the period 1 April through 30 April is contained herein. Efforts were devoted to both Concept 1 and Concept 2. Efforts of Concept 1 involved (1) development of equations to investigate system sensitivity for the approach reported last month and the availability of components that will satisfy the sensitivity predictions, and (2) possible new approaches for achieving a noise source technique. Efforts of Concept 2 involved (1) a survey of off-the-shelf instruments available for the sweep-frequency technique, and (2) the initiation of a simplified study for investigating sensitivity of an averaged antenna pattern over a specified bandwidth, with and without phase information preserved.

The receiving portion of the Concept 1 variable bandwidth noise approach presented last month has been examined in greater detail. Suitable equations have been developed relating transmitted power and bandwidth to the receiver sensitivity and bandwidth. Detector-averager-amplifier circuitry in the receiver have been compared for sensitivity and dynamic range. Typically, these circuits have taken the form of radiometer receivers using Dicke modulation. The RF portion may be a crystal video, tuned RF, or superheterodyne RF front end. The front end is followed by a square law crystal diode envelope detector. Averaging is done in tuned and low pass audio amplifiers.

The results of this preliminary examination showed that both the tuned RF (TRF) and the superheterodyne configurations have comparable sensitivity, which is about 15 dB better than that of the crystal video configuration. One advantage of the TRF approach is that it needs no local oscillator, and hence, has fewer parts than a superheterodyne receiver. Sensitivity for a TRF detector-averager-amplifier, using a silicon Schottky diode detector, is
about -110 dBm with a 40-dB dynamic range. This dynamic range is achievable since the sensitivity allows working at low signal levels which are well within the square law region of the diode. Such a detector-averager-amplifier can be assembled in the form of a TRF Dicke type radiometer using off the shelf hardware. Typical components would include the following: an octave bandwidth tunnel diode RF amplifier, a ferrite or diode modulator switch, a silicon Schottky diode detector, and a synchronized low noise lock-in amplifier. The averaging process yields a direct voltage output that is compatible with a Scientific-Atlanta DC amplifier plug-in used in place of their crystal bolometer amplifier in the 1520 pattern recorder.

The recording speed of the pattern recorder enters into the design of the detector-averager-amplifier. The sensitivity of the receivers discussed above improves as the square root of the integration or averaging time. A radiometer integrating the signal for 10 seconds, for example, is about 3 times more sensitive than one integrating the signal for 1 second. To obtain recording speeds near that achievable with conventional operation of a SA 1520 recorder, integration times of 0.1 second for medium speed and 1 second for slow speed would be required. To obtain higher recording speeds means trading off some sensitivity by using an integration time of less than 0.1 second.

This month an approach for amplifying thermally generated broadband noise to usable power levels was also investigated in Concept 1. The necessary noise power, using travelling wave tube (TWT) amplifiers, appears to be obtainable up through about 15 GHz. However, continuous control of the bandwidth and center frequency of a broadband noise spectrum with tunable filters is impractical because tuning of wideband microwave filters is a very tedious process. If a maximum absolute bandwidth rather than a percentage bandwidth is specified, then a scheme for controlling the bandwidth and center frequency using fixed-tuned filters and heterodyne techniques appears possible and is described as follows. A thermal noise source is amplified to an intermediate power level and filtered by a fixed-tuned bandpass filter. This signal is then translated in frequency and filtered through another fixed-tuned filter with the same bandwidth as the previous filter. By adjusting the frequency of the local oscillator, the bandwidth of the noise spectrum can be adjusted from zero to the maximum bandwidth of the band-pass filters. This adjustable bandwidth spectrum is then translated in frequency by a second mixer. Adjustment of this oscillator controls the center frequency of the noise spectrum. The signal is then filtered and amplified to a useful power level. The frequencies of the local oscillators would be selected to minimize the required number of filters. The receiver needed for this system is conventional, except that the pre-detector bandwidth must be at least as broadband as the transmitted spectrum.
Another approach under Concept 1, which is presently being considered because of its simplicity, utilizes a swept frequency transmitter and an integrating receiver. This concept hinges on the two following possibilities: (1) a CW oscillator that is frequency swept as a linear function of time has constant output power per unit bandwidth as a time sequence of frequencies, and (2) the time integral of the response of a linear network to this swept frequency signal is equivalent to the average response of the network to broadband noise, provided of course, that the noise and swept frequency signals cover the same frequency band. The transmitter of this system consists of a sweep generator and an RF modulator driven by a square wave with timing such that the sweep generator retrace occurs during the RF off portion of the modulating square wave. The receiver is similar to the conventional broad RF-bandwidth crystal-video receiver used for antenna pattern recording except that the usual very narrow post detector bandpass filter and envelope detector are replaced by an integrator and a sample and hold circuit, which are synchronized with the transmitter through a communication link. The output from the sample and hold circuit, which is recorded by the pattern recorder, is the average radiation pattern of an antenna over the specified frequency band.

The survey of off-the-shelf instruments that are currently available for a sweep-frequency antenna measurement system under Concept 2 has been completed. Instruments for a phase/amplitude system typically exhibit approximately -80 to -90 dBm sensitivity, whereas instruments for an amplitude system typically exhibit approximately -110 to -120 dBm sensitivity over the frequency range of 1 to 15 GHz. Since typical sweep frequency oscillators have output powers on the order of 20 - 50 milliwatts, amplification is necessary for long antenna range distances. Traveling wave tube (TWT) amplifiers typically cover octave bandwidths; thus, several amplifiers are needed to cover the 1 - 15 GHz frequency range. Therefore, the relatively low sensitivity of phase/amplitude network analyzer receivers coupled with the necessity of high-power travelling wave tube amplifiers would make this sweep-frequency phase/amplitude system relatively more expensive with respect to the amplitude-only system.

Also in work performed under Concept 2 this month, a software simulation of an idealized measurement system has been initiated to investigate the critical parameters which affect broadband antenna measurements. The results for the noise and the sweep-frequency techniques of Concepts 1 and 2 will be compared. The equations which describe the response of a uniformly illuminated rectangular aperture antenna to a broadband leveled noise signal or to an amplitude-only swept frequency signal have been derived. These relationships give the integrated (time average) received power as a function of antenna size, far-field angular direction, and noise (or sweep) bandwidth. An in-house computer program is being modified to use these equations to simulate the composite antenna pattern as a function of noise bandwidth. The calculations are
expected to indicate the necessary bandwidth required to obtain a good approximation to the mean-squared voltage pattern of the antenna over the desired frequency range.

Next month, work on Concepts 1 and 2 will continue. In Concept 1, receiving and transmitting antenna requirements will be examined for compatibility with the sensitivity requirements of the receivers, the system gain variations as a function of frequency, and specified far-field distances. In addition, the effects of interference and multipath on the performance of a radiometer receiver will be assessed. In Concept 2, a computer simulation of the phase/amplitude sweep-frequency system will be developed. The noise and sweep frequency system simulations will then be used to compare the measurement techniques.

During this reporting period, it is estimated that 440 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 576 and that approximately 18% of the technical effort has been completed.

Respectfully submitted,

Fred L. Cain
Project Director

PLC:sp

Approved:

H. A. Ecker
Chief, Radar Division
Rome Air Development Center  
Griffiss Air Force Base, 
New York 13441

Attention: Mr. Martin Jaeger, OCTS

Reference: Contract No. F30602-73-C-0194  
Project No. 4506

Title: "Broadband Antenna Measurement Techniques"

Subject: Monthly Status Report No. 3

Dear Sir:

A summary of progress on the referenced contract for the period 1 May through 31 May is contained herein. Efforts related to both Concept 1 and Concept 2 were continued. Under Concept 1, the evaluation of the "synthetic noise" approach and the investigation of component availability for the controllable bandwidth thermal noise approach continued. Under Concept 2, efforts included (1) initiation of an exploratory investigation into techniques for limiting the effects of multipath and external interference and (2) continuation of the development of a simplified computer simulation for predicting typical responses of selected antenna types. On 31 May, a meeting was held at Georgia Tech to brief the sponsor on technical progress to date on the contract and to discuss plans for future activities.

In Monthly Status Report No. 2, a controllable bandwidth, controllable center frequency, broadband thermal noise approach was described for Concept 1. A survey has been made to determine if this approach could be implemented using commercially available components. Thermal noise generator power output is on the order of -50 dBm for a 1 GHz bandwidth and therefore will require several stages of amplification (more than 100 dB) to overcome system losses and produce adequate power output. Since microwave components typically operate over octave or waveguide bandwidths, coverage of the frequency range of interest (1-15 GHz) will require several sets of amplifiers, filters, and mixers. To obtain equal noise power across the frequency range, equalizers will be required to compensate for TWT gain variations with frequency. Also an isolator will be required in each octave or waveguide set to prevent power reflecting from the filters back into the TWT. In addition, to precisely control bandwidth and center frequency, a stable local oscillator will be required for each mixer because the mixing in certain stages of the system is performed at relatively high frequencies. Thus, a large collection of microwave components is required to build this system; however, all
of these components appear to be within state-of-the-art capability. Further, this system is quite complex and costly, as well as lacking some desired flexibility.

Also under Concept 1, investigations into the feasibility of the frequency-sweep/time-integration scheme for producing results equivalent to broadband noise were continued. To maintain present antenna pattern recording speed and thus prohibit undesired increases in data recording time, a video bandwidth of several hundred Hz is required. This data recording speed becomes especially important with the increased data which must be recorded for broadband analyses. The recording speed requirement prohibits the use of slow response thermal-type square law detectors. Semiconductor diode type square law detectors compared to thermal devices have about two orders of magnitude more sensitivity and several orders of magnitude faster response, but have the serious disadvantage of insufficient dynamic range. However, proven techniques are currently being used to extend the dynamic range by processing of the video signal. Instrumentation of this system appears feasible with available devices.

Under Concept 2, an examination of techniques to control or reduce measurement errors caused by multipath and by external interference has been initiated. These techniques apply only to reducing effects on the antenna under test. A subsequent investigation will be conducted to explore the potential interference problem to other equipment in nearby areas. The control of multipath and interference effects to the antenna under test is desirable to obtain increased accuracy when measuring such parameters as gain, beamwidth, polarization, far-out sidelobes, and null depth. Effective control of multipath and interference effects will allow full time utilization of measuring facilities and will allow measurements over a dynamic range limited only by the sensitivity of the receiving system electronics.

An effective technique that provides discrimination against both multipath and interference effects in a spatial manner involves the use of an additional reference receiving antenna and additional receiving electronics. This additional reference antenna should have very low side lobes over angular sectors in which interference and multipath sources are significant. Suppression of these undesired signals is provided in a signal processor that multiplies and correlates the signal from the antenna under test with a signal from the reference antenna. The processing has the desired effect of suppressing received signals in all regions where the response of the reference antenna is low.

The use of a rejection filter can provide an effective technique for interference discrimination against out-of-band frequencies. Various high,
low, and band pass filters can provide this interference rejection. By using a combination of both the spatial and frequency interference rejection using the above techniques, adequate rejection appears possible without placing extreme requirements on any one component. Several receiver circuit techniques are available for providing additional interference rejection, if required.

A computer simulation for predicting the antenna patterns which would be obtained using the sweep frequency amplitude-only and the sweep frequency amplitude-plus-phase techniques is presently being developed. The necessary analytical expressions have been developed and programmed and some results have been obtained for the amplitude-only case. These results include the antenna pattern of a uniformly illuminated rectangular aperture with a leveled noise excitation of various bandwidths. These simulations will be continued and expanded to include the amplitude-plus-phase case.

On 31 May, Mr. Martin Jaeger visited the Engineering Experiment Station at Georgia Tech for a briefing and technical discussions on the referenced contract. The briefing included a detailed summary of the status of the various potential broadband antenna measurement techniques that have been conceived. Following the briefing, discussions concerned with preliminary evaluations of the approaches and techniques were held to reduce the number of techniques that will be considered in future efforts; however, it was decided that a detailed comparison of the advantages and disadvantages of the various techniques should be made before reaching a final decision. Other areas of interest included (1) a discussion of the possible design goals and considerations of the radar system designer with respect to the expected wide bandwidth antenna performance and (2) the potential electromagnetic interference to nearby installations due to the transmission of swept signals over wide bandwidths. These areas of concern will be addressed in future efforts.

On 1 June, Mr. J. D. Adams will join the team working on the referenced contract. Mr. Adams, who has a wide technical background, will be a valuable asset to the team and will make a major contribution to the program.

Effort next month will concentrate on comparing advantages and disadvantages of the various techniques that have been investigated. This comparison will be used in concentrating future efforts on the most cost-effective techniques for broadband measurement. Effort also will be directed toward a preliminary analysis of the Electromagnetic Interference (EMI) which might be caused by the test pulse to other equipment in the vicinity of the test site. The extent of the EMI problem will depend on the installation and in some cases might dictate the form of the transmitted signal.
During this reporting period it is estimated that 342 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 918 and that approximately 29% of the technical effort has been completed.

Respectfully submitted,

Fred L. Cain
Project Director

FLC:bp

Approved:

H. A. Ecker
Chief, Radar Division
A summary of progress on the referenced contract for the period 1 June through 30 June is contained herein. Efforts related to both Concept 1 and Concept 2 were continued. Under Concept 1, broadband thermal noise, a preliminary analysis of interference effects due to radiation of the broadband noise was completed. Under both Concepts 1 and 2, several different implementations previously have been synthesized and examined in block diagram form for feasibility and component availability. Effort during this reporting period has been concentrated on generating more detailed data for comparison of the various approaches. Before selection and recommendation of systems for continued analyses, additional guidance from RADC is desired. A list of items for further discussion with RADC has been generated.

Under Concept 1, an analysis was conducted to determine the magnitude of the interference which might be caused by radiation of the broadband noise test signal. In this analysis noise power spectral density, due to the radiated test signal, was calculated as a function of range with bandwidth, center frequency, transmitted power, and antenna gains as parameters. From this analysis, which was based on free-space transmission, it was concluded that by proper control of the radiated power level, interference can be limited so that it will not be a severe problem to receiving sets of typical sensitivities unless they are within a few kilometers of the test site and they have main lobe antenna coupling to the transmitting antenna. Results of the analysis can be extrapolated to predict possible interference levels at specific test sites with known features.

Under both Concepts 1 and 2, a number of measurement systems, which appear feasible with available components, have been synthesized at the
block diagram level. These systems include simple octave and multiple octave noise techniques, bandwidth adjustable noise techniques, amplitude only swept frequency schemes, and computer controlled phase-plus-amplitude swept frequency schemes. Since a detailed analysis of each of these systems is not appropriate, selections must be made for a more detailed comparison. During this reporting period, effort has concentrated on producing general descriptions of the candidate systems with tabulations of their measurement capabilities, general equipment requirements, block diagrams, and their good and bad features. This effort is approximately 75% complete. At the conclusion of this effort, systems will be selected for detailed comparison of capabilities, equipment requirements and availability, and cost and complexity. An investigation of data link requirements has shown that ordinary telephone lines will be adequate for most applications.

During the past few months of the study and analysis program, several items have arisen on which clear-cut decisions need to be quickly made. In order to assure that the intended objectives of RADC are met, it is necessary that several items be mutually discussed. Accordingly, a list of these items for discussion has been generated and will be submitted to RADC, Code OCTS. It is anticipated that these items will be discussed between RADC and Georgia Tech in the near future.

During the coming month, generation of the overall comparison data between the various systems will be completed and tentative selection of systems for more detailed comparison will be made. Task for analyses by Scientific Atlanta will be defined and scheduled for discussion with Scientific Atlanta. In addition, simulation of the phase/amplitude swept-frequency concept will be continued.

During this reporting period it is estimated that 313 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 1231 and that approximately 38% of the technical effort has been completed.

Respectfully submitted,

Fred L. Cain
Project Director

FLC:am

Approved:

Fred L. Cain
Project Director

for H. A. Ecker

H. A. Ecker
Chief, Radar Division
Rome Air Development Center
Griffiss Air Force Base,
New York 13441

Attention: Mr. Martin Jaeger, OCTS

Reference: Contract No. F30602-73-C-0194
Project No. 4506

Title: "Broadband Antenna Measurement Techniques"

Subject: Monthly Status Report No. 5

Dear Sir:

A summary of progress on the referenced contract for the period 1 July through 31 July is contained herein. Efforts related to both Concept 1 and Concept 2 were continued. General descriptions of all candidate systems with tabulations of measurement capabilities, general equipment requirements, and good and bad features were completed. More detailed investigation of equipment requirements for these systems is continuing. Under Concept 2, computer simulation of a phase-plus-amplitude system was continued and preliminary results were obtained. A meeting was held with Scientific Atlanta technical personnel and topics for their initial investigation were defined.

Part of the effort on this project consists of specialized support services from Scientific Atlanta, Inc. Definition of candidate systems has progressed to the point that Scientific Atlanta's expertise can now be applied in the most efficient manner. On 20 July, Mr. Searcy Hollis of Scientific Atlanta met with Georgia Tech project personnel to discuss a group of items for Scientific Atlanta's initial efforts. These efforts are directed toward investigation of means to satisfy instrumentation requirements of a phase/amplitude swept frequency system. Possible modification of existing equipment to meet measurement requirements will be examined first. If this approach appears infeasible, development of new equipment with the desired features will be investigated. In addition, requirements for, and means of obtaining a reference signal between transmitting and receiving sites will be investigated.

General descriptions of all candidate systems which have been examined under both Concepts 1 and 2 have been completed. As stated in Monthly
Status Report No. 4, these descriptions include a discussion of general equipment requirements and measurement capabilities, block diagrams, and a summary of good and bad features. Further efforts are being devoted to define these systems in more complete hardware detail by examination of equipment requirements and by investigations of satisfactory component availability. In most of the systems which are under consideration, availability of only one or two required components are likely to be future problems.

A number of the broadband measurement systems which are being examined on this program require a data link between the transmitting and receiving sites for synchronization and/or remote control. Preliminary estimates of data rate requirements indicate that ordinary telephone lines can meet these requirements and provide considerable margin for growth. Required data rates are expected to be between 300 and 600 bits/second while 1200 bits/second can be achieved with ordinary telephone lines and relatively simple modulation equipment.

Under Concept 2, a preliminary investigation is being conducted to identify and estimate error sources when making phase measurements. Results of this investigation indicate that transmission-line-induced phase errors in phase comparison circuits can be held to within $\pm 10^\circ$, up to 18 GHz. Also under Concept 2, the phase/amplitude swept frequency simulation studies were continued. Several simulations for a typical aperture antenna operating in X-band have been performed. These simulations indicate that the distortion effects on the transmitted pulse which are caused by differential phase shift in the antenna transmission line are more severe than distortion caused by gain variations. These distortion effects are also more severe as the bandwidth of the incident pulse signal is increased since the differential phase shift of the transmission line is greater for wider-band signals. This type of simulation, coupled with amplitude and phase sweep frequency measurements, would be useful in predicting performance of an antenna for pulse radar systems.

During the coming month, investigations of equipment requirements and component availability for the candidate systems will continue. The phase/amplitude swept frequency simulation studies will continue and initial results from Scientific Atlanta's investigations are expected. A visit by Georgia Tech personnel to RADC is anticipated during the later part of August. During this visit, Georgia Tech will brief RADC on project work-to-date, make recommendations on systems for more detailed work on the present project, and discuss recommended additional efforts as a follow-on to the present program.
During this reporting period it is estimated that 304 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 1535 and that approximately 47% of the technical effort has been completed.

Respectfully submitted,

Joseph D. Adams
Project Director

JDA:am

Approved:

F. L. Cain
Technical Area Manager,
Electromagnetic Effectiveness
A summary of progress on the referenced contract for the period 1 August through 31 August is contained herein. Efforts during this reporting period were primarily directed toward Concept 2 Systems. Scientific Atlanta examined the feasibility of modifying the S-A Series 1750 receiver for wideband swept frequency phase/amplitude measurements. Investigation into the applicability of group velocity measurements for Concept 2 was begun. On 30 August a meeting was held at RADC to brief the RADC technical monitor and other RADC personnel on contract status and to discuss plans for the remainder of this contract period.

On 15 August, Georgia Tech project personnel met with Mr. Searcy Hollis and Dr. Larry Clayton of Scientific Atlanta to discuss initial findings of their supporting investigations. These initial investigations were directed toward the problem of developing a swept frequency phase/amplitude receiver with satisfactory tracking rate and detection sensitivity for Concept 2 Systems. Scientific Atlanta reports that their Series 1750 receiver can be modified to incorporate rapid, phase-locked, frequency tracking over the frequency range of 1 - 18 GHz. A frequency tracking rate up to 10 GHz/sec at all frequencies within the 1 - 18 GHz range can be provided. Required changes in the Series 1750 mixer circuitry would result in the modified receiver having approximately 10 dB less sensitivity than that of the existing Series 1750 receivers in the coherent mode of operation. In addition, any increase above the current (approximately 100 Hz) information bandwidth of the output circuits would cause a corresponding decrease in sensitivity. Based on these projected receiver characteristics, it is concluded that
development of a satisfactory receiver for phase/amplitude swept frequency measurements is feasible. However, analyses to date indicate that one of the more demanding problems in implementing a swept frequency phase/amplitude system will be calibrating and correcting for phase variations in the phase reference channel so that measured relative phase between the test and the reference channel can be correctly interpreted.

During this reporting period, investigation into the application of group velocity measurements for wideband antenna characterization was begun. Group velocity measurements represent another approach toward obtaining the phase/amplitude Concept 2 information. An apparent advantage of the group velocity measurement is that a separate reference channel is not required. However, the RF transmission path from the transmitter to the receiver must be calibrated versus frequency before insertion phase shift of the test antenna can be determined. Group velocity measurements have been used to determine phase dispersion in other applications; further investigations are required to determine the usefulness of this type of measurement in the present wide-band application.

On 30 August, Messrs. F. L. Cain and J. D. Adams of Georgia Tech visited RADC to brief the project technical monitor and other interested RADC personnel on investigations to date and to discuss areas of concentration for the remainder of this contract period. Block diagrams of both Concept 1 and Concept 2 transmitters and receivers and of selected approaches toward Concept 1 and Concept 2 systems were presented along with summary charts of each system's "good" and "bad" features. The added measurement and analysis capabilities of a phase/amplitude Concept 2 system over the Concept 1 system was discussed. These added capabilities make a phase/amplitude swept frequency system very attractive. However, cost and operational complexity of this system would generally be greater than that of a Concept 1 system. It was pointed out that Georgia Tech's proposed "Hybrid" system could be used as a building block toward a full swept frequency phase/amplitude system while at the same time providing equivalent measurement capability to both a wideband noise system and an amplitude-only Concept 2 system. It was agreed that the Hybrid system and Concept 2 techniques would be further investigated during the remainder of this contract period. This includes further investigation into applicability of group velocity measurements. Achieving RADC's long range goal of a comprehensive wideband antenna measurement capability in the most efficient and timely manner was briefly discussed in terms of priority of needs for follow-on efforts.

During the coming month, efforts will be directed toward both Concept 1 and Concept 2 systems. Applicability of group velocity measurements will be further investigated. Investigation of techniques for reference channel calibration in a swept frequency phase/amplitude system will be pursued. A more detailed design of the Hybrid system will be begun.
During this reporting period it is estimated that 363 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 1898 and that approximately 58% of the technical effort has been completed.

Respectfully submitted,

Joseph D. Adams
Project Director

JDA:am

Approved:

F. L. Cain
Technical Area Manager,
Electromagnetic Effectiveness
Air Force Systems Command  
Rome Air Development Center  
Griffiss Air Force Base  
New York 13441

Attention: Mr. Martin Jaeger, OCTS

Reference: Contract No. F30602-73-C-0194

Title: "Broadband Antenna Measurement Techniques"

Subject: Monthly Status Report No. 7

Dear Sir:

A summary of progress on the referenced contract for the period 1 September through 30 September is contained herein. Efforts during this reporting period were directed toward the Hybrid and the group velocity measurement systems. Efforts on the Hybrid system were directed toward defining requirements and investigating techniques for compensation of test system gain variations with frequency. Investigations into the applicability of group velocity measurements for wideband antenna characterization were begun.

It may be recalled that with the Hybrid measurement system, the test antenna response (on receive) is obtained by electronically integrating received power over the time required to sweep the transmitter over the frequency range of interest. In general, the power received depends on transmitter power, transmit antenna gain, space loss, test antenna gain, and receiver response. Therefore, if the integrated power is to reflect only the test antenna's band-pass characteristics, all other amplitude variations must be eliminated, or compensation must be made for them. It may be shown that by using two identical receiver channels in parallel, the transmit antenna gain variation, the space loss variation, and the receiver gain variation can be ideally compensated. The "reference" channel of this dual channel receiver incorporates a receiving antenna which is identical to the transmitting antenna. Effects due to both transmit antenna gain variation and receiver gain variation are eliminated within the degree to which the two receiver channels are matched, by subtracting the reference channel signal from the test channel signal. Space loss compensation requires knowledge of space loss as a function of frequency. An advantage of this receiver compensation approach is that it is not required to know the transmit antenna gain variation with frequency and thus periodic re-calibration of the transmit antenna is not required. The measurement ac-
accuracy which is achievable with this receiver compensation approach will be established as receiver design progresses and component values are defined.

Proper operation of the dual channel compensation receiver requires constant transmitter power versus frequency. For a swept frequency transmitter, as opposed to a noise transmitter, transmitter power leveling may be accomplished in a straightforward manner. Basically, the TWTA (traveling-wave-tube amplifier) output is sampled to derive an error signal for adjusting power level at the TWTA input. Investigation of component availability and comparison of power flatness versus circuit complexity is in progress. This approach to achieving constant transmitter power will allow a readily achievable specification on gain flatness of the TWT amplifiers.

At the 30 August RADC briefing, an example was presented of one method of making group velocity measurements. Since there are several different approached (various modulations and swept carrier) which could be taken for making group velocity measurements, the relative advantages and disadvantages of the various approaches, when applied to wideband antenna measurements, must be determined. On 12 September, a meeting was held with Scientific Atlanta personnel to discuss task assignments for their support in evaluating feasibility of the different group velocity measurement approaches. These supporting investigations are now in progress.

During the coming month, efforts to evaluate circuitry for the Hybrid system will continue. Initial conclusions regarding the different group velocity approaches are expected and general investigations into operating procedures with the Hybrid system will be initiated. This last item will include such things as spatial and frequency sampling requirements.

During this reporting period it is estimated that 214 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 2112 and that approximately 65% of the technical effort has been completed.

Respectfully submitted,

Joseph D. Adams
Project Director

F. L. Cain
Technical Area Manager
Electromagnetic Effectiveness
Dear Sir:

A summary of progress on the referenced contract for the period 1 October through 31 October is contained herein. Efforts during this reporting period were devoted to consideration of operational procedures for use of wideband antenna measurement systems and to continue investigation of phase measurement calibration requirements. On 25 October, a meeting was held at RADC to plan the next phase of effort on a follow-on contract and to discuss task descriptions for that effort.

Recently several questions have been raised regarding use of wideband antenna measurement systems. Some of these questions have been addressed during this reporting period. For applications such as calculating power radiated or received, two types of absolute antenna gain information may be desired. First, conventional CW-type data showing antenna gain at specified frequencies across a wide frequency range may be desired. Second, an "average" antenna gain which indicates antenna response to an instantaneous wideband pulse is desired.

The first type of data can be obtained by substitution measurements at each of the frequencies of interest using a known standard gain antenna. In substitution measurements, received power is first measured with the standard gain antenna substituted for the test antenna and then received power is measured with the test antenna in place. The absolute gain of the test antenna at each frequency is then calculated from the product of
measured test antenna-to-standard gain antenna received power ratio multiplied by gain of the standard gain antenna at that frequency. For these power measurements, a conventional super-hetrodyne receiver, a sweep frequency phase/amplitude receiver, or the Hybrid receiver operated in the CW mode could be used. More than one standard gain antenna may be required to cover the entire frequency range of interest for a particular test antenna.

The second type of information, average antenna gain, may be determined from a series of single frequency measurements, or with one measurement using the Hybrid system if a suitable standard gain antenna is available. Average gain from single frequency measurements is determined simply by taking the arithmetic average of closely spaced individual gain values. Individual values are determined by the substitution method described above. Average gain is determined from a single Hybrid system measurement as follows. First, average gain of a standard antenna is calculated arithmetically from known signal frequency data across the desired bandwidth. Then a substitution measurement is made. In this substitution measurement, the Hybrid system receiver measures total integrated received power across the desired bandwidth for both the standard gain antenna and the test antenna. Average test antenna gain is now calculated from the product of measured test antenna-to-standard gain antenna integrated received power ratio multiplied by calculated average gain of the standard gain antenna.

For sweep-frequency antenna data recording, both spatial and frequency sampling is required. Basically, there are two approaches which should be considered. Either the antenna is continuously and slowly rotated with rapid frequency sweeps at specified time intervals, or the antenna is incrementally stepped to new positions and remains stationary during the frequency sweep. There are no practical frequency or spatial sampling limitations with the step-scanned approach. Since the time required to sweep the transmitter at each antenna position is short relative to the antenna stepping time, the primary practical consideration is one of spatial resolution versus total time required to record and process the data.

For situations in which step-scanning of the antenna is not desirable, compatible continuous antenna scan rate and frequency scan rate are achievable with typical hardware. Assuming a frequency sweep rate of 10 MHz/msec (SA projected maximum track rate of wideband phase/amplitude receiver), frequency sampling every 10 MHz and spatial sampling each 0.1 deg (while recording frequency, angle, amplitude, and phase) is compatible with a ± 90 degree pattern recording time of 20 minutes. A/D conversion and data recording rates implied by the above are readily available.

