AFCRL

AD627879

# VISTAR - A PULSE DOPPLER AIRBORNE WEATHER RADAR


Prepared by: G. Kinzer

Motorola Inc. Military Electronics Division
Western Center
Scottsdale, Arizona



Contract No. AF 19(628)-2382

Project No. 6672




FINAL REPORT

20 January 1964



Code 1

Prepared for

Geophysics Research Directorate
Air Force Cambridge Research Laboratories
Office of Aerospace Research
United States Air Force
Bedford, Massachusetts

AFCRL

WF-2923

# VISTAR - A PULSE DOPPLER AIRBORNE WEATHER RADAR

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ABSTRACT

This Abstract is Unclassified.

The weather surveillance objectives for the VISTAR airborne weather radar are briefly described. The basic operational requirements for VISTAR are outlined and a signal flow presented. Detailed system and circuit operation are given and calibration procedures noted. (Schematic drawings are not a part of this document.) Criticisms of the system are presented and recommendations given. A block diagram is included.

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SECTION I

# 1. INTRODUCTION

This report constitutes the final report document required for Air Force Contract No. AF19(628)-2382 covering the contract period 1 October 1962 through 31 March 1963, (as amended).

## 1.1 PURPOSE OF PROGRAM

The program is aimed at providing a high altitude airborne Doppler radar instrumentation suitable for gathering and recording data from vertically moving air masses containing precipitation or hail, that are associated with storm phenomena. The instrumentation concept is that of utilizing a downward-pointing antenna beam to intercept the storm cloud formation while the aircraft moves along a track over the formation. The backscattered microwave energy from the vertically moving water and/or ice particles contained within the cloud formation is accumulated within the receiver processing system for subsequent Doppler analysis, display, and recording. This process, repeated many times over storm formations, will then yield records of vertical profiles of the storm evolution, over various tracks, in terms of height, intensity, and velocity. These records, when correlated with other simultaneous instrumentation of the same storms, will help to formulate more exact models of storm development.

## 1.2 USE OF ASTAR EQUIPMENT

Motorola had previously developed for the Air Force an experimental Doppler processing radar for an application different from this program requirement. That program, AF Contract No. AF33(616)-5114, resulted in a breadboard equipment known as ASTAR. It was determined in discussions between Dr. David Atlas of AFCRL and Motorola engineering personnel that there existed in the residual ASTAR equipment sufficient basic capability to undertake a program for the conversion of ASTAR to VISTAR, the code name for the required radar instrumentation. Accordingly the subject contract was awarded to Motorola about 1 October 1962.

## 1.3 CONTRACT SUMMARY

Preliminary effort on the contract was divided between defining and ordering long lead time components and a study, in conjunction with AFCRL, of the applicability of several aircraft for the mission. In the second effort, three aircraft were studied in some detail; these were: RB-47, RB-57, RB-66. The RB-66 was eliminated and the RB-57 was further explored. It was finally decided that, because of the severe difficulty in repackaging equipment, the RB-57 would not be suitable. Hence, the RB-47 was selected for the vehicle.

A considerable effort was placed upon the mechanical aspect of the ASTAR modification, in order that the equipment could be fitted into the available space in the selected RB-47. This was especially true with respect to the antenna assembly. A special parabolic dish and feed were procured in order that the existing radome assembly of the RB-47 could be utilized without modification.

While electrical modifications were in process on the equipment, detailed drawings of the mechanical layout of the system were prepared for AFCRL and ASD, the cognizant agency for the aircraft and installation of equipment therein.

Because of a radical change in the repetition frequency for the new application, a new high power modulator was required. A subcontract was let to the Quantatron Company for this modulator development. Unfortunately, the modulator produced was not acceptable and a new subcontract had to be let to the Magnetic Research Company for the modulator. This modulator proved to be unusable when installed in the field and was returned to Magnetic Research for modification. It has since been returned to AFCRL, presumably in operating condition.

Meanwhile, because of the modulator and other technical difficulties of modification, together with an increased requirement for safety-of-flight in terms of wiring, the program for installation in the aircraft and flight data gathering was cancelled for 1963.

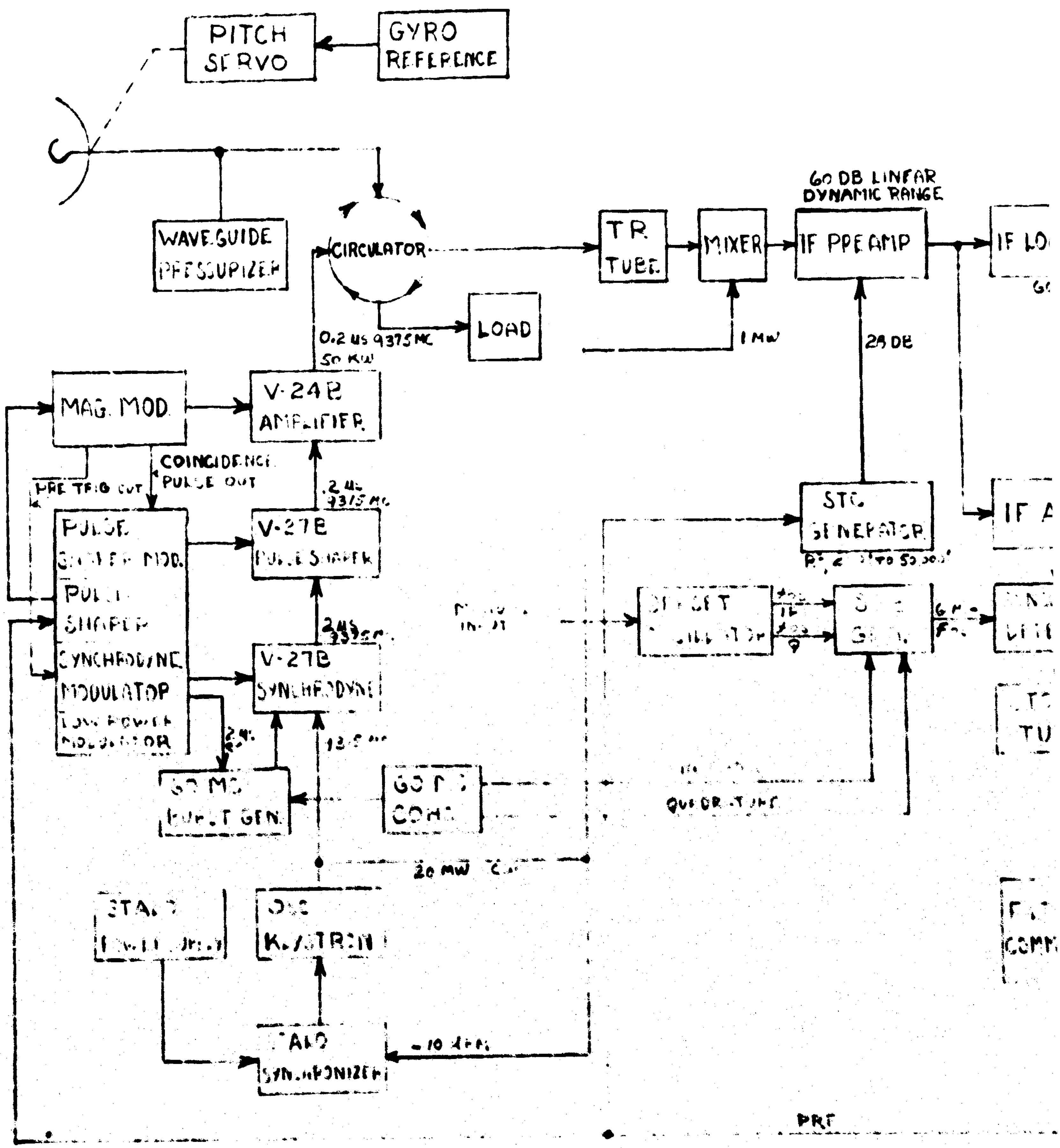
The modification program continued until 30 June 1963, at which time the VISTAR equipment was shipped to AFCRL, Field Weather Station, Sudbury, Mass. At this time the greater portion of the modifications had been performed and the receiving and data processing portions of the system had been operated. The transmitter had not been operated, however, due to the lack of a successful modulator and also because of the ultimate failure of both the residual V-24B power output klystrons used in the transmitter.

A new V-24C to replace the V-24B has been ordered by AFCRL but has not been delivered as yet.

### 1.4 REPORT CONTENTS

The principal objective of this report is to serve as a working document for personnel required to work with the VISTAR equipment. As such, it contains no theoretical analyses of problems of a fundamental nature concerning the application of VISTAR to the weather surveillance mission, but instead dwells on the descriptions of subsystems and circuits contained in the VISTAR system.

Descriptive material is grouped in two major sections: Functional Description, and Detailed Description. The Functional Description gives a view of the requirements of the over-all system plus a signal flow analysis. The Detailed Description is concerned with subsystem and circuit operation. For comprehensive use of this material, reference must be made to the several schematic drawings, graphs, technical memos, etc, that have been included with the VISTAR equipment delivery. These drawings have not been included here because of their bulk. However, a block diagram (Figure 1) has been included for general reference and orientation in the system.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MAG. MOD
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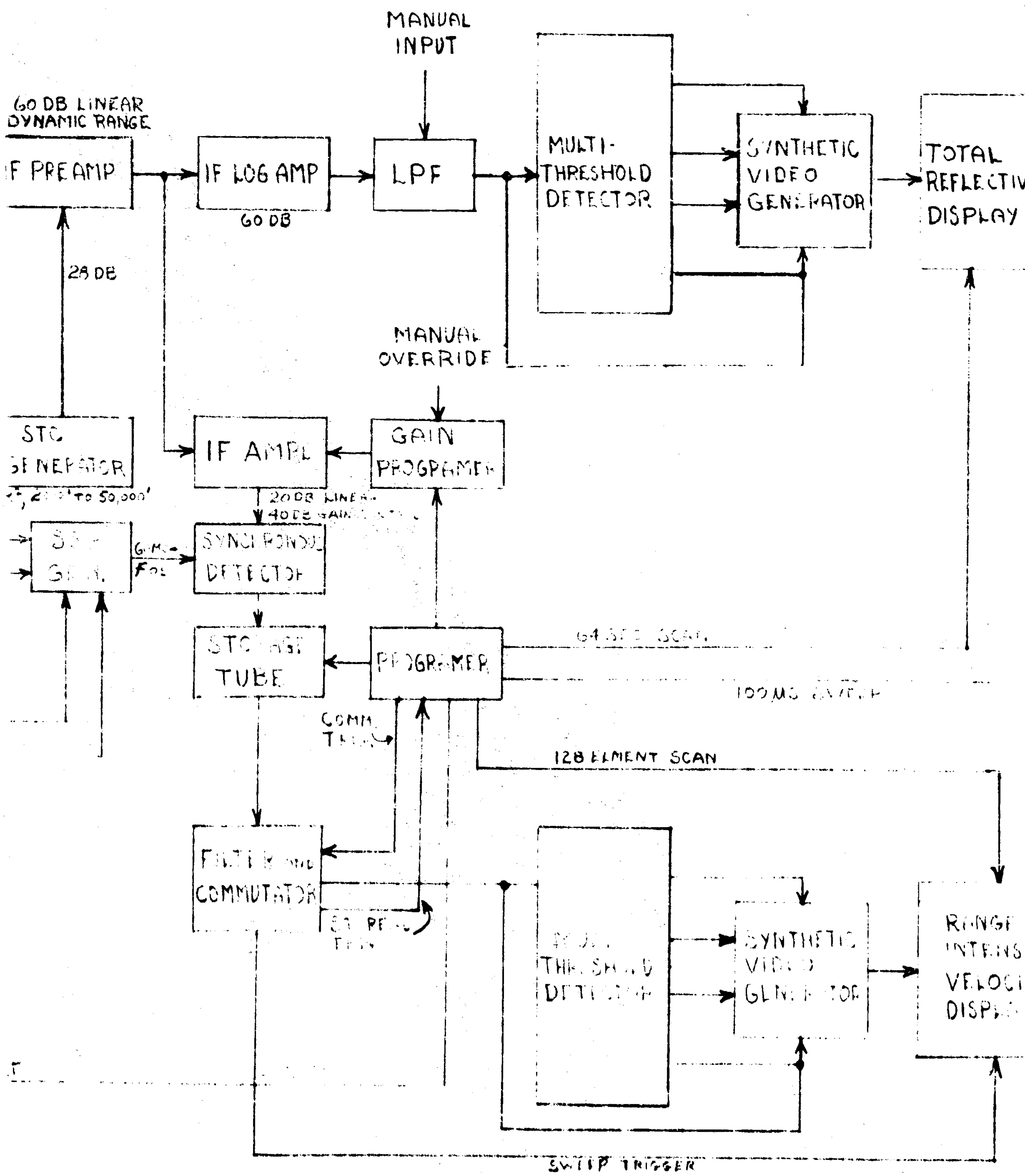

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SYNCHRONOUS DETECTOR
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PROGRAMER
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128 ELMENT SCAN
FILTER AND COMMUTATOR
THRESHOLD DETECTOR
SYNTHETIC VIDEO GENERATOR
RANGE
VELOCI
SWEEP TRIGGER

## SECTION II

## 2. FUNCTIONAL DESCRIPTION OF VISTAR SYSTEM

### 2.1 BASIC REQUIREMENTS OF THE VISTAR SYSTEM

Several aspects of the VISTAR mission impose constraints upon the instrumentation philosophy employed in the system. In particular, the fact that the velocity spectrum analysis is performed from a high speed aircraft results in a time limitation on data accumulation from a given spatial volume. Multiple range element processing, also a requirement, contributes to the instrumentation complexity. Considered as a whole, the mission requires a three-dimensional analysis consisting of range, intensity, and velocity, in which each dimension has a finite range of interest and desired resolution.

In view of the airborne application and complexity of the analysis, some of the more straightforward Doppler analysis techniques must be eliminated from consideration. For example, the single range-gated filter that must be swept in both range and frequency is hopelessly slow for this use. It was from these considerations that the applicability of the ASTAR Doppler processing philosophy was studied. The ASTAR processor has the theoretical capability of performing an amplitude versus Doppler analysis for many range increments with a <u>single</u> set of transmitted pulses of microwave energy, i.e., the number of pulses usually required for <u>one</u> range element taken at <u>one</u> frequency in the case of a range-gated single filter. This type of processing is made possible through use of a storage-tube analog memory and a "comb" filter for Doppler frequency analysis. The basic requirements of operation may now be outlined from the viewpoint of using the ASTAR radar system to serve as the foundation of the required instrumentation.

A pulsed, coherent X-band transmitter and receiver system is utilized to acquire the Doppler data samples to be processed. Details of this portion of the radar are presented in a later section.

The composite weather signal received by a pulsed, coherent radar must undergo the following processing elements to provide the required estimate of range, intensity, and velocity:

1. Adjustment of a sampling rate (prf) to yield unambiguous samples of all Doppler frequencies of interest.

2. Range separation (time gating) of the received signal into elements of required length (range resolution).

3. Accumulation of signal samples (at each range element) such that sufficient energy is contained for each Doppler frequency of interest as represented by the ordered collection of samples.

