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RADC-TR-80-275
Final Technical Report
August 1980



TROPOSCATTER ANTENNA POSITIONER

University of Wisconsin

W. P. Birkemeier
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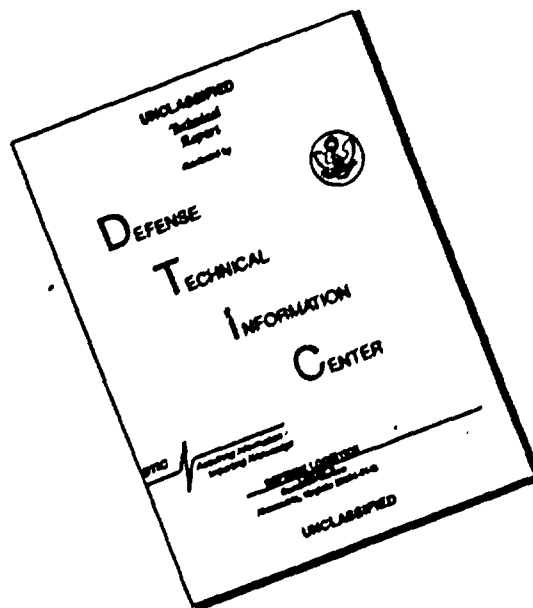
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1. REPORT NUMBER RADCR-80-275	2. GOVT ACCESSION NO. AD-H091778	3. RECIPIENT'S CATALOG NUMBER
4. TITLE (and Subtitle) TROPOSCATTER ANTENNA POSITIONER	5. TYPE OF REPORT & PERIOD COVERED Final Technical Report, Sep 77 - Feb 80	6. AUTHORING ORG. REPORT NUMBER N/A
7. AUTHOR W. P. Birkemeier A. B./Fontaine	8. CONTRACT OR GRANT NUMBER(s) F30602-77-C-0148	
9. PERFORMING ORGANIZATION NAME AND ADDRESS The University of Wisconsin Madison Wisconsin 53706	10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS 62702F 45192146	
11. CONTROLLING OFFICE NAME AND ADDRESS Rome Air Development Center (DCCT) Griffiss AFB NY 13441	12. REPORT DATE August 1980	
14. MONITORING AGENCY NAME & ADDRESS (if different from Controlling Office) Same	13. NUMBER OF PAGES 275	15. SECURITY CLASS. (of this Report) UNCLASSIFIED
16. DISTRIBUTION STATEMENT (of this Report) Approved for public release; distribution unlimited.		15a. DECLASSIFICATION/DOWNGRADING SCHEDULE N/A
17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report) Same		
18. SUPPLEMENTARY NOTES RADCR Project Engineer: Paul Sierak (DCCT)		
19. KEY WORDS (Continue on reverse side if necessary and identify by block number) Troposcatter Doppler Spectrum Alignment Techniques Computer Programs Software		
20. ABSTRACT (Continue on reverse side if necessary and identify by block number) This report covers results of a contract to provide for a rapid and accurate alignment procedure of troposcatter antennas. An experimental model embodying a micro-processor based computation routine at the receive antenna and a stable transmitted frequency allows for alignment of antennas along the great circle path. The test data shows that an alignment accuracy of better than $\pm 1/3$ degree or approximately $\pm 1/10$ of a beamwidth could be accomplished in approximately 10 minutes.		

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ABSTRACT

Under a previous effort "Antenna Alignment Study," F30602-74-C-0276, a technique was formulated to align troposcatter antennas by swinging the beams while noting the symmetry of the received Doppler spectrum. The measure of symmetry was computed from the cross-correlation of the quadrature components of the received complex envelope signal. A concern in this study was the need for synchronization of the transmitter and receiver clocks. This report details the computation of the Doppler spectrum skew which is fairly insensitive to clock offset. A microprocessor was designed and programmed to provide a real-time skew estimate that could be used as an alignment indicator. Tests on the RADC Verona-Ontario link employing TRC-97 radios showed that alignment accuracy better than $\pm 1/3^\circ$ or approximately $\pm 1/10$ beam-width could be accomplished in about 10 minutes. It was found necessary to set the average frequency shift equal to zero by adjusting the receiver clock in order to estimate spectrum skew correctly using the algorithm presented here.

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Program Name	Description	Pages for FORTRAN Program	Pages for 8085 Program
MAIN	1) Input data (from tape if FORTRAN; A/D or PROM in 8085 2) Pass x,y data samples to other routines PROCESS, DISPLAY, UPDATE, etc.	B.2-3	B.17-18
PROCES	Process an xy data sample and accumulate averages for 1 sec. Includes: 1) Test of xy data sample for minimum signal value 2) Derivative calculations $\dot{x}, \dot{y}, \ddot{x}, \ddot{y}$ 3) One second averages for correlation, skew, and slope variables	B.4-6	B.59-65
DISPLAY	Convert long-term average of correlation, skew, and slope variables to a form suitable for display	B.6-7	B.26-29
INITIAL	Initialize all system variables and parameters	B.7-8	B.23
UPDATE	Add one second averages fo alignment variables formed in the routine PROCES to long-term averages	B.8-11	B.66-72
STARTUP	Program begins here after a manual reset ..	N.A. (not applicable)	B.57-59
ASCHED	Schedules 8085 routines: A/D every 1/100 sec. Front panel scan every 1/10 sec. Display update every sec.	N.A.	B.18
RDFRNT	Scan front panel switches	N.A.	B.19
ATOD	Input analog data	N.A.	B.20-23
DISBCD	Convert a floating point number to BCD format and display it	N.A.	B.23-26
	Debugging monitor used for program development	N.A.	B.31-48
	9511 arithmetic routines	N.A.	B.48-50

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Evaluation

This report summarizes work aimed at providing a rapid and accurate alignment technique for troposcatter antennas. An initial program, "Antenna Alignment Study," performed under contract F30602-74-C-0276 established a basis for a new alignment procedure that is based on received spectrum symmetry rather than received power level techniques presently employed. This initial work though was validated using a Hewlett Packard correlator which was not optimized for inclusion into existing troposcatter radio systems.

The work under the present contract was aimed at providing for an alternate implementation that will ultimately provide for a fieldable piece of hardware that will accomplish the goals of rapid and accurate alignment. Besides providing for rapid alignment it is important to realize that the technique provides for accurate alignment on the great circle path while power level based techniques can only provide for short term alignment which is not necessarily the great circle path and thus requires updating.

Before fielding several issues require further work, these are validation of the alignment technique over the full breadth of troposcatter medium variations, reworking of the experimental model to eliminate erroneous results due to aircraft and further optimizing the implementation.

Based on results and the recommendations for future work the improvement to troposcatter alignment is viable and much closer to delivery to the field.



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Project Engineer

Section 1

INTRODUCTION AND SUMMARY

Objectives of the Program

Until recently, azimuthal alignment of troposcatter antennas was accomplished by averaging received power to find the proper pointing angle of each antenna. The procedure is arduous, slow, and often inaccurate because of the deep, persistent fading present in the channel. In earlier work, Birkemeier and Sill [1] established a new technique for alignment. According to their method, a monochromatic signal is transmitted and the shape of the received Doppler spectrum is used to determine alignment. The original method involved moving the receiving antennas until the Doppler spectrum was symmetric about zero frequency. When this was accomplished, the receiving antenna was defocused and both antennas moved synchronously clockwise or counter-clockwise until the Doppler spectrum was again symmetric. The beams were then shown to be pointing along the great circle path.

The first Doppler measurements were obtained with a mini-computer that performed an FFT on the incoming data. This method required expensive equipment and operator interpretation of the shape of the spectrum. A less expensive method requiring little operator skill was sought. The next approach made use of the fact that if the spectrum is not symmetric about zero frequency, there will be a non-zero average frequency. The average frequency can be used as a measure of the symmetry of the Doppler spectrum and thus of antenna alignment. There are several known algorithms for the estimation of the average frequency. Birkemeier, Sill, and Fontaine adapted the correlation method [2] so that it could be run in real time on a microprocessor. However, the scheme has a particular disadvantage in that the average Doppler frequency $\bar{\omega}$ will be in error by an amount equal to any frequency difference between the transmitter and receiver clocks. Thus a substantial alignment error would occur with this method if the clocks were not in synchronism as shown by Sill [2].

In this program we provide a microprocessor design together with software to estimate the 3rd central moment or the skew of the Doppler spectrum. The skew provides an alignment indicator that is a direct measure of the symmetry of the spectrum about its mean frequency and therefore does not depend on exact clock synchronization.

In order to use the microprocessor with the TRC-97 radio it was also necessary to provide coherent signal sources for the TRC-97 equipment. This report also details the design of this auxilliary equipment.

Technical Approach to Doppler Moment Computations

In order to compute the skew, we will need to determine various spectral moments of the input process and it is worthwhile to review the significance of these moments.

The first moment, $\bar{\omega}$, is the sum of two independent frequency terms. The first component is due to the frequency difference between the receiver and transmitter clocks, and the second term is due to the lack of symmetry in the spectrum.

The second spectral moment $\overline{(\omega - \bar{\omega})^2}$ is a measure of the spread of the spectrum about the mean frequency.

The third central moment $\overline{(\omega - \bar{\omega})^3}$ is used to estimate the symmetry of the Doppler spectrum about its mean frequency. If the moment is zero, the spectrum is symmetric about the mean; if positive, the spectrum is skewed positively; if negative, the spectrum is skewed negatively.

We begin by defining the skew to be:

$$\text{Skew} = \frac{\int_{-\infty}^{\infty} (\omega - \bar{\omega})^3 S(\omega) d\omega}{\int_{-\infty}^{\infty} S(\omega) d\omega} \quad (1)$$

where $\bar{\omega}$ is the average frequency defined by [4]

$$\bar{\omega} = \frac{\int_{-\infty}^{\infty} \omega S(\omega) d\omega}{\int_{-\infty}^{\infty} S(\omega) d\omega} \quad (2)$$

Expanding (1) yields

$$\text{Skew} = \frac{\int_{-\infty}^{\infty} \omega^3 S(\omega) d\omega - 3\bar{\omega} \int_{-\infty}^{\infty} \omega^2 S(\omega) d\omega + 3\bar{\omega}^2 \int_{-\infty}^{\infty} \omega S(\omega) d\omega - \bar{\omega}^3 \int_{-\infty}^{\infty} S(\omega) d\omega}{\int_{-\infty}^{\infty} S(\omega) d\omega} \quad (3)$$

We now note that the received signal can be written as a narrow band process [5].

$$\begin{aligned} S(t) &= X(t) \cos \omega_c t + Y(t) \sin \omega_c t \\ &= \text{Re}\{[X(t) + jY(t)]e^{j\omega_c t}\} \end{aligned} \quad (4)$$

The quantity $X(t) + jY(t)$ is known as the complex envelope of the received signal and $X(t)$ and $Y(t)$ are known as it's "inphase" and "quadrature components" respectively. Calling the complex envelope $Z(t)$; we now proceed to determine the Doppler spectrum of the correlation functions of $X(t)$ and $Y(t)$ as follows:

$$Z(t) = X(t) + jY(t) \quad (5)$$

$$\begin{aligned} R_{ZZ}(\tau) &= E\{[X(t) + jY(t)][X(t+\tau) - jY(t+\tau)]\} \\ &= R_{XX}(\tau) + R_{YY}(\tau) + jR_{YX}(\tau) - jR_{XY}(\tau) \end{aligned} \quad (6)$$

If we assume that the amplifiers in the quadrature channels have equal gain and phase characteristics, the assumption $R_{XX}(\tau) = R_{YY}(\tau)$ can be made. If we further assume that $X(t)$ and $Y(t)$ are uncorrelated random variables, it can be implied that $R_{XY}(\tau) = -R_{YX}(\tau)$ [5]. Using these assumptions, (6) can be simplified to yield

$$R_{ZZ}(\tau) = 2(R_{XX}(\tau) - jR_{YX}(\tau)) \quad (7)$$

If we define $S(\omega) = F(R_{ZZ}(\tau))$, then (8) follows:

$$\int_{-\infty}^{\infty} S(\omega) e^{j\omega 0} d\omega = 2\pi R_{ZZ}(0) = 4\pi R_{XX}(0) \quad (8)$$

Papoulis [5] has shown that

$$R_{XY}(\tau) = -R_{XY}'(\tau) \quad (9)$$

This result will be used to calculate $\bar{\omega}$. The differentiation relationship of the Fourier transform allows us to write [7]:

$$\frac{1}{2\pi} \int_{-\infty}^{\infty} j\omega S(\omega) e^{j\omega \tau} d\omega = R_{ZZ}'(\tau) \quad (10)$$

Using this relationship, it follows that

$$\int_{-\infty}^{\infty} j\omega S(\omega) e^{j\omega 0} d\omega = 2\pi R_{ZZ}'(0) \quad (11)$$

$$\int_{-\infty}^{\infty} \omega S(\omega) d\omega = \frac{2\pi}{j} R_{ZZ}'(0)$$

We combine (9), (10), and (11) and use the fact that $R_{XX}(\tau)$ is even to find $\bar{\omega}$.

Thus

$$\bar{\omega} = 4\pi R_{XX}(0) \quad (12)$$

Using (10) twice, we get the second moment

$$\int_{-\infty}^{\infty} \omega^2 S(\omega) e^{j\omega 0} d\omega = -2\pi R_{ZZ}''(0) \quad (13)$$

Papoulis [5] has also shown that

$$R_{ZZ}(\tau) = R_{ZZ}''(\tau) \quad (14)$$

Since $R_{XY}(\tau)$ is an odd function, even derivatives at $\tau=0$ are zero. This allows us to combine (13) and (14) to yield:

$$\bar{\omega}^2 = 4\pi R_{XX}(0) \quad (15)$$

Finally, the third moment is computed in the same fashion.

$$\int_{-\infty}^{\infty} \omega^3 S(\omega) e^{j\omega 0} d\omega = \frac{-2\pi}{j} R_{ZZ}'''(0) \quad (16)$$

$$R_{ZZ}'''(0) = R_{ZZ}'''(0) \quad (17)$$

$$\bar{\omega}^3 = 4\pi R_{XY}(0) \quad (18)$$

We now combine (8), (12), (15), (18), and (3) to yield the formula for the skew:

$$\text{Skew} = \frac{R_{XY}'''(0)}{R_{XX}(0)} - \frac{3R_{XY}'(0)R_{XX}''(0)}{R_{XX}(0)R_{XX}'(0)} + \frac{2[R_{XY}(0)]^3}{[R_{XX}(0)]^3} \quad (19)$$

This is the theoretical formula for the relative skew. Several assumptions about the process have been made. They are $R_{XX}(\tau) = R_{YY}(\tau)$, $R_{XY}(\tau) = -R_{YX}(\tau)$ and that the $X(t)$ and $Y(t)$ processes are jointly stationary. For the computation algorithm to function properly, all of these assumptions must hold.

Practical Implementation of the Skew Estimator

The validity of each of the assumptions made earlier must be examined in light of the data. The most basic assumption made earlier is that the in-phase and quadrature components of the signal are uncorrelated random variables, that is $R_{XY}(0) = 0$. It is clear that if both channels possess a DC term, this assumption cannot hold. All amplifiers, no matter how carefully designed, will have some DC offset. This is especially true of amplifiers that must operate at frequencies close to zero. The software must preprocess the data to remove the offset.

The second major assumption is that $R_{XX}(0) = R_{YY}(0)$. This statement implies that the statistics of the channels are equal and that the gains of the amplifiers in both channels are equal. If this is true, we should be able to interchange the roles of $X(t)$ and $Y(t)$ and get the same results. This does happen with data that is not properly preprocessed. Certain terms are immune to differences in the X and Y channel gains. In practice, terms that contain both X and Y are immune to interchanging roles. Terms such as $R_{XX}(\tau)$ or $R_{YY}(\tau)$ do not have mixed terms and will be vulnerable to both short term and long term variations in the gain. These terms are particularly sensitive because they contain derivatives which accent the temporal variations in channel gain.

The long term variation in the difference in the amplifier gains can be taken care of in software by correcting the gain of the Y channel by the ratio of the rms value of $X(t)$ and $Y(t)$. This is slow agc which cannot compensate for the short term variations in $R_{XX}(0)$ and $R_{YY}(0)$. To correct for these short term variations we have obtained good results by taking the geometric mean of $R_{XX}(0)$ and $R_{YY}(0)$ and using this number to replace $R_{XX}(0)$ in the 2nd term in (19).

The following section describes the microprocessor implementation of these programming objectives.

Summary of Results

Tests of the Slope algorithm on the RADC link indicated that alignment could be accomplished to within about 0.3° or about $1/10$ of a beam width with a period of about ten minutes. The Slope Algorithm actually measures $\bar{\omega}$, the average Doppler shift and its success depends on exactly synchronized clocks.

The skew algorithm indicated that the antennas could be aligned to 1/3 degree or better, also in a time period of about ten minutes. These measurements were based on single observations of displayed skew which occur once each second. Visual averaging would lead to substantial improvement in expected accuracy, but such improvement is not quantified in this report. Exact clock synchronization was not required in alignment using skew indication. Instead, $\bar{\omega}$ was set equal to zero by varying the receiver clock before each skew measurement was made.

Organization of Report

The microprocessor hardware and software design is summarized in Section 2. Section 3 details the design of the frequency synthesizers and detector components necessary to interface the microprocessor to the TRC-97 and the TRC-170 equipment. The antenna alignment tests on the RADC link are presented in section 4. Appendix A contains the microprocessor drawings. Appendix B lists processor program. Appendix C presents complete drawings concerned with the design and fabrication of the synthesizer modules together with parts lists. Appendix D contains a more detailed theoretical description of the nature of troposcatter Doppler spectrum and the estimation of its parameters.

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Section 2

MICROPROCESSOR HARDWARE AND SOFTWARE

Introduction

A block diagram of the microprocessor hardware is shown in Fig. 2.1. The major components consist of the 8085 microprocessor, 4-2716 PROMS (one with test data, one with a debugging monitor and two with the alignment programs), AM9511 arithmetic processor, AD571 10-bit A/D converter, 8251 serial terminal and phone line interfaces for the program development and front panel switch, light and BCD display hardware. The schematics for the microprocessor system are in Appendix A. The system operates in five modes (reset, clock, slope, correlation, and skew) that are front panel push-button selectable by the operator.

The reset mode resets all hardware in the system and then uses the slope algorithm to process five seconds of digitized test data that is stored in a PROM. This provides a self test for the system. The clock mode is used to measure and display the offset frequency between the transmitter and receiver oscillators while the slope, correlation and skew modes are the three alignment algorithms. Front panel switches are scanned at a ten per second rate by the program and whenever a switch is found pressed the program is restarted using the algorithm that is indicated by the switch.

A five digit BCD display provides the operator with a floating point (four-digit mantissa, one-digit exponent) digital readout of alignment information. In all cases the displayed quantity represents an average that has been exponentially averaged with a time constant of several seconds. Although the time constant is several seconds the display is updated once a second with the new value of the running average.

The arithmetic processor allows arithmetic expressions to be programmed very easily -- much like a hand calculator. The 10-bit digitized values for the in phase and quadrature signals are immediately converted

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to a 32-bit floating point format by the arithmetic processor and all subsequent processing of these is done using the floating point format.

Two programs are listed in Appendix B. One in FORTRAN and one in 8085 assembly, are included. The FORTRAN program was used on a Raytheon 704 with tape recorded data to develop the processing algorithms. A deliberate attempt was made to maintain a close correspondence between the FORTRAN and the 8085 code and in most cases each line of the FORTRAN program is included as a comment in the 8085 instructions that are used to implement the FORTRAN expression.

General Signal Considerations

The inphase and quadrature signals are filtered with analog filters before they are digitized by the A/D system on the microprocessor board. The pass band of these filters is 0 to 30 Hz when the clock and skew algorithms are used and 0.3 to 20 Hz when the slope and correlation algorithms are used. In this section and in the program listings the inphase and quadrature signals will be called X and Y respectively.

The tape recorded data used in the FORTRAN program was sampled at a 100 sample per second rate. The programs were written with the intent of using a timer-drive microprocessor interrupt to sample X and Y every 0.01 seconds, accumulate data from 100 samples, and update the display once a second with the new data included in the running average. The timer interrupt was not implemented because the microprocessor program when operating at full speed result in approximately a 100 sample per second sampling rate.

The displayed variables for clock or antenna alignment algorithms are averaged quantities and hence statistical in nature. Two conditions exist that can introduce large errors in these averages. One situation is when X and Y simultaneously become very small and the other is reflections from

aircraft. The first situation is easily programmed: the incoming data are ignored whenever the magnitude of both X and Y are less than 2% of the maximum value.

The aircraft contamination is more difficult to detect. Aircraft introduce higher than normal frequencies into X and Y and these are particularly troublesome in the skew algorithm that requires the second derivatives of X and Y. In an attempt to eliminate data with aircraft contamination the frequency X and Y is estimated from each successive pair of samples and the data ignored whenever the frequency exceeds 8 Hz. The first and second derivatives are calculated using

$$\Delta X_N = \frac{X_{N+1} - X_{N-1}}{2} \quad \Delta^2 X_N = X_{N+1} - 2X_N + X_{N-1}$$

Thus whenever data are ignored, three valid samples are required before the derivative calculations can continue.

Measurements made on the inphase and quadrature detector outputs indicated that both contained non-zero dc components and that the gain of the two detectors varied slightly. Both effects are removed in the processing by computing long time dc and mean square values of the detector outputs and equalizing the sampled values with them:

$$X_i = X_i - X_{dc} \quad Y_i = (Y_i - Y_{dc}) \sqrt{\frac{\overline{X^2}}{\overline{Y^2}}}$$

As mentioned earlier an exponential averaging technique is used to form a running average of X_{dc} , $\overline{X^2}$, $\dot{X}Y$, skew, etc. If A_{N-1} is the current value of the running average and I_N is the new input to be added to the average the new average is computed from

$$A_N = A_{N-1} + \frac{I_N - A_{N-1}}{N}$$

8085 Processing - Overall View

The four FORTRAN subroutines described above perform identical processing in the 8085 microprocessor system. A brief description of additional routines is given in this section. The X, Y sample pair is obtained by the subroutine ATOD. The routine will return either test data that is stored in PROM or will sample live data with the A/D system. Test data is returned if the AC power system has just been turned on or if the operator has pushed the RESET button; otherwise, sampled live data is returned.

The front panel switches are scanned with the routine RDFRNT. If a switch is found depressed the system is reset and a logical flag corresponding to the switch is set. Other subroutines use the flags to determine the processing that is required.

A scheduler routine, SCHED, is used to direct the real-time processing. Every call to SCHED results in a call to ATOD for a new data sample and a call to PROCES to process the data sample. Every 10th call to SCHED results in a call to RDFRNT to scan the front panel switches and every 100th call produces a call to UPDATE and DISPLY. As mentioned earlier the system was designed for a programmable timer to produce a call to SCHED every .01 sec but this is not implemented. Instead each return from SCHED is immediately followed by another call to this routine.

Program Implementation of Clock and Alignment Algorithms

Under ideal conditions of noise free, zero doppler shift, transmission with the transmitter and receiver oscillators operating at a frequency difference of ω_d , the X and Y signals become:

$$x(t) = \cos \omega_d t \quad \text{and} \quad y(t) = \sin \omega_d t .$$

where N is the number of samples that have been taken. This algorithm acts like an RC averaging circuit with a time constant of NT seconds where T is the time between new inputs I_N . Thus new information is weighted heavily and old information is forgotten. In addition N is varied during the measurement starting at a value of one and increasing to a maximum of 16 or 60 depending on the variable that is averaged.

FORTTRAN Processing - Overall View

The FORTRAN program has four subroutines INITIAL, PROCES, UPDATE, and DISPLY for processing the data and each of these has a counterpart in the 8085 program. A program listing of the FORTRAN program is in Appendix A. The data used in the FORTRAN is read in from magnetic tape in blocks of 1024 digitized (12-bit), pairs of samples X_i, Y_i by the main program. The main program then passes one pair of samples at a time to the subroutine PROCES just as the A/D subroutine in the 8085 system would pass one pair to the 8085 PROCES routine. Each subroutine contains the processing necessary for all four algorithms: clock, slope, correlation, and skew but uses logical flags to select only the processing that is necessary for the algorithms in use when the subroutine is called.

The subroutine INITIAL is used to initialize all arrays, exponential averaging parameters, etc. each time a run is started or whenever the operator decides to reset or clear the program of accumulated data.

The PROCES routine is called once for each X, Y pair of samples. Data are tested for validity, derivatives calculated, one-second sums of dc and mean square values of X and Y are tabulated for the 100 samples that are passed to PROCES. After 100 samples (equivalent to 1 second of data) are processed, the accumulated runs are used by UPDATE to add to the running averages of these variables and the parameters in PROCES are reset to allow for the accumulation of the next 100 samples. After the running averages are updated the routine DISPLY is called to display the new values.

The frequency ω_d can be obtained by defining $\theta = \tan^{-1}(y/x)$ and calculating

$$\frac{d\theta}{dt} = x \frac{dy}{dt} - y \frac{dx}{dt} = \omega_d.$$

In terms of signals that are sampled every Δt sec the difference frequency is estimated as

$$f_d \approx \frac{\Delta\theta_i}{2\pi\Delta t} = \frac{x_i \Delta y_i - y_i \Delta x_i}{2\pi\Delta t}$$

The running average (16 sec time constant) of the quantity $\frac{\Delta\theta_i}{2\pi\Delta t}$ is continuously displayed when the clock algorithm is used.

The slope algorithm displays a running average of the quantity $\frac{(\Delta x_i) Y_i}{x^2 \Delta t}$. The X_i and Y_i used in this expression and in all calculations are the sample values with the dc removed and normalized to the same mean square value. The $\overline{x^2}$ term is a running average of the mean square value of x that is used to normalize the slope statistic so that it is independent of signal strength.

The cross correlation between continuous function of time $x(t)$ and $y(t)$ is defined as $R_{xy}(\tau) = \overline{x(t)y(t-\tau)}$. For sampled signals the cross correlation becomes

$$R_{xy}(m\Delta\tau) = \overline{x(k\Delta t)y(k\Delta t-m\Delta\tau)}.$$

The cross correlation is computed for $-1.5 < \tau < 1.5$ seconds by using $\Delta t = \Delta\tau = 0.03$ sec and a table to store the 100 values for $R_{xy}(m\Delta\tau)$ that correspond to $-50 < m < 50$. Since .03 is three times the sampling interval, every third sample of raw data is used in calculating the cross correlation. These data samples are placed in a "delay line", first in first out type table, 50 samples long for X and 100 samples long for Y . As each X sample reaches the bottom of the table it is correlated with the 50 samples of Y forward and backward in time and then discarded. Since the alignment criteria

with the correlation algorithm is to minimize the absolute slope about $\tau=0$, the slope of a line from the origin to the maximum value of $R_{xy}(m\Delta\tau)$ is the quantity that is displayed to the operator.

The equation for skew was developed earlier; and although it is a straightforward arithmetic calculation, it is a sensitive measure with a larger variance than the slope or correlation measures. This is due to the subtraction of terms in the expression that frequently are very large and nearly equal.

Summary and Recommendations

The hardware and software have proven to be very trouble free and reliable. The higher doppler frequencies that were present in data on the RADC link, but not on the UW link, frequently caused the system to reject the data as aircraft contamination. Other means of detecting aircraft reflections should be tried and the most promising of these is to look for rapid fluctuations in envelope that exceed a peak to peak threshold.

Section 3

INTERFACING WITH TRC-97 AND TRC-170

Frequency Synthesizer Design

The Air Force troposcatter TRC-97 and TRC-170 equipment requires a frequency synthesizer that will convert a precision frequency reference into coherent 70 MHz and 50 MHz outputs. The receiver also requires a quadrature detector for its 70 MHz intermediate frequency signal. A block diagram of the system is shown in Appendix C, Figures C1 and C2. As the figures show the radios are to use their own internal 50 MHz and 70 MHz sources when operated in the non-alignment mode and the synthesizer's coherent oscillator signals in the alignment mode. Coaxial relays are included in the aligner for switching these signals. The 50 MHz and 70 MHz synthesizer outputs must have a frequency stability of at least 1 part in 10^{10} , and both the 50 MHz and the 70 MHz outputs must provide 18 dbm to a 50 ohm resistive load. Harmonic content within 10 MHz of the fundamental was to be down at least 20 db.

These specifications were met with an "indirect" frequency synthesizer, employing phase locked loops (PLL) for frequency multiplication. The basic designs for these synthesizers are shown in Figure C3 and C4.

The Precision Quartz Clock

The precision 5 MHz crystal oscillator which provides the source for the phase lock loops is a double-oven Vectron model CO-246V-1 oscillator with frequency stability of 1 part in 10^{10} . The oscillator frequency can be externally controlled by a DC voltage. The connection diagram for the frequency standard is shown in Figure C5.

The 25 MHz and 35 MHz Phase Lock Loops

The 5 MHz reference signal is power divided between the 50 MHz and 70 MHz phase locked loops. The basic phase locked loop design is shown in Figure C6. The voltage controlled oscillators, VCO's, oscillate at half

the desired output frequency (35 MHz for the 70 MHz unit and 25 MHz for the 50 MHz unit) because integrated circuit PLL's are limited to a maximum frequency of about 35 MHz. A balanced mixer is used to double the VCO frequencies to the desired 70 MHz or 50 MHz values.

The XR215 PLL integrated circuit chosen consists of a phase detector and VCO. The lowpass filter corner frequency for the control signal and the VCO frequency range are determined by an external capacitance connected to the XR215 chip. The frequency divider is the 93S05 variable modulo counter. For the 70 MHz circuit it is wired to frequency divide by 7; for the 50 MHz circuit it is wired to divide by 5.

The 70 MHz phase locked loop schematic is shown in figure C7. Due to the variation in characteristics from one VCO to another the VCO capacitor C107 must be selected to give the correct VCO frequency range. C107 should be no larger than 12 pF. C105 and C106 determine the corner frequency of the tuning voltage low pass filter. The filter transfer function has a single pole at $1.33 \times 10^6 \frac{\text{rad}}{\text{sec}}$ or 212 KHz. R100, the VCO free run frequency adjusting resistor, should be adjusted to allow the PLL to achieve lock. Q100 is part of a common emitter amplifier that raises the VCO signal amplitude to the level required to drive the Schmidt Trigger, U101. The Schmidt Trigger converts the analog 35 MHz signal to a TTL compatible signal to prevent erratic operation of the frequency divider which follows it. The frequency divider, U102, takes the 35 MHz signal from the Schmidt trigger and divides its frequency by 7. Thus the output of U102 has a frequency of 5 MHz. This signal is coupled to the phase detector in U100 where it is phase compared with the 5 MHz frequency standard. The output of the Schmidt trigger is also coupled to U101 which is part of an emitter follower power amplifier. This power amplifier drives a double balanced mixer, U103, with the 35 MHz signal. The mixer is used as a frequency doubler to generate the desired 70 MHz signal. This signal is passed to the 70 MHz amplifier for amplification and band pass filtering.

The 50 MHz phase locked loop schematic diagram is shown in Figure C8. and its parts list follows. The 50 MHz PLL is similar to the 70 MHz PLL. The only differences are a larger coupling capacitance, C214, a different wiring of the frequency divider to divide by 5 instead of 7 and a resistor, R218, in series with the 50 MHz output. This resistor attenuates the 50 MHz output to prevent it from saturating the 50 MHz amplifier and generating excessive harmonic levels.

The 70 MHz Amplifier

The original design with minor modifications was used for the 70 MHz amplifier. The schematic diagram is shown in Figure C9. The output of the amplifier goes to the detector in the receiver and to jack J1 on the transmitter A-17 assembly by way of the 70 MHz coax relay.

The original design with minor modification was also used for the 50 MHz amplifier. The schematic diagram is shown in Figure C10. R400 was added to the amplifier input to eliminate self oscillation. L400 and C400 were also added to the amplifier input as a series tuned circuit to reduce the 75 MHz spurious output from the 50 MHz phase locked loop.

The Quadrature Detector and Low-Pass Filter

The receiver also requires a synchronous quadrature detector. This detector must demodulate the 70 MHz intermediate frequency signal into its quadrature components. These signals must then be low pass filtered with a corner frequency of 20 Hz. The schematic diagram of the detector is shown in Figure C11. The low pass filter contained in the detector was designed to have a voltage gain of 10. This should result in 2 volt peak-to-peak output levels from the X and Y channels under nominal channel conditions. This is the level required to drive the following stage which is a DC blocking filter. The low pass filter is a Sallen and Key configuration designed to have the following transfer function:

$$H(s) = \frac{10}{s^2 + 178\left(\frac{\text{rad}}{\text{sec}}\right)s + (1.58 \cdot 10^4)\left(\frac{\text{rad}}{\text{sec}}\right)^2}$$

This results in a two pole low pass filter with a corner frequency of 20 Hz and a Q of 0.707. This Q gives the steepest possible skirt without any passband ripple.

Low-Pass, Band-Pass Filter

The output of the detector module is next fed to the Low-Pass, Band-Pass Filter. One section of this filter additionally attenuates frequencies above 20 Hz by 6 db/octave and outputs this low-pass signal to the microprocessor to be used in the "slope" program to compute the average Doppler frequency or the slope of the cross-correlation function at zero lag. The band-pass filter section adds a high pass filter in cascade to the low-pass section to remove frequency components below 0.3 Hz. The band-pass filtered outputs are fed separately to the microprocessor to be used to compute the average frequency whenever a strong specular signal at zero Doppler shift exists in the signal. In tests at RADC this output was not used. The schematic diagrams and parts layout are found in Fig. C11 and Fig. C23, respectively.

The Power Supply

The frequency synthesizer power requirements are 5 volts for the phase locked loops, 12 volts for the phase locked loops, amplifiers and detector, 24 volts for the frequency standard and -12 volts for the detector. The power supply schematic diagram is shown in Figure C12. The power supply module contains the bridge rectifiers, filter capacitors and 24 volt regulator. The outputs of the module are 24 volts regulated for the frequency standard and relay coils, 40 volts unregulated for the 12 volt and 5 volt regulators and -20 volts unregulated for the -12 volt regulator. The 12 volt, 5 volt, and -12 volt regulators are preceded by voltage dropping resistors to reduce the voltage regulator dissipation. The 12 volt, 5 volt, and -12 volt

regulators are mounted on the frequency synthesizer base plate for heat sinking. Because the transmitter does not have a detector circuit, components C601, R602, CR601 and U603 are not used in the transmitter power supply.

Each phase locked loop, amplifier and detector circuit was built on its own printed circuit board and mounted in its own box. All power supply connections to the boxes are made with feed through capacitors. All other connections to or between boxes are made with TNC connectors or feedthrough terminals. This connection scheme is used to eliminate electro-magnetic interference between the frequency synthesizer and other system components.

Test Results

When the synthesizers were tested, the phase locked loop selected capacitors and VCO tuning potentiometers were adjusted while observing the 50 MHz and 70 MHz signal outputs with a spectrum analyzer. These components were adjusted to values that allowed the phase locked loops to achieve lock at 50 MHz and 70 MHz after the internal temperatures stabilized with an ambient temperature of 25° C. The 50 MHz and 70 MHz output spectral purities and signal levels are recorded in photographs C13 and C14. The units were each driving a 50 ohm resistive load for these results. It can be seen from the photographs that the signal output level requirements are met with these synthesizers.

Additional Suggestions

The spectral purity of the frequency synthesizers could be improved at an added cost. The noise sidebands on the 50 MHz and 70 MHz signals could be improved by replacing the RC VCO's in the phase locked loops with voltage controlled crystal oscillators, VCXO's. VCXO frequencies would be more stable with age and temperature variations. The use of VCXO's should also reduce the VCO frequency variations from IC to IC. This would eliminate the need for selected capacitors in the PLL circuits. The XR215 can be used as a VCXO.

The spectral components that are spaced every 5 MHz about the 50 MHz and 70 MHz desired signals could be greatly attenuated by installing crystal filters between the phase locked loops and amplifiers. An additional stage would need to be added to compensate the insertion loss.

Printed Circuit Boards

Art work for PC boards used in the synthesizers are shown in Appendix C Figures C15-C18.

Parts Layout

Layout of parts for the various synthesizer modules are shown in Figures C19-C22. The 50 MHz PLL is identical to the 70 MHz parts layout shown, except its parts are numbered in the two-hundreds.

Parts Lists for Synthesizer and Filter Modules

See pages C.28 through C.36, Appendix C.

TRC-170 Interfacing Requirement

The TRC-170 equipment is fully coherent using automatically controlled clocks to drive its frequency synthesizers. As such there is no problem in interfacing this radio with the aligner's microprocessor. The aligner's detector package, however, must be driven by the TRC 170's detector, which in turn must be driven by a 70 MHz source. This source is provided by the aligner's synthesizer using its own internal 5 MHz standard or by substituting the TRC-170's 5 MHz standard. The latter would be preferred, since it eliminates the need to set the clock according to skew estimation procedure outlined in Section 4.8.

Section 4

TEST RESULTSUW Link Tests

The antenna aligner tests were performed in accordance with the Proposed Test Plan/Procedures (clin item 0004). Test efforts on the 940 MHz University of Wisconsin link held on 12/14/79 and 12/15/79 were largely unsuccessful as a result of more-or-less continuous aircraft intrusions on both days. Moreover, it proved impossible to skew the Dopple spectrum by azimuthally displacing the receiving antenna. This is possibly explained by the occurrence of equal and opposite cross-path wind fields in scattering layers of equal cross-section, although we have no meteorological proof of such a situation existing.

UW Test of Skew Algorithm

We observed the skew algorithm to correctly reflect the changing slope of the Doppler spectrum when an aircraft crossed the common volume, but unfortunately, the skew measure also was influenced by the clock setting contrary to theoretical expectations. We speculated that this was the result of a program deficiency that causes the 1st and 2nd derivatives of the quadrature components of the received data to be underestimated when the fading is too slow. Fading on these days was extremely slow due to a strong specular component in the received signal. We expect faster fading at 4 GHz to mitigate this problem.

UW Test of Slope Algorithm

Aircraft traversals provided the only way to test the slope algorithm. The slope changed sign as the aircraft crossed the great-circle in every case qualitatively indicating proper operation.

UW Test of Clock Setting Algorithm

This algorithm was tested by varying the 5 MHz clock frequency in the receiver. The average frequency displayed agreed with the amount of observed clock offset.

UW Test of Correlation Algorithm

The correlation algorithm is protected from false data during on aircraft intrusion by a scheme that rejects an entire 30 second data stream if the phase of the received signal changes by more than eight degrees from one sample to the next (0.1 sec). Since aircraft were particularly numerous, this algorithm succeeded only on brief occasions and could not be evaluated realistically. Later it turned out that fast fading on the RADC link was able to activate the protective device and therefore defeat the estimation algorithm.

Conclusions from UW Tests

While these tests were far from satisfactory they indicated that we could expect proper operation under the conditions to be expected on the shorter, higher frequency RADC link. We therefore begin tests in Verona, New York on 1/17/79.

RADC Link Tests

The RADC TRC-97 link from Ontario, New York to Verona, New York was operated at 4880 MHz using Cesium-beam clocks and HP synthesizers to establish all signal sources in the transmitter and receivers. Detection was accomplished using the coherent quadrature detector package of the Antenna Aligner together with the 70 MHz reference signals established by the 5 MHz Vectron oscillator also contained in the aligner. Observation

of the X-Y display of the quadrature components on an oscilloscope revealed random phase jitter of approximately $\pm 45^\circ$ on the received signal. This was pure phase jitter that apparently was generated in the TRC-97 transmitter. (No phase jitter was observed when the receiver was fed the 4880 output from a coherent source). This phase noise caused us no problem because most of its frequency components were above 20 Hz and were therefore eliminated by the low-pass filters in the aligner. Nevertheless, such jitter constitutes a potential source for error if the filter bandwidths are expanded as explained later. Other troubles in getting the link up were due to cold environmental temperatures ($< 40^\circ \text{ F}$) at the Verona site. In particular the aligner's phase-lock loops were reluctant to lock and the microprocessor clock chip refused to operate until raised to room temperature.

Test of Slope Algorithm, Step 1

The slope algorithm measures the presence or absence of a cross-correlation function by measuring its slope at the origin. Zero slope implies that the Doppler spectrum is symmetrical about zero frequency shift. In order to provide unambiguous results, the shift due to clock error must be zero. In these tests the signal contained no zero-Doppler-shifted component so the receiver clock could not be set to zero beat. We therefore set the aligner's clock to zero beat with the Cesium-beam standard supplying the receiver. The slope display was sampled and recorded each second for 60 seconds. Three such runs were made -- the first two about one hour apart, and the third on the next day. The data summary is presented in Table 1 below. In each case the transmitting antenna was set on the great-circle and both antennas were focused. The slope, S , in $\text{volts}^2/\text{sec}$ is given for the receiving antenna displaced $\pm 1^\circ$ from the great-circle.

Table 1. Slope Estimation for Step 1 of Alignment Process							
Rec. Ant. = 1° N of great-circle					Rec. Amt. = 1° S of Great Circle		
Run 1	3:00 EST 12/21/79	$\bar{S} = -1.68$	$\sigma_S = .27$	$\bar{S}/\sigma_S = 6.2$	$\bar{S} = 1.09$	$\sigma_S = .14$	$\bar{S}/\sigma_S = 7.8$
Run 2	4:30 EST 2/21/79	$\bar{S} = -1.08$	$\sigma_S = .33$	$\bar{S}/\sigma_S = 3.20$	$\bar{S} = 2.56$	$\sigma_S = .22$	$\bar{S}/\sigma_S = 11.6$
Run 3	8:30 AM 12/22/79	$\bar{S} = -1.008$	$\sigma_S = .33$	$\bar{S}/\sigma_S = 3.05$	$\bar{S} = 1.46$	$\sigma_S = .39$	$\bar{S}/\sigma_S = 3.75$

From the table above we estimate the alignment accuracy for the first step in the procedure by noting that the ratio of the mean slope $\bar{S}$ to the standard deviation of the data, σ_S , has a minimum value of 3.05 in Run 3. This means that 1 degree of receiving antenna offset resulted in a mean slope reading about three times larger than the rms fluctuations in the displayed slope data. The expected accuracy is thus about 1/3 of a degree. That is, treating the 1 second samples as independent data, a single sample of the slope measure has probability of .64 of lying within $\pm$ one standard deviation of the true mean. Note that in Run 1 on the previous day about twice this accuracy would have been obtained. Note, however, that the increase appears to be due to increased $\bar{S}$ when the receive antenna was 1° North and due to decreased rms fluctuations when the antenna was 1° South. Presumably spatial and temporal nonstationarities in the scattering structure were responsible for these variations. The minimum obtained accuracy of 1/3° is less than the back-lash in the TRC-97 antenna gearing and is therefore probably sufficient in this case. The actual obtainable accuracy is much better, since the operator of the aligner mentally averages as he observes the displayed numbers. The rms value of the fluctuations drop with $1/\sqrt{N}$ where N is the number of independent observations averaged. While the one-second data values displayed are not entirely independent, a substantial improvement is certain to be obtained in actual practice by a experienced operator.

Test of Slope Algorithm, Step 2

Step one of the alignment procedure accomplishes straddling antenna beams. In step two of the procedure the receiving antenna is defocused and the beams are synchronously swung until the cross-correlation slope is again averaging zero. The beams are then in great-circle alignment. Several runs were made using about 60 samples of slope data with the beams displaced by various equal and opposite incremental angles starting at a maximum of $\pm 1\ 1/2^\circ$. The results are tabulated in Table 2 below. All five runs were made within one hour from 12:10 EST to 13:10 EST on 12/21/79.

Table 2. Slope Estimation for Step 2 of Alignment Procedure						
Run	Antenna Setting from Calibration on Mast.		$\bar{S}$	σ_S	$\bar{S}/\sigma_S$	Remarks
1	R = $1\ 1/2^\circ$ N Focused	T = $1\ 1/2^\circ$ S Focused	-.04	.308	---	Both antennas focused
2	R = $1\ 1/2^\circ$ N Defocused	T = $1\ 1/2^\circ$ S Focused	+1.19	.22	5.4	Receiving antenna defocused
3	R = 1° N Defocused	T = 1° S Focused	+.98	.20	4.9	
4	R = $1/2^\circ$ N Defocused	T = $1/2^\circ$ S Focused	+.11	.24	---	Both antennas very close to great circle
5	R = 0° Defocused	T = 0° Focused	-.783	.32	---	

Note from Run 4 in Table 2 that the data implies the correct antenna settings would be about $1/3^\circ$ N for receiving antenna and $1/3^\circ$ S for the transmitting antenna. This is just the amount of the apparent back-lash in the antenna pointing gear train and accounts for the displacement due to the wind loading on each antenna. (A north to south wind prevailed during the experiment.)

Test of Skew Algorithm, Step 1

It was noted in preliminary checks of the skew estimation algorithm that the measure was sensitive to the receiver clock setting which is contrary to theoretical expectations. It was possible to align the antennas with an improper initial clock setting using the skew algorithm, however. To accomplish this we simply set the clock to force the average "Doppler" frequency $\bar{\omega}$ equal to zero as indicated by the clock mode display of the aligner. With $\bar{\omega}$ equal to zero the skew mode of the aligner then displays the true skew measure of the received spectrum. To estimate the alignment accuracy we measured skew with the receive antenna set off the great circle. Data for two antenna positions are shown in Table 3. The transmit antenna was set to the zero degree mark or a true position of about $1/3^\circ$ N. (The wind now was south or north across the link). The receive antenna was set at 1° S or a true position of about $2/3^\circ$ S for Run 1. For Run 2 the receive antenna was set at 1° N or a true position of about $1\ 1/3^\circ$ N. The skew measures correctly reflect these antenna positions as shown in Table 3 below.

Table 3. Skew Estimation for Step 1 of Alignment Procedure					
Run	Antenna Setting Both Antennas Focused		$\bar{SK}$	σ_{SK}	$\bar{SK}/\sigma_{SK}$
1	R = $2/3^\circ$ S true	T = $1/3^\circ$ N true	-176	3.95	4.4
2	R = $1\ 1/3^\circ$ N true	T = $1/3^\circ$ N true	+428	114	3.75

In order to estimate the alignment accuracy from this data note from Run 1 that a displacement of the receiving antennas of only $1/3^\circ$ from its correct straddle point produced a significant average skew value $\bar{SK}$ that was 4.4 times larger than its standard deviation. This implies an alignment accuracy of about

1/12 degree. In Run 2 the receive antenna was displaced twice as far as in Run 1 and also toward the displaced transmit antenna. As a result the Doppler spectrum is strongly skewed in the opposite direction to that of Run 1. Note $\overline{SK}$ has correspondingly changed sign and increased in magnitude over Run 1. The standard deviation of the skew has also increased so that the ratio of average skew to its standard deviation remains close to 4. More data would be required to determine whether or not this ratio tends to remain constant with displacement of the composite antenna pattern. Nevertheless, these data suggest that the Skew estimator is as good or better than the Slope estimator as an alignment indicator, but with the advantage that absolute clock setting accuracy is not a requirement with Skew. The Skew approach does depend on the extra step of setting the clock to make $\overline{\omega}$ equal to zero before each Skew measure is attempted. This step is not time consuming and we estimate from this experience that complete Step 1 and Step 2 alignment procedures could be completed in about ten minutes with the TRC-97 as configured in these experiments. Because of time limitations it was not possible to attempt Step 2 of the alignment procedure. There is no reason to believe that the results would differ greatly from those experienced with the slope algorithm since ratios of mean to standard deviation of about 4 were observed with both Slope and Skew measurements.

Overview of Test Results

Both the "slope" and "skew" estimation techniques were shown to provide a sensitive measure of antenna misalignment and a fast and accurate method of alignment based on these estimates was demonstrated. The "correlation" technique could not be tested because the program was devised to eliminate fast fading data as a safeguard against aircraft signals. The fading rate on the RADC link was so fast that the scheme was nearly always defeated. However, both "slope" and "skew" were sufficiently successful to perhaps negate the need to test the correlation approach.

Accuracy and Alignment Time

Both the successful modes of operation provided azimuth alignment to within $\pm 1/3$ degree of the great circle, which is about the same as the backlash on the TRC-97 antenna gear drive. This degree of accuracy is implied by the fact that at $1/3$ degree error, the average value of the displayed quantity equaled the rms value of its 1-sec fluctuation. The displayed quantity is derived from 16 seconds of data, and this is the amount of time required for the rms fluctuation to reach their minimum value after data taking is started. It is important to point out once again, that an operator of the aligner would tend to mentally average the 1-second displayed numbers before making an estimate of the misalignment. If 15 seconds of observations are permitted, a total of 30 seconds would elapse before each adjustment of the beam could be made. Assuming about ten such adjustments are necessary to reach final alignment, 300 seconds of data processing time would be required to align one antenna using the slope technique. Data averaging by the operator will also tend to improve the alignment accuracy, so it would be reasonable to predict that accuracy at least twice as good as $\pm 1/3^\circ$ could be achieved in practice.

Comparison of Slope and Skew Methods

The tests indicated that the slope and the skew methods achieved essentially the same degree of accuracy. However, the slope method is not usable unless some method is available to synchronize the transmitter and receiver clocks. If the propagation medium provides a strong specularly reflected signal with zero Doppler shift the aligner's clocks can be synchronized by zeroing the average frequency read on the display using the aligner's "clock" mode. In the absence of a specular signal it is necessary to use the skew mode and to correct the receiver clock at each new antenna position. Correcting the clock takes an additional 10 or 15 seconds at each antenna position which about doubles the time required to align

the antennas. An estimate of the total time to align using the skew mode is thus about 600 seconds.

Our experience at troposcatter frequencies near 1 GHz indicates specular layers are usually present. At the TRC-97 frequency near 5 GHz specular reflections are much less likely to occur. Based on this the skew algorithm is the only choice for the TRC-97 radios providing alignment is about 10 minutes for each antenna. Newer coherent radios such as the TRC-170 with atomic frequency control would be able to use the slope algorithm and achieve alignment in about 5 minutes for each antenna.

Connecting the Aligner to the TRC-97

The overall block diagrams in Fig. C.1 shows the aligner attached to the TRC-97 transmitter and receiver when the system is in the alignment mode. This mode is accomplished by placing the align-nonalign front panel switch in the align position which applies the IF output to the aligner's product detectors and substitutes the aligner's coherent frequency sources for the TRC-97's incoherent sources. The operating frequency in the alignment mode is 4880 MHz. Figure C.2a and C.2b show details of the relay switching of the signal lines from the TRC-97 to the aligner and back. In Fig. C.2a the port labeled "INCOHERENT 50 MHz IN" receives the TRC-97's 50 MHz synthesizer output that normally connects to J3A17. The port labeled "INCOHERENT 70 MHz IN" receives the TRC-97's 50 MHz synthesizer output that normally connects to J1 of the upconverter A17A4. The remaining signal lines and ports are self-explanatory and should be followed in installing the aligner with the TRC-97.

When the align-nonalign switch is in the nonalign position or when the aligner is powered-off the TRC-97 is returned to its normal configuration.

Step-by-Step Operating Procedure for Skew Method

1. Test the aligner's microprocessor by depressing the "test" pushbutton. The display should read 019(-2) which represents the decimal number 19×10^{-2} . If it reads correctly proceed to step 2.

2. With the aligner installed according to the preceding description and the align-nonalign switch in the align position the tropo signal should be acquired following the standard procedures outlined in the TRC-97 instruction manual. If desired, the inphase and quadrature outputs from the back panel of the aligner can be displayed on an oscilloscope to aid in visualizing received signal strength.
3. With both antennas focused the aligner should be placed in the clock mode by depressing the front panel pushbutton labeled "clock." The displayed average frequency, $\bar{\omega}$, should then be set to zero with the front panel clock frequency adjustment control.
4. With $\bar{\omega}$ set at zero the skew-estimate is made by depressing the "skew" pushbutton. The numbers displayed will vary randomly each second around an average value. (If an aircraft intrudes the common volume the skew measure will vary widely with time as the aircraft Doppler varies. The "reset" pushbutton will clear the memory allowing a new skew estimate to be obtained when the aircraft clears the antenna beams.)
5. When a stable skew estimate is obtained the receive antenna should be rotated about $1/2^\circ$ in one direction and fixed.
6. The clock mode is again used to set $\bar{\omega}$ equal to zero.
7. The skew mode is next used to reestimate the skew. If the average value has been reduced the correct direction of antenna rotation was chosen. Continue to reset the antenna in small increments until the skew display changes sign. This is evidence that the receiving beam has gone through the symmetrical straddling beam position.
8. With the receiving beam positioned to give zero average skew, the defocuser is attached. If the skew remains zero the beams are correctly aligned. If the skew becomes non-zero proceed to step 9.

9. The transmitter and receiver beams are now swung synchronously in, say, $1/4^\circ$ increments, both clockwise or both counter clockwise. An order-wire voice channel back to the transmitter is necessary to coordinate this synchronous beam swinging procedure. At each position $\bar{\omega}$ is set to zero and the skew measured. If the correct direction of rotation (CW or CCW) has been chosen, the skew measure will fall with each new increment until it reaches zero. At this point the two beams are producing a non-skewed Doppler spectrum indicating that they have met at the great-circle.
10. The receiving beam is then refocused to complete the alignment procedure.
11. It should be noted that if the incorrect direction of rotation had been chosen the skew would have increased as the beams diverged from the great-circle.

Step-by-Step Operating Procedure for the Slope Method

1. The signal should be acquired by following the standard methods outlined in the TRC instruction manual.
2. The atomically controlled 5 MHz receiver clock should be substituted for the 5 MHz quartz standard contained in the aligner. It is also assumed that the transmitter and receiver atomic standards are in zero-beat.
3. The aligner's microprocessor is tested by depressing the "test" pushbutton. The correct display is 019-2 or 19×10^{-2} .
4. Depressing the "slope" pushbutton will cause the aligner to read the slope of the crosscorrelation function of the quadrature components of the received signal. The slope will be non-zero when the antennas are misaligned.
5. The receiving beam should now be moved in azimuth about $1/2$ degree and fixed.
6. The slope is remeasured. If it has decreased the antenna movement was in the correct direction.
7. Continue to increment the antenna in the direction that decreased slope until the slope equals zero.

8. The receiving antenna is next defocused. If the slope remains zero both beams are correctly aligned. If the slope becomes non-zero proceed to step 9.
9. The transmitting and receiving beams are now swung synchronously, both clockwise or both counter-clockwise in say, 1/4 degree steps. If the correct direction is chosen the slope will decrease with each increment until it reaches zero. At this point the beams are correctly aligned.
10. The defocuser is removed to complete the alignment procedure.

Suggestions For Further Study

1. Deficiency of Skew Estimation Program

For reasons that are not entirely clear, the skew algorithm displays sensitivity to clock error. As noted in the introduction this is contrary to theoretical expectations, but may be due to the fact that the Doppler spectrum is bounded by an 8 Hz low-pass digital filter. Therefore when the $X(t)$ or $Y(t)$ spectrum is spread by offsetting the clock, either the upper or lower edges of the Doppler spectrum are attenuated depending on whether the clock is raised or lowered in frequency from zero beat. This could induce positive or negative skew respectively. Widening the filter bandwidth should be tried as a solution to this problem.

Another factor that should be considered is the possibility that the skew programming of the microprocessor is deficient in computing the first and second derivatives of the quadrature components of the received signal. Since these signals are first quantized by an A/D converter, successive samples falling in the same level bin will yield a first derivative of zero and three successive samples in the same bin will yield a second derivative of zero. Thus, when the signal derivatives tend to be small, they may often be under estimated. Slowly fluctuating signals may therefore lead to errors in the various terms that make up the Skew

equation. This equation yields a small difference from large numbers and is therefore quite sensitive to errors associated with the various terms. See equation (19).

During UW tests Doppler spreads were much narrower than on the RADC link. We experienced skew estimates sensitive to clock-offset particularly when the Doppler spectrum was extremely narrow as evidenced by a zero-Doppler spike in the spectrum, suggesting the derivatives as the primary cause. The 8 Hz filter cutoff apparently was no problem on the UW link, since we were able to obtain skew estimates insensitive to clock-offset on two occasions when the Doppler spectrum was narrower than 8 Hz, but free from a specular component.

If the above two problems prove incorrect as the cause of skew sensitivity to clock offset, it might be expedient to attempt to make clock correction (to force $\bar{\omega}$ to equal zero) an automatic process. Since $\bar{\omega}$ is outputted by the slope algorithm an estimate of it could be used to correct the receiver clock by feedback to the clock control voltage input. This would remove the extra step that otherwise must be provided by the operator.

2. Additional Testing Requirements

Time permitted a fairly limited amount of testing on the RADC link. During these tests the fading rate was high and signal characteristics were consistent for the several days of the tests. Testing under a variety of signal conditions should be attempted before decisions concerning filter bandwidths are finalized.

The skew algorithm was tested only with focused beams. It is reasonable to expect the final accuracy with a defocused beam to parallel the results of the slope algorithm test, however. Nevertheless, it would be appropriate to check this when additional tests are run under different signal conditions.

