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TECHNICAL NOTE

INTERFERENCE MONITOR FEASIBILITY STUDY

(PHASE II)

RADIATION INCORPORATED
MELBOURNE, FLORIDA

CONTRACTOR'S REPORT NO. 1593-2

CONTRACT NO. AF 30(602)-2695

TECHNICAL NOTE NO. 2

PREPARED FOR

ROME AIR DEVELOPMENT CENTER

RESEARCH AND TECHNOLOGY DIVISION

AIR FORCE SYSTEMS COMMAND

UNITED STATES AIR FORCE

GRIFFISS AIR FORCE BASE

NEW YORK

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RADC-TDR-63-198

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TECHNICAL NOTE NO. 2

PROJECT NO. 4540

TASK NO. 454001

RADIATION INCORPORATED
MELBOURNE, FLORIDA

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FOREWORD

The study and investigation reported in this Technical Note have been supported by the Applied Research Branch of the Rome Air Development Center, Griffiss Air Force Base under Contract AF 30(602)-2695. The study is included in Project 4540, Task 454001 under the direction of:

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ABSTRACT

Approaches which will permit monitoring the RF Spectra emitted by pulse radar transmitters from 200 MC to 40 GC have been investigated. Techniques for monitoring levels of peak effective radiated power down to the permissible levels of MIL-R-27055 have been sought. The monitor approach which is described employs multiband swept tuned receivers, each covering approximately octave bandwidths simultaneously. Identification of the source of received spurious signals is provided by correlation with trigger pulses derived from the radars via coaxial cable.

A breadboard monitor has been constructed in order to permit demonstration of the approach. The breadboard, operating in the range from 2 to 4 GC, will be tested at the Verona Test Site.

PUBLICATION REVIEW

This report has been reviewed and is approved.

Approved: Louis F. Moses
LOUIS F. MOSES
Task Engineer
Applied Research Branch

Approved: Samuel D. Zaccari
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1.0 INTRODUCTION

The Interference Monitor Feasibility Program is a study for the purpose of developing techniques for the continuous monitoring of the spectrum radiated by pulse radar transmitters. Techniques are sought which will permit recording only when the levels of peak effective radiated power exceed the values specified in Paragraph 3.5 of MIL-R-27055 (USAF).

1.1 Objective of the Program

The purpose of this program is to define the problems in monitoring the spectrum from 200 MC to 40 GC and investigate possible technical approaches. A breadboard monitor operating from 2 GC to 4 GC will be constructed and utilized to demonstrate the feasibility of the most promising monitor technique.

The breadboard monitor will be tested at the Verona site and test results will be obtained from operation in the Electromagnetic Field created by several energized radar transmitters.

A final report will summarize the findings of the program and present the results of the breadboard tests. Based on the study effort and the breadboard tests, detailed recommendations for spectrum monitoring will be included.

1.2 Previous Work

The first Technical Note, RADC-TDR-62-505 covered the work on the program to 20 August.

Effort to that time consisted of study in the areas of defining the expected spectral characteristics, determining the factors to be considered in locating the monitor, and assimilating and organizing other information relative to RF component availability and performance.

Conclusions reached during the Task I work were the following:

1. An open field measurement approach would be employed. The measurement distance was found to be non-critical and a range of $R = \frac{KD^2}{\lambda}$ (with K equal 1 or 2), was adequate, and produced acceptable results. (K is a constant at the fundamental frequency; D is the transmitter antenna aperture; λ is the fundamental wavelength.)

2. The maximum expected levels of spurious received signals were defined based on the results of available data.

$$Pn = Po - 20 - 46 \log \frac{fn}{fo},$$

where Pn is the effective radiated power in dbw at frequency (fn), and Po is the power in dbw at the fundamental frequency (fo).

3. The monitor receiver bandwidth would be a constant 5 MC.

4. Precision of amplitude and frequency resolution would be secondary to rapidity of gathering data.

1.3 Requirements of the Monitor

The monitoring techniques being sought must provide the following:

1. Frequency coverage must be continuous from 200 MC to 40 GC.

2. A recorded output must be provided when the radiated spurious signal levels exceed the values specified by MIL-R-27055 (USAF).

3. Detection of levels and changes in levels at harmonic and spurious frequencies is required.

4. The recording and display techniques should permit rapid visual determination of the radiated spectrum as well as data reduction by computer techniques if desired.

5. Identification of the source of each received spurious radiation is required.

6. Operation in the fields produced by several energized radar equipments is required.

7. The monitoring process should not require that the radar equipments be operated in any unusual modes. Ideally, the radar sets can be monitored with no interference with normal missions.

1.4 Scope of this Report

This Technical Note summarizes the work done on Tasks II and III to the completion of the breadboard fabrication.

The Task II effort was a system study covering system analysis and design. Task III includes both the breadboard development and the testing at the Verona site. The results of the Verona tests to be performed will be reported in the final report.

2.0 TECHNICAL APPROACH

In order to provide a capability for monitoring levels of peak effective radiated power, a receiving system, including antennas is required. The sensitivities, dynamic range, bandwidth, and other characteristics have been determined for a monitor to be employed at the Verona Test Site.

In the consideration of possible approaches the availability of major components has been weighted heavily, since it is not intended that a component development be required to implement the chosen monitor approach.

In order to permit simultaneous reception of MIL-R-27055 (USAF) threshold signals from multiple radar transmitters, widebeam monitor receiving antennas and sensitive receivers are required.

A much simplified block diagram of the monitor is shown in Figure 1. The monitor is basically a nine band superhetrodyne receiver. Each of the frequency bands is associated with one or more antennas designed to provide the required beam coverage, gain and polarization characteristics.

A sweep generator simultaneously tunes each monitor band through its frequency range and provides the time base for the recording.

The identification and display subsystem permits direct operator observation and permanent storage of received data. Provision is included for timing and pulse width adjustments to permit automatic identification of the source of received spurious signals.

2.1 RF Subsystem

The studies undertaken during the Task I Phase of the program have shown that the monitor RF section should meet the following basic requirements:

a) Provide an over-all sensitivity on the order of -90 dbm for a 10 db signal-to-noise ratio.

b) Provide octave bandwidth tuning capabilities.

c) Large dynamic range.

A block diagram of the chosen design is shown in Figure 2.

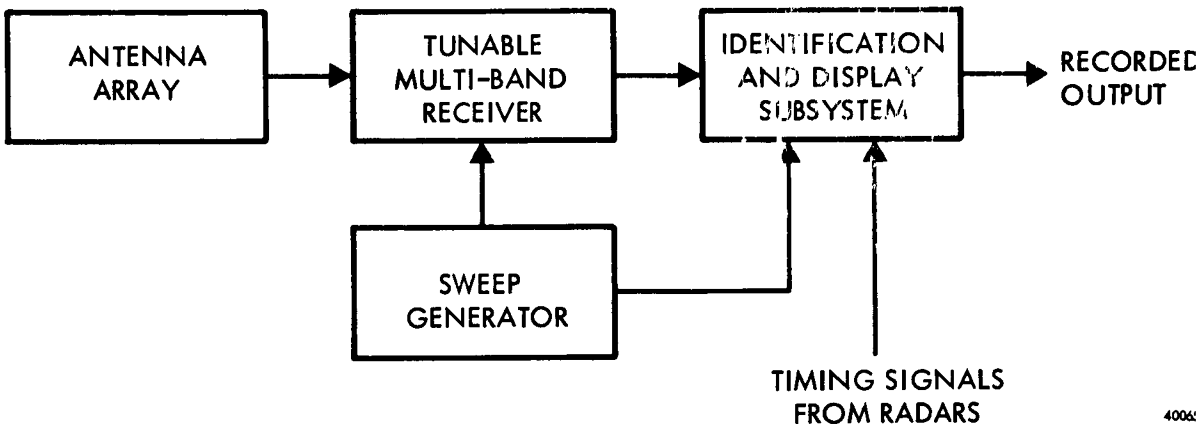

Figure 1. Simplified Monitor Block Diagram

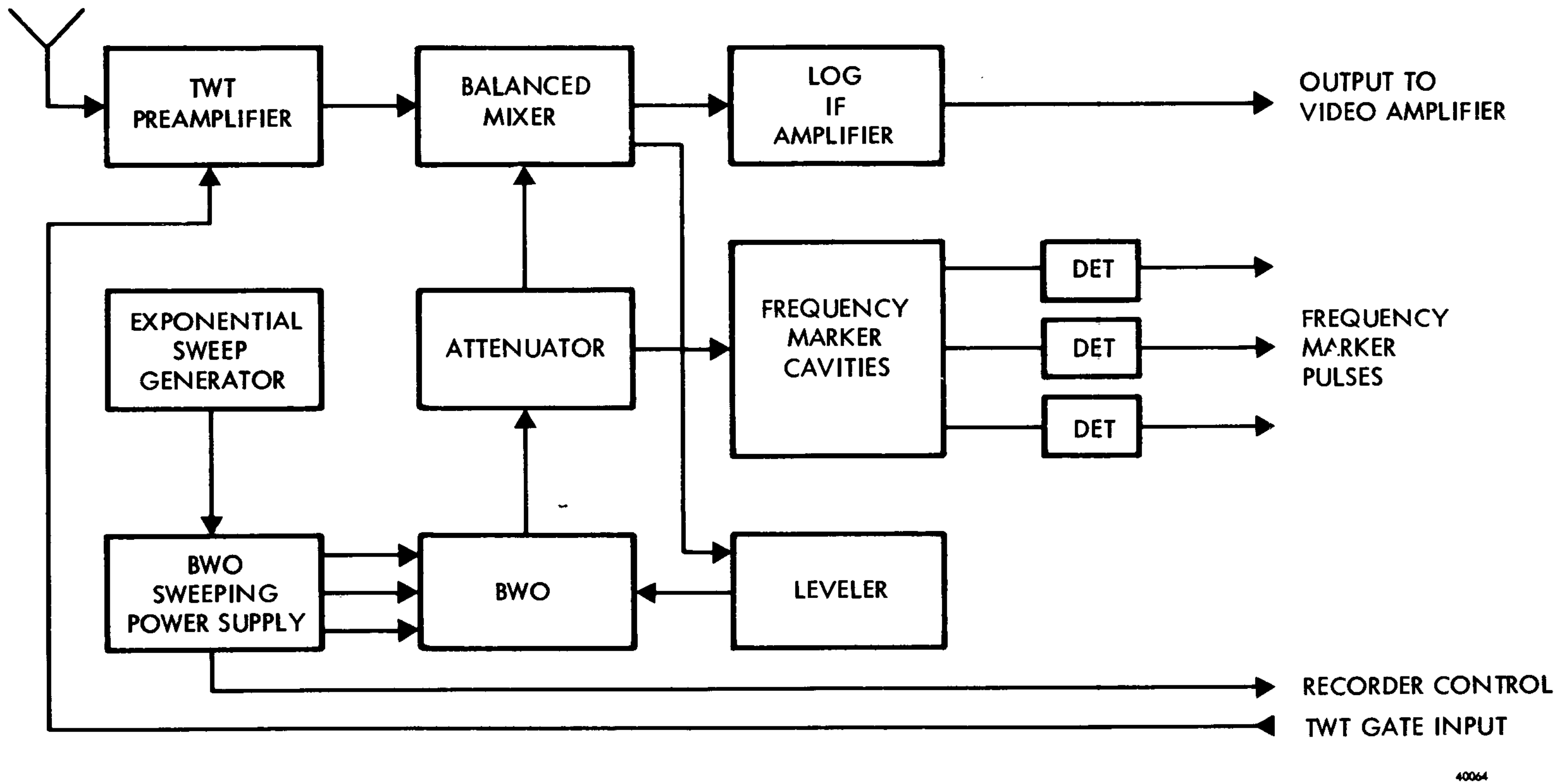

Figure 2. Block Diagram – Monitor RF Subsystem

2.1.1 Antenna Subsystem

It was noted in the previous Technical Note that the following factors had to be considered in determining the specific type of antennas to satisfy the monitor function:

a) Frequency

b) Polarization requirements

c) Site considerations

d) Gain and bandwidth

2.1.1.1 Frequency

The monitor antennas must provide essentially the same frequency coverage as the individual receivers.

2.1.1.2 Polarization Requirements

A survey of radar transmitting antenna characteristics indicates that the monitor antenna must be capable of receiving both linear horizontal, linear vertical, and right circular polarization. This can be accomplished by using right circularly polarized monitor antennas and accepting the three db loss which would occur when the antennas being monitored are linearly polarized.

2.1.1.3 Site Considerations

The monitor antennas will measure the peak effective power radiated by the antennas under test at the site of the monitor antennas, thus eliminating the problem of site considerations since they now become a part of the measurement.

2.1.1.4 Gain and Beamwidth

At the site presently under consideration for location of the monitor an azimuth angular coverage of approximately 130 degrees is required to monitor all of the desired radar antennas. Further, based upon available receiver sensitivity and required system sensitivity, the following gain at the receiver input will be required for the ultimate monitor antennas:

Frequency GC	Gain db
0.25 - 0.5	3
0.5 - 1.0	6
1.0 - 2.0	2
2.0 - 4.0	2.3
4.0 - 8.0	8
8.0 - 12.0	8
12.0 - 18.0	12
18.0 - 26.5	11
26.5 - 40.0	13

It is obvious that the requirement for broad angular coverage with relatively high gain requires a very specialized antenna design. For the very high frequency systems one antenna such as a biconical horn could be designed to provide the required performance. However, the further requirements of circular polarization and octave bandwidth frequency coverage begin to make such a design somewhat impractical. The most feasible approach to provide the required monitor antenna performance is the utilization of several different antennas for each frequency band where required. By either switching between the antennas or employing multiple monitor channels in a frequency band, the design of the required antennas is simplified.

In the event that radar transmitters to be monitored are ever to be situated away from the presently employed locations, coverage of the new equipments can be provided by similarly employing multiple monitor antennas.

2.1.2 Preamplifier

2.1.2.1 Monitor RF Sensitivity

In order to define the sensitivity requirements of the monitor, it is necessary to define the maximum range between the transmitting antenna and the monitor receiving antenna. This requires that an optimum physical location be chosen for the monitor at the Verona Test Site. Due to the fact that the physical locations of the radars at Verona change from time to time, coverage of the entire one mile strip must be considered in determining an optimum monitor location. Since the entire strip must be covered, the monitor should be located at a point equidistant from each end, somewhere along the perpendicular bisector of the strip. The greater the distance between the monitor and the strip, the less measurement error due to differences in path length and the smaller the coverage angle required, however, sensitivity requirements also increase with greater distance. Therefore, a compromise must be made between measurement error, coverage angle and sensitivity.

Figure 3 shows a rough map of the Verona site. It may be seen from this map that if the monitor location is to be inside the fenced area of the test site, then the point on the perpendicular bisector of the radar strip which will produce the least measurement error due to range differences is located just inside the southwest boundary fence. This location is approximately 1950 feet from the center line of the radar strip and about 3060 feet from either end. The maximum measurement error due to range difference above would be $20 \log \frac{3060}{1950} \approx 4$ db.

The required monitor sensitivity for this location may be calculated in the following manner:

From MIL-R-27055 (USAF) the transmitted power at specification level is

$P_T = 1.11\ (10)^{-9}\ f^2$ watts where f is in megacycles.

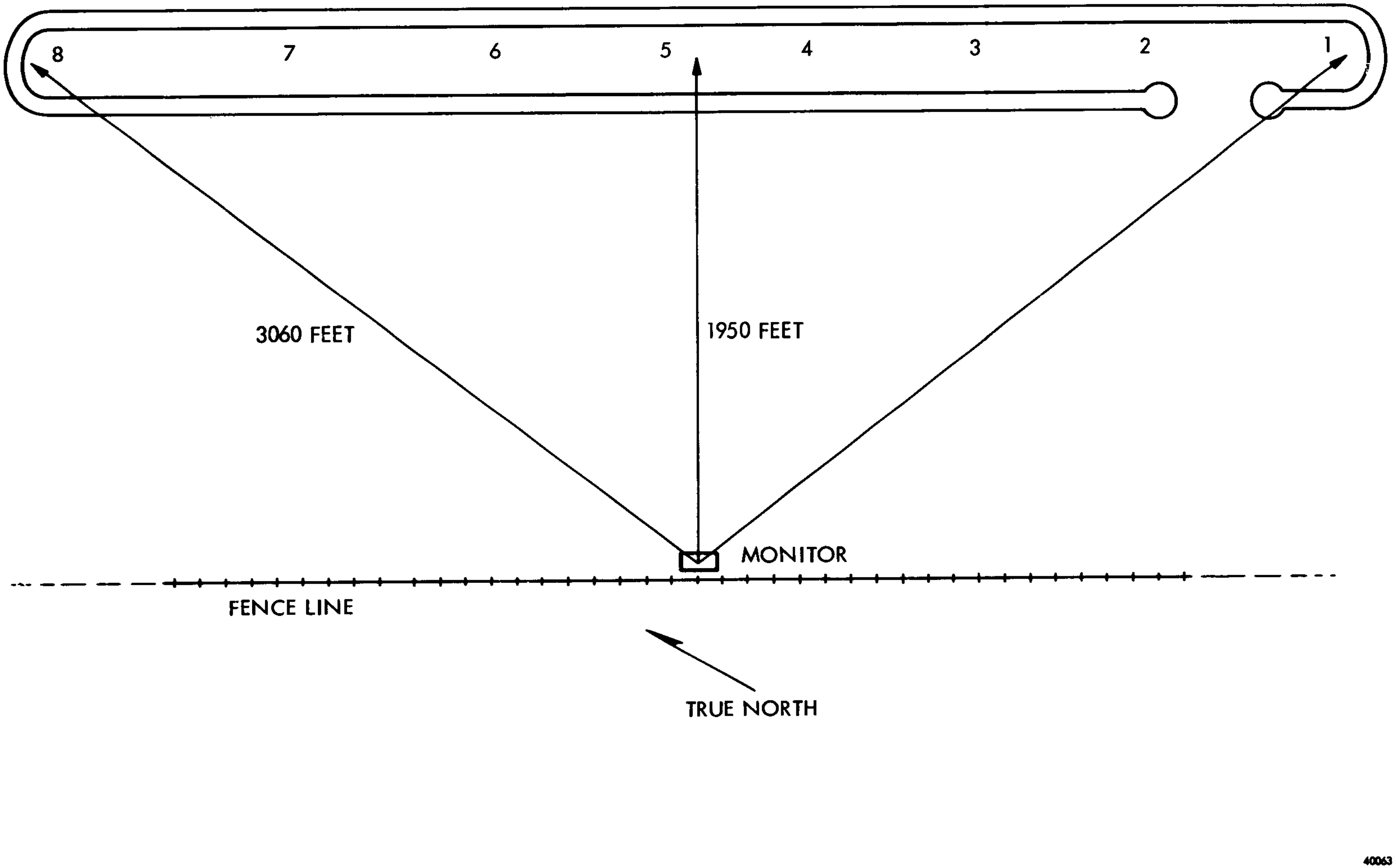

Figure 3. Verona Test Site - Equipment Layout

The power received at the monitor location due to path attenuation:

$$P_R = \frac{P_T}{4.56(10)^3 f^2 d^2} \qquad (1)$$

where f is in megacycles and d is in miles

then

$$P_R = \frac{1.11(10)^{-9}}{4.56(10)^3 d^2} = \frac{2.44(10)^{-13}}{d^2}$$

From Figure 3 the maximum slant range is 3060 feet or approximately 0.58 miles then for d = 0.58 miles,

$$P_R = \frac{2.44(10)^{-13}}{\left[5.8(10)^{-1}\right]^2} = \frac{2.44(10)^{-13}}{3.37(10)^{-1}} = 7.24(10)^{-13} \text{ watts}$$

P_R may be expressed in dbm as:

$$P_{R_{dbm}} \approx 20 \log \left[6.4(10)^4 d\right]$$ where d is in miles or for d = 0.58 miles

$$P_{R_{dbm}} = 20 \log \left[6.4(10)^4 5.8(10)^{-1}\right] = -91.4 \text{ dbm.}$$

In order to discuss the practical problems associated with implementing a receiver which will meet the desired minimum sensitivity requirement of -92 dbm, it is necessary to define what is meant by a minimum detectable signal. For the monitor receiver, a minimum detectable signal will be defined as being 10 db greater than the average noise level. This means that the power of a received pulse measured at the output of the IF amplifier before detection will be 10 db greater than the average noise power.