4. Integration of signal energy (for each range element and each Doppler frequency element) over enough samples to yield an estimate of signal amplitude to the required accuracy.

5. Separation of Doppler frequencies into calibrated elements (filtering) of the required resolution (for each range element).

6. Prepare the processed signal such that it is suitable for display and/or recording.

The VISTAR modification of the ASTAR system is designed to accommodate the above-listed requirements as follows:

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1. The sampling rate (prf) is adjusted to allow unambiguous measurements of a 50 meters-per-second target velocity, which at X-band (9375 Mc) results in a Doppler frequency of 3127 cycles per second. The prf is set accordingly at 6880 cps, which is twice the highest Doppler frequency (required by the sampling theorem) plus a safety margin.

2. Range separation in VISTAR is accomplished in the storage-tube memory by "cross-reading" the stored input data at constant range intervals. Since the line-raster used to store data received from the radar represents range lines, which are in registration, any line orthogonal to this set constitutes a line of constant range for all lines of the raster.

3. The accumulation of signal samples for the required Doppler analysis is provided for by the storage-tube memory as mentioned above. The line raster is composed of stored received data, one line for each transmission, laid down at equally spaced intervals. The line length represents 50,000 feet, the range required by the VISTAR mission. A total of 440 lines are written on the storage tube, which provides at least two cycles of the lowest non-zero Doppler frequency of interest. It should be noted that this number of cycles is less than that required for good spectrum distribution, i.e., the $\frac{\sin \omega}{\omega}$ frequency distribution of the pulse of Doppler frequency generated when the storage tube is read out at any given range element. However, the number of stored cycles increases with Doppler frequency and the spectrum distribution improves correspondingly.

4. Integration of signal samples in VISTAR is provided by the filters used for the spectrum analysis. By using high

Q ceramic filters driven by high impedance circuits, use is made of the initial portion of the exponential transient response of the filter to the on-frequency signal, to approximate a true integration of the signal energy. The number of samples integrated is determined by the 440 lines in the written raster, plus any signal averaging in the range dimension caused by intentional low-pass filtering of the video signal stored from the receiver.

5. Separation of Doppler frequencies into components (spectrum analysis) is accomplished by a "comb" filter bank system employing 99 fixed-frequency ceramic crystal filters. Because of certain constraints on the filter system, special handling of the signals derived from the storage tube are required. In fact, to allow flexibility in the absolute calibration of the spectrum under analysis, modification of the received radar signal prior to storage is performed. Details of these procedures are given in the text following this tabulation. Basically, they consist of frequency translation of the received spectrum prior to storage for calibration purposes, and frequency multiplication of the stored spectrum to drive the filter bank at a higher frequency.

6. Several steps are employed in VISTAR to prepare the filtered Doppler for display and recording. Since, for each ensemble of 440 transmissions, a complete three-dimensional set of data is produced, a recording scheme must operate to preserve this data. In VISTAR, a crt/camera system is utilized in which a single frame of exposure constitutes a range/intensity/velocity analysis of the 440 transmitted pulses. The two linear dimensions of the frame are range and velocity, with the contrast dimension used for intensity. Because of the restricted dynamic range of the crt and for an aid in data interpretation, a special amplitude coding is employed for intensity.

To supplement the range/intensity/velocity analysis (riv), a second channel of data processing known as a "total-reflectivity" (t-r) channel is instrumented. The purpose of this channel is to provide a calibrated record of the intensity of the composite (incoherent) return signal, in which all 440 transmissions (for each riv analysis) are integrated to form one line of the display and recording. A total of 60 db dynamic range has been provided for in this channel, to preserve the wide dynamic range of the weather signal. As with the riv system, an amplitude coding scheme is used to preserve the wide dynamic range. A logarithmic receiver is associated with the t-r channel; therefore, a "db" scale results from a linear interval coding of the output amplitude. The dimensions of the film record of the t-r display represent range and intensity as in the riv display, with the second linear dimension representing distance along the flight path. This produces a vertical profile "slice" of the storm area under surveillance. Timing is such that each frame of the t-r recording can accommodate approximately one minute of flight path, (64 transmission groups of 440 transmissions each starting at one second intervals).

Some special attention is now given to the Doppler frequency spectrum during processing. As mentioned in (5) above, the spectrum undergoes both translation and multiplication during the spectrum analysis. Translation is required for two purposes: (1) in order to recover simultaneously the Doppler frequencies caused by updrafts and downdrafts, i.e., frequencies both above and below the carrier frequency, the synchronous demodulation frequency must be "offset" sufficiently such that no frequency ambiguity exists in the output; and (2) a settable offset frequency allows the fixed-frequency comb filter spectrum to be placed in an optimum position relative to the actual Doppler spectrum. In the VISTAR system, the offset in frequency is accomplished with a settable offset oscillator together with a single sideband generator.

Multiplication of the spectrum (after translation) is required to allow the use of a set of higher frequencies for filtering. Since high Q filters are required in the comb filter bank, and since the Q must be proportional to frequency for equal channel frequency spacing and uniform amplitude response, the ratio of upper to lower frequency of the filter bank should be minimized. For a practical filter bank, this results in the requirement of a higher frequency band for the filters. In the VISTAR system, the comb filter is composed of two sets of filters ranging from 350 kc through 595 kc. These two sets of filters, when used with associated mixers and oscillators, provide an effective comb filter bank ranging from 35 kc through 595 kc.

Actual spectrum multiplication is accomplished in VISTAR by adjusting the readout time of the storage tube relative to the prf. The over-all relationship between Doppler frequency, $f_d$, and the output frequency of the storage tube memory used to drive the comb filter can be summarized in the following equation:

$$\text{frequency}_{out} = \frac{(f_d + f_{os})\ (1.1)\ (10^6)}{\text{prf}}$$

where $f_{os}$ = offset frequency.

Also, the change in output frequency due to a change in Doppler frequency or offset frequency is given by:

$$\Delta f_{out} = \frac{(1.1)\ (10^6)}{\text{prf}}\ \Delta f_d \quad \text{for Doppler changes}$$

or

$$\Delta f_{out} = \frac{(1.1)\ (10^6)}{\text{prf}}\ \Delta f_{os} \quad \text{for offset oscillator changes.}$$

These expressions are useful in calibrating the offset oscillator unit. They may also be normalized in terms of numbers of filters by knowing that the filters are spaced 5 kc.

A second mode of operation is utilized wherein the prf is reduced to one-half of normal, i.e., to 3440 pps. This allows a given width of Doppler spectrum to be spread over twice the number of filters as normal, as is evident from the above equations. In this mode, the offset frequency must also be altered to correctly translate the spectrum into the filter bank.

A final basic requirement of the VISTAR system results from the limited dynamic range of the storage tube used for the memory. Since the expected dynamic range of signal is in the order of 60 db, a sequential cycling of the system gain is effected in 20 db increments such that the total dynamic range can be covered in a three-step cycle. It should be noted that spurious Doppler sidebands will result when the signal level is high enough to exceed the memory amplitude capability on higher gain settings. However, it was deemed more desirable to accept this malady for portions of the processing rather than risk losing lower-level returns for lack of sensitivity.

## 2.2 SIGNAL FLOW: R-F AND I-F

The VISTAR equipment includes a pulsed, coherent transmitter and a coherent receiver operating in X-band at 9375 Mc. This section will discuss the general functions taking place in the transmitter and the receiver. The details of the operation and the circuits that perform them are described in a later section.

The cycling of the transmitter and receiver are rather unique due to the other signal processing operations in the VISTAR cycle. The transmitter is triggered once for each "write" sweep of the electrostatic storage tube which stores the received returns as a function of time in position which represents range on the storage surface of the memory. The next return from the succeeding transmission is stored adjacent to the first and so on until the storage surface is filled with 440 range lines. The transmitter then remains off until the next transmission burst which occurs at a one-per-second rate.

The transmitter can be triggered at either of two pulse repetition rates, nominally 6880 or 3440 pps. The triggers are supplied from the programmer which also controls the other VISTAR functions. The trigger is amplified and shaped to provide a 2-microsecond pulse to trigger the timing circuits in the high power (magnetic) modulator. After an internal timing delay, the high power modulator provides a trigger output to the low power modulator driving the synchrodyne klystron. During this 2-microsecond pulse, the low power modulator driving the pulse shaper klystron is triggered in coincidence with the high power modulator pulse to the output klystron.

The transmitter r-f reference is generated in a crystal stabilized reflex klystron whose output frequency is 9315 Mc. This klystron and the stabilizing circuit comprise the stable local oscillator (Stalo). The c-w output of the Stalo is pulsed and shifted 60 Mc in frequency in the synchrodyne stage whose output is a 2.0-microsecond pulse at 9375 Mc. The pulse shaper stage shapes and amplifies the portion of the synchrodyne pulse occurring after 1.5 microseconds with the resulting pulse duration, 0.2-microsecond, synchronized with the 0.3-microsecond high voltage pulse on the output klystron. The pulse out of the output klystron drives the antenna through a Rantec 4-port ferrite circulator. All of the klystrons are decoupled from each other using ferrite isolators. The transmitter amplifier chain is instrumented throughout with directional couplers and crystal detectors, to enable the observation of the various r-f waveforms in checkout and to aid in the adjustment of the klystrons.

The return signal from the antenna is directed into the receiver by the circulator. Because of reflections from the antenna which are also directed into the receiver arm of the circulator, protection of the detector mixer crystal is required. This is provided by a Microwave Associates MA 581 TR Tube. This tube has a typical recovery time of 3 microseconds, which is the controlling factor on the receiver dead time.

The received signal is combined with the Stalo output in an r-f detector and the difference frequency is amplified. This signal contains the 60-Mc difference frequency, originally used to shift the Stalo to 9375 Mc, and the Doppler frequencies of the radar scatterers. These Doppler frequencies, $f_d$, can be either positive, when the particles are approaching the radar, or negative when they are moving away from the radar. Therefore, the output of the first detector contains the frequencies

60 Mc + $f_d$ No inversion has taken place, since the input signals to the mixer are

(9315 Mc + 60 Mc + $f_d$) and 9315 Mc.

The Doppler frequency is related to the vertical velocity of the particles and the transmitter frequency by

$f_d = \frac{2V}{\lambda_c}$ where V = particle velocity in meters per second

$\lambda_c$ = carrier wavelength in meters = $3.1976 \times 10^{-2}$m.

The specifications for VISTAR require a velocity resolution of 0.5 meters per second or a Doppler frequency resolution of

$$f_d = \frac{2 \times 0.5}{3.1976 \times 10^{-2}} = 31.2734 \text{ cps per 0.5 meter/sec}$$

The VISTAR specifications call for normalization in range of the return signal. This normalization is such that any returns from the same target would have uniform amplitude no matter at what range it appeared within ranged from 2000 ft. to 50,000 ft. This normalization is obtained by reducing the i-f amplifier gain at near range (zero time) and then increasing gain with time after transmission. The power ratio of 2,000 to 50,000 is 13.98 db and, if the target is assumed to intercept the entire radar beam, the return varies with the square of the range, or, for this case, 20 db. The gain correction is performed on the input i-f preamplifier through the agc input. The control signal is derived in the programmer and is synchronized with the transmitter output pulse.

The range normalized coherent return at the output of the i-f preamplifier can have a linear dynamic range of 60 db, (the limits of the preamplifier). This output drives two i-f amplifier systems. One is a 60 db logarithmic i-f detector and incoherent video amplifier; the other a 20 db linear i-f amplifier. The log amplifier output is displayed as a function of time and produces a radar presentation of radar signal return as a function of range. The output of the linear amplifier contains the Doppler information. Because the storage tube and filter circuits have only 20 db linear dynamic range, the 60 db dynamic range is handled in 20 db increments by this amplifier. The programmer produces a gain control signal, as selected by the operator, that places this i-f amplifier in one of three 20 db portions of the input 60 db dynamic range. Signals above this portion are limited and those below are lost in the noise.

The output of the linear 20 db amplifier is coherent and does preserve the input signal, 60 Mc + $f_d$. This is synchronously detected by mixing the 60-Mc coherent reference signal, shifted a discrete frequency $f_{os}$, and 60 Mc + $f_d$ signal. The purpose of the $f_{os}$ frequency is to shift the Doppler into the frequency band of the filter bank used to analyze the velocity spectrum. The 60 Mc - $f_{os}$ signal is generated in the single sideband generator which combines 60 Mc and quadrature 60 Mc with $f_{os}$ and quadrature $f_{os}$ to obtain 60 Mc - $f_{os}$. The synchronous detector output goes to the input of the storage tube.

## 2.3 SIGNAL FLOW: STORAGE, FILTERING AND COMMUTATION

The storage tube receives the output of the synchronous detector and the data are stored on one line of a 440-line raster during each prf interval. Data received from the detector during any given prf interval represent samples of backscattered energy from weather targets. For moving targets that produce a Doppler frequency, $f_d$, the sample represents, in spectrum, $f_{os} + f_d$ for any particular range increment. The return stored is bipolar and phase coherent.

An accumulation of 440 such lines of data is stored in the storage tube in the form of an equally-spaced line raster. At the end of the data accumulation period, a data readout of the tube is performed by "cross-reading" the stored line raster at successive lines of constant range. This action produces a time function from the ordered collection of target phase samples, at a constant range, which is related to the Doppler frequency spectrum by a constant offset and multiplication factor, i.e., Storage tube output = $k\ (f_{os} + f_d)$. It is this modified frequency spectrum that is analyzed by the following comb filter bank.

The action of the programmer relative to the storage tube is to cycle the tube through the required modes of operation: ERASE, PRIME, WRITE, and READ. These modes are synchronized with other radar system functions. The programmer also generates sweep and scan waveforms for the storage tube deflections.

A comb filter system is driven by the storage tube output signal in a manner such that a spectrum analysis is performed for each "cross-sweep" of the stored raster. The filter system consists of 99 ceramic filters with equal bandwidths that are driven in parallel by the input signal. To recover the amplitude of the particular spectral components of the input signal, the output of each filter is rectified and sampled by a commutator. Storage for the commutator "wait" time is provided by the high Q characteristic of the filters themselves, that is, the filters "ring" after the drive energy is removed.

An electronic commutator is used to sample the filter bank output and thus form the "serialized" video for display. The commutator contains 99 gates that are associated with the filters. Each gate is enabled, in turn, by a pulse derived from a tap on a delay line string that has 99 taps. As a pulse is propagated through the line, each tap, in turn, is enabled by the pulse. The total commutation time is 50 microseconds. The gate outputs are summed to form a single video line for encoding and display.

From the block diagram, it may be noted that the commutator and programmer are linked in a closed loop for triggering purposes during the spectrum analysis part of the radar cycle.

## 2.4 DISPLAYS

Both the total reflectivity (t-r) display and the range/intensity/velocity (riv) display utilize synthetic video produced by multithreshold detectors and synthetic video generators. Each video channel, log i-f video for t-r display and commutator video for riv display, drives a similar threshold-video generator circuit. The purpose of these circuits is to encode the video signal according to amplitude such that the display cathode ray tube dynamic range may be more fully utilized for differentiating between signal levels.

