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Techniques for Suppressing Pulse Interference In UHF Receivers

By

R. E. Meek and W. B. Warren Jr.

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FINAL REPORT-VOLUME II

PROJECT NO. A-348

Contract No. AF 30 (602)-1789

Prepared for
Rome Air Development Center
Air Research and Development Command
United States Air Force
Griffiss Air Force Base, New York
1 June 1959

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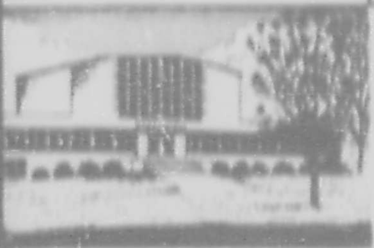

Engineering Experiment Station
Georgia Institute of Technology
Atlanta, Georgia

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By

R. E. Meek and W. B. Warren

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of the Georgia Institute of Technology
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FOREWORD

This is Volume II of a three volume Final Report on the work performed under Contract AF 30(602)-1789. Volume I, "The Performance of Communication Systems in the Presence of Interference," deals with a series of experiments performed at Georgia Tech concerning the effects of various types of interference on laboratory models of voice and digital communication systems, and with other general aspects of the communications interference problem.

The present volume treats the problem of suppressing pulse type interference to systems of conventional design, in particular narrow band voice systems. It includes a description of blankers developed on the project, remedial techniques for desensitization, and comments on standard suppression techniques described in the literature.

This volume is unclassified; however, certain pages in which reference is made to classified work reported by other agencies are classified, and will be found in Volume III.

ABSTRACT

Several schemes for reducing the effects of pulse type interference in UHF communications receivers are described and their principles of operation analyzed. Schemes investigated under the project include (1) a pulse-controlled cavity UHF blanker, which was carried to the prototype stage of development, (2) a broad band blanker, [redacted] (3) internal receiver circuit modifications which reduce desensitization effects, (4) a sampling filter, and (5) a non-linear capacitor scheme. The latter ~~two~~ received only very preliminary investigation. In addition, for the sake of completeness, an analysis of some techniques discussed in the literature is presented.

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I. Introduction

The advent of extremely high power radar equipment utilizing a large portion of the spectrum, and the necessity of locating communications equipment in the proximity of such radars, have created an interference problem for communications equipment of conventional design. A need exists for either auxiliary suppression devices or internal modification to ensure satisfactory operation of communications equipment in this environment.

Emissions have been catalogued for typical radar installations which fall in the HF through UHF portions of the spectrum, as well as in the microwave region.[5]

This report is primarily concerned with the effects of pulse type interference on UHF receivers.

The tone produced by radar interference is highly annoying to the operator of a UHF receiver when he is trying to understand code or speech. Also errors in the output of teletype or facsimile equipment may be produced by this type of interference. In addition to the annoying effect of the tone, a receiver may also suffer desensitization of its response to a desired signal. Other deleterious effects attributable to the large amplitude of pulses may occur in receivers of conventional design. The practical effect is that reception of a normal signal is made difficult, and in many cases impossible.

Operational experience with pulse type interference has prompted the development of various techniques for suppressing its effect on the receiving equipment either by deliberately blanking (cutting off) the receiver during the period of the pulse, or by limiting and filtering out the interference by special circuitry within the receiver.

These suppression techniques which are discussed in this report may be identified briefly as follows:

Blanking and gating techniques are those schemes by which the receiver is turned off for the duration of each interfering pulse and turned on again between pulses.

Limiter techniques are those which exploit a difference in the amplitude of the interfering pulse and desired signal. They provide a limiting action on the output level of the receiver at a predetermined level which results in an apparent improvement in the S/I ratio.

Guard Band Receiver techniques exploit spectral differences in the desired and undesired signals. Interfering energy "spilled over" or otherwise present in the upper and lower adjacent channels of a receiver, is used to generate deadening pulses applied to the desired signal channel.

Audio Filtering techniques reduce the level of the audio tone identified with repetition rate of the pulse source by employing comb filters in the audio channel of the receiver.