This definition of a minimum detectable signal provides approximately a 2 db margin over the normal definition of a tangential signal. This arbitrary margin is chosen to improve the accuracy of signal detection. Signal detection in the monitor receiver will not be visual, rather the signal will be detected by a threshold detector which must sense the voltage level of the signal with reference to a predefined threshold voltage.

Task I studies defined the minimum required signal bandwidth as 5 MC. For the purpose of sensitivity discussion it will be assumed that the effective noise bandwidth of the monitor receiver is also 5 MC. In order to further simplify the sensitivity discussion it will be assumed that the input vs. output characteristics of the monitor are linear over the dynamic range required by a minimum detectable signal.

As a first approximation if zero db antenna gain is assumed, then the maximum permissible noise figure for the monitor receiver is:

$$NF_{db} = KTB_{db} - S/N_{db} - \text{MINIMUM SIGNAL LEVEL}_{db}$$

$$NF = 107-10-92 = 5 \text{ db}$$

A 5 db noise figure is beyond the present state-of-the-art in the top five bands (4-40 GC) of the monitor when octave bandwidth coverage is desired. Antenna gain can be traded for noise figure, however, but this increases the antenna design problem since wide angle coverage is desired. A 5 db noise figure can be approached in the four lower bands (.25-4 GC) by the use of a low noise preamplifier.

2.1.2.2 Low Noise TWT Preamplifier

Ideal characteristics for a monitor preamplifier would be:

a) High gain

b) Low noise

c) Large dynamic range

d) Octave bandwidth tuning capability (preferably by electronic means)

e) 3 db bandwidth on the order of 7 MC

The first four characteristics can be approached closely by the traveling wave tube amplifier. Unfortunately narrow band TWT's such as backward wave amplifiers and dispersive amplifiers (wave velocity varies as a function of frequency) are not presently available in low noise types and are also not available to cover the entire 0.2 to 40 GC frequency range. This makes the broadband low noise TWT the only logical choice to meet the monitor preamplifier requirements.

In the top four bands, 8 GC to 40 GC, the maximum attainable sensitivity is approximately the same with or without the use of a traveling wave tube preamplifier. This is due to the relatively high noise figure of presently available TWT's.

In the five lower bands the system noise figure can be cut approximately in half by the use of presently available low noise TWT preamplifiers.

The use of TWT preamplifiers in the top four bands is still an advantage from the standpoint of system noise figure since it removes the stringent low noise requirements placed on the mixer-IF combination. This allows more freedom of choice for the IF frequency and reduces the design problems of obtaining minimum noise figure with a logarithmic IF amplifier characteristic.

For example, for the 26-40 GC band, a GE ZM3116# TWT may be used which has 20 db gain and a 15 db maximum noise figure. Using this preamplifier, a mixer-IF noise figure of 20 db would only raise the over-all noise figure 0.13 db.

Some of the other properties of the TWT preamplifier are perhaps more important from the standpoint of monitor system requirements. Based on the studies made in Task I a practical monitor may be exposed to signals over a dynamic range of 130 db or maximum signal levels on the order of +40 dbm (10 watts). Signals of this magnitude could cause a severe overload problem and depending on duty factors, possible mixer crystal damage. The need for a broadband limiter is immediately apparent.

The low noise TWT provides the necessary limiting action with very fast recovery time. The TWT limiting characteristic is not ideal, however, due to the fact that increases in input beyond the saturation point cause a reduction in output. This negative slope complicates the problem of amplitude measurement since a maximum signal level is not well defined. As the input signal is further increased the power output will continue to drop until a minimum is reached and then it will increase again to a maximum much lower than the first maximum.

Further increases in input power will cause the power output to go through more maximum and minimum points with the maximum points never approaching the original maximum power output. This overload behavior of the TWT preamplifier makes it possible for the monitor to accommodate future radar power increases without modifications.

Perhaps the greatest advantage gained by employing a TWT preamplifier in the monitor system is the ability to rapidly change the receiver gain. The change in gain is accomplished by decreasing the voltage on one or more of the TWT anodes. Due to the wide bandwidth of the TWT it is possible to very rapidly gate the tube off and on by this method. When the tube is gated off the attenuation is approximately the same as the cold attenuation of the tube. A typical cold attenuation is on the order of 20 to 30 db.

This gating ability provides the monitor system with a means of preventing overload and video feedthrough due to strong signals without loss of data from other signals on the same frequency which are not time coherent. Of course, it is necessary to have firing information (pretrigger pulses) from the offending signal at the monitor site.

Another factor in favor of employing a preamplifier on the monitor receiver is the reduction of radiation of the monitor local oscillator fundamental and its harmonics. For example, a typical local oscillator power level is one milliwatt. With a local oscillator rejection of 20 db from the mixer and an antenna gain of 10 db, the effective radiated power of the local oscillator would be -10 dbm. Since the local oscillator would be sweeping through an octave band, a radiated power of -10 dbm could constitute a serious interference problem.

The choice of a broadband low noise TWT for the preamplifier has some unfortunate disadvantages also. Because of the wide open octave bandwidth, noise at the image frequency is not attenuated and a 3 db degradation in system noise figure results.

The extreme broadband response of the TWT preamplifier can be a problem in a high signal level environment. A single high level pulse signal or combination of lower level pulse signals at different frequencies which are time coherent may cause collector current saturation which will result in a gain reduction at all frequencies within the band. The amount of gain reduction is a function of the level of collector current saturation. A 10 db reduction is a typical example. A more sophisticated monitor would require a low insertion loss bandpass filter preceding the TWT preamplifier to eliminate the problem of overload due to strong signals outside the band of interest. The overload problem within the band of interest is not as severe as it appears since a saturation level signal would ordinarily not be time coherent with any other spurious signal within the band. Therefore, unless a low level signal occurs during the recovery time of the TWT, there will be no loss of sensitivity and reception of low level signals will not be seriously affected by the saturation level signal.

If pre-trigger pulses are available from the radar which is causing system overload, a negative gate may be applied to the TWT to control the amplifier gain during the period of the overload pulse. The TWT may be completely cut off so that the "cold" tube attenuation is available - typically 20 to 30 db. Recovery time of the TWT is very short, in the order of 10 nanoseconds.

The TWT for or against argument has several other factors on the against side which must be considered. Foremost among these factors is cost. The initial cost amounts to an appreciable percentage of the total cost of a monitor system and when spare tubes and increased maintenance costs are added to this the total cost is a significant factor.

The cost of employing TWT preamplifiers is further increased if such factors as increased cooling problems are taken into account. The average solenoid for a low noise TWT dissipates approximately 500 watts. This power is in addition to the heat losses in the high voltage supply for each tube.

The original packaging of a system employing TWT's is complicated by the tube's susceptibility to stray magnetic fields which can cause beam defocusing and amplitude modulation of the output.

2.1.3 Mixer

The basic requirements for the mixer used in a monitor receiver are:

a) Octave bandwidth

b) Low conversion loss

c) Low noise figure

d) Low sensitivity as a direct video detector with no local oscillator input.

The requirements are best met by a balanced crystal mixer. Unfortunately, balanced crystal mixers are not presently available to cover the entire 0.2 to 40 GC frequency range and single ended types will have to be used above 8 GC. For a given frequency range, the balanced mixer will have a lower noise figure than the single ended mixer due to the cancellation of local oscillator generated noise. The carrier rejection feature of the balanced mixer tends to reduce its sensitivity as a video detector in the absence of local oscillator input. In the area of bandwidth, however, the single ended mixer is superior to the balanced mixer.

An alternate approach to the mixer, which has been previously discussed in the Technical Note on Task I, is the TWT mixer.

As discussed previously, this approach has no great advantage unless a VHF or UHF IF amplifier is desired for image rejection in a particular application. Im rejection by using a high IF frequency would not be feasible in the present monitor due to lack of preselection.

2.1.4 IF Amplifier

Previous study under Task I has indicated that a minimum RF bandwidth of 5 MC will be required. This corresponds to a bandwidth of $\frac{1}{t}$ where the minimum pulse width, t, is 0.2 microseconds. A bandwidth of $\frac{0.7}{t}$ will reproduce the peak amplitude of a 0.2 μsec pulse with approximately 1 db error.

Since the TWT preamplifier will have a wide open octave bandwidth the monitor RF bandwidth will be determined by the bandwidth of the IF amplifier.

2.1.4.1 IF Amplifier Center Frequency

The choice of an optimum IF center frequency for the monitor receiver is a complex problem involving careful consideration of several conflicting factors. The following discussion will cover the more important factors which affect the monitor receiver requirements.

2.1.4.1.1 The Image Factor

It is well known that any superheterodyne receiver will have an image response which is separated from the local oscillator frequency by twice the IF frequency.

The amplitude of this image response will depend on the degree of preselection preceding the mixer. In the proposed monitor receiver with an octave bandwidth preamplifier preceding the mixer, the image response will be essentially the same as the desired response. With no preselection the approach of raising the IF frequency to obtain image rejection is not a practical solution since IF frequencies from 150 MC in the low band (200-500 MC) to 7 GC in the high band (26-40 GC) would be required before any image rejection would be obtained for signals at the lower edge of each band.

Narrow band low noise preselectors that can be tuned over octave bandwidths and can be made to properly track the local oscillator are not available at the present time and most certainly present some formidable design problems.

Another approach is to accept the presence of the image and try to minimize its effect as a confusion factor to signal identification. If the IF frequency is lowered to a frequency where the separation between signal and image becomes insignificant when observed on the particular visual display employed then the confusion factor of the image is eliminated. Unfortunately, other factors must be considered when the IF frequency is lowered.

2.1.4.1.2 The Video Frequency IF

The theoretical limit to lowering the IF center frequency is the point where the IF pass band is centered at zero frequency or DC. The IF amplifier then becomes a video amplifier whose bandwidth is half the original IF bandwidth. This bandwidth reduction is possible only if the signal and the local oscillator are at exactly the same frequency and are phase coherent. Under these conditions there is no image response and no reduction in sensitivity due to image channel noise. Unfortunately for a sweeping receiver application it is not feasible to phase lock the local oscillator to each received signal, especially when the received signal might be a single pulse of 0.2 microseconds duration.

For the swept receiver application two factors must be considered when the IF center frequency is lowered into the video region:

a) The effect on pulse fidelity

b) Mixer noise figure increase

For purposes of discussion, assume a video IF amplifier whose pass band extends from DC to 5 MC. Further assume that the received signal is a 0.2 microsecond burst of transmitter carrier frequency which, for example, may be 3 GC.

When this video IF amplifier is used in a swept receiver it would appear that the output pulse fidelity is seriously affected. This becomes apparent when it is realized that the translated carrier frequency of the received signal which is the difference between the local oscillator frequency and the received signal frequency, may be anywhere from DC to a frequency determined by the particular signal frequency and the limits of the local oscillator sweep. However, if the difference frequency falls within the assumed IF pass band it may be any frequency between DC and 5 MC and will be determined by the relative timing between the local oscillator sweep and the transmitter firing time. For a 0.2 microsecond pulse this translated carrier becomes a maximum of one cycle when the translated frequency is 5 MC and only portions of a cycle for any lower frequency. For the purposes of the monitor receiver where pulse fidelity is relatively unimportant so long as a measure of the peak amplitude is retained, the situation is not as bad as it appears.

Since there is no direct correlation between the frequency of the transmitted signal carrier and the pulse modulation, the amplitude of a small portion of a cycle of the difference frequency between the transmitted carrier and the local oscillator is random. This in effect would cause a jitter in the received pulse amplitude which would be insignificant at high difference frequencies and would increase to a maximum as the difference frequency approached zero. This effect can cause considerable error in pulse amplitude measurement when the number of pulses falling within the IF pass band is very low. However, due to the random nature of the jitter in pulse amplitude, when the number of received pulses is a reasonable value, say ten or more, the probability of receiving a level which represents the peak amplitude is greatly increased. As discussed earlier under monitor sensitivity requirements, another factor to be considered when the IF frequency is lowered is the effect on mixer noise figure. Below about 100 kc the noise temperature of the mixer diodes varies as $\frac{1}{f}$, which results in an increase in mixer noise at low frequencies. This effect places the practical low frequency limit for the IF pass band at approximately 50 kc.

2.1.4.1.2.1 Video Feedthrough

A more serious problem which is aggravated by the use of a video frequency IF is that of direct video feedthrough. Due to non-linearity in the mixer conversion characteristic, high level signals cause video detection directly in the mixer. When the mixer is followed by a video frequency IF, a video signal produced by the mixer is passed on through the IF and the receiver develops a video output which is independent of local oscillator frequency. This difficulty can be eliminated if the signal level into the mixer is controlled so that video detection does not take place. It originally appeared that gating the traveling wave tube preamplifier coincident with the high level signal pulse would accomplish the desired signal level control and the design of the breadboard feasibility model was based on this approach.

2.1.4.1.2.2 TWT Preamplifier Gating

Subsequent tests revealed a serious problem associated with TWT gating. Since the TWT output VSWR is not 1 but 2 or more, mixer local oscillator leakage at the signal input terminal will travel down the line between TWT output and mixer input, be reflected due to the TWT output VSWR and appear back at the mixer input terminal as a signal at the local oscillator frequency but differing in phase. Under these conditions the mixer becomes a phase detector and will produce a DC output proportional to the phase difference between the local oscillator signal and the reflected local oscillator leakage signal. The DC output is also a function of the amplitude of the reflected signal.

A variation in the complex TWT output impedance would cause a phase variation as well as an amplitude variation in the reflected local oscillator leakage signal and a corresponding change in the DC output of the mixer. Herein lies the problem. Gating of the TWT is accomplished by pulsing the first anode voltage to reduce the beam current of the tube. Unfortunately, this change in beam current produces a change in the complex output impedance, and consequently the video gating pulse is reproduced at the output of the mixer. Figure 4 shows TWT output VSWR as a function of frequency for various anode 1 voltages. Plus fifteen volts corresponds to normal CW operation and at zero volts the tube is cut off. For any particular frequency the length of the transmission line between TWT output and mixer input can be adjusted to minimize the pulse output, but since this method is frequency sensitive it is not a practical solution to the problem.

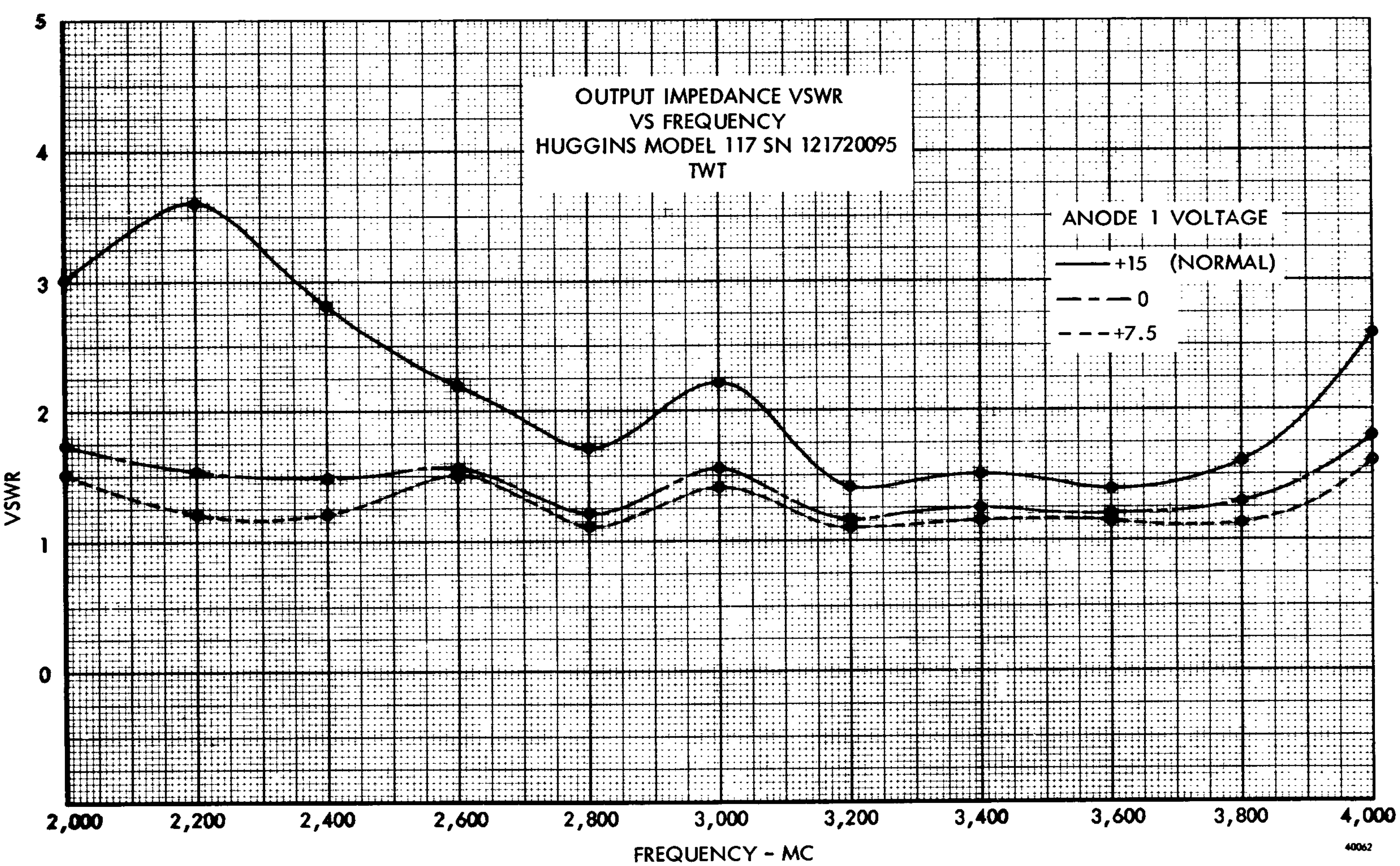

Figure 4. Traveling Wave Tube - VSWR Versus Frequency

Another possible solution would be to use a broadband isolator between the TWT output and mixer input with sufficient attenuation such that the local oscillator leakage signal which reaches the TWT output and is reflected back, is reduced in amplitude to the point where the reproduced gating pulse is below the system threshold.