In the t-r channel, the video output is logarithmic; therefore, equal spacing in the threshold circuits results in a decibel scale coding on the display. Five levels are employed, with the lowest level being a nonencoded application of the video to the display in order to preserve a realistic "noise-like" background.

The riv channel is quite similar to the t-r channel, except that the commutator video is linear and, therefore, the threshold settings must be adjusted logarithmically to produce a decibel scale at the display output. Five levels are also employed in the riv system.

The programmer provides sweep and scan information for the displays as well as film advance signals. For the t-r display, the sweep and scans are generated in the programmer proper, and also for the riv display scan. The riv sweep is generated in the display scope and is triggered by the commutator.

SECTION III

# 3. DETAILED DESCRIPTION OF VISTAR SYSTEM

## 3.1 TRANSMITTER AND RECEIVER

The VISTAR transmitter and receiver are a combination of oscillators, amplifiers, shapers, isolators, modulators, mixers, detectors, and associated timing circuits. The basic control is supplied by the programmer with internal timing circuits supplying the subroutines required. The receiver is considered up to the two i-f outputs; one a noncoherent video signal and the other a coherent video signal with a long-term characteristic representative of any Doppler shift which the r-f transmission has experienced. Various control signals are supplied by the control panel and the programmer which are considered in the discussion of the units.

The description of the circuits in the transmitter and receiver are grouped as follows:

Oscillators
- R-F Stable Local Oscillator (Stalo)
- 60-Mc Coherent Oscillator (Coho)
- Offset Oscillator

Amplifiers
- R-F Klystron Amplifier Chain
- I-F Amplifiers

Mixers
- Single Sideband Mixer (SSB)
- Synchronous Detector

Miscellaneous

Modulators and Timing

### Stalo

The r-f Stalo in the VISTAR transmitter is a stabilized reflex klystron oscillator operating at 9315 Mc. The STALO consists of the klystron, a klystron power supply, and a Dymec DY-2650A Oscillator Synchronizer. The reflex klystron is a Varian VA-242E reflex klystron which operates at the desired frequency. This tube is capable of 750 mw output power and is conservatively operated in this application to assure longlife and reliability.

The power supply for the klystron is a Hewlett-Packard 716A Klystron Power Supply modified for operation at a line frequency of 400 cps. This supply offers the superior regulation for operation from aircraft primary power along with d-c filaments, low noise, hum, and ripple.

The oscillator synchronizer controls the absolute frequency of the klystron by phase-locking the klystron oscillator to a crystal reference which has high short-term stability. The synchronizer is essentially a crystal-controlled superheterodyne receiver terminating in a phase comparator. A sample of the r-f output of the klystron is mixed with a harmonic of the crystal reference to produce an intermediate frequency of 30 Mc, which is compared in phase with the 30-Mc reference. For stabilizing the klystron, the resultant phase error voltage is added in series with the klystron reflector power supply voltage.

The r-f reference frequency is controlled by a quartz crystal, oven-mounted for temperature stability. The oven and circuit accommodate a fifth-overtone crystal ground for a frequency near 100 Mc, which gives a harmonic 30 Mc away from 9315 Mc.

60-Mc Coho

The 60-Mc oscillator is a doubler synchronized to a 30-Mc reference crystal in an oven. The oscillator stage drives a tuned amplifier and the result is 60 Mc at one watt power level.

Klystron Amplifier Chain

The Stalo output is a c-w r-f signal which must be frequency shifted so that when the received signal is mixed with the local oscillator signal in the first detector, the resulting carrier will be at the intermediate frequency. In order to achieve discrimination in range, the transmitted signal must be time modulated. In addition to these operations, the Stalo power must be amplified before transmission. All of this is performed in the klystron amplifier chain.

The frequency shift takes place in the first stage of the amplifier chain. The successive stages then serve to reject the unwanted sidebands of the mixing operation. A Varian V-27B used in a synchrodyne mode shifts the frequency from 9315 Mc to 9375 Mc. Synchrodyning occurs when the klystron beam is modulated and the output signal is equal to the input signal plus the modulation products. The beam modulation is obtained by mixing 60 Mc with the high voltage pulse applied to the cathode by the modulator. In the VISTAR amplifier chain, the input klystron, V-27B is a two-cavity klystron amplifier with the input cavity tuned to 9315 Mc and the output cavity tuned to 9375 Mc. To prevent the high level 60-Mc signal from interfering with the i-f amplifiers the 60-Mc oscillator signal is amplified in a burst generator during the synchrodyne pulse. Therefore, the high level 60-Mc signal is present during transmission only and the possibility of 60-Mc interference with the receiver from this source is eliminated.

The pulse of 9375 Mc is amplified in another V-27B which also shapes the r-f pulse. Rather than rely upon the modulating pulse in the final amplifier stage, the transmitted pulse shape is determined in this second V-27B. Finally, the r-f pulse is power amplified in a Varian V-24B.

The Stalo output power is about 100 mw, which is enough power to adequately bias the first detector. With cable loss and 8 db gain through the two V-27B's, the 125-mw input requirement to the V-24B is met. There is a variable attenuator to permit adjustment of the r-f drive to the output klystron. The V-27B has a gain of 10 as an amplifier but the loss in the synchrodyne stage results in the combined gain of 8 db.

The time modulation in the output is obtained by pulsing the transmitter. The transmitted pulse length is 0.2 microsecond and optimum pulse shape is obtained in the pulse shaper stage.

The triggering of the pulse shaper occurs during the end of the synchrodyne pulse. It has been noted that satisfactory synchrodyne operation does not occur immediately upon triggering but in about 1 microsecond after the beginning of the pulse. The pulse length chosen for the synchrodyne stage is 2.0 microseconds.

## I-F Amplifiers

The i-f amplifiers used in the VISTAR equipment consist of a linear i-f preamplifier which drives a logarithmic amplifier and linear amplifier in parallel. All of the i-f amplifiers operate at a center frequency of 60 Mc.

The i-f preamplifier has a gain of 45 db. The range compensation signal drives the agc input of this preamp, giving 28 db of gain variation. Over this compensation range, the preamplifier has a dynamic range of 60 db that is linear to within 1 db. The i-f preamp drives a dual cathode follower and each cathode follower drives one of the remaining two amplifiers.

The log i-f amplifier has a noncoherent video output. The dynamic range of this amplifier is 60 db and the output video signal is quantized and displayed on the t-r Display. The amplifier has two stages of i-f amplification, one stage of video amplification, and a cathode follower output. There is a requirement for some range integration of this video and this is obtained by smoothing the video in a filter with a variable time constant. The time constants specified are 1.0, 1.5, and 2.0 microseconds and these are obtained by switching several capacitors in the video stage of this amplifier. The capacitors are incorporated into the switch on the control panel.

The remaining linear i-f amplifier drives the synchronous detector and the storage tube in the data processor. The linear dynamic range of the data processor is about 20 db and this linear i-f amplifier is designed to have this capability. The amplifier is made to operate over the 60 db dynamic range of the input preamp through use of a three-step gain program. A control signal determines which of the three 20 db segments is to be used. The 3 db bandwidth of this amplifier is 3 Mc and the maximum output voltage is about 3 volts.

Offset Oscillator

The purpose of the offset oscillator is to translate the Doppler frequency spectrum into the frequency range of the filter band used for Doppler analysis. Reference to the diagram for the offset oscillator will illustrate the arrangement used to accomplish the required d-c control voltage to frequency conversion. Basically, the conversion involves heterodyning a voltage controlled oscillator with a crystal controlled local oscillator. The difference frequency is measured with a frequency meter, producing a voltage proportional to the frequency. This analog equivalent of $f_{os}$ is summed with the controlled oscillator. All units are transistorized.

An astable multivibrator is used as the voltage controlled oscillator. The operating frequency is made variable by controlling the base-return potentials. Output frequency is very nearly

$$f_{vco} = (99 + 2.5\ E)\ \text{kc}$$

where

$E$ = output voltage of operational amplifier.

The effective operating range is 94 to 103 kc. In-phase and quadrature outputs are emitter follower coupled to the in-phase and quadrature mixers.

The local oscillator is a conventional Clapp type crystal controlled oscillator operating at a frequency of 93 kc. The output is also follower coupled to the in-phase and quadrature mixers.

The in-phase mixer consists of a carrier-rejection diode mixer driven by the outputs of the voltage controlled and local oscillators. The difference frequency is selected by a low-pass filter eliminating the individual oscillator frequencies and also the sum frequency. The difference frequency component is amplified by a feedback-pair amplifier followed by an emitter follower. This output is termed in-phase $f_{os}$ and is supplied to the single sideband mixer.

The quadrature mixer differs from the in-phase mixer only to the extent that the 93-kc local oscillator frequency is shifted $90^{\circ}$ by a phase-shift network before being used by the diode mixer. The output voltage is a sinusoid having the same amplitude as the in-phase output but differing in phase by $90^{\circ}$. The two outputs are matched in amplitude and phase relation to within ± 1 db and $\pm 1^{\circ}$ over the frequency range of 1 to 10 kc. The quadrature mixer output is also supplied to the single sideband mixer.

The frequency meter is the linearity determining component of the offset oscillator. It consists of an overdriven amplifier, a monostable multivibrator, two stages of pulse shaping, and a discriminator network. The objective is to convert each cycle of the in-phase waveform into a pulse of constant energy. This pulse is applied to the discriminator which is a conventional RC type based on the principle of successive capacitor discharges having equal volt-time areas. These pulses are smoothed by a low-pass filter; the output voltage should be exactly proportional to the input frequency. By this method, a linearity of better than 0.1 per cent is obtained reconverting frequency to voltage. This analog voltage proportional to $f_{os}$ is summed with the offset oscillator command voltage at the operational amplifier network.

Sufficient loop gain is provided by an operational amplifier to make the offset oscillator characteristics essentially those of the frequency meter. The positive control voltage is summed with the negative frequency meter output voltage and the small difference voltage is amplified and used by the voltage controlled oscillator causing its frequency to change in such a manner as to null the voltage at the summing point.

Additional circuitry in the operational amplifier network provides a means for adjusting the offset oscillator sensitivity and offset. A relay is also included to reduce the amplifier gain momentarily at the time the unit is first energized to allow the system to "lock on".

## Single Sideband Mixer

The function of the single sideband (SSB) mixer is to add the offset oscillator compensation frequency to the 60-Mc reference oscillator frequency. A phase-balanced modulator configuration is utilized in this application because of the excellent lower sideband rejection capabilities.

Two balanced modulators, each using two 5725/6AS6 suppressor control tubes, are connected to a common load or, in this case, a combining circuit. Refer to the block diagram of the SSB mixer using two balanced modulators. The designated input signals are those supplied to the tubes after being shifted in phase.

The carrier frequency is $\alpha$ and the modulation frequency is $\beta$. A length of cable cut to $\frac{1}{4}$ wavelength at 60 Mc is connected between modulator No. 1 and modulator No. 2. Modulation frequency is applied to each suppressor grid out of phase with the other tube in each modulator. This is accomplished by an audio transformer with a balanced secondary; the two quadrature audio sources are the offset oscillator outputs. The output of each balanced modulator is applied to the output circuit or combining network.

The output current of each half of a balanced modulator can be expressed by a power series of the form

$$i_0 = C_0 + C_1 e_1 + C_2 e_1{}^2 + \text{----} \tag{1}$$

where

$$e_1 = A_1 \cos \alpha t + B_1 \sin \beta t, \quad \text{for one case.}$$

Expansion of equation (1) yields

$$i_{01} = C_1 A_1 \cos \alpha t + 2 C_2 A_1 B_1 [\sin (\alpha + \beta) t - \sin (\alpha - \beta) t] \tag{2}$$

Similarly, the other half of the balanced modulator will produce a current

$$i_{02} = C_1 A_1 \cos \alpha t - 2 C_2 A_1 B_1 [\sin (\alpha + \beta) t - \sin (\alpha - \beta) t] \quad (3)$$

The resultant current will represent the sum of the sideband currents without carrier

$$i_{02} - i_{o1} = 4 C_2 A_1 B_1 [\sin (\alpha + \beta) t - \sin (\alpha - \beta) t] \quad (4)$$

The second balanced modulator will produce an output current of the form

$$i_{02} - i_{03} = 4 C_2 A_1 B_1 [\sin (\alpha + \beta) t + \sin (\alpha - \beta) t] \quad (5)$$

These currents are combined in the common output or combining circuit to give

$$E_0 = K \sin (\alpha + \beta) t \quad (6)$$

Equation (6) represents the desired frequency derived from the two balanced modulators.

The above derivation is complicated by the fact that phase and/or amplitude unbalances may be present. It is sufficient to state here the restrictions imposed on these quantities to obtain a 30-db rejection figure.

For 30-db rejection of the unwanted sideband, the components $A_1$ and products $B_1$ must be within 3 per cent of the same magnitude, and the deviation from $90^\circ$ in the modulating signal must be less than $3.6^\circ$.

In the single sideband generator, the operation of each tube is controlled by two potentiometer adjustments. One adjustment controls the transconductance of the tube by varying the screen voltage, and the other controls the modulation by varying the audio drive.

Without audio applied to the suppressor grids but with 60 Mc applied to the control grids, the carrier can be rejected by adjustment of the screen potentiometers. The extent of this rejection is on the order of 50 db. The modulation controls are used to achieve a smooth output signal, i.e., no large peaks or dips.

A two-stage tuned amplifier couples the output signal of the single sideband mixer to the synchronous demodulator.

Synchronous Demodulator

The synchronous demodulator performs the function of detecting the radar return signal using the compensated reference signal from the single sideband mixer. The amplified output is provided to the data processing section of the radar.

Two inputs are applied to the demodulator. The reference input is the output of the single sideband mixer at a frequency of 60 Mc + $f_{os}$. The signal input is the output of the radar receiver and is the actual radar return consisting of samples of frequencies 60 Mc + $f_d$.

A synchronous demodulator is a linear detector, provided sufficient drive is present for the signal to be demodulated. This means that the modulating diodes must be driven from full cutoff to full conduction. Under this condition, the diodes are switches operating at the rate of the reference frequency, and hence the output demodulated signal is proportional to the input signal. As used in the coherent radar, this characteristic is fortunate in that the reference voltage may vary as long as its amplitude is above a value determined by the demodulator requirements.

A pentode line driver couples the detected output from the synchronous demodulator to the data processing equipment.

Modulators and Timing

The low power modulator, residual ASTAR contract, was modified for this application. The major modification involved changes to allow the 7000-pps pulse rate necessary for this application.

The high power modulator for the V-24B required a new design and a magnetic modulator was selected. The operation of this magnetic modulator is such that the timing of the other klystrons must be controlled by the output klystron modulator and this resulted in minor changes to the timing sequence. Because of the relationship between the magnetic modulator and the timing for the transmitter, timing is discussed along with the operation of the modulators.

A detailed description of the magnetic modulator is contained in a technical manual supplied by the Magnetics Research Corp. for the modulator used.

