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EMERGENCY VEHICLE ALERT SYSTEM

PHASE II REPORT
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Prepared for:
George C. Marshall Space Flight Center
Attn: Mr. James Currie
Marshall Space Flight Center, AL 35812

Prepared by:
Applied Research, Inc
P. O. Box 11220
Huntsville, AL 35814-1220

Tom Barr
Warren Harper
Bill Reed
David Wallace

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Executive Summary
Emergency Vehicle Alert System (EVAS)
Phase I & II

The Emergency Vehicle Alert System (EVAS) is designed to warn hearing impaired drivers of potential dangers of approaching emergency vehicles. The concept was suggested by Gallaudet University of Washington DC. The rudiments of EVAS are to provide a warning device and display for drivers to show the relative approach angle and distance of emergency vehicles. Gallaudet University approached NASA with this problem. It was circulated to the NASA centers. Mr. Jim Currie of the Marshall Space Flight Center proposed a solution based on radio (RF) communications between emergency and private vehicles. The emergency vehicle would transmit a message sequence then broadcast its direction of travel and identity. The private or receiving vehicle would receive the signal and compare it's direction with that of the emergency vehicle and thus determine the actual approach direction. Range would be derived from the strength of the received signal.

A contract was let to Applied Research Incorporated of Huntsville to study this concept, simulate various vehicle approach situations and build a conceptual system. The simulation showed that some situations of the emergency vehicle to private vehicle movement could not be resolved. For example, cases where the two vehicles were traveling in the same direction, but on parallel streets. Problems of this type led to the concept of sending a signal from the emergency vehicle that could be used to derive relative direction information of the private vehicle as related to the emergency vehicle.

The approach to direction determination was to use a coded beam antenna somewhat similar to an aircraft direction finding system, but based on a unique approach using digital generated and coded antenna patterns. For the prototype system, three monopoles on a ground plane are programmed with a set of phase and amplitude sequences that are controlled by a digital generator (computer). The sequence is selected to produce a sweeping antenna beam from the emergency vehicle. The amplitude and dwell time of each beam sweep generates a code at the receiver. This code has a pattern that changes depending on the relative angular position of the private vehicle to the emergency vehicle. The accuracy of the angle measurement is related to the number of antenna elements. This, in turn, defines the fineness of the antenna code. Although designed for the EVAS it is adaptable to other applications such as spacecraft rendezvous or aircraft collision avoidance.

An initial problem of a widely used public emergency radio system is licensing of the transmission device and congestion of multiple user radio bands. The recent Federal Communication Commission resolution to allow the unlicensed use of the 900, 2500 and 5000 MHz radio bands with RF power of up to 1W with no restrictions on location or antenna gain came at a fortunate time for this program. The FCC rule 15 provided a easy solution for EVAS communications. The only restriction was the requirement that all transmissions use spread spectrum modulation. Spread spectrum provided a solution to possible jamming of signals by other users. Spread spectrum jamming resiliency comes from the fact that the signal is spread over a wide frequency band, it's dwell time at any one frequency is small and not likely to collide with other users. Jamming isolation and accommodation of many users is related to the bandwidth and coding length. Additionally spread spectrum is now economical since public sector involvement has generated research and marketing of complete spread spectrum coding and decoding electronics on a single integrated circuit. Current prices of IC's are $40 with the price expected to drop fast as more user applications develop.

Data transmission is one way, from the emergency vehicle to the private vehicle. The basic message is a sequential message that contains vehicle identification, magnetic heading, antenna code synchronizing and related information such as vehicle speed and possibly special warning announcements. This message sequence and antenna control code is generated by a small microprocessor. Reception and decoding of this signal by the private vehicle is also managed by a microprocessor. Modulation and demodulation is handled by the spread spectrum receiver.

The conclusion of the Phase I and Phase II studies was to show that it was possible to design and build a EVAS. During Phase I a spread spectrum transmitter and receiver were developed and tested. The frequency of operation was 912 MHz. The testing showed that the dynamic range and sensitivity of the system would be adequate for EVAS. In Phase II the transmitter and receiver were refined and a coded beam antenna system was developed to allow relative direction between the transmitter and receiver to be determined. Field testing of the system under static and dynamic (moving vehicles) was performed with good results. Using a simple signal processing algorithm and only performing one pass, direction between the transmitter and receiver was determined correctly in 66 out of 69 cases. Follow on work will be to concentrate on building a more robust transmitter, antenna and receiver system and integration of a conceptual control processor. In addition, testing will be performed in congested areas where multipath and signal blockage are likely to occur.

1.0 INTRODUCTION

The Emergency Vehicle Alert System (EVAS) project was initiated by the Marshall Space Flight Center Technical Utilization Office to provide a warning device for hearing impaired drivers. The alert is to provide a visual display of an approaching emergency vehicle. The display is to show the approach path and distance so that the hearing impaired driver can take appropriate action. During Phase 1 a radio frequency concept per Mr. Jim Currie's direction was developed to accomplish the above. The license free 902-928 MHz band using spread spectrum modulation was selected for use. A transmitting and receiving system was developed and preliminary testing performed. The reader is referred to the Phase I final report for further details.

1.1 EVAS PHASE II REQUIREMENTS

The purpose of Phase II effort has been to refine the transmitter and receiver developed during Phase I and to develop an antenna system that would allow direction to be determined. The specific tasks to be accomplished are as follows:

Task 1: Transmitter/Receiver and Antenna Switch

This task was to develop a transmitter and receiver suitable for field testing that met Part 15 of the FCC rules. Interfaces between the transmitter and receiver and the microcontroller were to be defined and developed. In addition, a switching assembly for a three element directional antenna was to be designed and developed.

Task 2: Direction Determining Antenna/System Test

This task was to design, fabricate and test a three element directional antenna with the transmitter and receiver developed in Task 1. The testing was to include pattern measurements of the antenna system and field testing of the EVAS system under dynamic conditions at various distances and directions.

1.2 SYSTEM DESCRIPTION

The Emergency Vehicle Alert System is designed to inform the driver of a vehicle that an emergency vehicle, such as an ambulance or a fire truck is in the local area and to give the direction and approximate distance to the emergency vehicle. The system employs the use of a transmitting unit on each emergency vehicle and a receiving unit on any vehicle whose driver wishes to use the system. A radio frequency in a license-free band is used for transmission from the emergency vehicle to the user vehicle. The system provides direction and distance information to the driver of the user vehicle by the following means:

1. Amplitude of the received signal is used to determine the approximate distance from the emergency vehicle to the receiving vehicle.

2. A directional antenna subsystem on the emergency vehicle is used to determine the direction of the user vehicle, relative to the body coordinate system of the emergency vehicle.

3. Telemetry data, including magnetic compass heading, are transmitted to the user vehicle to enable computation of direction relative to the user vehicle body axes to be made.

1.2.1 DIRECTION DETERMINATION

The method of determining direction of the user vehicle relative to the emergency vehicle body axes will now be explained. The transmitting system on the emergency vehicle uses a three-element antenna array in which the amplitudes and phases of the voltages driving the three elements are electronically controlled. By selecting some desired combination of these amplitudes and phases, a radiation pattern of some desired shape can be formed. The relative amplitude of the received signal at the user's vehicle will then be dependent on his direction relative to the emergency vehicle.

The EVAS, as currently designed, uses six radiation patterns of selected shapes. These patterns are sequentially activated in rapid succession, producing a sequence of amplitude values at the receiver. An example of such a sequence is shown in Figure 1.2.1-1. For this example, it is assumed that the user vehicle direction is 60 degrees relative to the emergency vehicle coordinate system.

It may be seen by inspection of Figure 1.2.1-1 that the shape of the waveform produced by the sequential activation of the six radiation patterns is determined by the direction of the transmission path. Any given direction produces a unique waveform; hence, direction can be determined by observation of the waveform.

Combination of this directional information with compass headings and other relevant telemetered information by the processor in the user vehicle produces a set of output voltages which activate a visual display unit on the user's instrument panel, giving the user the required direction and distance information.

1.2.1.1 Antenna Array

The antenna array on the emergency vehicle consists of three monopoles, mounted on a ground plane. They are spaced one half wavelength apart, as shown in Figure 1.2.1.1-1. Designating the monopole elements as A, B and C, the six selected excitations of the elements are described below.

1. All of the power is divided between element A and element B. Elements A and B are driven in phase. No power is delivered to element C.

2. All of the power is divided between element B and element C. Elements B and C are driven in phase. No power is delivered to element A.

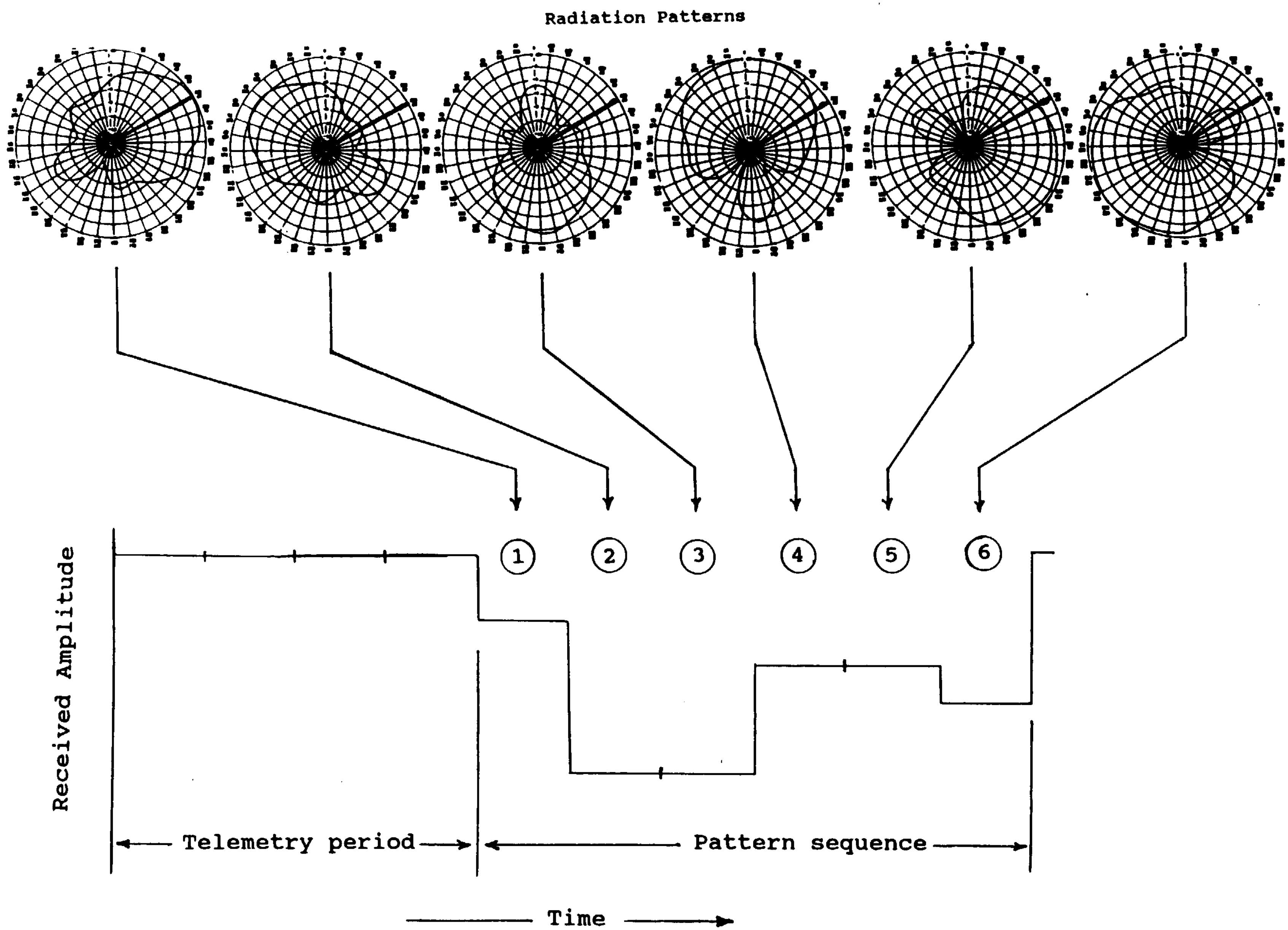


Figure 1.2.1-1 Received Amplitude Sequence for 60° Direction

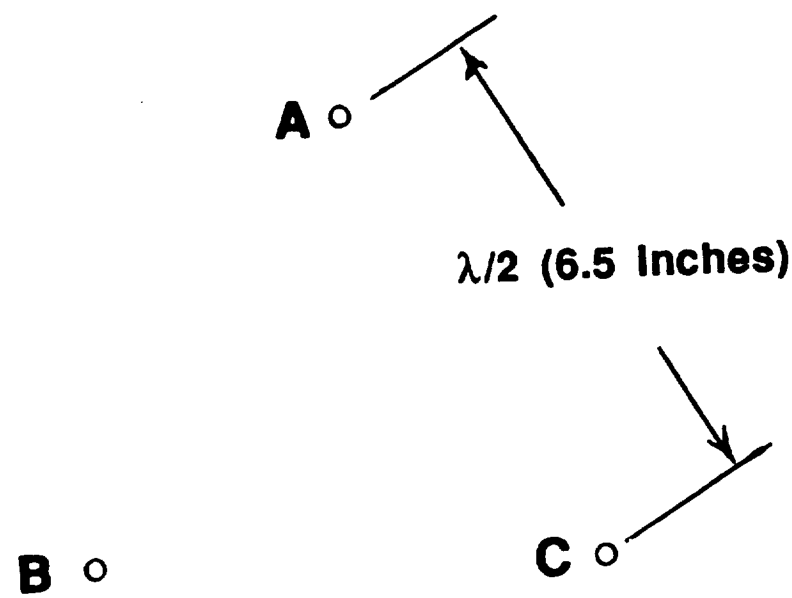


Figure 1.2.1.1-1 Antenna Element Arrangement for Monopole Array

3. All of the power is divided between element C and element A. Elements C and A are driven in phase. No power is delivered to element B.

4. Part of the power is equally divided between element A and element B. Elements A and B are driven in phase. The remaining power is delivered to element C. The phase of the voltage driving element C is 60 to 70 degrees relative to that driving elements A and B.

5. Part of the power is equally divided between element B and element C. Elements B and C are driven in phase. The remaining power is delivered to element A. The phase of the voltage driving element A is 60 to 70 degrees relative to that driving elements B and C.

6. Part of the power is equally divided between element C and element A. Elements C and A are driven in phase. The remaining power is delivered to element B. The phase of the voltage driving element B is 60 to 70 degrees relative to that driving elements C and A.

The six radiation patterns that result from the six element excitations described above are sequentially activated by control circuit which continuously repeats the six-pattern sequence. Each sequence also includes a time period during which telemetry data are transmitted. The element excitation for this time period provides equal amplitude and equal phase for the three elements.

2.0 TRANSMITTER/RECEIVER

Based upon the favorable results and experience obtained with the OCI 100 Spread Spectrum ASIC during Phase 1, it was decided to use it in the transmitter and receiver. The OCI 100 can be used to synthesize both the transmit frequency and the receiver local oscillator frequency and direct sequence spread and despread the data to be communicated in compliance with Part 15 of the FCC Rules. A copy of the OCI-100 data sheet is enclosed as Appendix A. The data sheet describes the OCI-100 in the transmission mode and the reception mode.

2.1 TRANSMITTER DESCRIPTION

Figure 2.1-1 is a schematic diagram of the transmitter. VCO, D-8914, is phase locked to 912.26 MHz using the OCI 100 and the 4.9152 MHz oscillator as a reference. Figure 2.1-2 shows how the transmitter frequency is synthesized. The 912.26 MHz VCO is divided by 64 (PT) by MC12073 and input to the divide by TXFREQ counter of the OCI 100. The 4.9152 MHz oscillator is input to the divide by TXREF counter of the OCI 100. The outputs of these two counters are then phase compared to generate an error signal to phase lock the D-8914 VCO. The transmit frequency is governed by the following expression

FTX = PT * TXFREQ * FREF / TXREF
= 64 * TXFREQ 8 4.9152 / TXREF MHz

The value of RXFREQ and FREF are set by a serial bit stream clocked into the OCI 100 as described in the data sheet. Normally this serial bit stream would be generated by the controlling microprocessor. In order to facilitate testing without a microcontroller, a programmable arithmetic logic (PAL) device was programmed to generate the serial bit stream to set the counters to phase lock the VCO to 912.26 MHz. The PAL is activated by the momentary switch. TXFREQ is set to 29 and TXREF is set to 10.

Data to be transmitted is clocked into the OCI 100 on pin (5), Figure 2.1-1, where it is spread by a 16 chip sequence. The spread data is then output on pin 18 and summed with the phase detector output, pin 26, and presented to pin 2 of the VCO. The VCO is phase modulated by the data chip sequence since the chip rate is outside the loop bandwidth. The deviation of the data is set by the 10K potentiometer. The output of the VCO is then amplified by MC 5809 to 15 dBm and presented to the transmitting antenna array. A portion of the output is coupled to the divide-by-64 counter (MC 12073) whose output is amplified by the MRF901 and then input to the OCI-100 on pin 20. This completes the loop to cause the VCO to be phase locked to the input to the OCI 100 on pin 9 according to the above equation.

2.1.1 TRANSMITTER/MICROPROCESSOR INTERFACE

Figure 2.1.1-1 shows a block diagram of the transmitter/microprocessor interface. The transmitter/microprocessor interface consists of the following signals.

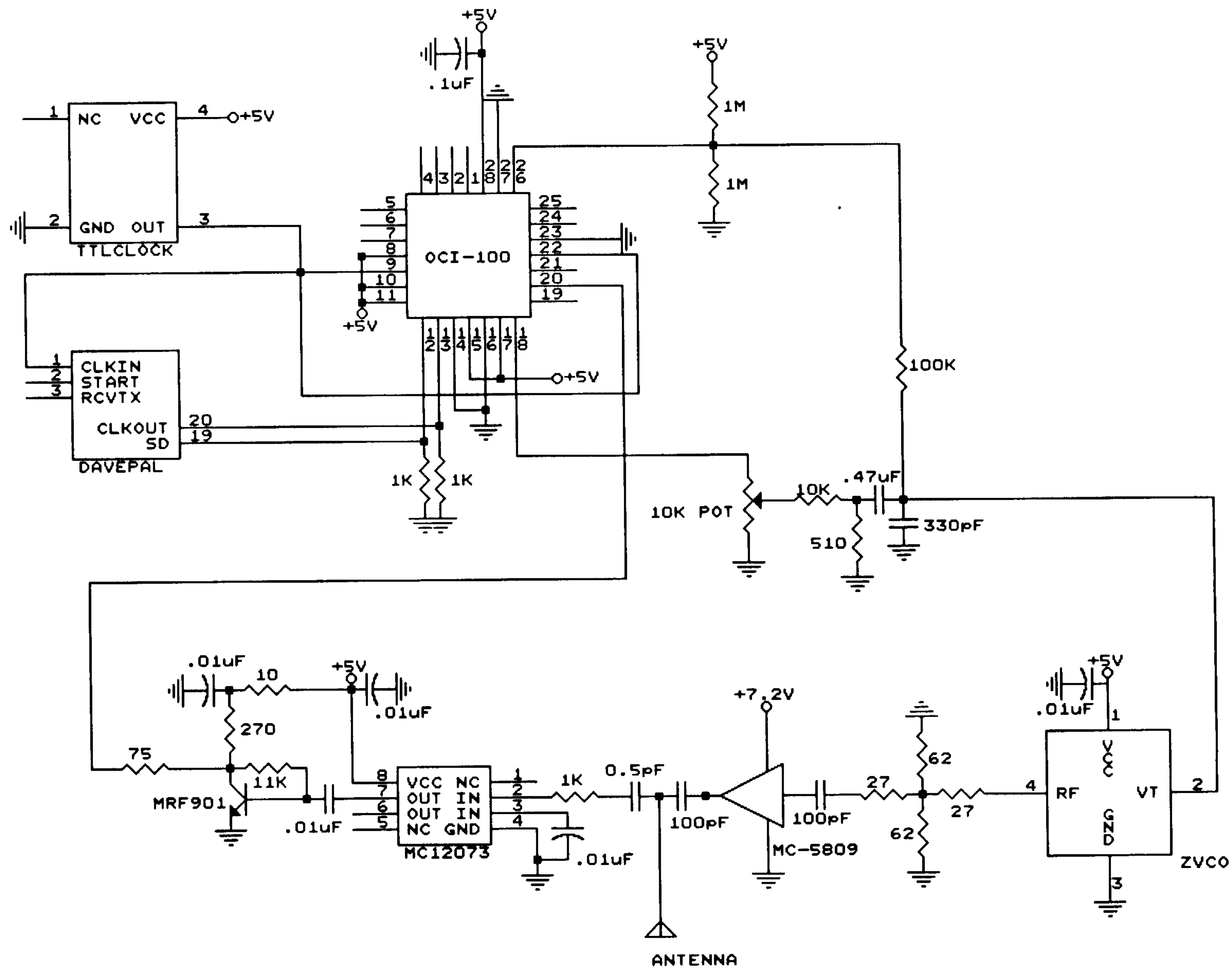
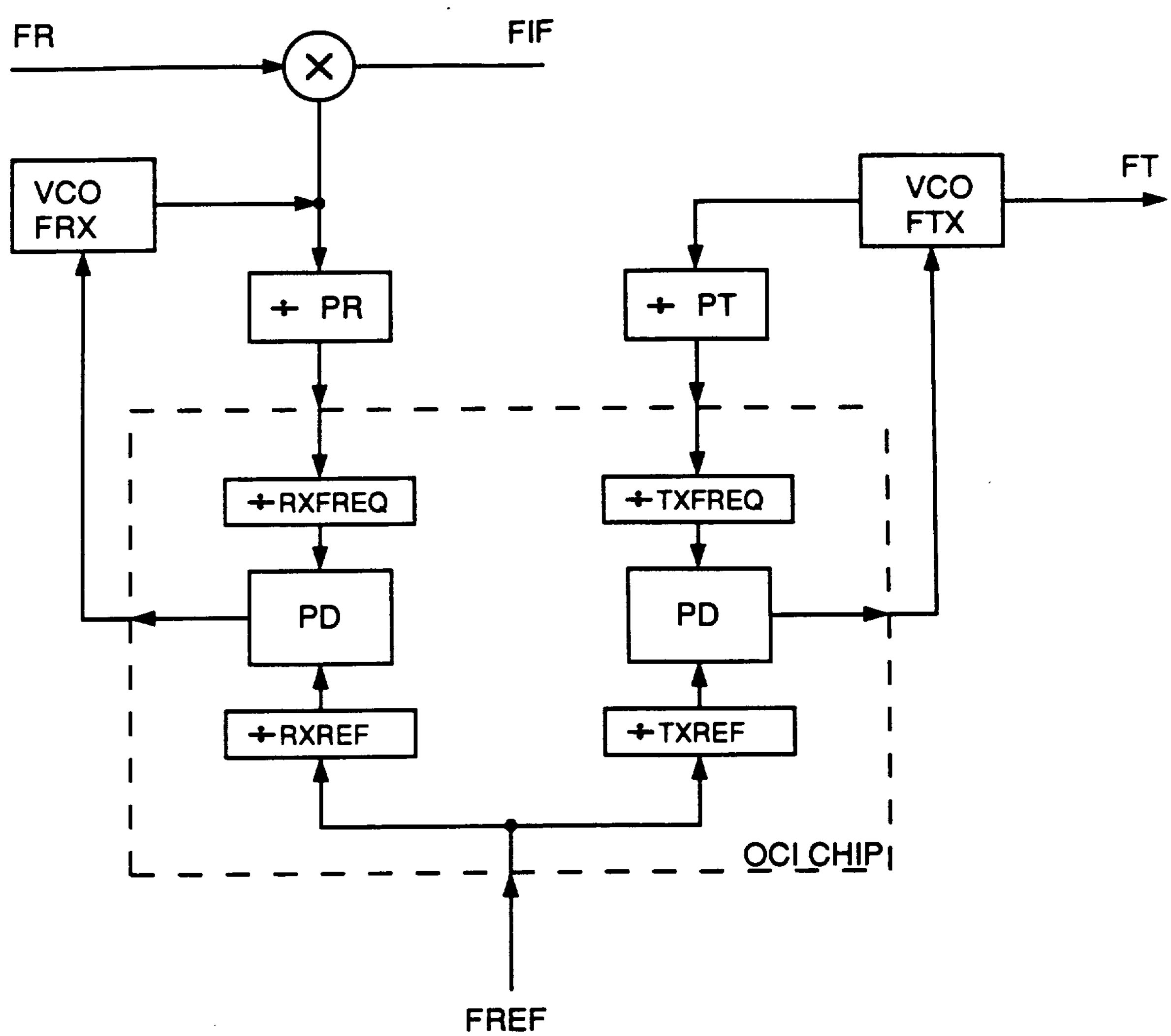


Figure 2.1-1 Transmitter Schematic



$$FIF = FRX - FR$$

$$FT = FR$$

$$\frac{FRX}{PR * RXFREQ} = \frac{FREF}{RXREF}$$

$$\frac{FTX}{PT * TXFREQ} = \frac{FREF}{TXREF}$$

Figure 2.1-2 Frequency Synthesis

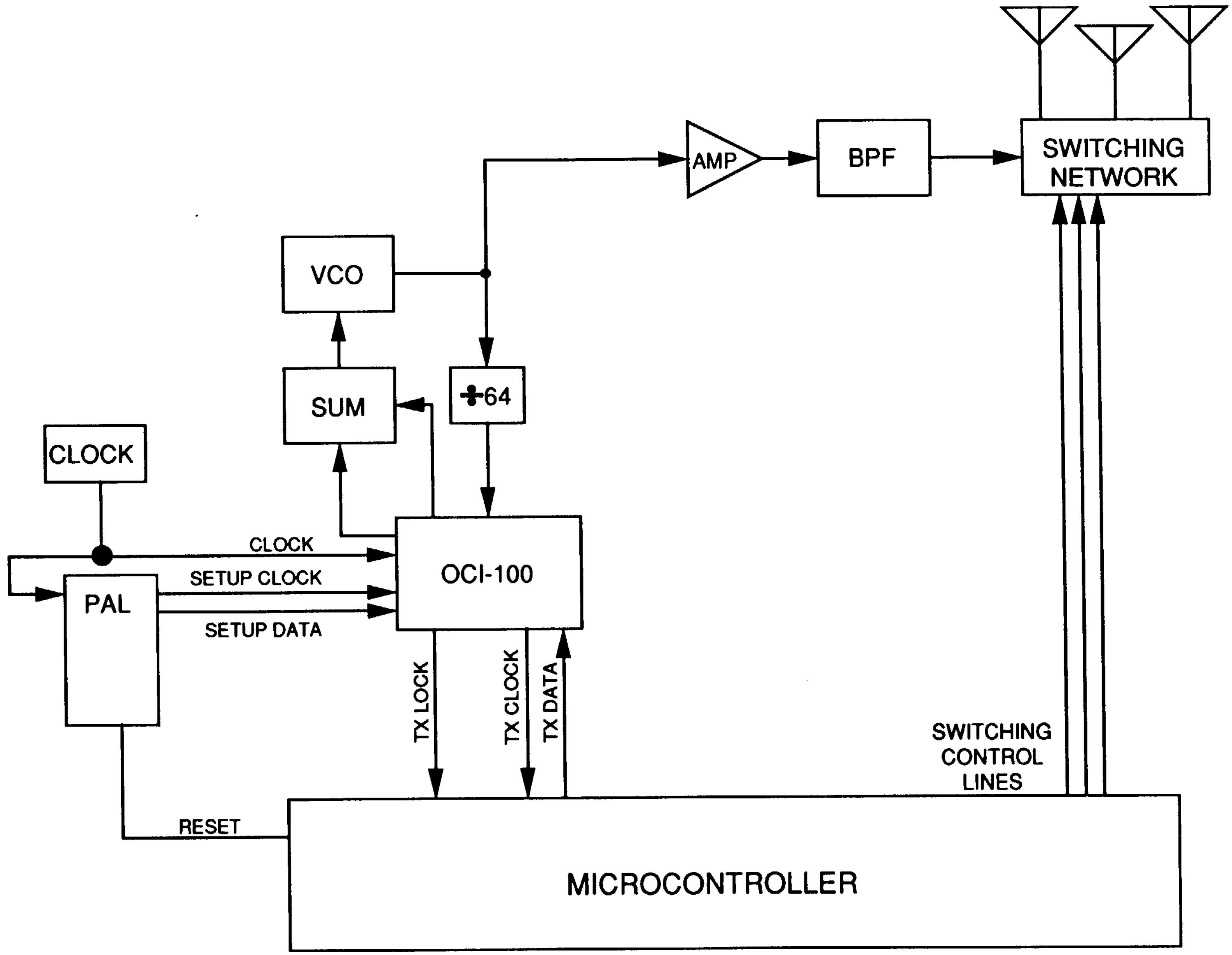


Figure 2.1.1-1 Transmitter Block Diagram with Microprocessor Interface

118-1041-1192-dwg

Transmitter -> Microprocessor

LOCK - Logical signal giving the status of the VCO. If the VCO is locked the signal is a logical 1 else it is a logical 0.

TXCLOCK - A clock signal that can be used to clock data into the transmitter.

Microprocessor -> Transmitter

DATA - Logical data to be transmitted by the transmitter.

ANTENNA CONTROL - *** logical lines used to control the stepping pattern of the antenna.

RESET - Resets the transmitter to initial conditions.

2.2 <u>RECEIVER DESCRIPTION</u>

The electrical design of the receiver developed in Phase I was selected to be used in the deliverable receiver since its performance was very good. Some physical and electrical modifications were made as will be described later. The receiver in Phase I had the down convertor and IF amplifier/detector in separate modules to provide good isolation. It was decided that the receiver could be assembled in one module if care in parts placement and internal shielding were employed. A modified down convertor, ATV4 in Figure 2.2-1, from Communication Concepts was employed due to low cost. The ATV4 did not have sufficient RF or IF selectivity. A helical filter was added to the input of the ATV to reject signals from the public telephone service at 835 to 888 MHz. A narrow bandpass filter was added to the output of the ATV4 to improve sensitivity. A MC 13055 is used to demodulate the data plus spreading code and present it to the OCI 100 where it is despread and presented to the microcontroller. The MC 13055 has a linear signal detector with a wide dynamic range. This detector output is amplified and filtered in two separate circuits. One circuit (antenna level) has a time constant consistent with the rate that the antenna patterns change (0.1 milliseconds) and is used to determine direction. This second detector output (signal level) has a time constant greater than the period of the pattern switching rate. This allows it to be used to determine average signal level for use in distance determination. Either time constant may be changed easily by altering the value of a capacitor.

Figure 2.2-2 is a calibration curve of the receiver signal level detector. The dynamic range is from -35dBm to -95dBm.

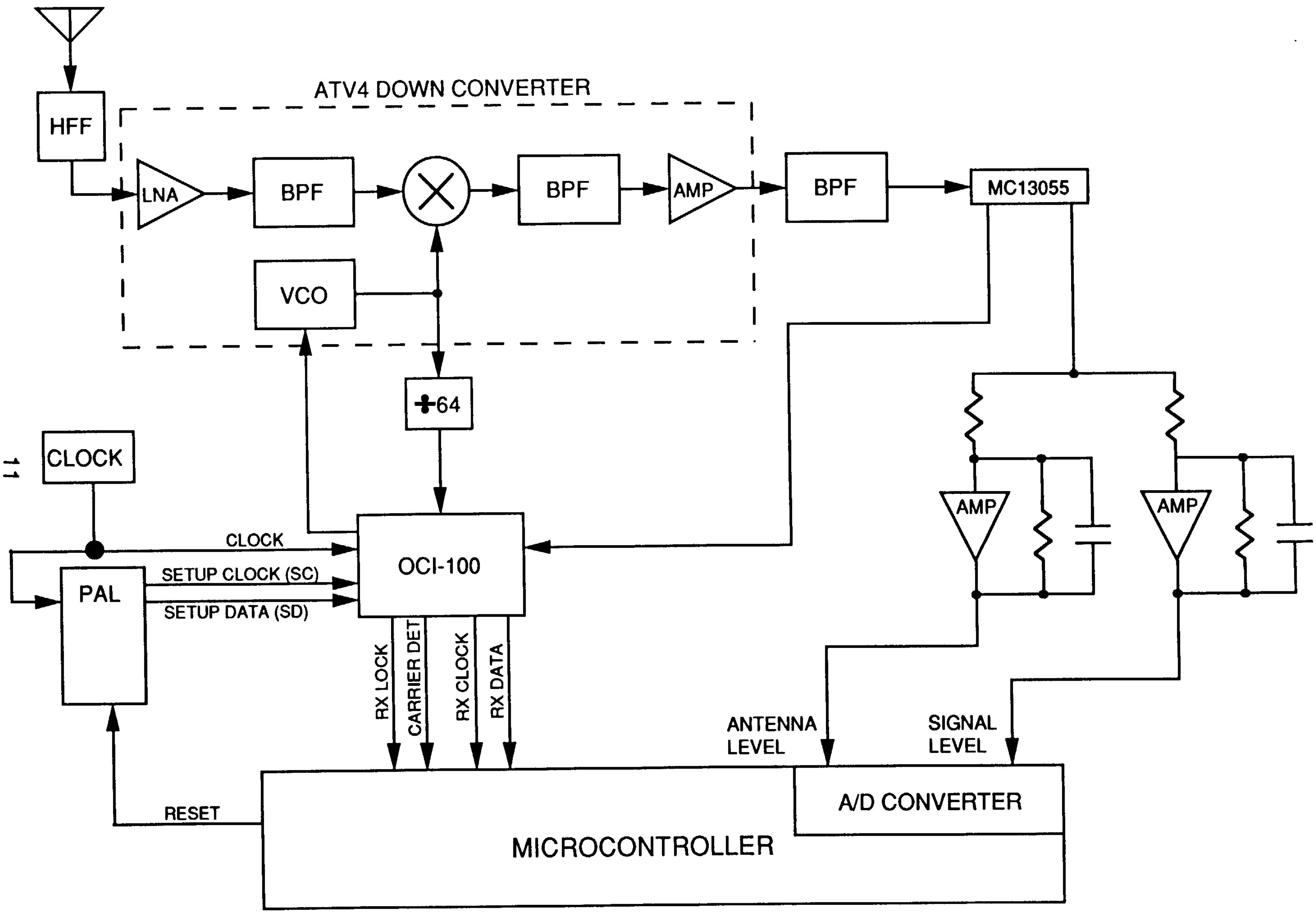


Figure 2.2-1 Receiver Block Diagram with Microprocessor Interface

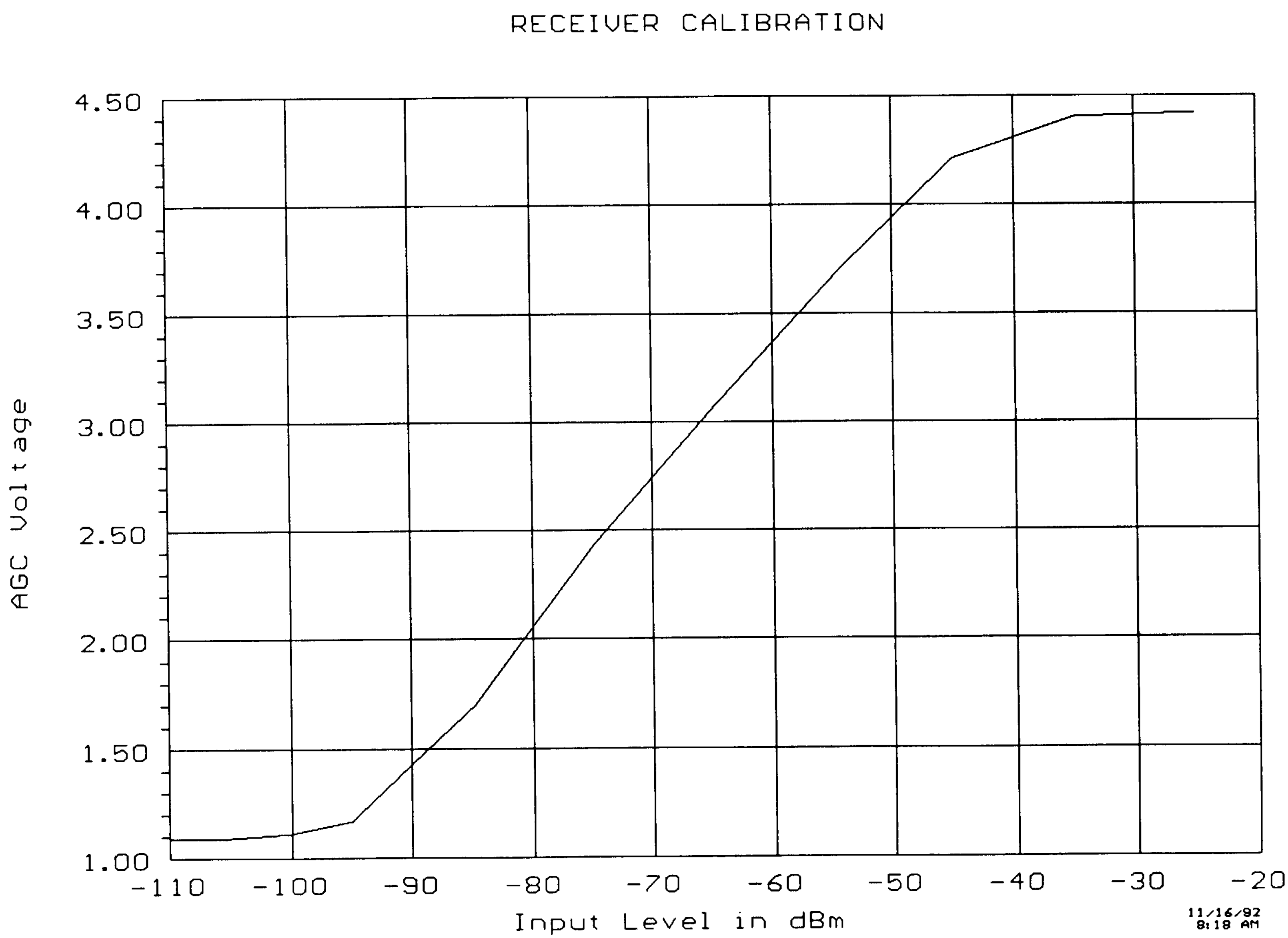


Figure 2.2-2 Receiver Antenna Level Calibration

The ATV4 down convertor has a voltage controlled oscillator with a potentiometer for adjusting the frequency. The ATV4 was modified in three ways to allow it to be phase locked. A capacitor was added to the oscillator to change its frequency range, a portion of the oscillator output is coupled out and the tuning potentiometer was disconnected and the control line brought out. Referring to Figure 2.1-2 the ATV4 VCO is divided by 64 (MC 12073) and presented to the OCI 100 (pin 20) where it is further divided and presented to a phase detector. The 4.192 MHz reference oscillator is presented to the OCI 100 where it is further divided and presented to the other input of the phase detector. The output of the phase detector is connected to the VCO to complete the phase locked loop. The receiving frequency is governed by the following expression:

$$FRX = PR * RXFREQ * FREF / RXREF$$

$$= 64 * RXFREQ * 4.9152 / RXFEF \text{ MHz}$$

The internal RXFREQ and RXFEF counters are set by the PAL or the microcontroller as described for the transmitter in Section 2.1. Figure 2.2-3 shows the ATV4 schematic and Figure 2.2-4 shows the remainder of the circuitry.

2.2.1 RECEIVER/MICROPROCESSOR INTERFACE

The receiver/ microprocessor interface consists of the following signals and was depicted in a previous figure (see Figure 2.2-1).

Receiver -> Microprocessor

LOCK - Logical signal giving the status of the VCO. If the VCO is locked the signal is a logical 1 else it is a logical 0.

CD - Carrier Detect. This is a logical signal that goes low when valid data is being received.

DATA - Logical data to be received by the receiver.

RXCLOCK - A clock signal that can be used to clock data into the microprocessor.

ANTENNA LEVEL - An analog signal representing input signal level of receiver. This signal is to be used to determine direction.

SIGNAL LEVEL - An analog signal representing the input signal level of receiver. This signal is heavily filtered. It is to be used to determine range.

Microprocessor -> Receiver

RESET - Resets the transmitter to initial conditions.

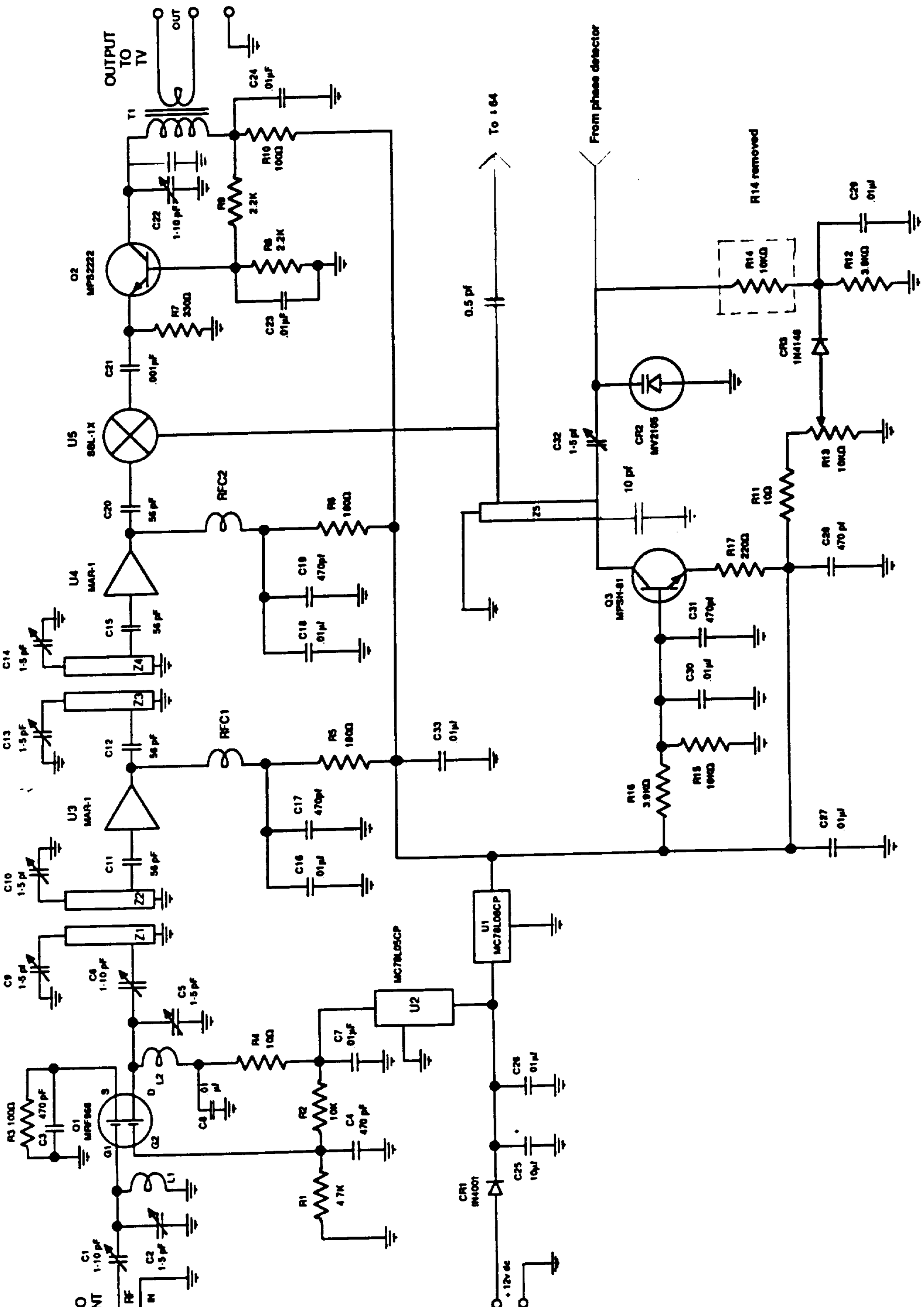


Figure 2.2-3 ATV4 Schematic

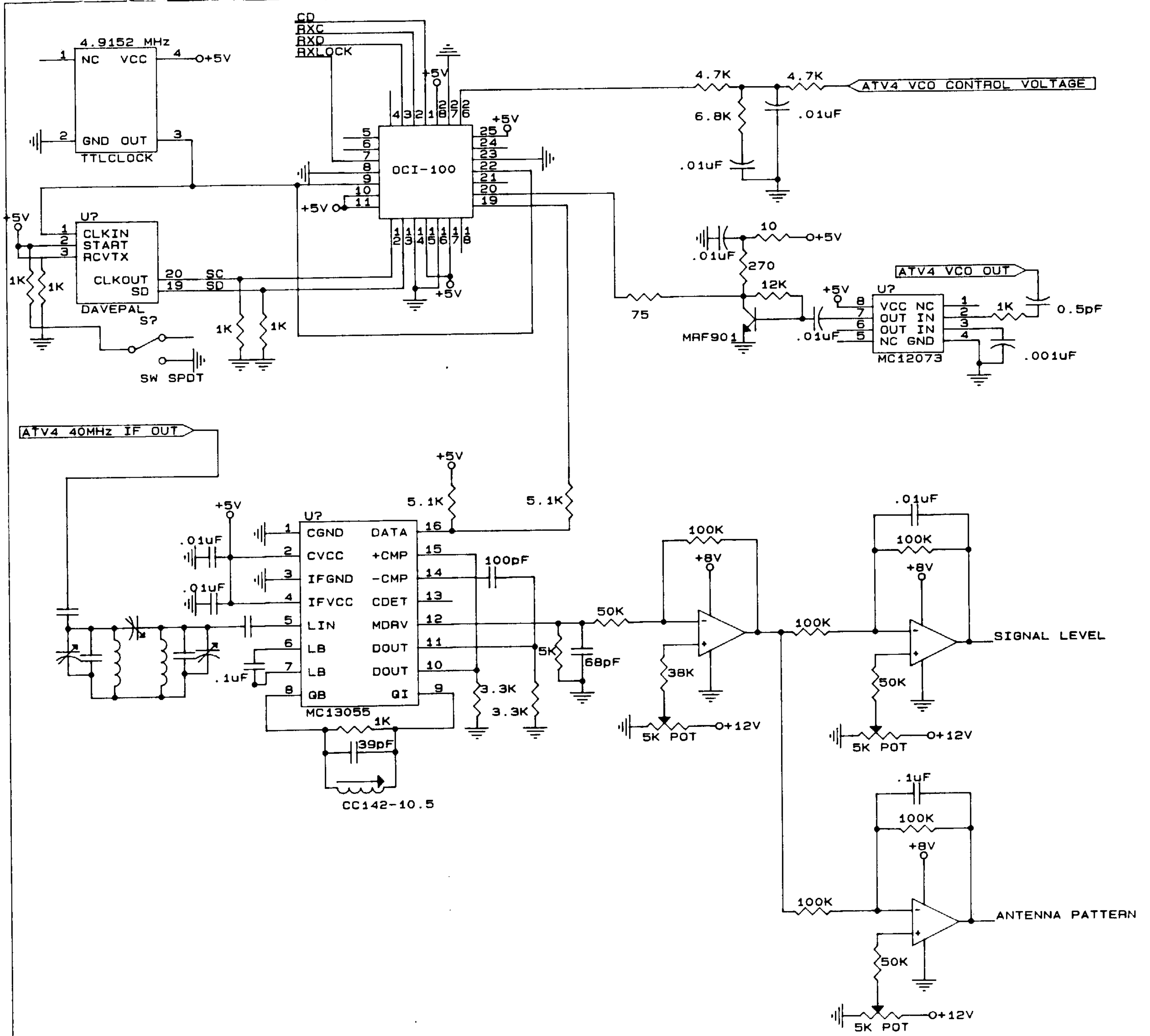


Figure 2.2-4 Receiver Schematic

3.0 ANTENNA SYSTEM

3.1 DESCRIPTION

The antennas for the EVAS system are the receiving antenna, which will be mounted on the user's vehicle, and the transmitting antenna assembly, which will be used on the emergency vehicle. The receiving antenna is a simple monopole, and requires little discussion. The transmitting antenna assembly, on the other hand, is fairly complex in both design and operation, and will be the subject of discussion for this part of the report.

The method of direction determination, using a three-element monopole array has been described in the Phase I report. It employs three vertical monopoles above a circular ground plane, as shown in Figure 3.1-1. By controlling the phases and amplitudes of the voltages delivered to these elements, radiation patterns of various shapes can be generated. If a set of such patterns is created, and the patterns are sequentially activated, then the amplitude of the received signal will abruptly change from one level to another as the transmitting-antenna patterns are sequentially stepped from pattern to pattern throughout the set. When the sequence is completed, the first pattern is again activated, and the sequence is continuously repeated. The sequence of increases and decreases in the received signal level is dependent on the direction of the receiver from the transmitter, expressed in the coordinate system of the emergency vehicle. Thus, the stepped sequence of signal levels provides information on the basis of which the receiving-vehicle direction can be determined. A transmitting antenna unit which uses this method has been designed, constructed and tested. It consists of three sections, each section containing a phase/amplitude controller, three amplifier stages and a monopole antenna element. The drive power is equally divided among the three sections by an impedance-transforming line section.

3.2 PHASED ARRAY

The radiating elements are quarter-wavelength monopoles mounted on a circular ground plane. The elements are spaced one half wavelength apart. They are coupled to the microstrip line through a metal block as shown in Figure 3.2-1. This method of coupling provides a 40-ohm coaxial transmission line section between the microstrip line and the antenna element and prevents excitation of a spurious parallel-plate transmission mode between the antenna ground plane and the stripline plate. The dimensions of the block are shown in Figure 3.2-2.

The size of the antenna ground plane is a compromise between a need to control the radiation pattern and a desire to minimize the physical size of the antenna for mounting purposes. A compromise diameter of 26 inches (2 wavelengths) was selected for the unit that was constructed. It is understood that this size can be reduced by making use of the mounting surface of the emergency vehicle as a ground plane, assuming that the surface is sufficiently planar and reasonably free of objects that would corrupt the radiation pattern.