Investigation of phase reference requirements for wideband phase measurements has continued. In particular, feasibility of using a direct coaxial cable link as the reference channel for phase calibration of an antenna on a short range has been investigated. This phase calibrated antenna could then be used in the reference channel for very long range measurements. Special
phase-compensated coaxial cable with good phase stability is available. On a relatively short range (on the order of 1000 ft), comparable space loss and cable attenuation can be achieved and they are not prohibitive. Thus, this approach to reference antenna phase calibration appears feasible.

On 25 October a meeting was held at RADC to plan for the follow-on effort on the Wideband Antenna Measurements Program and to discuss task descriptions for that effort. These tasks cover three general areas of investigation: (1) development and laboratory verification of Hybrid System, (2) continued analysis, design, and breadboard verification of phase/amplitude system, and (3) investigation into feasibility of predicting wideband, far-field properties of very large reflectors from measured phase/amplitude feed data. Task content in each of these areas was defined during the RADC meeting. The need to expedite contractual arrangements for the follow-on effort was emphasized. Following the meeting, Georgia Tech has generated and supplied to RADC specific task descriptions, including measurement systems specification as appropriate.

During the coming month, design efforts on the Hybrid system and analyses of phase/amplitude systems will continue.

During this reporting period it is estimated that 210 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 2322 and that approximately 71% of the technical effort has been completed.

Respectfully submitted

Joseph D. Adams
Project Director

JDA: jb

Approved:

F. L. Cain
Technical Area Manager
Electromagnetic Effectiveness
Air Force Systems Command  
Rome Air Development Center  
Griffiss Air Force Base  
New York  13441

Attention: Mr. Martin Jaeger, OCTS

Reference: Contract No. F30602-73-C-0194

Title:  "Broadband Antenna Measurement Techniques"

Subject:  Monthly Status Report No. 9

Dear Sir:

A summary of progress on the referenced contract for the period
1 November through 30 November is contained herein. Efforts during this
reporting period were devoted toward analyses of Concept II systems. Material on the proposed NASA propagation experiment was obtained and reviewed, and a general computer controlled phase/amplitude system was examined in greater detail. As a guideline for development of a phase and amplitude measurement system, resolution requirements for phase error measurements were investigated.

Documents related to the proposed NASA atmospheric propagation experiment were obtained and reviewed.¹ ² This experiment, designated the ATS-E Millimeter Wave Propagation Experiment, was designed to determine the propagation characteristics of the earth/spacecraft channels centered in the atmospheric windows at 15.3 GHz and at 31.65 GHz. The measurements were to be made between the ATS-E spacecraft and several ground stations. Since the experiment was designed to measure phase dispersion by a group velocity measurement, the system design has been reviewed and evaluated for applicability to wideband antenna measurements. This system uses a phase modulation


Note: This last reference refers to several other documents on the NASA program.
technique to produce two sidebands which are spaced equidistant from the center. For both the 15.3 GHz and the 31.65 GHz channels, the sideband separation is selectable up to a maximum of 100 MHz. The ground based 15.3 GHz receiver uses a phase-lock loop so that a narrow (≤ 1 kHz) pre-detection bandwidth can be used. Use of a narrow pre-detection bandwidth tends to minimize receiver noise and, consequently, maximize the probability of detecting weak signals from the spacecraft. Over a dynamic range of 25 dB, the absolute power measurement is accurate to within ± 1 dB, and the differential sideband phase measurement is accurate to within ± 2.5 degrees. Although a 25 dB dynamic range is not usually adequate for antenna pattern recording, the amplitude and phase accuracies are approaching those required for antenna measurements. The most serious limitation of the NASA design for our purposes is its 100 MHz maximum phase measurements range at a fixed carrier frequency. Due to amplifier bandwidth requirements, it is apparent that this type of design (fixed carrier frequency) is not realistically extendable to 1 GHz bandwidth at the present time. It may be recalled from previous status reports and discussions that for group velocity measurements, there are essentially two different design approaches. One approach is to fix the carrier frequency and vary the sideband separation, and the other approach is to fix the sideband separation and sweep the carrier. Since greater bandwidth can be obtained with this latter approach, it is more appropriate for very wideband measurements and it is currently under study for wideband antenna measurements.

During the past month, discussions were held between Georgia Tech and Scientific Atlanta personnel concerning design and implementation of a group velocity phase and amplitude measurement system. One fact which has been re-emphasized by these discussions is the need to define the resolution and accuracy required in making antenna phase distortion measurements. Measurement of phase non-linearities as small as 15 degrees was postulated as being realistic, and Georgia Tech has examined the effect of this degree of phase error on a rectangular 1 GHz bandwidth pulse. Under these conditions, a small amount of pulse shape distortion was discernible, but there was negligible effect on pulse rise time. However, since other signal analysis or processing may require more stringent phase linearity specifications, knowledge of current or projected requirements which RADC has identified in this area would be beneficial in developing the most widely applicable system.

The computer-controlled phase/amplitude system has been defined in more detail, particularly in the area of the various interfaces between equipments. Two or three computer and/or recorder interfaces which may require design and development of new hardware have been identified. The special hardware design requirements arise from the need to preserve dynamic range and accuracy throughout the measurement, recording, and display of data. Other hardware interface requirements appear achievable through off-the-shelf type of equipment. Investigations into component and interface requirements are continuing.
As a result of the 25 October meeting at RADC, task descriptions of the follow-on program, as discussed at RADC, have been prepared and submitted. These task descriptions define three general areas of activity: (1) complete design, construct, and checkout Hybrid system, (2) study feasibility of predicting the complete wideband phase and amplitude response of very large secondary radiators from measured phase and amplitude data on the primary radiator, and (3) complete design and breadboard test Critical Concept II system areas.

During the coming month, design efforts and comparison of a group velocity system versus a separate reference channel phase/amplitude system shall continue.

During this reporting period it is estimated that 184 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 2506 and that approximately 83% of the technical effort has been completed.

Respectfully submitted,

Joseph D. Adams
Project Director

JDA:lb

Approved:

F. L. Cain
Technical Area Manager
Electromagnetic Effectiveness
Air Force Systems Command  
Rome Air Development Center  
Griffiss Air Force Base,  
New York 13441

Attention: Mr. Martin Jaeger, OCTS

Reference: Contract No. F30602-73-C-0194

Title: "Broadband Antenna Measurement Techniques"

Subject: Monthly Status Report No. 10

Dear Sir:

A summary of progress on the referenced contract for the period 1 December through 31 December is contained herein. Efforts during this reporting period were devoted to further design of the Hybrid system and to continuing the investigation and comparison of phase/amplitude measurement systems.

It may be recalled (see Monthly Status Report No. 7) that part of the Hybrid design concept is to compensate for amplitude variations in the measurement system. Specifically, any variations in the transmit antenna gain, space loss, receiver response, and transmitter power must be accounted for to determine the response of the test antenna. One approach for accomplishing this compensation is a design based on a constant-power transmitter and a dual channel receiver which would compensate for variations in the transmit antenna gain, space loss, and receiver response. Since this dual channel approach requires duplication of relatively expensive RF equipment, a single channel receiver which is switched on alternate transmitter sweeps between the test antenna and a reference antenna has been considered. Practical advantages of the single channel receiver over the dual channel receiver include elimination of duplicate RF equipment (which must be matched over the frequency measurement range) and reduction of synchronization requirements between the transmitter sweep rate and the compensation circuitry. With the single channel receiver, only the beginning and end of transmitter sweep must be provided to the receiver.

With the exception of the compensation circuitry, preliminary design of the Hybrid single channel system is nearly complete. For the compensation circuitry, a detailed block diagram with functional specifications on the required components has been generated. Availability of suitable components meeting these required specifications is now being investigated. Except
for possibly one or two components such as a sample-and-hold circuit and the integrator, it appears that these requirements can be met with standard components. Of course, special automatic timing and control circuitry will be required.

Investigations of the applicability of a group velocity measurement system to wideband antenna measurements have continued. A preliminary block diagram of a system which is based on an FM transmitter and a sideband differential-phase processor has been generated. Basically, this system would transmit an upper and lower sideband centered about a carrier which could be swept across the desired measurement range. The receiver would separate the two sidebands and determine their relative phase. From this measured sideband phase data as a function of carrier center frequency, phase characteristics of the test antenna can be determined. The feasibility of implementing this type of system and the definition of the system's expected measurement capabilities are now under investigation.

As part of the continuing tradeoff studies between a group velocity system and a separate reference channel swept-CW system, calibration procedures with the two systems are being compared. With a separate reference channel system, the calibration problem reduces essentially to determining the phase characteristics of the reference antenna. Other items which must be characterized (transmission links from the test and the reference antennas to the receiver mixers) can be calibrated under laboratory conditions. For small reference antennas, characterization of the reference antenna with the measurement system itself is practical. This is true since a short range can be used, and consequently the reference signal for calibration can be transmitted via, for example, a coaxial cable which has been laboratory tested. Other methods of characterizing reference antennas are also under investigation.

With a group velocity system, sequential measurements allow calibration of the complete measurement system under the assumptions that two electrically identical antennas are available for calibration measurements and that atmospheric phase dispersion is zero. Sequential measurements for calibration consists of making phase and amplitude measurements on increasing portions of the test system. A practical approach appears to be to first connect the transmitter directly to the receiver (no antennas) in the laboratory and make calibration measurements under these conditions. Next, the test system is assembled on the antenna range with two electrically identical antennas, one of which will become the transmit antenna for future measurements. The difference between calibration measurements made on the antenna range and those previous ones made in the laboratory are due to the combined effect of the two identical antennas and the atmosphere. Two electrically identical antennas are required since one-half of the difference in calibration data must be assigned to each of the two antennas. If two un-calibrated antennas were used with this sequential calibration approach, it would not be possible to separate the final calibration data measured on the antenna range into a
part due to the atmosphere and a part due to the antennas. Any atmospheric effect must be neglected under these circumstances. If two antennas whose phase properties have been determined are available, atmospheric dispersion at the measurement site may be included in calibration of the group velocity system. Further comparisons and evaluations of the operational advantages and disadvantages of the two systems are continuing.

During the coming month, design efforts on the Hybrid system and comparison of a group velocity system versus a separate reference channel phase/amplitude system shall continue.

During this reporting period it is estimated that 216 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 2702, and that approximately 88% of the technical effort has been completed.

Respectfully submitted,

Joseph D. Adams
Project Director

JDA: jb

Approved:

Fred L. Cain
Technical Area Manager
Electromagnetic Effectiveness
Air Force Systems Command  
Rome Air Development Center  
Griffiss Air Force Base,  
New York 13441

Attention: Mr. Martin Jaegar, OCTS

Reference: Contract No. F30602-73-C-0194

Title: "Broadband Antenna Measurement Techniques"

Subject: Monthly Status Report No. 11

Dear Sir:

A summary of progress on the referenced contract for the period 1 January through 31 January is contained herein. Efforts during this reporting period were devoted to further design of the compensation circuitry in the Hybrid system, to analyses of reference antenna phase calibration methods for phase/amplitude systems, and to an investigation to identify methods which might be feasible for predicting wideband far-field parameters of large reflectors from measured phase/amplitude feed data.

The single-channel amplitude-compensated Hybrid receiver design concept was discussed in Monthly Status Report No. 10. Design of circuitry for compensation of variations in transmit antenna gain, space loss, and receiver response has continued. This compensation achieves the effect of constant power density at the receiving antenna aperture. Modules, or module designs, for all the required functions have been tentatively selected, and an error analysis which is currently in progress has shown that the selected compensation approach has a beneficial effect on measurement accuracy. It may be recalled that compensation is effected through electronically deriving the ratio of integrated power (power versus frequency) received through the test antenna to integrated power received through a reference antenna. A consequence of taking this ratio is that any system non-linearities tend to cancel, and thus measurement accuracy may be improved if the reference antenna is accurately characterized. Design approaches for generating the required timing and control signals also are being investigated.
With the separate reference channel phase/amplitude system, a reference receiving antenna is required to obtain the phase reference signal. This reference antenna must be phase calibrated to interpret test antenna data correctly. Calibration of relatively small reference antennas by a direct phase comparison measurement has been discussed in previous status reports. Use of the phase/amplitude system itself to make these phase comparison calibration measurements is possible if a direct transmission link (e.g., coaxial cable) is used to provide the calibration reference signal. During this reporting period, other reference antenna calibration techniques have been considered. These other techniques include the use of measured antenna impedance and S-parameters to calculate the complex (phase and amplitude) gain function and the use of measured antenna backscatter data to calculate the complex gain function. The antenna impedance and S-parameter approach consists of wideband measurements of the antenna input impedance and of the transfer voltage gain between a pair of "identical" antennas which are separated by a known distance. The measured data and the known separation are used to compute individual antenna gain in both amplitude and phase. This method requires a TEM mode transmission link between the antennas and the test instrument. Loss between the device under test and the test instrument leads to degraded S-parameter measurement accuracy. Since TEM transmission lines (coaxial cable) typically exhibit tenths of dB per foot loss, this type of measurement becomes less desirable for larger antennas. The requirement of two electrically identical antennas also becomes an undesirable feature of the approach for large antennas. It, therefore, appears that this approach would only be practically useful for relatively small antennas for which short ranges could be used. In this respect, the antenna impedance/ S-parameter approach has a similar limitation to the direct phase comparison approach. In addition, the S-parameter approach requires special S-parameter test instrumentation.

The free-space scattering method for the determination of an antenna's complex gain function exploits the property that if the antenna terminals are mismatched, the antenna scattered field is proportional to its voltage gain squared. If the amplitude and phase of this scattered field are measured, the complex voltage gain can be determined. A method of measuring the amplitude and phase of the scattered field from the antenna is to first measure the scattered field from a sphere of known size and then to measure the amplitude and phase of the field scattered by the antenna under short-circuit conditions. A potential difficulty of this approach for wideband applications is the cancellation of the background signal over the wide bandwidth. The single frequency background cancellation technique (which utilizes a nulling loop) is not easily adaptable to wideband sweep-frequency measurements. A possible solution is to first record the background signal as a function of frequency and then subtract this recorded background signal from the antenna-plus-background measurement. This recorded background approach has been used at Georgia Tech for single frequency measurements. Experimentation to assess wideband application of the free-space scattering method is desirable.
It has been suggested previously (see Monthly Status Reports Nos. 8 and 9) that wideband large-reflector pattern data may be predicted from measured phase and amplitude data of the antenna's feed system. This predictive approach would be primarily applicable to very large reflectors for which far-field measurements are impractical. For example, with an existing reflector system for which wideband data are desired, it might not be feasible to move that reflector to an antenna range. The feasibility of the predictive approach depends essentially upon the feasibility of implementing a suitable technique for transforming wideband phase and amplitude aperture data to obtain the desired far-field information. During this reporting period, various mathematical formulations for performing the required transformations have been identified and briefly characterized. The complexity of the mathematical formulation increases with the desired accuracy and completeness of the far-field data. Obtaining complete pattern information (including wide-angle sidelobes) is a considerable numerical task for large apertures since sample points spaced across the aperture at less than one-half wavelength are required when using, as is typical, a Fast Fourier Transform algorithm. If only the main beam and a small angular region about the main beam are of interest, the number of sample points can be reduced. If RMS sidelobe information is sufficient, a further reduction in computational complexity can be obtained. Accuracy, resolution, and angular coverage versus data requirements must be further investigated. Main beam and peak sidelobe parameters can be bounded by comparing the actual aperture distribution with aperture distributions which approximate the actual distribution and for which far-field parameters are well known. Further investigation and analyses will be required to quantitatively define the obtainable information versus the numerical complexity of the various prediction approaches.

During the coming month, an error analysis of the Hybrid system and design of Hybrid system timing and control circuitry will continue. Analyses and evaluation of phase/amplitude systems shall continue.

During this reporting period it is estimated that 444 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 3146, and that approximately 94% of the technical effort has been completed.

Respectfully submitted,

Joseph D. Adams
Project Director

JDA:1b
Approved:

Fred L. Cain
Technical Area Manager
Electromagnetic Effectiveness
Air Force Systems Command  
Rome Air Development Center  
Griffiss Air Force Base  
New York 13441

Attention: Mr. Martin Jaeger, OCTS
Reference: Contract No. F30602-73-C-0194
Title: "Broadband Antenna Measurement Techniques"
Subject: Monthly Status Report No. 12

Dear Sir:

A summary of progress on the referenced contract for the period 1 February through 28 February is contained herein. Efforts during this reporting period were devoted to completion of the Hybrid system error analysis, selection of a timing and control technique for the Hybrid system, investigation of wideband transmitting antennas, further definition of the phase/amplitude system, and a survey to determine availability of major components required for wideband antenna measurements outside the 1-15 GHz range which has previously received major emphasis.

The single-channel amplitude-compensated Hybrid receiver design concept has been discussed in Monthly Status Reports Nos. 10 and 11. Recall that compensation is effected through electronically deriving the ratio of integrated power (integrated over frequency) received through the test antenna to integrated power received through a calibrated reference antenna. As implied in Status Report No. 11, it has been determined that a consequence of this compensation method is the improvement of overall system accuracy, if the reference antenna is accurately calibrated. The least accurate element in the Hybrid system is expected to be a dc logarithmic-amplifier module. Consequently, a worst-case estimate has been made for the error contributed by this module. The error in system output (ratio of test antenna to reference antenna response) due to this component is expected to be less than ± 0.5 dB. This worse case error occurs when the response of the two antennas differ by 40 dB. When the response of the test and reference antenna is equal, measurement inaccuracies of the system cancel completely. It must be realized that this error estimate is not based on a complete model of system operation. The system error under various conditions could be obtained by computer modeling of the system. However, a
more logical approach for this system is to obtain accuracy data from laboratory tests on the system when it is constructed.

Proper operation of the Hybrid system requires that the electronic processes for obtaining integrated power and power ratio be synchronized with transmitter sweep cycles. Techniques for obtaining this synchronization have been considered and an approach has been tentatively selected. Five control signals (to integrator, antenna-select switch, and 3 sample-and-hold modules) are required. These five control signals are all derivable from the basic transmitter-cycle signal. Appropriate delays will be required in the individual timing circuits to maintain the proper sequence of operation. It is expected that synchronization can be maintained by driving a Schmitt Trigger circuit with a detected sample of rf at the leading edge of each transmitter sweep cycle. This sample would be obtained from the output of the reference antenna at the receiving site. This type of synchronization would have the advantage of not requiring the transmission of timing signals between the transmitter and receiver sites. This approach is currently under further evaluation to determine if sufficient flexibility to accommodate a variety of sweep rates and bandwidths is practical.

Previous feasibility-of-implementation investigations have concentrated on the 1-15 GHz frequency range. However, the broader frequency range from 200 MHz to 75 GHz is of interest; during this reporting period, efforts have been devoted to assessing the feasibility of implementing broadband antenna measurement systems over this entire frequency range. This investigation has included the major rf equipment required for a hybrid or a phase/amplitude system. Of course, other items of equipment, such as the computer, pattern recorder, analog-to-digital converters, etc., are not sensitive to the measurement frequency, and they can be used across the entire measurement range. Major rf equipment items include signal sources, signal power amplifiers, low noise amplifiers, detectors, and mixers. For both a phase/amplitude system and a hybrid system, suitable components are readily available from a few MHz up to about 18 GHz. Above this frequency range, components have been demonstrated up to 75 GHz but they are not readily available for certain frequency ranges. Most difficulty would likely arise in obtaining power amplifiers for some frequency ranges in which little demand for the equipment exist. In summary, although it is practical to implement both hybrid and phase/amplitude systems to cover the 200 MHz to 75 GHz range, above about 18 GHz some equipment will not be readily available.

Evaluation of phase/amplitude receiver types has continued. Discussions have been held with Scientific Atlanta personnel concerning the impact of requiring various degrees of precision in phase measurements. Providing a phase resolution down to a few degrees is not very stressing with current technology. However, providing a phase measurement precision of a few degrees does have
significant impact on receiver design and cost. Phase resolution and precision requirements are currently being factored into Scientific Atlanta's receiver design evaluations.

A broadband transmitting and/or reference antenna will be required in the Hybrid and the phase/amplitude systems. There are three general types of microwave antennas for broadband operation: (1) frequency independent antennas such as the spiral and log periodic, (2) broadband horn antennas, such as the ridged horn or pinwall, and (3) reflector antennas with broadband feeds. To obtain the desired gain, a reflector antenna with a broadband feed is the most appropriate choice for the current applications. Reflector antenna systems with good performance properties over octave or greater bandwidths can be achieved. Due to change in electrical size, the gain of a reflector antenna increases by 6 dB per octave of frequency increase. Feed horn effects may modify this change somewhat. However, since space loss also increases by the same factor, the two effects would tend to cancel and produce a more nearly constant power density at the receiving antenna than would have been produced if a transmitting antenna with constant gain versus frequency had been used. Of course, the two effects will not exactly cancel over an octave bandwidth, but both the Hybrid and the phase/amplitude systems can accommodate changes in power density. The Hybrid system electronically compensates for these changes. For the phase/amplitude system, calibration data will be used to computer-correct the measured test data and account for any variations in power density at the test antenna. Thus, suitable broadband antennas for both the Hybrid and the phase/amplitude system are available.

During the month, a 30-day no-cost contract extension was granted. As a result, technical effort will now extend to 19 March and the approval draft of the Final Technical Report will be due 19 April.

During the coming month, timing circuitry for the Hybrid system will be further evaluated, a technical report from Scientific Atlanta will be received and evaluated, and work on the contract Final Technical Report will begin.

During this reporting period, it is estimated that 198 professional man-hours were expended. It is estimated that the cumulative professional man-hours expended since contract initiation is 3344, and that approximately 98% of the technical effort has been completed.

Respectfully submitted,

Joseph D. Adams
Project Director

JDA:lb

Approved:

Fred L. Cain
Technical Area Manager
Electromagnetic Effectiveness
Air Force Systems Command  
Rome Air Development Center  
Griffiss Air Force Base  
New York 13441

Attention: Mr. Martin Jaeger, OCTS

Reference: Contract No. F30602-73-C-0194

Title: "Broadband Antenna Measurement Techniques"

Subject: Monthly Status Report No. 13

Dear Sir:

A summary of progress on the referenced contract for the period 1 March through 31 March is contained herein. Technical effort on the contract has been completed, and preparation of the Final Technical Report has begun. On 19 March, Mr. J. D. Adams visited RADC to discuss plans for the Final Technical Report and to review technical accomplishments on the contract.

A report from Scientific-Atlanta, which presented the results of their investigations under subcontract, was received and evaluated. Scientific-Atlanta has defined the required changes to their existing wide-range receivers for incorporating phase-locked frequency tracking over the 1 to 15 GHz range. In addition, the sideband phase processing circuits required for implementing the group velocity concept have been defined. With the modified receiver and the sideband processing circuits, either the group velocity or the separate reference channel direct phase measurements could be made. The sideband phase processing circuits can be constructed separately from the modified receiver. Calibration and reference requirements for phase/amplitude measurements have been further investigated. These requirements, and means for satisfying the requirements, will be fully discussed in the technical report now in preparation.

Technical efforts were completed, and on 19 March, Mr. J. D. Adams visited the RADC project engineer to review technical accomplishment on the contract. Also discussed during this visit were plans for the Final Technical Report. An outline of the report was reviewed, and the technical content in each area was described.
The report will contain a description of all efforts under the contract. Conclusions and recommendations for implementing broadband antenna measurement systems will be presented.

During the coming period, the Final Technical Report will be completed and sent to RAIC for approval. Completion of the report is scheduled for 19 April.

During this reporting period, it is estimated that 330 professional man-hours were expended, and that the cumulative professional man-hours expended since contract initiation is 3674. The technical effort has been 100% completed.

Respectfully submitted,

Joseph D. Adams
Project Director

Approved:

Fred L. Cain
Technical Area Manager
Electromagnetic Effectiveness
Air Force Systems Command  
Rome Air Development Center  
Griffiss Air Force Base  
New York 13441

Attention: Mr. Martin Jaeger, OCTS

Reference: Contract No. F30602-73-C-0194

Title: "Broadband Antenna Measurement Techniques"

Subject: Monthly Status Report No. 14

Dear Sir:

Progress on the referenced contract for the period 1 April through 19 April is described herein. The Final Technical Report was completed on 19 April 1974. The original reproducible and three copies were mailed to RADC on 22 April 1974. This final report is currently under RADC review. As specified in the referenced contract, Georgia Tech will make required corrections, if any, upon return of the reproducible and notification of any required corrections.

Subject to RADC approval of the Final Technical Report which has been submitted, the contract requirements have been fully met. During this reporting period it is estimated that 118 professional man-hours were expended and that a cumulative total of 3792 professional man-hours were expended on the contract. Attachment I to this status report is the contracturally required certificate of man-hours expended on the contract.

Respectfully submitted,

Joseph D. Adams  
Project Director

JDA/lb

Attachment

Approved:

Fred L. Cain  
Technical Area Manager  
Electromagnetic Effectiveness
CERTIFICATE OF TOTAL MAN-HOURS EXPENDED. This is to certify that based on monthly estimates of man-hours expended, the cumulative total of professional man-hours expended from contract outset to date as set forth in this report is correct. This certificate covers only the professional man-hours expended in support of work for which a level of effort is specified in the contract schedule.

M. W. Long  
Secretary,  
Georgia Tech Research Institute

Date: May 16, 1974
BROADBAND ANTENNA MEASUREMENT TECHNIQUES

by


Radar Division
Engineering Experiment Station
Georgia Institute of Technology

TECHNICAL REPORT NO. RADC-TR-

Distribution of this document is unlimited

Rome Air Development Center
Air Force Systems Command
Griffiss Air Force Base, New York
When US Government drawings, specifications, or other data are used for any purpose other than a definitely related government procurement operation, the government thereby incurs no responsibility nor any obligation whatsoever; and the fact that the government may have formulated, furnished, or in any way supplied the said drawings, specifications, or other data is not to be regarded by implication or otherwise, as in any manner licensing the holder or any other person or corporation, or conveying any rights or permission to manufacturer, use, or sell any patented invention that may in any way be related thereto.

Do not return this copy. Retain or destroy.
BROADBAND ANTENNA MEASUREMENT TECHNIQUES

by

J. D. Adams, F. L. Cain, and C. E. Ryan, Jr.

Engineering Experiment Station
Georgia Institute of Technology

Distribution of this document is unlimited
This final technical report describes the work performed under Contract F30602-73-C-0914, Project No. 4506, Task No. 450604 during the period from 16 February 1973 to 19 March 1974.

The work on this contract was performed by the Radar Division of the Engineering Experiment Station at Georgia Institute of Technology, Atlanta, Georgia under Project A-1517, with the support of Scientific-Atlanta, Inc., under sub-contract. The RADC Project Engineer was Mr. Martin Jaeger (OCTS).

In addition to the authors, contributors to the work described herein include D. C. Griffin, T. M. Miller, Jr., and J. M. Schuchardt at Georgia Tech, J. S. Hollis and R. E. Pidgeon, Jr. at Scientific-Atlanta, Inc.

This technical report has been reviewed and it is approved.

