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LINEAR FM VERTICAL SOUNDER FOR IONOSPHERIC DISTORTION CORRECTION

D. J. Belknap,
D. R. Bungard,
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B. D. Perry

DECEMBER 1969

Prepared for

DIRECTORATE OF PLANNING AND TECHNOLOGY
ELECTRONIC SYSTEMS DIVISION
AIR FORCE SYSTEMS COMMAND
UNITED STATES AIR FORCE
L. G. Hanscom Field, Bedford, Massachusetts

Project 7160
Prepared by
THE MITRE CORPORATION
Bedford, Massachusetts
Contract AF 19(628)-68-C-0365

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FOREWORD

This document was prepared in connection with Project 7160, OTH Backscatter Systems and Technology, by The MITRE Corporation, Bedford, Massachusetts, under Contract No. F19 (628)-68-C-0365. The project involved a vertical HF sounder using a linear FM waveform and employing computer-controlled real time correction for ionospheric distortions.

REVIEW AND APPROVAL

This technical report has been reviewed and is approved.

JOSEPH N. ALLRED, Capt, USAF
Technical Project Officer
Development Engineering Division
Directorate of Planning and Technology

ABSTRACT

A linear FM vertical sounder has been designed and implemented in order to demonstrate the feasibility of real-time correction of ionospheric distortions. The system is monostatic, relying on chopping for transmitter/receiver isolation. All signal waveforms, local oscillator and correction signals are derived from a specially-modified, programmed Hewlett Packard frequency synthesizer. Results indicate a capability of making real-time corrections over bandwidths of up to 900 kHz and time periods of several seconds.

ACKNOWLEDGMENTS

The authors wish to again thank Dr. O. G. Villard, chairman of ESD's Division Advisory Group, and his associates at Stanford University. Their cooperation in the earlier stages of this program is greatly appreciated. We also wish to thank Dr. J. V. Evans of MIT's Lincoln Laboratory for his kind offer to supply us with vertical ionograms taken at the time of our tests.

The cooperation of Charles J. Beanland of MITRE in the area of transmitters and antennas has also been a great help.

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SECTION I

INTRODUCTION

Two years ago Department D-85 undertook a program in adaptive signal processing for ionospheric distortion correction. Early in the program a theoretical technique was devised.[1] To determine the feasibility of this technique, data was gathered on an oblique ionospheric link operated by Stanford University and then the data was analyzed by computer at MITRE. This non-real time computer simulation demonstrated that the basic technique was feasible.[2]

Last year a vertical HF sounder using a linear FM waveform and employing computer-controlled real time correction for ionospheric distortions was designed and implemented here at MITRE Bedford. This equipment provides a means for demonstrating the practicality of real-time correction, and in a convenient way since the transmitter and receiver are co-located. The equipment can also be used as a "building block" for future oblique experiments. The vertical sounder has thus been employed as an experimental test bed rather than as an operational data-gathering station.

SECTION II

THEORY OF OPERATION

The basic theory of operation of linear FM sounders, as well as further background information on this project, can be found in references 1 and 2. Briefly stated, a linear FM signal is transmitted vertically, reflected off the ionosphere where it suffers complex distortions, received, and correlated with the transmitted waveform. The linear FM modulation maps time into frequency. After correlation, the mean of the spectrum of the signal is a measure of the range to the ionosphere and the spreading of the spectrum is a measure of the path distortions. Range gating is accomplished by adjusting the receiver bandwidth and center frequency. Range resolution, in absence of path distortions, depends on the extent of the frequency sweep.

Although some transmitter/receiver isolation is achieved through the use of a passive hybrid or by having separate, cross-polarized antennas, most of the required isolation is due to time-multiplexing or "chopping". The chopping rate is chosen so that the receiver is turned on at the appropriate time. The returned signal thus arrives as a many-time-around echo (in the radar sense). The resulting range ambiguity is resolved by prior knowledge of the approximate layer height.

A high chopping rate is desirable in order to be able to remove the sidebands due to the chopping frequency and its harmonics by filtering in the receiver. On the other hand a low chopping rate permits a higher duty cycle which conserves transmitter peak power. This is so because for optimum reception the receiver on-time should equal the transmitter on-time, τ, plus the range window of the receiver. If the receiver has a bandwidth of BHz (at IF) and the waveform is a linear FM of slope W/T Hz/sec then the range window will be BT/W sec. In order to be able to receive returns separated by as much as BT/W seconds the receiver should be connected to the antenna for τ + BT/W seconds thus making the total interpulse interval 2τ + BT/W. The desired high duty cycle will be achieved if BT/W is a small fraction of this interval since the duty cycle is

$$D = \frac{\tau}{2\tau + BT/W} \tag{1}$$

The chopping rate is the inverse of the interpulse interval or

$$f_r = \frac{1}{2\tau + \frac{BT}{W}} \tag{2}$$

For the equipment being described in this report B = 300 Hz and although W/T is adjustable a typical value is 500 KHz/sec. Thus a typical BT/W is 600 μsec. In order to achieve a good compromise between chopping rate filtering and duty cycle τ should also be 600 μsec which gives

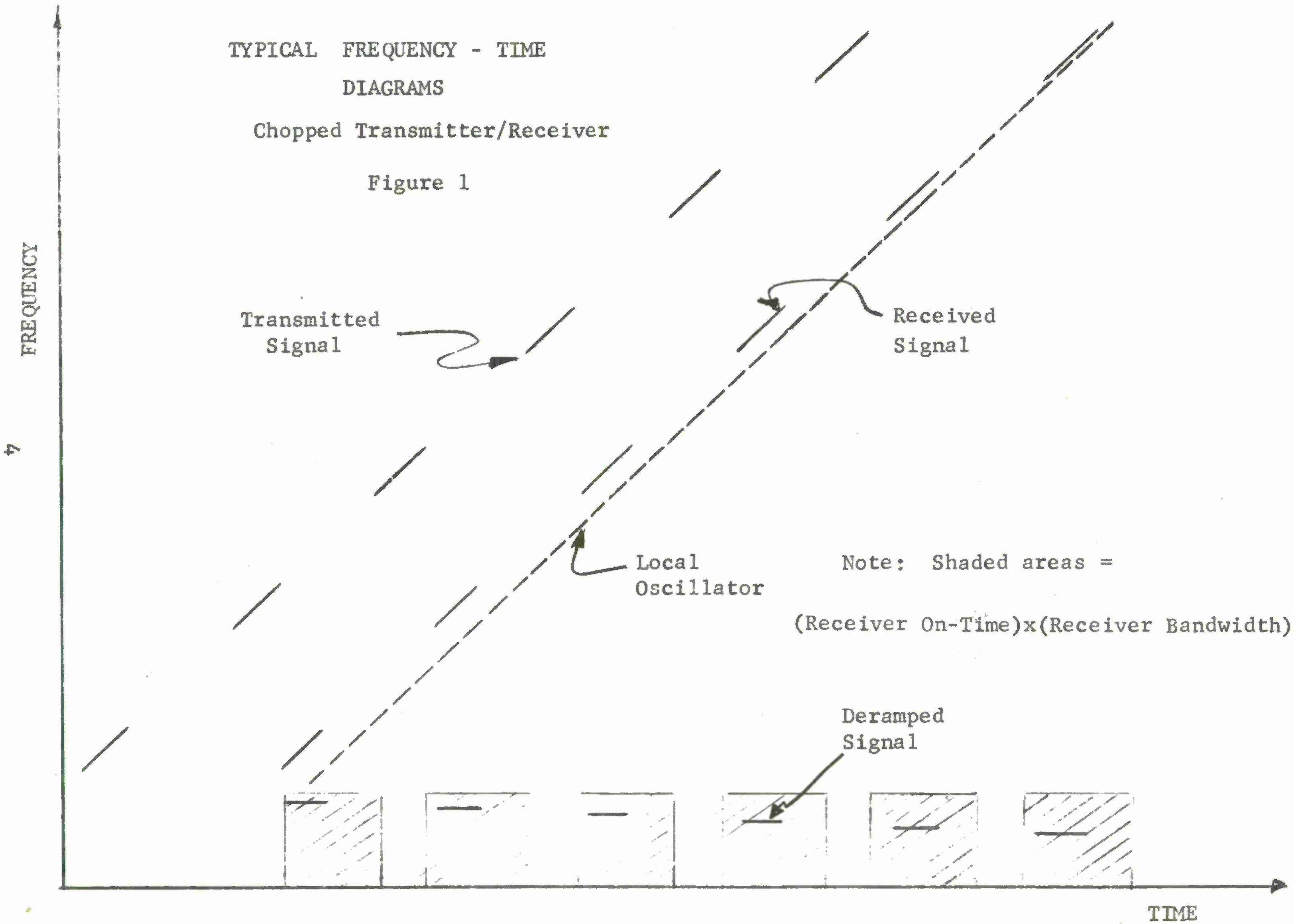
$$D = \frac{600}{2(600) + 600} = .33$$

and

$$f_r = \frac{10^6}{2(600)+600} = 555 \text{ Hz}$$

Thus the first sidebands due to chopping are at $\pm$ 555 Hz which is sufficiently far out to be removed by the 300 Hz wide receiver filter.

Naturally, if one is willing to mismatch the receiver on-time and the range window, higher duty cycles and/or lower chopping rates can be achieved. Figure 1 is a sketch of frequency-time plots for typical signal. For this example the duty cycle is about 0.30. Also,



TYPICAL FREQUENCY - TIME DIAGRAMS

Chopped Transmitter/Receiver

Figure 1

the spread in frequency of the received signal, due to dispersion in the path, can be clearly seen (after de-ramping to remove the linear FM modulation). The delay to and from the ionosphere (range) is such that the signal returns as a second-time-around echo.

SECTION III

DESCRIPTION OF EQUIPMENT

General

A simplified block diagram of the equipment is shown in Figure 2. The heart of the system is a specially modified Hewlett-Packard frequency synthesizer which generates the linear FM waveform to be transmitted, provides the linear FM local oscillator signal to the correlation mixer in the receiver and is programmed to correct for the measured phase-perturbations of the ionospheric path. The received data is quadrature-detected, sampled and converted to digital form. The first linear FM sweep in a sequence is used to measure the transfer function of the path. On subsequent sweeps the computer provides outputs to compensate the receiver in amplitude and phase for this transfer function. All results are recorded for later analysis. (see Section V below).

The equipment, exclusive of the high power transmitter amplifiers is housed in two relay racks as shown in Figure 3.

Digital Control Circuits

The objective of the digital control circuits are to control the timing, waveform generation, and data acquisition of the ionospheric sounder. A linear FM waveform is transmitted vertically to the ionosphere and back. The received signal is sampled and converted to digital information for processing and storage. The computer determines the ionospheric transfer function and computes corrections to be applied by the digital control circuits to subsequent received signals.