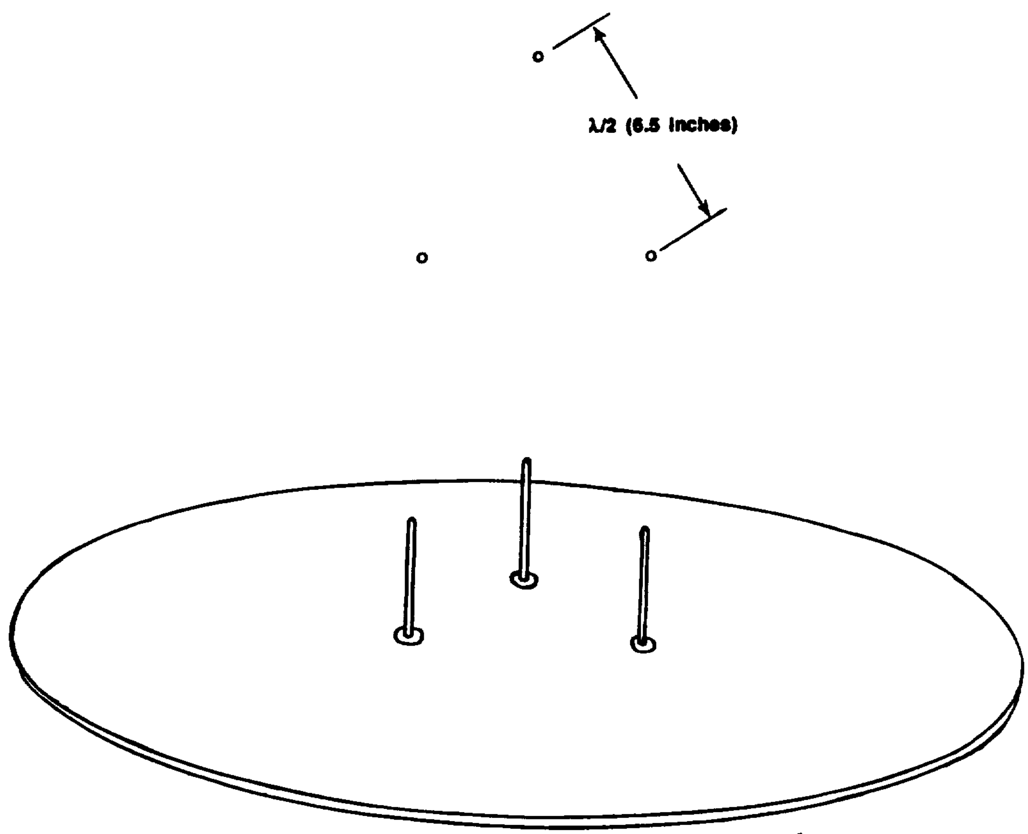


Figure 3.1-1 Monopole Array Antenna

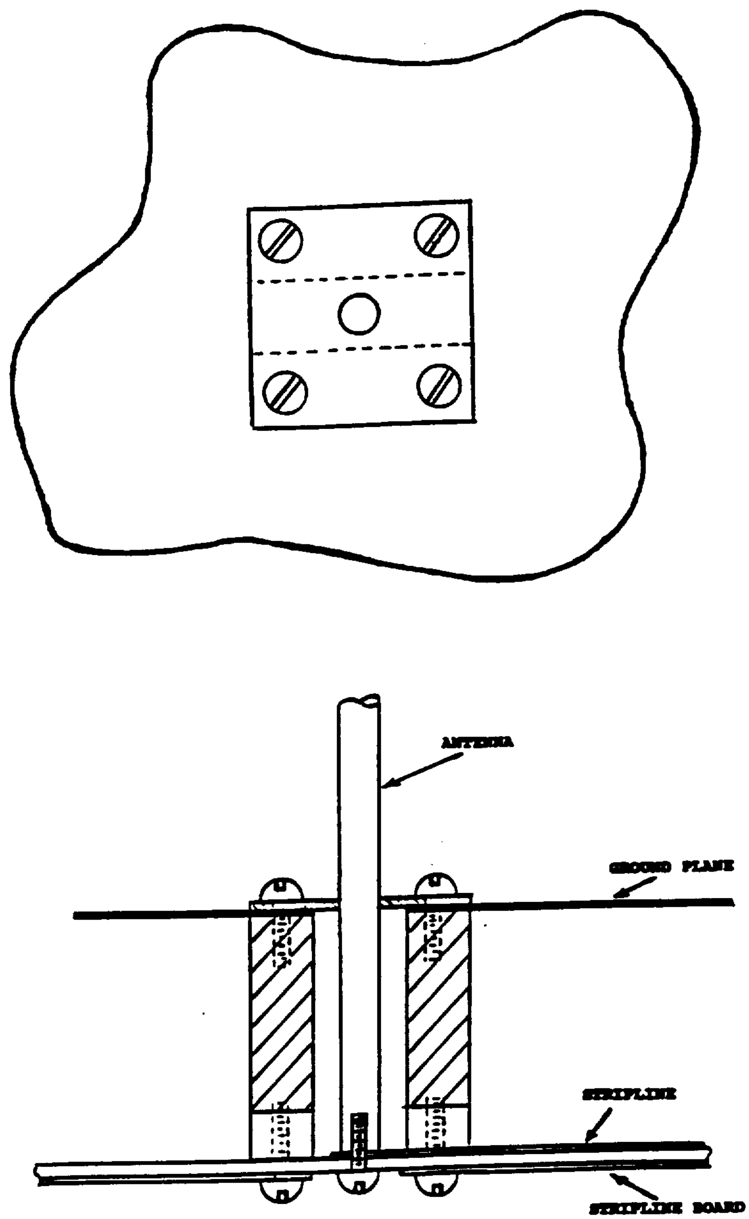
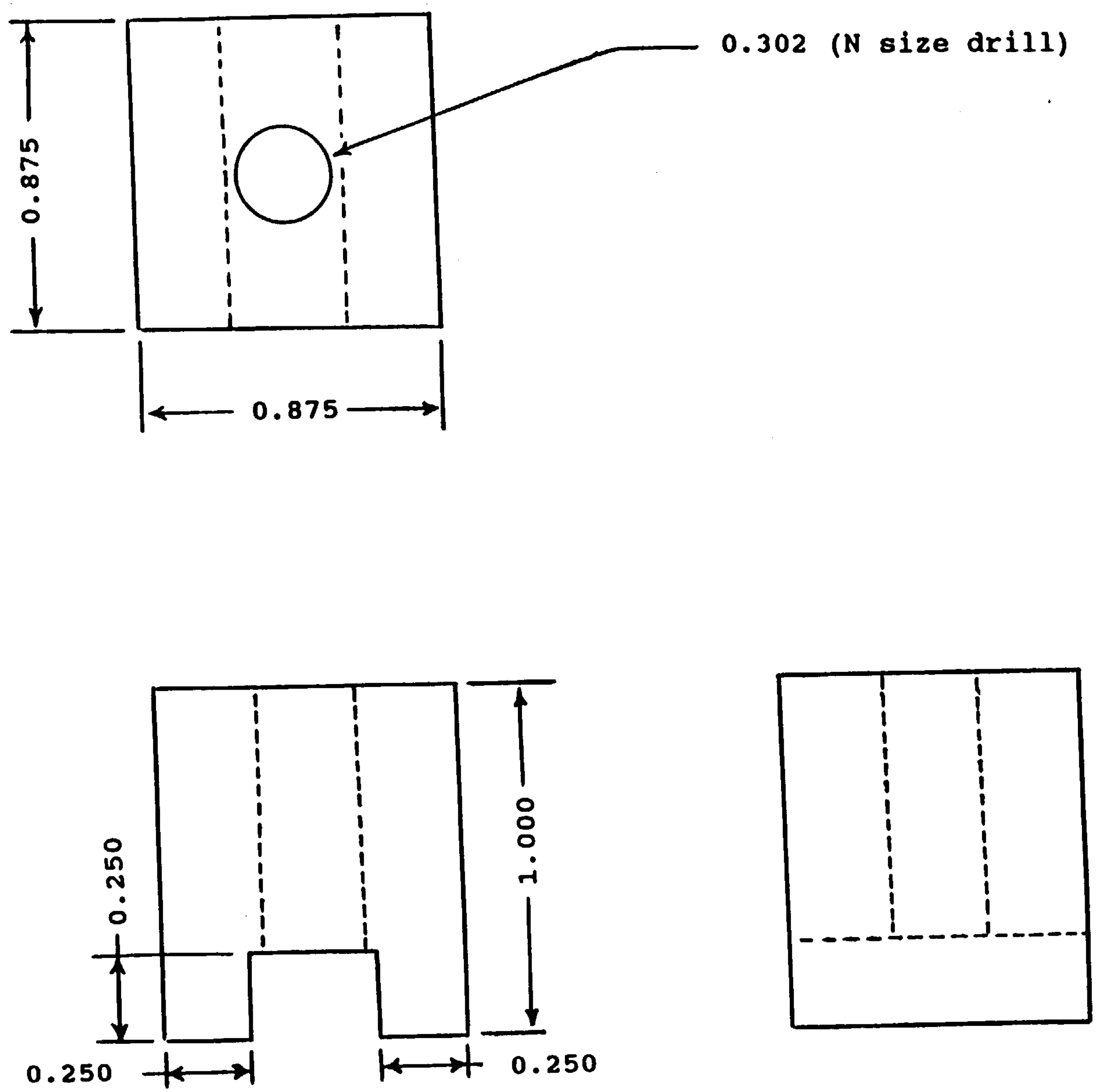


Figure 3.2-1 Method for Connecting Stripline Output Terminal to Antenna



Scale: 2:1

Material: Brass

Dimensions in Inches

Figure 3.2-2 Antenna Support Block

3.2.1 Amplitude and Phase Control

The method to be used for controlling the excitation of the antenna elements will now be discussed. The basic method is depicted in Figure 3.2.1-1. The RF input signal to the controller is applied to a 90-degree hybrid, dividing the power equally between two output paths. The voltages appearing at these outputs have a phase difference of 90 degrees. Thus, we may speak of having an in-phase component and a quadrature component (of equal magnitude) at these output terminals. If we independently control the amplitude of each of these components and add the controlled values, then both the amplitude and the phase of the resultant voltage can be set to any desired value. The method for controlling the amplitude of the two components, as seen in Figure 3.2.1-1, is to use two variable attenuators. When the amplitude of the quadrature component is reduced to zero, then the phase of the resultant voltage is zero, and its amplitude may be adjusted to any desired value. If, on the other hand, the in-phase component is set to zero, the resultant signal will have a phase of 90 degrees, and its amplitude may be adjusted to any desired value. It follows that any combination of these component levels can be used, and any amplitude and phase of the resultant signal can be produced.

The output signals from the two attenuators are input to a 180-degree hybrid ring. This hybrid again divides the input power equally between two output ports. In doing so, it isolates the two input sources from each other, preventing any interdependence effects. In this application of the 180-degree hybrid, one of its output ports is terminated in a 51-ohm resistor. The other port delivers its half of the power to a set of cascade amplifiers, where the power level is raised to that required to be radiated. It is then delivered to one of the antenna elements and radiated.

The variable attenuators used in the control unit are designed to be used as RF mixers. Essentially, they use a set of diodes to multiply one input voltage by another input voltage. In our application, one of these input voltages is the RF signal. The DC control voltage is applied to the other terminal. By use of this device, a means of rapidly setting and resetting the amplitude of the RF signal is afforded. A schematic diagram of the control system is shown in Figure 3.2.1-2. The unit used as an attenuator is the Mini-Circuits frequency mixer model ASK-2, shown in Figure 3.2.1-3.

Three sets of the circuitry described above are used, each set delivering power to one of the three antenna elements. The layout of the system on the microstrip board is shown in Figures 3.2.1-4 and 3.2.1-5.

It was decided in Phase 1 that six unique radiation patterns would be sufficient to determine direction to the nearest quadrant. The circuit that provides sets of control voltages for the variable attenuators is shown in Figure 3.2.1-6. Since each of the three antenna elements requires an in-phase and quadrature input and there are six patterns, 36 unique control voltages are required (six for each of the three in-phase and the three quadrature inputs). Q_1 provides the control voltage to the in-phase input of the controller for antenna element Ar. A voltage divider is made of the 390 ohm resistor at the base of Q_1 and one of six potentiometers attached to the open

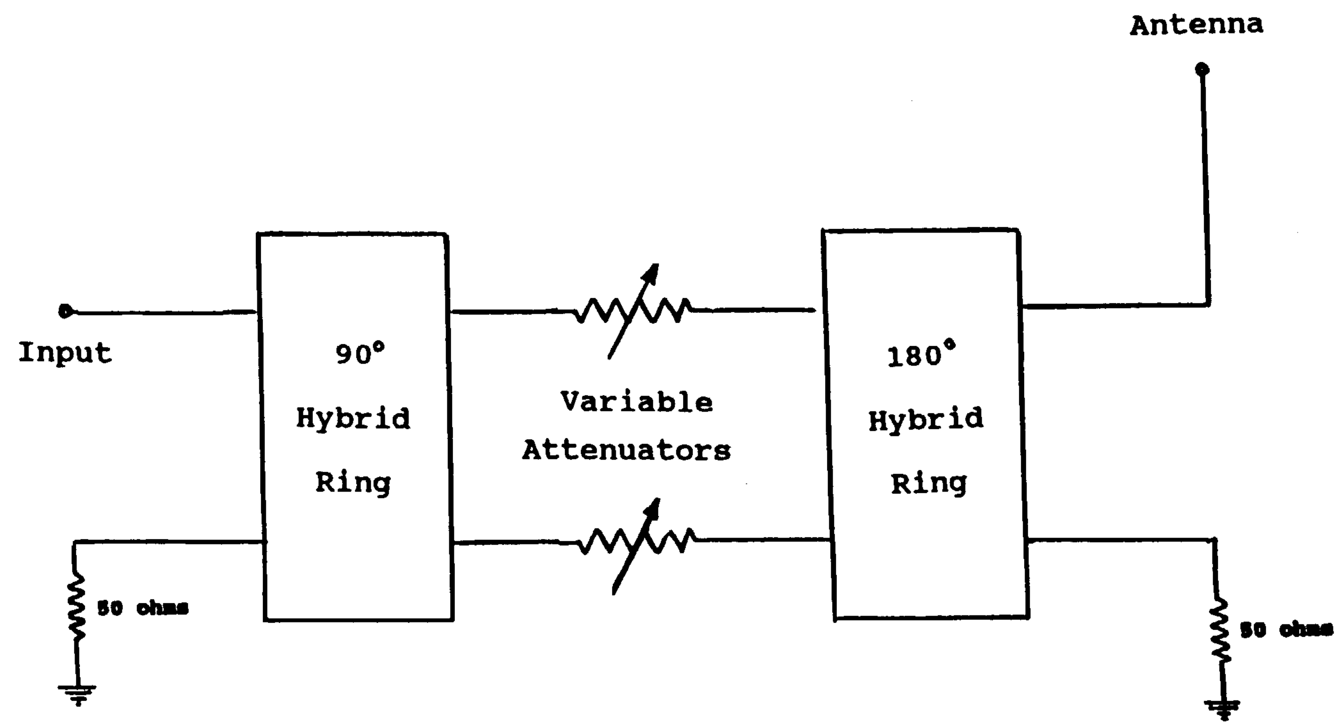


Figure 3.2.1-1 Method for Controlling Amplitude and Phase of Antenna Excitation

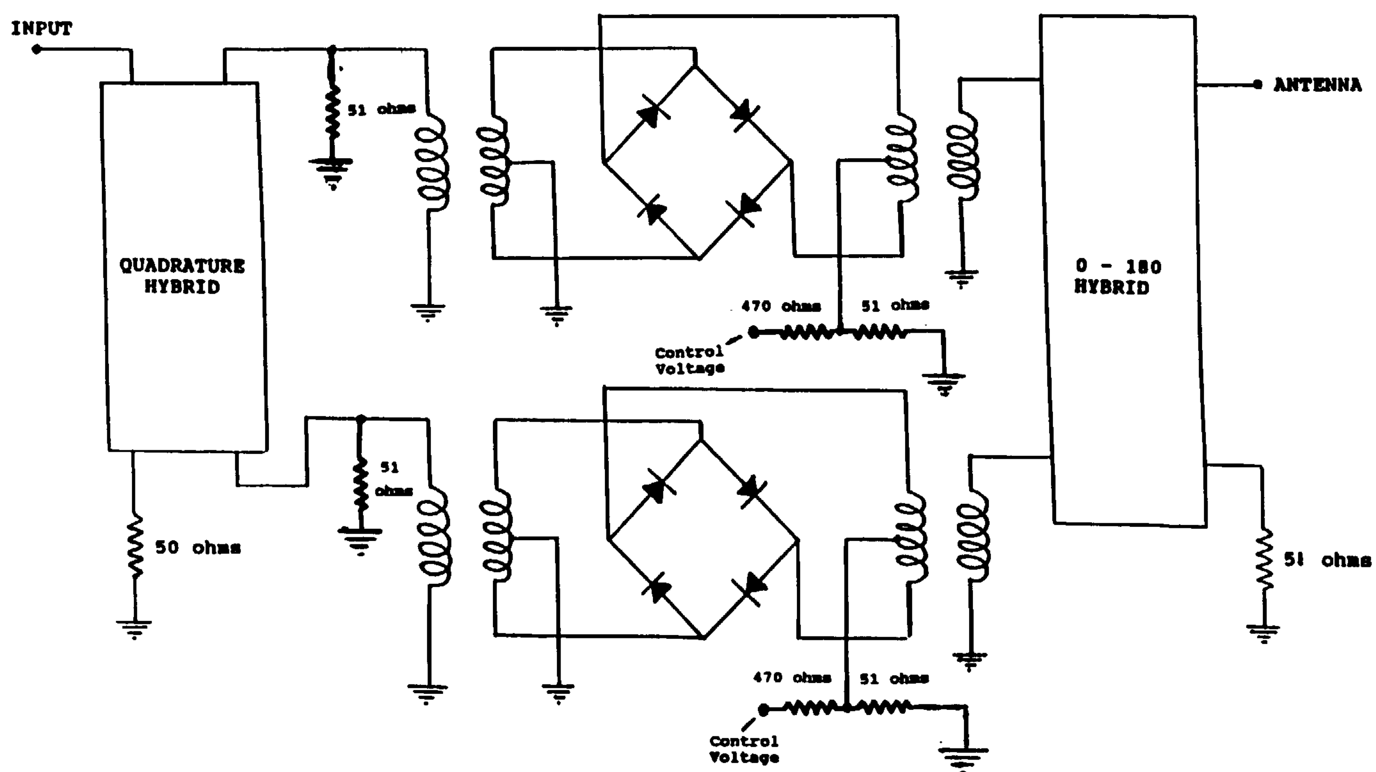


Figure 3.2.1-2 Amplitude and Phase Control System for EVAS Antenna System

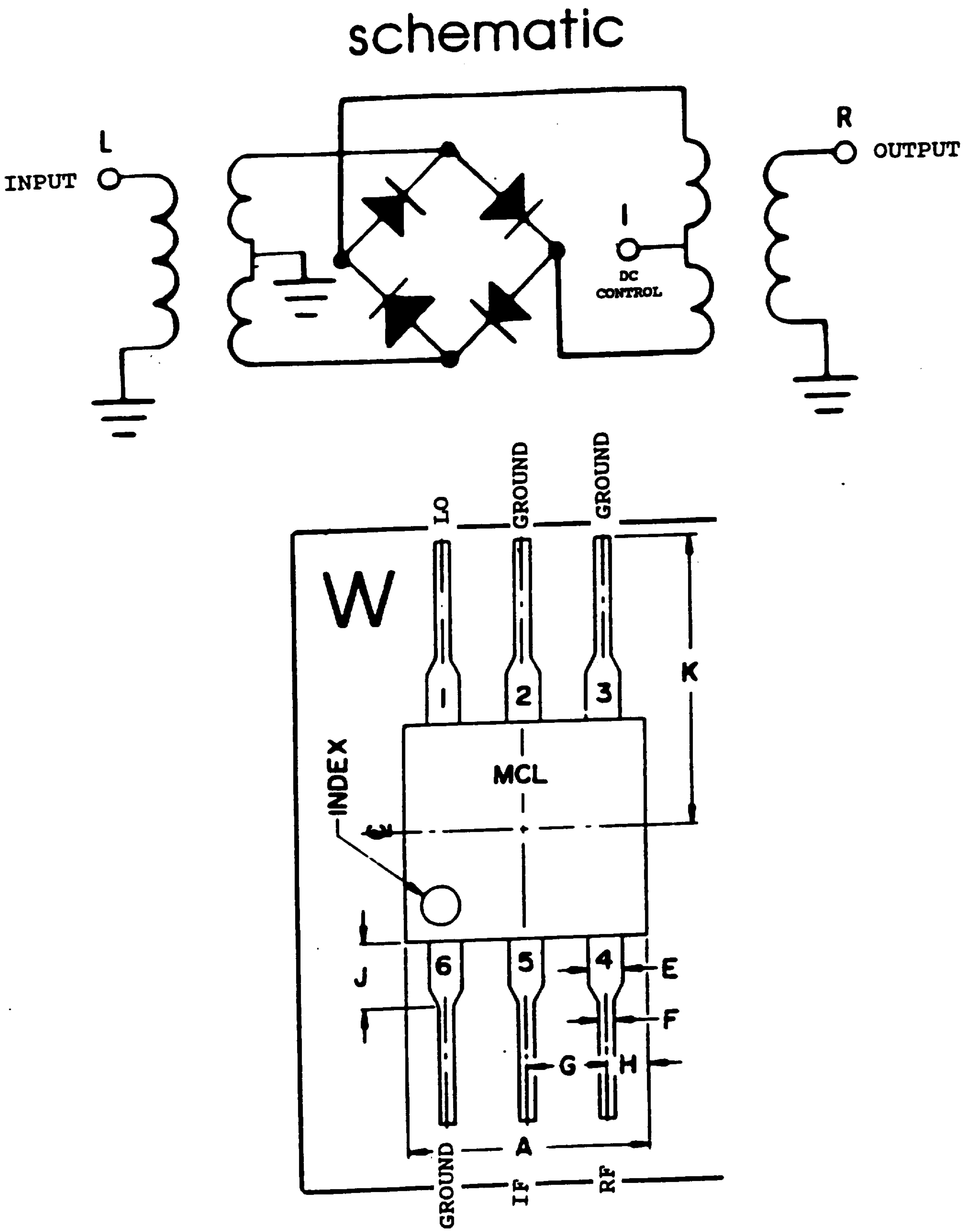


Figure 3.2.1-3 Mixer unit used as Attenuator in Amplitude/Phase Controller

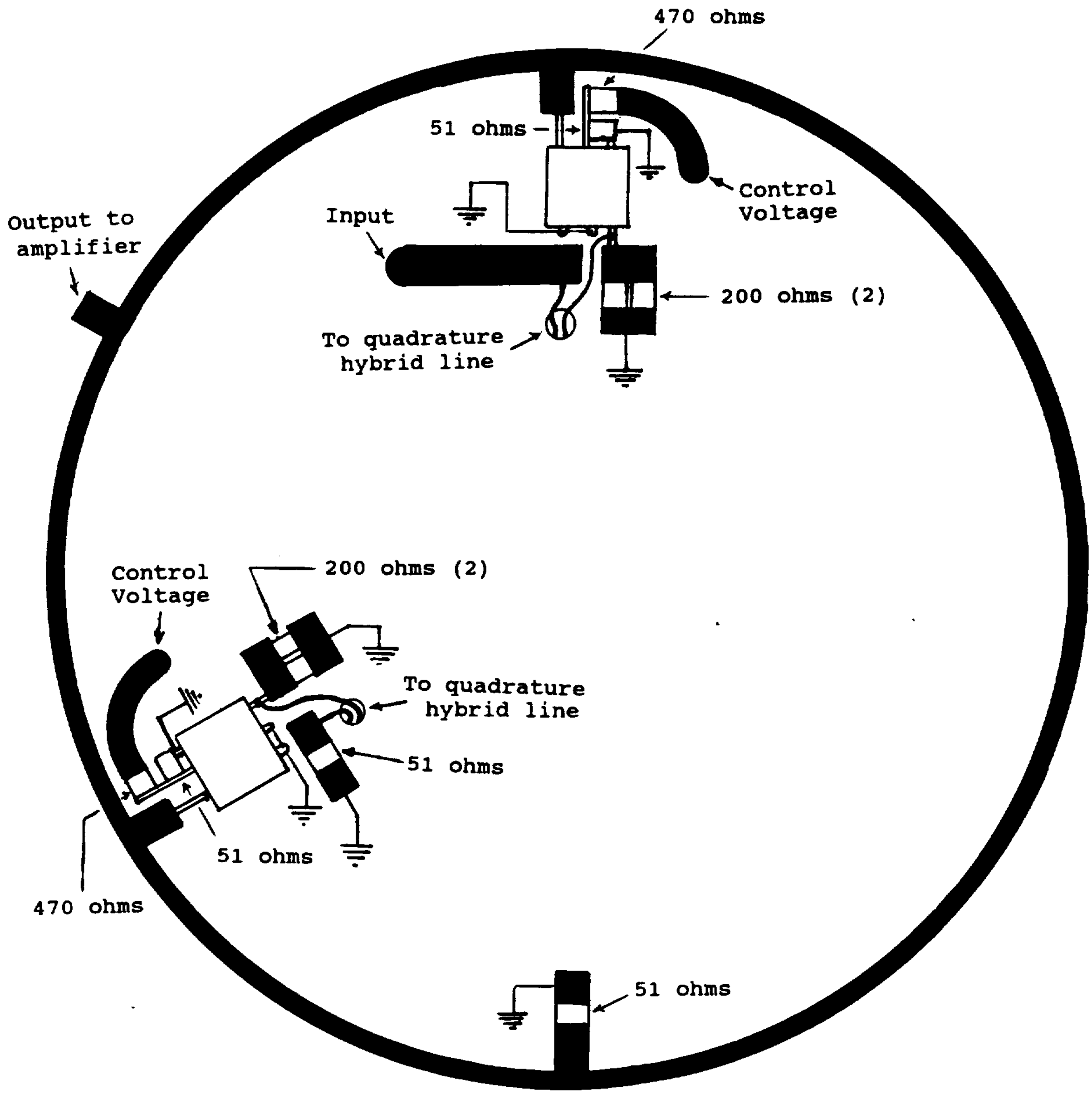


Figure 3.2.1-4 Amplitude/Phase Controller

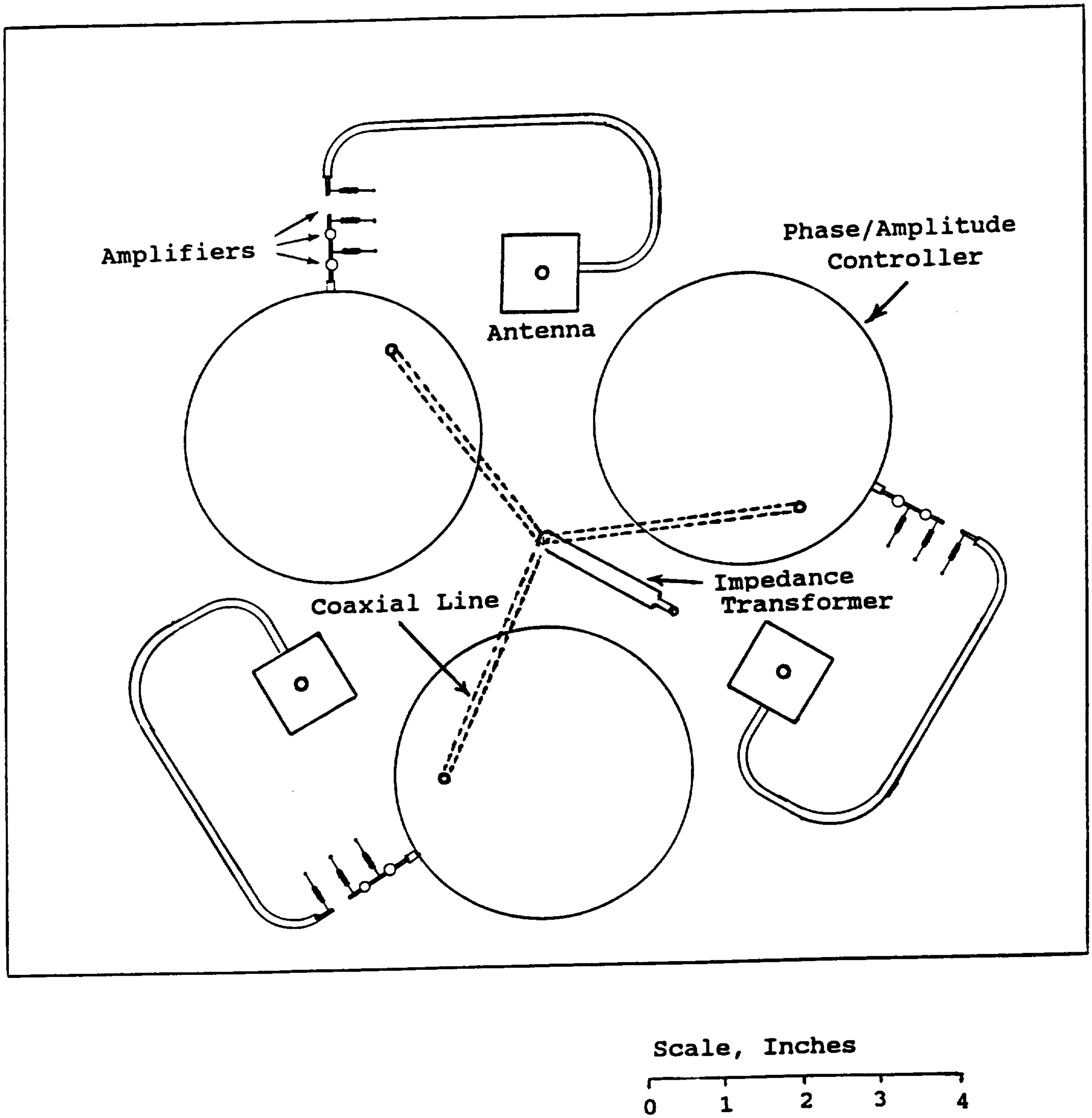
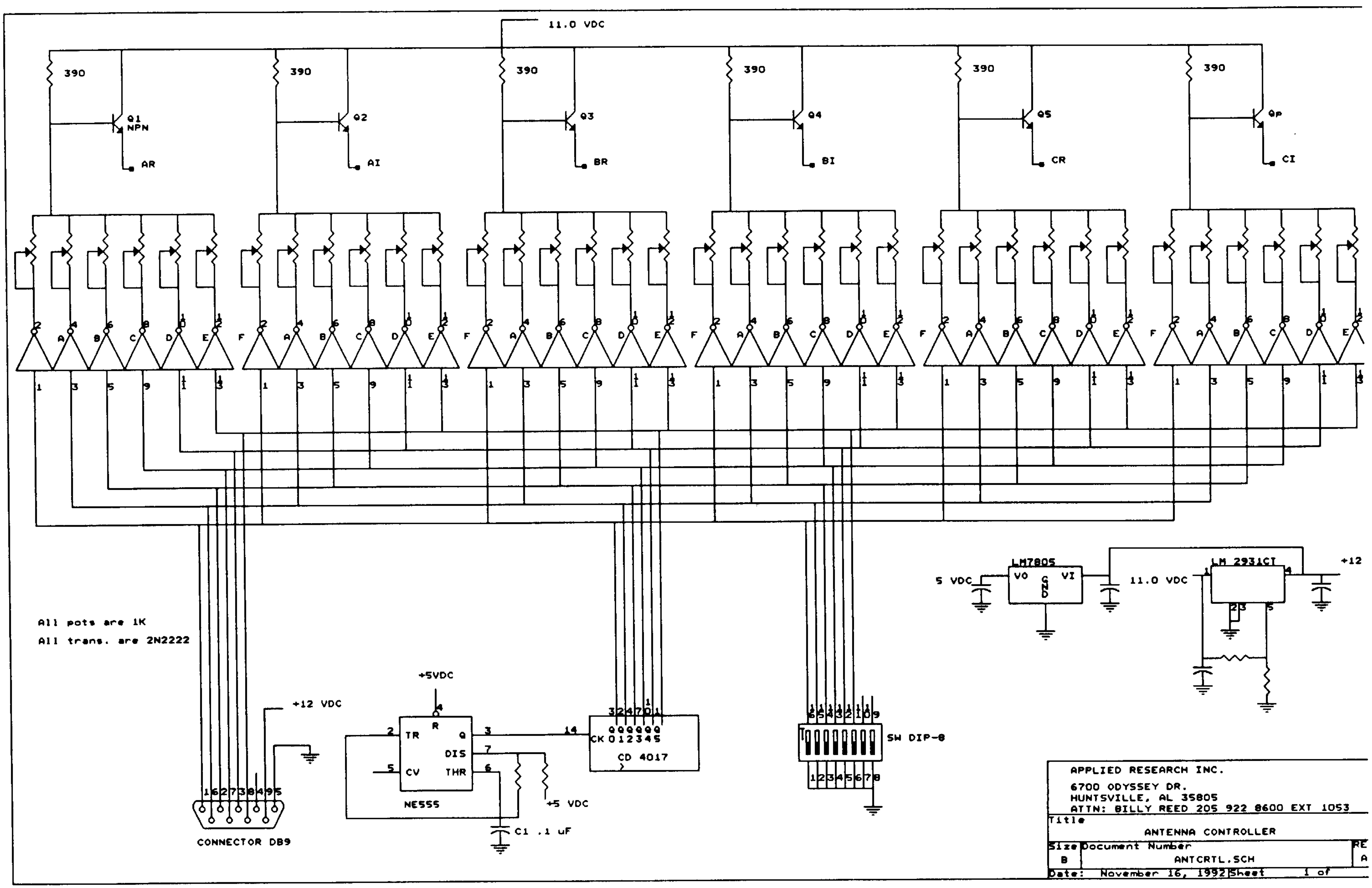


Figure 3.2.1-5 Layout for EVAS Transmitting Assembly

Figure 3.2.1-6 Antenna Controller



collector outputs of the SN 7405. Only one of the six potentiometers is switched to ground at a time as determined by the inputs on the DB9 connector. Notice that like inputs of each SN 7405 are connected. For pattern number 1, pin 1 of the DB9 connector would be logic 0 and pins 6, 2, 7, 3 and 8 would be logic 1. This causes the first open collector switch of each SN 7405 to be switched to ground with the other five open. For pattern 2, pin 1 would be a logic 0 and the others a logic 1. In this manner six different sets of in-phase and quadrature control voltages may be generated for each of the three antenna elements under control of the microprocessor. A DIP switch is also connected to the SN 7405 inputs. In normal operation these switches are open and have no effect. When adjusting the potentiometers, each of the six patterns can be activated by closing one switch at a time. For field testing without a microprocessor a CD 4107 decade counter with decoded outputs was added to provide inputs to the SN 7405's. The clock input is provided by the NE 555 connected as an astable multivibrator. The first six of the ten decoded outputs are connected to the SN 7405. When using this feature six patterns will be generated as determined by the potentiometers and one pattern for four counts with full voltage applied to the in-phase and quadrature controllers (all SN 7405 outputs open). This causes equal power to be delivered to each of the three elements during the four counts. When not operating in this mode the CD 4017 is removed so that it does not interfere with the microprocessor inputs on the DB 9 connector.

3.3 RADIATION PATTERNS

A previous EVAS report, ARI/92-R-004Z, showed a set of patterns for the three-element array. These patterns were given only for illustration of the basic principle of direction-determination, and did not include the effects of interelement coupling. This coupling affects the shape of the radiation pattern, and the measured patterns shown in this report include the effect.

Patterns for the array were measured by mounting the antenna assembly on a 4-foot by 4-foot ground plane and recording the level of the signal received at a distant point as the antenna is rotated in angular steps. Sets of control voltages for the six patterns were preset in the control unit prior to measurement of the patterns. Each of the six patterns was made after activating the control-voltage set associated with that pattern. The measurements were made at an outdoor location at the rear of ARI Building 6700. The layout of the measurement system is shown in Figure 3.3-1. Patterns were measured first with a directional receiving antenna and then with an omnidirectional monopole, such as will be used on a user vehicle. Detrimental effects that might be produced by the nearby building when using the omnidirectional antenna were found to be small enough to be disregarded.

The measured patterns for the six amplitude/phase settings are shown in Appendix B.

3.4 ANTENNA SYSTEM CALCULATIONS & TESTS

Calculations and tests performed in the course of development of the EVAS antenna system are included as appendices to this report. They are identified and described in the following paragraphs.

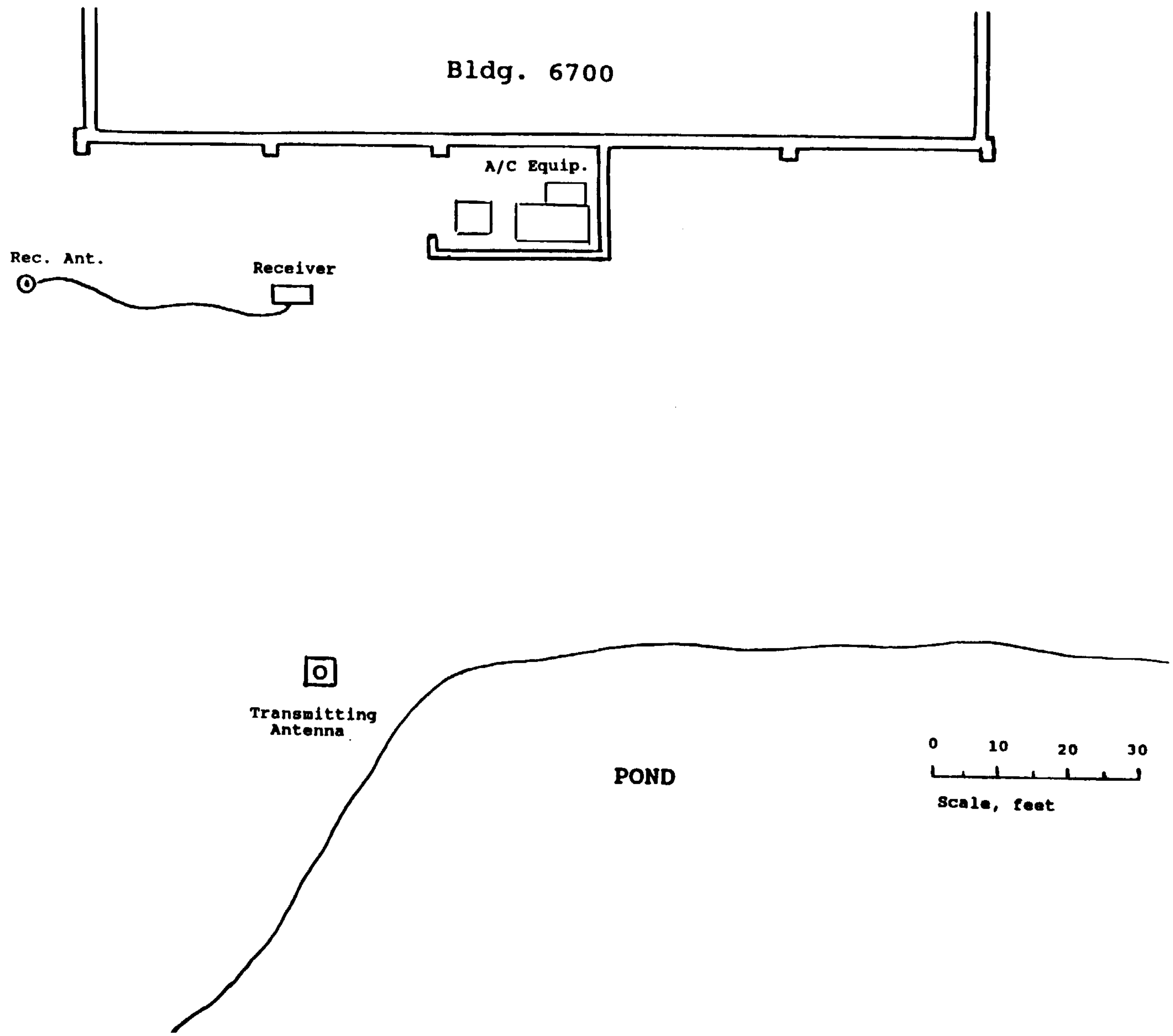


Figure 3.3-1 Antenna Pattern Measurement Layout

1. Selection of Microstrip Substrate.

Because of the desire to minimize production cost of the ultimate design, the least expensive substrate that would support the RF requirements was sought. The choice was an epoxy/fiberglass type, NEMA grade FR-4, ED130. The epoxy binder is a rather poor dielectric material at radio frequencies, but for this application, the losses are sufficiently low at 912 MHz and for the short line lengths to satisfy the requirements. The loss for a 50-ohm microstrip line is slightly less than 1 dB per foot. The cost difference between this material and the higher-quality substrates is very great: the teflon-based types cost more than 10 times as much as the type that was used. The small degradation in performance is considered acceptable in view of the cost saving.

2. Computation of Monopole Antenna Parameters.

The input impedance of one of the monopole antenna elements was calculated, taking into account the presence of the other two elements and their terminating impedances. The mutual impedance of the elements was also calculated. The computations were performed by use of a program based on the method of moments.

3. Calculation and Testing of Control Circuit Parameters.

Design of the controller circuitry required calculation of the microstrip line parameters such as characteristic impedance, wavelength on the line and power loss. The transmission-line parameters for microstrip depend on the dielectric constant of the substrate, its loss tangent at the frequency being used and on the ratio of the line width to the substrate thickness (w/h). The values of these parameters were calculated for the FR-4 material, and the circuit design was based on these calculated values. Tests were performed to validate the calculations. Records of these calculations and tests are included in Appendix C.

4. Amplitude/phase Controller Tests.

The amplitude/phase control circuitry was tested at two levels,

(a) as an individual unit, serving one antenna element.

(b) as a complete assembly, driving all three antenna elements.

Test results for an individual unit are provided in Appendix D. Those for the entire assembly are shown in Appendix E.

A description of the power divider that is used to distribute the drive power to the three antenna units is shown in Appendix F.

4.0 SYSTEM TEST DATA

Three types of tests data were recorded for analysis of the receiver system. The first was reference waveform data, the second was static road data, and the third was dynamic road data. The tests were performed using the transmitter, receiver, and directional antenna described in Sections 2.0 and 3.0 of this report. The receiving antenna was an omni antenna on a two foot square ground plane. The antenna waveform output of the receiver was recorded using a FLUKE 97 Scope Meter. The Scope Meter is a portable 50 MHz Digital Storage Oscilloscope capable of storing waveforms and transferring them to a computer through a RS-232 port. Software was developed to receive the waveforms from the Scope Meter and process the data. Figure 4.0-1 shows a block diagram of the hardware test setup.

4.1 REFERENCE WAVEFORM DATA

Reference waveform data was recorded in order to build a set of reference waveforms to be used in determining direction. A layout of the reference test area was given in the previous section (see Figure 3.3-1). The transmitting and receiving antennas were separated by approximately 75 feet. The receiving antenna was separated from the receiver by approximately 40 feet to minimize interference from the person operating the receiver. While this was not the ideal test range, the data collected did closely resemble the waveform projected from the measured antenna patterns.

To record the reference waveform data, both the transmitter and the receiver were held in stationary positions. The transmitter was pointed a known angle from the receiver and allowed to step through the antenna patterns. The receiver recorded these steps for several cycles and stored the recorded data to disk. Reference data was recorded with the transmitting antenna positioned at 0 degrees to 360 degrees in increments of 10 degrees. Figure 4.1-1a to 4.1-1f shows plots of these recorded waveforms.

4.2 STATIC ROAD DATA

Static road data was recorded during field test of the transmitter and receiver systems. Figure 4.2-1 shows a layout of the test area. The light poles along the road were used as markers for test distances.