Approved:

Martin Jaeger  
Project Engineer  
Radiation and Signal Processing Section
This report presents the results of a program to study and investigate advanced measurement techniques for evaluating the performance of broadband antenna systems for use in high resolution radar systems. New techniques to measure, record, and analyze antenna gain and pattern performance were studied, and the feasibility of developing the necessary instrumentation to perform these measurements was investigated. Systems based on the use of broadband noise signal sources and systems using sweep frequency techniques were studied. It was concluded that systems using broadband noise signal sources would not be cost-effective. Preliminary design of a broadband amplitude-only electronically integrating sweep frequency system was completed. It was concluded that this system could be implemented immediately and that it would provide an effective first step in realization of the ultimate phase plus amplitude broadband antenna measurement system. The implementation and operation of sweep frequency amplitude plus phase systems were studied, and an effective approach to the realization of a phase/amplitude system was identified. Recommendations for implementing this approach are presented.
# TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>I. INTRODUCTION</strong></td>
<td><strong>1</strong></td>
</tr>
<tr>
<td>A. GENERAL</td>
<td>1</td>
</tr>
<tr>
<td>B. PROGRAM OBJECTIVES</td>
<td>2</td>
</tr>
<tr>
<td>C. SUMMARY OF WORK PERFORMED</td>
<td>3</td>
</tr>
<tr>
<td>1. Concept I--Broadband Noise</td>
<td>4</td>
</tr>
<tr>
<td>2. Concept II--Phase and Amplitude System</td>
<td>8</td>
</tr>
<tr>
<td>3. Hybrid System</td>
<td>13</td>
</tr>
<tr>
<td><strong>II. CONCEPT I SYSTEMS</strong></td>
<td><strong>17</strong></td>
</tr>
<tr>
<td>A. INTRODUCTION</td>
<td>17</td>
</tr>
<tr>
<td>B. AMPLIFIED THERMAL NOISE SYSTEM</td>
<td>17</td>
</tr>
<tr>
<td>C. NOISE-MODULATED SYNTHESIZED BROADBAND SYSTEM</td>
<td>21</td>
</tr>
<tr>
<td>D. INTERFERENCE BY NOISE SYSTEM</td>
<td>22</td>
</tr>
<tr>
<td>E. SUMMARY AND EVALUATION OF CONCEPT I SYSTEMS</td>
<td>28</td>
</tr>
<tr>
<td><strong>III. HYBRID SYSTEM</strong></td>
<td><strong>31</strong></td>
</tr>
<tr>
<td>A. INTRODUCTION</td>
<td>31</td>
</tr>
<tr>
<td>B. AMPLITUDE COMPENSATION</td>
<td>33</td>
</tr>
<tr>
<td>C. SENSITIVITY ANALYSIS</td>
<td>36</td>
</tr>
<tr>
<td>1. Minimum Detectable Signal</td>
<td>36</td>
</tr>
<tr>
<td>2. Transmitter Power Required</td>
<td>39</td>
</tr>
<tr>
<td>3. Detector Requirements</td>
<td>40</td>
</tr>
<tr>
<td>D. HARDWARE DESCRIPTIONS</td>
<td>41</td>
</tr>
<tr>
<td>1. Transmitter</td>
<td>41</td>
</tr>
<tr>
<td>2. Receiver</td>
<td>44</td>
</tr>
<tr>
<td>3. Data Processor</td>
<td>44</td>
</tr>
<tr>
<td>4. Timing and Control</td>
<td>55</td>
</tr>
<tr>
<td>5. Transmit and Reference Antennas</td>
<td>58</td>
</tr>
<tr>
<td>E. CALIBRATION</td>
<td>59</td>
</tr>
</tbody>
</table>
# TABLE OF CONTENTS (Continued)

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>F. SYSTEM ACCURACY</td>
<td>61</td>
</tr>
<tr>
<td>G. DATA</td>
<td>65</td>
</tr>
<tr>
<td><strong>IV. CONCEPT II SYSTEMS</strong></td>
<td>67</td>
</tr>
<tr>
<td>A. INTRODUCTION</td>
<td>67</td>
</tr>
<tr>
<td>B. SEPARATE REFERENCE CHANNEL CONCEPT</td>
<td>74</td>
</tr>
<tr>
<td>C. GROUP VELOCITY CONCEPT</td>
<td>77</td>
</tr>
<tr>
<td>D. CALIBRATION METHODS</td>
<td>83</td>
</tr>
<tr>
<td>1. Three-Antenna Group Delay</td>
<td>86</td>
</tr>
<tr>
<td>2. Separate Reference Channel Direct</td>
<td>88</td>
</tr>
<tr>
<td>3. Antenna Impedance and S-Parameter</td>
<td>89</td>
</tr>
<tr>
<td>4. Free Space Scattering</td>
<td>93</td>
</tr>
<tr>
<td>5. Comparison of Methods</td>
<td>95</td>
</tr>
<tr>
<td>E. GENERAL HARDWARE REQUIREMENTS</td>
<td>96</td>
</tr>
<tr>
<td>F. SYSTEM OPERATION</td>
<td>103</td>
</tr>
<tr>
<td>G. MEASUREMENT SYSTEM ACCURACY</td>
<td>109</td>
</tr>
<tr>
<td>H. LARGE REFLECTOR APPLICATIONS</td>
<td>112</td>
</tr>
<tr>
<td><strong>V. CONCLUSIONS AND RECOMMENDATIONS</strong></td>
<td>117</td>
</tr>
<tr>
<td><strong>VI. REFERENCES</strong></td>
<td>121</td>
</tr>
<tr>
<td><strong>VII. APPENDICES</strong></td>
<td>123</td>
</tr>
<tr>
<td>I. COMPONENT DESCRIPTIONS</td>
<td>124</td>
</tr>
<tr>
<td>II. TEST FOR IDENTICAL ANTENNAS</td>
<td>139</td>
</tr>
<tr>
<td>III. THE ANTENNA AS A TRANSMITTER AND RECEIVER</td>
<td>143</td>
</tr>
<tr>
<td>IV. THE ANTENNA AS A SCATTERER</td>
<td>147</td>
</tr>
<tr>
<td>Figure</td>
<td>Description</td>
</tr>
<tr>
<td>--------</td>
<td>-----------------------------------------------------------------------------</td>
</tr>
<tr>
<td>1</td>
<td>Simplified Block Diagram of Thermal Noise Transmitter</td>
</tr>
<tr>
<td>2</td>
<td>Broadband Receiver for Thermal Noise Transmitter</td>
</tr>
<tr>
<td>3</td>
<td>Noise-Modulated Synthesized Broadband System</td>
</tr>
<tr>
<td>4</td>
<td>Operation of Synthesized Broadband System</td>
</tr>
<tr>
<td>5</td>
<td>Direct Phase/Amplitude Separate Reference Channel System</td>
</tr>
<tr>
<td>6</td>
<td>Group Velocity Concept Phase/Amplitude System</td>
</tr>
<tr>
<td>7</td>
<td>Simplified Block Diagram of Hybrid System</td>
</tr>
<tr>
<td>8</td>
<td>Amplified Thermal Noise Transmitter</td>
</tr>
<tr>
<td>9</td>
<td>Double-Balanced Mixer Response</td>
</tr>
<tr>
<td>10</td>
<td>Schematic of Interference Condition</td>
</tr>
<tr>
<td>11</td>
<td>Hybrid System S-Band Transmitter Block Diagram</td>
</tr>
<tr>
<td>12</td>
<td>Hybrid System S-Band Receiver Block Diagram</td>
</tr>
<tr>
<td>13</td>
<td>Hybrid System Data Processor Block Diagram</td>
</tr>
<tr>
<td>14</td>
<td>Integrator For Hybrid System Data Processor</td>
</tr>
<tr>
<td>15</td>
<td>Typical Detector and Integrator Waveforms</td>
</tr>
<tr>
<td>16</td>
<td>Logarithmic Amplifier With Voltage-To-Current Converter</td>
</tr>
<tr>
<td>17</td>
<td>Illustration of Hybrid System Timing and Control Requirements</td>
</tr>
<tr>
<td>18</td>
<td>Broadband Pattern for Uniformly Illuminated Rectangular Aperture, 1 GHz Bandwidth</td>
</tr>
<tr>
<td>19</td>
<td>Broadband Pattern For Uniformly Illuminated Rectangular Aperture, 2 GHz Bandwidth</td>
</tr>
<tr>
<td>20</td>
<td>Broadband Pattern For Uniformly Illuminated Rectangular Aperture, 5 GHz Bandwidth</td>
</tr>
<tr>
<td>21</td>
<td>Schematic ω - β Diagram</td>
</tr>
<tr>
<td>22</td>
<td>General Antenna Phase Measurement System</td>
</tr>
</tbody>
</table>
LIST OF FIGURES (Continued)

Figure                      Page
23. Three Antenna Group Delay Calibration Method       87
24. Basic Setup For Measuring Reflection Coefficient $S_{11}$   91
25. Setup For Measuring Forward Transmission Coefficient $S_{21}$  92
26. Tracking Phase-Locked Receiver                      97
27. Sideband Differential-Phase Processor               98
28. Phase-and-Amplitude Measurement System              100
29. Paraboloid of Revolution Geometry                   114
30. Equivalent Bridge Circuit For Antenna Scattering     140
31. Thevenin Equivalent Circuit Antenna Analog          144
32. General Antenna Operating Conditions               148

LIST OF TABLES

Table                      Page
1. Received Noise Power Density                           26/27
2. Component Summary For S-Band Hybrid System Transmitter 45
3. Component Summary for S-Band Hybrid System Receiver    47
4. Component Summary of Hybrid System Data Processor      56
5. Actual Log Amp Output                                   63
6. Subtractor Output Voltage                              63
7. Typical Noise Sources                                   125
8. Common Minicomputer Features                           135
9. Broadband Dual Polarized Feed Data                      138
SECTION I
INTRODUCTION

A. GENERAL

New antenna-range measurement techniques must be developed to characterize antennas that are designed to meet the requirements of modern-day high-resolution radars. Satisfying the mission objectives of planned or existing modern-day radars requires meeting stringent performance specifications on, among other things, range resolution and range accuracy. Range resolution of 6 inches or less is desired in some applications. Obtaining this type of range resolution, along with good range accuracy, requires a very narrow pulse with a fast rise time to provide a well defined leading-edge. Obtaining increased resolution and accuracy requires increasing the bandwidth of several stages of the overall radar system, including that of the antenna. The old rule of thumb that a 1-microsecond pulsewidth, which requires a 1-MHz bandwidth, will provide a 500-foot resolution must now be extended to a 1-nanosecond pulsewidth, which requires a 1-GHz bandwidth, provides a 6-inch resolution. In order to obtain a one-half inch resolution, as apparently some users are proposing, a one-tenth nanosecond pulsewidth and a bandwidth of 10 GHz will be required. To radiate these broadband pulses with essentially no distortion, the antenna must have a wide instantaneous bandwidth.

Because short-pulse directive antennas must have a wide instantaneous bandwidth, new techniques for quickly measuring and evaluating antenna performance must be developed. Not only must the antenna have good broadband spatial amplitude characteristics, but it must also have good phase properties over the broad bandwidths required. Antenna phase and/or amplitude errors can lead to such undesirable effects such as degraded spatial patterns, pulse
distortion, and limitations on pulse compression performance. In many cases, if the specific phase and amplitude errors were known, corrective measures could be implemented to achieve more nearly ideal performance. Thus, it becomes imperative that the instantaneous broadband phase and amplitude properties of the antenna be known.

The measurement of instantaneous broadband phase and amplitude properties of an antenna to determine its effects on short pulses is a deviation from the conventional antenna pattern measurement technique. The additional measurement requirements that must be met to characterize short-pulse radar antennas demand a significant increase in antenna measurement capability that does not now exist. Currently available broadband measurement techniques which are applicable to high gain radar antennas typically rely on radio stars or noise generators as signal sources; as a consequence, they suffer such deficiencies as lack of practical bandwidth control, lack of phase information, limited dynamic range, and usually no polarization discrimination. Although time-domain impulse response techniques have also been used for testing wideband antennas [1,2], these techniques are generally not well suited to the high-gain narrow-beam microwave antennas. Under this program, Georgia Tech has investigated new techniques for achieving the desired broadband phase and amplitude measurement capabilities.

B. PROGRAM OBJECTIVES

The general objectives of this contract are to explore new techniques to measure, record, and analyze broadband antenna performance and to determine the feasibility of developing necessary instrumentation to perform such measurements. Although emphasis has been given to the 1-15 GHz frequency range, availability of major instrumentation components for the 200 MHz to 75 GHz region was investigated.
As a minimum in achieving the objectives of this program two basic concepts, which are designated as Concepts I and II, were considered as well as any variations from these two basic concepts. The first concept involves the use of a broadband leveled noise source to provide average antenna gain and pattern information. The second concept involves a swept frequency transmitter and a suitable receiver for measuring and recording sampled phase and amplitude data over a broad bandwidth. The recorded phase and amplitude data would permit the investigation of antenna response to any particular frequency component or to any combination of frequency components; hence, by proper selection of frequencies, the antenna effect on pulses of various bandwidths and modulations can be simulated and diagnosed. A variation from Concepts I and II, designated as the Hybrid approach, was formulated and investigated. This Hybrid approach has some similarities to both Concepts I and II. In all approaches, the system sensitivity should be sufficient to permit system use on very long outdoor antenna ranges with a dynamic range of at least 40 dB. A transmitter power output capability of approximately 1 to 10 watts CW is desired.

C. SUMMARY OF WORK PERFORMED

Initially, efforts on this contract were devoted to the analysis and evaluation of instrumentation for the wideband measurement systems of Concepts I and II. Of the various Concept I systems which were considered, two of the most promising ones were selected for evaluation. These two were identified as (1) a Broadband Amplified Thermal Noise System, and (2) a Noise-Modulated Synthesized Broadband Noise System. Under Concept II, both computer controlled amplitude-only systems and computer controlled amplitude-plus-phase systems were evaluated. A third approach, designated as the Hybrid system because of its similarities to both Concept I and Concept II systems, was also synthe-
sized and evaluated. Relative evaluations among all of the Concepts I and II systems indicated that the Concept I systems would not be cost effective due to their measurement limitations and equipment complexities. Consequently, with sponsor concurrence, succeeding efforts were concentrated on completing design of the Hybrid system and on further evaluation of techniques for implementing a Concept II system. A brief summary of the results of the investigations in all of the various areas is given below.

1. Concept I—Broadband Noise

From the various approaches that were considered for implementing a Concept I system, two different measurement systems involving noise techniques were selected for evaluation. One of these two noise techniques involves the use of a very broadband noise source, octave bandwidth TWT power amplifiers, a broadband transmitting antenna, and a broadband receiver whose pre-detection bandwidth is as broad as the transmitted spectrum. A simplified block diagram of the transmitter for this system is shown in Figure 1. Continuous center frequency control from 1 to 15 GHz and bandwidth adjustment from nearly 0 to 1 GHz are achieved through heterodyne techniques with adjustable frequency local oscillators and fixed-tuned filters. The most appropriate receiver type for this application is the wideband RF which consists of a detector-video receiver proceeded by a low-noise RF amplifier. A block diagram of this receiver is shown in Figure 2. This receiver has adequate sensitivity and bandwidth, but a separate RF amplifier is required for each octave of frequency or waveguide frequency band.

The second noise technique is based on sequential transmission of band-limited frequency-contiguous noise which is averaged in the receiver to synthesize a very broadband noise response. A block diagram of this system is shown in Figure 3. This system consists basically of a conventional voltage
Figure 1. Simplified block diagram of thermal noise transmitter
Figure 2. Broadband receiver for thermal noise transmitter
Figure 3. Noise-modulated synthesized broadband system
controlled oscillator, noise generator, and TWT power amplifier at the trans-
mitting site with an octave bandwidth receiver and averager at the receiving
site. Operation of this system may be understood by reference to Figure 4.
The antenna response to each of the relatively narrow-band noise pulses is de-
termined by the receiver, and this response is integrated over the time re-
quired to step the noise pulses over the full bandwidth desired. Consequently,
an adjustable bandwidth capability is achieved.

For the received power to reflect only the test antenna response over the
selected bandwidth, a constant power density versus frequency is required at
the test antenna. With a broadband noise spectrum, satisfying this requirement
is complicated by the fact that transmit antenna gain and space loss both
depend on frequency. Thus, even if a uniform amplitude noise spectrum at
the required power level is generated, the required constant power density at
the test antenna may not be achieved. One possible approach to a solution of
this problem is to tailor antenna gain to vary inversely as space loss so that
these two effects cancel. However, tailoring antenna gain to accurately
follow a prescribed gain function over an octave-bandwidth is a difficult
task. The Concept I broadband noise systems would provide average amplitude
information only, and a new spatial pattern must be measured for each desired
bandwidth.

2. Concept II--Phase and Amplitude System

The function of the phase-amplitude system is to measure and digitally
record the phase and amplitude response of the test antenna as the transmitter
is swept over a wide frequency range (typically up to 1 octave). These phase
and amplitude data as a function of frequency can be digitally processed to
determine antenna response to wideband signals, such as various short pulses
or FM coded longer pulses. Two different phase measurement techniques have
Figure 4. Operation of synthesized broadband system
been investigated for this system. With one technique, the transmitter produces a single-tone signal which is swept across the desired frequency range. With this technique, phase delay is measured directly, and a separate RF reference channel is required to obtain a phase reference signal at the receiver. With the second technique, the sweep oscillator is modulated to produce a carrier plus an upper and a lower sideband. These three signal components are swept across the measurement range while the sideband separation remains constant. With this second technique, the receiver separates the two sidebands whose phase separation has been modified after passing through the test antenna and then measures the phase between them. From this sideband phase difference as a function of carrier frequency, the test antenna phase characteristics can be obtained by integration. This second technique of measuring sideband phase information is based on the concept of group velocity, i.e., measurement of group delay. A phase/amplitude system using the direct (separate reference channel) phase measurement technique is diagrammed in Figure 5, while a system using the group velocity technique is diagrammed in Figure 6.

The two basic differences between these phase/amplitude systems are that (1) the group velocity system requires the addition of a modulator in the transmitter, and (2) the group velocity system requires no separate calibrated reference channel. A more subtle difference between the two systems involves the procedures required for system calibration and for data reduction. With the group velocity system, the measured sideband differential phase data will depend, of course, on the phase properties of the entire transmission path from signal generation to detection. To obtain test antenna phase properties, the measured data must be corrected for any phase dispersion introduced by the transmitter or the transmit antenna and for any errors in the receiver. The
Figure 5. Direct phase/amplitude separate reference channel system
Figure 6. Group velocity concept phase/amplitude system
measured data can be computer corrected if calibration data versus frequency is determined and stored for this purpose. With the separate reference channel system, the calibration involves determining the phase properties of the reference antenna and determining any difference in phase properties of the two transmission paths from the receiver to the test antenna and from the receiver to the reference antenna. Again, calibration data versus frequency can be used to computer-correct the measured data. Techniques for calibration and operation of these systems are discussed in Section IV.

3. Hybrid System

As indicated previously, measurement flexibility of a noise-based system is limited, and in addition, the instrumentation for a uniform amplitude adjustable bandwidth, noise transmitter is quite complex. A system which provides increased measurement capability with decreased instrumentation complexity is illustrated by the simplified block diagram shown in Figure 7. This new system uses a sweep frequency transmitter, such as would be used for the phase/amplitude system, and it provides results that are equivalent to the broadband noise systems. This system can also provide CW amplitude-only single frequency data. Thus, it is identified as the Hybrid system.

Operation of the Hybrid system may be easily understood by consideration of how an average antenna response is obtained with a noise system. The receiver output for the noise system is proportional to the integral of the amplitude transfer function of the antenna over the frequency range of the noise source. Consequently, the integration process of the noise system is instantaneous by virtue of the fact that all of the frequency components are simultaneously present in the incident signal and that a power detector provides a voltage which is proportional to the total power of all of these frequency components of the incident signal. The Hybrid system produces "average" antenna gain,
Figure 7. Simplified block diagram of hybrid system
which is equivalent to that of the noise system, by integrating the response of the antenna to a signal which is swept over the desired frequency measurement range. That is, in the Hybrid system, time-domain integration of a swept frequency signal replaces frequency integration of a noise signal.

Important waveform considerations at various circuit points in the Hybrid system are depicted by the sketches with dotted lines indicating the appropriate circuit point on the block diagram in Figure 7. An important feature of this Hybrid system is that amplitude variations (such as those due to variation in transmit antenna gain and space loss) can be compensated automatically in the receiver. This compensation is equivalent to obtaining constant power density at the test antenna aperture. Basically, compensation is based on electronically determining the ratio of test antenna response to the response of a reference antenna whose gain versus frequency characteristics are known. This system can provide average antenna response over any bandwidth up to the transmitter sweep range, or if desirable, the transmitter can be operated in a nonsweep mode so that the antenna response at any selected single frequency can be recorded. Preliminary design of this system is complete, and it is essentially ready for implementation.

All aspects of these systems, such as hardware requirements and system operation, are described in detail in following sections of this report. The body of this report describes the Concept I systems which were evaluated. Design and operation of the hybrid system is discussed, and the results of the feasibility studies of phase/amplitude systems are presented. To preserve report continuity and to facilitate reading of the report, basic component descriptions are presented in Appendix I.
A. INTRODUCTION

Initial efforts on this contract were devoted to investigation and evaluation of both Concept I and Concept II systems. Several Concept I systems were postulated, and two were investigated to the detail required to evaluate their potential as a broadband antenna measurement system. A hardware description of these two systems is given in this section, along with a discussion of the factors which leads to a conclusion that a Concept I system is not a cost-effective approach to the broadband antenna measurements problem.

B. AMPLIFIED THERMAL NOISE SYSTEM

A block diagram of the amplified thermal noise transmitter is shown in Figure 8. This transmitter may be considered as being composed of two sections. One section provides for adjustable bandwidth, and the other section provides adjustable center frequency. The particular frequency range of the devices has been selected for ease of implementation. This transmitter is based on use of a broadband uniform noise source, a low-noise TWT, and a broadband TWT amplifier as described in Appendix I, along with fixed tuned filters and heterodyne techniques to achieve an adjustable center frequency and bandwidth. Also shown in Figure 8 is a frequency scale upon which the bandpass or spectral output characteristics of the various components are shown. These simplified characteristics are identified with the respective components by connecting lines. For discussion, the bandpass and spectral output characteristics have been idealized and integer frequencies are used.

The thermal noise is amplified first by a low-noise TWT amplifier and then filtered by a bandpass filter such that the spectrum into the first mixer
Figure 8. Amplified thermal noise transmitter
has a power level of about 0 dBm (which is generally required for good mixer performance) and a constant amplitude from 3 to 4 GHz. Because the purpose of this stage of the system is only to provide an adjustable bandwidth, the frequency as well as the frequency range has been chosen to simplify mixer and local oscillator requirements. The output from oscillator 1 is mixed with this spectrum in a double-balanced mixer. The frequency of oscillator 1 is tunable from 5 GHz to 6 GHz. The double-balanced mixer suppresses the two input signals (original noise spectrum and oscillator 1 signal) relative to the sum and difference of these two signals. The first mixer output spectrum consists of a lower and an upper sideband. A low pass filter having an upper cut-off frequency of 2 GHz is used to pass the lower sideband and reject the upper.

If oscillator 1 is set at 5 GHz, then the lower or difference sideband spectrum will cover from 1 GHz to 2 GHz. Now, if the frequency of oscillator 1 is increased, the first mixer output spectrum will shift up in frequency and the lower sideband will be partially outside the passband of the low pass filter. By increasing the frequency of oscillator 1 to 6 GHz, the lower sideband will be completely outside the filter passband. As shown by the notation on the frequency scale, the upper limit of the noise spectrum at the output of the low pass filter remains at 2 GHz and the lower limit is adjustable from 1 to 2 GHz. Thus, the bandwidth is adjustable from 0 to 1 GHz.

The next stage of the transmitter is for shifting the adjustable bandwidth spectrum to other portions of the 1-15 GHz frequency range of interest. To accomplish this shift, the adjustable bandwidth spectrum is applied to a second mixer. This mixer is used as a double-balanced modulator where the spectrum is up-converted by mixing with a 16 GHz signal from the fixed-frequency oscillator 2. The output spectrum is a double sideband suppressed carrier signal, and only the lower sideband is passed by a bandpass filter. At the output of this
The bandpass filter, the spectrum bandwidth is adjustable from 0 to 1 GHz across the frequency range from 14 to 15 GHz. The third mixer and adjustable oscillator 3 are used to down-convert this spectrum to the desired center frequency. By adjusting oscillator 3 output from 16 to 30 GHz, the lower sideband from mixer 3 can be translated to any center frequency from 1 to 15 GHz. The upper sideband and the oscillator 3 signals are rejected by a bandpass filter. The final amplifier (or amplifier chain) must have sufficient gain (50 to 60 dB) to amplify the noise spectrum to a level of 1 to 10 watts.

Because microwave components, especially amplifiers, typically operate over octave or waveguide bandwidths, coverage of the frequency range of interest (1-15 GHz) would require that the third mixer, the third oscillator, the final amplifier, and the bandpass filter be duplicated for each octave or waveguide band. Several other components, not shown in the simplified block diagram of Figure 8, would be required for proper system operation. To obtain equal noise power across the frequency bands, equalization of TWT gain variations would be required. To prevent reactive mismatches between components, an isolator would be required before each filter to absorb reflected power.

All oscillators in the system must be stabilized to a low percentage of error because the spectrum bandwidth and the center frequency are controlled by the frequency differences between the three relatively high frequency oscillators. A large collection of microwave components is thus required to build this transmitter. A broadband transmitting antenna is required to radiate this signal. A transmitting antenna gain function which will provide constant power density at the test antenna is desirable.

A suitable broadband receiver is necessary to record the average radiation pattern of an antenna over a broadband of frequencies. Assume for the moment that the broadband noise signal arrives at the test antenna aperture with
constant spectral density, i.e., it has constant power per unit bandwidth. If such a signal in incident on an antenna, the output spectral density is determined by the amplitude-only frequency response characteristics, $H(s)$, of the antenna. Broadband detection of this output signal followed by low-pass filtering with a filter having an integration time sufficient to remove the random amplitude fluctuations of the noise source results in an output that is proportional to $\int H(s) \, ds$. As indicated in Appendix I, the most suitable receiver type is the broadband RF receiver. This type of receiver, without the low-noise amplifier, is used, for example, in the HP 415 SWR Meter and the SA 1554-2 Crystal-Bolometer Amplifier plug-in for the SA 1520 Rectangular Pattern Recorder.

C. NOISE-MODULATED SYNTHESIZED BROADBAND SYSTEM

A simplified block diagram for implementation of this type of system was shown in Figure 3. The system would consist basically of a conventional voltage controlled oscillator (e.g., a BWO), noise generator, double-balanced mixer, power amplifier (TWT), and amplitude leveler at the transmit site with octave bandwidth receiver (RF front end and detector) at the receiving site. The wideband antenna response would be synthesized by integrating response to narrow band noise modulated signals.

The low level portion of the noise transmitter considered here consists of a voltage-tunable oscillator, a double balanced mixer, and a low frequency noise source. The operation of the low-level portion of the noise source is most influenced by the effectiveness of the double balanced mixer (DBM) as a Double Sideband Suppressed Carrier up-converter. If the DBM functions effectively, the oscillator signal is suppressed, and the low frequency noise is up-converted to produce well characterized noise sidebands centered about the oscillator frequency.
Figure 9 shows the spectrum of the signals involved. A low frequency spectrum extending from DC to \( f_n \) is generated by the noise source. This low frequency noise is translated upward and is centered around the CW oscillator frequency \( f_o \). The actual CW carrier is suppressed by the double balanced mixer. This double sideband suppressed carrier modulation (DSBSC) in effect doubles the noise spectrum bandwidth.

To cover the desired antenna measurement range, the voltage tuned oscillator signal is slowly swept over the required frequency range, and the broadband noise response is determined electronically at the receiver. This procedure was illustrated in Figure 4.

**D. INTERFERENCE BY NOISE SYSTEM**

A potential problem with use of a broadband noise system is interference by the radiated signal with other receivers in the surrounding area. Because the radiated signal could be of relatively high power density (10 watts radiated by high gain antenna) and very broadband, interference could be significant. Calculations were made to determine the expected received noise power for various receiver bandwidths and center frequencies.

The noise power per unit bandwidth, \( N_d \), at the receiving antenna terminals is given by

\[
N_d (\phi, \theta) = \frac{P_t G_t (\theta) G_r (\phi) \lambda_c^2}{(4\pi)^2 R^2 B_t}
\]

where

- \( P_t \) = power transmitted,
- \( G_t (\theta) \) = power gain of transmitting antennas as a function of angle \( \theta \),
- \( G_r (\phi) \) = power gain of receiving antenna as function of angle \( \phi \),
Figure 9. Double-balanced mixer response
\[ \lambda_c = \text{wavelength at the carrier frequency}, \]
\[ R = \text{range between transmitting and receiving antennas, and} \]
\[ B_t = \text{noise bandwidth}. \]

The geometry of the interference situation is shown schematically in Figure 10. For analysis, the above parameters were assumed to have values typical of those to be expected under measurement conditions. Noise bandwidths of 250 MHz and 1 GHz at carrier frequencies between 0.5 and 14.5 GHz and a transmitter power of +15 dBm were assumed. Any variations of the transmitting and receiving antenna gains over the noise bandwidths were neglected. Although several different transmitting antennas could be used over the 1-15 GHz range, the peak gains of the transmitting and receiving antennas were assumed to be equal and to have a value of 20 dB up to 8 GHz and of 30 dB from 8 GHz to 15 GHz. In summary, the following parameters were assumed:

\[ B_t = 250 \text{ MHz and 1 GHz} \]
\[ P_t = +15 \text{ dBm, and} \]
\[ G_t(0) = G_r(0) = 20 \text{ dB @f < 8 GHz or} \]
\[ 30 \text{ dB @f > 8 GHz}. \]

Calculations were made for several different center frequencies, and the results are given in Table 1. These data show the noise power density at the receiving antenna terminals for main lobe-to-main lobe coupling with the assumed maximum gains. Sidelobe coupling can be accounted for by subtracting the sidelobe level in dB from the value of \( N_d \) which is presented. The receiver bandwidth must be accounted for as follows. If the bandwidth of the receiver up through the detector is greater than or equal to the noise bandwidth \( B_t \), the full noise power will be detected and would be equal to the noise power per unit bandwidth \( N_d \) multiplied by the ratio of \( B_t \) to 1 MHz. For a superheterodyne re-
Figure 10. Schematic of interference condition
<table>
<thead>
<tr>
<th>( f_c ) = 0.5 GHz</th>
<th>( f_c ) = 0.5 GHz</th>
<th>( f_c ) = 5 GHz</th>
<th>( f_c ) = 5 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>( B_t = 250 \text{ MHz} )</td>
<td>( B_t = 1.0 \text{ GHz} )</td>
<td>( B_t = 250 \text{ MHz} )</td>
<td>( B_t = 1 \text{ GHz} )</td>
</tr>
<tr>
<td>( N_d (\text{dBm/MHz}) )</td>
<td>( R (\text{km}) )</td>
<td>( N_d (\text{dBm/MHz}) )</td>
<td>( R (\text{km}) )</td>
</tr>
<tr>
<td>-45</td>
<td>1.0</td>
<td>-51</td>
<td>1.0</td>
</tr>
<tr>
<td>-51</td>
<td>2.0</td>
<td>-57</td>
<td>2.0</td>
</tr>
<tr>
<td>-57</td>
<td>4.0</td>
<td>-63</td>
<td>4.0</td>
</tr>
<tr>
<td>-65</td>
<td>10.0</td>
<td>-71</td>
<td>10.0</td>
</tr>
<tr>
<td>-71</td>
<td>20.0</td>
<td>-77</td>
<td>20.0</td>
</tr>
</tbody>
</table>
**TABLE (Concluded)**

**RECEIVED NOISE POWER DENSITY**

<table>
<thead>
<tr>
<th>$f_c = 10$ GHz</th>
<th>$f_c = 10$ GHz</th>
<th>$f_c = 14.5$ GHz</th>
<th>$f_c = 14.5$ GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>$B_t = 250$ MHz</td>
<td>$B_t = 1$ GHz</td>
<td>$B_t = 250$ MHz</td>
<td>$B_t = 1.0$ GHz</td>
</tr>
<tr>
<td>$N_{d dBm/MHz}$</td>
<td>$R$(km)</td>
<td>$N_{d dBm/MHz}$</td>
<td>$R$(km)</td>
</tr>
<tr>
<td>$N_{d dBm/MHz}$</td>
<td>$R$(km)</td>
<td>$N_{d dBm/MHz}$</td>
<td>$R$(km)</td>
</tr>
<tr>
<td>$N_{d dBm/MHz}$</td>
<td>$R$(km)</td>
<td>$N_{d dBm/MHz}$</td>
<td>$R$(km)</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th></th>
<th>-51</th>
<th>-57</th>
<th>-55</th>
<th>-61</th>
<th>-57</th>
<th>-63</th>
<th>-61</th>
<th>-67</th>
<th>-71</th>
<th>-77</th>
<th>-83</th>
<th>-81</th>
<th>-87</th>
<th>-81</th>
<th>-87</th>
<th>-81</th>
<th>-87</th>
<th>-81</th>
<th>-87</th>
</tr>
</thead>
<tbody>
<tr>
<td>R (km)</td>
<td>1.0</td>
<td>1.0</td>
<td>1.0</td>
<td>1.0</td>
<td>2.0</td>
<td>2.0</td>
<td>2.0</td>
<td>2.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
</tr>
</tbody>
</table>
receiver with an IF bandwidth which is less than $B_t$, the detected noise is the indicated noise power density multiplied by the ratio of the IF bandwidth to 1 MHz.