The overall block diagram of the control circuits is shown in Figure 4. These circuits consist of a twenty-four hour clock and

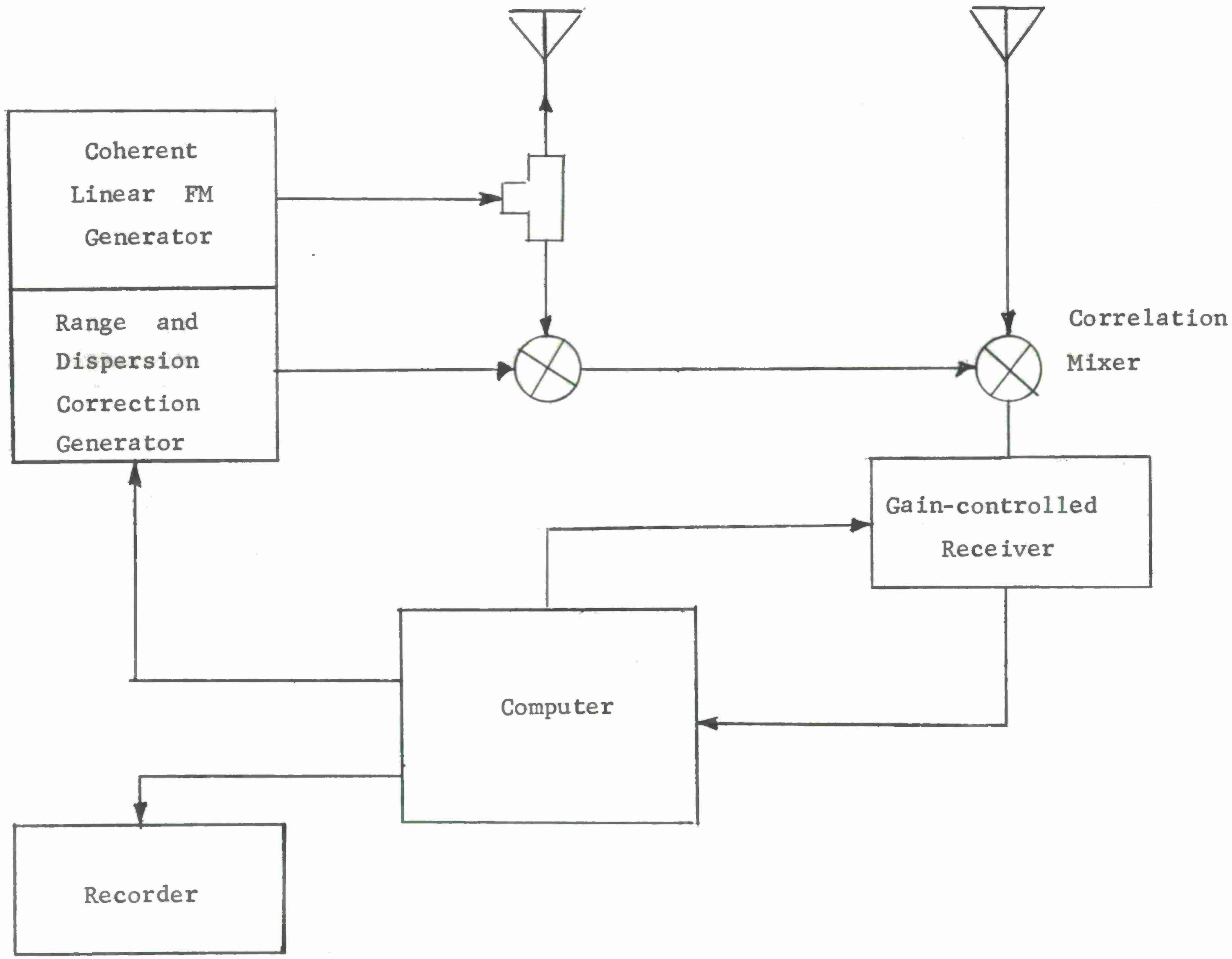


Figure 2. Simplified Block Diagram

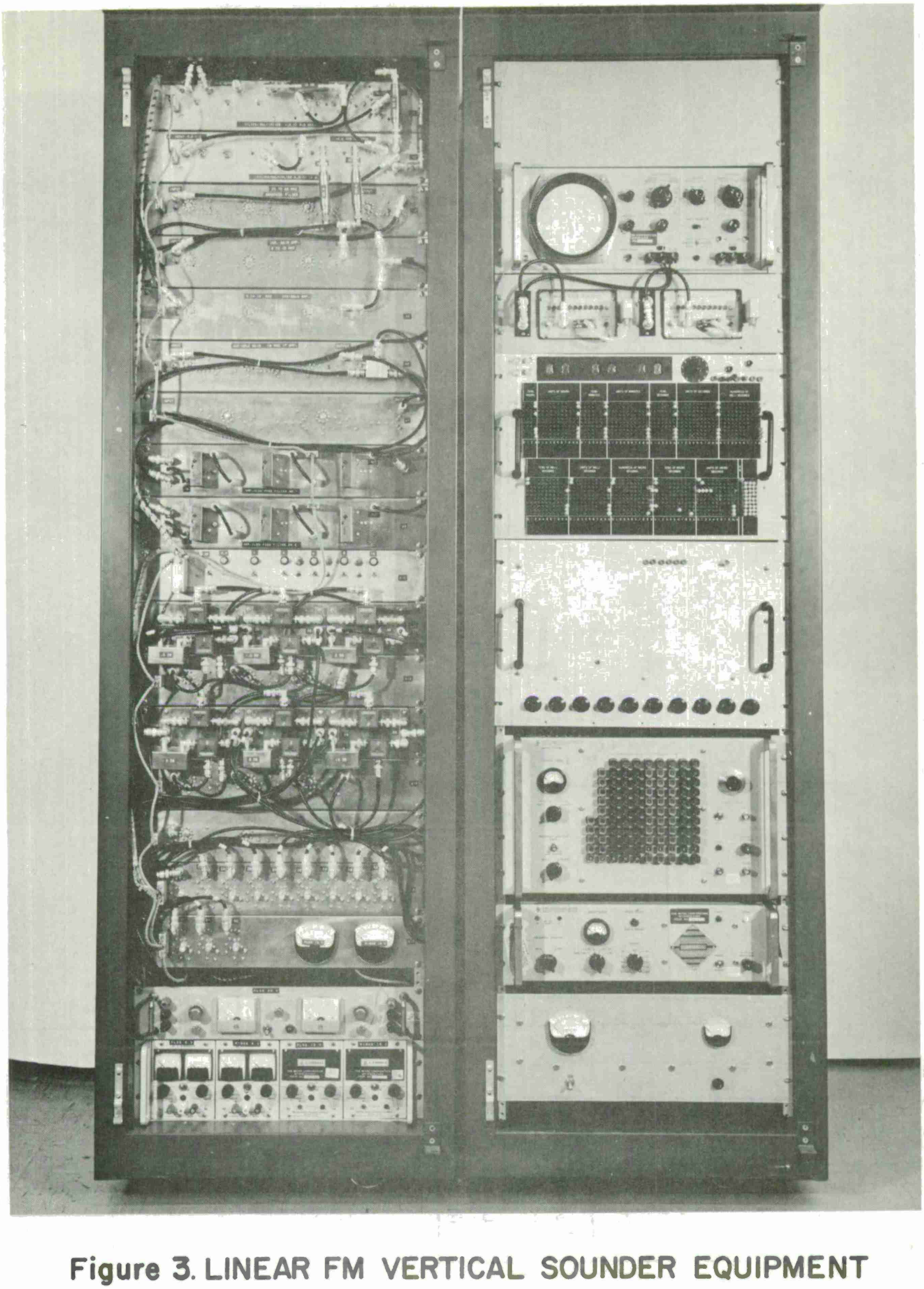


Figure 3. LINEAR FM VERTICAL SOUNDER EQUIPMENT

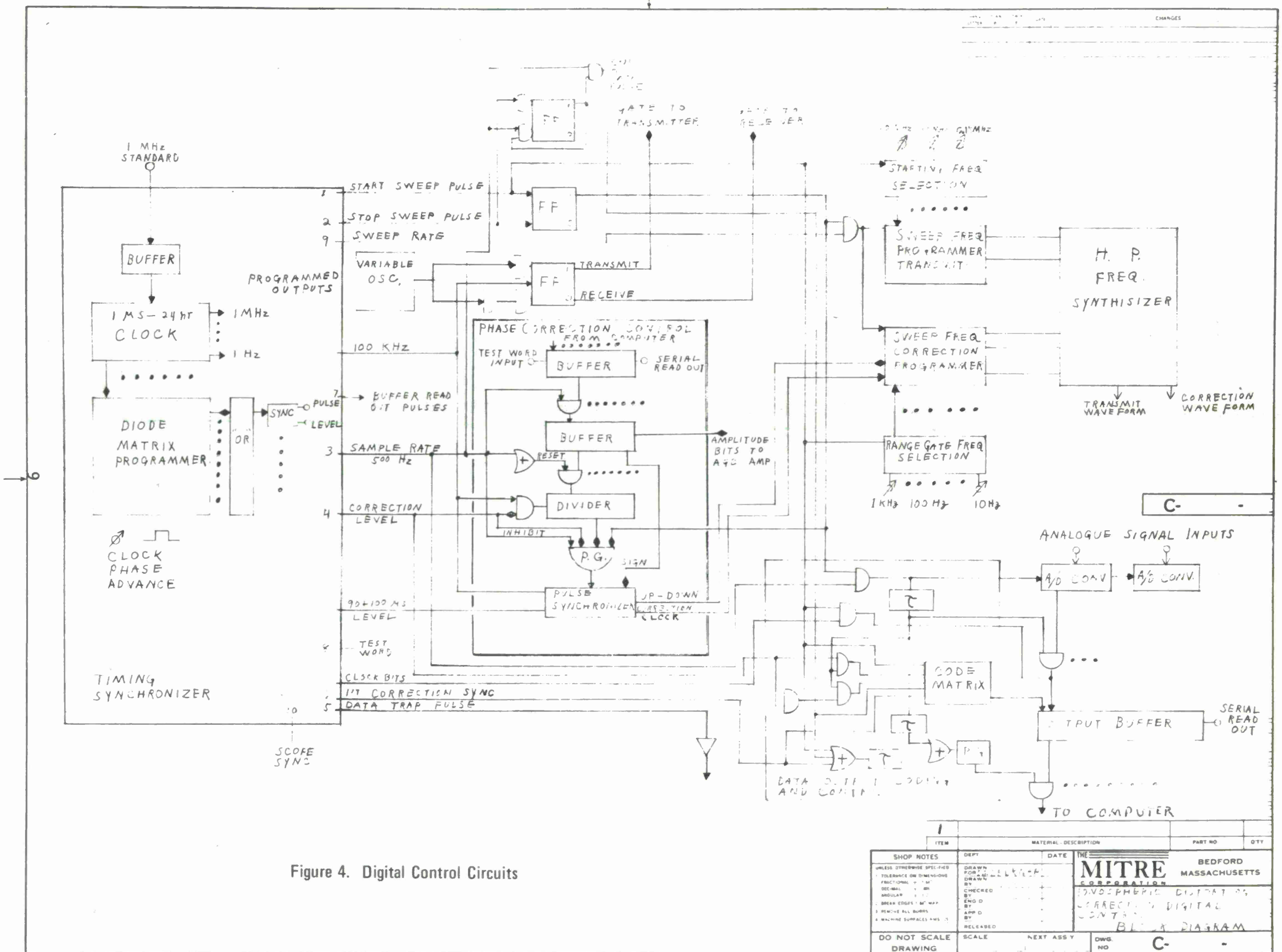


Figure 4. Digital Control Circuits

timing generator, a modified Hewlett-Packard frequency synthesizer, programmers and drivers for the synthesizer, data sampling and storage circuits, and a phase and amplitude correction controller.

Clock and Timing Generator

The twenty-four hour digital clock and timing generator controls all timing to synchronize the transmitted and received local oscillator waveforms, sample the received signal, transfer the data to the computer, and apply computer generated corrections to the received signals.

A plug-in diode programmer allows timing marks to be selected at any time of the day to a resolution of one microsecond. Ten different programs can be selected. The flexibility of each program is best described by the following example: Timing marks might be desired every hour, every minute, in the zero and five seconds interval, to occur every two milliseconds for 900 milliseconds. In additon, any two programs can be "or'ed" together. The pulse outputs are synchronized with the basic system clock to within 20 nanoseconds. They are negative going and 200 nanoseconds wide. A synchronized level output is also available. Pulses are also available at basic frequencies from 1 Hz to 1 MHz in decade steps. The clock and timing generator are shown in the block diagram of Figure 5.

Dividers and decoders generate the clock digits from the 1 MHz reference of the Hewlett Packard synthesizer. Diodes select any digit in any decade which then logical "and" with selections from all other decades to generate a particular timing signal. The outputs of the diode gates are synchronized to the system clock with a flip-flop.