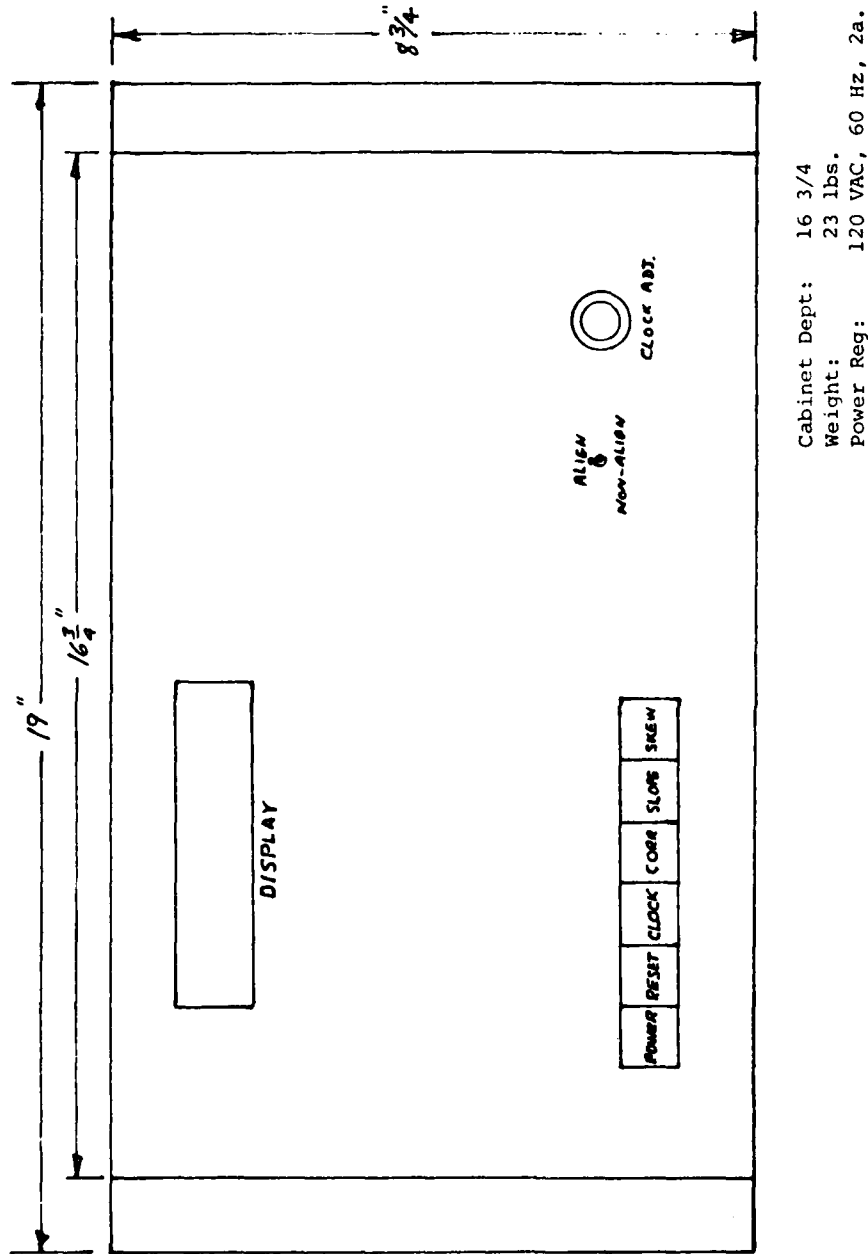


Fig. 4.1 Mechanical and Electrical Specifications

APPENDIX A

MICROPROCESSOR SCHEMATICS

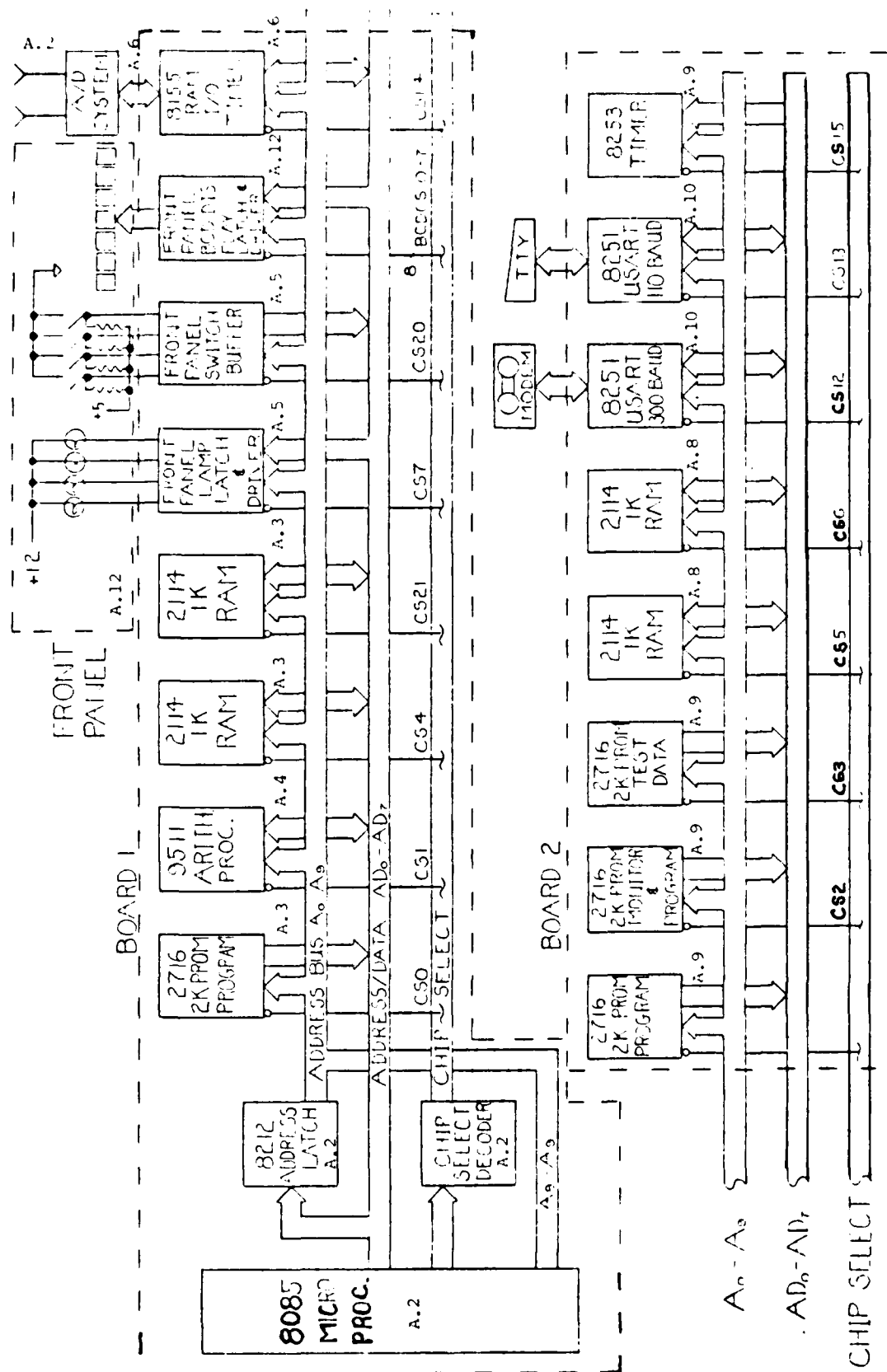
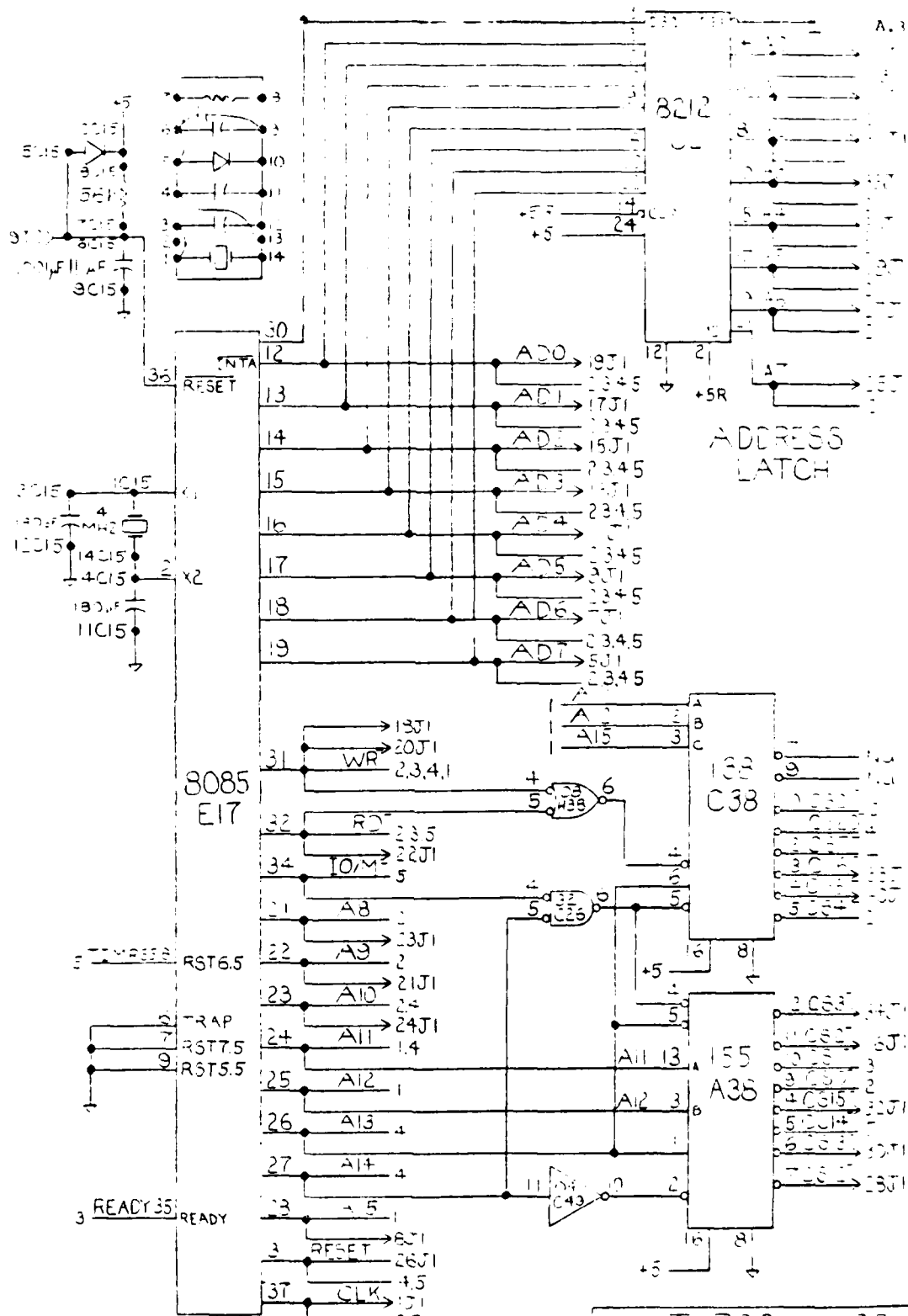


Fig. A1 BLOCK DIAGRAM OF MICROPROCESSOR SYSTEM



CHIP SELECT DECODING

Fig.A2

155 A38 138 C38
155 A38 138 C38

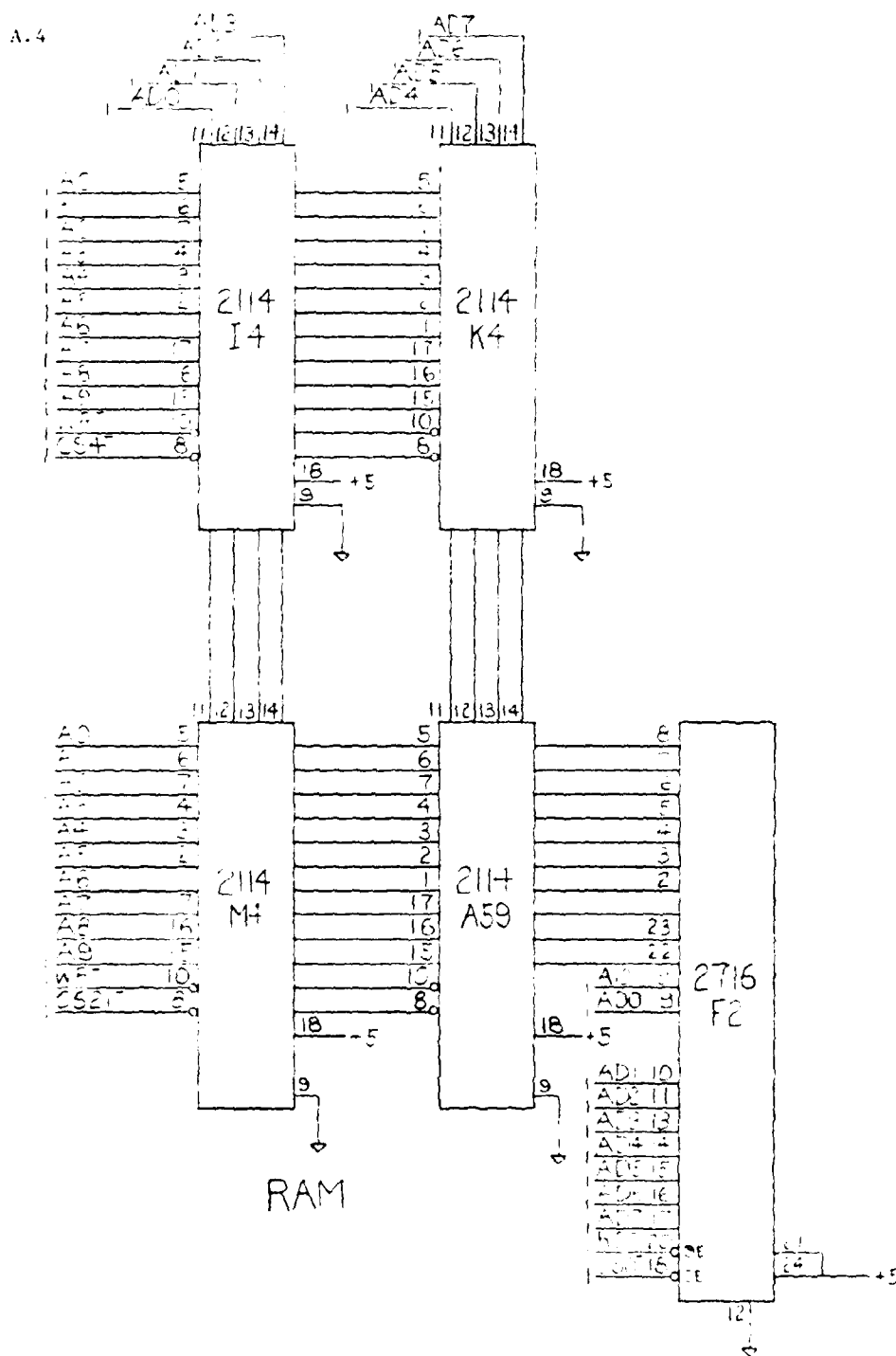
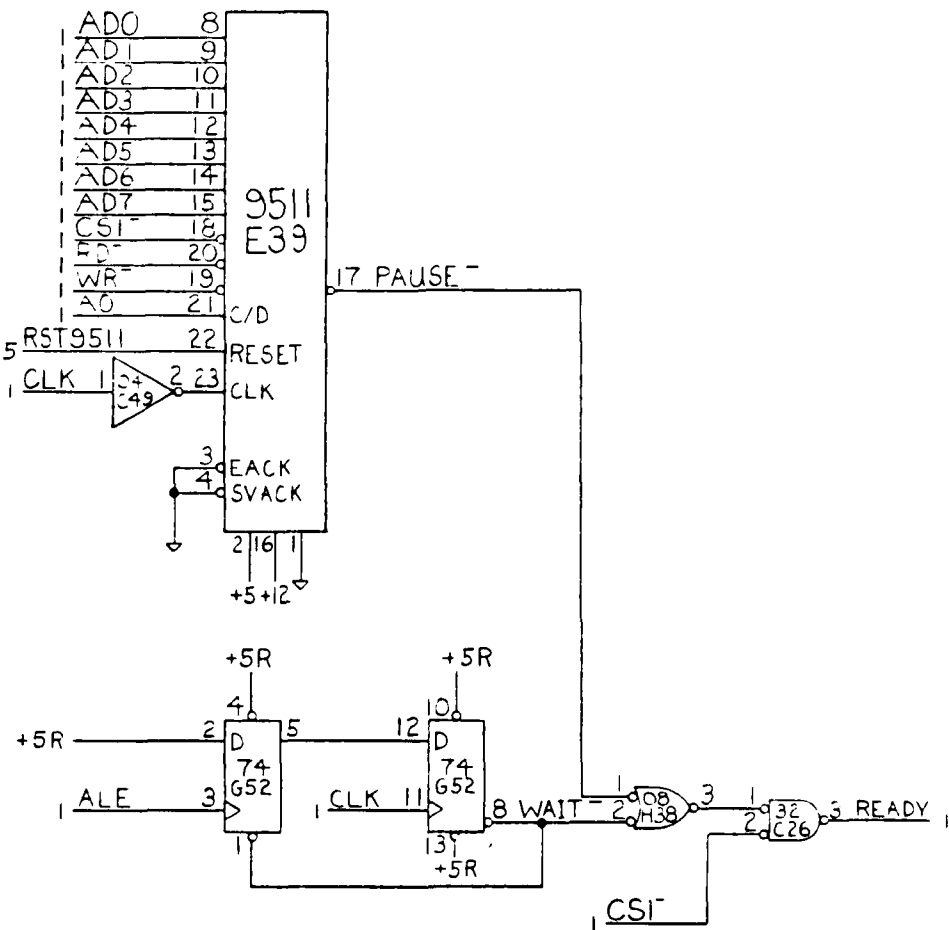


Fig. A3 PROGRAM FROM

ANT. POS.	BDI
SHEET 1	OF 379
4BF	



ARITHMETIC PROCESSOR AND WAIT STATE GENERATOR

Fig. A4

ANT. POS. -		BDI
SHEET 3	07-23-79	ABF

A.6

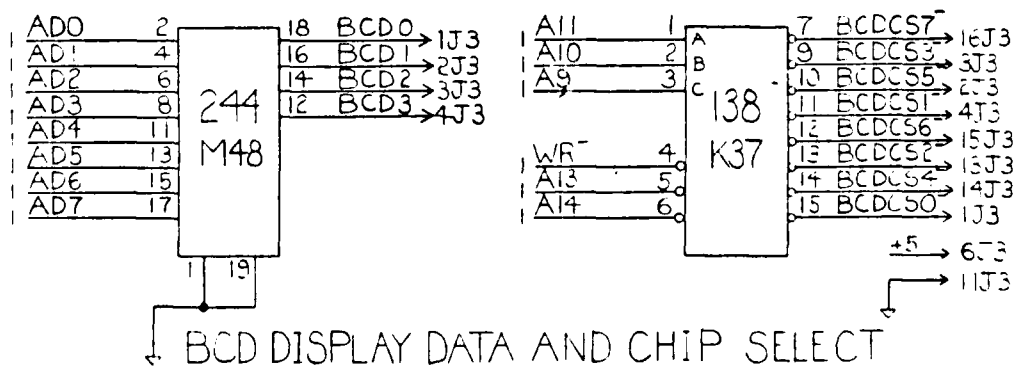
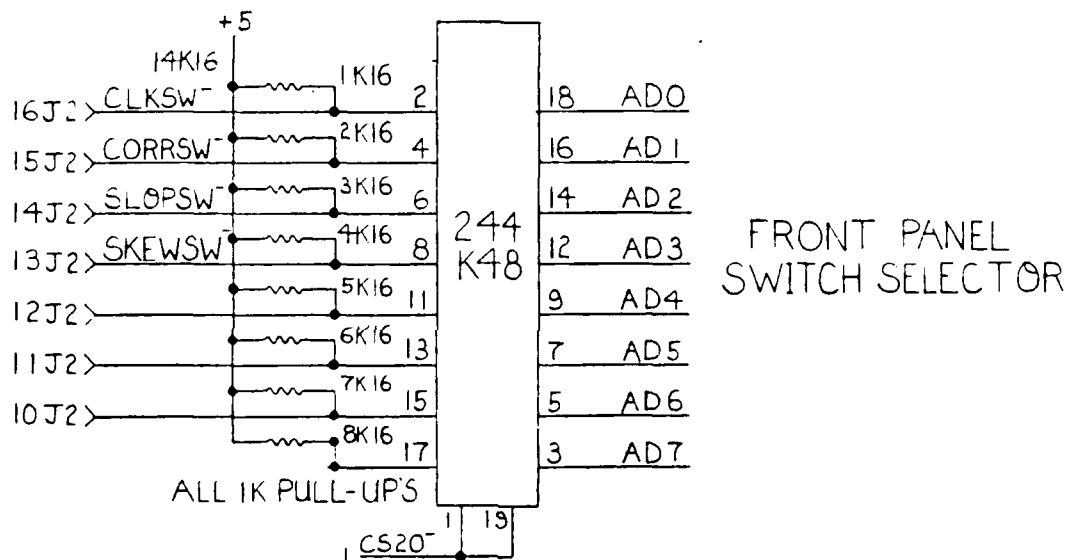
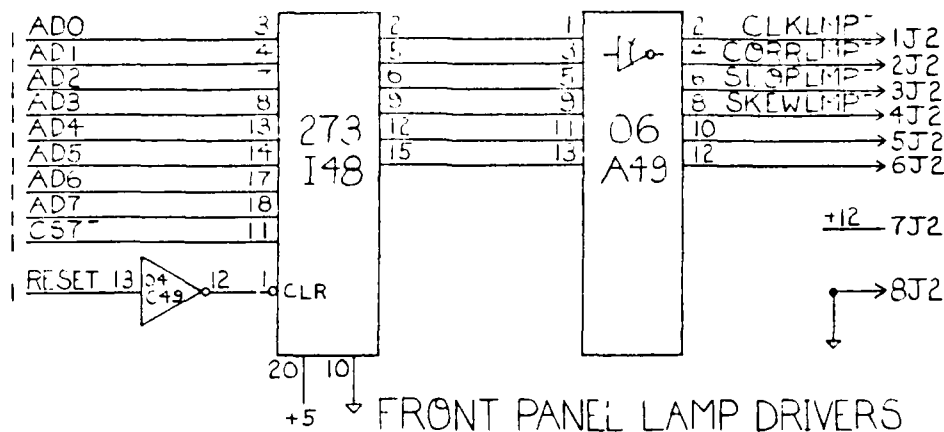


Fig.A5

ANT. POS.	BDI
SHEET 4	07-23-79 ABF

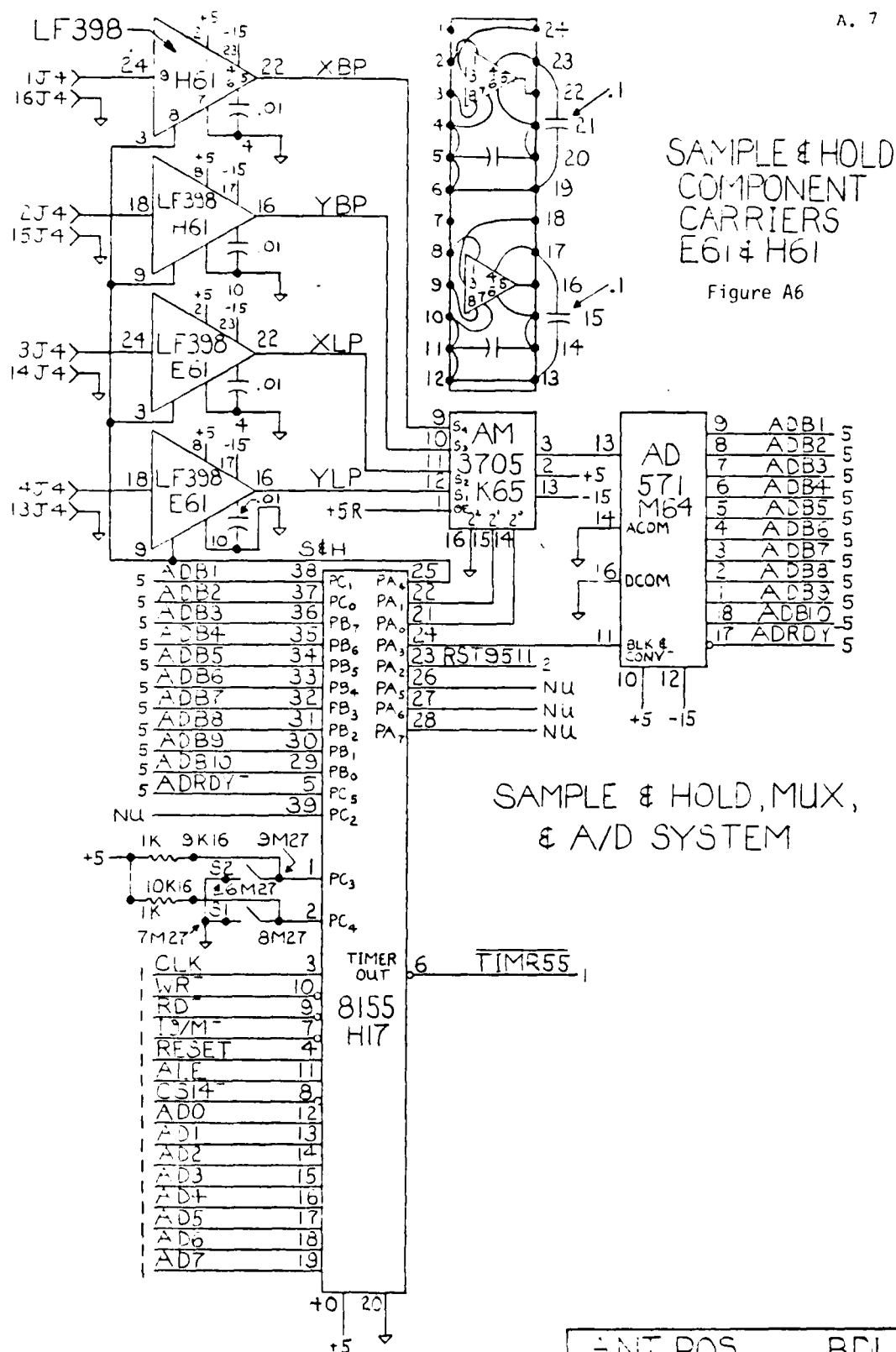
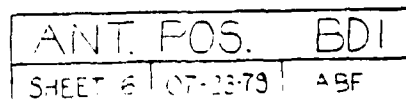
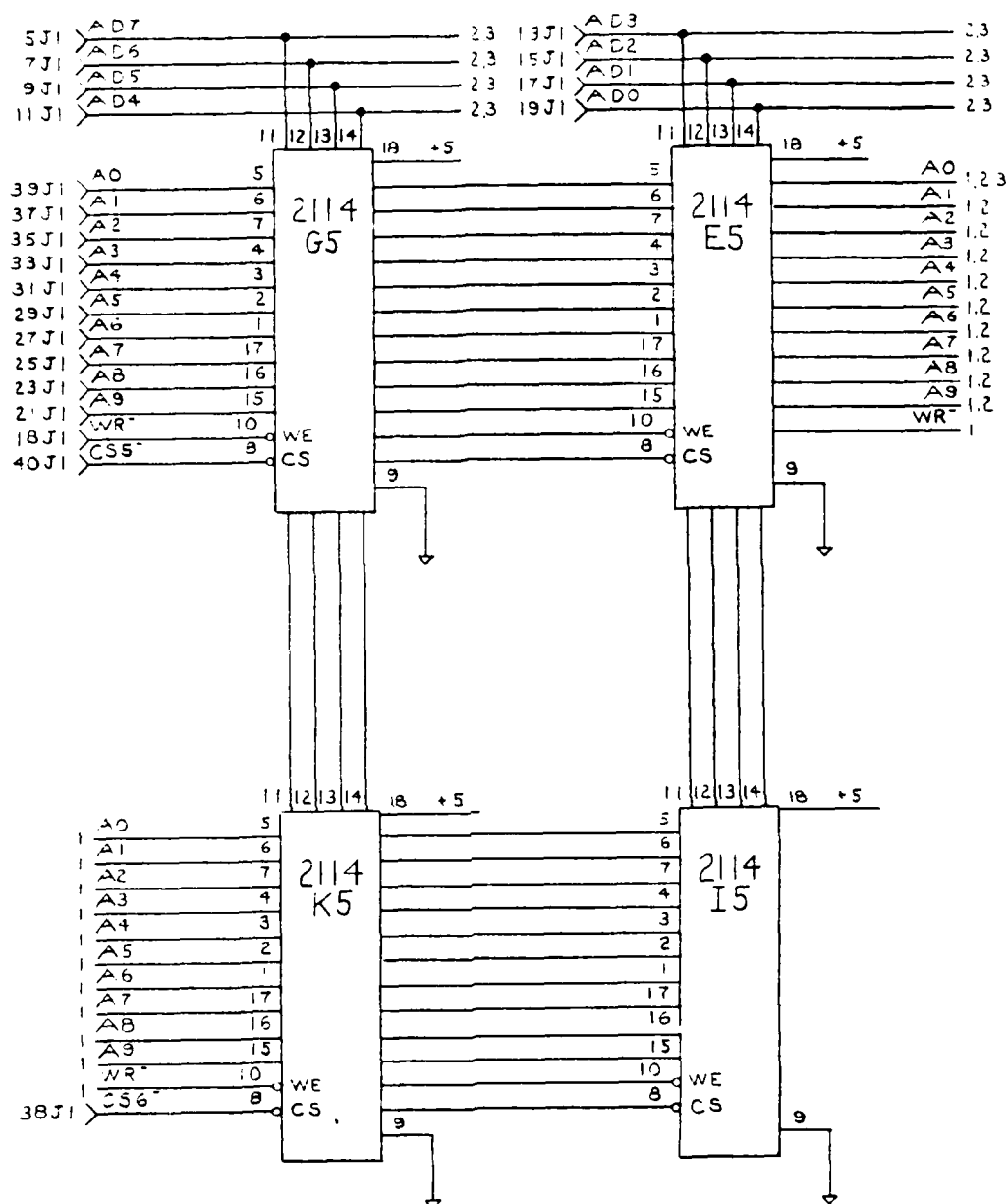


Figure A7

N M L K J I H G F E D C B A



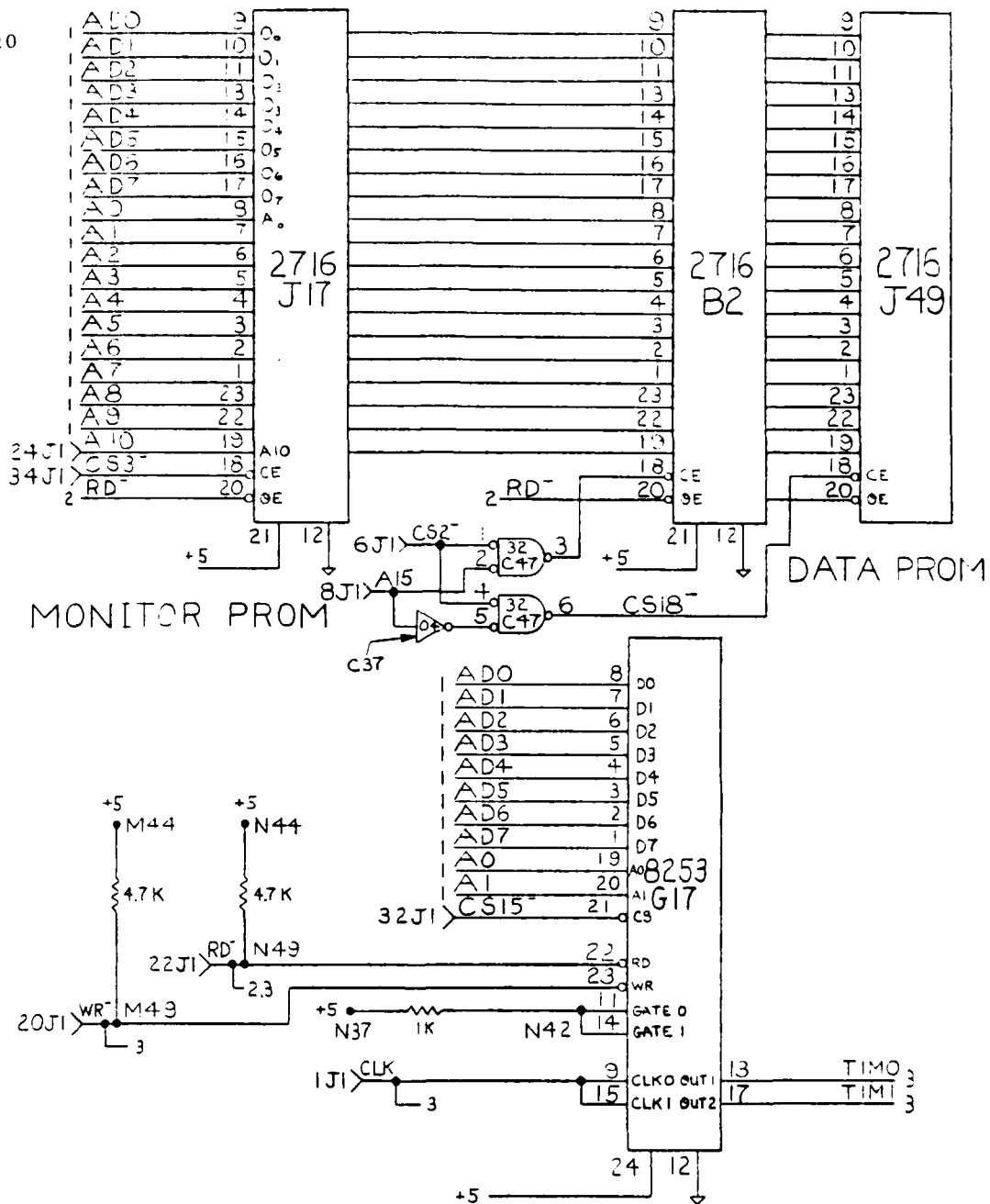


RAM

Figure A8

ANT. POS		BD2
SHEET 1	07-23-79	ABF

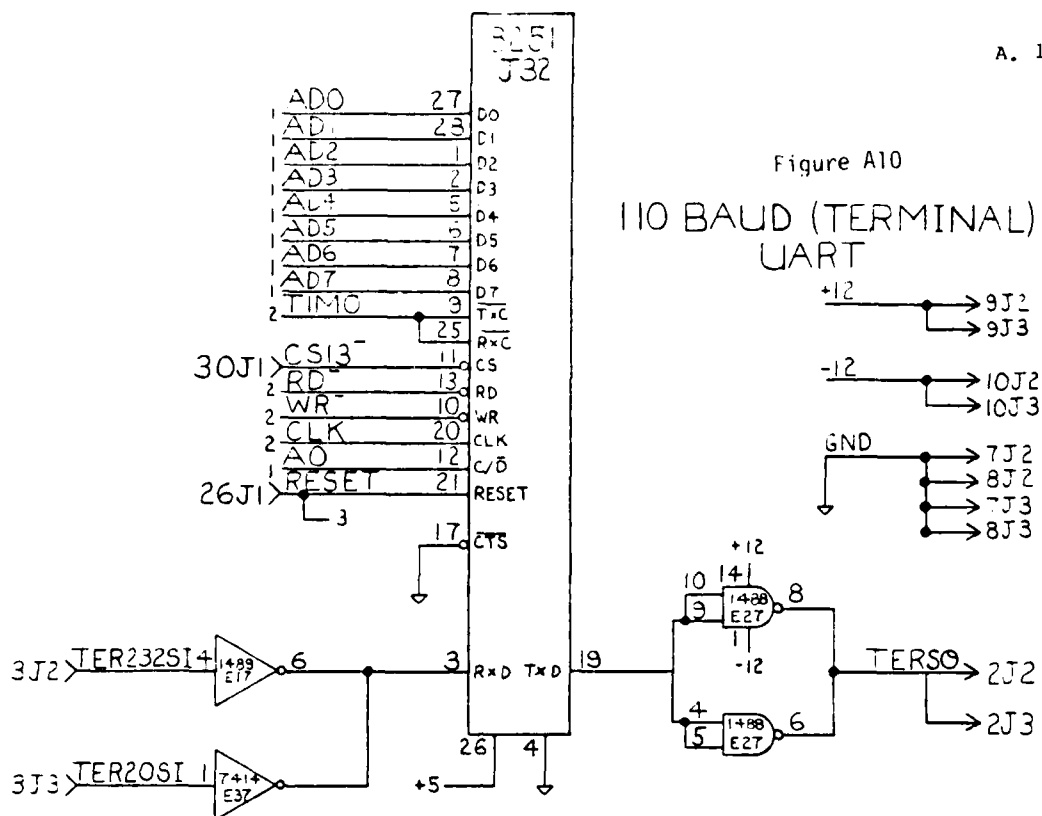
A.10



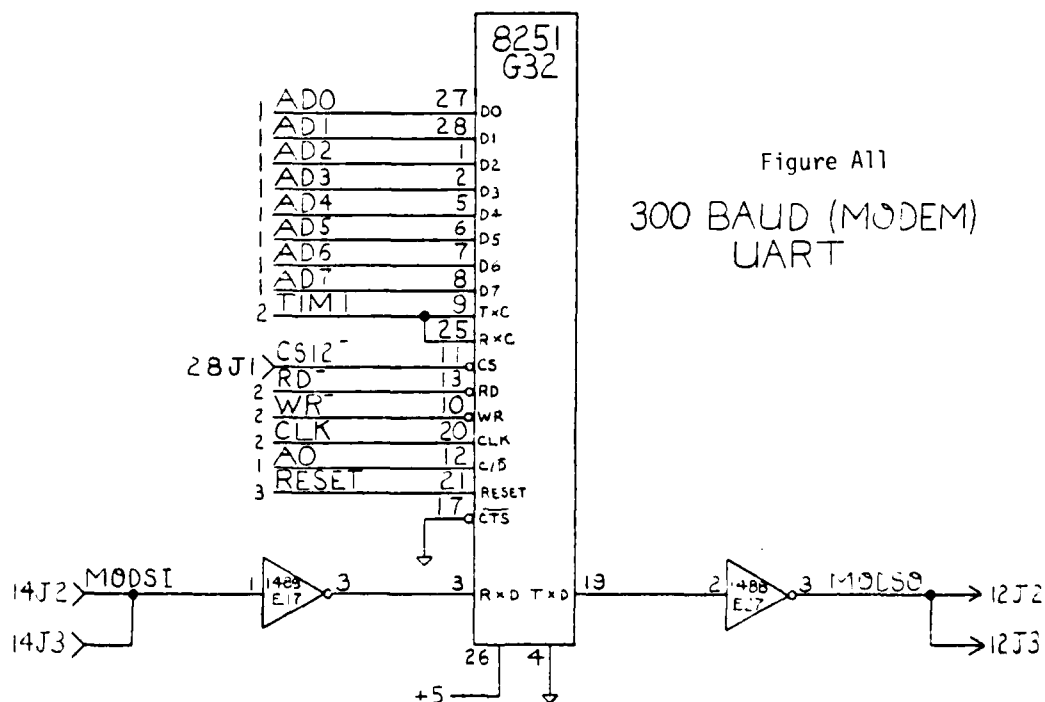
UART PROGRAMMABLE TIMER

Figure A9

ANT. PCS.	BD2
SHEET 2	07-23-79
ABF	



300 BAUD (MODEM)
UART



ANT. POS. BD2

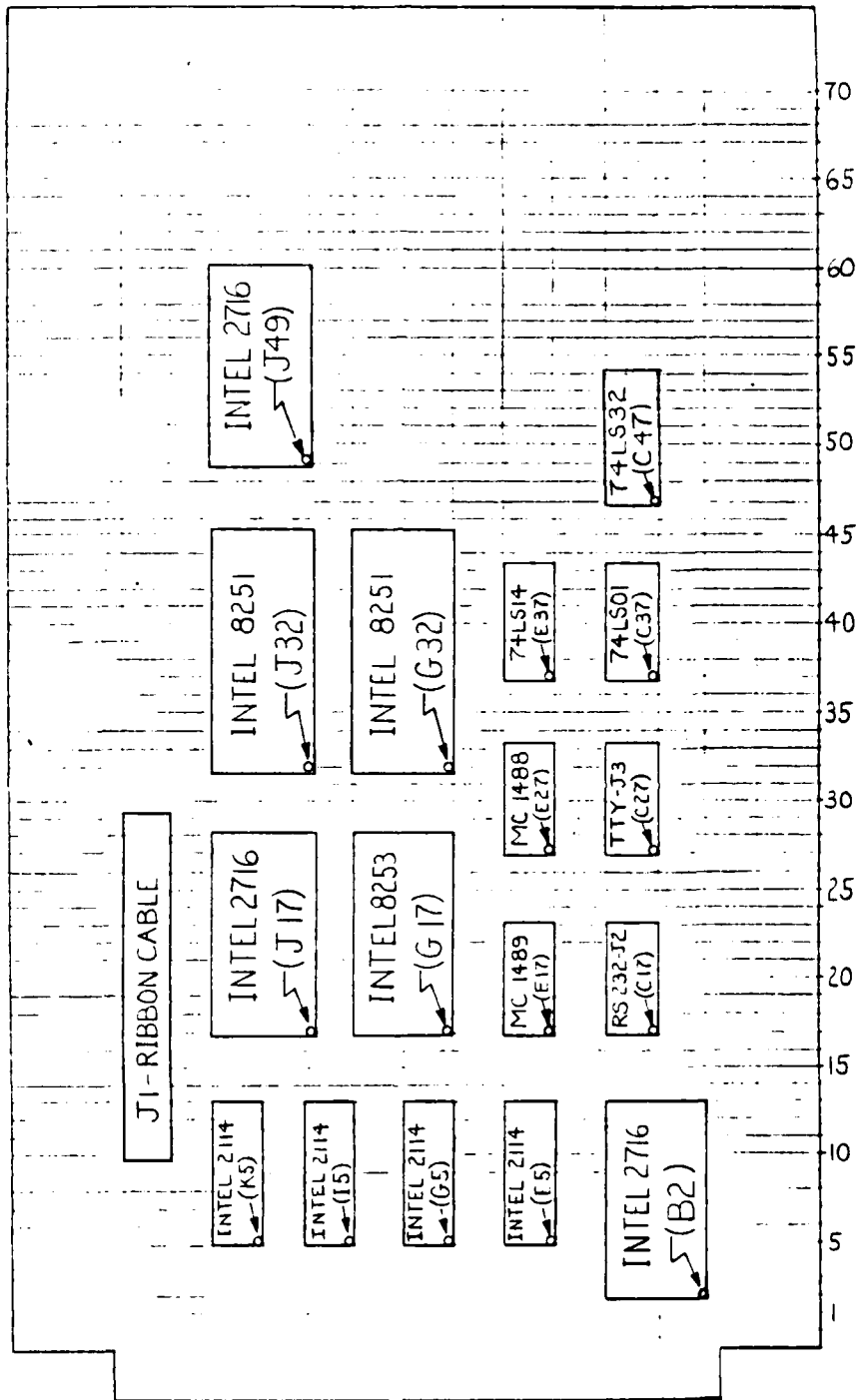
SHEET 3	07-23-79	ARF
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COMPONENTS

Figure A12

N M L K J I H G F E D C B A

A.12



ANT. POS. BD2		
SHEET 4	07-23-79	ABF

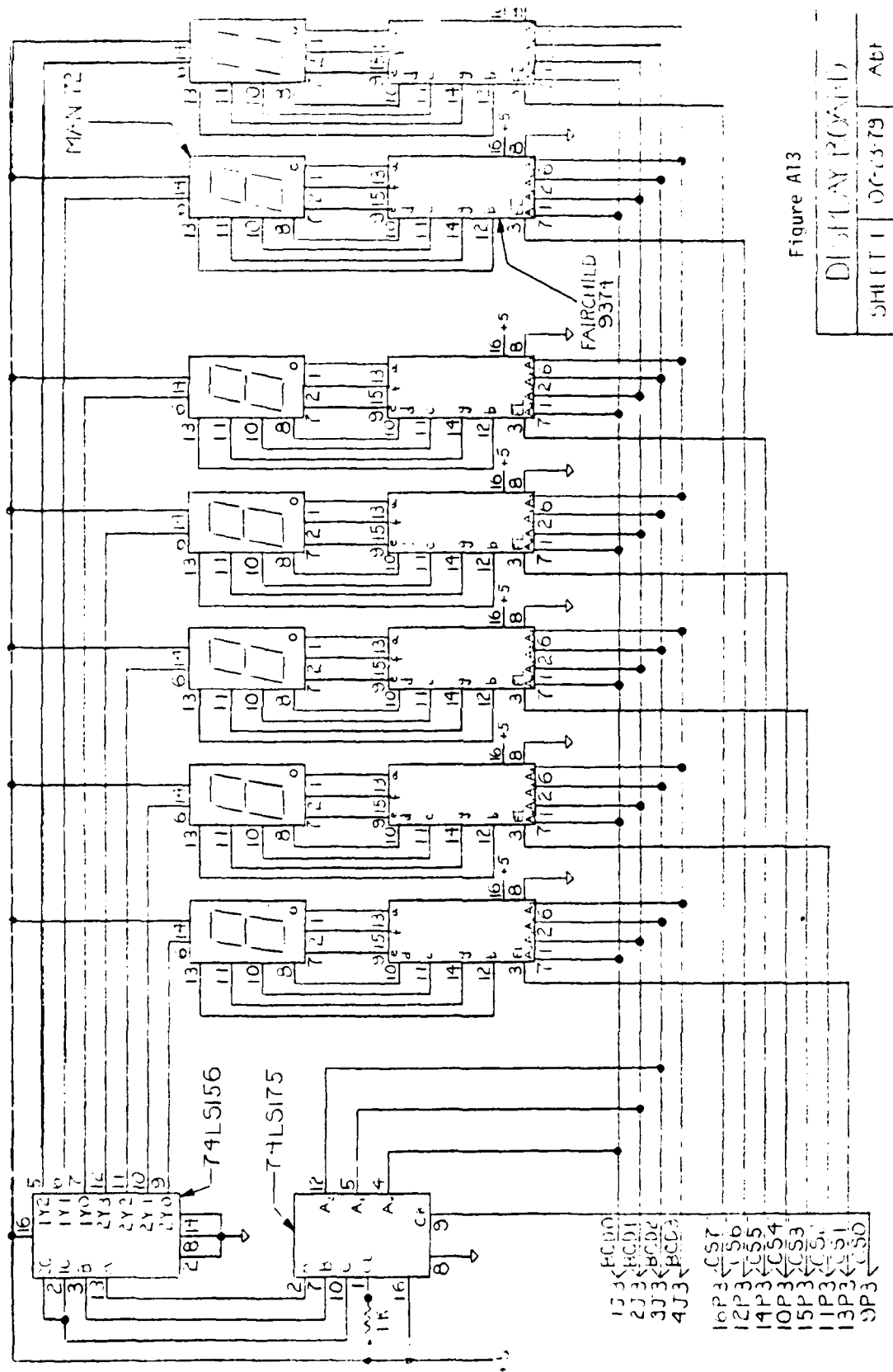


Figure A13

DISPLAY PCARD

SHEET 1 OF 3-79 ABT

A.14

Parts List Microprocessor Board #1

<u>Location</u>	<u>Item</u>
MH	Intel 2114, IC
KH	Intel 2114, IC
I4	Intel 2114, IC
F2	Intel 2716, IC
C2	Intel 8212, IC
K16	CTS 760-1-R1K, Resistor Network
H17	Intel 8155, IC
E17	Intel 8085
C26	National DM 74LS32, IC
K37	National DM 74LS138, IC
H39	National DM 74LS08, IC
C38	National DM 74LS138, IC
A38	National DM 74LS155, IC
M48	National DM 74LS244, IC
K48	National DM 74LS244, IC
I48	National DM 74LS273, IC
G52	National DM 74LS74, IC
C49	National DM 74LS04, IC
A49	National DM 7406, IC
E39	American Micro AM9511, IC
K65	National AM 3705CN, IC
M64	Analog Devices AD571JD, A/D converter
A59	Intel 2114, IC
H61	National LF398, S/H IC
F61	National LF398, S/H IC
C15	56K Resistor 1/4 W, 5%
C15	4 MHz Crystal
C15	180 pF Capacitor-Ceramic
C15	1 μ F Capacitor
C15	1N44 Diode
H61	.01 μ F Capacitor-Ceramic
H61	0.1 μ F Capacitor-Ceramic
E61	.01 μ F Capacitor-Ceramic
E61	0.1 μ F Capacitor-Ceramic
K16	1K Resistor 1/4 W, 5%
M27	Dip Switch 8 section, SPST

Parts List Microprocessor Board #2

A.15

K5	Intel 2114, IC
I5	Intel 2114, IC
G5	Intel 2114, IC
E5	Intel 2114, IC
J17	Intel 2716, IC
G17	Intel 8253, IC
E17	Motorola MC1489, IC
E27	Motorola MC 1488, IC
J32	Intel 8251, IC
G32	Intel 8251, IC
E37	National DM74LS14, IC
C37	National DM74LS01, IC
C47	National DM74LS32, IC
J49	Intel 2716, IC
N42	1K Resistor 1/4 W, 5%
M49	4.7 K Resistor 1/4 W, 5%
N49	4.7K Resistor 1/4 W, 5%

Parts List Microprocessor Display

- 1 - National DM74LS156, IC
- 1 - National DM74LS175, IC
- 7 - Monsanto MAN72, Display
- 7 - Fairchild 9374, IC

Power Supplies

- +15 Volt, 100 mA Power Mate MD15A
- + 5 Volt, 6 A Power Mate ETR-142E

APPENDIX B - MICROPROCESSOR PROGRAMMING

Program Name	Contents			Pages for FORTRAN Program	Pages for 8085 Program
	Description	Memory Locations (8085 hexadecimal)			
MAIN	1) Input data (from tape if FORTRAN; A/D or PROM in 8085) 2) Pass x,y data samples to other routines PROCESS, DISPLY, UPDATE, etc.	9039-9069	B.2-3		B.17-18
PROCESS	Process an xy data sample and accumulate averages for 1 sec. Includes: 1) Test of xy data sample for minimum signal value 2) Derivative calculations $\dot{x}, \dot{y}, \ddot{x}, \ddot{y}$ 3) One second averages for correlation, skew, and slope variables	00A3-043A	B.4-6		B.59-65
DISPLY	Convert long term average of correlation, skew, and slope variables to a form suitable for display	94A1-9677	B.6-7		B.26-29
INITIAL	Initialize all system variables and parameters	925D-929B	B.7-8		B.23
UPDATE	Add one second averages of alignment variables formed in the routine PROCES to long term averages	043B-0703	B.8-11		B.66-72
STARTUP	Program begins here after a manual reset	0000-00A0	N.A. (not applicable)		B.57-59
ASCHED	Schedules 8085 routines: A/D every 1/100 sec. Front panel scan every 1/10 sec. Display update every sec.	908D-90D0	N.A.		B.18
RDFRNT	Scan front panel switches	90D1-9128	N.A.		B.19
ATOD	Input analog data	912B-925C	N.A.		B.20-23
DISBCD	Convert a floating point number to BCD format and display it	92F3-93FF	N.A.		B.23-26
	Debugging monitor used for program development	1800-1CB5	N.A.		B.31-48
	9511 Arithmetic routines	1CB6-1D33	N.A.		B.48-50

1000

```
ISAM-L=0
NIENRS=0
VALULL=0
EXPNIK=0
KYPINZ=0
CLTIME=FLA(ILLUNS),ST-LFN
RU 1/5=1,100
```

175
NY 100-10000

$\bar{K}(X, Y) \neq \emptyset$.
 $\bar{K}(Y, Y) \neq \emptyset$.
 $\bar{K}(X, X) \neq \emptyset$.
 $\bar{K}(Y, X) \neq \emptyset$.

UNION

Shew - Annette S.

W

[illegible]

U.S. MARSHALS

```

RINS=0
NACEP=0
NNELUND=1
TIME=0
TWOP=2.33E-05
UPTIME=50000
ZCUMIN=100
RETURN
END

```

[illegible]

UPDATE AND FLAT VARIABLES PAGE 4 FROM

[illegible]

```

108A 100000000
108B 100000000
108C 100000000
108D 100000000
108E 100000000
108F 100000000
108G 100000000
108H 100000000
108I 100000000
108J 100000000
108K 100000000
108L 100000000
108M 100000000
108N 100000000
108O 100000000
108P 100000000
108Q 100000000
108R 100000000
108S 100000000
108T 100000000
108U 100000000
108V 100000000
108W 100000000
108X 100000000
108Y 100000000
108Z 100000000
109A 100000000
109B 100000000
109C 100000000
109D 100000000
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109F 100000000
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109H 100000000
109I 100000000
109J 100000000
109K 100000000
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109O 100000000
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109R 100000000
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110T 100000000
110U 100000000
110V 100000000
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111C 100000000
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111F 100000000
111G 100000000
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113U 100000000
113V 100000000
113W 100000000
113X 100000000
113Y 100000000
113Z 100000000

```


[illegible]

[illegible]

1874 179 NPEAR: DS 4
 1875 180 NPEMIN: DS 2
 1876 181 NPEMAX: DS 2
 1877 182 SCALE1: DS 4
 1878 183 SCALE2: DS 4
 1879 184 SKEW1: DS 4
 1880 185 SKEW2: DS 4
 1881 186 SLOPE1: DS 4
 1882 187 SLOPE2: DS 4
 1883 188 TETA1: DS 4
 1884 189 TETA2: DS 4
 1885 190 VALLOC: DS 4
 1886 191 VALEAT: DS 4
 1887 192 WNSLUP: DS 4
 1888 193 X1: DS 4
 1889 194 X2: DS 4
 1890 195 X3: DS 12
 1891 196 X4: DS 4
 1892 197 Y1: DS 4
 1893 198 Y2: DS 4
 1894 199 Y3: DS 4
 1895 200 Y4: DS 12
 1896 201 Y5: DS 4
 1897 202 Z50: DS 4
 1898 203 Z50U: DS 4
 1899 204 Z50L: DS 4
 1900 205 Z50M: DS 4
 1901 206 Z50N: DS 4
 1902 207 Z50O: DS 4
 1903 208 Z50P: DS 4
 1904 209 Z50Q: DS 4
 1905 210 Z50R: DS 4
 1906 211 Z50S: DS 4
 1907 212 Z50T: DS 4
 1908 213 Z50U: DS 4
 1909 214 Z50V: DS 4
 1910 215 Z50W: DS 4
 1911 216 Z50X: DS 4
 1912 217 Z50Y: DS 4
 1913 218 Z50Z: DS 4
 1914 219 Z50A: DS 4
 1915 220 Z50B: DS 4
 1916 221 Z50C: DS 4
 1917 222 Z50D: DS 4
 1918 223 Z50E: DS 4
 1919 224 Z50F: DS 4
 1920 225 Z50G: DS 4
 1921 226 Z50H: DS 4
 1922 227 Z50I: DS 4
 1923 228 Z50J: DS 4
 1924 229 Z50K: DS 4
 1925 230 Z50L: DS 4
 1926 231 Z50M: DS 4
 1927 232 Z50N: DS 4
 1928 233 Z50O: DS 4
 1929 234 Z50P: DS 4
 1930 235 Z50Q: DS 4
 1931 236 Z50R: DS 4
 1932 237 Z50S: DS 4
 1933 238 Z50T: DS 4
 1934 239 Z50U: DS 4
 1935 240 Z50V: DS 4
 1936 241 Z50W: DS 4
 1937 242 Z50X: DS 4
 1938 243 Z50Y: DS 4
 1939 244 Z50Z: DS 4
 1940 245 Z50A: DS 4
 1941 246 Z50B: DS 4
 1942 247 Z50C: DS 4
 1943 248 Z50D: DS 4
 1944 249 Z50E: DS 4
 1945 250 Z50F: DS 4
 1946 251 Z50G: DS 4
 1947 252 Z50H: DS 4
 1948 253 Z50I: DS 4
 1949 254 Z50J: DS 4
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 1951 256 Z50L: DS 4
 1952 257 Z50M: DS 4
 1953 258 Z50N: DS 4
 1954 259 Z50O: DS 4
 1955 260 Z50P: DS 4
 1956 261 Z50Q: DS 4
 1957 262 Z50R: DS 4
 1958 263 Z50S: DS 4
 1959 264 Z50T: DS 4
 1960 265 Z50U: DS 4
 1961 266 Z50V: DS 4
 1962 267 Z50W: DS 4
 1963 268 Z50X: DS 4
 1964 269 Z50Y: DS 4
 1965 270 Z50Z: DS 4
 1966 271 Z50A: DS 4
 1967 272 Z50B: DS 4
 1968 273 Z50C: DS 4
 1969 274 Z50D: DS 4
 1970 275 Z50E: DS 4
 1971 276 Z50F: DS 4
 1972 277 Z50G: DS 4
 1973 278 Z50H: DS 4
 1974 279 Z50I: DS 4
 1975 280 Z50J: DS 4
 1976 281 Z50K: DS 4
 1977 282 Z50L: DS 4
 1978 283 Z50M: DS 4
 1979 284 Z50N: DS 4
 1980 285 Z50O: DS 4
 1981 286 Z50P: DS 4
 1982 287 Z50Q: DS 4
 1983 288 Z50R: DS 4
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 1985 290 Z50T: DS 4
 1986 291 Z50U: DS 4
 1987 292 Z50V: DS 4
 1988 293 Z50W: DS 4
 1989 294 Z50X: DS 4
 1990 295 Z50Y: DS 4
 1991 296 Z50Z: DS 4
 1992 297 Z50A: DS 4
 1993 298 Z50B: DS 4
 1994 299 Z50C: DS 4
 1995 300 Z50D: DS 4
 1996 301 Z50E: DS 4
 1997 302 Z50F: DS 4
 1998 303 Z50G: DS 4
 1999 304 Z50H: DS 4
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 2032 337 Z50O: DS 4
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 2034 339 Z50Q: DS 4
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 2037 342 Z50T: DS 4
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 2105 410 Z50J: DS 4
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 2107 412 Z50L: DS 4
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 2111 416 Z50P: DS 4
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 2114 419 Z50S: DS 4
 2115 420 Z50T: DS 4
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 2252 557 Z50A: DS 4
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 2255 560 Z50D: DS 4
 2256 561 Z50E: DS 4
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 2261 566 Z50J: DS 4
 2262 567 Z50K: DS 4
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 2278 583 Z50A: DS 4
 2279 584 Z50B: DS 4
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 2282 587 Z50E: DS 4
 2283 588 Z50F: DS 4
 2284 589 Z50G: DS 4
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 2316 621 Z50M: DS 4
 2317 622 Z50N: DS 4
 2318 623 Z50O: DS 4
 2319 624 Z50P: DS 4
 2320 625 Z50Q: DS 4
 2321 626 Z50R: DS 4
 2322 627 Z50S: DS 4
 2323 628 Z50T: DS 4
 2324 629 Z50U: DS 4
 2325 630 Z50V: DS 4
 2326 631 Z50W: DS 4
 2327 632 Z50X: DS 4
 2328 633 Z50Y: DS 4
 2329 634 Z50Z: DS 4
 2330 635 Z50A: DS 4
 2331 636 Z50B: DS 4
 2332 637 Z50C: DS 4
 2333 638 Z50D: DS 4
 2334 639 Z50E: DS 4
 2335 640 Z50F: DS 4
 2336 641 Z50G: DS 4
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 2339 644 Z50J: DS 4
 2340 645 Z50K: DS 4
 2341 646 Z50L: DS 4
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 2343 648 Z50N: DS 4
 2344 649 Z50O: DS 4
 2345 650 Z50P: DS 4
 2346 651 Z50Q: DS 4
 2347 652 Z50R: DS 4
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 2349 654 Z50T: DS 4
 2350 655 Z50U: DS 4
 2351 656 Z50V: DS 4
 2352 657 Z50W: DS 4
 2353 658 Z50X: DS 4
 2354 659 Z50Y: DS 4
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 2356 661 Z50A: DS 4
 2357 662 Z50B: DS 4
 2358 663 Z50C: DS 4
 2359 664 Z50D: DS 4
 2360 665 Z50E: DS 4
 2361 666 Z50F: DS 4
 2362 667 Z50G: DS 4
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 2364 669 Z50I: DS 4
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 2366 671 Z50K: DS 4
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 2368 673 Z50M: DS 4
 2369 674 Z50N: DS 4
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 2372 677 Z50Q: DS 4
 2373 678 Z50R: DS 4
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 2379 684 Z50X: DS 4
 2380 685 Z50Y: DS 4
 2381 686 Z50Z: DS 4
 2382 687 Z50A: DS 4
 2383 688 Z50B: DS 4
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 2389 694 Z50H: DS 4
 2390 695 Z50I: DS 4
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 2392 697 Z50K: DS 4
 2393 698 Z50L: DS 4
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 2396 701 Z50O: DS 4
 2397 702 Z50P: DS 4
 2398 703 Z50Q: DS 4
 2399 704 Z50R: DS 4
 2400 705 Z50S: DS 4
 2401 706 Z50T: DS 4
 2402 707 Z50U: DS 4
 2403 708 Z50V: DS 4
 2404 709 Z50W: DS 4
 2405 710 Z50X: DS 4
 2406 711 Z50Y: DS 4
 2407 712 Z50Z: DS 4
 2408 713 Z50A: DS 4
 2409 714 Z50B: DS 4
 2410 715 Z50C: DS 4
 2411 716 Z50D: DS 4
 2412 717 Z50E: DS 4
 2413 718 Z50F: DS 4
 2414 719 Z50G: DS 4
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 2429 734 Z50V: DS 4
 2430 735 Z50W: DS 4
 2431 736 Z50X: DS 4
 2432 737 Z50Y: DS 4
 2433 738 Z50Z: DS 4
 2434 739 Z50A: DS 4
 2435 740 Z50B: DS 4
 2436 741 Z50C: DS 4
 2437 742 Z50D: DS 4
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 2439 744 Z50F: DS 4
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 2443 748 Z50J: DS 4
 2444 749 Z50K: DS 4
 2445 750 Z50L: DS 4
 2446 751 Z50M: DS 4
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 2448 753 Z50O: DS 4
 2449 754 Z50P: DS 4
 2450 755 Z50Q: DS 4
 2451 756 Z50R: DS 4
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 2453 758 Z50T: DS 4
 2454 759 Z50U: DS 4
 2455 760 Z50V: DS 4
 2456 761 Z50W: DS 4
 2457 762 Z50X: DS 4
 2458 763 Z50Y: DS 4
 2459 764 Z50Z: DS 4
 2460 765 Z50A: DS 4
 2461 766 Z50B: DS 4

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235  INIT:  DS  4
240  ACPRT:  DS  4
241  D71:   DS  4
242  TOLFT: DS  4
243  TUALT: DS  4
244  TUALT: DS  4
245  TUALT: DS  4
246  TUALT: DS  4
247  TUALT: DS  4
248  TUALT: DS  4
249  TUALT: DS  4
250  TUALT: DS  4
251  TUALT: DS  4
252  TUALT: DS  4
253  TUALT: DS  4
254  TUALT: DS  4
255  TUALT: DS  4
256  TUALT: DS  4
257  TUALT: DS  4
258  TUALT: DS  4
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261  TUALT: DS  4
262  TUALT: DS  4
263  TUALT: DS  4
264  TUALT: DS  4
265  TUALT: DS  4
266  TUALT: DS  4
267  TUALT: DS  4
268  TUALT: DS  4
269  TUALT: DS  4
270  TUALT: DS  4
271  TUALT: DS  4
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273  TUALT: DS  4
274  TUALT: DS  4
275  TUALT: DS  4
276  TUALT: DS  4
277  TUALT: DS  4
278  TUALT: DS  4
279  TUALT: DS  4
280  TUALT: DS  4
281  TUALT: DS  4
282  TUALT: DS  4
283  TUALT: DS  4

```

VARIABLES OF PROCESS UNITARY AND DISPLY UNIT ARE INITIALIZED TO A
FORWARD VALUE BY THE SUBROUTINE INITIAL.