Several conclusions can be made with the aid of the data which are presented in Table 1. If recording an antenna pattern in conjunction with a broadband receiver, the total noise power received would be $N_d$ increased by 14 dB for the 250 MHz case and by 30 dB for the 1 GHz case. For pattern recording under either the 250 MHz or the 1 GHz condition at ranges of 10 km (6.2 miles) or less, the received power under the assumed conditions is adequate for observing at least -40 dB sidelobes with a typical receiver sensitivity.

For interference considerations without main-lobe-to-main-lobe coupling, $N_d$ would decrease greatly for most high gain antennas (25 to 50 dB depending on the orientation of the receiving antenna with respect to the transmitting antenna). If the victim receiver were a typical superheterodyne with a 10 MHz IF bandwidth, then the detected power at a range of 4 km for the $f_c = 0.5$ GHz $B_t = 250$ MHz case would be $-57 + 10 - 25 = -72$ dBm and $-57 + 10 - 50 = -97$ dBm for antenna coupling reduced by 25 and 50 dB, respectively. At range of 20 km these levels decrease to $-86$ dBm and $-111$ dBm, respectively. Thus, depending on the orientation and the gain of the antenna at the victim reciever site, interference could very well be a problem. A specific analysis would be required for each anticipated test range to determine permissible transmitted power levels.

E. SUMMARY AND EVALUATION OF CONCEPT I SYSTEMS

Concurrent with the investigation and evaluation of the broadband noise systems which have been described in this section, swept frequency phase and/or amplitude systems were also under investigation. These concurrent investigations for realization of broadband antenna measurement instrumentation were
required in order to define the relative merit of the two basic concepts under consideration. Based on these initial investigations, the following conclusions and evaluations of Concept I systems are made.

(1) Because the noise spectrum across the entire measurement bandwidth is simultaneously present, spectrum shaping must be accomplished by passive devices, and therefore, achieving uniform power density at the test antenna would be difficult.

(2) Only average antenna patterns can be obtained, and a new spatial pattern must be recorded for each desired bandwidth.

(3) Achieving adjustable bandwidth, adjustable center frequency, and adequate transmitter power requires a large collection of microwave equipment.

(4) Because radiation at potentially interfering frequencies cannot be easily controlled, a broadband high power noise transmitter could cause significant interference.

On the other hand, a Concept II system or the Hybrid system (Sections IV and III, respectively) can provide average patterns at selected bandwidths and center frequencies by simply changing sweep limits at the transmitter. These other systems can also provide single frequency data at any frequency within the transmitter sweep range. In addition, these other systems (as discussed later) do not require uniform power density at the test antenna. Moreover, implementation of the Hybrid system, which yields results equivalent to the noise systems, is considerably less complex than implementation of the noise systems. Further, because recorded broadband phase and amplitude data would provide complete flexibility in antenna diagnostics (CW spatial patterns, average patterns, pulse diagnostics, FM distortion prediction), development of a phase/
amplitude computer controlled system is the ultimate goal of the broadband antenna measurements program. Realization of the Hybrid system will be a positive step in the realization of the phase/amplitude system, while at the same time, it will provide more capability than a noise-based system could provide. Based on the conclusion that a more detailed design analyses of Concept I systems was not appropriate, efforts were thereafter devoted to only the Hybrid and Concept II systems. The remainder of this report deals with the design, capabilities, and operation of these other systems.
A. INTRODUCTION

As indicated in the previous section, Concept I systems provide only limited measurement capabilities while having complex instrumentation requirements. The Hybrid system, which is less complex than the noise systems of Concept I, can provide both the average pattern measurement capability of a Concept I system and the single frequency capability of an amplitude-only Concept II system. Measurement of broadband spatial patterns can be considered as a first step in the characterization of broadband radar antennas. If the antenna under test does not have a well behaved broadband pattern, it may not provide satisfactory radar performance, regardless of its phase properties which affect the pulse characteristics. In addition, broadband spatial characterization of an antenna is important from the electromagnetic compatibility (EMC) and electronic countermeasure (ECM) viewpoints. Thus, measurement of average broadband spatial patterns is an important consideration. The Hybrid system will provide this capability, and because of its relative simplicity, the Hybrid system can be configured as a mobile system for field test. Thus, the Hybrid system will provide an important measurement capability while serving as a building block for the phase/amplitude system.

A simplified block diagram of the Hybrid system has been given in Figure 7. The Hybrid system produces "average" antenna gain by integrating the antenna response to a signal which is swept over the desired frequency measurement range. That is, in the Hybrid system, time-domain integration of a swept frequency signal replaces frequency integration of a noise signal. An important feature of this Hybrid system is that amplitude variations (such as
those due to variations in transmit antenna gain and space loss) can be compensated automatically in the receiver. However, because a constant-power transmitter may be desirable for some applications and because this transmitter might be used with other receivers, this design includes provisions for obtaining a constant-power transmitter.

Referring to Figure 7, operation of the hybrid system may be described as follows. The reference antenna switch position is selected and the sweep oscillator begins to sweep over a pre-selected frequency range. The power amplifier (a travelling wave tube) provides necessary power gain to produce adequate signal-to-noise ratio at the receiver. Up to 10 watts output from the power amplifier is available. A feedback loop provides a leveling voltage to the sweep oscillator so that transmitter power variations versus time or frequency can be eliminated. At the receiving site, a bandpass filter over the desired operating frequency range can be used, if necessary, to reject interfering signals. Consequently, at the output of the bandpass filter, undesired signals are reduced to the noise level. The power versus frequency at this point reflects the bandpass characteristics of the receiving antenna. A low-noise broadband amplifier is used before the detector to improve receiver sensitivity. The detector voltage versus time is integrated over one transmitter sweep period, and the integrated voltage represents average antenna power response over the transmitter sweep range. This integrated voltage is "read" at the end of the integration time and converted to logarithmic (dB) form. This converted value now represents average response of the reference antenna over the desired frequency range, expressed in dB form, at one spatial point. This value is temporarily stored until the above described process is repeated for the test antenna. After completion of this process for the test antenna, the two values are subtracted, and the result
is the average gain of the test antenna expressed in dB which is referred to the average gain of the reference antenna over the same bandwidth.

In typical operation, the test antenna would rotate continuously while the transmitter sweep cycle is repeated each 10 msec. At an antenna rotation rate of 360 degrees per one-half hour, for example, the test antenna would move only $2 \times 10^{-3}$ degrees during a transmitter sweep, and the antenna could be considered stationary. Thus, one obtains an essentially continuous spatial record of the average gain of the test antenna, referenced to the peak average gain of the reference antenna. If the absolute gain versus frequency characteristic of the reference antenna is known, its "average absolute" gain and the test antenna's "average absolute" gain can be calculated for any desired bandwidth up to the full sweep capability of the transmitter (typically one octave).

The following portions of this section present an analysis of the amplitude compensation approach, a system sensitivity analysis, hardware descriptions, and a discussion of applications of the Hybrid system.

B. AMPLITUDE COMPENSATION

If the power density of the signal which illuminates the test antenna varies with frequency and this received signal were integrated directly, an erroneous integrated gain would be obtained. Thus, either the incident power density must remain constant over the integration time or any variations must be compensated. As indicated previously, maintaining constant power density at the test antenna over an octave bandwidth would be difficult. Fortunately, electronic compensation can be achieved in a relatively straightforward manner.

The average gain of the test antenna can be determined from measurements of the average power of the antenna under test and of a reference antenna. Consider a general antenna test system consisting of a transmitter, a transmit
antenna, a test antenna, and a receiver. The power received through the test antenna as a function of frequency may be expressed as

\[ P_{\text{test}}(f) = k G_{\text{TR}}(f) P_T L(f) G_{\text{test}}(f), \]

where

- \( P_{\text{test}}(f) \) = received power through test antenna as function of frequency,
- \( k \) = constant,
- \( G_{\text{TR}}(f) \) = gain of transmit antenna as function of frequency,
- \( P_T \) = transmitter power (assumed constant),
- \( L(f) \) = system losses (space attenuation) as function of frequency,
- \( G_{\text{test}}(f) \) = gain of test antenna as function of frequency.

It is desired to measure \( P_{\text{test}}(f) \) and thereby determine the test antenna gain, \( G_{\text{test}}(f) \). Therefore, all the other variables in the above equation must either be known as a function of frequency so that their effect can be removed mathematically, or the variables may be compensated electronically. Electronic compensation may be accomplished in real time, and it does not require knowledge of the functions which describe how each of the variables depend on frequency. Electronic compensation is accomplished through use of an additional reference antenna. The power received through the reference antenna, \( P_{\text{ref}}(f) \), may be written as

\[ P_{\text{ref}}(f) = k G_{\text{TR}}(f) P_T L(f) G'_{\text{ref}}(f), \]

where \( G'_{\text{ref}}(f) \) is the peak gain of the reference antenna as a function of frequency, and the other symbols are as previously defined. The hybrid system determines the average test antenna gain over some frequency range \( \Delta f \). In terms of average quantities the two previous equations can be expressed as

\[ \overline{P}_{\text{ref}} = k \overline{G}_{\text{TR}} G'_{\text{ref}} \overline{L} P_T, \]

and

\[ \overline{P}_{\text{test}} = k \overline{G}_{\text{TR}} \overline{G}_{\text{test}} \overline{L} P_T. \]
\[ \overline{P}_{\text{test}} = k \overline{G}_{\text{TR}} \overline{G}_{\text{test}} L P_T, \]  

where the bar indicates an average over a specified \( \Delta f \). Dividing \( \overline{P}_{\text{test}} \) by \( \overline{P}_{\text{ref}} \) and solving for \( \overline{G}_{\text{test}} \) yields

\[ \overline{G}_{\text{test}} = \frac{\overline{G}_{\text{test}}}{\overline{G}_{\text{ref}}} \overline{P}_{\text{test}} \overline{P}_{\text{ref}}. \]  

Thus, average test antenna gain may be determined from measurements of \( \overline{P}_{\text{test}} \) and \( \overline{P}_{\text{ref}} \) if the average gain of the reference antenna is known. The effect of all other variables has been eliminated by taking the ratio, \( \overline{P}_{\text{test}}/\overline{P}_{\text{ref}} \).

The logarithmic or decibel form of the above Equation (6) is

\[ 10 \log \frac{\overline{G}_{\text{test}}}{\overline{G}_{\text{ref}}} = 10 \log \overline{P}_{\text{test}} - 10 \log \overline{P}_{\text{ref}}. \]  

The average gain of the test antenna, expressed in dB and referred to the average gain of the reference antenna, is obtained by taking the difference of the logarithms of the integrated received powers. The basic compensation concept* is shown in block diagram form as follows:

![Block Diagram](image)

*An alternate compensation concept also was considered. The end result of this compensation concept was the variation of test antenna gain only, referred to the gain of the test antenna at one end of the frequency range. However, this concept required two electrically identical antennas and a voltage ramp at the receive site which was exactly time coincident over the entire sweep time with the transmitter frequency sweep ramp. In addition, this concept required constant transmitter power. Due to practical problems in meeting all these requirements, this concept is not attractive.
This circuit can be simplified and hardware savings can be realized by time-sharing one channel from the two antennas to the subtractor. The time-shared system becomes

On alternate transmitter sweeps, test and reference antennas are selected. It should be recognized that the log output must be "held" for one transmitter sweep. A completely analog implementation of the Hybrid system which incorporates this time-shared receiver channel has been designed, and it is discussed in following portions of this section.

C. SENSITIVITY ANALYSIS

1. Minimum Detectable Signal

The hybrid system is based on a readily realizable design which meets the measurement goals of a Concept I system, provides CW measurement capability, and will provide growth capability by incorporation of major portions of the transmitter into a Concept II system. Accordingly, meeting performance requirements with a relatively simple receiver design was a design goal which has been accomplished. As discussed in Appendix I, the most straightforward receiver design for obtaining broad instantaneous
bandwidth (no tuning or LO tracking) with good sensitivity is the wideband RF receiver with a separate low-noise RF amplifier for each octave band.

The minimum detectable signal for a receiver is limited by thermal noise. Thermal noise power $N$ is given by

$$N = k T_o B,$$  \hspace{1cm} (8)

where

$k = \text{Boltzmann's constant},$

$T_o = \text{reference temperature} = 290^\circ K,$ and

$B = \text{receiver bandwidth}.$

The specific Hybrid system which is considered here is designed for the 2 to 4 GHz octave. Thermal noise in this 2-GHz band is given by

$$N = \left(4 \times 10^{-21} \text{ joules}\right) \left(2 \times 10^9 /\text{sec}\right)$$

$$= 8 \times 10^{-12} \text{ watts} = 8 \times 10^{-9} \text{ mW}, \text{ or}$$

$$N = -81 \text{ dBm}.$$ 

Because typical semiconductor detectors have minimum detection levels on the order of -60 dBm, a signal level of -60 dBm would provide a 21-dB signal-to-noise ratio (SNR). However, since a SNR of 21 dB is not required, the detection of lower level signals can be achieved by use of a low-noise RF amplifier before the detector.

The effect of adding the RF amplifier, as illustrated in the receiver block diagram shown below, must be considered. For compensation, switching between the test and reference antenna is required. A bandpass filter has also been included for rejection of out-of-band interfering signals.
The switch and filter losses as well as the gain and noise figure of the RF amplifier must be considered in calculating sensitivity of this receiver configuration. These factors can be accounted for by the effective receiver noise figure, which is equivalent to the degradation in the signal-to-noise ratio (SNR) from the antenna terminals through the detector. This effective noise figure for this receiver configuration is

$$F_{\text{eff}} = L_s + L_f + (L_s + L_f) (F_{\text{amp}} - 1)$$

where

- $L_s$ = switch insertion loss,
- $L_f$ = filter insertion loss, and
- $L_{\text{amp}}$ = noise figure of the RF amplifier.

The insertion losses of the filter and the switch are estimated at 1.0 dB each. For the 2-4 GHz range, typical amplifier specifications are the following.

- Maximum noise figure: 4.5 dB
- Minimum small signal gain: 20 to 35 dB
- Power out @ 1-dB gain compression: 5 mW
With the above switch, filter, and amplifier specifications, the effective receiver noise figure would be

$$F_{\text{eff}} = 1.26 + 1.26 + (1.26 + 1.26) (1.82) = 7.65.$$  

The noise figure, expressed in dB, is 8.84 dB.

Based on the results obtained from Equations (8) and (9), the received signal power $P_r$ at the antenna terminals can be determined if a minimum detection requirement of a 10-dB SNR is assumed. For an assumed SNR of 10 dB, the required signal power at the antenna terminals must be at least 10 dB above the effective noise level, $k T_0 F_{\text{eff}} B$. Therefore, received power must be

$$P_r \geq \text{SNR} (k T_0) F_{\text{eff}} B$$

$$= 10(4 \times 10^{-21}) (7.65) (2 \times 10^9)$$

$$= 6.12 \times 10^{-7} \text{ mW}, \text{ or } \approx -62 \text{ dBm}.$$  

With an RF amplifier gain of, say 25 dB, power into the detector would be a minimum of -37 dBm, or greater than 20 dBm above its noise level.

2. **Transmitter Power Required**

The transmitter design is based on providing up to 10 watts at the transmit antenna terminals so that adequate transmitter power is available for a variety of antenna conditions and for growth to the Concept II system. Antenna test ranges of up to 7000 feet are postulated. Besides the transmitter power and test range considerations, other factors which will affect the received power level are space loss, test frequency, and antenna size. Because the greater space loss occurs at higher frequencies, the received power for a 10-watt transmitter will be calculated for test operation at the high frequency edge of S-band, 4 GHz. Also, since a fairly directive transmit antenna should be used to minimize multipath problems, a 30-dB gain antenna will be assumed. For a typical aperture illumination function, the 3-dB
beamwidth of this transmitting antenna will be approximately 5 degrees. A gain of 30 dB for the antenna under test will also be assumed. If smaller antennas were used, the test range could be considerably reduced. For example, the typical far-field range requirement \((R \geq 2 D^2/\lambda)\) for a 30-dB gain 4 GHz antenna is satisfied at a range of about 100 ft. Therefore, the above set of numbers represents an expected worst-case condition. The received power under these conditions would be

\[
P_r = \frac{P_t G_t G_r \lambda^2}{(4 \pi R)^2}
\]

\(\approx 8 \times 10^{-2} \text{ mW}, \) or

\[
(P_r)_{\text{dB}} \approx -11 \text{ dBm},
\]

where

- \(P_t = \) power delivered to the transmitting antenna = 10 watts,
- \(G_t = \) gain of transmitting antenna = 1000 (30 dB),
- \(G_r = \) gain of receiving antenna under test = 1000 (30 dB),
- \(\lambda = \) operating wavelength at 4.0 GHz = 0.25 feet, and
- \(R = \) range between the two antennas = 7000 feet.

This -11 dBm is the received power in the main beam of the test antenna. Since -62 dBm of Equation (11) provides a 10-dB SNR, sidelobes of the test antenna as low as -51 dB below the peak of the main beam could be observed at a 10-dB SNR with the assumed transmitter power, test range, and antenna gains.

### 3. Detector Requirements

A minimum of 40-dB instantaneous dynamic range is required. This is approximately the maximum range which can be covered by a semiconductor detector, while maintaining square-law response, without special electronic compensation schemes. Typical semiconductor detectors cover the RF power
range from 0 dBm to -40 dBm with square-law response and with corresponding
detector output from 25 μV to 250 mV. The minimum received power at the test
antenna terminals has been specified as -62 dBm. Switch and filter insertion
losses have been estimated at a total of 2 dB maximum. Therefore, to obtain
a minimum of -40 dBm at the detector input, the low noise amplifier must have
a gain of at least 23 dB. Specifying a minimum gain of 25 dB would allow
inserting a coupler for power monitoring and a limiter for receiver protection
after the RF amplifier. Inserting these devices after the 25-dB gain low-
noise amplifier will not materially degrade the effective noise figure of the
receiver.

D. HARDWARE DESCRIPTIONS

Preliminary design of the hybrid system and component specification for
the 2-4 GHz range have been completed. This subsection contains a description
of the hardware necessary to implement this 2-4 GHz system.

1. Transmitter

A block diagram of the 2-4 GHz (S-band) transmitter is given in
Figure 11. This transmitter, which will provide constant power at the 10-watt
level across this 2 to 4 GHz range, consists of a basic signal source (solid-
state sweep oscillator), power amplifier (helix traveling wave tube, TWT), and
a power sampler to provide a leveling voltage for the sweep oscillator. Power
output across the 2-4 GHz range is typically 10 mW ± 0.5 dB for solid-state
sweepers. Thus, the TWT must have a gain of 30 dB to provide a 10-watt output.
Octave bandwidth TWTs typically have small signal gain variations versus fre-
quency. In a well constructed TWT, gross gain variations across the band can
be held to about 3 dB with a fine grain variation about this gross curve of
± 1 dB. For the entire transmitter chain, power variations of 4-5 dB could
occur. Thus, to achieve constant transmitter-power leveling is required.
Figure 11. Hybrid system S-band transmitter block diagram
For long term reliability, solid-state sweep oscillators are preferred over conventional BWO (backward wave oscillator) sweep oscillators. For a typical 2-4 GHz sweeper, the output power variation with external leveling may be held to ± 0.1 dB plus leveling signal error. A TWT output of 10 watts will be obtained for an input of 10 mW if the TWT gain is 30 dB. If the upper limit of the TWT's small signal region is at 10 watts, more than a 10 mW output from the sweeper would drive the TWT into saturation. For signal purity, it is desirable to operate a TWT in the small signal region; therefore, a sweeper output level-set capability is desired. This level-set capability is also desirable to reduce the transmitter power to only that required for pattern measurements under varying range conditions. With a TWT noise figure of 35 dB and a gain of 30 dB, the noise power output in a 2 GHz band would be 56 dB below the 10-watt level so that a very broad dynamic range in the TWT output is available. (Noise power output may be calculated from Equation A-3 of Appendix I with the ENR term being equivalent here to the sum of gain plus noise figure.)

Remote programming is an option which many oscillators have available and is attractive for the hybrid system and for growth to a phase/amplitude system. Up to 1000 discrete frequencies may be remotely selected through a 12-line 8421 BCD input. A standard feature (RF Enable) permits the remote on/off control of the RF power. These features would allow remote selection of single frequencies for CW testing.

Octave bandwidth TWTs may have significant second harmonic output, depending on drive level and frequency. A low-pass filter is incorporated to remove any harmonics generated in the TWTA. Since low-pass filters are not very expensive ($40), the option of selecting a filter with the minimum possible cut-off for a particular sweep range is recommended. For example, if a 4 GHz
low-pass filter were used when sweeping from 2 GHz to 3 GHz, the harmonic of the low end of the sweep range would not be cut-off. A 3-GHz cut-off would be appropriate, and a minimum of 2 low-pass filters per octave range is desirable. Pertinent specifications for the components of this transmitter are summarized in Table 2.

2. Receiver

The basic receiver design concept and performance of the critical components have been discussed in Subsection III-C, Sensitivity Analysis. A block diagram of the receiver is shown in Figure 12. This receiver accepts the RF signals from either the test or reference antenna and provides a dc output which is proportional to the power level of the RF signal. The receiver instantaneous dynamic range for square-law response is a minimum of 40 dB over the detector input range from 0 dBm to -40 dBm. The typical detector output voltage range for this power range is from 25 μV to 250 mV. However, this type of detector has sufficient sensitivity for use at power levels as low as -50 dBm, and as discussed later, the reference antenna compensation scheme tends to eliminate system non-linearities so that the useful dynamic range will be 50 dB or greater. Laboratory tests will be required to define the system accuracy over this detection range.

A limiter and power sampler (coupler) are incorporated for control of the maximum power level into the detector and for receiver calibration. Adjustment of the power level into the receiver is through the level-set attenuator in the sweep oscillator at the transmitting site. Characteristics of major components of the receiver are summarized in Table 3.

3. Data Processor

A block diagram of the data processor is given in Figure 13. The data processor is that part of the system which acts on the detector output
### TABLE 2

**COMPONENT SUMMARY*** FOR S-BAND HYBRID SYSTEM TRANSMITTER

<table>
<thead>
<tr>
<th><strong>Sweep Oscillator</strong></th>
<th><strong>TWT Amplifier</strong></th>
<th><strong>Level Sensor</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Power output: 10 mW</td>
<td>Small-Signal Output Capability: 10 watts CW</td>
<td>Frequency Range: 2-4 GHz</td>
</tr>
<tr>
<td>Frequency Range: 2-4 GHz</td>
<td>Nominal Gain: 30 dB</td>
<td>Output Voltage Flatness: ± 0.3 dB</td>
</tr>
<tr>
<td><strong>Sweep Modes:</strong></td>
<td>Frequency Range: 2-4 GHz</td>
<td>Input Power: ≤ 10 W</td>
</tr>
<tr>
<td>Repetitive Sweep between Any Present Limits within 2-4 GHz; Fixed Bandwidth Sweeps about Selectable Center Frequency</td>
<td>Beam Supply Regulation: ± 0.02%</td>
<td>Directivity: ≥ 20 dB</td>
</tr>
<tr>
<td><strong>Non-Sweep CW Mode</strong></td>
<td>Helix Voltage Regulation: ± 0.1%</td>
<td>Line Voltage Variation: ± 10%</td>
</tr>
<tr>
<td>Remote On/Off Control</td>
<td>Interlocks/Interrupts: Time Delay Beam/Helix Overcurrent Thermal Cutout Power Line Interrupt</td>
<td></td>
</tr>
<tr>
<td>Remote CW Frequency Selection</td>
<td>Load VSWR: 2.5:1</td>
<td></td>
</tr>
<tr>
<td>Nominal Power Variation: ± 5 dB (Without External Leveling)</td>
<td>Noise Figure: ≤ 35 dB</td>
<td></td>
</tr>
<tr>
<td>Sweep Linearity: ± 1%</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Sweep Time: Adjustable from 10 msec to 100 sec per sweep</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Level-Set Attenuator</td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Level Sensor</strong></td>
<td></td>
<td></td>
</tr>
<tr>
<td>External Level Accuracy: ± 0.1 dB plus level signal error</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

*Nominal specifications are illustrated.
Figure 12. Hybrid system S-band receiver block diagram
## Table 3
### Component Summary* for S-Band Hybrid System Receiver

<table>
<thead>
<tr>
<th>Component</th>
<th>Switch</th>
<th>Filter</th>
<th>Low-Noise Amplifier</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Switch</strong></td>
<td>Frequency: 2-4 GHz</td>
<td>Passband: 2-4 GHz</td>
<td>Type: Transistor</td>
</tr>
<tr>
<td></td>
<td>Maximum Insertion Loss: 1 dB</td>
<td>Rejection: 40 dB below 1.2 GHz and above 5.2 GHz</td>
<td>Frequency: 2-4 GHz</td>
</tr>
<tr>
<td></td>
<td>Switching Time: ( \leq 0.5 \mu s)</td>
<td>Passband Maximum VSWR: 2:1</td>
<td>Maximum Noise Figure: 4.5 dB</td>
</tr>
<tr>
<td></td>
<td>Isolation: ( \leq 40 ) dB</td>
<td>Insertion Loss: 1 dB max</td>
<td>Minimum Gain: 25 dB</td>
</tr>
<tr>
<td></td>
<td>Type: SPDT with integral driver</td>
<td></td>
<td>Gain Flatness: ( \pm 1 ) dB</td>
</tr>
<tr>
<td><strong>Limiter</strong></td>
<td>Frequency: 2-4 GHz</td>
<td></td>
<td>Power out at 1 dB</td>
</tr>
<tr>
<td></td>
<td>Power Handling: 1 W</td>
<td></td>
<td>Gain Compression: ( \geq 0 ) dBm</td>
</tr>
<tr>
<td></td>
<td>CW Leakage: ( \leq 100 ) mW</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>Insertion Loss @ 0 dBm or less: ( \leq 1.0 ) dB</td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Coupler</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>Insertion Loss: 0.2 dB</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>Coupling: 20 dB</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>Directivity: ( \geq 20 ) dB</td>
<td></td>
</tr>
<tr>
<td><strong>Filter</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Detector</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

*Nominal specifications are illustrated.
Figure 13. Hybrid system data processor block diagram
voltage to provide compensated average gain data. The time-varying signal from the detector represents the frequency dependent response of the antenna under test, plus variations due to any other frequency dependent elements in the entire measurement network from the transmitter through the detector. As previously discussed in Subsection B, the effect of all of these other variations is removed by amplitude compensation in the data processor. The data processor consists essentially of a circuit which integrates the detector output over one sweep period, a sample-and-hold device which samples the integrator output at the end of each sweep period and holds this value until the logarithm can be performed by the log amplifier. The logarithm signal is fed to a sample-and-hold network combined with a subtractor. One of the sample-and-hold modules in this network always contains the most recent value of the logarithm of the reference antenna's average response over the desired frequency range, while the other sample-and-hold contains the most recent value of the logarithm of the test antenna's average response over this sweep range. The subtractor output is the difference of these two log signals stored in the sample-and-hold circuits. This difference (which corresponds to a power ratio) is the integrated (average) response of the test antenna, which is expressed in dB and is referred to the average response of the reference antenna.

A typical transmitter sweep time for the maximum data recording rate would be 10 msec. Therefore, sample-and-hold (S/H) Module #1, which contains the integrator output, must be updated each 10 msec. Since the test and the reference antennas are selected on alternate transmitter sweeps, sample-and-hold (S/H) Modules #2 and #3, which contain the test and reference antenna responses, respectively, must be updated each 20 msec. Since the reference
antenna will be stationary, the output of S/H Module #3 will remain essentially constant for a given frequency sweep range.