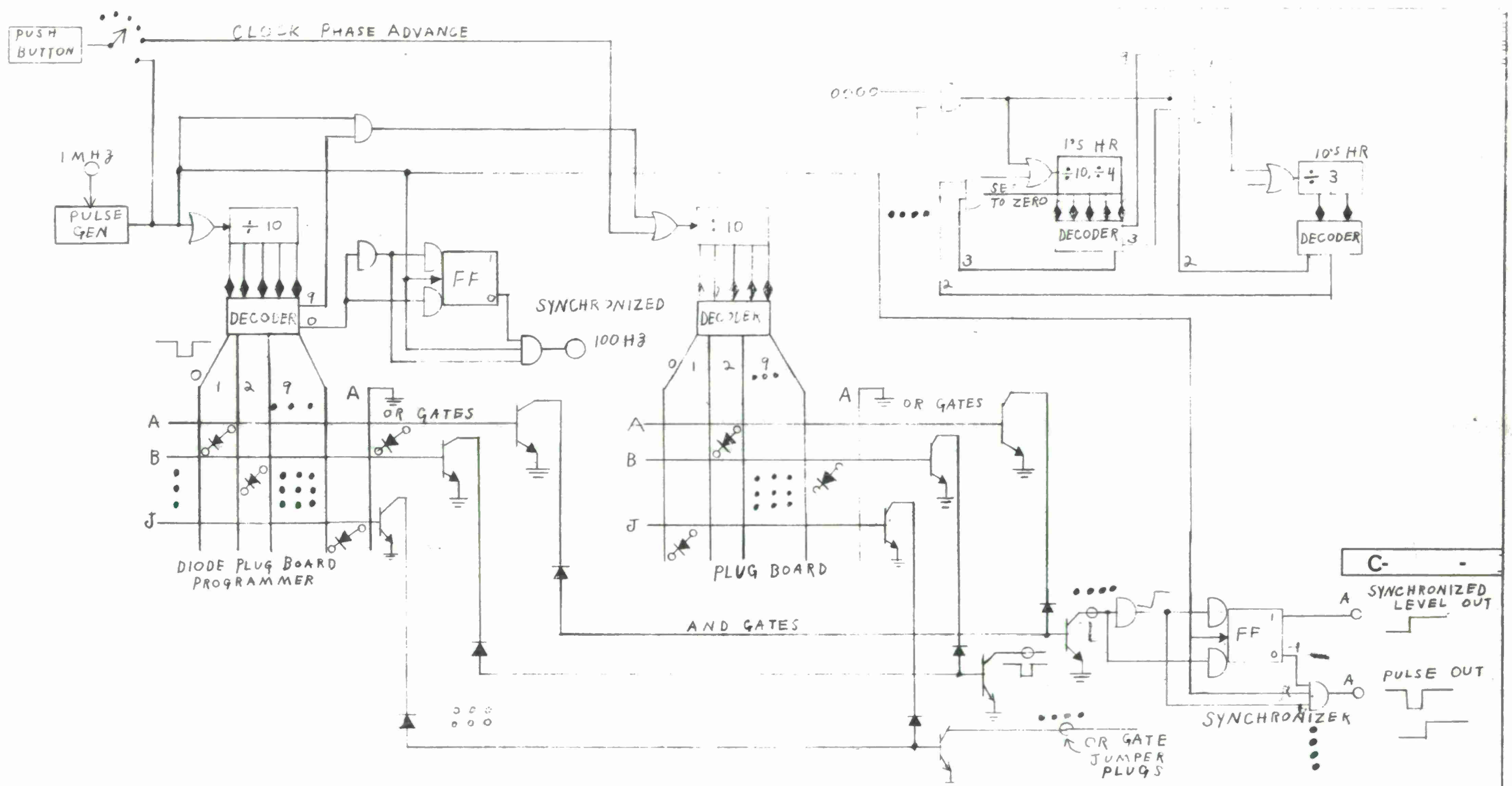


Figure 5. 24 Hr Clock and Timing Generator

SHOP NOTES	DEPT	DATE	THE MITRE CORPORATION BEDFORD MASSACHUSETTS
DO NOT SCALE DRAWING	DRAWN FOR BJ BELKNAP		CLOCK AND PROGRAMMER BLOCK DIAGRAM
	SCALE	NEXT ASS'Y	DWG. NO. C- -

MODERN BLUE PRINT CO. NO. 1050 CLEARPRINT

The outputs from the timing generator are:

1. a "start of sweep" pulse which starts the linear FM waveform,
2. a "stop sweep pulse" which stops the linear FM waveform,
3. the received signal "sample rate" (500 samples per second),
4. a "correction level" which determines which FM sweeps are to have computer generated corrections applied to them,
5. a "data trap pulse" to allow read out of words at the computer to equipment interface,
6. "send first correction" synchronization pulse,
7. buffer serial shift pulses to allow serial read out of the input and output buffer registers,
8. a serial test word which can be inputed to the buffer registers,
9. the linear FM "sweep rate" (100 Hz step every 100 μsec, for example)
10. A scope "sync" pulse.

Frequency Synthesizer

A Hewlett-Packard phase coherent frequency synthesizer model H-75-5110A/5110A is used to generate a linear FM phase coherent transmitter waveform and a local oscillator de-ramping waveform. It is a modified version of a standard 50 MHz frequency synthesizer. All switched frequencies (3.0 to 3.9 MHz) are multiples of 100 KHz and are adjusted to go through zero phase angle at ten microsecond intervals, thus allowing phase coherent switching at these times. A front panel switch on the synthesizer driver unit allows the 3.0 to 3.9 MHz frequencies to be adjusted in phase to within $\pm$ 18^{o} of each other.

A further refinement of the phase coherence of these reference frequencies and the 30 to 39 MHz frequencies is accomplished by trimming the channel output filters. Since the phase of these frequencies will drift due to component aging and temperature changes, a trimming capacitor is added across the output stage of each of the filters. A periodic alignment of these frequencies can be made without removing the filters from the synthesizer.

A test was conducted to determine the phase stability of the 30 to 39 MHz reference frequencies. The phase change with respect to the 1 MHz standard was only $\pm$ 3 degrees in eight hours.

To make this frequency synthesizer phase coherent when frequency transitions occur in more than one decade at the same time, the frequency switching transition must arrive at certain mixers in time synchronization. The dividers which follow these mixers would otherwise be driven by incorrect frequencies for an interval and could slip in phase. To realize this synchronization, delay lines have been inserted to bring the delay of each decade to ten microseconds while each successively higher decade is programmed to switch after a ten microsecond delay.

A further significant modification has been to separate the four lowest order decades from the higher order decades and to program them independently to generate the ionospheric phase correction waveform and frequency (range) offset. The higher order decades generate the linear FM for both the transmit waveform and receiver local oscillator. The correction waveform is added to the local oscillator before signal de-ramping.

Diode clippers have been added to the synthesizer switching matrix to decrease the energy in some of the switching transients.

A linear FM phase coherent waveform has been generated. This waveform has been transmitted and signal returns received from the ionosphere. Phase coherency tests have shown that no phase slipping occurs in this waveform. Another test has shown that the synthesizer can be made to switch large increments in frequency in a phase coherent manner, although it is not required to do so in the present sounder experiment.

Synthesizer Programmers

The Hewlett-Packard frequency synthesizer is controlled by digital decade counters. Two programmers exist, one to control the transmitted signal waveform and receiver local oscillator and one to generate a phase correction waveform. The required control signals are applied to the synthesizer through three fifty-pin connectors. Each pin corresponds to one push-button switch on the front panel.. The block diagram of these circuits is shown in Figure 6.

The controller programmer is driven by a clock signal which causes it to step the synthesizer in a 100 hertz step for each clock pulse.

Since frequency changes must occur at zero phase points, all timing must be precisely controlled. To achieve this timing accuracy, the programmers are synchronous ring counters. Low impedance high speed drivers drive the cable and synthesizer switches.

The switching signals are synchronized with the 10 microsecond carry delays from one decade to the next. The starting frequency selection must also be delayed 10 microseconds from decade to decade as must the direction of counting control (count up or down).

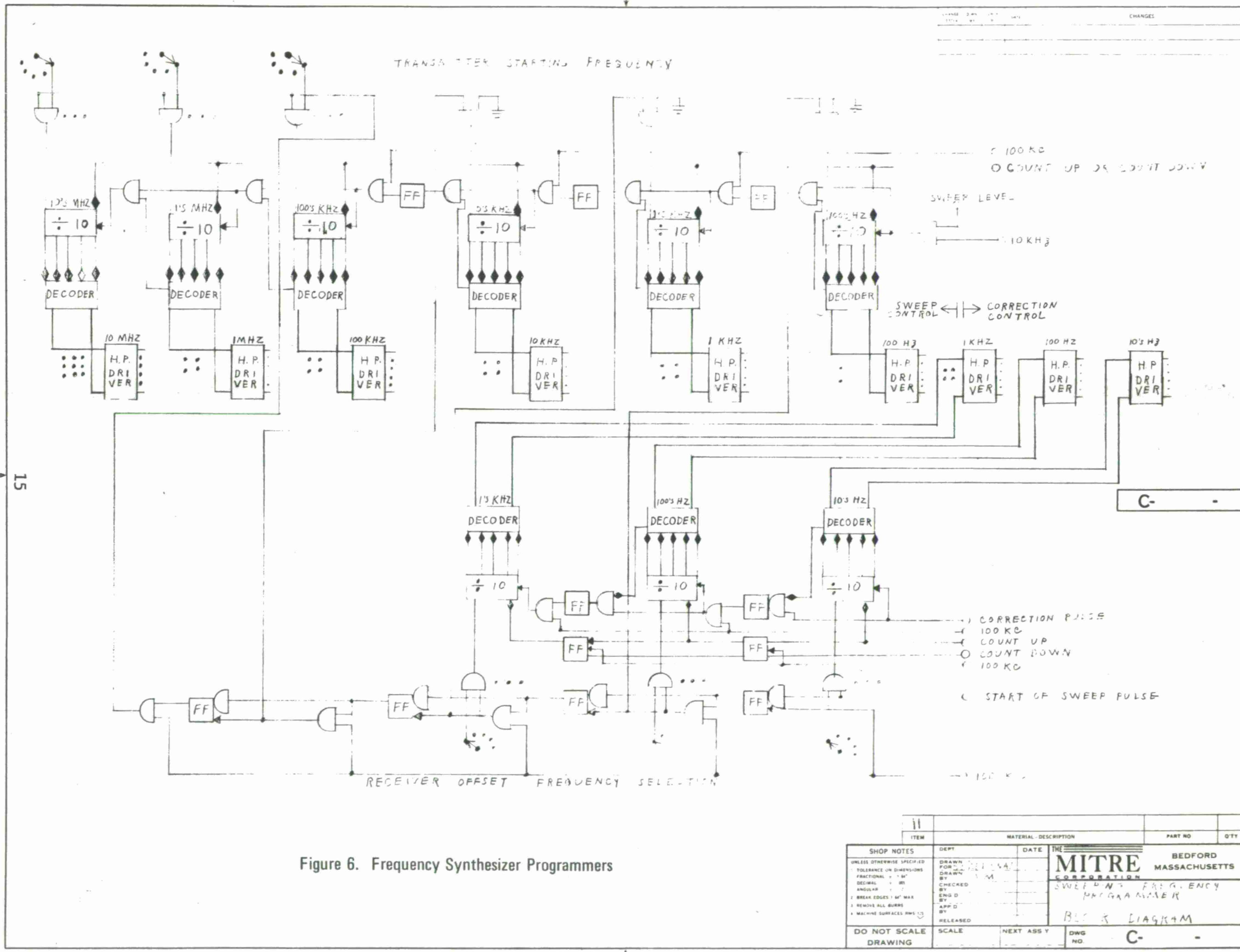


Figure 6. Frequency Synthesizer Programmers

Data Output Coding and Control

The analog signals from quadrature phase detectors are sampled and converted to digital form. The digital data is transferred through the output buffer register to the SDS 930 computer. In addition, the clock time at the start of each transmission is transferred to the computer. Appropriate coding is generated to allow the computer to recognize (1) raw data, (3) data which has been amplitude and phase corrected, (3) clock signals, and (4) to synchronize correction information from the computer back to the system.

Phase and Amplitude Correction Control (figure 7)

Phase correction is accomplished by quickly slewing the frequency synthesizer to a new frequency. Since the time required for the slewing is always a small fraction of the time between corrections, the synthesizer output approximates a step in frequency. Thus the phase progresses linearly between correction, arriving at the desired value (so as to compensate for the measured phase error) just as the next correction is applied. The phase correction word from the computer equals the number of 10 Hertz frequency steps required to bring the frequency synthesizer to the correct frequency. The word enters a buffer storage register. At a precise time, it is jammed to another holding register where it causes a counter to count the correct number of 100 KHz clock pulses. These clock pulses are synchronized and sent to the correction programmer. The direction of the frequency change is controlled by the sign of the correction word.