```

251  JUDGE:  DS  4
252  JUDGE:  DS  4
253  JUDGE:  DS  4
254  JUDGE:  DS  4
255  JUDGE:  DS  4
256  JUDGE:  DS  4
257  JUDGE:  DS  4
258  JUDGE:  DS  4
259  JUDGE:  DS  4
260  JUDGE:  DS  4
261  JUDGE:  DS  4
262  JUDGE:  DS  4
263  JUDGE:  DS  4
264  JUDGE:  DS  4
265  JUDGE:  DS  4
266  JUDGE:  DS  4
267  JUDGE:  DS  4
268  JUDGE:  DS  4
269  JUDGE:  DS  4
270  JUDGE:  DS  4
271  JUDGE:  DS  4
272  JUDGE:  DS  4
273  JUDGE:  DS  4
274  JUDGE:  DS  4
275  JUDGE:  DS  4
276  JUDGE:  DS  4
277  JUDGE:  DS  4
278  JUDGE:  DS  4
279  JUDGE:  DS  4
280  JUDGE:  DS  4
281  JUDGE:  DS  4
282  JUDGE:  DS  4
283  JUDGE:  DS  4

```

END OF MASTER LIST OF EQUATED GLOBAL VARIABLES AND END ASSIGNMENTS

```

280  JUDGE:  DS  4
281  JUDGE:  DS  4
282  JUDGE:  DS  4
283  JUDGE:  DS  4

```

MISC ROUTINES TO UL/ THIS FROM TO ASSEMBLE

```

280  JUDGE:  DS  4
281  JUDGE:  DS  4
282  JUDGE:  DS  4
283  JUDGE:  DS  4

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[illegible]

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0402 11 1550 157 000 0000
0403 11 1560 158 000 0000
0404 11 157
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723 ; CALLING SEQUENCE:
724 ; THIS SUBROUTINE EXHIBITS THE USE OF REGISTERS AND THE MATH IN THE
725 ; AT THE NUMBER TO BE DISPLAYED.
726 ;
727 ; USES ALL REGISTERS EXCEPT L.
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729 ;
730 ;-----
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734 ; 732 MOV R0,R0
735 ; 733 MOV R0,R0
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9330 32 104C 184 LAI MATEN
9331 32 001F 185 CALL PUSHA
9332 32 2F 186 STA MATC
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9334 32 0108 188 STA MATC
9335 32 0108 189 CALL SW1CH
9336 32 104C 191 LAI MATEN
9337 32 001F 192 CALL PUSHA
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9340 32 2D 196 STA MATC
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9342 32 0108 198 STA MATC
9343 32 1170 199 CALL SW1CH
9344 32 1170 200 STA MATC
9345 32 1170 201 CALL BLANK
9346 32 0F70 202 LUI 001
9347 32 004A 203 STA LED6
9348 32 1170 204 LUI 001
9349 32 004B 205 STA LED5
9350 32 1170 206 LUI 001
9351 32 004C 207 STA LED4
9352 32 004D 208 STA LED3
9353 32 004E 209 STA LED2
9354 32 004F 210 STA LED1
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9774 32 0469 630 CALL SW1CH
9775 32 0470 631 STA MATC
9776 32 0471 632 MVI A,AF10D
9777 32 0472 633 STA MATC
9778 32 0473 634 CALL SW1CH
9779 32 0474 635 STA MATC
9780 32 0475 636 MVI A,AF10D
9781 32 0476 637 STA MATC
9782 32 0477 638 CALL SW1CH
9783 32 0478 639 STA MATC
9784 32 0479 640 MVI A,AF10D
9785 32 0480 641 STA MATC
9786 32 0481 642 CALL SW1CH
9787 32 0482 643 STA MATC
9788 32 0483 644 MVI A,AF10D
9789 32 0484 645 STA MATC
9790 32 0485 646 CALL SW1CH
9791 32 0486 647 STA MATC
9792 32 0487 648 MVI A,AF10D
9793 32 0488 649 STA MATC
9794 32 0489 650 CALL SW1CH
9795 32 0490 651 STA MATC
9796 32 0491 652 MVI A,AF10D
9797 32 0492 653 STA MATC
9798 32 0493 654 CALL SW1CH
9799 32 0494 655 STA MATC
9800 32 0495 656 MVI A,AF10D
9801 32 0496 657 STA MATC
9802 32 0497 658 CALL SW1CH
9803 32 0498 659 STA MATC
9804 32 0499 660 MVI A,AF10D
9805 32 0500 661 STA MATC
9806 32 0501 662 CALL SW1CH
9807 32 0502 663 STA MATC
9808 32 0503 664 MVI A,AF10D
9809 32 0504 665 STA MATC
9810 32 0505 666 CALL SW1CH
9811 32 0506 667 STA MATC
9812 32 0507 668 MVI A,AF10D
9813 32 0508 669 STA MATC
9814 32 0509 670 CALL SW1CH
9815 32 0510 671 STA MATC
9816 32 0511 672 MVI A,AF10D
9817 32 0512 673 STA MATC
9818 32 0513 674 CALL SW1CH
9819 32 0514 675 STA MATC
9820 32 0515 676 MVI A,AF10D
9821 32 0516 677 STA MATC
9822 32 0517 678 CALL SW1CH
9823 32 0518 679 STA MATC
9824 32 0519 680 MVI A,AF10D
9825 32 0520 681 STA MATC
9826 32 0521 682 CALL SW1CH
9827 32 0522 683 STA MATC
9828 32 0523 684 MVI A,AF10D
9829 32 0524 685 STA MATC
9830 32 0525 686 CALL SW1CH
9831 32 0526 687 STA MATC
9832 32 0527 688 MVI A,AF10D
9833 32 0528 689 STA MATC
9834 32 0529 690 CALL SW1CH
9835 32 0530 691 STA MATC
9836 32 0531 692 MVI A,AF10D
9837 32 0532 693 STA MATC
9838 32 0533 694 CALL SW1CH
9839 32 0534 695 STA MATC
9840 32 0535 696 MVI A,AF10D
9841 32 0536 697 STA MATC
9842 32 0537 698 CALL SW1CH
9843 32 0538 699 STA MATC
9844 32 0539 700 MVI A,AF10D
9845 32 0540 701 STA MATC
9846 32 0541 702 CALL SW1CH
9847 32 0542 703 STA MATC
9848 32 0543 704 MVI A,AF10D
9849 32 0544 705 STA MATC
9850 32 0545 706 CALL SW1CH
9851 32 0546 707 STA MATC
9852 32 0547 708 MVI A,AF10D
9853 32 0548 709 STA MATC
9854 32 0549 710 CALL SW1CH
9855 32 0550 711 STA MATC
9856 32 0551 712 MVI A,AF10D
9857 32 0552 713 STA MATC
9858 32 0553 714 CALL SW1CH
9859 32 0554 715 STA MATC
9860 32 0555 716 MVI A,AF10D
9861 32 0556 717 STA MATC
9862 32 0557 718 CALL SW1CH
9863 32 0558 719 STA MATC
9864 32 0559 720 MVI A,AF10D
9865 32 0560 721 STA MATC
9866 32 0561 722 CALL SW1CH
9867 32 0562 723 STA MATC
9868 32 0563 724 MVI A,AF10D
9869 32 0564 725 STA MATC
9870 32 0565 726 CALL SW1CH
9871 32 0566 727 STA MATC
9872 32 0567 728 MVI A,AF10D
9873 32 0568 729 STA MATC
9874 32 0569 730 CALL SW1CH
9875 32 0570 731 STA MATC
9876 32 0571 732 MVI A,AF10D
9877 32 0572 733 STA MATC
9878 32 0573 734 CALL SW1CH
9879 32 0574 735 STA MATC
9880 32 0575 736 MVI A,AF10D
9881 32 0576 737 STA MATC
9882 32 0577 738 CALL SW1CH
9883 32 0578 739 STA MATC
9884 32 0579 740 MVI A,AF10D
9885 32 0580 741 STA MATC
9886 32 0581 742 CALL SW1CH
9887 32 0582 743 STA MATC
9888 32 0583 744 MVI A,AF10D
9889 32 0584 745 STA MATC
9890 32 0585 746 CALL SW1CH
9891 32 0586 747 STA MATC
9892 32 0587 748 MVI A,AF10D
9893 32 0588 749 STA MATC
9894 32 0589 750 CALL SW1CH
9895 32 0590 751 STA MATC
9896 32 0591 752 MVI A,AF10D
9897 32 0592 753 STA MATC
9898 32 0593 754 CALL SW1CH
9899 32 0594 755 STA MATC
9900 32 0
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1805      0000 -
ns
xc
1 1 MONITOR
2 1 11/12/79
3 1 CHANGES
4 1 11/12/79 - POWER UP AND DISPLAY ETC MOVED TO ROMC
5 1 11/11/79 - RECOVERED 11/07/79 FROM THE FILE. SOME SHOULD BE UNRECOVERED
6 1 DISPLAY MODE
7 1 ROM DISPLAY, CONSTANTS, AND VOLT READINGS ETC
8 1
9 1
10 1
11 1 *****
12 1 11/12/79 - POWER UP AND DISPLAY ETC MOVED TO ROMC
13 1 11/11/79 - RECOVERED 11/07/79 FROM THE FILE. SOME SHOULD BE UNRECOVERED
14 1 *****
15 1 *****
16 1 *****
17 1 11/12/79 - POWER UP AND DISPLAY ETC MOVED TO ROMC
18 1 11/11/79 - RECOVERED 11/07/79 FROM THE FILE. SOME SHOULD BE UNRECOVERED
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111 ; ADDRESS ENTRY POINT.
112 ;
113 ;
114 ; COMPUTE THE TOP OF USER MEMORY AVAILABLE IN THIS SYSTEM
115 ; AND CALCULATE THE USER REGISTER SIZE (USER REG SIZE) AND
116 ; POSITION SIMUL AT THE TOP OF RAM.
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1853 2Y	171	MOV	D	1	ADD TO TABLE AND
1854 21	172	POP	D	1	RESTORE PC
1855 7E	173	MOV	AM	1	GET LSR OF ADDRESS
1856 23	174	IRX	M	1	POINT TO NEXT ADDRESS
1857 65	175	MOV	M	1	GET HIGH OF ADDRESS
1858 4F	176	MOV	LH	1	LSR TO L
1859 05 02	177	MOV	LH	1	LSR TO L
1859 25	178	CALL	LH	1	LSR TO L
	179				LSR TO L
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2170 DB 1232 MVI
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2172 DD 1234 MVI
2173 DE 1235 MVI
2174 DF 1236 MVI
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2184 E9 1246 MVI
2185 EA 1247 MVI
2186 EB 1248 MVI
2187 EC 1249 MVI
2188 ED 1250 MVI
2189 EE 1251 MVI
2190 EF 1252 MVI
2191 F0 1253 MVI
2192 F1 1254 MVI
2193 F2 1255 MVI
2194 F3 1256 MVI
2195 F4 1257 MVI
2196 F5 1258 MVI
2197 F6 1259 MVI
2198 F7 1260 MVI
2199 F8 1261 MVI
2200 F9 1262 MVI
2201 FA 1263 MVI
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2206 FF 1268 MVI
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2208 01 1270 MVI
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2216 09 1278 MVI
2217 0A 1279 MVI
2218 0B 1280 MVI
2219 0C 1281 MVI
2220 0D 1282 MVI
2221 0E 1283 MVI
2222 0F 1284 MVI
2223 10 1285 MVI
2224 11 1286 MVI
2225 12 1287 MVI
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2235 1C 1297 MVI
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2238 1F 1300 MVI
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2240 21 1302 MVI
2241 22 1303 MVI
2242 23 1304 MVI
2243 24 1305 MVI
2244 25 1306 MVI
2245 26 1307 MVI
2246 27 1308 MVI
2247 28 1309 MVI
2248 29 1310 MVI
2249 2A 1311 MVI
2250 2B 1312 MVI
2251 2C 1313 MVI
2252 2D 1314 MVI
2253 2E 1315 MVI
2254 2F 1316 MVI
2255 30 1317 MVI
2256 31 1318 MVI
2257 32 1319 MVI
2258 33 1320 MVI
2259 34 1321 MVI
2260 35 1322 MVI
2261 36 1323 MVI
2262 37 1324 MVI
2263 38 1325 MVI
2264 39 1326 MVI
2265 3A 1327 MVI
2266 3B 1328 MVI
2267 3C 1329 MVI
2268 3D 1330 MVI
2269 3E 1331 MVI
2270 3F 1332 MVI
2271 40 1333 MVI
2272 41 1334 MVI
2273 42 1335 MVI
2274 43 1336 MVI
2275 44 1337 MVI
2276 45 1338 MVI
2277 46 1339 MVI
2278 47 1340 MVI
2279 48 1341 MVI
2280 49 1342 MVI
2281 4A 1343 MVI
2282 4B 1344 MVI
2283 4C 1345 MVI
2284 4D 1346 MVI
2285 4E 1347 MVI

```

[illegible]

```

: DATA IF NULL ENTRY
: IF ENTRY = 0 THEN GO TO CONTINUE
: SAVE FOURIER TO ADDR
: SAVE PRECISION
: GET NEW NEW VALUE
: A = PRECISION
: B = DELIMITER CHAR
: A = PRECISION
: STORE 256 IN REGISTER AREA
: RETRIEVE PRECISION
: B BITS ONLY
: STORE MSB IN REGISTER AREA
: RETRIEVE DELIMITED CHAR
: TEST FOR END OF TABLE
: TEST DELIMITER
: OUTPUT A SPACE
: GET THE CHARACTER
: POINT AT DISPLACEMENT
: SET CONDITION CODES
: ALL DONE
: PRINT CHARACTER
: PRINT EQUAL SIGN
: FETCH DISPLACEMENT
: POINT AT PRECISION
: SAVE HL IN DE
: COMPUTE ADDRESS OF VALUE
: HL = CURRENT TABLE ADDRESS
: FETCH PRECISION
: POINT TO NEXT DISPLAY
: GET LSP OF DATA
: DISPLAY IT
: B BIT PRECISION, ALL DONE
: POINT TO USE OF 16 BIT VALUE
: FETCH IT
: DISPLAY IT
: CONTINUE

```

[illegible]

```

111 : 1.0 MS DELAY
712 :
1853 C5 713 RELAY: PUSH R
1854 C6 714 MVI R, R
1855 C7 715 MLO: MCR M
1856 C8 716 JNZ DLO
1857 C9 717 MVI D, 0
1858 CA 718 MVI D, 0
1859 CB 719 : RETURN
720 :
721 :
722 :
723 :
1860 C5 724 CURV: MVI R, 0
1861 C6 725 DAA
1862 C7 726 DAA
1863 C8 727 DAA
1864 C9 728 MOV R, R
1865 CA 729 MVI R, 0
1866 CB 730 : RETURN
731 :
732 :
733 :
734 :
1867 C5 735 CALL C
1868 C6 736 CALL C
1869 C7 737 CALL C
1870 C8 738 CALL C
1871 C9 739 JNZ C
740 :
741 :
742 :
743 :
1872 C1 0000 EXP: LXI R, 0
1873 C2 0001 EXP: LXI R, 0
1874 C3 0002 EXP: LXI R, 0
1875 C4 0003 EXP: LXI R, 0
1876 C5 0004 EXP: LXI R, 0
1877 C6 0005 EXP: LXI R, 0
1878 C7 0006 EXP: LXI R, 0
1879 C8 0007 EXP: LXI R, 0
1880 C9 0008 EXP: LXI R, 0
1881 CA 0009 EXP: LXI R, 0
1882 CB 0010 EXP: LXI R, 0
1883 CC 0011 EXP: LXI R, 0
1884 CD 0012 EXP: LXI R, 0
1885 CE 0013 EXP: LXI R, 0
1886 CF 0014 EXP: LXI R, 0
1887 D1 0015 EXP: LXI R, 0
1888 D2 0016 EXP: LXI R, 0
1889 D3 0017 EXP: LXI R, 0
1890 D4 0018 EXP: LXI R, 0
1891 D5 0019 EXP: LXI R, 0
1892 D6 0020 EXP: LXI R, 0
1893 D7 0021 EXP: LXI R, 0
1894 D8 0022 EXP: LXI R, 0
1895 D9 0023 EXP: LXI R, 0
1896 DA 0024 EXP: LXI R, 0
1897 DB 0025 EXP: LXI R, 0
1898 DC 0026 EXP: LXI R, 0
1899 DD 0027 EXP: LXI R, 0
1900 DE 0028 EXP: LXI R, 0
1901 DF 0029 EXP: LXI R, 0
1902 E1 0030 EXP: LXI R, 0
1903 E2 0031 EXP: LXI R, 0
1904 E3 0032 EXP: LXI R, 0
1905 E4 0033 EXP: LXI R, 0
1906 E5 0034 EXP: LXI R, 0
1907 E6 0035 EXP: LXI R, 0
1908 E7 0036 EXP: LXI R, 0
1909 E8 0037 EXP: LXI R, 0
1910 E9 0038 EXP: LXI R, 0
1911 EA 0039 EXP: LXI R, 0
1912 EB 0040 EXP: LXI R, 0
1913 EC 0041 EXP: LXI R, 0
1914 ED 0042 EXP: LXI R, 0
1915 EE 0043 EXP: LXI R, 0
1916 EF 0044 EXP: LXI R, 0
1917 F1 0045 EXP: LXI R, 0
1918 F2 0046 EXP: LXI R, 0
1919 F3 0047 EXP: LXI R, 0
1920 F4 0048 EXP: LXI R, 0
1921 F5 0049 EXP: LXI R, 0
1922 F6 0050 EXP: LXI R, 0
1923 F7 0051 EXP: LXI R, 0
1924 F8 0052 EXP: LXI R, 0
1925 F9 0053 EXP: LXI R, 0
1926 FA 0054 EXP: LXI R, 0
1927 FB 0055 EXP: LXI R, 0
1928 FC 0056 EXP: LXI R, 0
1929 FD 0057 EXP: LXI R, 0
1930 FE 0058 EXP: LXI R, 0
1931 FF 0059 EXP: LXI R, 0

```



```

12E0 LY      031      RET      ; RETURN
12E1      032      ;
12E2      033      ;
12E3      034      ; TEST FOR FULL INPUT PARAMETER
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12F0      047      ;
12F1      048      ;
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1573      690      ;
1574      691      ;
1575      692      ;
1576      693      ;
1577      694      ;
1578      695      ;
1579      696      ;
157A      697      ;
157B      698      ;
157C      699      ;
157D      700      ;
157E      701      ;
157F      702      ;
1580      703      ;
1581      704      ;
1582      705      ;
1583      706      ;
1584      707      ;
1585      708      ;
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1587      710      ;
1588      711      ;
1589      712      ;
158A      713      ;
158B      714      ;
158C      715      ;
158D      716      ;
158E      717      ;
158F      718      ;
1590      719      ;
1591      720      ;
1592      721      ;
1593      722      ;
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1595      724      ;
1596      725      ;
1597      726      ;
1598      727      ;
1599      728      ;
159A      729      ;
159B      730      ;
159C      731      ;
159D      732      ;
159E      733      ;
159F      734      ;
15A0      735      ;
15A1      736      ;
15A2      737      ;
15A3      738      ;
15A4      739      ;
15A5      740      ;
15A6      741      ;
15A7      742      ;
15A8      743      ;
15A9      744      ;
15AA      745      ;
15AB      746      ;
15AC      747      ;
15AD      748      ;
15AE      749      ;
15AF      750      ;
15B0      751      ;
15B1      752      ;
15B2      753      ;
15B3      754      ;
15B4      755      ;
15B5      756      ;
15B6      757      ;
15B7      758      ;
15B8      759      ;
15B9      760      ;
15BA      761      ;
15BB      762      ;
15BC      763      ;
15BD      764      ;
15BE      765      ;
15BF      766      ;
15C0      767      ;
15C1      768      ;
15C2      769      ;
15C3      770      ;
15C4      771      ;
15C5      772      ;
15C6      773      ;
15C7      774      ;
15C8      775      ;
15C9      776      ;
15CA      777      ;
15CB      778      ;
15CC      779      ;
15CD      780      ;
15CE      781      ;
15CF      782      ;
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15D1      784      ;
15D2      785      ;
15D3      786      ;
15D4      787      ;
15D5      788      ;
15D6      789      ;
15D7      790      ;
15D8      791      ;
15D9      792      ;
15DA      793      ;
15DB      794      ;
15DC      795      ;
15DD      796      ;
15DE      797      ;
15DF      798      ;
15E0      799      ;
15E1      800      ;
15E2      801      ;
15E3      802      ;
15E4      803      ;
15E5      804      ;
15E6      805      ;
15E7      806      ;
15E8      807      ;
15E9      808      ;
15EA      809      ;
15EB      810      ;
15EC      811      ;
15ED      812      ;
15EE      813      ;
15EF      814      ;
15F0      815      ;
15F1      816      ;
15F2      817      ;
15F3      818      ;
15F4      819      ;
15F5      820      ;
15F6      821      ;
15F7      822      ;
15F8      823      ;
15F9      824      ;
15FA      825      ;
15FB      826      ;
15FC      827      ;
15FD      828      ;
15FE      829      ;
15FF      830      ;
1600      831      ;
1601      832      ;
1602      833      ;
1603      834      ;
1604      835      ;
1605      836      ;
1606      837      ;
1607      838      ;
1608      839      ;
1609      840      ;
160A      841      ;
160B      842      ;
160C      843      ;
160D      844      ;
160E      845      ;
160F      846      ;
1610      847      ;
1611      848      ;
1612      849      ;
1613      850      ;
1614      851      ;
1615      852      ;
1616      853      ;
1617      854      ;
1618      855      ;
1619      856      ;
161A      857      ;
161B      858      ;
161C      859      ;
161D      860      ;
161E      861      ;
161F      862      ;
1620      863      ;
1621      864      ;
1622      865      ;
1623      866      ;
1624      867      ;
1625      868      ;
1626      869      ;
1627      870      ;
1628      871      ;
1629      872      ;
162A      873      ;
162B      874      ;
162C      875      ;
162D      876      ;
162E      877      ;
162F      878      ;
1630      879      ;
1631      880      ;
1632      881      ;
1633      882      ;
1634      883      ;
1635      884      ;
1636      885      ;
1637      886      ;
1638      887      ;
1639      888      ;
163A      889      ;
163B      890      ;
163C      891      ;
163D      892      ;
163E      893      ;
163F      894      ;
1640      895      ;
1641      896      ;
1642      897      ;
1643      898      ;
1644      899      ;
1645      900      ;
1646      901      ;
1647      902      ;
1648      903      ;
1649      904      ;
164A      905      ;
164B      906      ;
164C      907      ;
164D      908      ;
164E      909      ;
164F      910      ;
1650      911      ;
1651      912      ;
1652      913      ;
1653      914      ;
1654      915      ;
1655      916      ;
1656      917      ;
1657      918      ;
1658      919      ;
1659      920      ;
165A      921      ;
165B      922      ;
165C      923      ;
165D      924      ;
165E      925      ;
165F      926      ;
1660      927      ;
1661      928      ;
1662      929      ;
1663      930      ;
1664      931      ;
1665      932      ;
1666      933      ;
1667      934      ;
1668      935      ;
1669      936      ;
166A      937      ;
166B      938      ;
166C      939      ;
166D      940      ;
166E      941      ;
166F      942      ;
1670      943      ;
1671      944      ;
1672      945      ;
1673      946      ;
1674      947      ;
1675      948      ;
1676      949      ;
1677      950      ;
1678      951      ;
1679      952      ;
167A      953      ;
167B      954      ;
167C      955      ;
167D      956      ;
167E      957      ;
167F      958      ;
1680      959      ;
1681      960      ;
1682      961      ;
1683      962      ;
1684      963      ;
1685      964      ;
1686      965      ;
1687      966      ;
1688      967      ;
1689      968      ;
168A      969      ;
168B      970      ;
168C      971      ;
168D      972      ;
168E      973      ;
168F      974      ;
1690      975      ;
1691      976      ;
1692      977      ;
1693      978      ;
1694      979      ;
1695      980      ;
1696      981      ;
1697      982      ;
1698      983      ;
1699      984      ;
169A      985      ;
169B      986      ;
169C      987      ;
169D      988      ;
169E      989      ;
169F      990      ;
16A0      991      ;
16A1      992      ;
16A2      993      ;
16A3      994      ;
16A4      995      ;
16A5      996      ;
16A6      997      ;
16A7      998      ;
16A8      999      ;
16A9      1000      ;
16AA      1001      ;
16AB      1002      ;
16AC      1003      ;
16AD      1004      ;
16AE      1005      ;
16AF      1006      ;
16B0      1007      ;
16B1      1008      ;
16B2      1009      ;
16B3      1010      ;
16B4      1011      ;
16B5      1012      ;
16B6      1013      ;
16B7      1014      ;
16B8      1015      ;
16B9      1016      ;
16BA      1017      ;
16BB      1018      ;
16BC      1019      ;
16BD      1020      ;
16BE      1021      ;
16BF      1022      ;
16C0      1023      ;
16C1      1024      ;
16C2      1025      ;
16C3      1026      ;
16C4      1027      ;
16C5      1028      ;
16C6      1029      ;
16C7      1030      ;
16C8      1031      ;
16C9      1032      ;
16CA      1033      ;
16CB      1034      ;
16CC      1035      ;
16CD      1036      ;
16CE      1037      ;
16CF      1038      ;
16D0      1039      ;
16D1      1040      ;
16D2      1041      ;
16D3      1042      ;
16D4      1043      ;
16D5      1044      ;
16D6      1045      ;
16D7      1046      ;
16D8      1047      ;
16D9      1048      ;
16DA      1049      ;
16DB      1050      ;
16DC      1051      ;
16DD      1052      ;
16DE      1053      ;
16DF      1054      ;
16E0      1055      ;
16E1      1056      ;
16E2      1057      ;
16E3      1058      ;
16E4      1059      ;
16E5      1060      ;
16E6      1061      ;
16E7      1062      ;
16E8     
```



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1091 0F
1092 01
1093 01
1094 10
1095 02
1096 30
1097 13
1098 02
1099 33
109A 07
109B 02
109C 1F

      DB      'M',  HLUC,  2
      DB      'P',  PLUC,  2
      DB      'S',  SLUC,  2

      DB      -1
      *****
001  *****
002  ; THIS IS TO SET UP THE CDS TIMER FOR THE UART
003  ; AT A 16X 110 BAUD RATE. THIS IS DONE BY DIVIDING
004  ; THE 2MHz 8085 CLK BY 1136.
005  ;
006  TIMER:  MVI A,  FMODE1      ; SET MODE FOR COUNTER 0
007           MVI 7BH           ; READ LST. SIG. BYTE THEN
008           MVI A,  00H       ; MUST SIG. BYTE SQUARE WAVE
009           MVI 7BH           ; AND SINKY OUTPUT
010           MVI A,  04H       ; INTR SEND CONTROL WORD
011           MVI 7BH           ; AND COUNT DOWN TO 8003
012  ;
013  ; THIS IS FOR SECOND UART
014  ; DIVIDE BY 1136 FOR 110 BAUD 0160H
015  ; DIVIDE BY 409 FOR 300 BAUD 0194H
016  ;
017  MVI A,  FMODE1      ; SET UP MODE FOR TIMER 2
018  MVI 7BH
019  MVI A,  00H
020  MVI 79H
021  MVI A,  04H
022  MVI 79H
023  RET

024  ;
025  *****
026  ;
027  ; 9511 ANTIMELIL
028  ;
029  *****
030  FMODE1:  XCHG
031           CALL PUSH4
032           XCHG
033  FMODE2:  CALL PUSH4
034           MVI A, 10H
035           STA MATHC
036           RET
037  ;
038  FMODE3:  XCHG
039           CALL PUSH4
040           XCHG
041           JMP FMODE1
042  FMODE4:  CALL PUSH4
043           MVI A, 04H
044           STA MATHC
045           RET
046  ;
047  FMODE5:  XCHG
048           CALL PUSH4
049           XCHG
050           JMP FMODE1
051  FMODE6:  CALL PUSH4
052           MVI A, 04H
053           STA MATHC
054           RET
055  ;
056  FMODE7:  XCHG
057           CALL PUSH4
058           XCHG
059           JMP FMODE1
060  FMODE8:  CALL PUSH4
061           MVI A, 04H
062           STA MATHC
063           RET
064  ;
065  FMODE9:  XCHG
066           CALL PUSH4
067           XCHG
068           JMP FMODE1
069  FMODEA:  CALL PUSH4
070           MVI A, 04H
071           STA MATHC
072           RET
073  ;
074  FMODEB:  XCHG
075           CALL PUSH4
076           XCHG
077           JMP FMODE1
078  FMODEC:  CALL PUSH4
079           MVI A, 04H
080           STA MATHC
081           RET
082  ;
083  FMODED:  XCHG
084           CALL PUSH4
085           XCHG
086           JMP FMODE1
087  FMODEE:  CALL PUSH4
088           MVI A, 04H
089           STA MATHC
090           RET
091  ;
092  FMODEF:  XCHG
093           CALL PUSH4
094           XCHG
095           JMP FMODE1
096  FMODEG:  CALL PUSH4
097           MVI A, 04H
098           STA MATHC
099           RET
100  ;
101  FMODEH:  XCHG
102           CALL PUSH4
103           XCHG
104           JMP FMODE1
105  FMODEI:  CALL PUSH4
106           MVI A, 04H
107           STA MATHC
108           RET
109  ;
110  FMODEJ:  XCHG
111           CALL PUSH4
112           XCHG
113           JMP FMODE1
114  FMODEK:  CALL PUSH4
115           MVI A, 04H
116           STA MATHC
117           RET
118  ;
119  FMODEL:  XCHG
120           CALL PUSH4
121           XCHG
122           JMP FMODE1
123  FMODEM:  CALL PUSH4
124           MVI A, 04H
125           STA MATHC
126           RET
127  ;
128  FMODEN:  XCHG
129           CALL PUSH4
130           XCHG
131           JMP FMODE1
132  FMODEO:  CALL PUSH4
133           MVI A, 04H
134           STA MATHC
135           RET
136  ;
137  FMODEP:  XCHG
138           CALL PUSH4
139           XCHG
140           JMP FMODE1
141  FMODEQ:  CALL PUSH4
142           MVI A, 04H
143           STA MATHC
144           RET
145  ;
146  FMODER:  XCHG
147           CALL PUSH4
148           XCHG
149           JMP FMODE1
150  FMODES:  CALL PUSH4
151           MVI A, 04H
152           STA MATHC
153           RET
154  ;
155  FMODET:  XCHG
156           CALL PUSH4
157           XCHG
158           JMP FMODE1
159  FMODEU:  CALL PUSH4
160           MVI A, 04H
161           STA MATHC
162           RET
163  ;
164  FMODEV:  XCHG
165           CALL PUSH4
166           XCHG
167           JMP FMODE1
168  FMODEW:  CALL PUSH4
169           MVI A, 04H
170           STA MATHC
171           RET
172  ;
173  FMODEX:  XCHG
174           CALL PUSH4
175           XCHG
176           JMP FMODE1
177  FMODEY:  CALL PUSH4
178           MVI A, 04H
179           STA MATHC
180           RET
181  ;
182  FMODEZ:  XCHG
183           CALL PUSH4
184           XCHG
185           JMP FMODE1
186  FMODEAA:  CALL PUSH4
187           MVI A, 04H
188           STA MATHC
189           RET
190  ;
191  FMODEB:  XCHG
192           CALL PUSH4
193           XCHG
194           JMP FMODE1
195  FMODEBB:  CALL PUSH4
196           MVI A, 04H
197           STA MATHC
198           RET
199  ;
200  FMODEC:  XCHG
201           CALL PUSH4
202           XCHG
203           JMP FMODE1
204  FMODECB:  CALL PUSH4
205           MVI A, 04H
206           STA MATHC
207           RET
208  ;
209  FMODED:  XCHG
210           CALL PUSH4
211           XCHG
212           JMP FMODE1
213  FMODEDC:  CALL PUSH4
214           MVI A, 04H
215           STA MATHC
216           RET
217  ;
218  FMODEE:  XCHG
219           CALL PUSH4
220           XCHG
221           JMP FMODE1
222  FMODEDE:  CALL PUSH4
223           MVI A, 04H
224           STA MATHC
225           RET
226  ;
227  FMODEF:  XCHG
228           CALL PUSH4
229           XCHG
230           JMP FMODE1
231  FMODEDF:  CALL PUSH4
232           MVI A, 04H
233           STA MATHC
234           RET
235  ;
236  FMODEG:  XCHG
237           CALL PUSH4
238           XCHG
239           JMP FMODE1
240  FMODEDG:  CALL PUSH4
241           MVI A, 04H
242           STA MATHC
243           RET
244  ;
245  FMODEH:  XCHG
246           CALL PUSH4
247           XCHG
248           JMP FMODE1
249  FMODEDH:  CALL PUSH4
250           MVI A, 04H
251           STA MATHC
252           RET
253  ;
254  FMODEI:  XCHG
255           CALL PUSH4
256           XCHG
257           JMP FMODE1
258  FMODEDI:  CALL PUSH4
259           MVI A, 04H
260           STA MATHC
261           RET
262  ;
263  FMODEJ:  XCHG
264           CALL PUSH4
265           XCHG
266           JMP FMODE1
267  FMODEDJ:  CALL PUSH4
268           MVI A, 04H
269           STA MATHC
270           RET
271  ;
272  FMODEK:  XCHG
273           CALL PUSH4
274           XCHG
275           JMP FMODE1
276  FMODEDK:  CALL PUSH4
277           MVI A, 04H
278           STA MATHC
279           RET
280  ;
281  FMODEL:  XCHG
282           CALL PUSH4
283           XCHG
284           JMP FMODE1
285  FMODEDL:  CALL PUSH4
286           MVI A, 04H
287           STA MATHC
288           RET
289  ;
290  FMODEM:  XCHG
291           CALL PUSH4
292           XCHG
293           JMP FMODE1
294  FMODEDM:  CALL PUSH4
295           MVI A, 04H
296           STA MATHC
297           RET
298  ;
299  FMODEN:  XCHG
300           CALL PUSH4
301           XCHG
302           JMP FMODE1
303  FMODEDN:  CALL PUSH4
304           MVI A, 04H
305           STA MATHC
306           RET
307  ;
308  FMODEO:  XCHG
309           CALL PUSH4
310           XCHG
311           JMP FMODE1
312  FMODEDO:  CALL PUSH4
313           MVI A, 04H
314           STA MATHC
315           RET
316  ;
317  FMODEP:  XCHG
318           CALL PUSH4
319           XCHG
320           JMP FMODE1
321  FMODEDP:  CALL PUSH4
322           MVI A, 04H
323           STA MATHC
324           RET
325  ;
326  FMODEQ:  XCHG
327           CALL PUSH4
328           XCHG
329           JMP FMODE1
330  FMODEDQ:  CALL PUSH4
331           MVI A, 04H
332           STA MATHC
333           RET
334  ;
335  FMODER:  XCHG
336           CALL PUSH4
337           XCHG
338           JMP FMODE1
339  FMODEDR:  CALL PUSH4
340           MVI A, 04H
341           STA MATHC
342           RET
343  ;
344  FMODES:  XCHG
345           CALL PUSH4
346           XCHG
347           JMP FMODE1
348  FMODEDS:  CALL PUSH4
349           MVI A, 04H
350           STA MATHC
351           RET
352  ;
353  FMODET:  XCHG
354           CALL PUSH4
355           XCHG
356           JMP FMODE1
357  FMODEDT:  CALL PUSH4
358           MVI A, 04H
359           STA MATHC
360           RET
361  ;
362  FMODEU:  XCHG
363           CALL PUSH4
364           XCHG
365           JMP FMODE1
366  FMODEDU:  CALL PUSH4
367           MVI A, 04H
368           STA MATHC
369           RET
370  ;
371  FMODEV:  XCHG
372           CALL PUSH4
373           XCHG
374           JMP FMODE1
375  FMODEDV:  CALL PUSH4
376           MVI A, 04H
377           STA MATHC
378           RET
379  ;
380  FMODEW:  XCHG
381           CALL PUSH4
382           XCHG
383           JMP FMODE1
384  FMODEDW:  CALL PUSH4
385           MVI A, 04H
386           STA MATHC
387           RET
388  ;
389  FMODEX:  XCHG
390           CALL PUSH4
391           XCHG
392           JMP FMODE1
393  FMODEDX:  CALL PUSH4
394           MVI A, 04H
395           STA MATHC
396           RET
397  ;
398  FMODEY:  XCHG
399           CALL PUSH4
400           XCHG
401           JMP FMODE1
402  FMODEDY:  CALL PUSH4
403           MVI A, 04H
404           STA MATHC
405           RET
406  ;
407  FMODEZ:  XCHG
408           CALL PUSH4
409           XCHG
410           JMP FMODE1
411  FMODEDZ:  CALL PUSH4
412           MVI A, 04H
413           STA MATHC
414           RET
415  ;
416  FMODEAA:  XCHG
417           CALL PUSH4
418           XCHG
419           JMP FMODE1
420  FMODEDAA:  CALL PUSH4
421           MVI A, 04H
422           STA MATHC
423           RET
424  ;
425  FMODEB:  XCHG
426           CALL PUSH4
427           XCHG
428           JMP FMODE1
429  FMODEDB:  CALL PUSH4
430           MVI A, 04H
431           STA MATHC
432           RET
433  ;
434  FMODEC:  XCHG
435           CALL PUSH4
436           XCHG
437           JMP FMODE1
438  FMODEDC:  CALL PUSH4
439           MVI A, 04H
440           STA MATHC
441           RET
442  ;
443  FMODED:  XCHG
444           CALL PUSH4
445           XCHG
446           JMP FMODE1
447  FMODEDD:  CALL PUSH4
448           MVI A, 04H
449           STA MATHC
450           RET
451  ;
452  FMODEE:  XCHG
453           CALL PUSH4
454           XCHG
455           JMP FMODE1
456  FMODEDE:  CALL PUSH4
457           MVI A, 04H
458           STA MATHC
459           RET
460  ;
461  FMODEF:  XCHG
462           CALL PUSH4
463           XCHG
464           JMP FMODE1
465  FMODEDF:  CALL PUSH4
466           MVI A, 04H
467           STA MATHC
468           RET
469  ;
470  FMODEG:  XCHG
471           CALL PUSH4
472           XCHG
473           JMP FMODE1
474  FMODEDG:  CALL PUSH4
475           MVI A, 04H
476           STA MATHC
477           RET
478  ;
479  FMODEH:  XCHG
480           CALL PUSH4
481           XCHG
482           JMP FMODE1
483  FMODEDH:  CALL PUSH4
484           MVI A, 04H
485           STA MATHC
486           RET
487  ;
488  FMODEI:  XCHG
489           CALL PUSH4
490           XCHG
491           JMP FMODE1
492  FMODEDI:  CALL PUSH4
493           MVI A, 04H
494           STA MATHC
495           RET
496  ;
497  FMODEJ:  XCHG
498           CALL PUSH4
499           XCHG
500           JMP FMODE1
501  FMODEDJ:  CALL PUSH4
502           MVI A, 04H
503           STA MATHC
504           RET
505  ;
506  FMODEK:  XCHG
507           CALL PUSH4
508           XCHG
509           JMP FMODE1
510  FMODEDK:  CALL PUSH4
511           MVI A, 04H
512           STA MATHC
513           RET
514  ;
515  FMODEL:  XCHG
516           CALL PUSH4
517           XCHG
518           JMP FMODE1
519  FMODEDL:  CALL PUSH4
520           MVI A, 04H
521           STA MATHC
522           RET
523  ;
524  FMODEM:  XCHG
525           CALL PUSH4
526           XCHG
527           JMP FMODE1
528  FMODEDM:  CALL PUSH4
529           MVI A, 04H
530           STA MATHC
531           RET
532  ;
529  FMODEN:  XCHG
533           CALL PUSH4
534           XCHG
535           JMP FMODE1
536  FMODEDN:  CALL PUSH4
537           MVI A, 04H
538           STA MATHC
539           RET
540  ;
541  FMODEO:  XCHG
542           CALL PUSH4
543           XCHG
544           JMP FMODE1
545  FMODEDO:  CALL PUSH4
546           MVI A, 04H
547           STA MATHC
548           RET
549  ;
550  FMODEP:  XCHG
551           CALL PUSH4
552           XCHG
553           JMP FMODE1
554  FMODEDP:  CALL PUSH4
555           MVI A, 04H
556           STA MATHC
557           RET
558  ;
559  FMODEQ:  XCHG
560           CALL PUSH4
561           XCHG
562           JMP FMODE1
563  FMODEDQ:  CALL PUSH4
564           MVI A, 04H
565           STA MATHC
566           RET
567  ;
568  FMODER:  XCHG
569           CALL PUSH4
570           XCHG
571           JMP FMODE1
572  FMODEDR:  CALL PUSH4
573           MVI A, 04H
574           STA MATHC
575           RET
576  ;
577  FMODES:  XCHG
578           CALL PUSH4
579           XCHG
580           JMP FMODE1
581  FMODEDS:  CALL PUSH4
582           MVI A, 04H
583           STA MATHC
584           RET
585  ;
586  FMODET:  XCHG
587           CALL PUSH4
588           XCHG
589           JMP FMODE1
590  FMODEDT:  CALL PUSH4
591           MVI A, 04H
592           STA MATHC
593           RET
594  ;
595  FMODEU:  XCHG
596           CALL PUSH4
597           XCHG
598           JMP FMODE1
599  FMODEDU:  CALL PUSH4
600           MVI A, 04H
601           STA MATHC
602           RET
603  ;
604  FMODEV:  XCHG
605           CALL PUSH4
606           XCHG
607           JMP FMODE1
608  FMODEDV:  CALL PUSH4
609           MVI A, 04H
610           STA MATHC
611           RET
612  ;
613  FMODEW:  XCHG
614           CALL PUSH4
615           XCHG
616           JMP FMODE1
617  FMODEDW:  CALL PUSH4
618           MVI A, 04H
619           STA MATHC
620           RET
621  ;
622  FMODEX:  XCHG
623           CALL PUSH4
624           XCHG
625           JMP FMODE1
626  FMODEDX:  CALL PUSH4
627           MVI A, 04H
628           STA MATHC
629           RET
630  ;
631  FMODEY:  XCHG
632           CALL PUSH4
633           XCHG
634           JMP FMODE1
635  FMODEDY:  CALL PUSH4
636           MVI A, 04H
637           STA MATHC
638           RET
639  ;
640  FMODEZ:  XCHG
641           CALL PUSH4
642           XCHG
643           JMP FMODE1
644  FMODEDZ:  CALL PUSH4
645           MVI A, 04H
646           STA MATHC
647           RET
648  ;
649  FMODEAA:  XCHG
650           CALL PUSH4
651           XCHG
652           JMP FMODE1
653  FMODEDAA:  CALL PUSH4
654           MVI A, 04H
655           STA MATHC
656           RET
657  ;
658  FMODEB:  XCHG
659           CALL PUSH4
660           XCHG
661           JMP FMODE1
662  FMODEDB:  CALL PUSH4
663           MVI A, 04H
664           STA MATHC
665           RET
666  ;
667  FMODEC:  XCHG
668           CALL PUSH4
669           XCHG
670           JMP FMODE1
671  FMODEDC:  CALL PUSH4
672           MVI A, 04H
673           STA MATHC
674           RET
675  ;
676  FMODED:  XCHG
677           CALL PUSH4
678           XCHG
679           JMP FMODE1
680  FMODEDD:  CALL PUSH4
681           MVI A, 04H
682           STA MATHC
683           RET
684  ;
685  FMODEE:  XCHG
686           CALL PUSH4
687           XCHG
688           JMP FMODE1
689  FMODEDE:  CALL PUSH4
690           MVI A, 04H
691           STA MATHC
692           RET
693  ;
694  FMODEF:  XCHG
695           CALL PUSH4
696           XCHG
697           JMP FMODE1
698  FMODEDF:  CALL PUSH4
699           MVI A, 04H
700           STA MATHC
701           RET
702  ;
699  FMODEG:  XCHG
703           CALL PUSH4
704           XCHG
705           JMP FMODE1
706  FMODEDG:  CALL PUSH4
707           MVI A, 04H
708           STA MATHC
709           RET
710  ;
701  FMODEH:  XCHG
711           CALL PUSH4
712           XCHG
713           JMP FMODE1
714  FMODEDH:  CALL PUSH4
715           MVI A, 04H
716           STA MATHC
717           RET
718  ;
703  FMODEI:  XCHG
719           CALL PUSH4
720           XCHG
721           JMP FMODE1
722  FMODEDI:  CALL PUSH4
723           MVI A, 04H
724           STA MATHC
725           RET
726  ;
705  FMODEJ:  XCHG
727           CALL PUSH4
728           XCHG
729           JMP FMODE1
730  FMODEDJ:  CALL PUSH4
731           MVI A, 04H
732           STA MATHC
733           RET
734  ;
707  FMODEK:  XCHG
735           CALL PUSH4
736           XCHG
737           JMP FMODE1
738  FMODEDK:  CALL PUSH4
739           MVI A, 04H
740           STA MATHC
741           RET
742  ;
709  FMODEL:  XCHG
743           CALL PUSH4
744           XCHG
745           JMP FMODE1
746  FMODEDL:  CALL PUSH4
747           MVI A, 04H
748           STA MATHC
749           RET
750  ;
711  FMODEM:  XCHG
751           CALL PUSH4
752           XCHG
753           JMP FMODE1
754  FMODEDM:  CALL PUSH4
755           MVI A, 04H
756           STA MATHC
757           RET
758  ;
713  FMODEN:  XCHG
759           CALL PUSH4
760           XCHG
761           JMP FMODE1
762  FMODEDN:  CALL PUSH4
763           MVI A, 04H
764           STA MATHC
765           RET
766  ;
715  FMODEO:  XCHG
767           CALL PUSH4
768           XCHG
769           JMP FMODE1
770  FMODEDO:  CALL PUSH4
771           MVI A, 04H
772           STA MATHC
773           RET
774  ;
717  FMODEP:  XCHG
775           CALL PUSH4
776           XCHG
777           JMP FMODE1
778  FMODEDP:  CALL PUSH4
779           MVI A, 04H
780           STA MATHC
781           RET
782  ;
719  FMODEQ:  XCHG
783           CALL PUSH4
784           XCHG
785           JMP FMODE1
786  FMODEDQ:  CALL PUSH4
787           MVI A, 04H
788           STA MATHC
789           RET
790  ;
721  FMODER:  XCHG
791           CALL PUSH4
792           XCHG
793           JMP FMODE1
794  FMODEDR:  CALL PUSH4
795           MVI A, 04H
796           STA MATHC
797           RET
798  ;
723  FMODES:  XCHG
799           CALL PUSH4
800           XCHG
801           JMP FMODE1
802  FMODEDS:  CALL PUSH4
803           MVI A, 04H
804           STA MATHC
805           RET
806  ;
725  FMODET:  XCHG
807           CALL PUSH4
808           XCHG
809           JMP FMODE1
810  FMODEDT:  CALL PUSH4
811           MVI A, 04H
812           STA MATHC
813           RET
814  ;
727  FMODEU:  XCHG
815           CALL PUSH4
816           XCHG
817           JMP FMODE1
818  FMODEDU:  CALL PUSH4
819           MVI A, 04H
820           STA MATHC
821           RET
822  ;
729  FMODEV:  XCHG
823           CALL PUSH4
824           XCHG
825           JMP FMODE1
826  FMODEDV:  CALL PUSH4
827           MVI A, 04H
828           STA MATHC
829           RET
830  ;
731  FMODEW:  XCHG
831           CALL PUSH4
832           XCHG
833           JMP FMODE1
834  FMODEDW:  CALL PUSH4
835           MVI A, 04H
836           STA MATHC
837           RET
838  ;
733  FMODEX:  XCHG
839           CALL PUSH4
840           XCHG
841           JMP FMODE1
842  FMODEDX:  CALL PUSH4
843           MVI A, 04H
844           STA MATHC
845           RET
846  ;
735  FMODEY:  XCHG
847           CALL PUSH4
848           XCHG
849           JMP FMODE1
850  FMODEDY:  CALL PUSH4
851           MVI A, 04H
852           STA MATHC
853           RET
854  ;
737  FMODEZ:  XCHG
855           CALL PUSH4
856           XCHG
857           JMP FMODE1
858  FMODEDZ:  CALL PUSH4
859           MVI A, 04H
860           STA MATHC
861           RET
862  ;
739  FMODEAA:  XCHG
863           CALL PUSH4
864           XCHG
865           JMP FMODE1
866  FMODEDAA:  CALL PUSH4
867           MVI A, 04H
868           STA MATHC
869           RET
870  ;
741  FMODEB:  XCHG
871           CALL PUSH4
872           XCHG
873           JMP FMODE1
874  FMODEDB:  CALL PUSH4
875           MVI A, 04H
876           STA MATHC
877           RET
878  ;
743  FMODEC:  XCHG
879           CALL PUSH4
880           XCHG
881           JMP FMODE1
882  FMODEDC:  CALL PUSH4
883           MVI A, 04H
884           STA MATHC
885           RET
886  ;
745  FMODED:  XCHG
887           CALL PUSH4
888           XCHG
889           JMP FMODE1
890  FMODEDD:  CALL PUSH4
891           MVI A, 04H
892           STA MATHC
893           RET
894  ;
747  FMODEE:  XCHG
895           CALL PUSH4
896           XCHG
897           JMP FMODE1
898  FMODEDE:  CALL PUSH4
899           MVI A, 04H
900           STA MATHC
901           RET
902  ;
749  FMODEF:  XCHG
903           CALL PUSH4
904           XCHG
905           JMP FMODE1
906  FMODEDF:  CALL PUSH4
907           MVI A, 04H
908           STA MATHC
909           RET
910  ;
751  FMODEG:  XCHG
911           CALL PUSH4
912           XCHG
913           JMP FMODE1
914  FMODEDG:  CALL PUSH4
915           MVI A, 04H
916           STA MATHC
917           RET
918  ;
753  FMODEH:  XCHG
919           CALL PUSH4
920           XCHG
921           JMP FMODE1
922  FMODEDH:  CALL PUSH4
923           MVI A, 04H
924           STA MATHC
925           RET
926  ;
755  FMODEI:  XCHG
927           CALL PUSH4
928           XCHG
929           JMP FMODE1
930  FMODEDI:  CALL PUSH4
931           MVI A, 04H
932           STA MATHC
933           RET
934  ;
757  FMODEJ:  XCHG
935           CALL PUSH4
936           XCHG
937           JMP FMODE1
938  FMODEDJ:  CALL PUSH4
939           MVI A, 04H
940           STA MATHC
941           RET
942  ;
759  FMODEK:  XCHG
943           CALL PUSH4
944           XCHG
945           JMP FMODE1
946  FMODEDK:  CALL PUSH4
947           MVI A, 04H
948           STA MATHC
949           RET
950  ;
761  FMODEL:  XCHG
951           CALL PUSH4
952           XCHG
953           JMP FMODE1
954  FMODEDL:  CALL PUSH4
955           MVI A, 04H
956           STA MATHC
957           RET
958  ;
763  FMODEM:  XCHG
959           CALL PUSH4
960           XCHG
961           JMP FMODE1
962  FMODEDM:  CALL PUSH4
963           MVI A, 04H
964           STA MATHC
965           RET
966  ;
765  FMODEN:  XCHG
967           CALL PUSH4
968           XCHG
969           JMP FMODE1
970  FMODEDN:  CALL PUSH4
971           MVI A, 04H
972           STA MATHC
973           RET
974  ;
767  FMODEO:  XCHG
975           CALL PUSH4
976           XCHG
977           JMP FMODE1
978  FMODEDO:  CALL PUSH4
979           MVI A, 04H
980           STA MATHC
981           RET
982  ;
769  FMODEP:  XCHG
983           CALL PUSH4
984           XCHG
985           JMP FMODE1
986  FMODEDP:  CALL PUSH4
987           MVI A, 04H
988           STA MATHC
989           RET
990  ;
771  FMODEQ:  XCHG
989           CALL PUSH4
990           XCHG
991           JMP FMODE1
992  FMODEDQ:  CALL PUSH4
993           MVI A, 04H
994           STA MATHC
995           RET
996  ;
773  FMODER:  XCHG
997           CALL PUSH4
998           XCHG
999           JMP FMODE1
1000  FMODEDR:  CALL PUSH4
1001           MVI A, 04H
1002           STA MATHC
1003           RET
1004  ;
775  FMODES:  XCHG
1005           CALL PUSH4
1006           XCHG
1007           JMP FMODE1
1008  FMODEDS:  CALL PUSH4
1009           MVI A, 04H
1010           STA MATHC
1011           RET
1012  ;
777  FMODET:  XCHG
1013           CALL PUSH4
1014           XCHG
1015           JMP FMODE1
1016  FMODEDT:  CALL PUSH4
1017           MVI A, 04H
1018           STA MATHC
1019           RET
1020  ;
779  FMODEU:  XCHG
1021           CALL PUSH4
1022           XCHG
1023           JMP FMODE1
1024  FMODEDU:  CALL PUSH4
1025           MVI A, 04H
1026           STA MATHC
1027           RET
1028  ;
781  FMODEV:  XCHG
1029           CALL PUSH4
1030           XCHG
1031           JMP FMODE1
1032  FMODEDV:  CALL PUSH4
1033           MVI A, 04H
1034           STA MATHC
1035           RET
1036  ;
783  FMODEW:  XCHG
1037           CALL PUSH4
1038           XCHG
1039           JMP FMODE1
1040  FMODEDW:  CALL PUSH4
1041           MVI A, 04H
1042           STA MATHC
1043           RET
1044  ;
785  FMODEX:  XCHG
1045           CALL PUSH4
1046           XCHG
1047           JMP FMODE1
1048  FMODEDX:  CALL PUSH4
1049           MVI A, 04H
1050           STA MATHC
1051           RET
1052  ;
787  FMODEY:  XCHG
1053           CALL PUSH4
1054           XCHG
1055           JMP FMODE1
1056  FMODEDY:  CALL PUSH4
1057           MVI A, 04H
1058           STA MATHC
1059           RET
1060  ;
789  FMODEZ:  XCHG
1061           CALL PUSH4
1062           XCHG
1063           JMP FMODE1
1064  FMODEDZ:  CALL PUSH4
1065           MVI A, 04H
1066           STA MATHC
1067           RET
1068  ;
791  FMODEAA:  XCHG
1069           CALL PUSH4
1070           XCHG
1071           JMP FMODE1
1072  FMODEDAA:  CALL PUSH4
1073           MVI A, 04H
1074           STA MATHC
1075           RET
1076  ;
793  FMODEB:  XCHG
1077           CALL PUSH4
1078           XCHG
1079           JMP FMODE1
1080  FMODEDB:  CALL PUSH4
1081           MVI A, 04H
1082           STA MATHC
1083           RET
1084  ;
795  FMODEC:  XCHG
1085           CALL PUSH4
1086           XCHG
1087           JMP FMODE1
1088  FMODEDC:  CALL PUSH4
1089           MVI A, 04H
1090           STA MATHC
1091           RET
1092  ;
797  FMODED:  XCHG
1093           CALL PUSH4
1094           XCHG
1095           JMP FMODE1
1096  FMODEDD:  CALL PUSH4
1097           MVI A, 04H
1098           STA MATHC
1099           RET
1100  ;
799  FMODEE:  XCHG
1101           CALL PUSH4
1102           XCHG
1103           JMP FMODE1
1104  FMODEDE:  CALL PUSH4
1105           MVI A, 04H
1106           STA MATHC
1107           RET
1108  ;
801  FMODEF:  XCHG
1109           CALL PUSH4
1110           XCHG
1111           JMP FMODE1
1112  FMODEDF:  CALL PUSH4
1113           MVI A, 04H
1114           STA MATHC
1115           RET
1116  ;
803  FMODEG:  XCHG
1117           CALL PUSH4
1118           XCHG
1119           JMP FMODE1
1120  FMODEDG:  CALL PUSH4
1121           MVI A, 04H
1122           STA MATHC
1123           RET
1124  ;
805  FMODEH:  XCHG
1125           CALL PUSH4
1126           XCHG
1127           JMP FMODE1
1128  FMODEDH:  CALL PUSH4
1129           MVI A, 04H
1130           STA MATHC
1131           RET
1132  ;
807  FMODEI:  XCHG
1133           CALL PUSH4
1134           XCHG
1135           JMP FMODE1
1136  FMODEDI:  CALL PUSH4
1137           MVI A, 04H
1138           STA MATHC
1139           RET
1140  ;
809  FMODEJ:  XCHG
1141           CALL PUSH4
1142           XCHG
1143           JMP FMODE1
1144  FMODEDJ:  CALL PUSH4
1145           MVI A, 04H
1146           STA MATHC
1147           RET
1148  ;
811  FMODEK:  XCHG
1149           CALL PUSH4
1150           XCHG
1151           JMP FMODE1
1152  FMODEDK:  CALL PUSH4
1153           MVI A, 04H
1154           STA MATHC
1155           RET
1156  ;
813  FMODEL:  XCHG
1157           CALL PUSH4
1158           XCHG
1159           JMP FMODE1
1160  FMODEDL:  CALL PUSH4
1161           MVI A, 04H
1162           STA MATHC
1163           RET
1164  ;
815  FMODEM:  XCHG
1165           CALL PUSH4
1166           XCHG
1167           JMP FMODE1
1168  FMODEDM:  CALL PUSH4
1169           MVI A, 04H
1170           STA MATHC
1171           RET
1172  ;
817  FMODEN:  XCHG
1173           CALL PUSH4
1174           XCHG
1175           JMP FMODE1
1176  FMODEDN:  CALL PUSH4
1177           MVI A, 04H
1178           STA MATHC
1179           RET
1180  ;
819  FMODEO:  XCHG
1181           CALL PUSH4
1182           XCHG
1183           JMP FMODE1
1184  FMODEDO:  CALL PUSH4
1185           MVI A, 04H
1186           STA MATHC
1187           RET
1188  ;
821
```


[illegible]


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-39 00ZNY: US 4
-40 00ZNY: US 4
-41 00ZNY: US 4
-42 00ZNY: US 4
-43 00ZNY: US 4
-44 00ZNY: US 4
-45 00ZNY: US 4
-46 00ZNY: US 4
-47 00ZNY: US 4
-48 00ZNY: US 4
-49 00ZNY: US 4
-50 00ZNY: US 4
-51 00ZNY: US 4
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-93 00ZNY: US 4
-94 00ZNY: US 4
-95 00ZNY: US 4
-96 00ZNY: US 4
-97 00ZNY: US 4
-98 00ZNY: US 4
-99 00ZNY: US 4

```

VARIABLES UP PROCESS, UPDATE, AND DISPLAY THAT ARE INITIALIZED TO A
 NON-ZERO VALUE AT THE BEGINNING INITIAL.