Techniques for reducing desensitization involve reducing the time constant of AGC networks at critical points in the receiver so that circuit charging caused from large amplitude pulses may not reduce the sensitivity of the receiver to the desired signal.

Two techniques are described whose principle of operation may fall in one of the above classes, but which involve unconventional type circuit [illegible] or circuit operation. One scheme makes use of the new diode capacitor* to facilitate the shifting of tuned circuit response during the interfering pulse, thus effectively reducing the gain of the associated amplifier stage. The other scheme employs a sampling technique with narrow sampling pulses originating from a jittered trigger. Pulse interference reduction is obtained by the low probability of coincidence between the sample pulse and interfering pulses, the desired signal being essentially unaffected.

Several of the techniques mentioned above were evaluated experimentally during the contract period. Investigation of one technique (Blankers) culminated in the construction of a field test model of a narrow band blanker for UHF communication receiver. This unit has been delivered to RADC for field tests.

2. Blanking

2.1 Blanking and Gating Techniques

High level pulse interference may cause severe [illegible] desensitization, or complete blocking. Excessive pulse amplitude overloads amplifier circuits, principally r-f and i-f amplifiers and mixers, and causes charging of resistance capacitance networks that have long discharge time-constants. Because r-f amplifiers are vulnerable to high level pulse signals, a scheme to gate the signal path preceding these stages is required. With moderately high level pulse signals, overloading of the r-f amplifiers may not be noticeable or damaging, but mixers or i-f amplifiers following these stages may be subject to excessive overload and a scheme that gates these circuits during an interfering pulse is effective in preventing degradation.

2.1.1 Narrow Band Blankers.(1) The narrow band blanker is essentially two highly selective resonant elements that have crystal diode shorting devices across their high impedance terminals, the diodes being triggered into the conduction condition by a positive gate pulse coincident with the interfering pulse. In the conducting condition these diodes produce a high attenuation in the transfer characteristic of the resonant network.

The resonant elements are two coaxial resonant cavities constructed in an integral cylindrical assembly. Crystal diodes are mounted between the end of each inner conductor and the adjacent outside wall (outer conductor) of the cavity. As long as each diode is back biased, its impedance is essentially a high resistance in parallel with a small capacitance, and the two cavities act as a conventional, low insertion-loss, selective filter. However, application of a forward biasing voltage to the diode causes them to appear as very low resistances effectively short circuiting each inner conductor to the outer wall of the cavity. This action alters the resonant frequencies of the two cavities and lowers their Q to such a degree that the device appears as a very high insertion loss attenuator (blanker) between the antenna and the receiver. Figure 1 shows the simplified constructional and electrical design of the cavity blanker. The cavity cylinder is constructed from a six-inch, silver-plated, brass pipe 16 inches long. The separator across the middle of the cylinder divides the space into two sections which constitute the two resonant cavities. Figure 2 shows a photograph of the internal construction of