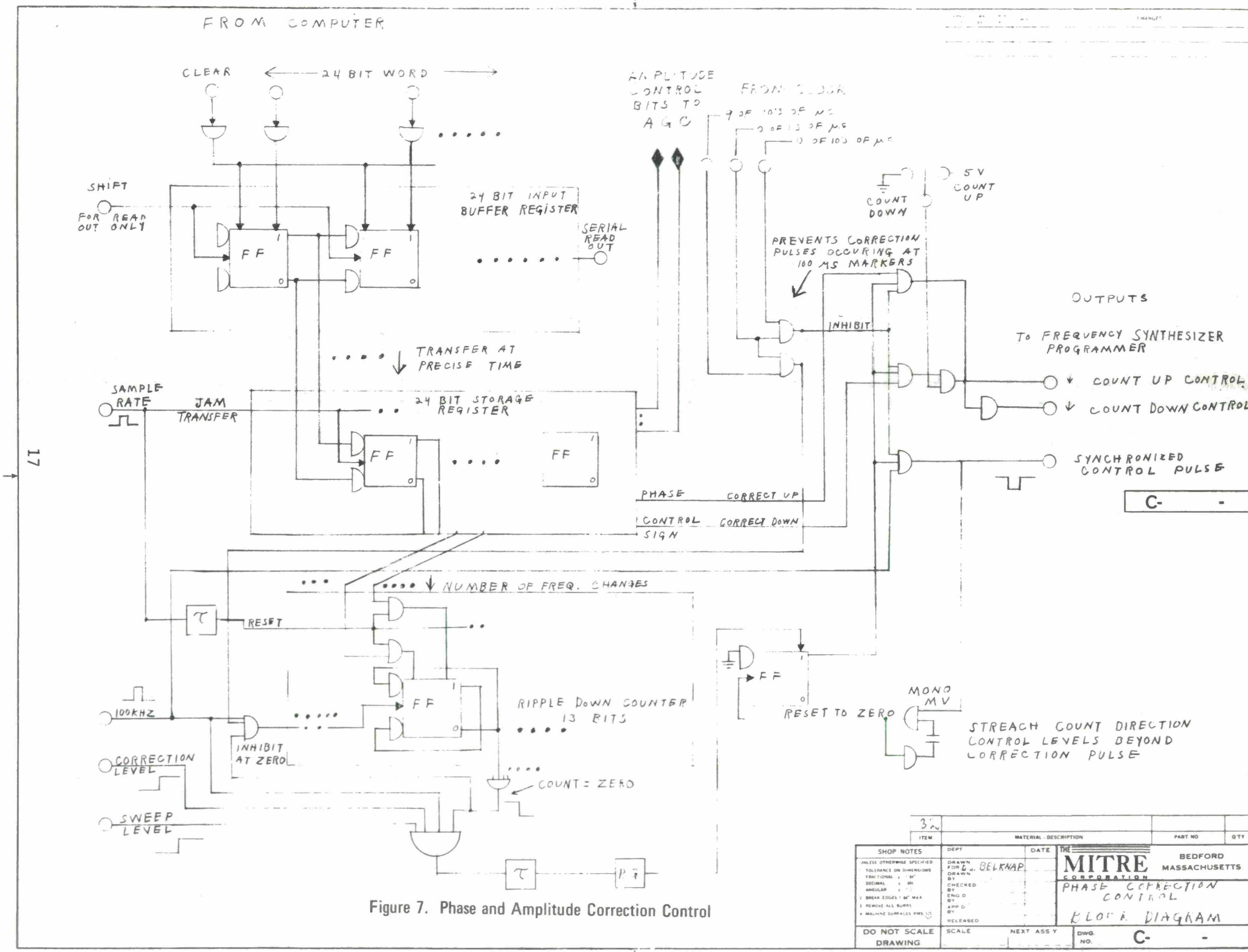


Figure 7. Phase and Amplitude Correction Control

The phase correction is computed as follows:

$$\Delta m_n = \frac{\phi e_n - \phi c_{(n-1)}}{2\,\pi\,\Delta f \Delta t} - \sum_{i=1}^{n-1} \Delta m_i$$

where Δm is the number to be sent to the synthesizer

ϕe_n is the phase residual in radians at sample n

$\phi c_{(n-1)}$ is the computed phase correction at sample (n-1) as explained below.

Δf is the frequency step (10 Hz)

Δt is the time between samples (2×10^{-3} sec)

and

$\sum_{i=1}^{n-1} (\Delta m_i)$ is the sum of previous correction words (after rounding off to nearest whole numbers).

(Δm_n is rounded off to the nearest whole number before being sent to the synthesizer.)

The actual phase correction, ϕc, differs from the phase error, ϕe, because of round off quantizations, and is computed by

$$\phi c_n = \sum_{i=1}^{n} (\Delta m_i) \quad (2\pi \Delta f \Delta t)$$

where Δm_i is now the rounded whole numbers for each of the previous corrections. Initially $\phi c_1 = 0$ and $\Delta m_0 = 0$.

The amplitude correction is also contained in the word from the computer. It is sent from the holding register directly to the appropriate gain controlled amplifier. Both linear gain control and a logarithmic gain control are available. The number which is sent to the equipment from the computer is determined as follows:

(1) Compute the average amplitude

$$A_{ave} = \frac{1}{N} \sum_{i=1}^{N} A_i$$

(2) For the linear gain case, compute the amplitude correction

$$A_{corr} = 32 \frac{A_{ave}}{A_n}$$

and if it is greater than 63 set to 63, where A_{corr} is the number (in 6 bit binary) sent to the gain controlled amplifier.

(3) For the logarithmic case, compute the amplitude correction

$$A_{corr} = 40 \log_{10} \left(\frac{A_n}{A_{ave}}\right) + 32$$

and if $A_{corr} > 63$ set to 63 and if $A_{corr} < 0$ set to zero. A_{corr} is a 6 bit binary number.

Analog Equipment

Transmitter and Antennas (Figure 8)

The power level at the output of the Hewlett-Packard frequency synthesizer is 20 mw. A 10 MHz video driver amplifier compensates for the losses in the 3 db power splitter at the synthesizer output (not shown), and the RF switch. The output power level is about 40 mw in normal operation. The 4 watt driver amplifier has a gain of 20 db and a 30 MHz bandwidth. It either drives one of the dipole antennas via the T/R Hybrid or the East-West Log Periodic Antenna via the final amplifier. The final amplifier is capable of 1KW peak or CW output power but was rarely operated over the 50 watt level as this provided a sufficiently high signal-to-noise ratio for all experiments conducted.

Four simple dipole antennas were erected, tuned at 5.5, 6.6, 7.8, and 8.6 MHz. All four antennas are fitted with stub-matching cables which provide a proper impedance match for most of

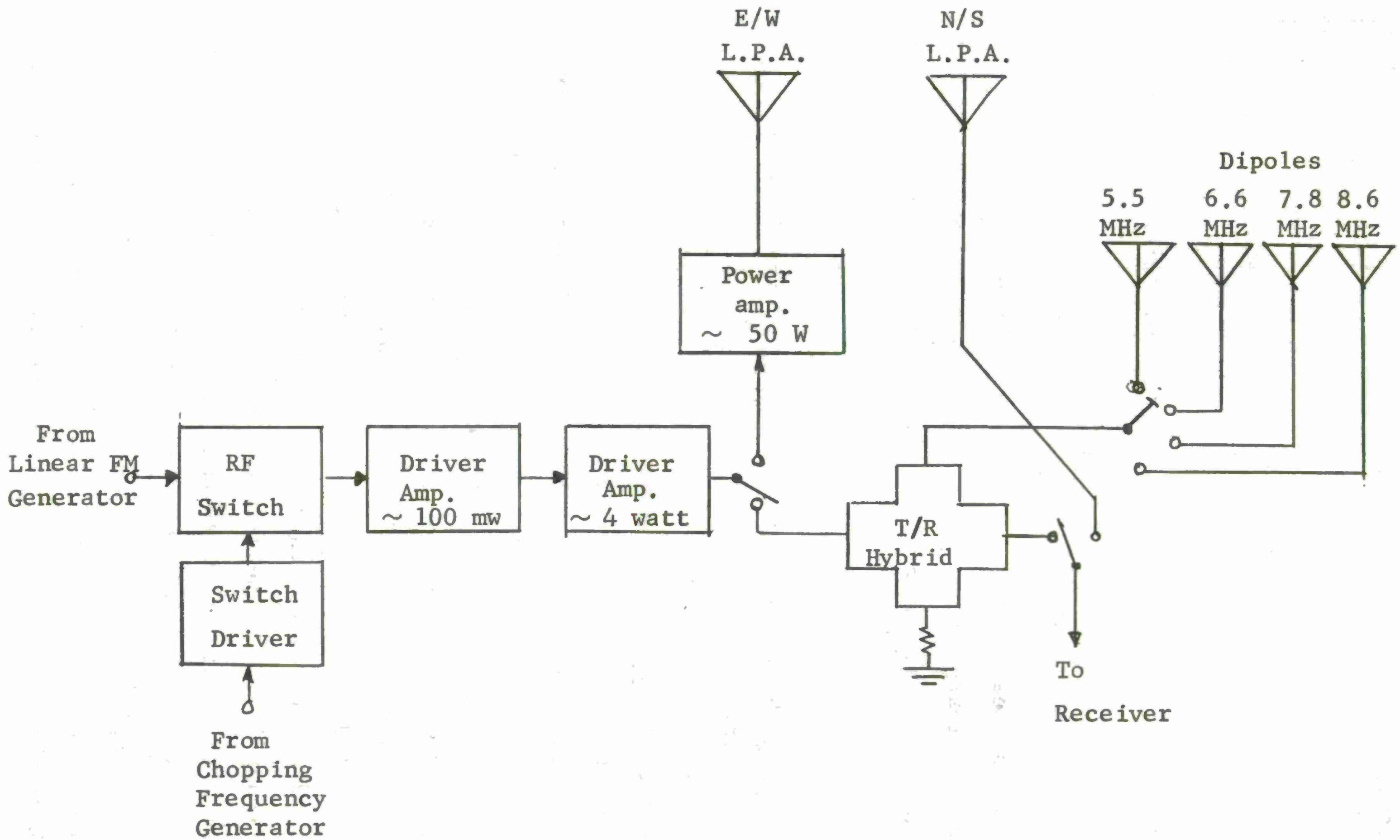


Figure 8. Transmitter Amplifiers and Antennas

the transmission line at the appropriate center frequency. The resulting worst point VSWR's over $\pm$150KHz either side of bandcenter are 1.3, 1.3, 1.45 and 1.28 respectively. The isolation afforded by the T/R hybrid is 35 dB when all ports are terminated in 50 Ω. In the actual system, isolation of 30 dB was measured.

Later in the project an alternate antenna/transmitter arrangement became available. It consists of two cross-polarized log-periodic antennas and a 1 Kilowatt amplifier having a 1.0 MHz "flat" bandwidth and a mechanically adjustable center frequency. The isolation between the two antennas was measured as 60 dB. The VSWR of this system, while not as good as the dipole set, was less than 1.5 everywhere between 4 and 10 MHz for each antenna.

Receiver-General Description:

A block diagram of the receiver is shown in figures 9 and 10. The requirements of coherent and stable frequency operation were accomplished by deriving all local oscillator frequencies from the stable 1 MHz frequency standard in the Hewlett Packard synthesizer. The entire receiver is of solid state design, including the power supplies.

An rf switch is included to protect the receiver from the high power transmit signal. This switch is turned off during transmission of the FM signal to prevent the receiver from saturating. Another rf switch is included for increased receiver isolation from the transmitter as well as other transmissions in the area. These switches provide greater than 100 dB of isolation.