```

-100 00ZNY: US 4
-101 00ZNY: US 4
-102 00ZNY: US 4
-103 00ZNY: US 4
-104 00ZNY: US 4
-105 00ZNY: US 4
-106 00ZNY: US 4
-107 00ZNY: US 4
-108 00ZNY: US 4
-109 00ZNY: US 4
-110 00ZNY: US 4
-111 00ZNY: US 4
-112 00ZNY: US 4
-113 00ZNY: US 4
-114 00ZNY: US 4
-115 00ZNY: US 4
-116 00ZNY: US 4
-117 00ZNY: US 4
-118 00ZNY: US 4
-119 00ZNY: US 4
-120 00ZNY: US 4
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-122 00ZNY: US 4
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-124 00ZNY: US 4
-125 00ZNY: US 4
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-188 00ZNY: US 4
-189 00ZNY: US 4
-190 00ZNY: US 4
-191 00ZNY: US 4
-192 00ZNY: US 4
-193 00ZNY: US 4
-194 00ZNY: US 4
-195 00ZNY: US 4
-196 00ZNY: US 4
-197 00ZNY: US 4
-198 00ZNY: US 4
-199 00ZNY: US 4

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0001
 0001

AB25	174	MSHE: US	4	CRITICAL AND TEMPORARY STORAGE FOR USFLT
AB26	180	PLP12: US	4	TEMPORARY STORAGE FOR USFLT
AB27	181	PL1: US	4	TEMPORARY STORAGE FOR USFLT
AB28	182	PL2: US	4	TEMPORARY STORAGE FOR USFLT
AB29	183	PL3: US	4	TEMPORARY STORAGE FOR USFLT
AB30	184	PL4: US	4	TEMPORARY STORAGE FOR USFLT
AB31	185	PL5: US	4	TEMPORARY STORAGE FOR USFLT
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AB33	187	PL7: US	4	TEMPORARY STORAGE FOR USFLT
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AB35	189	PL9: US	4	TEMPORARY STORAGE FOR USFLT
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0994 JA 2971 994
0995 JA 2972 995
0996 JA 2973 996
0997 JA 2974 997
0998 JA 2975 998
0999 JA 2976 999
1000 JA 2977 1000

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0319	7C	704	RVV	MM
031A	ES	705	ARI	80H
031B	L2	706	JM2	LCV
031F	84	707	URA	M
0320	CC	708	JM2	CCB
0321	1L	709	RGV	ALL
0322	7A	710	QVW	QVW
0323	7A	711	JZ	LCV
0324	7A	712	SLM	INDEX2
0325	7A	713	SLM	INDEX1
032F	11	714	LXI	DIRY
0331	19	715	MM	M
0332	7A	716	CALL	PUSHA
0333	7A	717	CALL	INDEX2
0334	19	718	LXI	DIRY
033C	19	719	MM	M
033D	LD	720	CALL	PSUM1
033E	21	721	LXI	INDEX0
0340	LD	722	CALL	PAUL1
0341	21	723	LXI	INDEX
0342	7A	724	MM	M
0343	7A	725	MM	M
0344	7A	726	JM2	ES
0345	7A	727	JM2	ES
0346	7A	728	CALL	PAUL1
0347	LD	729	CALL	PAUL4
0348	7A	730	MM	M
0349	7A	731	MM	M
034A	7A	732	CF3	51
034B	7A	733	JM1	INDEX3
034C	7A	734	MM	M
034D	7A	735	MM	M
034E	7A	736	MM	M
034F	7A	737	MM	M
0350	7A	738	MM	M
0351	7A	739	MM	M
0352	7A	740	MM	M
0353	7A	741	MM	M
0354	7A	742	MM	M
0355	7A	743	MM	M
0356	7A	744	MM	M
0357	7A	745	MM	M
0358	7A	746	MM	M
0359	7A	747	MM	M
035A	7A	748	MM	M
035B	7A	749	MM	M
035C	7A	750	MM	M
035D	7A	751	MM	M
035E	7A	752	MM	M
035F	7A	753	MM	M
0360	7A	754	MM	M
0361	7A	755	MM	M
0362	7A	756	MM	M
0363	7A	757	MM	M
0364	7A	758	MM	M
0365	7A	759	MM	M
0366	7A	760	MM	M
0367	7A	761	MM	M
0368	7A	762	MM	M
0369	7A	763	MM	M
036A	7A	764	MM	M
036B	7A	765	MM	M
036C	7A	766	MM	M
036D	7A	767	MM	M
036E	7A	768	MM	M
036F	7A	769	MM	M
0370	7A	770	MM	M
0371	7A	771	MM	M
0372	7A	772	MM	M
0373	7A	773	MM	M
0374	7A	774	MM	M
0375	7A	775	MM	M
0376	7A	776	MM	M
0377	7A	777	MM	M
0378	7A	778	MM	M
0379	7A	779	MM	M
037A	7A	780	MM	M
037B	7A	781	MM	M
037C	7A	782	MM	M
037D	7A	783	MM	M
037E	7A	784	MM	M
037F	7A	785	MM	M
0380	7A	786	MM	M
0381	7A	787	MM	M
0382	7A	788	MM	M
0383	7A	789	MM	M
0384	7A	790	MM	M
0385	7A	791	MM	M
0386	7A	792	MM	M
0387	7A	793	MM	M
0388	7A	794	MM	M
0389	7A	795	MM	M
038A	7A	796	MM	M
038B	7A	797	MM	M
038C	7A	798	MM	M
038D	7A	799	MM	M
038E	7A	800	MM	M
038F	7A	801	MM	M
0390	7A			


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008 ; UPDATE DATA FOR THE DISPLAY
009 ; *****
010 ; *****
011 ; *****
012 ; *****
013 ; *****
014 ; *****
015 ; *****
016 ; *****
017 ; *****
018 ; *****
019 ; *****
020 ; *****
021 ; *****
022 ; *****
023 ; *****
024 ; *****
025 ; *****
026 ; *****
027 ; *****
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030 ; *****
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033 ; *****
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039 ; *****
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042 ; *****
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046 ; *****
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074 ; *****
075 ; *****
076 ; *****
077 ; *****
078 ; *****
079 ; *****
080 ; *****
081 ; *****
082 ; *****
083 ; *****
084 ; *****
085 ; *****
086 ; *****
087 ; *****
088 ; *****
089 ; *****
090 ; *****
091 ; *****
092 ; *****
093 ; *****
094 ; *****
095 ; *****
096 ; *****
097 ; *****
098 ; *****
099 ; *****
100 ; *****

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0481 11 851F 854    LTI M177
0482 11 851F 855    CALL F50B1
0483 11 851F 856    LTI M177
0484 11 851F 857    CALL F50B1
0485 11 851F 858    LTI M177
0486 11 851F 859    CALL F50B1
0487 11 851F 860    LTI M177
0488 11 851F 861    CALL F50B1
0489 11 851F 862    LTI M177
0490 11 851F 863    CALL F50B1
0491 11 851F 864    LTI M177
0492 11 851F 865    CALL F50B1
0493 11 851F 866    LTI M177
0494 11 851F 867    CALL F50B1
0495 11 851F 868    LTI M177
0496 11 851F 869    CALL F50B1
0497 11 851F 870    LTI M177
0498 11 851F 871    CALL F50B1
0499 11 851F 872    LTI M177
0500 11 851F 873    CALL F50B1
0501 11 851F 874    LTI M177
0502 11 851F 875    CALL F50B1
0503 11 851F 876    LTI M177
0504 11 851F 877    CALL F50B1
0505 11 851F 878    LTI M177
0506 11 851F 879    CALL F50B1
0507 11 851F 880    LTI M177
0508 11 851F 881    CALL F50B1
0509 11 851F 882    LTI M177
0510 11 851F 883    CALL F50B1
0511 11 851F 884    LTI M177
0512 11 851F 885    CALL F50B1
0513 11 851F 886    LTI M177
0514 11 851F 887    CALL F50B1
0515 11 851F 888    LTI M177
0516 11 851F 889    CALL F50B1
0517 11 851F 890    LTI M177
0518 11 851F 891    CALL F50B1
0519 11 851F 892    LTI M177
0520 11 851F 893    CALL F50B1
0521 11 851F 894    LTI M177
0522 11 851F 895    CALL F50B1
0523 11 851F 896    LTI M177
0524 11 851F 897    CALL F50B1
0525 11 851F 898    LTI M177
0526 11 851F 899    CALL F50B1
0527 11 851F 900    LTI M177
0528 11 851F 901    CALL F50B1
0529 11 851F 902    LTI M177
0530 11 851F 903    CALL F50B1
0531 11 851F 904    LTI M177
0532 11 851F 905    CALL F50B1
0533 11 851F 906    LTI M177
0534 11 851F 907    CALL F50B1
0535 11 851F 908    LTI M177
0536 11 851F 909    CALL F50B1
0537 11 851F 910    LTI M177
0538 11 851F 911    CALL F50B1
0539 11 851F 912    LTI M177
0540 11 851F 913    CALL F50B1
0541 11 851F 914    LTI M177
0542 11 851F 915    CALL F50B1
0543 11 851F 916    LTI M177
0544 11 851F 917    CALL F50B1
0545 11 851F 918    LTI M177
0546 11 851F 919    CALL F50B1
0547 11 851F 920    LTI M177
0548 11 851F 921    CALL F50B1
0549 11 851F 922    LTI M177
0550 11 851F 923    CALL F50B1