During the static road test, the transmitter was placed on the top of one vehicle and the receiver was placed on the top of another. The transmitter was pointed at a known angle from the receiver and allowed to step through the antenna patterns. The vehicle with the receiver was then driven a given distance from the transmitter and several cycles of received waveform were recorded and stored to disk. Several tests were made with the transmitting antenna set to different angles and the receiver at different distances. Figures 4.2-2a to 4.2-2h shows plots of the received waveforms recorded during the static road test. The data collected during this test showed that the

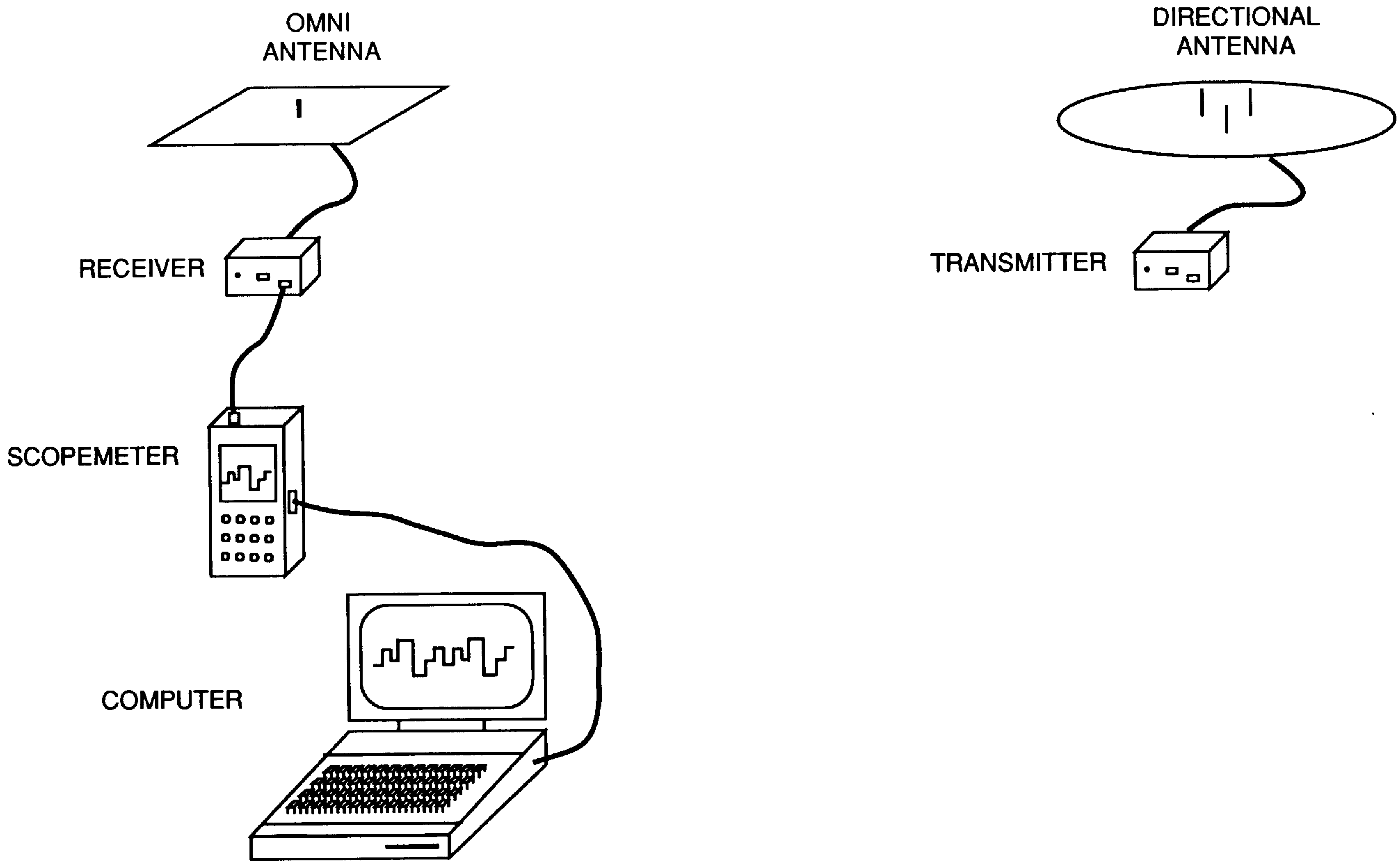


Figure 4.0-1 Test Setup Hardware

118-1043-1192-dw

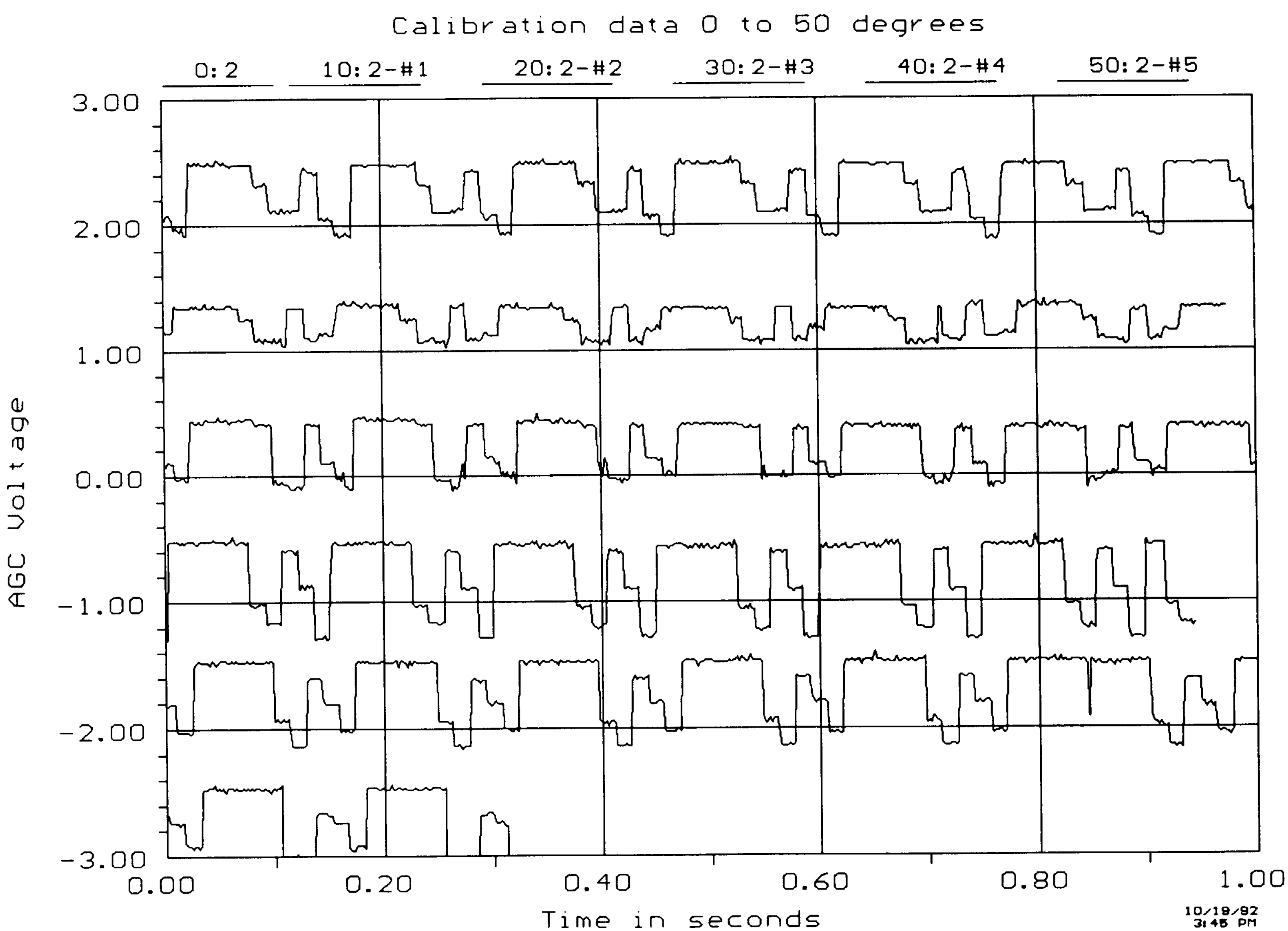


Figure 4.1-1a Reference Curve Plots

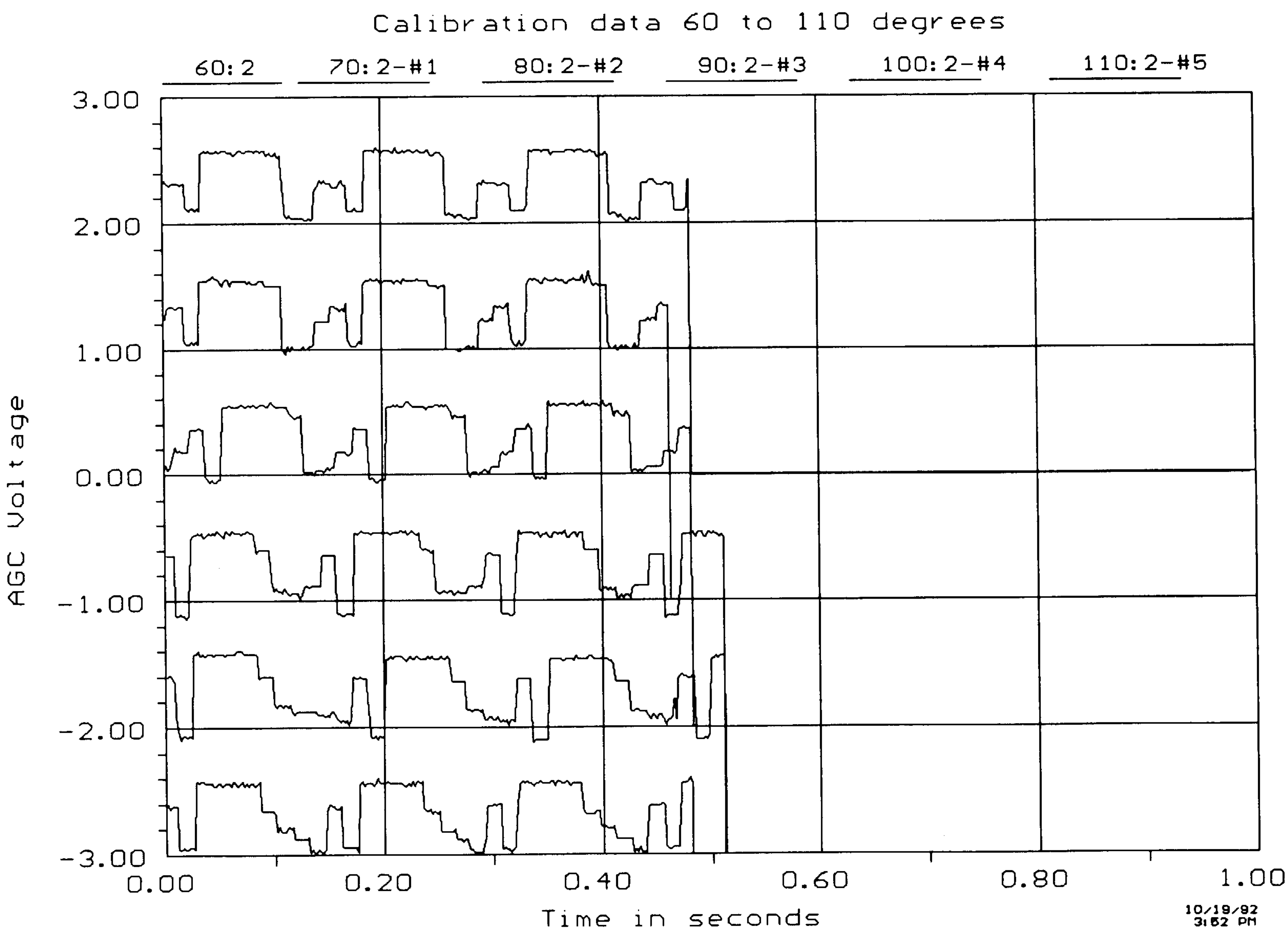


Figure 4.1-1b Reference Curve Plots

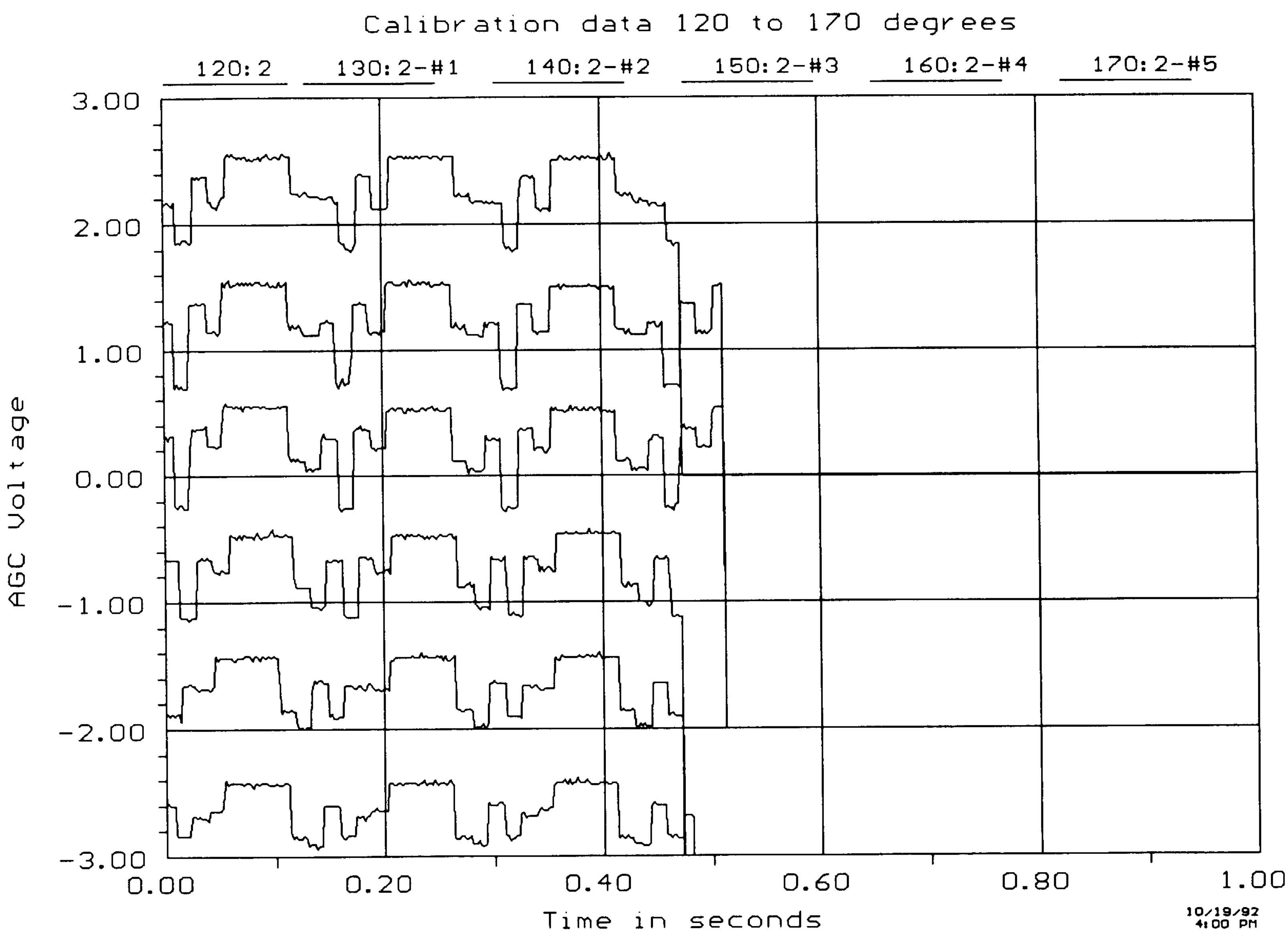


Figure 4.1-1c Reference Curve Plots

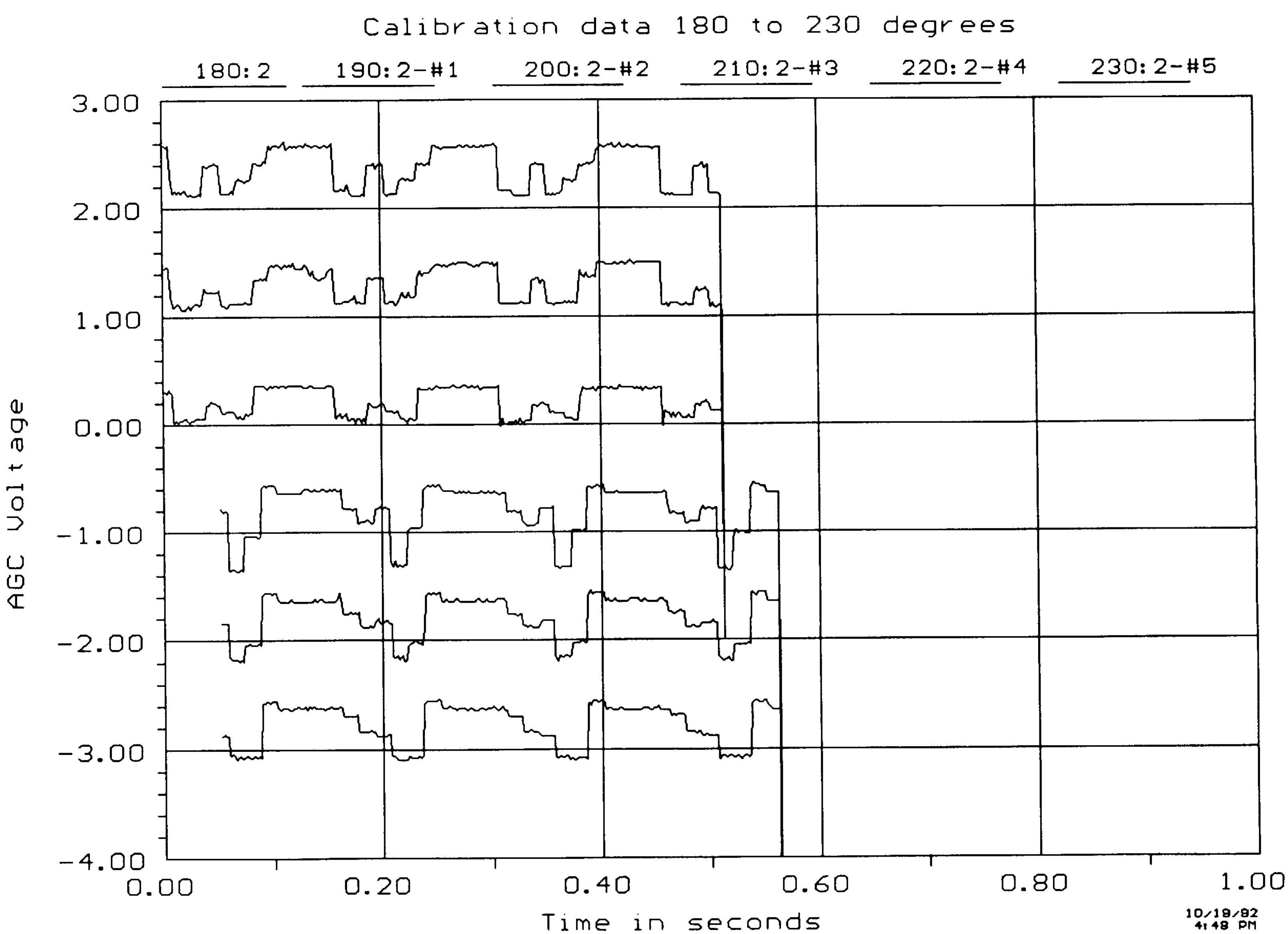


Figure 4.1-1d Reference Curve Plots

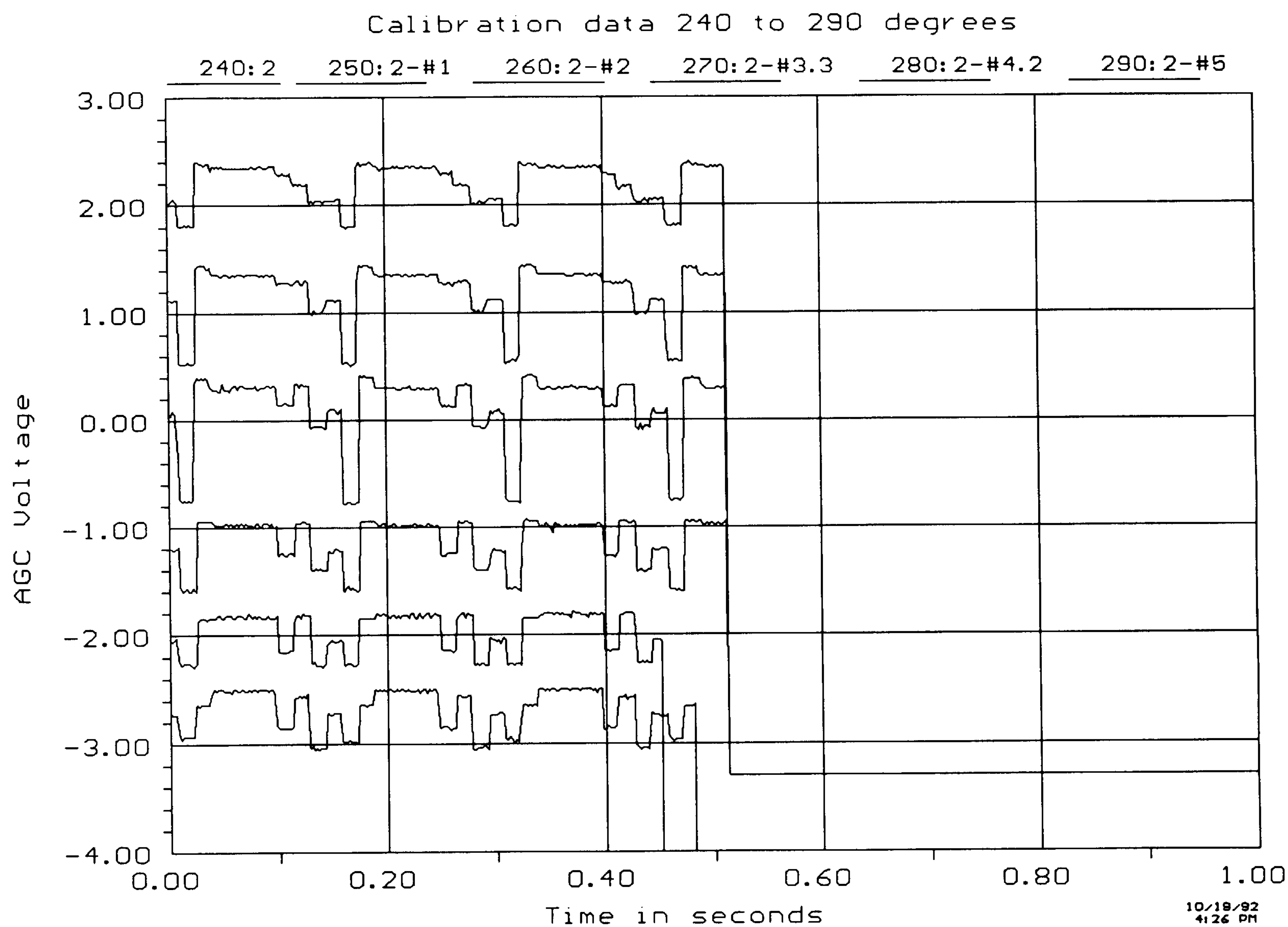


Figure 4.1-1e Reference Curve Plots

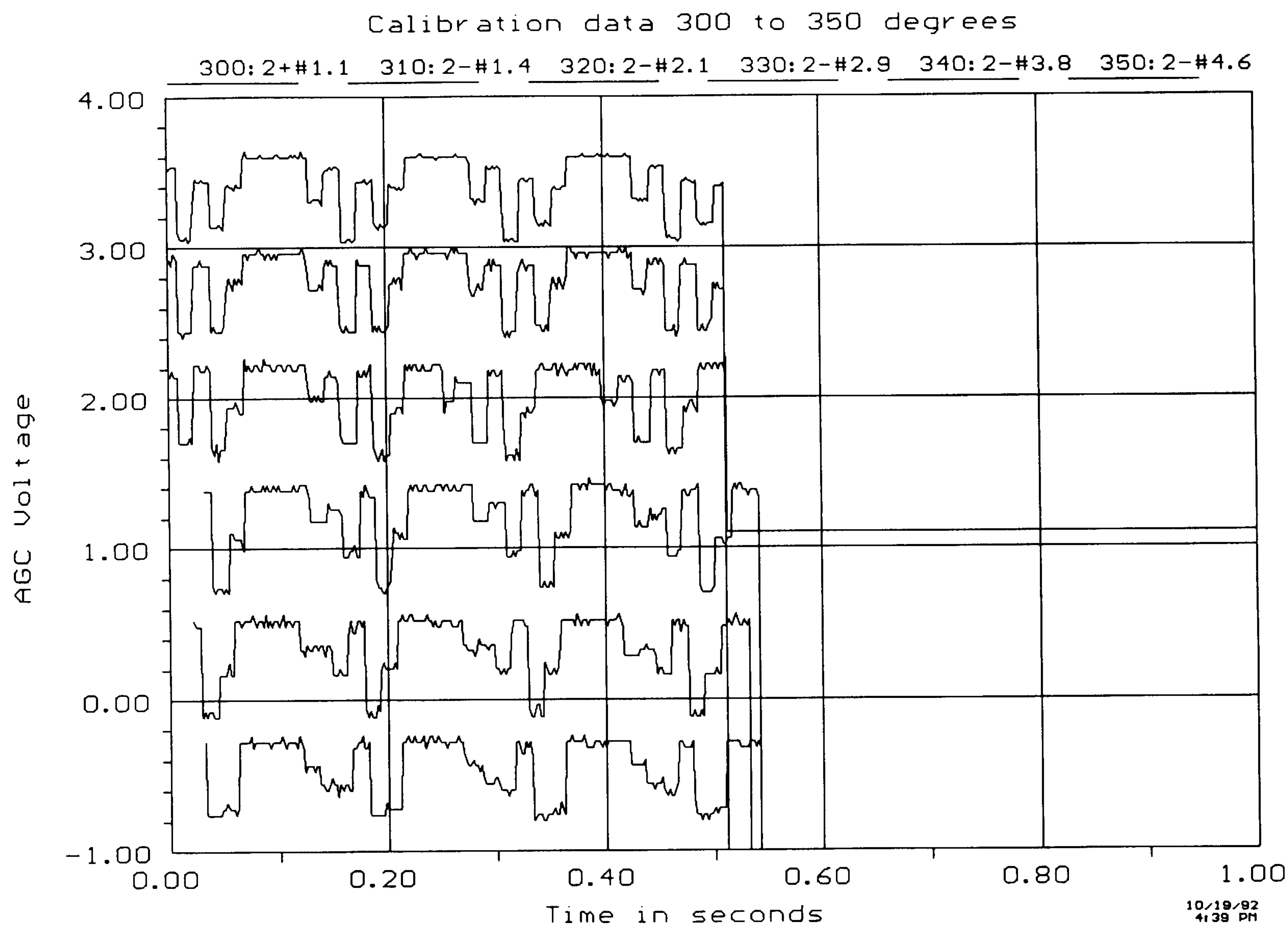


Figure 4.1-1f Reference Curve Plots

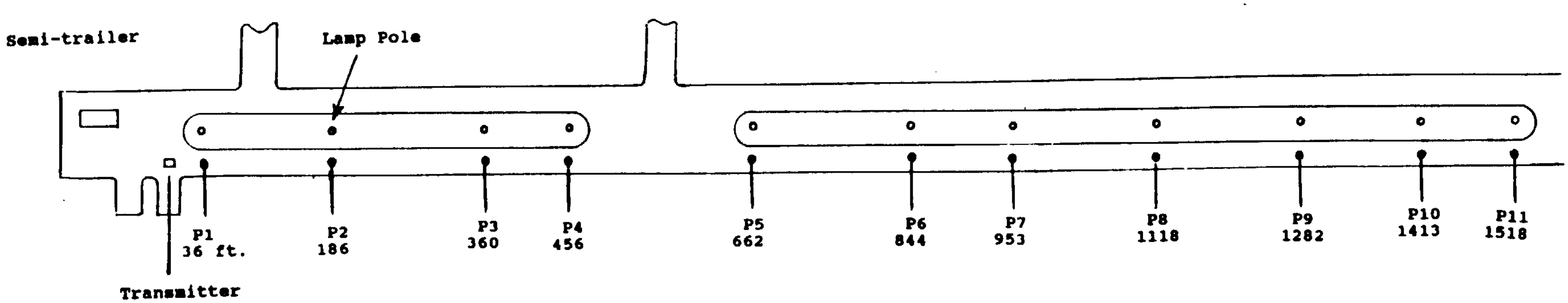


Figure 4.2-1 EVAS Field Test Layout

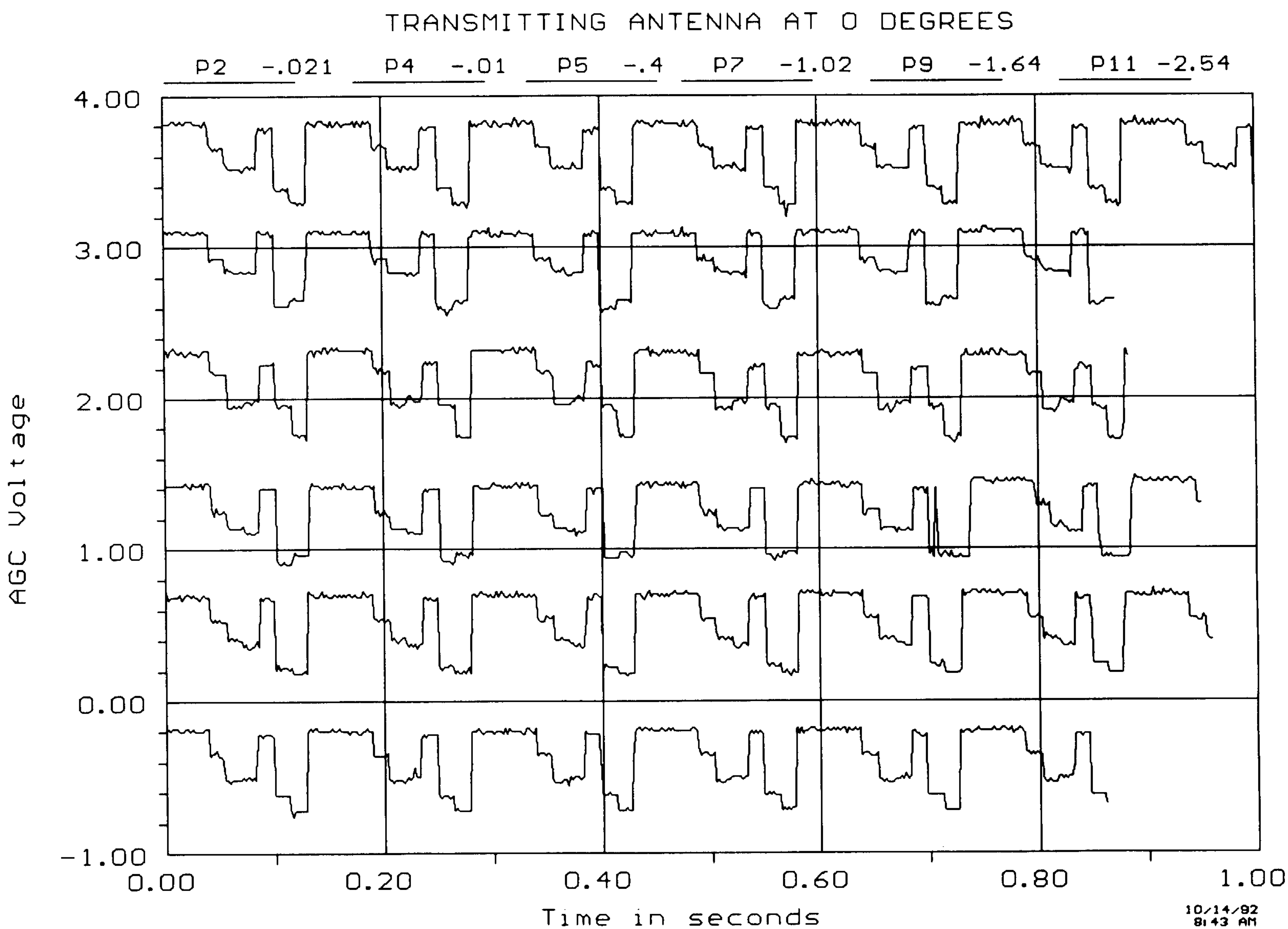


Figure 4.2-2a Static Test Waveforms

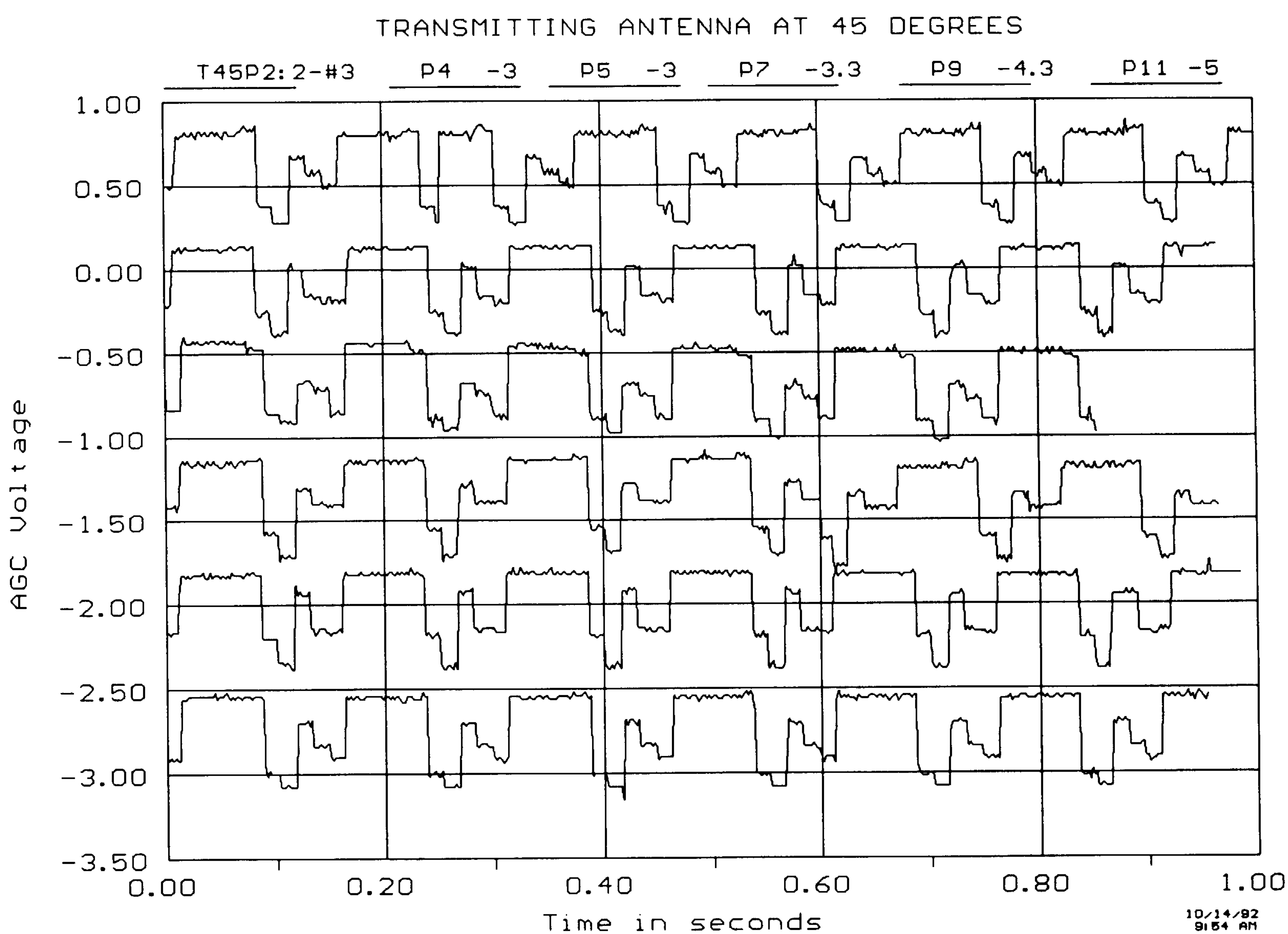


Figure 4.2-2b Static Test Waveforms

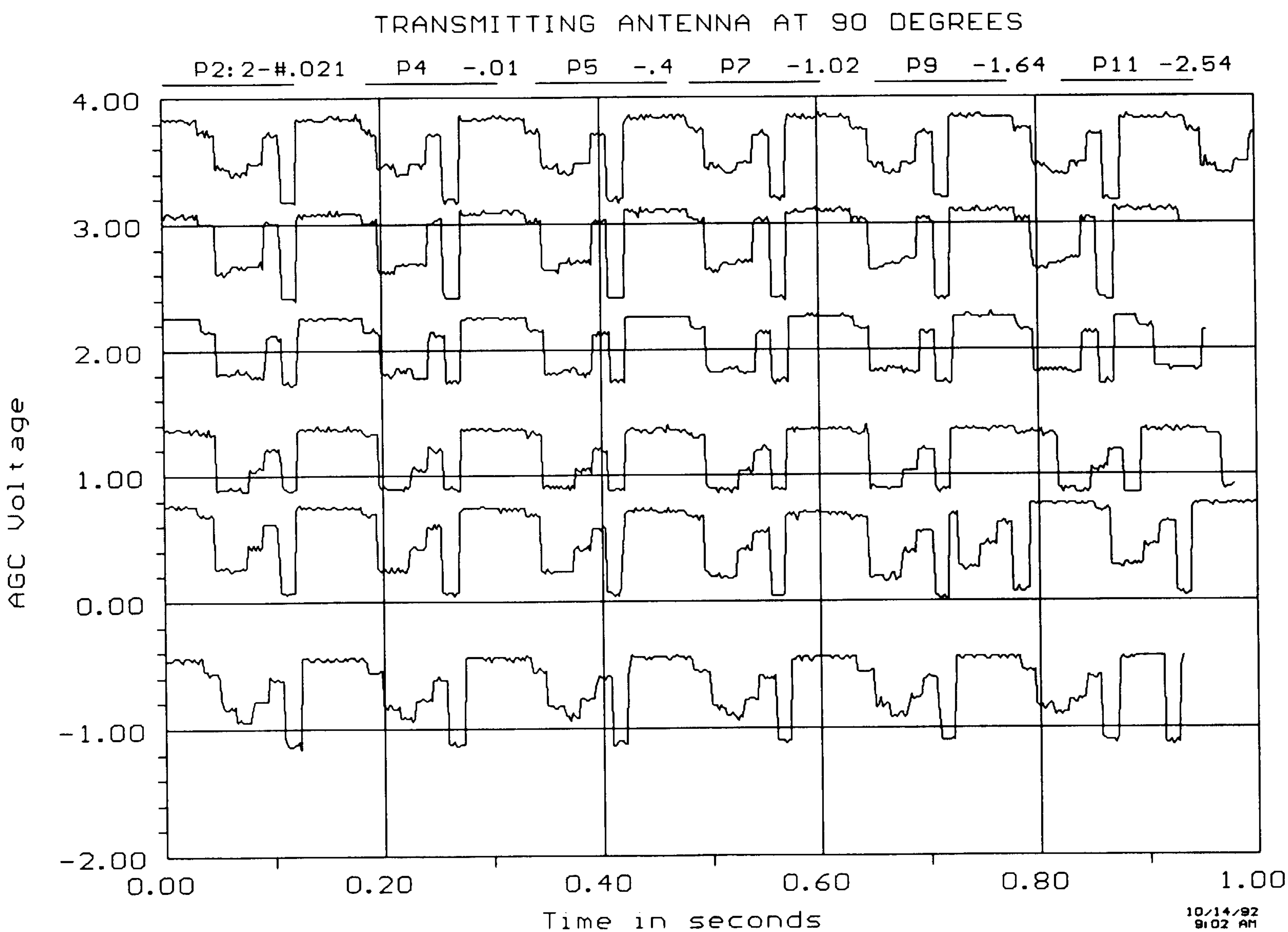


Figure 4.2-2c Static Test Waveforms

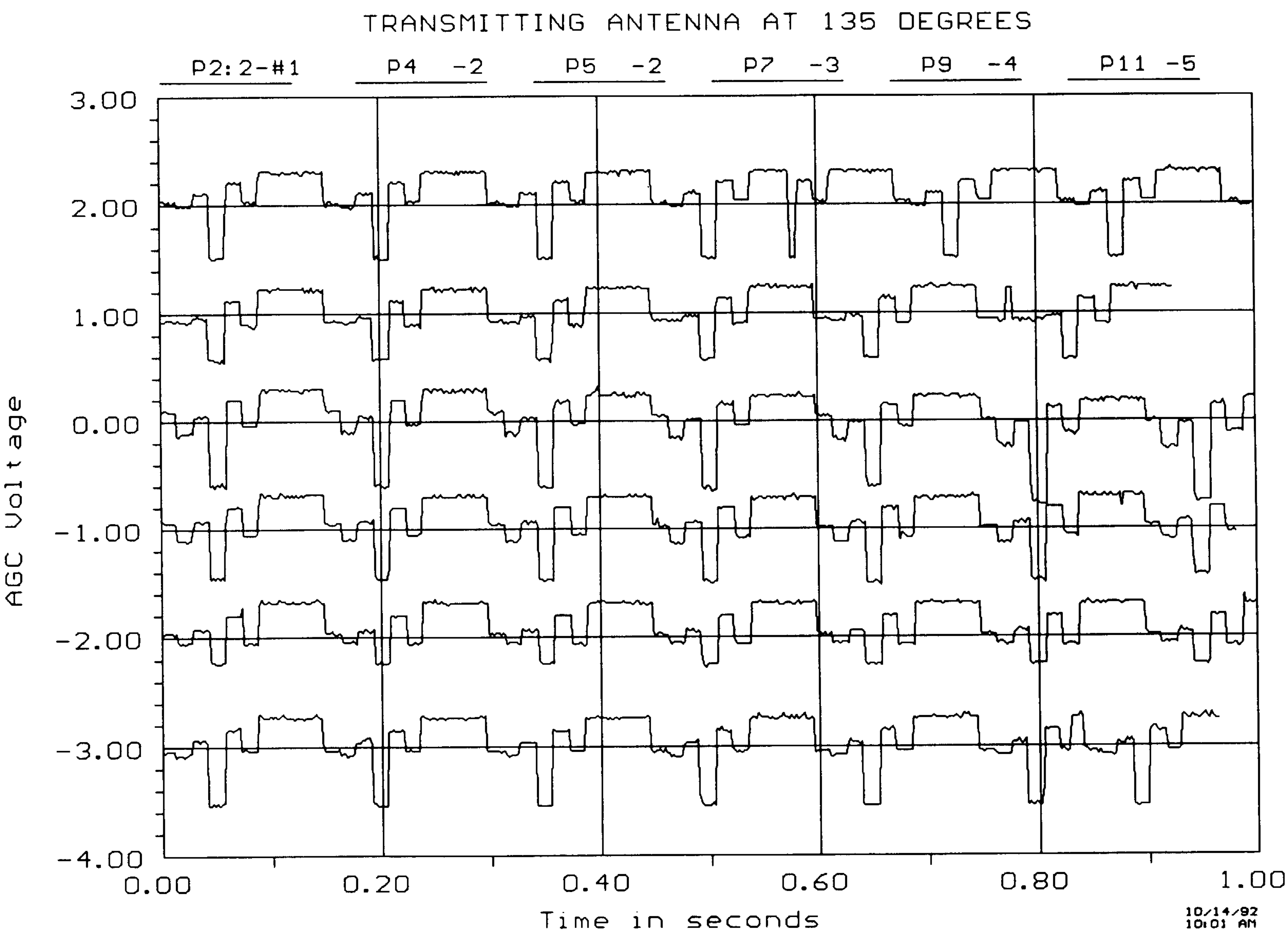


Figure 4.2-2d Static Test Waveforms

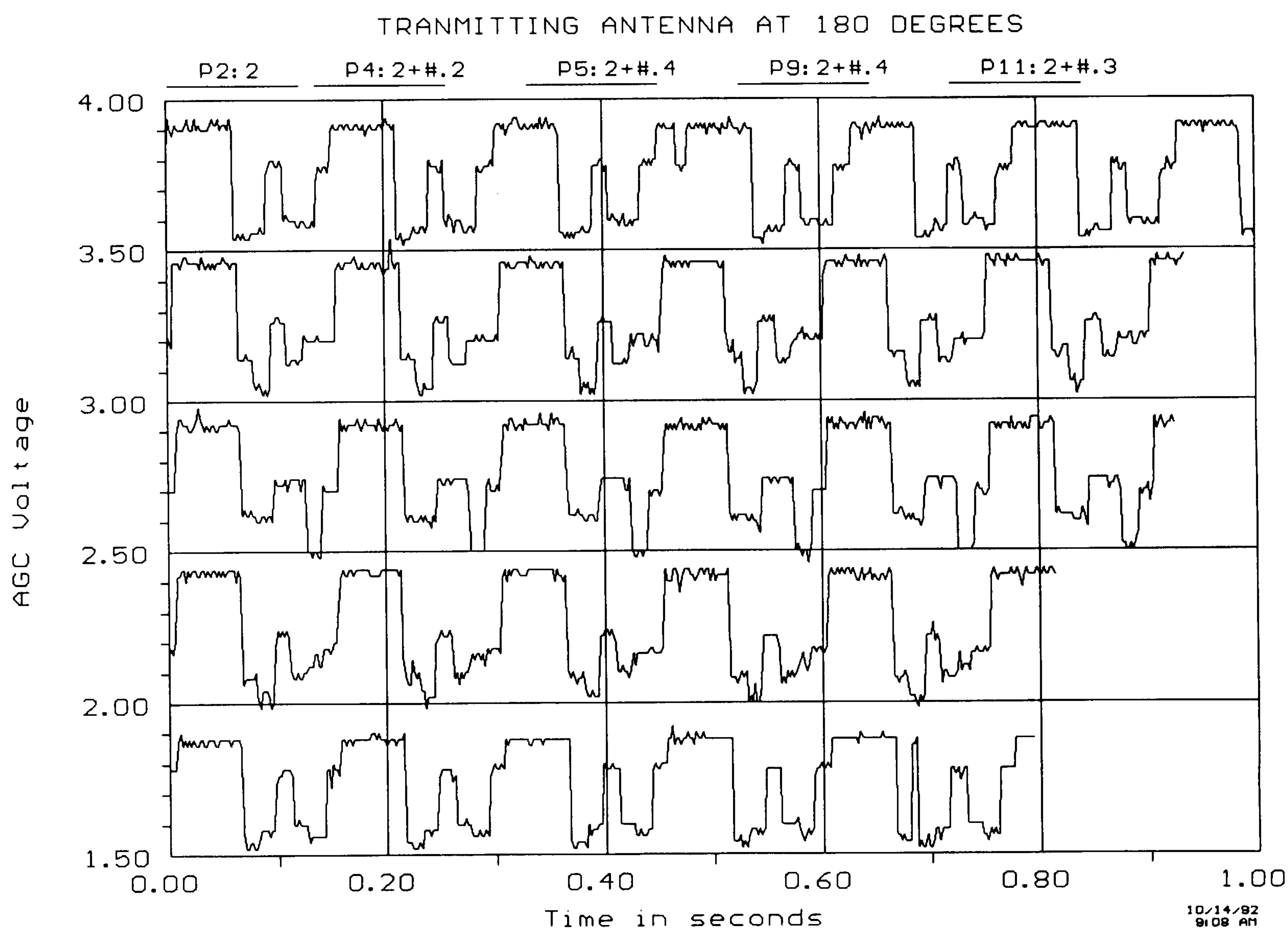


Figure 4.2-2e Static Test Waveforms

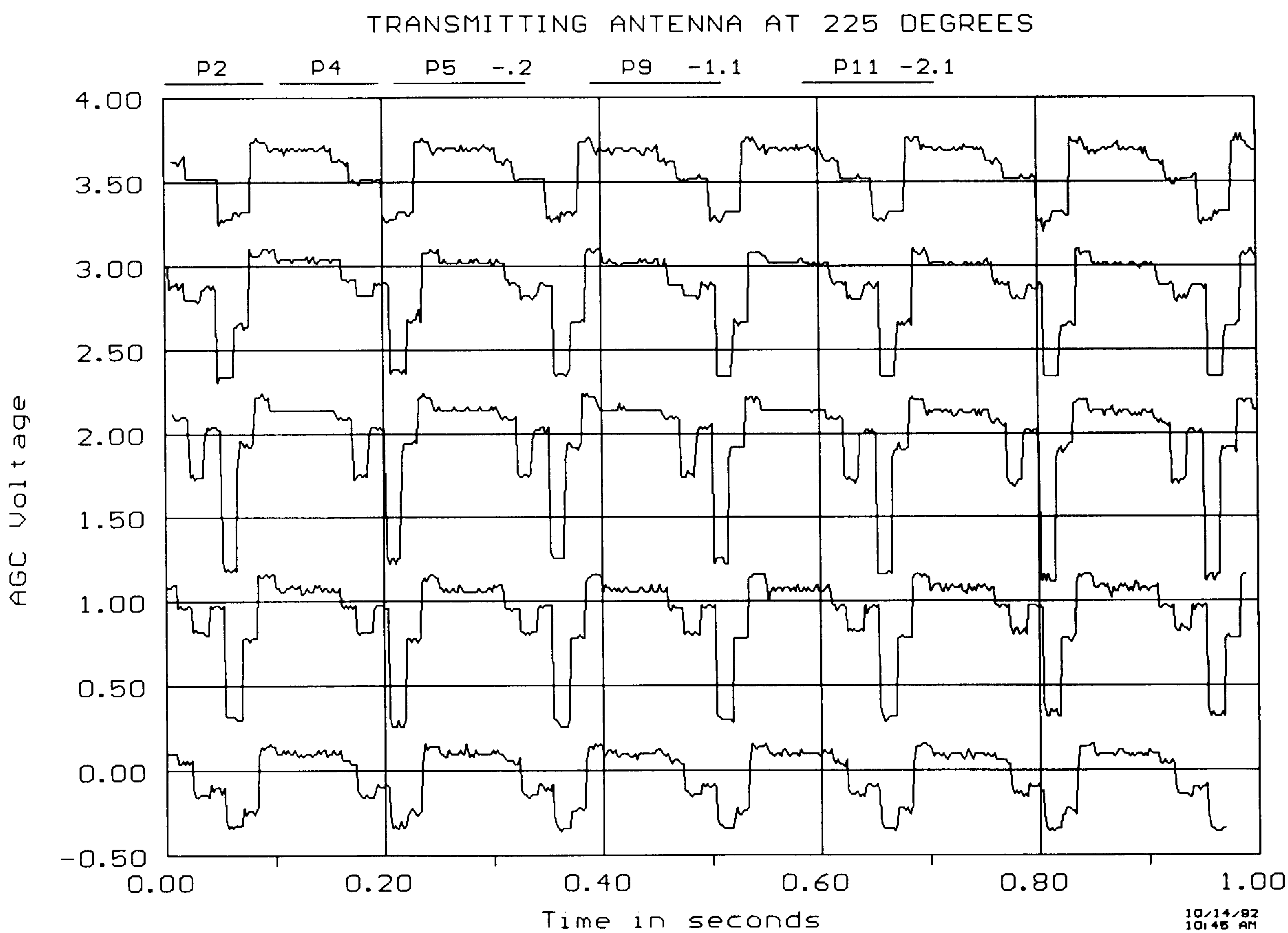


Figure 4.2-2f Static Test Waveforms

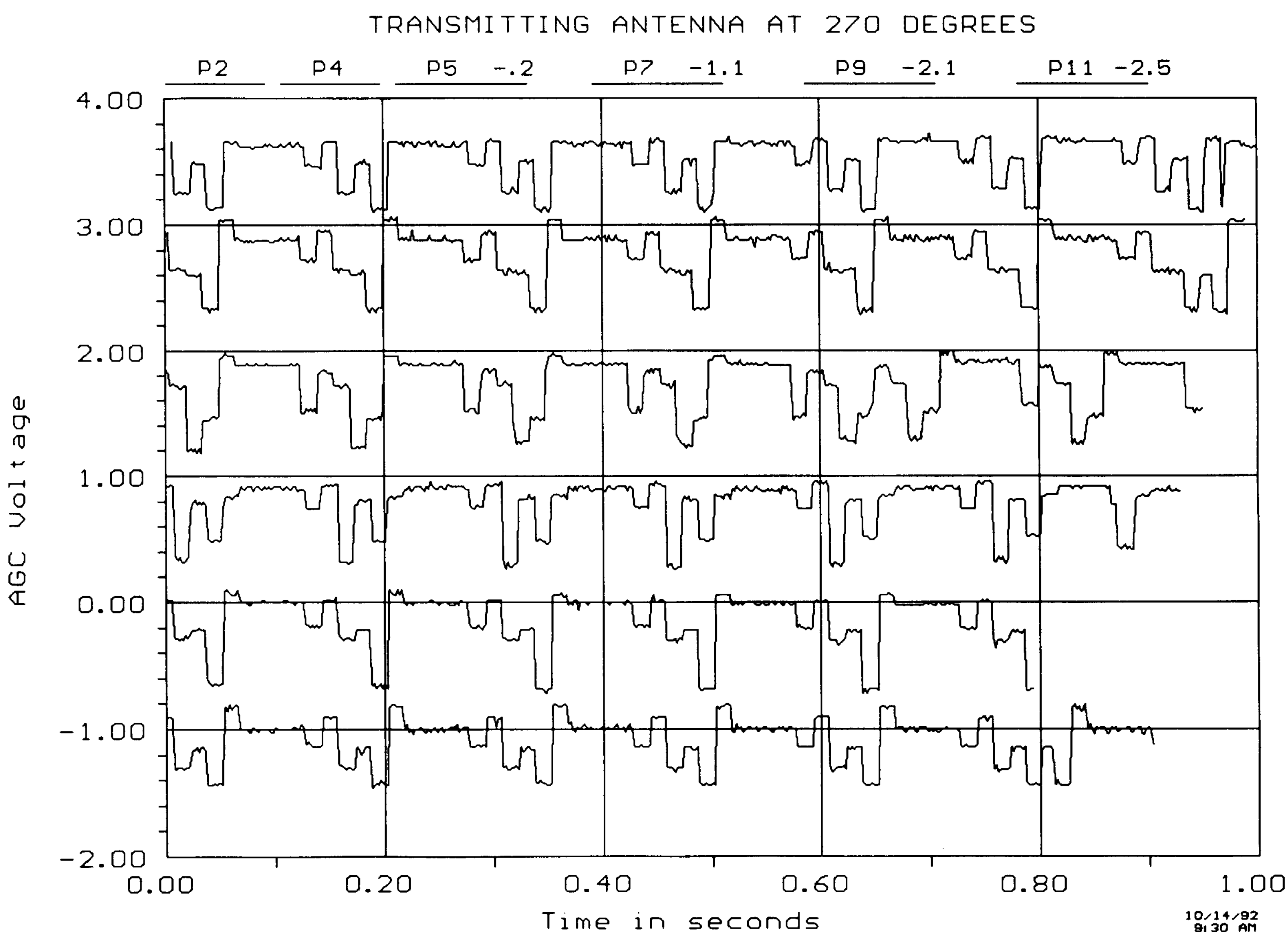


Figure 4.2-2g Static Test Waveforms

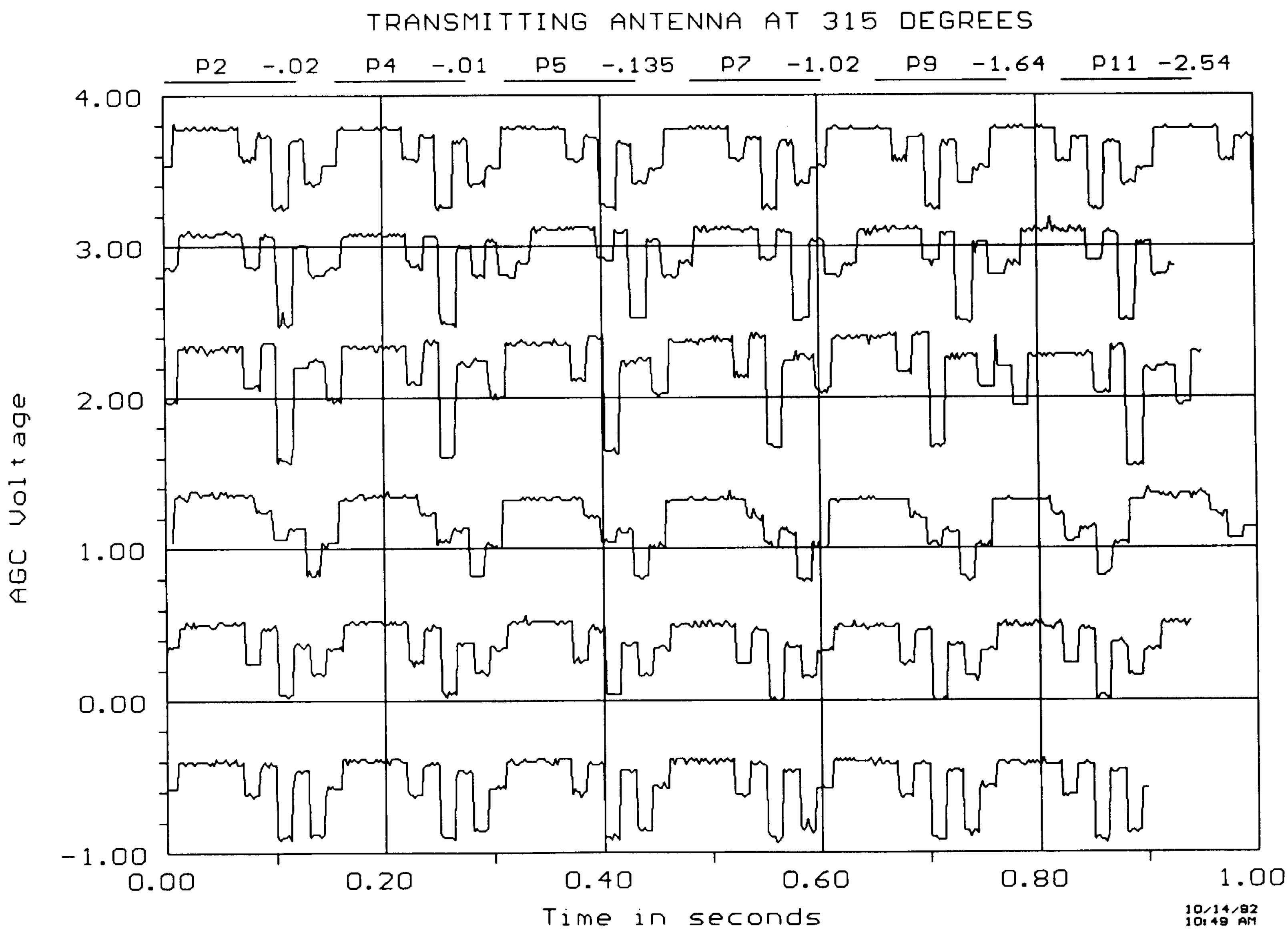


Figure 4.2-2h Static Test Waveforms

waveforms could be distinguished at distances greater than a quarter of a mile with obstacles present. Signals at distances greater than 1/2 mile were received and had good recognizable waveforms as can be seen in Figure 4.2-3. The peak power transmitted was less than 0.4 watts. Increasing the power to the authorized limit of one watt would further increase this range. Several inexpensive amplifiers in the one watt range have recently been introduced in the cellular telephone market.