Very large antenna positioners have maximum scan rates on the order of 0.5 rpm. Even at this maximum scan rate, in 20 msec the test antenna would rotate through only 0.060 degree. Broadband average antenna response will not change dramatically over this angular increment (broadband pattern characteristics are discussed in the following section). Therefore, even for the maximum angular rotational speed of the antenna positioner, there would not be appreciable changes in successive values of either S/H Module #2 or S/H Module #3 output, and the subtractor output would be a smooth function of time (or test antenna spatial coordinate). The impact of various test antenna characteristics on spatial rotation rate is also discussed in the following section.

The integrator is shown schematically in Figure 14. The integrator consists of a commercially-available low-noise (3 $\mu$V rms) operational amplifier plus the appropriate external resistor and capacitor to obtain the required response time. The output voltage $e_o$ is given by

$$e_o = \frac{1}{RC} \int_0^{t_1} e_i(t)dt,$$

where R and C are defined by Figure 14, $t_1$ is the integration time, and $e_i(t)$ is the input voltage to the integrator. Consider a typical situation diagramed in Figure 15. The upper sketch shows a waveform which might appear at the output of the detector. A voltage which varies in time between $\approx 0$ V and 250 mv is shown. The area under the curve represents an average gain-bandwidth measure of the antenna under test so that the integral of this voltage, which represents the gain-bandwidth measure of the test antenna, might appear as shown in the lower sketch of Figure 15.
Figure 15. Typical detector and integrator waveforms

Figure 14. Integrator for Hybrid system data processor
If $e_i(t)$ has a cosine-type variation over the 10 msec measurement interval and a peak value of 0.250 volt, then $e_i(t)$ can be represented by

$$e_i(t) = 0.125 (1 - \cos 200\pi t). \quad (13)$$

The value at the output of the integrator after a 10 msec period is given by

$$e_o = \frac{1}{RC} \int_0^{0.01} 0.125 (1 - \cos 200\pi t) \, dt = 0.00125/RC. \quad (14)$$

It is desirable that the integrator output voltage $e_o$ be in the optimum range of 1 mv to 10 volts, which is necessary for the subsequent operation of the S/H Module #1 and log amplifier. Therefore, the RC product must be selected and controlled by the operator in order to provide an optimum voltage range for a wide range of frequency sensitive antennas that may be tested. For the present example involving a cosine-type detector output function, an integrator output of 10 volts is desired when the input voltage covers the 0-250 mv range. From Equation (14), the RC factor is calculated to be

$$RC = \frac{.00125}{10v} = 1.25 \times 10^{-4}.$$ 

Therefore, selecting $R = 10^4$ ohms and $C = 1.25 \times 10^{-8}$ farads will satisfy the condition on the output voltage range.

The sample-and-hold Module #1 must sample the integrator output at the end of each integration period and hold that value until the end of the following integration period. Standard commercially available modules that are well suited for this type of application are available. For example, one suitable unit has an acquisition time (time necessary to acquire the input signal and approximately settle to its steady-state value) of 15 µsec and an output voltage droop factor of 5 µV/msec. This value of acquisition time is short compared to changes in the integrator output so that no error is introduced. Although the droop factor must be considered, a voltage droop of 5 µV/msec over a 10 msec period will only be 50 µV. Therefore, because the
signal voltages cover four decades (40 dB) from 1 mV to 10 V, the maximum percentage droop will occur for the minimum signal (1 mV) and it will be only 5%.

The accuracy and dynamic range of standard logarithmic amplifiers (log amp) can be considerably improved by operating them in a current input mode rather than in a voltage input mode. Consequently, a voltage-to-current converter (VCC) is utilized as the input element to the logarithmic amplifier, as shown in Figure 16. Both the operational amplifier in the VCC and the entire log amp are standard commercial products. The voltage to current conversion is expressed by

\[ i_L = -\frac{e_i}{R_2}, \text{ provided} \]
\[ R_3/R_2 = R_F/R_1, \]

where

- \( i_L \) is the current input to the log amp,
- \( e_i \) is the voltage input to the VCC, and
- \( R_2, R_3, R_F, \) and \( R_1 \) are indicated in Figure 16.

Four decades of input current (corresponding to the four decades of signal voltages of 1 mV to 10 V) to the log amp are required. Since the midrange of a typical log amp's input capability is from \( 10^{-4} \) amperes to \( 10^{-8} \) amperes, and since \( e_i \) will range from 1 mV to 10 V, from Equation (15) \( R_2 \) should be \( 10^5 \) ohms. The log amp output voltage range will be from -6 V to +2 V for the \( 10^{-8} \) amperes to \( 10^{-4} \) amperes current input, respectively. This -6 V to +2 V is shifted into the 0-10 V range by an operational amplifier with a fixed bias input.

Sample-and-hold Modules #2 and #3 are identical to S/H Module #1. At the end of the test and reference antenna integration intervals, S/H Module #2 and S/H Module #3, respectively, sample the log output and hold the values
Figure 16. Logarithmic amplifier with voltage-to-current converter
until each respective output is updated on alternate transmitter sweeps. The output value of the subtractor, which is a standard operational amplifier module, typically ranges from 0 to 10 volts.

For use with the Hybrid system, the Scientific-Atlanta 1520 Rectangular Pattern Recorder would be equipped with the Series 1556 dc-chopper preamplifier. When so equipped, dc input voltages of up to 100 volts are permitted. Full scale sensitivity is 0.01 volt and the noise level is 10 μV rms. Thus, the 0-10 V input signal will interface well with this recorder.

Characteristics of major components of the data processor are summarized in Table 4. It should be noted that the data processor is not frequency dependent, and it can be used for various RF ranges.

4. Timing and Control

Five control lines are required to synchronize and control the operation of the various data processor modules. These lines control the front end switch, the integrator, and the three sample/hold modules. The front end switch selects either the calibrated reference antenna or the test antenna. The integrate signal starts and stops the integration of the detected video signal, and the sample/hold lines select and store the appropriate signals from the integrator and the log amp.

In Figure 17, each of the signals between processing modules is illustrated schematically in the top half of the figure, and the control signals on each of the five control lines are shown in the bottom half of the figure. A transmitter sweep rate of 100 Hz is indicated by the video signal directly from the detector. The reference antenna integrated response is stored at t = 10 ms, and the test antenna integrated response is stored at t = 20 ms. The control signals SH2 and SH3 transfer the voltages at the log amp output to either the subtractor (+) port or the subtractor (-) port. These inputs
TABLE 4

COMPONENT SUMMARY* OF HYBRID** SYSTEM DATA PROCESSOR

<table>
<thead>
<tr>
<th>Operational Amplifier (1)</th>
<th>Sample-and-Hold</th>
<th>Logarithmic Amplifier</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output: ±10 V @ 5 mA</td>
<td>Acquisition Time: 5 μsec to 0.01% of steady state</td>
<td>Rated Output: ±10 V @ 5 mA</td>
</tr>
<tr>
<td>Input Impedance: 10^11 ohms (3.5 pF)</td>
<td>Droop Rate: 5 μV/msec</td>
<td>Accuracy of Log: ≤ 1%, referred to input</td>
</tr>
<tr>
<td>Input Noise: Voltage, 3 μV rms; Current, 0.1 pA peak-to-peak</td>
<td>Input Range: ±10 V</td>
<td>Dynamic Range: 60 dB, 10^-9-10^-3 amp</td>
</tr>
<tr>
<td>Input Voltage: +8 to -10 V</td>
<td>Input Impedance: 10^9 ohms</td>
<td>Aperture Delay: 40 nsec</td>
</tr>
<tr>
<td>Operating Temperature Range: 0 to +70°C</td>
<td>Aperture Jitter: 4 nsec</td>
<td></td>
</tr>
</tbody>
</table>

*Nominal Specifications are illustrated.

**Used in integrate, log, and subtract functional modules.
Figure 17. Illustration of Hybrid system timing and control requirements
at the subtractor lead to the self-calibration of the measured data on a sweep-by-sweep basis.

The five control signals are derived from the basic square wave signal controlling the front end switch. The integrator control signal, SH2, and SH3 are simple monostable multivibrator output signals which are triggered with appropriate delays relative to the front end switch signal. SH1 is a square wave at twice the rate of the square wave which controls the front end switch.

The front-end-switch signal may be generated in several ways. The simplest way, conceptually, is to derive the switch signal from the transmitter sweep waveform by transmitting the timing signal via a direct data link of suitable capacity or by telephone service with modems at both sites. However, for mobility and field use it is desirable that operation of the Hybrid system not be dependent on such a link, and an alternate approach has been selected. With the alternate approach, the basic timing signal is synchronized with the test and reference transmitter sweeps by detecting the beginning of the sweep cycles from a sample of the receiver signal. This detected sample drives a Schmidt Trigger circuit which forms a rectangular pulse which is shaped to be a square wave and is appropriately delayed to drive the front end switch between transmitter re-sets. Synchronization is maintained by detecting the leading edge of each transmitter sweep cycle.

Pushbutton switches for manual selection of the test or reference antenna position are required for the single frequency mode of operation.

5. Transmit and Reference Antennas

As described in Appendix I-C, broadband multipolarized antennas suitable for the hybrid system transmit and reference antennas are available. For a feasibility demonstration, it is expected that the initial 2-4 GHz system will employ relatively simple transmit and reference antennas and that
they will be "identical." A broadband linearly polarized feed with a parabolic reflector is a practical choice which can meet the requirements of this demonstration system. Furthermore, the system could be upgraded to a multipolarized system by replacement of the feeds.

As indicated in the sensitivity analysis, the reflector should be sized to provide a gain of approximately 30 dB at 4 GHz. The power gain of a circular aperture with a typical illumination function (-25 dB first sidelobe) can be expressed as

\[
G = K \frac{A \cdot 4\pi}{\lambda^2} \tag{16}
\]

where

- \( A \) = aperture area,
- \( \lambda \) = wavelength, and
- \( K \) = efficiency factor.

Expressing aperture area in terms of diameter \( d \), and solving for \( d \) at a wavelength of 2.95 inches (\( f = 4 \) GHz), an efficiency of 0.5, and a gain of 1000 (30 dB), one gets \( d \approx 42 \) inches. The 3-dB beamwidth achieved with this illumination function is about 5 degrees at 4 GHz.

The antenna feed can be simply an open ended ridged waveguide with a waveguide-to-coax adapter for interface to the transmitter and receiver. The use of ridged waveguide will offer improved broadband matching over that obtainable with conventional rectangular guide. For example, a VSWR of 1.06:1 is typical for a coaxial to dual-ridged transformation over the 2-4 GHz range.

This simple antenna system will provide a cost-effective approach for realization of the initial 2-4 GHz hybrid system.

E. CALIBRATION

Since the Hybrid system provides integrated broadband test antenna data which are referred to the broadband response of a reference antenna, proper
Interpretation of the test data requires knowledge of the broadband properties of the reference antenna. The reference antenna can be calibrated versus a set of standard gain antennas covering the measurement range of the particular Hybrid system. Calibration of a 2-4 GHz reference antenna is described here. Calibration for other frequency ranges would be similar. The experimental setup is sketched below.

The reference antenna should be calibrated (have its absolute gain measured) at small frequency increments across the frequency range for which it is to serve as the reference. Since the gain of the reference antenna may not be a linear function of frequency, it is important to have a relatively large number of samples from which to determine the average gain over any desired bandwidth. Calibration of the reference antenna gain at each 100 MHz interval across the 2-4 GHz range will be assumed. Since standard gain horn antennas cover waveguide bands, two standards will be required to cover the 2-4 GHz range: one covering from 2.0 to 2.60 GHz, and one covering from 2.60 to 4.0 GHz. Each standard would be supplied with a gain curve showing its gain across its entire bandwidth.
The reference antenna would then be calibrated as follows. The antenna which is to be calibrated as a reference and one of the standard gain horns are mounted at the receiver site. The Hybrid system transmitter and receiver are operated in their non-sweep mode, sequentially covering a set of frequencies separated by 100 MHz across the range of the standard gain horn. At each of these frequencies, the gain of the antenna under calibration is determined relative to the gain of the standard. This procedure is followed with the second standard also until a set of absolute gain values is obtained for the reference antenna every 100 MHz \((G_{2.0}, G_{2.1}, \ldots, G_{3.9}, G_{4.0})\) across the 2-4 GHz range. Now, when this calibrated antenna is used as the reference in broadband measurements on an unknown antenna, its known gain at some frequency within the desired measurement range is used as the reference to convert measured data to absolute data. Average gain could also be calculated by converting the dB data into linear form and taking their average. The average gain of the reference antenna across the 3.0 to 3.5 GHz range would be

\[
\bar{g}_{\text{ref}} = \frac{g_{3.0} + g_{3.1} + g_{3.2} + g_{3.4} + g_{3.5}}{5},
\]

where \(g\) represents a respective gain in absolute (non-dB) units.

**F. SYSTEM ACCURACY**

Since the logarithmic amplifier is expected to be the least accurate element in the hybrid system, the impact of errors in the log module has been examined. As indicated previously, a consequence of the electronic compensation scheme is the improvement of overall system accuracy. This improvement is illustrated by the following analyses of the impact of log module errors on system accuracy.

The log module ideally performs the following transform

\[
V_{\text{out}} = -2 \log_{10} \left( \frac{V_{\text{in}}}{V_{\text{ref}}} \right),
\]
where \( V_{\text{ref}} \) is a fixed reference voltage supplied to the log amp and \( V_{\text{in}} \) is the input voltage. The transform error is taken into account by the following equation.

\[
V_{\text{out}} + \epsilon_o = -2 \log_{10} \left[ \frac{(V_{\text{in}} + \epsilon_i)}{V_{\text{ref}}} \right], \tag{19}
\]

where \( \epsilon_o \) accounts for departures from ideal \( V_{\text{out}} \) for various errors \( \epsilon_i \) in the input. The error at the output of the log module (effectively, the input to the subtractor) has been calculated for input errors of \( \pm 5\% \) and \( \pm 10\% \) and tabulated in Table 5 for a reference voltage of 100 mV.

In this error analysis, log amp errors of \( \pm 10\% \) and \( \pm 5\% \) referred to the log input voltage were assumed. The system output error was calculated at representative points across the 1 mV to 10 V log module input range. The assumptions for this estimate are the following:

1. the log module error is defined in terms of the input voltage to the module,
2. the overall processor error is defined in terms of the output voltage from the subtractor,
3. no errors are considered except log module error,
4. for worst case conditions, a positive extreme error is assumed for 1 mV input, no error is assumed for 100 mV input, and negative extreme error is assumed for 10 V input (40 dB dynamic range of interest), and
5. the log module transfer function is stationary, not time-varying.

The error at the output of the log module (effectively, the input to the subtractor) has been calculated for input errors \( \epsilon_i \) of \( \pm 5\% \) and \( \pm 10\% \) and tabulated in Table 5 for a reference voltage of 100 mV.

The subtractor output voltage is equal to the difference between successive measurements of the gain-bandwidth product for the test and reference
### TABLE 5

**ACTUAL LOG AMP OUTPUT VOLTAGE**

<table>
<thead>
<tr>
<th>$V_{in}$</th>
<th>$e_i = 0$</th>
<th>$e_i = +5%$</th>
<th>$e_i = -5%$</th>
<th>$e_i = +10%$</th>
<th>$e_i = -10%$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 mV</td>
<td>+4</td>
<td>3.958</td>
<td>4.045</td>
<td>3.917</td>
<td>4.092</td>
</tr>
<tr>
<td>100 mV</td>
<td>0</td>
<td>-.042</td>
<td>.045</td>
<td>-.083</td>
<td>.092</td>
</tr>
<tr>
<td>10 V</td>
<td>-4</td>
<td>-4.042</td>
<td>-3.955</td>
<td>-4.083</td>
<td>-3.980</td>
</tr>
</tbody>
</table>

### TABLE 6

**SUBTRACTOR OUTPUT VOLTAGES**

<table>
<thead>
<tr>
<th>$V_2$</th>
<th>$e_i$</th>
<th>$V_1 - 1$ mV ($e_i = +10%$)</th>
<th>$100$ mV ($e_i = 0$)</th>
<th>$10$ V ($e_i = 10%$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 mV</td>
<td>+10%</td>
<td>0</td>
<td>-3.917(2.1%)</td>
<td>-7.825(2.2%)</td>
</tr>
<tr>
<td>100 mV</td>
<td>0</td>
<td>3.917(2.1%)</td>
<td>0</td>
<td>-3.908(2.3%)</td>
</tr>
<tr>
<td>10 V</td>
<td>-10%</td>
<td>7.825(2.2%)</td>
<td>3.908(2.3%)</td>
<td>0</td>
</tr>
</tbody>
</table>
antennas. The ideal equation for the subtractor output is

\[ S_{out} = -2 \log_{10} \left( \frac{V_1}{V_{ref}} \right) + 2 \log_{10} \left( \frac{V_2}{V_{ref}} \right), \]

where \( V_1 \) and \( V_2 \) represent the integrated detector voltages due to the reference antenna and the antenna under test, respectively. Based on the worst case assumption (4), the subtractor output voltage and the percent error for each error case have been calculated. These data are given in Table 6.

The errors are expressed as a percent of subtractor output with no errors in the log module. It may be seen that with the same gain-bandwidth product for the test and reference antennas \( (V_1 = V_2) \), there is no error in subtractor output, regardless of the percent of log transform error. The maximum error (2.3%) occurs for the maximum difference in gain-bandwidth response of the two antennas. The log transform error of 10% is larger than that expected when operating the log module in the current mode under nominal conditions. However, changes in operating temperature, bias voltage drifts, etc., could lead to errors of this magnitude. Even under these conditions, the error in the system output is not large. Laboratory tests will be required to establish actual system accuracy under a variety of conditions. Under all conditions, when \( V_1 \) and \( V_2 \) are equal, no error occurs in the subtractor output.

To obtain maximum accuracy from the Hybrid system, there are certain operational constraints which must be observed. The system has been designed for optimum operation over specific power levels and dynamic ranges. For example, the power level into the detector should not be more than 0 dBm maximum, at the peak gain of the test and reference antennas. In order to avoid saturating the system, the transmitter power level should be reduced to meet this requirement. In addition, the peak detector output voltages should be approximately equal for the test and the reference antennas. This requirement means that the test and reference antennas should be oriented for their
maximum gains, and the power level equalized through use of a broadband attenuator immediately following the higher gain antenna. Of course, this attenuation value must be accounted for in converting the measured relative gain into absolute gain.

As indicated previously, the Hybrid system is configured to accommodate a typical maximum positioner rotation rate of 0.5 rpm with a resulting spatial uncertainty of 0.030 degree in a 10-msec sweep time. If more precise spatial data are desired, the antenna rotation rate can be reduced with an increase in data-recording time. For example, if the rotation rate is reduced to 360 degrees/one-half hour, the resulting spatial uncertainty is 0.002 degree.

G. DATA

The Hybrid system is designed to produce broadband-average gain and pattern data as well as CW data at any selected frequency within its operating range. For single frequency operation, the sweep oscillator is switched to CW and tuned to the desired frequency. (Tuning may be either manual or remote if a data link between the transmit and receiving sites is available.) At the receiver site, timing of the processor functions continues but transmitter sync is not required. The basic 100-MHz timing signal is generated by a free running oscillator. This timing feature and CW operation are selected by pushbutton control and data to the pattern recorder remains in dB form and is fully compensated and normalized to the reference antenna.

In its primary mode of operation, the Hybrid system produces average gain/pattern information over a selected bandwidth. This bandwidth can be from zero up to an octave. The type of data produced is discussed and illustrated in Section IV-A, which is concerned with applications of swept-frequency phase/amplitude systems.
SECTION IV
CONCEPT II SYSTEMS

A. INTRODUCTION

Concept II systems are based on the use of a broadband tunable signal source and a synchronously tuned receiver to provide sampled data which can be either displayed in real time or stored on magnetic tape for diagnostic processing. Because of the large amounts of data which are involved, Concept II systems lend themselves naturally toward an automatic computer-controlled implementation. Initially, both amplitude-only and amplitude-plus-phase Concept II systems were considered. However, it was found that except for no receiver requirement to measure phase, the implementation of a Concept II swept-frequency sampled-data amplitude-only system requires equipment similar to that needed for the implementation of a Concept II phase-plus-amplitude system. However, because broadband amplitude-only data are provided by the less complex Hybrid system, development of a Concept II type of system would not be an effective utilization of resources for instrumentation if the end item is to be used strictly for obtaining amplitude-only data. Therefore, only the phase-plus-amplitude Concept II systems were further investigated.

Phase information can be important in determining the performance of very broadband systems. However, the importance of the phase information depends on the application. For example, both phase and amplitude information would be necessary in determining the antenna effect upon pulse shape or in defining pulse compression performance limitations. A new area in which phase information would be useful is the application of measured target backscatter phase characteristics to aid in target identification. In addition, if the antennas phase properties are known, compensation networks can
be added to achieve more nearly ideal system performance. Areas where phase information is not required is in determination of broadband average gain and broadband spatial patterns. An application of the broadband spatial pattern is to determine the received power from a sidelobe noise jammer so that the system signal-to-noise (interfering) ratio can be found. For this application, the RMS sidelobe levels over the frequency band of the interfering signal may be used to compute the interference power level. The RMS sidelobe levels can be determined over the desired bandwidth from amplitude-only measurements. Complete amplitude and phase information in the sidelobe regions is required if the antenna effect on the pulses received through a sidelobe is of interest. Since the distortion of pulses received through the sidelobes is not usually of interest, because the radar system is designed for main-beam operation, it appears that sidelobe phase information is of limited practical value.

The types of antenna analyses which can be performed using amplitude-only and amplitude-plus-phase data can be examined analytically. If amplitude-only data are used, mathematically the patterns which are obtained are equivalent to those which would be obtained with a band limited white noise measurement system. The spatial pattern for this case may be expressed as

$$ P(\theta, \Delta \omega) = \int_{\Delta \omega} |G_v(\omega, \theta)|^2 \hat{s}(\omega) \, d\omega , $$

(21)

where

$$ G_v(\omega, \theta) = \text{complex voltage gain}, \text{ and} $$

$$ \hat{s}(\omega) = \text{incident signal spectrum}. $$

That is, the average received power over a band of frequencies can be obtained by integrating the power gain of the antenna over the frequency band
of interest, weighted by the incident-signal spectrum. Of course, the antenna gain at a single frequency, or as a function of frequency for any given spatial angle, can also be extracted from measured amplitude data. Figures 18 through 20 show the effect on the antenna spatial pattern of increasing the integration bandwidth for a rectangular aperture antenna with uniform illumination. For small bandwidths, the pattern does not deviate significantly from the monochromatic pattern. However, as the bandwidth is increased, the pattern nulls are obscured, and the sidelobe peaks approach their average values. This type of information from amplitude-only data can be used in determining the susceptibility of the antenna to sidelobe noise jammers.

In order to account for pulse distortion effects, the complex voltage gain of the antenna must be known as a function of frequency. From these data, the received pulse waveform can be computed if the spectrum of the incident pulse is known. Thus, pulse distortion affects can be computed as a function of spatial angle. This may be expressed as

\[ V_{rcv}(\theta, t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} G_v(\theta, \omega) \tilde{\phi}_i(\omega) e^{j\omega t} d\omega \]  

where

- \( V_{rcv} \) = received pulse voltage at spatial angle \( \theta \), as a function of time,
- \( G_v(\theta, \omega) \) = complex voltage gain of antenna,
- \( \tilde{\phi}_i(\omega) = \int_{-\infty}^{\infty} \varphi(t) e^{-j\omega t} dt \), and
- \( \varphi(t) \) = incident pulse.

The antenna effect on various pulse shapes \( \varphi(t) \) can be computed from the above equation after the required amplitude and phase data are measured for
Figure 18. Broadband pattern for uniformly illuminated rectangular aperture, 1 GHz bandwidth
Figure 19. Broadband pattern for uniformly illuminated rectangular aperture, 2 GHz bandwidth
Figure 20. Broadband pattern for uniformly illuminated rectangular aperture, 5 GHz bandwidth.
As discussed in Section I-C, two techniques for phase measurement have been considered. Instrumentation of both techniques is feasible with the current state-of-the-art. Adaptation of the conventional separate reference channel technique for sweep frequency applications can be accomplished through modification of the existing Scientific-Atlanta Series 1750 receivers to incorporate phase-locked frequency tracking between the transmitter and the receiver. The group velocity concept can be instrumented through the addition of sideband processing circuits to the modified Series 1750 receiver. Thus, although the instrumentation of a sweep frequency phase/amplitude system is currently feasible, it may be seen from Figures 5 and 6 that a great deal of relatively complex equipment is required. In addition, implementation of a sweep frequency phase/amplitude system requires a system analysis, a definition of computer interface requirements, a software development, an analysis of system calibration requirements, and a definition of system calibration procedures. Thus, a relatively modest expenditure of time and money is required to implement the desired phase/amplitude system. Accordingly, the approach which has been defined through the studies under this contract is based on (1) the completion of the preliminary design of the relatively inexpensive Hybrid system to the component level so that it can be implemented immediately, and (2) the completion of the feasibility investigation of the phase/amplitude system. This second item includes the evaluation of the application of both the separate reference channel direct phase measurement technique and the group velocity measurement technique, to the broadband antenna measurements problem, the investigation of calibration requirements and techniques, the estimation of system accuracy, the preliminary system
design (not to the component level), and the definition of critical areas where breadboarding may be desired. The Hybrid system design has been presented in Section III, and the results of the phase/amplitude system investigations are presented in this section.

**B. SEPARATE REFERENCE CHANNEL CONCEPT**

As indicated above, implementation of the separate reference channel phase/amplitude system would employ a modified version of the Scientific-Atlanta Series 1750 receiver. These receivers measure amplitude and phase of a test signal, relative to a reference channel signal. For complete antenna diagnostics, both amplitude and phase information are required. Amplitude versus frequency information which is relative to absolute gain versus frequency can be provided by use of a standard gain antenna in the reference channel. Measurement of the test antenna's phase properties places additional phase calibration requirements on the reference channel. These additional calibration requirements are defined in the following paragraphs.

Conventional laboratory phase measurements may be performed with a setup such as that sketched below.
Phase measurements may be grouped into three categories - insertion, incremental, or comparative. In an insertion phase measurement, the relative phase shift between the test channel and the reference channel is determined both before and after the test device is inserted in the channel. The difference in the two readings is the insertion phase of the device for that particular set of conditions. In incremental measurements, the test device is inserted in the test channel, and relative phase between the test and reference channels is measured as some parameter, e.g. frequency, is changed. Since only incremental phase shift is of interest in this measurement, zero is arbitrarily established at some point in the measurement. In a comparative measurement, phase shift of the test device is determined relative to that of a designated standard.

In the antenna applications under consideration here (pulse diagnostics, antenna effect on FM pulse compression, etc.), it is the test antenna's incremental phase shift over a specified frequency range which is of interest. The measurement setup for measuring the incremental phase shift of the test antenna is diagramed below. It will be shown that to determine incremental phase shift of the test antenna versus frequency, incremental phase shift properties of the reference channel must also be known, and any difference between physical path length in the test and reference channels must be accounted for.
The symbols to be used in defining the calibration requirements are illustrated in the above test setup and are defined below:

- \( I_T \) = insertion phase of test antenna from aperture plane to antenna terminals,
- \( I_R \) = insertion phase of reference antenna from aperture plane to antenna terminals,
- \( L_T \) = physical length from test antenna terminals to receiver input, and
- \( L_R \) = physical length from reference antenna terminals to receiver input.

To measure the variation of \( I_T \) with frequency, the calibration requirements for the reference channel must be defined. The absolute value of \( I_T \) is not required; that is, the phase measurement of interest here is the manner in which \( I_T \) varies as a function of frequency across some specified frequency range.

At some beginning frequency, the measured phase difference between the test channel and the reference channel, \( \Delta \phi_1 \), can be expressed as

\[
\Delta \phi_1 = \frac{\left[I_T(1) + \frac{2\pi}{\lambda_1} L_T - \phi R(1)\right]}{N},
\]

where \( \phi R(1) \) is total phase delay in the reference antenna plus the transmission line \( L_R \), and the integer \( N \) accounts for the multiple-of-360° ambiguity in the measured phase difference. If the frequency is now changed by some amount to \( \omega_2 \) such that the measured \( \Delta \phi \) does not change by more than 360°, then,

\[
\Delta \phi_2 = \frac{\left[I_T(2) + \frac{2\pi}{\lambda_2} L_T - \phi R(2)\right]}{N}.
\]

Combining these two expressions yields

\[
\Delta \phi_1 - \Delta \phi_2 = I_T(1) - I_T(2) + \frac{2\pi L_T}{\lambda_1} - \frac{2\pi L_T}{\lambda_2} - \phi R(1) + \phi R(2).
\]

By extending the method used to derive this last expression to additional frequencies, it is clear that the change in measured phase difference
as a function of frequency change can be expressed as

\[ \Delta \Phi(w) = \Delta L_T(w) + \Delta \Delta L_T(w) - \Delta \Phi_R(w) \] (26)

That is, an increase in measured phase difference between the test and reference channels is due to increased phase delay in the test antenna plus an increased electrical length \(\Delta \Delta L_T\) of the test antenna line link, minus any increased phase delay in the reference antenna channel. As written, this equation is algebraically correct for increased phase delay in any term of the equation. A decreased phase delay would change the sign of that term.