All of the timing signals used in the low power modulator are differentiated and used to trigger vaccuum tube circuits. All of the signals are positive going with amplitude of 40 volts. Experience has shown that 40-volt positive pulses from a 100-ohm source, with rise time of less than 1 microsecond, are sufficient to trigger the circuits in the low power modulator. This is the characteristic of the prf trigger pulse from the programmer.

The prf pulse from the programmer to the low power modulator chassis triggers a low impedance pulse generator whose output to the high power modulator is a 20-volt pulse, 2.0 microseconds long, at 10 ohms. This appears at the modulator trigger BNC. After a delay of approximately 7.5 microseconds the high power modulator produces an output trigger of the required 40-volt characteristic. This trigger is used to enable a 2.0-microsecond synchrodyne blocking oscillator. The blocking oscillator drives a dual cathode follower with two outputs, one for the 60-Mc burst generator and the other for the modulator tube. The 2.0-microsecond burst of low level 60-Mc energy forms another input to the low power modulator chassis and drives a high power 60-Mc amplifier whose output is mixed into the cathode of the V-27B synchrodyne. Therefore, at this stage of the transmitter we have the high power modulator in its cycle and the synchrodyne stage is triggered ON and will remain so for 2.0 microseconds.

The high power modulator modulates the V-24B with a 36-kv, 6-ampere pulse, 0.3 microsecond long, at a time 1.5 microseconds after the synchrodyne trigger. A trigger pulse of the proper characteristic is generated by the high power modulator in synchronism with the current pulse through the V-24B klystron. This coincident trigger pulse is the trigger input to the pulse shaper klystron modulator. The 0.2 microsecond pulse of r-f energy out of the pulse shaper klystron occurs a few nanoseconds after the trigger during the high voltage pulse on the output klystron. The shape of the output rf is thus determined by the fidelity of the r-f drive pulse of the output stage.

### 3.2 PROGRAMMER

The primary function of the programmer is to enable the functioning of storage tube through the operations of WRITE, READ, ERASE, and PRIME. Auxiliary subroutines controlling the VISTAR equipment which are associated with these storage tube operations are:

WRITE: PRF Triggers to the Transmitter
Range Normalization (STC) of the 60 db I-F Preamplifier
T-R Display Sweep
Gate Reference Marks to T-R Display

READ: Commutator Sync
RIV Display Sweep
Flash Instrument and Indicator Lamps
Gate Reference Marks to the RIV Display

ERASE: None

PRIME: Advance Cameras

Commercial logic modules are used in the programmer and special circuits have been fabricated on breadboard cards which are mounted in the card holders with the other logic cards. The intercard wiring is made with taper-pins allowing a semipermanent installation with the option to rewire with a minimum of effort.

The purpose of this section is to describe the VISTAR programmer by the various subroutines. When necessary for clarity and to aid in any troubleshooting, specific cards are indicated by position in the card holders. The numbering convention is that marked on the card holders and indicated in the logic schematics furnished with the VISTAR equipment. These schematics are referenced here, but not included as part of the report.

The VISTAR programmer uses logic referenced to two clocks. The programmer cycle and some of the subroutines are synchronized to the 1-second clock. Intra 1-second clock functions are synchronized to the 880 kc clock.

1 Second Clock

The 1-second clock pulse, $t_o$, is derived by counting the 115-v, 400-cycle, 1-phase primary power. The primary voltage is attenuated in a resistor divider and decoupled from the logic by a transformer. This circuit is mounted on Card 16. The waveform is squared, using two stages of inverting amplifiers, and used to trigger a one-shot to derive pulses compatible with the logic. The one-shot output is counted in a 9-stage binary counter, Cards 23 and 24, with the appropriate stages of the binary counter combined to decode the count of 400. The 400 count are the outputs of the 265, 128, and 16 count stages and the combination occurs at the counter input rate divided by 400 or once every 1 second. The stability and accuracy is dependent upon the 400 cycle primary power.

The count of 400 is shaped in a one-shot and amplified to achieve a "fan out" capability for loading. This amplified output is the $t_o$ pulse used to trigger certain subroutines. In addition, it is used to reset the counter in the 1-second clock in preparation for the decoding of the next 400 counter inputs. The 1-second clock pulse is available on Card 10, pin S.

AND Gate 2, Card 21 is wired to permit the decoding of some other count. The necessity for this feature has been eliminated with the magnetic high power modulator but the wiring remains. This logic is available for other applications if required.

## 880-kc Clock and Counter

The 880-kc Clock is a standard Navcor 1-Mc Clock with the crystal reference changed to 880 kc. The clock has a start input that is driven from $t_o$ and a auxiliary microswitch to turn it on and off. The experience to date with the VISTAR programmer has indicated that the clock does not always start on $t_o$ and it may be necessary to start the clock manually using the microswitch when power is initially applied. There is an indicator lamp on the clock card, card 1, which can be used as a visual indicator.

The 880-kc clock drives the count input of an 8-stage counter on Cards 25 and 26. The first five stages are 1-Mc counters, required because of the input frequency. The outputs of the seventh and eighth stages are inputs to the prf selector switch on the control panel. Therefore, the prf control consists of selecting either of two signals which differ in frequency by a factor of 2, both referenced to the 880-kc Clock. One frequency is the clock frequency divided by 128, the other by 256. The crystal frequency 880.675 kc and the prf frequencies are, respectively, 6880 cps and 3440 cps.

The output of the prf selector switch is shaped in a Navcor 338 Pulse Standardizer and is the synchronizing clock for all of the high speed subroutines in the programmer. The output of the first stage of the counter is at a 440-kc rate and this output is used for the t-r display hatching signal applied to the t-r threshold detectors. The output of the third counter stage is amplified and shaped in a 338 Pulse Standardizer. The output pulses occur at a 110-kc rate and are used for the t-r display range markers. The radar range represented by these markers is 4,470 feet. Since the prf, t-r display hatching and t-r display range marks are all synchronized to the 880-kc clock, they have the same registration on successive sweeps and/or frames of the t-r display. The hatching appears as horizontal stripes and the range marks appear at one side of the display, at zero range and repeat at 4,470-foot intervals.

The prf is used to trigger the linear sweeps in the WRITE, ERASE, and PRIME modes and also triggers the transmitter in the WRITE mode. The prf output is obtained from terminals S and 15 of a 338 Pulse Standardizer, Card 27. The hatching signal is obtained from terminal U, Card 25 and the range marks are obtained from terminal 6 of a 338 Pulse Standardizer, Card 7.

Operations Counter

The 1-second clock pulse, $t_o$, sets flip-flop 307, Card 9 and the false output is inverted producing a true output at terminal T of Card 10. This signal is designed $G_o$, a control signal. The $G_o$ signal remains true until the programmer has completed the four storage tube cycles and, at the end of the PRIME period, goes false for the remainder of the 1-second interval.

Other control signals required for the operation of the storage tube circuits are derived from the operations counter, Card 14. This counter consists of two stages of a 310 advanced one count after each storage tube operation. The outputs of the operations counter are designated $OP_1$, $\overline{OP}_1$, $OP_2$, and $\overline{OP}_2$ for the 1 and 0 output of stages 1 and 2, respectively. The programmer is in the WRITE, READ, ERASE, and PRIME mode when the outputs of the counter are true and decoded as follows:

| Code | Storage Tube Operation |
| --- | --- |
| $\overline{OP}_1$ $\overline{OP}_2$ | WRITE |
| $OP_1$ $\overline{OP}_2$ | READ |
| $\overline{OP}_1$ $OP_2$ | ERASE |
| $OP_1$ $OP_2$ | PRIME |

The detailed logic for the storage tube cycle is explained below. The cycle incorporates a general reset bus to assure starting no matter where the logic is when the programmer has power applied to it. Once the operation has started the cycle repeats at the 1-second clock rate going through, in order, WRITE, READ, ERASE, and PRIME and then returning to WRITE condition until the next 1-second clock trigger. Waiting in WRITE assures that all of the storage tube voltages have stabilized and the storage tube is ready for the WRITE sequence.

At $t_o$, after the programmer is turned on, the operations counter is reset through the common reset bus as are all of the counters in the programmer. The operations counter is in the $\overline{OP}_1$, $\overline{OP}_2$ state, which defines the WRITE mode. Before the cycling of the operation counters can be discussed further, some of the subroutines should be considered.

<u>WRITE Routine</u>

At $t_o$, three inputs to an AND gate become true: $G_o$, prf, and $\overline{OP}_1$ or $OP_2$ (denoted by $\overline{OP}_1 + OP_2$). This AND gate is on Card 13 with output at terminal C. It is seen that $\overline{OP}_1 + OP_2$ is true for every storage tube operation except READ. The output at terminal C, Card 13 is inverted and used to trigger a 338 Pulse Standardizer whose output at terminal 9, Card 27 triggers the 100-microsecond sweep for WRITE/ERASE/PRIME, located on Card 2. The 100-microsecond sweep circuit schematic is that of the sweep generator in the set of schematics and consists of a monostable multivibrator whose ON time is nominally 100 microseconds and adjustable about this value. The output of the multivibrator is available at terminals 13 and P. The 100-microsecond gate is level shifted and impedance varied with polarity through a feedback circuit to improve the reset time of the sweep. The resulting gate is used to enable a modified bootstrap circuit used for sweep generation. The resulting sawtooth output is at terminal 5.

An additional output from the multivibrator is a trigger produced at the end of the 100-microsecond gate and designated $SE_1$. The trigger results in a standardized pulse at terminal M, Card 7. $SE_1$ is combined in AND gates with the outputs of the operations counter such that pulses synchronized to $SE_1$ and occurring only during the WRITE/ERASE/PRIME operations are obtained. These pulses are then counted in a 9-stage binary counter contained on Cards 6 and 5. Each $SE_1$ advances the counter one count. The outputs of counter stages 4, 5, 6, 8, and 9 are combined in an AND gate with output at terminal C, Card 8. The true condition triggers a 304B One-Shot on Card 12, which occurs at the count 440, the required number of sweeps for the storage tube. The one-shot output, terminal D, Card 8 is designated $CT_1$ for count completed. At the trailing edge of the one-shot pulse, the 9-stage counter is reset to zero. This resetting pulse is designated $CT_1$ (delayed).

The 9-stage counter serves the dual purpose of providing an output to be decoded for the count of 440 and is also used to drive a digital to analog converter. The output of the digital-to-analog converter during the accumulation of the 440 count is a voltage stairstep with 440 steps of uniform voltage difference. This stairstep is generated using a 341C Ladder Network (digital-to-analog converter (dac)) located at Card 4, the output of which is summed in a 3PA6A1 Operational Amplifier, Card 3. To assure the repeatability of the staircase, the ladder network power supply is an SP 303 module power supply at Card 39, which is a stabilized reference supply designed for operation with this ladder network. The stairstep is the scan waveform to the storage tube during the WRITE, ERASE, and PRIME operations. The sweep for these operations is the 100 microsecond sweep at terminal 5, Card 2.

The $CT_1$ (delayed) is combined with $t_o$ in an OR gate whose output, with proper shape and drive capability, is used to reset the 440 Counter. The $CT_1$ (delayed) + $t_o$ signal is combined with $\overline{OP}_1$ + $OP_2$ in an AND gate with an output at terminal 17, Card 8. The decoded operations counter signal is true, since the cycle is in the WRITE mode and there is an output from the AND gate. This is inverted and used to trigger a 338 pulse standardizer through an AND gate with output at terminal D, Card 8. The inversion of $(CT_1 \text{ (delayed)} + t_o) \cdot (\overline{OP}_1 + OP_2)$ is required, since the 338 trigger must be positive. The AND gate on Card 8 followed by the 338 is equivalent to an OR gate. The output of the 338 is the count input to the operations counter; therefore, at the 440 count this counter is advanced to the READ state.

READ Routine

At the change in state of the operation counter $OP_1$ $\overline{OP_2}$ is true and $G_o$ is still true. These three signals are combined in an AND gate with the trigger from the filter band circuit denoting the end of the commutation cycle. The output of this AND gate is terminal 16, Card 21, which triggers a 338 pulse standardizer which, in turn, is used to trigger the 400 micro-second sweep, Card 2. To obtain the first trigger from the commutator, a special subroutine is incorporated which causes the RIV Display to cycle one sweep. Velocity markers are also recorded during this time without filter-bank video.

The operation counters change of state to READ is the true condition for an AND gate with output on terminal 17, Card 13. This is inverted with output on terminal C and 3, Card 22. The positive going waveform at the beginning of READ triggers a 338 pulse standardizer, Card 27. The output of the 338 on terminal 12 sets a 307 Flip-Flop with ZERO output on terminal 12, Card 9. This ZERO output is inverted and the logic ONE is an input to two AND gates. In one case, the ONE is combined with the velocity marks and the AND gate output terminal E, Card 21 goes to the video combining circuit of the RIV Display. In the other case the ONE and prf are combined and used to trigger the filter-bank circuits. The 307 flip-flop is reset by the first trigger to the 400-microsecond sweep. Therefore, the velocity marks produced by the filter trigger are gated to the riv encoded video only during this pre-READ routine and the operation producing the display of the velocity marks generates the first commutator output trigger.

The commutator trigger, $G_o$ $OP_1$ and $\overline{OP_2}$ are the input to an AND gate with output at terminal 16, Card 21. The commutator trigger is the negative going 30-microsecond commutation gate. When this gate returns positive, the 338 pulse standardizer is triggered and, in turn, there is an output pulse at terminal J, Card 27 which triggers the 400-microsecond sweep. The 338 pulse, as described above, resets the flip-flop controlling the velocity markers.

Associated with the READ operation is the 400-microsecond sweep. The multivibrator in the gate circuit of the 400-microsecond sweep triggers a 338 pulse standardizer whose output pulse on terminal 12, Card 7 is designated $SE_2$. The $SE_2$ signals are the input to a 7-stage counter on Cards 15 and 14. The counter output is also converted to a stairstep scan in a dac with ladder network on Card 17, operational amplifier on Card 18, and power supply on Card 19. The scan output is on terminal L, Card 18. This 7-stage counter counts to 128 and the output of the seventh stage is used to trigger a 304B One-Shot whose output is designated $CT_2$. The $\overline{CT}_2$ signal advances the operations counter one count. $CT_2$ is the READ end signal. The first stage of the 7-stage counter provides the hatching signal to the riv video threshold detectors.

## ERASE and PRIME Routines

With the advance of the operations counter by $\overline{CT}_2$, the three inputs to the AND gate on Card 13, output at terminal C, are again true as they were during WRITE. In PRIME, the operations counter has outputs $OP_1$ and $OP_2$. These are combined with $CT_1$ at the end of the PRIME operation in an AND gate and the true output at terminal 15, Card 8 resets the $G_o$ flip-flop on Card 9 stopping the $G_o$ until the flip-flop is again set at time $t_o$. The $CT_1$ signal at the end of the PRIME routine advances the operations counter to the WRITE condition thereby allowing the voltages on the storage tube to stabilize prior to $t_o$.