4.3 DYNAMIC ROAD DATA

Dynamic road data was recorded during advanced field testing of the transmitter and receiver systems. The test area layout was the same as for the static road test.

During the dynamic road test the transmitter was placed on the top of a vehicle, pointed at a known angle, and allowed to step through the antenna patterns. The receiver was placed on the top of another vehicle and driven toward and away from the transmitter. Received waveform data was collected while the receiving vehicle was in motion, and the data was stored to disk. Several tests were made with the transmitting antenna set to different angles and the receiver at different locations. Figures 4.3-1a through 4.3-1d show the data received during the dynamic road test.

Tests were also performed to observe the multipath effects during the dynamic test. During these tests the transmitting antenna was set up to transmit omni-directional. The receiver then moved away from the transmitter at a constant speed. The received signal level waveform was recorded and stored to disk. The effects of multipath from the road and nearby objects can be seen in the waveforms. Figures 4.3-2a through 4.3-2c show these waveforms. Section 5.2 will look further at the multipath effects.

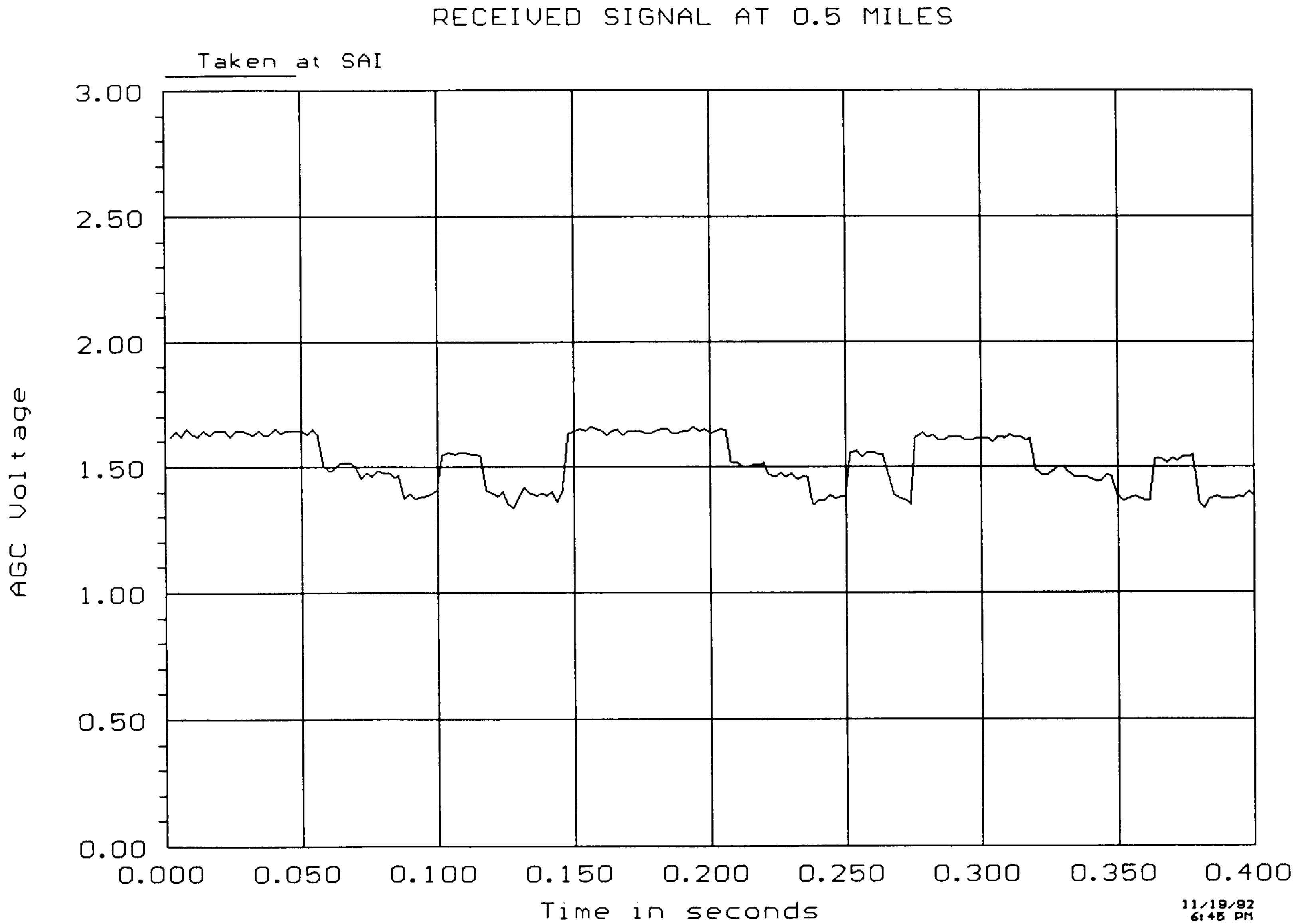


Figure 4.2-3 Received Signal at 0.5 Miles

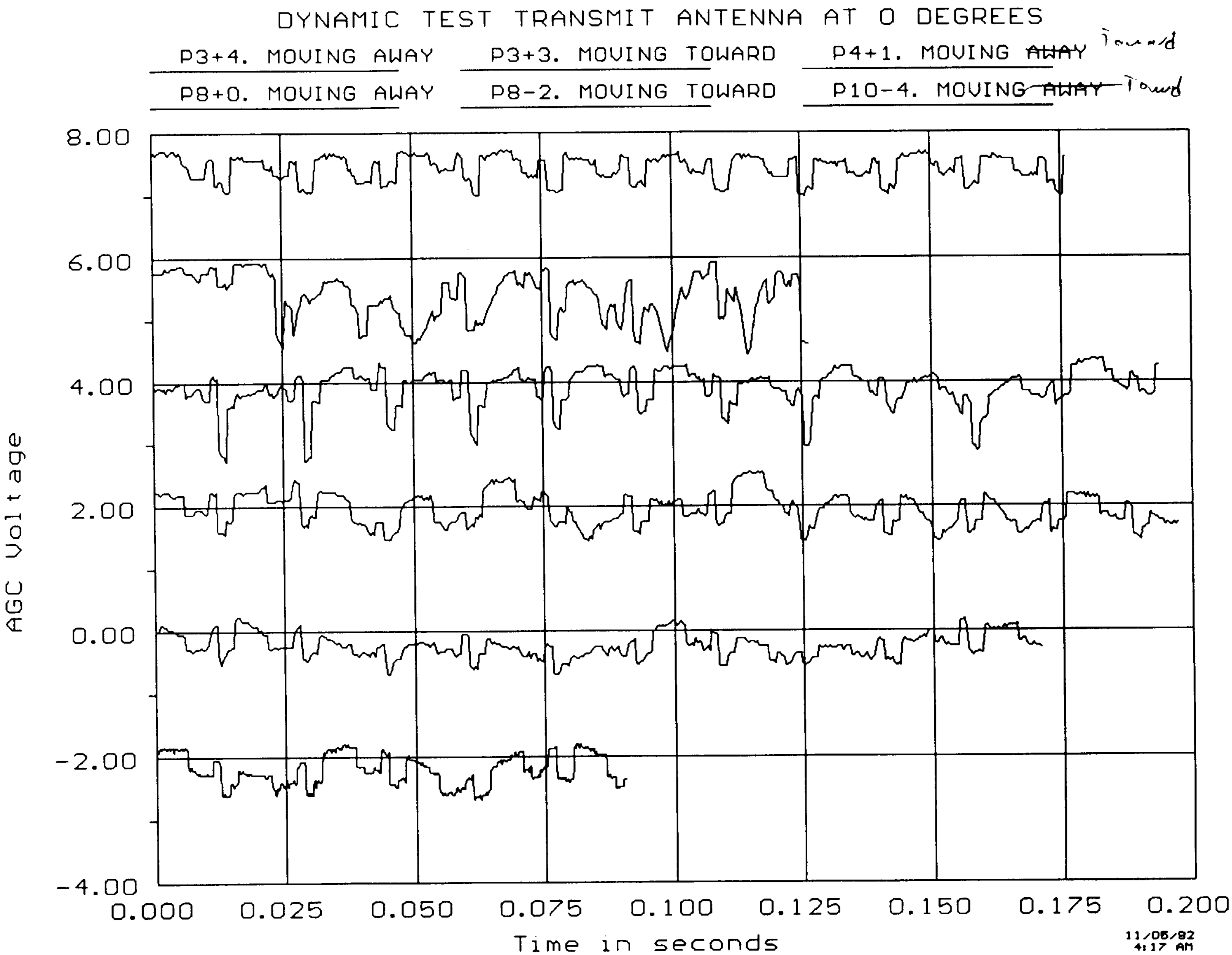


Figure 4.3-1a Dynamic Road Test Data

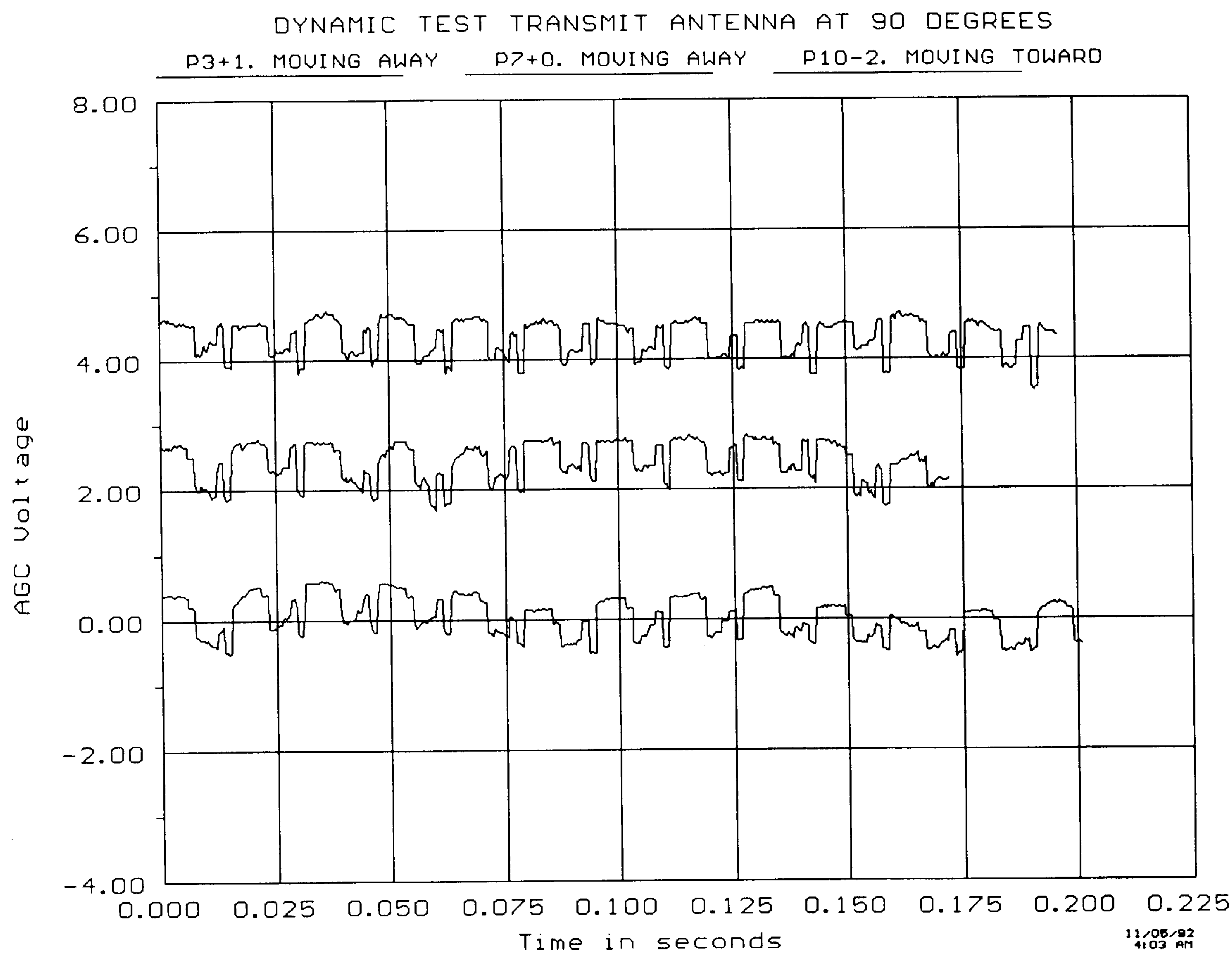


Figure 4.3-1b Dynamic Road Test Data

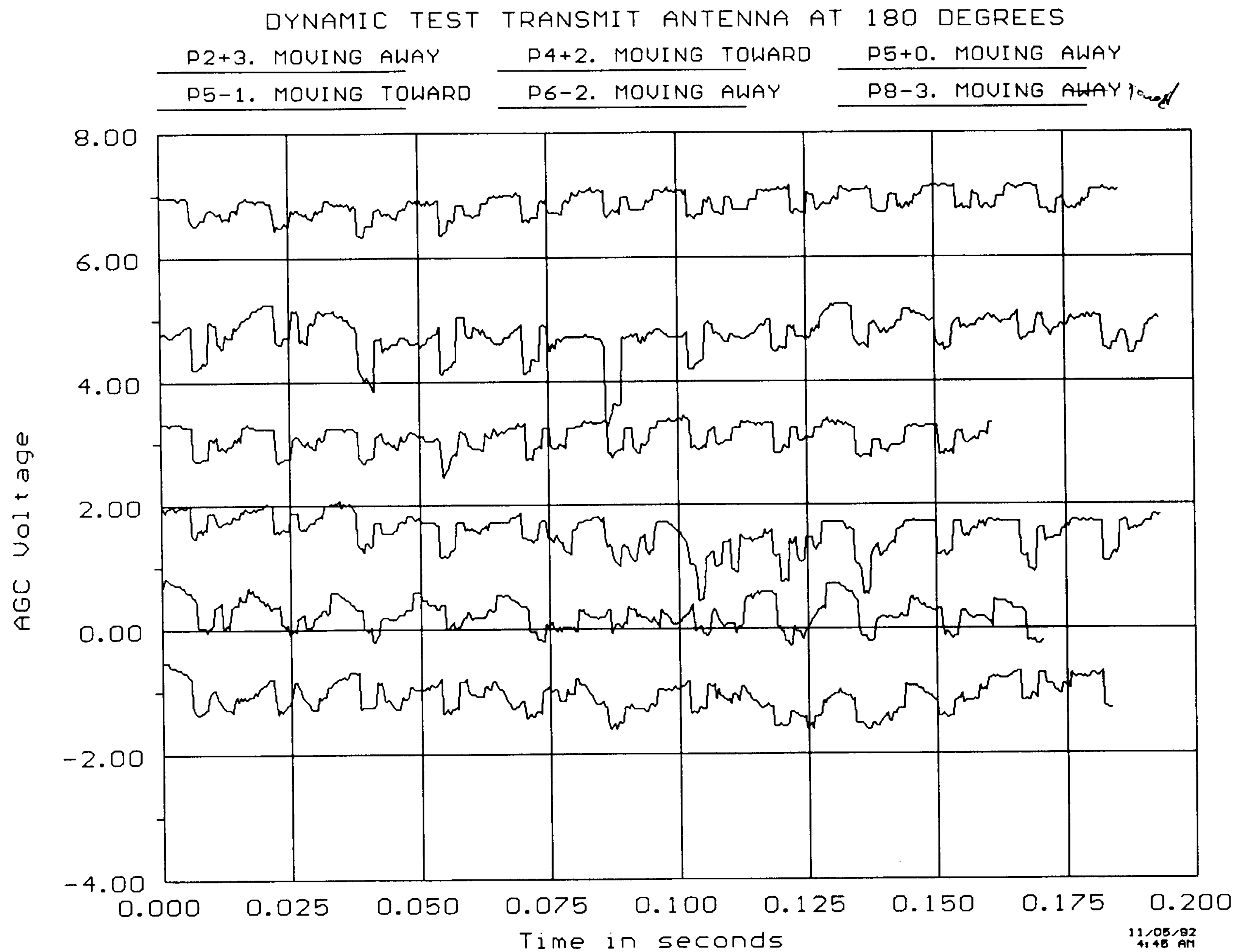


Figure 4.3-1c Dynamic Road Test Data

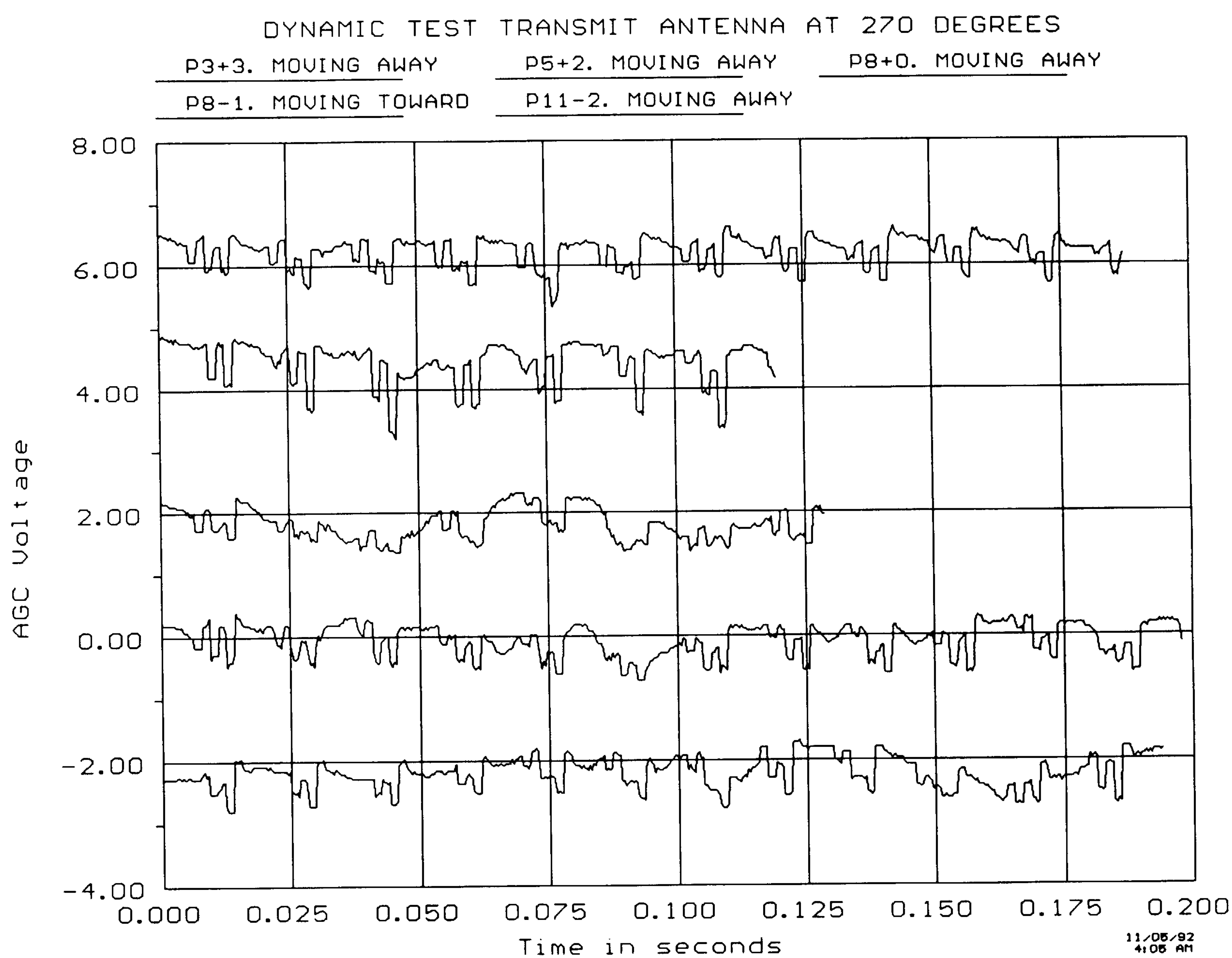


Figure 4.3-1d Dynamic Road Test Data

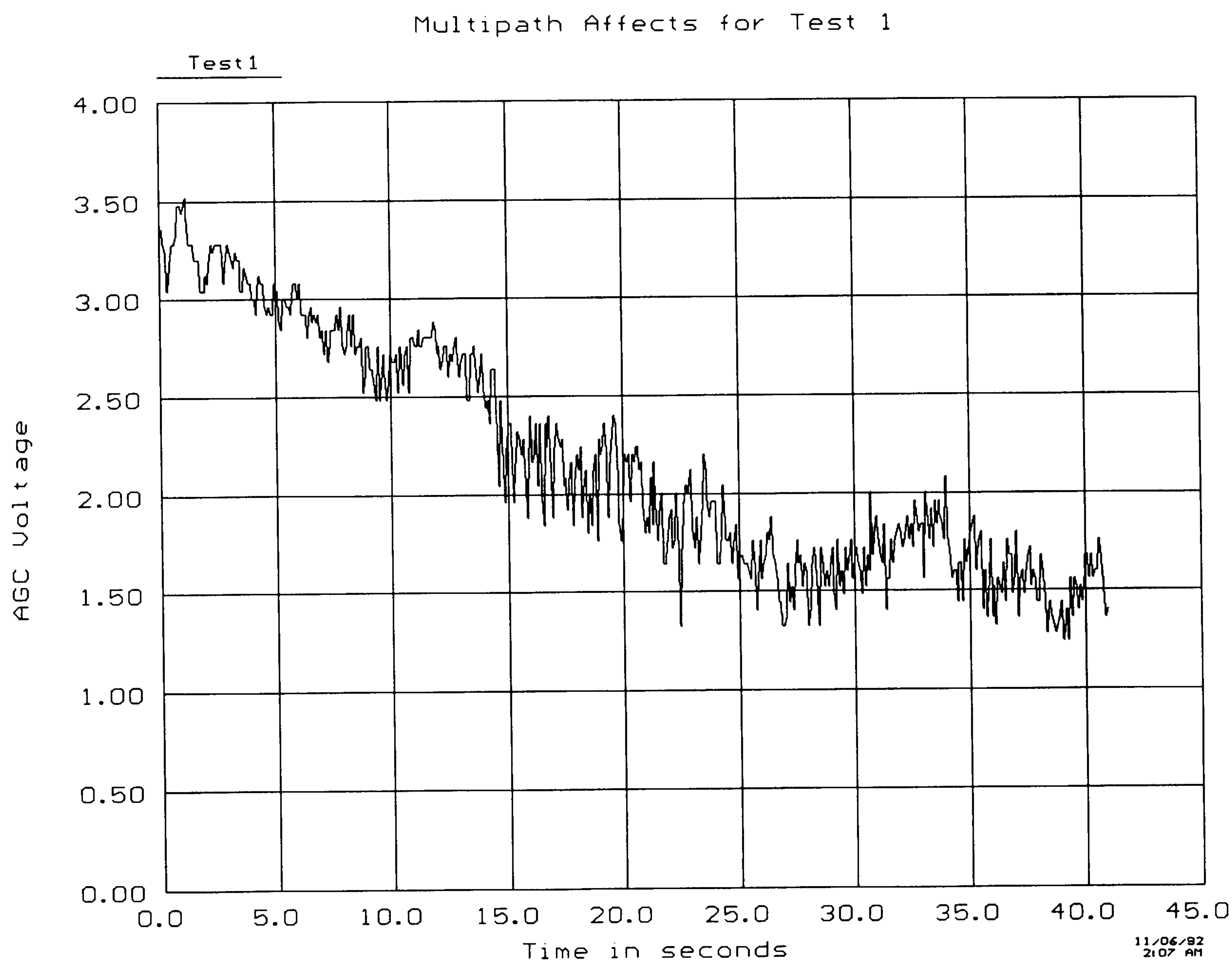


Figure 4.3-2a Multipath Effects on Signal Level, Test 1

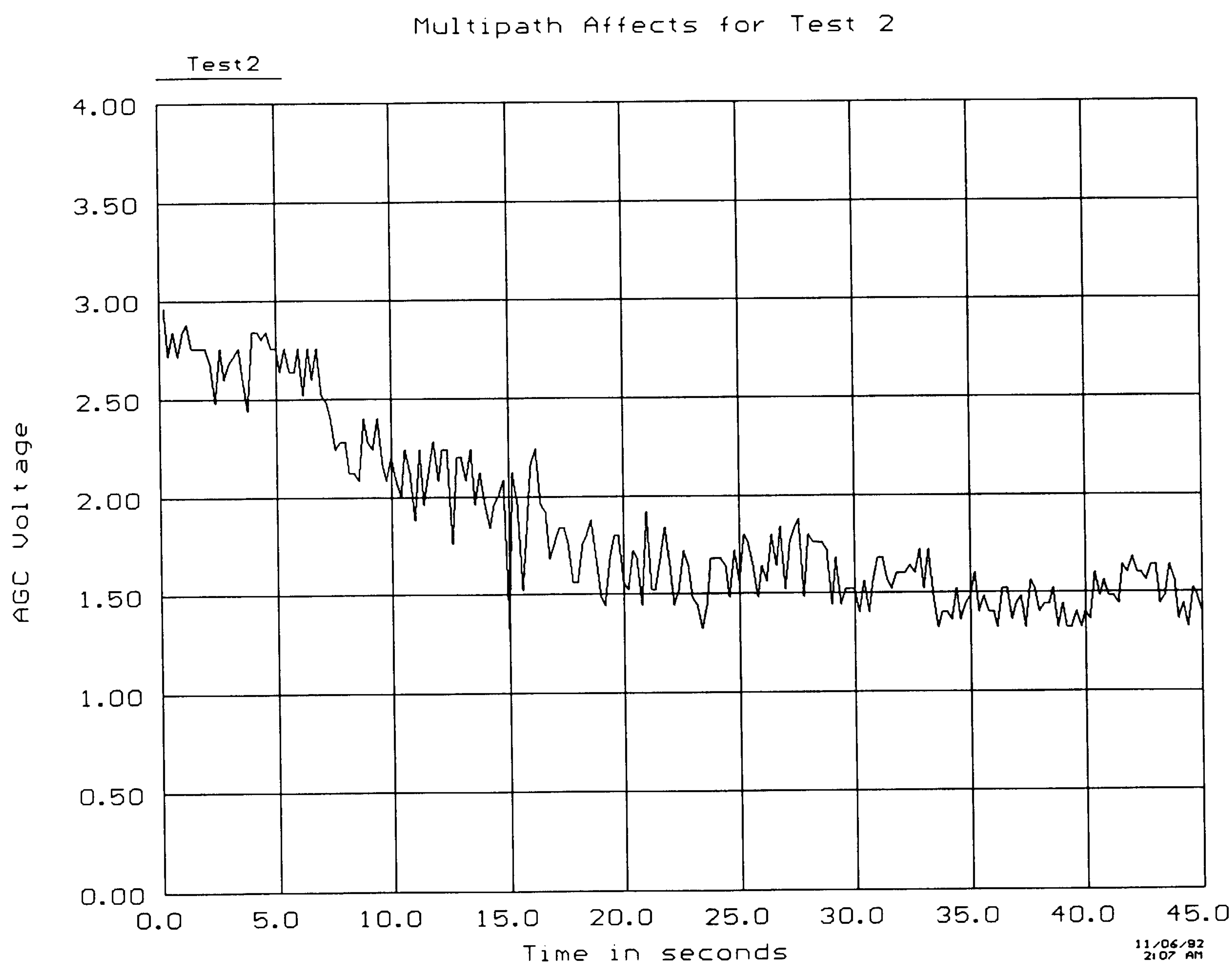


Figure 4.3-2b Multipath Effects on Signal Level, Test 2

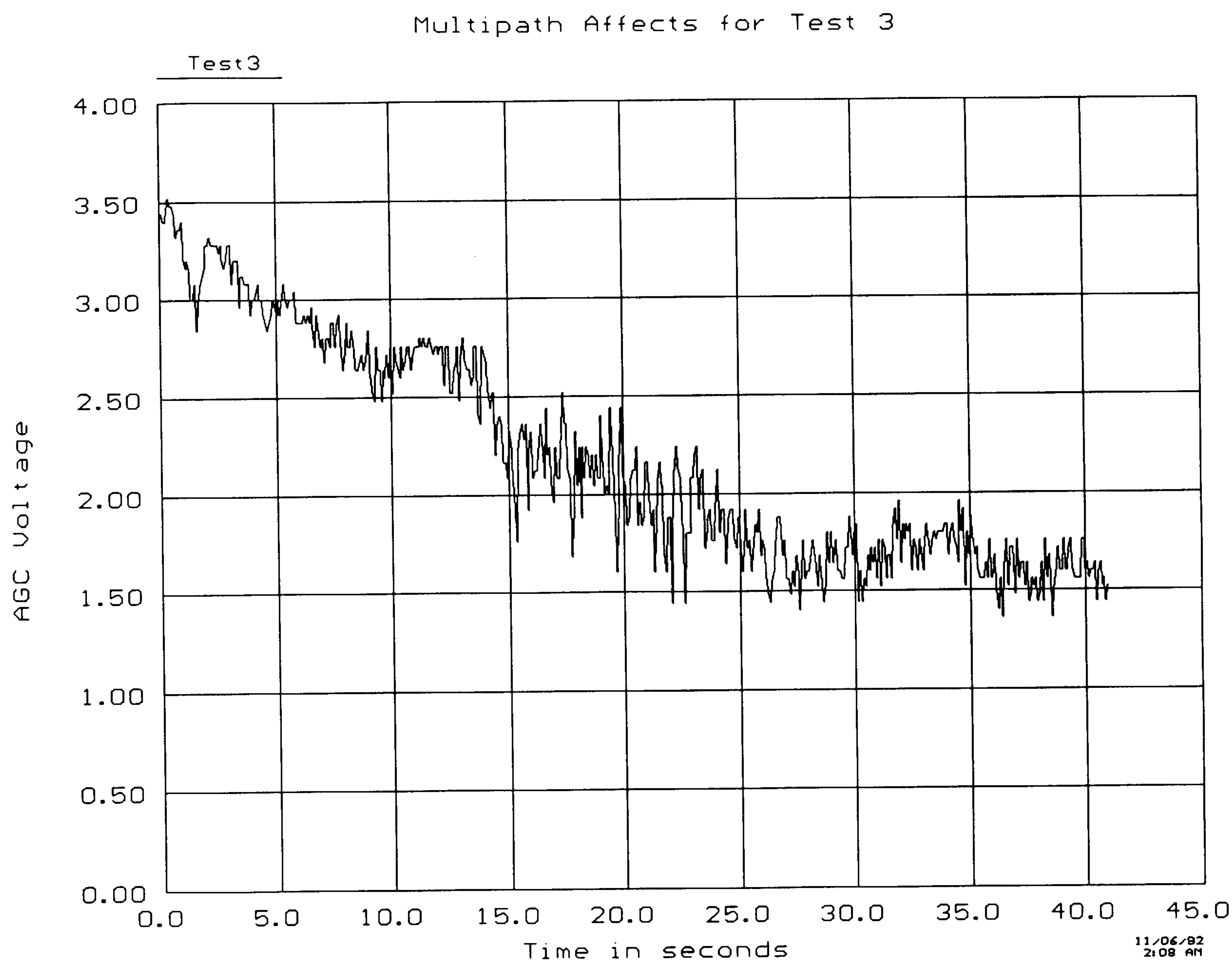


Figure 4.3-2c Multipath Effects on Signal Level, Test 3

5.0 DATA ANALYSIS

This section will describe the manner in which the data collected during the reference, static and dynamic tests described earlier was used to derive direction. Two methods of data analysis were looked at to determine a match to an antenna direction.

5.1 DIRECTION DETERMINATION

It can be seen by observation of the previous waveforms that the field test waveforms are very similar to reference waveforms of like angles. The data that produced the plots of these waveforms is contained in ASCII files on floppy discs. Ten data points, one for each of the four rest periods and the six antenna patterns, were extracted from each of the ASCII files. Ten point data sets for each of events in the field tests and the reference tests were captured. The data sets were synchronized in time by using the four rest periods as a sync. In the operational system the synchronization can easily be obtained by the data transmitted during this period. The field test data sets were then correlated or compared to reference data to determine a best match.

5.1.1 WAVEFORM MATCHING

Comparison was performed between the test data sets and the reference data sets. There were 8 reference waveforms used in the comparison: 0 degree waveform, 40 degree waveform, 90 degree waveform, 130 degree waveform, 180 degree waveform, 220 degree waveform, 270 degree waveform, and 320 degree waveform. Figure 5.1.1-1a, b, shows a graphical representation of the reference waveforms and Table 5.1.1-1 lists the waveforms in tabulated format.The test data set was then processed through the reference data sets with a Minimum Absolute Difference (MAD) routine. With this routine each point of the received test data set is subtracted from the reference data set; then, the absolute values of the differences are summed together. The reference data set producing the minimum sum is declared the best match. Figure 5.1.1-2 shows a sample MAD comparison to determine the direction using a test data set.

A cross correlation method of matching the data was also developed. This correlation method multiplied the reference waveform and the received waveform and summed the result. It produced a peak when the signals were correlated. The MAD comparison method produced the same results and used only additions and subtractions. An analysis of the MAD routine showed that it could be easily implemented in a a Motorola MC68HCII microcontroller and would produce the 8 results for comparison in approximately 4000 clock cycles of the microprocessor. If the microcontroller is clocked at 2 MHz, it would take only 2 milliseconds to produce the result.

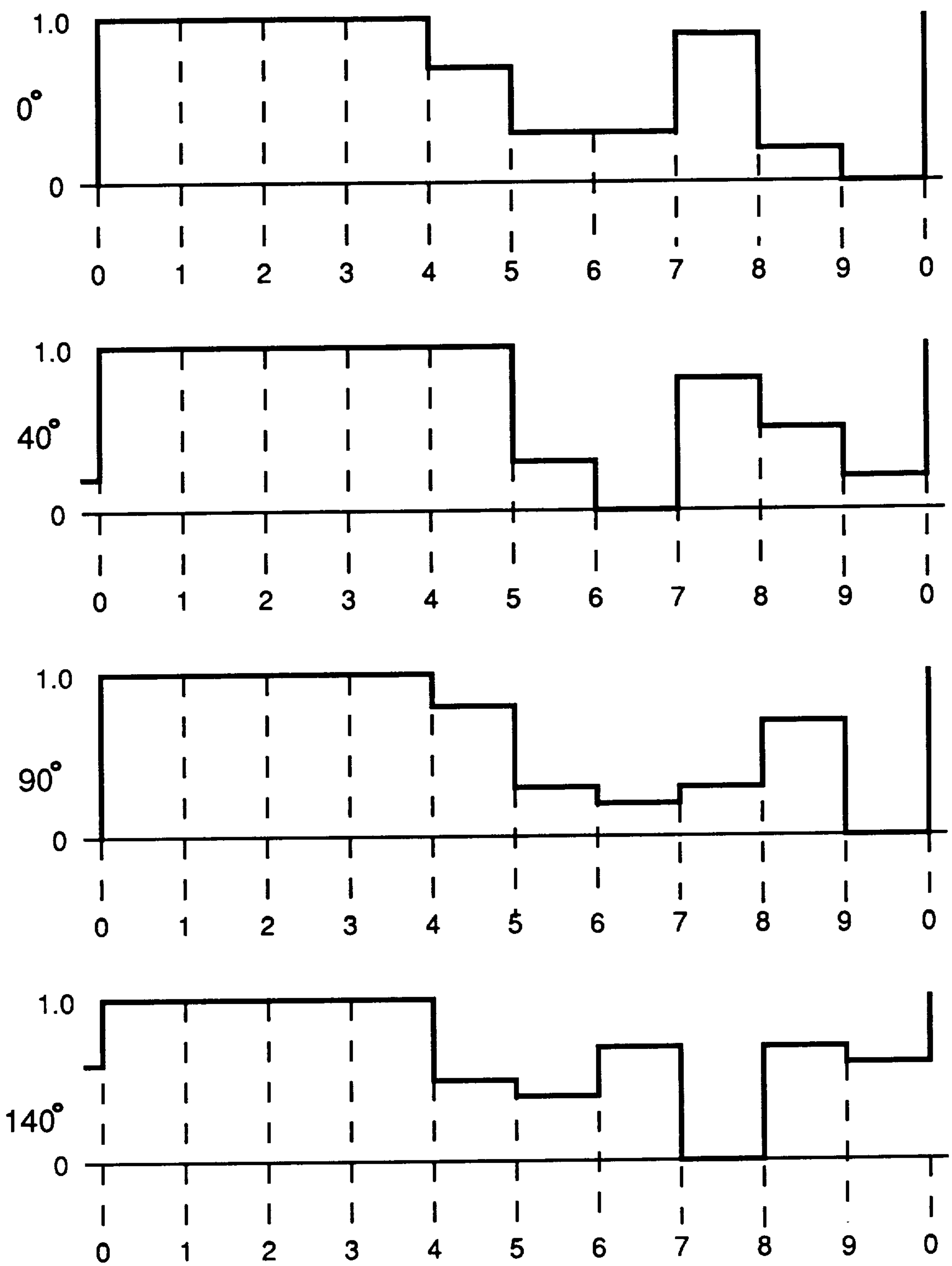


Figure 5.1.1-1a Scaled Reference Curves

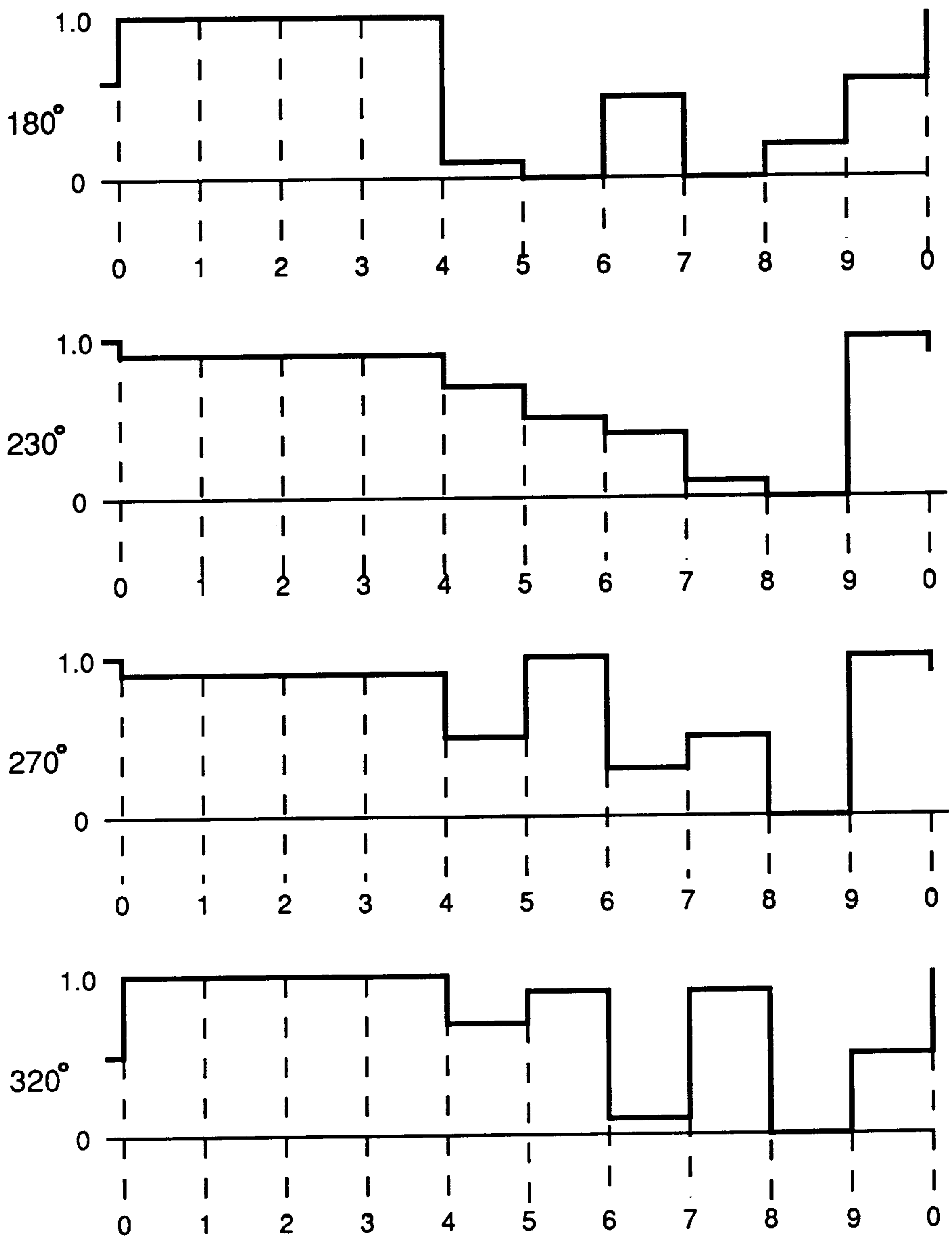


Figure 5.1.1-1b Scaled Reference Curves

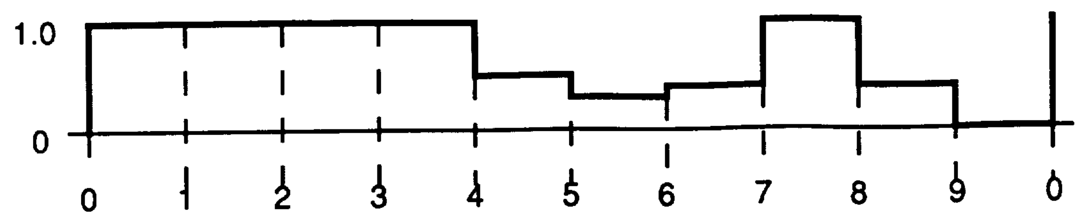
118-1044-1192-dw6

	STEP									
ANGLE	0	1	2	3	4	5	6	7	8	9
0°	1.0	1.0	1.0	1.0	0.7	0.3	0.3	0.9	0.2	0.0
40°	1.0	1.0	1.0	1.0	1.0	0.3	0.0	0.8	0.5	0.2
90°	1.0	1.0	1.0	1.0	0.8	0.3	0.2	0.3	0.7	0.0
140°	1.0	1.0	1.0	1.0	0.5	0.4	0.7	0.0	0.7	0.6
180°	1.0	1.0	1.0	1.0	0.1	0.0	0.6	0.0	0.3	0.6
230°	0.9	0.9	0.9	0.9	0.7	0.5	0.4	0.1	0.0	1.0
270°	0.9	0.9	0.9	0.9	0.5	1.0	0.3	0.5	0.0	1.0
320°	1.0	1.0	1.0	1.0	0.7	0.9	0.1	0.9	0.0	0.5

REST PERIOD

Table 5.1.1-1 Scaled Reference Waveform Values

118-1045-1192-dwg



SCALED INPUT WAVEFORM

STEP	SCALED INPUT WAVEFORM VALUES
0	1.0
1	1.0
2	1.0
3	1.0
4	0.5
5	0.3
6	0.4
7	1.0
8	0.4
9	0.0

STEP	0	1	2	3	4	5	6	7	8	9	
0 DEG. REF.	1.0	1.0	1.0	1.0	0.7	0.3	0.3	0.9	0.2	0.0	
INPUT	1.0	1.0	1.0	1.0	0.5	0.3	0.4	1.0	0.4	0.0	TOTAL
ABS DIFF	0.0	0.0	0.0	0.0	0.2	0.0	0.1	0.1	0.2	0.0	0.5

STEP	0	1	2	3	4	5	6	7	8	9	
40 DEG. REF.	1.0	1.0	1.0	1.0	1.0	0.3	0.0	0.8	0.5	0.2	
INPUT	1.0	1.0	1.0	1.0	0.5	0.3	0.4	1.0	0.4	0.0	TOTAL
ABS DIFF	0.0	0.0	0.0	0.0	0.5	0.0	0.4	0.2	0.1	0.2	1.4

STEP	0	1	2	3	4	5	6	7	8	9	
90 DEG. REF.	1.0	1.0	1.0	1.0	0.8	0.3	0.2	0.3	0.7	0.0	
INPUT	1.0	1.0	1.0	1.0	0.5	0.3	0.4	1.0	0.4	0.0	TOTAL
ABS DIFF	0.0	0.0	0.0	0.0	0.3	0.0	0.2	0.7	0.3	0.0	1.5

STEP	0	1	2	3	4	5	6	7	8	9	
140 DEG. REF.	1.0	1.0	1.0	1.0	0.5	0.4	0.7	0.0	0.7	0.6	
INPUT	1.0	1.0	1.0	1.0	0.5	0.3	0.4	1.0	0.4	0.0	TOTAL
ABS DIFF	0.0	0.0	0.0	0.0	0.0	0.1	0.3	1.0	0.3	0.6	2.3

STEP	0	1	2	3	4	5	6	7	8	9	
180 DEG. REF.	1.0	1.0	1.0	1.0	0.1	0.0	0.6	0.0	0.3	0.6	
INPUT	1.0	1.0	1.0	1.0	0.5	0.3	0.4	1.0	0.4	0.0	TOTAL
ABS DIFF	0.0	0.0	0.0	0.0	0.4	0.3	0.2	1.0	0.1	0.6	2.6

STEP	0	1	2	3	4	5	6	7	8	9	
230 DEG. REF.	0.9	0.9	0.9	0.9	0.7	0.5	0.4	0.1	0.0	1.0	
INPUT	1.0	1.0	1.0	1.0	0.5	0.3	0.4	1.0	0.4	0.0	TOTAL
ABS DIFF	0.1	0.1	0.1	0.1	0.2	0.2	0.0	0.9	0.4	1.0	2.2

STEP	0	1	2	3	4	5	6	7	8	9	
270 DEG. REF.	0.9	0.9	0.9	0.9	0.5	1.0	0.3	0.5	0.0	1.0	
INPUT	1.0	1.0	1.0	1.0	0.5	0.3	0.4	1.0	0.4	0.0	TOTAL
ABS DIFF	0.1	0.1	0.1	0.1	0.0	0.7	0.1	0.5	0.4	1.0	2.2

STEP	0	1	2	3	4	5	6	7	8	9	
320 DEG. REF.	1.0	1.0	1.0	1.0	0.7	0.9	0.1	0.9	0.0	0.5	
INPUT	1.0	1.0	1.0	1.0	0.5	0.3	0.4	1.0	0.4	0.0	TOTAL
ABS DIFF	0.0	0.0	0.0	0.0	0.2	0.6	0.3	0.1	0.4	0.5	2.1

THE SMALLEST TOTAL IS WITH THE 0 DEGREE REFERENCE; THEREFORE, THE SIGNAL IS BEST MATCHED TO 0 DEGREES.

Figure 5.1.1-2 MAD Example

118-1046-1192-dw6

5.1.2 DIRECTION DETERMINATION RESULTS

Three types of data were recorded for analysis. The first was reference waveform data, the second was static road data, and the third was dynamic road data. See Sections 4.1, 4.2, and 4.3 for further information on the three types of system test data. Using the described data processing techniques with the data obtained in the field test, the direction was determined correctly 95% of the time, for static road tests and 90% of the time for dynamic road test. The results of 95% and 90% accuracy were felt to be impacted primarily by transient multipath effects. Therefore an attempt was made to show that this was the case for the road test data. For the road test data several consecutive transmitted waveform directions were averaged together before a result was output. Using this approach the accuracy of determining direction increased to 100%. Table 5.1.2-1 lists the results of the static and dynamic test. Appendix G shows graphic representation of the matching of static road test waveforms and reference waveforms.