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144 : JUNIT FREQUENCY: AVERAGE
145 :
146 :
0522 11 3148 747 LXI D:MKXP:
0526 21 2190 748 LXI M:MKXP:
0534 00 7017 749 CALL PSUR2
0535 00 7117 750 CALL PSUR2
0537 00 7417 751 CALL PSUR2
0541 00 7517 752 CALL PSUR2
0544 11 3148 753 LXI D:MKXP:
0547 11 3148 754 LXI M:MKXP:
0549 00 7017 755 CALL PSUR2
0550 00 7117 756 CALL PSUR2
0551 00 7417 757 CALL PSUR2
0553 11 3148 758 LXI D:MKXP:
0556 21 2190 759 LXI M:MKXP:
0559 00 7017 760 CALL PSUR2
0562 00 7117 761 CALL PSUR2
0565 00 7417 762 CALL PSUR2
0568 00 7517 763 CALL PSUR2
0571 11 3148 764 LXI D:MKXP:
0574 21 2190 765 LXI M:MKXP:
0577 00 7017 766 CALL PSUR2
0580 00 7117 767 CALL PSUR2
0583 00 7417 768 CALL PSUR2
0586 00 7517 769 CALL PSUR2
0589 00 7617 770 CALL PSUR2
0592 11 3148 771 LXI D:MKXP:
0595 21 2190 772 LXI M:MKXP:
0598 00 7017 773 CALL PSUR2
0599 00 7117 774 CALL PSUR2
0602 00 7417 775 CALL PSUR2
0605 00 7517 776 CALL PSUR2
0608 00 7617 777 CALL PSUR2
0611 11 3148 778 LXI D:MKXP:
0614 21 2190 779 LXI M:MKXP:
0617 00 7017 780 CALL PSUR2
0620 00 7117 781 CALL PSUR2
0623 00 7417 782 CALL PSUR2
0626 00 7517 783 CALL PSUR2
0629 00 7617 784 CALL PSUR2
0632 11 3148 785 LXI D:MKXP:
0635 21 2190 786 LXI M:MKXP:
0638 00 7017 787 CALL PSUR2
0641 00 7117 788 CALL PSUR2
0644 00 7417 789 CALL PSUR2
0647 00 7517 790 CALL PSUR2
0650 00 7617 791 CALL PSUR2
0653 11 3148 792 LXI D:MKXP:
0656 21 2190 793 LXI M:MKXP:
0659 00 7017 794 CALL PSUR2
0662 00 7117 795 CALL PSUR2
0665 00 7417 796 CALL PSUR2
0668 00 7517 797 CALL PSUR2
0671 00 7617 798 CALL PSUR2
0674 11 3148 799 LXI D:MKXP:
0677 21 2190 800 LXI M:MKXP:
0680 00 7017 801 CALL PSUR2
0683 00 7117 802 CALL PSUR2
0686 00 7417 803 CALL PSUR2
0689 00 7517 804 CALL PSUR2
0692 00 7617 805 CALL PSUR2
0695 11 3148 806 LXI D:MKXP:
0698 21 2190 807 LXI M:MKXP:
0701 00 7017 808 CALL PSUR2
0704 00 7117 809 CALL PSUR2
0707 00 7417 810 CALL PSUR2
0710 00 7517 811 CALL PSUR2
0713 00 7617 812 CALL PSUR2
0716 11 3148 813 LXI D:MKXP:
0719 21 2190 814 LXI M:MKXP:
0722 00 7017 815 CALL PSUR2
0725 00 7117 816 CALL PSUR2
0728 00 7417 817 CALL PSUR2
0731 00 7517 818 CALL PSUR2
0734 00 7617 819 CALL PSUR2
0737 11 3148 820 LXI D:MKXP:
0740 21 2190 821 LXI M:MKXP:
0743 00 7017 822 CALL PSUR2
0746 00 7117 823 CALL PSUR2
0749 00 7417 824 CALL PSUR2
0752 00 7517 825 CALL PSUR2
0755 00 7617 826 CALL PSUR2
0758 11 3148 827 LXI D:MKXP:
0761 21 2190 828 LXI M:MKXP:
0764 00 7017 829 CALL PSUR2
0767 00 7117 830 CALL PSUR2
0770 00 7417 831 CALL PSUR2
0773 00 7517 832 CALL PSUR2
0776 00 7617 833 CALL PSUR2
0779 11 3148 834 LXI D:MKXP:
0782 21 2190 835 LXI M:MKXP:
0785 00 7017 836 CALL PSUR2
0788 00 7117 837 CALL PSUR2
0791 00 7417 838 CALL PSUR2
0794 00 7517 839 CALL PSUR2
0797 00 7617 840 CALL PSUR2
0800 11 3148 841 LXI D:MKXP:
0803 21 2190 842 LXI M:MKXP:
0806 00 7017 843 CALL PSUR2
0809 00 7117 844 CALL PSUR2
0812 00 7417 845 CALL PSUR2
0815 00 7517 846 CALL PSUR2
0818 00 7617 847 CALL PSUR2
0821 11 3148 848 LXI D:MKXP:
0824 21 2190 849 LXI M:MKXP:
0827 00 7017 850 CALL PSUR2
0830 00 7117 851 CALL PSUR2
0833 00 7417 852 CALL PSUR2
0836 00 7517 853 CALL PSUR2
0839 00 7617 854 CALL PSUR2
0842 11 3148 855 LXI D:MKXP:
0845 21 2190 856 LXI M:MKXP:
0848 00 7017 857 CALL PSUR2
0851 00 7117 858 CALL PSUR2
0854 00 7417 859 CALL PSUR2
0857 00 7517 860 CALL PSUR2
0860 00 7617 861 CALL PSUR2
0863 11 3148 862 LXI D:MKXP:
0866 21 2190 863 LXI M:MKXP:
0869 00 7017 864 CALL PSUR2
0872 00 7117 865 CALL PSUR2
0875 00 7417 866 CALL PSUR2
0878 00 7517 867 CALL PSUR2
0881 00 7617 868 CALL PSUR2
0884 11 3148 869 LXI D:MKXP:
0887 21 2190 870 LXI M:MKXP:
0890 00 7017 871 CALL PSUR2
0893 00 7117 872 CALL PSUR2
0896 00 7417 873 CALL PSUR2
0899 00 7517 874 CALL PSUR2
0902 00 7617 875 CALL PSUR2
0905 11 3148 876 LXI D:MKXP:
0908 21 2190 877 LXI M:MKXP:
0911 00 7017 878 CALL PSUR2
0914 00 7117 879 CALL PSUR2
0917 00 7417 880 CALL PSUR2
0920 00 7517 881 CALL PSUR2
0923 00 7617 882 CALL PSUR2
0926 11 3148 883 LXI D:MKXP:
0929 21 2190 884 LXI M:MKXP:
0932 00 7017 885 CALL PSUR2
0935 00 7117 886 CALL PSUR2
0938 00 7417 887 CALL PSUR2
0941 00 7517 888 CALL PSUR2
0944 00 7617 889 CALL PSUR2
0947 11 3148 890 LXI D:MKXP:
0950 21 2190 891 LXI M:MKXP:
0953 00 7017 892 CALL PSUR2
0956 00 7117 893 CALL PSUR2
0959 00 7417 894 CALL PSUR2
0962 00 7517 895 CALL PSUR2
0965 00 7617 896 CALL PSUR2
0968 11 3148 897 LXI D:MKXP:
0971 21 2190 898 LXI M:MKXP:
0974 00 7017 899 CALL PSUR2
0977 00 7117 900 CALL PSUR2
0980 00 7417 901 CALL PSUR2
0983 00 7517 902 CALL PSUR2
0986 00 7617 903 CALL PSUR2
0989 11 3148 904 LXI D:MKXP:
0992 21 2190 905 LXI M:MKXP:
0995 00 7017 906 CALL PSUR2
0998 00 7117 907 CALL PSUR2
1001 00 7417 908 CALL PSUR2
1004 00 7517 909 CALL PSUR2
1007 00 7617 910 CALL PSUR2
1010 11 3148 911 LXI D:MKXP:
1013 21 2190 912 LXI M:MKXP:
1016 00 7017 913 CALL PSUR2
1019 00 7117 914 CALL PSUR2
1022 00 7417 915 CALL PSUR2
1025 00 7517 916 CALL PSUR2
1028 00 7617 917 CALL PSUR2
1031 11 3148 918 LXI D:MKXP:
1034 21 2190 919 LXI M:MKXP:
1037 00 7017 920 CALL PSUR2
1040 00 7117 921 CALL PSUR2
1043 00 7417 922 CALL PSUR2
1046 00 7517 923 CALL PSUR2
1049 00 7617 924 CALL PSUR2
1052 11 3148 925 LXI D:MKXP:
1055 21 2190 926 LXI M:MKXP:
1058 00 7017 927 CALL PSUR2
1061 00 7117 928 CALL PSUR2
1064 00 7417 929 CALL PSUR2
1067 00 7517 930 CALL PSUR2
1070 00 7617 931 CALL PSUR2
1073 11 3148 932 LXI D:MKXP:
1076 21 2190 933 LXI M:MKXP:
1079 00 7017 934 CALL PSUR2
1082 00 7117 935 CALL PSUR2
1085 00 7417 936 CALL PSUR2
1088 00 7517 937 CALL PSUR2
1091 00 7617 938 CALL PSUR2
1094 11 3148 939 LXI D:MKXP:
1097 21 2190 940 LXI M:MKXP:
1100 00 7017 941 CALL PSUR2
1103 00 7117 942 CALL PSUR2
1106 00 7417 943 CALL PSUR2
1109 00 7517 944 CALL PSUR2
1112 00 7617 945 CALL PSUR2
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1118 21 2190 947 LXI M:MKXP:
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1124 00 7117 949 CALL PSUR2
1127 00 7417 950 CALL PSUR2
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1136 11 3148 953 LXI D:MKXP:
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1607 00 7117 1110 CALL PSUR2
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1628 00 7117 1117 CALL PSUR2
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1847 00 7617 1190 CALL PSUR2
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2132 00 7117 1285 CALL PSUR2
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2138 00 7517 1287 CALL PSUR2
2141 00 7617 1288 CALL PSUR2
2144 11 3148 1289 LXI D:MKXP:
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2162 00 7617 
```


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0066	21	0000	066	CALL	0000	U6
0067	21	0000	067	CALL	0000	U6
0068	21	0000	068	CALL	0000	U6
0069	21	0000	069	CALL	0000	U6
0070	21	0000	070	CALL	0000	U6
0071	21	0000	071	CALL	0000	U6
0072	21	0000	072	CALL	0000	U6
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0118	21	0000	118	CALL	0000	U6
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0121	21	0000	121	CALL	0000	U6
0122	21	0000	122	CALL	0000	U6
0123	21	0000	123	CALL	0000	U6

[illegible]

APPENDIX C

SYNTHESIZER DRAWINGS AND PARTS LISTS

C.2

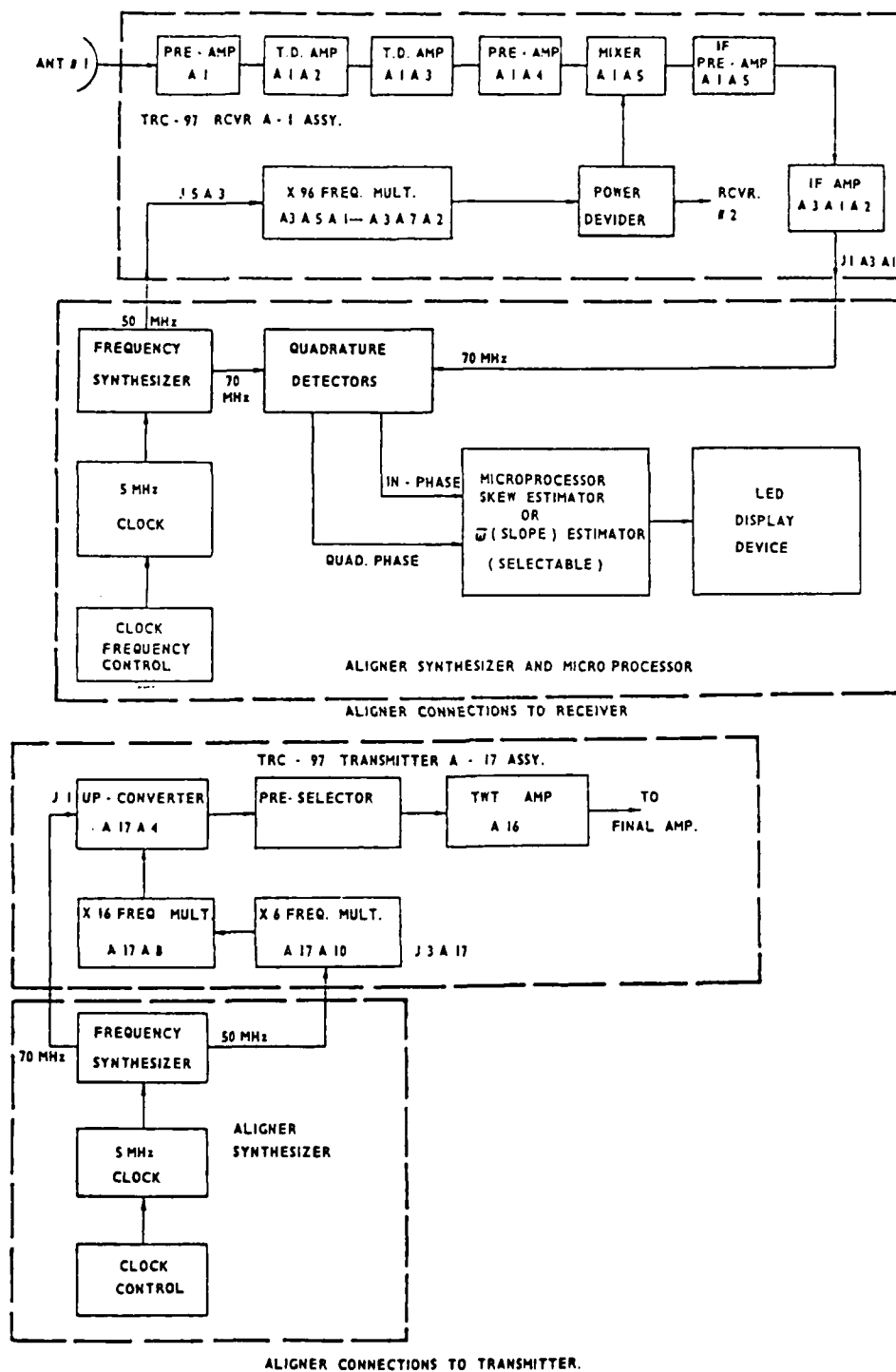


Figure C.1 INTERCONNECTIONS OF ALIGNER AND TRC-97

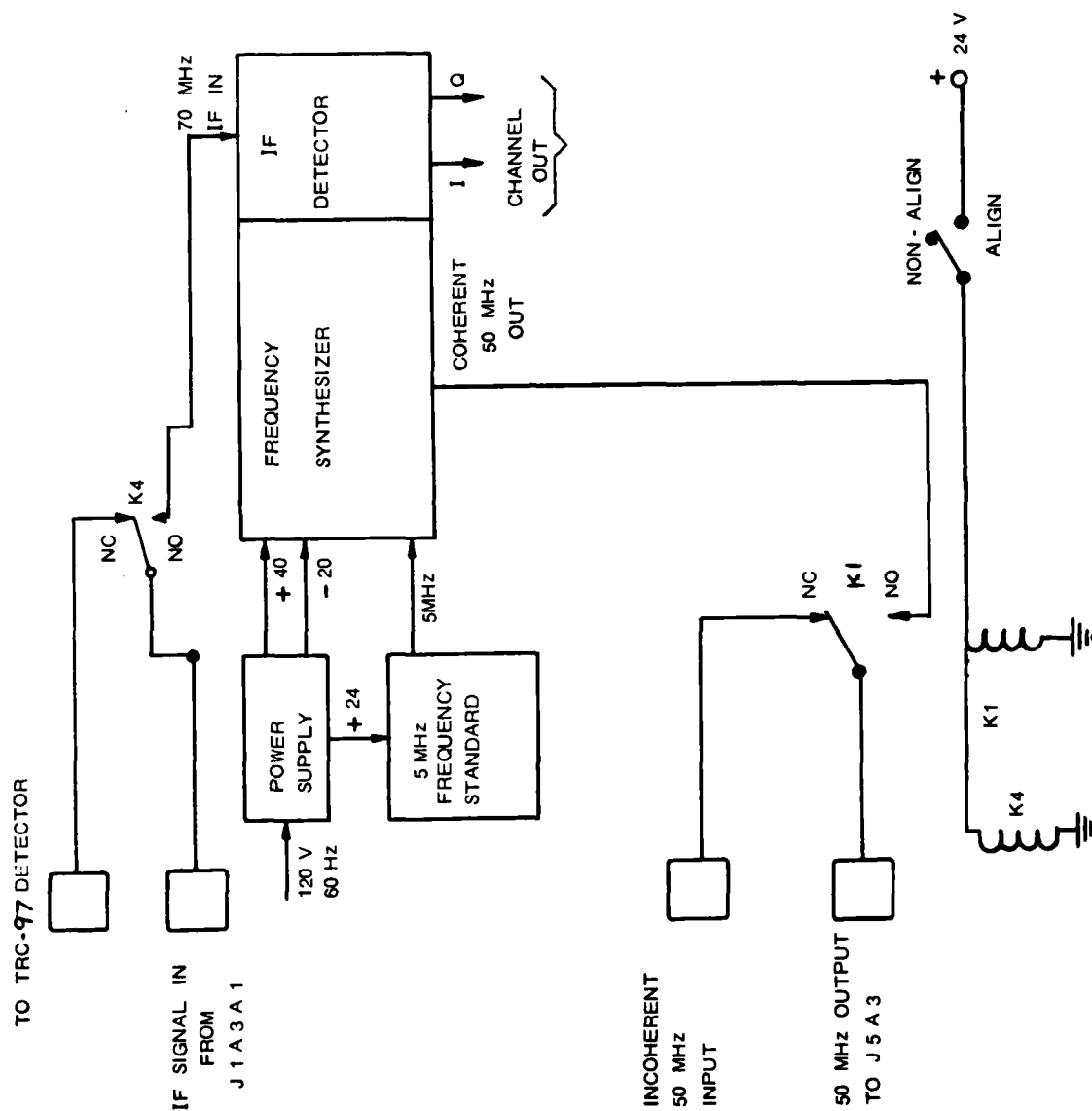


Fig. C.2a SIGNAL LEAD RELAY SWITCHING FOR TRC-97 RECEIVER APPLICATION

C.4

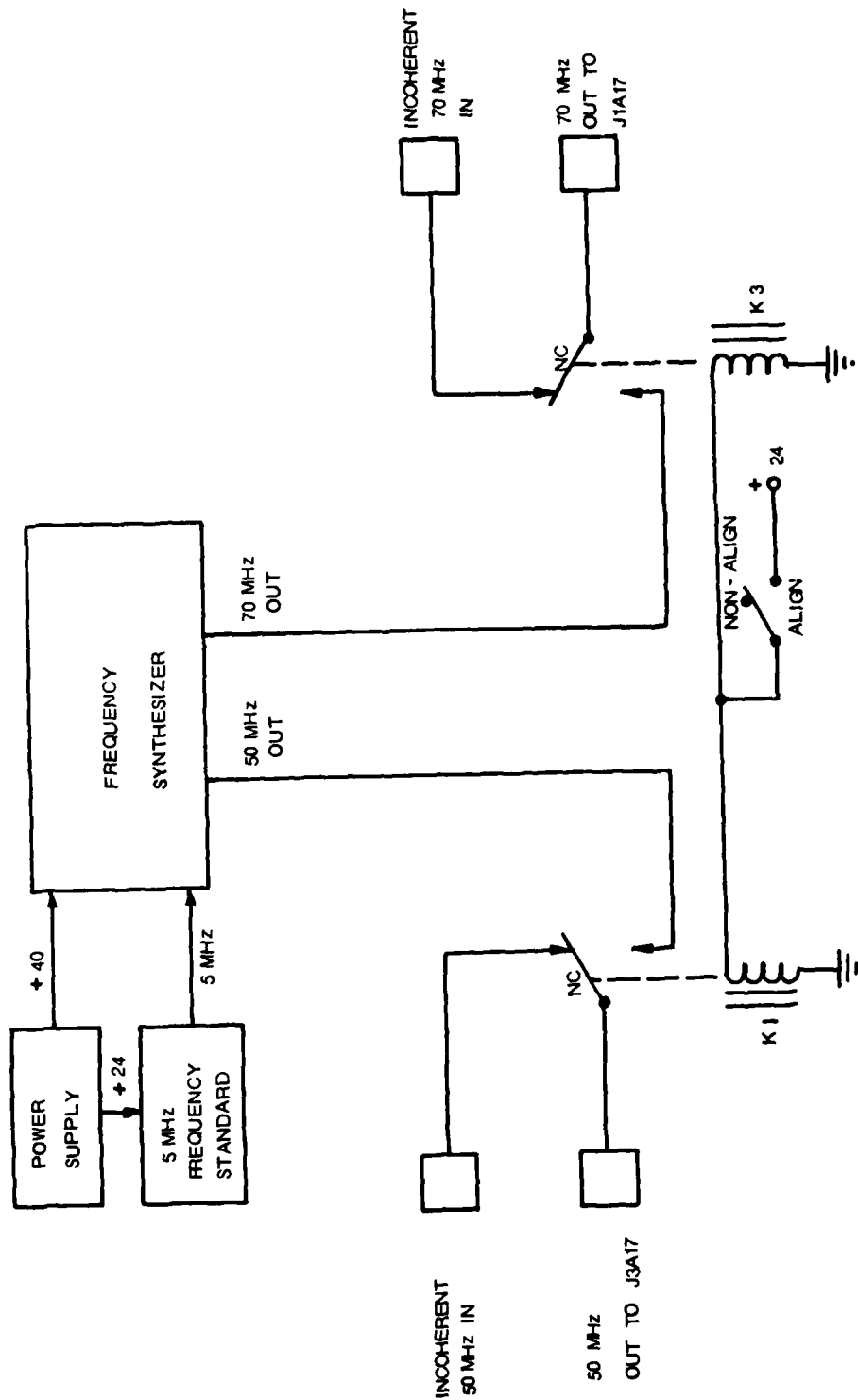


Fig.C2b SIGNAL LEAD RELAY SWITCHING FOR TRC-97 TRANSMITTER APPLICATION

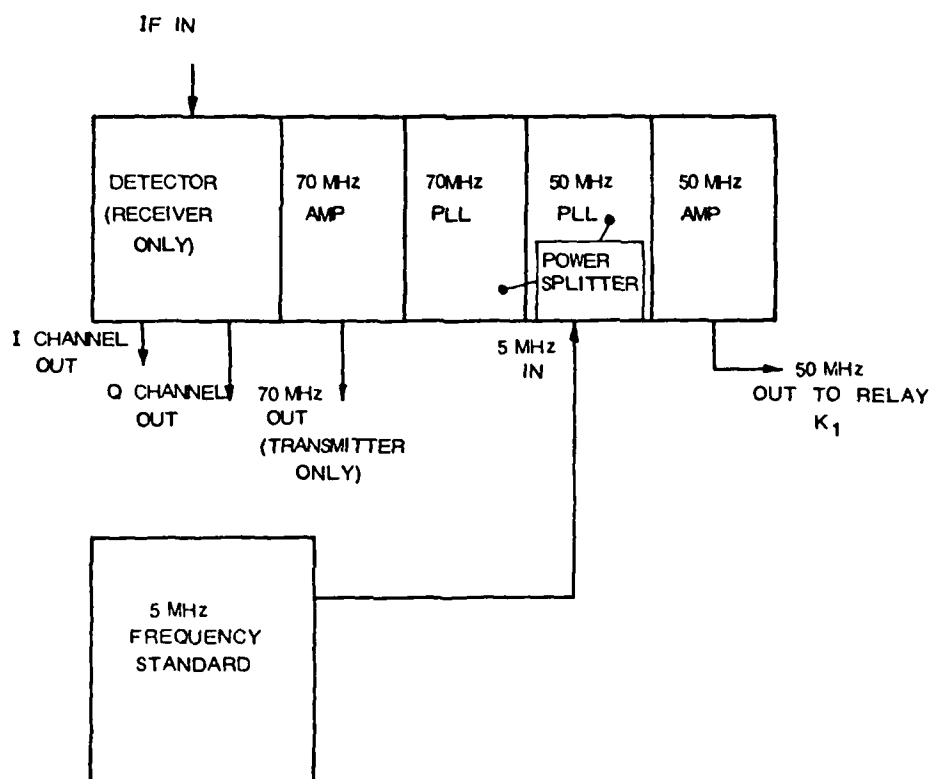


Figure C.3. RECEIVER FREQUENCY SYNTHESIZER BASIC DESIGN

C.6

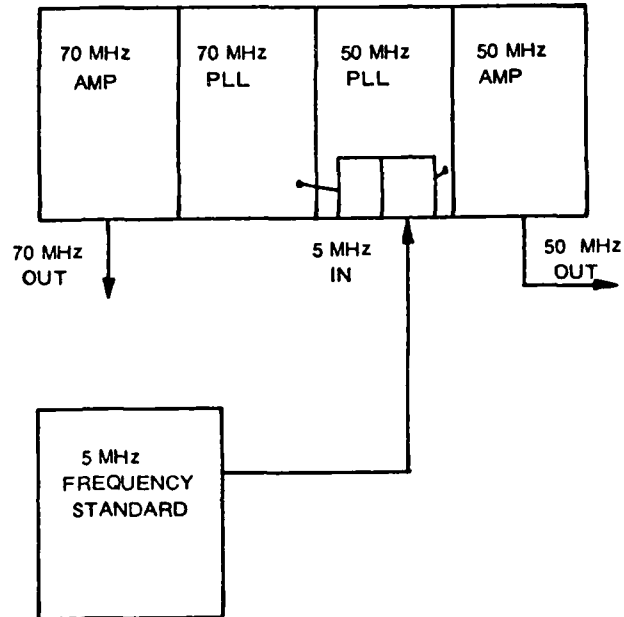


Figure C.4. TRANSMITTER FREQUENCY SYNTHESIZER BASIC DESIGN

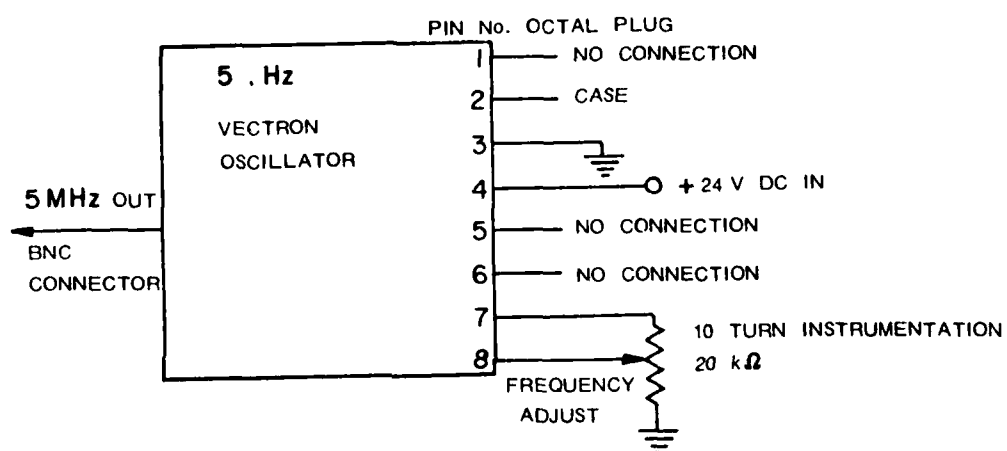


Figure C.5. 5 MHz FREQUENCY STANDARD CONNECTION
DIAGRAM

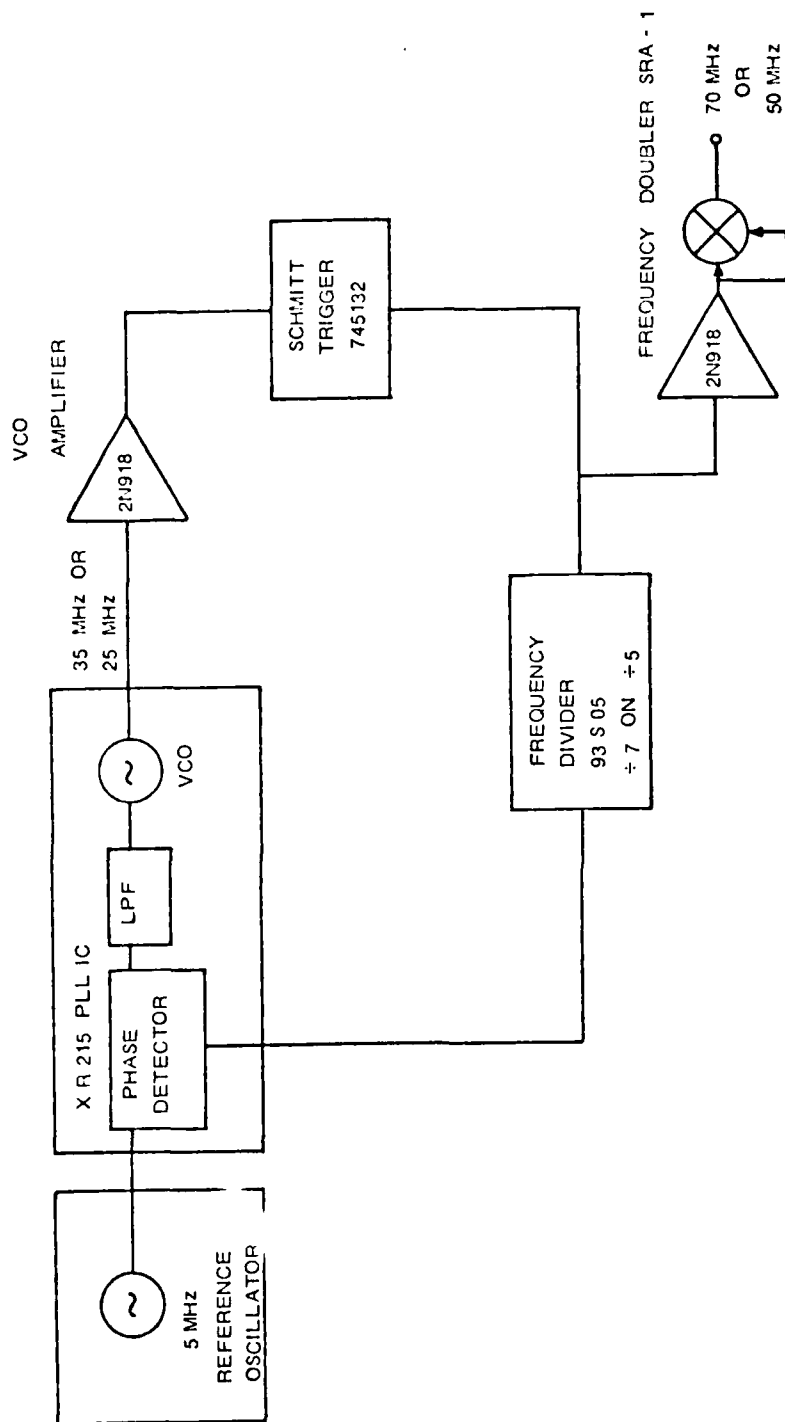


Figure C.6.

BASIC PHASED LOCKED LOOP DESIGN

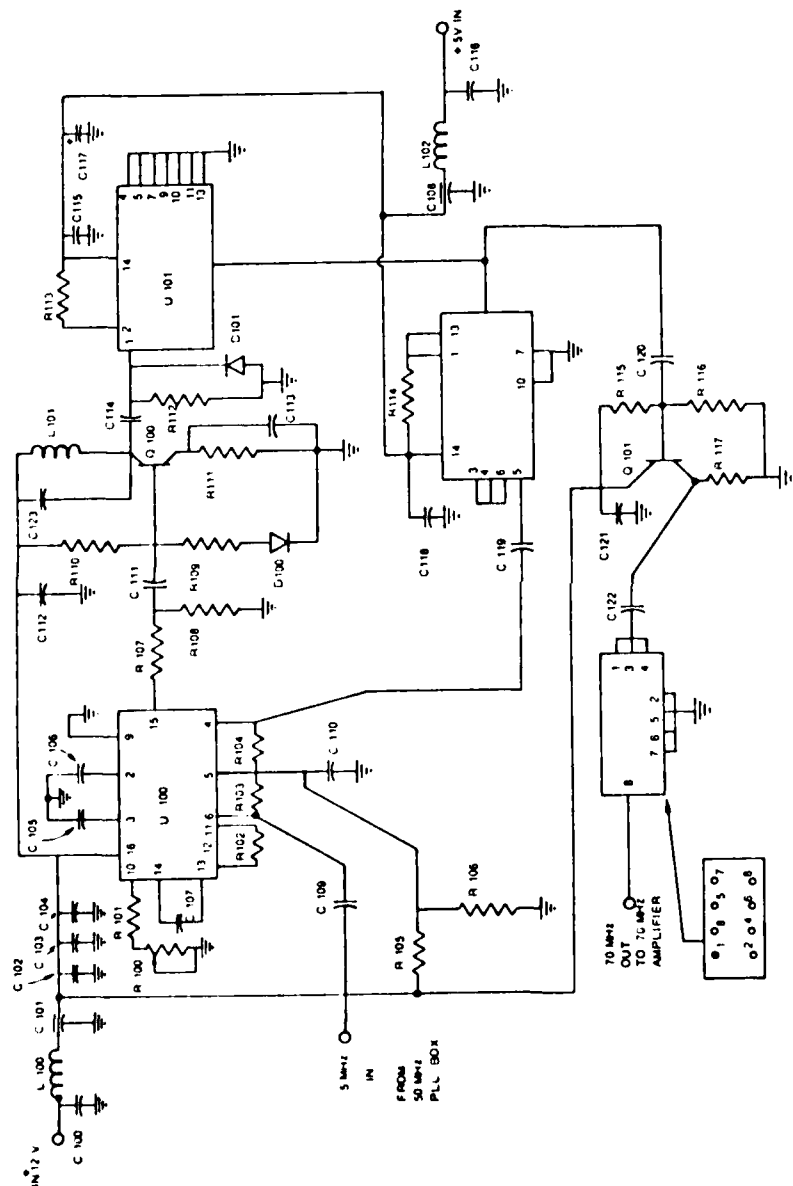


Figure C.7. 70 MHz PHASE LOCKED LOOP

Figure C.8. 50 MHz PHASE LOCKED LOOP

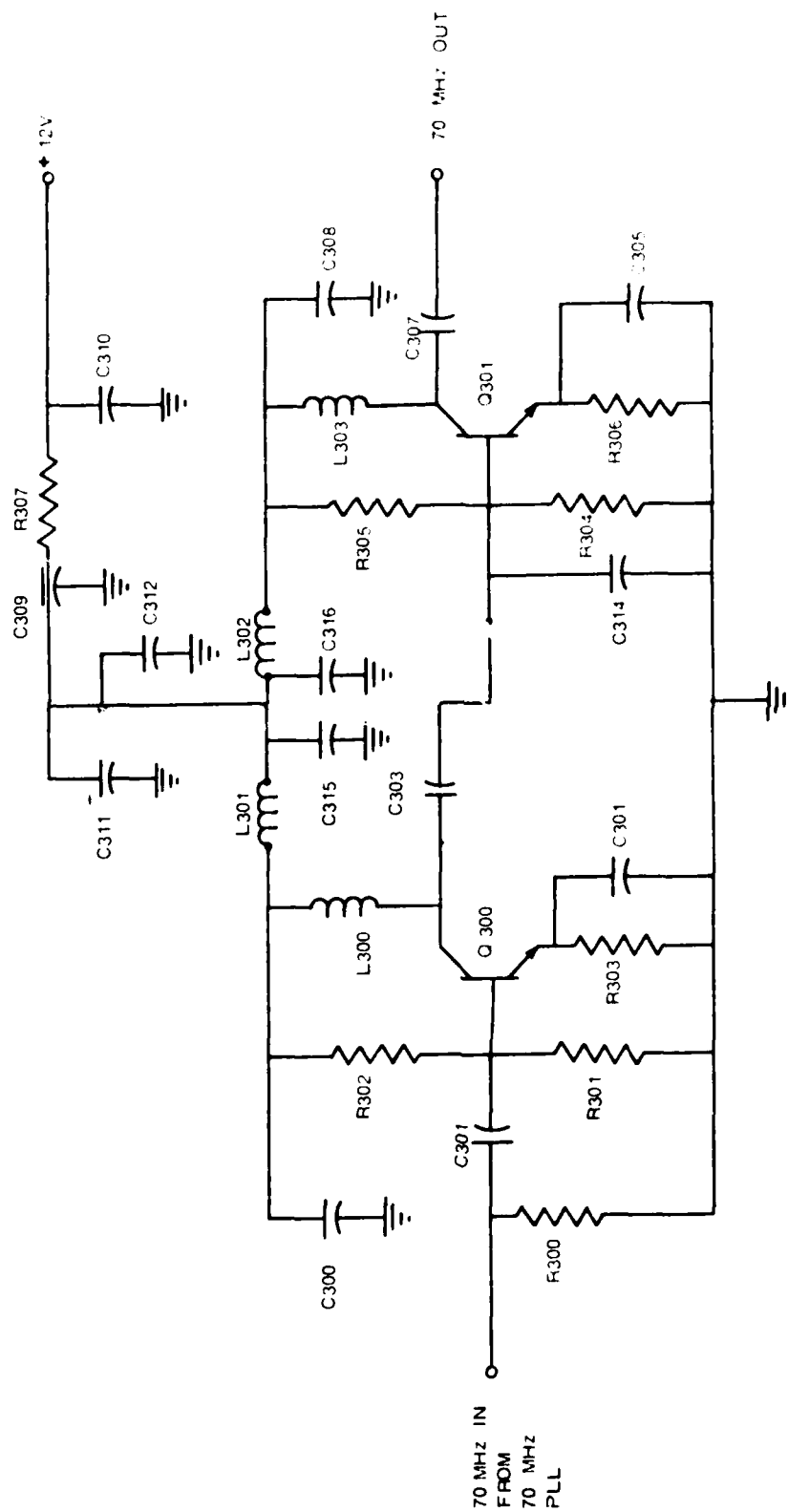


Figure C.9. 70 MHz AMPLIFIER

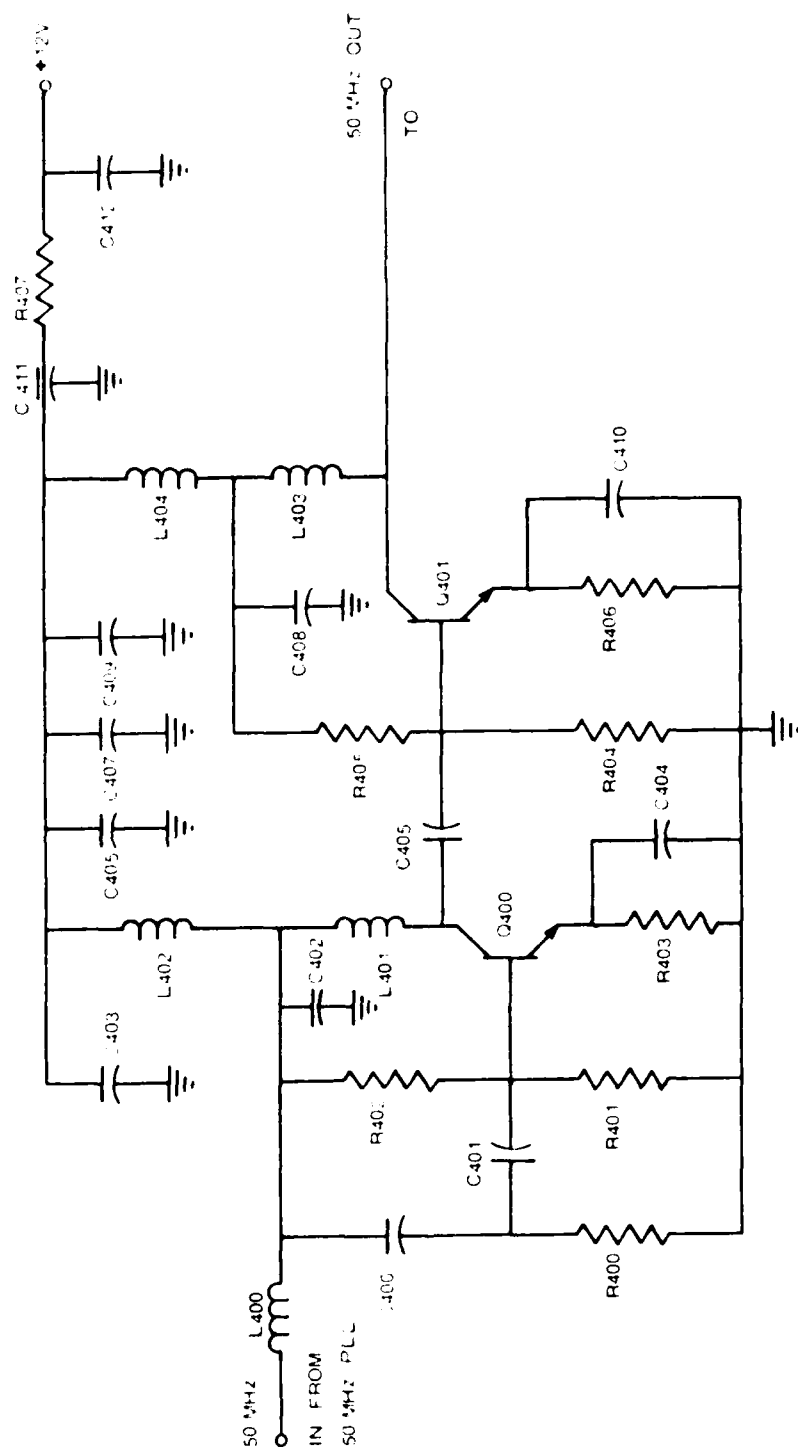


Figure C.10. 50 MHz AMPLIFIER

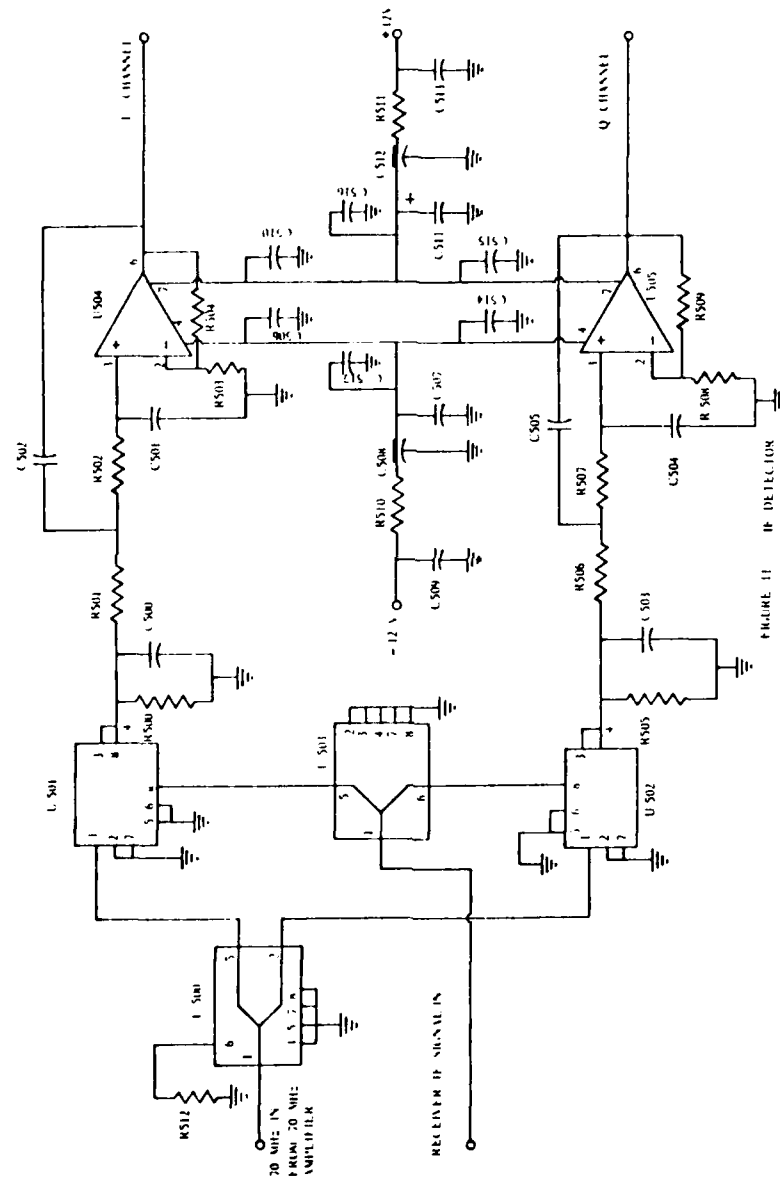


Figure C.11a. DETECTOR AND LOW-PASS FILTER

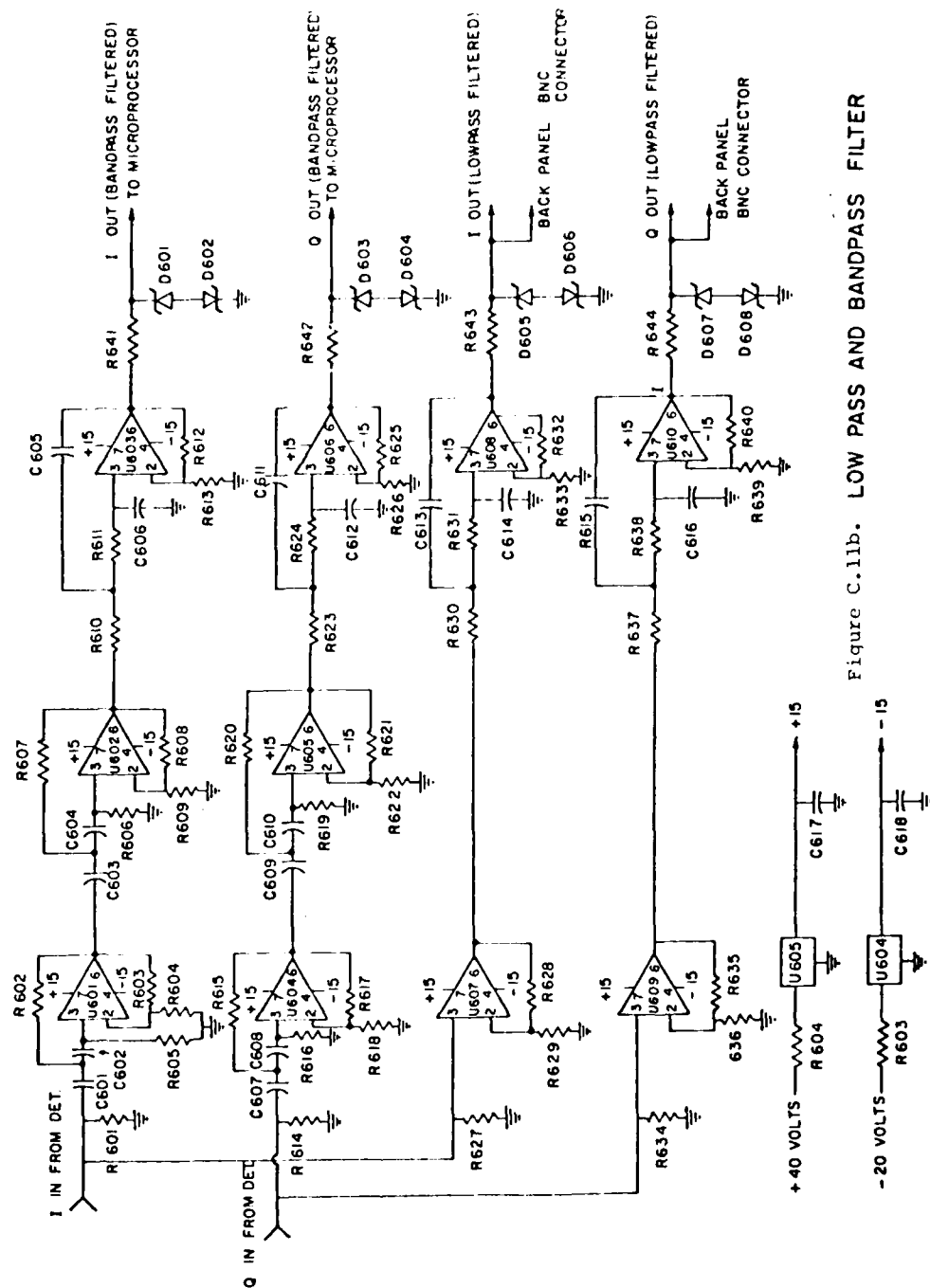


Figure C.11b. LOW PASS AND BANDPASS FILTER

C.16

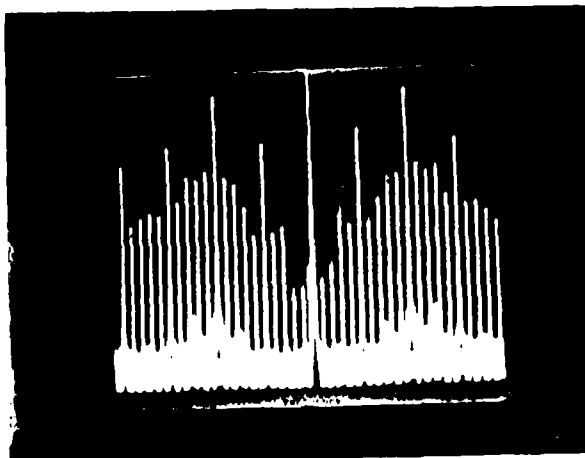


Figure C.13. 50 MHz Amplifier Output Spectrum. Full Scale = 20 dbm;
Vertical Scale: 10 db/Div; Horizontal Scale: 10 MHz/Div;
Center Freq. = 0 MHz.

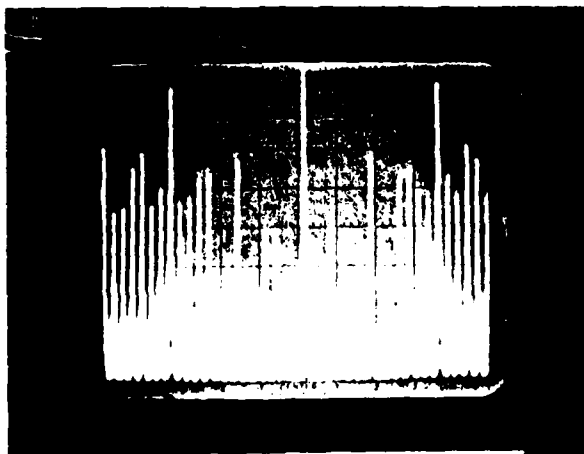


Figure C.14. 70 MHz Amplifier Output Spectrum. Full Scale = 20 dbm;
Vertical Scale: 10 db/Div; Horizontal Scale: 10 MHz/Div;
Center Freq. = 0 MHz.

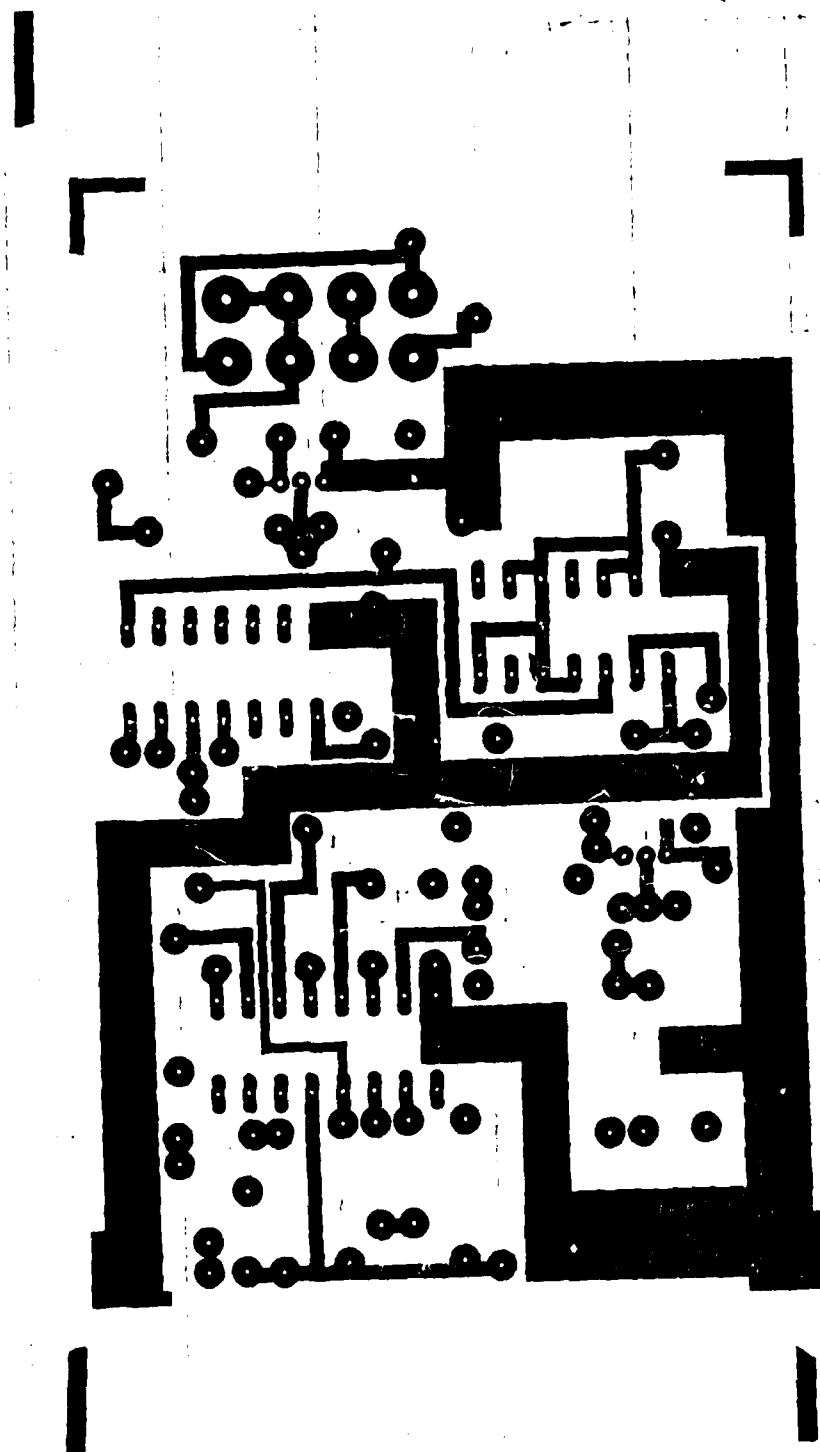


Figure C.17. 74100 10-Input NAND

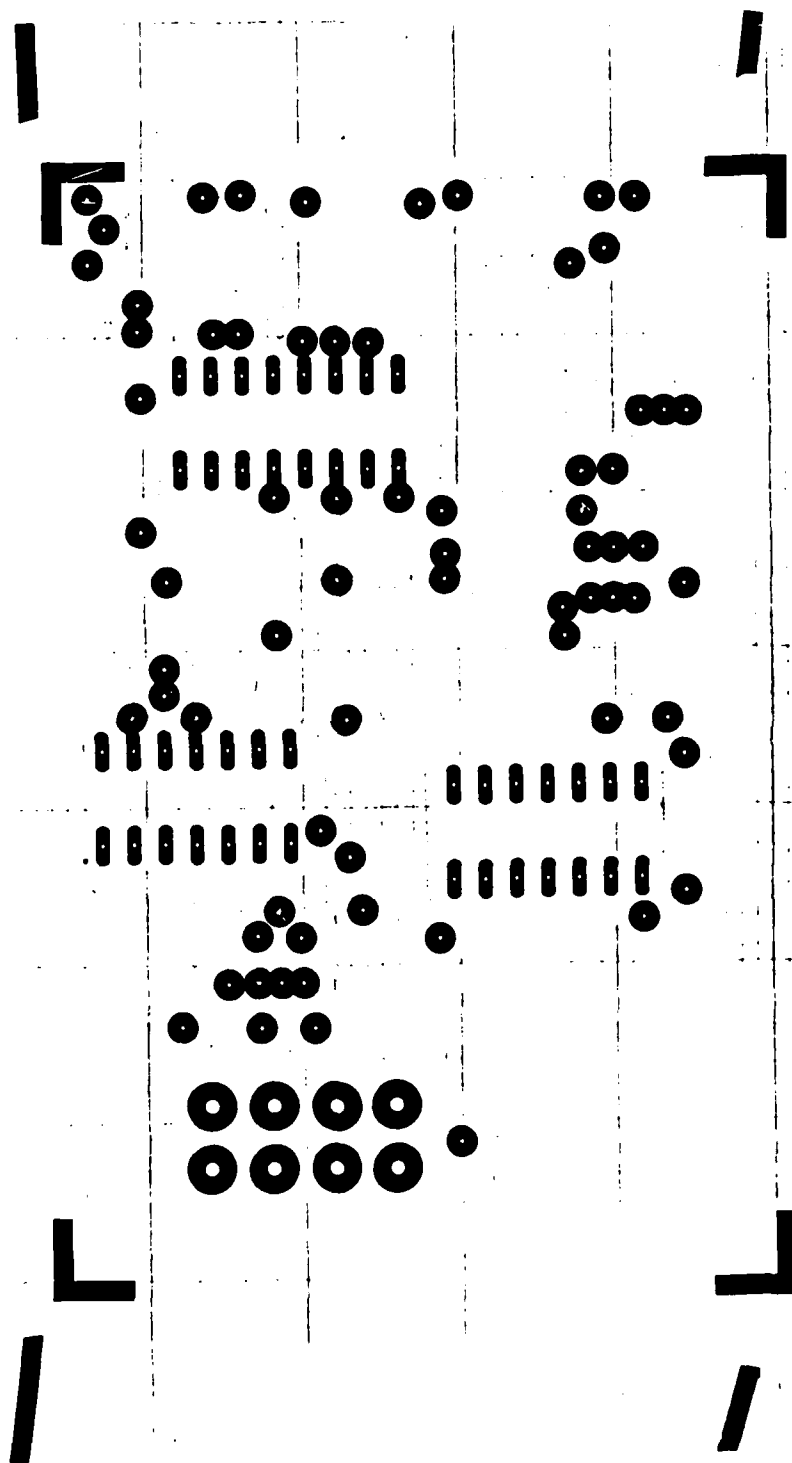


Figure C.15b. PHASE LOCK LOOP GROUND PLANE

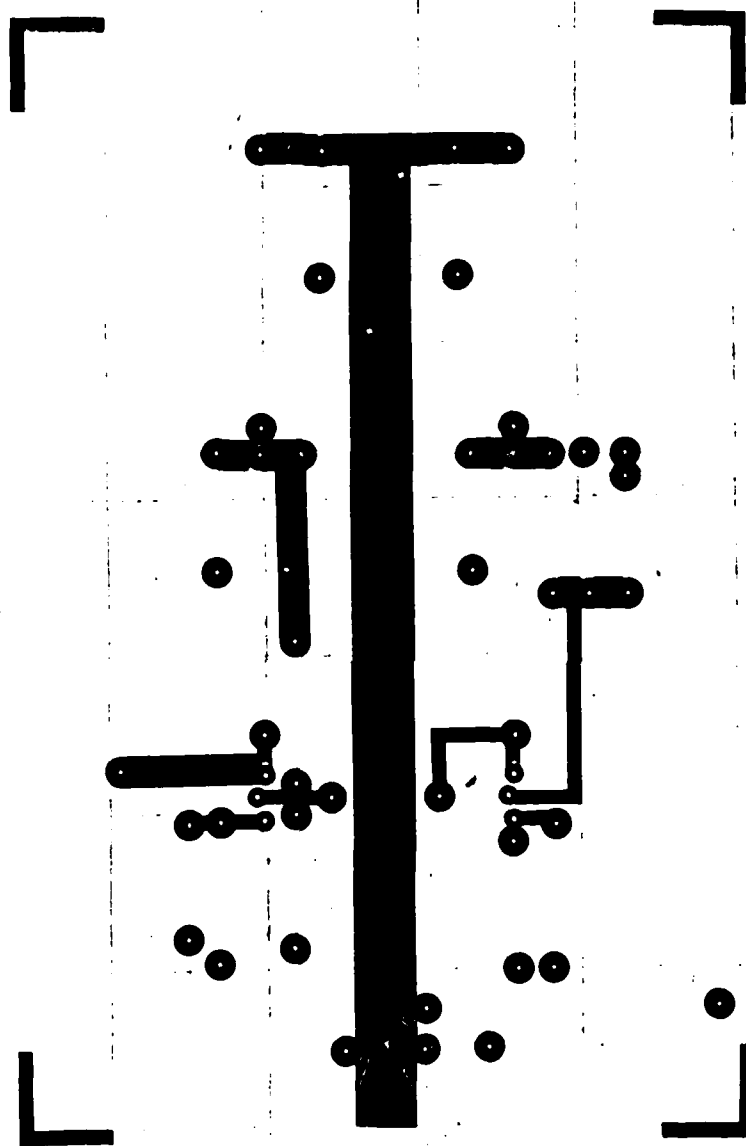


Figure C.16a. 50 MHz AMPLIFIER

C.20

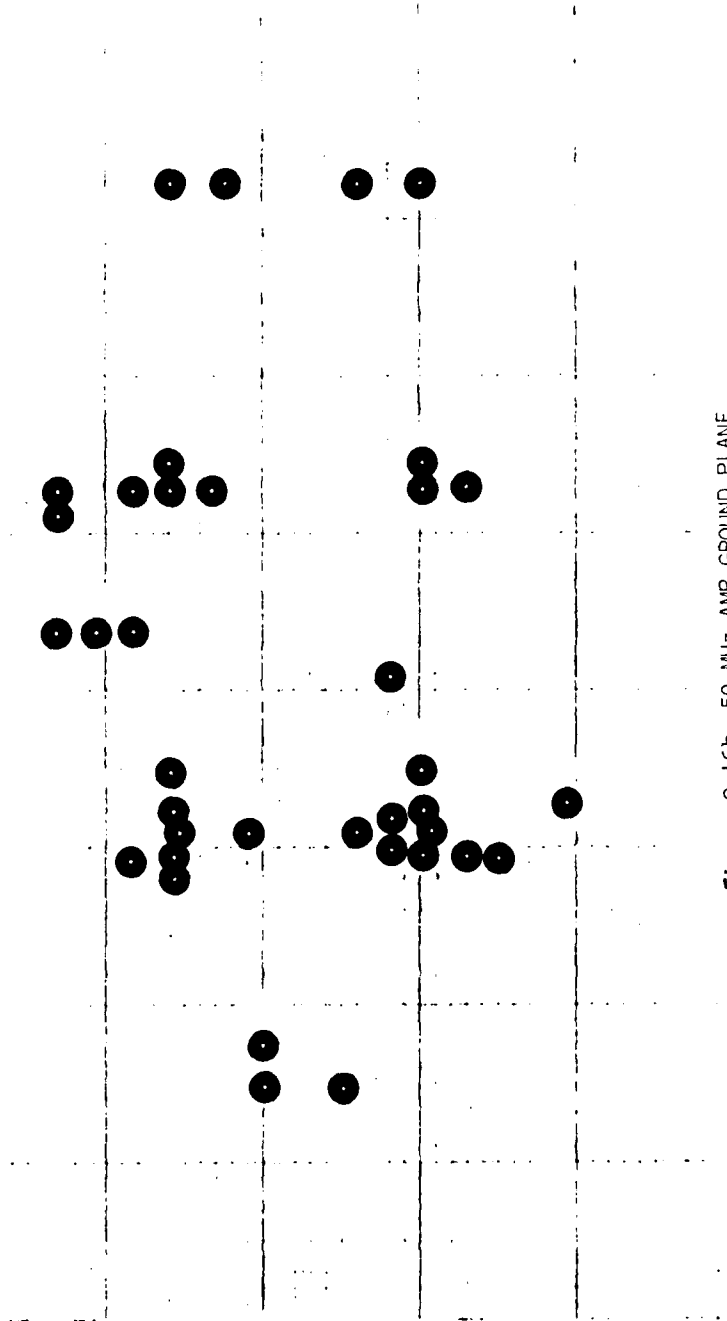


Figure C.16b. 50 MHz AMP GROUND PLANE

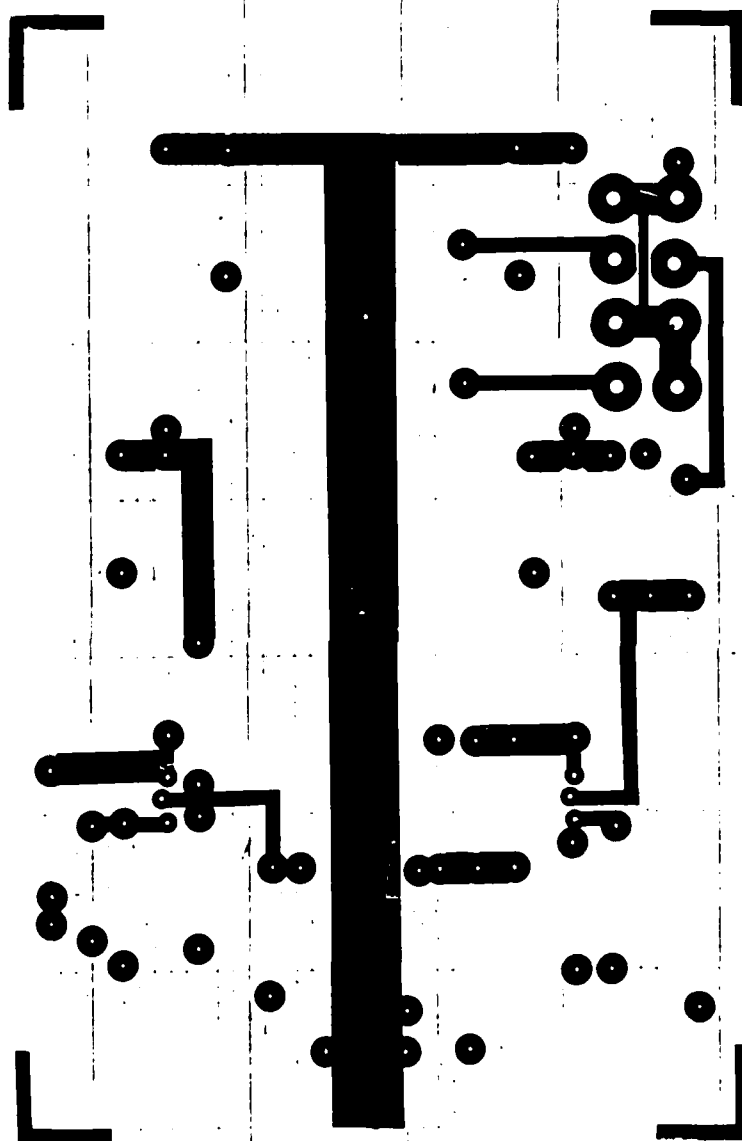


Figure C.17a. 70 MHz AMPLIFIER

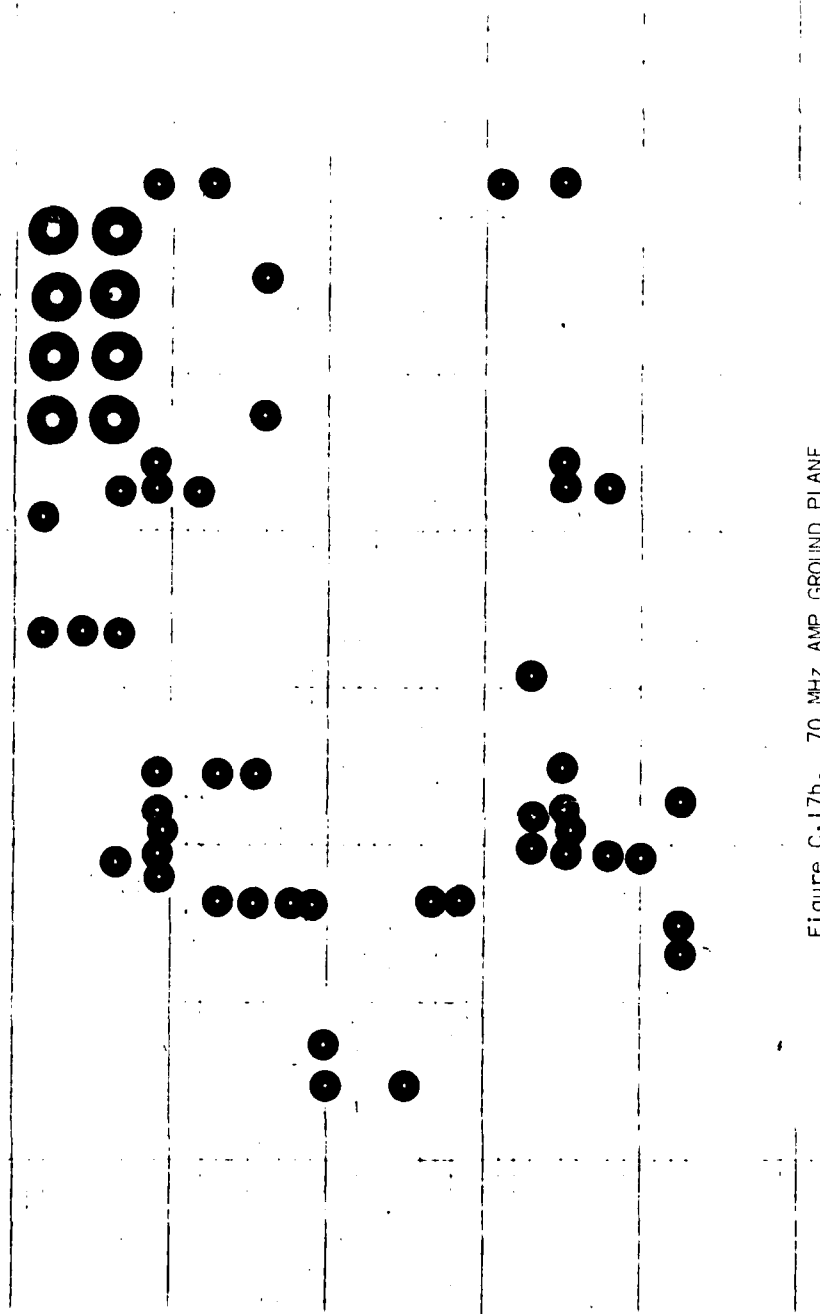


Figure C.17h. 70 MHz AMP GROUND PLANE

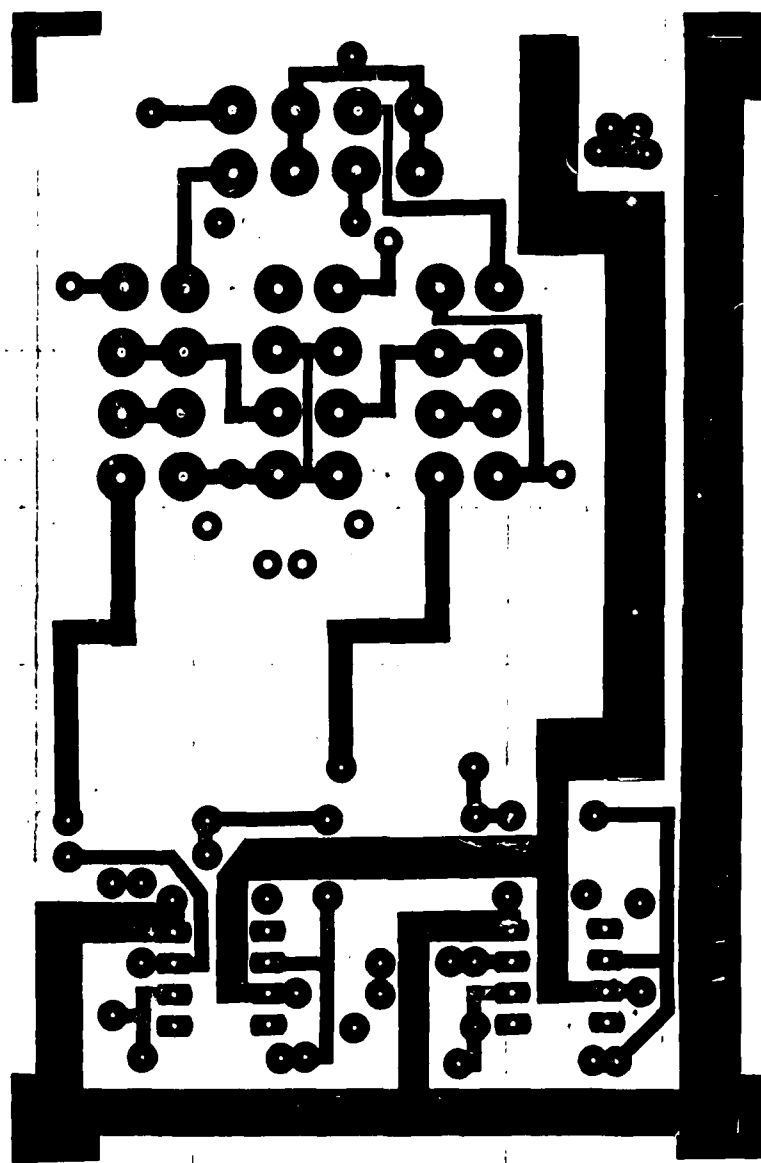


Figure C-13a. DETECTOR AND LOW-PASS FILTER

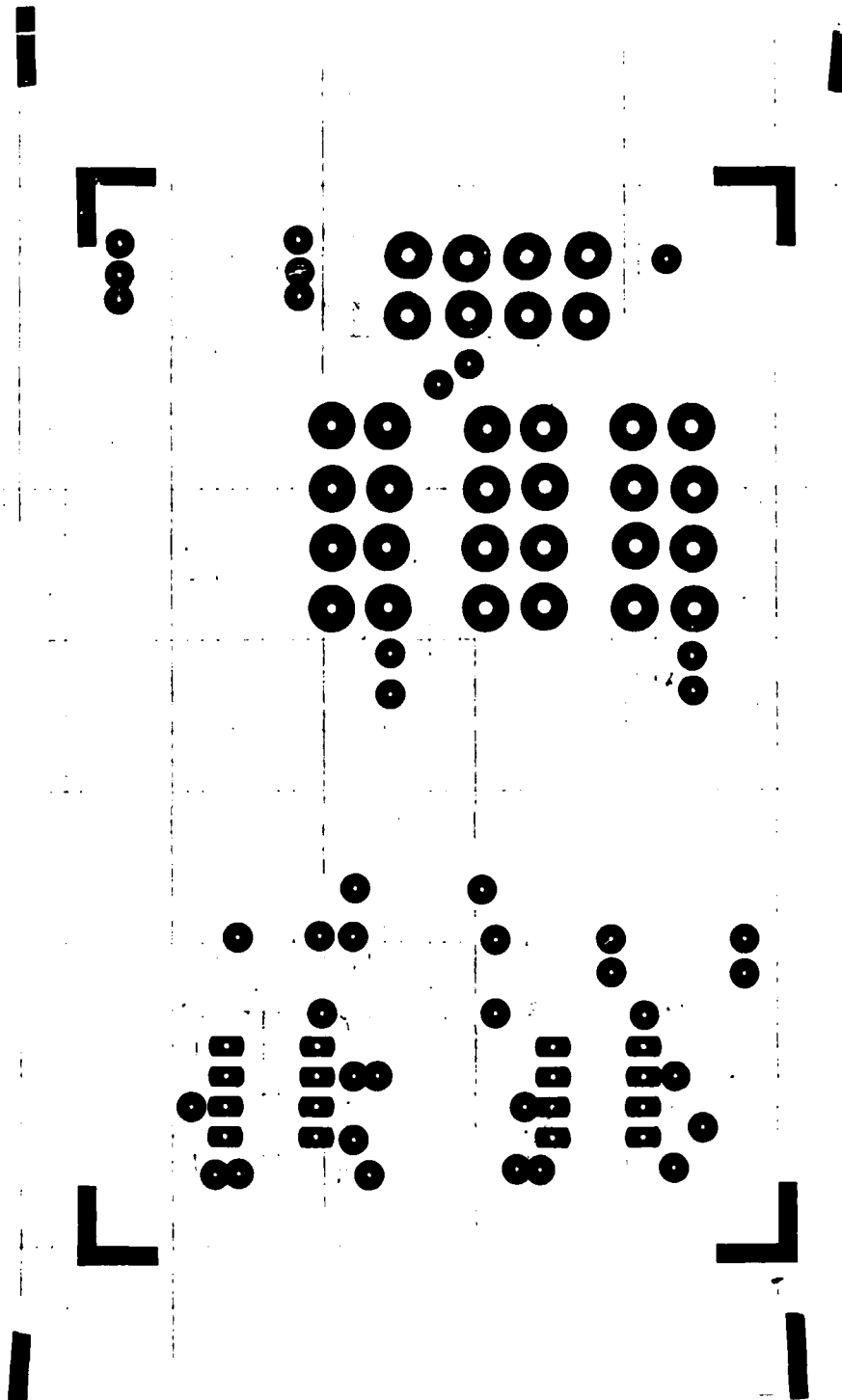
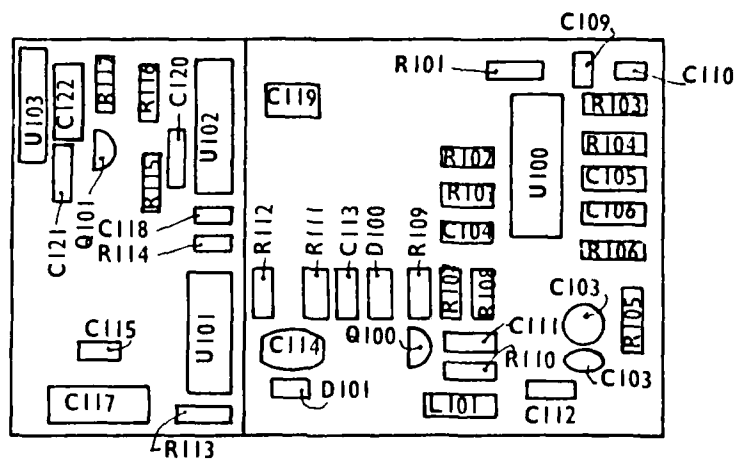


Figure C.18b. DETECTOR AND LOW-PASS FILTER GROUND PLANE



70 MHz PLL

C100, 101, 108, 116
L100, L102, R100 } ON WALL OF BOX

Figure C-19. 70 MHz Phase Lock Loop. The 50 MHz PLL is identical except for part numbers in the two-hundreds.

C. 20

70 MHz AMP

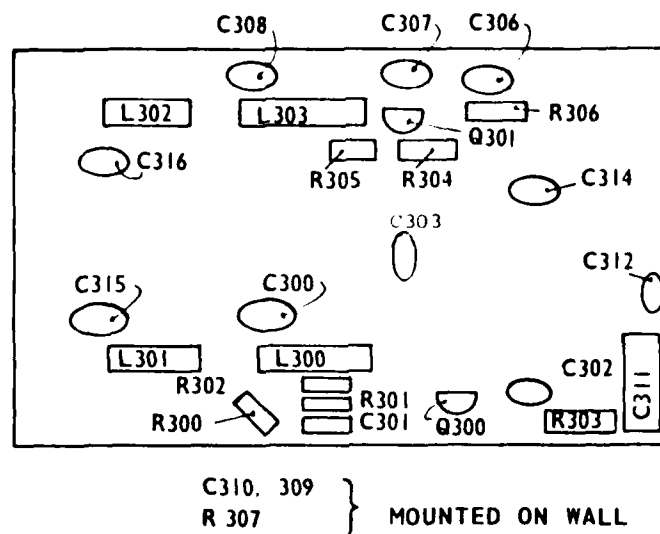
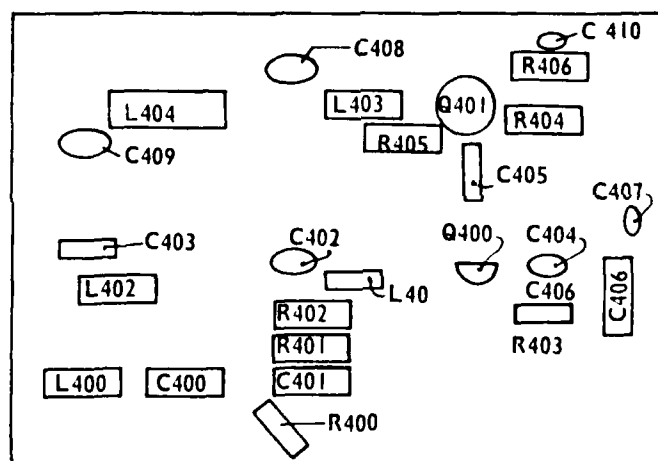


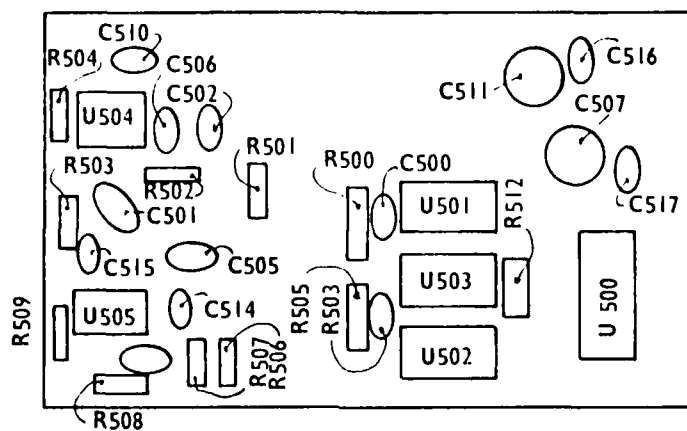
Figure C-20. 70 MHz amplifier.



50 MHz AMP

C411, R407, C4 ON BOX WALL

Figure C-21. 50 MHz amplifier.



C513, 512, 508, R510, R511
MOUNTED ON WALL OF BOX

Figure C-22. Detector and Low-Pass Filter.

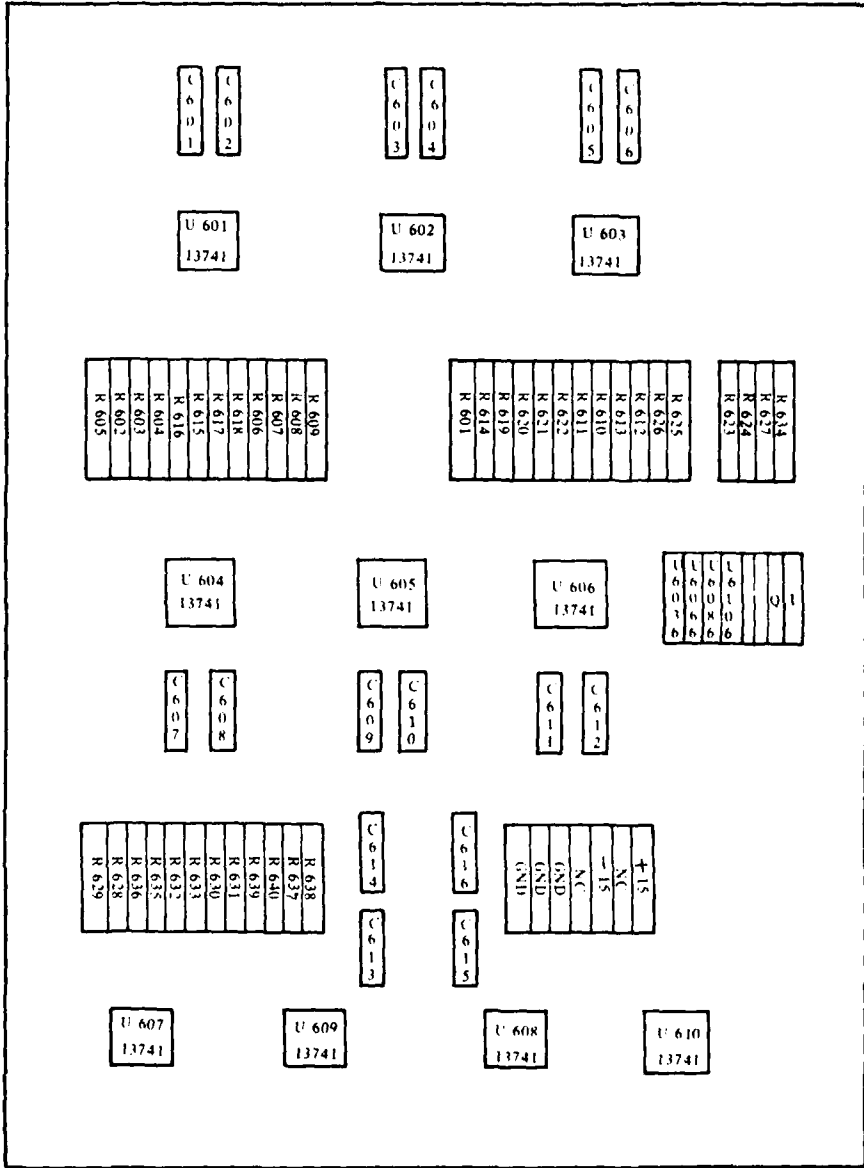


Figure C-23. Parts layout for low-pass and band-pass filter. Zener diodes are not shown.

Parts List 70 MHz PLL

Capacitors:

C100 1200 pF
C101 1200 pF Feedthrough
C102 10 microFarad, 25 Volt, Electrolytic
C103 0.01 microFarad
C104 330 pF
C105 120 pF
C106 120 pF
C107 selected
C108 1200 Feedthrough
C109 0.01 microFarad
C110 330 pF
C111 330 pF
C112 330 pF
C113 330 pF
C114 20 pF
C115 330 pF
C116 1200 pF
C117 47 microFarad, 25 Volt, Electrolytic
C118 330 pF
C119 330 pF
C120 330 pF
C121 330 pF
C122 330 pF
C123 4.7 pF

Transistors:

Q100 2N918
Q101 2N3866

Resistors: All $\frac{1}{4}$ Watt, 5%

R100 500 Ohm, 20 turn potentiometer
R101 470 Ohm
R102 2k Ohm
R103 2k Ohm
R104 2k Ohm
R105 5.1k Ohm
R106 5.1k Ohm
R107 510 Ohm
R108 20k Ohm
R109 33k Ohm
R110 18k Ohm
R111 2.2k Ohm
R112 2.7k Ohm
R113 1k Ohm
R114 1k Ohm
R115 12k Ohm
R116 47k Ohm
R117 330 Ohm

Diodes:

D100 1N4148
D101 1N4148

C.30

Inductors:

L100 4.7 microHenry
L101 4.7 microHenry
L102 4.7 microHenry

Integrated Circuits:

U100 XR215
U101 74S132
U102 93S05
U103 Mini Circuits Laboratory SRA-1 balanced mixer

Parts List 50 MHz PLL

Capacitors:

C200 1200 pF
C201 1200 pF Feedthrough
C202 10 microFarad, 25 Volt, Electrolytic
C203 0.01 microFarad
C204 330 pF
C205 120 pF
C206 120 pF
C207 selected
C208 1200 pF Feedthrough
C209 0.01 microFarad
C210 330 pF
C211 330 pF
C212 330 pF
C213 330 pF
C214 43 pF
C215 330pF
C216 1200 pF
C217 47 microfarad, 25 Volt, Electrolytic
C218 330 pF
C219 330 pF
C220 330 pF
C221 330 pF
C222 330 pF

Transistors:

Q200 2N918
Q201 2N918

Resistors: All $\frac{1}{4}$ Watt, 5%

R200 500 Ohm, 20 turn potentiometer
R201 470 Ohm
R202 2k Ohm
R203 2k Ohm
R204 2k Ohm
R205 5.1k Ohm
R206 5.1k Ohm
R207 510 Ohm
R208 20k Ohm
R209 33k Ohm
R210 18k Ohm
R211 2.2k Ohm
R212 2.7k Ohm
R213 1k Ohm
R214 1k Ohm
R215 12k Ohm
R216 47k Ohm
R217 330k Ohm

Diodes:

D200 1N4148
D201 1N4148

C.32

Inductors:

L200 4.7 microHenry
L201 4.7 microHenry
L202 4.7 microHenry

Integrated Circuits:

U200 XR215
U201 74S132
U202 93S05
U203 Mini Circuits Laboratory SRA-1 balanced mixer
U204 Mini Circuits Laboratory PSC-2 power splitter

Parts List 70 MHz Amplifier

Capacitors:

C300 1000 pF
C301 62 pF
C302 1000 pF
C303 16 pF

C305 1000 pF
C307 18 pF
C308 1000 pF
C309 1200 pF Feedthrough
C310 1200 pF
C311 47 microFarad, 25 Volt, Electrolytic
C312 0.01 microFarad

C314 12 pF
C315 1000 pF
C316 1000 pF

Resistors: All $\frac{1}{4}$ Watt, 5%

R300 100 Ohm
R301 2.2k Ohm
R302 3.9k Ohm
R303 330 Ohm
R304 2.2k Ohm
R305 2.7k Ohm
R306 330 Ohm
R307 10 Ohm

Inductors:

L300 0.28 microHenry
L301 2.2 microHenry
L302 2.2 microHenry
L303 0.28 microHenry

Transistors:

Q300 2N918
Q301 2N918

Parts List 50 MHz Amplifier

Capacitors:

C400 27 pF
C401 62 pF
C402 1000 pF
C403 1000 pF
C404 1000 pF
C405 27 pF
C406 47 microFarad, 25 Volt, Electrolytic
C407 0.01 microFarad
C408 1000 pF
C409 1000 pF
C410 1000 pF
C411 1200 pF Feedthrough
C412 1200 pF Feedthrough

Resistors: All $\frac{1}{4}$ Watt, 5%

R400 100 Ohm
R401 2.2k Ohm
R402 2.7k Ohm
R403 330 Ohm
R404 3.9k Ohm
R405 2.2k Ohm
R406 120 Ohm
R407 10 Ohm

Inductors:

L400 0.39 microHenry
L401 0.39 microHenry
L402 2.2 microHenry
L403 0.39 microHenry
L404 2.2 microHenry

Transistors:

Q400 2N918
Q401 2N3866

Parts List Detector

Capacitors:

C500 0.001 microFarad
 C501 0.1 microFarad
 C502 0.1 microFarad
 C503 0.001 microFarad
 C504 0.1 microFarad
 C505 0.1 microFarad
 C506 0.01 microFarad
 C507 100 microFarad, 15 Volt, Electrolytic
 C508 1200 pF Feedthrough
 C509 1200 pF
 C510 0.01 microFarad
 C511 100 microFarad
 C512 1200 pF Feedthrough
 C513 1200 pF
 C514 0.01 microFarad
 C515 0.01 microFarad
 C516 0.01 microFarad
 C517 0.01 microFarad

Resistors: All 1/4 Watt, 5%

R500 51 Ohm
 R501 820 Ω
 R502 12K Ω
 R503 10k Ohm
 R504 100k Ohm
 R505 51 Ohm
 R506 1K Ω
 R507 12K Ω
 R508 10k Ohm
 R509 100k Ohm
 R510 10 Ohm
 R511 10 Ohm
 R512 51 Ohm

Integrated Circuits:

U500	Mini Circuits Laboratory	PSCQ2	quadrature power splitter
U501	"	"	SRA-1 balanced mixer
U502	"	"	SRA-1 balanced mixer
U503	"	"	PSC2-1 power splitter
U504	LM741		
U505	LM741		

Parts List Low-Pass and Band-Pass Filter

1. Operational Amplifiers U601-U610 13741 (LM 741)
 U611 LM 7815 +15 volt regulator
 U612 LM 7915 -15 volt regulator
2. Capacitors C619, C620 100 microfarad, 56 volt DC
 Capacitors C601 - C618, 1 microfarad, 50 volt
3. Resistors \$641, 642, 150 Ω , 10 watt

All resistors are 1/4 watts

R601, R614, R627, R634 --- 313 k Ω

R602, R605, R606, R707 } --- 510 k Ω
 R615, R616, R619, R620 }

R603, R617 --- 12 k Ω

R628, R635 --- 15 k Ω

R604, R609, R613, R618 } --- 10 k Ω
 R622, R626, R629, R633 }
 R636, R639 }

R608, R621 --- 1.5 k Ω

R610, R611, R623, R624 } --- 6.8 k Ω
 R630, R631, R637, R638 }

R612, R625, R632, R640 --- 5.6 k Ω

Parts List Power Supply

Capacitors:

C700 1000 microFarad, 56 Volt, Electrolytic
•C701 1000 microFarad, 56 Volt, Electrolytic
C702 1200 pF Feedthrough

Resistors:

R700 100 Ohm, 10 Watt
R701 150 Ohm, 10 Watt
•R702 200 Ohm, 5 Watt

Rectifiers:

CR700 2 Ampere, 50 Volt PIV, Diode Bridge
•CR701 2 Ampere, 50 Volt PIV, Diode Bridge

Integrated Circuits:

U700 LM7824, 24 Volt Regulator
U701 LM7812, 12 Volt Regulator
U702 LM7805, 5 Volt Regulator
•U703 LM7912, -12 Volt Regulator

T700 Power Transformer, Primary Voltage = 117 VRMS, 60 Hz
Two secondaries, 28 VRMS, 1 Ampere each
eg. Signal Transformer Co. Inc. Model 56-1

*Not required for the transmitter power supply.

Interrelations of Synthesizer Drawings and Parts Lists

Module	Schematic	Printed Board	Parts Layout	Parts List
70 MHz PPL	Fig. C.7	Fig. C.15a,b	Fig. C.19	Page C29
50 MHz PPL	Fig. C.8	Fig. C.15a,b	Fig. C.19	Page C31
70 MHz Amp	Fig. C.9	Fig. C.17a,b	Fig. C.20	Page C33
50 MHz Amp	Fig. C.10	Fig. C.16a,b	Fig. C.21	Page C34
Detector-Filter	Fig. C.11a	Fig. C.18a,b	Fig. C.22	Page C35
LP-BP Filter	Fig. C.11b	Fig. not used	Fig. C.23	Page C36
Power Supply	Fig. C.12	Fig. not used		Page C37
50 MHz Spectrum	Fig. C.13			
70 MHz Spectrum	Fig. C.14			

Appendix D

Theory of Doppler Alignment Estimators
and
Tropo-Channel Simulation

Contents

Part I

"Doppler Spectra as a Function of Tropo-Link Parameters and
Accuracy Analysis of 'Slope' Estimators"
Interim Report for 12/11/77 to 1/11/78.....D.2

Part II

"Estimation of Quadrature Correlation Function for all Delays by
Analog and Polarity-Coincidence Methods"
Interim Report for 1/11/78 to 2/11/78.....D.26

Part III

"A New Complex Gaussian Troposcatter Channel Simulator"
Interim Report for 6/11/78 to 7/11/78.....D.49

Summary of Appendix D

Part I reviews the relationship between troposcatter link geometry, the meteorology of the scattering structure, and the Doppler spectrum imparted by the mean cross-path wind. It is shown that Doppler spectrum symmetry or its lack can be used to test for correct antenna alignment. Various estimators for measures of symmetry are then discussed including average Doppler shift and skew. The variance of these estimators and their sensitivity to link parameters are calculated and tabulated for several types of signal processing.

Part II is essentially devoted to the analysis of the variance of the cross-correlation or slope estimator when one or both quadrature component signals are hard clipped. The increase in variance of 6 db due to clipping was the main reason for using non-distorted signals in the final version of the aligner.

Part III is a description of a tropochannel simulator in which delay-line tap gains are Gaussian random variables with quadrature statistics that produce Doppler effects analogous to those found in true channels. Skewed spectra can be synthesized and thus the effects of aircraft intrusion can be modeled.

Doppler Spectra as a Function of Tropo-Link Parameters
and Accuracy Analysis of 'Slope' Estimators

Introduction

When a monochromatic carrier is transmitted, the received signal is normally spread about the carrier frequency because of atmospherically-induced Doppler shifts. In this chapter, the relative accuracy of the alignment scheme by 'slope method' is related to the Doppler spread function which is shown to depend on the antenna patterns and their position relative to the great-circle, the angular dependence of scattering, the cross path wind speed, and the link geometry. Next we show that the accuracy of the slope estimators is degraded by factors which increase the fade-rate such as beam-width or cross-path wind speed. Finally, we show that the accuracy is further degraded if 'hard-clipped' estimator is used instead of 'true' estimator of slope of quadrature correlation function at zero lag.

Review of Link Geometry: The paths taken by the transmitted and scattered signals are slightly bent toward the earth due to the average decrease in refractive index with height. Bean and Dutton (1966) have shown that under normal conditions the average refractivity variation near the earth's surface is linear, thereby permitting the use of an earth model whose radius is approximately $4/3$ that of the true earth radius. The equivalent earth model has several distinct practical advantages: 1) the refractive index surrounding the earth is uniform and constant, 2) the curved radio paths become straight lines (or ray-paths), and 3) the ray paths have the same length as their curved counterparts. Assuming this equivalent earth model, the Doppler spectrum is written in terms of angular dependence and

antenna function using the Doppler-shift model of Birkemeier (1968).

To each point in space there corresponds a ray path length $L(x,y,z)$ which is defined by a ray from the transmitter to the receiver passing through the given point. Surfaces of constant path or phase are approximately ellipsoids of revolution having the radio terminals as foci (Fig. 2). A signal obtained via any single moving point is shifted in frequency by an amount proportional to the rate at which the point cuts equiphase surfaces. Thus the 'single scatterer' Doppler-shift f is given by

$$f = -\frac{1}{\lambda} \frac{dL}{dt} = -\frac{1}{\lambda} \cdot v \cdot \nabla L \quad (1)$$

where λ is the wavelength and v is the velocity of the moving scatterer. Using the coordinate system of Fig. 1 with origin at the midpoint of the straight (chordal) line between the terminals:

$$L = [(d+x)^2 + y^2 + z^2]^{\frac{1}{2}} + [(d-x)^2 + y^2 + z^2]^{\frac{1}{2}} \quad (2)$$

Applying (1) and (2) yields, Birkemeier (1968),

$$f = -\frac{4}{\lambda L} \frac{ux(1-4d^2/L^2) + vy + wz}{1-16x^2d^2/L^4} \quad (3)$$

where (u,v,w) are the (x,y,z) components of v . The relative importance of the various terms in (3) depends upon the position of the scatterer and on the relative magnitude of the velocity components. The along path (u) contribution is negligible because the ellipsoids are longitudinally very 'flat'. The factor $(1 - \frac{4d^2}{L^2})$ is small, and equal to $4z_m^2/L^2 = 4h_m^2/d^2$ where z_m and h_m are the midpath heights of an ellipsoid above the chordal axis and spherical earth, respectively. This factor ranges from 10^{-4} to 10^{-2} from the bottom to the top of the common volume. Since at a given

point, the average horizontal velocity normally greatly exceeds the average vertical velocity, Eq. (3) can be simplified to

$$f = -\frac{2}{\lambda} v \left(\frac{2y}{L} \right) = -\frac{2}{\lambda} v \sin \alpha \approx -\frac{2}{\lambda} v \alpha \quad (4)$$

where α is the azimuthal angle of the single scatterer from the great circle. For a given crosspath wind speed, pointing the antennas to azimuthal angles away from the great circle tends to favor scatterers at larger angles α , which consequently increases the observed Doppler shift. There are many scatterers at various positions, however, that create a spectrum of Doppler shifts in the received signal. Birkemeier (1968) specified the dependence of Doppler spectrum upon azimuthal pointing angle α_a in terms of the multiplied azimuthal pattern function $G(\alpha)$ of the antennas and the scattered power per unit azimuthal angle $A(\alpha)$. For synchronously pointed antennas with boresights at α_a ,

$$S(f) = S_{\alpha}(\alpha, \alpha_a) = G(\alpha - \alpha_a) A(\alpha) \quad (5)$$

where $S(f)$ is the Doppler spectrum, $S_{\alpha}(\alpha, \alpha_a)$ is Doppler spectrum written in terms of α . And α and f are related by Eq. (4).

Composite Antenna Function

The antenna functions used in the simulation are contrived to be a simple but accurate representation of the actual antennas used in the radio experiments. The parameters of link antennas at the UW link between Arlington, Wisconsin and Cedar Rapids, Iowa and RADC link between Youngstown and Verona, New York are given in Table 1. Under normal operating conditions, the antennas are bore-sited at the horizon. Thus, rays from the lower half of the beam are directed at the foreground and are both absorbed by the

terrain and diffusely reflected. Rays from the upperhalf of each beam are directed over the horizon into the troposphere and their intersections are referred to as the common volume.

Representation of the antenna field pattern by Gaussian functions is analytically convenient and reasonably accurate. If the transmitting antenna is assumed pointed at the horizon over the great circle but the receiver antenna is azimuthally pointed at an angle α_a (azimuthal height y_a) with respect to great circle then the nominal voltage gain at the receiver from any scattering point at a given height z_0 is given by

$$G_R(y)G_T(y) = \frac{1}{r_T^2} \exp\left[-\frac{y^2}{2\sigma_T^2}\right] \exp\left[-(z_0 - z_{gr})^2/2\sigma_T^2\right] \\ \frac{1}{r_R^2} \exp\left[-\frac{(y - y_a)^2}{2\sigma_R^2}\right] \exp\left[-(z_0 - z_{gr})^2/2\sigma_R^2\right] \quad (6)$$

where r_R, r_T = distance to scattering point from receiving and transmitting antennas respectively.

σ_R^2, σ_T^2 = variance of Gaussian approximation determined by antenna beam width.

Within the common volume

z_{gr} = grazing ray height above link axis, and

$$\sigma_R^2 = \frac{(r_R \alpha_3)^2}{(4 \ln 2)} \quad , \quad \sigma_T^2 = \frac{(r_T \alpha_3)^2}{(4 \ln 2)} \quad \text{and} \quad y_a = d \alpha_a$$

where α_3 is -3db beam-width of a single antenna pattern and $2d$ is the length of the link axis as shown in Fig. 1.

Assuming $r_R = r_T = d$, at midpath and grouping the constants in Eq. (6) by $G_0(z_0, d)$, antenna function $G(\alpha)$ can be written as

$$G(\alpha) = G(y) = [G_R(y)G_T(y)]^2 = G_0^2(z_0, d) \exp\left[-\frac{y^2}{\sigma^2}\right] \exp\left[-\frac{(y-y_a)^2}{\sigma^2}\right] \quad (7)$$

where

$$\sigma^2 = \frac{(d\alpha_3)^2}{4\ln 2} \quad \text{and} \quad y = d\alpha \quad \text{with} \quad y_a = d\alpha_a \quad (8)$$

Substituting Eq. (8) into Eq. (7), one obtains:

$$G(\alpha) = G_0^2(z_0, d) \exp\left[-\frac{\alpha^2}{\sigma_0^2}\right] \exp\left[-\frac{(\alpha-\alpha_a)^2}{\sigma_0^2}\right] \quad (9)$$

where $\sigma_0^2 = \alpha_3^2/4\ln 2$.

In the above discussion, single Gaussian function has been used to represent the main lobe of the composite. For better refinement with side lobe structure, bi-Gaussian or penta-Gaussian function (Merrill, 1968) can be used.

Angular Dependence

A major ingredient of turbulence scattering theory is angular dependence. The signal received from a scatterer is a function of the angle between transmitted rays incident upon the scatterer and the rays from the scatterer to the receiver. The angle involved is called the "scattering angle". The reflection coefficient of the layer theories (Friis, Crawford and Hogg, 1957; Ducastel, et al., 1960) involves the same "reflection angle". Merrill (1968) called it "deflection angle" which does not explicitly indicate a particular mechanism.

A combination of turbulence theory and wave theory results in the scattering cross section

$$\sigma = \frac{C}{\lambda^{-p}} \sin^2 \beta \sin^{-p}(\theta/2) \quad (10)$$

where θ is the scattering angle and β is the angle between the electric field vector and the ray to the receiver, β is also a function of position but is always nearly 90° . The factor C includes the effect of wavelength, scale of turbulence and mean-square refractivity fluctuation intensity. For a thin layer with homogeneous statistical properties, C is assumed to be independent of position. The power law p depends upon the particular version of turbulence theory advocated. The prominent Kolmogorov-Tatarski theory (Tatarski, 1961) gives $p = 11/3$.

Based upon the scattering cross section function of Eq. (10), one obtains the angular dependence function $A(\alpha)$

$$A(\alpha) = A(y) = A_0(\sin \theta/2)^{-11/3} \quad (11)$$

where

$$\sin \theta/2 \approx \sqrt{\frac{z_0^2 + y^2}{d^2}} \quad \text{at midpath (x=0) and } A_0 \text{ is a constant}$$

depending upon the wavelength and intrinsic strength of the scatterers. y and α are related as in Eq. (8). Equation (11) can be rewritten as

$$A(\alpha) = A_0(\alpha^2 + z_0^2/d^2)^{-11/6} \quad (12)$$

Frequency dependence of $A(\alpha)$ can be obtained from the relation between α and f as in Eq. (4).

Approximation of Angular Dependence
for Small Antenna Pointing Angles

For small antenna pointing angles, the angular dependence function can be approximated by the Gaussian function of the form

$$A(\alpha) = A(y) = C_0 \exp\left[-\frac{2}{\lambda^2 d^2} (y^2 + z_0^2)\right] \quad (13)$$

where C_0 and λc are constants depending upon the 3-dimensional space correlation length. For a given thin layer at a height z_0 , one can rewrite Eq. (13) as

$$A(\alpha) = A_0(z_0) \exp\left[-\frac{\alpha^2}{\sigma_A^2}\right] \quad (14)$$

where σ_A^2 specifies the severity of rolloff of $A(\alpha)$ and is dependent upon wavelength λ .

Functional Representation of Doppler Spectrum
When no Clock-offset is Present

From Eqs. (5), (9) and (14) one can write the spectral density of the received signal as:

$$\begin{aligned} S(f) \equiv S(\alpha, \alpha_a) &= G_0^2(z_0, d) A_0(z_0) \exp\left[-\left(\frac{\alpha^2}{\sigma_0^2} + \frac{(\alpha - \alpha_a)^2}{\sigma_0^2} + \frac{\alpha_a^2}{\sigma_A^2}\right)\right] \\ &= S_0(z_0, d) \exp\left[-\frac{2\alpha^2 + \alpha_a^2 - 2\alpha\alpha_a}{\sigma_0^2} - \frac{\alpha_a^2}{\sigma_A^2}\right] \end{aligned} \quad (15)$$

Equation (15) can be simplified by simple manipulation as:

$$S(f) = S(\alpha) = [S_0(z_0, d) \exp\left[-\frac{\alpha_a^2(1-k_0)}{\sigma_0^2}\right]] \exp\left[-\frac{(\alpha - k_0\alpha_a)^2}{\sigma_0^2 k_0}\right] \quad (16)$$

where $k_0 = \frac{\sigma_A^2}{\sigma_0^2 + 2\sigma_A^2}$ and α and f are linearly related as in Eq. (4).

As seen from Eq. (16), the shape of the spectrum is Gaussian but shifted by an amount f_a where f_a is

$$f_a = -\frac{2}{\lambda} v \alpha_a \left(\frac{\sigma_A^2}{\sigma_0^2 + 2\sigma_A^2} \right) \quad (17)$$

This is the single scatter value, as in Eq. (4), reduced by the factor in brackets. For a given wind speed, the narrower $A(\alpha)$ is, the smaller is the slope of f_a vs α_a . The Gaussian model for $A(\alpha)$, however, yields a linear f_a vs. α_a relationship that agrees well only for small antenna pointing angles (Birkemeier, 1968).

Functional Representation of Doppler Spectrum When Clock-offset is Present

As mentioned in an earlier report that when the transmitting clock (w_c) and receiver clock (w_0) are not identical then the received Doppler spectrum gets shifted by an amount equal to clock offset Δw ($=w_c - w_0$). So if $\Delta\alpha = \left(\frac{\lambda}{4\pi v}\right) \Delta w$, then the shifted spectrum $S(f)$ can be obtained from Eq. (16) as

$$S(f) = S(\alpha) = S_0(z_0, d, \alpha_a) \exp\left[-\frac{(\alpha - \Delta\alpha - k_0 \alpha_a)^2}{\sigma_0^2 k_0^2}\right] \quad (18)$$

$$\text{where } S_0(z_0, d, \alpha_a) = S_0(z_0, d) \exp\left[-\left(\frac{\alpha_a^2}{\sigma_0^2}\right) \left(\frac{\sigma_0^2 + \sigma_A^2}{\sigma_0^2 + 2\sigma_A^2}\right)\right]$$

The received spectrum is still Gaussian but shifted by an amount $\bar{f}$ where $\bar{f}$ is:

$$\bar{f} = f_a + 2\pi \Delta\omega \quad (19)$$