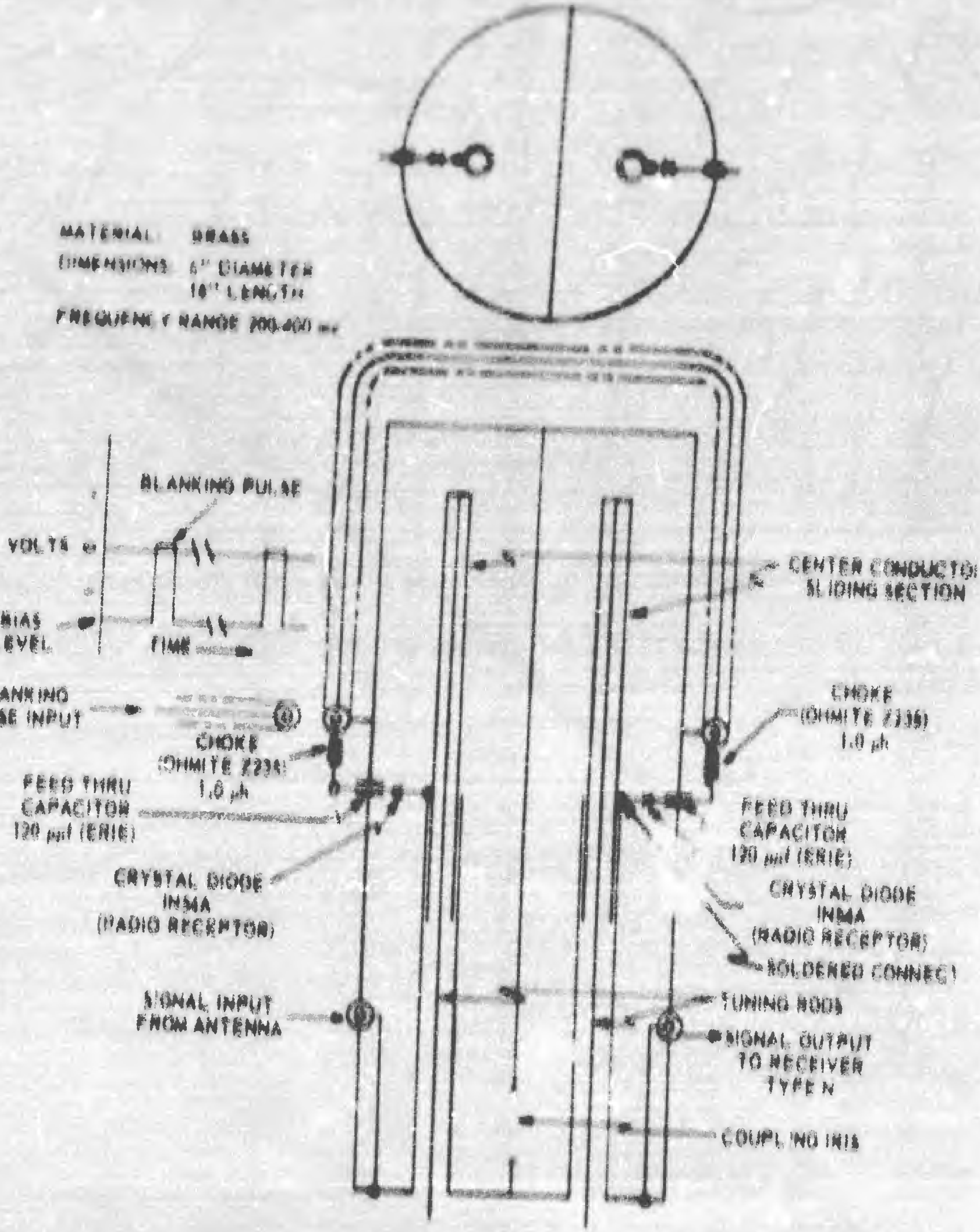

Figure 1. Pulse Controlled Cavity Blanker Model 1A.