Measurements were made on the S-band feasibility breadboard to determine the approximate attenuation necessary to reduce the pulse output below threshold. Since calibrated precision attenuators are not available, non-calibrated attenuators were placed in the line between TWT output and mixer input and the line length (using a line stretcher) was adjusted for maximum pulse output. Sufficient attenuation was added until the maximum pulse output was below threshold (tangential signal or less); then the VSWR as seen by the mixer input terminal was measured. The required VSWR was very close to 1.1 over the 2.3-4 GC band. This would correspond to a worst case (output shorted) pure resistive attenuator of approximately 14 db. Using a practical attenuator the required attenuation would be greater.

These measurements indicate that the characteristics of presently available broadband isolators are marginal but this approach might provide a satisfactory solution. For example, an S-band isolator such as the Sperry D44S5 provides a minimum isolation of 10 db over the 2-4 GC range with a maximum insertion loss of 1 db. Further investigation and study would be necessary in this area, in order to provide an optimum design. Better mixer balance along with improved isolator characteristics would probably solve the problem.

Another possible solution to the gating problem would be the use of two traveling wave tube preamplifiers in cascade where the gate is applied to the first tube. Although a solution to the problem, the cost of this method would appear to be prohibitive if gating were required in all bands.

2.1.4.1.3 Conventional IF Frequency

If the frequency of the IF is raised to relieve the problem of direct video feedthrough, either from a high level signal which causes video detection in the mixer or from the TWT gating process, the problem of determining an optimum IF frequency arises. From the standpoint of reducing the confusion factor of the image, it would be desirable to keep the frequency of the IF as low as possible. It would then be desirable to have some method of estimating the level of video feedthrough as a function of pulse level, rise time, IF bandwidth, and center frequency. A suitable method involves calculation of the peak amplitude of the sum of the harmonic components which fall within the IF pass band. This may be accomplished by use of the Fourier analysis for a symmetrical trapezoidal pulse train which approximates the shape of a received radar pulse train. Several simplifying assumptions are made which will tend to give pessimistic results and thus provide a built in safety factor.

1) The receiver IF pass band is assumed to be ideal.

2) The center frequency of the IF pass band is assumed to always fall at a frequency where the $\frac{\sin \theta}{\theta} \frac{\sin \phi}{\phi}$ distribution of the harmonics of the pulse train is at a peak or where $\sin \theta \sin \phi$ approaches 1.

3) The harmonic components which fall within the IF pass band are assumed to be equal in amplitude and phase to the component which falls in the center of the pass band.

Using the foregoing simplifying assumptions an equation for the expected peak level within a given IF pass band may be derived.

The Fourier analysis for a trapezoidal wave train gives the peak value for a given harmonic, n of the pulse repetition frequency:

$$E_n = 2A \frac{(t_0 + t_1)}{T} \left(\frac{\sin \pi n \frac{t_1}{T}}{\pi n \frac{t_1}{T}} \right) \left(\frac{\sin \pi n \frac{(t_1 + t_0)}{T}}{\pi n \frac{(t_1 + t_0)}{T}} \right) \qquad (1)$$

where

A = the peak amplitude of the pulse

t_0 = pulse width

t_1 = pulse rise (and fall) time

T = pulse period or the reciprocal of the pulse repetition frequency. For the center of the IF pass band, f_0, the harmonic, $n = \frac{f_0}{PRF} = f_0 T$

substituting in equation (1).

$$E_{f_0} = \frac{2A}{T} \frac{\left[\sin \pi f_0 t_1\right]}{\pi f_0 t_1} \frac{\left[\sin \pi f_0 (t_1 + t_0)\right]}{\pi f_0} \tag{2}$$

For a given IF bandwidth, the total number of harmonic components

$$= \frac{BW_{IF}}{PRF} = BW_{IF}\ T$$

therefore the peak value of interference for a given IF bandwidth, where all the harmonic components are assumed to have the same peak value and phase as the center frequency component, is equal to $E_{f_0}\ (BW)_{IF}\ T$ or

$$E_{peak} = 2A\ BW_{IF} \left(\frac{\sin \pi f_0 t_1}{\pi f_0 t_1}\right) \left(\frac{\sin \pi\ f_0 (t_1 + t_0)}{\pi\ f_0}\right) \tag{3}$$

A worse case approximation may be made by assuming that $\sin \pi f_0 t_1$ and $\sin \pi f_0 (t_1 + t_0) = 1$ then equation (3) reduces to

$$E_{peak} \approx \frac{2A\ BW_{IF}}{\pi^2 f_0^2 t_1} \tag{4}$$

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(1) Reference Data for Radio Engineers, Fourth Edition, International Telephone and Telegraph Corporation

Equation (4) is discontinuous at $t_1 = 0$, however, if equation (3) is examined at $t_1 = 0$ it reduces to:

$$E_{peak} = 2A\ BW_{IF} \quad \frac{\sin \pi f_0 t_0}{\pi f_0} \tag{5}$$

which is the equation for a zero rise time pulse train. An examination of equation (4) shows that the interference level is independent of pulse width as would be expected if the pulse rise time is a constant. The approximations used in the foregoing derivations are suggested in a paper by Lawrence W. Beard, "Laboratory Substantiation of Interference Prediction Techniques", Sprague Technical Paper No. 62-14.

In order to choose an optimum IF center frequency in a practical situation equation (4) may be solved for f_0:

$$f_0 \approx \frac{1}{\pi} \sqrt{\frac{2A\ BW_{IF}}{E_p\ t_1}} \tag{6}$$

Unfortunately there are two unknowns in equation (6) which must be determined: A, the peak value of the video pulse at the IF input, and E_p, the peak value of pulse feedthrough which can be tolerated.

E_p may be calculated from knowledge of the desired IF sensitivity by assumming a signal to noise ratio which can be tolerated.

The value of A is best determined by direct measurement in a practical situation. Where the IF is driven by the output of a mixer, a worst case measurement can be made if a video output coupling transformer is used on the mixer output. Measurements on the S-band feasibility breadboard were made in this manner. Maximum video output for a high level input signal (determined by the TWT saturation characteristic) was approximately 50 millivolts. Maximum video output due to TWT gating was approximately 10 millivolts.

In choosing an IF frequency for the monitor system, the maximum video level due to the gating pulse is the determining factor since the gating pulse may be used to prevent video output due to high level input signals. Also the gating pulse rise time may be controlled to reduce the level of video feedthrough. For example, using a gating pulse rise time of 0.1 microseconds, a value of A = 10 millivolts peak, E_p = 30 microvolts peak, and BW_{IF} = 5 MC

$$f_0 = \frac{1}{\pi} \sqrt{\frac{2(10)^{-2}\ 5(10)^6}{3(10)^{-5}\ 10^{-7}}} = \frac{1}{3.14} \sqrt{3.3(10)^{16}}$$

$$f_0 = \frac{(1.82)\ (10)^8}{3.14} = 58.2 \text{ MC}$$

Sufficient measurements have not been made as yet to determine the validity of equation (6) under practical conditions. However, measurements made on the breadboard feasibility model indicate that the results of equation (6) provide a good first approximation for the optimum choice of IF center frequency based on the criterion of minimum video feedthrough.

2.1.5 Local Oscillator

Local oscillator requirements for octave bandwidth tuning, tuning linearity and repeatability are best met over the greatest portion of the 0.2 to 40 GC frequency range by the backward-wave oscillator. Suitable backward-wave oscillators or BWO's are available to cover the entire frequency range with the exception of the two lowest bands 0.2 to 0.5 and 0.5 to 1 GC. Voltage tunable magnetrons will be used to cover these bands.

The average BWO will have an output power variation as great as 10 db over its octave tuning range. This power variation will cause variations in receiver sensitivity and even possible mixer crystal damage unless a power leveling system is employed. The most desirable leveling system would employ some means of controlling the BWO power external to the BWO, such as a voltage controlled attenuator between the BWO output and mixer input. However, at present, voltage controllable attenuators are not available except for narrow frequency ranges primarily in X-band.

It appears that the most practical system, assuming that a component development program is not feasible, is BWO power control by varying grid voltage. This system may generate additional random noise and reduce over-all receiver sensitivity[1]; however, actual tests will have to be made to determine the extent of this noise.

Power level sensing may be accomplished by sampling a portion of the RF output of the BWO through a directional coupler and comparing the detected sample with a DC reference voltage. However, this system will not compensate for variations in VSWR of the mixer input over the band or variations in coupling to the mixer crystals. A more satisfactory system monitors the BWO output by comparing the mixer crystal current with a reference. The difference or error signal is then amplified and used to control the BWO grid. This system is proposed for the monitor receiver.

2.1.5.1 Frequency Markers

Referring to the RF subsystem block diagram shown in Figure 2, frequency markers are obtained by sampling a portion of the BWO output and coupling it to several tuned cavities which may be tuned to selected portions of the tuning range. The output of each cavity is detected by a conventional diode video detector so that as the BWO sweeps through the resonant frequency of each cavity, a video pulse is produced. This pulse is then used to trigger a one-shot multivibrator whose output provides the frequency marker pulse for recording.

2.2 Data and Display Subsystem

2.2.1 Technical Approach

The effect of the display and recording mechanism on the requirements of the data processing equipment preceding it is of sufficient importance to require

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1. Final Report - Development of AN/GRR-9(XW-1) RADC-TDR-62-149 1 March 1962, American Electronic Laboratories, Inc., Contract AF 30(602)-1894

a selection of the display system first. The basic requirements of the display system are as follows:

a) Display and permanently record the amplitude of any interfering signal which exceeds the specification limit in the frequency band, 200 MC to 40 kmc.

b) Display and permanently record the frequency of the interfering signal.

c) Display and permanently record the identity of the interfering radar.

Requirement a) dictates the need for nine display channels to accommodate the nine octave bands within the over-all band, 200 MC to 40 kmc.

Requirement b) appears to dictate the need for nine additional channels to accommodate frequency identification for each octave band. However, this depends to a large extent upon the presentation method employed and the tie-in of the display with the RF portion of the monitor. For example, if the display is graphical as in the case of a chart recorder or CRT presentation, and real time is plotted along the abscissa and if real time is linearly related to the sweep frequency in the RF section of the monitor then the abscissa is really a plot of frequency and no additional channels are required. The linear relationship between time and frequency, however, implies that the time sweep mechanism on the display is synchronized to the frequency sweep mechanism in the RF section. Hence, additional hardware is required to accomplish the synchronization. An alternate solution is to generate frequency markers at appropriate intervals within the RF section and to present the markers on a separate channel underneath the interference amplitude information. This has the effect of calibrating the real time axis in terms of frequency. The penalty for taking this approach appears to be the need for nine additional channels on the display, one for each octave band. However, if only one display is used for all of the octave bands (as opposed to a separate display for each band) then the frequency markers associated with each band can be super imposed on one channel. Assigning a unique amplitude to frequency markers occurring within the same octave band provides the identification necessary to associate a particular frequency marker with the correct octave band. If a separate display is used for each octave band, then a separate channel is required for each set of frequency markers. An exception to this statement occurs if the frequency markers are super imposed on the amplitude channel without degrading the amplitude information. Intensity modulating the Z axis of a CRT display with frequency markers is an illustration of this exception. In essence then, anywhere from zero to nine additional channels are required to accommodate frequency calibration dependent upon the approach taken.

Requirement c) dictates the need for a minimum of one channel to identify the interfering radar. Since twenty radars may be causing interference, if amplitude is used as a means of discerning one radar from another on a one channel presentation, then twenty unique amplitude levels are required. Accurately resolving this many discrete amplitudes on one channel is impractical. A minimum of five channels for radar identification is more realistic, resulting in four unique amplitudes per channel. If separate displays are used for each octave band, then nine times five or forty-five channels are required for radar identification. In summary, the number of channels required for a one display system covering all bands is between fourteen and twenty three. The number of channels required for a separate display system for each octave band is between fifty four and sixty three. Aside from the fact that it is extremely desirable to have all the information available on one presentation for cross-reference checks between bands, the number of channels required for separate displays makes this approach prohibitive.

The three general classifications of display devices investigated having both "quick-look" and permanent storage capability are:

(1) Storage tubes

(2) Print-out devices

(3) Chart recorders

During the early stages of the study program when a fast sweep of the frequency spectrum was considered, the storage tube technique held considerable promise because of its inherently fast frequency response. The decision to sweep slowly, however, renders this advantage inconsequential. The disadvantages of a storage tube display are many:

(1) Assuming a one display system for all bands, at least fourteen channels are required to present all the information. This means that elaborate beam switching techniques must be employed to secure all fourteen traces during one horizontal sweep.

(2) The largest standard storage tube size is ten inches. Hence, the frequency spectrum of each octave band must be compressed to less than a ten inch length with attendant loss of frequency resolution.

(3) Permanent retention of data requires elaborate photographic or other storage techniques.

Print-out devices such as the Clary Printer offer considerable promise for future monitor systems. The amplitude, frequency, and radar identification information is converted to binary form and stored in registers. The printer acquires the binary information from the storage register and activates typewriter keys which print out the decimal equivalent information on paper. A set of information consisting of one interference amplitude and its associated frequency and radar identification is printed out in three columns on one line. Succeeding sets of information are printed out underneath the original set resulting in a neat columnized log of interference data. Another column indicating the time of occurrence of each information set may also be added, for further interference definition. The most noteworthy advantage of this device is the efficiency with which it records. Unlike CRT and chart recording devices, the printer only logs information that is essential. It remains inoperative during that portion of the frequency spectrum where interference threshold levels are not exceeded. Also, since supporting circuitry is needed to accomplish binary conversion of data, the system is compatible with digital computer reduction and storage techniques for long term trend analysis and statistical studies. The major disadvantage of this device is the requirement for control and binary conversion support circuitry.

A third mechanism for displaying and recording data is a multi-channel chart recorder. Attributes of this device are its simplicity of operation, its multi-channel capability and its minimal requirement for support circuitry. Of particular interest is the Honeywell 1508 Visicorder Oscillograph. This recorder can handle twenty four information channels, exceeding the maximum requirement for twenty three discussed previously. It has a bandwidth of 5 kc which minimizes the pulse stretching requirement of the IF detector (discussed subsequently). Writing is accomplished by impregnating an emulsified paper with a narrow beam of ultra-violet light, the displacement of which is controlled by the amplitude of the information signal. Since mechanical styli are not employed in the recording, overlapping of channel information (and hence increased amplitude resolution) can occur without penalty. Specifications for the 1508 are as follows:

Channels:	12 or 24
Galvanometers:	Type M, sub-miniature
Record width:	8"
Record length:	100' standard
Record speeds:	12, pushbutton selected as follows: 0.1, 0.2, 0.4, 0.8, 1.0, 2.0, 4.0. 8.0, 10, 20, 40, 80"/second
Frequencies:	DC to 5,000 cps
Writing speed:	Greater than 50,000"/sec
Time lines:	4-interval system with .01, 0.1, 1.0, and 10-second intervals
Grid lines:	0.1" with 5th line heavy, or 2 mm with 1 cm heavy
Optical arm:	11.8" (30 cm) standard
Trace identifier:	45° slope every 8", spaced .032" max., .02" min.,
Power:	117V, 60 cycle, 5-6 amp
Dimensions:	19" wide x 7" high x 17-1/2" deep. Weight approximately 50 lbs.

The disadvantage of a CRT presentation as enumerated previously are serious enough to disqualify it as a first choice, barring the necessity for a fast sweep system. The selection, then, is between a digital printer and a chart recorder. As a growth item the digital printer takes precedence over a chart recorder because of its compatibility with computer technology. However, the chart recorder satisfies the minimum requirements of the monitor as presently stated and requires the least development of support circuitry. It is therefore recommended as the display and recording medium for the first generation RFI monitor. The specific recorder proposed is the Honeywell 1508 Visicorder.

2.2.2 Subsystem Description

2.2.2.1 Chart Recorder Channel Allocation

The Honeywell 1508 Visicorder is capable of twenty four channel operation. The proposed channel allocation appears in Figure 5. Channels 1 through 3, 5 through 7, and 9 through 11, are used to record the amplitude of interference signals emanating from the nine sub-band monitors. Channel 4 is allocated to frequency markers from sub-band monitors 1 through 3. Frequency markers associated with a given sub-band are distinguished from frequency markers of the remaining sub-bands by pulse amplitude. Similarly, channels 8 and 12 are allocated to frequency markers from sub-band monitors 4 through 6 and 7 through 9 respectively. Channels 13 through 17 are allocated to radar identification. Channel 13 accommodates the identification of radars 1 through 4. A given radar is distinguished from the remaining radars by pulse amplitude. Similar statements can be made regarding channels 14 through 17. Channels 18 through 22 are used as spares to accommodate an increase in the number of radars and/or an increase in the number of sub-bands.

2.2.2.2 Radar Identification

In order to identify the source of an interference signal, the monitor must know precisely when each radar is transmitting. By comparing the time coincidence of an interference pulse at the monitor output to the transmitted pulse of each radar, identification can be made. The transmission interval of each radar can be acquired at the monitor in a number of ways. One way is to establish an RF link between each radar and the monitor. This approach requires a transmitter and receiver for each radar. An alternate approach is to utilize a coaxial cable connection between each radar and the monitor along which a video pulse is transmitted several micro-seconds ahead of the main band RF pulse. Pre-triggering of the video pulse is necessary to offset the longer propagation time of video along coax cable relative to RF through free space and hence to maintain time coincidence of the two pulses at the monitor input. The second method is chosen over the first because it involves less complexity, less development, lower cost to obtain the same result, and requires no wide frequency channel allocations.