Equation (18) can be written in frequency domain as:

$$S(f) = S_0(f_a, z_0, d) \exp\left[-\frac{(f-m_1)^2}{2m_2^2}\right] \quad (20)$$

$$\text{where } m_1 = \bar{f}, \quad m_2^2 = \frac{1}{2} \left(\frac{\sigma_0^2 + \sigma_A^2}{\sigma_0^2 + 2\sigma_A^2} \right) \left(\frac{2v}{\lambda} \right)^2$$

$$\text{and } S_0(f_a, z_0, d) = S_0(z_0, d) \exp\left[-\left(\frac{\sigma_0^2 + \sigma_A^2}{\sigma_0^2 + 2\sigma_A^2}\right) \left(\frac{\lambda}{2v}\right)^2 \left(\frac{f_a}{\sigma_0}\right)^2\right]$$

The resulting spectra is still Gaussian. It is well known that symmetrical Gaussian is represented by two moments of the spectrum where the first moment (m_1) gives the centroid (mean) and is defined by:

$$m_1 = \frac{\int f S(f) df}{\int S(f) df} \quad (21)$$

and the 2nd central moment (m_2) gives the variance of the spectrum and is defined by

$$m_2 = \frac{\int (f-m_1)^2 S(f) df}{\int S(f) df} \quad (22)$$

The expression of Doppler spectrum in Eq. (20) was derived based on the assumption that combined antenna pattern and angular dependence is Gaussian. This results in symmetrical Doppler spectra with mean m_1 and variance m_2 . However, experimental evidence (Birkemeier, 1968) shows that the spectrum tended to be asymmetrical, when the transmitting and receiving antennas are not pointed at the great circle. At small antenna offsets from boresight, few correction terms in Gram-charlier expansion of the nearly Gaussian spectrum can account for this asymmetry factor, e.g.,

$$S(f) = \frac{1}{\sqrt{2\pi}m_2} \exp\left(-\frac{1}{2}\left(\frac{f-m_1}{m_2}\right)^2\right) \left[1 + \frac{s_k}{3!} H_3\left(\frac{f-m_1}{m_2}\right) + \frac{Ex}{4!} H_4\left(\frac{f-m_1}{m_2}\right) + \frac{s_k^2}{2!3!3!} H_6\left(\frac{f-m_1}{m_2}\right)\right] \quad (23)$$

Equation (23) is the corrected form of Eq. (20) such that total input power $[\int S(f)df = 1]$ is unity. In Eq. (23) s_k is defined as skew coefficient or asymmetry coefficient by:

$$s_k = \frac{m_3}{m_2^{1.5}} \quad (24)$$

where $m_3 = \left(\frac{1}{2\pi}\right)^3 \frac{\int (\omega-m_1)^3 S(\omega) d\omega}{\int S(\omega) d\omega}$ and is defined as coefficient of excess, e.g.

$$Ex = \frac{m_4}{m_2^2} - 3 \quad (25)$$

where $m_4 = \left(\frac{1}{2\pi}\right)^4 \frac{\int (\omega-m_1)^4 S(\omega) d\omega}{\int S(\omega) d\omega}$

For normal spectral slope $Ex = 0$. For other spectra having the same variance as the normal spectra, Ex will indicate whether the distribution has a flatter ($Ex < 0$) or sharper ($Ex > 0$) peak than the normal spectrum.

One should note that skew (s_k) and excess (Ex) can be obtained from the third (m_3) and fourth (m_4) central moments of the Doppler spectrum if non-Gaussian angular dependence (Eq. 12) is considered and thus related to tropolink parameters. In Eq. (23) $H_n(y)$ is a Hermite polynomial of degree n defined by:

$$H_n(y) = (-1)^n e^{y^2/2} \frac{d^n}{dy^n} (e^{-y^2/2}) \quad (26)$$

with $H_3(y) = y^3 - 3y$ and $H_4(y) = y^4 - 6y^2 + 3$ and $H_{n+1}(y) = y H_n(y) - n H_{n-1}(y)$.

Autocorrelation Function of Doppler Spectrum

The autocorrelation function $R(\tau)$ of the spectrum density (Doppler spectrum) $S(\omega)$ is defined as:

$$R(\tau) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S(\omega) e^{j\omega\tau} d\omega \quad (27)$$

If $S(f)$ is as given as in Eq. (23) then $R(\tau)$ can be obtained by simple manipulation as

$$R(\tau) = \left(\exp(jm_1\tau - \frac{m_2^2\tau^2}{2}) \right) \left(1 - \frac{j s_k m_2^3 \tau^3}{6} + \frac{(Ex)m_2^4 \tau^4}{24} - \frac{s_k^2 m_2^6 \tau^6}{72} \right) \quad (28)$$

m_1, m_2, s_k, Ex are the moments of the Doppler spectrum $S(f)$ defined earlier (Eqs. 21, 22, 24, 25).

The first and second derivatives of Eq. (28), which will later be used in alignment accuracy analysis, is given as:

$$\begin{aligned}
 \dot{R}(\tau) = & \left(\exp(jm_1\tau - \frac{m_2^2\tau^2}{2}) \right) \left[jm_1 - m_2^2\tau - j \frac{s_k m_2^3}{2} \tau^2 + \left(\frac{m_1 s_k m_2^3}{6} + \frac{(Ex)m_2^4}{6} \right) \tau^3 \right. \\
 & + j \left(\frac{(Ex)m_2^4 m_1}{24} + \frac{s_k m_2^5}{6} \right) \tau^4 - \left(\frac{(Ex)m_2^6}{24} + \frac{s_k m_2^6}{12} \right) \tau^5 \\
 & \left. - j \frac{m_1 s_k m_2^6}{72} \tau^6 + \frac{s_k m_2^8}{72} \tau^7 \right] \quad (29)
 \end{aligned}$$

$$\text{and } \ddot{R}(\tau) = (-B_1 + B_2\tau^2 - B_3\tau^4 + B_4\tau^6 - B_5\tau^8) - j(c_1\tau + c_2\tau^3 - c_3\tau^5 + c_4\tau^7) \quad (30)$$

$$\text{where } B_1 = m_1^2 + m_2^2$$

$$B_2 = m_2^4 + m_1 s_k m_2^3 + \frac{(Ex)m_2^4}{2}$$

$$B_3 = \frac{9}{24} (Ex)m_2^6 + \frac{5}{12} s_k m_2^6 + \frac{(Ex)m_2^4 m_1^2}{24} + \frac{m_1 s_k m_2^5}{3}$$

$$B_4 = \frac{13}{72} s_k m_2^8 + \frac{m_1^2 s_k m_2^6}{72} + \frac{(Ex)m_2^8}{24}$$

$$B_5 = \frac{s_k m_2^{10}}{72}$$

$$c_1 = s_k m_2^3 + 2m_1 m_2^2$$

$$c_2 = \frac{m_1 s_k m_2^4}{3} + \frac{2 s_k m_2^5}{3} + \frac{s_k m_2^5}{2} + \frac{m_1^2 s_k m_2^3}{6}$$

$$c_3 = \frac{m_1 s_k m_2^6}{6} + \frac{m_1 (Ex)m_2^6}{12} + \frac{s_k m_2^7}{6}$$

$$c_4 = \frac{m_1 s_k m_2^8}{36}$$

Accuracy Analysis of the Slope Method

In the earlier report (Nov., 1977), the error analysis of the slope of the quadrature correlation function at zero lag has been shown. The formulas are summarized below for two estimators of the slope:

- a) True Estimator: In this estimator, the slope (S) is estimated from the expression:

$$\hat{S} = \frac{1}{T} \int_0^T \beta(t) \frac{d}{dt} \alpha(t) dt \quad (31)$$

with $E[\hat{S}] = \dot{R}_{\alpha\beta}(0)$

where 'E' means "the expected value of" and $\alpha(t)$ and $\beta(t)$ are the quadrature components of the complex process $(\alpha(t) + j\beta(t))$ whose spectral density is given by Eq. (23). The variance of this estimator is given by:

$$\text{var}[\hat{S}] = \frac{2}{T} \int_0^T \left(1 - \frac{\tau}{T}\right) [R_{\alpha\alpha}(\tau) R_{\beta\beta}(\tau) + [R_{\alpha\beta}(\tau)]^2] d\tau \quad (32)$$

where $R_{\alpha\alpha}(\tau) = \text{Re}[R(\tau)]$ and $R_{\alpha\beta}(\tau) = \text{Im}[R(\tau)]$.

- b) Estimator Hard Limiting: In this estimator, the input signals are first clipped by a 'sqn' operator $\begin{bmatrix} +1 & x(t) > 0 \\ 0 & x(t) = 0 \\ -1 & x(t) < 0 \end{bmatrix}$, where $x(t)$ is the input signal] and then one of the clipped signals is differentiated, e.g., the slope of the quadrature correlation function at zero lag is estimated from:

$$\hat{S}_{CL} = \frac{\pi}{2T} \int_0^T [\text{sgn } \beta(t) \frac{d}{dt} [\text{sgn} \alpha(t)]] dt$$

$$\text{with} \quad E[\hat{S}_{CL}] = \dot{R}_{\alpha\beta}(0) \quad (33)$$

$$\text{and} \quad \text{var}[\hat{S}_{CL}] = \frac{\pi^2}{4T^2} - \frac{1}{T^2} [\sin^{-1}(\rho(\tau))]^2 \quad (34)$$

$$+ \frac{2}{T} \int_0^T (1 - \frac{\tau}{T}) \frac{\dot{\gamma}^2(0)[\gamma^2 + \rho^2] + \dot{\rho}^2 + \dot{\gamma}^2 + 2\dot{\gamma}(0)[\dot{\rho}\gamma - \rho\dot{\gamma}]}{1 - \rho^2 - \gamma^2} d\tau$$

where $\rho(\tau) = \text{Re}[\frac{R(\tau)}{R(0)}]$ and $\gamma(\tau) = \text{Im}[\frac{R(\tau)}{R(0)}]$

where 'Re' implies 'Real' and 'Im' implies 'Imaginary'.

Equations 28, 29 and 30 were substituted in Eqs. (32) and (34) and the variance is compared for both estimators with different spectral moments (m_1, m_2, m_3 and m_4 is assumed zero for simplicity). The results are summarized below:

Effect of Fade Rate on Skew: As mentioned before, the spectral moments are functions of tropolink parameters such as wind speed, antenna beam width, angular dependence, transmitted frequency, the distance between the transmitter and the receiver etc. The antenna elevation angle is assumed zero. The variations in the link parameters causes these coupled spectral moments to change accordingly. So, simply varying these moments will mimic the different link conditions, the net effect of which will result in changed moments. In table 2, with $m_1 = 0$, the effect of fade rate under different 'skew' (asymmetric) conditions is shown. Although with different 'skew' conditions, m_1 should correspondingly be non-zero, but in calculations

first m_1 is assumed zero (Table 2) and then for a given m_2 , m_1 is varied (Table 3). For a given skew coefficient, the difference in m_1 can account for the 'residual' clock-offset (see Eq. 19). From Table 2, it could be observed that the variance of both estimators increases in proportion to the spectral width. If the spectral width (or the fade rate) increases by a factor of 2, the variance also increases by the same factor. The theoretical explanation is given in Appendix 1. The clipping loss (defined as the ratio of $\frac{\text{variance of true estimator of slope}}{\text{variance of clipped estimator of slope}}$) is in the range of -4 to -5 db. If the large amount of clock-offset is present, then the clipping loss decreases accordingly (see Table 3). If the Doppler spectrum is symmetrical ($m_3 = 0$), then it is shown in Table 4 that for large clock-offset (as m_1 increases) the clipping loss could result in clipping gain.

Path	UW Link	RADC Link
Antenna parameters	Symmetrical, 28' steerable diameter paraboloids with circular nominal -3 db bandwidth of 2.8°, 0° elevation $\sigma_0^2 = 2.82$ (degree ²)	Symmetrical, 28' steerable diameter paraboloids with circular nominal -3 db bandwidth of .54°, 0° elevation $\sigma_0^2 = .105$ (degree ²)
Path length (2d)	231 km	168 miles, 268.80 km
Frequency, wavelength	810 MHz, .37 m	4.86 GHz, .0625 m
Size of common volume (2Lx*2Ly*Lz) (lengths in km)	110.*5.6*2.8	34.9*1.26*.63
Z _{gr} , grazing ray height from link axis	1.57 km	2.12 km
Z _c , height of earth surface from link-axis	.78 km	1.06 km
Z _{up} , height of upper beam from link axis	4.39 km	2.75 km
Z _b , height of the intersection of the transmitter upper beam and receiver from link axis	2.31 km	2.40 km
Z ₀ , single layer height assumed from link axis	2.5 km	2.5 km
σ_A^2 , variance (degree ²) of Gaussian angular dependence for fade rate (m_2) of 2,4,6,8 & 10 Hz and wind speed 2.5 m/sec	.045, .2, .547, 1.394 4.905	.001, .005, .014, .032, .077 (and .135 for $m_2 = 11$ Hz)
θ , angle of antenna central beam to the link axis	1.55°	1.81°

Table D.1

Link Parameters for UW and RADC Links

Table 2

Comparison of Two Estimators of Slope at Different Fade Rates and Antenna Misaligned Positions ($m_1=0$, $E_x=0$, $T=5$).

S_k skew coefficient	VARTR variance of true estimator			VARCLP variance of clipped estimator			LOSS (VARTR/VARCLP)			DB LOSS $10 \log_{10} (\text{Loss})$		
	m_2											
	.5	1.0	2.0	.5	1.0	2.0	.5	1.0	2.0	.5	1.0	2.0
-.50	.0751	.1615	.3507	.2697	.5126	.9986	.2784	.3151	.3512	-5.5540	-5.0160	-4.5447
-.40	.0770	.1617	.3451	.2678	.5065	.9840	.2875	.3192	.3507	-5.4133	-4.9588	-4.5501
-.30	.0793	.1636	.3431	.2663	.5018	.9728	.2980	.3259	.3526	-5.2586	-4.8686	-4.5267
-.25	.0807	.1651	.3432	.2657	.5000	.9685	.3036	.3301	.3544	-5.1772	-4.8133	-4.5054
-.20	.0821	.1669	.3441	.2652	.4985	.9650	.3095	.3343	.3566	-5.0938	-4.7523	-4.4782
-.15	.0836	.1690	.3457	.2648	.4974	.9623	.3156	.3399	.3593	-5.0088	-4.6066	-4.4459
-.10	.0852	.1715	.3480	.2646	.4965	.9603	.3219	.3454	.3623	-4.9227	-4.6169	-4.4089
0	.0886	.1772	.3543	.2644	.4959	.9588	.3351	.3573	.3696	-4.7489	-4.4691	-4.3232
.1	.0923	.1838	.3630	.2646	.4965	.9603	.3487	.3702	.3779	-4.5754	-4.3152	-4.2258
.15	.0942	.1875	.3680	.2648	.4974	.9623	.3557	.3769	.3825	-4.4896	-4.2376	-4.1740
.20	.0962	.1913	.3736	.2652	.4985	.9650	.3627	.3837	.3872	-4.4049	-4.1606	-4.1208
.25	.0982	.1952	.3797	.2657	.5000	.9685	.3697	.3904	.3920	-4.3216	-4.0845	-4.0667
.30	.1003	.1993	.3862	.2663	.5018	.9728	.3767	.3972	.3970	-4.2398	-4.0100	-4.0121
.40	.1046	.2079	.4006	.2678	.5065	.9840	.3907	.4105	.4071	-4.0820	-3.8672	-3.9030
.50	.1090	.2169	.4166	.2697	.5126	.9986	.4043	.4231	.4172	-3.9331	-3.7352	-3.7963

Table 3

Effect of non-zero m_1 in comparing the performance of two estimators of slope at a given fade rate and different antenna misaligned positions for $m_2=1$, $E_x=0$ and $T=5$.

skew coefficient	VARTR variance of true estimator			VARCLP variance of clipped estimator			LOSS (VARTR/VARCLP)			DB LOSS 10 log ₁₀ (Loss)		
	$m_1=0$	.1	.3	0	.1	.3	0	.1	.3	0	.1	.3
-.50	.1615	.1767	.2257	.5126	.5126	.5126	.3151	.3446	.4403	-5.0160	-4.6263	-3.5626
-.40	.1617	.1743	.2185	.5065	.5065	.5065	.3192	.3440	.4315	-4.9588	-4.6339	-3.6500
-.30	.1636	.1736	.2131	.5018	.5018	.5018	.3259	.3460	.4246	-4.8686	-4.6098	-3.7200
-.25	.1651	.1739	.2109	.5000	.5000	.5000	.3301	.3478	.4219	-4.8133	-4.5871	-3.7479
-.20	.1669	.1745	.2092	.4985	.4985	.4985	.3348	.3501	.4196	-4.7523	-4.5592	-3.7715
-.15	.1690	.1775	.2078	.4974	.4974	.4974	.3399	.3529	.4177	-4.6866	-4.5236	-3.7408
-.10	.1715	.1768	.2067	.4965	.4965	.4965	.3454	.3561	.4163	-4.6169	-4.4840	-3.8062
0	.1772	.1803	.2055	.4959	.4959	.4959	.3573	.3637	.4144	-4.4691	-4.3927	-3.8257
.10	.1838	.1849	.2055	.4965	.4965	.4965	.3702	.3724	.4138	-4.3152	-4.2894	-3.8318
.15	.1875	.1876	.2059	.4974	.4974	.4974	.3769	.3772	.4140	-4.2376	-4.2348	-3.8304
.20	.1913	.1904	.2065	.4985	.4985	.4985	.3637	.3820	.4143	-4.1606	-4.1790	-3.8265
.25	.1952	.1935	.2075	.5000	.5000	.5000	.3904	.3870	.4149	-4.0845	-4.1226	-3.8202
.30	.1993	.1968	.2086	.5018	.5018	.5018	.3972	.3921	.4157	-4.0100	-4.0661	-3.8119
.40	.2079	.2038	.2116	.5065	.5065	.5065	.4105	.4023	.4178	-3.8672	-3.9566	-3.7900
.50	.2169	.2114	.2155	.5126	.5126	.5126	.4231	.4123	.4205	-3.7352	-3.8477	-3.7601

Table 4
Comparison of two estimators of slope of quadrature correlation at zero lag
(or center frequency of symmetrical doppler spectrum) with varying clock offset and antenna
positions ($S_k=0$, $E_x=0$, $T=5$)

fade rate m_2 , spectral width	VARTR variance of true estimator					VARCLP variance of clipped estimator					LOSS = $\frac{\text{VARTR}}{\text{VARCLP}}$					DB LOSS = $10 \log_{10} (\text{Loss})$				
	$m_1=0$	.1	.3	.5	0	.1	.3	.5	0	.1	.3	.5	0	.1	.3	.5	0	.1	.3	.5
.1	.0092	.0260	.1266	.2604	.0617	.0771	.1083	.0838	.1494	.3371	1.1690	3.1070	-8.2556	-4.7228	.6702	4.9234				
.2	.0299	.0419	.1220	.2505	.1186	.1230	.1355	.1253	.2519	.3404	.9006	1.9995	-5.9881	-4.6795	-.4545	3.0092				
.3	.0514	.0598	.1225	.2389	.1699	.1709	.1743	.1715	.3023	.3497	.7031	1.3926	-5.1952	-4.5632	-1.5301	1.4384				
.4	.0706	.0771	.1286	.2300	.2178	.2179	.2185	.2180	.3240	.3537	.585	1.0547	-4.8944	-4.5132	-2.3028	0.2316				
.5	.0886	.0341	.1381	.2258	.2644	.2644	.2644	.2644	.3351	.3559	.5221	.8542	-4.7489	-4.4872	-2.8228	-.6846				
1.0	.1772	.1803	.2055	.2558	.4959	.4959	.4959	.4959	.3573	.3637	.4144	.5159	-4.4691	-4.3927	-3.8257	-2.8747				
1.5	.2658	.2680	.2854	.3204	.7274	.7274	.7274	.7274	.3654	.3684	.3929	.4405	-4.3723	-4.3367	-4.0623	-3.5606				
2.0	.3543	.3560	.3694	.3961	.9588	.9588	.9588	.9588	.3696	.3713	.3852	.4131	-4.3232	-4.3028	-4.1426	-3.8390				
2.5	.4429	.4442	.4550	.4767	1.1902	1.1902	1.1902	1.1902	.3721	.3732	.3823	.4005	-4.2936	-4.2804	-4.1758	-3.9729				
3.0	.5314	.5325	.5416	.5598	1.4216	1.4216	1.4216	1.4216	.3738	.3746	.3810	.3938	-4.2739	-4.2645	-4.1911	-4.0477				
4.0	.7083	.7092	.7161	.7299	1.8844	1.8844	1.8844	1.8844	.3759	.3764	.3800	.3873	-4.2493	-4.2440	-4.2020	-4.1193				

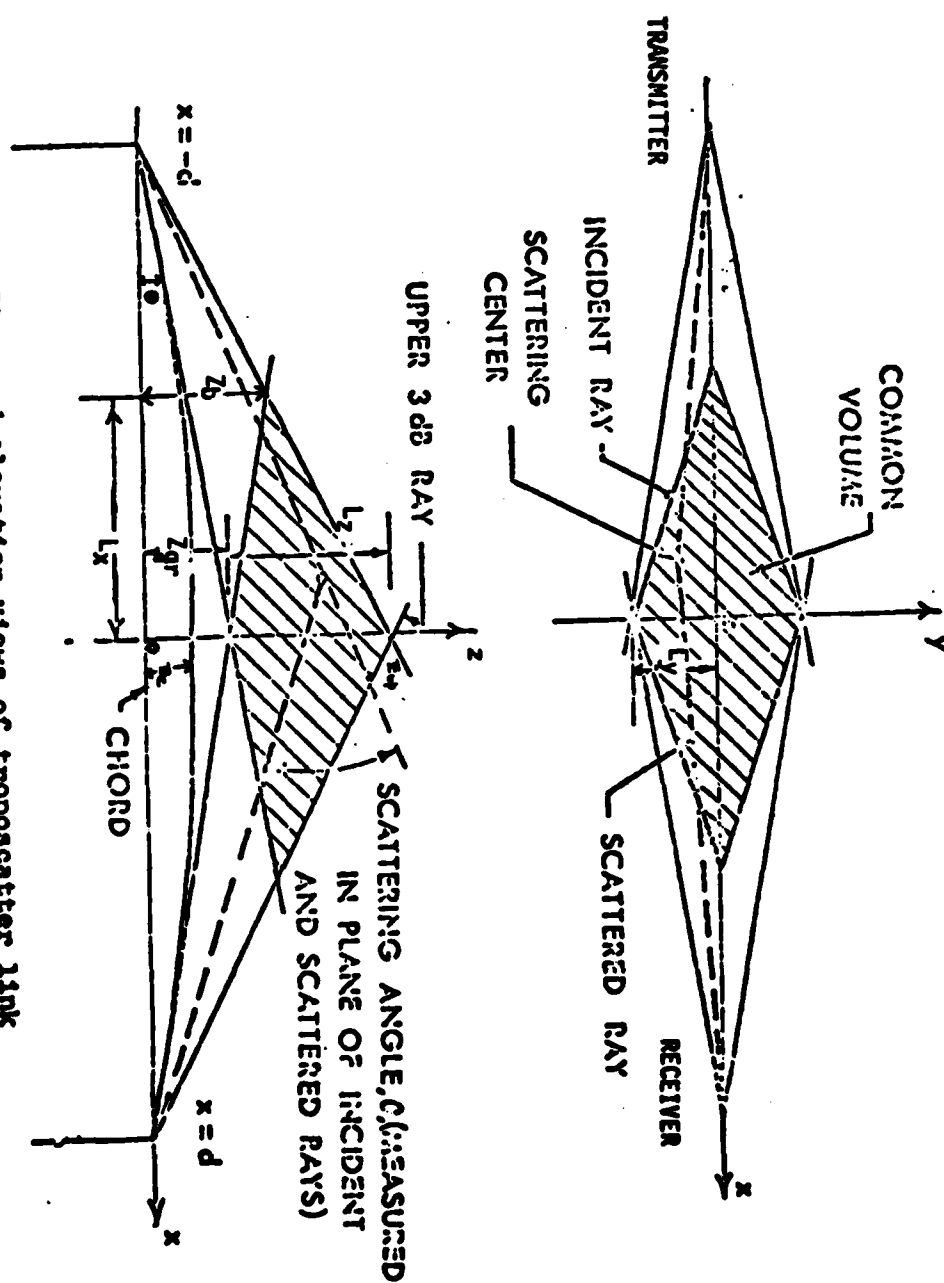


Figure 1 Plan and elevation views of troposcatter link

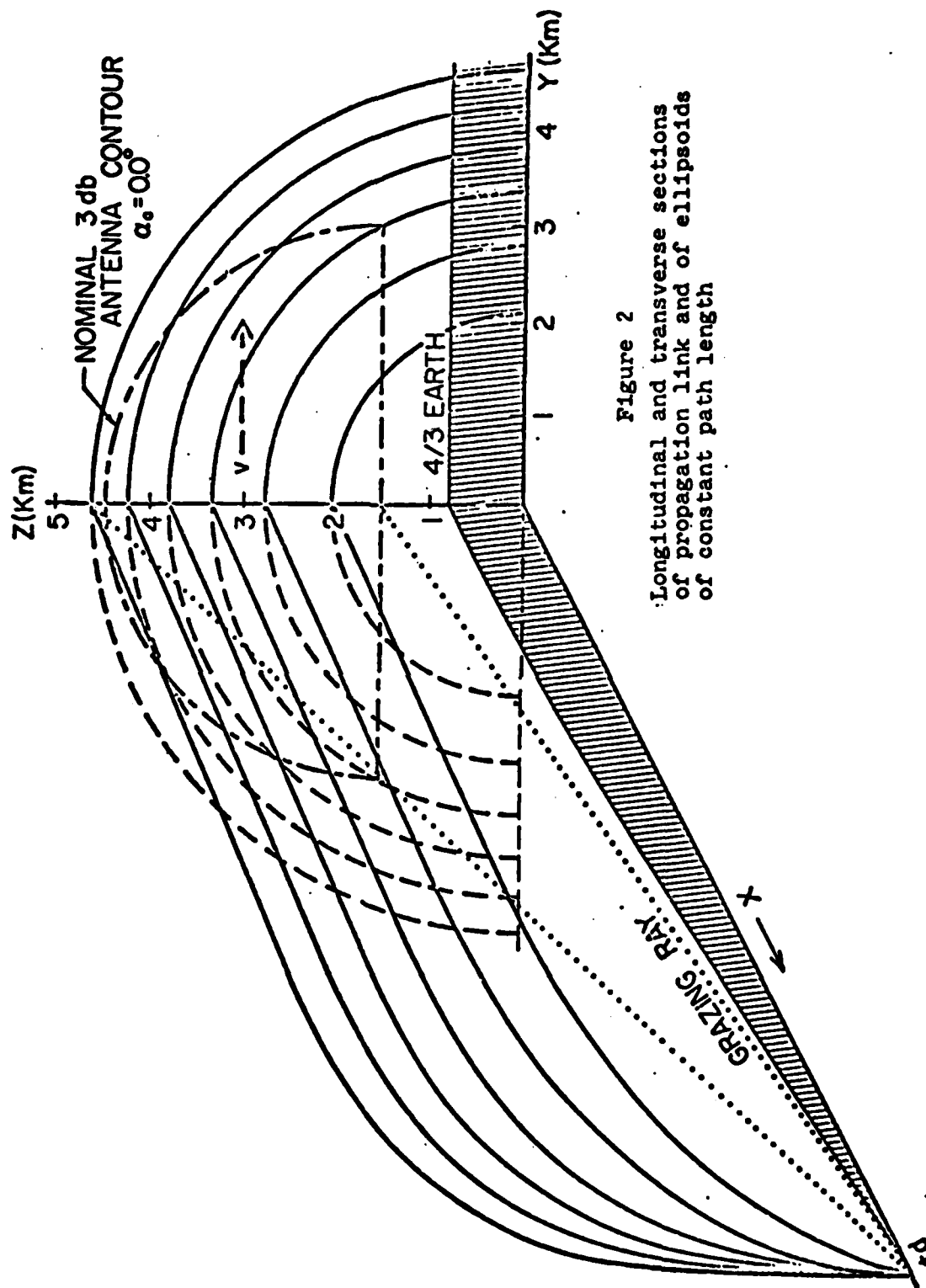


Figure 2
Longitudinal and transverse sections
of propagation link and of ellipsoids
of constant path length

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Appendix D1

In this appendix the effect of the change in the width of the Doppler spectrum on the variance of the two estimators of the slope of the quadrature correlation function at zero lag is considered. Since the variance is inversely proportional to the averaging time and the spectral width is an indicator of fade rate, so the effect of the different fade rates (due to different wind velocity, antenna patterns, etc.) on the required averaging time is estimated by assuming that the total received power is the same under various fading conditions, e.g., if the Doppler spectrum increases in width by a factor of k , then

$$R_{\alpha\beta}(\tau) \rightarrow R_{\alpha\beta}(k\tau)$$

and

(A1-1)

$$R_{\alpha\alpha}(\tau) \rightarrow R_{\alpha\alpha}(k\tau)$$

where $R_{\alpha\beta}(\tau)$ is the quadrature cross-correlation function and $R_{\alpha\alpha}(\tau)$ is the quadrature auto-correlation function of quadrature processes $\alpha(t)$ and $\beta(t)$. In deriving Eq. (A1-1), the well-known Fourier transform scaling relationship is utilized, e.g.

$$\frac{1}{a} F\left(\frac{\omega}{a}\right) \leftrightarrow f(a\tau)$$

Substituting Eq. (A-1) in equations (32) and (34), one obtains

$$\text{var}[\hat{S}] = \frac{2k}{T} \int_0^T \left(1 - \frac{n}{kT}\right) [R_{\alpha\alpha}(n)R_{\beta\beta}(n) + [R_{\alpha\beta}(n)]^2] dn \quad (\text{A1-2})$$

and

$$\begin{aligned} \text{var}[\hat{S}_{CL}] &= \frac{\pi^2}{4T^2} - \frac{1}{T^2} [\sin^{-1}(\rho(T))]^2 \\ &+ \frac{2k}{T} \int_0^T \left(1 - \frac{n}{kT}\right) \left(\frac{\dot{\gamma}^2(0)(\gamma^2 + \rho^2) + \dot{\rho}^2 + \dot{\gamma}^2 + 2\dot{\gamma}(0)[\dot{\rho}\gamma - \rho\dot{\gamma}]}{1 - \gamma^2 - \rho^2} \right) dn \end{aligned} \quad (A1-3)$$

For large 'T', Eq. (A1-2) and (A1-3) reduce to

$$\text{Var}[\hat{S}] = \frac{2k}{T} \int_0^T [R_{\alpha\alpha}(n)R_{\beta\beta}(n) + [R_{\alpha\beta}(n)]^2] dn$$

and

$$\text{Var}[\hat{S}_{CL}] = \frac{2k}{T} \int_0^T \frac{\dot{\gamma}^2(0)(\gamma^2 + \rho^2) + \dot{\rho}^2 + \dot{\gamma}^2 + 2\dot{\gamma}(0)[\dot{\rho}\gamma - \rho\dot{\gamma}]}{1 - \rho^2 - \gamma^2} dn \quad (A1-4)$$

For Eq. (A1-4) it is evident that the variance of both estimators increase by the same factor as the width of the Doppler spectrum. In other words, an increase in the width of the Doppler spectrum requires the increase in the averaging time by the same factor.

Estimation of Quadrature Correlation Function for all Delays by Analog and Polarity-Coincidence Method:

In the report (9/11/77-10/11/77), the expected value and the variance of the estimator of the quadrature correlation function by analog method have been derived which are summarized below.

If $\alpha(t)$ and $\beta(t)$ are the quadrature processes of unit power then the quadrature correlation function $R_{\alpha\beta}(\tau)$ is estimated by:

$$\hat{R}_{\alpha\beta}(\tau) = \frac{1}{T} \int_0^T \alpha(t)\beta(t-\tau)dt \quad (1)$$

the expected value and variance of $\hat{R}_{\alpha\beta}(\tau)$ is given by:

$$E[\hat{R}_{\alpha\beta}(\tau)] = R_{\alpha\beta}(\tau) \quad (2)$$

and

$$\text{var}[\hat{R}_{\alpha\beta}(\tau)] = \frac{1}{T} \int_{-T}^T \left(1 - \frac{|n|}{T}\right) [R_{\alpha\alpha}^2(n) + R_{\alpha\beta}(n+\tau)R_{\beta\alpha}(n-\tau)]dn \quad (3)$$

For $\tau=0$, zero delay (lag), Eq. (2) and (3) reduce to:

$$E[\hat{R}_{\alpha\beta}(0)] = 0 \quad (4)$$

and

$$\text{var}[\hat{R}_{\alpha\beta}(0)] = \frac{1}{T} \int_{-T}^T \left(1 - \frac{|n|}{T}\right) [R_{\alpha\alpha}^2(n) - R_{\alpha\beta}^2(n)]dn \quad (5)$$

In the same report, the quadrature correlation function is estimated by the polarity-coincidence method as

$$\hat{r}_c(\tau) = \frac{1}{T} \int_0^T \text{sgn}[\alpha(t)]\text{sgn}[\beta(t-\tau)]dt \quad (6)$$

where sgn is the 'hard-clipped' operator, defined as

$$\begin{aligned} &1 \quad x > 0 \\ \text{sgn}(x) &= 0 \quad x = 0 \\ &-1 \quad x < 0 \end{aligned}$$

the expected value of $\hat{r}_c(\tau)$ is given by Van Vleck [1966], e.g.

$$E[\hat{r}_c(\tau)] = \frac{2}{\pi} \arcsin(R_{\alpha\beta}(\tau)) \quad (7)$$

In the above mentioned report, the variance of $\hat{r}_c(0)$ had been derived as

$$\text{Var}[\hat{r}_c(0)] = \frac{4}{\pi^2 T} \int_{-T}^T [\{\arcsin \rho(n)\}^2 - \{\arcsin r(n)\}^2] dn \quad (8)$$

In Eq. (8), $\rho(n)$ is the autocorrelation function of in-phase or quadrature processes and $r(n)$ is the normalized quadrature correlation function given by Eq. (1).

In this report, the variance of the estimator of the polarity-coincidence quadrature correlation function is computed for all delays and the two estimators of correlation function are compared on the basis of signal to noise ratio (SNR). If SNR is defined as

$$\text{SNR} = \frac{(\text{DC output of the correlation at a given lag})^2}{(\text{variance of the output at that lag})} \quad (9)$$

then clipping loss or gain [Cheng, 1968] is defined as

$$\frac{(\text{SNR of polarity coincidence estimator})}{(\text{SNR of analog estimator})} \quad (10)$$

Variance of Polarity Coincidence Quadrature Correlator

In this correlator, the quadrature correlation function ($r_c(\tau)$) is estimated by Eq. (6). The variance of this estimator is given by:

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$$\text{Var}[\hat{r}_c(\tau)] = \frac{1}{T^2} \iint_{00}^{TT} E[\text{sgn}\alpha(u)\text{sgn}\alpha(v)\text{sgn}\alpha(u-\tau)\text{sgn}\beta(v-\tau)]dudv - E^2[\hat{r}_c(\tau)] \quad (11)$$

In Eq. (11), one needs to obtain fourth product moment [McFadden, 1958]:

$$E[\text{sgn}\alpha(u)\text{sgn}\alpha(v)\text{sgn}\beta(u-\tau)\text{sgn}\beta(v-\tau)],$$

where α and β are jointly gaussian with zero mean and equal variance (σ).

To evaluate this equation, we write:

$$y_1 = \text{sgn}\alpha(u)$$

$$y_2 = \text{sgn}\beta(u-\tau)$$

$$y_3 = \text{sgn}\alpha(v)$$

$$y_4 = \text{sgn}\beta(v-\tau)$$

and

$$P_4 = P[\alpha(u) \geq 0, \beta(u-\tau) \geq 0, \alpha(v) \geq 0, \beta(v-\tau) \geq 0] .$$

Then it is known that, for example, [McFadden, 1966] and [Childs, 1967] that

$$P_4 = \frac{1}{16} [1 + \sum_{j>i \geq 1}^4 E[y_i y_j] + E[y_1 y_2 y_3 y_4]] \quad (12)$$

The problem now is to evaluate P_4 . To do this, we let

$$z_1 = \alpha(u)/\sigma$$

$$z_2 = \beta(u-\tau)/\sigma$$

$$z_3 = \alpha(v)/\sigma$$

$$z_4 = \beta(v-\tau)/\sigma$$

Then one can obtain (see report 9/11/77-10/11/77) the covariance matrix (C) of the Gaussian variables z_1, z_2, z_3, z_4 with zero mean and unit variance.

If $\eta = u-v$, then

$$C = \begin{vmatrix} 1 & r(\tau) & \rho(n) & r(n+\tau) \\ r(\tau) & 1 & r(\tau-n) & \rho(n) \\ \rho(n) & r(\tau-n) & 1 & r(\tau) \\ r(n+\tau) & \rho(n) & r(\tau) & 1 \end{vmatrix}$$

$$= \begin{vmatrix} \rho_{11} & \rho_{12} & \rho_{13} & \rho_{14} \\ \rho_{12} & \rho_{22} & \rho_{23} & \rho_{24} \\ \rho_{13} & \rho_{23} & \rho_{33} & \rho_{34} \\ \rho_{14} & \rho_{24} & \rho_{34} & \rho_{44} \end{vmatrix} \quad (13)$$

It is well known that P_4 for any covariance matrix can not be obtained in closed form. In the following discussion P_4 is computed by Plackett's method (1954) which reduces to closed form expression (Eq. 8) for the zero lag case.

We now make use of the result of Appendix 1, Eq. (12) and Van Vleck's (1966) formula $E[y_i y_j] = \frac{2}{\pi} \arcsin \rho_{ij}$ for $i, j=1, 2, 3, 4$ and obtain the expression for fourth-product moment, i.e.

$$E[y_1 y_2 y_3 y_4] = 1 + \frac{4}{\pi^2} [2I_1 + 2I_2 + I_3 + I_4] \quad (14)$$

where I_1 , I_2 , I_3 and I_4 are given in the Appendix 1 (equations A1-15, 16, 17, 18).

Substituting Eq. (14) into Eq. (11), one obtains

$$\text{Var}(\hat{r}_c(\tau)) = \left[\frac{1}{T} \int_{-T}^T \left(1 - \frac{|n|}{T}\right) E[y_1 y_2 y_3 y_4] dn \right] - E^2[\hat{r}_c(\tau)] \quad (15)$$

$$= 1 + \left[\frac{4}{\pi^2} \int_{-T}^T \left(1 - \frac{|n|}{T}\right) [2I_1 + 2I_2 + I_3 + I_4] dn \right] - E^2[\hat{r}_c(\tau)] \quad (16)$$

Since

$$\int_{-T}^T \left(1 - \frac{|n|}{T}\right) I_3 dn = \int_{-T}^T \left(1 - \frac{|n|}{T}\right) I_4 dn$$

Eq. (16) can be simplified to

$$\text{var}[\hat{r}_c(\tau)] = 1 + \left[\frac{4}{\pi^2 T} \int_{-T}^T \left(1 - \frac{|n|}{T}\right) [2I_1 + 2I_3 + 2I_4] dn \right] - E^2[\hat{r}_c(\tau)] \quad (17)$$

Estimation of the Slope of the Quadrature Correlation Function ($R(\tau)$) at Zero Delay, When the Maximum Value of $R(\tau)$ is used as an Estimator.

In this section, the 'slope' of the quadrature correlation function at zero lag is estimated by the maximum value of the quadrature correlation function, i.e.

$$\text{Slope} = \hat{S}_m = \frac{\hat{R}_m(\tau_0) - \hat{R}(0)}{\tau_0} \quad (18)$$

Here $\hat{R}_m(\tau_0)$ is the estimated maximum value of $R(\tau)$ occurred at lag τ_0 .

If $R(\tau)$ is estimated by the Analog method, (see Fig. 1), then the slope estimator is given by:

$$\hat{S}_{AN} = \frac{\hat{R}_{\alpha\beta}(\tau_0) - \hat{R}(0)}{\tau_0} \quad (19)$$

the expected value of this estimator is given by

$$E[\hat{S}_{AN}] = \frac{E[\hat{R}_{\alpha\beta}(\tau_0)]}{\tau_0} = \frac{R_{\alpha\beta}(\tau_0)}{\tau_0} \quad (20)$$

If $R(\tau)$ is estimated by the Polarity-coincidence method, then

$$\hat{S}_{PCC} = \frac{r_c(\tau_0) - r_c(0)}{\tau_0} \quad (21)$$

the expected value of this estimator is given by

$$E[\hat{S}_{PCC}] = \frac{E[r_c(\tau_0)]}{\tau_0} = \frac{1}{\tau_0} r_c(\tau_0) = \frac{2}{\pi} \frac{1}{\tau_0} \arcsin R_{\alpha\beta}(\tau_0) \quad (22)$$

Variance Calculations:

The variance of the slope estimator is given by:

$$\tau_0^2 \text{var}[\hat{S}_m] = \text{var}[\hat{R}_m(\tau_0)] + \text{var}[\hat{R}(0)] - 2E[\hat{R}_m(\tau_0)\hat{R}(0)] \quad (23)$$

The variance of $\hat{R}_m(\tau_0)$ and $\hat{R}(0)$ can be computed by the methods discussed in the earlier section for both analog and polarity-coincidence cases. Now a method is derived to compute the third term for both cases. When $\hat{R}_m(\tau_0)$ and $\hat{R}(0)$ are computed by Analog method, then

$$\hat{R}(0) = \frac{1}{T} \int_0^T \alpha(t) \beta(t) dt$$

and

$$\hat{R}_m(\tau_0) = R_{\alpha\beta}(\tau_0) = \frac{1}{T} \int_0^T \alpha(t) \beta(t - \tau_0) dt \quad (24)$$

So $E[\hat{R}_m(\tau_0)\hat{R}(0)]$ is given by:

$$\begin{aligned} E[\hat{R}_m(\tau_0)\hat{R}(0)] &= \frac{1}{T^2} E\left[\int_0^T \alpha(u) \beta(u) du \int_0^T \alpha(v) \beta(v - \tau_0) dv\right] \\ &= \frac{1}{T^2} \iint_{00}^{TT} E(\alpha(u) \beta(u) \alpha(v) \beta(v - \tau_0)) du dv \end{aligned} \quad (25)$$

If α and β are jointly Gaussian, then

$$\begin{aligned} E[\alpha(u) \beta(u) \alpha(v) \beta(v - \tau_0)] &= R_{\alpha\beta}(u - v + \tau_0) R_{\beta\alpha}(u - v) \\ &\quad + R_{\alpha\alpha}(u - v) R_{\beta\beta}(u - v + \tau_0) \end{aligned} \quad (26)$$

If $u - v = \eta$, then Eq. (25) can be rewritten as

$$E[\hat{R}_m(\tau_0)\hat{R}(0)] = \frac{1}{T} \int_{-T}^T \left(1 - \frac{|\eta|}{T}\right) \left(R_{\alpha\alpha}(\eta) R_{\beta\beta}(\eta + \tau_0) - R_{\alpha\beta}(\eta) R_{\alpha\beta}(\eta + \tau_0)\right) d\eta \quad (27)$$

If $\hat{R}_m(\tau_0)$ and $\hat{R}(0)$ are computed by polarity coincidence method, then

$$\hat{R}(0) = \hat{r}_c(0) = \frac{1}{T} \int_0^T \text{Sgn } \alpha(t) \text{Sgn } \beta(t) dt \quad (28)$$

$$\text{and } \hat{R}_m(\tau_0) = r_c(\tau_0) = \frac{1}{T} \int_0^T \text{Sgn } \alpha(t) \text{Sgn } \beta(t-\tau_0) dt \quad (28)$$

So $E[r_c(\tau_0)\hat{r}_c(0)]$ is given by:

$$E[\hat{r}_c(\tau_0)\hat{r}_c(0)] = \frac{1}{T^2} \int_0^T \int_0^T E[\text{Sgn } \alpha(u) \text{Sgn } \beta(u) \text{Sgn } \alpha(v) \text{Sgn } \beta(v-\tau_0)] du dv \quad (29)$$

$$\text{If } yy_1 = \text{Sgn } \alpha(u)$$

$$yy_2 = \text{Sgn } \beta(u)$$

$$yy_3 = \text{Sgn } \alpha(v)$$

$$yy_4 = \text{Sgn } \beta(v-\tau_0)$$

then, using Eq. (12) and Eq. (A3-12), one can obtain the expression for the fourth-product moment $E[yy_1 yy_2 yy_3 yy_4]$ (similar to Eq. (14)),

$$E[yy_1 yy_2 yy_3 yy_4] = 1 + \frac{4}{\pi^2} [VI_1 + VI_2 + VI_3 + VI_4] \quad (30)$$

where VI_1 , VI_2 , VI_3 , and VI_4 are given in Appendix 3 (Equations A3-8,9,10,11).

Substituting Eq. (30) in Eq. (29), one obtains

$$\begin{aligned} E[\hat{r}_c(\tau_0)\hat{r}_c(0)] &= \frac{1}{T} \int_{-T}^T \left(1 - \frac{|n|}{T}\right) E[yy_1 yy_2 yy_3 yy_4] dn \\ &= 1 + \left(\frac{4}{\pi^2}\right) \frac{1}{T} \int_{-T}^T \left(1 - \frac{|n|}{T}\right) (VI_1 + VI_2 + VI_3 + VI_4) dn \quad (31) \end{aligned}$$

Substitution of Eqs. (27) and (31) in Eq. (23), one obtains the variance of the slope estimator $[\hat{S}_m]$ for both analog and polarity-coincidence methods and using Eqs. (3,5,8,17,20,22,23) and Eqs. (9) and (10), one can compare the two estimators on the basis of signal to noise ratio.

Functional Form of Doppler Spectrum

In the report (12/11/77 - 1/11/78), the asymmetrical Doppler spectrum, at small antenna offsets from boresight, was represented by Gram-Charlier expansion. This expansion has been modified here in such a way that the first four central moments of the expansion are identical to that of the spectrum derived from the combined antenna pattern and angular dependence. The modified Gram-Charlier expansion of the spectrum $S(f)$ is written as

$$S(f) = \frac{1}{\sqrt{2\pi m_2}} \left(\exp \left(-\frac{1}{2} \left(\frac{f-m_1}{\sqrt{m_2}} \right)^2 \right) \right) \left[1 + \frac{\left(s_k - \frac{3m_1^3}{m_2^{1.5}} \right)}{3!} H_3 \left(\frac{f-m_1}{\sqrt{m_2}} \right) \right. \\ \left. + \left(E_x + \frac{14m_1^4}{m_2^2} - \frac{12m_1^2}{m_2} \right) H_4 \left(\frac{f-m_1}{\sqrt{m_2}} \right) + \frac{s_k^2}{2!3!3!} H_6 \left(\frac{f-m_1}{\sqrt{m_2}} \right) \right] \quad (32)$$

The autocorrelation function $R(\tau)$ of the spectrum $S(f)$ defined as

$$R(\tau) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S(\omega) e^{j\omega\tau} d\omega \quad (33)$$

is obtained by manipulation as:

$$R(\tau) = \left(\exp \left(jm_1 \tau - \frac{m_2 \tau^2}{2} \right) \right) \left[1 - j \left(\frac{s_k m_2^{1.5}}{6} - \frac{3m_1^3}{6} \right) \tau^3 + \right. \\ \left. \left((E_x) + \frac{14m_1^4}{m_2^2} - \frac{12m_1^2}{m_2} \right) \frac{m_2^2 \tau^4}{24} - \frac{s_k^2 m_2^3}{72} \tau^6 \right] \quad (34)$$

The first derivative of Eq. (34), which will be used in alignment accuracy analysis is given as:

$$\begin{aligned} \dot{R}(\tau) = & \left(\exp(j m_1 \tau - \frac{m_2 \tau^2}{2}) \right) \left[j m_1 - m_2 \tau - j \left(\frac{s_K m_2^{1.5}}{2} - \frac{3}{2} m_1^3 \right) \tau^2 \right. \\ & + \left(\frac{m_1 s_K m_2^{1.5}}{6} + \frac{(E_x) m_2^2}{6} - \frac{m_1^4}{2} \right) \tau^3 \\ & + j \left(\frac{(E_x) m_2^2 m_1}{24} + \frac{s_K m_2^{2.5}}{6} \right) \tau^4 - \left(\frac{(E_x) m_2^3}{24} + \frac{s_K^2 m_2^3}{12} \right) \tau^5 \\ & \left. - j \frac{m_1 s_K^2 m_2^3}{72} \tau^6 + \frac{s_K^2 m_2^4}{72} \tau^7 \right] \end{aligned} \quad (35)$$