5.2 MULTIPATH TRANSMISSION

During the field test it was noticed that multipath caused the signal amplitude to shift up and down. The affect of this multipath led to distorted signals at some points in the dynamic road test.

Two sources of multipath were present during the dynamic testing: ground multipath and object multipath. Ground multipath is a result of the signal being reflected from the ground. Figure 5.2-1 is an illustration of ground multipath. The effect of ground multipath is a slow variation in the signal amplitude. This is due to the fact that the difference between the path lengths of direct signal and the reflected signal is changing at a slow rate.

The second source of multipath is caused by the signal being reflected from objects, such as buildings, poles, signs, and curbs. Figure 5.2-2 is an illustration of this type of multipath. Multipath from objects causes rapid signal level variations. This is due to the rapidly changing path length difference of the direct and reflected signals.

The affects of multipath on signal amplitude are unpredictable and unavoidable in this system because of the unknown dynamic environment. However, the multipath causes only intermittent problems that were overcome by relatively simple signal processing. Further testing under a more severe dynamic environment should be performed to better verify the system.

TYPE OF FIELD TEST	NUMBER OF DIFFERENT WAVEFORMS TO COMPARE	NUMBER OF WAVEFORMS COMPARED CORRECTLY USING ONLY 1 CYCLE	NUMBER OF WAVEFORMS COMPARED CORRECTLY USING 5 CYCLES AND MAJORITY VOTE
STATIC ROAD TEST	49	48 (98%)	NA
DYNAMIC ROAD TEST	20	18 (90%)	20 (100%)

Table 5.1.2-1 System Test Results

118-1047-1192-dwd

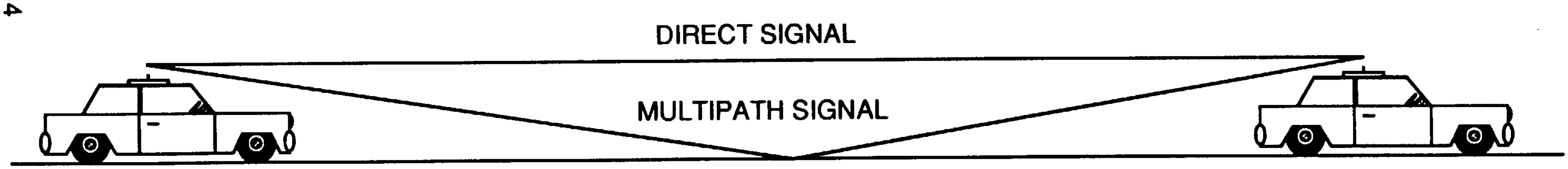


Figure 5.2-1 Ground Multipath

118-1048-1192-dw

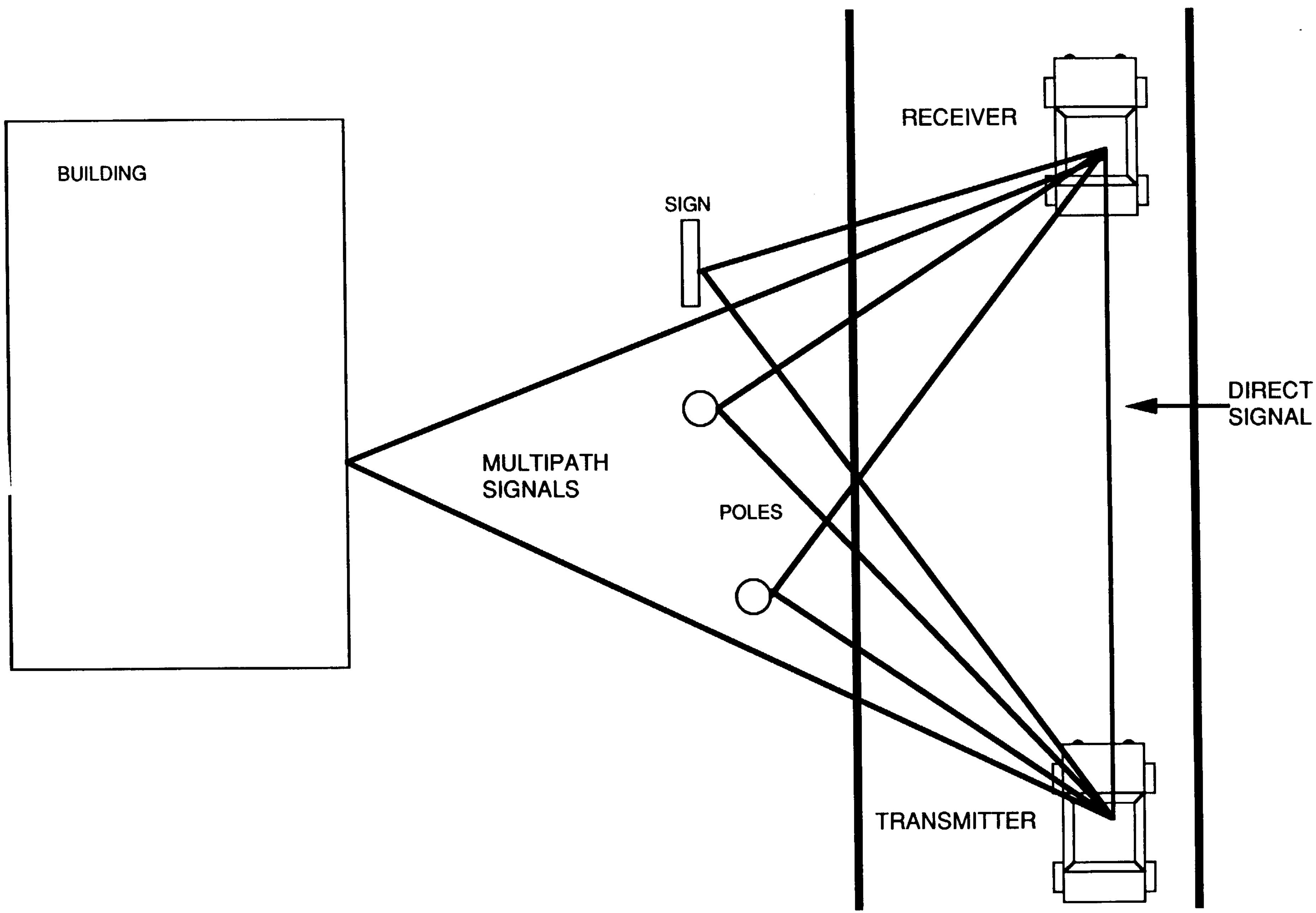


Figure 5.2-2 Object Multipath

118-1049-1192-dwg

6.0 CONCLUSIONS

This second phase effort of the EVAS system program has shown that the MSFC-proposed method of determining an emergency vehicle's direction and range can be achieved using a low cost radio link.

A transmitter and receiver was developed for use in the license free 902 to 928 MHz band per FCC Part 15 regulations. A three element phased array transmitting antenna system was developed that would generate six unique patterns to be used to determine vehicle direction.

Static and dynamic field tests were performed with the direction determined to the nearest quadrant over 90% of the time with a single comparison.

7.0 RECOMMENDED PHASE 3 WORK

For the Phase 3 EVAS work we recommend three tasks with the goal to fully prove the EVAS concept under adverse environmental conditions.

TASK #1 Transmitter/Receiver/Antenna

Two spread spectrum transmitters/receivers and two direction determining antennas will be fabricated for the extended field testing. The design will be based on the Phase 3 units with necessary improvements dictated by experimental field testing. Units will be packaged into as small as possible enclosures using available commercial components.

TASK #2 Design and Test Microcontroller System

Design, build and test two microcontrollers for operational control and signal processing of the PV and EV data. This will include display devices, antenna control and algorithms to process position and range data. Units will be packaged to interface with the transmitter/receiver and antenna system.

TASK #3 Field Testing and Data Analysis

Field testing of the two EVAS systems will begin immediately after construction and test of the transmitter, receiver, antenna and controller/display. Tests will be conducted to determine the overall ability of the EVAS to provide consistent and reliable warning to hearing-impaired PV operators. Extensive testing will be conducted to determine the effects of buildings, terrain and traffic density on the EVAS operation. Tests will also be made to determine effects of multiple EV operation in close proximity to each other.

APPENDIX A

OCI-100
SPREAD SPECTRUM ASIC

PRELIMINARY DATA SHEET

OCI Spread Spectrum ASIC

Overview

The OCI spread spectrum ASIC can be used to provide a low cost direct sequence spread spectrum communication system when used with simple external circuitry. While the ASIC was developed for use in UHF communications systems which need to comply with Part 15 of the FCC Rules, it may also be used in other applications, such as spread spectrum carrier current transmission.

Refer to Fig. 1 for a block diagram of a typical spread spectrum system using the ASIC. The ASIC provides the following functions:

1) radio helpers, which provide low cost frequency control for a 915 MHz radio to work with the ASIC and to comply with Part 15 of the FCC Rules.
2) transmit functions, which code a transmitted data stream into a baseband direct sequence "chip"stream
3) receive functions, which recover the received data and data clock from the received chip stream, and provide a digital carrier detection function for use in CSMA systems.

The ASIC uses post-detection despreading. That allows it to be easily used with limiter-discriminator integrated circuits, such as the Motorola MC13055, while still providing spread spectrum operation. The physical modulation can be frequency shift keying at the spread spectrum chip rate. Binary phase shift keying (phase reversal keying) can also be used, but the user must provide the conversion from RF to demodulated chips. A theoretical discussion of the relative merits of FSK and BPSK is included in the appendix.

Radio Helpers

The radio helpers provide circuits necessary to control the frequencies of two external vco's used in two frequency synthesizers. The divide by n numbers are loaded into a 24 bit serial to parallel register. The outputs of the shift register are used as presets to four divide by n counters.

The receiver frequency counter is an 8 bit counter with an initial divide by n of 153. This is achieved by preloading with 256-153. The receiver reference counter is a 7 bit divide by n with an initial divide by n of 50. The transmit frequency counter is a 5 bit counter with an initial divide by n of 21. The transmit reference counter is a 4 bit divide by n with an initial divide by n of 8. These numbers provide operation at 910.983 MHz when used with a 4.9152 MHz crystal, an external divide by 64 prescaler and a 51.61 MHz first IF. The counters all divide by 2^n-k where n is the number of bits and k is the load number. The

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100 Thanet Circle
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(609) 497-6800

Receiver

The despreader operates on a baseband received chip stream (post-IF despreading) to recover the received data and clock. The despreader converts a stream of digital samples taken at the 4XCLK rate to an output bit stream at 1/64th of that rate. An output bit stream clock and carrier detection signals are also generated from information in the sampled stream. The despreader has four major blocks , a descrambler, a correlator, a clock extractor, and a carrier detector.

The descrambler performs the inverse function of the scrambler.

Despreading is performed by a digital matched filter, which is matched to the chip code used to spread the signal. The code is fixed in the ASIC. When the received signal is matched in the filter, there is a maximum amplitude response of the filter, which is used to recover the transmitted data and to recover clock information.

Clock extraction is performed by state machines in the ASIC and a digital PLL, which adjusts the receive clock in 1/4 chip increments. Data to be transmitted by the ASIC should be preceded by a preamble of at least 8 bits to allow the DPLL to lock onto the data. Longer preambles are preferable if there are transmission impairments in other parts of the system, such as transmitter key-up time, receiver threshold settling time, etc. In normal operation in the recommended 915 MHz radio system, a 3 ms preamble is used.

The carrier detector is a state machine which determines when carrier is present. When a clock has been recovered, the occurrence of three more correlations in the expected time window sets the carrier detect true. The loss of correlation for 15 bit times clears the carrier detect state.

Eleictrical Data

Absolute Maximum Ratings

Supply Voltage	7.0 Volts
power dissipation	500 mW
Ambient Temperture	70 C

Normal Operating Ranges

PARAMETERS	MINIMUM	MAXIMUM	CONDITIONS
Supply Voltage Vdd	4.5	5.5	
VIL	0	0.3Vdd	
VIH	0.7Vdd	Vdd	

LOOP\ - This signal enables local loopback of chipout to chipin when it is a 0. It is included for test purposes, which may included in the final products. This is a static signal.

SCRAMBLE - This signal enables the scramble and descramble circuits when a 1 . It is included to allow for error rate and spectral testing with and without scrambling. This is a static signal.

RESET\ - The reset signal is used to put the chip into a known state. Specific registers and flip flops that are reset are defined in the design documentation. This is a Schmitt trigger input.

STANDBY\ - This pin is used to reduce the operating power of the ASIC to a minimum. It does this by inhibiting the clock to all circuits on the chip when a 0, including the synthesizer counters.

CLOCK - The main clock for the spreader/despreader portion of the ASIC. It has a maximum frequency of 10 Mhz and will have approximately 50% duty cycle.

DIV2 - This signal when a 1 enables a divide by 2 circuit for the CLOCK before passing it on the rest of the devices on the ASIC.

TXD - Transmit data from the controlling microprocessor peripheral. The data rate may be up to 156000 bits per second. The design is such that timing of TXD with respect to TXC should not be a problem with common uP support devices, independent of the edge of TXC that the data changes.

CHIPIN - Scrambled spread spectrum data input. The maximum data rate is 2.5 Megachips per second. The timing of this data is asynchronous with respect to clock. This is a Schmitt trigger input.

Outputs

CHIPOUT - This is spread spectrum data output from the ASIC to be sent into an external device. The timing is asynchronous to the external device.

CD\ - Carrier detect. Indicates that valid data appears on the CHIPIN pin. CD\ is active after 3 valid data bits have been detected and goes inactive after 16 bits are missing.

RXD - Received data. This is the demodulated data from the chipin pin. The data rate corresponds to the data rate at the TXD pin. The data occurs at least one sixteenth of a RXC before RXC. The data is valid on both edges of RXC.

RXC - Received data clock. Both edges of the clock signal are valid as references to the data signal. The clock is approximately a square wave in normal operation. While

The SD is transferred in this order:

first bit	TXREF least significant bit
	.
	.
	TXREF ms bit
	TXFREQ ls bit
	.
	.
	TXFREQ ms bit
	RXREF ls bit
	.
	.
	RXREF ms bit
	RXFREQ ls bit
	.
	.
last bit	RXFREQ ms bit

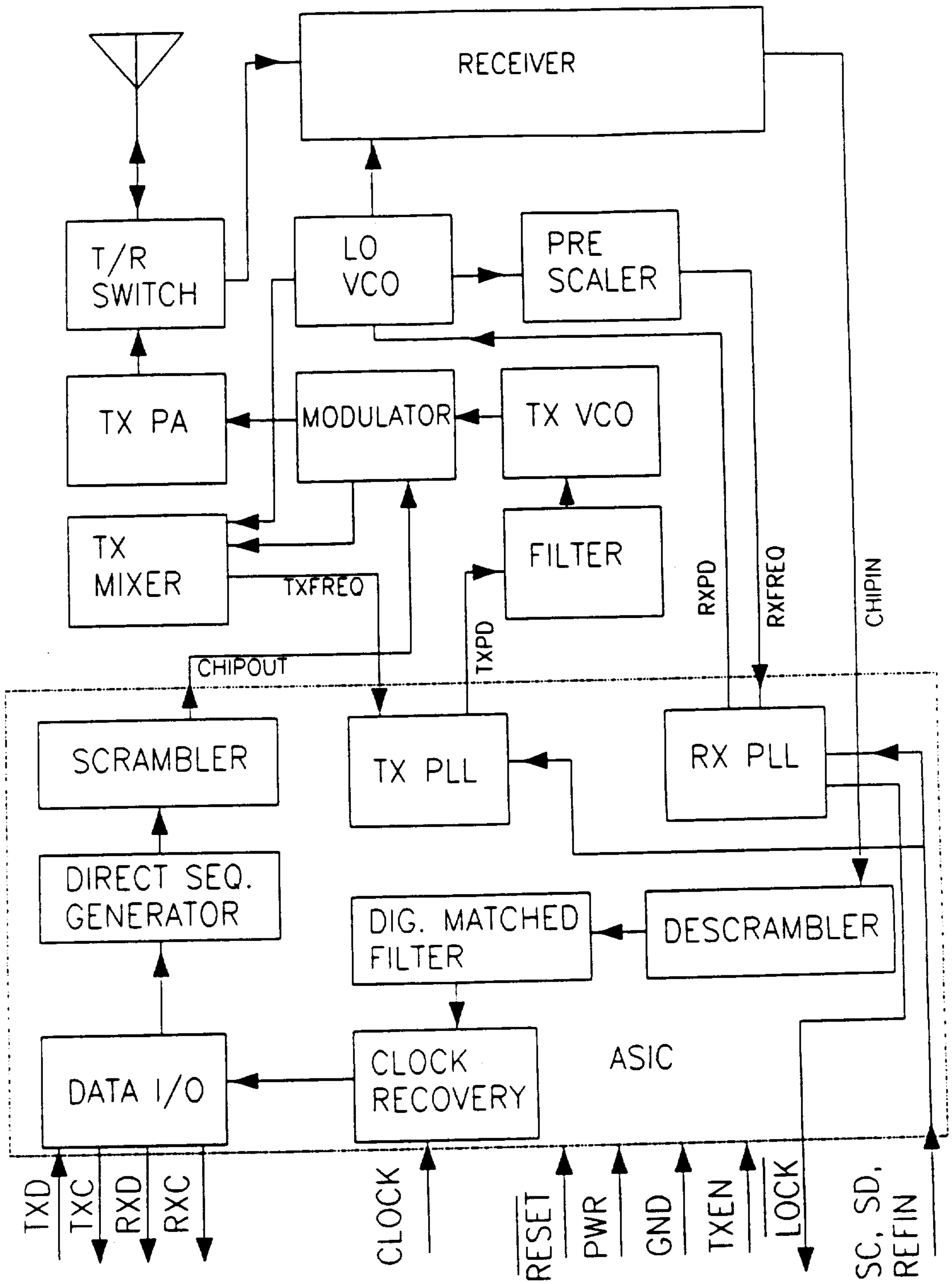
The data present on the serial data line is clocked into the frequency synthesizer on the low to high transition of the serial clock line.

The synthesizer consists of 4 up counters of bit size B. The counters are programmed by setting the count to maxcount - N where maxcount is 2 to power B, and N is the desired divide by number. M is the number of bits in the counter.

The following is sample code for setting up tables to load the counters for 4 channels.

```
;       table of synthesizer values for oci100 syntesizer
;
;       rxf? counters are 8 bits
;       rxr? counters are 7 bits
;       txf? counters are 5 bits
;       txr? counters are 4 bits
;
;       the counters are preloaded from a shift register that is
;       organized
;       as follows - all 24 bits must be loaded to change channel
;
;             rxf                 rxr                 txf                 txr
;          |           |     |           |     |           |     |           |
; SD ->    |  8 bits   |-->|  7 bits   |-->|  5 bits   |-->|  4 bits   |
;          |           |     |           |     |           |     |           |
;         msb        lsb   msb        lsb   msb        lsb   msb        lsb
;
;
```

310002 Rev. C Nov 9, 90



RECEIVER
T/R SWITCH
LO VCO
PRE SCALER
TX PA
MODULATOR
TX VCO
TX MIXER
FILTER
TXFREQ
TXPD
RXPD
RXFREQ
CHIPIN
CHIPOUT
SCRAMBLER
TX PLL
RX PLL
DIRECT SEQ. GENERATOR
DIG. MATCHED FILTER
DESCRAMBLER
DATA I/O
CLOCK RECOVERY
ASIC
TXD
TXC
RXD
RXC
CLOCK
RESET
PWR
GND
TXEN
LOCK
SC, SD, REFIN

APPENDIX

The performance of FSK direct sequence spread spectrum with post detection despreading is compared to the performance of optimum non-coherent signalling with an ideal despreader. Operation at a spreading ratio of 16 chips per bit, as in the OCI-100 is assumed, but the technique may be used for any ratio.

Let E_{ch} be the energy transmitted per chip and $N_0/2$ be the two sided noise power spectral density. Then for optimum non-coherent detection of FSK with independent decisions on each chip, the probability of a chip error is given by the well known result[1]

$$P_{ch} = \frac{1}{2}\exp\left(-\frac{E_{ch}}{2N_0}\right) \tag{1}$$

The probability of a bit error is the probability of having bit errors on more than half of the chips in the bit. Therefore, the probability of a bit error is given by

$$P_e = 1 - \sum_{n=0}^{8} \binom{16}{n} P_{ch}^{n} (1-P_{ch})^{16-n} \tag{2}$$

and the probability of a packet being received correctly is

$$P_{pok} = (1-P_e)^N \tag{3}$$

where N is the number of bits in the packet. For BPSK with ideal despreading and optimum coherent detection, the probability of chip error is irrelevant as all the chip energy is combined into one bit before the bit decision is made. Thus the probability of bit error is the same as for BPSK with spreading (spread spectrum provides no gain in AWGN):

$$P_e = Q(\sqrt{2E_b/N_0}) \tag{4}$$

$$Q(x) = \int_x^{\infty} \frac{1}{\sqrt{2\pi}} e^{-y^2/2} dy \tag{5}$$

Similarly, for optimum non-coherent signalling, with ideal despreading the probability of bit error is

$$P_e = \frac{1}{2}\exp\left(-\frac{E_b}{2N_0}\right) \tag{6}$$

Now consider the case of the non-coherent FSK spread spectrum receiver operating with a single tone interferer in AWGN. The general case is difficult to analyze in detail, but since the optimum non-coherent receiver makes the decision on the basis of a comparison of the energy in the two filters, it is clear that the worst case is an interferer exactly on one of the two transmission frequencies, say the mark frequency. This worst case is more easily

---

[1] R. E. Blahut, Digital Transmission of Information, Addison-Wesley, 1990

second type is one half the conditional probability.

The two types of error are mutually exclusive, so the total probability of error is the probability of error of the first type plus the probability of an error of the second type.

$$P_c = P_{c1} + P_{c2} \tag{12}$$

The value of P_c found above is substituted into equation 5 to calculate the probability of bit error, and that result substituted in equation 6 to calculate the probability of successful packet transfer.

The bit error rate vs SNR is plotted, for the case of a 3 dB S/I ratio, in figure 3. It can be seen that at the higher bit error rates, an ideal non-coherent spread spectrum system, such as one using BPSK spreading, provides slightly better performance. A chip rate of 16 chips per bit has been assumed in each case. At the lower bit error rates, where one would prefer to operate a wireless data system, the FSK system provides better performance, because the ideal non-coherent system converts the interferer power in to noise. As a result, it can never achieve a better despread signal to noise ratio than 3db + 10log16 = 15 dB, so it has an irreducible BER equivalent to that of non-coherent detection with a 15 SNR.

Addendum to rev C ASIC data sheet:

ASIC Current Drain depends on the clock frequency, as with all CMOS integrated circuits. Typical current drain at 5 volts and 5 MHz is 5 to 6 ma.

Interfacing the OCI - 100 IC with Radio Circuits

Application Note

1.0 Introduction

This application note provides information and experience-based hints for developing low cost, Part 15 compliant spread spectrum radio systems. The topics covered include using the OCI-100 frequency synthesizer (section 2), selecting the modulation type and receiver and transmitter type (sections 3 and 4), controlling noise and emissions (section 5) and using the OCI-100 in voice systems (section 6).

2.0 Frequency Synthesizer Hints

2.1 How to Design the Synthesizer

The OCI-100 ic is designed to allow frequency control of both the transmitter and receiver. As normally used, the receiver local oscillator is synthesized, and the transmitter oscillator is synthesized by reference to the receiver local oscillator. That is, the transmitter synthesizer holds the transmitter at a fixed frequency difference from the receiver LO. This scheme results in a few significant performance and cost advantages:

- Only one prescaler is needed
- The transmitter frequency divisors are smaller, resulting in reduced lock up time for the transmitter loop
- Different receiver reference divisors are possible, with a constant IF, resulting in a higher receiver PLL reference rate, therefore simpler PLL design for equivalent performance.

Normally, the transmitter offset is selected to equal the IF, then the transmit frequency is equal the receive frequency. Of course, the user in not constrained to use the above scheme. For instance, if the user may choose a direct conversion receiver design, in which case the transmitter loop would not be used. Or, for a dedicated link, the transmitter offset could be chosen differently at each end of the link to allow duplex operation.

As explained above, the OCI-100 includes circuitry to support two frequency control loops, one for the transmitter and one for the receiver. Usually, for operation in the 915 MHz band a divide by 64 prescaler, such as the Siemens SD 2211 can be used. This gives an output frequency in the neighborhood of 15 MHz (depending on the IF chosen and whether the LO is above or below the operating frequency).

Many low cost prescalers provide only ECL level output. In this case, a level shifter or amplifier will be needed to interface to the OCI-100. OCI has had excellent results using both single bipolar transistors and logic gates (such as the 74HCU04) as level shifters. The level shifter may be ac coupled as its signal is continuous. If the 74HCU04 is used, remember to include a series input resistor for stability (100 ohms and 100 to 1000 pf coupling capacitor is a good combination).

2.2 Phase Detector Characteristics

The design of PLL's is too lengthy a subject to review in detail in this note. Several excellent texts on the subject are available, including *Frequency Synthesizers Theory and Design* by Vadim Manassewitsch (Wiley-Interscience), *Digital PLL Frequency Synthesizers* by Ulrich L. Rohde (Prentice-Hall) and *Phaselock Techniques* by Floyd M. Gardner (Wiley-Interscience). Some elementary circuits which work well with the OCI-100 are presented below

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(609) 497-6800

and can be used as starting points.

The phase detector characteristics are important input data for PLL design. The OCI-100 uses simple digital frequency-phase detectors. In the phase detector mode, their sensitivity, in volts per radian, is approximately

$$K_{\phi}=\frac{V_{dd}}{4\pi} \tag{1}$$

The phase detector is of the charge pump variety, so it can be used with an all passive loop filter as shown in figure 1 or with active filter components. The circuit in figure 1 assumes a voltage controlled oscillator (VCO) tuning sensitivity of 20 MHz/volt, with a very high impedance input (typical of varactor tuned oscillators).

Note that the transmitter phase detector is in a high impedance state when the TXEN signal is a logic 0. This allows the use of two techniques to improve transmitter frequency acquisition time. First, a coarse tuning voltage can be applied to most loop filter designs through a high series resistance that will not affect the loop performance but will pretune the VCO to near the operating frequency. Second, if the system design calls for repetitive transmissions at short intervals, the loop filter can hold the previous voltage (by virtue of charge stored on the filter capacitor and the high impedance output), which will provide an automatic pretuning of the VCO.

2.3 Programming the Operating Frequency

The OCI-100 provides two frequency control loops, one for the transmitter and one for the receiver. Each contains two counters, a reference counter and a frequency counter. The counters all divide by 2^n-k, where n is the number of bits and k is the load number. The receive frequency is given by:

$$F_{rx}=PF_{refin}\frac{RXFREQ}{RXREF}-F_{IF} \tag{2}$$

where P is the prescaler ratio (typically 64 or 128), F_{refin} is the synthesizer reference frequency, RXFREQ is the receiver counter divisor, RXREF is the receiver reverence counter divisor, and F_{IF} is the receiver intermediate frequency (51.61 MHz. in the example below). NOTE: the intermediate frequency is not a free variable, but depends on how the transmitter PLL is programmed, for systems using the same transmit and receive frequency.

The transmitter is programmed by setting the IF as follows:

$$F_{IF}=P_{tx}F_{refin}\frac{TXFREQ}{TXREF} \tag{3}$$

Where TXFREQ is the transmitter frequency divisor, P_{tx} is the transmitter prescaler if used (if not, P_{tx} is 1) and TXREF is the transmitter reference divisor.

The bit sizes (n values) for the counters are:

RXFREQ	8 bits
RXREF	7 bits
TXFREQ	5 bits
TXREF	4 bits

The bits are transmitted to the IC (using SD and SC) MSB first. The following example illustrates the programming for an operating frequency of 910.983 MHz, with a reference frequency of 4.9152 MHz, an IF frequency of 51.61 MHz, a receiver LO prescaler of 64 and a transmitter prescaler of 4. (These are the default settings to which the IC is set after a RESET). From equation (2) above,

$$RXFREQ/RXREF = \frac{910.983+51.61}{64*4.9152} = 3.06 = 153/50 \qquad (4)$$

so we can use RXFREQ = 153 and RXREF = 50. Recall that the effective division ratio is 2^n-k, so we must calculate the k values and use them to load the IC. Thus for RXFREQ, the value of k is 103 (decimal), and it is the 103 (decimal) which is converted to binary (67 hex or 01100111 binary) and loaded into the IC using the SC and SD inputs. The values for the transmitter synthesizer are calculated from equation (3), resulting in TXFREQ = 21 and TXREF = 8. The load values are then TXFREQ load values = 32-21 = 11 and TXREF = 16-8 = 8 respectively.

The SD is transferred in this order:

first bit TXREF least significant bit
.
.
TXREF ms bit
TXFREQ ls bit
.
.
TXFREQ ms bit
RXREF ls bit
.
.
RXREF ms bit
RXFREQ ls bit
.
.
last bit RXFREQ ms bit

The data present on the serial data line is clocked into the frequency synthesizer on the low to high transition of the serial clock line. All 24 bits must be loaded to change frequency.

3.0 Transmitter and Modulation Selection

The type of modulation selected affects both the transmitter and receiver design. Frequency shift keying (FSK) offers significant advantages in simplicity and economy. In some cases, BPSK can provide better interference rejection at a cost in complexity. Other types of modulation are also possible, but are beyond the scope of this note.

3.1 FSK Modulation.

If FSK modulation is selected, it is possible to use a VCO operating on the output frequency as the transmitter, possibly with power amplifier stages if high power is desired. Oscillator transmitters require a degree of load control to provide stable, repeatable results, but can have low parts count and high DC to RF conversion efficiency. They also tend to be rich in harmonics which should be removed by an appropriate RF low pass between the oscillator and antenna.

With the OCI-100, the chip stream may be added to the transmit loop filter as shown in the diagram. Note that in this diagram, the loop filter is very narrowband to prevent the PLL from canceling the modulation. This is more important if the OCI-100 is used in the scrambler mode. If rapid frequency stabilization of the transmitter is also needed a switching circuit should be used to short out part of the loop filter input resistor during frequency acquisition. This switch can be timed from the rising edge of TXEN.

Low Pass filtering the chip stream before adding it to the PLL control voltage will reduce the sidelobe content of the transmitted signal (sometimes called a splatter filter).

3.2 BPSK and DPSK

If the designer chooses BPSK or DPSK modulation, the chip stream can be directly applied to the modulator (the use of a splatter filter or other means of sidelobe control should still be considered in the design). The OCI-100 can be used to set the transmitter frequency as described above.

4.0 Receiver Types

4.1 FSK Receivers

The OCI-100 interfaces easily to FSK receiver ICs such as the Motorola MC13055 and MC3356. These ICs have a digital output from their internal data "slicer" which is directly compatible with the OCI-100 if the receiver slicer is operated from a 5 volt source. (Be sure to decouple the two ICs if a common +5 volt supply is used, see below). We have found that including a series resistor of about 5K Ohms between the receiver slicer output and the OCI-100 input improves the receiver sensitivity slightly in some circumstances.

Other types of FM receiver integrated circuits may be applicable, including phase locked loop detectors. We have also observed that carefully designed external slicers can outperform the on-chip slicer at a cost of additional complexity.

4.2 BPSK and DPSK

Reception of BPSK/DPSK is more complex. OCI is not aware of any suitable single IC solutions. It is possible to use the above mentioned ICs to provide the IF amplification, and provide additional circuitry to operate on the limited output signal, which is available at the quadrature detector coil pin of these ICs, to demodulate the BPSK signal. If differential coding is used, a delay line and a mixer can be used. Squaring loop and Costas loop types of phase locked loop detectors can also be used. The reader is referred to any good communication theory text for additional information on these topics. Additional advice in these matters is also available under the OCI engineering assistance contract.

5.0 Noise Considerations

The OCI-100 is implemented in 1.25 micron CMOS, so that it is capable of generating pulses with rise times in the 2-3 ns. range. These pulses have frequency components well into the RF spectrum which can couple into the radio circuitry as noise or interfering signals. Also, if these signals are allowed to radiate, they can generate sufficient radiated power to create difficulties in meeting standards for electromagnetic compatibility. CMOS circuits also generate current spikes in their V_{dd} and V_{ss} circuits when they switch. These spikes can also generate electromagnetic noise. These potential problems can be avoided by proper design and layout as explained below.

It is important to bypass the OCI-100 properly. A 0.1 uF ceramic surface mount capacitor mounted very close to the V_{dd} pin and returned directly to ground is the first step. Note that the total inductance of the current loop form the V_{dd} pin to the V_{ss} pin is in series with the bypass capacitor. This inductance should be minimized. Consider that the impedance of a 0.125" wide trace above a ground plane on 0.062" FR-4 fiberglass circuit board is approximately 1.25 Ohms per inch at 10 MHz[1]. A milliampere of switching current through a one inch "ground" trace will produce 1.25 mV of noise across the "ground" trace, corresponding to a -45 dBm signal.

We have found the addition of a tantalum or aluminum electrolytic capacitor as a second bypass, located a little farther from the chip than the ceramic bypass provides additional suppression, but its effect is secondary to that of the ceramic capacitor. Inclusion of a decoupling network in the V_{dd} circuit is helpful, especially if the same power source is used for some of the RF circuitry.

The inclusion of small resistors in series with inputs and outputs attenuates high frequency noise which may couple onto printed circuit traces and decreases the Q of any traces which may be long enough to exhibit resonant properties (in conjunction with lumped capacitances in the circuit). The only OCI-100 lines which need to be routed to the radio area are the phase detector outputs, CHIPOUT, CHIPIN and TXEN. Most loop filter designs will accept series damping resistors of a few hundred ohms with no ill effects. Often the filter input resistor can be split into two parts, so the first resistor, combined with the trace stray capacitance, forms a low pass filter for the unwanted high frequency components.

Similarly, low pass filtering the CHIPOUT line will decrease the transmitter sidelobes and decrease unwanted coupling into other circuitry. Reducing the transmitter sidelobes (not its main lobe spread bandwidth) tends to increase the useful power in transmitted signal and reduces out of band emissions. The degree of filtering which is acceptable without excessively creating intersymbol interference depends on the overall system design. Usually, we find the use of a 2 or three pole filter with a rise time 20-30% of the chip time provides a simple, non-critical design.

If possible, keep the ASIC away from low level portions of the RF system, especially the IF amplifier inputs. Remember that because of integer ratio relationships between the clock frequency and the IF, there will be some spectral components of the OCI-100 digital noise at the IF. If it is necessary to place the OCI-100 near the IF amplifier, consider the possibility of including a metal shield between the two as part of the design. Use of a solid ground plane in the area of the OCI-100 is helpful in controlling noise. If the design doesn't permit a complete ground plane, at least include the low impedance grounding path as discussed above.

6.0 Operation with Digitized Voice

The OCI-100 bit rate capabilities make it a natural match to most popular voice rate digital coding schemes, allowing low cost realizations of digital cordless telephones, wireless microphones and other audio applications. The low complexity schemes for which there are readily available support ICs include pulse code modulation (PCM), continuously variable slope delta modulation (CVSD), and adaptive differential pulse code modulation (ADPCM). CVSD is particularly favorable for simple systems because all bits carry the same weight, and it is tolerant of single bit errors. CVSD can provide good quality telephone speech at 40 kb/s and communications quality speech at 16 kb/s. PCM uses unequal bit weights, so an error in a more significant bit sounds worse. Similar considerations apply to ADPCM.

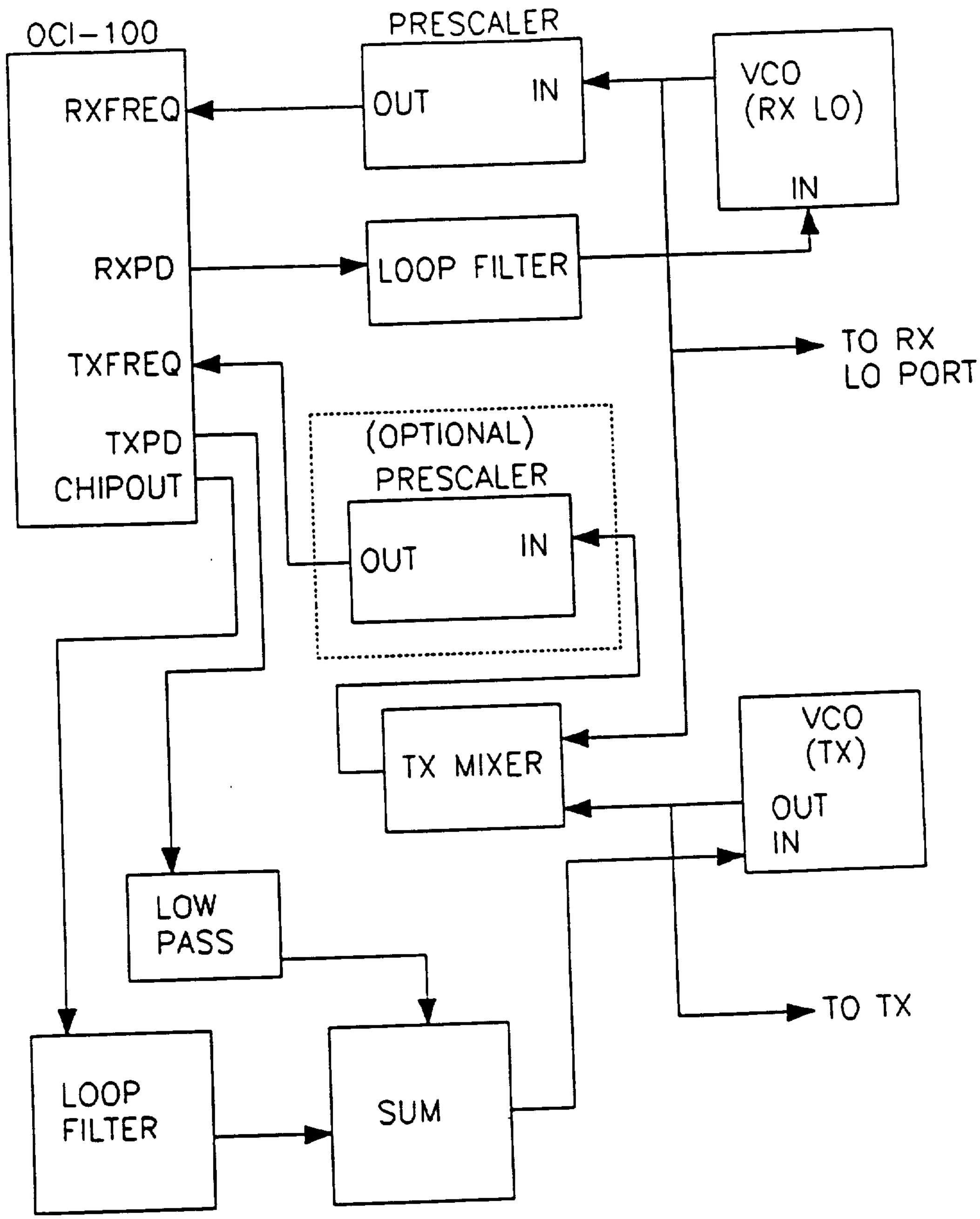
CVSD ICs are available from Motorola (MC3417/18 MC3517/18) and Harris (HC-55564). ADPCM ICs are available from Dallas Semiconductor and PCM coder-decoder integrated circuits are available from Texas

[1]I. Strauss, "Designing PC board for EMC Compliance", *Compliance Engineering*, Summer 1990, p18.

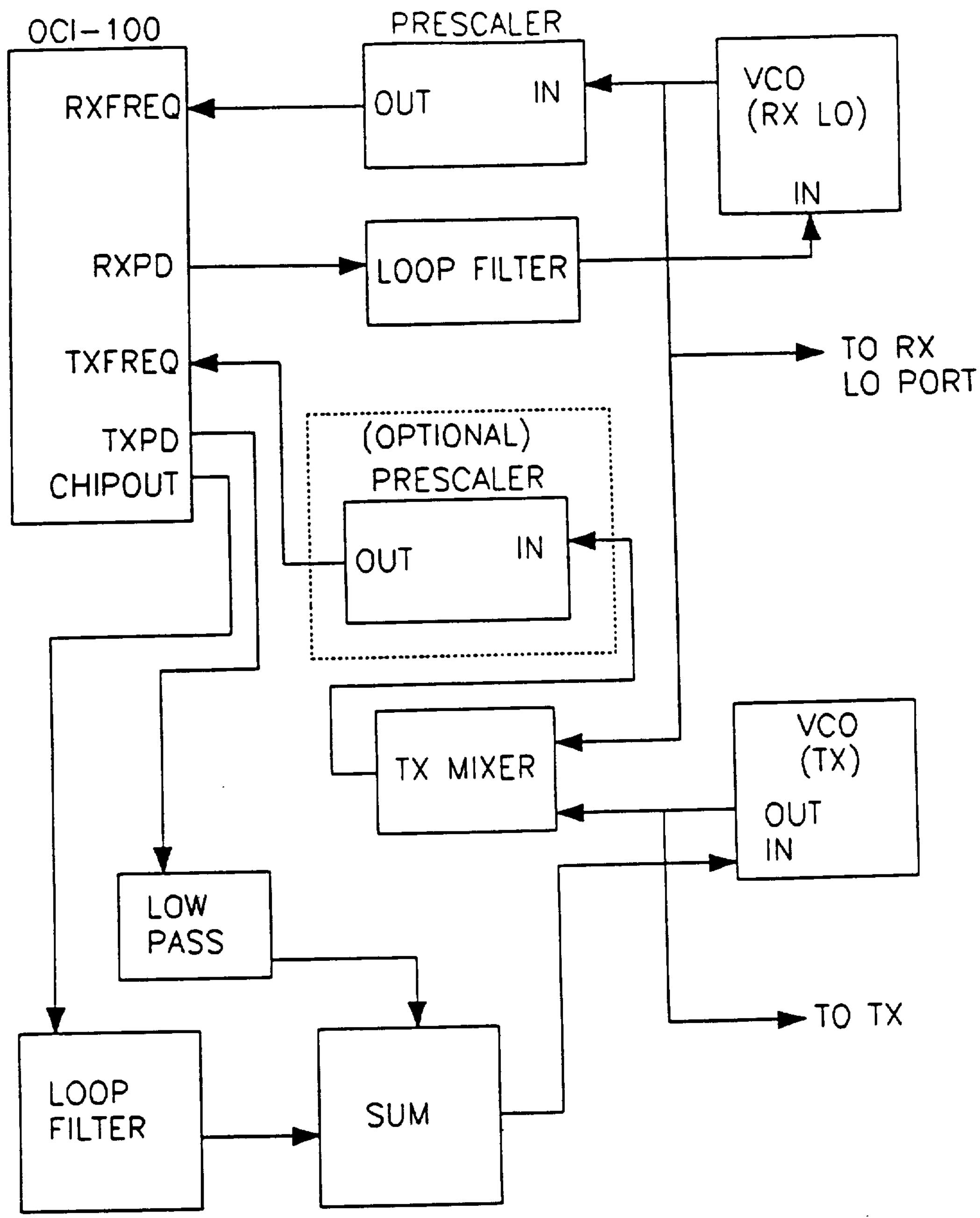
Instruments. An example of a CVSD to OCI-100 interface application is given in Figure 3.

These digitized voice systems have the advantages of privacy and superior sound quality (assuming bit errors have been controlled by proper link design, error control coding or the like). Unlike conventional analog cordless telephones, these signals will sound unintelligible if tuned in on a scanning radio. In fact, few scanning radios are able to "tune in" a spread spectrum signal because it exceeds the IF filter bandwidth of most (but not all) scanners.

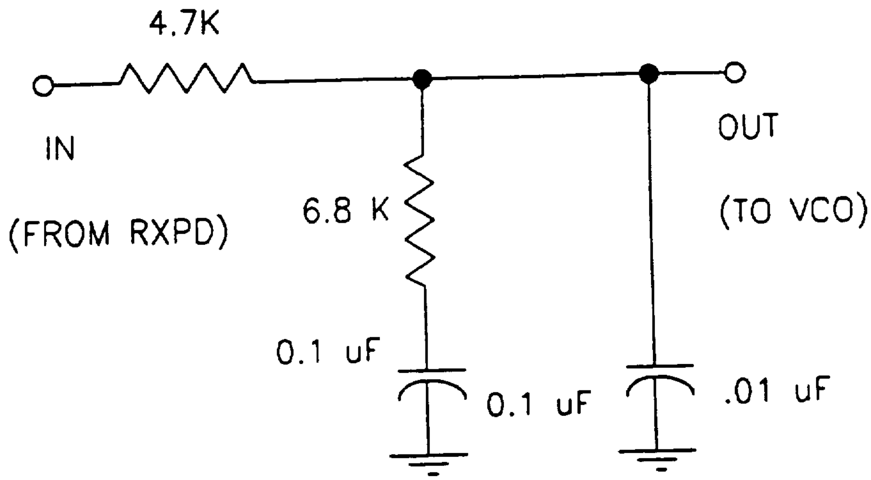
However, a more sophisticated eavesdropper could be expected to use a radio of the same spread spectrum design (probably by buying one of the product and modifying it if required). Therefore, the digital speech encoding should be augmented by proper encryption if protection from eavesdropping by a more sophisticated adversary is required.



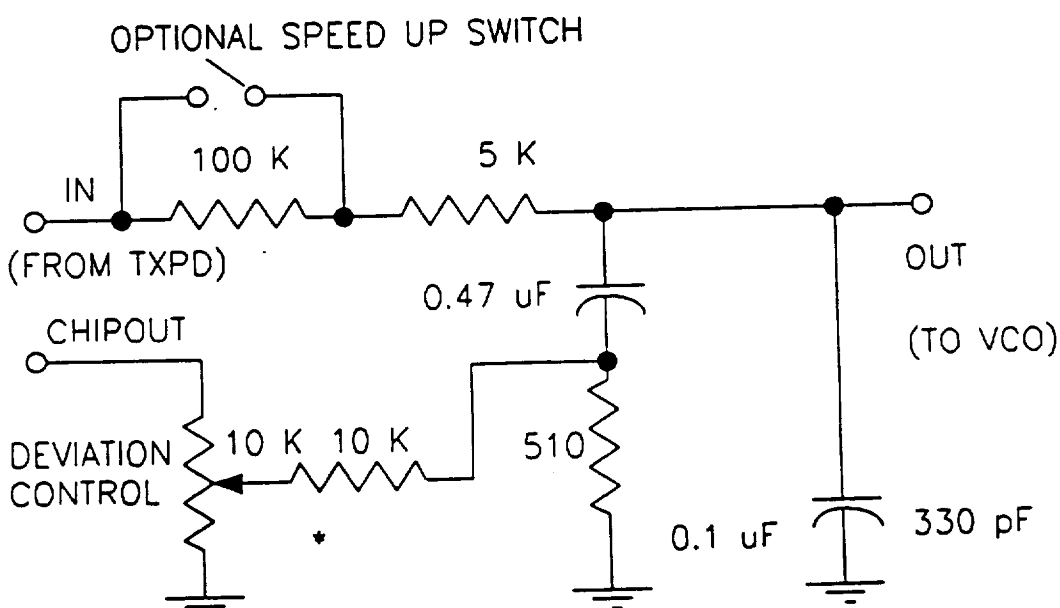
OCI-100 to Radio Interfacing,
Typical Block Diagram



OCI–100 to Radio Interfacing,
Typical Block Diagram

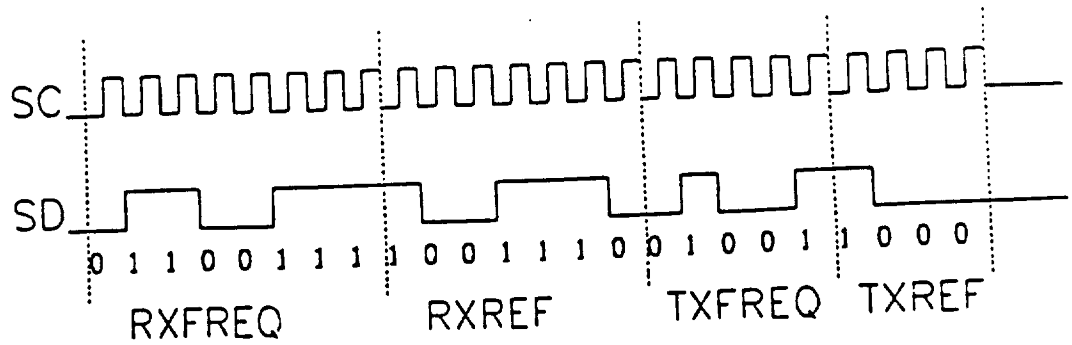


TYPICAL RECEIVER L.O. LOOP FILTER

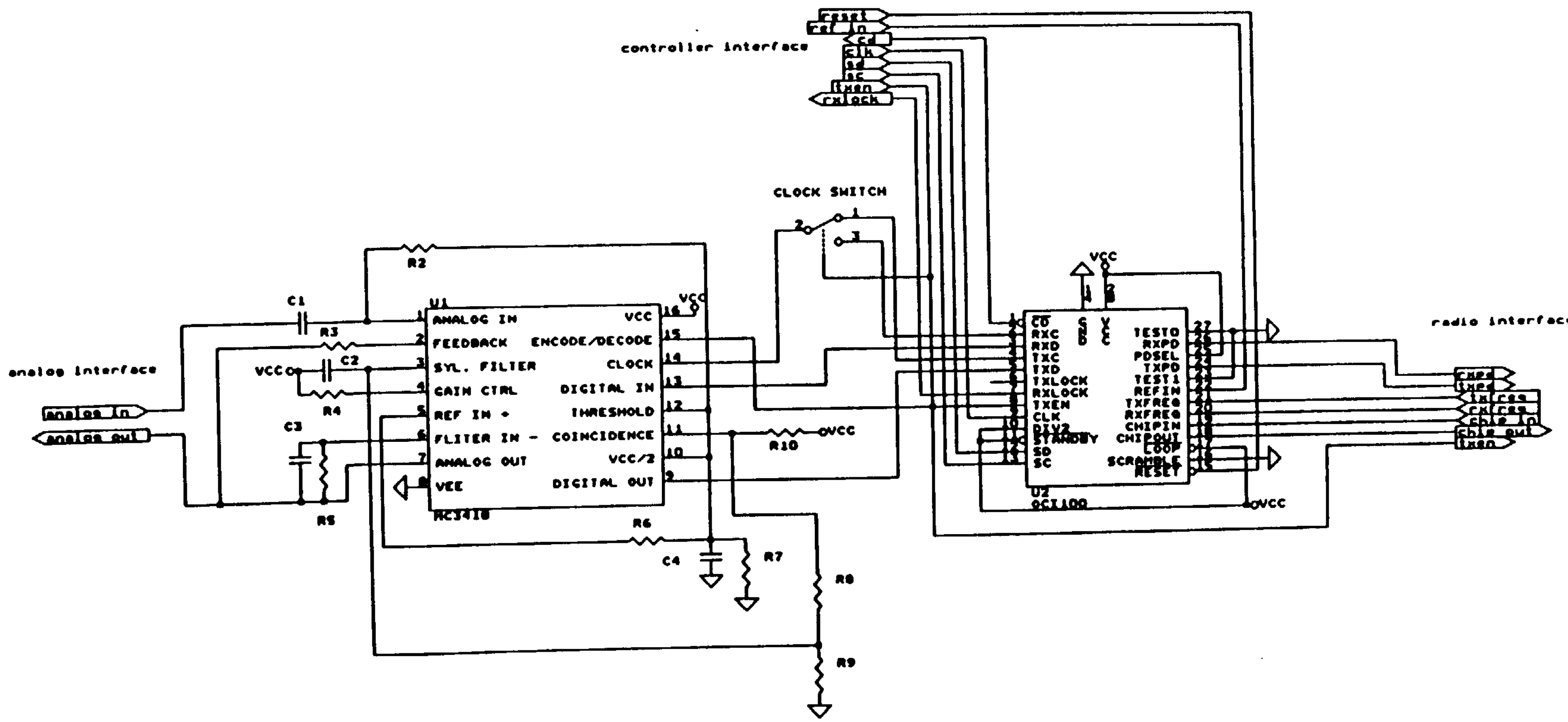


* SUBSTITUTE LOW PASS FILTER FOR THIS RESISTOR IF DESIRED FOR MORE FILTERING OF SIDELOBES

TYPICAL TRANSMITTER LOOP FILTER AND SUM



Frequency Synthesizer Load Waveforms



O'NEILL COMMUNICATIONS, INC
100 THANET CIRCLE
PRINCETON, NJ 08540

Title	OCI100-CVSD	
Size B	Document Number PRELIMINARY	REV 0
Date: November 19, 1990	Sheet 1 of 1	

SPECIFICATION
OCI Spread Spectrum RF Module

The RF Module is a complete spread spectrum data transceiver based on the RF technology used in OCI's successful LAWN product. These modules permit the rapid development and economical production of wireless digital communication systems that can be operated without need for licensing if the communication system complies with Part 15 of the FCC Rules (see note below on FCC rules).