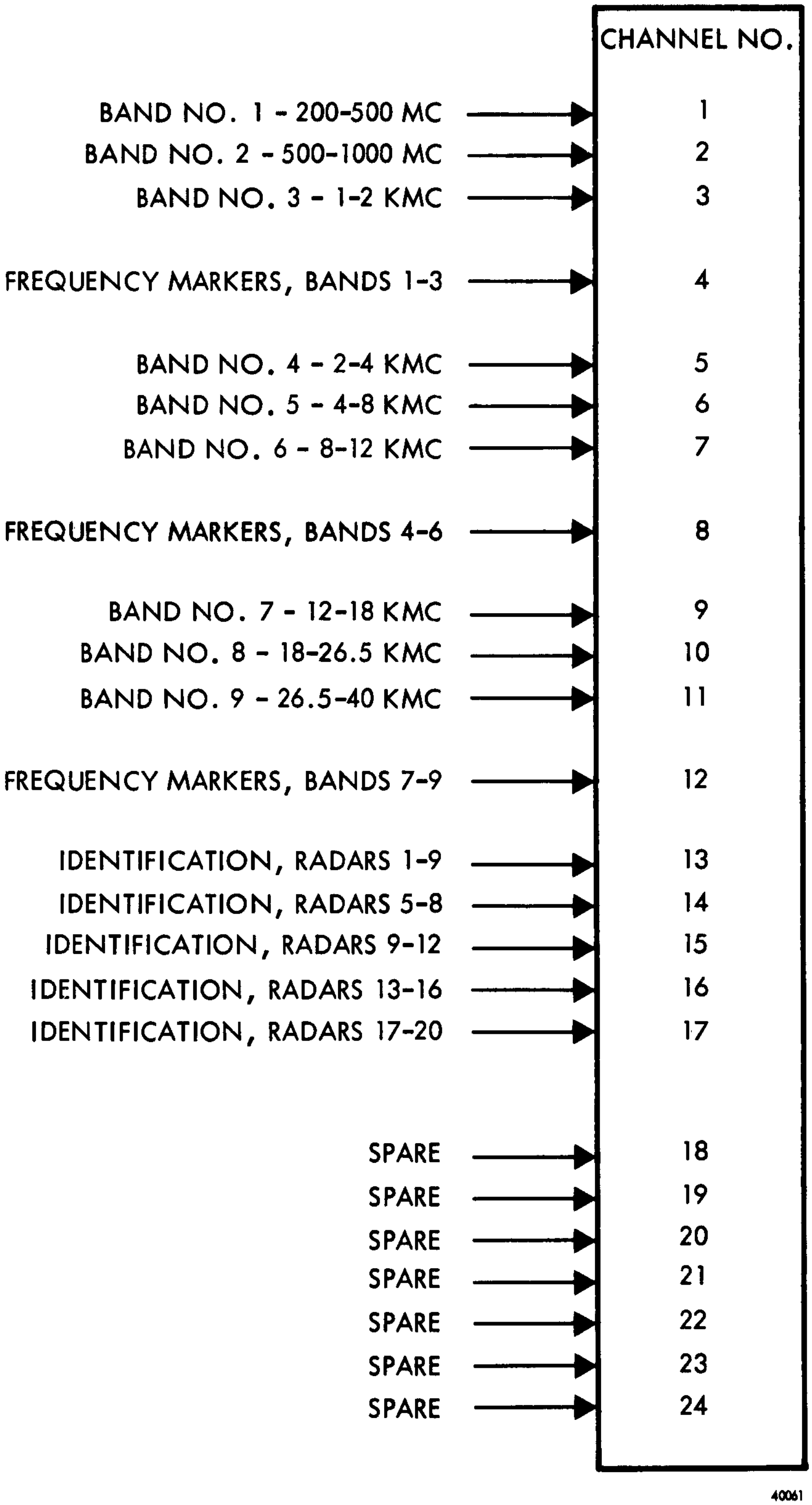


Figure 5. Chart Recorder Channel Allocation

2.2.2.2.1 Radar-Monitor Interface

A block diagram of the Radar-Monitor interface appears in Figure 6. For the sake of simplicity only four radar sets are shown in the diagram, it being understood that four similar groupings of four radars per group are likewise present yielding a total of twenty radar sets. The four radars shown correspond to the group allocated channel 13 on the chart recorder (reference Figure 1). It is anticipated at this writing that pre-trigger video pulses are available from each radar set. Each pre-trigger pulse enters a pulse shaping network which standardizes the pulse amplitude, pulse width, and output impedance prior to transmission along coax cable to the monitor. The time delay, Δt, shown on the diagram, is a result of the pre-triggering. A time delay network at the monitor delays the leading edge of the incoming video pulse until it corresponds in time to the leading edge of the RF pulse. A width control network then reconstructs the video pulse until is slightly overlaps the RF pulse. The video pulses thus derived are labeled R1 to R4, indicating their relationship to Radar 1 through Radar 4 respectively. Similarly, video pulses derived from Radar 5 through Radar 20 (not shown) are labeled R5 to R20.

2.2.2.2.2 Identification Logic

Before discussing the identification logic, it is well to define and label the pulses emanating from the output of the monitor. A block diagram showing the outputs of the first sub-band monitor appears in Figure 7. The interference pulse is labeled A1 indicating the amplitude of the interference signal. The frequency marker pulse is labeled F1 indicating a frequency calibration point. Derived from the A1 pulse through a one-shot multivibrator is a B1 pulse which indicates the time of occurrence of an interference signal and the band in which it occurs. Hence, if interference in Band 1 is caused by Radar 1 then time coincidence exists between the R1 pulse and the B1 pulse. For each of the sub-band monitors, 2 through 9, there is a set of amplitude pulses, A2 through A9; frequency pulses, F2 through F9, and band pulses, B2 through B9.

A radar identification logic diagram appears in Figure 8. All of the B pulses, 1 through 9, are "ORed" at OR Gate 1. The output thus derived is "ANDed" with each of the R pulses, 1 through 20. Since the presence of a B pulse must be accompanied by a time coincident R pulse, one of the AND gates yields an output which triggers a one-shot multivibrator having a unique identifiable output amplitude. Each set of four one-shots are "ORed" to one recorder channel. Hence,

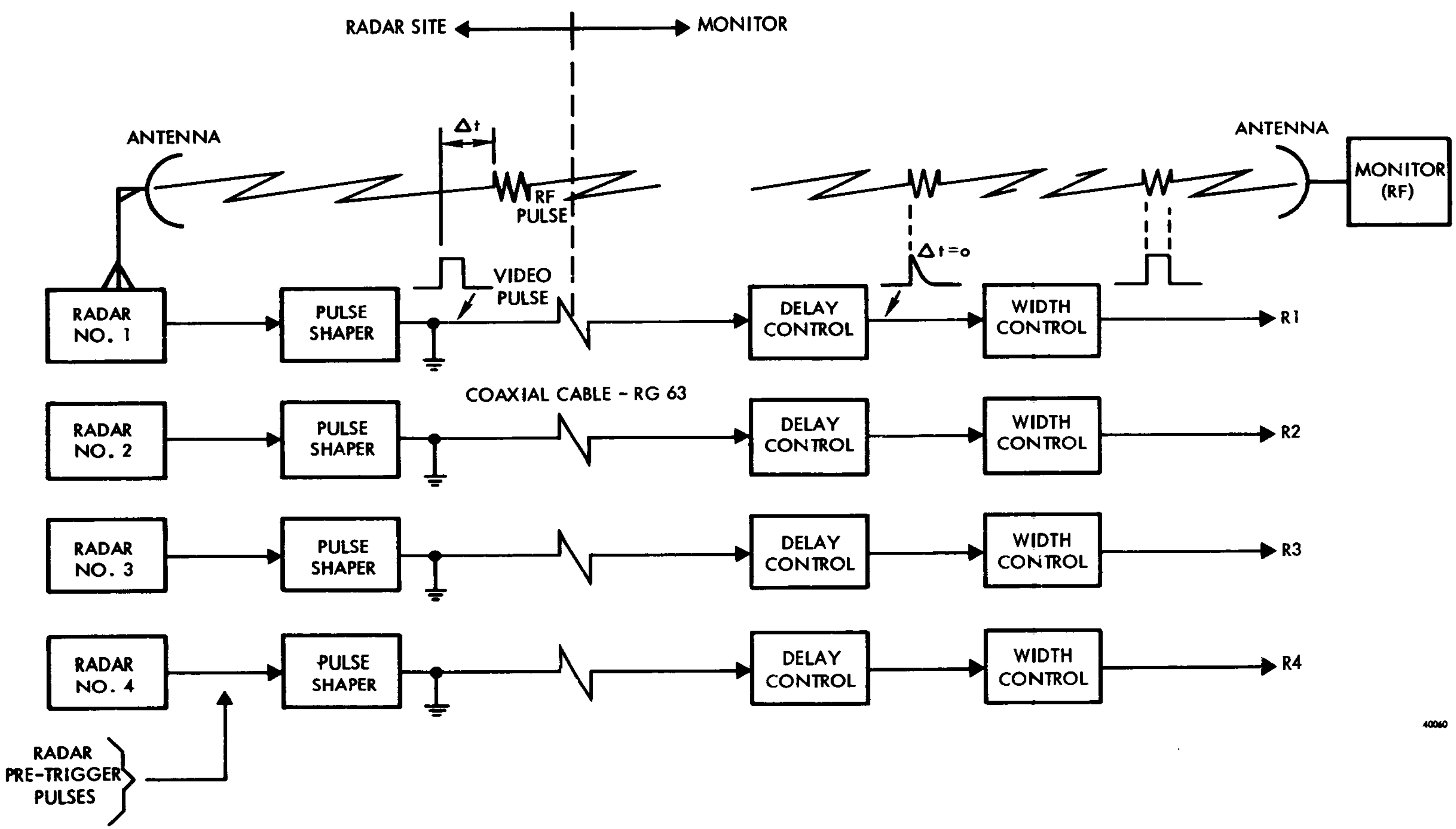


Figure 6. Radar - Monitor Interface Diagram

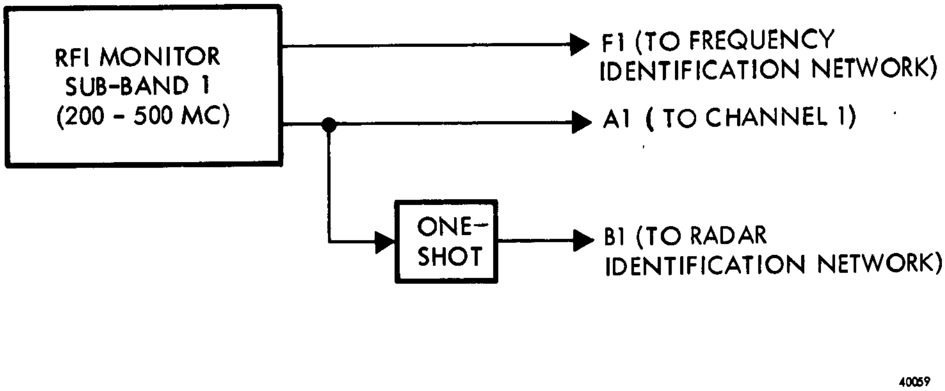


Figure 7. Typical Monitor Band Outputs

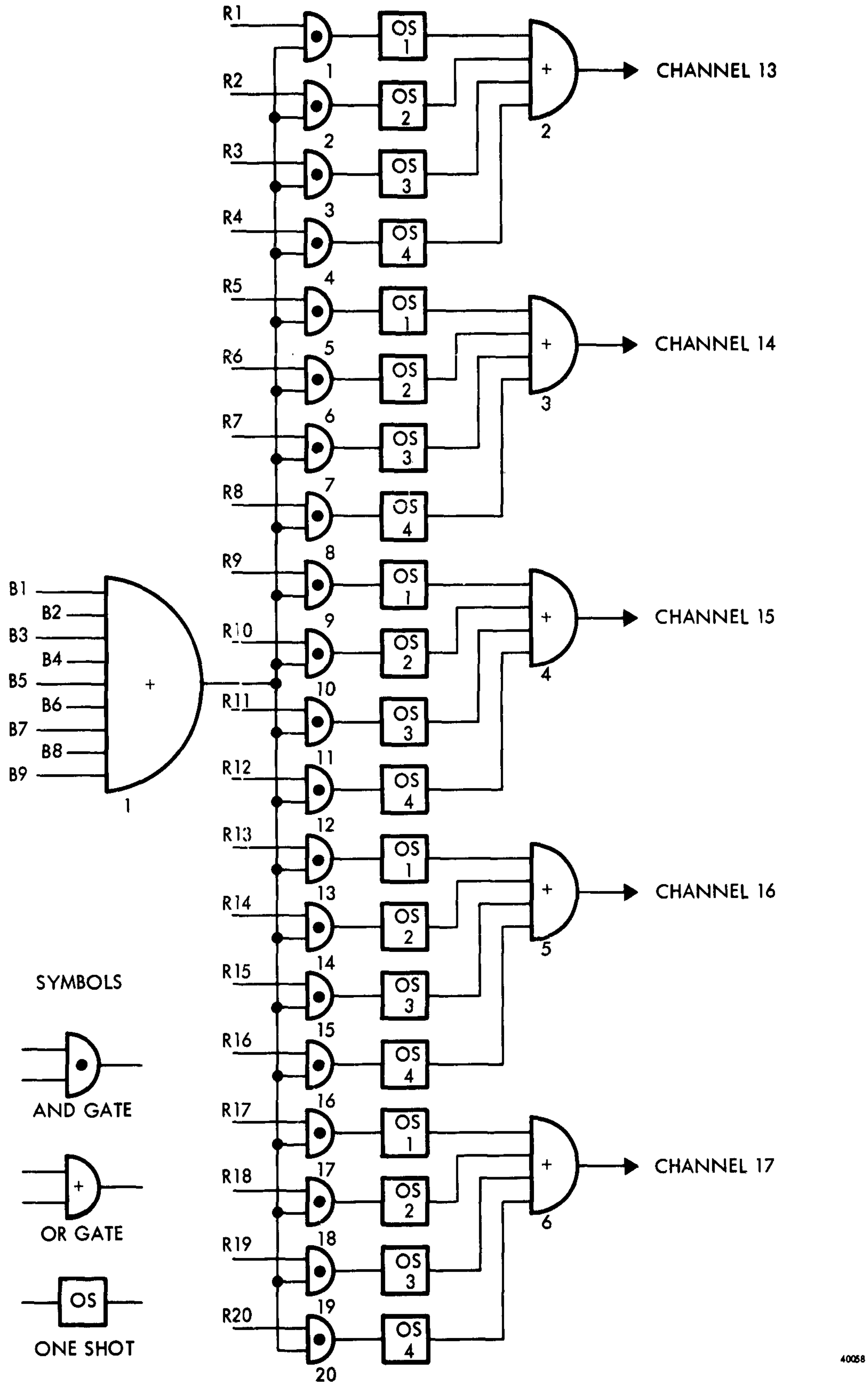


Figure 8. Radar Identification Logic Diagram

four unique amplitudes are required for identification. This is indicated on the diagram by numbering each set of one-shots 1 to 4. To illustrate the logic, assume that interference is generated in Band 9 by Radar 5. A B9 pulse is thus generated which appears at the input and hence the output of OR Gate 1. Since this pulse is time coincident with the R5 pulse associated with Radar 5, a pulse is generated at the output of AND Gate 5. A one-shot is thus triggered having a normalized amplitude of 1. After processing through OR Gate 3, this amplitude appears on Channel 14 identifying radar 5 as the cause of the interference.

2.2.3 Frequency Calibration and Identification

As discussed in reference to Figure 7, each sub-band generates frequency markers as a means of calibrating the real time axis of the display in terms of frequency. The frequency markers are labeled F1 to F9 indicating their relationship to Bands 1 to 9 respectively. In order to minimize the number of channels required for frequency calibration, consistent with the number of channels available, a set of three frequency markers share one channel as shown in the logic diagram of Figure 9. Each F pulse triggers a one-shot multivibrator having a unique identifiable output amplitude. Each set of three one-shots are "ORed" to one recorder channel. Hence, three unique amplitudes are required for identification. This is indicated on the diag.am by numbering each set of one-shots 1 to 3.

2.2.4 Interference Data-Chart Recorder Interface

As discussed in connection with Figure 7, each sub-band monitor carries amplitude information on interference signals occurring within its band. The amplitude data is labeled A1 through A9 indicating its relationship to bands 1 through 9 respectively. On the recorder, Channels 1, 2, and 3 are allocated to A1, A2, and A3; Channels 5, 6, and 7 are allocated to A4, A5, and A6; Channels 9, 10, and 11 are allocated to A7, A8, and A9 (reference Figure 5). There are three main interface problems to be solved. One involves the selection of optimum recorder speed. The second involves the amount of amplitude pulse stretching required to produce full deflection on the recorder, consistent with the recorder bandwidth. The third involves the amount of buffering required between the detected interference signal and the low input impedance of the recorder.

The selection of recorder speed is a trade-off between the amount of information crowding and frequency resolution that is tolerable and the desire to keep

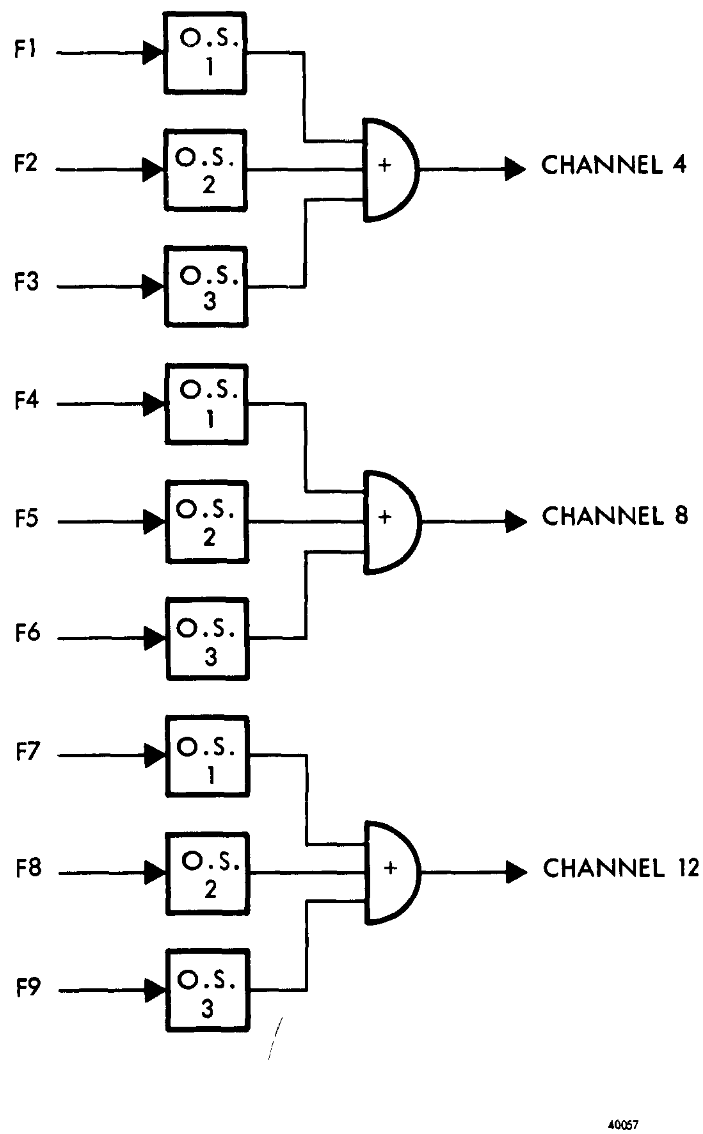


Figure 9. Frequency Marker Identification Diagram

the amount of paper produced during a measurement run to a reasonable length. As presently planned, the time required for one measurement run is approximately 18 minutes. The slowest speed of the 1508 Visicorder is 0.1 inches per second yielding 9 feet of paper during an 18 minute run. The frequency resolution at this speed in the worst case (band 9) is 130 Mc per inch. Assuming the human eye can resolve position accurately to better than a tenth of an inch, frequency can be resolved to within 13 Mc which is 0.1% of the total band (13 KMC). A recorder speed of 0.1 inches per second satisfies the apparent resolution requirements and at the same time yields minimum paper length. Hence, this operating speed is selected for the display.