## Storage Tube Control

The buffered output of the operations counter, Card 10, pins K, L, F, and H, are decoded by AND gates on Cards 13 and 14 to serve as storage tube mode control signals. Since the logic in the storage tube unit is positive, as opposed to the negative logic of the programmer, a sign inversion and level change is required for these control signals. This is accomplished on Card 16 with inverting amplifiers and level changing circuitry, which produces the ERASE, PRIME, WRITE, and READ command signals in a form compatible with the storage tube logic.

Deflection signals for the storage tube are generated by the 100- and 400-microsecond sweep generators and by the two digital-to-analog converters. The polarity and amplitude are suitable for direct application to the deflection circuits.

## PRF Subroutine

The purpose of the prf subroutine is to trigger the transmitter during the WRITE operation only in order to keep the receiver terminals quiet during the other storage tube operations. The prf pulses from the 880-kc clock are gated to the transmitter through an AND gate with output at terminal C, Card 21 by way of a pulse amplifier which inverts the polarity and provides a positive going, 40-volt, low-impedance pulse. Logic is provided to obtain the several pre-triggers required of a high power modulator used previously, but the present configuration produces only 440 prf pulses to the transmitter. The other leg to the AND gate driving the pulse amplifier is the output of a 307 flip-flop at terminal 13, Card 9. This flip-flop is set by $G_o$ and reset by $CT_1$, therefore, it is ON during the first 440 pulses after $G_o$, or during the WRITE operation.

## Display Subroutines

The t-r display operates during the WRITE operation of the storage tube in association with a subroutine that generates a 64 second scan. The line sweep used for the t-r display is the 100-microsecond sweep. The 100-microsecond sweep is triggered only during the WRITE operation, the operation of interest in the t-r display. The scan for this display is a staircase waveform obtained by converting the output of a 6-stage counter to an analog voltage. This counter is on Cards 31 and 32 and counts pulses synchronized with $t_o$. Again, a 341C ladder network is used and is located on Card 33. The output staircase voltage is decoupled from the ladder network with a 3PA6A1 operational amplifier on Card 34. This waveform is the 64-second scan for the t-r display. The full count capability of the counter is used and, therefore, the normal reset occurs after the count of 64. Because the stability requirements of this display are not as stringent as those for the storage tube, the reference power supply, SP 303 is replaced by the -15 vdc.

The hatching signal for the t-r video encoder is obtained from the first stage of the counter associated with the 880 kc clock. The range marks for the t-r display are derived from the same counter. The output of the third stage occurs at a 110-kc rate and this output is shaped in a 338 Pulse Standardizer with output at terminal 6, Card 7. The requirement on the inputs to the display video mixer is that the reference markers and the encoded video be exclusive events, i.e., the encoded video must occur during the WRITE operation and therefore the reference marks must occur at some other time. The t-r display cycles at the 64 second interval and the range marks are required once during each cycle. To prevent the range marks from interfering with the video during the 64-second period and to cause their display once each t-r scan, the ZERO output of the last stage of the 6-stage 64-second counter is used to enable the range mark display during the following one-second cycle period. This output is differentiated and used to set a 307 Flip-Flop with a ONE output on terminal 9, Card 49. The count of 64 is coincident with $t_o$ and therefore the flip-flop output is true during the following one-second cycle until it is reset at the end of the PRIME period. To prevent the range marks from occuring at the time the camera is advanced during PRIME, the true combination of the flip-flop ONE output on terminal 9, Card 49, the range marks at terminal 6, Card 7, and the ERASE gate at terminal 16 Card 13 are used as inputs to an AND gate at terminals K, L, and M, Card 45 to form a gated range mark output at terminal D which occurs during the ERASE period only. These gated range markers are the appropriate input to the t-r display mixer.

The other reference marks for the t-r display are time marks which occur every 10 seconds, (10 second markers). These are obtained by counting one second pulses in a 4-stage counter located on Card 31. The second and fourth stage outputs are decoded at the count of 10 by combining them in an AND gate with output on terminal E, Card 29. The AND gate output triggers a 304B One-Shot whose output at terminal H, Card 28 resets the 4-stage counter. The ZERO outputs of the count-of-10 counter are combined in an AND gate with the output of the first stage of the 128-count counter in the READ scan. The true output is at terminal 17, Card 45 and is present every 10 seconds for 64 pulses. Since the 100-microsecond sweep is disabled during the RFAD routine, there is no range deflection of the crt and hence no interference with the t-r video or range marks.

The signal to flash the instrument lights of the t-r display is in the true state between the count of 64 seconds and the end of the next READ operation. This is obtained by setting a 307 Flip-Flop with the differentiated ZERO output of the sixth stage in the count-of-64 counter and resetting the flip-flop with $CT_2$, the READ End Pulse. The light logic is such as to require a logic ONE output to keep the lights OFF and this is obtained at terminal 8, Card 49, the flip-flop ZERO output.

The t-r camera advance signal is generated once each 64 second scan by combing the range mark control flip-flop output from Card 48, pin C with the PRIME gate output from Card 48, pin D in an AND gate, Card 47, pins 8 and 9. The gate output, pin 6 is amplified and inverted in the 3PA2A on Card 46, pin J to H. This advance signal is present only during the first PRIME interval following the completion of a 64 second t-r scan interval.

The neon lights indicating the setting of the low-pass filter are also flashed during the first one-second cycle period following the 64 second scan. In this case the enabling period is during READ as provided by flip-flops on Card 49, pin 13 and Card 49, pin 10 combined pin an AND gate on Card 47, pins 12, 13 and 15.

The signal to advance the riv camera is the PRIME gate complement obtained at terminal 6, Card 48. The camera advance circuit is such that the logic state nearest zero advances the camera. The flash indicator signal occuring during the READ operation, is obtained by setting a 307 Flip-Flop with the WRITE end signal, at terminal R, Card 27 and resetting the flip-flop with the READ end signal, $CT_2$. The flip-flop ONE output is obtained at terminal 13, Card 49. The amplified ONE is obtained at terminal 15, Card 46. The same signal is obtained at terminal T, Card 46 and this is used to flash a portion of the indicator lights. The output at terminal 15, Card 46 is also the flash instrument lamp signal.

The reference marks for the riv display must be exclusive events from each other and from the encoded video, as for the t-r display. The operation of gating the velocity marks in to the video mixer is required during the READ routine and discussed in that section. The riv range markers are the decoded output of a 4-stage counter on Card 30 that counts $SE_2$ pulses. The counter output is decoded for a count of 10 in an AND gate with output at terminal 16, Card 16. The count of 10 and the output of a 304B One-Shot synchronized with the start of the filter trigger are combined in an AND gate with output at terminal 15, Card 29. The true condition is the RIV Range Mark or the WRITE end signal at terminal N, Card 22 are used to reset the 4-stage counter.

<u>Receiver Range Compensation</u>

The requirement for range compensation in the i-f preamplifier was discussed in the functional description of the receiver signal processing. This compensation, sometime called sensitivity, time control (stc) is caused by modulating the preamp AGC input with a control signal generated in the programmer. The gain characteristic for the receiver preamplifier as a function of the agc voltage is plotted for the desired gain versus radar range or time compensation. The result is a voltage-time function which must be generated.

The time of interest to VISTAR is 100 microseconds with negative agc voltage acting to reduce preamplifier gain. The required compensation curve is one that will give the maximum attenuation at the near range with increasing gain proportional to time. The voltage curve of interest and other applicable calibration information were delivered with the VISTAR equipment. This discussion is concerned with the manner in which the stc function is generated and adjusted.

The stc function is synchronized with the 100-microsecond sweep gate which occurs during the WRITE operation. This gate, at terminal P, Card 2, is the input gate to the stc circuit on Card 43. The 100-microsecond gate causes the generation of a negative going linear sweep in a circuit similar to the storage tube sweep generators. As with the others, this sweep is adjustable in amplitude and linearity by the length is a function of the input gate. The linear sweep is amplified by a modified 3PA6A1 Operational Amplifier on Card 44. The modification consists of adding a variable resistor in the feedback loop of the amplifier to adjust the gain. The negative going linear sweep drives a circuit that varies the gain of the operational amplifier according to the preamplifier output voltage. The gain is varied in three steps and the compensation curve is generated by the three resulting straight line approximations. The gain change is obtained in the following manner. The gain of the amplifier is determined by the input impedance and by the amplifier feedback resistance. For the initial portion of the linear sweep the input resistance is adjustable using series shunt variable resistances. In combination with the adjustable feedback resistance, the output voltage can be adjusted. As the output voltage increases a diode in the feedback circuit is turned on and the resistance in series with the diode adds in parallel to the feedback resistance decreasing the gain of the amplifier. As the output voltage increases further another diode turns on to further decrease the amplifier gain. The non-linear sweep out of the amplifier is the stc function. During the absence of the 100 microseconds gate this output is at its most negative value, reducing the preamplifier sensitivity to the lowest level for the READ routine. During the 100 microsecond gate the output sweep increases to zero vdc where maximum gain is obtained at the far range.

The amplitude of the linear sweep is 10 vdc corresponding to the stc function range of -2.7 to 0 vdc. The excursion of the stc signal causes the preamplifier gain to cover a 28 db range. The smoothness of the stc control characteristic is such that the range compensation is accurate to within ± 1 db.

Gain Programmer

The gain programmer generates the control signals and the appropriate indicator lamp signals for the 20 db linear i-f amplifier in the Doppler processor. The gain programmer is controlled by a selector switch on the control panel and decodes the switch setting to produce the amplifier control voltages. The gain control causes 0 db, 20 db, or 40 db attenuation in the gain of this amplifier and the control input selects one of the above attenuations or causes the cycling of the gain programmer through all three or only the first two at the usual one cycle per second rate.

The selection of either of the automatic cycles causes a true output from an OR gate at terminal D, Card 11. The true output is combined in an OR gate with the $\overline{CT}_2$ pulse and the true condition is counted in a 2-stage counter on Card 26. The outputs of the counter are decoded in AND gates according to the following combinations where the convention $1^1$ refers to the ONE output of the first stage, $0^1$ the ZERO output and so forth.

| Code | Attenuation |
|---|---|
| $0^1 \cdot 0^2 \cdot$ Automatic Cycle | 0 db |
| $1^1 \cdot 0^2$ | 20 db |
| $1^1 \cdot 1^2$ | 40 db |

It should be noted that the 0 db signal will be true whenever the counter is reset, therefore the output from terminal D, Card 11 is a required input for automatic operation. The -20 db signal is combined with the 0 db, -20 db automatic cycle signal in an AND gate whose true output at terminal D, Card 35 causes the rest of the counter at the next $\overline{CT}_2$. Otherwise the counter is reset after the -40 db signal. Selection of any of the non-cycle inputs resets the counter through the t-r bus. The automatic cycle outputs and the non-cycle outputs are combined in OR gates and the outputs amplified to provide a drive capability. The control voltages are logic ONE's at the output of these amplifiers. The amplifier outputs are also decoded in AND gates to obtain indicator lamp signals.

The control signals are exclusive events and are combined in an OR gate on Card 41 after the -20 db and -40 db signals are provided with level setting potentiometers. The OR output is decoupled and provided with a drive capability in an emitter follower stage whose output goes to the linear i-f amplifier.

3.3 STORAGE TUBE UNIT

As described in the functional description of the VISTAR System, a cathode ray recording storage tube is employed to provide: (1) the necessary storage media required to accumulate the samples of the Doppler history of the backscattered microwave energy for the ranges of interest, and (2) the necessary integration of signal samples required to yield an estimate of signal amplitude. The first requirement above stems from the necessity to accumulate Doppler samples at any given range increment over some time period in order to analyze the composite ensemble for the spectral components. The second requirement above, integration, is basic to the detection of weather signals.

Prior to giving detailed descriptions of the circuitry, a brief description of the theory of storage tube action will be presented.

The storage tube employed in VISTAR is know as a "recording storage tube", made by the Raytheon Corporation. Though not utilized in VISTAR, the unique feature of the tube is a <u>nondestructive</u> readout of the stored data. A model number QK685 tube is used, which is a single-gun device, (i.e., data must be sequentially stored and retrieved as contrasted with a dual-gun tube in which data may be simultaneously entered and retrieved). The storage assembly in the QK685 consists of an accelerator screen, storage screen, and collector electrode composed in three-layer "sandwich" with about 50 mills spacing between each element. The assembly is mounted in a plane orthogonal to the tube axis, and at the opposite end of the tube from the gun. Electron beam focus is at the storage surface of the storage screen. This surface is formed by a dielectric material deposited on the gun side of the storage screen. Data is stored on the storage surface in the form of an electric charge pattern laid down by the action of a modulated electron beam and beam deflection over the storage surface.

The storage screen is placed between the collector and the decelerator screen, which is placed toward the electron gun end of the tube. The decelerator screen has two functions, one to act as the final element of a three-element electron lens system in the tube (the other two elements are formed by the anode coating and two cylindrical lens coatings), and the other to maintain a uniform and constant field in front of the storage screen. This latter purpose is important because the potential on the storage surface varies during the various operating modes of the tube, which would result in severe beam defocusing and change of beam energy without the decelerator screen. The collector is the final element of the three-larger sandwich that collects that fraction of beam current which is allowed to pass through the storage screen. During the READ mode of operation, this current is modulated by the charge pattern on the storage surface and thus constitutes the data "signal".

The fundamental phenomenon which allows a charge pattern to be stored, read, and erased on an insulating dielectric is that of "secondary emission". In particular, it is required that the secondary emission ratio, (i.e., the ratio of average emitted secondary electrons to average incident primary electrons) for the particular dielectric employed for the storage screen coating be a function of beam energy in a practical range, and at the same time have a low leakage characteristic. In the recording type storage tube, operation both above and below the unity secondary emission ratio point (critical point) is utilized to selectively remove or add electronic charge.

The operation of the storage tube in a practical system can be broken down into four operating modes: (1) ERASE, (2) PRIME, (3) WRITE and (4) READ. The ERASE mode, as the name implies, clears previously stored data and is accomplished by application of a moderately high voltage (+400 vdc) to the storage screen, such that the beam energy is well above the critical point. Sweeping the beam over the storage surface allows the storage surface potential to approach that of the metallic backing, i.e., very low potential difference between the surface and the metallic mesh. It should be noted that full erasure is an equilibrium condition in that the potential at the storage surface may not be changed by further erasing.

"Priming" the storage tube is the inverse operation of erasure. An unmodulated beam is again caused to scan the storage surface; however, in this mode, the storage screen is adjusted to a low potential (+20 vdc) such that operation is well below the critical point. Under this action, the storage surface will accumulate primary electrons and the surface potential will shift negative until it approaches the cathode potential. The result of this is a uniform negative surface potential relative to the metallic storage screen backing. The tube is now ready for writing.

The WRITE mode is similar to erase in that the storage screen is operated well above the critical point. In this mode, however, the beam is modulated by the electron gun control grid (input signal); therefore, the data storage is accomplished by removal of electronic change according to the modulation of the beam. In opposition to the ERASE mode, writing is not an equilibrium process and, therefore, beam current and modulation amplitude must be carefully adjusted in an operating system. After writing, the storage surface potential relative to the storage screen at any particular point will be, according to the modulation, somewhere between zero (for full writing) and the priming level voltage, e.g., 20 v (for no writing).