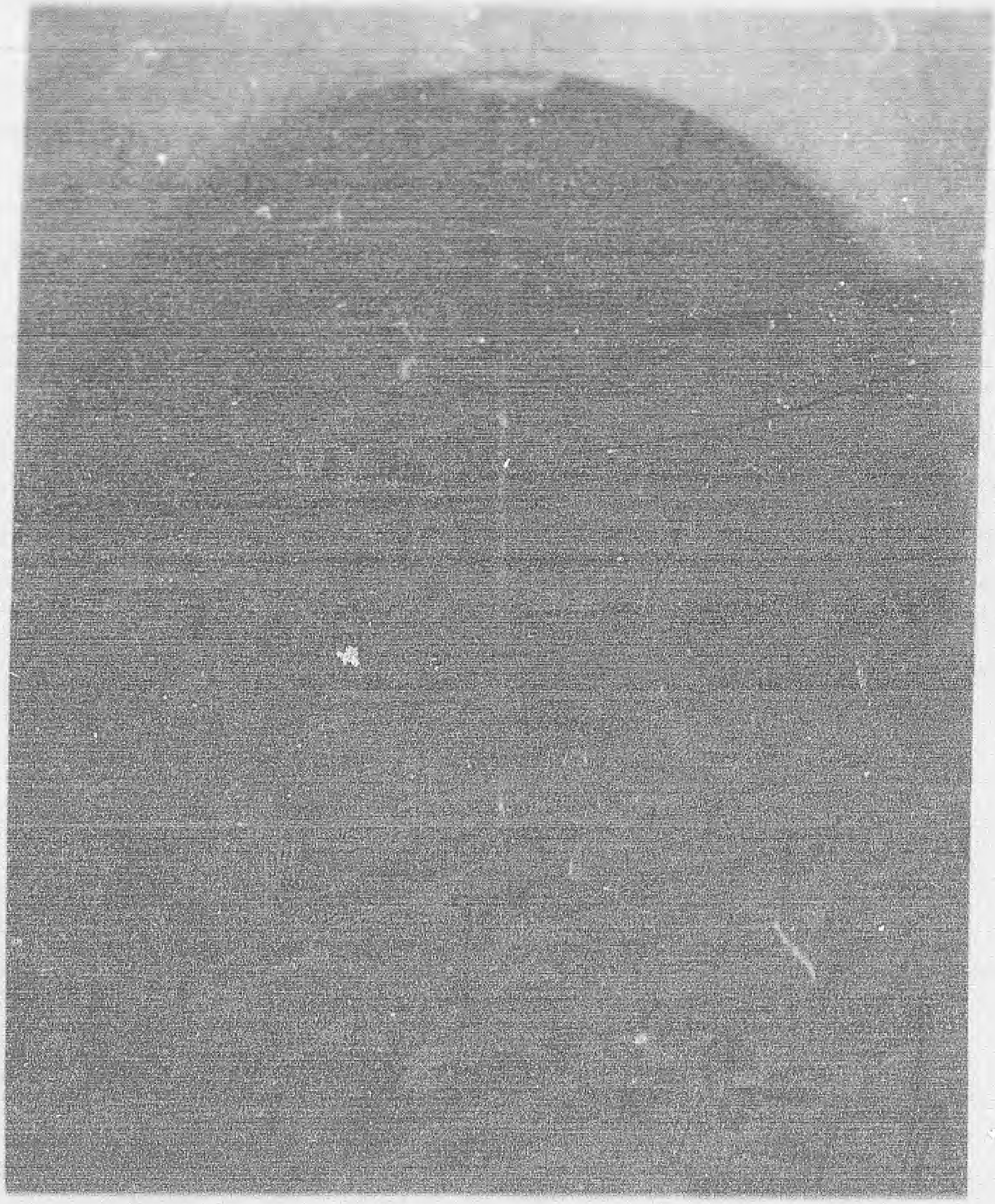

Figure 2. [illegible] Blaster Model 1A - Internal Construction.

the blanker. The center conductor of each cavity is a one-inch, silver-plated, brass tube with a snug-fitting, thin-walled, copper tube sliding inside of the brass tube. This sliding copper tube is the tuning element for adjustment of the resonant frequency of the cavities. A mechanical tuning assembly with a counter provides a facility for actual tuning and frequency calibration of the cavities. Each cavity is tuned independently of the other. Figures 3 and 4 are photographs of the mechanical details of the tuning assembly and the mounting scheme for the blanker.

Measurements were made to determine the selectivity and insertion loss of the blanker when the diodes were forward biased and back biased respectively. Figure 5 shows the insertion loss characteristics under both conditions. Selectivity characteristics of the blanker are shown in Figure 6, which also shows a comparison of the blanker selectivity (Curve a) with that of a Collins 156-C2 Filter (Curve b). Connecting these two devices in cascade produces Curve c, and reveals a considerable improvement in the r-f selectivity of a communications receiver and a Collins Filter when the narrow band blanker is added. The possibility of closer channel spacing than is usually permitted makes the use of this device more attractive for existing installations than it would be if its blanking function alone were considered.