The two synthesizer outputs are combined as shown in figure 10 so as to produce a signal at the correlation mixer which contains the linear FM deramping local oscillator as well as the appropriate phase corrections. Since the receiver IF is 18 MHz it is necessary to translate the output of the synthesizer containing the phase corrections from 3 MHz to 18 MHz. For this purpose a 21 MHz local oscillator is derived by way of a x21 frequency

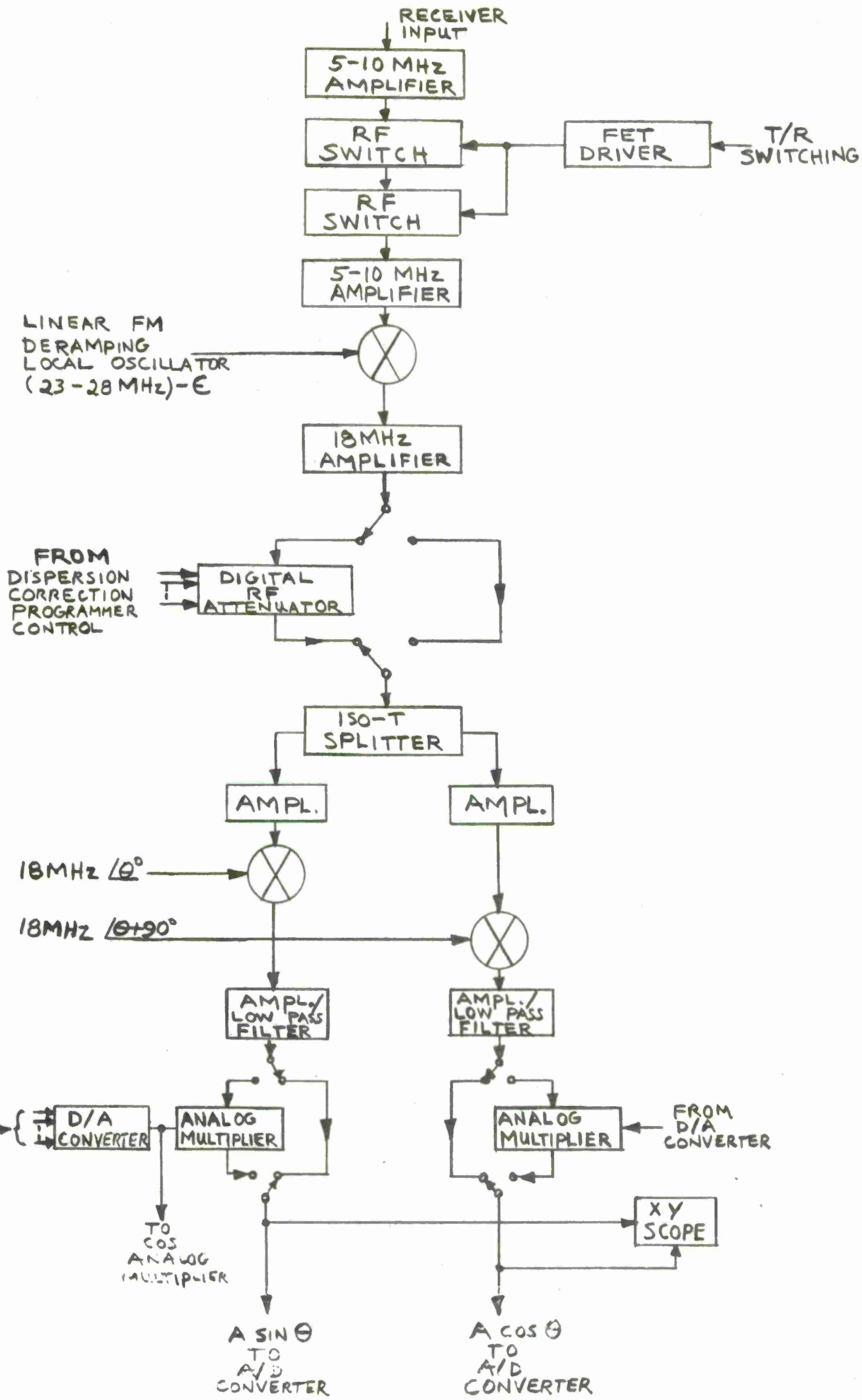


Figure 9. Receiver

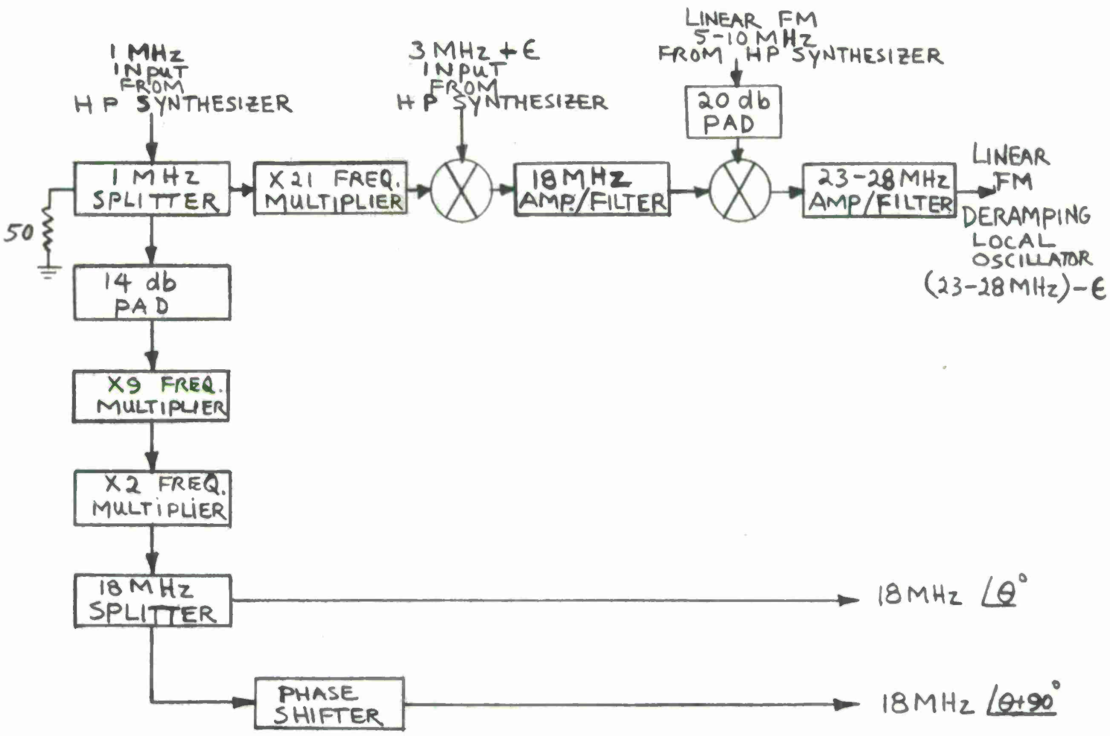


Figure 10. Frequency Synthesis for Receiver

multiplier and a synchronously-turned narrow band IF from the synthesizer driver's 1 MHz reference frequency. The correction signal at 18 MHz is then mixed with the linear FM output of the synthesizer and the desired sideband selected by a 23 to 28 MHz amplifier. Measurement of the amplitude and phase of the 18 MHz IF signal is made with a pair of quadrature phase detectors. The two output signals are A cos θ and A $\sin\theta$, where A is the true amplitude and θ is the phase of the received signal with respect to the local oscillator signal. The local oscillator in this case, is derived via a x 18 multiplier and tuned amplifier from the 1MHz reference.

The low pass filters consist of two cascaded third-order Butterworth filters having a sharp cutoff and a flat band-pass region. The amplitude response is flat to 0.1 db from DC to 100 Hz with a -3 db bandwidth of 150 Hz. The response is down 60 db at 500 Hz. The bandwidth of these filters can be changed by increasing or decreasing the RC time constant of each filter stage.

A six bit digitally controlled rf attenuator can be inserted after the 18 MHz narrow band amplifier to compensate for ionospheric amplitude errors. The attenuation is controlled by the Dispersive Correction Programmer in binary steps of 0.5 db; maximum attenuation is 31.5 db. An alternate method of compensation was also constructed. This second method uses a D/A converter and low frequency analog multiplier. The output signal level is kept constant by digitally controlling the amplitude of the reference input to the multiplier, (D/A converter output). Detailed descriptions of the more critical sections of the receiver follow below.

RF Amplifiers

The basic i.f. amplifier building block used throughout the receiver is a broadband feedback pair or doublet. This unit utilizes both source/load impedance control and feedback to achieve stable power gain. The feedback makes the gain virtually independent of transistor parameters, and was chosen to make the gain also independent of the load. The feedback pairs are hybrid circuits consisting of integrated circuits in combination with discrete thin-film deposited resistors.

The principal parameters of interest include:

1. 3 db bandwidth of 1 KHz - 100 MHz.
2. Gain stability, 1 db over a temperature range of -55 degrees C, to 100 degrees C.
3. Power gain per stage, 10 to 25 db per stage depending on feedback.
4. Linear small signal power capability, 0 dbm.
5. Noise figure, 8 db.

Frequency Multipliers

The reference oscillator is the 1 MHz frequency standard from the Hewlett Packard synthesizer. Two steps of frequency multiplication, x9 and x2, are used to achieve the 18 MHz reference frequency used for the phase detectors. Similarly, two steps of multiplication, x7 and x3, are used to generate the 21 MHz reference. The power gain of each multiplier is approximately one. The parameters of interest include short term stability and harmonic rejection. The multiplier characteristics that contribute to excellent short term stability include: 1) over-driven class A amplifier/harmonic generator, 2) class A voltage amplifier and buffer, and 3) magnetically shielded transformers.

The harmonic suppression, (greater than 60 db), referenced to the output frequency was achieved by interstage and output filtering.

RF Switching

The r.f. switch described here can meet most signal processing needs from 200 KHz to over 100 MHz and compares very favorably with, and in many cases exceeds, the performance obtained in a special-purpose switch.

High speed switches are often plagued by coupling of the gating signal to the output. The result is an output signal that rides on a "pedestal". Several methods have been employed to eliminate this spurious signal; a bridge circuit can be used to balance out the undesirable gating signal. However, even in pedestal free circuits, extraneous signals in the form of voltage spikes are often coupled to the output. This is caused by gating signals which are not precisely matched in shape as well as amplitude.

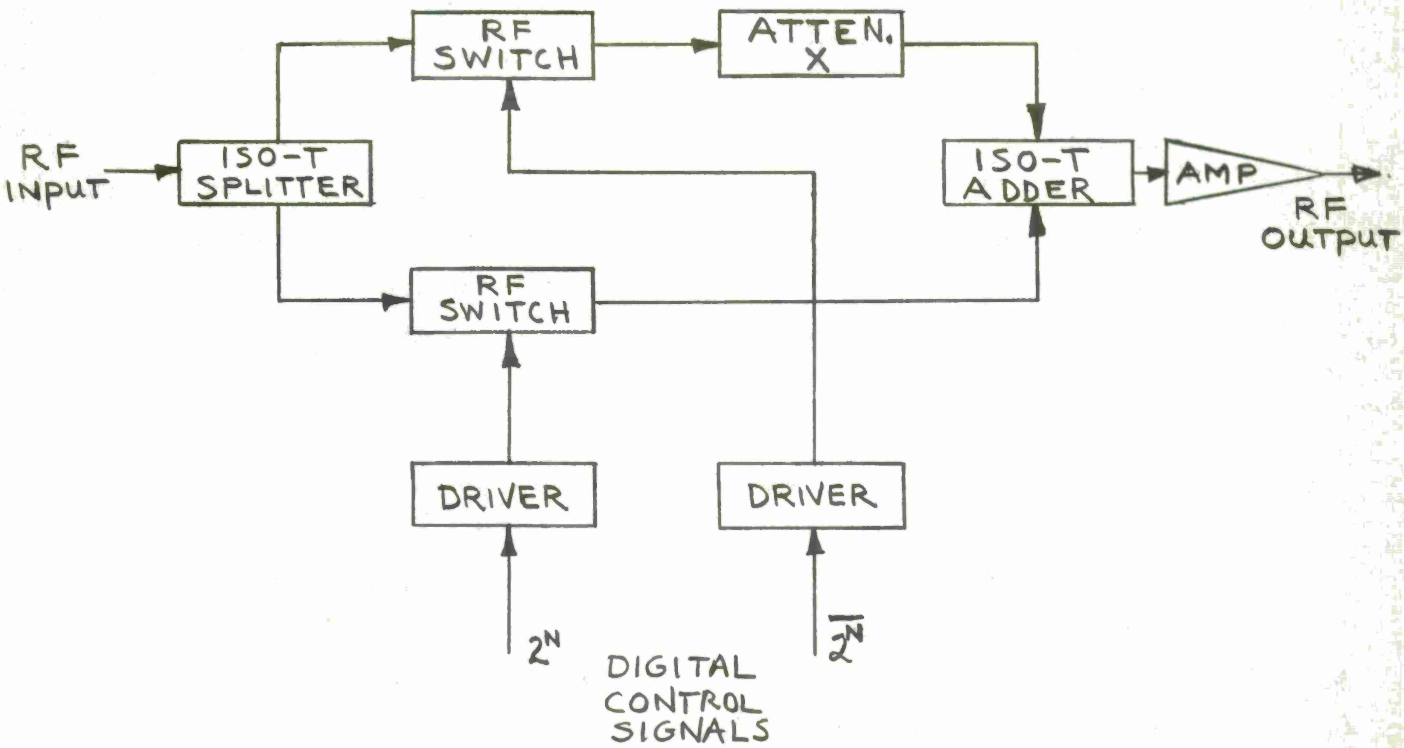
An r.f. switch functions as a variable impedance connected between a source and a load. In diode gates the impedance varies from a few ohms, when the diodes are forward biased, to a few megohms, when the diodes are back-biased. The ratio of the output voltage when the gate is on to the output voltage when the gate is off is one characteristic of the diode gate's performance. Other desirable output characteristics are small gate transients and fast rise and fall times.