A READ mode is utilized to recover the data stored on the storage surface during WRITE. Reading is accomplished by reducing the storage screen electrode potential to a value such that the maximum storage surface potential is less than zero relative to the cathode. This allows the storage surface to act as a negative control element on an unmodulated beam. As the beam is swept over the storage surface, a signal is generated in the collector circuit in conformance to the potential distribution at the storage surface. This constitutes the electrical output signal of the storage system.

In the VISTAR system, the operation of the storage tube is commanded by the programmer, including the sweeps and scans for the raster deflection. The commands for the particular mode of operation, erase, prime, write, or read, is generated in the programmer as a mutually exclusive, positive voltage signal occurring on one of four command lines. These lines are fed to the storage tube unit (refer to storage tube schematic diagram) to activate the various switching functions with the operational modes of the tube.

The switching logic associated with the storage tube unit is a positive voltage logic with the "high" state for "true" and the "low" state for "false". The type of logic is NOR, which is used in the required combinations to drive the three switching circuits, one for adjusting the bias on the storage tube grid and the other two for switching the storage screen potential.

Three NOR gates are used to form the logical states: READ, ERASE + PRIME and WRITE. This is accomplished by combining the command line inputs to the respective NOR gates as the complement of the required output. In the case of WRITE and READ this is redundant, as the command lines themselves contain this information; however, there is power amplification in the NOR circuits which assures a full logic level for the switches.

The grid bias switch (Beacon-2) accepts the three logic states outlined above and, in turn, produces three distinct and adjustable levels for bias. Note that ERASE and PRIME are common levels and, in general, are adjusted for reasonably high beam current levels. Storage screen levels are set by the action of two switches, one a 500-volt switch and the other a 75-volt switch. The results of the switches are mixed with diodes on a common line to the storage screen. The 500-volt switch is a shunt-type switch which disables the 500 volts when a true state is applied to either of two inputs. For this reason the complement of the modes requiring 500 volts at the storage screen are applied to the inputs, i.e., READ and PRIME. When 500 volts is applied to the screen, the 75-volt switch output is disabled via a disconnect diode within the 75-volt switch. In the opposite situation, i.e., the 500 volts shunted by switch action, the 75-volt switch output is active and the 500-volt line is disconnected via external disconnect diodes. The output of the 75-volt switch can assume two values, corresponding to PRIME and READ, the latter being adjustable. The PRIME voltage level is fixed at +20 volts.

The tube is held blanked by a cathode blanking circuit which is disabled during any one of the various sweep intervals. Two inputs are applied to the blanking card; a sweep gate from the 100-microsecond ERASE, PRIME and WRITE sweep generator, and a sweep gate from the 400-microsecond READ sweep gate. Either of these inputs will disable the blanking of the storage tube.

Input video to the storage tube is transformer coupled into the grid circuit from the synchronous detector via a two-stage driving amplifier. This amplifier, mounted external to the storage tube assembly, has a gain control to allow video amplitude adjustment and to serve as a termination for the coax line from the synchronous detector.

Since the QK685 is a tetrode gun tube, a screen grid potential is required. This is derived from the +500-volt supply by a fixed resistor voltage divider. Two lens elements in the tube, L and $L_2$, require adjustable voltages which are supplied by potentiometer and padding resistor from the +500 supply. The tube also has an electrostatic focus anode, $A_2$, which is tied to the high voltage anode, $A_1$, when the tube is operated with magnetic focus.

Magnetic focussing is accomplished with a focus coil that has its current supplied through an adjustable resistor plus padding resistor tied to the +500-volt supply. This use of a high voltage source and series resistance approximates a constant current source which, in turn, minimized the dependence of focus current upon coil temperature. The focus coil has a dynamic focus winding that is not used in this application.

Deflection for the storage tube is supplied by two identical deflection amplifiers which drive orthogonal windings of a low-inductance deflection yoke. The yoke has four windings, two on each axis, which allows the beam to be off-centered by fixed currents in one orthogonal set. This allows the deflection amplifiers to be designed for unipolar drive to the alternate set of orthogonal windings. Offset current is supplied by a variable and fixed resistor network tied to the +20-volt bus. The deflection amplifiers are a four-stage transistorized feedback amplifier that has provisions for two inputs. Feedback is supplied from a resistor that samples the deflection coil current and is compared to the sweep or scan voltage waveform supplied as an input. The waveform error is amplified to provide the drive to the deflection coil. In this manner, a deflection current is generated in accordance with a sweep or scan waveform. Recalling that the sweep and scan waveforms supplied to any given axis of the tube are mutually exclusive events, it can be seen that only one of the two inputs to the amplifiers is active at one time. Therefore, each input is scaled to provide full deflection on the output. There are two adjustments on the deflection amplifiers, one for equalizing the input gains between each input, and the other for biasing the input with direct current such that the output stage is just biased on without signal input.

Output from the storage tube is in the form of a current source with amplitude of the order of one microampere or less. The output amplifier must build this up to a usable level (about 0.5 volt) prior to leaving the storage tube unit. An amplifier consisting of two feedback pairs is mounted in close proximity to the target end of the tube. In order to maintain the required bandwidth, the input circuit of the amplifier is swamped with resistance (about 10K ohms). Since the signal level is quite low at the input stage, care must be taken to assure that power sources supplying the amplifier and collector circuit are quiet. There are decoupling filters on each of the power sources to accomplish this. Power for the amplifier is ±12 volts, and +500 volts for the signal collector element of the tube.

The high voltage anode, $A_1$, is normally operated at +3500 volts which is supplied by a regulated power supply. Since beam currents vary widely during tube operation, regulation of this potential is important to maintain correct focussing.

A "BNC" connector matrix is located on the front cover of the plug-in board case on the front of the unit. These terminals provide monitoring points for sweep, scan, and gating waveforms associated with the storage tube operation. Also located on the front panel is a NORMAL-TEST switch plus a bushbutton to facilitate storage tube setup. The switch breaks the automatic continuous cycling mode, (by breaking the line between Card 12 pin 13 and Card 8 pin L), and allows the system programmer to recycle continuously in READ. The pushbutton simply shorts the switch so that the system may be cycled through a single cycle at a time while the storage system is being adjusted.

Alignment of the storage tube unit is a somewhat tedious task in that final adjustments consist of compromise settings of several parameters A rough procedure will be outlined here; however, some practice is necessary to obtain the maximum inherent capability from the unit.

The first task in tube alignment is to verify that all operating potentials, gates, sweeps, and scans are present. Sweep gates should be adjusted for correct time intervals; 100 microseconds for the erase/prime/write sweep gate, and 400 microseconds for the read sweep gates. These adjustments are on Card 2. Next all of the scans and sweeps should be equalized in amplitude such that an approximately square raster will result, and also to assure that the various rasters generated will coincide. Since the scan generators have the least range of adjustment, they should be equalized with each other first. The adjustment for this is a trim pot that has been added to the summing resistor on the output of the digital-to-analog converter ladder network. These potentiometers are on cards 4 and 17 for the 440 and 128 step scans. After this equalization, the sweeps may be matched in amplitude with the appropriate adjustment of controls on Card 2. There is a linearity adjustment on each of the sweep circuits. This degree of refinement is not required for the erase/prime/write sweep; however, the read sweep linearity is extremely important in order to extract the spectral components of Doppler by integration of the phase-coherent samples distributed along a sweep line of the read raster. Evaluation of the linearity adjustment for the read sweep should be made after satisfactory tube alignment by observing the filter system directly, (i.e., a scope waveform of the responding filter), with a single frequency component stored in the tube. The linearity may then be adjusted to maximize the on-frequency filter and simultaneously equalize the periods of the response envelopes of filters adjacent to the on-frequency filter.

Verification of the various levels applied to the control grid, $G_1$, and to the storage screen during the ERASE, PRIME, WRITE and READ modes should be made prior to attemps at focussing the tube. When these fixed and adjustable levels have been verified, the collimation and focussing adjustments may be made. Preliminary adjustments should be made without writing on the storage surface, using a procedure similar to the following: (1) reduce the bias level on $G_1$ for the WRITE and READ modes until the beam current is cut off; (2) adjust the ERASE/PRIME beam current to a nominal value (e.g., bias voltage -5 vdc); (3) reduce storage screen READ voltage to minimum, i.e., below beam current cutoff; (4) with a scope synchronized to the 400-microsecond READ - sweep gate, monitor the output of the storage tube output amplifier, and apply all operating voltages. (5) while varying the READ screen voltage through its range, increase the $G_1$ READ bias level until beam current flows during the READ portion of the cycle, (system cycling can be in the "normal' condition for this), as evidenced by a finite waveform amplitude on the scope during READ. Note that at a particular $G_1$ bias setting, the range of waveform amplitude at the output will be some particular value. As the beam current is increased, this range of output will increase. Set the $G_1$ bias level such that this output range is a reasonable value compared to the noise level, e.g., one volt. The system cycling may be stopped for closer examination of READ waveforms, and cycled manually by pushbutton. Care should be exercised in the interpretation of amplitudes while varying the READ screen voltage, as a potential shift will result on the storage surface, if the surface potential is allowed to exceed zero relative to the cathode. (6) With the READ screen voltage set for a nominal output waveform, the electron lens, $L_1$, $L_2$, and focus current adjustments can be made. These controls, which are interacting, should be adjusted for maximum amplitude and "flatness" of the pedestal waveform. Readjustment of both current and READ screen voltage should be made as the lens and focus adjustments are brought into close range of the correct setting, to maintain the output waveform in the linear range. (7) When the above adjustments have been performed to a reasonable degree of satisfaction, a WRITE-in of data should be attempted. To accomplish this, the READ screen voltage is reduced until the output waveform is reduced to zero. Then the WRITE beam current bias applied to $G_1$ should be increased until the waveform is again visible within the linear output range. To be able to make refinements in the various settings for maximum resolution and uniform response, a modulated WRITE waveform is desired. It is also desirable that the stored modulation appear stationary during readout, to facilitate waveform interpretation. This may be accomplished by writing a raster of lines of uniform amplitude, such that when they are "cross-swept" during READ, a stationary waveform is produced. A raster such as just described can be written into the tube by selectively unblanking the storage tube during WRITE according to a periodic sequence determined by the WRITE scan counter. A set of logic has been built into the programmer to accomplish this and is covered in a technical memorandum. By appropriate choice of bit positions in the scan counter, rasters may be stored to produce output frequencies, related by a factor of 2, up to the maximum possible, i.e., every other raster line written. Using this technique, refinements in collimation and focus adjustments can be made to optimize the unit. (8) WRITE current bias will have to be altered somewhat for actual video input writing; however, other settings should remain substantially constant. Some touch up of the READ screen may be required.

### 3.4 FILTERS AND COMMUTATOR

The filters and commutator in VISTAR form the interface processing between the storage tube and the range-intensity-velocity (riv) display. The output of the storage tube is a composite of all of the processed Doppler returns and, at each successive range interval, these frequency elements are sorted in parallel through a bank of piezoelectric filters, and then commutated to put the parallel data into serial form for eventual display. The filter bank, commutator, and peripheral circuits are the subject of this section.

The filter bank used in the VISTAR consists of 99 piezoelectric filters. Each filter is a wafer of barium titinate hermetically sealed, with a trimming resistor external to the encapsulating module. Each filter has its Q adjusted to be proportional to the filter frequency. Therefore the filter bank is a comb filter in which each filter has the same bandwidth. The final step in the alignment of the filter bank is the adjustment of the Q which requires the trimming resistor in most instances.

Because ceramic filters have several modes of resonance they display peculiar overtone response and it is, therefore, difficult to make a filter bank of this type cover a decade of frequency.

This filter bank of 99 filters consists of two filter banks of 49 filters each, with a filter-frequency spacing of 5 kc, covering frequencies from 350 kc to 590 kc, plus one filter with center frequency 595 kc. Now the offset oscillator has an output frequency such as to shift the Doppler frequencies of interest into the frequency band 0 to 545 kc. As explained in the secion on Basic Requirements in the Functional Description of the VISTAR, the frequencies at the bottom and the top of this range are discarded, since some buffer is required for possible ambiguous frequencies. The result is that the Doppler frequencies of interest, those required in the specifications, are in the frequency band 35 kc to 525 kc. The filter bank is made to cover this band with the existing filters by the use of two transistor oscillators and vacuum tube mixers employed in hetrodyning circuits. The oscillator frequencies are 625 kc and 875 kc. The filter banks are ordered 590 kc to 350 kc and 595 kc to 350 kc, respectively, for these oscillator frequencies. Therefore, the 590-kc filter responds to the mixer product of 35 kc and the 625 kc oscillator and the 350-kc filter in the same group responds to 275 kc. The next input frequency in line is 280 kc, which produces 595 kc when mixed with 875 kc and so forth. The case for the mixers and unwanted mixer products is discussed below.

The carrier frequency lies outside of the filter frequencies and the balanced mixer products are the odd harmonics. The higher order harmonics have low enough amplitude so that they do not cause appreciable filter response. The 35-kc fundamental has its main response at 590 kc and the third, fifth, and seventh harmonics lie within the filter band. This is the worst-case frequency; however, experience has shown that harmonic response in the filters is produced only by overdriving the input.

The mixers are described above as being vacuum tube devices with local oscillator inputs from transistor oscillators. The output from the storage tube is the input to two identical mixers. The input tube operates as a wideband inverter which drives the primary of a transformer. The secondary is center-tapped and the local oscillator input is through this center tap. The mixer itself is a diode bridge whose output contains the carrier and balanced mixer products. The outputs are amplified and form the inputs to the filter bank.

The filters are mounted on nine printed circuit boards with 11 filters on each board. The center board has two separate inputs, one in parallel with the first group of boards and the other with the second group of boards. Each board input from the mixer circuit consists of a transistor emitter follower for power gain and each of these emitter followers drives an emitter follower on the input of each filter.

The output of each filter is detected in a doubler circuit whose time constant is compatible with the filter transient characteristic. The detected filter outputs are then sampled in order by the electronic commutator. The sampled outputs are common to a single video line forming the basic riv display input video.

The filter bank circuits include those circuits and the logic responsible for the timing of the commutator and filter bank. The filters perform the frequency sampling in parallel but the filter bank circuits synchronize the commutator to the programmer and form the video output to the display video encoder.

The programmer provides a trigger output to the filter bank circuit, which occurs at the end of the 400-microsecond sweep of the storage tube. At this time, the filters have been driven and those with signal have responded and the detectors associated with responding filters have a voltage present. The trigger is delayed in a one-shot within the filter bank circuitry after which a flip-flop is set. The delay allows any transients in the storage tube output to decay and at the time the flip-flop is set one of nine identical blocking oscillators is triggered. The output of the first blocking oscillator drives a tapped delay line with 10 taps with a delay of 0.3 microsecond per tap, -- a total delay of 3.0 microseconds. The input to the line is coincident with the sample command on the first filter and the successive filters are sampled by command from the succeeding taps. The output from the end of the line is the trigger input to the next blocking oscillator, which causes the same sequence until all 99 filters have been sampled. The output from the last tap resets the flip-flop and causes an output trigger to the programmer indicating the end of the commutation period. The blocking oscillator outputs are combined in an OR circuit whose output forms the velocity marks for the riv display.