The module provides all the radio and spread spectrum processing functions necessary to transmit and receive (half duplex) user bit streams. The user needs to supply only power, a crystal controlled clock signal, and a transmit enable signal.* The module is supplied in circuit board form.

Interface - Data Rates: 38 kb/s to 125 kb/s direct, depends on clock frequency, 0 to 38 kb/s can be accommodated by allowing bits to be transmitted multiple times
Note: For other rates, contact OCI.
Connector: 2 x 13 header, .025" sq. pins on .100" centers
Note: For other connectors, contact OCI.

Power Requirements - +5 Volts, regulated, current: receive mode 140 ma. nominal
transmit mode 170 ma. nominal
+8 to +14 Volts, filtered
current, receive mode 3.5 ma. nominal
current, transmit mode 65 ma. nominal

Antenna Port - 50 Ohms nominal impedance
Not damaged by any VSWR (operation may be degraded at high VSWR)
Connector: SMA standard, others by special order

RF Power Output - 100 mW nominal at 13 V dc supply (see above)

Receiver Sensitivity - -90 dBm for 10-3 BER at 38.4 kb/s

Dimensions - Circuit board, 3.8" x 4.5" x 0.8"

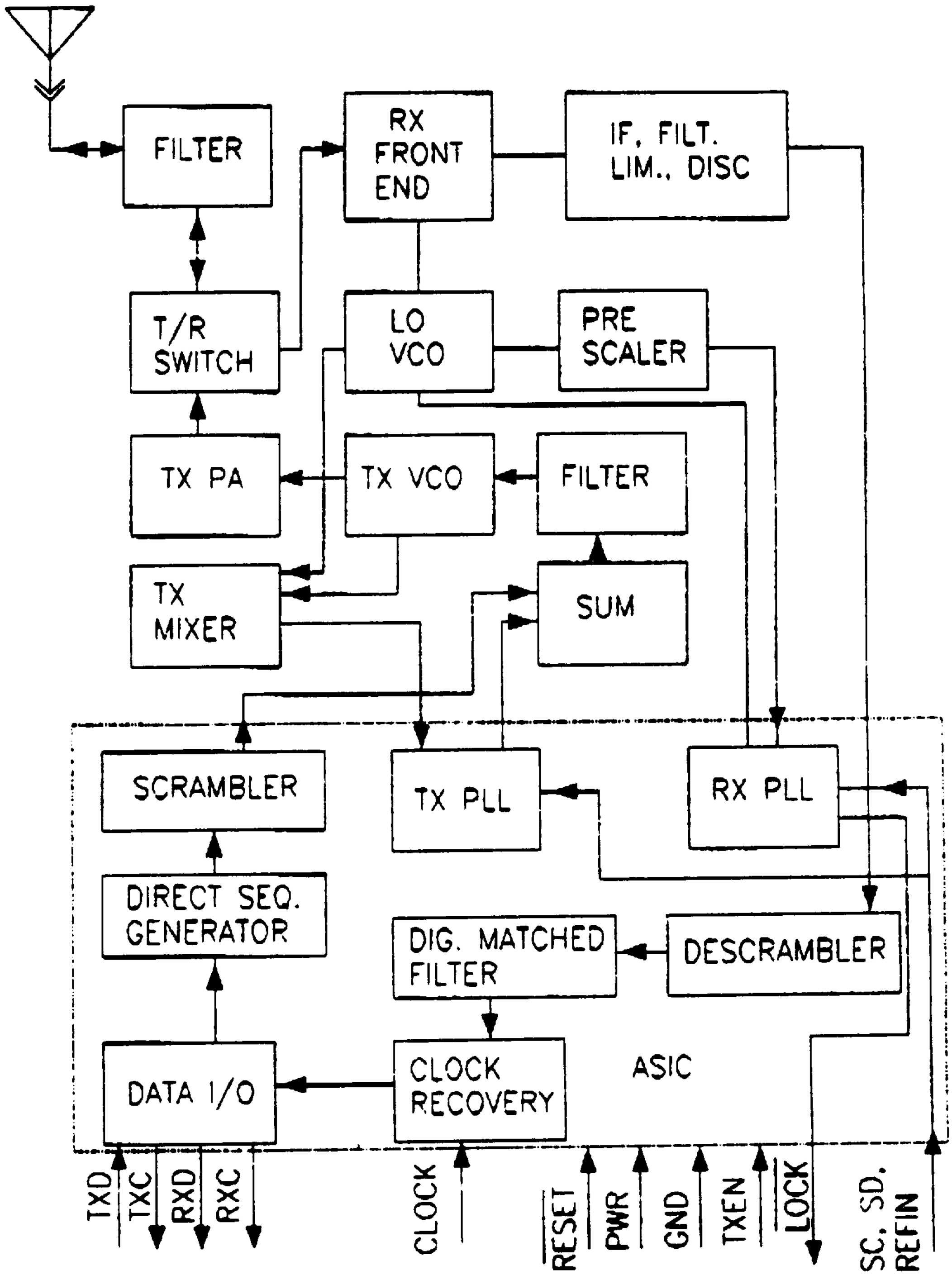
FCC Matters - Under section 15.101(e) of the Commission's Rules, this subassembly is not subject to certification under Part 15 by itself, but the equipment in which it is used must comply with the Rules.

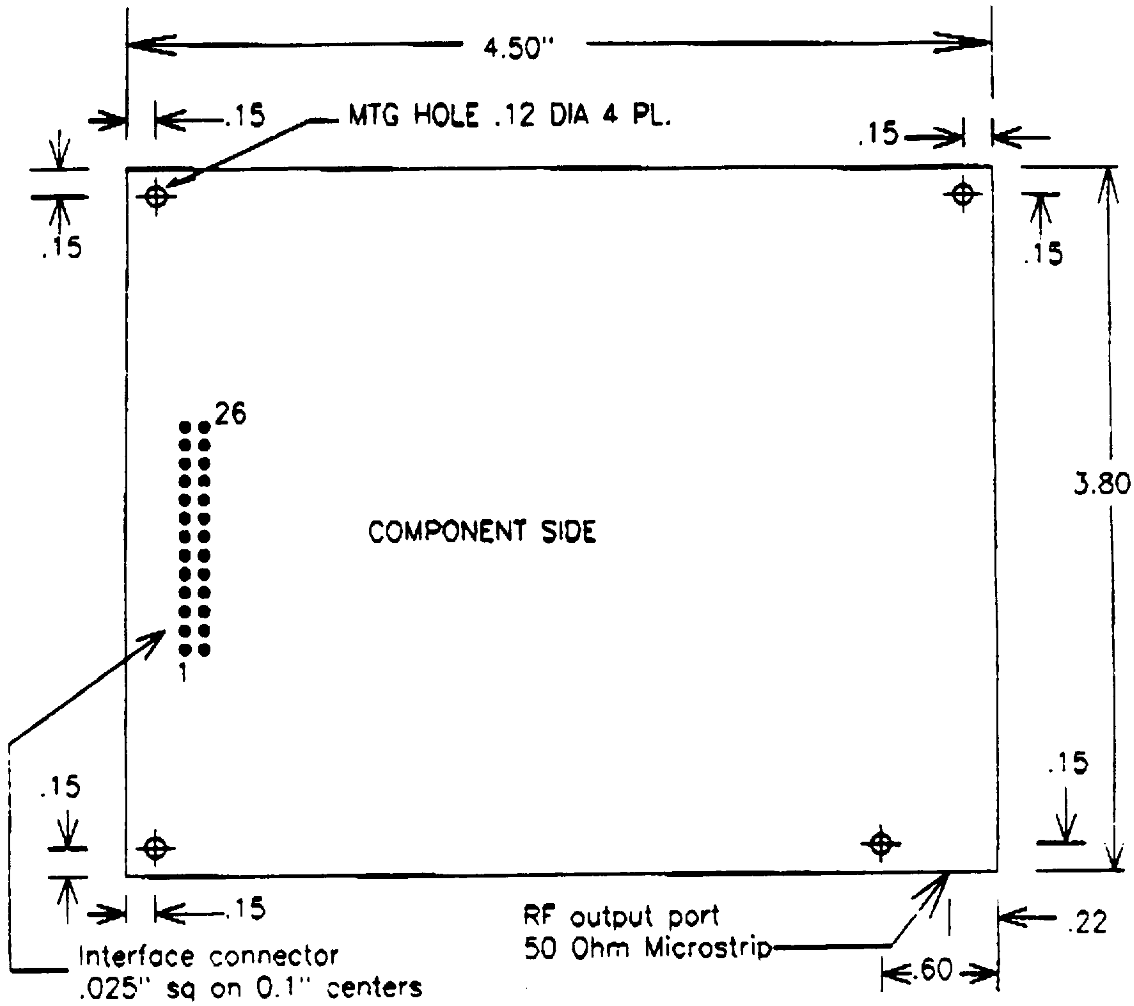
The RF Module is intended for inclusion by a manufacturer in his product. Since OCI has no control over the use and connection of the module, FCC compliance of the resulting product with the requirements of Part 15 cannot be guaranteed. OCI is available to assist manufacturers in achieving this certification. The RF properties of the module are known to be capable of meeting FCC Rules, because they are extracted from OCI product(s) that have been certified.

* FOR MORE DETAILED INFORMATION, REQUEST APPLICATION NOTE 300000.

310001 - Rev. C

O'Neill Communications, Inc.
100 Thanet Circle
Princeton, NJ 08540
(609) 497-6800

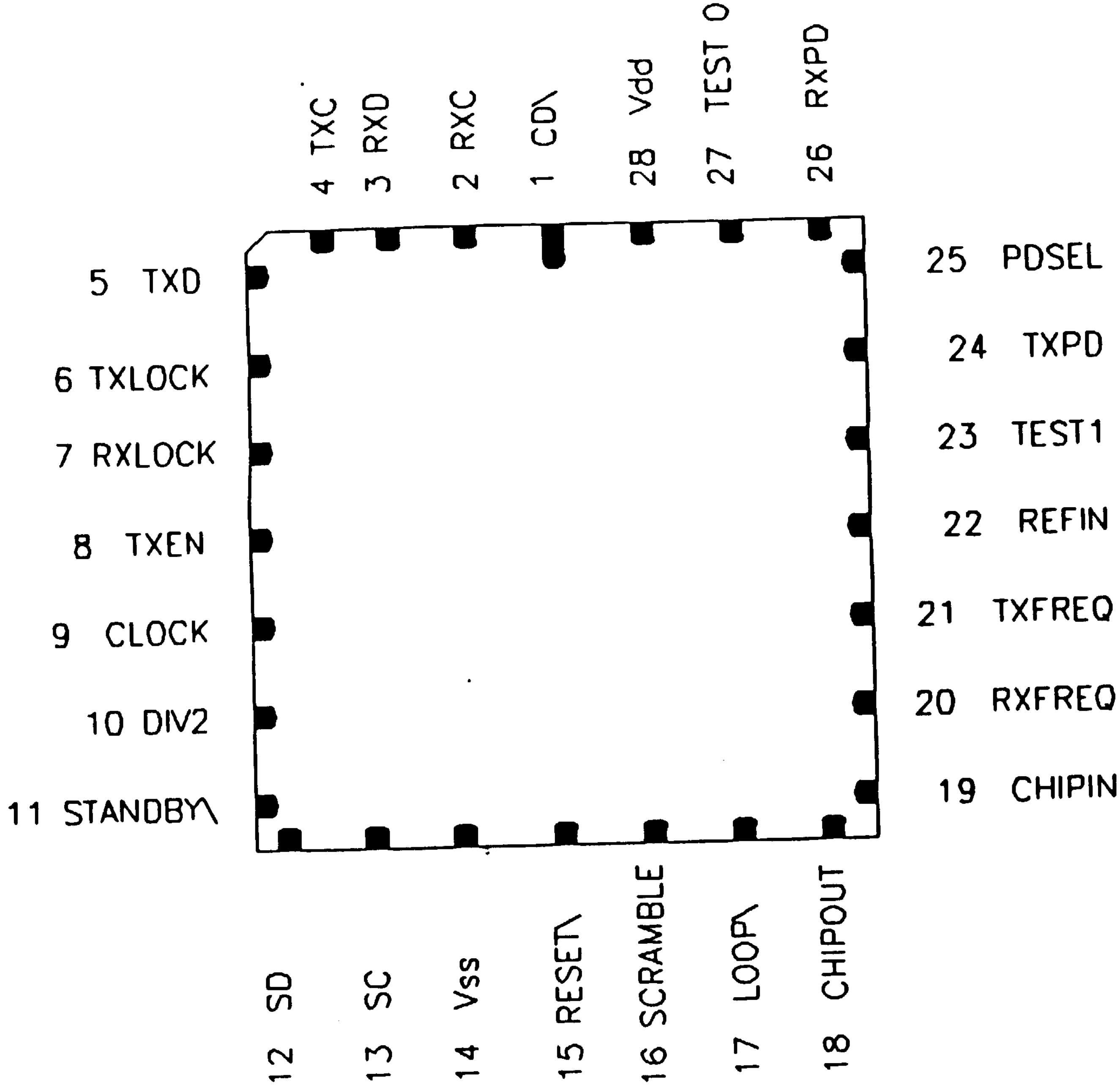

FILTER
RX FRONT END
IF, FILT. LIM., DISC
T/R SWITCH
LO VCO
PRE SCALER
TX PA
TX VCO
FILTER
TX MIXER
SUM
SCRAMBLER
TX PLL
RX PLL
DIRECT SEQ. GENERATOR
DIG. MATCHED FILTER
DESCRAMBLER
DATA I/O
CLOCK RECOVERY
ASIC
TXD
TXC
RXD
RXC
CLOCK
RESET
PWR
GND
TXEN
LOCK
SC, SD, REFIN



* * *

Although information in this data sheet has been carefully checked, no responsibility for inaccuracies can be assumed by OCI.

OCI reserves the right to make changes without further notice to any products herein to improve reliability, function or design. OCI does not assume any liability arising out of the application or use of any product or circuit described herein; neither does it convey any license under its patent rights nor the rights of others. OCI products are not authorized for use as components in life support devices or systems intended for surgical implant into the body or intended to support or sustain life. Buyer agrees to notify OCI of any such intended end use whereupon OCI shall determine availability and suitability of its product or products for the use intended.



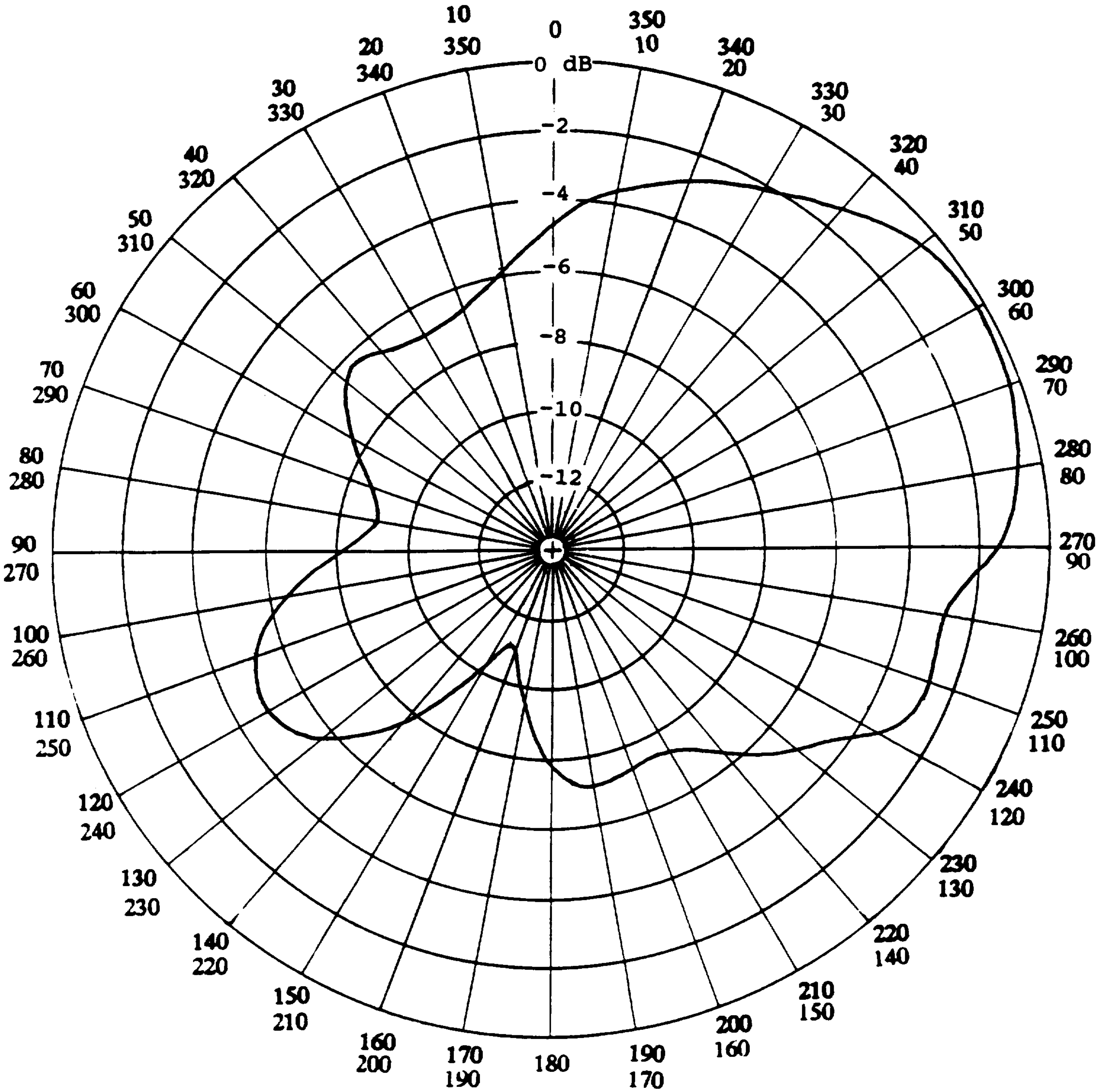
VIEWED FROM ABOVE

C-2

APPENDIX B

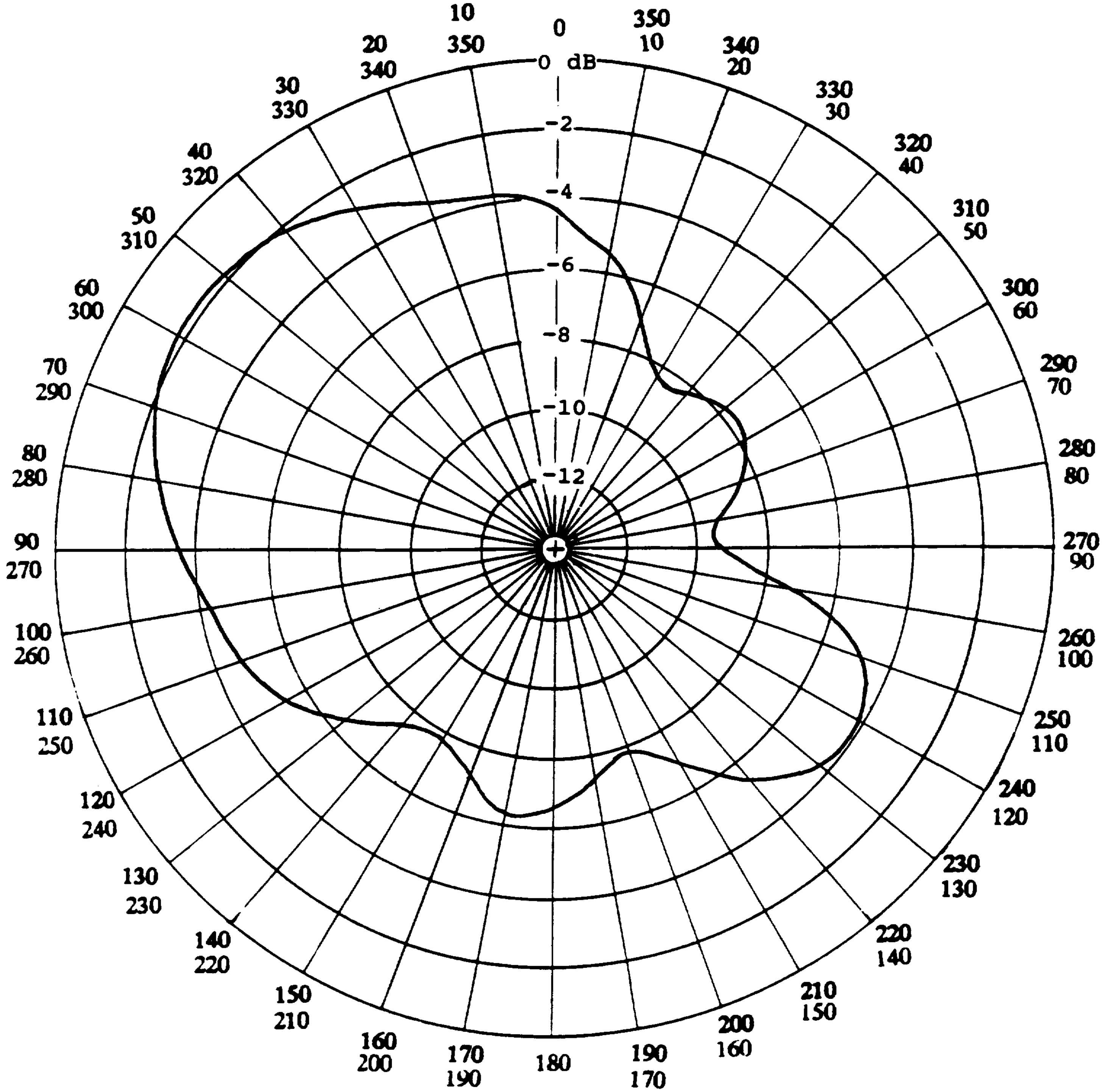
MEASURED ANTENNA PATTERNS

PATTERN #1



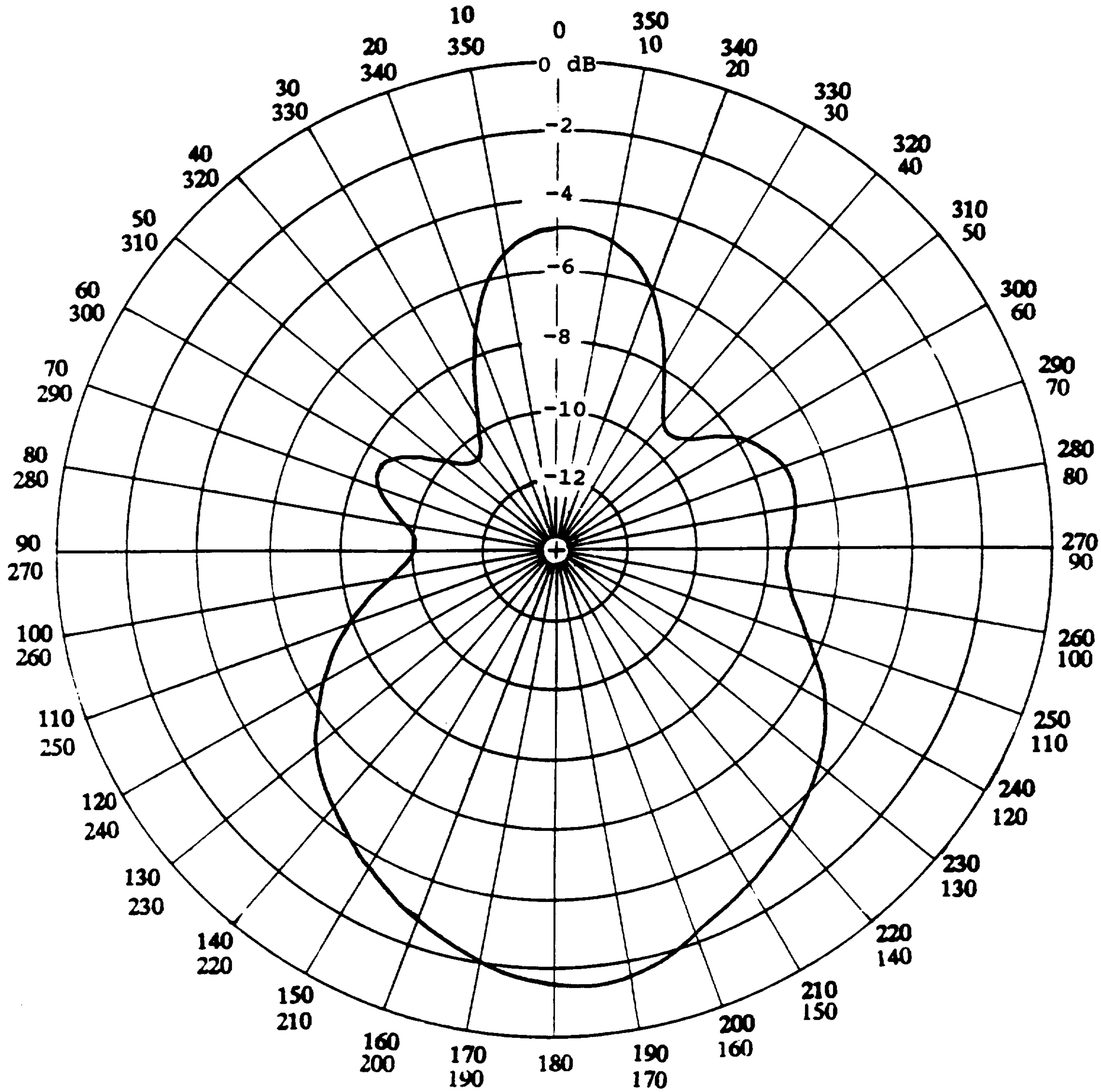
0
0 dB
-2
-4
-6
-8
-10
-12
10 350
20 340
30 330
40 320
50 310
60 300
70 290
80 280
90 270
100 260
110 250
120 240
130 230
140 220
150 210
160 200
170 190
180
190 170
200 160
210 150
220 140
230 130
240 120
250 110
260 100
270 90
280 80
290 70
300 60
310 50
320 40
330 30
340 20
350 10

PATTERN #2



0
0 dB
-2
-4
-6
-8
-10
-12
10 350
20 340
30 330
40 320
50 310
60 300
70 290
80 280
90 270
100 260
110 250
120 240
130 230
140 220
150 210
160 200
170 190
180
350 10
340 20
330 30
320 40
310 50
300 60
290 70
280 80
270 90
260 100
250 110
240 120
230 130
220 140
210 150
200 160
190 170

PATTERN #3



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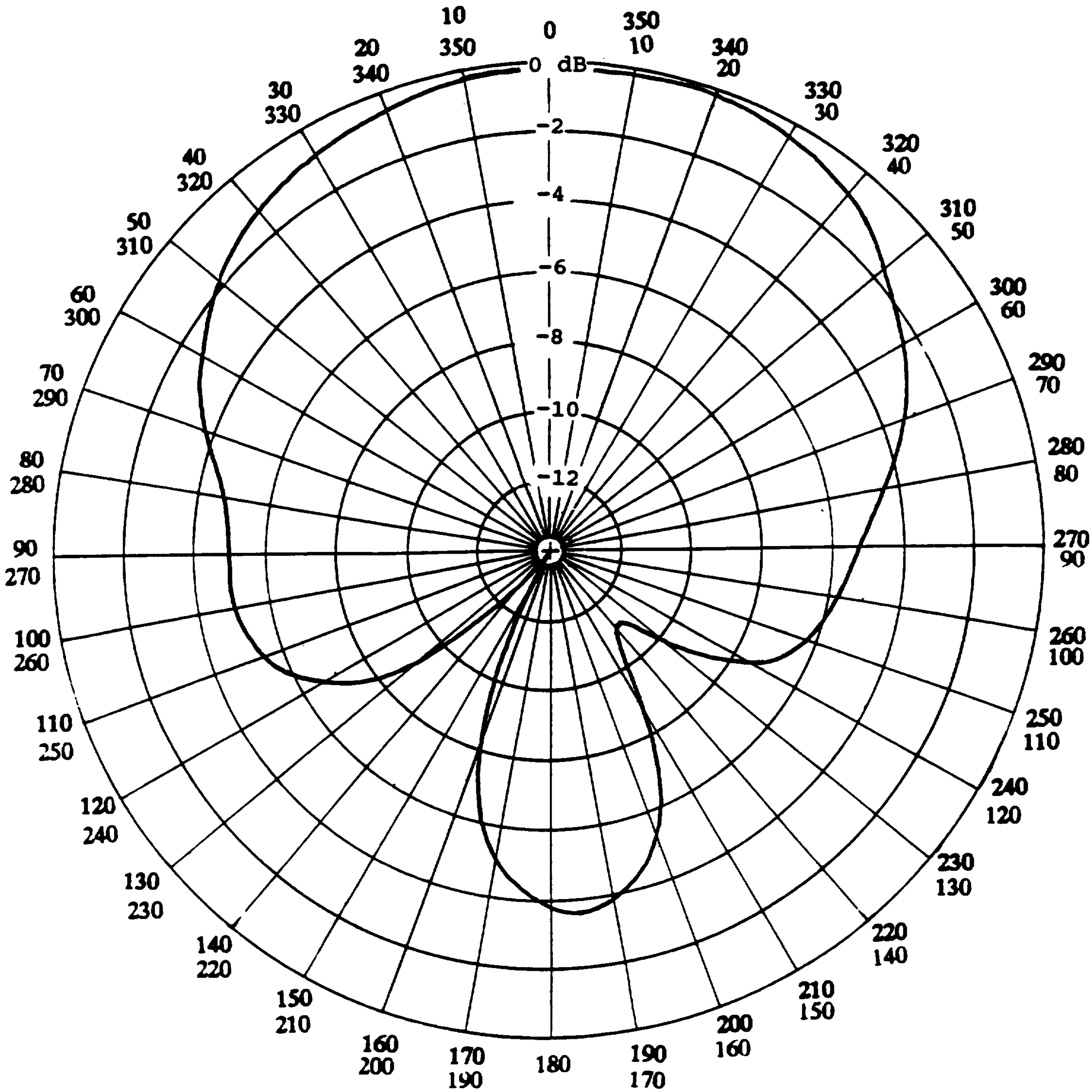
-10
-12

PATTERN #4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100
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260
100
110
250
250
110
120
240
240
120
130
230
230
130
140
220
220
140
150
210
210
150
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0 dB
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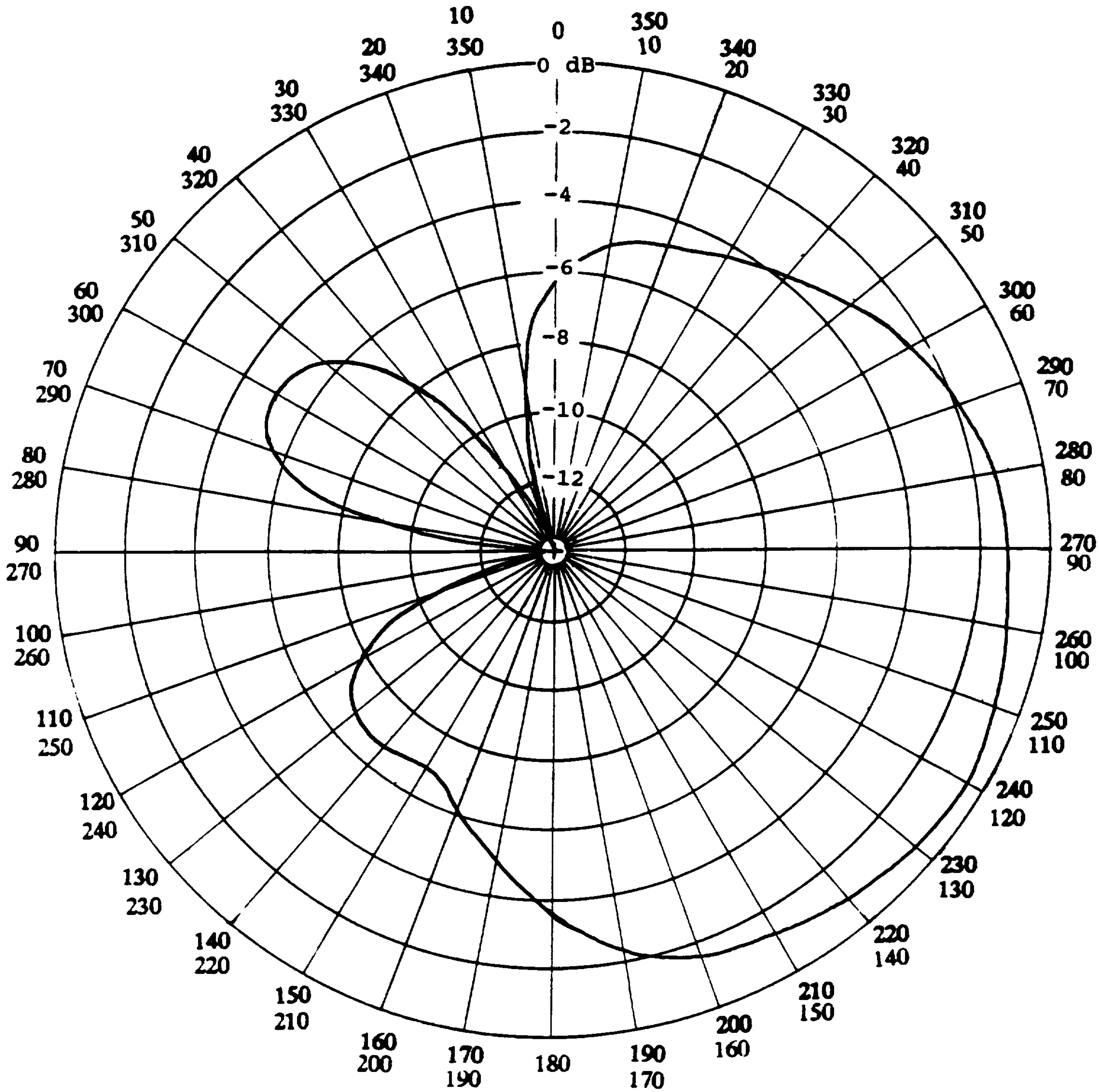
-6
-8
-10
-12

PATTERN #5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-8
-10
-12
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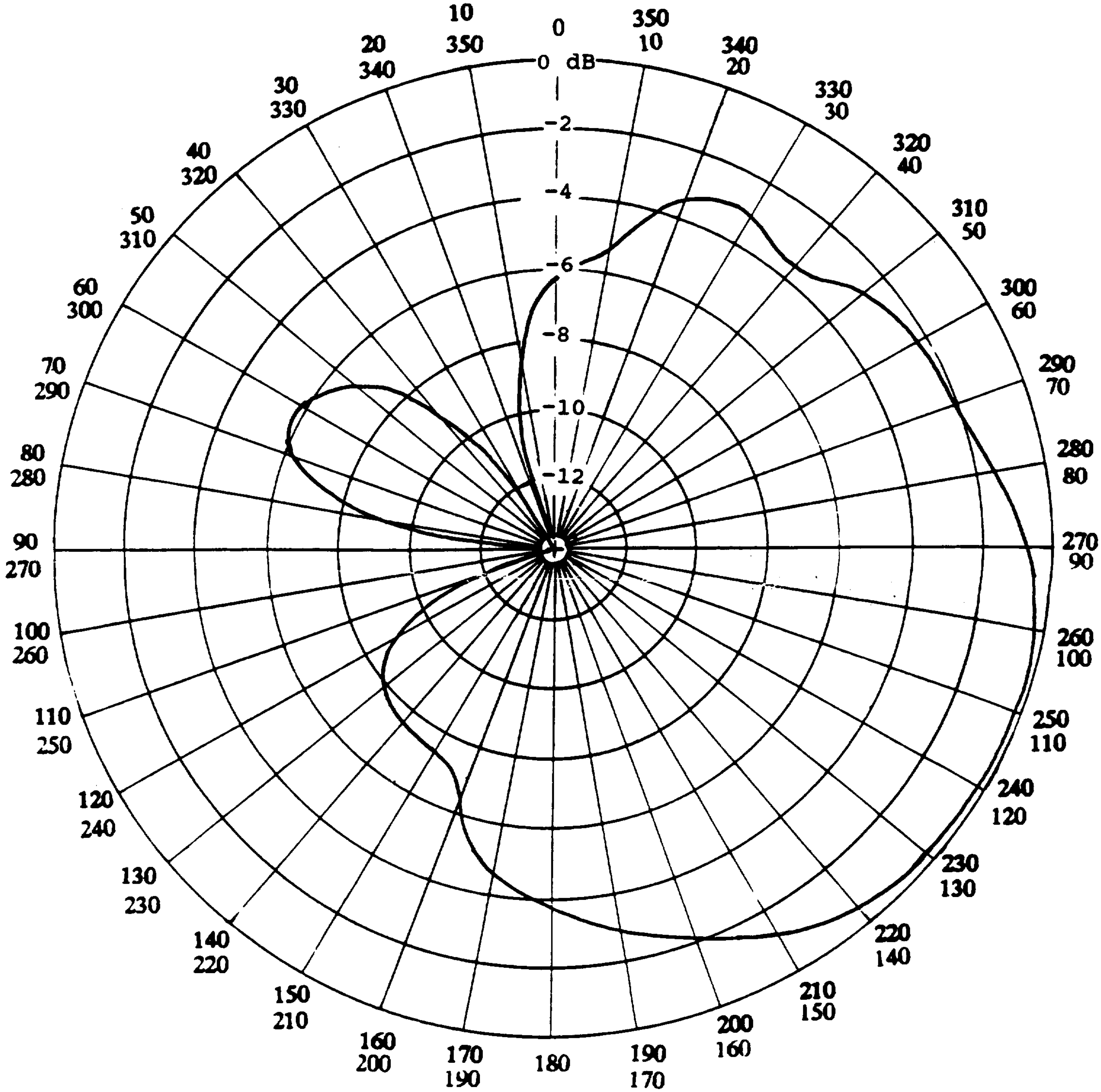
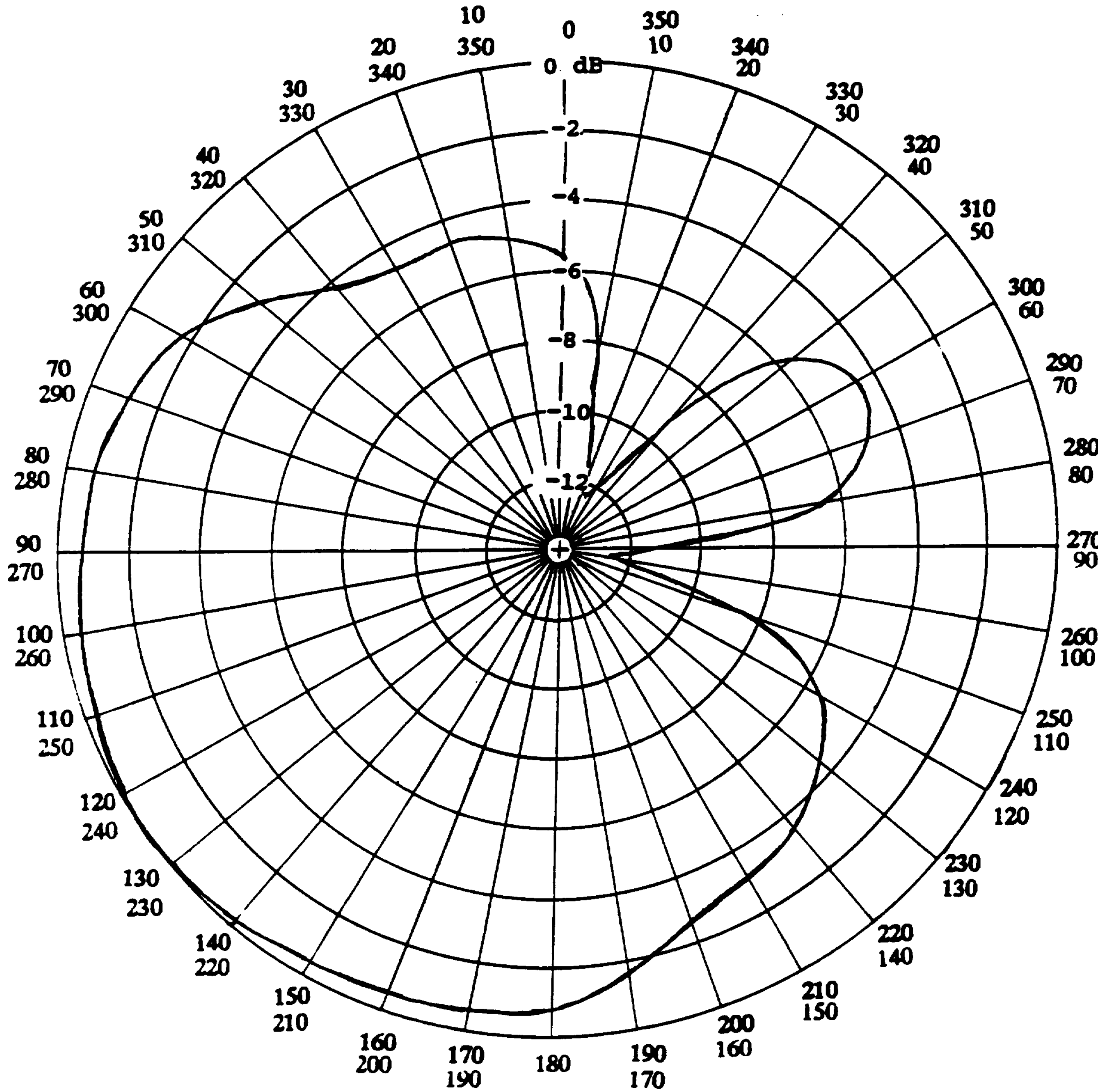
PATTERN #5
USING OMNIDIRECTIONAL SOURCE ANTENNA



PATTERN #6


0
0 dB
-2
-4
-6
-8
-10
-12
10
350
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330
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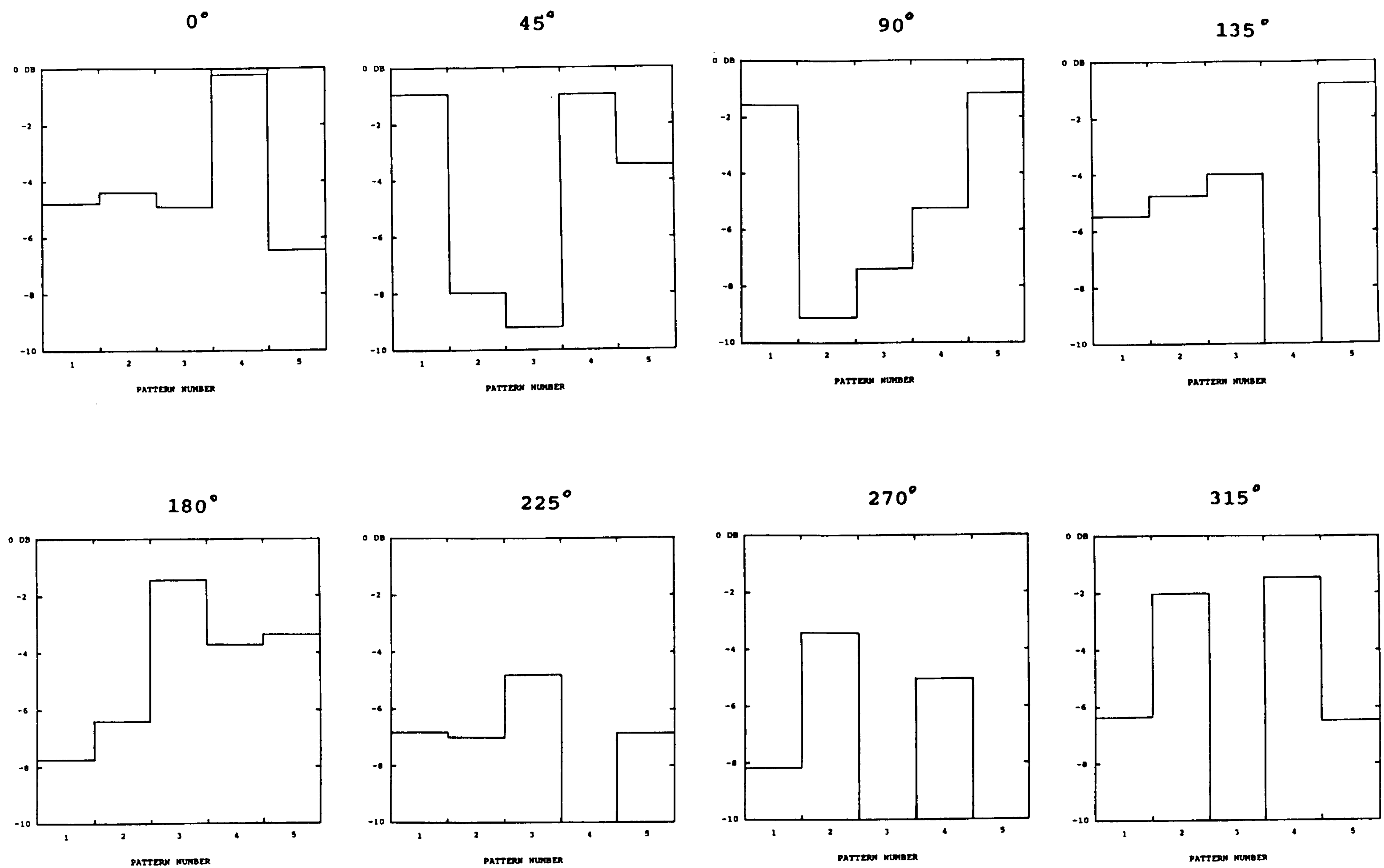


AMPLITUDE SEQUENCES VERSUS DIRECTION

ANTENNA PATTERN #1

ANGLE	GAIN DB	GAIN VOLTAGE
0	-5.8	.512
5	-4	.630
10	-3.8	.645
15	-3.2	.691
20	-3	.707
25	-2.5	.749
30	-2.2	.776
35	-2.2	.776
40	-1	.891
45	-.6	.933
50	-.3	.966
55	-.2	.977
60	-.2	.977
65	-.2	.977
70	-.3	.966
75	-.5	.944
80	-1	.891
85	-1.1	.881
90	-1.6	.831
95	-3	.707
100	-3.2	.691
105	-2.6	.741
110	-2.9	.716
115	-3.2	.691
120	-3.5	.668
125	-4.6	.588
130	-5.1	.555
135	-5.7	.518
140	-6.4	.478
145	-7.1	.441
150	-7.6	.416
155	-7.6	.416
160	-7.6	.416
165	-7.8	.407
170	-6.8	.457
175	-7.3	.431
180	-8	.398
185	-8.3	.384
190	-9.5	.334
195	-10.2	.309
200	-11	.281
205	-11	.281
210	-10	.316
215	-8.8	.363
220	-7.5	.421
225	-6.5	.473
230	-5.8	.512
235	-5.2	.549
240	-4.8	.575
245	-4.8	.575
250	-5	.562
255	-5.8	.512
260	-6.5	.473
265	-7.2	.436
270	-8.1	.393
275	-8.8	.363
280	-9	.354
285	-8.8	.363
290	-8.5	.375
295	-7.8	.407
300	-7.8	.407
305	-7	.446
310	-6.5	.473
315	-6.2	.489
320	-6.5	.473
325	-6.8	.457
330	-6.8	.457
335	-7	.446
340	-6.3	.484
345	-6	.501
350	-6	.501
355	-5.5	.530