The rf switch used in the receiver and transmitter consists of a double-balanced mixer and field effect transistor driver. The switch functions as a variable impedance connected between the source and load. The amount of signal passing through the mixer from the L port to the R port is a function of the DC control current inserted at the X port. Maximum attenuation is achieved with no DC current and corresponds to the isolation of the mixer. Minimum attenuation is achieved with a DC control current of 10 ma. or greater. Performance of these switches include a greater than 60 db on-off ratio and an output rise time of less than 20 nanoseconds.

Digital RF Attenuator

The digital attenuator provides a method of applying amplitude corrections to the ionospheric data. The attenuation is digitally controlled in 32 binary steps of 0.5 dB (max. attenuation 31.5dB) by the dispersive correction programmer, thus providing a logarithmic gain control. The attenuator can be used at rf frequencies from 500 KHz to greater than 100 MHz. It is presently being used at 18 MHz.

The digital attenuator's basic elements consist of balanced ISO-T's for splitting and adding the rf signal and rf switches for selecting the appropriate path, i.e., attenuation or no attenuation. The rf switches are the same as those used in the front end of the receiver. The pads or attenuators consist of precision resistors in a "T" configuration. Wide band amplifiers are inserted between attenuator stages to compensate for the loss in the ISO-T's and rf switches. Cascading six of these stages in a binary steps of 0.5dB (i.e., 0.5dB, 1.0dB, 2.0dB, 4.0dB, 8.0dB and 16.0dB) provides a maximum attenuation of 31.5 dB for a control signal N= 111111 and an attenuation of 0 dB for N= 000000. A block diagram of a typical section of the attenuator is shown in figure 11.



SECTION N	ATTENUATION X
1	0.5 db
2	1.0 db
3	2.0 db
4	4.0 db
5	8.0 db
6	16.0 db

Figure 11. Digital RF Attenuator Typical Section

SECTION IV

ON-LINE PROGRAM FOR CONTROL AND RECORDING

The on-line program required for operation of the ionospheric correction equipment is designed to ease the installation of modems and the remoting of system hardware should this become desirable. This capability is built in by requiring all input data to be coded to indicate its type. There are four codes provided, time, sounding data, correction data required, and corrected data.

The data recorded is all stored on magnetic tape for further processing. Block diagrams of the programs are shown in Figure 12, 13, 14, 15, and 16. The recorded output is stored in a buffer region according to the following format:

Title, Date, Etc.

	Time of Sweep
Data	Sound Data: A_s, A_c, A, ϕ
Block	M, B, A_{corr}(I)
	Sweep data: A_s, A_c, A, ϕ

The data block repeats several times to fill an ITABLE, then the ITABLE will be written on the output tape while an alternate buffer is filling.

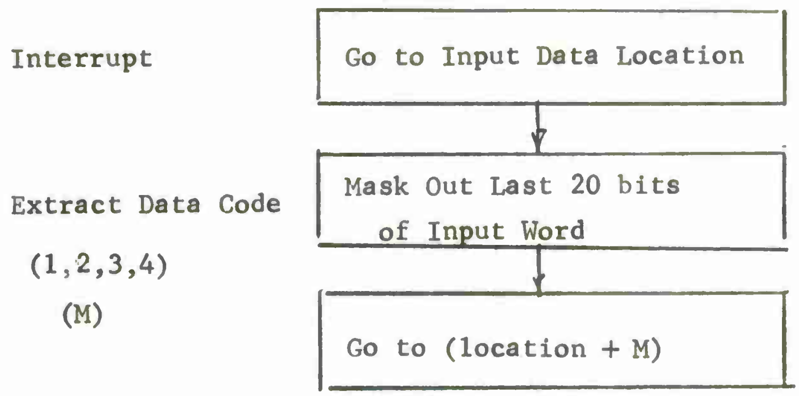


Figure 12   Interrupt and Data Code Logic

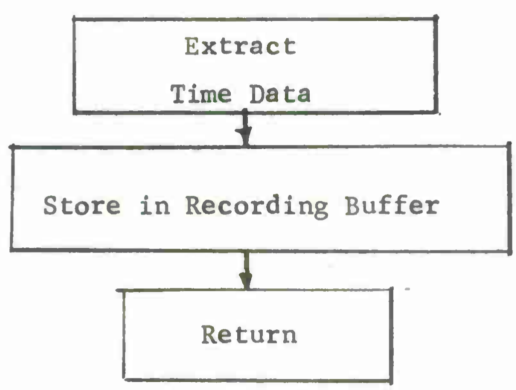
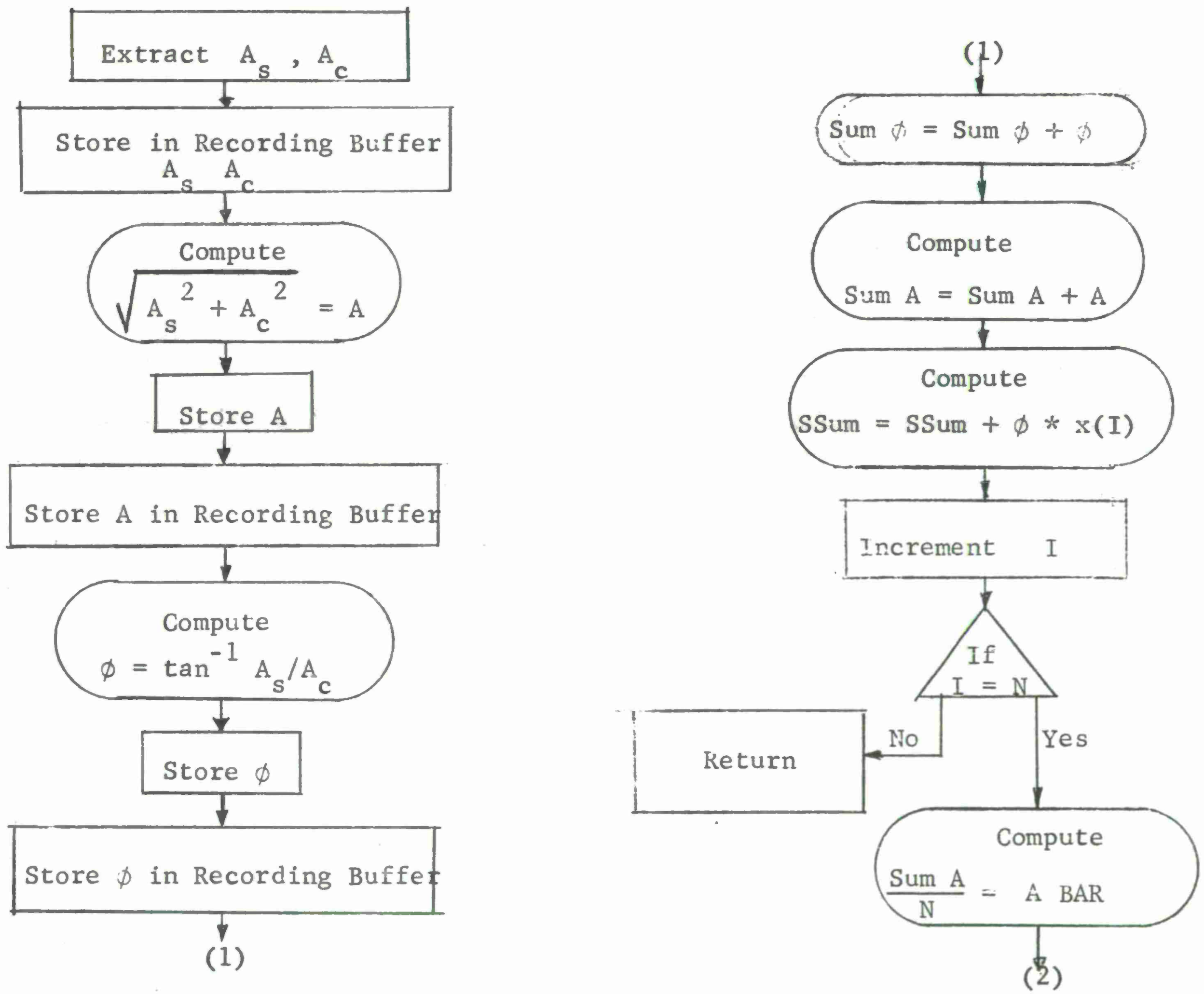


Figure 13 Block Diagram for Time Input Code



Extract A_s , A_c
Store in Recording Buffer A_s A_c
Compute $\sqrt{A_s^2 + A_c^2} = A$
Store A
Store A in Recording Buffer
Compute $\phi = \tan^{-1} A_s/A_c$
Store ϕ
Store ϕ in Recording Buffer
(1)
(1)
Sum ϕ = Sum ϕ + ϕ
Compute
Sum A = Sum A + A
Compute
SSum = SSum + ϕ * x(I)
Increment I
If I = N
No
Yes
Return
Compute
$\frac{\text{Sum A}}{\text{N}}$ = A BAR
(2)