Accuracy Analysis of the Correlation Method of Alignment

In the report 12/11/77-1/11/78, it was shown that the spectral moments are functions of tropolink-parameters such as wind-speed, antenna beam width, angular dependence, transmitted frequency, the distance between the transmitter and the receiver, etc. The antenna elevation angle is assumed zero. The variation in the link parameters cause these coupled spectral moments to change accordingly. So simply varying these moments will mimic the different link conditions, the net effect of which will result in changed moments. In Table 1, with $m_1=0$, the effect of fade rate under different 'skew' (asymmetric) conditions is shown. Although with different 'skew' conditions, m_1 should correspondingly be non-zero, but in calculations, first m_1 is assumed zero (Table 1) and then for a given m_2 , m_1 is varied (Table 2). For a given skew coefficient, the difference in m_1 can account for the 'residual clock-offset'. From Table 1, it could be observed that if the spectral width (or the fade rate) increases by a factor of 2, the variance decreases by the same factor. The theoretical explanation is given

in Appendix 2. The clipping loss, defined as in Eq. (10), is in the range of -1.5 to -3 db. In Table 3, the case of the symmetrical Doppler spectrum is considered. In Table 2 and 3, it should be noted that in some cases the clipping loss could result in the clipping gain. It would be of significant interest to identify those cases a priori. From Tables 1, 2, and 3 it can be concluded that the variance decreases by the same factor as the averaging time is increased.

If the slope is estimated from the correlation function (Eq. 18), then it could be observed from the equations 23, 27 and 31, that the variance of the slope in both cases will decrease as the spectral width (or fade rate) is increased which is different than that of the slope estimation by True and hard-limiter method (report 11/12/77-1/11/78) where the variance increases in proportion to the fade rate. So it can be concluded that the slope estimation by Eq. (18) is 'better' than the estimation by True and hard-limiter methods.

References

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Appendix 1

In this Appendix the expression for P_4 (quadrivariate normal orthant probability) is derived. For 4 correlated normal variables, with zero mean and unit variance, the probability that all 4 variables will be positive is

$$P_4 = \int_0^{\infty} \int_0^{\infty} \int_0^{\infty} \int_0^{\infty} \phi(x_1, x_2, x_3, x_4) dx_1 dx_2 dx_3 dx_4 \quad (A1-1)$$

Where $\phi(x_1, x_2, x_3, x_4)$ is the joint normal density of the Gaussian variables x_1, x_2, x_3, x_4 of zero mean and covariance matrix (c) given by equation (13). Since this matrix is symmetrical about the diagonal, only the upper-half is written as:

$$c = \begin{bmatrix} 1 & a & b & c \\ & 1 & d & b \\ & & 1 & a \\ & & & 1 \end{bmatrix} = \begin{bmatrix} \rho_{12} & \rho_{13} & \rho_{14} \\ & 1 & \rho_{23} & \rho_{24} \\ & & 1 & \rho_{34} \\ & & & 1 \end{bmatrix} \quad (A1-2)$$

where

$$\begin{aligned} a &= r(\tau) \\ b &= \rho(n) \\ c &= r(\tau+n) \\ d &= r(\tau-n) \end{aligned} \quad (A1-3)$$

The inverse of the covariance matrix (c) is given by:

$$c^{-1} = \begin{bmatrix} c_{11} & c_{12} & c_{13} & c_{14} \\ & c_{22} & c_{23} & c_{24} \\ & & c_{33} & c_{34} \\ & & & c_{44} \end{bmatrix} \quad (A1-4)$$

where

$$\begin{aligned}
 c_{11} &= c_{44} = \frac{1-a^2-d^2-b^2+2abd}{|C^{-1}|} \\
 c_{12} &= c_{34} = -\frac{(a-a^3-bd+acd+ab^2-bc)}{|C^{-1}|} \\
 c_{13} &= c_{24} = \frac{ad-a^2b-b+ac+b^3-bcd}{|C^{-1}|} \\
 c_{14} &= -\frac{a^2d-2ab+c+b^2d-cd^2}{|C^{-1}|} \\
 c_{22} &= c_{33} = \frac{1-a^2-b^2-c^2+2abc}{|C^{-1}|} \\
 c_{23} &= -\frac{d-2ab+a^2c+b^2c-c^2d}{|C^{-1}|}
 \end{aligned} \tag{A1-5}$$

and $|C^{-1}| = 1-2a^2-d^2+4abd-2b^2+a^4-2a^2cd-2a^2b^2+4abc+b^4-2b^2cd-c^2+c^2d^2$.

The method followed to compute P_4 is the application of the general method given by Plackett [1954]. Let the correlation coefficients be denoted by ρ_{ij} (Eq. A1-2), then since the density function obeys the relation:

$$\frac{\partial \phi}{\partial \rho_{ij}} = \frac{\partial^2 \phi}{\partial x_i \partial x_j}$$

we have, for example

$$\begin{aligned}
 \frac{\partial P_4}{\partial \rho_{12}} &= \int_0^\infty \int_0^\infty \left[\int_0^\infty \int_0^\infty \frac{\partial^2 \phi}{\partial x_1 \partial x_2} dx_1 dx_2 \right] dx_3 dx_4 \\
 &= \frac{1}{2\pi(1-\rho_{12})^{\frac{1}{2}}} P_2(c_{33}, c_{34}, c_{44})
 \end{aligned} \tag{A1-6}$$

There will be 6 such equations, one for each correlation coefficient.
Equation (A1-6) is a special case of Eq. (6) given by Plackett [1954].

Since

$$P_2[c_{33}, c_{34}, c_{44}] = \frac{1}{4} + \frac{1}{2\pi} \arcsin \left(\frac{-c_{34}}{(c_{33}c_{44})^{1/2}} \right) \quad (\text{A1-7})$$

combining Eq. (A1-6) and (A1-7), we have

$$\frac{\partial P_4}{\partial \rho_{12}} = \frac{1}{8\pi(1-\rho_{12}^2)} \left[1 + \frac{2}{\pi} \arcsin \left[\frac{-c_{34}}{(c_{33}c_{44})^{1/2}} \right] \right] \quad (\text{A1-8})$$

There are six such equations, upon combining them we obtain

$$P_4 = P_4[k_0] + \frac{1}{8\pi} \sum_{i,j} \rho_{ij} \left[1 + \frac{2}{\pi} \arcsin \left(\frac{-c_{pq}}{(c_{pp}c_{qq})^{1/2}} \right) \right] \frac{d\lambda_{ij}}{\sqrt{1-\lambda_{ij}^2}} \quad (\text{A1-9})$$

where (p, q, i, j) is a permutation of $(1, 2, 3, 4)$.

If $P_4[k_0]$ is the normal orthant probability of four Gaussian variables of zero mean, unit variance and covariance matrix

$$\begin{bmatrix} 1 & 0 & 1 & 0 \\ & 1 & 0 & 1 \\ & & 1 & 0 \\ & & & 1 \end{bmatrix} \quad (\text{A1-10})$$

then

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$$\begin{aligned}
 P_4 = P_4[k_0] &+ \frac{1}{8\pi} \left[2 \int_0^a \left[1 + \frac{2}{\pi} \arcsin \frac{(-c_{34})}{(c_{33}c_{44})^{1/2}} \right] \frac{d\lambda_{12}}{\sqrt{1-\lambda_{12}^2}} \right. \\
 &+ 2 \int_1^b \left[1 + \frac{2}{\pi} \arcsin \frac{(-c_{24})}{(c_{22}c_{44})^{1/2}} \right] \frac{d\lambda_{13}}{\sqrt{1-\lambda_{13}^2}} \\
 &+ \int_0^c \left[1 + \frac{2}{\pi} \arcsin \frac{(-c_{23})}{(c_{22}c_{33})^{1/2}} \right] \frac{d\lambda_{14}}{\sqrt{1-\lambda_{14}^2}} \\
 &\left. + \int_0^d \left[1 + \frac{2}{\pi} \arcsin \frac{(-c_{14})}{(c_{11}c_{44})^{1/2}} \right] \frac{d\lambda_{23}}{\sqrt{1-\lambda_{23}^2}} \right] \quad (A1-11)
 \end{aligned}$$

$P_4[k_0]$ can be computed from Cheng [1968] e.g.

$$P_4[k_0] = \phi(0,1) \text{ of Cheng} = \frac{1}{4} \quad (A1-12)$$

Combining (A1-11) and (A1-12), one can obtain

$$P_4 = \frac{1}{8} + \frac{1}{8\pi} [2\sin^{-1}a + 2\sin^{-1}b + \sin^{-1}c + \sin^{-1}d] + \frac{1}{4\pi^2} \text{FPM} \quad (A1-13)$$

where

$$\begin{aligned}
 \text{FPM} = & \left[2 \int_0^a \arcsin \left[\frac{-c_{34}}{(c_{33}c_{44})^{1/2}} \right] \frac{d\lambda_{12}}{\sqrt{1-\lambda_{12}^2}} \right] + \left[2 \int_1^b \arcsin \left[\frac{-c_{24}}{(c_{22}c_{44})^{1/2}} \right] \frac{d\lambda_{13}}{\sqrt{1-\lambda_{13}^2}} \right] \\
 & + \left[\int_0^c \arcsin \left[\frac{-c_{23}}{(c_{22}c_{33})^{1/2}} \right] \frac{d\lambda_{14}}{\sqrt{1-\lambda_{14}^2}} \right] + \left[\int_0^d \arcsin \left[\frac{-c_{14}}{(c_{11}c_{44})^{1/2}} \right] \frac{d\lambda_{23}}{\sqrt{1-\lambda_{23}^2}} \right] \quad (A1-14)
 \end{aligned}$$

Substitution of c_y from (A1-5) into (A1-14), one obtains:

$$FPM = 2I_1 + 2I_2 + I_3 + I_4$$

where

$$I_1 = \int_0^a \frac{dA}{\sqrt{1-A^2}} \arcsin \left[\frac{(A-A^3-BD+ACD+AB^2-BC)}{(1-A^2-B^2-D^2+2ABD)^{\frac{1}{2}}(1-A^2-B^2-C^2+2ABC)^{\frac{1}{2}}} \right]$$

$$I_2 = \int_1^b \frac{dB}{\sqrt{1-B^2}} \arcsin \left[\frac{(B+A^2B-AD-AC-B^3+BCD)}{(1-A^2-B^2-D^2+2ABD)^{\frac{1}{2}}(1-A^2-B^2-C^2+2ABC)^{\frac{1}{2}}} \right]$$

$$I_3 = \int_0^c \frac{dC}{\sqrt{1-C^2}} \arcsin \left[\frac{(D-2AB+A^2C+B^2C-C^2D)}{(1-A^2-B^2-C^2+2ABC)} \right]$$

and

$$I_4 = \int_0^d \frac{dD}{\sqrt{1-D^2}} \arcsin \left[\frac{(A^2D-2AB+C+B^2D-CD^2)}{(1-A^2-B^2-D^2+2ABD)} \right]$$

Since A, B, C, D are interrelated variables as in Eq. (A1-3), so variation in one results in the variation of another. So the above integrals I_1 , I_2 ,

I_3 , I_4 can be changed as:

$$I_1 = \int_0^{\tau} \frac{\dot{r}(t') dt'}{\sqrt{1-r^2(t')}} \arcsin \left[\frac{(r(t')-r^3(t')-br(t'-n)+r(t')r(t'+n)r(t'-n))}{+r(t')^2-br(t'+n))} \right] \quad (A1-15)$$

$$\left[\frac{(1-r^2(t')-b^2-r^2(t'-n)+2r(t')(b)r(t'-n))^{\frac{1}{2}} \times (1-r^2(t')-b^2-r^2(t'+n)+2r(t')(b)r(t'+n))^{\frac{1}{2}}}{(1-r^2(t')-b^2-r^2(t'-n)+2r(t')(b)r(t'-n))^{\frac{1}{2}} \times (1-r^2(t')-b^2-r^2(t'+n)+2r(t')(b)r(t'+n))^{\frac{1}{2}}} \right]$$

$$I_2 = \int_0^n \frac{\dot{\rho}(n') dn'}{\sqrt{1-\rho^2(n')}} \arcsin \left[\frac{(\rho(n')+a^2\rho(n')-a(r(\tau-n')+r(\tau+n'))-\rho^3(n'))}{+\rho(n')r(\tau+n')r(\tau-n'))} \right] \quad (A1-16)$$

$$\left[\frac{(1-a^2-\rho^2(n')-r^2(\tau-n')+2a\rho(n')r(\tau-n'))^{\frac{1}{2}}}{(1-a^2-\rho^2(n')-r^2(\tau+n')+2a\rho(n')r(\tau+n'))^{\frac{1}{2}}} \right]$$

The integrals I_3 and I_4 are computed using the chain rule:

$$dC = \left. \frac{dr(\tau'+n)}{d\tau'} d\tau' \right|_{n=\text{constant}} + \left. \frac{dr(\tau+n')}{dn'} dn' \right|_{\tau=\text{constant}}$$

$$\text{and } dD = \left. \frac{dr(\tau'-n)}{d\tau'} d\tau' \right|_{n=\text{constant}} - \left. \frac{dr(\tau-n')}{dn'} dn' \right|_{\tau=\text{constant}}$$

So

$$I_3 = \int_0^{\tau+n'} \frac{\dot{r}(t') dt'}{\sqrt{1-r^2(t')}} \arcsin \left[\frac{r(t'-2n)-2r(t'-n)b+r^2(t'-n)r(t') + b^2 r(t')-r^2(t')r(t'-2n))}{(1-r^2(t'-n)-b^2-r^2(t')+2r(t'-n)br(t'))} \right] \\ + \int_0^{\tau+n} \frac{\dot{r}(n') dn'}{\sqrt{1-r^2(n')}} \arcsin \left[\frac{r(2\tau-n')-2ap(n'-\tau)+a^2 r(n')+r(n')}{\rho^2(n'-\tau)-r^2(n')r(2\tau-n'))} \right] \quad (A1-17)$$

Similarly

$$I_4 = \int_0^{\tau-n} \frac{\dot{r}(t') dt'}{\sqrt{1-r^2(t')}} \arcsin \left[\frac{(r(t'+2n)-2br(t'+n)+r(t')(r^2(n+t')+b^2) - r^2(t')r(t'+2n))}{(1-r^2(t'+n)-b^2-r^2(t')+2br(t')r(t'+n))} \right] \\ + \int_0^{\tau-n} \frac{\dot{r}(n') dn'}{\sqrt{1-r^2(n')}} \arcsin \left[\frac{(r(2\tau-n')-2ap(\tau-n')+(a^2+\rho^2(\tau-n'))r(n') - r^2(n')r(2\tau-n'))}{(1-a^2-\rho^2(\tau-n')-r^2(n')+2ap(\tau-n')r(n'))} \right] \quad (A1-18)$$

Substituting equations (A1-15,16,17,18) in Eq. (A1-13), one obtains

$$P_4 = \frac{1}{16} \left[1 + \frac{2}{\pi} [2\sin^{-1}a + 2\sin^{-1}b + \sin^{-1}c + \sin^{-1}d] \right. \\ \left. + \frac{4}{\pi^2} [2I_1 + 2I_2 + I_3 + I_4] \right] \quad (A1-19)$$

Appendix 2

In this Appendix, the effect of the change in the width of the Doppler spectrum on the variance of the two estimators of the quadrature correlation function at all lags is considered. Since the variance is inversely proportional to the averaging time and the spectral width is an indicator of fade rate, so the effect of the different fade rates (due to different wind velocities, antenna patterns, etc.) on the required averaging time is estimated by assuming that the total received power is the same under various fading conditions, e.g. if the Doppler spectrum increases in width by a factor of k , then

$$R_{\alpha\beta}(\tau) \rightarrow R_{\alpha\beta}(k\tau)$$

and

$$R_{\alpha\alpha}(\tau) \rightarrow R_{\alpha\alpha}(k\tau)$$

(A2-1)

where $R_{\alpha\beta}(\tau)$ is the quadrature cross-correlation function and $R_{\alpha\alpha}(\tau)$ is the quadrature auto-correlation function of quadrature processes $\alpha(t)$ and $\beta(t)$. In deriving Eq. (A2-1), the well known Fourier transform scaling relationship is utilized, e.g.

$$\frac{1}{a} F\left(\frac{W}{a}\right) \leftrightarrow f(a\tau)$$

Substituting Eq. (A2-1) into Eqs. (3) and (15), one obtains, for large T ,

$$\text{Var}[\hat{R}_{\alpha\beta}(\tau)] = \frac{1}{KT} \int_{-T}^T (R_{\alpha\alpha}^2(n) + R_{\alpha\beta}(n+\tau)R_{\beta\alpha}(n-\tau))dn \quad (\text{A2-2})$$

and

$$\text{Var}[\hat{r}_c(\tau)] = \frac{1}{kT} \int_{-T}^T E[y_1 y_2 y_3 y_4] dn \quad (\text{A2-3})$$

where $\hat{r}_c(\tau)$ is the clipped quadrature correlation function.

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From Eqs. (A2-2 and A2-3), it is evident that the variance of both estimators decreases by the same factor as the width of the Doppler spectrum increases. In other words, an increase in the width of the Doppler spectrum requires the decrease in the averaging time by the same factor.

Appendix 3

In this Appendix, the expression for P_4 (quadrivariate normal orthant probability) is derived which is required in computing the variance of the slope estimator by polarity coincidence method. The method is described in detail in Appendix 1 and has been applied in computing the variance of the quadrature correlation function estimator by polarity coincidence method.

In the case under discussion, the covariance matrix (c) is given by

$$c = \begin{bmatrix} 1 & 0 & a & b \\ & 1 & c & d \\ & & 1 & g \\ & & & 1 \end{bmatrix} \quad (A3-1)$$

$$\begin{aligned} \text{where } a &= \rho(n) \\ b &= r(n+\tau_0) \\ c &= -r(n) = r(-n) \\ d &= \rho(n+\tau_0) \\ g &= r(\tau_0) \end{aligned} \quad (A3-2)$$

The inverse of the covariance matrix (c) is given by:

$$c^{-1} = \begin{bmatrix} c_{11} & 0 & c_{13} & c_{14} \\ & c_{22} & c_{23} & c_{24} \\ & & c_{33} & c_{34} \\ & & & c_{44} \end{bmatrix} \quad (A3-3)$$

where

$$c_{11} = \frac{1-g^2-c^2+2cdg-d^2}{|c^{-1}|}$$

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$$\begin{aligned}
 c_{13} &= \frac{-a+bg+ad^2-bcd}{|c^{-1}|} \\
 c_{14} &= \frac{ag-b-acd+bc^2}{|c^{-1}|} \\
 c_{22} &= \frac{1-a^2-b^2-g^2+2abg}{|c^{-1}|} \\
 c_{23} &= -\frac{c-gd+abd-b^2c}{|c^{-1}|} \\
 c_{24} &= \frac{cg-d+a^2d-abc}{|c^{-1}|} \\
 c_{33} &= \frac{1-b^2-d^2}{|c^{-1}|} \\
 c_{34} &= -\frac{g-cd-ab}{|c^{-1}|} \\
 c_{44} &= \frac{1-a^2-c^2}{|c^{-1}|}
 \end{aligned} \tag{A3-4}$$

and

$$|c^{-1}| = 1-g^2-c^2+2cgd-d^2+(-a^2+2abg+a^2d^2-2abcd)-b^2+b^2c^2$$

Similar to Eq. (A1-13), one can write

$$P_4 = \frac{1}{8} + \frac{1}{8\pi} [\text{arc sin } a + \text{arc sin } b + \text{arc sin } c + \text{arc sin } d + \text{arc sin } g] + \frac{1}{4\pi^2} \text{FPM} \tag{A3-5}$$

where

$$\begin{aligned}
 \text{FPM} &= \left[\int_1^a \text{arc sin} \left[\frac{-c_{24}}{(c_{22}c_{44})^{1/2}} \right] \frac{d\lambda_{13}}{\sqrt{1-\lambda_{13}^2}} \right] + \left[\int_0^b \text{arc sin} \left[\frac{-c_{23}}{(c_{22}c_{33})^{1/2}} \right] \frac{d\lambda_{14}}{\sqrt{1-\lambda_{14}^2}} \right] \\
 &+ \left[\int_0^c \text{arc sin} \left[\frac{-c_{14}}{(c_{11}c_{44})^{1/2}} \right] \frac{d\lambda_{23}}{\sqrt{1-\lambda_{23}^2}} \right] + \left[\int_1^d \text{arc sin} \left[\frac{-c_{13}}{(c_{11}c_{33})^{1/2}} \right] \frac{d\lambda_{24}}{\sqrt{1-\lambda_{24}^2}} \right]
 \end{aligned} \tag{A3-6}$$

Substituting c_{ij} from (A3-4) into (A3-6), one obtains

$$FPM = VI_1 + VI_2 + VI_3 + VI_4 \quad (A3-7)$$

where

$$VI_1 = \int_1^a \frac{dA}{\sqrt{1-A^2}} \arcsin \left| \frac{(D+ABC-CG-A^2D)}{(1-A^2-B^2-G^2+2ABG)^{\frac{1}{2}}(1-A^2-C^2)^{\frac{1}{2}}} \right|$$

$$VI_2 = \int_0^b \frac{dB}{\sqrt{1-B^2}} \arcsin \left| \frac{(C-GD+ABD-B^2C)}{(1-A^2-B^2-G^2+2ABG)^{\frac{1}{2}}(1-B^2-D^2)^{\frac{1}{2}}} \right|$$

$$VI_3 = \int_0^c \frac{dC}{\sqrt{1-C^2}} \arcsin \left| \frac{(B+ACD-AG-BC^2)}{(1-G^2-C^2-D^2+2CDG)^{\frac{1}{2}}(1-A^2-C^2)^{\frac{1}{2}}} \right|$$

$$VI_4 = \int_1^d \frac{dD}{\sqrt{1-D^2}} \arcsin \left| \frac{(A+BCD-BG-AD^2)}{(1-G^2-C^2-D^2+2CDG)^{\frac{1}{2}}(1-B^2-D^2)^{\frac{1}{2}}} \right|$$

Since A,B,C,D are interrelated variables as in Eq. (A3-2), the variation in one results in the variation of another. So the above integrals I_1 , I_2 , I_3 , and I_4 can be changed as

$$VI_1 = \int_1^n \frac{\dot{\rho}(n')dn'}{\sqrt{1-\rho^2(n')}} \arcsin \left[\frac{(\rho(n'+\tau_0)-\rho(n')r(n'+\tau_0)r(n')+g\rho(n')-\rho^2(n')\rho(n'+\tau_0))}{((1-\rho^2(n')-r^2(n'+\tau_0)-g^2+2g\rho(n')r(n'+\tau_0))^{\frac{1}{2}}(1-\rho^2(n')-r^2(n'))^{\frac{1}{2}})} \right] \quad (A3-8)$$

$$VI_2 = \int_0^{\tau_0+n} \frac{\dot{r}(t')dt'}{\sqrt{1-r^2(t')}} \arcsin \left[\frac{(c-r(t'-n)\rho(t')+ar(t')\rho(t')-r^2(t')c)}{((1-a^2-r^2(t')-r^2(t'-n)+2ar(t')r(t'-n))^{\frac{1}{2}}(1-r^2(t')-\rho^2(t'))^{\frac{1}{2}})} \right] \\ + \int_0^{\tau_0+n} \frac{\dot{r}(n')dn'}{\sqrt{1-r^2(n')}} \arcsin \left[\frac{(-r(n'-\tau_0)-g\rho(n')+\rho(n'-\tau_0)r(n')\rho(n')+r^2(n')r(n'-\tau_0))}{((1-\rho^2(n'-\tau_0)-r^2(n')-g^2+2\rho(n'-\tau_0)r(n')g)^{\frac{1}{2}}(1-r^2(n')-\rho^2(n'))^{\frac{1}{2}})} \right] \quad (A3-9)$$

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$$VI_3 = - \int_0^{+n} \frac{\dot{r}(n') dn'}{\sqrt{1-r^2(n')}} \arcsin \left| \frac{(r(n'+\tau_0)-\rho(n')r(n')\rho(n'+\tau_0)-\rho(n')g-r^2(n')r(n'+\tau_0))}{((1-g^2-r^2(n')-\rho^2(n'+\tau_0)-2gr(n')\rho(n'+\tau_0))^{\frac{1}{2}})(1-r^2(n')-\rho^2(n'))^{\frac{1}{2}})} \right| \quad (A3-10)$$

$$VI_4 = \int_0^{\tau_0+n} \frac{\dot{\rho}(t') dt'}{\sqrt{1-\rho^2(t')}} \arcsin \left| \frac{(a+r(t')c\rho(t')-r(t')r(t'-n)-a\rho^2(t'))}{((1-c^2-\rho^2(t')-r^2(t'-n)+2c\rho(t')r(t'-n))^{\frac{1}{2}})(1-r^2(t')-\rho^2(t'))^{\frac{1}{2}})} \right|$$

$$+ \int_0^{\tau_0+n} \frac{\dot{\rho}(n') dn'}{\sqrt{1-\rho^2(n')}} \arcsin \left| \frac{(\rho(n'-\tau_0)-r(n')r(n'-\tau_0)\rho(n')-r(n')g-\rho(n'-\tau_0)\rho^2(n'))}{((1-g^2-r^2(n'-\tau_0)-\rho^2(n')-2gr(n'-\tau_0)\rho(n'))^{\frac{1}{2}})(1-r^2(n')-\rho^2(n'))^{\frac{1}{2}})} \right| \quad (A3-11)$$

Substituting equations (A3-8,9,10,11) in Eq. (A3-5), one obtains

$$P_4 = \frac{1}{16} \left[1 + \frac{2}{\pi} [\arcsin a + \sin \arccos b + \arcsin c + \arcsin d + \arcsin g] \right]$$

$$+ \frac{4}{\pi^2} [I_1 + I_2 + I_3 + I_4] \quad (A3-12)$$

Part III

A New Complex Gaussian Troposcatter Channel Simulator

At the present there is an increasing interest in the use of communication channel simulators for the checkout and comparative testing of modems. The cost and difficulty of obtaining operating links and the impossibility of comparative modem testing with the same link (at the same time) have been the factors stimulating this interest. Another, less obvious factor, is the need for a simulator as a "standard channel" to be used in the calibration of channel measurement equipment. The only meaningful checkout of channel measurement equipment (and also modems) prior to on-the-air use involves a synthetic channel which can be set up to simulate known channel parameters.

The mathematical model around which the simulators, in the past, have been built is that of a band-limited uniform tapped delay line [Baghdady, 1969; Bello, 1969; Daly, 1964; Fitting, 1967; Klein, 1967; Hancock, 1966; Freudberg, 1965; Rose, 1966; Stein, 1965; Walker, 1966] with each tap undergoing independent fading according to complex Gaussian statistics. The choice of complex Gaussian statistics can be justified on the grounds of central limit theorem as applied to the contribution at each path delay caused by scattering from a large number of atmospheric refractive index inhomogeneities. Actually, it has been noted [Bennet, 1948] that as few as six sine waves with independently fluctuating random phase will give a fluctuating result whose envelope closely follows Rayleigh statistics and whose phase is uniformly distributed. Statistically, the problem of determining the resultant of a set of random phasors is familiar as that of two-dimensional random walk. A more basic view of the two-dimensional random walk follows

by regarding the constituent independently varying phasors to be resolved into quadrature components. The simultaneous values of the quadrature components of the resultant at any instant are then readily shown to be uncorrelated, and, by the central limit theorem, to independently approach Gaussian variates as the number of steps in the random walk (the number of contributing phasors) becomes large.

With regard to experiments, the short-term fading statistics of a received signal have often been observed to have a nearly Rayleigh distribution which is consistent with the complex Gaussian assumptions. It is well known that the received signal is the linear combination of the signals from the different paths (delays). The distinguishability of several paths implies that they are resolvable, i.e., their electric lengths are sufficiently different. More concretely, the situation usually described as 'multipath' is one in which more than one propagation path exists, with the time differentials between the several paths larger than the period of the highest 'significant' frequency components in the modulation of the transmitted signal.

Shortcomings of the Past Simulators. The above mentioned simulators, though, give adequate performance in the laboratory, but they have several shortcomings from a meteorological and communication point of view.

1. Aligned antennas are assumed. Normally antennas are less than optimally aligned, so the Doppler spectrum from the single tap is not necessarily symmetrical and the quadrature components therefore become statistically dependent processes.
2. Zero clock-offset is assumed. Normally there is clock-offset between the transmitting and receiving clocks, e.g., the clock frequencies are not identical. Also there is some drift present

in the clock, though the clocks used are usually fairly stable (1 in 10^{10} Hz). Both the effects, clock-offset and drifts of the clocks, should be included in the effective channel simulator.

3. There is no provision to include the average horizontal drift of the scatterers--a factor which is vitally important to a realistic model.
4. There is no provision to include the angular dependence of the scatterer which assigns a signal strength to the scatter as a function of the scattering angle.
5. There is no provision to directly include meteorological conditions. These simulators are not responsive to a change from turbulent conditions (fast fading) characterized by a Rayleigh envelope to an anisotropic medium (slow-fading) usually characterized by a Rician distributed envelope except by "brute force" method, i.e., forcing one scatterer to be very strong.
6. Airplane free common volume is assumed. Normally, while link is in operation, there are airplanes flying in the common-volume. Reflections from airplanes can be viewed as the reflections from a "large scatter" resulting in the 'v' shape instantaneous frequency of the return complex signal.

In the following description, a channel model is proposed in which the above-mentioned effects are included. This discussion will deal with the simulation of single tap in the tapped delay line model (or single path) of the channel. This model, in general, can be used for other channels (UHF, VHF, ionospheric channel, underwater channel, etc.). This channel model is divided into three parts as shown in Fig. 1.

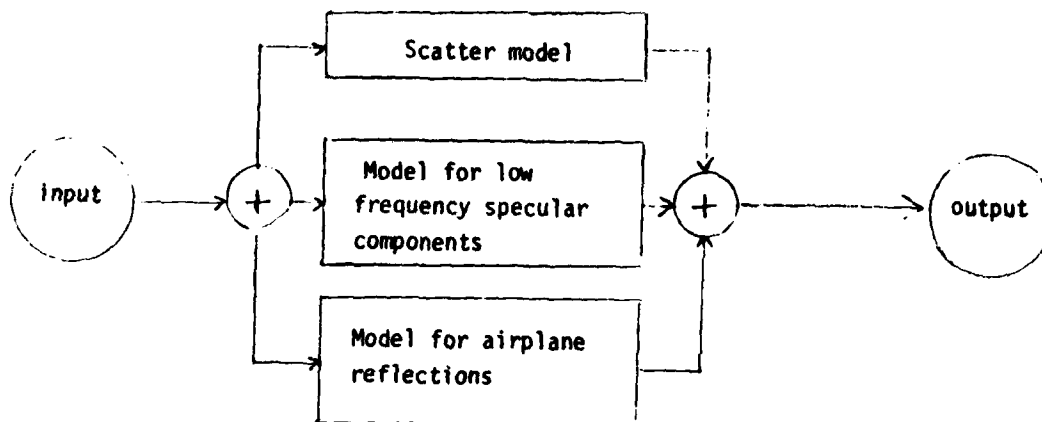


Fig. 1

Models for airplane reflections will be the topic of further research and will be discussed at a later date. In the following, a model for diffuse scattering such as would occur from a single turbulent atmospheric layer is discussed, and then a model for low frequency specular component is described.

Scatter Model. As mentioned earlier, the models in the past utilized the symmetrical Doppler spectrum with no clock-offset, the tap-gains were varied to represent the different fading conditions. In the proposed model, the other parameters of the tropo-channel are also included which mimic the effects of the clock-offset, misalignment of the antennas, spatial non-stationarities in the drifting scattering layers, etc. In the following discussion, only single tap-delay line (single path) model is considered for simplistic reasons.

In the report 12/11/77 - 1/11/78, it was shown that the received Doppler spectrum $S(f)$ could be represented as:

$$S(f) = \text{COF}[G_T(f) A(f) G_R(f)] \quad (1)$$

where $G_T(f)$ and $G_R(f)$ are assumed Gaussian transmitter and receiver gain functions, $A(f)$ is the Angular Dependence function and COF is receiver 'clock-offset shift' operator. If there is no offset then received spectrum $S_0(f)$ can be represented as:

$$S_0(f) = G_T(f) A(f) G_R(f) \quad (2)$$

$S(f)$ and $S_0(f)$ are related by: $S_0(f) = S(f+\Delta f)$ where Δf is the clock-offset.

If the transmitter and receiver antenna beam-widths are identical with Gaussian beam width "measure" of σ_0^2 and the transmitter is assumed pointed at the horizon over the great circle but the receiver antenna is azimuthally pointed at an angle α_a , then one obtains:

$$G_T(f) = G_0 \exp\left(-\frac{f^2}{\sigma_0^2}\right) \quad (3)$$

$$\text{and } G_R(f) = G_0 \exp\left(-\frac{(f-f_a)^2}{\sigma_0^2}\right) \quad (4)$$

In equations (3) and (4), (Merril, 1968)

$$f_a = -\frac{2}{\lambda} v \alpha_a \quad \text{and} \quad \sigma_0^2 \equiv \frac{(d\alpha_3)^2}{(4\ln 2)} \quad (5)$$

Where v is the cross-path wind velocity, λ is the wave length of the transmitted CW signal, $2d$ is the length of the link axis, α_3 is -3db beam-width of a single antenna pattern and G_0 is constant.

Similarly angular dependence function is written as: (Merril, 1968)

$$A(f) = A_0 [f^2 + k_0^2]^{-11/6} \quad (6)$$

Where A_0 is constant depending upon the cross-path wind-velocity and intrinsic

strength of the scatterers and k_0 is given by:

$$k_0 = \frac{z_0}{d} \cdot \frac{2v}{\lambda} \quad (7)$$

z_0 is the height of the layer in consideration. For small antenna misalignment and narrow beams, $A(f)$ can be approximated by:

$$A(f) = A_0 [f^2 + k_0^2]^{-2} \quad (8)$$

Simulation of the Complex Channel: In this discussion, the tropo-channel will be simulated and then the clock-offset effect will be included in the overall simulation. It is well known that a process with spectral density $S_0(f)$ can be generated by passing white noise through the filter of transfer function $H(S)$ such that:

$$|H(j\omega)|^2 = S_0(f) = G_T(f) A(f) G_R(f) \quad (9)$$

Similarly, if

$$\begin{aligned} |H_T(j\omega)|^2 &= G_T(f) \\ |H_A(j\omega)|^2 &= A(f) \end{aligned} \quad (10)$$

$$\text{and} \quad |H_R(j\omega)|^2 = G_R(f)$$

then one can write

$$|H_T(j\omega)|^2 = |H_T(j\omega)|^2 |H_A(j\omega)|^2 |H_R(j\omega)|^2 \quad (11)$$

Using equations (3), (4), (6), (10) and (11), one obtains

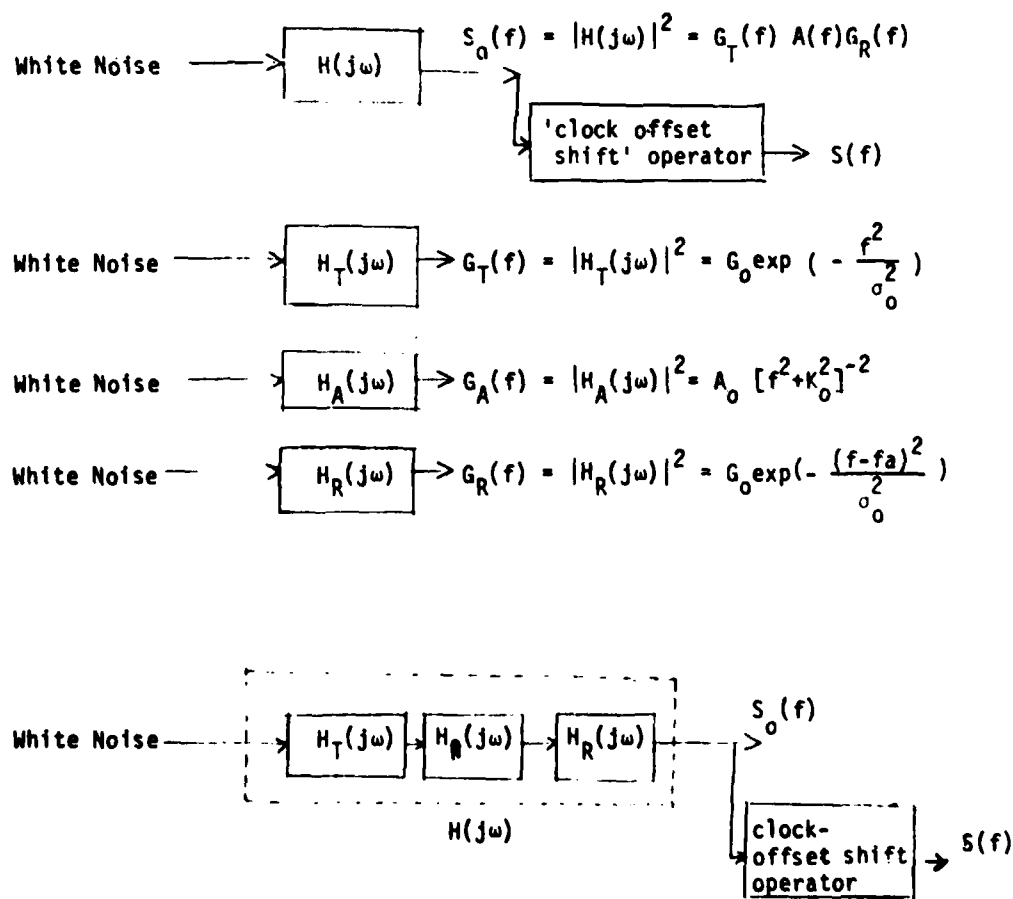
$$|H_T(j\omega)|^2 = G_0 \exp \left[- \frac{f^2}{\sigma_0^2} \right] \quad (12)$$

$$|H_R(j\omega)|^2 = G_0 \exp \left[-\frac{(f-f_a)^2}{\sigma_0^2} \right] \quad (13)$$

$$|H_A(j\omega)|^2 = A_0 [f^2 + K_0^2]^{-2} \quad (14)$$

$$\text{and } |H(j\omega)|^2 = G_0^2 A_0 [f^2 + K_0^2]^{-2} \exp \left[-\frac{f^2 + (f-f_a)^2}{\sigma_0^2} \right] \quad (15)$$

eqs. (12) to (15) are shown in block-diagram form in Figure 2.



Block diagram forms of equations 12 - 15

Fig. 2

Some Properties of $|H(j\omega)|^2$. Here it is shown that the asymmetric filter $H(S)$ causes the asymmetrical Doppler spectrum which results in the cross-correlation function of the output (of the filter) quadrature processes (inphase and quadrature). It has been mentioned in earlier reports that the misalignment of the antennas and the clock-offset both result in the output quadrature processes being correlated. It will be shown in the later section that by decomposing $H(S)$ in even and odd components, one can directly relate the odd part of $H(S)$ to the cross-correlation of the output processes.

The even and odd parts of $|H(j\omega)|^2$ are defined as:

$$Ev|H(j\omega)|^2 = \text{Even part of } |H(j\omega)|^2 = (1/2)|H(j\omega)|^2 + (1/2)|H(-j\omega)|^2 \quad (16)$$

$$\text{and } OD|H(j\omega)|^2 = \text{Odd part of } |H(j\omega)|^2 = (1/2)|H(j\omega)|^2 - (1/2)|H(-j\omega)|^2$$

From equation (16), it is clear that the odd part of $|H(j\omega)|^2$ will be zero if $|H(j\omega)|^2$ is symmetrical about the origin (i.e., zero clock-offset and aligned antennas). If the input process to the filter $H(j\omega)$ is symmetrical about the origin with spectral density $U(\omega)$ [constant if white noise] then the auto-spectrum, $S_{\alpha\alpha}(\omega)$ and cross-spectrum $S_{\alpha\beta}(\omega)$ of the output quadrature processes $\alpha(t)$ and $\beta(t)$, are given by [Frank, 1969]

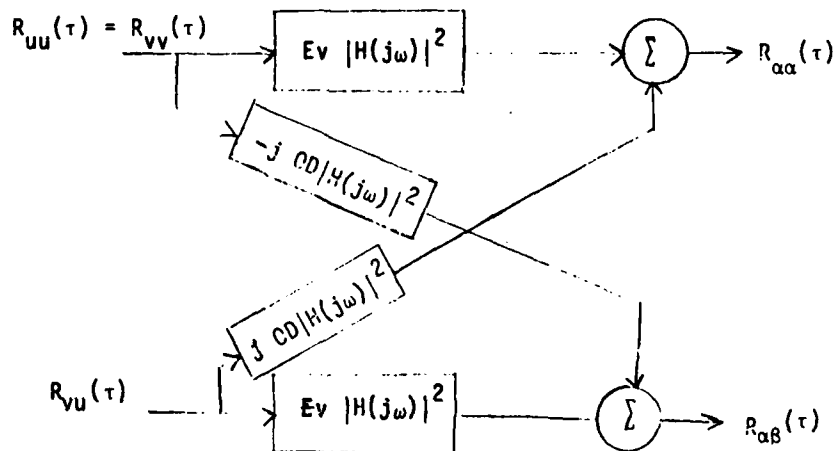
$$S_{\alpha\alpha}(\omega) = S_{\beta\beta}(\omega) = Ev |H(j\omega)|^2 U(\omega) \quad (17)$$

$$\text{and } S_{\alpha\beta}(\omega) = -j OD|H(j\omega)|^2 U(\omega)$$

If input complex white noise is correlated and is of asymmetric spectrum (which results in auto spectral density $S_{uu}(\omega)$ and cross-spectral density $S_{uv}(\omega)$ of the input quadrature processes $u(t)$ and $v(t)$), then [Frank, 1969]

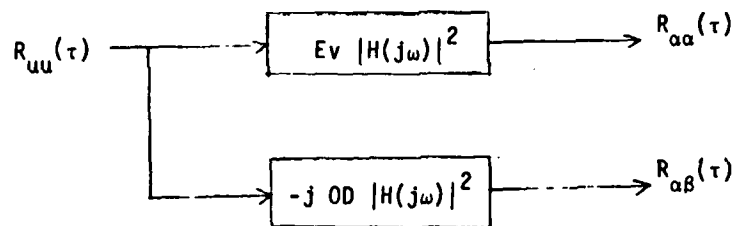
$$\begin{aligned}
 S_{\alpha\alpha}(\omega) &= E_v |H(j\omega)|^2 S_{uu}(\omega) + j OD |H(j\omega)|^2 S_{vu}(\omega) \\
 S_{\alpha\beta}(\omega) &= -j OD |H(j\omega)|^2 S_{uu}(\omega) + E_v |H(j\omega)|^2 S_{vu}(\omega)
 \end{aligned}
 \quad (18)$$

These linear transformations on auto-correlation and cross-correlation functions, not on the signals themselves, are shown symbolically in Fig. 3(a) and (b). Fig. 3(b) results from 3(a) when input process is white uncorrelated noise. These diagrams make it clear how non-symmetrical band-pass characteristics of the channel introduces correlation between inphase and quadrature components of the band pass process. Now a method is discussed to obtain $H(S)$ from $|H(j\omega)|^2$ and the eigen-values of $H(S)$ are then related to the tropo-link parameters. First, a general complex model will be derived, and this model is then applied to the tropo-channel.



Low-pass equivalent of transformation relating input and output into correlation functions when input Gaussian noise is correlated

Fig. 3(a)