The amount of amplitude pulse stretching required to produce full deflection on the recorder is a function of the recorder bandwidth. If a step function is applied to a recorder input channel having a bandwidth B, then the time required to produce 99% deflection is given by the relationship:

$$(1) \qquad t = 4T$$

where

$$(2) \qquad T = \frac{1}{2\pi B}$$

Hence,

$$(3) \qquad t = \frac{4}{2\pi B}$$

Since the bandwidth of the 1508 is 5 KC,

$$t = \frac{4}{2\pi(5000)} = 0.127 \text{ milliseconds}$$

Thus, the pulse width of the detected interference signal must be stretched to approximately 0.13 milliseconds in order to insure full deflection on the recorder.

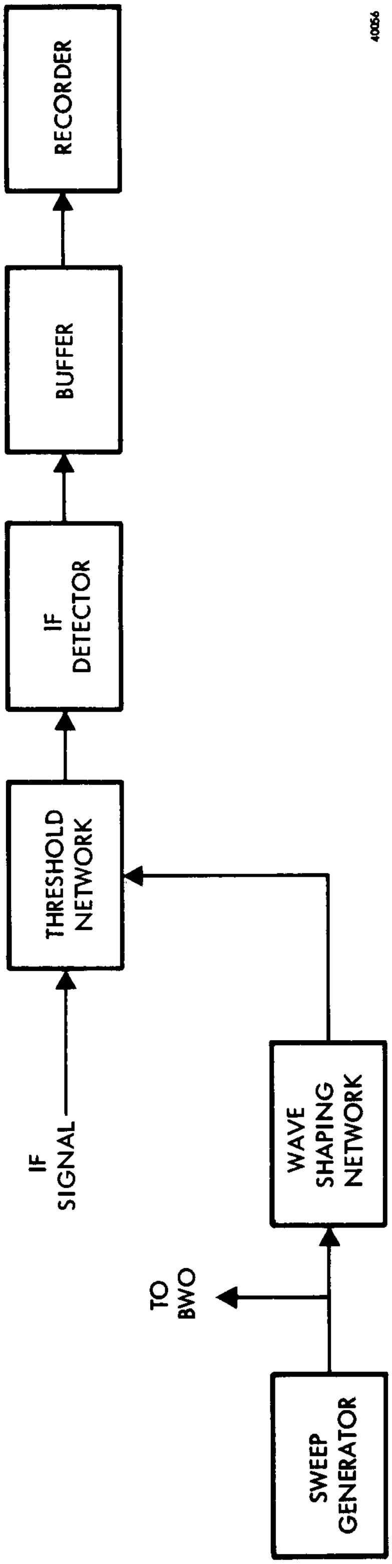
The third interface problem, namely the buffering requirement between the detected interference signal and the low input impedance of the recorder, is a result of the need to control the pulse stretching of the signal. This is discussed in the following section.

2.2.5 Threshold Level Setting and IF Detection

A threshold sensing device is required somewhere between the monitor receiver and the recorder such that only interference signals which exceed their specification limit are recorded and identified. If the receiver presented a constant system gain to all incoming signals regardless of frequency, a fixed threshold level would be sufficient. However, variations in TWT gain with signal frequency dictate the need for a frequency dependent variable threshold level. A block diagram of the back end of the monitor receiver appears in Figure 10. A variable threshold level is derived from the sweep generator through a wave shaping network designed to compensate the threshold level for variations in TWT gain with frequency. That portion of the IF signal which exceeds the threshold level passes through the threshold network to the IF detector. The peak difference between the IF signal and the threshold level is detected and time stretched to allow the recorder time to respond. A circuit diagram of the threshold network and IF detector appears in Figure 11. The IF amplifier feeding the network input terminals is replaced in the diagram by its Thevenins' equivalent circuit. The threshold network consists of diode CR1 and resistors R1 and R2. A dc blocking capacitor C and resistor R3 form the coupling circuit to the peak detector consisting of CR2 and C1. The input resistance of the buffer amplifier appears as a load resistance R_L at the detector output. Circuit operation is as follows. The threshold signal establishes a positive voltage, E_T, at the cathode of diode CR1. Hence, CR1 remains back biased until the IF signal goes more positive than E_T. When this happens, CR1 is forward biased and the IF information passes through coupling capacitor C and diode CR1 to charge capacitor C1 to the peak difference between the threshold level and the IF signal. The peak voltage thus developed across C1 leaks off at a rate determined by the C1-R_L time constant. This time constant is controlled to allow the recorder sufficient time to reach full deflection as discussed previously. The selection of capacitor C1 is critical. It needs to be large enough to maintain a charge while the recorder responds and small enough to charge up within a few cycles of the input IF signal. For a fixed value of C1, the charge time is determined by the source resistance RG and the dynamic resistance of diodes CR1 and CR2. The decay time is determined by R_L. As determined previously, the time required for the 1508 recorder to reach full deflection is 0.13 milliseconds. Hence, the peak voltage detected must not decay appreciably during this time. If a 20% decay is considered tolerable, it can be shown that the C1-R_L time constant required is

$$(4) \qquad T_1 = C_1 R_L = .00058 \text{ sec.}$$

The charge time required depends upon the pulse width envelope of the IF signal and the number of pulses present at the input. The narrowest pulse width expected is


RECORDER
BUFFER
IF DETECTOR
THRESHOLD NETWORK
IF SIGNAL
WAVE SHAPING NETWORK
TO BWO
SWEEP GENERATOR
40056

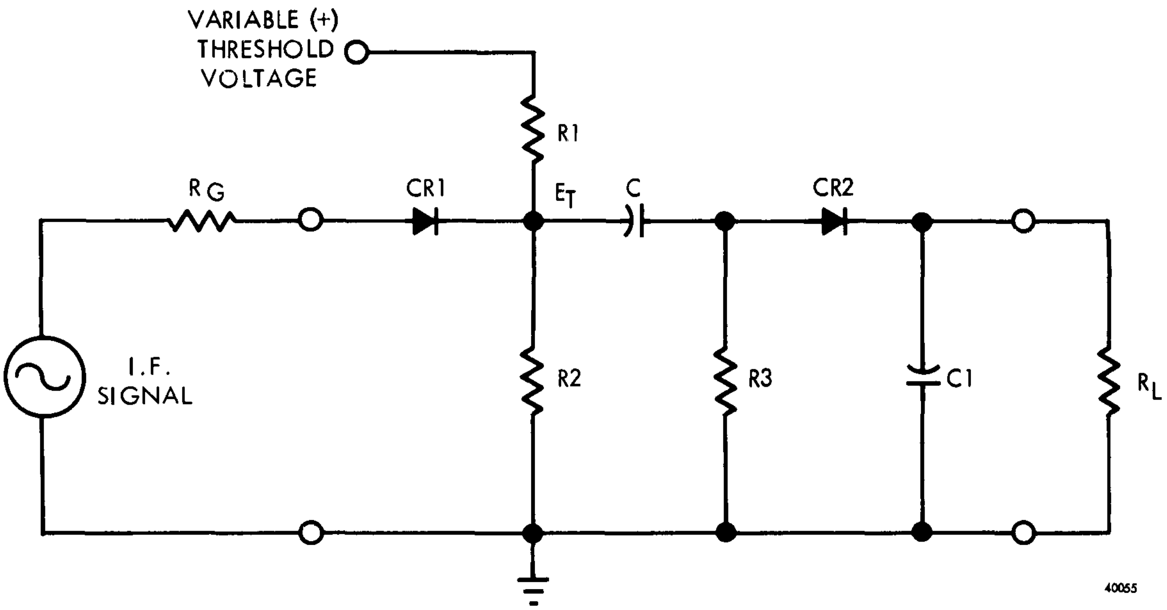


Figure 11. Threshold Network and IF Detector, Circuit Diagram

0.2 microseconds. The number of pulses expected is at least 10. Hence, an effective pulse width of 10 x 0.2 = 2 microseconds appears at the input terminals. The charge time is determined by

$$(5) \qquad T = (R_G + R_D)\, C_1 = 2 \times 10^{-6} \text{ sec}$$

where R_D is the dynamic resistance of diodes CR1 and CR2 (approximately 40 ohms).

From $$(5) \qquad T = (50 + 40)\, C_1 = 2 \times 10^{-6}$$

Hence,

$$(6) \qquad C_1 = .022 \text{ uf}$$

Substituting this value into (4) yields the input resistance required of the buffer amplifier.

$$(7) \qquad R_L = 26 \text{ K}$$

3.0 FEASIBILITY BREADBOARD

A block diagram of the 2-4 Gc feasibility breadboard is shown in Figure 12. This breadboard does not represent a physical prototype of one band of the ultimate monitor system. Its purpose is to demonstrate the feasibility of a chosen design approach and consequently the choice of components to implement the circuitry is based on availability and cost rather than most desirable electrical characteristics.

3.1 Preamplifier

Referring to Figure 12, the TWT preamplifier used is a Huggins Model 117. This amplifier is packaged in two rack mounted 7" panels and includes all required power supplies. The general specifications for this amplifier are: nominal gain of 25 db from 2 to 4 Gc with a maximum noise figure of 15 db. Nominal saturated power output is 0 dbm. Actual measured data on small signal gain, noise figure, saturated power output and power gain is shown in Figures 13 and 14. Choice of this particular amplifier was based mainly on cost and availability. Better noise figure (approximately 5 db) and more uniform gain characteristics are available at greater cost and longer delivery time. TWT gating may be accomplished by applying a negative pulse to the first anode. This input is not standard and circuit modifications are necessary. A pulse level of approximately -15 volts is sufficient to cut the tube off.

3.2 Mixer

The mixer employed is a standard Sage Model 2531 coaxial balanced type. The Model 2531 is a strip line type broadband hybrid mixer with local oscillator leakage down approximately 20 db over the 2-4 Gc range. Type 1N416E crystals are used. Nominal conversion loss at video IF frequencies including output coupling transformer is approximately 10 db. Conversion loss at IF frequencies on the order of 30 Mc will be somewhat less with a well designed output coupling transformer. This particular mixer was chosen because it is a standard off-the-shelf item and is representative of the best commercial mixers of this type.

3.3 Local Oscillator

The local oscillator employed in the feasibility breadboard is a Raytheon QKB914 backward wave oscillator. This particular BWO was chosen primarily for

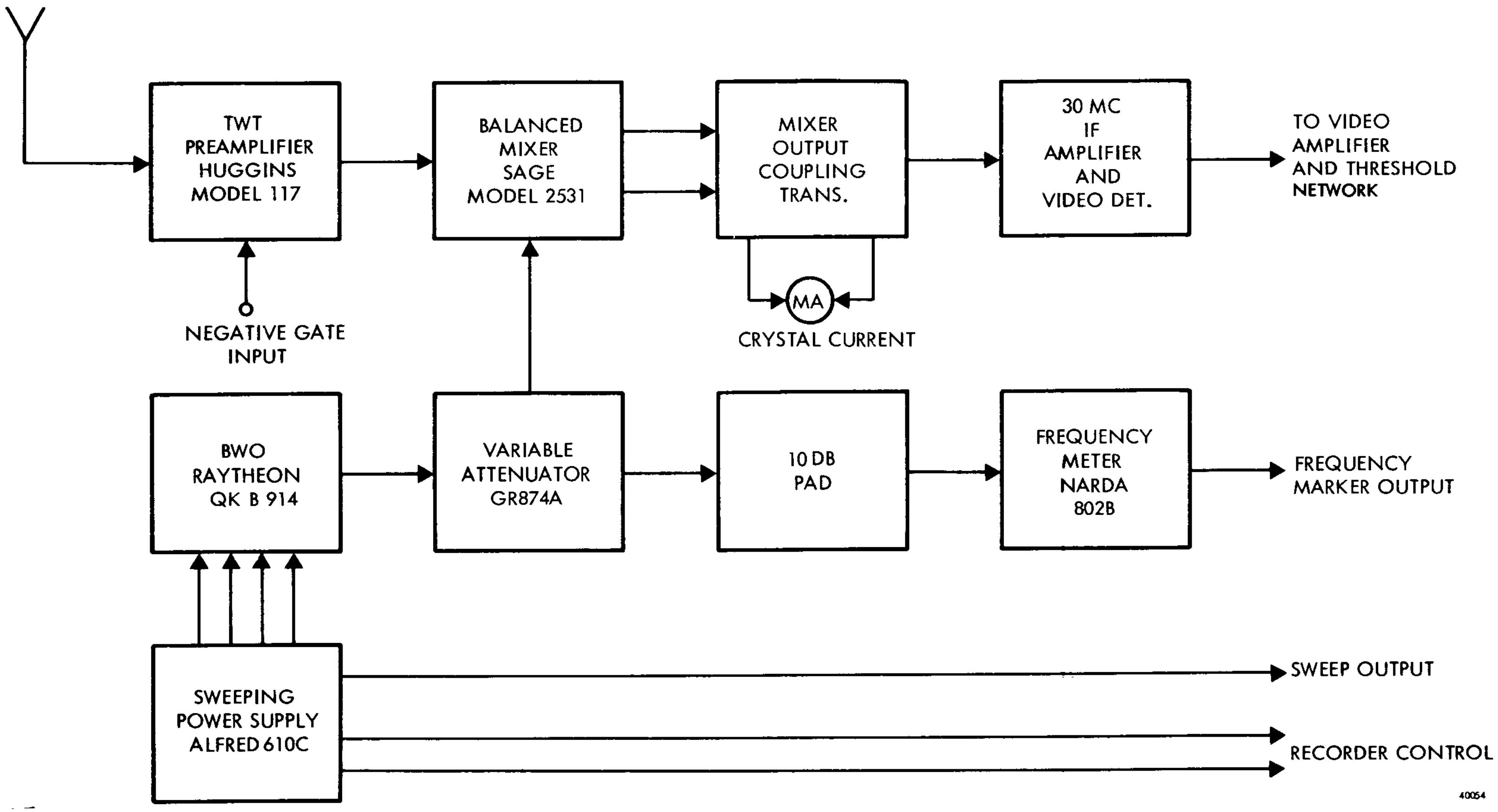


Figure 12. Block Diagram - 2-4 Gc Breadboard Monitor

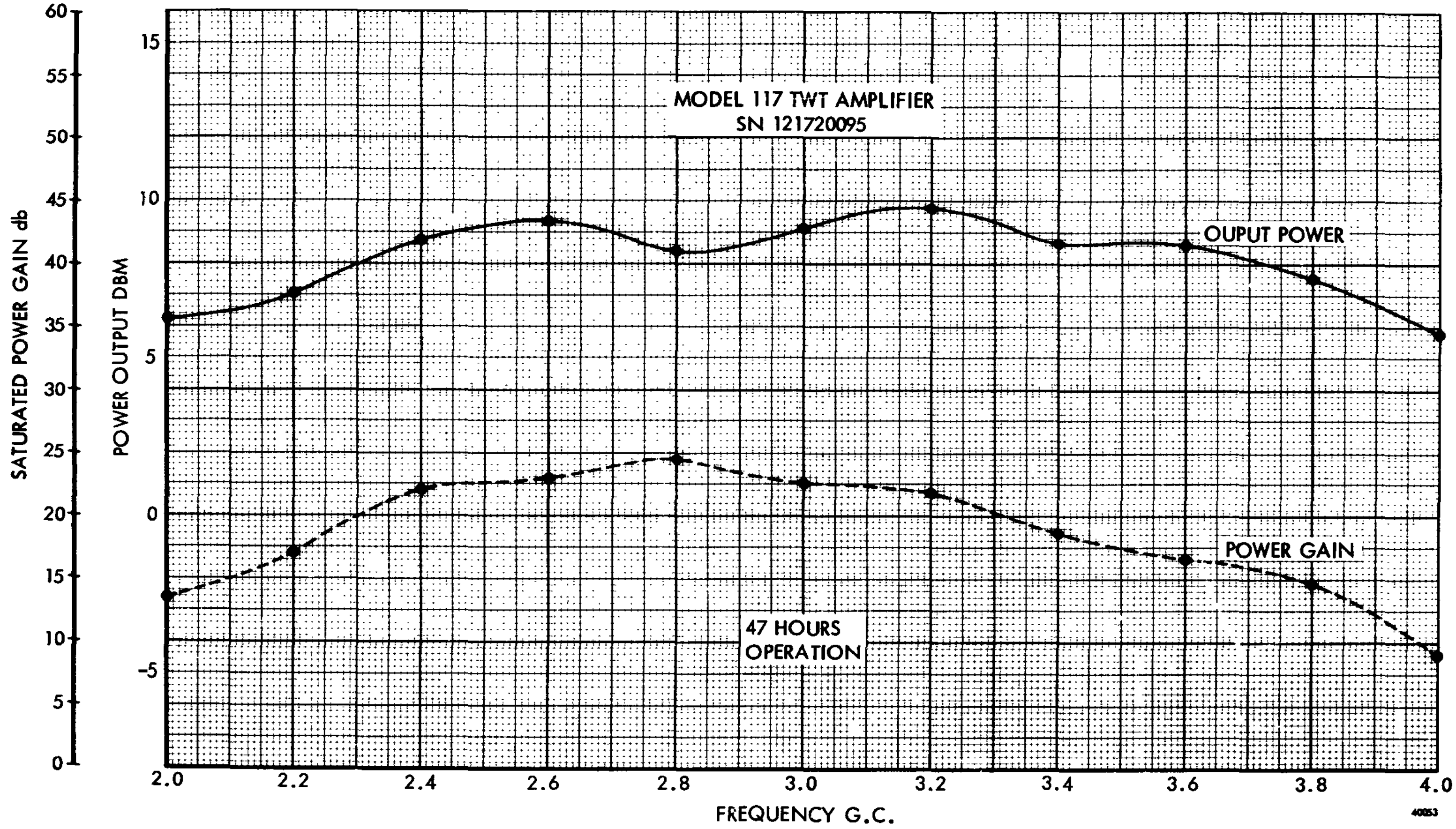


Figure 13. Traveling Wave Tube - Saturated Power Output and Power Gain Versus Frequency