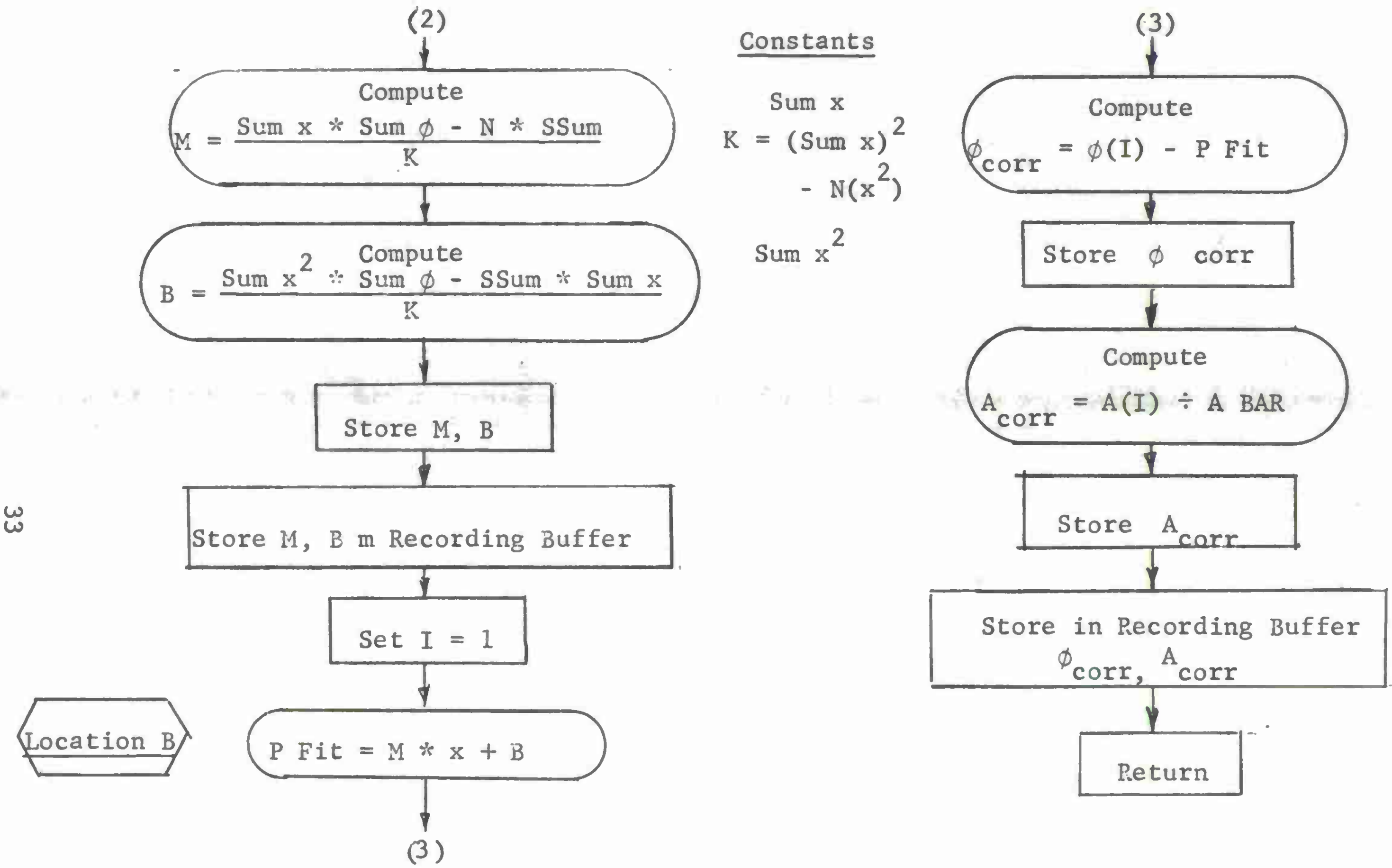


Figure 14. Correction Computation

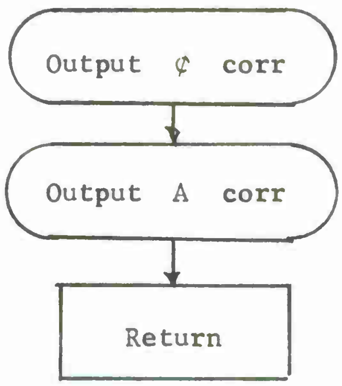


Figure 15 Data Output

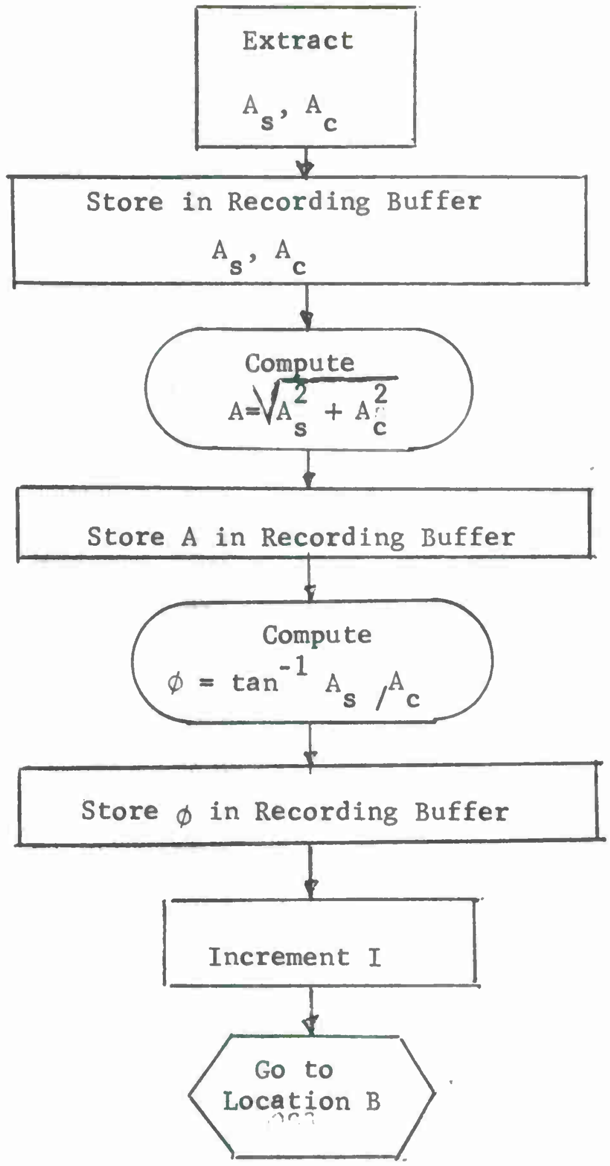


Figure 16 Recording Program

SECTION V

RESULTS

The recorded output of the system, consisting of samples (in phase quadrature) of the uncorrected and corrected sweeps was further processed off-line. This processing consisted of a computation of the Fourier transform of each sweep via the so-called "fast-Fourier-transform" algorithm (no weighting was employed). Three examples of the results obtained to date will now be described.

In the first example, the signal was swept from 4.5 to 4.85 MHz at a rate of 500 kHz per second. This sweep rate maps the 300 Hz receiver bandwidth into a 600μ sec range gate. As shown in Figure 17, the upper record is the transform of an uncorrected sweep and the lower record is the transform of a real-time corrected sweep, one second later. The ideal resolution of 2.2 μsec is indicated on the plot. Figure 18 shows an ionogram taken with a vertical sounder at Millstone Hill, Mass.* at the time the data were recorded. The vertical scale is 100Km/division or 667μsec/division.

The second example, shown in Figure 19 is for a sweep from 5.1 MHz to 6.0 MHz. The sweep rate of 1 MHz per second results in a 300 μsec range window. Again the upper record is the transform of a corrected sweep one second later. The ideal resolution is now 1.1 μsec. Both the uncorrected transform and the ionogram of figure 20 indicate that returns via both ordinary and extraordinary ray paths were present within the range window. The elimination of the dual return on the uncorrected transform is quite advantageous in applications where "sorting" and/or "deghosting" problems exist.

In the third example more than one corrected sweep is shown. In this case the sweep extended from 5.3 MHz to 5.75 MHz and was executed at a rate of 500 kHz per second. Again the range window was

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* These vertical ionograms were furnished courtesy of Dr. J. V. Evans, Lincoln Lab.