Low pass equivalent of transformation when input Gaussian noise is uncorrelated

Fig. 3(b)

A Complex Model. The motivation to simulate the band-pass channel (filter) into the low-pass equivalents (or complex filter) stems from the fact that the inherent low-pass nature and speed limitations of the analog computer make its operation in the low-pass mode highly desirable. In the following discussion, we shall show how a band-pass system can be simulated equivalently by two inter-connected low-pass systems.

Consider the band-pass filter with transfer function $R(f)$ and its low-pass equivalent $H(f)$. The input signal is given by:

$$\begin{aligned} x_1(t) &= \text{Re} \left[z_1(t) \exp(j2\pi f_0 t) \right] \\ &= (1/2)z_1(t) \exp(j2\pi f_0 t) + (1/2)z_1^*(t) \exp(-j2\pi f_0 t) \end{aligned} \quad (19)$$

$$\text{where } z_1(t) = u(t) - jv(t) \quad (20)$$

$$\text{Hence } X_1(f) = (1/2)Z_1(f-f_0) + (1/2)Z_1^*(-f-f_0) \quad (21)$$

The transfer function of the filter can be expressed in terms of $H(f)$ in a similar manner,

$$R(f) = (1/2)H(f-f_0) + (1/2)H^*(-f-f_0) \quad (22)$$

$$\text{where } H(f-f_0) = (1+\text{sgn}f) R(f) \quad (23)$$

Now for the output signal,

$$\begin{aligned}
 X_2(f) &= R(f) X_1(f) \\
 &= (1/4)H(f-f_0) Z_1(f-f_0) + (1/4)H^*(f-f_0) Z_1^*(-f-f_0) \\
 &= (1/2)Z_2(f-f_0) + (1/2)Z_2^*(-f-f_0)
 \end{aligned} \tag{24}$$

From eq. (24), it is evident that

$$Z_2(f) = (1/2)H(f) Z_1(f)$$

$$\text{or } z_2(t) = (1/2)h(t) \otimes z_1(t) \tag{25}$$

Output $z_2(t)$ in complex form is written as

$$z_2(t) = \alpha(t) - j\beta(t) \tag{26}$$

Now we decompose $h(t)$ into its real and imaginary parts so that

$$(1/2)h(t) = h_1(t) + j h_2(t), \tag{27}$$

where $h_1(t)$ and $h_2(t)$ are real and can be considered as impulse responses of real low-pass filters. The corresponding transfer functions are given by

$$H_1(f) = 1/4 [H(f) + H^*(-f)] \tag{28}$$

$$\text{and } H_2(f) = (1/4j)[H(f) - H^*(-f)]$$

i.e., by the symmetrical and anti-symmetrical parts, respectively. It follows that $H_1(f)$ and $H_2(f)$ are related directly to the parts of the band-pass function $R(f)$ which are symmetrical and anti-symmetrical, respectively about the center frequency f_0 . $H_1(f)$ and $H_2(f)$ are related to even and odd parts of $|H(j\omega)|^2$ (eq. 16) by:

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$$E|H(j\omega)|^2 = |H_1(f)|^2 + |H_2(f)|^2$$

$$\text{and } OD|H(j\omega)|^2 = jH_2(f) H_1^*(f) - jH_2^*(f) H_1(f) \quad (29)$$

From examples (20), (25), (26) and (27), one writes:

$$\begin{aligned} z_2(t) &= \alpha(t) j\beta(t) = (u-jv) \otimes (h_1 + j h_2) \\ &= (u \otimes h_1 + v \otimes h_2) - j [v \otimes h_1 - u \otimes h_2] \end{aligned} \quad (30)$$

or in frequency domain

$$\begin{aligned} \alpha(f) &= U(f) H_1(f) + V(f) H_2(f) \\ \beta(f) &= V(f) H_1(f) - U(f) H_2(f) \end{aligned} \quad (31)$$

So by this approach, a single input, single output band pass filtering operation is replaced by a double input, double output low-pass filtering operation which characterizes separately the transfer of inphase and quadrature components. The equivalent low-pass system is represented schematically by the lattice arrangement shown in Fig. 4.

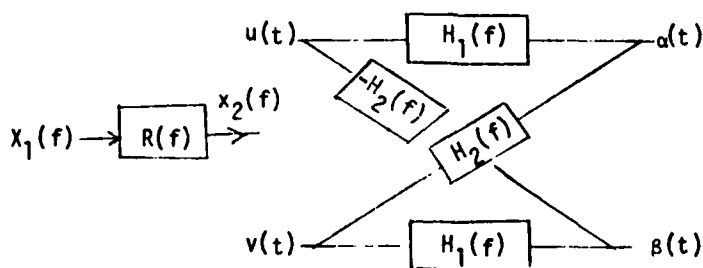


Fig. 4. Equivalent low-pass filtering.

A cascade combination of band-pass filters can be handled directly in terms of low-pass equivalents as shown in Fig. (5). [Frank, 1969]

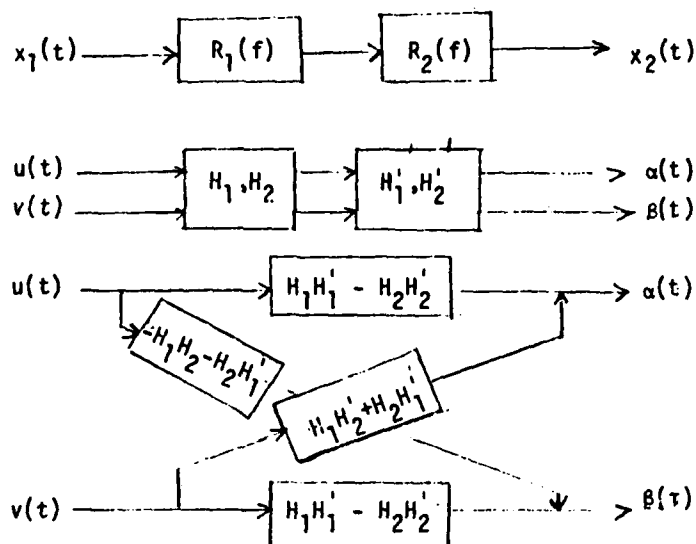


Fig. 5 Low-pass equivalents of cascade combination of band-pass filters

We should note that if a band-pass filter exhibits symmetry about f_0 , then $H_2(f) = 0$ and there is no cross-coupling between the in-phase and quadrature components. This is the kind of transmission required to avoid amplitude to phase modulation conversion on a band pass signal.

In Appendix 1, a first order and second order complex system is derived which when driven by the complex white noise outputs the complex narrow-band process. In the following discussion, the eigen-values of the complex system are estimated from the first three spectral moments of the output Doppler spectrum (or complex spectrum). First, the individual blocks $[H_T(j\omega), H_R(j\omega), A(j\omega)]$ will be simulated, then the combined system $[H(j\omega)]$, simulating the tropo-link, will be synthesized.

Simulation of the Transmitter Gain Function (equation 12). Simulation of the transmitter gain function is the approximation of the Gaussian-magnitude approximation given by equation 12. Given the ideal Gaussian magnitude characteristic:

$$|H_T(j\omega)| = \exp(-\omega^2/2)$$

where the unit bandwidth has been selected as a matter of convenience, let us find an all-pole transfer function

$$H(s) = \frac{1}{G_n(s, \bar{a})}$$

where $G_n(s, \bar{a})$ is a real Hurwitz polynomial with a coefficient vector $\bar{a}$, whose spectrum approximates $|H_T(j\omega)|^2$ best in the Taylor sense at the origin, defining the error function as:

$$E(s, \bar{a}) = G_n(s, \bar{a}) G_n(-s, \bar{a}) - \exp[-s^2]$$

We readily see that the optimum $G_n(s, \bar{a}) G_n(-s, \bar{a})$ corresponds to the truncated Maclaurin expansion:

$$G_n(s, \bar{a}) G_n(-s, \bar{a}) = 1 - s^2 + \frac{s^4}{2!} - \frac{s^6}{3!} + \dots + (-1)^n \frac{s^{2n}}{n!}$$

Finally, $G_n(s, \bar{a})$ is isolated from the product $G_n(s, \bar{a}) G_n(-s, \bar{a})$ by factoring and retaining only LHP factors. Based upon the above analysis and renormalizing w.r.t. unit 3-db bandwidth, Dishal [1959] and Humphry [1970] obtained the pole locations up to the order $n = 10$. Increasing n , obviously, would give better 'matching' characteristics. But only the case for $n = 2$ would give symmetrical spectrum. So a simple second-order complex model is considered and it is suggested that further work is necessary to obtain better 'matching' characteristics without distorting the symmetry. By considering the pole locations at $a + jb$, where $a = -1.2841100\sigma_0$ and $b = \pm j.531895\sigma_0$ (σ_0 is 3-db bandwidth given by equation 5), one obtains the second-order complex

model (see Appendix 1, equation A1-7) as in Fig. 6

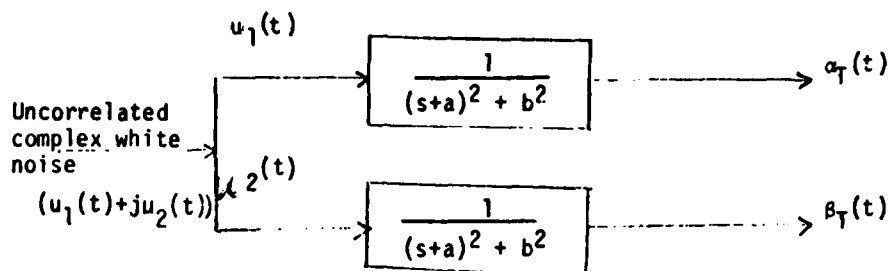


Fig. 6

Simulation of Angular Dependence (equation 14). The angular dependence function is given by equation 14, e.g.,

$$|H_A(j\omega)|^2 = \frac{A_0}{(f^2 + K_0^2)^2}$$

where K_0 is given by equation (7). From the above equation, one obtains

$$H_A(s) = \frac{2\pi^{3/2} \sqrt{A_0/K_0}}{(s + 2\pi K_0)^2} \quad (32)$$

The simulation of $H_A(s)$ in quadrature form can be obtained by assuming $a = c$, $b = d = 0$ in equation A1-3 of Appendix 1, and then substituting in equation A1-7, one obtains:

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$$H_1(s) = H_A(s) \text{ and } H_2(s) = 0$$

and

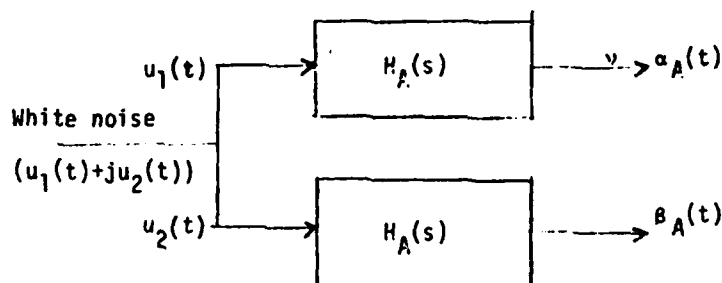


Fig. 7

Simulation of Receiver Antenna Gain Function (equation 13). Two methods are discussed to simulate the receiver antenna gain function given by equation (13). It is noted that the transmitter and receiver antennas are identical except that the receiver antenna is not pointed at the great-circle but is offset by an amount α_a corresponding to a Doppler offset of f_a .

Method 1. In this method, first the spectrum $|H_T(j\omega)|^2 |H_A(j\omega)|^2$ ($G_T(f) A(f)$) would be shifted left by an amount f_a . The shifted spectrum would, then, be multiplied by $|H_R(j\omega)|^2$ with $f_a = 0$ (it is equivalent to $|H_T(j\omega)|^2$). Finally, the spectrum would be shifted right by f_a . The overall simulation is shown in Fig. 8. The spectral-shift unit shown in Fig. 9 performs a single side-band modulation directly at baseband. The mathematics of the signal processing is outlined below.

[Manske, 1968]

Consider the case in which a narrow-band signal, expressed in the form of $x(t) = I(t) \cos \omega_c t + Q(t) \sin \omega_c t$ (33)

is passed through a filter having characteristics centered at a frequency

ω_1 , which is different from ω_c by a small amount ($\omega_1 - \omega_c = f_a$). The signal can then be expressed in terms of ω_1 , rather than ω_c in the form of:

$$X(t) = \alpha_R(t) \cos \omega_1 t + \beta_R(t) \sin \omega_1 t \quad (34)$$

From equations (33) and (34), one can obtain (see Fig. 9)

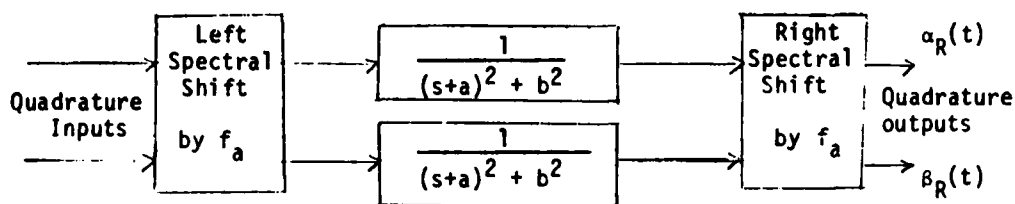


Fig. 8. Simulation of receiving antenna gain function (method 1)

$$\alpha(t) = I(t) \cos f_a t - Q(t) \sin f_a t \quad (35)$$

and
$$\beta(t) = I(t) \sin f_a t + Q(t) \cos f_a t$$

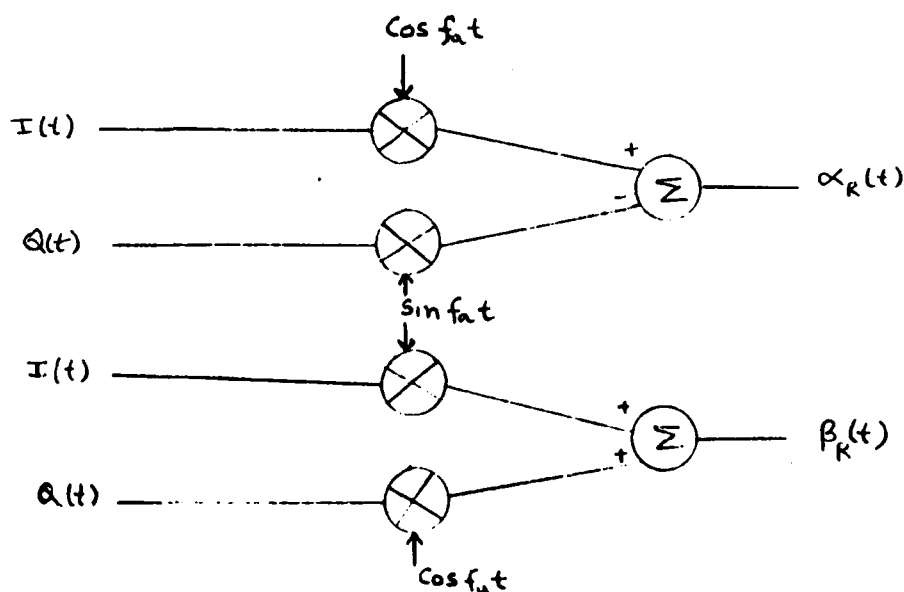


Fig. 9. Spectral-shift unit

Method 2. In this method, the receiver antenna gain function is simulated by the complex model as described in Appendix 1. If in equation (A1-3), one assumes $a = c = a_0$, $d = 0$, $b = b_0$, then one can obtain from equation A1-7 that: [See Fig. 10]

$$H_1(s) = \frac{1}{(s+a_0)^2 + b_0^2} \quad \text{and} \quad H_2(s) = \frac{b_0}{(s+a_0)[(s+a_0)^2 + b_0^2]} \quad (36)$$

and from equation (A1-11 and 12), one obtains

$$\dot{R}(0) = -j \frac{b_0}{2} \quad \text{and} \quad \ddot{R}(0) = - \left(a_0^2 + \frac{b_0^2}{2} \right) \quad (37)$$

Since the first moment (gives the average frequency of the spectrum) is equal to $\frac{1}{2\pi} \dot{R}(0)$ and the second central moment (related to $\ddot{R}(0)$) gives the 3-db band-width, one derives from equation 37, that

$$f_a = \left(-\frac{b_0}{2} \right) \frac{1}{2\pi} \quad \text{or} \quad b_0 = -4\pi f_a \quad (38)$$

Similarly,

$$\sigma_0^2 = \frac{1}{(2\pi)^2} \left[a_0^2 + \frac{b_0^2}{2} - \frac{b_0^2}{4} \right] = \frac{1}{(2\pi)^2} \left[a_0^2 + \frac{b_0^2}{4} \right] \quad (39)$$

From equations (38) and (39), one obtains

$$a^2 = (2\pi)^2 \left[\sigma_0^2 - f_a^2 \right] \quad (40)$$

For $a^2 > 0$, one requires that $|\sigma_0| > |f_a|$ which is intuitively obvious.

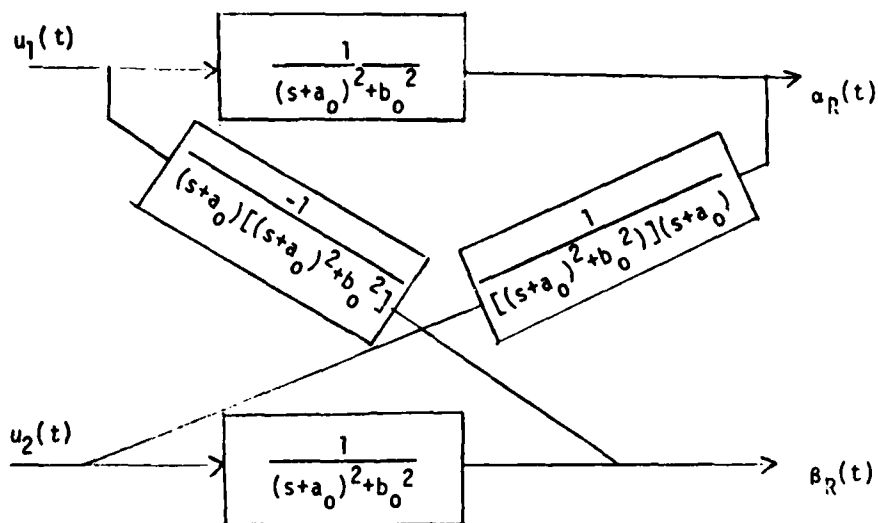


Fig. 10. Simulation of receiver antenna gain function (method 2)

Simulation of Clock-Offset [COF' operator, Equation 1]. It has been mentioned earlier that the clock-offset (Δf) merely shifts the spectrum either to the left or right depending upon the positive or negative clock-offset. The spectral-shift unit can be implemented similar to that shown in Fig. 9, by substituting Δf instead of f_a . Drift in the clock can be incorporated by varying Δf accordingly.

Over-all Tropo-Channel Transfer Function (Diffuse Scattering)

The overall channel transfer function in quadrature form is the cascaded simulation of various terms in equation 1 and is shown in Fig. 11.

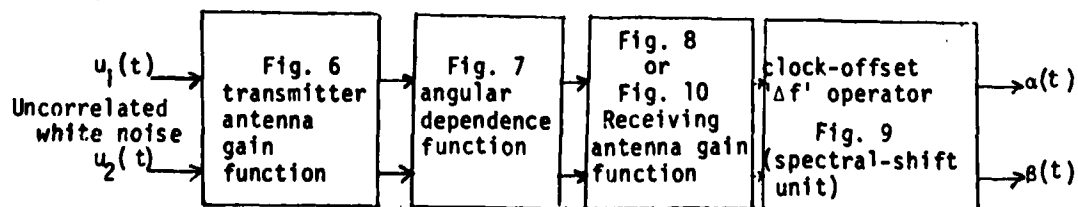


Fig. 11. Over-all tropo-channel simulator for diffused scattering

Using Fig. 5, one can obtain the cascade form of the transfer functions of transmitter and receiver antenna gain functions (Figs. 6 & 10) and angular dependence function (Fig. 7) as shown in Fig. 12, where

$$H_{cr} = \frac{1}{[(s+a)^2+b^2][s+k]^2[(s+a_0)^2+b_0^2]} \quad (40)$$

and
$$H_{un} = \frac{1}{[(s+a)^2+b^2][s+k]^2(s+a_0)[(s+a_0)^2+b_0^2]}.$$

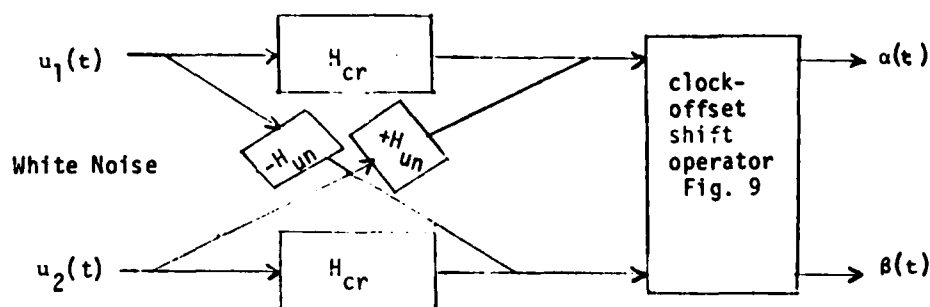


Fig. 12. Cascade form of over-all tropo channel simulator as shown in Fig. 11.

Models for Low Frequency Specular Components:

Duvoisin (1969) points out that in the presence of multiple scattering layers, the total Doppler power spectrum is the sum of the spectral contributions of the individual layers. The shape of the portion of the Doppler (power) spectrum of the received signal caused by a given horizontal scattering layer can therefore be shown to be controlled by the large numbers of scattering points drifting across the scattering volume and is functionally represented by the equation 2. Considering equations (3 to 7), Duvoisin (1969) concluded

that the narrow power-peaks near zero Doppler could be produced by any one of the following conditions:

1. Very low values of cross-path wind, V .
2. Very narrow beam antennas pointed down the great circle azimuth.
3. Very narrow angular dependence function.

Since neither conditions (1) or (2) were satisfied during most of the Barrow, et al. (1968) rake sounding experiments, the specular peaks, in most cases, were probably caused by condition 3. The angular dependence function used in Birkemeier, et al. (1968) to simulate the narrow specular peaks observed in the rake data was of the form

$$A_n(f) = \left[\frac{2\pi A_n \ell_z}{V} \right]^{-p} \left[f^2 + \frac{k_0^2}{A_n^2} \right]^{-p/2} \quad (41)$$

where k_0 is given by eq. 7, A_n is the degree of anisotropy, ℓ_z is the spatial correlation length in a vertical direction and p is $\frac{11}{3}$ for Tatar ski scattering.

For $A_n = 1$ (isotropic case) and $p = \frac{11}{3}$, equation 41 reduces to the equation 6, where

$$A_0 = \left[\frac{2\pi \ell_z}{V} \right]^{-11/3} \quad (42)$$

In eq. (41), the narrow-peaks can be obtained by either increasing p or increasing A_n . According to Duvoisin (1969), maintaining $A = 1$ and increasing p could produce sufficiently narrow specular peaks only when p was increased by almost two orders of magnitude. However, a power law of $p = 11/3$ and a degree of anisotropy of $A = 20$ produced peaks having the necessary Doppler bandwidth. So equation 41 can be rewritten as:

$$A_n(f) = A_0 A_n^{-11/3} \left[f^2 + \frac{k_0^2}{A_n^2} \right]^{-11/6} \quad (43)$$

Similar to equation 8, for small antenna misalignment, $A(f)$ can be approximated by:

$$A_n(f) = A_{on} [f^2 + k_{on}^2]^{-2} \quad (44)$$

$$\text{where } A_{on} = A_o A_n^{-11/3} \quad \text{and} \quad k_{on} = k_o/A_n \quad (45)$$

Now similar to equation 32 and Fig. 7, one can obtain

$$H_{An}(s) = \frac{2\pi^{3/2} \sqrt{(A_{on}/k_{on})}}{(s + 2\pi k_{on})^2} \quad (46)$$

and

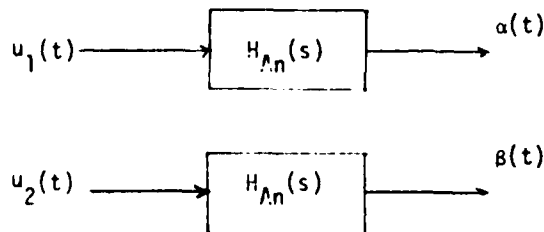


Fig. 13

Now similar to Fig. 11, one can obtain the overall tropo-channel simulator for low frequency specular components i.e.,

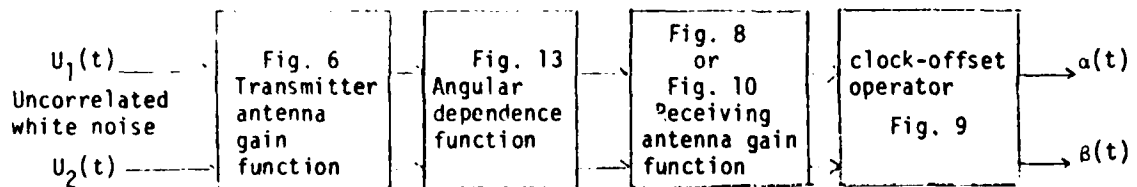


Fig. 14. Overall tropo-channel simulator for low-frequency specular components.

Overall tropo-channel transfer function can be obtained by combining Fig. (11) and (14) and is given in Fig. 15.

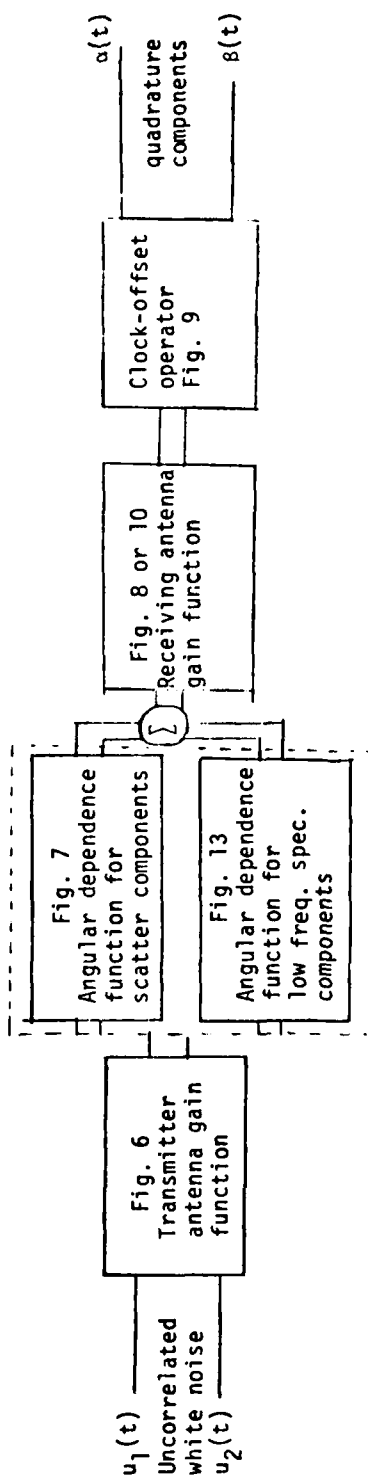


Fig. 15. Overall tropo-channel simulator for diffused and specular components

Model for the Average Drift of the Scatters:

In the simulation of the angular dependence function, symmetrical form of the function is assumed. Since we would expect on the average as many up-Doppler signals as down-Doppler signals and therefore, the average total doppler-shift for aligned antennas would be zero. If there are more scatters on one side of the great circle, it results in the average frequency (f_d) to be non-zero and the simulation of the angular dependence function (as shown in Fig. 7 and 13) would be modified. Using the results of the Appendix 1 (A1-15), one obtains:

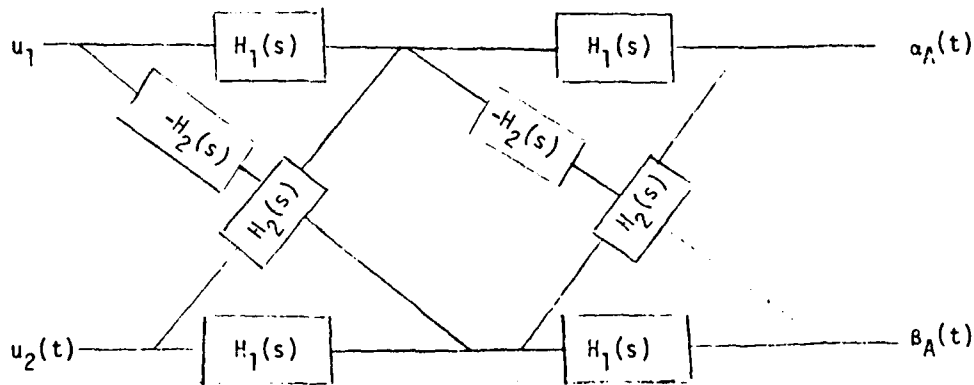


Fig. 16

where

$$H_1(s) = \left(2\pi^{3/2} \sqrt{A_{on}/k_{on}} \right) \frac{(s+2\pi k_{on})}{(s+2\pi k_{on})^2 + \Delta A^2}$$

$$H_2(s) = \left(2\pi^{3/2} \sqrt{A_{on}/k_{on}} \right) \frac{\Delta A}{(s+2\pi k_{on})^2 + \Delta A^2}$$

(46)

Now similar to equation (38), one can obtain:

$$\Delta A = 2\pi f_d \quad (47)$$

If the horizontal drift of the scattering is occurring in a deterministic or random fashion, then ΔA (the imaginary part of the complex pole $-\tilde{k} = 2\pi k_{on} + j\Delta A$) can be varied accordingly.

Examples:

In Fig. 12 a tropo-scatter channel simulator for diffused layer is shown. The poles of this channel transfer function are related to the tropo-link parameters. The variations in the pole locations would correspond to various link conditions, resulting in different spectral moments as described in earlier reports. In the following, for illustrative purposes, the pole locations $[-\tilde{k}_1, -\tilde{k}_2]$ of a 2nd-order complex system are varied to simulate various link conditions and their effect on the spectral properties of the output quadrature processes are tabulated in Table 1. The Doppler spectrum (eq. A1-10) and auto and cross-correlation functions (real and imaginary parts of the equation A1-9) are plotted in Fig. 17 and 18 for all the cases described in Table 1. Some of the conclusions from this analysis are summarized below.

1. For symmetrical spectrum with no clock-offset, the cross-correlation function is zero for all τ (case 1,2,7,8).
2. For symmetrical spectrum, the clock-offset (average freq.) is obtained by averaging the imaginary part of the complex poles (case 3,3c,6,6c).
3. Slope of the cross-correlation function at zero by $[\dot{R}_{\alpha\beta}(0)]$ is obtained from the equation A1-11, and is equivalent to the average frequency of the spectrum, so the sign of $\dot{R}_{\alpha\beta}(0)$ can be computed from the average frequency of the process (1st moment).

4. For asymmetrical spectrum, the slope, $\dot{R}_{\alpha\beta}(0)$, could be zero. This results due to the fact that the average frequency of the asymmetrical spectrum due to the 'asymmetry' alone is equal and opposite in sign of the clock-offset. So the net average frequency of the spectrum is the sum of these two effects and is therefore zero (case 4,4c,9).
5. The auto correlation function of the quadrature processes after hard-clipping decays faster than the case of the unclipped signals.
6. Hard-clipping of the processes does not change the zero-crossings in auto and cross-correlation function.

Case no.	Pole locations ($-k_1, -k_2$)	Spectrum symmetrical & asymmetrical	Average frequency	Cross-correlation $R_{\alpha\beta}(\tau)$ for all τ	Sense of $R_{\alpha\beta}(0)$
1	$-1+j, -1-j$	symmetrical	zero	zero	zero
2	$-1+2j, -1-2j$	symmetrical	zero	zero	zero
3	$-1, -1-j$	symmetrical	(-ve)	non-zero	(-)ve
3c	$-1, -1+j$	symmetrical	(+)ve	non-zero	(+)ve
4	$-2+2j, -1-j$	asymmetrical	zero	non-zero	zero
4c	$-2-2j, -1+j$	asymmetrical	zero	non-zero	zero
5	$-2, -1-j$	asymmetrical	(-)ve	non-zero	(-)ve
5c	$-2, -1+j$	asymmetrical	(+)ve	non-zero	(+)ve
6	$-2, -2-j$	symmetrical	(-)ve	non-zero	(-)ve
6c	$-2, -2+j$	symmetrical	(+)ve	non-zero	(+)ve
7	$-2+j, -2-j$	symmetrical	zero	zero	zero
8	$-2+2j, -2-2j$	symmetrical	zero	zero	zero
9	$-3+3j, -2-2j$	asymmetrical	zero	non-zero	zero
10	$-3+2j, -2-j$	asymmetrical	(+)ve	non-zero	(+)ve

Table 1. Summary of the Spectral Analysis of Output of the Complex System for Different Pole Locations.

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Appendix 1

In this Appendix, a new concept of complex modeling is introduced. This complex modeling (for complex processes or orthogonal processes) is shown to be a special form of modeling bivariate processes. The complex representation of narrow-band processes has already been described in earlier reports. Here the state variable description of these processes is considered and its application to modeling the scatter portion of the channel is described.

In this complex state variable description [Baggeroer, 1970], a complex random process is generated by exciting a linear system having a complex state variable description with a complex white noise. The only assumptions made are on $\tilde{u}(t)$ and $\tilde{x}(T_0)$ where $\tilde{u}(t)$ is the complex white noise and $\tilde{x}(T_0)$ is the complex initial conditions of the linear system which could be assumed to have the following state variable description.

$$\frac{d}{dt} \tilde{x}(t) = \hat{F}(t)\tilde{x}(t) + \hat{G}(t)\tilde{u}(t) \quad [\text{Linear State Equation}]$$

$$\tilde{z}(t) = \tilde{C}(t)\tilde{x}(t) \quad [\text{Linear Observation Equation}] \quad (\text{A1-1})$$

and
$$\tilde{z}(t) = \alpha(t) + j\beta(t)$$

where all the coefficient matrices may be complex.

With the assumptions on $\tilde{u}(t)$ and $\tilde{x}(T_0)$, one can obtain the covariance function $k_{\tilde{z}(\tau)}$ of the output [Baggeroer, 1970]. From $k_{\tilde{z}(\tau)}$, the complex spectrum of $\tilde{z}(t)$, $s_z(\omega)$, can be obtained. It can be shown that the complex output $\tilde{z}(t) = \alpha(t) + j\beta(t)$ satisfies the conditions of narrow-band processes (orthogonal processes).

In this way, a complex process can be generated. It will be shown below that the different coefficient matrices could result in different complex models which will mimic the different channel conditions resulting due to clock-offset, misaligned antennas, varying fade rate and spectral shape of the channel. In the description below, an example (2nd order system) illustrates how these conditions of the channel can be simulated. This example is described in [Baggeroer, 1970], but its application in channel modeling is pointed out here. Finally, 1st-order system is considered which would be useful in simulating angular dependence function.

Derivation of the Channel Transfer Function

Let us consider a second-order system. In the steady-state, it corresponds to a system with two poles. The pole locations $-\tilde{k}_1$ and $-\tilde{k}_2$ need not be complex conjugates of one another. For this system, the state and observation equations are:

$$\frac{d}{dt} \begin{bmatrix} \tilde{x}_1(t) \\ \tilde{x}_2(t) \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ -\tilde{k}_1\tilde{k}_2 & -(\tilde{k}_1+\tilde{k}_2) \end{bmatrix} \begin{bmatrix} \tilde{x}_1(t) \\ \tilde{x}_2(t) \end{bmatrix} + \begin{bmatrix} 0 \\ 1 \end{bmatrix} \tilde{u}(t)$$

$$\text{and } \tilde{z}(t) = [1:0] \begin{bmatrix} \tilde{x}_1(t) \\ \tilde{x}_2(t) \end{bmatrix}$$

By simple manipulations, it can be shown that the above equation corresponds to

$$\ddot{\tilde{z}}(t) = -\tilde{k}_1\tilde{k}_2 \tilde{z}(t) - (\tilde{k}_1+\tilde{k}_2) \dot{\tilde{z}}(t) + \tilde{u}(t) \quad (A1-2)$$

Let $\tilde{z}(t) = \alpha(t) + j\beta(t)$

$$\tilde{k}_1 = a + jb$$

$$\tilde{k}_2 = c + jd$$

$$A = ac - bd$$

$$B = bc + ad$$

$$E = a + c$$

$$F = b + d$$

(A1-3)

The above equation (A1-2) can be rewritten as:

$$\ddot{\alpha}(t) + j\ddot{\beta}(t) = -(E + jF)(\dot{\alpha}(t) + j\dot{\beta}(t)) -$$

(A1-4)

$$[A + jB] \cdot [\alpha(t) + j\beta(t)] + U_1(t) + jU_2(t)$$

Separating real and imaginary parts in Eq. (A1-4), one can obtain

$$\ddot{\alpha}(t) + E\dot{\alpha}(t) - F\dot{\beta}(t) + A\alpha(t) - B\beta(t) = U_1(t)$$

(A1-5)

$$\ddot{\beta}(t) + E\dot{\beta}(t) + F\dot{\alpha}(t) + B\alpha(t) + A\beta(t) = U_2(t)$$

Equation (A1-4) can be written in transfer function form. Let us assume that $U_1(t)$, $U_2(t)$, $\alpha(t)$, $\beta(t)$ are general input-output of a bivariate system, such that their Laplace Transforms can be taken. Then by simple manipulations, one can show that

$$\alpha(s) = \frac{(s^2 + ES + A)}{(s^2 + ES + A)^2 + (FS + B)^2} U_1(s) + \frac{(FS + B)}{(s^2 + ES + A)^2 + (FS + B)^2} U_2(s)$$

and

(A1-6)

$$\beta(s) = \frac{s^2 + ES + A}{(s^2 + ES + A)^2 + (FS + B)^2} U_2(s) - \frac{(FS + B)}{(s^2 + ES + A)^2 + (FS + B)^2} U_1(s)$$

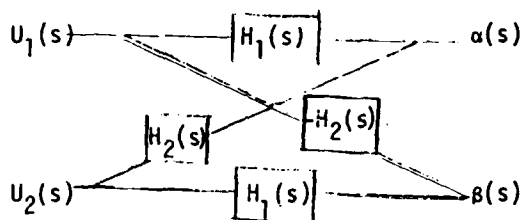
$$\text{If } H_1(s) = \frac{s^2 + ES + A}{(s^2 + ES + A)^2 + (FS + B)^2} \quad \text{and } H_2(s) = \frac{FS + B}{(s^2 + ES + A)^2 + (FS + B)^2} \quad (\text{A1-7})$$

then, equation (A1-6) can be written as:

$$\alpha(s) = H_1(s)U_1(s) + H_2(s)U_2(s) \quad (\text{A1-8})$$

and
$$\beta(s) = H_1(s)U_2(s) - H_2(s)U_1(s)$$

Equation (A1-8) can be written in block-diagram form as:



In the above block diagram, it can be observed that the forward transfer function $H_1(s)$ in both the paths are the same and the transfer function in the cross-paths $[H_2(s)]$ are equal but of opposite sign. The author points out here that this is the condition for modeling complex processes (orthogonal processes) in quadrature form.

If, in equations (A1-6) and (A1-8), $B=0$ and $F=0$, then $H_2(s)$ is zero. For this condition there is no cross-coupling between two quadrature components $\alpha(t)$ and $\beta(t)$, and these two processes are uncorrelated. As discussed earlier, zero correlation implies symmetrical spectrum with no clock-offset. If $B=0$ and $F=0$, then $a=c$ and $b=-d$ e.g., both the poles are mirror images of one another about σ -axis in $s(=\sigma+j\omega)$ plane.

For non-zero, B and F constants, $H_2(s)$ is not zero and $\alpha(t)$ and $\beta(t)$ processes are correlated. Since B and F are functions of the poles of the system (k_1, k_2) , that implies that the pole position determines the shape of the power spectral density of the complex process. Equation (A1-5) can easily be simulated on an analog computer. By varying pole locations $(\tilde{k}_1, \tilde{k}_2)$, one can obtain different spectrums. Also note that the differential equations involving $\alpha(t)$ and $\beta(t)$ in Eq. (A1-5) are coupled equations. So 'coupled' systems (e.g. non-ideal transformer) can be analyzed by this complex or phasor analysis.

The covariance function $k_z(t, t)$ of the output is given by [Baggeroer, 1970]: (P is the input noise power).

$$k_z(\tau) = \begin{cases} \frac{P}{(\tilde{k}_2 - \tilde{k}_1)^*} \left[\begin{aligned} & \left(\tilde{k}_2^* e^{\tilde{k}_1^* \tau} - \tilde{k}_1^* e^{\tilde{k}_2^* \tau} \right) \\ & + j \frac{\text{Im}[\tilde{k}_1 \tilde{k}_2]}{\text{Re}[\tilde{k}_1 + \tilde{k}_2]} \left(e^{\tilde{k}_1^* \tau} - e^{\tilde{k}_2^* \tau} \right) \end{aligned} \right] & \tau < 0 \\ \frac{P}{(\tilde{k}_2 - \tilde{k}_1)} \left[\begin{aligned} & \left(\tilde{k}_2 e^{-\tilde{k}_1 \tau} - \tilde{k}_1 e^{-\tilde{k}_2 \tau} \right) \\ & - j \frac{\text{Im}[\tilde{k}_1 \tilde{k}_2]}{\text{Re}[\tilde{k}_1 + \tilde{k}_2]} \left(e^{-\tilde{k}_1 \tau} - e^{-\tilde{k}_2 \tau} \right) \end{aligned} \right] & \tau > 0 \end{cases} \quad (\text{A1-9})$$

From e.g. (A-9), one can compute $s_z(\omega)$ as:

$$\begin{aligned} \text{Let } \tilde{A}_1 &= \frac{\tilde{k}_2 + \tilde{k}_1^*}{(\tilde{k}_2 - \tilde{k}_1)} \cdot \frac{\text{Re}[\tilde{k}_2]}{\text{Re}[\tilde{k}_1 + \tilde{k}_2]} \\ \tilde{A}_2 &= - \frac{\tilde{k}_1 + \tilde{k}_2^*}{(\tilde{k}_2 - \tilde{k}_1)} \cdot \frac{\text{Re}[\tilde{k}_1]}{\text{Re}[\tilde{k}_1 + \tilde{k}_2]} \end{aligned}$$

then

$$S_{\tilde{z}}(\omega) = 2P \left[\frac{\text{Re}[\tilde{A}_1]\text{Re}[\tilde{k}_1] + \text{Im}[\tilde{A}_1](\omega + \text{Im}[\tilde{k}_1])}{\text{Re}^2[\tilde{k}_1] + [\omega + \text{Im}(\tilde{k}_1)]^2} + \frac{\text{Re}[\tilde{A}_2]\text{Re}[\tilde{k}_2] + \text{Im}[\tilde{A}_2](\omega + \text{Im}[\tilde{k}_2])}{\text{Re}^2[\tilde{k}_2] + [\omega + \text{Im}(\tilde{k}_2)]^2} \right] \quad (\text{A1-10})$$

From e.g. (A1-9) one can obtain:

$$\dot{R}(0) = -j \left[\frac{bc+ad}{a+c} \right] = -j \frac{B}{E} \quad (\text{A1-11})$$

and

$$\ddot{R}(0) = - \frac{a(c^2+d^2)+c(a^2+b^2)}{a+c} = - \left(A + \frac{BF}{E} \right) \quad (\text{A1-12})$$

First-order Complex System:

The equations that describe the 1st-order system with one pole $(-\tilde{k})$ are:

$$\frac{d\tilde{x}(t)}{dt} = -\tilde{k} \tilde{x}(t) + \tilde{u}(t) \quad (\text{state equation}) \quad (\text{A1-13})$$

$$\text{and} \quad \tilde{z}(t) = \tilde{x}(t) \quad (\text{observation equation})$$

Following the steps similar to 2nd-order system, one can obtain by letting

$$\tilde{k} = a + jb \quad (\text{A1-14})$$

$$\text{and} \quad \tilde{z}(t) = \alpha(t) + j\beta(t)$$

D.84

that

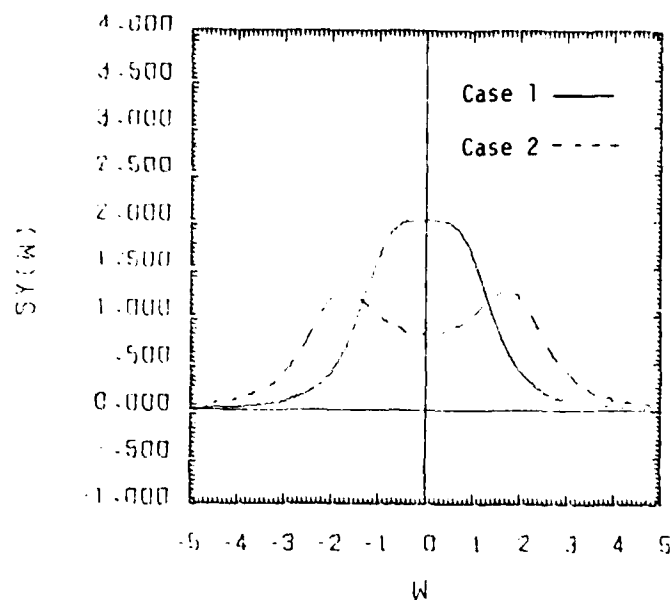
$$H_1(s) = \frac{s+a}{(s+a)^2+b^2} \quad \text{and} \quad H_2(s) = \frac{b}{(s+a)^2+b^2} \quad (\text{A1-15})$$

Similar to equations (A1-9) and (A1-10) one obtains [Baggeroer, 1970]

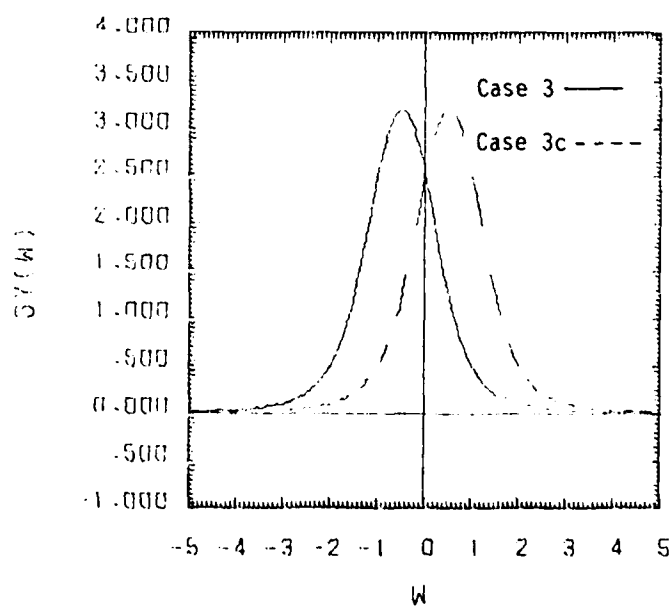
$$k_z^-(\tau) = p e^{-\text{Re}[\tilde{k}]\tau} e^{-j\text{Im}[\tilde{k}]\tau} \quad (\text{A1-16})$$

$$\text{and} \quad s_z^-(\omega) = \frac{2\text{Re}[\tilde{k}]p}{(\omega - \text{Im}[\tilde{k}])^2 + \text{Re}[\tilde{k}]^2} \quad (\text{A1-17})$$

The auto and cross spectra of $\alpha(t)$ and $\beta(t)$ can easily be found in terms of even and odd parts of $s_z^-(\omega)$. Note that the net effect of allowing a complex gain in e.g. (A1-13) is that the spectrum has a frequency shift equal to the imaginary part of the gain.



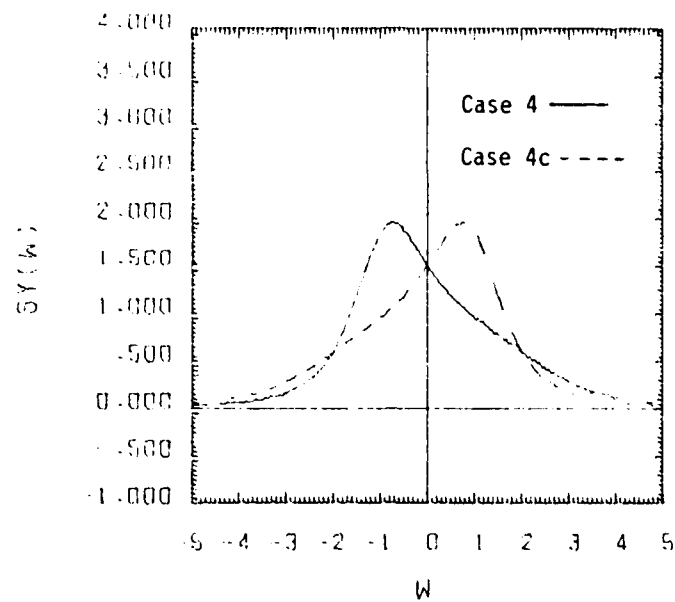
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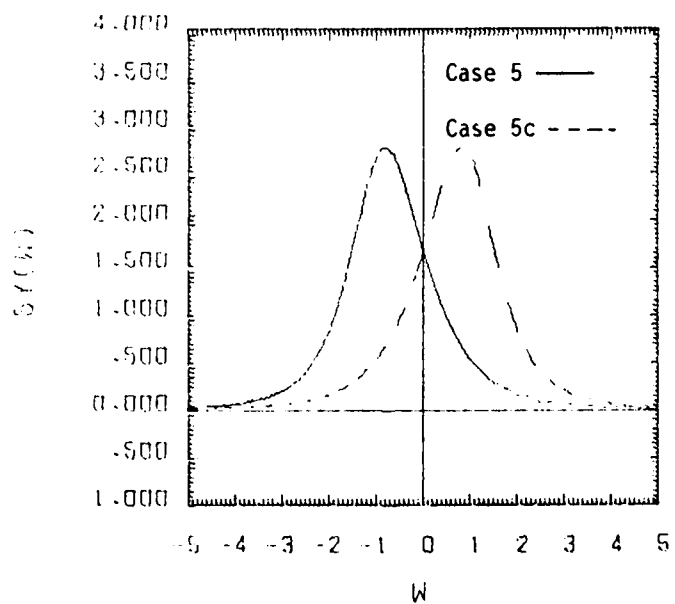
DOPPLER SPECTRUM

Fig. 17a

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DOPPLER SPECTRUM

Fig. 17b

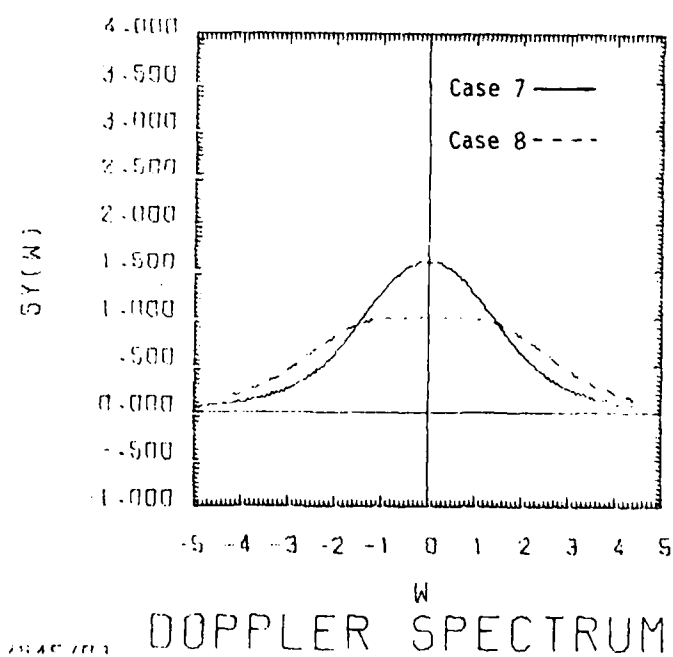
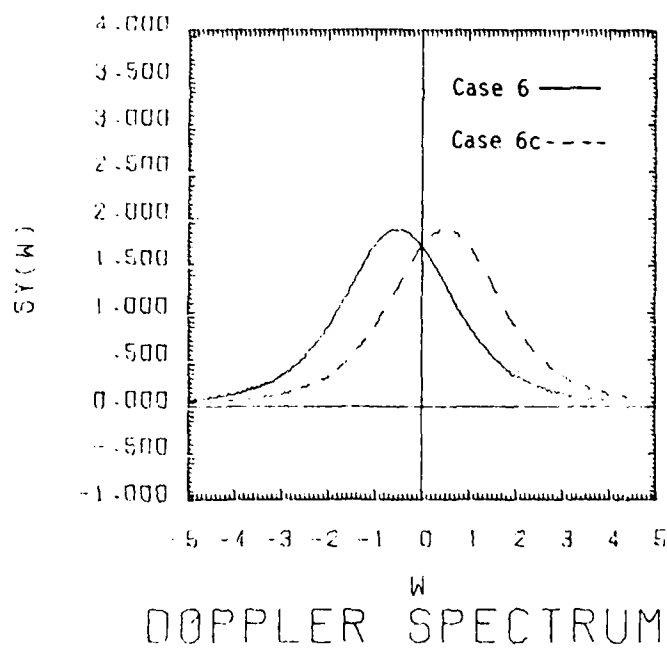


Fig. 17c

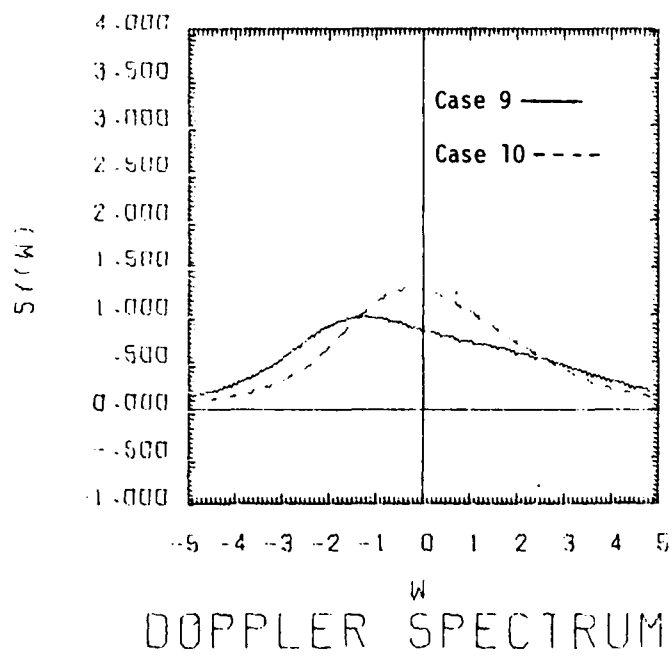
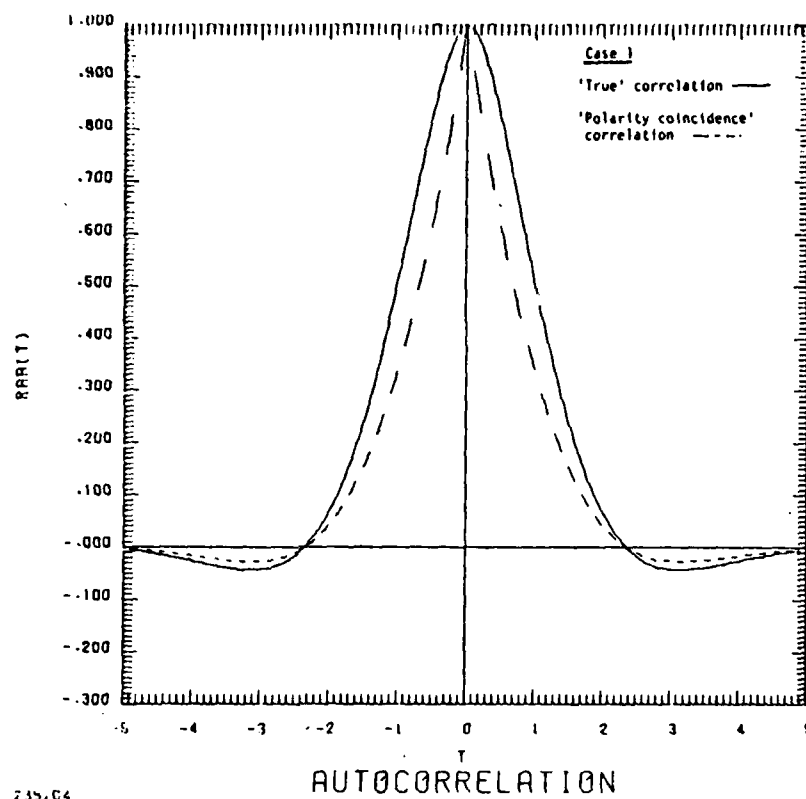


Fig. 17d



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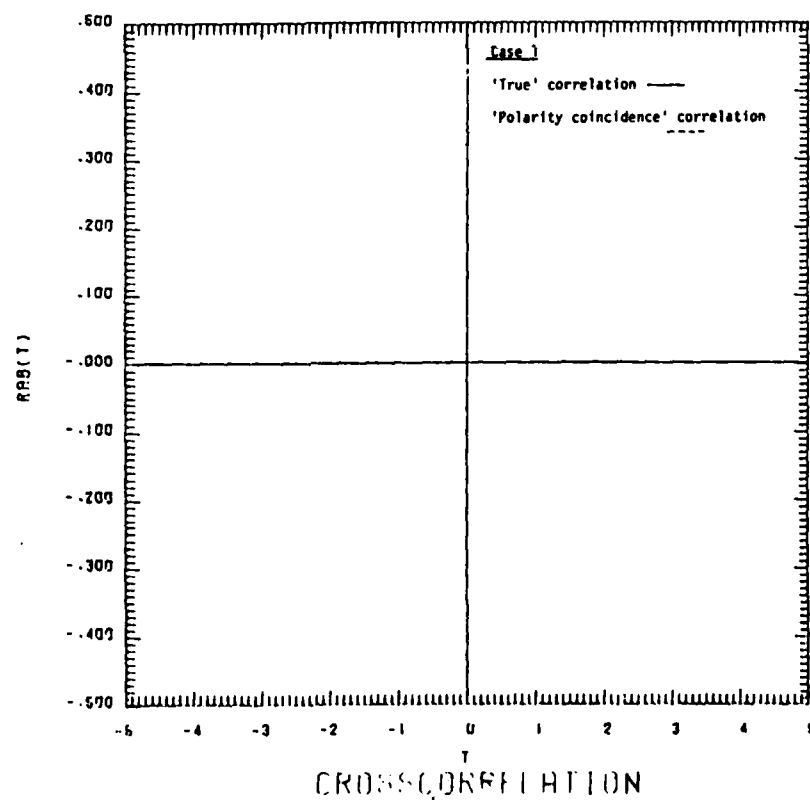


Fig. 18a

D. 90

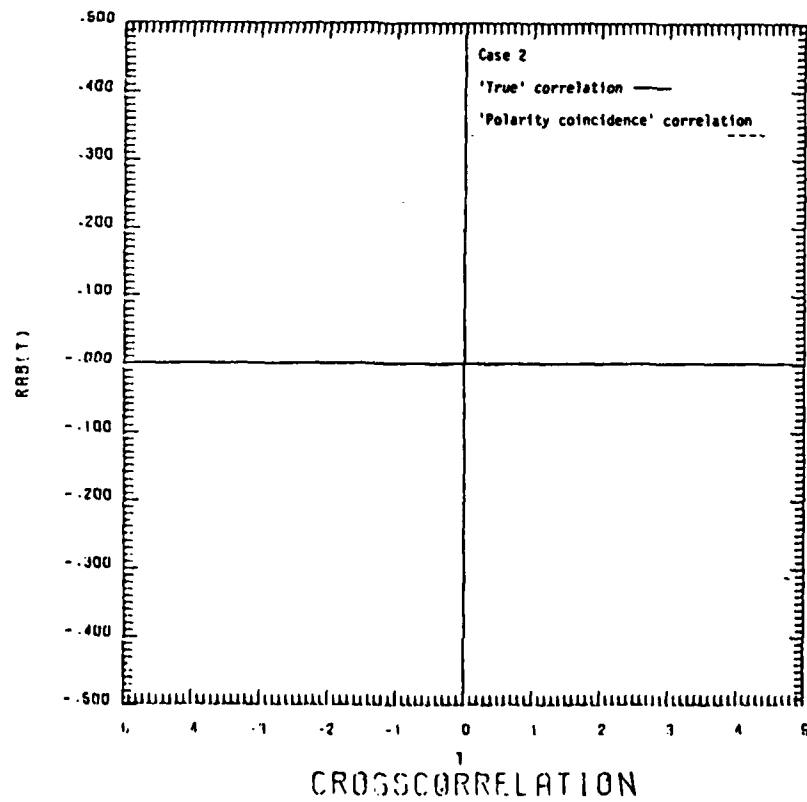
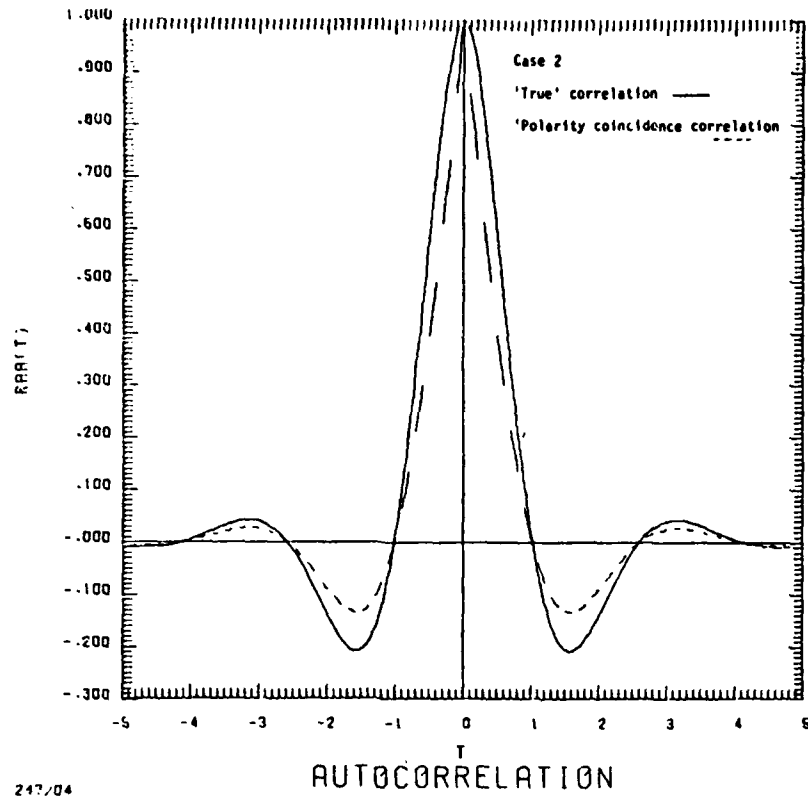
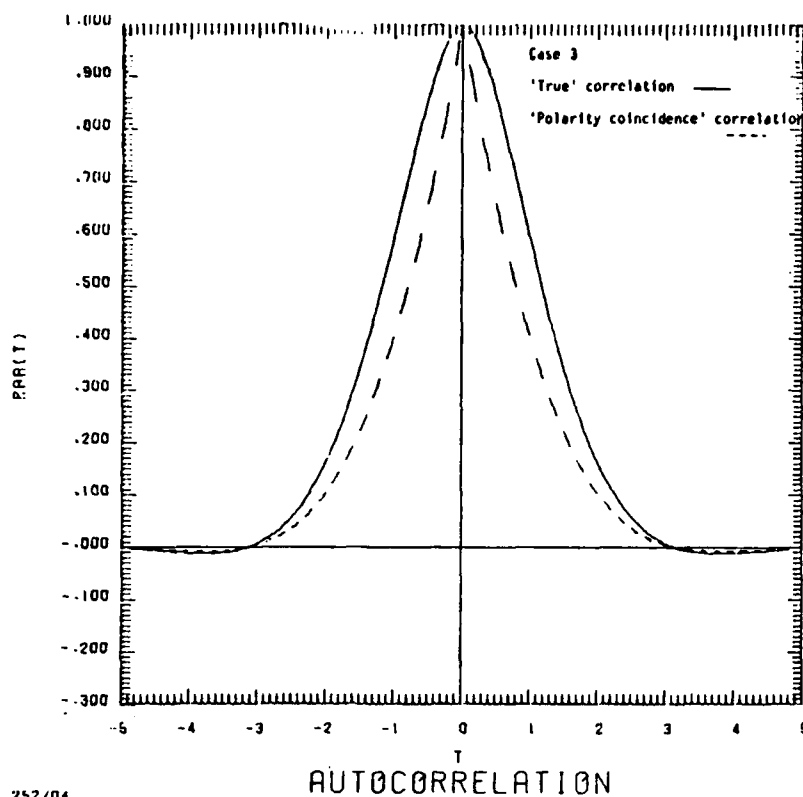


Fig. 18b

D.91



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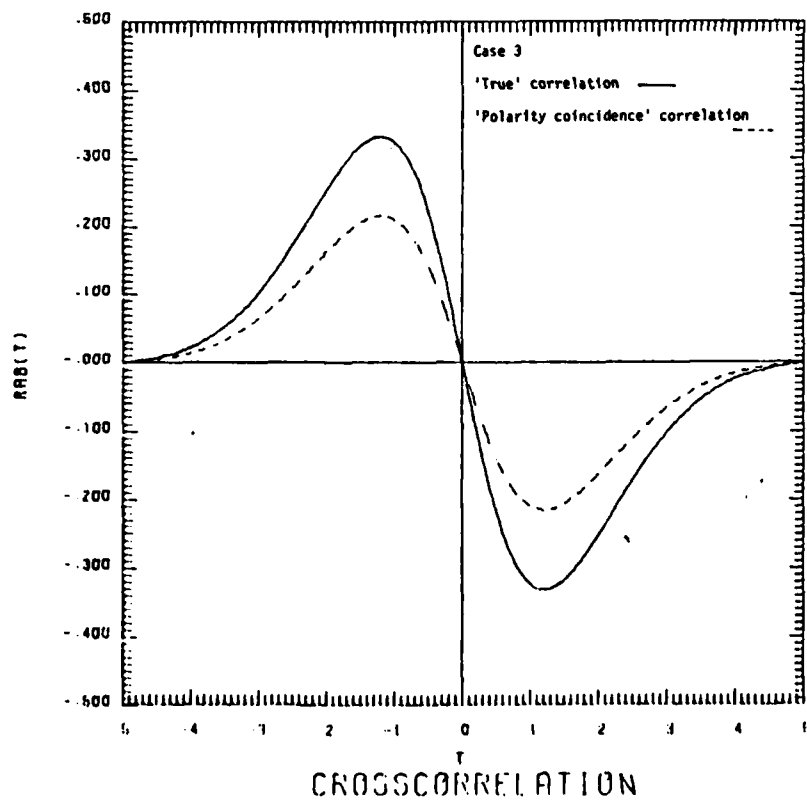


Fig. 18c

D.92

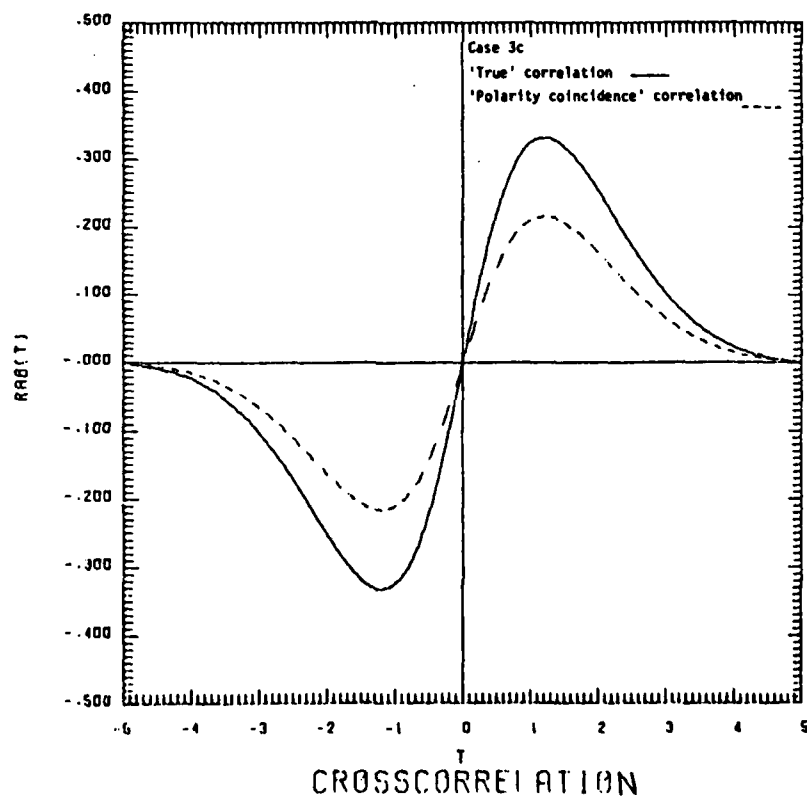
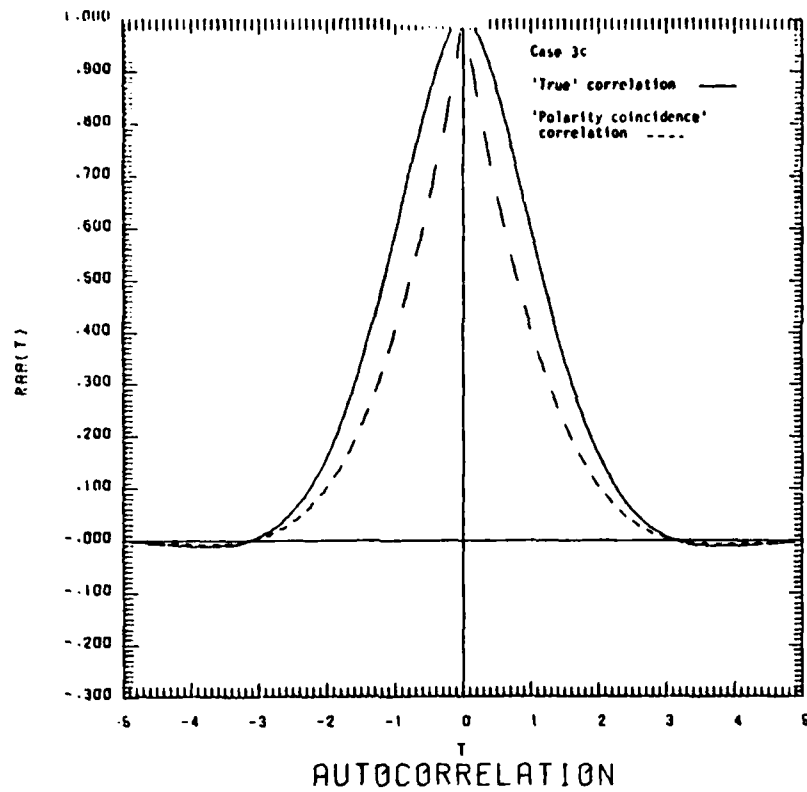
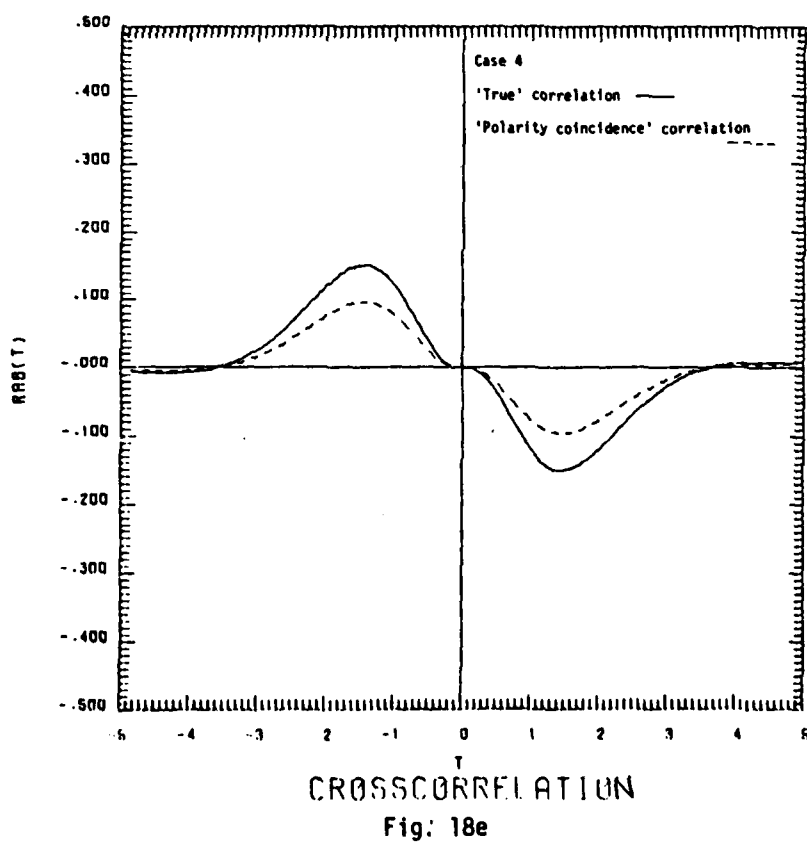
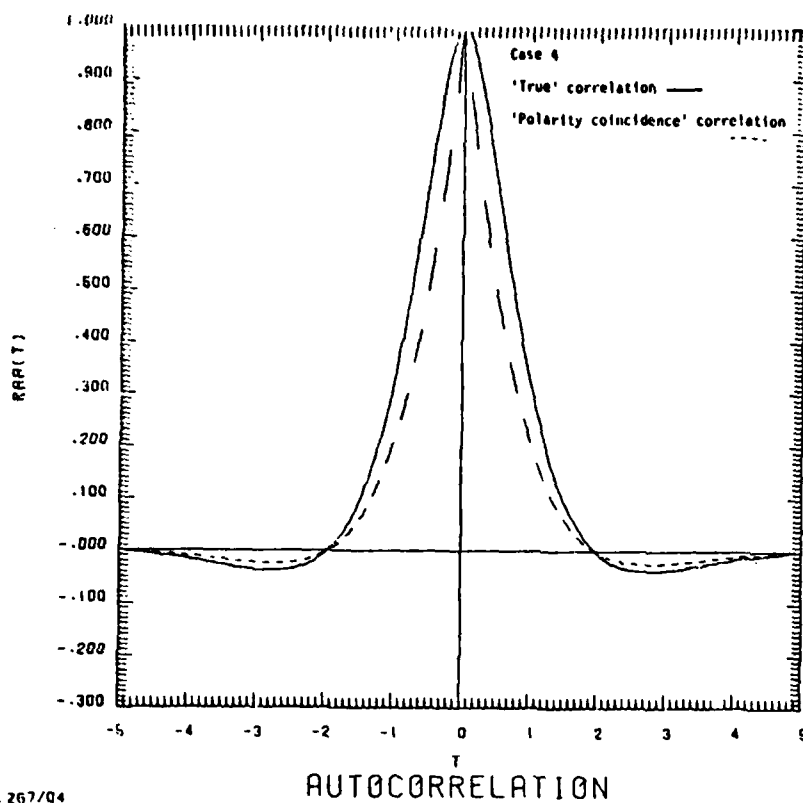


Fig. 18d



D. 94

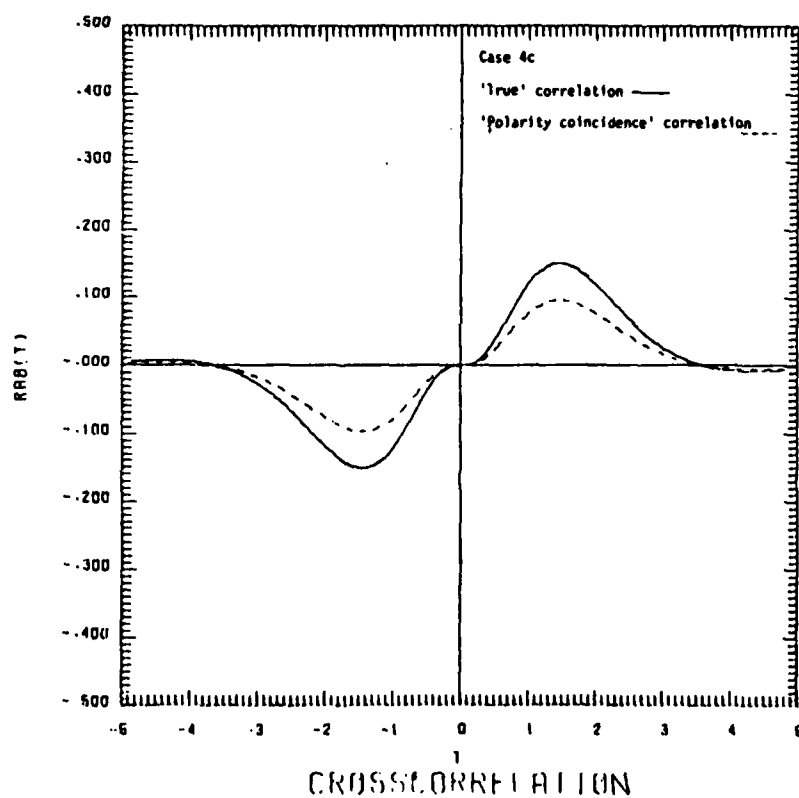
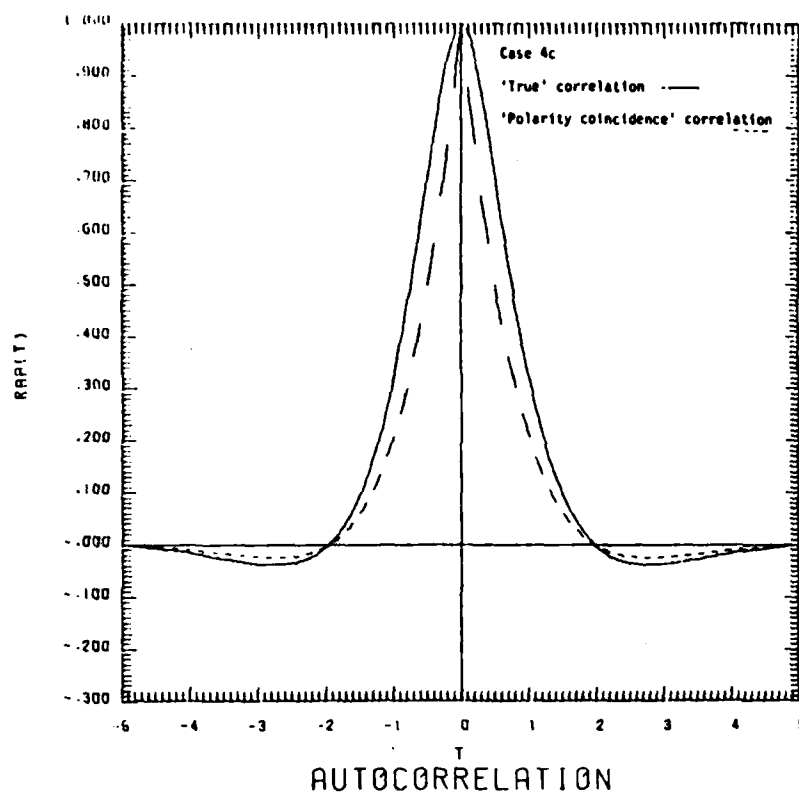


Fig. 18f

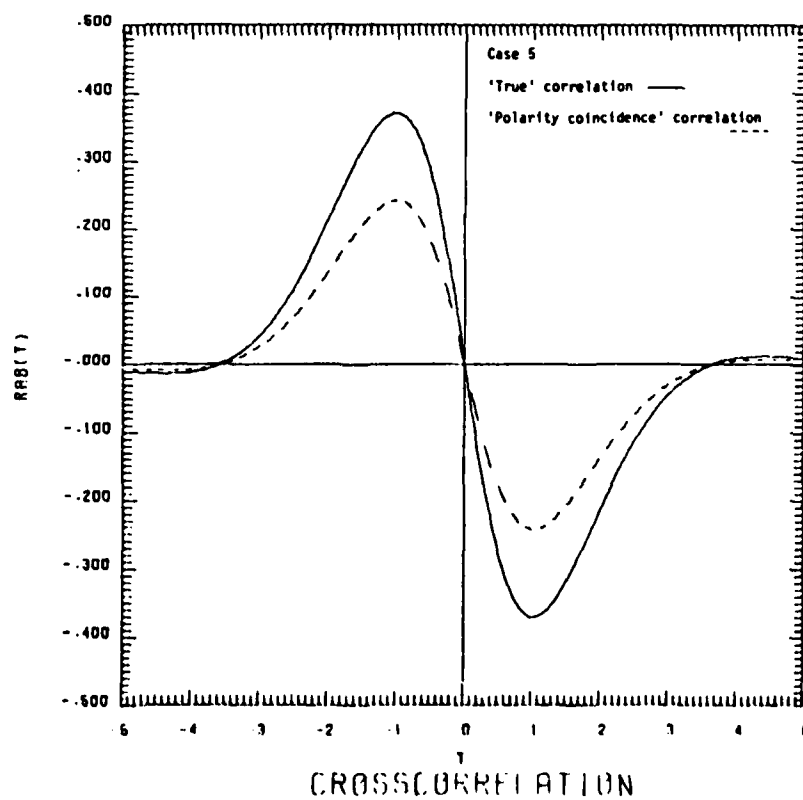
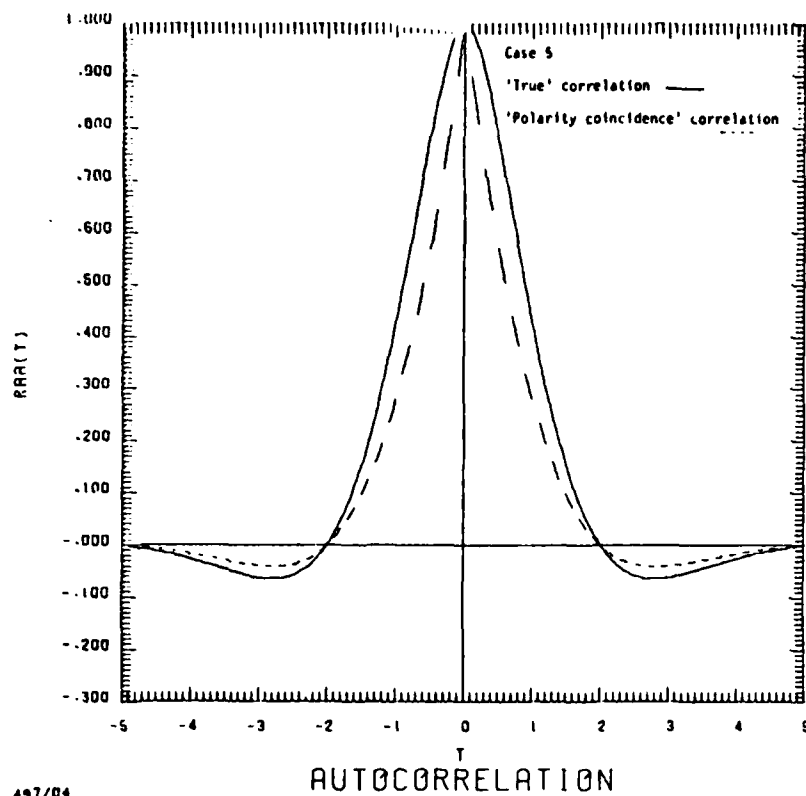


Fig. 18g

D.96

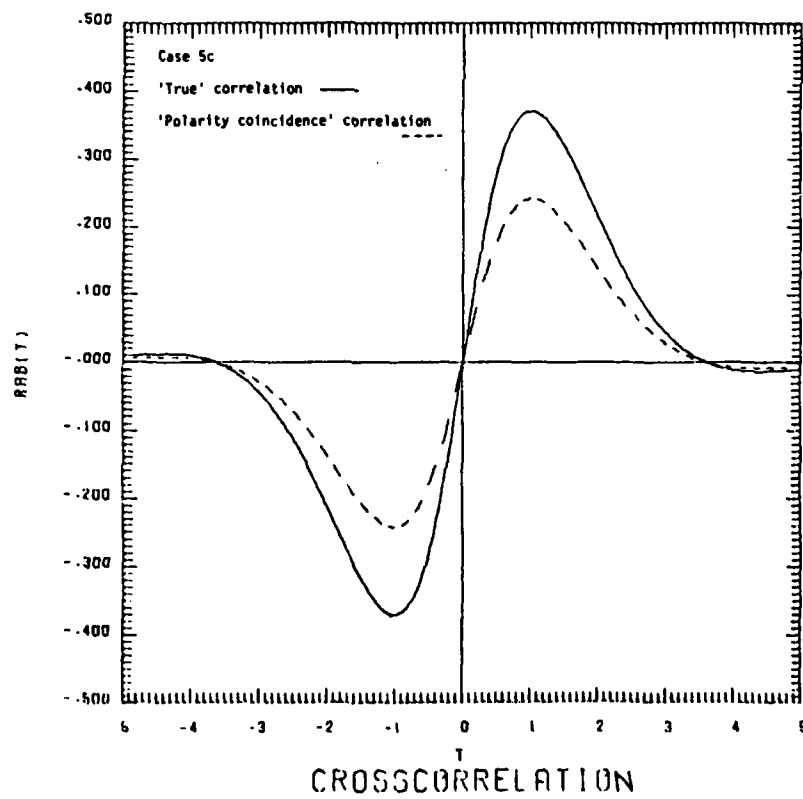
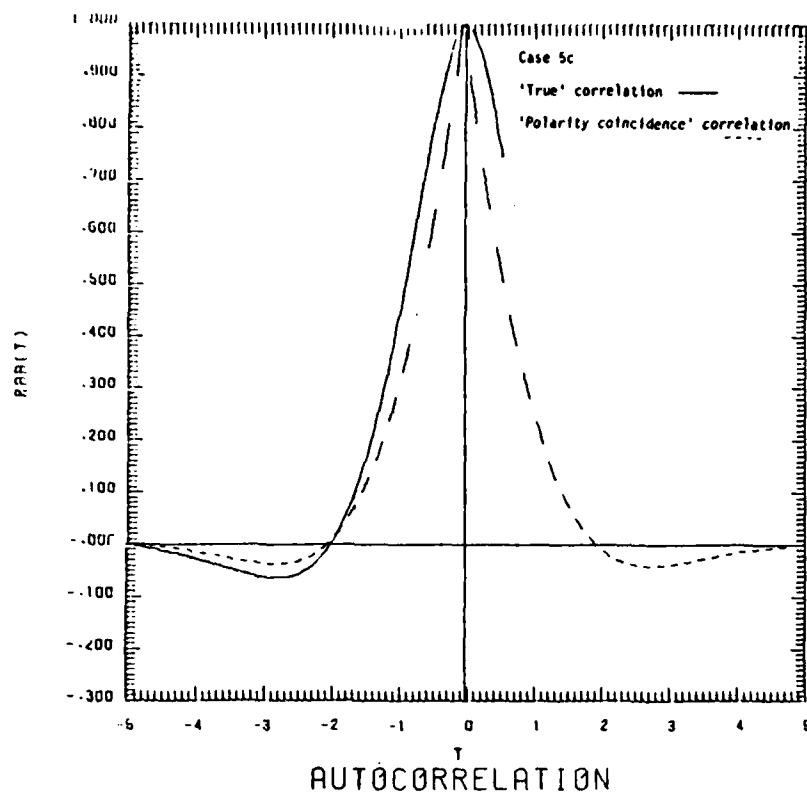


Fig. 18h

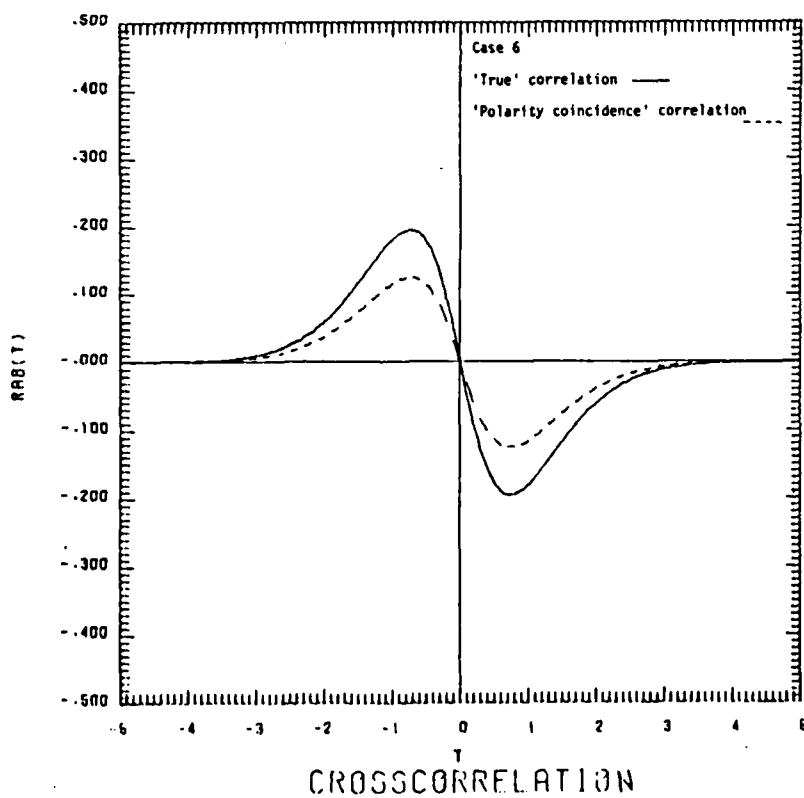
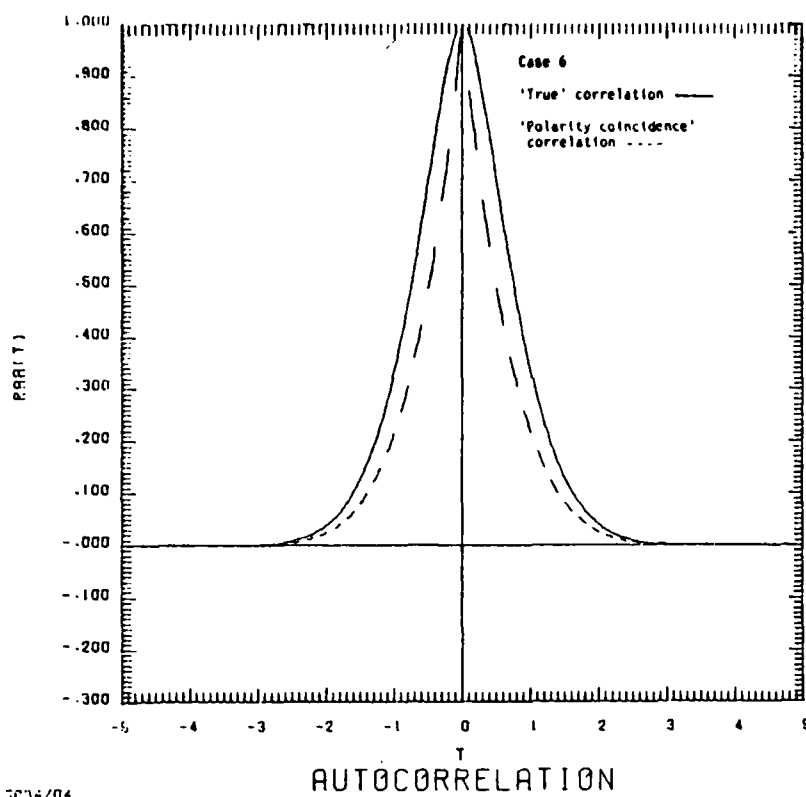


Fig. 181

D.98

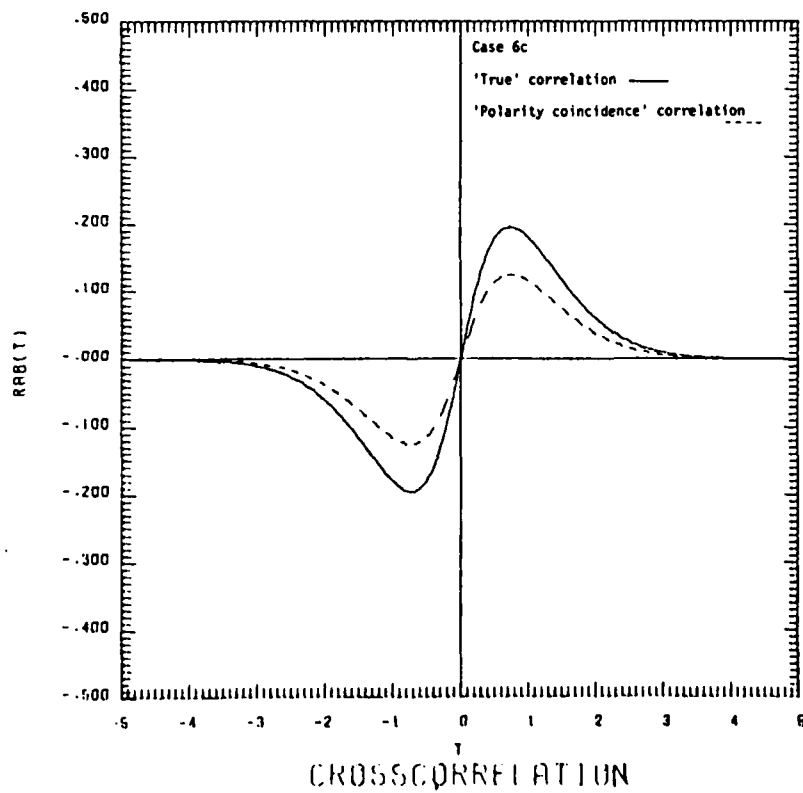
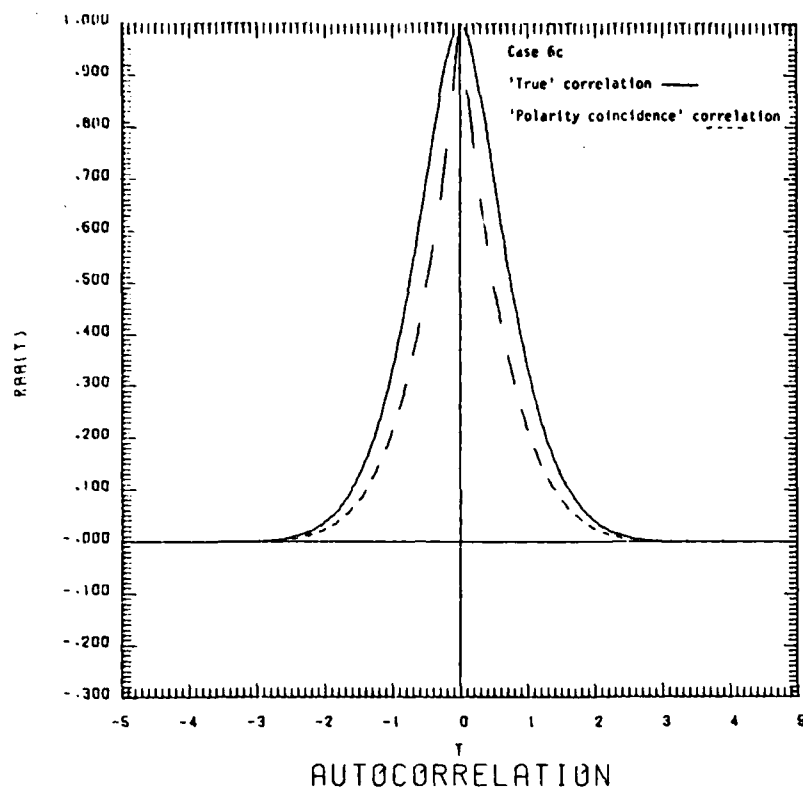
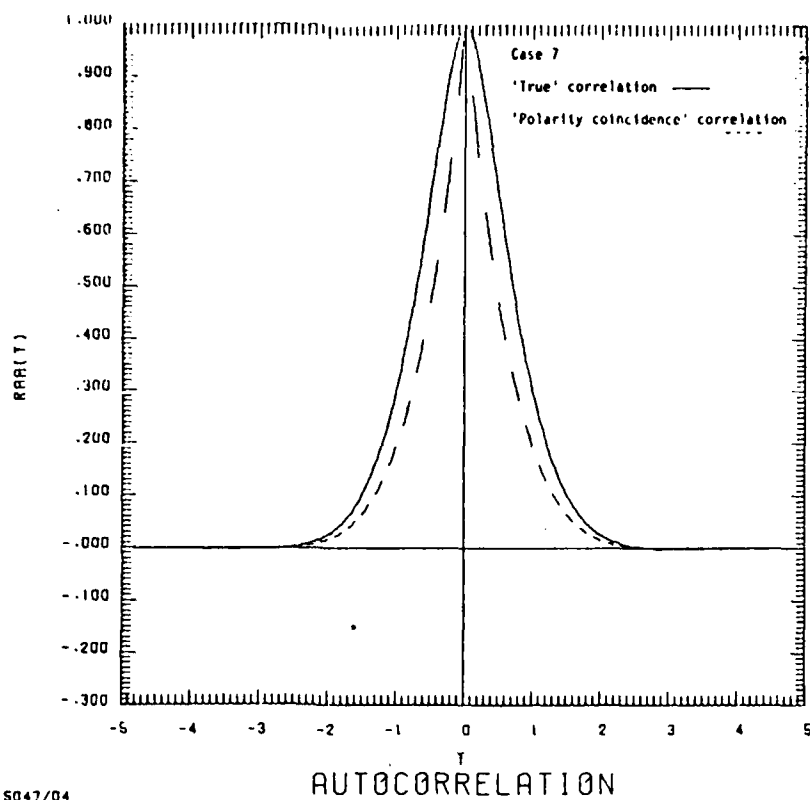


Fig. 18j



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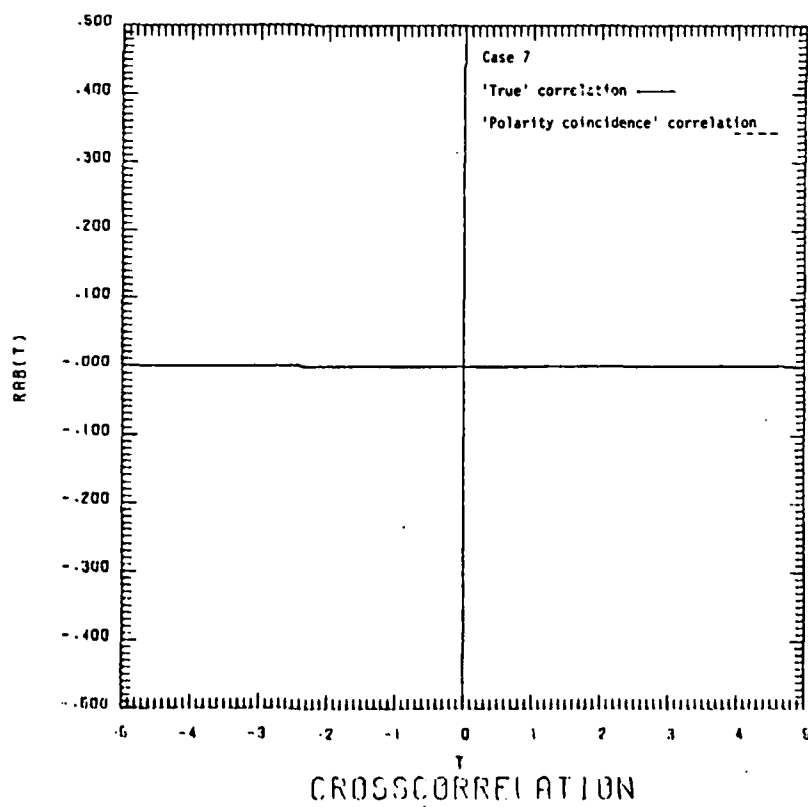


Fig. 18k

D.100

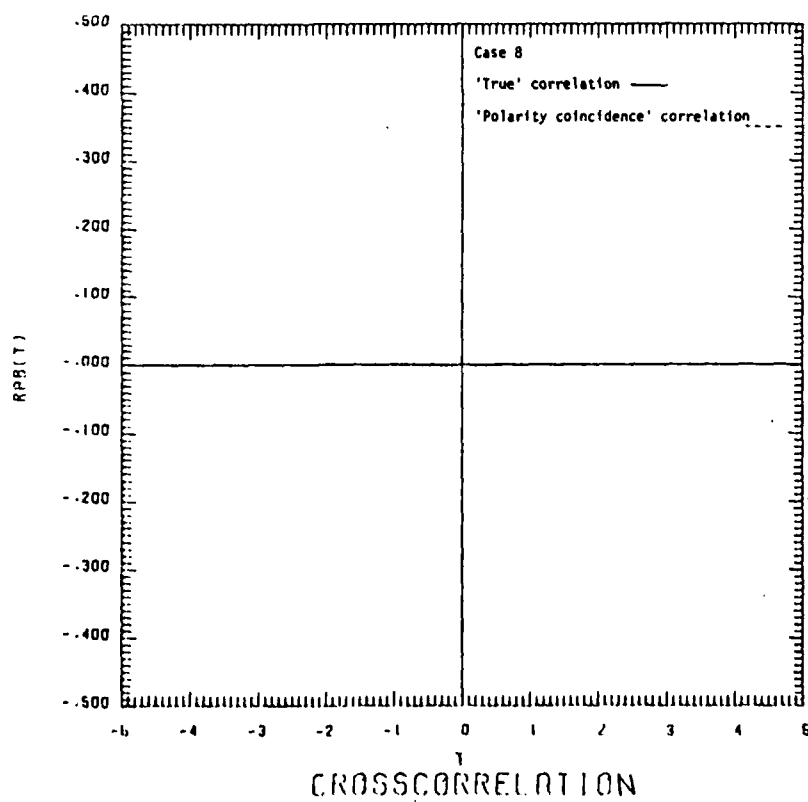
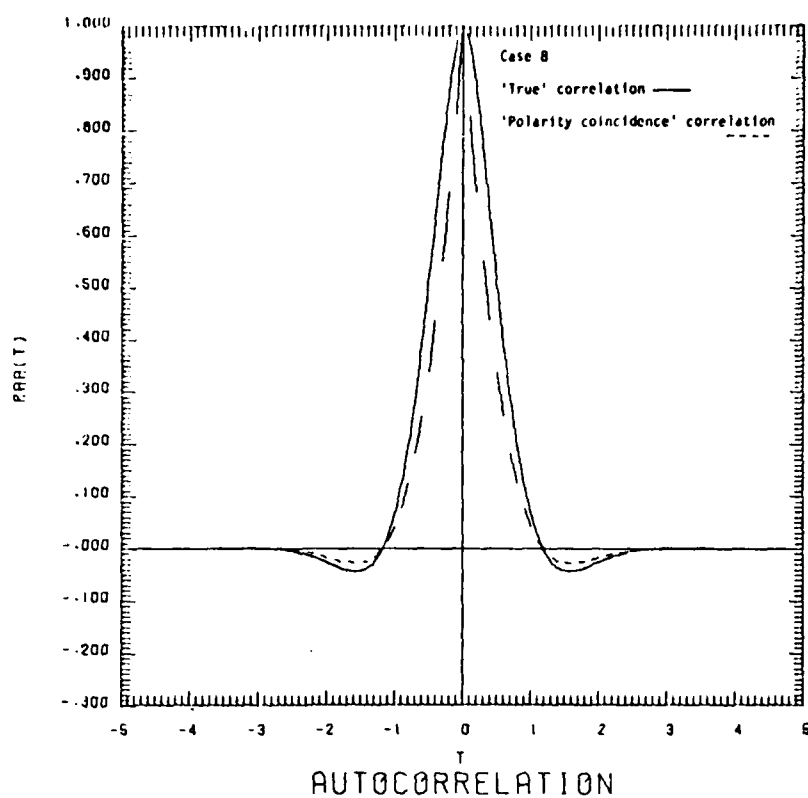
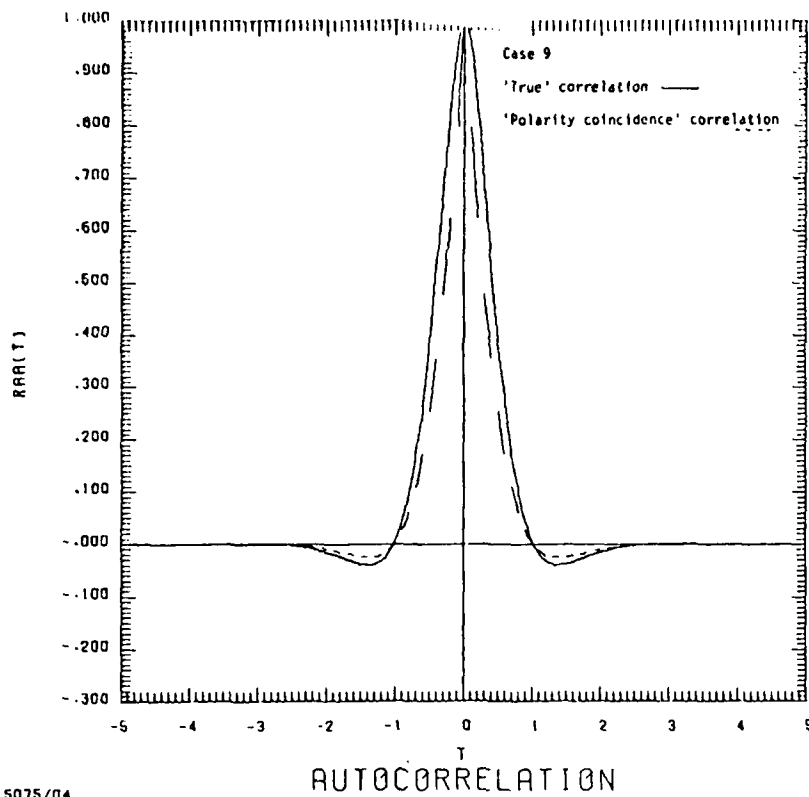


Fig. 181

D.101



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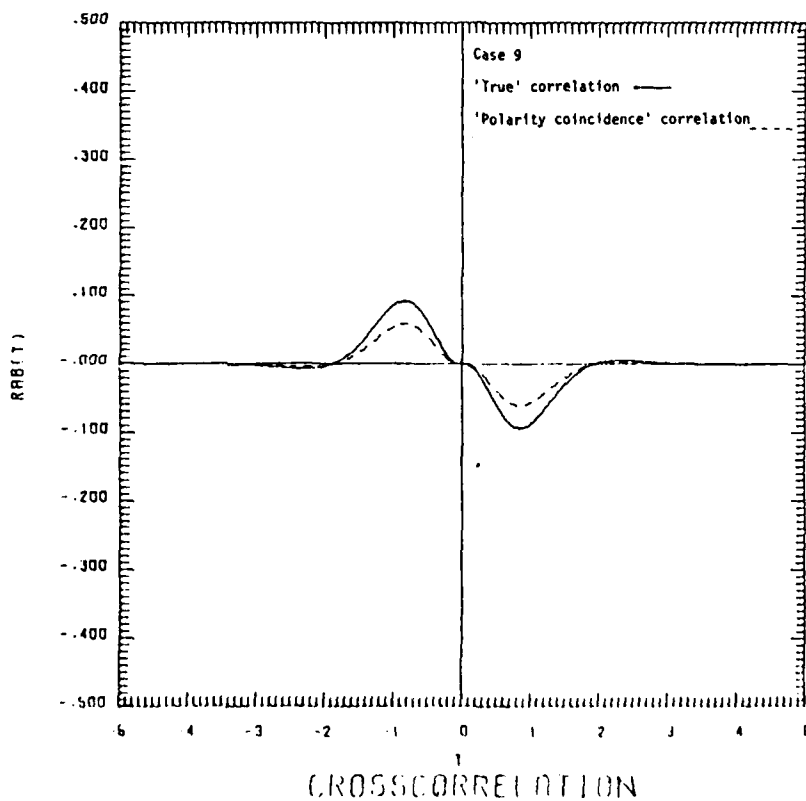


Fig. 18m

D.102

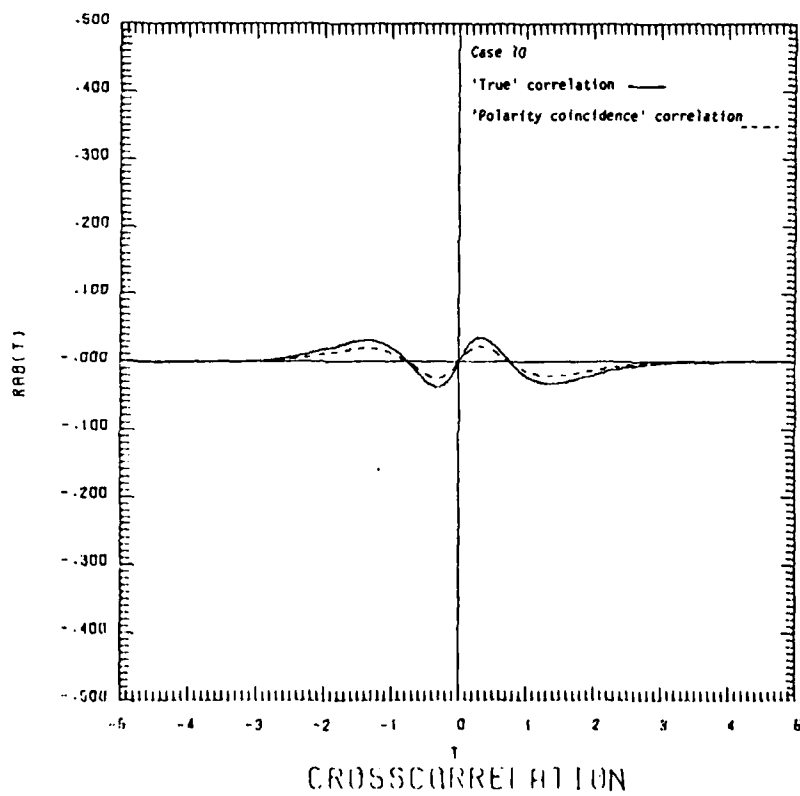
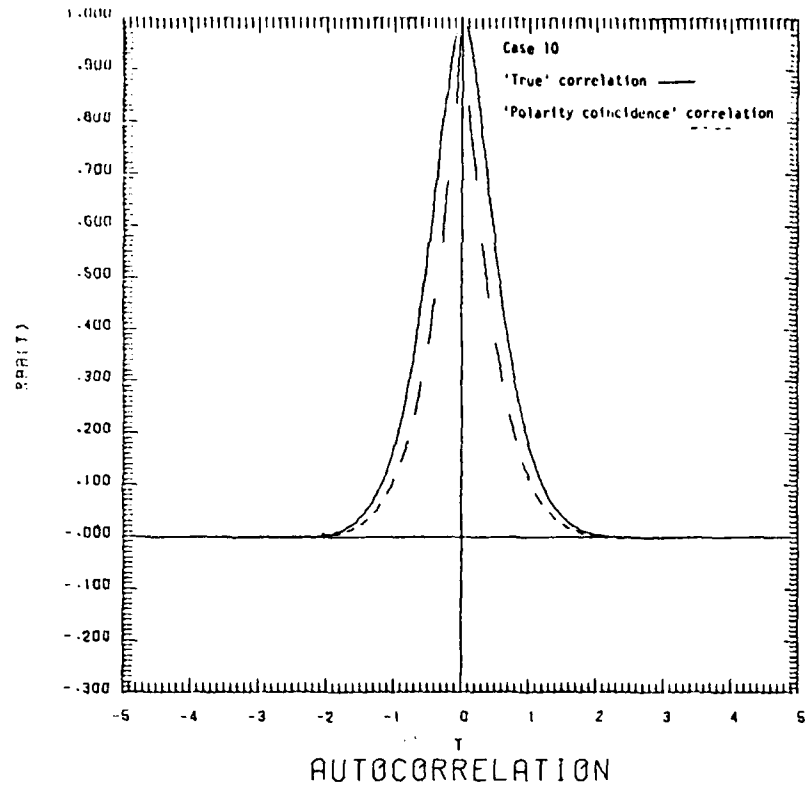


Fig. 18n