ANTENNA PATTERN #2

ANGLE	GAIN DB	GAIN VOLTAGE
0	-4.2	.616
5	-5.2	.549
10	-5.4	.537
15	-6.2	.489
20	-7.4	.426
25	-8.099999	.393
30	-8.5	.375
35	-8.5	.375
40	-8.4	.380
45	-8.099999	.393
50	-8	.398
55	-7.9	.402
60	-8.099999	.393
65	-8.2	.389
70	-8.5	.375
75	-9.099999	.350
80	-9.7	.327
85	-9.8	.323
90	-9.2	.346
95	-8.2	.389
100	-7	.446
105	-6	.501
110	-4.8	.575
115	-4.2	.616
120	-4.2	.616
125	-4.2	.616
130	-4.4	.602
135	-4.7	.582
140	-5.5	.530
145	-6.3	.484
150	-7.2	.436
155	-7.7	.412
160	-8	.398
165	-7.4	.426
170	-7.2	.436
175	-6.5	.473
180	-6.5	.473
185	-6.4	.478
190	-6.4	.478
195	-6.7	.462
200	-7.2	.436
205	-7.4	.426
210	-7.7	.412
215	-8	.398
220	-7.7	.412
225	-7.2	.436
230	-6.5	.473
235	-6	.501
240	-5.4	.537
245	-5.2	.549
250	-4.8	.575
255	-4.4	.602
260	-4.3	.609
265	-4	.630
270	-3.4	.676
275	-4	.630
280	-2.5	.749
285	-2.2	.776
290	-2	.794
295	-1.8	.812
300	-1.7	.822
305	-1.7	.822
310	-1.8	.812
315	-2.1	.785
320	-2.2	.776
325	-2.4	.758
330	-2.7	.732
335	-3.1	.699
340	-3.4	.676
345	-3.3	.683
350	-3.5	.668
355	-3.9	.638

ANTENNA PATTERN #3

ANGLE	GAIN DB	GAIN VOLTAGE
0	-4.8	.575
5	-4.7	.582
10	-5.099999	.555
15	-5.4	.537
20	-6.3	.484
25	-7.099999	.441
30	-7.9	.402
35	-8.8	.363
40	-9.3	.342
45	-9.4	.338
50	-10.8	.288
55	-8.4	.380
60	-7.7	.412
65	-7.5	.421
70	-7.2	.436
75	-7	.446
80	-7.2	.436
85	-7.3	.431
90	-7.4	.426
95	-7.4	.426
100	-7.2	.436
105	-7	.446
110	-6.5	.473
115	-5.9	.506
120	-5.4	.537
125	-5	.562
130	-4.5	.595
135	-4.2	.616
140	-3.8	.645
145	-3.3	.683
150	-3.2	.691
155	-3	.707
160	-2.7	.732
165	-2.2	.776
170	-1.7	.822
175	-1.4	.851
180	-1.9	.803
185	-2	.794
190	-2.1	.785
195	-2.3	.767
200	-2.7	.732
205	-3.2	.691
210	-3.5	.668
215	-4	.630
220	-4.3	.609
225	-4.7	.582
230	-5.2	.549
235	-5.7	.518
240	-6.2	.489
245	-7.099999	.441
250	-7.8	.407
255	-8.599999	.371
260	-9.599999	.331
265	-10.1	.312
270	-10.2	.309
275	-10.1	.312
280	-9.7	.327
285	-9.099999	.350
290	-8.9	.358
295	-8.8	.363
300	-9.2	.346
305	-9.8	.323
310	-10.2	.309
315	-10.7	.291
320	-10.9	.285
325	-10.4	.301
330	-9.4	.338
335	-8.3	.384
340	-7.2	.436
345	-6.2	.489
350	-5.4	.537
355	-4.9	.568

ANTENNA PATTERN #4

ANGLE	GAIN DB	GAIN VOLTAGE
0	-.3	.966
5	-.1999999	.977
10	-.1999999	.977
15	-.3	.966
20	-.3	.966
25	-.3999999	.954
30	-.5999999	.933
35	-.8	.912
40	-.8999999	.901
45	-1.3	.860
50	-1.5	.841
55	-2.2	.776
60	-2.6	.741
65	-3.2	.691
70	-3.8	.645
75	-4.4	.602
80	-4.8	.575
85	-5.4	.537
90	-5.7	.518
95	-5.9	.506
100	-6.1	.495
105	-6.3	.484
110	-6.400001	.478
115	-7.1	.441
120	-8.1	.393
125	-9.3	.342
130	-10.7	.291
135	-11.3	.272
140	-11.3	.272
145	-10.3	.305
150	-8.3	.384
155	-6.6	.467
160	-5.5	.530
165	-4.7	.582
170	-4.3	.609
175	-4.1	.623
180	-4.2	.616
185	-4.5	.595
190	-5.3	.543
195	-6.3	.484
200	-8.3	.384
205	-11.3	.272
210	-13.6	.208
215	-13.8	.204
220	-13.8	.204
225	-11.3	.272
230	-9.1	.350
235	-7.8	.407
240	-6.8	.457
245	-6.1	.495
250	-5.8	.512
255	-5.5	.530
260	-5.5	.530
265	-5.5	.530
270	-5.5	.530
275	-5.4	.537
280	-5.1	.555
285	-4.5	.595
290	-4.1	.623
295	-3.4	.676
300	-3.3	.683
305	-2.6	.741
310	-2.3	.767
315	-2	.794
320	-1.5	.841
325	-1.3	.860
330	-1.2	.870
335	-1	.891
340	-.8	.912
345	-.5999999	.933
350	-.3999999	.954
355	-.3	.966

ANTENNA PATTERN #5

ANGLE	GAIN DB	GAIN VOLTAGE
0	-6.5	.473
5	-5.7	.518
10	-5.3	.543
15	-5	.562
20	-4.9	.568
25	-4.9	.568
30	-4.5	.595
35	-4.2	.616
40	-4	.630
45	-3.6	.660
50	-3.1	.699
55	-2.8	.724
60	-2.5	.749
65	-2.1	.785
70	-2	.794
75	-1.7	.822
80	-1.2	.870
85	-1.2	.870
90	-1.2	.870
95	-1.2	.870
100	-1	.891
105	-.9000001	.901
110	-.8	.912
115	-.5	.944
120	-.5	.944
125	-.5999999	.933
130	-.7	.922
135	-.8	.912
140	-.9000001	.901
145	-1	.891
150	-1	.891
155	-1.3	.860
160	-1.6	.831
165	-1.7	.822
170	-2.1	.785
175	-2.5	.749
180	-3.3	.683
185	-4	.630
190	-4.6	.588
195	-5.4	.537
200	-6	.501
205	-6.3	.484
210	-6.5	.473
215	-6.5	.473
220	-6.3	.484
225	-6.5	.473
230	-6.5	.473
235	-7	.446
240	-8	.398
245	-9.3	.342
250	-11.5	.266
255	-13.5	.211
260	-14.5	.188
265	-14.5	.188
270	-12.5	.237
275	-10	.316
280	-8.3	.384
285	-6.6	.467
290	-5.5	.530
295	-5.2	.549
300	-5	.562
305	-5.3	.543
310	-5.8	.512
315	-6.7	.462
320	-8.2	.389
325	-10.2	.309
330	-12.5	.237
335	-13.5	.211
340	-12.7	.231
345	-10.7	.291
350	-9	.354
355	-7.7	.412

ANTENNA PATTERN #6

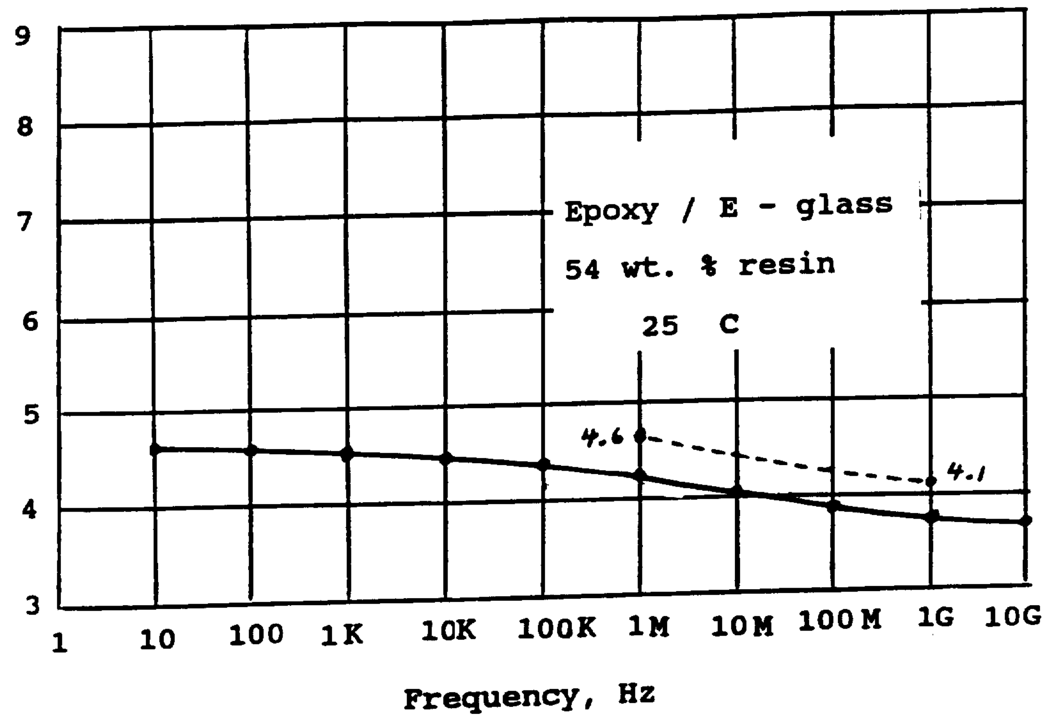
ANGLE	GAIN DB	GAIN VOLTAGE
0	-5.7	.518
20	-12.1	.248
40	-9.5	.334
60	-4.5	.595
80	-6.5	.473
100	-12.5	.237
120	-5.7	.518
140	-3.5	.668
160	-2.5	.749
180	-.7	.922
200	-.5	.944
220	-.1	.988
240	0	1
260	-.5	.944
280	-.7	.922
300	-1.7	.822
320	-4.1	.623
340	-4.5	.595

APPENDIX C

CALCULATIONS AND TESTS FOR MICROSTRIP LINE

The value of dielectric constant of the FR-4 epoxy fiberglass substrate at 910 mHz could not be found in the available literature. The value of 4.6 at 1.0 mHz is given in the table on the following page, together with a graph which shows the variation with frequency for a 0.005-inch thick substrate. The 4.6 value taken from the table was superimposed on the graph at the 1 mHz frequency and an offset curve was used to approximate the value for the 0.059-inch thickness at 910 mHz. The value so obtained is 4.1.

This curve is for the 0.005 - inch core thickness



General Information

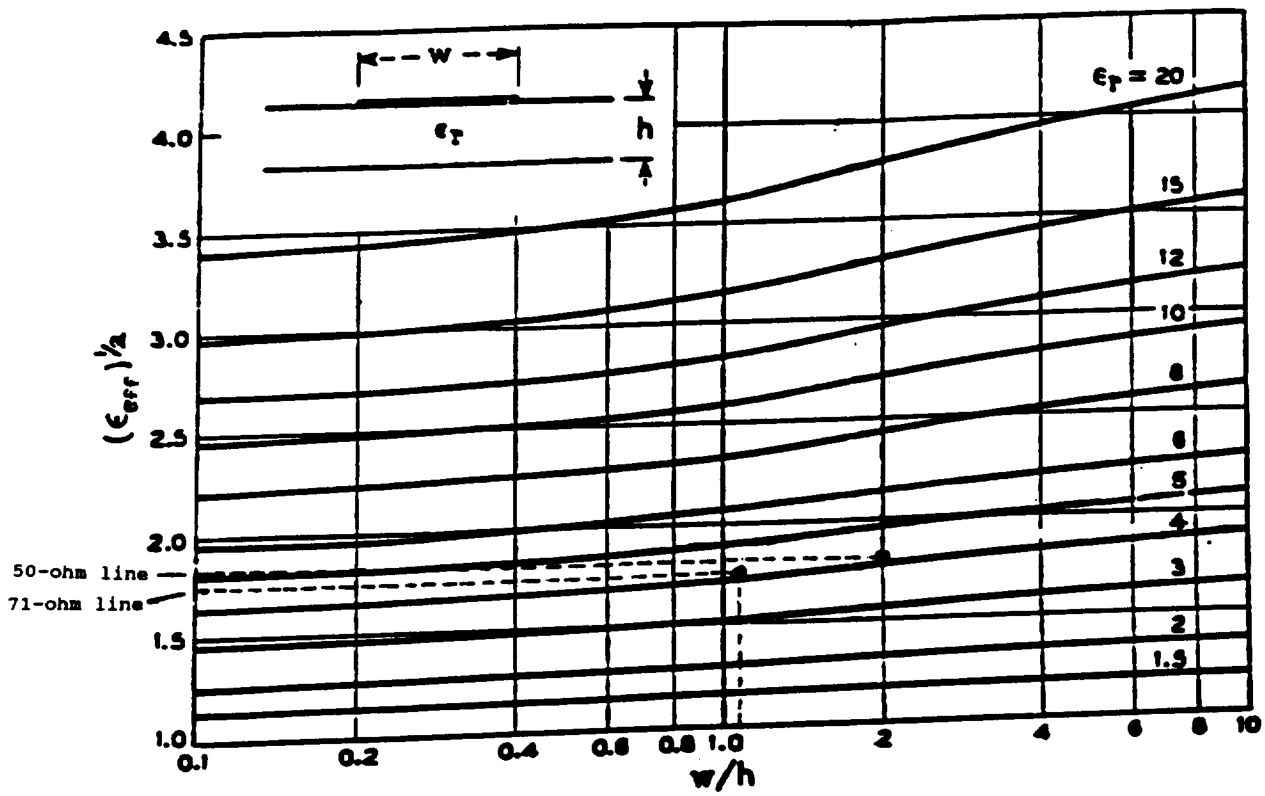
ED-130

NEMA Designation	FR-4
MIL-P-13949G Designation	GFN
U.L. Recognition: Plastics	E37002
Logo (upon request)	Material over .031
Thicknesses	.0035" to 0.125"

Electrical Properties	**Nominal Values 0.005" Core**	**Nominal Values 0.059" Overall**
Dielectric Constant[2] @ 1MHz (Permittivity)	4.2	4.6
Dissipation Factor @ 1MHz	0.020	0.020
Volume Resistivity		
Humidity	10^6 megohm-cm	10^8 megohm-cm
At Elevated Temperature	10^8 megohm-cm	10^6 megohm-cm
Surface Resistivity		
Humidity	——	10^4 megohm
At Elevated Temperature	10^8 megohm	10^8 megohm
Arc Resistance	120 seconds	100 seconds
ielectric Breakdown	——	>55 kV
lectric Strength	900	——

[1] Dimensional Stability is dependent upon laminate construction.
[2] Dielectric Constant is dependent upon resin content.

Only a part of the electric field between the line and the ground plane passes through the dielectric material of the substrate. Therefore, the effective dielectric constant is less than the dielectric constant of the substrate material. The set of curves shown below reveal the manner in which the effective dielectric constant varies with the ratio w/h.



The values of effective dielectric constant to be used in the design of the microstrip circuits was calculated as follows:

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2\sqrt{1 + \frac{10}{w/h}}}$$

The values of w/h that were used in the equation were obtained by use of program MSTRIP.BAS. The results are shown on the following page.

CALCULATION OF PARAMETERS FOR HYBRID RING

71-OHM LINE FOR MAIN RING:

W/H? 1.05

DIELECTRIC CONSTANT OF SUBSTRATE? 4.1
EFFECTIVE DIELECTRIC CONSTANT = 3.027799

FREQUENCY IN MEGAHERTZ? 910

MICROSTRIP WAVELENGTH =	18.94595	CENTIMETERS
OR	7.459035	INCHES

ZO = 71.01374

SUBSTRATE THICKNESS? .057
STRIP WIDTH = .05985"

50-OHM LINE FOR INPUT AND OUTPUT:

W/H? 2.0

DIELECTRIC CONSTANT OF SUBSTRATE? 4.1
EFFECTIVE DIELECTRIC CONSTANT = 3.182785

FREQUENCY IN MEGAHERTZ? 910

MICROSTRIP WAVELENGTH =	18.4789	CENTIMETERS
OR	7.275159	INCHES

ZO = 50.12747

SUBSTRATE THICKNESS? .057
STRIP WIDTH = .114"

```
REM       **************** PROGRAM MSTRIP.BAS ***************************
REM
REM       THIS PROGRAM COMPUTES THE CHARACTERISTIC IMPEDANCE OF
REM       MICROSTRIP TRANSMISSION LINE AS A FUNCTION OF W/H
REM       AND DIELECTRIC CONSTANT OF THE SUBSTRATE
REM
          CLS
          INPUT "W/H"; WH
          PRINT
          INPUT "DIELECTRIC CONSTANT OF SUBSTRATE"; EPS
REM
REM       ******** CALCULATE EFFECTIVE DIELECTRIC CONSTANT
REM
          A = (EPS + 1) / 2
          B = (EPS - 1) / 2
          C = SQR(1 + 10 / WH)
          EPSEFF = A + B / C
          SQTEPS = SQR(EPSEFF)
          PRINT "EFFECTIVE DIELECTRIC CONSTANT = "; EPSEFF
          PRINT
REM       ************ CALCULATE MICROSTRIP WAVELENGTH ********************
          INPUT "FREQUENCY IN MEGAHERTZ"; FREQ
          LAMDA = 30000 / FREQ
          LAMEFF = LAMDA / SQR(EPSEFF)
          INCHES = LAMEFF / 2.54
          PRINT "MICROSTRIP WAVELENGTH = "; LAMEFF, "CENTIMETERS"
          PRINT "                        OR  "; INCHES, "INCHES"
          PRINT
REM
REM       ********* CALCULATE CHARACTERISTIC IMPEDANCE FOR AIR ***********
REM
          IF WH > 1 GOTO 10
          ZO = 60 * LOG((8 / WH) + WH / 4)
          GOTO 20
10        ZO = 377 / (WH + 2.42 - .44 / WH + (1 - 1 / WH) ^ 6)
REM
REM       **DIVIDE BY THE SQUARE ROOT OF THE EFFECTIVE DIELECTRIC CONSTANT **
REM
20         ZO = ZO / SQTEPS
          PRINT "ZO = "; ZO
          PRINT
          INPUT "SUBSTRATE THICKNESS"; THK
          W = WH * THK
          PRINT "STRIP WIDTH = "; W
          END
```

TEST OF LOSS FOR 50-OHM MICROSTRIP LINE
APRIL 29, 1992

A 9-inch length of microstrip line was tested for loss at 910 mHz. A Hewlett Packard 614A signal generator and a Hewlett Packard 5342A frequency counter / power meter were used to perform the test. The output of the signal generator was initally applied, through two short cables to the power meter and the output power level was set to 0 dBm. The cables were then separated at the midpoint and the microstrip line was inserted. The power was again measured. The measured power was -0.9 dBm. The loss through the 9-inch length of microstrip line was thus 0.9 dB. The loss per inch is therefore 0.1 dB.

CALCULATION OF LOSS FOR 50-OHM MICROSTRIP LINE
APRIL 30, 1992

A computation of the loss of a 9-inch long microstrip line using 0.057-inch thick fiberglass/epoxy substrate was made. The equation that was used for computation of the loss in the dielectric substrate was that shown on page 112 of "Foundations of Microstrip Design" by Terry Edwards. The equation is

$$\alpha_d = 27.3\frac{\epsilon_r(\epsilon_{eff} - 1)\tan\delta}{\sqrt{\epsilon_{eff}}(\epsilon_r - 1)\lambda_0} \quad \text{dB/unit length}$$

The value of effective dielectric constant was previously calculated as 3.18, based on a dielectric constant of 4.1. The loss tangent of 0.02 was extracted from the literature. The free-space wavelength at 910 mHz is 0.3297 meter.

Using these parameter values,

$$\alpha_d = \frac{(27.3)(4.1)(2.18)(0.02)}{3.18\ (3.1)(0.330)} = 2.68\ dB/m.$$

Multiplying by 0.229 meter (9 inches), the loss for the 9-inch length is

$$(2.68)(0.229) = 0.614\ dB.$$

To calculate the loss in the copper line, an equation on the same page of the reference was used:

$$\alpha_c = 0.072\frac{\sqrt{f}}{wZ_0}\lambda_g \quad \text{dB/microstrip wavelength}$$

(f is in gigahertz and Z_0 is in ohms.)

Substituting the values of w, frequency and microstrip wavelength,

$$\alpha_c = \frac{(0.072)\sqrt{0.91}\ (0.185)}{(.00290)(50.0)} = 0.0876$$

Assuming surface roughness to increase the attuation by 60 percent (estimate from reference),

$$(1.60)(0.0876) = 0.1402$$

Multiplying by the length of the line in wavelengths,

$$(0.1402)\ \frac{0.229}{0.185} = 0.174$$

Adding the losses from conductor and dielectric,

$$\begin{array}{r} 0.614 \\ 0.174 \\ \hline 0.788 \end{array} \text{ dB total}$$

(measured value = 0.9 dB.)

4/30/92

APPENDIX D

TESTS OF AMPLITUDE/PHASE CONTROLLER (SINGLE UNIT)

TEST OF HYBRID-RING AMPLITUDE/PHASE CONTROL CIRCUIT

APRIL 8, 1992

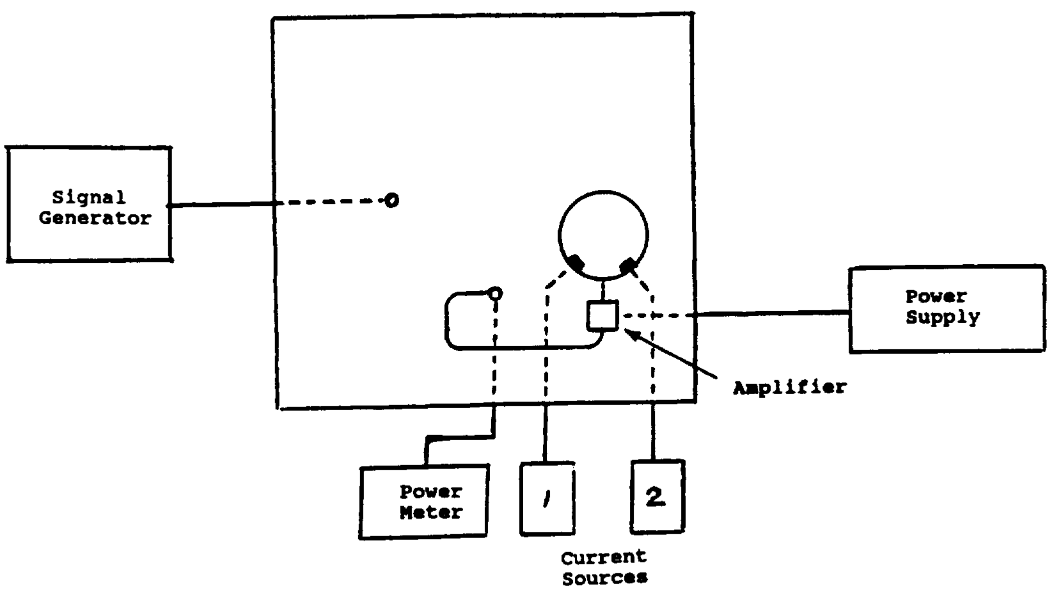
The circuit was activated by providing dc control currents to the two mixer attenuators and by putting a 900-mHz CW signal through the unit. The individual control currents were varied and the resulting output power levels were measured by observing them on a spectrum analyzer. The measured data are shown below.

CONTROL CURRENT milliamperes		INPUT CONTROL VOLTS		VOLTAGE ACROSS MIXER INPUT		OUTPUT POWER dBm
A (red)	B (green)	A	B	A	B	
17	17	8.34	8.97	0.31	0.31	0 (ref.)
10	17	4.94	8.98	0.18	0.31	0
8	17	3.87	8.98	0.26	0.31	-0.5
6	17	2.99	8.98	0.24	0.31	-1.0
4	17	2.00	8.98	0.18	0.31	-4.0
2.4	17	1.19	8.98	0.11	0.31	-5.5
17	10	8.32	5.49	0.31	0.28	-0.5
17	8	8.30	4.31	0.31	0.27	-0.8
17	6	8.30	3.22	0.31	0.24	-2.0
17	4	8.30	2.19	0.31	0.19	-3.8
17	1.0	8.30		0.31		-18.0

Note: The mixers saturate at about 15 milliamperes control current (at input terminals). They produce minimum output power at about about 2.5 milliamperes.

TEST OF EVAS AMPLIFIER/CONTROLLER ASSEMBLY
August 5, 1992

A test was performed on the circuit board containing the amplitude and phase control circuitry and the RF amplifiers after installation of the amplifiers. The test setup was as shown below. The signal generator was set to deliver 0 dBm (1 milliwatt). The power supply for the amplifiers was set at 11 volts. The amplifiers were tested one at a time, with no power on the unused amplifiers and no termination at their antenna connection.



The measured data were identical for all three units. The measured values were:

UNIT	OUTPUT POWER (DBM) MIN.	MAX.
1	-25.	+ 9.99
2	-25.	+ 9.99
3	-25.	+ 9.99

TEST OF AMPLITUDE/PHASE CONTROLLER
FOR
GENERATION OF HARMONIC FREQUENCIES

MAY 4, 1992

A test was performed to determine whether the non-linear characteristics of the diodes used in the controllable attenuators would generate significant power at multiples of the input frequency. A CW signal was provided by a Hewlett Packard signal generator, model 614A. The signal power was applied to the input terminal of the controller, and its output power was input to a Hewlett Packard model 8555A spectrum analyzer. Power levels at the carrier frequency of 910 mHz and at its second and third harmonics were observed on the spectrum analyzer. Observations were made for four conditions:

(1) Control currents set at 8 milliamperes for full output, input power set at -20 dBm.

(2) Control currents set at 8 milliamperes, input power set at 0 dBm.

(3) Control currents set at 4 milliamperes, input power set at -20 dBm.

(4) Control currents set at 4 milliamperes, input power set at 0 dBm.

RESULTS:

Power levels at the second harmonic frequency were observed to be approximately 45 dB below that of the primary frequency. Power at the third harmonic was about 34 db. down. These values were seen at both settings of the control currents.

The controller was then removed from the experiment and power was applied directly from the signal generator to the spectrum analyzer. The same power levels at the harmonic frequency levels were seen for this condition. They were apparently produced by the signal generator.

CONCLUSIONS:

It was concluded that, for the input power levels of 0 dBm or less, no spurious frequency components of significant amplitude are generated by the controller.

TEST OF HYBRID-RING CONTROLLER AT MSFC USING VECTOR VOLTMETER

APRIL 10, 1992

The test setup was as shown in the attached sketch.

I. Both control currents were set for maximum output. Then, the current for attenuator A was gradually reduced while the relative output power and the output phase were measured. The measured phase was the phase difference between points (1) and (2) on the sketch. The results are shown below. The input power was +10 dBm.

CONTROL CURRENT, ma. A (red)	B (green)	OUTPUT POWER dBm.	RELATIVE VOLTAGE	PHASE Deg.	
15	15	-30.4	0.030	167.1	(47.1)
10	15	-30.6	0.029	166.6	(46.6)
5	15	-32.9	0.023	156.2	(36.2)
4	15	-34.0	0.020	149.1	(29.1)
3	15	-35.2	0.017	134.4	(14.4)
2	15	-35.6	0.016	116.8	(-3.2)
1	15	-35.4	0.017	104.8	(-15.2)
0	15	-35.2	0.017	99.4	(-20.6)

II. Both control currents were reset to 15 milliamperes to obtain maximum output power. Then, current B was gradually reduced to zero while the same measurements were made.

15	15	-30.4	0.030	167.7	(47.7)
15	10	-30.8	0.029	169.5	(49.5)
15	5	-32.9	0.023	-168.4	(71.6)
15	4	-33.5	0.021	-156.5	(83.5)
15	3	-33.8	0.020	-144.2	(95.8)
15	2	-33.8	0.020	-132.5	(107.7)
15	1	-33.4	0.021	-123.2	(116.8)
15	0	-32.9	0.023	-118.0	(122.0)

III. Output power was reduced to a minimum achievable value (-55 to -60 dBm) by adjustment of the two control currents. The currents required for this condition were A: 2.0 ma., B: 3.2 ma. Control current B was then held constant while current A was increased in steps up to a value of 15 milliamperes. Output power and phase were measured at each step.

CONTROL CURRENT A ma.	OUTPUT POWER dBm	RELATIVE VOLTAGE volts	PHASE deg.
2	-55 to -60	0.002	4.8 (244.
2.5	-59.2	0.001	-83.5 (156.
3	-50.0	0.003	-128.9 (111.
4	-42.0	0.008	-139.8 (100.
5	-38.0	0.013	-145.8 (94.2
6	-35.9	0.016	-148.4 (91.6
7	-34.9	0.018	-148.6 (91.4
8	-34.4	0.019	-148.6 (91.4
9	-34.2	0.019	-148.6 (91.4
10	-34.0	0.020	-148.5 (91.5
15	-33.7	0.021	-148.1 (91.9

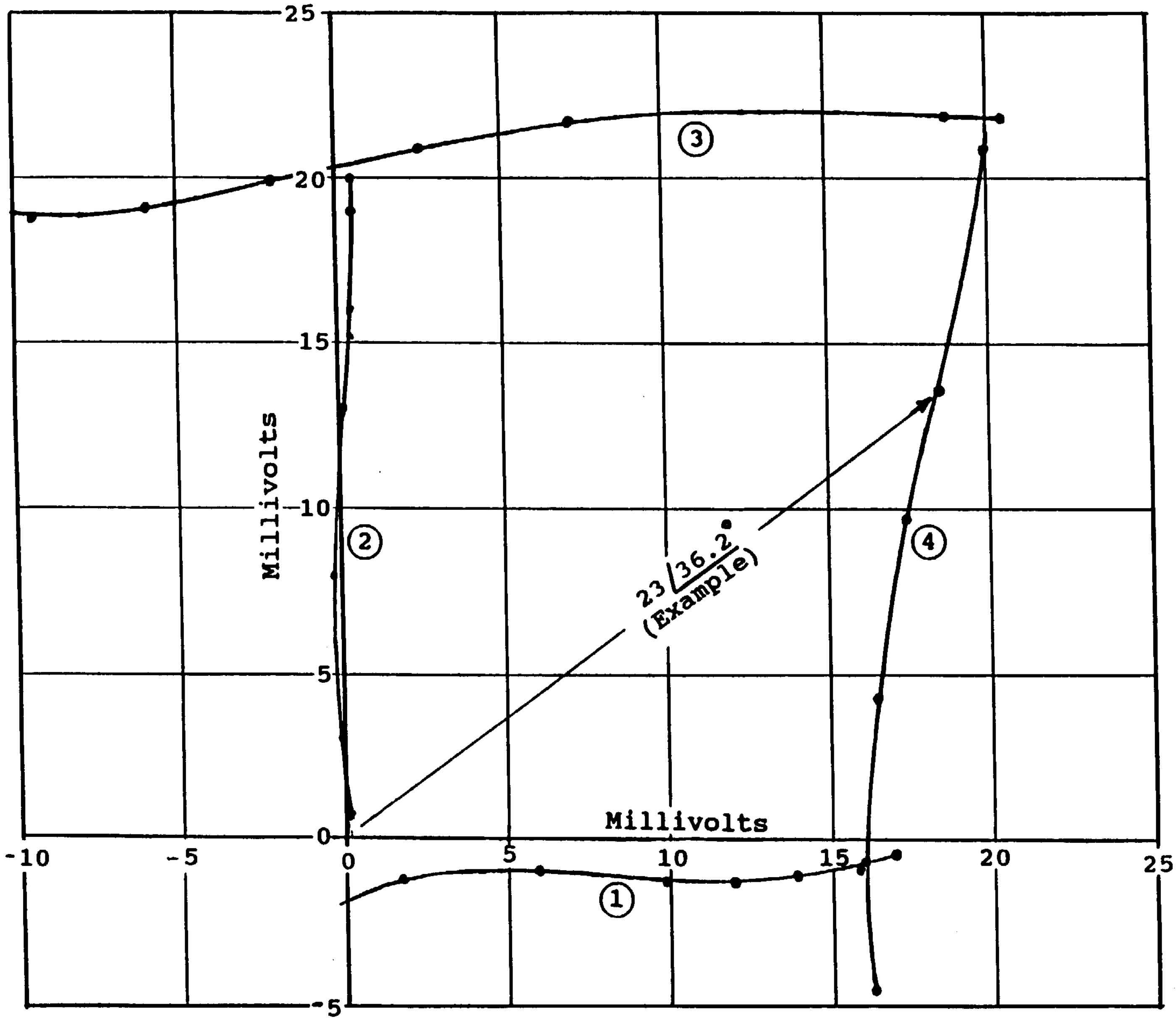
IV. The control currents were reset for minimum output power and current B was increased in steps to 16 milliamperes. The same measurements were made.

3.4	-56.1	0.002	18.2 (-101.
4	-52.3	0.002	89.8 (-30.
5	-44.1	0.006	110.7 (-9.3)
6	-40.4	0.010	113.2 (-6.8)
7	-38.3	0.012	114.4 (-5.6)
8	-37.0	0.014	115.6 (-4.4)
9	-36.5	0.015	116.4 (-3.3)
10	-36.1	0.016	117.1 (-2.9)

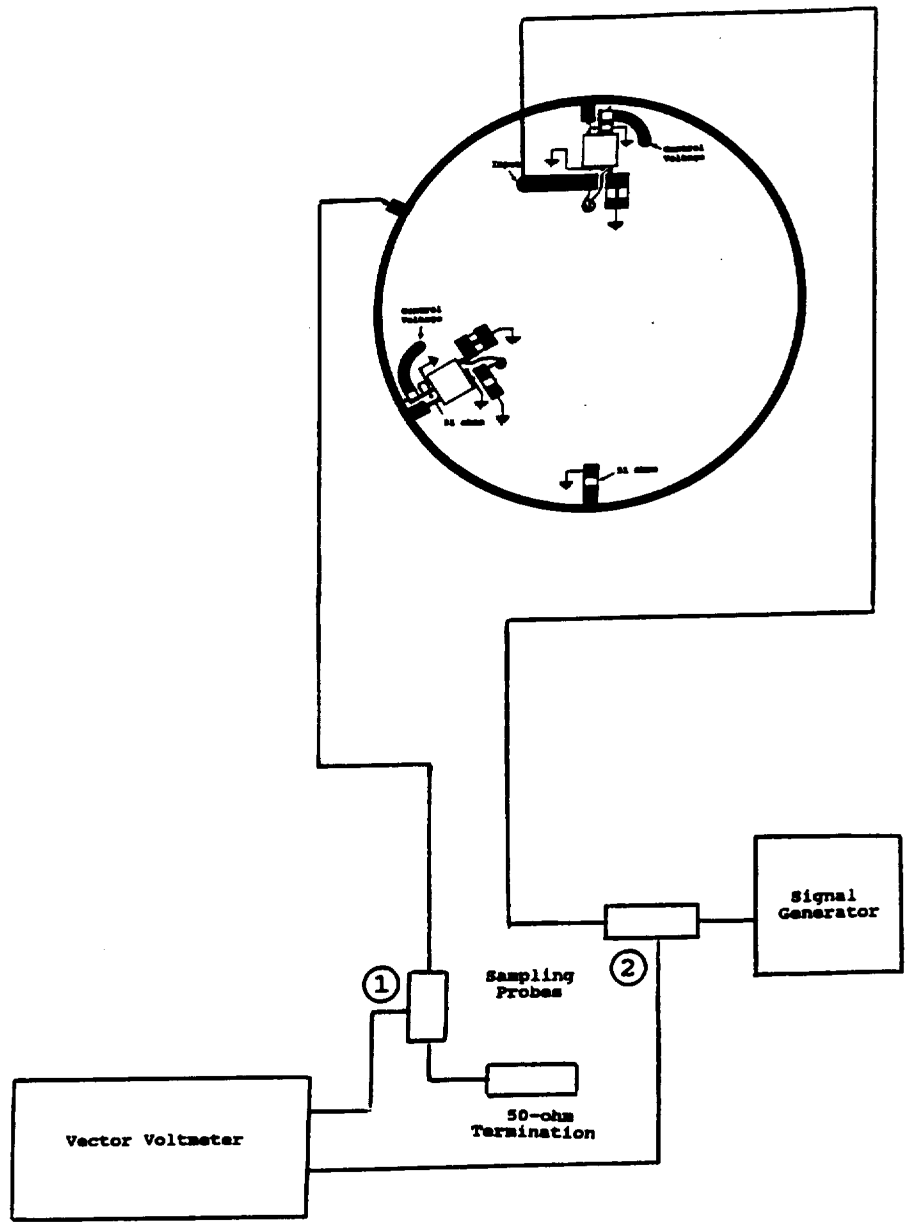
12	-35.8	0.016	117.8 (-2.2)
14	-35.6	0.017	118.2 (-1.8)
16	-35.5	0.017	118.6 (-1.4)

V. Repeat of run #III:

2.5	-62.1	0.001	-157.2 (82.8)
3	-50.4	0.003	-147.8 (92.2)
4	-41.7	0.008	-147.7 (92.3)
5	-37.8	0.013	-150.4 (89.6)
6	-35.7	0.016	-151.9 (88.1
7	-34.7	0.018	-152.0 (88.0
8	-34.3	0.019	-151.8 (88.2
9	-34.0	0.020	-151.7 (88.3
10	-33.9	0.020	-151.4 (88.6
12	-33.7	0.020	-151.3 (88.7



SUMMATION OF OUTPUT VOLTAGES



TEST SETUP FOR MAKING AMPLITUDE AND PHASE MEASUREMENTS

Note: The control-current sources are not shown.

TEST OF RATRACE (0, 180 DEG.) HYBRID RING

APRIL 17, 1992

The hybrid ring was tested without any other components attached, as shown in the sketch. Power was applied to one of the input ports and the power delivered at each of the output ports was measured, the other port being terminated in 50 ohms. Power was also measured at the other inport port with both of the output ports terminated in 50 ohms. The results are as shown below.

POWER APPLIED TO PORT #1 -

Measured input power: -3.60 dbm.

Power from output port #1: -7.0 dbm.
Power from output port #2: -7.1 dbm.
Power from input port #2: not measurable (<-25 dbm.)

POWER APPLIED TO PORT #2 -

Power from output port #1: -6.9 dbm.
Power from output port #2: -6.9 dbm.
Power from input port #1: not measurable (<-25 dbm.)

The differences between input and output powers are then calculated as follows:

FOR POWER INPUT TO PORT #1,

7.0 - 3.6 = 3.4 for output port #1
7.1 - 3.6 = 3.5 for output port #2

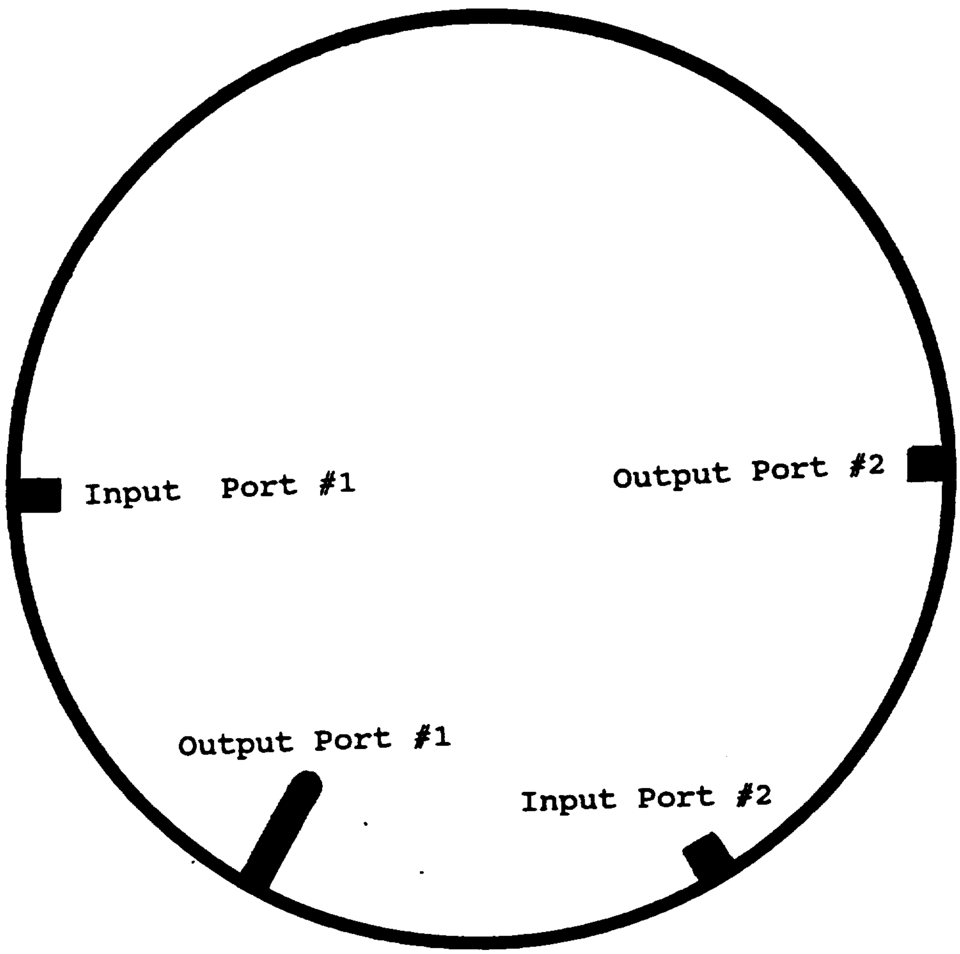
FOR POWER INPUT TO PORT #2,

6.9 - 3.6 = 3.3 for output port #1
6.9 - 3.6 = 3.3 for output port #2

The two input ports are clearly well isolated from each other.

A VSWR measurement was made for each of the two input ports with all other ports terminated in 50 ohms each. The results are as follows:

VSWR at port #1: 1.16
VSWR at port #2: 1.08



TEST CONFIGURATION FOR RATRACE (0, 180 DEG.) HYBRID RING

SECOND TEST OF AMPLITUDE/PHASE CONTROLLER

MSFC, BLDG. 4487

APRIL 21, 1992

A second test of the controller was performed, using the vector voltmeter. Two changes were involved in this test:

(1) The resistive pads between the output of the quadrature hybrid ring and the attenuators (mixers) was changed. The 200-ohm resistance was changed to 100 ohms in an effort to reduce the insertion loss of the unit.

(2) An input power level of 0 dBm (1.0 milliwatt) was used.

This test setup was the same as was used in the first test. The tests performed and the results are shown below. The relative-voltage values were calculated from the measured dBm values. The phase angles shown have been rotated from the measured values by a constant (72 degrees) to provide a more convenient phase reference.

I. The control currents were adjusted to reduce the total output power to a minimum level (-60 dBm). The current for attenuator A was then increased in steps to a maximum value of 10 milliamperes and the output power and phase measured at each step. The currents were then readjusted for minimum output power and the process was repeated for variation of the current to attenuator B.

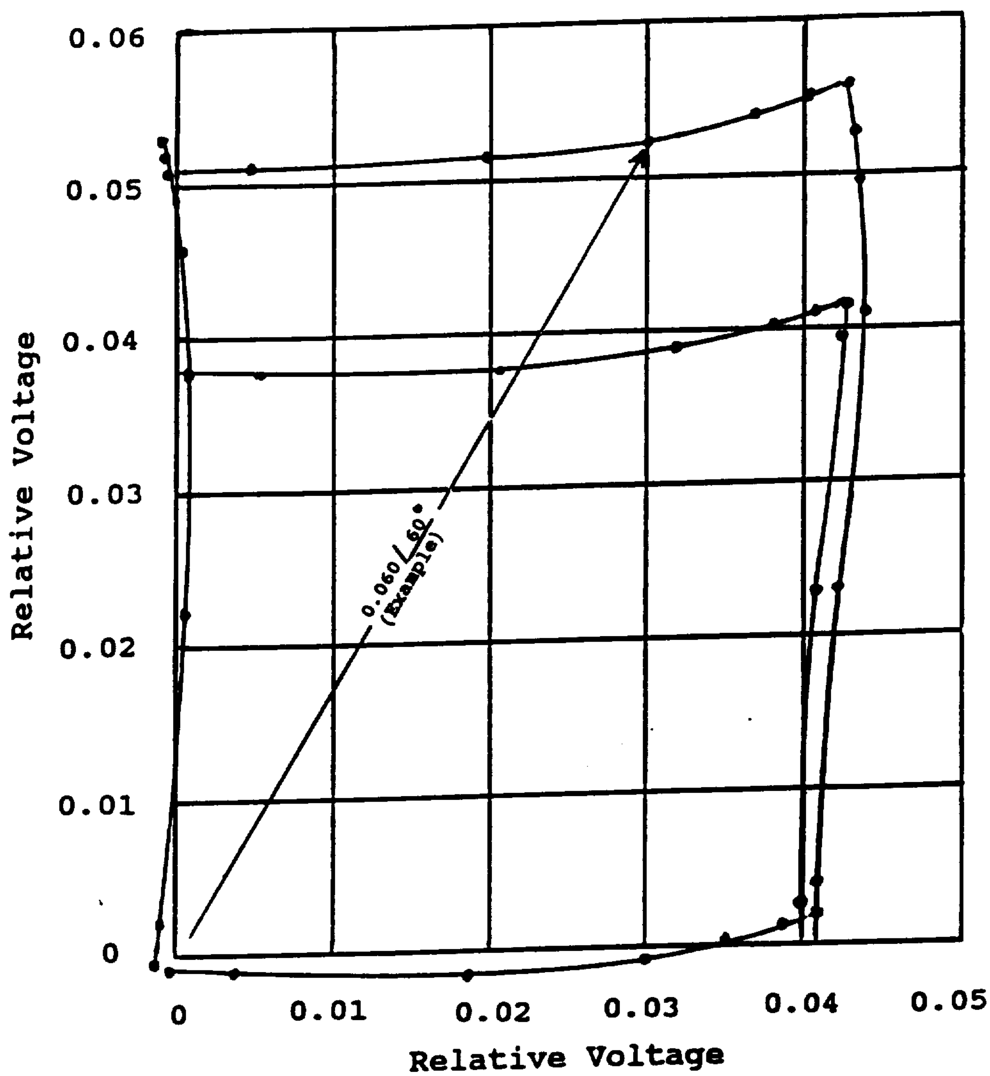
CONTROL CURRENT milliamperes A	B	OUTPUT POWER dBm	RELATIVE VOLTAGE	PHASE DEG
2.9	3.7	-60.0	0.001	-160.0
3.0	3.7	-53.7	0.002	-244.0
4.0	3.7	-33.1	0.022	89.2
5.0	3.7	-28.5	0.038	88.6
6.0	3.7	-26.8	0.046	89.4
7.0	3.7	-26.2	0.049	90.1
8.0	3.7	-25.8	0.051	90.6
9.0	3.7	-25.6	0.052	90.8
10.0	3.7	-25.5	0.053	91.0
2.9	3.7	-60.0	0.001	-133.0
2.9	4.0	-47.5	0.004	-18.0
2.9	5.0	-34.3	0.019	-5.0
2.9	6.0	-30.6	0.030	-2.0
2.9	7.0	-29.1	0.035	-0.1
2.9	8.0	-28.4	0.038	1.0
2.9	9.0	-28.1	0.039	2.0
2.9	10.0	-27.8	0.041	2.5

II. The control currents were then set for maximum output power. The current for attenuator A was then reduced in steps, leaving the current for attenuator B unchanged. Output power and phase were measured at each step. The currents were then reset for maximum output power and the process repeated for attenuator B.

CONTROL CURRENT milliamperes A	B	OUTPUT POWER dBm	RELATIVE VOLTAGE	PHASE DEG
8.0	10.0	-23.1	0.070	52.3
7.0	10.0	-23.3	0.068	50.9
6.0	10.0	-23.6	0.066	48.5
5.0	10.0	-24.5	0.060	43.3
4.0	10.0	-26.3	0.048	28.8
3.0	10.0	-27.8	0.041	5.1
2.5	10.0	-27.9	0.040	-7.7
2.0	10.0	-27.8	0.041	-17.3
8.0	10.0	-23.1	0.070	52.4
8.0	9.0	-23.2	0.069	52.9
8.0	8.0	-23.4	0.068	53.8
8.0	7.0	-23.8	0.065	55.6
8.0	6.0	-24.4	0.060	60.0
8.0	5.0	-25.2	0.055	69.0
8.0	4.0	-25.8	0.051	85.0
8.0	3.0	-25.6	0.052	101.7
8.0	2.0	-25.1	0.056	-248.2

III. The test performed under II was repeated with the initial power level set to a lower value.

CONTROL CURRENT milliamperes A	B	OUTPUT POWER dBm	RELATIVE VOLTAGE	PHASE DEG
5.0	9.0	-24.7	0.058	43.2
4.0	9.0	-26.5	0.047	29.1
3.0	9.0	-28.0	0.040	3.6
2.0	9.0	-27.9	0.040	-18.7
5.0	9.0	-24.5	0.060	44.1
5.0	8.0	-24.8	0.058	45.1
5.0	7.0	-25.2	0.055	46.5
5.0	6.0	-26.0	0.050	50.5
5.0	5.0	-27.3	0.043	61.3
5.0	4.0	-28.3	0.038	81.6
5.0	3.0	-28.1	0.039	104.2
5.0	2.0	-27.4	0.043	103.4



SUMMATION OF OUTPUT VOLTAGES

TEST OF AMPLITUDE/PHASE CONTROLLER AFTER SECOND MODIFICATION OF RESISTANCE PAD BETWEEN QUADRATURE HYBRID AND ATTENUATORS

APRIL 27, 1992

The minimum insertion loss of the controller had been measured as 26 dB after the first modification. That modification had been the reduction of resistance R_2 from 200 ohms to 100 ohms, with R_1 being 62 ohms. The second modification consisted of making R_2 zero ohms (direct connection) and increasing R_1 to 100 ohms. The test was performed as follows:

I. Input power from the signal generator was set at 0 dBm (one milliwatt). The two control currents were then adjusted to obtain minimum output power from the controller. The current values were found to be 1.5 milliamperes for each attenuator. The value of control current A was then varied in steps, leaving current B at the 1.5-ma value. The resulting output power was measured at each step. Current A was then reset to 1.5 ma. and current B was varied in steps, again measuring the resulting output power. The results are shown below.