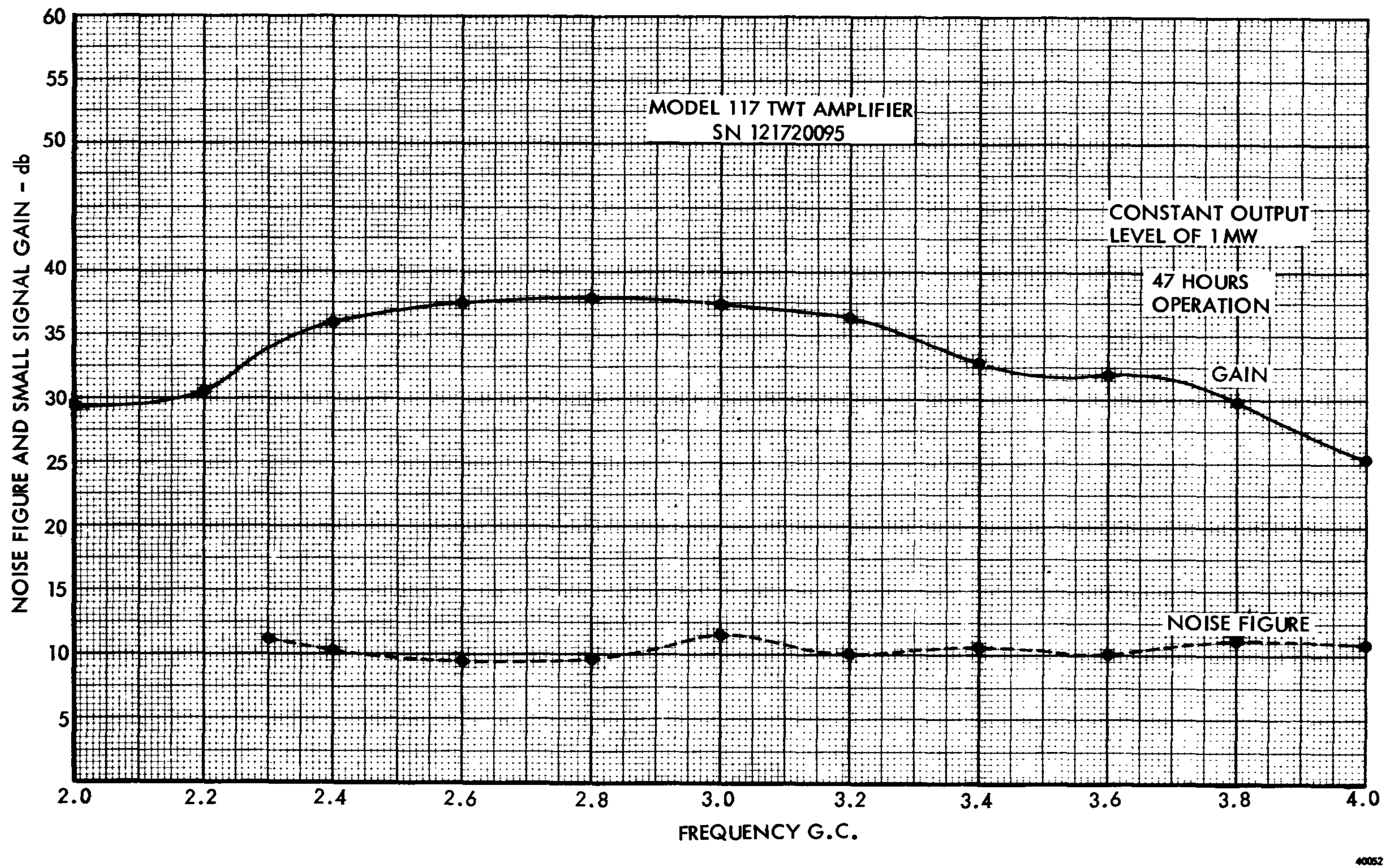


Figure 14 – Traveling Wave Tube – Noise Figure and Small Signal

delivery and price reasons but also because of its permanent magnet focusing. However, subsequent experience gained from testing and operating this type BWO has shown that for the monitor application its disadvantages far outweighed the advantage of permanent magnet focusing. Measured characteristics are shown in Figure 15.

The primary disadvantage of the type QKB914 lies in the fact that the helix or delay line is grounded to the tube housing which is also RF signal ground. This makes it necessary to operate the cathode negative with respect to ground over the full tuning voltage range approximately 195 to 1400 volts. The power output of the tube is controlled by the grid-cathode voltage and consequently when a leveler loop is used to maintain a constant power output, it is necessary to include DC isolation circuitry to couple the DC error signal to the grid. Obtaining leveler loop stability becomes a serious problem with a backward wave oscillator tube since the frequency is also a function of the grid-cathode voltage. For the particular QKB914 used in the breadboard the mid-band sensitivity is approximately 800 Kc per volt of grid cathode voltage change. Increasing grid bias (decreasing cathode current and power output) causes an increase in frequency. A more suitable tube for the monitor application in S-band is the Stewart OD 2-4. All DC voltages in this tube are isolated from both tube housing and r.f. output. It is then possible to ground the grid and apply error voltage between cathode and ground such that an increase in grid bias also represents a decrease in helix-cathode voltage; thus, the frequency shift due to grid bias change is counteracted by a shift due to the helix-cathode voltage change. A leveler is not used in the feasibility breadboard in the interest of simplicity and to avoid the aforementioned problems which would be solved by employing a more costly tube. Grid bias is supplied by a simple bias battery adjustable by varying the setting of a potentiometer voltage divider. Grid bias is adjusted to provide a compromise power output which will not cause excessive crystal current variations as the BWO is tuned over the 2-4 Gc range.

Sensitivity variations caused by variations in crystal current are compensated for in the calibration of the system threshold.

Power for the backward wave oscillator is supplied by an Alfred Model 610C sweeping power supply. This supply is designed to provide all of the necessary operating voltages to power voltage tunable magnetron type tubes. The Model 602 which is designed to power backward wave oscillators was not available. The Model 610C does not provide an exponential sweep voltage which is necessary to produce a linear frequency sweep from the BWO.

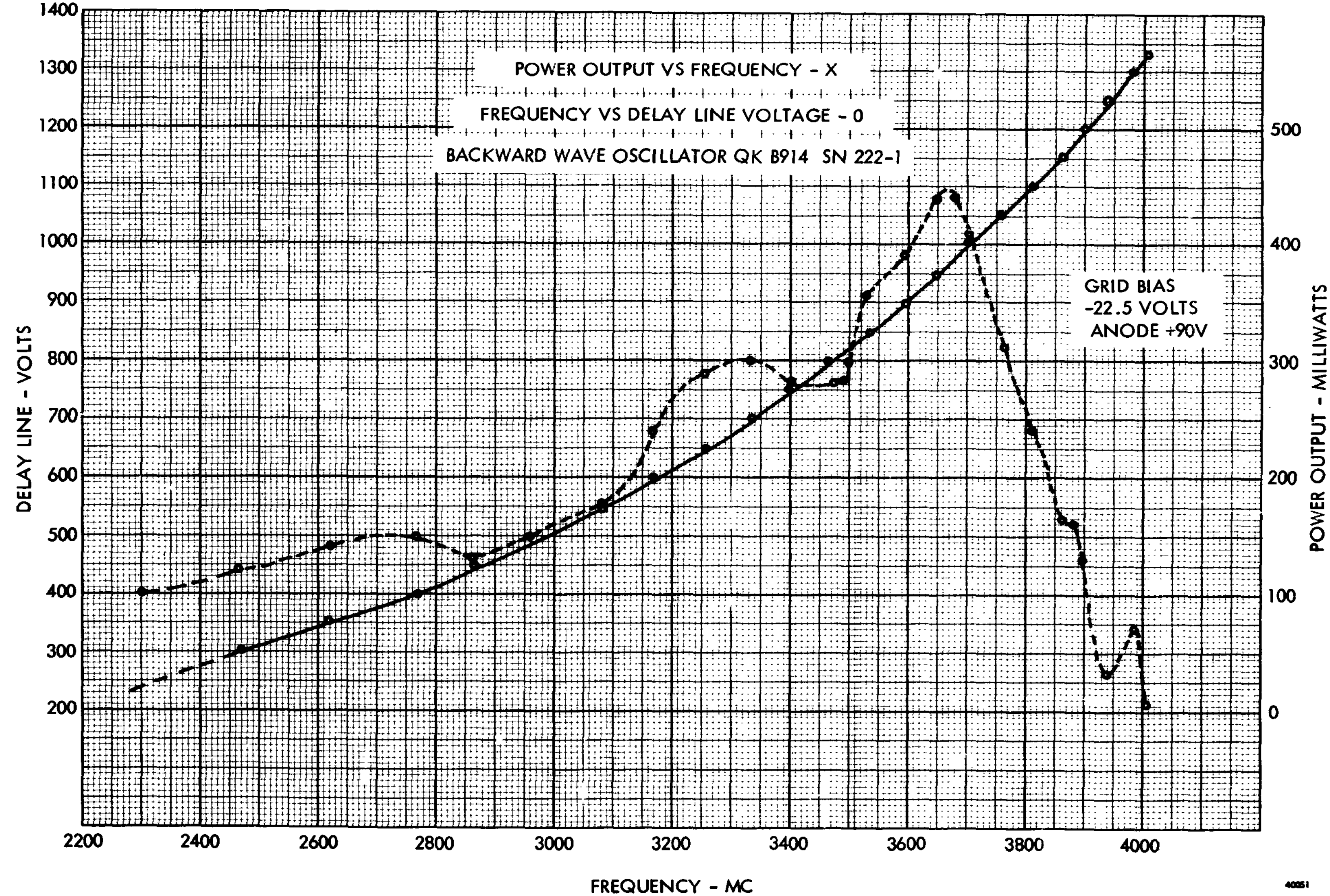


Figure 15. Backward Wave Oscillator Tuning Characteristic and Power Output Versus Frequency

Figure 15 shows a plot of BWO frequency vs. helix or delay line tuning voltage. It may be seen from this plot that the required shape of the tuning voltage curve approximates an exponentially increasing function. For the requirements of the breadboard feasibility model a linear sweep voltage does not present a problem since a portion of the BWO power is coupled to a Narda 802B frequency meter which provides a frequency marker pulse tunable to any frequency within the BWO tuning range. Accurate frequency identification is thus provided.

The capabilities of the Alfred 610C supply will allow the BWO to be swept over a complete tuning range of 2400 to 4000 Mc or any portion of this band at sweep frequencies from 0.01 cps to 10 cps. The low frequency end of the sweep (2400 Mc) is limited by the minimum output voltage of the Supply (approximately 240 volts). The supply is capable of supplying up to 2500 volts at 60 ma, therefore an adjustable over voltage limit which is provided is set for 1500 volts to protect the BWO. In addition, an external fuse has been provided in series with the BWO anode lead. This is necessary since the BWO anode which operates at approximately 90 volts is supplied from the built-in power supply normally used for the reflector of the voltage tunable magnetron. This supply is adjusted to 90 volts by a front panel control which will allow the output to be adjusted over a range from zero to approximatel 900 volts.

A small amount of frequency modulation on the order of ± 50 KC is present on the BWO output due to power supply ripple. This causes an increase in the low frequency noise level at the IF output due to slope detection of the FM on the local oscillator signal. This noise presents no particular problem since it can be removed with a simple high pass filter following the IF video detector.

3.4 IF Amplifier

The preliminary breadboard design employed a video frequency IF amplifier. The balanced mixer output was coupled to the IF input by a balanced video transformer which combined the IF outputs of the two crystals and provided an unbalanced 100 ohm output. The IF amplifier used was a Maxson type LVT-1.

This amplifier is a transistorized unit with a low level signal gain of approximately 86 db and a low level bandwidth of approximately 3 Mc. Input dynamic range is 70 db before hard clipping. Linear dynamic range is on the order of 55 db. Output dynamic range is approximately 25 db, therefore high level input signals are compressed about 45 db. Transfer characteristics of this amplifier are shown in Figures 16 and 17. A high pass filter with a low frequency cutoff at 50 Kc

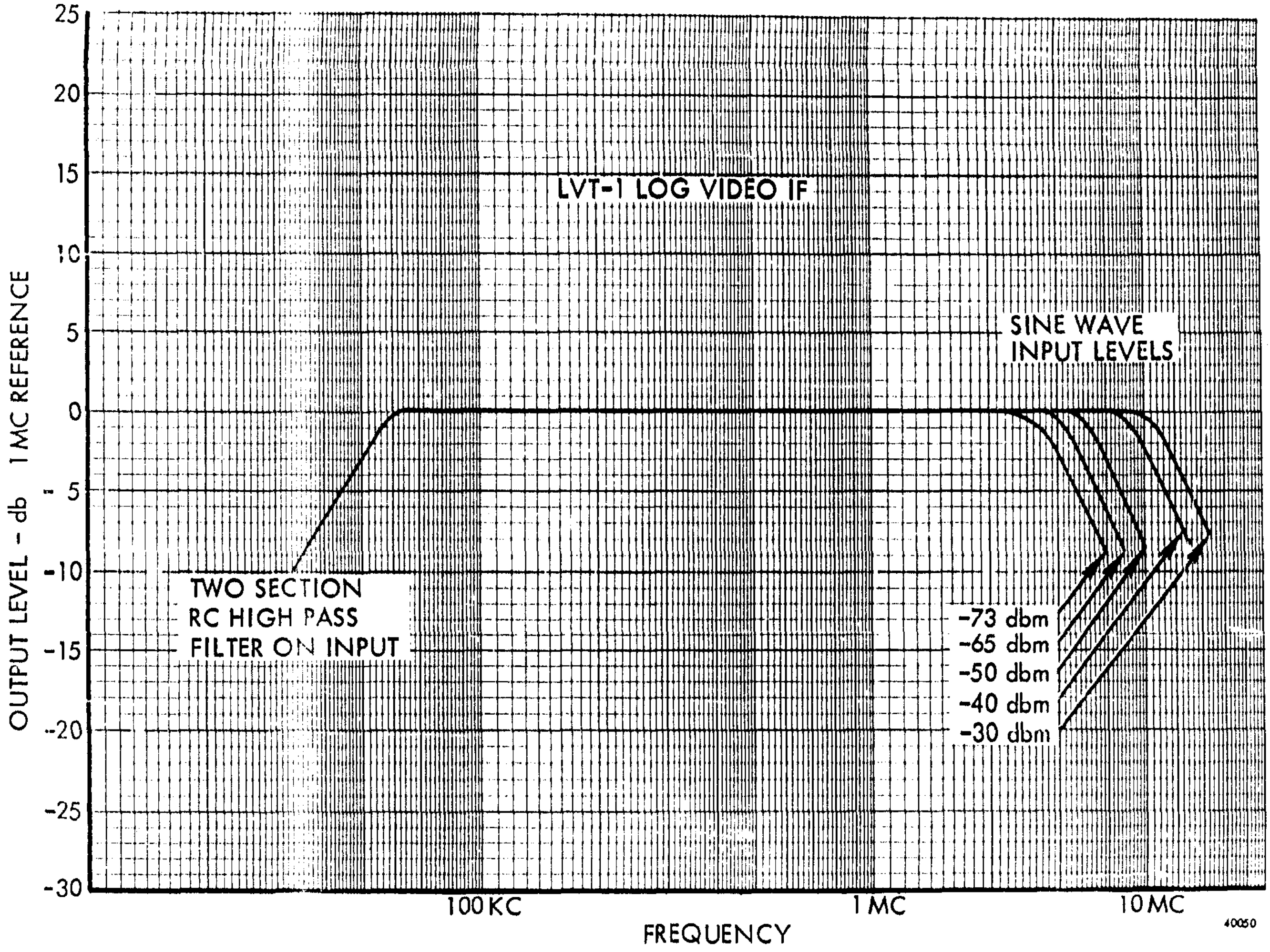
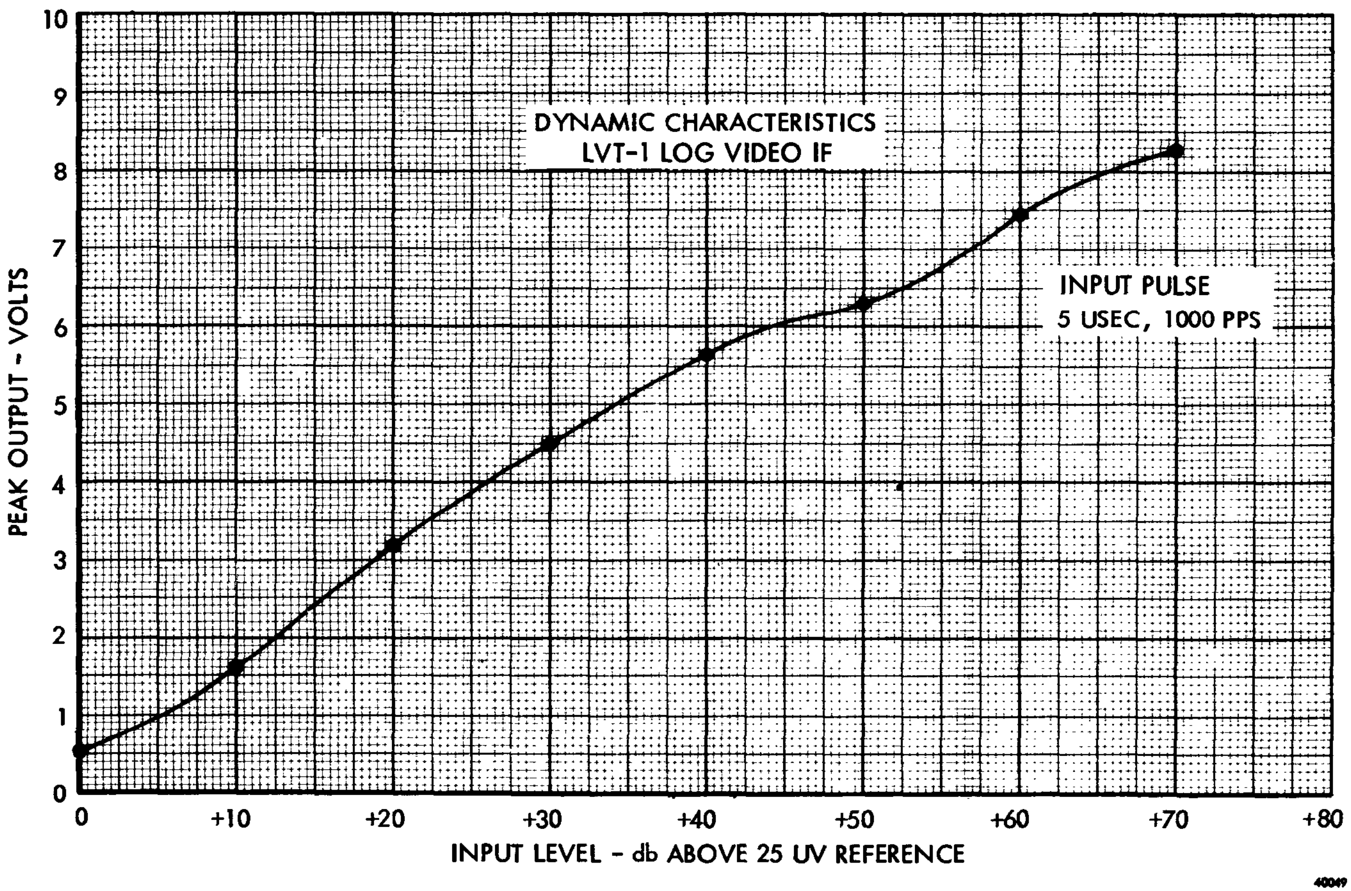


Figure 16. Log Video IF Amplifier - Frequency Response Versus Input Level


DYNAMIC CHARACTERISTICS
LVT-1 LOG VIDEO IF
INPUT PULSE
5 USEC, 1000 PPS
PEAK OUTPUT - VOLTS
0
1
2
3
4
5
6
7
8
9
10
0
+10
+20
+30
+40
+50
+60
+70
+80
INPUT LEVEL - db ABOVE 25 UV REFERENCE
40049

was placed in the amplifier input to reduce low frequency mixer noise. From the measured characteristics of the TWT preamplifier, the overall RF sensitivity of the breadboard for a 10 db signal to noise ratio at 3 Gc would be

$$-114 + 4.8 + 10 + 10 = -89.2 \text{ dbm}$$

The actual measured overall RF sensitivity for an estimated 10 db signal to noise ratio was approximately -87 dbm. The above calculation for overall sensitivity does not include the degradation of RF system noise figure due to image channel noise. A 3 db correction for the image channel noise would result in the calculated sensitivity being very close to the measured sensitivity.