The commutated video is amplified and the output is of the proper amplitude for video encoding. Emitter follower outputs are used to increase the drive capability of the circuits and NOR logic is used in the flip-flop. All of the circuits are on printed circuit plug-in cards and are packaged in a separate unit with the blocking oscillators which are individually shielded from each other and from the sampling circuits to reduce pick-up.

### 3.5 DISPLAYS

Photographic storage using two recording cameras is the method used for recording the large amount of information gathered in the application of the VISTAR equipment to storm analysis. These cameras photograph oscilloscope displays, various instruments, and indicator lights which visually show the VISTAR data, mode of operation, and pertinent aircraft and weather data. The signals used in the displays and camera cycling signals form the display logic.

The two displays used in VISTAR are the range intensity velocity (riv) display and the total-reflectivity t-r display. The riv display consists of:

1. Oscilloscope for RIV Display
2. Indicator Lights

| | |
|---|---|
| PRF Hi | (6880 pps) |
| PRF Low | (3440 pps) |
| Tilt | (Antenna error or excessive vertical acceleration) |
| Max Threshold | (Maximum Threshold has been exceeded) |

OS1
OS2 (Digitally encoded to show velocity
OS3 band displayed)

Gain 1 (Digital encoded to show which 20 db band
Gain 2 of dynamic range is displayed)

3. 24- hr Clock
4. Altitude
5. Frame Counter
6. Air Speed
7. Heading
8. Temperature
9. Data Card for Date, etc.

The T-R Display consists of:

1. Oscilloscope $f_o$ T-R Display
2. Indicator Lights

   Maximum Threshold Exceeded
   Filtering 1 (Digitally encoded to indicate the filter
   Filtering 2 time-constant)
   24-hr Clock
   Data Card for Date, etc.

The riv display oscilloscope is synchronized so that the X-axis sweep occurs during the 30-microsecond commutation period and the Y-axis scan is stepped in range with the storage READ tube scan. The Z-axis is intensity modulated by the encoded filter video and certain timing marks. The oscilloscope data are the signal return strength sorted with respect to Doppler velocity and range. Within the 20 db dynamic range of the data, the filter video is encoded so that the more limited dynamic range of the crt can show all of the signal dynamic range. The coding is as follows:

| Dynamic Range (db) | Displayed |
|---|---|
| Noise level to 5 | Linear with 5 db near phosphor saturation |
| 5 to 10 | White and Black alternating on successive range lines |
| 10 to 15 | Gray |
| 15 to 20 | Black |
| 20 db or greater | Black with the Max Thres Light lit |

The Gain 1 and 2 lights indicate which of the 20 db ranges within the 60 db range is displayed. Gain 1 only indicates the 0 db to 20 db range, Gain 2 indicates the 20 db to 40 db range, and both indicate the 40 db to 60 db range.

The velocity range of the filters and their spacing is a function of the prf and the offset oscillator setting. The theory is discussed under the Functional Description section and the velocity range and spacing are listed in the Table in the calibration section. The digital coding of the O.S. lights for the switch positions is:

| Switch Position | O.S. 1 | O.S. 2 | O.S. 3 |
|---|---|---|---|
| 1 | 1 | 0 | 0 |
| 2 | 0 | 1 | 0 |
| 3 | 1 | 1 | 0 |
| 4 | 0 | 0 | 1 |
| 5 | 1 | 0 | 1 |
| 6 | 0 | 1 | 1 |
| 7 | 1 | 1 | 1 |

where 1 in ON and 0 is OFF

The frame counter changes when the camera is cycled. Other instruments were incorporated for auxiliary instrumentation for the over-all program mission.

The mechanical installation of the riv display is around an Air Force eight-instrument box. The indicator lights and the instruments are mounted in the box and their images (on a mirror) are photographed. This reduces the over-all length of the installation, since the required focal length of the camera is folded. The oscilloscope face is placed at the same focus plane as the instruments by cutting a hold in the mirror in the instrument box. Provision is made for adjusting the oscilloscope through a half-silvered 45-degree mirror placed between the camera and the crt.

The t-r display is swept in range by the 100 μsec linear sweep used for the storage tube during the WRITE cycle. This sweep is along the vertical axis. The horizontal axis of the t-r oscilloscope has d-c response and the deflection is the 64-second scan from the programmer. The vertical sweep is repeated for each transmitted pulse or 440 times at each position of the horizontal deflection and the video intensity modulation of the phosphor is allowed to integrate the data. The horizontal deflection is then stepped for the next transmitter burst and the 440 sweeps repeated. After 64 such cycles, the horizontal deflection returns to the start and the camera is cycled. The frame of the camera film is filled in the longitudinal dimension by the 64-second crt scan deflection and in the next frame, the deflection is such as to continue the scan in the same manner so that the time histories of the radar returns are in order.

The indicator lights, clock, and data card are imaged in a 45-degree mirror at the same focal length as the crt and photographed in the remaining space of the transverse dimension of the frame. The crt deflections are such as to give a square presentation.

The video from the log i-f amplifier is encoded and used to intensity modulate this display. The intensity coding is:

| Dynamic Range (db) | Displayed |
|---|---|
| Noise to 20 | Linear video |
| 20 to 30 | White and Black switched at a 880 kc rate |
| 30 to 40 | Gray |
| 40 to 50 | Black |
| 50 and greater | Black and the Max Thres Light |

The setting of the filtering is indicated by two neon lights. Both OFF indicates 1.0 microsecond, one ON indicates 1.5 μsec. and both ON indicates 2.0 μsec. The clock is used to indicate the frame count, since there is no space available for a frame counter.

The video input for the riv display is the video output of the commutator and, for the t-r display, it is the video output of the log i-f amplifier. These video signals are encoded in transistor threshold circuits, adjusted in amplitude, combined with X and Y reference marks, and used to Z-axis modulate the oscilloscopes. These operations take place in identical circuits which are described below. The encoding and combining is performed in semiconductor circuitry installed in card holders of the programmer. The input and output is made through the BNC connector panel and the signals are taken in and out through coaxial cables.

The encoding cards are at positions 53 and 54 of the card holders. The two cards in these positions contain identical threshold circuits of four threshold circuits to a card. The threshold circuit is a low hysteresis, high speed, Schmitt trigger with a 10-millivolt ambiguous region. The Schmitt trigger output is phase-split and amplified to give logical one and zero outputs.

The outputs of the threshold circuits are decoded to determine the maximum threshold which has been exceeded. If the video signal is of sufficient amplitude to exceed several thresholds, the lower thresholds must be excluded so that the proper intensity encoding may take place. This is done by combining the "one" output for the threshold circuit with the "zero" output of the next higher threshold circuit thus forming the decoded output. The logic outputs have several other functions. The incoming video is applied both to the threshold circuits and a transmission gate on Card 51. Until the first threshold is exceeded, this transmission gate is enabled. When the first threshold is exceeded the associated logic switch the diodes in the transmission gate to block the video signal.

The AND gate decoding the first threshold is a three-leg gate with the third leg input being the hatching signal. The output of this gate is then switched by the hatching signal. The hatching signal for the riv display is derived from the first stage of the 128-count counter which generates the range scan for the READ operation of the storage tube. This causes the hatched output to be on alternate range lines of the riv display. The hatching signal for the t-r display is derived from the 880-kc clock which is used to produce the prf pulses. The hatching is, therefore, synchronized in range with the 100 μsec range sweep of the t-r display.

The output indicating that the second threshold has been exceeded is decoded and forms another output from the threshold logic. The output from the t-r second threshold detector is also used for the control panel logic light indicating there is a signal exceeding 40 db.

Although there are decoded outputs denoting that the third threshold has been exceeded, these outputs are not used for the present decoding scheme. The output from the forth threshold detector is used to set a flip-flop. In this manner the transient exceeding of this threshold is remembered and used to switch the maximum threshold light in the display. The flip-flop is reset before the next frame of the display.

The gated video, hatched synthetic video, and the second threshold output are combined on Card 41. These are called video, hatched, and gray for simplicity. The hatched and gray inputs are normalized to the video. The three inputs are exclusive, i.e., no two cannot occur simultaneously. They are combined in an OR gate whose output is changed from positive to negative logic and combined with the X and Y reference marks for the display.

The reference marks for the riv display are velocity markers along the X-axis and range markers along the Y-axis. The reference markers for the t-r display are 10-second timing marks along the horizontal axis and range marks along the vertical axis. These marks are generated in the programmer and occur during the time when there is no video.

The combined video outputs from Card 41 are the encoded video inputs to the displays. The two oscilloscopes in the present

VISTAR equipment differ in that the t-r display is intensity modulated at the crt grid and the riv display is intensity modulated at the crt cathode. The encoded video is negative going for the ON condition, is amplified in a variable gain amplifier, and inverted for the t-r display. Only level shifting is required for the riv display, which is accomplished in vacuum tube video amplifiers. These amplifiers are located on separate chassis and mounted in proximity to the oscilloscopes.

The timing signals for the camera advance is determined in the programmer. Two identical 35-mm Bell and Howell cameras with provision for single frame operation are used in VISTAR. The normal operation of the cameras, upon receiving the advance signal, is to rotate the shutter plate 360 degrees. When the shutter begins to rotate, the film is advanced one-half frame and stops while the opening in the shutter passes between the lens and the film. Then the shutter closes and advances the film another one-half frame and the shutter stops. The timing of this operation is not applicable to the VISTAR displays. The timing for the riv communtation would have to depend upon coincidence of the open shutter and the commutation. The multiple sweeps per frame used in the t-r display is even more foreign to this mode of camera operation. To eliminate this difficulty, the shutter plates of both cameras have been removed, the instruments and oscilloscope made light tight, and logic incorporated to flash the indicator lights and the instrument lights.

The signal for the camera advance is derived from the PRIME gate in the programmer. This occurs after the oscilloscopes have been cycled and the indicator lights and instrument lights have been flashed, and long enough before the next oscilloscope cycle for the camera to advance the film. The signal to advance the t-r camera is simultaneous with the riv camera advance command but is gated to occur on every 64th cycle of the riv camera. The programmer logic is discussed in detail in that section. The camera drive circuits are transistor power switches which control the 28-volt d-c power necessary to advance the cameras. These circuits are driven by programmer signals.

Incadescent lights are incorporated in the instrument boxes for the two displays to illuminate the various instruments and/or data cards and clocks. The drive circuits for these lights are also transistor power switches controlled by signals from the programmer. These signals occur during the READ gate of the programmer cycle but are logically conceived to allow a longer switched ON time for a longer "exposure". Again, the t-r display is synchronized to the end of the 64-second scan.

The indicator lights are flashed at the same time the instrument lights are flashed. These indicator lights are driven from AND gates which have the flash signal as one input and the control signal as the other input. The AND gates for the tilt, Gain 1 and Gain 2 lights are incorporated in the programmer and the signal input to these lights comes from the programmer. The gates for the prf Hi and Low, O.S. 1, 2, and 3 in the riv Display and the filtering lights in the t-r display are mounted in the chassis with the power switching circuits located at the displays. The maximum threshold lights are an exception to the AND gate concept. As stated above, the threshold detectors whose outputs determine the Max Thres condition drive flip-flop circuits. These flip-flops are reset with the prime gate prior to the camera advance. The flip-flop outputs drive the Max Thres indicator lights directly.

The indicator lights in the riv display are commercial products employing a neon indicator light, transistor drive circuitry, and lens. The same drive circuit is used in the t-r indicator lights but the limited space availabe in the instrument box did not permit the use of the other lights and self-contained drive circuits.

### 3.6 ANTENNA STABILIZATION

A comprehensive description and analysis of the antenna and associated stabilization system is contained in a separate technical memo, a summary of which is included here.

The antenna pedestal and pitch stabilization system was constructed by modifying components from both the APS-23 and the APN-81 Radar Systems. The vertical reference gyro, erection control amplifier, and slaving control used were from the APN-81 Radar System and were utilized without modification. The servo amplifier and the antenna pedestal were modifications of APS-23 components.

The vertical reference gyro utilizes a viscous damped pendulum and a-c pickoffs to establish a vertical reference. The gyro then acts as an integrator to smooth the pendulum disturbances. To cause the gyro originally to erect and maintain the vertical, the outputs from the pendulum pickoffs are amplified and applied to torquers on the appropriate gyro axes, thus causing the gyro to process until the pickoff signals are zero. Under static conditions the gyro vertical will be within ± 5' of arc of the local vertical.

Two gyro erection modes are possible, normal erection and fast erection. When the system is first turned on, the gyro may be in any position and must be initially erected to the vertical. The fast erection mode ($60^{\circ}$/minute) is automatically applied for the first 5 minutes after system turn on. At the end of this period, a time delay relay returns the system to its normal erection mode ($5^{\circ}$/minute with a 14-second time constant for angles less than 15 feet of arc). The fast erection mode may be initiated manually during flight by applying +28 vdc to the fast erection control wire at the operator's station (the fast to normal duty cycle should not exceed 1:5 with 3-minute maximum fast ON time).

To prevent the gyro from erecting to a false vertical during a sustained aircraft turn, a rate gyro (slaving control) is used to disable the roll erection system when the rate of turn exceeds $25^{\circ}$/minute.

The gyro has $360^{\circ}$ freedom in roll and $\pm 80^{\circ}$ freedom in pitch.

The antenna is a four-foot parabola with a calculated moment of inertia of 33 lb-ft$^2$ about the pitch pivot. A low inertia, 60-watt rvomotor drives the antenna through a gear-lever arm ratio o 0:1 and a control transformer senses movement about the pitch pivot point. The control transformer is energized by a transmitter in the gyro and delivers its output (system error signal) to the servo amplifier. The amplifier supplies both gain (~ 1000) and compensation (lead-lag) before applying the error signal to the servo motor.

As an operational aid, a positive going error signal is generated in the servo amplifier and presented to the operator (TILT light). It has a sensitivity of 8 volts/degree and is limited to a maximum of 4.2 volts.

The following tabulation summarizes the performance limitations and equipment complement of the antenna system.

| | | |
|---|---|---|
| 1. | Stabilization range from normal angle of attack $\pm 9^{o}$. | |
| 2. | Maximum systematic error | 13' of arc |
| 3. | RMS random error | 11' of arc |
| 4. | Velocity constant | .05 degree/sec |
| 5. | Velocity limit | 25 degree/sec |
| 6. | Acceleration limit | 60 degree/sec$^2$ |
| 7. | Gyro erection time constant | 14 sec. |
| 8. | Temperature range | -55C to + 70C |
| 9. | Power requirements | |
| | 115 v - 380 to 420 cps | 0.8 amp max., 0.6 amp normal |
| | 28 vdc | 0.3 amp max., 0.2 amp normal |
| 10. | Equipment complement | |
| | 1 - Antenna and Pedestal | APS-23 (Modified) |
| | 1 - Gyro | C-1160/APN-81 |
| | 1 - Gyro Erection Amplifier | AM-743/APN-81 |
| | 1 - Slaving Control | N-1 Compass System |
| | 1 - Servo Amplifier | AM-193A/APS-23 (Modified) |

## 3.7 SYSTEM CONTROL

The control panel in the VISTAR provides for the primary power distribution and system control functions at a central location. The purpose of this section is to explain the various control functions embodied within the control panel. The adjustment of any of the control functions is discussed in the section on Calibration Procedures.