The calibration required to determine \(\Delta L_T(w)\) from the measured \(\Delta \Phi(w)\) may be deduced from the last expression above. The change in electrical length of the test antenna transmission line \(L_T\) must be known as a function of frequency, and the phase characteristics of the reference channel must be known as a function of frequency. Ideally, the transmission line from the test antenna to the receiver input should have a linear phase shift versus frequency property, and the reference antenna-plus-transmission link \(L_R\) should be calibrated as a unit. In addition, any difference in \(L_T\) and \(L_R\) should be small to prevent large excursions in the phase difference between the test and the reference channels as the frequency is changing. Under these conditions, any measured phase dispersion would be the phase dispersion of the test antenna minus the dispersion of the reference channel. Methods of phase calibrating a reference antenna are discussed in Sub-section IV-D.

C. GROUP VELOCITY CONCEPT

Systems first considered for satisfying the program objective involved measuring the phase versus frequency directly; however, these systems required a separate microwave transmission channel to transmit a phase
coherent replica of the test signal between the antenna range transmitting
and receiving sites. Because the requirement of a phase reference channel
may limit the flexibility of these systems, an approach which does not re-
quire transmitting a separate replica of the test signal for making phase
measurements has also been considered. This approach is based on the concept
of group velocity.

Consider the following exponential form for a traveling EM wave
\[ E = A e^{j(\omega t - \beta z)} \]  (27)
where
- \( \omega \) = angular frequency, and
- \( \beta \) = phase constant.

For a single frequency, all points of the wave travel, in a homogenous media,
with equal phase velocity \( v_p \) \( (v_p = \omega/\beta) \) and the shape of the wave does not
change with time. However, to represent an arbitrary periodic time function,
an infinite number of monochromatic waves is required. The shape of such a
composite wave is no longer preserved in time unless the phase velocity,
i.e., the ratio of \( \omega \) to \( \beta \), remains the same for all constituent waves of the
time function. If this is the situation, the propagating medium is called
non-dispersive.

For dispersive media, the phase velocity does depend on \( \omega \). Dispersive
and non-dispersive media may be further distinguished through the concept of
group velocity. Group velocity may be defined as the phase velocity of a
wave envelope, i.e., the velocity of a constant phase point on the envelope.
Group velocity, \( u \), is related to phase velocity by
\[ u = v_p - \lambda \frac{d\beta}{d\lambda} = \frac{dw}{d\beta} \]  (28)
For a non-dispersive media, \( dv_p/d\lambda = 0 \), and the group velocity equals the
phase velocity. Since \( dw/d\beta \) is the slope of the tangent to an \( \omega \) versus \( \beta \) curve
at a point, both \( \frac{d\omega}{d\beta} \) and \( \omega/\beta \) may be graphically illustrated through an \( \omega \cdot \beta \) diagram, such as illustrated in Figure 21. It is clear from the diagram that the slope of a straight line connecting an arbitrary point on the curve with the origin gives the corresponding phase velocity, \( v_p = \omega/\beta \). In non-dispersive media, the \( \omega \cdot \beta \) curve is a straight line through the origin, and \( v_p = u \). However, for the arbitrary curve of Figure 21, the slope of the straight line to the point P is quite different from the slope of the curve at that point.

Thus, recapitulating, in a non-dispersive medium, \( v_p = u \) and the shape of a pulse, such as radar pulse travelling through free space, does not change. In a dispersive medium, such as a dispersive delay line used in pulse compression, the shape of the pulse changes. It is desirable, therefore, to determine dispersive properties of all radar system components to ensure that pulse distortions due to unwanted dispersion do not occur. The system under consideration arrives at a measure of dispersion through a measurement of sideband differential phase of a single-tone modulated carrier.

For illustration consider the following expression for an amplitude modulated wave:

\[
e(t) = (E_c + m E_c \cos \omega_a t) \cos (\omega_c t + \xi),
\]

(29)

where \( e(t) \) = instantaneous amplitude, 
\( t \) = time, 
\( m \) = modulation index \( < 1 \), 
\( E_c \) = peak carrier amplitude, 
\( \omega_a \) = amplitude modulation radian frequency, 
\( \omega_c \) = carrier radian frequency, and 
\( \xi \) = arbitrary phase difference between carrier and modulating signals.

*Other modulation methods which preserve phase coherency of resulting sidebands can be used. AM is used here for simplicity in the resulting equation.
Figure 21. Schematic $\omega - \beta^\prime$
This equation can be trigonometrically expanded and rewritten as
\[ e_c(t) = E_c \cos (\omega_c t + \delta) + \frac{mE}{2} \cos (\omega_c t - \omega_a t + \delta) + \frac{mE}{2} \cos (\omega_c t + \omega_a t + \delta). \] (30)

The last two terms represent the lower sideband (LSB) and the upper sideband (USB), respectively:
\[ \text{LSB} = \frac{mE}{2} \cos (\omega_c t - \omega_a t + \delta), \text{ and} \]
\[ \text{USB} = \frac{mE}{2} \cos (\omega_c t + \omega_a t + \delta). \] (31) (32)

The instantaneous phase of these sidebands is represented by
\[ \Phi(\text{LSB}) = \omega_c t - \omega_a t + \delta, \text{ and} \]
\[ \Phi(\text{USB}) = \omega_c t + \omega_a t + \delta. \] (33) (34)

In the present application, the test antenna will delay these sideband phases. The phase delay introduced may be expressed as \( \Phi_u \) and \( \Phi_\ell \) for the upper and lower sidebands, respectively. Modifying Equations 33 and 34 to incorporate the phase delays yields
\[ \Phi(\text{LSB}) = \omega_c t - \omega_a t + \delta + \Phi_\ell, \text{ and} \]
\[ \Phi(\text{USB}) = \omega_c t - \omega_a t + \delta + \Phi_u. \] (35) (36)

Although sideband phase information may be processed in various ways to obtain a measure of antenna dispersion, it is desired to obtain phase delay versus frequency data such as illustrated below.
The upper and lower sidebands are separated, and a measure of the difference in sideband phase delay $\Delta \varphi_d$ for a given sideband separation $\Delta \omega$ ($\Delta \varphi_d/\Delta \omega$) is obtained.

This $\Delta \varphi_d/\Delta \omega$ is the average group delay over the range $\Delta \omega$. If the $\varphi_d$ versus $\omega$ curve is linear over $\Delta \omega$ with constant slope $\Delta \varphi_d/\Delta \omega = -\tau$, then the pulse is simply delayed by $\tau$ with no distortion. Therefore, the interest lies in measuring the antenna frequency response $H(j\omega)$ which represents any deviation of antenna response from constant amplitude and linear phase over the bandwidth of interest (the subscribed bar indicates that $H$ is a complex quantity). The conventional phase measurement approach which employs a separate reference channel measures $\varphi$ versus $\omega$ directly, whereas the differential sideband measurement provides $\Delta \varphi/\Delta \omega$ so that an integration is required to obtain $\varphi(\omega)$.

Without a delay-calibrated reference path, one could either attempt to measure $\frac{d^2 \varphi}{d\omega^2}$ and integrate twice or to measure $\frac{d \varphi}{d \omega}$ and integrate once to obtain $\varphi(\omega)$. In practice, the first and second derivatives are approximated by first and second differences, designated $\varphi'$ and $\varphi''$. These approximations are defined by

$$\varphi' = \frac{\Delta \varphi}{\Delta \omega}, \quad \text{and}$$

$$\varphi'' = \frac{(\varphi'_1 - \varphi'_2)}{\Delta \omega}, \quad (37)$$

where $\varphi'_1$ and $\varphi'_2$ are successive values of $\varphi'$ taken at intervals of $\Delta \omega$. This approximation is necessary because with the modulation, or sideband, approach a finite sideband separation is required.

Without an independent link to the modulation source, it is possible to measure $\varphi'$ in a dynamic measuring process in which the frequency band is swept continuously at a linear rate. With a link to the modulation source, it is possible to measure $\varphi'$ directly by static measurements. Successive
measurements at incrementally spaced frequencies over the frequency band are required, but these can be made over arbitrarily selected time increments as long as the modulation reference is sufficiently stable. A link to the source can be obtained by hard-wiring, by a microwave or laser link, or by synchronization of oscillators at the transmitting and receiving sites with WWVB. To achieve the desired accuracy, resolution, and stability, the measuring system proposed herein employs a modulating-tone reference from the source. For relatively short ranges, a coaxial cable transmitting the modulation reference (5-MHz) is a practical approach. This cable does not have to be phase calibrated, but its delay characteristic should be stable over the measurement time. For long ranges or if a direct link is impractical, common synchronization of both the modulating signal at the transmitter and the modulating reference signal at the receiver with WWVB is the preferred alternative.

Because the measured phase delay will represent the total phase delay from the transmitter to the receiver, determining the test antenna phase delay will require calibration of the phase delay due to other elements in the network. These elements include the signal source, the transmission line from the source to the transmitting antenna, the transmitting antenna, the transmission line from the test antenna to the receiver, and a reference antenna. Calibration methods are discussed in the following subsection.

D. CALIBRATION METHODS

Consider the general phase measurement system diagrammed in Figure 22. This figure illustrates all the signal requirements for measurements of antenna phase response either by direct measurement of \( \varphi(\omega) \) or by measurement of group delay \( \varphi'(\omega) \). The direct phase measurement employs a two-channel phase-amplitude receiver with a reference channel input (channel B) in addition to the test signal input. The signal source is swept continuously across the
Figure 22. General antenna measurement system
desired frequency range with no modulation. By ensuring an adequate signal to noise ratio, the reference channel input can also be used for frequency tracking. The receiver tracks the sweep signal at a rate up to 10 MHz/msec. No additional tracking or reference inputs to the receiver are required for this type of measurement. The receiver outputs are the test frequency and the amplitude and phase response of the test antenna. The amplitude and phase data are relative to the reference channel response, and they must be corrected by calibration data for the reference antenna.

For group delay measurements, the channel B reference input of the receiver is terminated, the signal source is modulated with a 5-MHz (tentative) modulation tone, and the carrier frequency is stepped sequentially in, for example, 100-MHz steps, and external frequency tracking and modulation reference signals are provided. A separate frequency tracking channel is used since the amplitude of the carrier signal through the test antenna may vary rapidly as the test antenna is rotated. The output of the sideband processor is the sideband differential phase delay.

The measured complex (phase and amplitude) response, $G$, of the complete system may be generally represented by

$$G = F_S H_C H_S H_R H_D,$$  \hfill (39)

where

- $F_S = F_S(jw)$ is the response of the swept frequency signal source,
- $H_C = H_C(jw)$ is the response of the transmission path connecting the source with the transmitting antenna,
- $H_S = H_S(jw, \phi_0, \theta_0)$ is the response of the transmitting antenna at the peak $(\phi_0, \theta_0)$ of its beam pattern, assumed to be directed toward the receiving antenna(s),
- $H_R = H_R(jw, \phi, \theta)$ is the response of the receiving antenna, and
\( H_D = H_D(jw) \) is the response of the transmission path from the plane defining the terminals of the receiving antenna to the terminals at which the output response \( G = G(jw, \phi, \theta) \) is measured.

It may be seen that to obtain the antenna response \( H_R \) from a single measurement of \( G \), all other quantities in Equation 39 must be known individually, or their product must be known. For substitution measurements, a calibrated standard antenna must be available whose phase and amplitude properties are known. There are several possible techniques for phase and amplitude calibration of a reference antenna and these are discussed below. The conventional standard gain antenna has been calibrated in amplitude only.

1. Three-Antenna Group Delay Technique

First, a receiver measurement of the source and transmission line responses \( (E_S, H_C, H_D) \) must be made. This measurement is made by connecting the source transmission line directly into the receiver connecting line. The impedances of these various components must be matched so that reflections between them are negligible. Calibration of the source-to-receiver response is made by measuring

\[ G_C = F_S H_C H_D. \]  
(40)

These factors must now remain unaltered for calibration of the reference antenna.

The three-antenna method requires two "reference" antennas, \( R \) and \( R' \), and the transmitting antenna. Either \( R \) or \( R' \) can later serve as a reference antenna. Three successive measurements are made using three different pairs of these antennas. The required setups are illustrated in Figure 23. The following three equations represent the measured responses

\[ G_1 = F_S H_C H_S H_R H_D, \]  
(41)

\[ G_2 = F_S H_C H_S H'_R H_D, \]  
and

\[ G_3 = F_S H_C H_S H'_R H_D. \]  
(42)
Figure 23. Three antenna group delay calibration method
With $F_S H_C H_D = G_C$ known from the previous measurement, the three antenna responses are found to be

\begin{align*}
  H_R &= \frac{G_1 G_3}{G_2 G_C}, \\
  H'_R &= \frac{G_2 G_3}{G_1 G_C}, \text{ and} \\
  H_S &= \frac{G_1 G_2}{G_3 G_C}.
\end{align*}

It is seen that antenna calibration depends on four separate measurements, interchanging of antenna pairs from transmitter to receiver, and direct connection of transmitting and receiving hardware at one site. The accuracy of this calibration is critically dependent on maintaining good impedance matches with each test arrangement and on transmitter/receiver stability over the measurement period.

2. Separate Reference Channel Direct Calibration

The two-channel phase/amplitude receiver can be used for calibration of the reference antenna if a suitable link to the signal source can be provided. For electrically small antennas, the far-field requirement can be met at ranges for which direct coaxial cable links are practical. A phase-compensated cable with nearly linear phase delay characteristics is available. It appears practical to use this type of cable on antenna ranges of approximately 100 feet at frequencies up to about 10 GHz. A 100-foot range could be used for S-band reference antennas with gains of about 30 dB or less. At 4 GHz, the loss in 100 feet of this type of cable is about 5 dB. Cable assemblies with phase deviation of a few degrees (<10) across the 2-4 GHz range are available. With this technique, identical antennas are required (see Appendix II) for transmit and receive, either one of which will become the
reference antenna. The combined phase delay of the two antennas is measured with the reference signal provided via the coaxial cable. One-half of the measured phase delay is then assigned to each antenna. The calibration accuracy of this technique is limited by the phase stability of coaxial cable and by the requirement of two identical antennas.

3. Antenna Impedance and S-Parameter Method

The antenna impedance and S-parameter method has been used to determine the amplitude and phase properties of broadband antennas over 10 to 1 bandwidths. This method consists of broadband measurements of both the input reflection coefficient of an antenna and the forward transmission coefficient between a transmit/receive pair of identical antennas (see Appendix II). The amplitude and phase properties of a reference antenna can be determined from measurements of both the input reflection coefficient $S_{11}$ and the forward transmission coefficient $S_{21}$ over the band of frequencies for which the antenna will be used as a reference. The fundamental equations which describe the antenna as a transmitter and receiver and the expression for the voltage gain of a system composed of two identical antennas separated by a free-space transmission path of range $r$ are presented in Appendix III.

The first step in this method of calibrating a reference antenna consists of a measurement of the input reflection coefficient of the antenna. The antenna impedance $Z_a$ is related to the reflection coefficient $S_{11}$ by

$$S_{11}^o = \frac{Z_a - Z_c}{Z_a + Z_c},$$

where $Z_c$ is the characteristic impedance of the transmission line. The notation $S_{11}^o$ is adopted to denote that the antenna is radiating into free space. Solving for $Z_a$ yields

$$Z_a = Z_c \frac{1 + S_{11}^o}{1 - S_{11}^o}.$$
A basic setup for measuring the reflection parameter $S_{11}^0$ is shown in Figure 24. The line stretcher allows the plane at which the measurements are made to be extended to the terminals of the antenna as well as the equalization of the transmission line lengths between the reference and test channels. If these line lengths are equalized, no differential phase change is observed between the reference and test channels as the frequency is swept over the band.

The next step in calibrating a reference antenna is the determination of the forward voltage gain $A_v$ by measurements of the forward transmission coefficient $S_{21}$ and the input reflection coefficient $S_{11}$ for an arbitrary load impedance. In most cases, the load impedance $Z_L$ at the output terminals of the receiving antenna will be chosen for a matched condition such that $Z_L = Z_C$ and then $S_{11}' = S_{11}$. For most transmit/receive test configurations of interest, the transmit and receive antennas are separated by sufficiently large distance that the radiation resistance of the transmit antenna is insensitive to the presence of the receive antenna. Thus, the previous measurement of $S_{11}^0$ will suffice to determine the input reflection coefficient $S_{11}$ under arbitrary test conditions.

Figure 25 shows the setup for the measurement of the forward transmission coefficient $S_{21}$. As discussed in Appendix III, the forward voltage gain $A_v$ is related to $S_{11}$ and $S_{21}$ by

$$A_v = \frac{S_{21} (1 + \Gamma_R)}{(1 - S_{22}) \Gamma_R (1 + S_{11})},$$

where $\Gamma_R$ is the reflection coefficient looking out from the receiving antenna terminals. In most practical situations, $\Gamma_R$ is approximately zero so that

$$A_v = \frac{S_{21}}{1 + S_{11}}.$$
Figure 24. Basic setup for measuring reflection Coefficient $S_{11}$
Figure 25. Setup for measuring forward transmission coefficient $S_{21}$
The line stretcher is used to refer the measurement reference planes to the antenna terminals. The subsequent measurement of $S_{21}$ will then include the phase shift due to the free space range $r$. If precise range information is available, the effects of the free-space path can be removed by calculation. Also, accurate range information is required to precisely locate the terminal planes of the antenna.

These reflection coefficient and forward transmission coefficient measurements in principal can be performed using standard laboratory test equipment. However, for directive antennas the requirement that a TEM-mode transmission line connect the "identical" transmitting and receiving antennas to the test instrument may present a problem. This requirement may be a problem for directive antennas with far-field requirements since there would be a significant cable loss which would degrade the accuracy of the measurement. Thus, the limitations of this approach appear similar to those of the direct phase comparison approach described in Subsection D-2 above.

4. Free-Space Scattering Method

A free-space method for the determination of an antenna's complex gain (amplitude and phase) versus frequency characteristics exploits the fact that, if the antenna terminals are mismatched, the echo area of an antenna is proportional to the square of the antenna's power gain. Or stated another way, the field scattered from the antenna is proportional to its voltage gain squared. If the magnitude and phase of this scattered field are measured, the complex voltage gain can be determined. (The required equations are developed in Appendix IV.) In developing the method, the amplitude and phase of the scattered field of a sphere of known size is first measured. Next, the amplitude and phase of the scattered field of the antenna for a short-circuit terminal condition is measured, relative to the scattered fields of the sphere.
Finally, the absolute magnitude and phase of the antenna scattered field are
determined using the known Mie series solution for the scattered field of
the sphere.

A potential difficulty in applying this technique to broadband sweep-
frequency measurements is that the background noise cannot be cancelled using
a nulling loop as is done in single frequency applications. However, if both
the sphere return and the antenna return are much larger than the background
noise, the error introduced by the background will be small. Also, it is
possible to record the background noise as a function of frequency and then
subtract this recorded background noise from the signal-plus-background noise.
This method has been successfully employed at Georgia Tech for near-field
radar cross section measurements [3] at a single frequency, and it has been
determined that background-to-target signal ratios on the order of -3 dB can
be tolerated. Experiments have shown that the background signals in an anechoic
chamber are reasonably stationary with respect to time over periods of a few
hours. Thus, it appears that this "computer nulling" method can be employed
in a sweep frequency measurement.

An advantage of this method is that the transmission paths are common for
both the sphere and the antenna under calibration. Thus, no error is intro-
duced by the transmission paths. Also, the phase shift from the antenna aper-
ture to the output terminals is directly measured with respect to the sphere.
Since the sphere center and the antenna aperture can be optically aligned, a
precise measurement of the range from the transmitter to the target is not
necessary. This method thus appears to have good potential for determining
reference antenna amplitude/phase versus frequency characteristics. Experi-
mental evaluation of this method is desirable.
5. Comparison of Calibration Methods

Standard gain antennas have long been used for the absolute gain calibration of an unknown antenna in amplitude only. However, a reference antenna calibration including both amplitude and phase antenna measurements is more difficult since both the voltage gain and the phase shift versus frequency of the reference antenna must be determined. It should be noted that, to date, no experience in characterizing a reference antenna in both magnitude and phase over a wide frequency bandwidth has been obtained. Thus, the magnitude of the practicable measurement difficulties which may arise can not currently be pinpointed.

The optimum calibration method will depend on the electrical size of the intended reference antenna and on the required accuracy of calibration. For relatively small antennas, direct phase measurements are practical, and they should provide accuracy within approximately 10 degrees. Continuous broadband calibration will require a tracking broadband receiver. However, static measurements at selected frequencies across the desired calibration range can be made with currently available instrumentation.

The three-antenna group delay method is attractive because it can be used with any reasonably sized reference antenna, and identical antennas are not required. Assuming availability of a group delay measuring receiver, each individual measurement for this method can be made with high accuracy. Experimental tests are required to firmly establish the overall calibration accuracy which can be achieved.

The free-space antenna scattering concept is currently limited in application to antennas of relatively small physical size. The required measurements are conveniently made with proper instrumentation. Experimental evaluation of achievable accuracy is desirable.
The antenna impedance and S-parameter method is probably better suited to less directive low frequency antennas.

E. GENERAL HARDWARE REQUIREMENTS

During the investigations into the capabilities and requirements of direct-phase sweep-frequency measurement techniques and of group delay measurement techniques, the two concepts were analyzed separately. However, since the group delay measurement would require a basic receiver plus the sideband processor, this concept separation is not desirable from a hardware implementation point of view. Hardware for implementing both techniques will be discussed.

As discussed in the previous portions of this section, a sweep-frequency direct-phase measurement technique can be implemented by modification of the existing Scientific-Atlanta Series 1750 Phase/Amplitude receiver. The group delay measurement capability can be implemented by including the sideband processing circuits in the modified receiver. A block diagram of the receiver is shown in Figure 26.

The receiver is a phase-locked superheterodyne type which employs harmonic mixing. It is generally similar to the Scientific-Atlanta Series 1750 Receiver, except that it employs a frequency-agile local oscillator and has additional circuits which give it the capability of making group-delay measurements in addition to direct phase/amplitude measurements. All readouts from the receiver are available in digital form by routing the 1-kHz output signals to integral phase and amplitude display units for processing and conversion to digital form. Also shown in Figure 26 are the local-oscillator and phase-lock circuits and the output to the sideband processor, which is shown in Figure 27.
Figure 26. Tracking phase-locked receiver.
Figure 27. Sideband differential-phase processor
The sideband processor is required for measurement of $\Delta \phi / \Delta \omega$. The result of the various frequency translations is to provide output signals $(\varphi_1 - \varphi_2 - 2\alpha) \mid_{\omega_1}$, where $\omega_1$ represents the set of measuring frequencies separated by $\Delta \omega$. The output signals are processed to give the phase difference

$$\Delta \hat{\varphi} = (\varphi_1 - \varphi_2 - 2\alpha) \mid_{\omega_1} = \Delta \varphi - 2\alpha \mid_{\omega_1},$$

(51)

where $\alpha = \text{phase delay in the modulation reference}$. To avoid error, the $2\alpha$ term must remain constant during the measuring interval. If this term is constant but non-zero, digital integration of $\Delta \hat{\varphi} / \Delta \omega$ gives rise to a term which represents a time shift, which is of no consequence. For convenience, $\alpha$ can be varied in setting up the system.

The phase/amplitude measurement system is shown in Figure 28. As discussed above, the sideband processing circuits are incorporated in the receiver, and the transmitter contains the modulator required for generating the sidebands. For clarity, A/D converters are specifically indicated in the receiver output channels. However, the Scientific-Atlanta hardware contains its own digital display and output capabilities. (A multiplexer is indicated for ease in interfacing these outputs with the computer.) The receiver's digital phase unit, whose dynamic range at the input is at least 60 dB, converts the measured phase between the test signal channel and the reference signal (or between the two sidebands) to a digital signal.

The receiver's digital amplitude unit converts the analog amplitude signals from the basic receiver to a digital dB format. It also provides the ratio, in dB, of the inputs from the test and the reference channels, and it can be externally triggered for updating. The minimum dynamic range of this unit is also 60 dB.

The computer-to-pattern-recorder interface (D/A converter), routes selected pattern data in decibel form from the computer to the pattern recorder.
Figure 28. Phase-and-amplitude measurement system
for display. The nominal range of this signal is four decades. Resolution requirements can be met by a 10-bit D/A converter (provides resolution here of about 0.04 dB). Since only one selected frequency component is expected to be recorded for display, the D/A conversion speed is not a critical requirement, and this interface presents no hardware problem.

The computer-to-position-controller interface passes such data as the test antenna start and stop position and the desired pattern cut. Since present positioners cannot be stepped to less than 0.1 degree, extreme resolution is not required in this channel. A resolution of 0.1 degree out of 360 degrees (equivalent to positioner step resolution) can be provided by a 12-bit D/A converter.

Flexibility in the computer-to-transmitter interface which will allow specifying data recording modes with varying degrees of accuracy and with various data rates is desirable. For a "quick look" capability, a simple START SWEEP command which causes the transmitter to linearly frequency sweep at a high rate across its range setting is desirable. The frequency is swept at a constant rate, and the frequency at each sample point is determined by a counter in the receiver. In this mode, the accuracy to which data are obtainable is limited by the uncertainties imposed by the requirement for a finite measurement and conversion time. For more precise data, a slower data rate is required. At some spatial antenna positions, such as near a sidelobe null, frequency sweeping may cause the antenna phase response to undergo rapid changes. Since the receiver output circuits are relatively narrowband, the phase information might be significantly "smoothed." In such a situation, more precise information can be obtained by stepping the frequency in discrete steps with 20 to 50 msec between steps. This 20- to 50-msec time interval would allow the phase output circuit to stabilize before a reading is taken,
and the frequency would always be known within its step accuracy. This latter mode of operation can be implemented with a programmable transmitter which responds to the STEP SCAN command from the computer. In addition, for group delay measurements, discrete frequency steps are required with a highly stable frequency during the measurement interval. Data recording rates and sampling requirements are further discussed in the following subsection.

Computer requirements for this system can be met by the Raytheon 703 which has been described in the Statement of Work for this investigation. Based on experience in operating the computer-controlled near-field range at Georgia Tech, software instructions for computer control during data recording are expected to require approximately 4k of storage. Recording phase, amplitude, and frequency at 100 frequency points per spatial position requires 300 data words. Since the Raytheon 703 has a 32k-word memory, phase, amplitude and frequency data for approximately 90 spatial positions could be recorded in the computer memory. The Raytheon 703 has two types of input/output data communication channels, the Direct Input/Output (DIO) and the Direct Memory Access (DMA). Using DMA, 16-bit data words may be interchanged between 703 memory and up to six external devices simultaneously, interlaced with computation/executions. With the 703's basic memory cycle time of 1.75 μsec, data can be accepted at a faster rate than would typically be required (see the next subsection). The 703 computer can also be used to perform some on-site diagnostics; this application will be discussed in a following subsection.

The Statement of Work also specifies that the Hewlett-Packard magnetic tape recorder model 2020E be used for data recording. This is a 7-track machine (6 data bits plus parity) which records at 45 inch/sec and at a byte (6 data bits) density of 556 per inch. Two bytes of information will, therefore, be required to record data in more than six-bit words. Tapes made on
this machine can be processed readily by the Honeywell 600-line computers and their associated tape handling equipment. Under proper software control, this recorder can interface with the 703 computer for data storage, and it is suitable for this application.

F. SYSTEM OPERATION

Before describing a typical measurement sequence, guidelines for selecting sampling intervals and data rates will be presented. Since the sweep-frequency direct-phase measurements and the static group delay measurements have somewhat different sampling requirements they will be discussed separately. For sweep frequency antenna measurements, there are two sample variables to be considered. These variables are spatial coordinates and frequency. Basically, there are two sampling approaches which should be considered. With the first approach, the antenna under test would be slowly and continuously rotated. At specified time intervals, dependent on the speed of antenna rotation and the desired spatial resolution, the test signal would be rapidly swept through the frequency range of interest. The frequency sweep rate should be such that the antenna could be considered essentially stationary over the sweep time. On the second approach, the test antenna would be incrementally stepped to new positions and remain stationary while the test signal is swept over the desired frequency range. The relative desirability of the two approaches depends on the test antenna size and the required accuracy of the data. Since test signal sampling is required for either of these approaches, this sampling requirement will be discussed first.

The upper limit of the frequency track rate for the projected phase/amplitude receiver is 10 MHz/msec. If samples every 100 MHz are needed, this would require a sample rate of 1/10 msec, or 100 Hz. Recording phase, amplitude, and frequency increase the total data rate to 300 Hz. This 100 Hz per
data channel is compatible with the maximum data rate of the receiver output channels. This data rate also is easily accommodated by the 1.75 μsec memory cycle time of the Raytheon 703 minicomputer. Alternately, at this data rate, the data could be recorded directly on magnetic tape. A magnetic tape recording rate of 15 kHz at 12 bits is typical. Sampling every 100 MHz over a 1-GHz range results in 10 frequency sample points. Recording frequency, phase, and amplitude at each sample point would require storing 30 data words for each frequency sweep. (With proper software "bookkeeping," spatial angle need be recorded only at the beginning of each frequency sweep.)