Sensitivity was measured under the following conditions:

a. Input pulse 5 microseconds at 3 Gc, 1000 pps

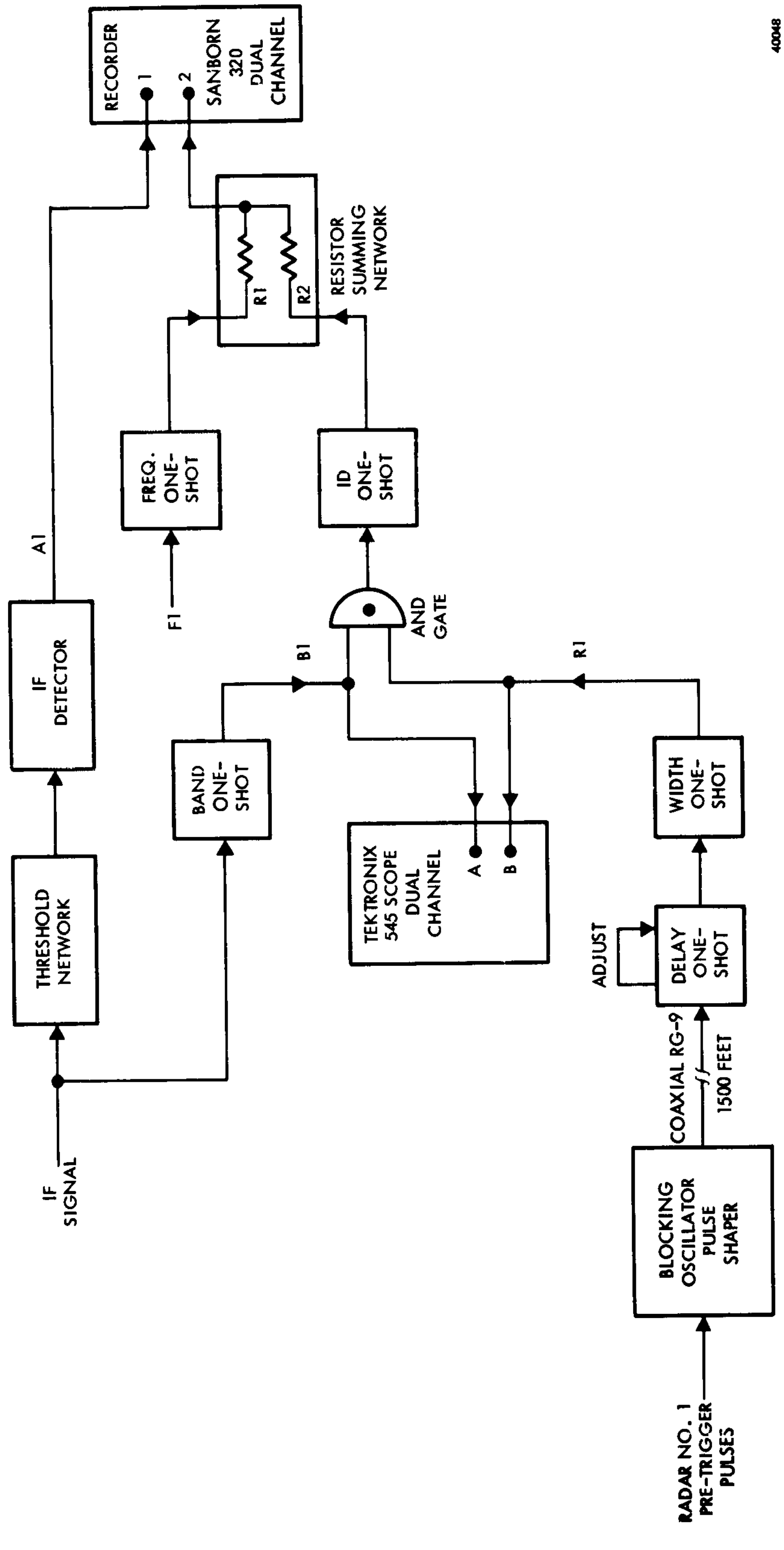
b. Local oscillator tuned approximately 2.5 Mc above or below 3 Gc

c. Signal observed at video IF output

The video IF amplifier approach was later discarded for reasons outlined in Section 2.14. A breadboard 30 Mc IF has been substituted for the video IF. This IF is an eight stage synchronously tuned amplifier employing single tuned capacitive coupled stages. Shunt diodes have been added across the grid resistors of the stages to provide compression of high level input signals. Low level IF bandwidth is approximately 5 Mc.

3.5 Data and Display Subsystem

A block diagram of the data and display portion of the breadboard system appears in Figure 18. To demonstrate interference source identification, a single coaxial cable connection between the radar site and the monitor is anticipated. Pre-trigger video pulses from a selected radar are conditioned by a blocking oscillator to produce a 20 volt, 1 microsecond pulse, having an output impedance of 50 ohms. The conditioned pulses are transmitted along 1500 feet of coaxial cable, RG-9, to a 50 ohm termination at the monitor video input. A 2 volt, noise limiting threshold level is established at the delay one-shot input to inhibit signals other than the conditioned video pulses from triggering the delay control. The delay one-shot pulse


RECORDER
1
2
SANBORN
320
DUAL
CHANNEL
R1
R2
RESISTOR
SUMMING
NETWORK
A1
FREQ.
ONE-
SHOT
F1
ID
ONE-
SHOT
AND
GATE
B1
R1
IF
DETECTOR
THRESHOLD
NETWORK
IF
SIGNAL
BAND
ONE-
SHOT
TEKTRONIX
545 SCOPE
DUAL
CHANNEL
A
B
WIDTH
ONE-
SHOT
ADJUST
DELAY
ONE-
SHOT
COAXIAL RG-9
1500 FEET
BLOCKING
OSCILLATOR
PULSE
SHAPER
RADAR NO. 1
PRE-TRIGGER
PULSES
40048

width is manually adjustable from 7 to 21 microseconds to allow leading edge alignment of the RF pulse and the reconstructed video pulse. After differentiating the output of the delay one-shot, the trailing edge is used to trigger the width one-shot. The width pulse is fixed at nominally 5 microseconds. Hence, the reconstructed video pulse thus derived is partially time coincident with the incoming main band RF pulse. The reconstructed video pulse is labeled R1 on the diagram indicating its relationship to Radar #1. Assuming an interference condition, an IF signal is generated in the monitor receiver which triggers the "band one-shot" yielding an output pulse of 5 microseconds and labeled B1 on the diagram. Since B1 is time coincident with R1, a pulse is generated at the output of the AND gate which triggers the I.D. (identification) one-shot. The I.D. one-shot output is directed to one channel of a dual channel Sanborn chart recorder (Model 320) through a resistor summing network. The resistor summing network is used to process two sources of independent data onto one channel of the recorder. Frequency markers derived from the receiver trigger a "frequency one-shot", the output of which is also directed to the recorder through the resistor summing network. Given identical one-shots, the information is distinguished on the recorder by choosing a different resistance value for each summing resistor yielding an apparent difference in amplitude on the recorder.

The remaining channel is used to record the amplitude of the interference signal. When the IF signal exceeds its specification limit, the excess amplitude passes through the threshold network to the IF detector where it is detected and time stretched to cause full deflection on the recorder.

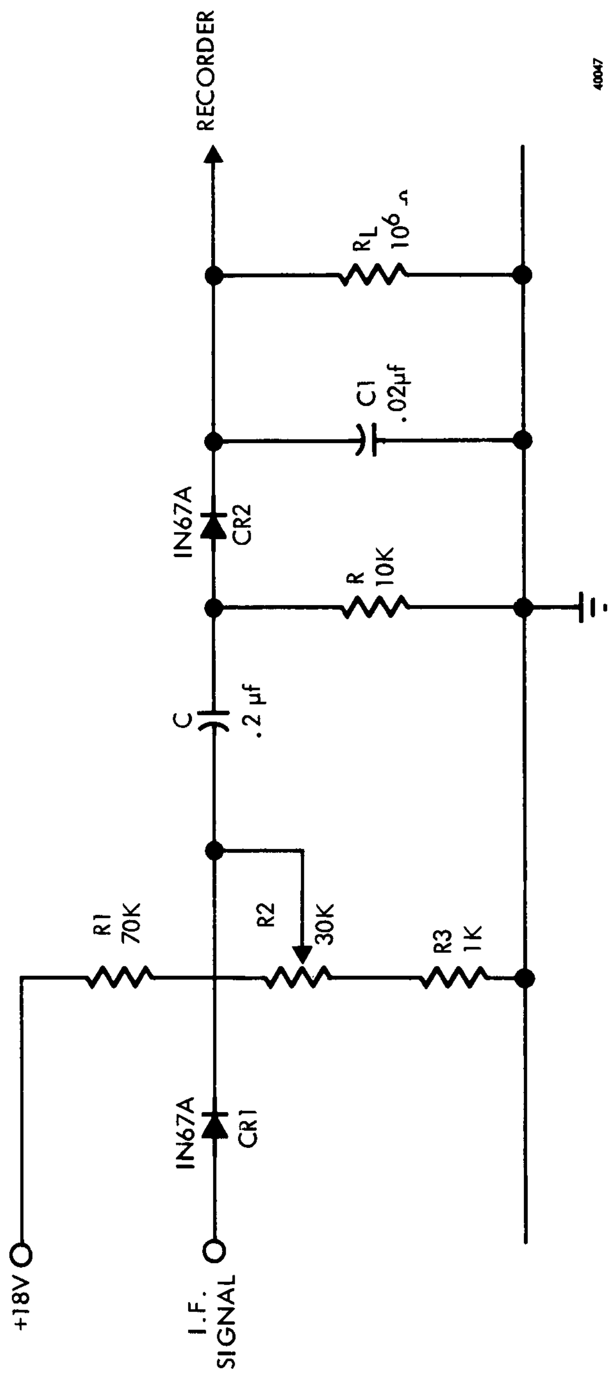
A Tektronix 545 scope with dual channel plug-in units is used for calibrating the delay one-shot. The B1 and R1 pulses are viewed on separate scope channels and the delay one-shot is adjusted until the leading edges of the pulses are aligned.

3.5.1 Circuit Diagram Description

3.5.1.1 Threshold Network and IF Detector

A circuit diagram of the threshold network and IF detector appears in Figure 19. The load resistance R_L is the input resistance to the Sanborn recorder, 1 megohm. The bandwidth of the recorder is 125 cps. Hence from (3) page 46,

$$(8) \quad t = \frac{4}{2\pi B} = \frac{4}{2\pi(125)} = 5 \text{ milliseconds.}$$


RECORDER
40047
RL
10^6 Ω
C1
.02µf
IN67A
CR2
R
10K
C
.2 µf
R1
70K
R2
30K
R3
1K
IN67A
CR1
+18V
I.F.
SIGNAL

For 20% decay during this time it can be shown that,

(9) $T_1 = C_1 R_L = .023$ sec.

thus,

$$(10)\ C_1 = \frac{.023}{R_L} = \frac{.023}{10^6} = .023 \text{ uf}$$

The threshold level is set by resistor divider network R_1, R_2, R_3, and is manually adjustable from 0.2 to 6 volts.

3.5.1.2 Band One-Shot

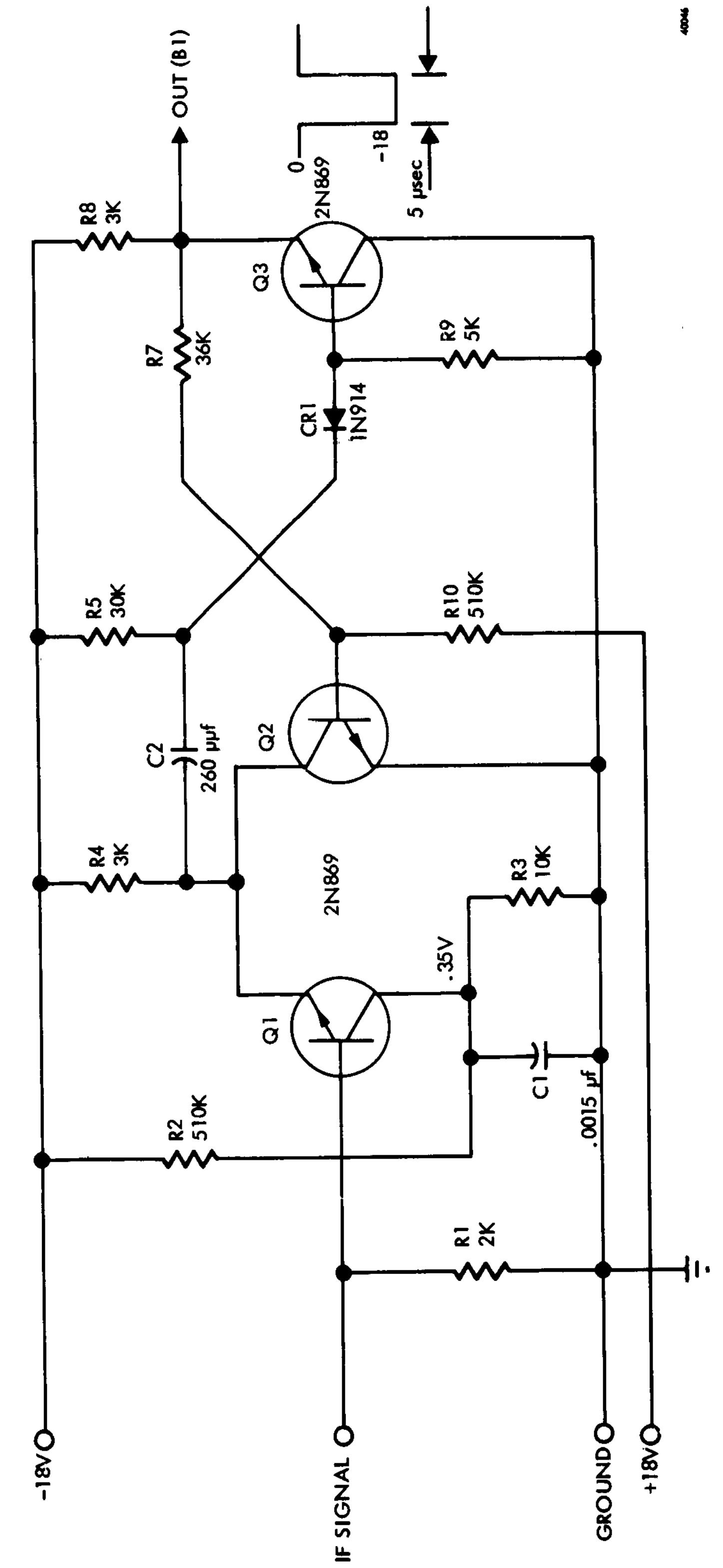
A circuit diagram of the "band one-shot" appears in Figure 20. The time constant R5-C2 is chosen to yield a 5 microsecond output pulse. The input is driven by the interference IF signal to which the base to emitter diode of transistor Q_1 and capacitor C_1 act as a detector. C1 charges to the negative peak of the input signal, triggering the one-shot initially and then preventing further triggering by subsequent negative excursions of the input signal. A triggering threshold level is established by resistor divider network R_2, R_3 which develops a back bias on the base to emitter junction of Q_1 of approximately 0.35 volts.

3.5.1.3 Delay One-Shot

A circuit diagram of the "delay one-shot" appears in Figure 21. The input is terminated in 50 ohms to match the coaxial cable from the radar site. A one volt back bias is established across the base to emitter junction of the input transistor and the input signal is divided down by a factor of two prior to the first stage. Hence, an effective 2 volt threshold level is established. The output pulse width is controlled by R_6 and is adjustable from 7 to 21 microseconds.

3.5.1.4 Width One-Shot

A circuit diagram of the "width-one-shot" appears in Figure 22. The delay pulse is differentiated at the input by capacitor C1 and resistor combination R1 and R2. The negative going trailing edge thus derived triggers the width one-shot. The time constant R4-C2 is chosen to yield a 5 microsecond output pulse.