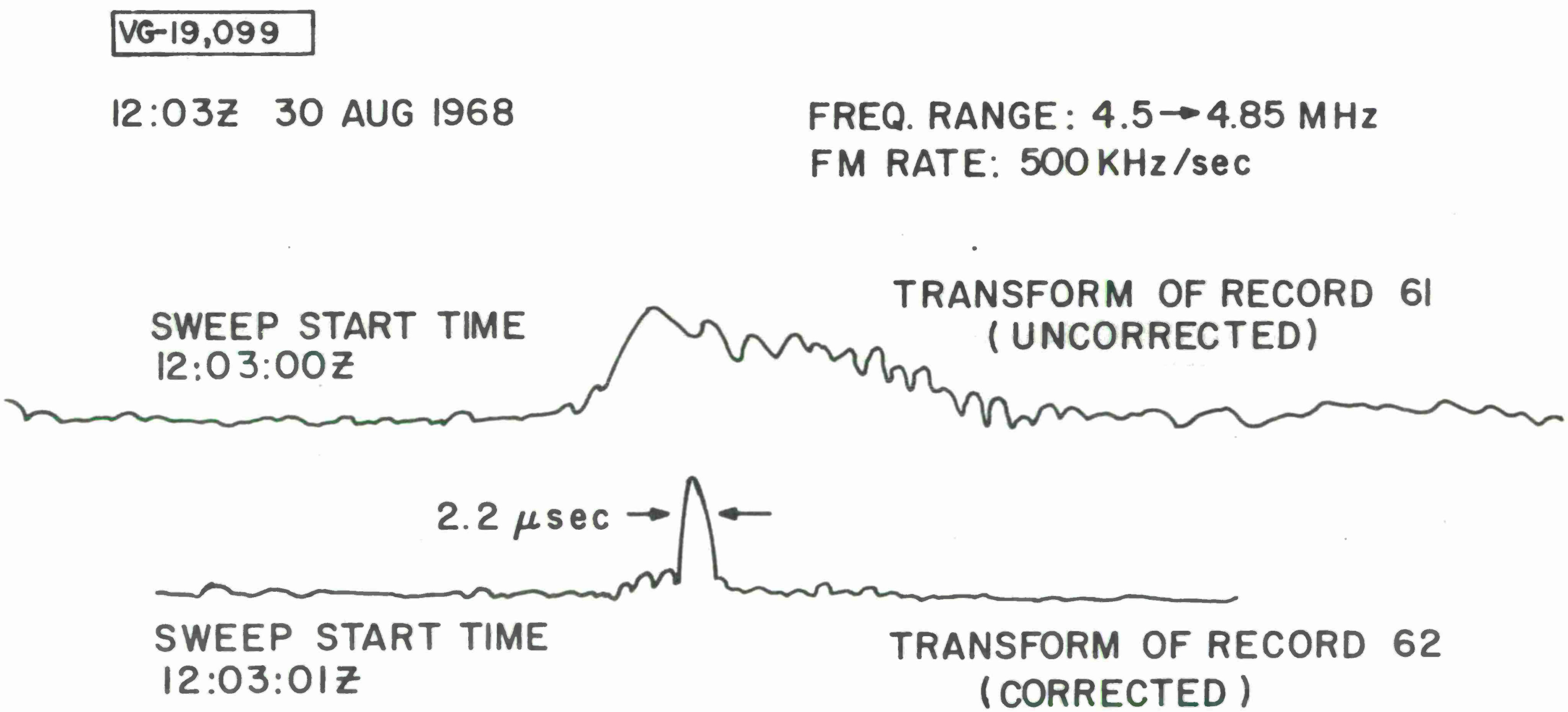

Figure 17. TRANSFORMS : FIRST EXAMPLE

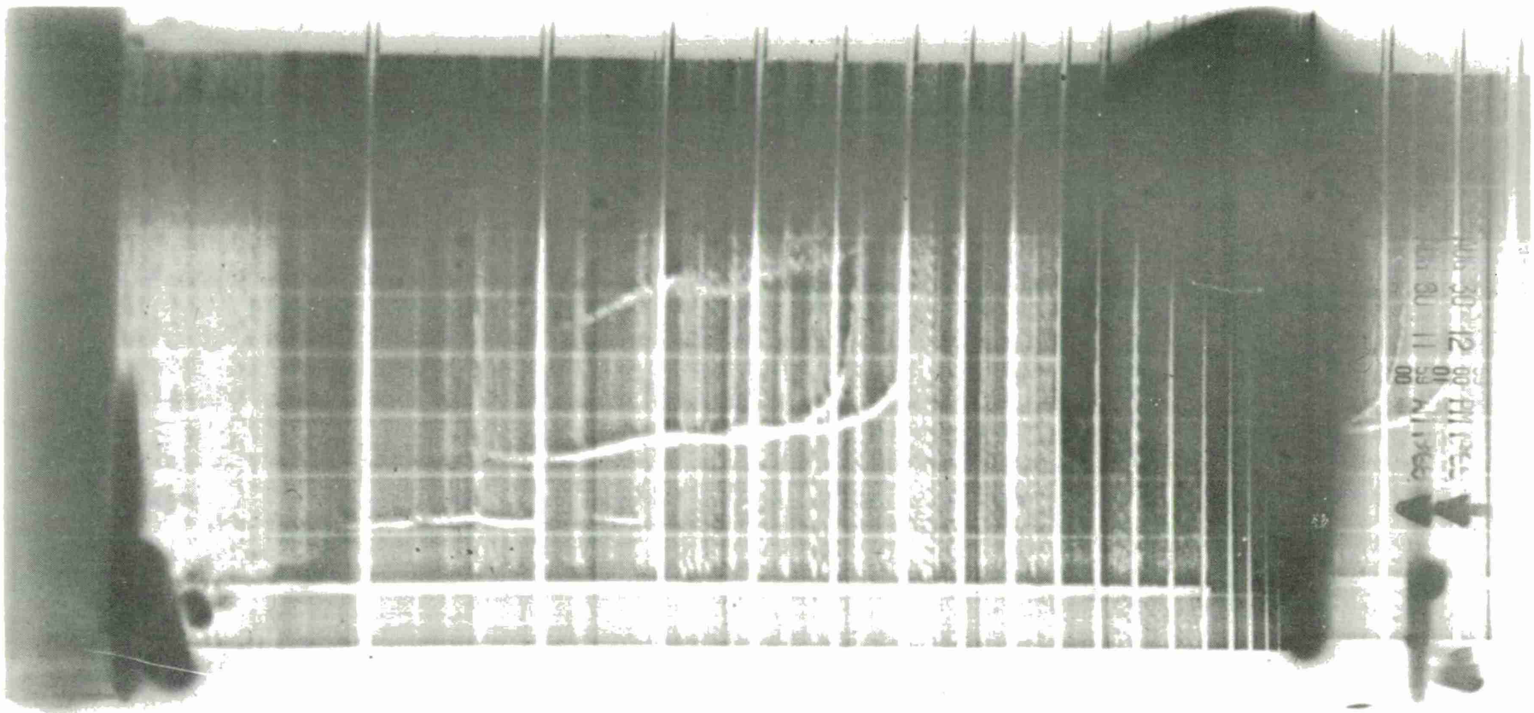

Figure 18. IONOGRAM 12:00 Z 30 AUG. 1968

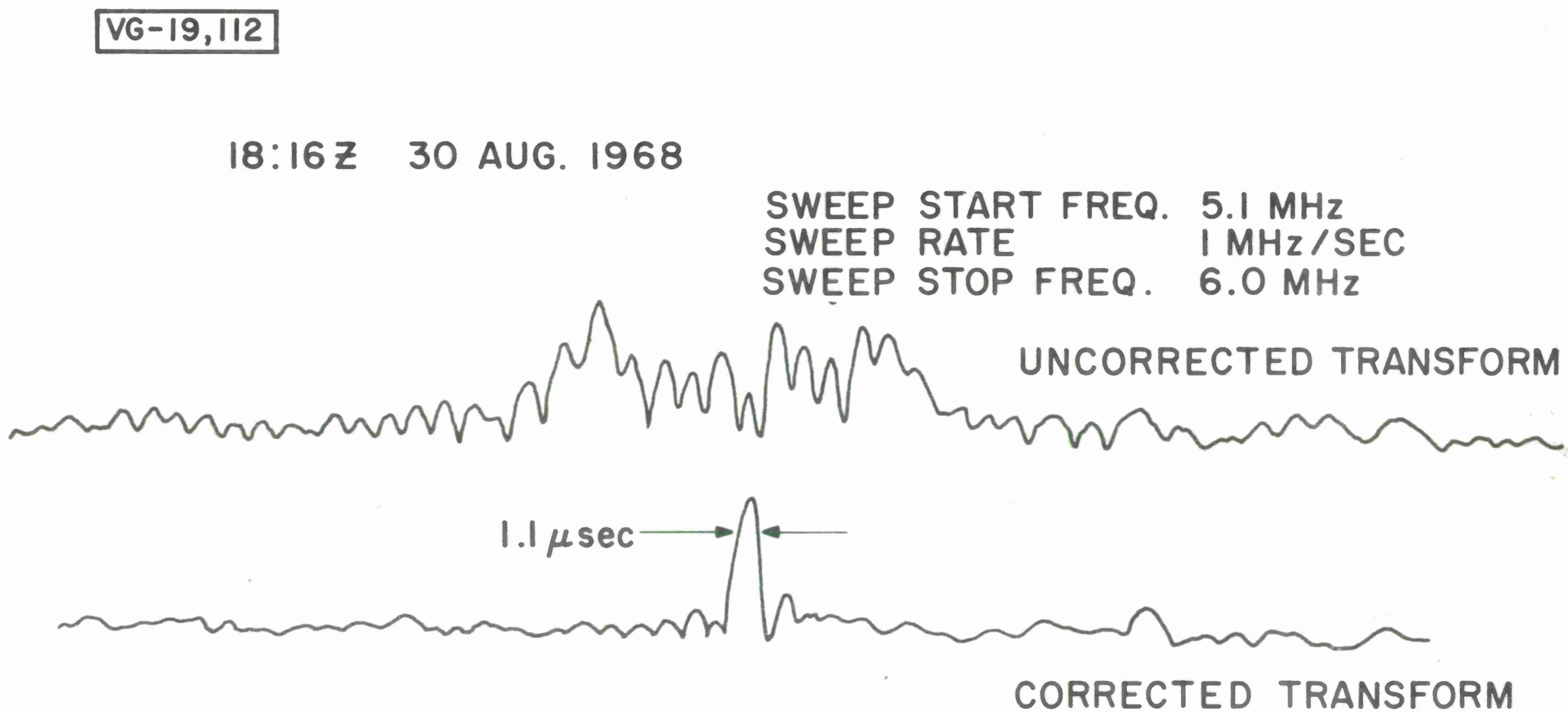

Figure 19. TRANSFORMS: SECOND EXAMPLE

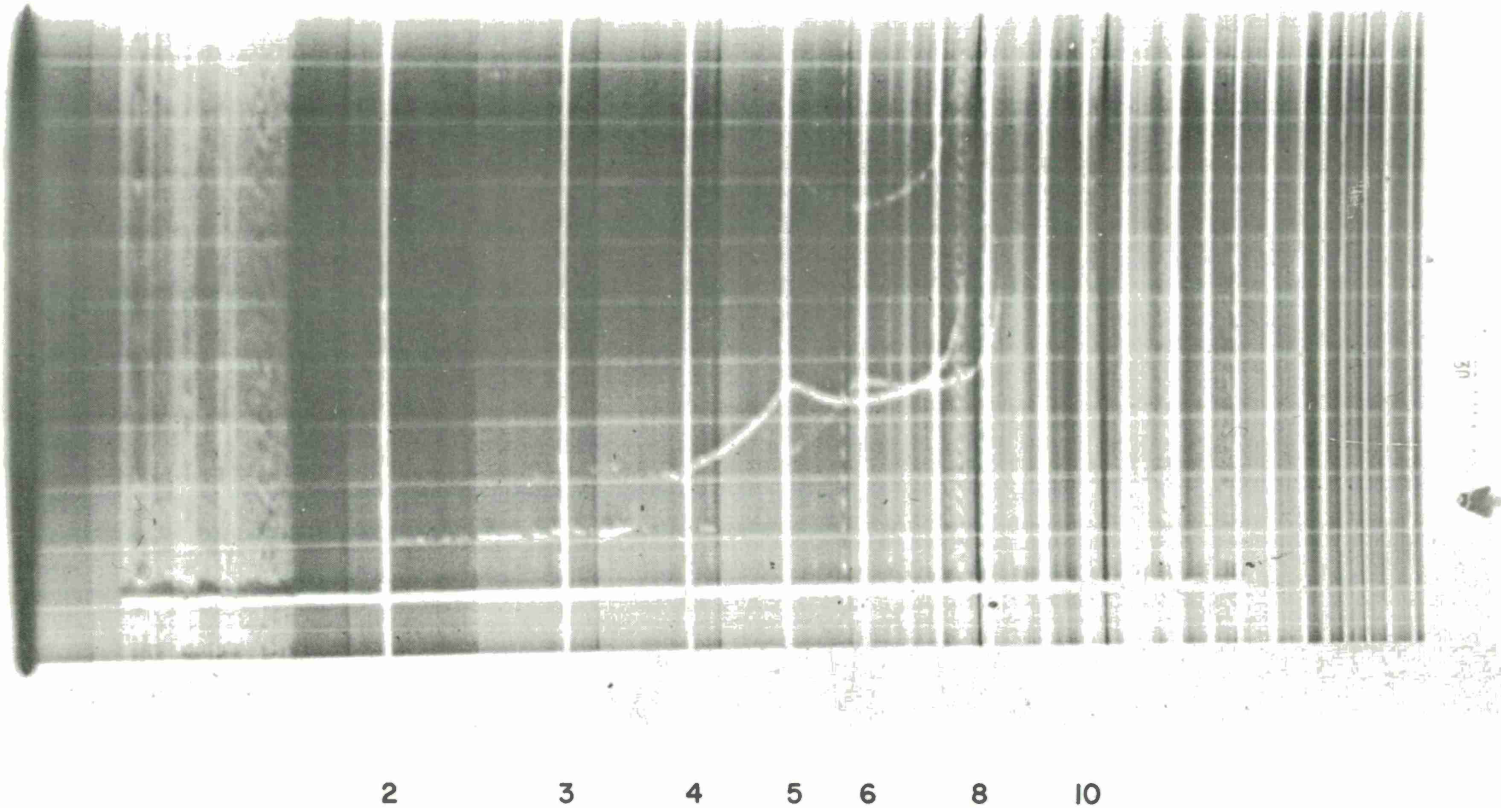

Figure 20. IONOGRAM 18:00 Z 30 AUG 1968

600 μsec and, as can be seen in the ionogram and uncorrected transform, two returns are present. The lowest record of Figure 21 is that of the first sweep in the sequence, and is an uncorrected transform at +10 seconds. Transforms of corrected transforms one, three, five and seven seconds later are shown above the first. In each case the corrections used were those computed from the first sweep.

These examples demonstrate the capability of the equipment to make a real-time correction of a vertically-directed ionospheric path over bandwidths of up to 900 kHz and time periods of several seconds.

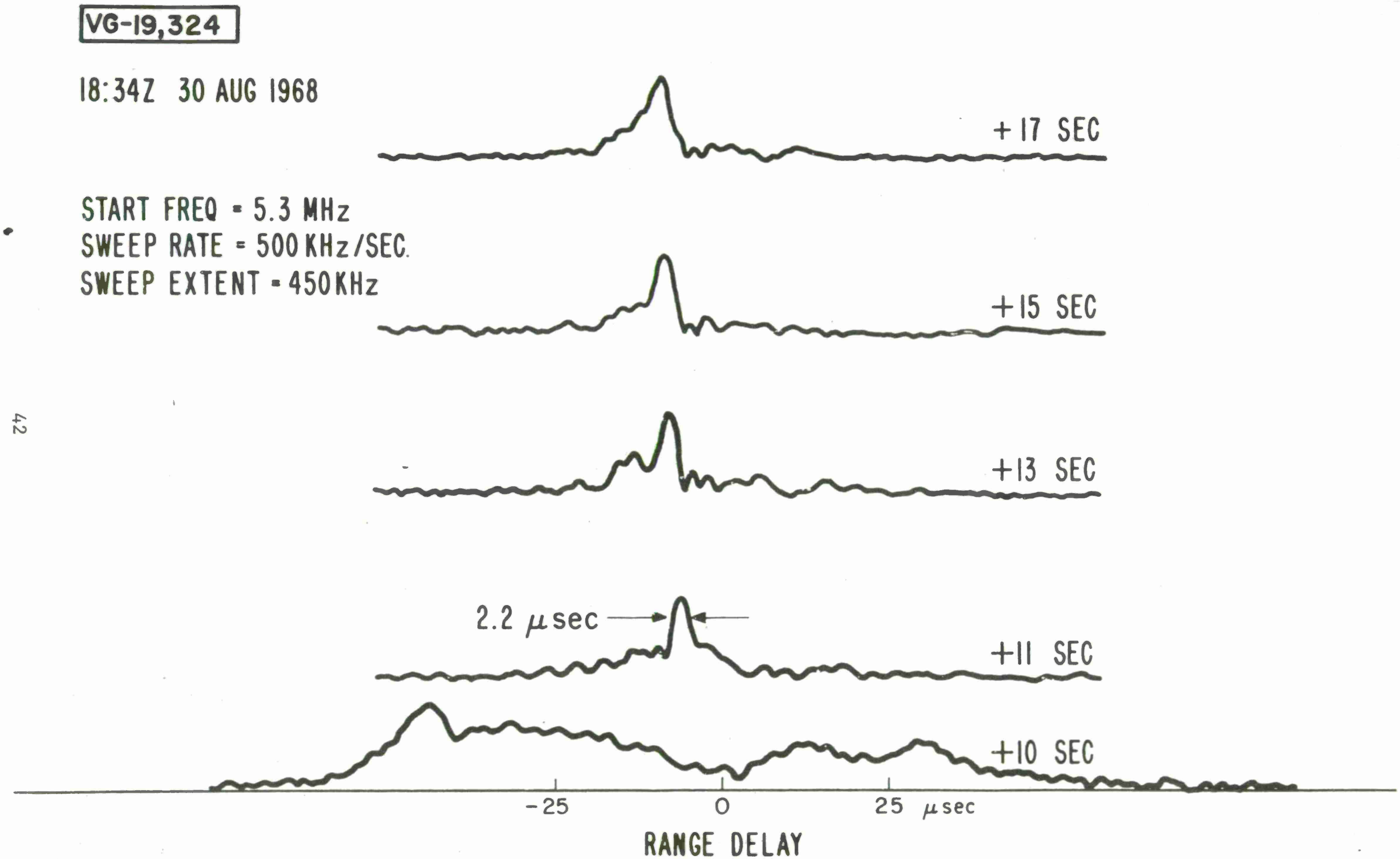

Figure 21. TRANSFORMS : THIRD EXAMPLE

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1. D. J. Belknap, R. D. Haggarty, B. D. Perry; "Six Months Report on Activities Conducted for project 7160 by Members of the Advanced Techniques Subdepartment". MITRE Working Paper WP-1444, 3 May, 1967.

2. D. J. Belknap, R. D. Haggarty, B. D. Perry; "Adaptive Signal Processing for Ionospheric Distortion Correction". MITRE Technical Report MTR-746, 1 August 1968.

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13. ABSTRACT

A linear FM vertical sounder has been designed and implemented in order to demonstrate the feasibility of real-time correction of ionospheric distortions. The system is monostatic, relying on chopping for transmitter/receiver isolation. All signal waveforms, local oscillator and correction signals are derived from a specially-modified, programmed Hewlett Packard frequency synthesizer. Results indicate a capability of making real-time corrections over bandwidths of up to 900 kHz and time periods of several seconds.

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