The most appropriate spatial sampling interval will depend on the test antenna size and the desired accuracy. For illustration, consider an antenna having a 3-dB beamwidth of one degree. (A 1-degree 3-dB beamwidth requires approximately a 25-foot diameter aperture at the S-band frequency of 3 GHz.) Assume that it is desired to record pattern cuts over ±180 degrees. The optimum positioner scan rate is primarily dependent on two competing factors. These are the desirability of minimizing the total pattern recording time (fast scan) and of minimizing movement of the test antenna during a frequency sweep (slow scan). Examination of these factors shows that a relatively slow antenna scan rate on the order of 0.15 deg/sec is appropriate. Thus, 40 minutes are required to record the entire ±180 degree pattern. With the previously considered transmitter sweep rate (10 MHz/msec), 0.1 second is required to sweep the full 1-GHz frequency range. During this time, the antenna would sweep through 0.015 degree. This much movement during a frequency sweep would amount to only one one-hundredth of the one-degree beamwidth, and the resulting data uncertainty would be tolerable. Additionally, 0.015 degree is comparable to the readout accuracy of typical positioners.
For a 1-degree total half-power beamwidth, the first nulls would occur near approximately \( +1.3 \) degree off beam center, and successive nulls would occur at further 1-degree increments. If 10 sample points per sidelobe provides sufficient resolution, a spatial sample is required each 0.1 degree; because the null-to-null width of the main beam is about 2 degrees, 20 sample points to define the main beam are necessary. At the previously assumed positioner rotation rate (0.15 deg/sec), a new frequency sweep must begin about every 0.67 second in order to sample every 0.1-degree spatial increment.

Since only 0.1 second is required per frequency sweep, the frequency sweep rate, spatial sample rate, and positioner scan rate are all compatible. There would be approximately one-half second of deadtime between each data recording interval. For small antennas which can be conveniently step scanned and stopped for data recording, the recording time per pattern may perhaps be reduced to less than 40 minutes. A programmable positioner controller is available which can provide scan increments as small as 0.1 degree (S/A Model 2004 Plus Model 4561 Servo Repeater-Converter). Based on antenna size (and beamwidths) and type of positioner available, the stepped-scan approach should be examined for specific applications. In addition, it may be necessary to use a stepped scan and a slower frequency sweep near null positions where phase and/or amplitude are rapidly changing.

With the group-delay measurement approach, a frequency-stepped approach is required. With the tentatively selected 5-MHz transmitter modulation tone, the sideband separation will be 10 MHz. If the carrier is stepped at 100-MHz increments, essentially the same sampling considerations apply as considered above. That is, in the frequency domain, 10 sample points are produced for each 1 GHz of frequency test range. At each of the 10 points, amplitude, \( \Delta \varphi/\Delta \omega \), and carrier frequency are measured for a total of 30 data words at
each spatial point. The minimum time between frequency steps should be 10 msec because of the bandwidth of the data channels. These guidelines are seen to be similar to those given for the sweep-frequency approach.

There is an additional consideration in antenna rotation rate which should be noted. A Doppler shift can occur due to antenna rotation during the measurement interval (discussed later under system accuracy). If this occurs, it causes the IF center frequency to deviate from its nominal value of 1 kHz and can lead to a measurement error if the IF shift is too great. If required, however, the antenna can be stepped to avoid any Doppler shift during measurement.

With a calibrated reference antenna available, amplitude and phase measurements can be made directly with the amplitude and phase measuring circuits of the receiver which has been described in Subsection E. Phase can also be measured by using the sideband processor to obtain the differential phase $\varphi$ and then integrating. For either case, the phase reference or frequency tracking antenna should be located in relatively close proximity but with negligible coupling to the antenna under test. The total electrical path length from the transmitting antenna through either reference antenna should be essentially equal to that through the antenna under test to prevent large phase excursions of the output with frequency.

Group delay measurements are made by first mounting the calibrated reference antenna on the test mount and taking a set of measurements $G_R (\omega)$. Then, a set of measurements $G_T (\omega)$ is taken with the test antenna in place. The test antenna transfer properties $H_T (\omega)$ are then given by

$$H_T (\omega) = \frac{G_T (\omega)}{G_R (\omega)} H_R (\omega),$$

where $H_R (\omega)$ is the calibrated reference antenna characteristics. In a typical
data recording sequence, the operator enters the following parameters into
the computer: 1) spatial antenna pattern cut, 2) data to be recorded (phase
and/or amplitude), 3) antenna rotation rate or step increment, 4) spatial
sample increments, 5) frequency sweep or step rate, 6) frequency increments,
7) direct phase or group delay measurements, and 8) frequency component to
be plotted in real time on the pattern recorder. The computer generates
initial antenna position data which are fed through the D/A converter to
the controller which initiates the antenna movement. As the antenna rotates,
position data are continuously received by the computer from the encoders on
the positioner. As each of the pre-set angular increments is reached, the
computer generates a sweep or step frequency command which is transmitted
via data link to the transmitter site. The transmitter responds to this
command by beginning a frequency sweep over prescribed limits or by se-
quently stepping frequency across the measurement range. The receiver
is continuously generating phase, frequency, and amplitude data which are
fed to the A/D processing units.

As each of the prescribed frequency sample points are reached, the com-
puter reads the phase, amplitude, and frequency data into the computer memory.
Antenna position data are recorded before each frequency sweep. The antenna
continues to scan or it is stepped after each complete frequency traversal,
and the computer repeats the data recording process each time a predetermined
angle is reached. In this way, a complete pattern cut with the broadband
frequency response at prescribed spatial points is stored in the computer.
At convenient intervals (such as the end of each pattern cut, or completion
of a frequency sweep), the data stored in computer memory are dumped onto
magnetic tape. For example, with the previously considered 30 data words per
spatial point and with approximately 28K of Raytheon 703 computer memory for
data recording, data for about 900 spatial positions could be stored in computer memory. If less than a 360° pattern or less than 0.1 degree resolution is sufficient, perhaps a full spatial pattern cut can be recorded before memory dump. The computer can select data at any specified frequency for plotting of a "real time" pattern. The amplitude data at each spatial point may be normalized to the maximum amplitude and converted to decibel form before routing to the recorder.

Although antenna diagnostics which require handling a large amount of data will be performed by a large general purpose off-site digital computer, some on-site diagnostics are practical. Tapes recorded on the HP 2020E digital tape recorder can be read directly by most computers, including the Honeywell 600-line, which would be suitable for off-site diagnostics. All of the data which could possibly be recorded for a single antenna probably will not be used for any one given diagnostic. For example, if phase and amplitude were recorded each 100 MHz over a 1 GHz range (10 x 2 data points) at each 0.1 degree over a full 360 degree sector, this would require recording 72,000 data words per pattern cut. At 14 bits per word, this is approximately 1 million bits of information per pattern cut. All of this information is available for processing if desired, but most diagnostics will not require all of this data. To obtain an average pattern response, only the amplitude data are required. Furthermore, in many cases, average response over angular sectors smaller than the full 360° sector will be of interest. In the case of phase response and time domain pulse diagnostics, many applications will be concerned with situations in which the antenna is pointed directly at the source of radar return. In such cases, only the main beam data are required, which greatly reduces the amount of data which must be handled. In summary, even though all the antenna information is available, sorting only that data
For on-site diagnostics, the Raytheon 703 can generate average antenna response at each spatial position since this requires handling only the amplitude samples over the desired bandwidth. Use of Fast Fourier Transform techniques may permit some pulse diagnostics as well.

G. MEASUREMENT SYSTEM ACCURACY

The accuracy of the measurement system is influenced by a number of factors, some of which are listed below and discussed briefly in the following paragraphs. The measurement system accuracy will depend, of course, on the final design and construction details as well as operational factors. Therefore, this subsection contains only a discussion of the major contributors to error with some general statements as to the accuracy that is achievable.

The measurement system accuracy is dependent on the following factors:

1. Reference antenna calibration accuracy,
2. Impedance,
3. Receiver measurement accuracy,
4. System stability and repeatability, including modulation reference system stability,
5. Dynamic errors,
6. Random noise and signal/noise ratio, and

Several methods for calibrating the reference antenna have been discussed. Since the test antenna measurement accuracy will depend on the reference antenna calibration accuracy, to obtain highly accurate test results, a highly accurate calibration method must be used. Assuming availability of a high accuracy measurement system, the three-antenna calibration method is potentially
very accurate. A source of error in this method is drifts and instabilities while the three antennas are being interchanged for measurements. Recall also that the transmitter must be transported to the receiver site and connected directly to the receiver for one of the measurements required by the three-antenna method. To obtain high accuracy, not only must the measurement system characteristics be repeatable for both locations of the transmitter, differences due to movement or substitution of RF cables must be avoided. The free space scattering method also is capable of high accuracy with relatively small antennas. Both the direct phase comparison and the antenna impedance/S-parameter methods are expected to be less accurate.

Mismatch errors during tests can be caused by reflections at the various interconnections. The magnitude of amplitude losses and phase uncertainties can be calculated if the reflections are known. Since the measurement system must operate over a broad frequency range, it will probably be necessary to use precision attenuators or broadband isolators to avoid multiple reflections.

The IF and the amplitude/phase measurement units of a modified Series 1750 receiver will be similar to those of the standard Scientific-Atlanta Series 1750 Phase-Amplitude Receiving System. With the Series 1750 system, high-accuracy measurements can be made over a wide dynamic range of the input signal level. Typical accuracies are as follows:

<table>
<thead>
<tr>
<th>Dynamic Range</th>
<th>Phase Accuracy</th>
<th>Amplitude Accuracy</th>
</tr>
</thead>
<tbody>
<tr>
<td>0-10 dB</td>
<td>0.25°</td>
<td>0.1 dB</td>
</tr>
<tr>
<td>0-20 dB</td>
<td>0.5°</td>
<td>0.2 dB</td>
</tr>
<tr>
<td>0-40 dB</td>
<td>2°</td>
<td>0.3 dB</td>
</tr>
<tr>
<td>0-60 dB</td>
<td>5°</td>
<td>1.0 dB</td>
</tr>
</tbody>
</table>

The Series 1750 system provides a phase resolution of 0.1°, and an amplitude resolution of 0.01 dB over a 20 dB range and of 0.1 dB over a 60 dB range. Further, because a modified Series 1750 is compatible with the group delay
measuring system, the accuracy for this system should also be comparable to that shown above. To achieve these results, however, care must be exercised in the design and implementation to prevent errors from RF or IF leakage between channels, pickup of internal reference signals, amplifier non-linearities etc. Particular care must be exercised with the RF circuitry to minimize error due to cross talk between channels. Excessive rates of amplitude and phase variation must be avoided during measurements.

For group delay measurements in which standard and test antennas are interchanged, the system must have good long-term stability in order not to require frequent recalibration of the system. Based on experience with the Series 1750 system, phase and amplitude stability of $0.5^\circ$ and $0.2$ dB over a $20$ dB range for a period of one to four hours appears reasonable for operation at a relative constant temperature after warm-up. For direct phase measurements with a calibrated antenna in the reference channel, system stability is not critical.

For extreme accuracy over limited angular segments, dynamic errors caused by scanning frequency while continuously rotating the antenna can be avoided. The frequency can be stepped and the dwell time adjusted as necessary to provide the maximum accuracy.

As indicated earlier, a frequency shift will occur between the signal channel input and the reference or frequency track channel while scanning in frequency, if there is a difference in RF or local oscillator path lengths between the channels, or if the antenna under test is not rotated about its phase center. For maximum accuracy, when these conditions exist the frequency shift should be limited to a few Hertz by limiting the rotation rate.

When making direct phase measurements with a reference antenna employed to receive a reference signal at the test frequency, atmospheric disturbances
could cause a random phase variation of the reference signal relative to the
signal from the test antenna. However, under normal weather conditions, if
the reference and test antennas are separated by not more than 100 feet, which
normally is the situation, and if the frequency is in the 1-18 GHz range,
the probable random phase variation would be less than one degree [4].

H. LARGE REFLECTOR APPLICATIONS

In many cases, it may not be practical to perform broadband measurements
on an entire antenna system. For example, the antenna's electrical size may
require a far-field range which is much greater than that which is available
or practical. In other cases, the antenna's structure may be such that assem-
bling the antenna on a test range would be impractical. In cases where the
entire-antenna system either cannot be directly measured or it would be ex-
tremely expensive, an alternate approach is desired. There are basically
two different types of information which might be required. These are (1)
far-field pattern information, and (2) peak gain and insertion phase as a
function of frequency. The desired patterns may be either broadband or CW.
Since CW information is a special case of the general broadband data, ap-
proaches for obtaining the broadband pattern data and the peak gain/phase
data are required. It is realized that the phase and amplitude system which
is described in this section may offer a possible solution. This approach
for obtaining far-field parameters consists of the following: (1) measure
broadband phase and amplitude data at the feed terminals with the system which
has been described in this section, (2) use this measured feed data to compute
aperture phase and amplitude distributions, and (3) transform the aperture
distributions to obtain the desired far-field parameters.

In a typical feed/reflect antenna system, the phase center of the feed
is not constant over a broad frequency range. Consequently, the phase of the
feed radiation pattern depends on frequency. Broadband phase measurements at the feed terminals can be referenced to a known phase center at a particular frequency, and used to predict variations in the feed radiation pattern with frequency.

From known feed amplitude, phase, and polarization data, antenna pattern parameters for a single frequency may be calculated using known computational procedures [5, 15]. For a paraboloidal reflector, the amplitude of the aperture field, \( E_a \), is related to the amplitude of the feed pattern \( E_f \) by

\[
|E_a(\theta,\varphi)| = \left(\frac{1 + \cos \theta}{2}\right) |E_f(\theta,\varphi)|
\]

where \( \theta \) and \( \varphi \) are conventional coordinates as illustrated by Figure 29. The aperture phase function \( \psi(\rho,\varphi) \) is determined from the phase of the feed pattern by

\[
\psi(\rho,\varphi) = \tan^{-1}\left(\frac{\text{Im} E_f(\theta,\varphi)}{\text{Re} E_f(\theta,\varphi)}\right)
\]

where \( \rho = \frac{2L \sin \theta}{1 + \cos \theta} \). Recently, a method for determining the aperture polarization by a stereographic mapping of the feed polarization has been developed [6]. Thus, techniques for determining aperture amplitude, phase, and polarization from single frequency feed data have been developed.

Once the complex aperture distributions \( E_a \) and \( H_a \) have been determined, the far-zone fields can be obtained by several methods. These methods generally differ in their computational complexity and in the completeness of data obtainable. The most accurate far-field data are obtained by using a vector formulation [7]. However, the evaluation of the required vector integrals is tedious and very time consuming for electrically large apertures. A less accurate but more tractable approach may be to calculate the far-zone using an approximate scalar relation. The scalar relation is in the form of a two-dimensional Fourier integral which can be efficiently evaluated using
Figure 29. Paraboloid of revolution geometry
the Fast Fourier Transform (FFT) algorithm. In order that the scalar evaluation be accurate, there are several restrictions which must be met [7]. The effect of these restrictions in the present application must be quantified.

Another method which may prove useful for calculation of far-removed sidelobes employs the high-frequency asymptotic methods of the geometrical theory of diffraction (GTD) [8, 9, 10]. Because the GTD method is presently not as accurate in the main beam region as in the sidelobe regions, the previously discussed integral formulations must be employed in the main beam region. Further, because previous applications of these computational procedures have been limited to narrow-band antennas, the feasibility of applying these procedures to wideband antennas must be investigated.

It may be seen that there are several problem areas to be investigated in order to define the feasibility of the three-step approach discussed above. The most efficient and direct method of obtaining feed radiation parameters from broadband phase and amplitude data must be determined. The feasibility of predicting broadband far-field parameters with methods previously used in single frequency analysis must be investigated. The required computational complexity (computer resources) and the resulting accuracy of the various methods must be determined.
The objectives of this contract were to explore new techniques to measure, record, and analyze broadband antenna patterns and to determine the feasibility of developing instrumentation to perform these measurements. These objectives have been fully met, and the investigations which have been described in this report establish the feasibility of broadband phase and amplitude measurement systems which are applicable to high-gain microwave antennas. Specific system concepts which were to be investigated included amplitude-only broadband noise systems and sweep frequency phase and/or amplitude systems. Variations from these two basic concepts were also to be investigated. Evaluations showed that the broadband noise concept is not a cost-effective approach for an amplitude-only system. A simpler sweep frequency system (called a Hybrid System) which employs electronic integration of a sweep frequency signal to obtain broadband amplitude data was synthesized and evaluated. This Hybrid system is an effective approach to an amplitude-only system. Preliminary design of the amplitude-only Hybrid system has been completed, and its construction is relatively straightforward with available components.

A computer-controlled sweep frequency phase-plus-amplitude system is required to completely characterize an antenna that requires a wide instantaneous bandwidth and to provide the necessary data to perform off-line diagnostics which define that antenna's frequency and time-domain performance under a wide variety of operating parameters. System operating parameters (frequency and spatial sample rates) and calibration techniques which permit a high measurement accuracy are practical. Although the computer controlled
sweep frequency phase/amplitude system is relatively complex, the individual components required for its instrumentation are feasible with today's technology. The primary hardware advance which is required involves incorporating a broadband phase-locked frequency tracking capability into an existing wide-range receiver.

Because of the relative simplicity of the Hybrid system, it can be constructed with a very modest expenditure of time and money. Construction of this amplitude-only system should begin immediately, thus demonstrating at an early date a significant increase in broadband antenna measurement capability. Amplitude-only spatial antenna patterns impact on a number of radar performance parameters (detection range, angular resolution, tracking accuracy, ECM susceptibility) and the ability to quickly measure these patterns over a broadband is greatly needed.

Because there are important applications in which amplitude-only information is not sufficient (such as off-line diagnostics to predict pulse distortion, and effects on FM modulated signals), completion of the tasks which are necessary to develop the computer-controlled sweep frequency phase-plus-amplitude system should be pursued. These additional tasks include (1) complete system design and component specification, (2) complete definition of system operating procedures, with specification of procedures for any required computer correction of both amplitude and phase data, (3) develop software necessary for computer-control and data recording with an operator-interactive capability, (4) breadboard a laboratory setup and experimentally demonstrate feasibility of the system, (5) analytically bound the reference antenna calibration accuracy which is practical, and (6) develop and demonstrate computer software to compute an antenna's effect on an arbitrary...
input pulse from measured phase/amplitude data on the antenna.

Because of prohibitive antenna size or other restraints on moving an antenna, in some important cases it may not be practical to measure antenna pattern parameters directly. Consequently, in addition to the above basic tasks, the feasibility of predicting broadband far-field parameters of electrically large apertures from measured feed phase/amplitude data should be investigated. Besides the antenna measurements problems of immediate interest to RADC, there are many other applications of broadband phase and/or amplitude measuring systems. Some of these other applications are identified below, along with the respective cognizant agency.

<table>
<thead>
<tr>
<th>APPLICATION</th>
<th>AGENCY</th>
</tr>
</thead>
<tbody>
<tr>
<td>Spread Spectrum Radar</td>
<td>U. S. Army MICOM</td>
</tr>
<tr>
<td></td>
<td>Huntsville, Alabama</td>
</tr>
<tr>
<td>Amplitude and Phase Characterization of Up to 10:1 Bandwidth Antennas</td>
<td>U. S. Army, MICOM</td>
</tr>
<tr>
<td></td>
<td>Huntsville, Alabama</td>
</tr>
<tr>
<td>RATSCAT Radar (330 pico-sec pulsewidth)</td>
<td>U. S. Air Force</td>
</tr>
<tr>
<td></td>
<td>Holloman AFB, New Mexico</td>
</tr>
<tr>
<td>Test Wideband Antennas for Guidance System</td>
<td>Naval Weapons Laboratory</td>
</tr>
<tr>
<td></td>
<td>Dahlgren, Virginia</td>
</tr>
<tr>
<td>Mine Detection Radar</td>
<td>U. S. Army MERDC</td>
</tr>
<tr>
<td></td>
<td>Fort Belvoir, Virginia</td>
</tr>
<tr>
<td>Short Pulse Transmitters</td>
<td>RADC</td>
</tr>
<tr>
<td></td>
<td>Griffiss AFB, New York</td>
</tr>
<tr>
<td>Real Time Pulse Compression</td>
<td>RADC</td>
</tr>
<tr>
<td></td>
<td>Griffiss AFB, New York</td>
</tr>
</tbody>
</table>

It may be concluded that there is great need for broadband phase and/or amplitude measuring systems. To fill this need, the following specific actions are recommended:

(1) construct a complete octave bandwidth (2-4 GHz) Hybrid system and verify its operation through laboratory test,
(2) complete system and component specification of the computer-controlled sweep frequency phase/amplitude system,

(3) completely develop system operation and computer correction procedures for the phase/amplitude system, including computer correction for reference antenna calibration,

(4) develop necessary operating software for the phase/amplitude system,

(5) demonstrate system feasibility through laboratory test,

(6) develop and demonstrate with selected examples the software required to predict an antenna's effect on an arbitrary input pulse, and

(7) investigate the feasibility of predicting broadband far-field parameters of electrically large apertures from measured antenna feed data.

These recommended actions will lead to a realization of the complete phase/amplitude system in an efficient manner with an early demonstration of a broadband amplitude-only system with a significant broadband measurement capability.
REFERENCES


15. ANTENNAS, Chapter 12, John D. Kraus, McGraw-Hill Book Company, Inc. (1950)
SECTION VII

APPENDICES
A. CONCEPT I SYSTEM COMPONENTS

1. General

The major components peculiar to implementation of a broadband noise system are discussed here. Other required components for actual assembly of a system are discussed in Section II, Concept I Systems. Since all types of broadband systems which have been investigated require a broadband transmitting antenna, antennas are discussed separately in Subsection C of this appendix.

2. Noise Sources

The noise power $P_n$ available from a thermal noise source is given as

$$ P_n = kTB, $$  \hspace{1cm} (A-1)

where

- $k = \text{Boltzmann's constant} = 1.38 \times 10^{-23} \text{ joules/degree Kelvin}$,
- $T = \text{temperature in degrees Kelvin}$, and
- $B = \text{bandwidth in Hertz}$.

The noise power density output of a resistor at room temperature ($290^\circ K$) is therefore $-114 \text{ dBm/MHz}$.

Power output of commercially available noise sources is specified by the term \text{Excess Noise Ratio (ENR)}, defined as

$$ \text{ENR(dB)} = 10 \log \frac{k(T - T_o)B}{kT_oB}, $$ \hspace{1cm} (A-2)

where

- $kTB = \text{available noise power at operating temperature}$,
\[ kT^B = \text{available noise power at } 290^\circ \text{K}, \text{ and} \]

\[ T = \text{equivalent absolute temperature of the noise source.} \]

In order for the noise source to be useful, \( T \) must be greater than twice \( T_0 \). Generally \( T \) is much greater than \( T_0 \). Thus, the noise power output of a commercial noise source is determined from the following equation:

\[ P_n = -114 + \text{ENR} + 10 \log \text{(Bandwidth in MHz)} = \text{noise power in dBm. (A-3)} \]

Typical data for various types of noise sources are shown in Table 7. For an ENR of 15 to 30 dB and a bandwidth of 1 GHz, corresponding noise power output would be from -69 to -54 dBm.

<table>
<thead>
<tr>
<th>TYPE</th>
<th>EXCESS NOISE RATIO (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Temperature - limited</td>
<td>6</td>
</tr>
<tr>
<td>Thermionic diode</td>
<td></td>
</tr>
<tr>
<td>Hot Wire Resistor (upper limit)</td>
<td>11</td>
</tr>
<tr>
<td>Argon discharge tube</td>
<td>15</td>
</tr>
<tr>
<td>Semiconductor diode</td>
<td>30</td>
</tr>
</tbody>
</table>

Thermal noise sources, by their inherent physical nature, generate noise power which has a uniform spectral distribution (white noise) over a frequency range well in excess of the frequencies of interest here. Bandwidth limitation is due to the transmission lines necessary for coupling. Coaxial cable coupled noise sources have uniform spectral distribution from VHF to about 12 GHz, while waveguide coupled ones are limited by the bandwidth of the particular waveguide used. The upper frequency limit of readily available noise sources appears to be about 40 GHz.

Since the transmitter power requirement here is on the order of 10 watts (+40 dBm), with -69 to -54 dBm of noise power available a gain of from 94 to 125
109 dB is necessary to amplify the noise source output to the required power level even if all other system losses are neglected. Amplifiers for this function are discussed in the following subsection.

3. Traveling Wave Tube (TWT) Amplifiers

As indicated above, amplification of from 94 to 109 dB is required to bring the noise source output to the proper power level. Both tube and solid-state microwave amplifiers are available. However, at the present state-of-the-art, obtaining 10 watts output with octave band coverage is not feasible with solid-state microwave amplifiers above about 1 GHz. Of the tube amplifiers, TWTs are well suited to this application since they have octave bandwidths along with high gain and more-than-adequate power capability.

Although each major element of a TWT performs a simple function, several different implementations of these elements are possible, and consequently, tubes of differing characteristics result. The major elements of a TWT are (1) the electron gun, (2) the RF interaction circuit, (3) the electron beam focusing device, and (4) the collector. Major differences in common TWTs are accounted for primarily by the choice of RF interaction circuit and beam focusing device. For broadband TWTs, the helix RF circuit is most appropriate, and bandwidths of 5:1 have been obtained. Beam focusing is accomplished by means of either permanent magnets or solenoid electro-magnets. Permanent magnet focusing is most convenient since no additional power supply is required. Electro-magnets are generally used in very high power tubes, but permanent magnet tubes have more than adequate power capability (up to several kW) for the present application. Thus, the most suitable power amplifier for this application is a TWT that utilizes a helix interaction circuit and permanent magnet focusing. Tube parameters of interest include gain, bandwidth, power output, and dynamic range. Typically available octave bandwidth TWTs
with a 10-watt output capability have gains from 35 to 50 dB, depending on frequency. Thus, because 94 to 109 dB of amplification is required, two or three stages of gain might be required to obtain 10 watts of noise output. In addition to amplifying the input signal, TWTs also generate excess noise. In particular, a high power TWT generates considerable noise. However, the noise available from a TWT does not have a broadband uniform spectrum. Hence, to preserve the required noise properties throughout the transmitter chain, the uniform noise spectrum input at each TWT stage should be large relative to the excess noise generated by the TWT. A low-noise TWT must be used in the low level portion of the transmitter.

It is within the state-of-the-art to produce TWTs at the required power level and with octave or waveguide bandwidths up to beyond 75 GHz. However, between about 18 GHz and 75 GHz, due to lack of demand, tubes may not be readily available for certain frequency ranges.

4. Broadband Receivers

The bandwidth of the receiver must be as wide as that of the transmitter in order to accurately measure the response of the test antenna. One of the properties of broadband levelled noise is that over its frequency band, it has constant spectral density, i.e., it has constant power per unit bandwidth. If this signal were passed through a frequency selective device such as the test antenna, the spectrum at the output of the device would correspond to the frequency response characteristics of that device. When this spectrum is in turn detected by a broadband square-law detector, the voltage produced is proportional to the total RF power within the spectrum bandwidth. Thus, frequency integration of the response characteristics of the device occurs and is expressed as
\[ V_o = \Phi K_d \int_{S_1}^{S_2} H(S)dS , \quad (A-4) \]

where,
\[ \Phi = \text{constant amplitude incident power spectrum}, \]
\[ H(S) = \text{amplitude transfer function of the device (function of frequency)} \]
\[ dS = \text{incremental frequency bandwidth}, \]
\[ K_d = \text{detector constant, and} \]
\[ V_o = \text{voltage output from the detector}. \]

Thus, the detector output is proportional to the integral of the device frequency response characteristics. Therefore, to accurately measure antenna response over the full transmitted spectrum bandwidth, a receiver of equal bandwidth is required.

General types of wideband receivers include the following:
(a) detector-video receiver,
(b) wideband RF receiver,
(c) heterodyne receiver with CW local oscillator and wideband IF, and
(d) heterodyne receiver with swept local oscillator and narrow-band IF.

Each of these receiver types will be discussed individually.

A detector-video receiver is one in which the incoming RF signal is converted directly to video by a detector, and the resulting video signal is amplified by a high gain video amplifier. The RF detector is usually a square-law device, i.e., a device that produces an output voltage proportional to the input RF power. Either a semiconductor diode or a thermal device such as a barretter may be used as the square-law detector. Semiconductor diodes have about 20 dB more sensitivity than the barretter, but the diodes are generally limited to about 40 dB in dynamic range. However, post-detection techniques can be used to extend the dynamic range to 60 or 70 dB by processing the video
signal. This type of receiver is extremely simple, the RF bandwidth is limited only by the detector, and the entire frequency range of interest (1-15 GHz) can be covered by a single detector. However, it has the lowest sensitivity of the four types.

The wideband RF receiver consists essentially of the above described detector-video receiver preceded by a low-noise wideband RF amplifier. This type of receiver has a very high sensitivity, but typically a separate amplifier is required for each octave or waveguide band of interest. Several different types of low-noise RF amplifiers are available. These include low-noise TWTs, transistor amplifiers, tunnel diode amplifiers, and parametric amplifiers. At the lower frequency ranges (up to X-band), the transistor amplifier offers good noise performance and broad bandwidth at a relatively low cost. Single transistors covering the frequency range from 200 MHz to 1 GHz are available, but over the frequency range of 1 GHz to 8 GHz, octave bandwidths are typical. Low-noise TWTs also offer octave bandwidths, but their noise figures at frequencies below X-band are generally slightly higher than those of the transistor amplifiers. At X-band or above, an octave bandwidth is difficult to achieve with any of the commonly available low-noise amplifiers except the TWTs. Above the frequency range of about 18 to 20 GHz, low-noise TWTs generally cover standard waveguide bands, up to 75 GHz. Thus, the best approach is a low-noise transistor amplifier up to 8 GHz with a low-noise TWT at 8 GHz and above.

In a conventional heterodyne receiver, the input signal is converted to an intermediate frequency with the same bandwidth regardless of the RF frequency. However, down-conversion is impractical for this application because an equal frequency range at a lower center frequency results in increased fractional bandwidth which would require an impossibly broad IF amplifier for
the bandwidth under consideration here. Up-conversion would result in a smaller percentage IF bandwidth, but because the IF amplifier would be at a very high frequency, it also is not a good solution.