OUT (B1)
0
-18
5 μsec
R8
3K
2N869
Q3
R7
36K
CR1
1N914
R9
5K
R5
30K
R10
510K
Q2
C2
260 μμf
R4
3K
2N869
R3
10K
.35V
Q1
C1
.0015 μf
R2
510K
R1
2K
-18V
IF SIGNAL
GROUND
+18V

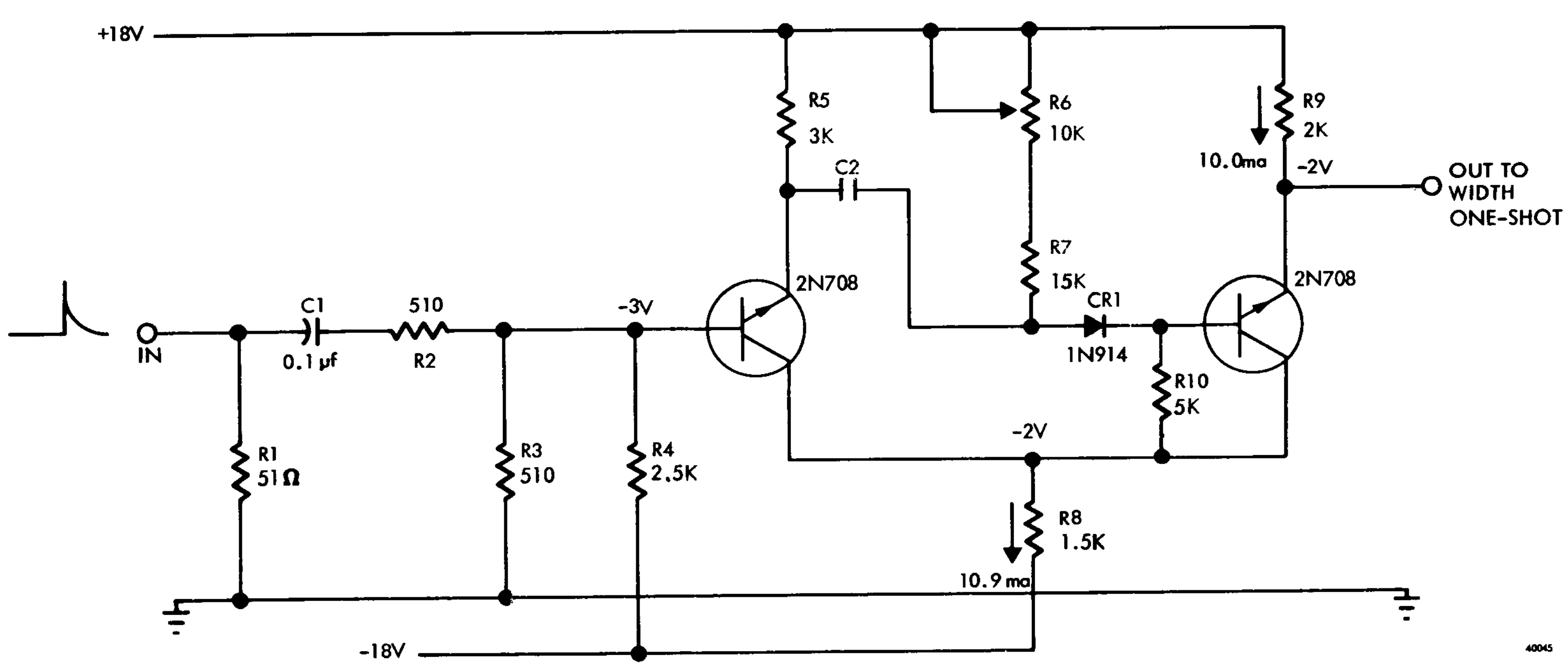
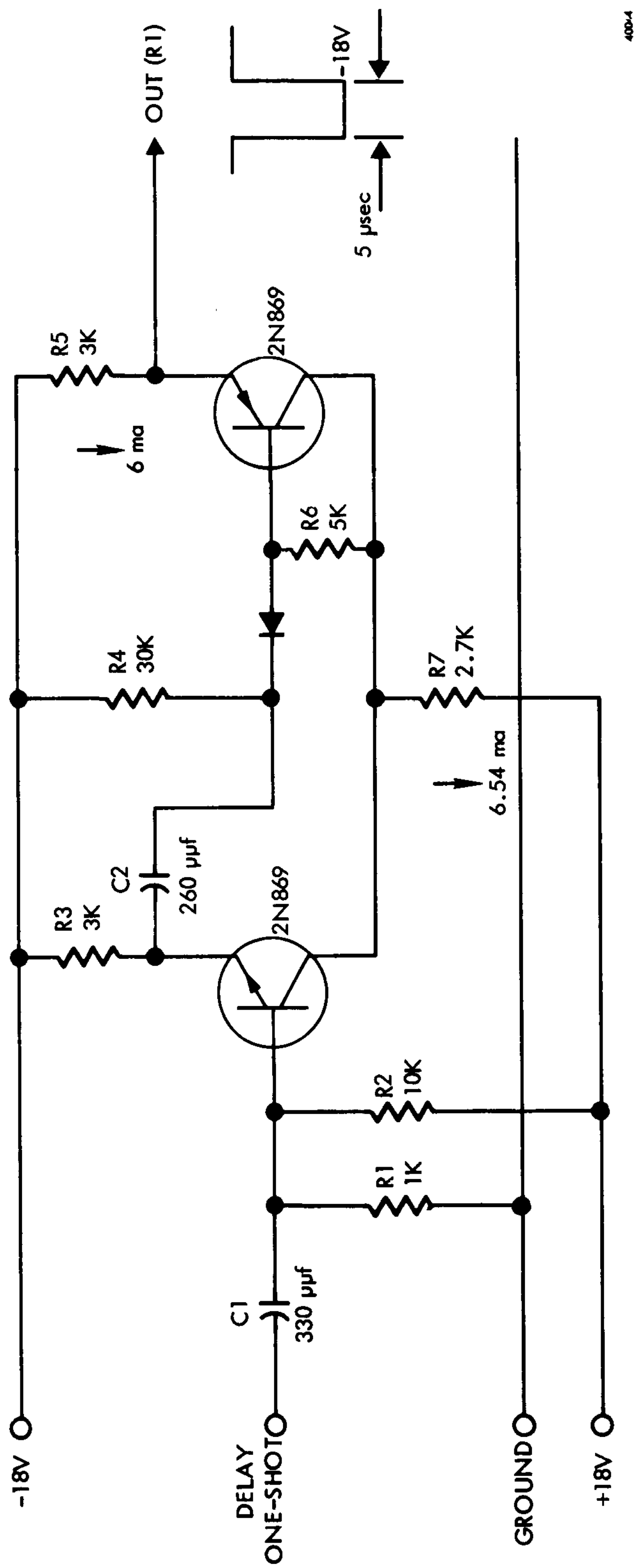


Figure 21. Schematic - Delay One-Shot


OUT (R1)
-18V
5 μsec
400-4
R5
3K
6 ma
2N869
R6
5K
R4
30K
R7
2.7K
6.54 ma
C2
260 μμf
R3
3K
2N869
R2
10K
R1
1K
C1
330 μμf
-18V
DELAY
ONE-SHOT
GROUND
+18V

3.5.1.5 "AND" Gate and I.D. One-Shot

A circuit diagram of the AND gate and I.D. one-shot appears in Figure 23. If either B1 or R1 is zero, point "a" is clamped to near ground through diode CR1 or CR2. When both B1 and R1 equal -18 volts simultaneously, point "a" is driven negative. The negative voltage thus derived is coupled through capacitor C1 to the base of transistor Q1 and triggers the one-shot. Since the I.D. one-shot feeds the recorder, its pulse width must be consistent with the recorder bandwidth. As determined previously, 5 milliseconds is required for 99% deflection. The R3-C2 time constant is chosen to yield this value.

3.5.1.6 Frequency One-Shot

A circuit diagram of the frequency one-shot appears in Figure 24. This circuit is identical to the preceding one except for the AND gate and output resistance.

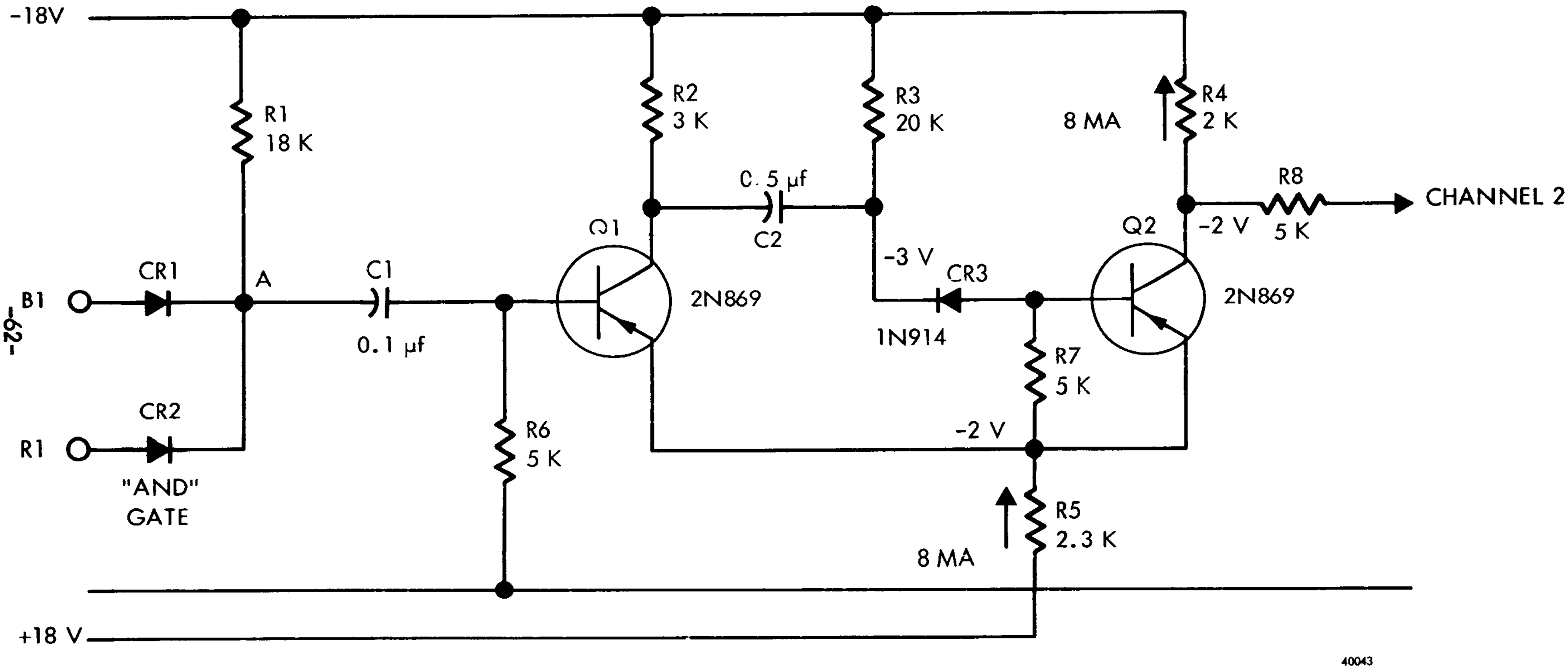
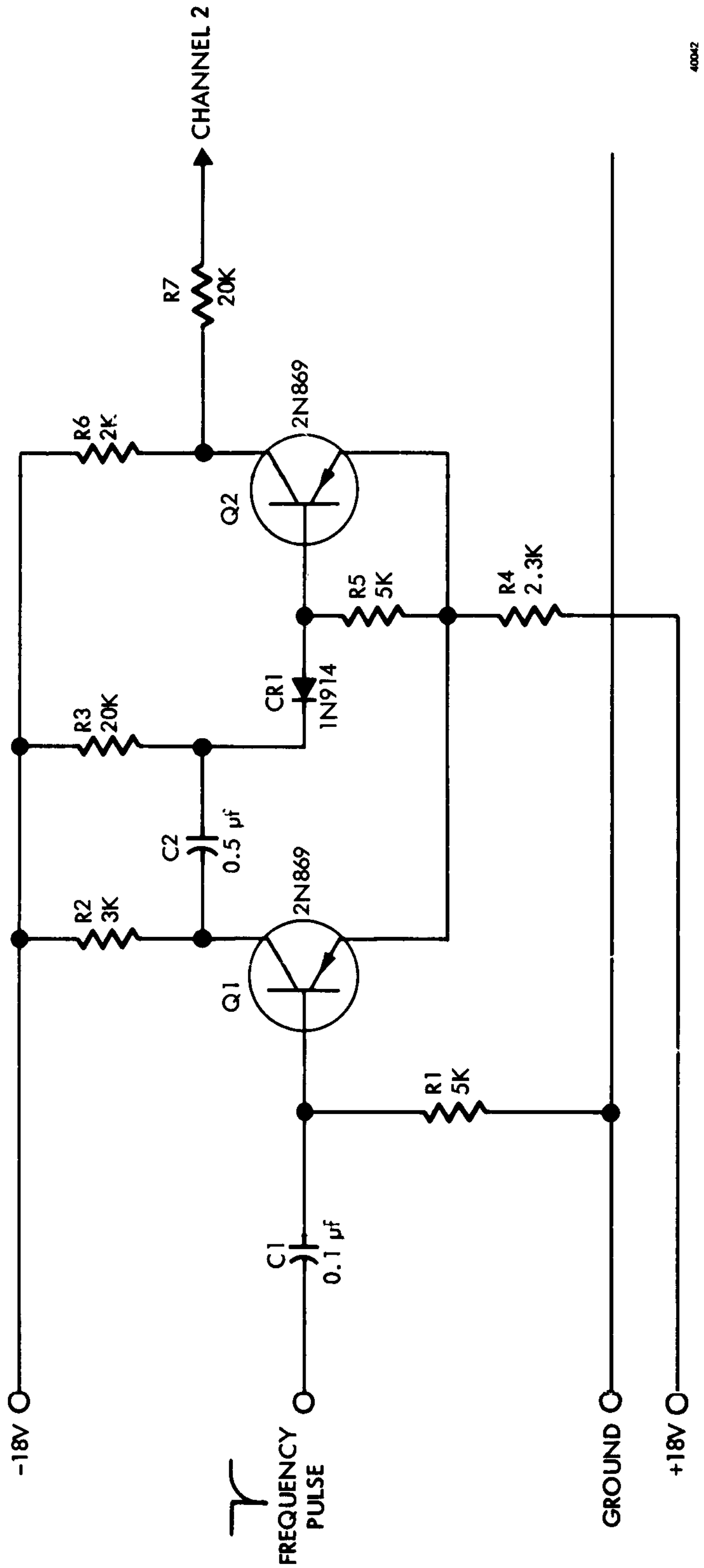


Figure 23. Schematic - AND Gate and Identification One-Shot


CHANNEL 2
R7
20K
R6
2K
2N869
Q2
R5
5K
R4
2.3K
CR1
1N914
R3
20K
C2
0.5 µf
R2
3K
2N869
Q1
R1
5K
C1
0.1 µf
-18V
FREQUENCY
PULSE
GROUND
+18V
40042

4.0 BREADBOARD TEST PLAN

The following tests are designed to demonstrate the feasibility of interference monitoring techniques which have evolved as a result of a study program supported by the Applied Research Branch of the Rome Air Development Center, Griffiss Air Force Base under Contract AF 30(602)-2695.

The performance of these tests will require the following test equipment and support from the Applied Research Branch:

Test Equipment

1. S-Band Signal Generator - HP 616 or equivalent
2. Oscilloscope - dual trace - Tektronix 545 with: Type C/A and Type H plug-in heads
3. Power Meter - HP 430C with HP 477B thermistor mount
4. Two Channel Recorder - Sanborn Model 320 or equivalent
5. BWO Power Supply - Alfred Model 602 - second choice Model 610C
6. Regulated Power Supply - Dual 0 to +300 at 150 ma
7. Regulated Power Supply - 0 to 30V at 100 ma, 3 required
8. Pulse Generator - HP 212A or equivalent
9. Pulse Generator - Rutherford B7B or equivalent
10. Frequency Meter - Narda 802B

Support

1. Personnel - The services of an engineer or technician familiar with the test site and radars to assist in obtaining pre-trigger pulse information from a selected radar to the monitor site is required. A qualified driver is required to move the vehicle used to transport the monitor equipment to the test site location. The driver will be available for other duties once the vehicle is located.

2. Transportation - A vehicle capable of transporting the monitor equipment to the site and providing protection from the weather. This vehicle preferably should be a shielded van with provision for coaxial cable entry and suitable power line filtering.

3. Power Supply - A portable power unit - 115V 60 cps capable of regulated output of approximately 6 KW.

4. Special Cabling - Coaxial cabling, preferably RG-9 or RG-8, is needed to carry pre-trigger video pulses from a selected radar site to the monitor. The minimum length required is 1500 feet, not necessarily in one section. The cable should be laid on top of the ground from the radar site chosen to the monitor located on Germany Road. When crossing roadways the cable should be protected from traffic by routing through pipe at the crossover point (or other suitable method). The cable will be laid for approximately a two-week period. Provision for getting the coax cable inside the selected radar and monitor vans is required.

5. Communication Equipment - Transceiving equipment permitting voice communication between radars and the monitor is required.

6. Miscellaneous

    a. Two 6 foot relay racks with mounting hardware for mounting equipment

    b. Tables or carts for supporting test equipment

    c. Miscellaneous RF fittings and cables as required

Basic Feasibility Breadboard Tests

1. Set Up and Calibration - Monitor equipment will be set up at a pre-determined site and calibrated.

   1.1 Calibration

   After a 30 minute warm up period, the monitor receiver threshold level will be calibrated at four frequencies: 2500, 3000, 3500, and 4000 MC. The threshold will be calibrated with a pulsed input signal from a HP 616B signal generator. A nominal pulse width of 5 microseconds at 1000 pps will be employed. The signal generator output will be adjusted to a value determined from the following equation:

   $$P_{SG}dbm = -10 \log (4.1(10)^9 d^2) + \text{antenna gain in db}$$

   Where d is the average range in miles from the monitor site to the radar transmitters. The antenna gain for each frequency will be determined from a previous free space measurement of the antenna pattern.

2. Band Monitor and Record - After the monitor equipment has been calibrated, received signals in the 2 to 4 Gc band will be recorded. During the recording period several radar equipments will be operated in a transmitting mode. The radar antennas will continuously rotate at the highest rate, and elevation angles will be nominally zero degrees.

3. Transmitter Identification - The proposed transmitter identification technique requires a pre-trigger pulse from each transmitter being monitored. For purposes of demonstrating the feasibility of this approach, the pre-trigger pulse from one radar will be employed. The validity of the transmitter identification provided by the monitor will be verified by turning off radar equipments until the source of a given spurious signal is established with certainty.

4. <u>Monitor Location</u> - During transmitter monitoring periods, the monitor will be situated at any of several convenient locations adjacent to Germany Road. Such positions must permit connection to the transmitters by the trigger cable provided. The path between the monitor and the transmitters being monitored should be unobstructed by trees or buildings.

5. <u>Duration of Tests</u> - Radiation Incorporated engineers will be in Rome for a two-week period. It is expected that the time required for actual testing will not exceed three days. The balance of the available time will be expended in equipment set-up and checkout and in review of test results with RAUMA personnel. Tentatively, this two-week period is from 15 to 26 April.

5.0 CONCLUSIONS

At the writing of this Interim Report the breadboard monitor has been constructed and tested in the laboratory. When operated in conjunction with pulsed signal generators simulating spurious signal sources, the monitor has consistently recognized threshold signals and provided the correct identification. The lack of high power pulsed signal sources placed a limitation on the dynamic range of laboratory tests, but small signal detection characteristics appear unchanged when interfering signals at the same frequency have amplitudes in excess of one milliwatt.

The results of tests to be performed at the Verona Test Site in April, 1963 will be reported in the Final Report. The Final Report will provide detailed recommendations on the ultimate monitor and indicate areas where the frequency coverage, dynamic range, and other measurement characteristics of the monitor could be extended.

It is anticipated that the monitoring techniques will permit the day-to-day observation of the spurious signal radiation characteristics of Verona radar equipments. Since the data can be readily stored in a format compatible with available computers, the behavior of radar transmitters at non-design frequencies can be more fully and rapidly determined.