### Power Distribution and Turn-on

The primary power input to the VISTAR equipment is made through the circuit breaker panel at the top of the control rack. The breakdown of circuit breaker functions is listed below.

| Circuit Breaker | Fuzed Primary Current | Secondary Power Enabled |
|---|---|---|
| CB1-A | 20 amp | ± 75 vdc |
| | | Filaments: +500 Power Supply |
| | | ±150 Power Supplies |
| | | Oscillator Synchronizer |
| B | 14 amp | Stalo Klystron Power Supply |
| | | Low Power Modulator Filaments |
| | | High Power Modulator Filaments |
| | | T-R Tube Keep Alive Supply |
| | | Programmer 400-cycle |
| CB2 | 11 amp | Storage Tube High Voltage Power Supply |
| | | + 15 vdc Power Supply |
| | | - 24 vdc Power Supply |
| | | T-R Display Oscilloscope |
| | | RIV Display Oscilloscope |
| CB3 | 15 amp | Plate Supply: +500 vdc Power Supply |
| | | ±150 vdc Power Supply |

| Circuit Breaker | Fuzed Primary Current | Secondary Power Enabled |
|---|---|---|
| CB4 | 23 amp | ±12 vdc Power Supplies |
| | | ±20 vdc Power Supplies |
| CB5 | 10 amp (peak) | High Power Modulator |

The turnon procedure commences with CB1 and CB2 turned on at the same time. Besides turning on the indicated power supplies a 3-minute thermal time delay is also actuated at this time. These first two circuit breakers light the first two indicator lights, one light operating on 60-cycle power and the other on 400-cycle power. Therefore, if the primary power is present, it is indicated on these lamps. After the time delay has cycled, the third indicator light comes on. This indicates that the transmitter can be turned on. The circuit breakers CB3 and CB4 can be turned on after a reasonable wait to allow the vacuum tube filaments to warm-up. Circuit breaker CB5 controls the 400-cycle, 115-volt, 3-phase power used in the high power modulator. This power is further controlled by the time delay relay to assure that the output klystron is allowed to warmup the required 3 minutes. The time relay output passes through a switch which controls a four-pole relay for 400-cycle, 3-phase power. The primary power switches on the control panel have the following functions:

| Switch | Function |
|---|---|
| S1 | Not Used |
| S2 | 28 vdc: Controls the camera cycling |
| S3 | Not Used - Wired for Storage Tube P/S Remote |
| S4 | High Power Modulator<br>Low Power Modulator High Voltage if -150 vdc is ON |

Since the remote turnon for the storage tube power supply is not wired in, this function must be manually controlled at the power supply and turned on only after the storage tube has been allowed to warm up and when the +500 vdc is ON. The low power modulator high voltage is interlocked with the -150 vdc which is used for bias; therefore, it is important that it be ON before the high voltage is applied. To assure this condition, the -150 vdc is switched through a two-pole relay that controls the high voltages.

The 400-cycle power controlled by CB1 is used in the programmer as an input to the 400-cycle clock and for the oven heaters in the reference power supplies. The 28 vdc is used in the camera drive circuits and in the oven heater of the frequency measuring circuit of the offset oscillator. Currently, switch S2 can be used to switch the cameras ON and OFF but this affects the stability of the offset oscillator. It is recommended that this be changed so that the oven in the offset oscillator is energized continuously.

The final step in the turnon procedure is concerned with starting the offset oscillator. After all the power is ON, especially the ±20-vdc and ±12-vdc, the starting toggle switch on the offset oscillator must be thrown to permit the control loop to close. It is suggested that the output of the offset oscillator be periodically monitored with an oscilloscope to assure that it is operating properly. If the loop has dropped out of lock, by simply cycling the start switch it can be locked in again. If difficulty is experienced, check the voltages to be sure that the +12-vdc used in the offset oscillator control circuit is present.

## Controls

Three microswitches are incorporated into the control panel and wired into the terminal boards. One of the microswitches is used in the antenna servo system. After an unusually sharp turn by the aircraft, this microswitch should be depressed to activate the fast erection circuit in the antenna gyro. The other two microswitches are manual camera advance switches and should be wired in parallel with the automatic camera drive circuits to permit the manual advance of the cameras when necessary.

The prf toggle switch controls the prf of the VISTAR and the positions are appropriately marked. This switch selects the appropriate stage of the counter determining the prf and also provides the proper signal to the prf indicator lights in the riv display.

The gain control switch is the manual control of the gain programmer. The gain programmer is a subroutine of the programmer which generates the voltages that determine the particular 20 db portion of the signal dynamic range to be processed by the storage tube and the filters. The control choices are the 0 - 20 db range, the 20 - 40 db range, and 40 - 60 db range, plus automatic cycling between the first two and automatic cycling among the three. Associated with this control, to aid in the proper range selection, is the >40 db indicator light. If there is a preponderance of signal above this level, the operation should be such that the 40 - 60 db range is examined.

The filtering control selects the time constant of the filtering in the video output to the t-r display and enables the appropriate filter indicator light.

The offset oscillator control controls the o.s. indicator lights and the control voltage to the offset oscillator. The offset velocity of the riv display is a function of the setting of this switch and of the prf. The relationship and the theory of operation are discussed in the Functional Description section and the calibration and setting of the adjustments are discussed in the Calibration Section.

## Metering

Several meters to monitor important functions of the VISTAR are on the control panel.

The secondary voltage meter and its associated selector switch is used to monitor the following secondary voltages in the VISTAR:

±12 vdc

±20 vdc

+15 vdc

-24 vdc

±150 vdc

The frequency meter is connected to one phase of the 3-phase power applied to the high power modulator to monitor the 400-cycle frequency used in VISTAR. This is to assure that the frequency is within the 5 per cent tolerance required by this unit.

The crystal current meter indicates that the Stalo is operating and along with the klystron current meter is indicative of the proper operation of the transmitter and receiver.

SECTION IV

## 4. ADJUSTMENTS AND SPECIAL NOTES

The various adjustments for the VISTAR components and subsystems are contained either in this section or in technical notes or schematics delivered with the VISTAR system. In the tabulation which follows, reference is made to the appropriate notes or schematic that is applicable, or a direct procedure is given. In many cases, the procedure is given elsewhere in this report and is so referenced.

### 4.1 LIST OF ADJUSTMENTS

#### 4.1.1 Transmitter

Stalo

See H.P. & Dymec Operating Instruction Manuals.

Klystrons

See Varian Instruction Manuals.

Driver Power

Variable attenuator adjusted for maximum output.

60 Mc Osc.

Tune slugs for maximum output.

60 Mc Amplifier

Tune slugs for maximum output.

#### 4.1.2 Receiver

Synchronous Detector

Separate technical note. (Tune butterfly capacitors for maximum signal and best balance.)

Single-Sideband

Tune screen potentiometer for minimum 60-Mc output without offset oscillator input. Tune butterfly capacitors for maximum output with offset oscillator input. (A 30 db, 60-Mc rejection is required.)

Burst Generator and Other Tuned Amplifiers

Tune slugs for maximum output.

Stalo Power

Adjust Stalo power for 1 ma. crystal current.

Preamplifier

See R.S. Electronics Schematic.

Log I-F & Lin I-F Amplifiers

See schematics for slug settings.

S.T.C.

See Programmer description (calibration curve supplied with VISTAR equipment). Adjustments are on cards 43 and 44.

Offset Oscillator

Monitor the offset oscillator frequency with a counter and adjust the potentiometers associated with the particular switch setting and center velocity according to the following table.

| Position | PRF | Offset Oscillator ($f_{os}$) Frequency (cps) | Center Velocity |
|---|---|---|---|
| 1 | Low | 95 | -12.5 meters/sec |
| 2 | Low | 885 | 0 meters/sec |
| 3 | Low | 1657 | +12.5 meters/sec |
| 4 | High | 188 | +25.0 meters/sec |
| 5 | High | 1751 | 0 meters/sec |
| 6 | High | 3315 | -25.0 meters/sec |
| 7 | High | 4878 | -50.0 meters/sec |

Note: The formula for computing this frequency is:

$$f_{os} = \left[ 56 - \frac{\text{center velocity}}{\Delta \text{ velocity}} \right] \Delta f_d$$

where: center velocity = Desired calibration of the center of the riv display

$\Delta f_d$ = Doppler resolution by filter bank

= 31.27 cps for high prf

= 16.64 cps for low prf

$\Delta$ velocity = System velocity resolution

= 0.5 for high prf

= 0.25 for low prf

### 4.1.3 Storage Tube, Filters, and Commutator

Storage Tube

See description of storage tube unit.

Filters

Individual adjustment for gain on each filter, however, is restricted in effective range. Use 400-μsec gated signal when adjusting for equal response.

Drive amplitude should be adjusted for linear operation and maximum dynamic range.

Local Oscillators

Adjust filter heterodying oscillators to 625 kc and 875 kc for the lower and upper input frequency bands, respectively.

RIV Gain Control

Pulse the linear i-f amplifier with 60-Mc pulse bursts and monitor gain. Adjust appropriate potentiometers on Card 41 for the -20 db and -40 db gain settings when the corresponding gain has been set on the control panel.

### 4.1.4 Displays

T-R Threshold Circuits

Use 60-Mc pulse bursts into Log i-f amplifier. Adjust threshold settings on Card 54 according to the appropriate input level.

RIV Threshold Circuits

Use video pulse generator directly into threshold circuits. Normalize calibration to riv noise level. Adjustments are on Card 53.

TILT Threshold Circuits

Calibrate antenna servo and accelerometer inputs with d-c voltage source. Note that the accelerometer input has two adjustments for bipolar settings. Adjustments are on Card 52.

Sweep and Scan Gain

Adjust t-r and riv deflection amplitude with external or internal gain controls as applicable (riv sweep uses internal sweep generator).

Video Levels

Adjust composite video level for t-r and riv displays with video amplifier gain control (external to scopes).

## 4.2 OVER-ALL SYSTEM CALIBRATION

Over-all system calibration should be broken down into several phases. Estimates of transmitter peak power can be made by direct measurement with a calibrated wattmeter, bolometer, and 50 db directional coupler in the antenna waveguide. A fixed duty cycle can be obtained to facilitate measurements by disabling the normal prf triggering to the transmitter and substituting a pulse generator with known repetition frequency in the general range of VISTAR operation, e.g., 5 kc.

Receiver system calibration should be started with the t-r channel. Most adjustments will have been accomplished with the setting of the threshold circuits. However care should be exercised to assure that the noise level at each point along the channel is governed by the effective noise at the preamp input to assure maximum sensitivity and dynamic range. Verification of t-r threshold settings should be made as the calibrated X-band test set is varied in amplitude over the dynamic range of the channel. Note that the test set must be pulsed to allow the signal to pass the video circuits.

A somewhat different procedure must be employed to calibrate the riv channel. A preliminary calibration test should be performed using a calibrated 60-Mc generator into the linear i-f amplifier. A disabling relay must be employed in the storage tube video input line, to remove extraneous modulation from the storage tube grid during READ. This relay is enabled by the WRITE mode signal. The logic and driver for this relay has been constructed and delivered with the VISTAR equipment. By displacing the frequency of the 60-Mc generator from the system 60-Mc oscillator, synthetic Doppler frequencies may be generated for processing. A mixer, low-pass filter, and counter would assist in the velocity calibration when used to monitor the difference frequency of the two 60-Mc sources. Several frequencies must be explored for a representative response from the filter system. Again, care must be taken to assure that the riv output noise level is represented by the receiver noise level and not some intermediate circuit.

An over-all calibration of the riv channel is difficult, as controlled Doppler generation at X-band is difficult without special test equipment. However, if accurate gain measurements of the mixer and i-f preamplifier are made, a composite calibration can be computed for the riv channel.

Antenna gain patterns should be verified by additional measurements at an antenna range. Also, the boresight axis should be determined to allow correct leveling of the servo platform in the aircraft.

It would be desirable to determine the over-all performance of the VISTAR through the use of a calibrated "weather-like" signal source. However, the design of that experiment has not been performed and will not be undertaken here.

SECTION V

## 5. CONCLUSIONS AND RECOMMENDATIONS

It is difficult to draw any significant conclusions with respect to the VISTAR program at this particular time, since the application of VISTAR to the mission for which it was intended has not as yet been accomplished. Therefore, only observations concerning portions of the VISTAR equipment that require criticism will be noted here.

From an over-all point of view, the VISTAR must suffer the criticism that is leveled at any instrumentation of breadboard caliber with several years of longevity. Although the microwave system has been modified to eliminate some of the original difficulties, it still reflects techniques and components now obsolete. Any future VISTAR would require a completely fresh approach to this portion of the system.

The heart of the VISTAR system lies in the storage tube and associated filter system. Unfortunately, this also represents the weakest link in the present instrumentation. Some of the more outstanding difficulties of the storage tube unit in terms of both component limitations and mission application are:

1. Limited dynamic range in amplitude.
2. Restricted sample capacity (resolution) reflecting a low number of possible samples for integration.
3. Degradation of those normalized Doppler frequencies requiring the maximum resolution of the tube.
4. A sampled output, as well as input, that requires precision timing and weighting for spectrum sideband control.
5. A requirement for many precision filters whose transient characteristics are well matched over a band of frequencies.
6. Degradation of normalized low Doppler frequencies due to an insufficient number of cycles stored for filter analysis.

Some of these maladies can be alleviated by replacing components with newer counterparts; however, some of the difficulties are inherent to the processing philosophy. Obviously in a long-term improvement program nearly all of the VISTAR equipment would require redesign for compactness, reliability, increased efficiency, etc. as well as a complete reassessment of system philosophy. However, for the present system goals, only a few changes can be made. Since the storage tube processor is the weakest element in the present system, it is recommended that an alternate method of accomplishing the task of this portion of the VISTAR be devised and the required equipment be developed. This procedure will allow a transition to a better instrumentation without complete redesign of the whole VISTAR, as well as provide a backup to the present processor, whose reliability and utility are questionable.

SECTION VI

# 6. PERSONNEL AND RELATED CONTRACTS

## 6.1 PERSONNEL

The following personnel were intimately associated with this contract and collaborated in the assembling of the information contained in this report.

Glenn E. Kinzer
John F. Class
Jack C. Lorden
Warren F. Christensen

## 6.2 RELATED CONTRACTS

A contract which is related to the present one is:

Automatic Cloud Measuring Indicator
Contract No. DA-36 SC-90825
U.S. Army Electronics Research and Development Laboratory