A heterodyne receiver with swept local oscillator and narrow-band IF amplifier is incompatible with Concept I because the instantaneous receiver bandwidth would not be great enough to process the broadband noise signal without loss of information. However, this type of receiver can be made compatible with Concept I by the use of additional hardware to process the detected video signal (a synchronized integrator and a sample and hold circuit).

Thus, the most straightforward receiver design for the broadband noise transmitter is a wideband RF receiver with a separate amplifier for each frequency band of interest.

5. Pattern Recorder

The selected broadband receiver must be compatible with the Scientific Atlanta, Inc., Series 1520 Rectangular Coordinate Pattern Recorder. This recorder is used in antenna testing to plot the received signal as a function of the angle of the test antenna. When this recorder is fed with a square-law detector (as described for the wideband RF receiver), either received power, received voltage, or their logarithms may be plotted. With a square-law detector, the recorder has a 40-dB dynamic range. When the recorder is to be used with a detector receiving an audio modulated carrier (e.g., 1 kHz), it is equipped with a narrow-band, crystal-bolometer amplifier. For recorder operation from a dc input signal, the crystal-bolometer amplifier is replaced with a dc-chopper preamplifier. Thus, the 1520 recorder can interface with practically any type of receiving system when furnished the appropriate amplifier and drive level.
B. CONCEPT-II SYSTEM COMPONENTS

Concept II systems, which rely on phase and/or amplitude samples over a wide frequency range, require a basic signal source which can be rapidly tuned over the desired measurement range. In addition to this basic signal source, the other major component of the transmitter is a microwave power amplifier. Major components required at the receiving site are a wideband phase and amplitude receiver, a computer for system control and interface, and data recorders. These major Concept II system components will be discussed in this section. Broadband transmitting antennas are discussed in Subsection C.

1. Signal Sources

Signal sources appropriate for Concept II systems include both sweep oscillators and frequency synthesizers. Sweep oscillators are more versatile for test instrumentation, and they provide greater frequency coverage for a given expenditure. However, the frequency synthesizer generally provides greater frequency stability, less phase noise, and consequently greater phase measurement accuracy with the group delay type of system. Hence, a frequency synthesizer may be required in the phase/amplitude system. Both types of signal sources are described below.

The microwave frequency synthesizer typically uses a YIG-tuned oscillator, and has its output derived directly from this source. YIG tuning is employed both for signal purity and linear tuning. Octave or greater tuning ranges are typical with linearity on the order of 0.1% and spurious radiation depressed by at least 60 dB below the desired signal. Other important features of frequency synthesizers are their frequency stability and freedom from phase noise. Phase noise (in a 1 Hz bandwidth 100 kHz from carrier) is typically 90 dB below the carrier, and frequency stability is on the order of $1 \times 10^{-8}$/day. Frequency resolution of less than 100 kHz is standard. Octave bandwidth units
from 1 GHz to 8 GHz are presently available with coverage to about 18 GHz practical at the present state-of-the-art.

Sweep oscillators, by eliminating the need of tedious point-by-point testing, enable practical broadband evaluation of frequency-dependent parameters, i.e., amplitude and phase. Many modern sweep oscillators use solid-state devices as the power source and use a modular construction in which the RF source is a plug-in unit. Plug-ins which cover the ranges from a few MHz to over 1 GHz, octave bands from 1 to 8 GHz, and waveguide bands above 8 GHz have been available for several years. However, systems are now available in which the outputs from individual sources are electronically multiplexed so that coverage of the entire 1-18 GHz range is provided at one output. Standard features of commercially available sweep oscillators include internal or external leveling, variable sweep widths, internal or external modulation capability, selectable sweep mode (single sweep, repetitive, triggered, or manual), single-frequency CW operation, variable sweep times (10 msec per sweep to 100 sec per sweep), and remote programability. Above 18 GHz, signal sources covering standard frequency bands (18-26.5 GHz, 26.5-40 GHz, 33-50 GHz, and 50-75 GHz) are catalog items. However, due to less demand for test instruments in these frequency ranges, these devices have not been designed and manufactured into such versatile sweep oscillator sources as described for the lower frequency bands. Moreover, these higher frequency sources are less readily available, and they are generally more expensive than the lower frequency ones. In some cases, it would be practical to use a versatile lower frequency sweep oscillator followed by a frequency multiplier to obtain the desired test signal at a higher frequency.
2. Broadband Amplifiers

Since the signal sources described above typically have power outputs on the order of 1 mW, amplification of 40 dB is required to obtain the desired 10 watts of transmitter power. Up to 1 GHz, wideband transistor amplifiers with a 10-watt power output capability are available. For example, over the frequency ranges of 50 MHz to 500 MHz and 500 MHz to 1 GHz, 40-dB gain, 10-watt units are available. Above 1 GHz, transistor amplifiers cannot presently provide the required power and bandwidth combination. However, as discussed in Subsection A-3, TWT-amplifiers of octave bandwidths whose power capabilities and gains are more than adequate are readily available up to about 18 GHz. Between 18 and 75 GHz, tubes offering waveguide band coverage have been demonstrated, although tubes covering some particular frequency ranges may not be readily available.

The phase stability of the amplifiers, in addition to the power, bandwidth, and gain, will be important for Concept II systems. Transistor amplifiers have demonstrated phase linearity and stability in phased-array applications (including pulse compression). In TWTs, phase shift is sensitive to such parameters as beam voltage, drive level, and heater voltage. Therefore, achieving a stable phase shift in the TWT requires regulation of these parameters. The most critical factor is that of beam voltage. A typical TWT exhibits a 3° phase shift for a 0.1% change in beam voltage. Regulation to this level or better is available in standard amplifier units suitable for this application. Consequently, suitable signal sources and amplifiers for the 200 MHz to 75 GHz range are feasible.

3. Receivers

Two different classes of commercially available receivers should be considered for sweep-frequency phase/amplitude systems. These classes are
(1) superheterodyne phase-locked receivers and (2) amplitude and phase network analyzers. These receivers were designed for different types of measurement problems, and therefore, they have differing characteristics. However, they have some overlap in their applications, and the feasibility of using either a standard or a modified version of both of these receiver classes was considered. (Amplitude-only spectrum analyzers were also considered initially, but it was concluded that a sweep-frequency computer-controlled amplitude-only system was not cost-effective and that spectrum analyzers are not applicable.)

Network analyzers and microwave receivers are both capable of providing the output phase and amplitude of a signal in a visual display and digital formats. There are several significant design differences between typical receivers and network analyzers (such as data bandwidths, methods of obtaining the LO signals, and detection methods), but the primary operational difference between the two types of devices is their sensitivity. Depending on the frequency range and bandwidth of interest, the sensitivity of network analyzers typically is 10 to 40 dB poorer than the sensitivity of microwave receivers. Network analyzers have been used for short-range antenna measurements (for example, in an anechoic chamber) by locating the reference signal generator and the network analyzer at the transmitting site and then routing the received test signal back to the network analyzer via a coaxial cable. Because of the requirement that the test and reference inputs must be fixed relatively close together, coupled with its poorer sensitivity, the network analyzer is not suited to the applications considered here. Although no commercially available microwave receiver currently has the desired swept frequency phase and amplitude capability, the modification of an existing microwave receiver to provide the required measurement capabilities is feasible. Scientific-
Atlanta, Inc., has defined the changes necessary to modify the Scientific-Atlanta Series 1750 Broadband Phase/Amplitude Receiver and to provide a tracking sweep-frequency measurement capability. In addition, the modifications required to also provide a group delay measurement capability have been investigated. These design modifications and the resulting receiver specifications are fully described in Section IV.

4 Minicomputer/Tape Recorders

The Concept II systems include a small computer as a data acquisition and storage and process control device. Such small computers are commonly termed minicomputers. Table 8 lists some typical characteristics of presently available minicomputers. Since the hardware, software, and architecture of these machines are in a present state of rapid commercial development, only general specifications are listed and these are often rapidly outdated. All of the features in Table 8 are available either as standard features or as standard options on the typical minicomputer.

The first five features of Table 8 are critical minicomputer requirements for real-time antenna measurements.

<table>
<thead>
<tr>
<th>TABLE 8. COMMON MINICOMPUTER FEATURES</th>
</tr>
</thead>
<tbody>
<tr>
<td>Program Interrupt</td>
</tr>
<tr>
<td>Real-Time Clock</td>
</tr>
<tr>
<td>Direct Memory Access (DMA)</td>
</tr>
<tr>
<td>4K-64K Memory Incorporating:</td>
</tr>
<tr>
<td>Fast Access Core Storage</td>
</tr>
<tr>
<td>Read-Only Memory (ROM)</td>
</tr>
<tr>
<td>Programmable ROM</td>
</tr>
<tr>
<td>Internal Registers</td>
</tr>
<tr>
<td>Scratch Pad Memory</td>
</tr>
<tr>
<td>Versatile Input/Output (I/O) Interconnection</td>
</tr>
<tr>
<td>Cycle Time: 400–64K Memory Incorporating:</td>
</tr>
<tr>
<td>Word Length: 12–18 Bit</td>
</tr>
<tr>
<td>&quot;Hardwired&quot; Arithmetic Units</td>
</tr>
</tbody>
</table>
The program interrupt feature allows Input/Output (I/O) devices to temporarily suspend an operating program so that I/O can be processed as required. The function of the real-time or external clock is to provide interrupts at prescribed rates. Thus, data can be acquired at precise sampling rates, or the CPU can execute a predetermined action at the correct instant in time. The direct memory access feature allows data to be transferred into and out of memory directly by allowing the I/O device to delay the CPU processing by one machine cycle while entering a single data word. This feature increases the data transfer rate. This can be significant when a large amount of data is required. The minicomputer should also provide for the interconnection of a variety of Input/Output devices. A specific I/O device required for the antenna measurement system is a digital tape unit for storage of measured data for subsequent processing. All minicomputers conform to the industry standard tape format. Tapes recorded on these machines interface with all large-scale computers which conform to the industry standard. Consequently, the Raytheon 703 Minicomputer for on-line operation and the Honeywell 600-line computers for off-line processing are both compatible with the Hewlett-Packard 2020E Digital Magnetic Tape Recorder.

C. TRANSMITTING ANTENNAS

A broadband transmitting antenna with selectable polarization (linear vertical, linear horizontal, or circular polarization) is desirable with all the systems which were considered in this study. In addition, the systems which have been selected for further development may require a reference antenna at the receiving site. With the Hybrid System, compensation for variations in power density at the receiving aperture is based on use of a reference antenna, and both of the phase/amplitude receivers require a frequency tracking channel. There are three general types of microwave antennas
for broadband operation: (1) frequency independent antennas such as the spiral and log periodic, (2) broadband horn antennas, such as the ridged horn or pin-wall, and (3) reflector antennas with broadband feeds. To obtain the desired gain, a reflector antenna with a broadband feed is the most appropriate choice for the current applications. Reflector antenna systems with good performance properties over octave or greater bandwidths can be achieved. Due to the increase of its electrical size, the gain of a reflector antenna increases by 6 dB per octave of frequency increase, but the feed horn effects may modify this change somewhat. However, since the space loss also increases by the same factor, the two effects tend to cancel. Consequently, this method would produce a more nearly constant power density at the receiving antenna than that which would be produced if a transmitting antenna with constant gain versus frequency were used. Because both the Hybrid and the phase/amplitude systems can accommodate changes in power density, it is not critical that the two effects do not exactly cancel over an octave bandwidth. The Hybrid system electronically compensates for these gain changes, while with the phase/amplitude system, calibration data will be used to computer-correct the measured test data and account for any variations in power density at the test antenna.

To limit ground reflections, a narrow-beam transmitting antenna is desired. Further, a fairly high gain transmitting antenna is desirable to produce an adequate received power level without an excessively high power transmitter. Calculations show (see Section III) that detection requirements can be met if the reflector is sized to provide a 30-dB gain transmitting antenna at 4 GHz. If the reflector is sized such as to provide 30 dB of gain at 4 GHz, the change of gain with frequency will compensate for the change of space loss with frequency. At frequencies below 4 GHz, the space loss decrease
to compensate for the reduction in gain; however, as the frequency is reduced from 4 GHz, the beamwidth of the reflector antenna broadens. Hence, for a reflector that has been sized at 4 GHz, the possibility of ground reflections increases as the frequency is reduced. Selectable polarization can be provided by employing a selectable polarization feed with a paraboloidal reflector. Common feed types for obtaining dual polarization include log-periodic and quad-ridge horns. These feeds, which can typically provide either dual linear (vertical and horizontal) or dual circular (right and left handed) polarizations, have good performance over octave or greater bandwidths. External power dividing and switching can be used with these dual polarized feeds to provide two orthogonal linear and two orthogonal circular polarizations from a single feed. Perhaps a more practical approach is to provide both a dual linear feed/reflector system and a dual circular/reflector system and then switch between the systems as required. Greater bandwidth and increased polarization isolation can be expected from the latter approach. Data on representative dual polarized feeds are shown in Table 9.

**TABLE 9. BROADBAND DUAL POLARIZED FEED DATA**

<table>
<thead>
<tr>
<th>Feed Type</th>
<th>Frequency Range</th>
<th>Polarization</th>
</tr>
</thead>
<tbody>
<tr>
<td>Quad-Ridge Horn</td>
<td>2-4</td>
<td>Dual linear</td>
</tr>
<tr>
<td>Quad-Ridge Horn</td>
<td>2-8</td>
<td>Dual linear</td>
</tr>
<tr>
<td>Quad-Ridge Horn</td>
<td>4-8</td>
<td>Dual linear</td>
</tr>
<tr>
<td>Quad-Ridge Horn</td>
<td>8-16</td>
<td>Dual linear</td>
</tr>
<tr>
<td>Log-Periodic</td>
<td>12-18</td>
<td>Dual linear</td>
</tr>
<tr>
<td>Log-Periodic</td>
<td>1-12</td>
<td>Circular</td>
</tr>
<tr>
<td>Log-Periodic</td>
<td>12-18</td>
<td>Circular</td>
</tr>
<tr>
<td>Log-Periodic</td>
<td>0.3-1</td>
<td>Dual linear</td>
</tr>
<tr>
<td>Log-Periodic</td>
<td>1-4</td>
<td>Dual linear</td>
</tr>
</tbody>
</table>

138
APPENDIX II

TEST FOR IDENTICAL ANTENNAS

This appendix presents the mathematical relations which allow one to define conditions for determining if two antennas are electrically identical. The fundamental equations which describe the performance of an antenna as a transmitter, receiver, and scatterer are presented in Appendix III. The circuit analog of an antenna's transmitting and receiving properties can be represented by a Thevenin equivalent circuit. However, the scattering properties cannot be described by a simple Thevenin equivalent since it is not generally possible to compute the scattered power from such a Thevenin equivalent circuit. The equivalent circuit of Figure 30 can be used as a circuit analog for antenna scattering [11].

In this circuit, \( V \) is the open circuit Thevenin equivalent voltage, \( Z_a \) is the antenna impedance, \( Z_L \) is the load impedance, \( Z_1 \) and \( Z_2 \) are two impedances to be determined such that the voltage \( V_s \) between terminals 1 and 2 is proportional to the scattered field, and \( V_L \) is the load voltage. The terminal voltage \( V_s \) may be expressed as

\[
V_s = V Z_2 \left( \frac{Z_2}{Z_1 + Z_2} + \frac{Z_L V}{Z_L + Z_a} \right)
\]  

(A-5)

Since the load current \( I_L \) is given by

\[
I_L = - \frac{V}{Z_a + Z_L}
\]  

(A-6)

and the scalar form of the scattering equation is

\[
E^s_k (Z_L) = E^s(0) - \frac{Z_L I(0)}{Z_L + Z_a} \cdot E^r_k
\]  

(A-7)

where

\[
E^s(0) = E^s_k (Z_L) @ Z_L = 0, \text{ and}
\]
Figure 30. Equivalent bridge circuit for antenna scattering
I(0) = I \ell @ Z_{\ell} = 0.

one has

\[ Z_{a} \frac{E^{S}(Z_{\ell})}{E^{r}} = \frac{Z_{a} E^{S}(0)}{E^{r}} + \frac{Z_{\ell} V}{Z_{\ell} + Z_{a}}. \]  

(A-8)

If one lets

\[ \frac{Z_{a} E^{S}(0)}{E^{r}} = - \frac{V Z_{2}}{Z_{1} + Z_{2}}, \]  

(A-9)

then

\[ E^{S}(Z_{\ell}) = \frac{E^{r}}{Z_{a}} V_{s}, \]  

(A-10)

where \( E^{r} \) is the radiated field strength per unit current. From Equations A-6 and A-7 one has

\[ \frac{E^{r}}{Z_{a}} = - \frac{E^{S}(0) - E^{S}(\infty)}{V}, \]  

(A-11)

where

\[ E^{S}(\infty) = E^{S}(Z_{\ell}) @ Z_{\ell} = \infty. \]

From Equations A-11 and A-9, one gets

\[ \frac{Z_{2}}{Z_{1}} = - \frac{E^{S}(0)}{E^{S}(\infty)}. \]  

(A-12)

This ratio relates the scattered field to the open circuit Thevenin equivalent voltage by the circuit of Figure 30 and Equation A-10. The relation given here can be used to determine whether two antennas are electrically identical. With respect to transmission and reception, two antennas whose impedance \( Z_{a} \) is identical as a function of frequency are identical. With respect to scattering, Equation A-12 must be satisfied by both antennas. Thus two antennas are electrically identical at a given angular frequency \( \omega \).
and aspect angle if

1) their input impedances are equal, and

2) the ratio of the scattered field for short and open circuit load conditions is equal.

\[ \frac{E_s(Z_L = 0, \omega)}{E_s(Z_L = \infty, \omega)} \]

A measurement of \( Z_a(\omega) \) and \( \frac{E_s(Z_L = 0, \omega)}{E_s(Z_L = \infty, \omega)} \) for all aspect angles of interest is required to determine whether two physically similar antennas are electrically identical.
APPENDIX III

THE ANTENNA AS A TRANSMITTER AND RECEIVER

The electric field \( \overline{E} \) at a great distance from an antenna is transverse to the direction of propagation and can be written as

\[
\overline{E} = \overline{E}_\theta + \overline{E}_\phi = -j \frac{\eta I_t}{2\lambda} \overline{h}_t \frac{e^{-jkr}}{r},
\]

(A-13)

where

- \( \eta \) = impedance of free space = 376.7 ohms,
- \( I_t \) = transmitting antenna input-terminal current,
- \( \lambda \) = wavelength,
- \( \overline{h}_t \) = effective transmitting antenna height,
- \( k = \frac{2\pi}{\lambda} \),

and \( r, \theta, \phi \) are conventional spherical coordinates. The effective transmitting antenna height contains both the polarization and pattern information.

For the transmission of EM energy between a pair of antennas, the electric field \( \overline{E}_r \) at the receiving antenna, at a range \( r \) from the transmitter, is given by Equation A-13, and this field induces an open circuit voltage at the receiver terminals given by [12]

\[
V_r = \overline{h}_r \cdot \overline{E}_r,
\]

(A-14)

where \( \overline{h}_r \) is the effective receiving antenna height.

The circuit analogy of the receiving properties of the receiving antenna can be represented by the Thevenin equivalent circuit shown in Figure 31. The current \( I_r \) through the load \( Z_L \) on the receiver terminals can be obtained from a Thevenin circuit as

\[
I_r = -\frac{V_o}{Z_a + Z_L}.
\]

(A-15)
Figure 31. Thevenin equivalent circuit antenna analog
Thus the voltage \( V_r \) across the load \( Z_L \) is given by

\[
V_r = \left( j \frac{\pi}{2\lambda} \right) \frac{e^{-jkr}}{r} I_t (\vec{h}_r \cdot \vec{h}_t) \left( \frac{Z_L}{Z_a + Z_L} \right).
\]  

(A-16)

The transmitting antenna terminal current \( I_t \) is related to the voltage \( V_t \) impressed on the terminals by the transmitting antenna impedance \( Z_t \) by

\[
V_t = I_t Z_t.
\]  

(A-17)

Thus the ratio between the transmitted and received voltage is given as

\[
\frac{V_r}{V_t} = \left( j \frac{\pi}{2\lambda} \right) \frac{e^{-jkr}}{r} (\vec{h}_r \cdot \vec{h}_t) \left( \frac{Z_L}{Z_a + Z_L} \right) \frac{1}{Z_t} = S_{12}.
\]  

(A-18)

We can relate this expression to the usual definition of the antenna gain by using the Friis transmission formula,

\[
\frac{W_r}{W_t} = \frac{A_{er}}{2\lambda} \frac{A_{et}}{2\lambda} = \left[ \frac{\lambda^2}{4\pi} D_r \right] \left[ \frac{\lambda^2}{4\pi} D_t \right] \frac{1}{2} \frac{1}{r^2}.
\]  

(A-19)

where \( A_{er}, A_{et} \) = effective aperture of the receiver and transmitter, respectively,

\( D_r, D_t \) = directivity of the receiver and transmitter, respectively, and

\( W_r, W_t \) = power received and power transmitted, respectively.

Assuming a lossless antenna structure, the power gains and the directivities are equal so that \( G_r = D_r \) and \( G_t = D_t \), with the result that

\[
\frac{W_r}{W_t} = \left( \frac{\lambda}{4\pi} \right)^2 \frac{G_r}{G_t} = \left| \frac{E_r}{E_t} \right|^2.
\]  

(A-20)

Thus, the antenna voltage gain transfer is given by

\[
\left| \frac{V_r}{V_t} \right| = \frac{\lambda}{4\pi} \left| G_r \right| \left| G_t \right|.
\]  

(A-21)

If the two antennas are identical, \( Z_a = Z_t \), and if they are aligned such that \( \vec{h}_r = \vec{h}_t (\theta, \phi) = \vec{h}_r (\theta, \phi) \), i.e., no polarization or pattern mismatch, one has
from Equation A-18

$$S_{12} = \left( \frac{\pi}{2\lambda} \frac{e^{-jkr}}{r} \right) (h)^2 \frac{Z_L}{Z_a + Z_L} \cdot \frac{1}{Z_a}.$$  \hspace{1cm} (A-22)

The scattering parameter $S_{12}$ can be measured using a network analyzer scattering parameter test set. If the range $r$ is known $h^2(w)$ and thus $h(w)$ can be determined. Once $h(w)$ is known, the phase and amplitude properties of the antenna can be calculated. The transmitted field due to an excitation current $I_t$ or the received voltage due to an incident electric field $E_i$ can be determined using Equations A-13 and A-14, respectively.
APPENDIX IV

THE ANTENNA AS A SCATTERER

This appendix presents the fundamental equations which describe an antenna as a free space scatterer. The Thevenin equivalent circuit which represents the antenna when acting as a receiver is not sufficient to fully describe both the receiving and scattering properties of the antenna when a plane wave is incident. In order to specify the properties of an antenna completely, it is necessary to consider the three operating conditions of the antenna illustrated in Figure 32. The surface S is designated as the terminal surface. It can be shown \([11, 13, 14]\) that the basic equation which describes the antenna's scattered electric field \(E^s\) in terms of the operating conditions of Figure 32 is

\[
E^s(Z) = E^s(0) - \frac{Z_a I(0) E^r}{Z_a + Z_L},
\]

where

\(Z_a\) = antenna impedance,

\(E^r\) = radiated field per unit current excitation, and

\(Z_L\) = load impedance, and the \((0)\) notation indicates \(Z_L = 0\).

This equation indicates that the field scattered by an antenna at a given frequency and direction is a function of both the load and the antenna impedance. Equation A-23 may be transformed to a more convenient form as follows. First, set \(Z_L = Z_a^*\) (complex conjugate of the antenna impedance) and then solve for \(E^s(0)\):

\[
E^s(0) = E^s(Z_a^*) + \frac{Z_a^* I(0) E^r}{2 R_a}.
\]

The terminal receiving current is given by

\[
I(Z_L) = \frac{-V}{Z_a + Z_L},
\]

(A-25)
Figure 32. General antenna operating conditions
where $V$ is the open circuit received voltage. Using Equation A-25 first with $Z_L = 0$ and then with $Z_L = Z_a^*$, and solving for $I(0)$, one gets
\[ I(0) = \frac{2R_a I(Z_a^*)}{Z_a} \quad (A-26) \]

Thus, Equation A-24 can be written:
\[ \overline{E}^S (0) = \overline{E}^S (Z_a^*) + \frac{Z_a^*}{Z_a} I(Z_a^*) \overline{E}^T \quad (A-27) \]

Substituting Equations A-26 and A-27 into Equation A-23 results in the following expression for the scattered field:
\[ \overline{E}^S (Z_L) = \overline{E}^S (Z_a^*) - \left[ I(Z_a^*) \overline{E}^T \right] \Gamma_m \quad (A-28) \]

where
\[ \Gamma_m = \frac{Z_L - Z_a^*}{Z_L + Z_a^*} \quad (A-29) \]

defines a modified voltage reflection coefficient. The first term of Equation A-28 may be interpreted as due to structural scattering and is a constant at a given frequency and direction; the second term is called the antenna mode scattering and is a function of the load impedance. The antenna mode scattering has the same pattern as the antenna when the antenna acts as a transmitter.

For most practical aperture type antennas, the maximum antenna mode scattering obtained for $|\Gamma_m| = 1$ is very much greater than the structural scattering in the "main beam" region. In short, the antenna is designed to be an efficient receiver of energy incident within the main beam. Thus assuming a short circuit at the antenna terminals ($Z_L = 0$, $\Gamma_m = -1$), the scattered field is
\[ \overline{E}^S (0) = \overline{E}^S (Z_a^*) + I(Z_a^*) \overline{E}^T \quad (A-30) \]
which is approximated as
\[ \overline{E}^S (0) = I(Z_a^*) \overline{E}^T \quad (A-31) \]
Recalling that $\overline{E^r}$ is the radiated field for a unit current excitation, one has

$$\overline{E^r} = -j\frac{\pi}{2\lambda} \overline{h_t} \frac{e^{-jkr}}{r} \quad (A-32)$$

Also, the received current due to an incident electric field $\overline{E^i}$ (See Appendix III, Equations A-14 and A-15) is given by

$$I_{a^*} = \frac{\overline{h_r} \cdot \overline{E^i}}{Z_a + Z_a^*} = \frac{v_r}{Z_a + Z_a^*} \quad (A-33)$$

Thus, the scattered field can be written as

$$\overline{E^s}(0) = \left(\frac{\overline{h_r} \cdot \overline{E^i}}{Z_a + Z_a^*}\right) \left(-j\frac{\pi}{2\lambda} \overline{h_t} \frac{e^{-jkr}}{r}\right) \quad (A-34)$$

If we assume no polarization or pattern mismatch between $\overline{h_r}$, $\overline{E^i}$, and $\overline{h_t}$ for transmit and receive, as would be true for the antenna boresight axis with the same sense polarization, $h_r = h_t = h$, and Equation A-34 becomes

$$\overline{E^s}(0) = \left(-j\frac{\pi}{2\lambda}\right) \frac{h^2}{2R_a} \frac{e^{-jkr}}{r} \overline{E^i} \quad (A-35)$$

This scattered field can be measured at a range $r$ from the antenna. For antenna calibration, let a sphere be substituted for the antenna at range $r$ and the scattered field of the sphere measured as

$$\overline{E^s}_{sphere} = \overline{E^i} \sqrt{\frac{\sigma_{sphere}}{4\pi}} \frac{e^{-jkr}}{r} \quad (A-36)$$

The ratio of the antenna and the sphere scattered field is

$$\frac{\overline{E^s}(0)}{\overline{E^s}_{sphere}} = \left(-j\frac{\pi}{2\pi}\right) \frac{h^2}{2R_a} \sqrt{\frac{4\pi}{\sigma_{sphere}}} \quad (A-37)$$

The factor $\sqrt{\frac{4\pi}{\sigma_{sphere}}}$, where $\sigma$ = the scattering cross-section of the sphere, is known exactly for the sphere. Thus a measurement of the ratio of the scattered fields in magnitudes and phase together with a measurement of the antenna
radiation resistance $R_a$ allows the antenna height $h$ to be determined. This measurement can be performed conveniently over a wide frequency band.
BROADBAND ANTENNA MEASUREMENT TECHNIQUES

This report presents the results of a program to study and investigate advanced measurement techniques for evaluating the performance of broadband antenna systems for use in high resolution radar systems. New techniques to measure, record, and analyze antenna gain and pattern performance were studied, and the feasibility of developing the necessary instrumentation to perform these measurements was investigated. Systems based on the use of broadband noise signal sources and systems using sweep frequency techniques were studied. It was concluded that systems using broadband noise signal sources would not be cost-effective. Preliminary design of a broadband amplitude-only electronically integrating sweep frequency system was completed. It was concluded that this system could be implemented immediately and that it would provide an effective first step in realization of the ultimate phase plus amplitude broadband antenna measurement system. The implementation and operation of sweep frequency amplitude plus phase systems were studied, and an effective approach to the realization of a phase/amplitude system was identified. Recommendations for implementing this approach are presented.
<table>
<thead>
<tr>
<th>KEY WORDS</th>
<th>LINK A</th>
<th>LINK B</th>
<th>LINK C</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>ROLE</td>
<td>WT</td>
<td>ROLE</td>
</tr>
<tr>
<td>Antenna Measurements</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Pulse Distortion</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Spatial Patterns</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Phase/Amplitude</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Broadband Patterns</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Group Velocity</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Group Delay</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>