I_A ma	I_B ma	POWER dBm
1.5	1.5	<-30
3.0	1.5	-30.1
4.0	1.5	-25.5
5.0	1.5	-22.6
6.0	1.5	-20.8
7.0	1.5	-19.6
8.0	1.5	-18.8
9.0	1.5	-18.5
10.0	1.5	-18.3
1.5	1.5	<-30
1.5	3.0	-30.8
1.5	4.0	-26.5
1.5	5.0	-23.4
1.5	6.0	-21.5
1.5	7.0	-20.5
1.5	8.0	-20.1
1.5	9.0	-19.9
1.5	10.0	-19.8

II. A test was then performed to check the quadrature relation of the A and B contributions. A value of control current A and its corresponding value of output power were selected from the tabulated value in the first part of test I. With control current A set to 1.5 milliamperes, current B was adjusted to produce the same power output as the selected tabulated value for A. Current A was then set at at the tabulated value and the resulting output power measured. The intent of this test was to set the two control currents to produce the same individual values of output power and measure the sum of the two contributions. It the amounts of power delivered by the two attenuators are equal and they are 90 degrees out of phase, then the sum of the two contributions should be 3 dB higher than either of the two individual power values. The results show this condition to be true.

(1) I_A value of 6.0 selected; corresponding power = -20.8. I_B adjusted to 6.5 ma to produce -20.9 dBm output (closest achievable).

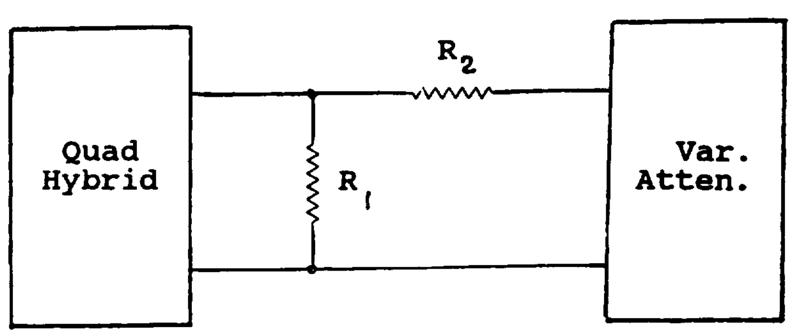
Measured output power: -17.9 dBm.

-20.9 - (-17.9) = 3.0 dB.

(2) I_A value of 5.0 selected; corresponding power = - 22.6. I_B adjusted to 5.4 ma to produce -22.6 dBm output power.

Measured output power: -19.7 dBm.

-22.6 - (-19.7) = 2.9 dB.



APPENDIX E

TESTS OF AMPLITUDE/PHASE CONTROLLER (COMPLETE ASSEMBLY)

TEST OF ANTENNA AND CONTROL CIRCUIT ASSEMBLY
July 21, 1992

The amplitude and phase control circuitry was tested for performance of all three antenna control modules to determine that each module was capable of setting the amplitude and phase of the antenna drive power to desired values. The test was performed by varying the control currents to the controllable attenuators and measuring the phase and relative amplitude of the voltage on the antenna line. The test points were used in making these measurments. The test setup is shown in the sketch.

The test results are shown below.

ANTENNA #1

CONTROL VOLTAGES		RELATIVE AMPLITUDE	RELATIVE PHASE
OUTER	INNER		
1.8 v.	0.6 v.	.047	-14 deg.
2.0	0.6	.193	-18
2.5	0.6	.477	-18
3.0	0.6	.675	-18
3.5	0.6	.780	-17
4.0	0.6	.838	-16
4.5	0.6	.872	-16
5.0	0.6	.885	-15
5.5	0.6	.904	-15
6.0	0.6	.892	-14
6.5	0.6	.900	-14
7.0	0.6	.902	-14
7.5	0.6	.908	-14
8.0	0.6	.910	-14
8.5	0.6	.911	-13
9.0	0.6	.911	-13
9.5	0.6	.915	-13
10.0	0.6	.918	-13
1.8	0.6	.040	-157
1.8	1.0	.213	89.3
1.8	1.5	.514	82.5
1.8	2.0	.838	80.3
1.8	2.5	1.113	79
1.8	3.0	1.365	78
1.8	3.5	1.515	78
1.8	4.0	1.588	79
1.8	4.5	1.632	79
1.8	5.0	1.653	80
1.8	5.5	1.677	80
1.8	6.0	1.678	80
1.8	6.5	1.692	80
1.8	7.0	1.692	80
1.8	7.5	1.700	80
1.8	8.0	1.700	80
1.8	8.5	1.702	80

1.8	9.0	1.706	80
1.8	9.5	1.706	80
1.8	10.0	1.706	80

ANTENNA #2

1.2	0.0	.008	130
1.5	0.0	.199	11
2.0	0.0	.530	9.4
2.5	0.0	.827	8
3.0	0.0	1.000	7.3
3.5	0.0	1.130	7.7
4.0	0.0	1.193	8
4.5	0.0	1.228	8.3
5.0	0.0	1.254	8.6
5.5	0.0	1.271	8.9
6.0	0.0	1.279	9.1
6.5	0.0	1.294	9.2
7.0	0.0	1.294	9.3
7.5	0.0	1.294	9.4
8.0	0.0	1.301	9.4
8.5	0.0	1.303	9.4
9.0	0.0	1.306	9.7
9.5	0.0	1.306	9.8
10.0	0.0	1.309	9.8

1.2	0.0	.008	12
1.2	0.5	.214	72
1.2	1.0	.487	77
1.2	1.5	.750	79
1.2	2.0	1.029	81
1.2	2.5	1.318	81
1.2	3.0	1.545	81
1.2	3.5	1.693	82
1.2	4.0	1.757	83
1.2	4.5	1.795	83
1.2	5.0	1.818	84
1.2	5.5	1.835	84
1.2	6.0	1.844	84
1.2	6.5	1.850	84
1.2	7.0	1.855	84
1.2	7.5	1.862	84
1.2	8.0	1.864	84
1.2	8.5	1.865	84
1.2	9.0	1.866	84
1.2	9.5	1.868	84
1.2	10.0	1.868	85

ANTENNA #3

1.4	0.8	.052	-177
1.5	0.8	.009	-143
2.0	0.8	.293	-3.0
2.5	0.8	.621	-3.3
3.0	0.8	.854	-3.0

3.5	0.8	.955	-2.2
4.0	0.8	1.023	-1.3
4.5	0.8	1.056	-0.8
5.0	0.8	1.080	-0.2
5.5	0.8	1.098	0.0
6.0	0.8	1.117	0.2
6.5	0.8	1.123	0.4
7.0	0.8	1.129	0.6
7.5	0.8	1.133	0.8
8.0	0.8	1.139	1.0
8.5	0.8	1.142	0.9
9.0	0.8	1.142	1.2
9.5	0.8	1.143	1.0
10.0	0.8	1.144	1.2
1.4	0.8	.052	-174
1.4	1.0	.095	+123
1.4	1.5	.395	97
1.4	2.0	.704	94
1.4	2.5	1.000	91
1.4	3.0	1.247	90
1.4	3.5	1.423	90
1.4	4.0	1.509	90
1.4	4.5	1.557	90
1.4	5.0	1.582	91
1.4	5.5	1.586	90
1.4	6.0	1.600	91
1.4	6.5	1.603	91
1.4	7.0	1.612	91
1.4	7.5	1.615	91
1.4	8.0	1.622	92
1.4	8.5	1.628	92
1.4	9.0	1.630	92
1.4	9.5	1.630	92
1.4	10.0	1.631	92

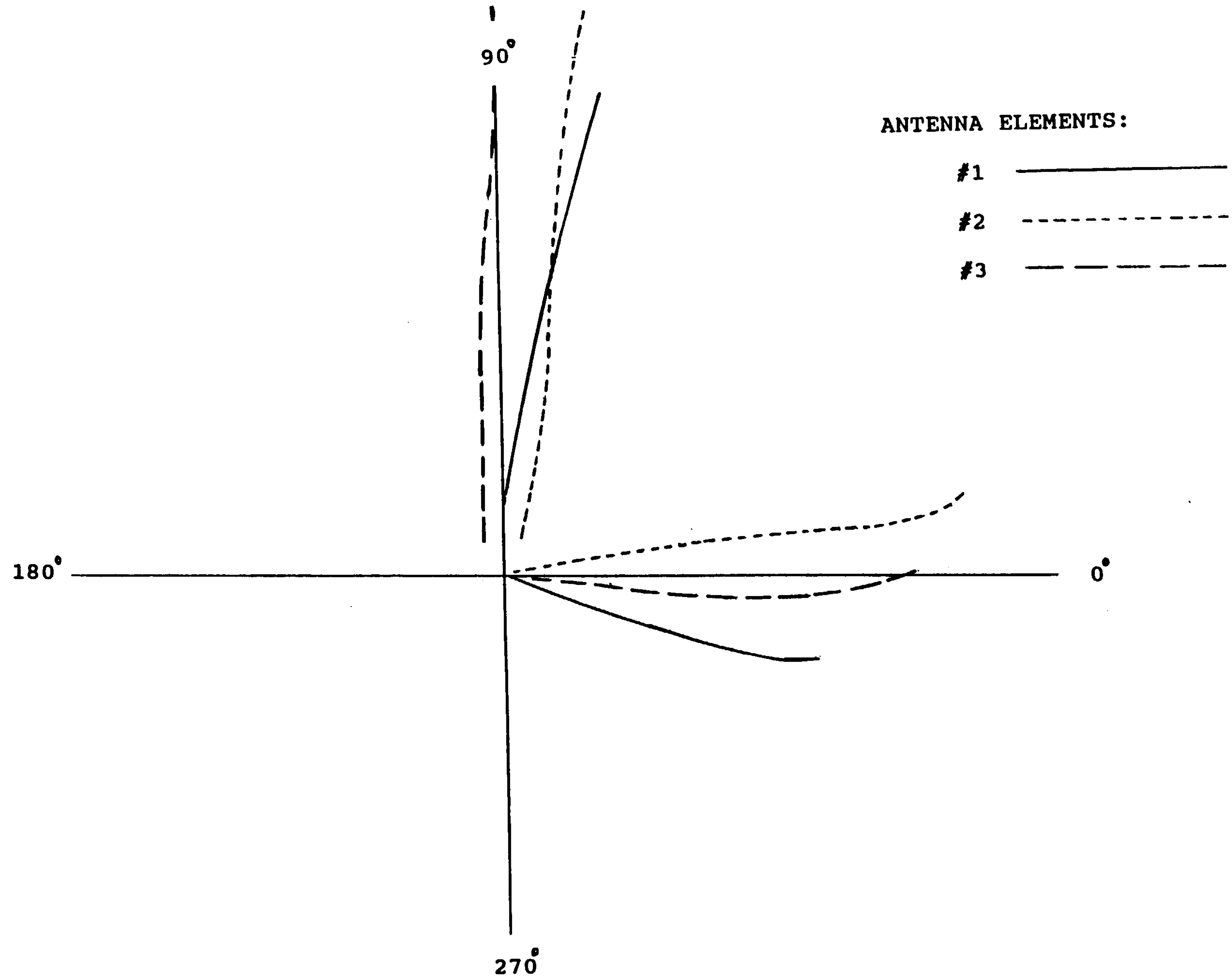
CURRENT VERSUS VOLTAGE TO ELEMENT #1

Outer Attenuator:

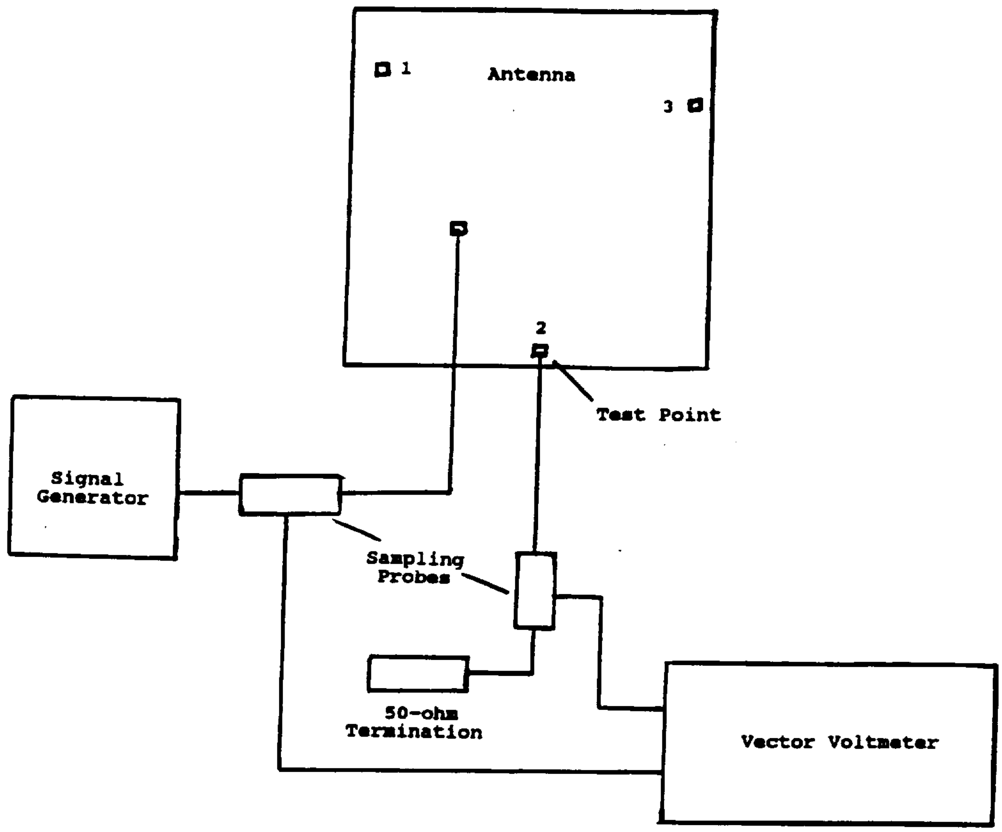
VOLTS	MILLIAMPERES
1.0	2.0
2.0	4.8
3.0	5.8
4.0	7.5
5.0	9.7
6.0	11.5
7.0	13.5
8.0	15.5
9.0	18.0
10.0	20.0

Inner Attenuator:

VOLTS	MILLIAMPERES
1.0	2.0
2.0	4.0
3.0	6.0
4.0	8.0
5.0	10.0
6.0	11.5
7.0	14.0
8.0	16.0
9.0	18.0
10.0	21.0



RELATIVE AMPLITUDES AND PHASES OF VOLTAGES TO ANTENNA ELEMENTS
WITH VARYING CONTROL CURRENTS



TEST SETUP FOR MAKING AMPLITUDE AND PHASE MEASUREMENTS

APPENDIX F

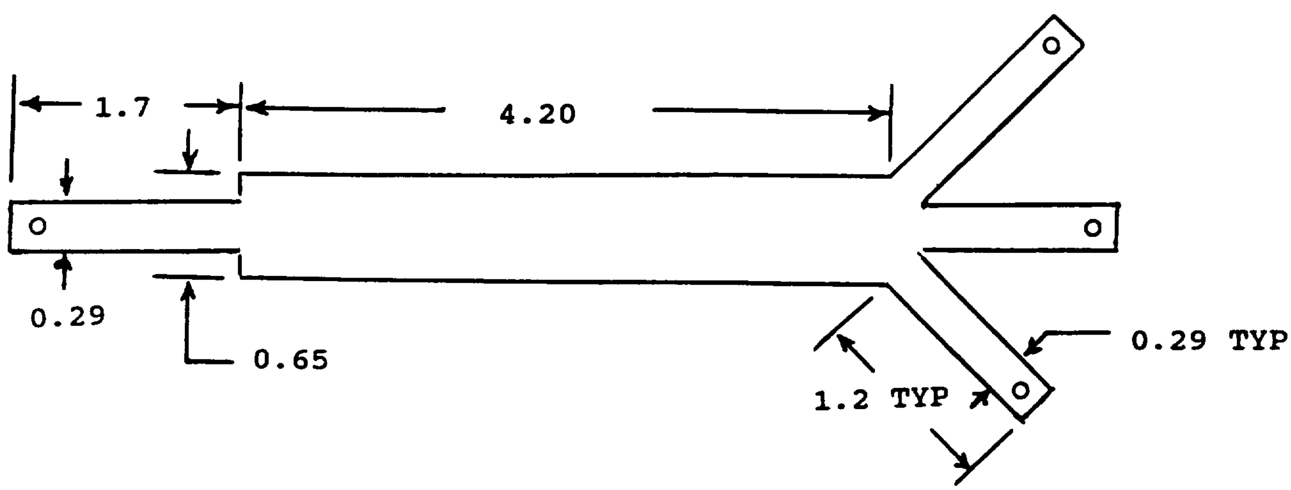
THREE-WAY POWER DIVIDER

IMPEDANCE TEST OF THREE-WAY POWER DIVIDER
MAY 1, 1992

A microstrip power divider was designed to distribute the RF power equally to the three amplitude/phase controllers. It consists of three 50-ohm output lines emanating from a common junction as shown in the sketch below. Each line will carry 1/3 of the input power to its controller. Impedance matching is accomplished by use of quarter-wavelength transforming section having a characteristic impedance of 50/ 3.

The divider was impedance-tested by terminating each of the three 50-ohm lines in a 50-ohm resistor and measuring the VSWR seen at the input end of the quarter-wavelength section. A Hewlett Packard 614A signal generator, a Hewlett Packard 805A slotted line and a Hewlett Packard 415D SWR meter were used to perform the test.

The measured VSWR at 910 mHz was 1.14:1.

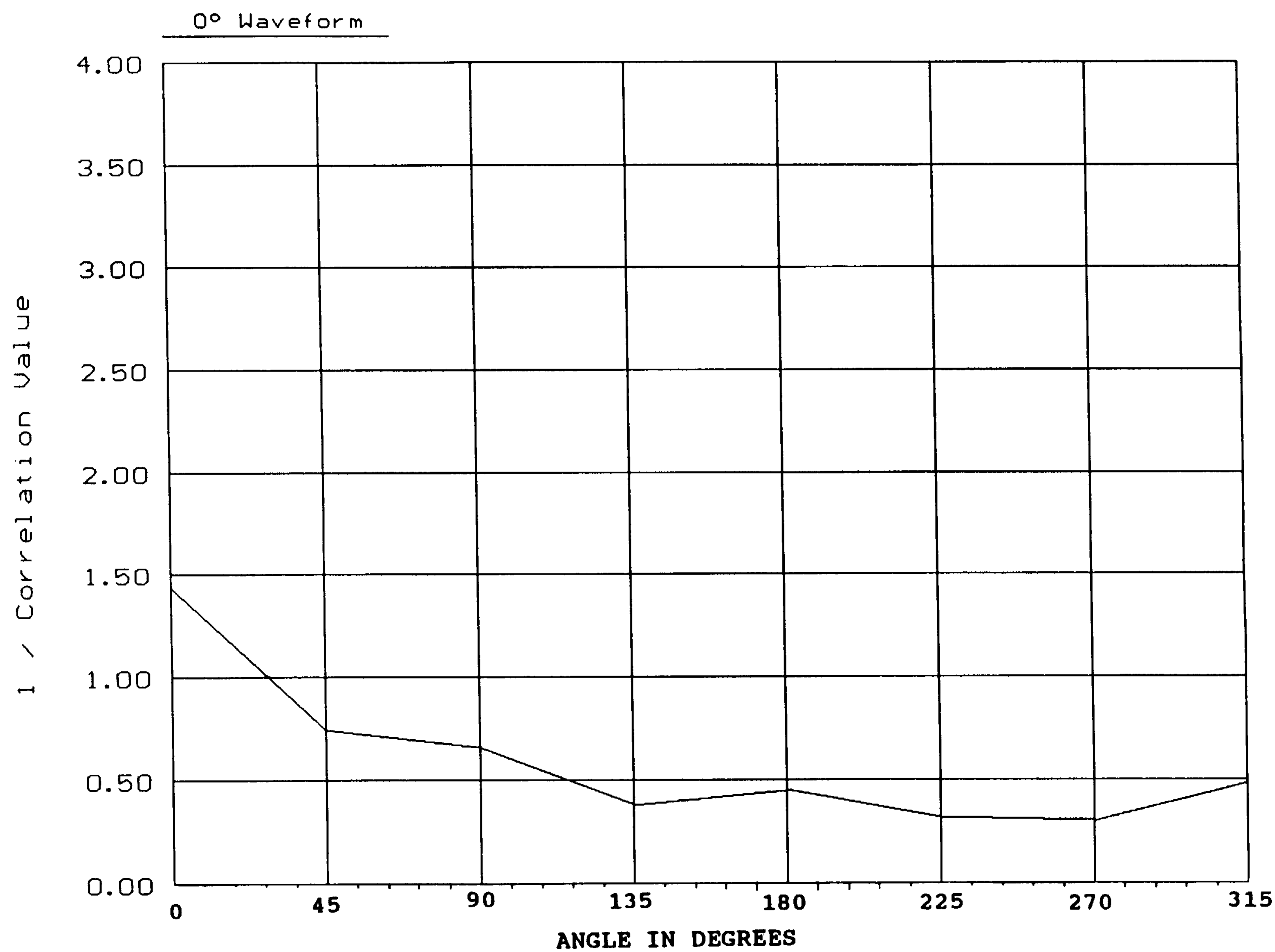


DIMENSIONS IN CENTIMETERS

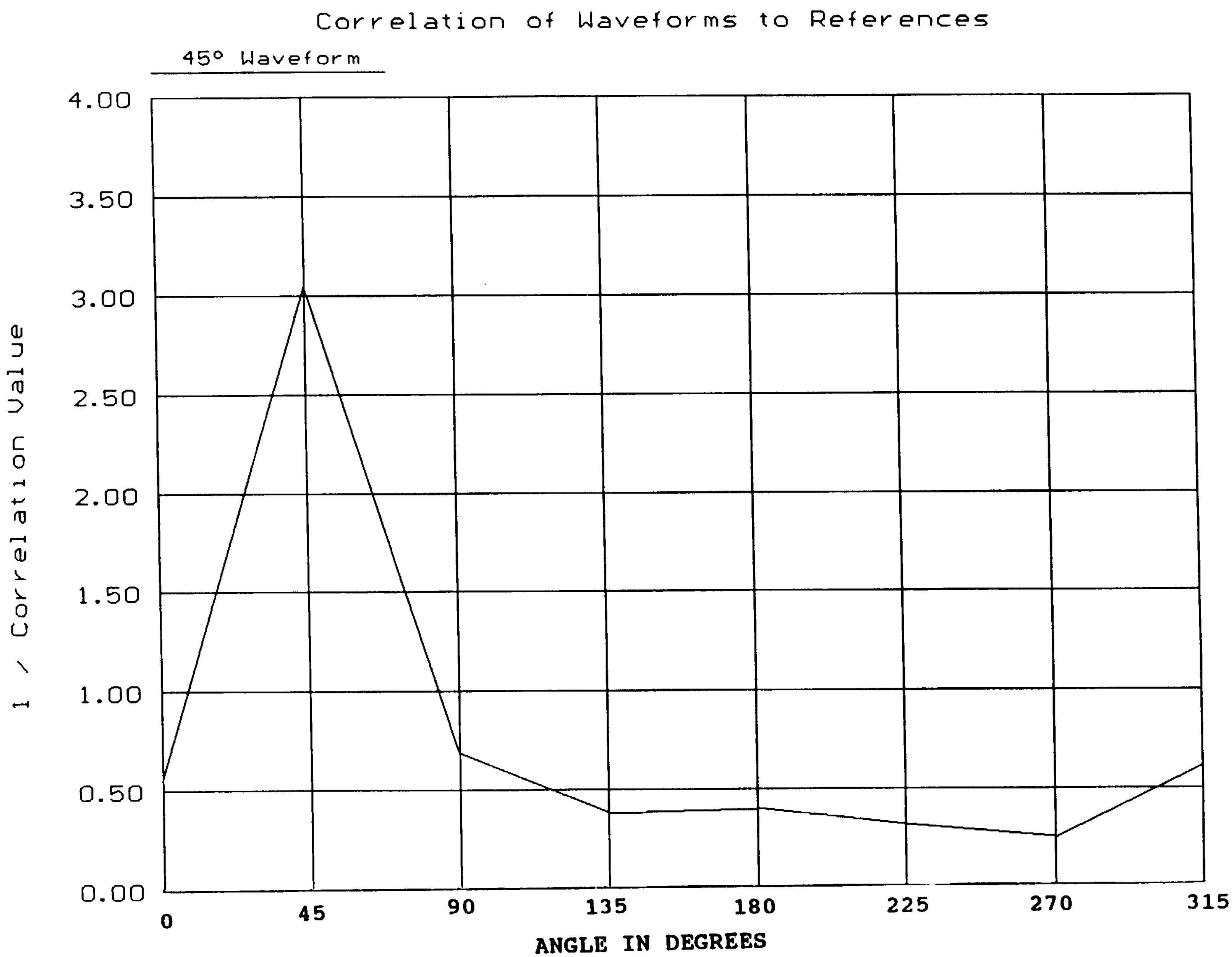
APPENDIX G

CORRELATION OF FIELD TEST DATA WITH REFERENCE WAVEFORMS

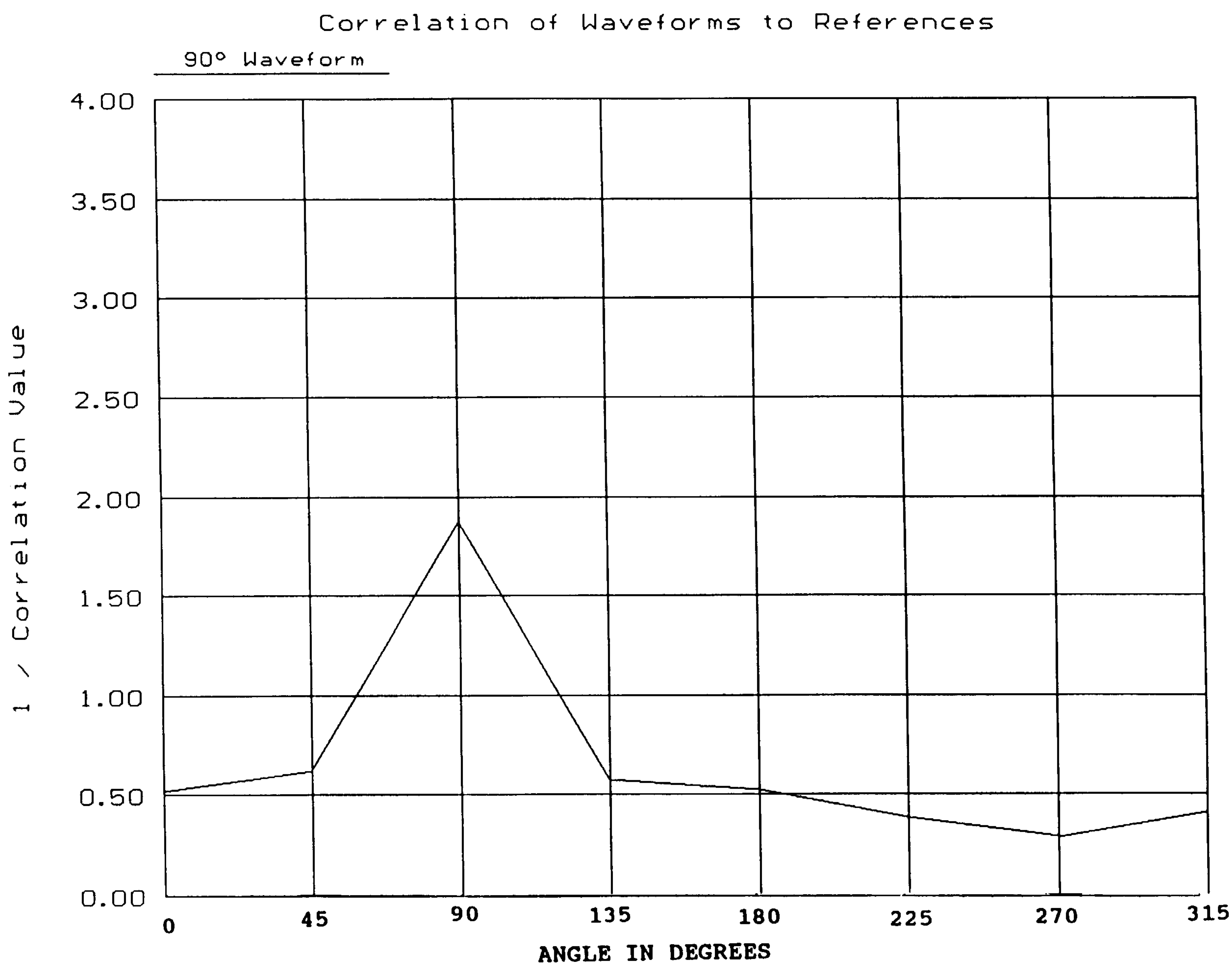
Correlation of Waveforms to References



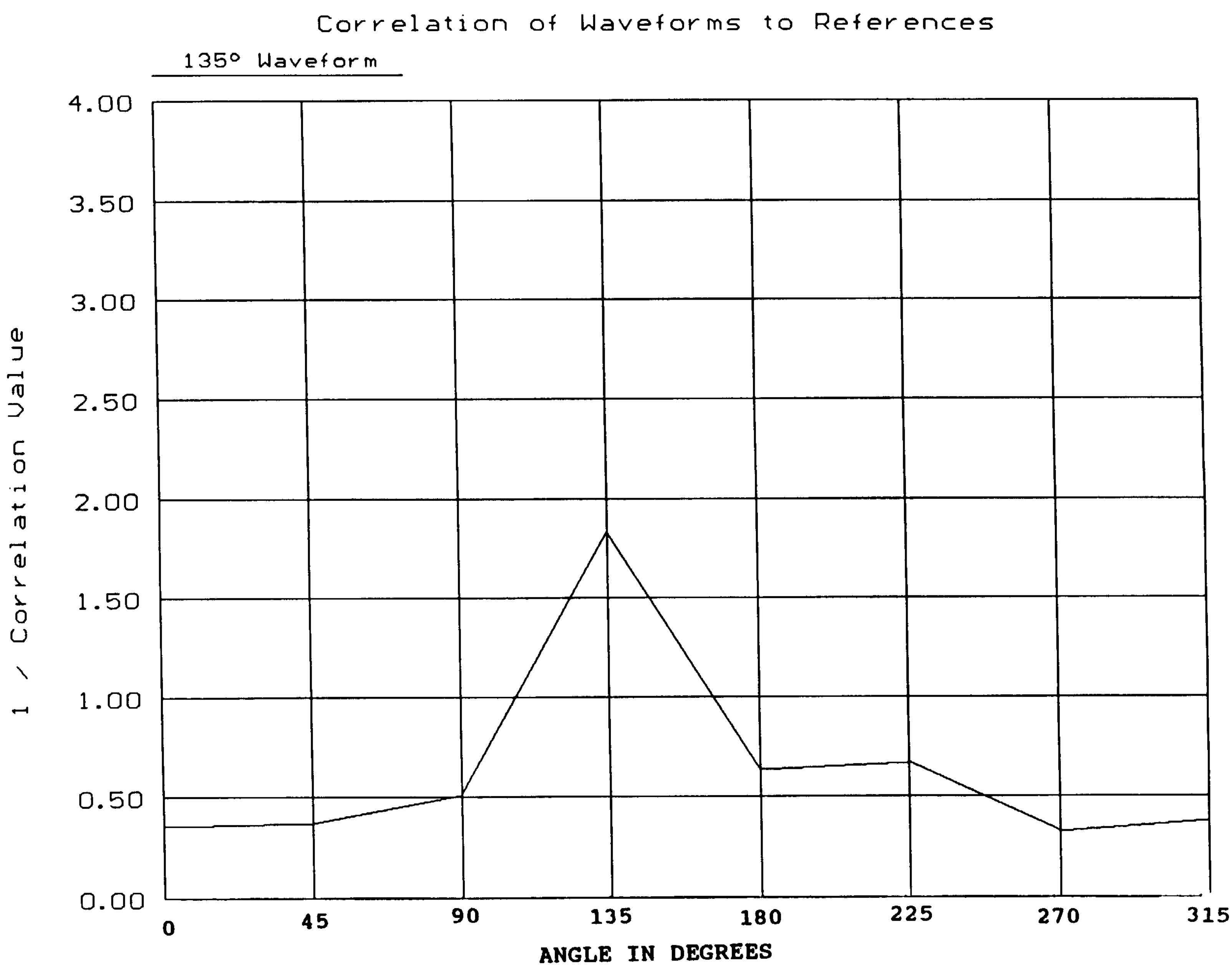
Appendix G(a)


Correlation of Waveforms to References
45° Waveform
4.00
3.50
3.00
2.50
2.00
1.50
1.00
0.50
0.00
1 / Correlation Value
0
45
90
135
180
225
270
315
ANGLE IN DEGREES

Appendix G(b)


Correlation of Waveforms to References
90° Waveform
1 / Correlation Value
4.00
3.50
3.00
2.50
2.00
1.50
1.00
0.50
0.00
0
45
90
135
180
225
270
315
ANGLE IN DEGREES

Appendix G(c)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135° Waveform
1 / Correlation Value
4.00
3.50
3.00
2.50
2.00
1.50
1.00
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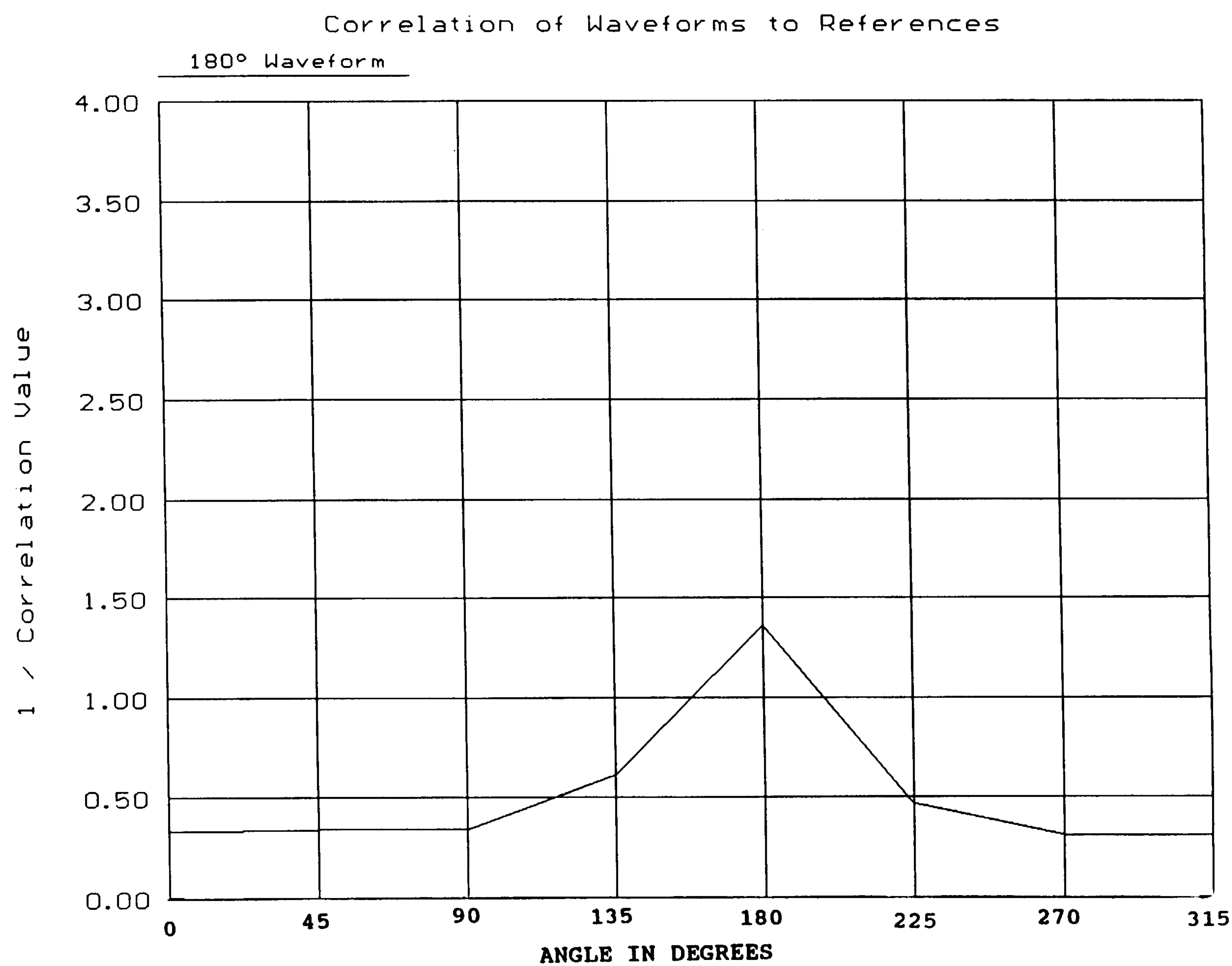
0.00
0
45
90
135
180
225
270
315
ANGLE IN DEGREES

Appendix G(d)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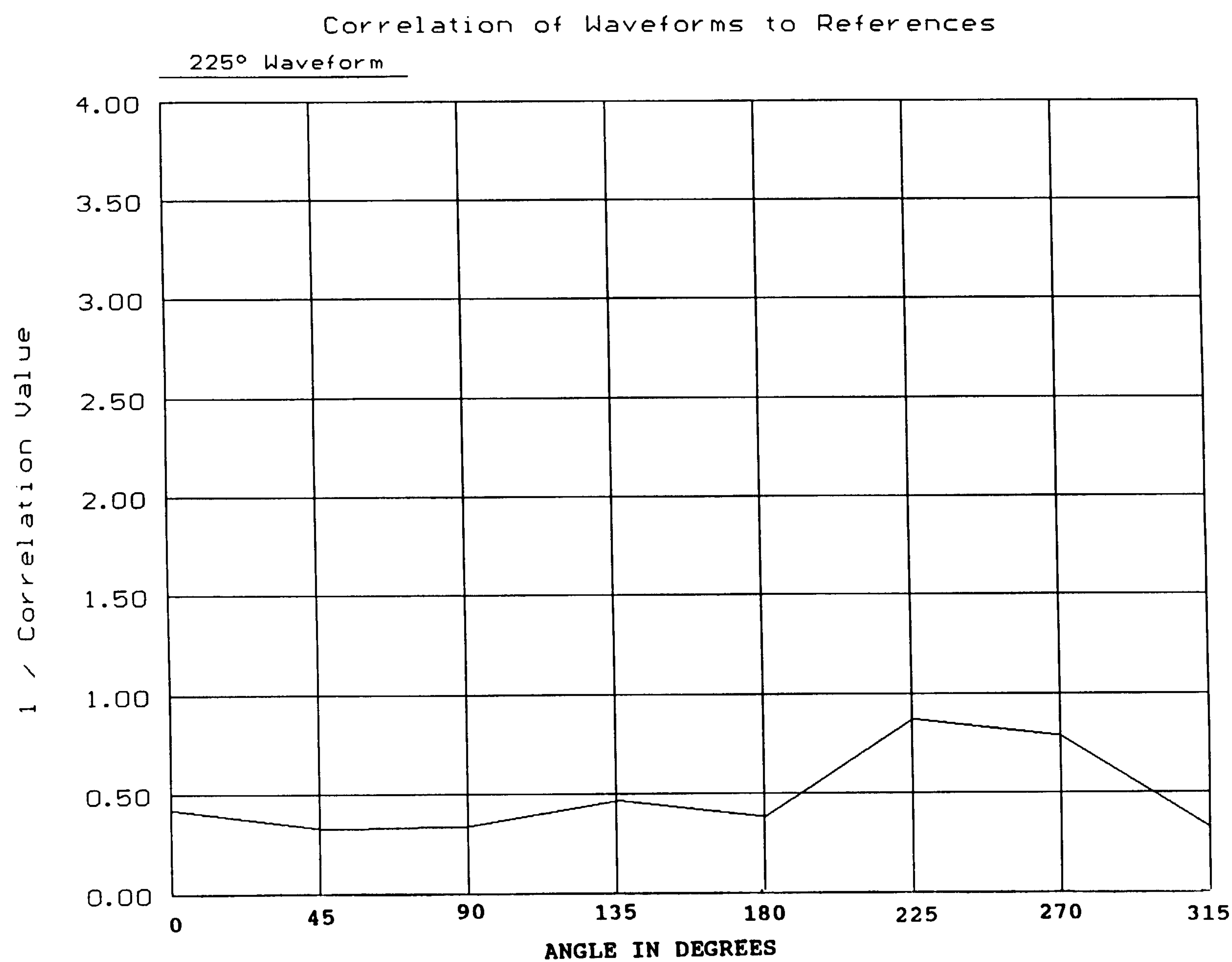
180° Waveform
4.00
3.50
3.00
2.50
2.00
1.50
1.00
0.50
0.00
1 / Correlation Value
0
45
90
135
180
225
270
315
ANGLE IN DEGREES

Appendix G(e)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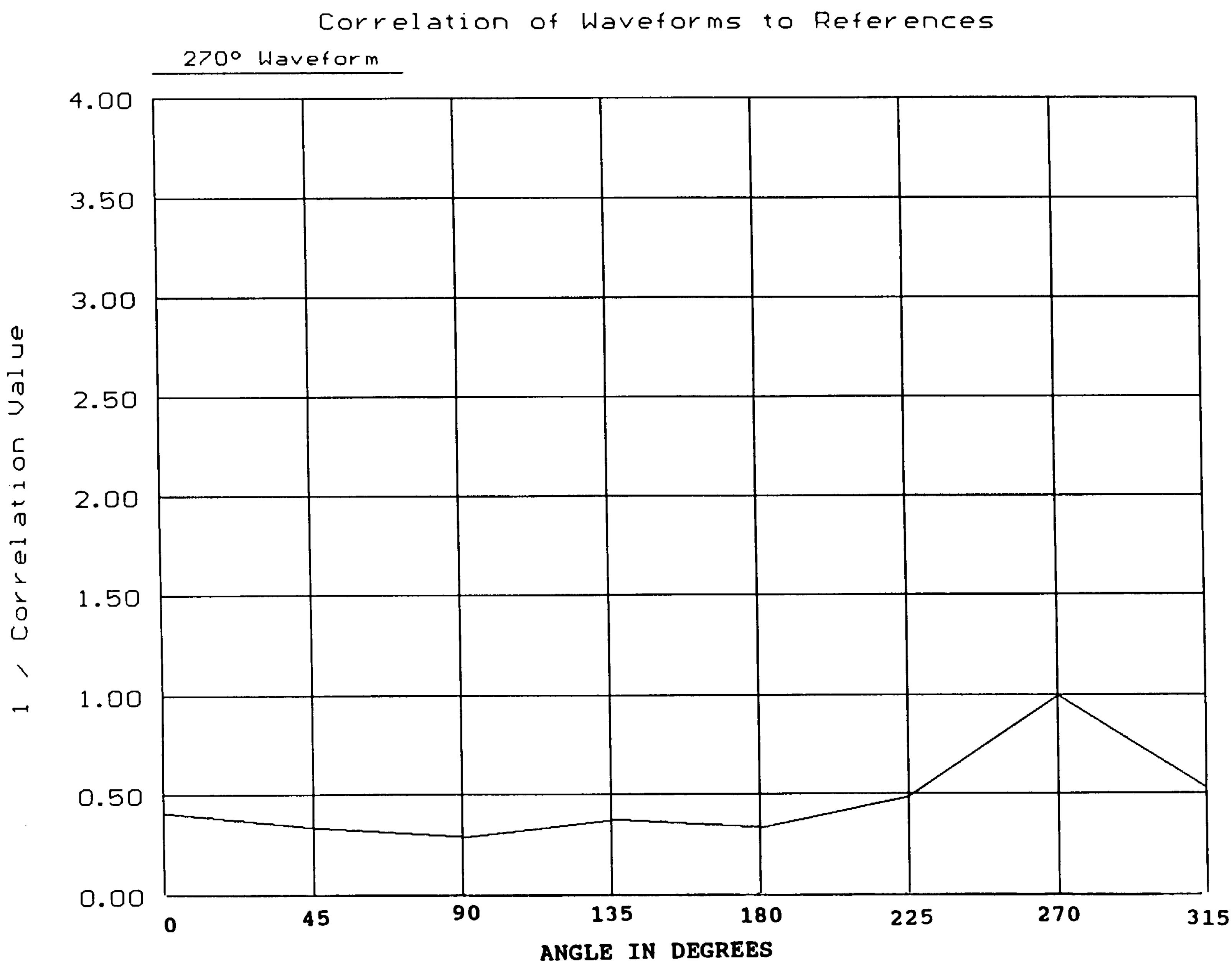
225° Waveform
4.00
3.50
3.00
2.50
2.00
1.50
1.00
0.50
0.00
1 / Correlation Value
0
45
90
135
180
225
270
315
ANGLE IN DEGREES

Appendix G(f)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270° Waveform
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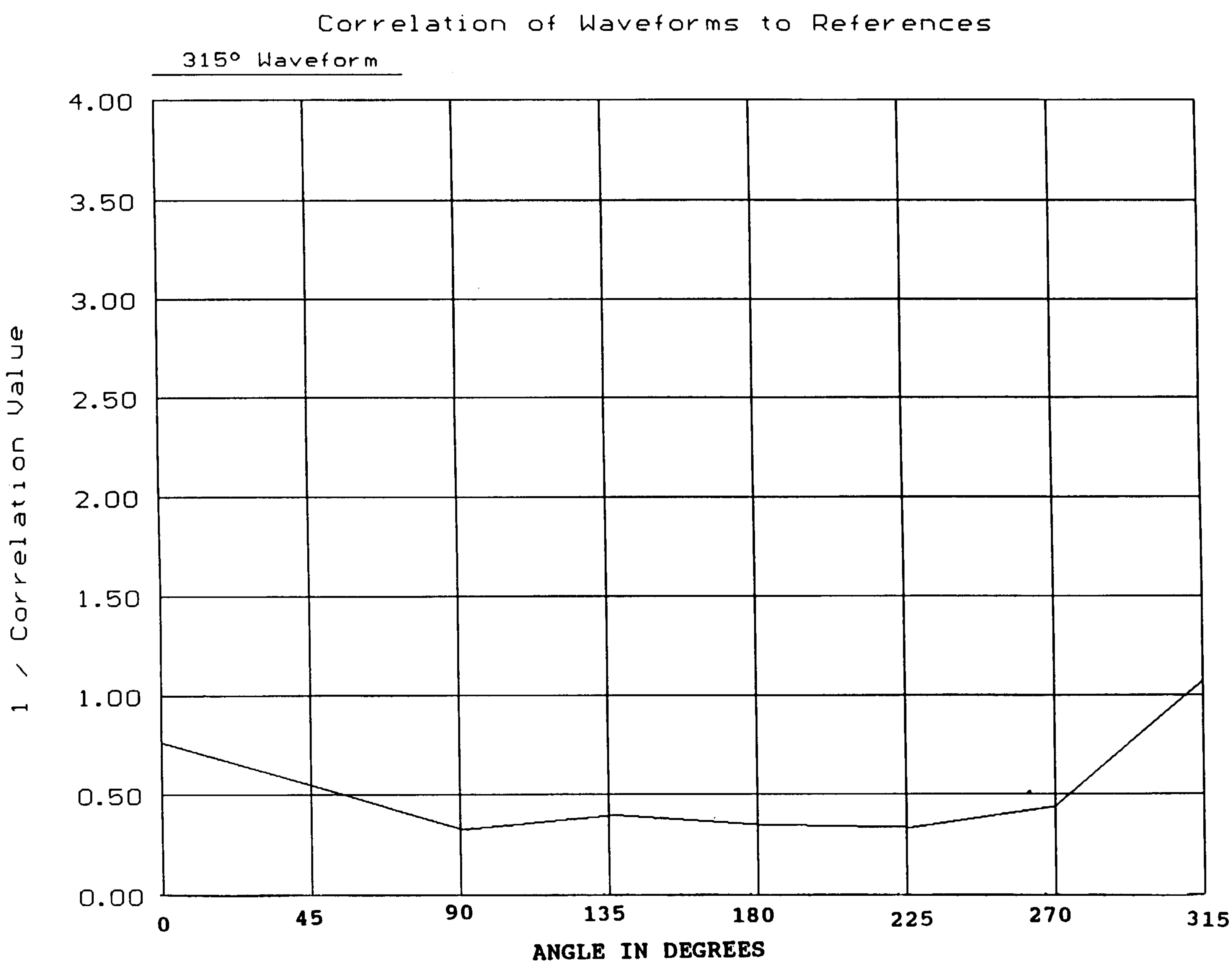
3.50
3.00
2.50
2.00
1.50
1.00
0.50
0.00
1 / Correlation Value
0
45
90
135
180
225
270
315
ANGLE IN DEGREES

Appendix G(g)



Correlation of Waveforms to References
315° Waveform
4.00
3.50
3.00
2.50
2.00
1.50
1.00
0.50
0.00
1 / Correlation Value
0
45
90
135
180
225
270
315
ANGLE IN DEGREES

Appendix G(h)

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13. ABSTRACT *(Maximum 200 words)*

The EVAS provides warning for hearing-impaired motor vehicle drivers that an emergency vehicle is in the local vicinity. Direction and distance to the emergency vehicle are presented visually to the driver. This is accomplished by a special RF transmission/reception system. During this phase the receiver and transmitter from Phase I were updated and modified and a directional antenna developed. The system was then field tested with good results. Static and dynamic (moving vehicle) tests were made with the direction determined correctly 98% of the time.

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