In utilizing the narrow band blanker a blanking pulse generator is required to provide the proper type of triggering information. Figure 7 is a block diagram of a system showing the basic components and principal signals involved. The blanker system is designed to be synchronized with a trigger from the source of interference. Figure 8 shows the functional block diagram of the Blanking Pulse Generator, and a schematic of this unit is given in Appendix I. The repetition rate of the incoming trigger pulse is multiplied by ten before the blanking pulse is generated and applied to the blanker so that no audible tone appears in the receiver output due to modulation of the desired signal by the blanker. A test was made using a trigger rate of 1,000 pps to generate a blanking pulse of the same rate. The receiver in the test arrangement was attached to a high power pulse source (100 watt peak) and a desired signal source (10 microvolts). The blocking condition in the receiver previously produced by the pulse source was alleviated, but the 1,000 cps

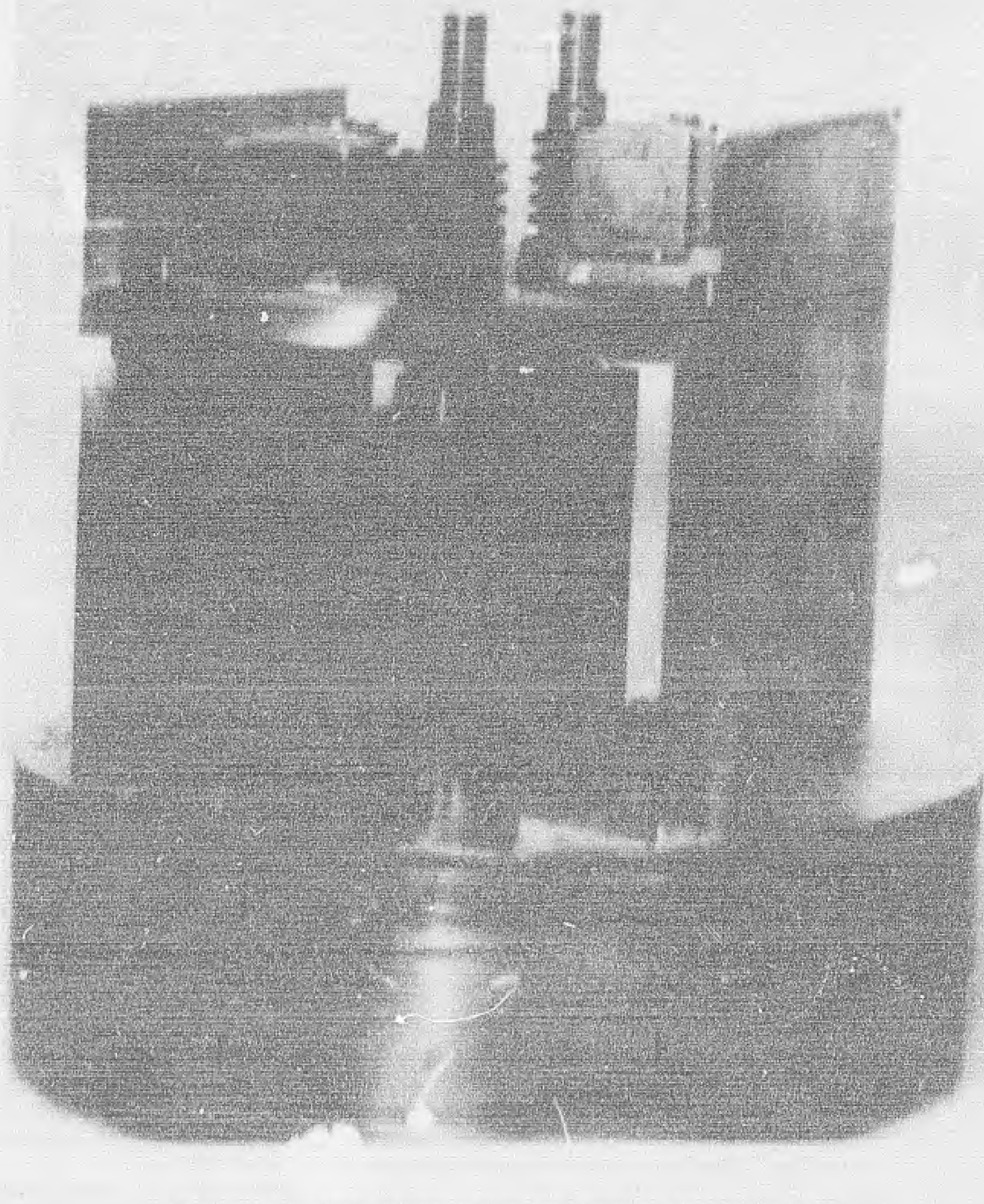

Figure 4. UHF Cavity Blocker Tuning Assembly View

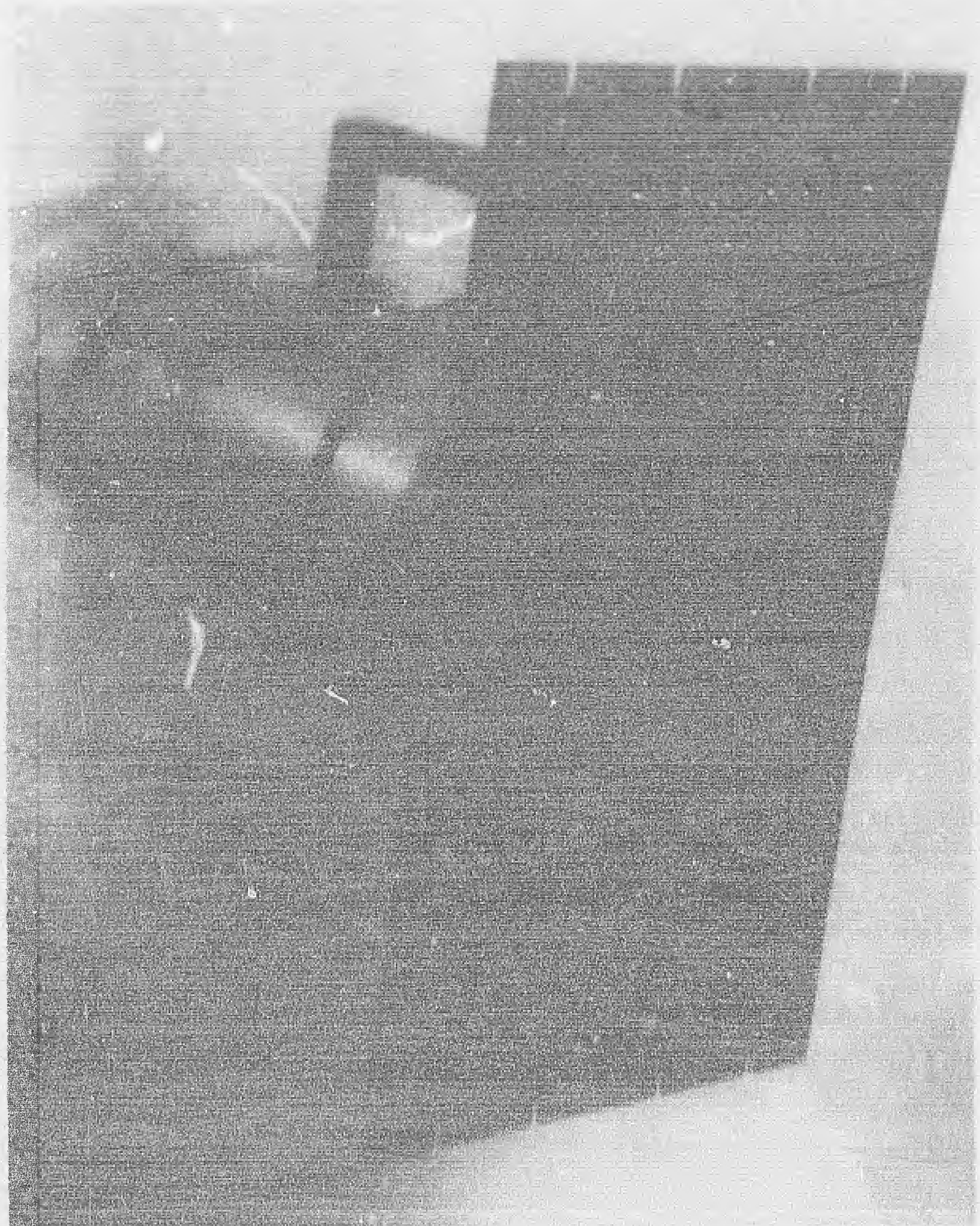

Figure 4. [illegible] Cavity Blanket Mounted Assembly.

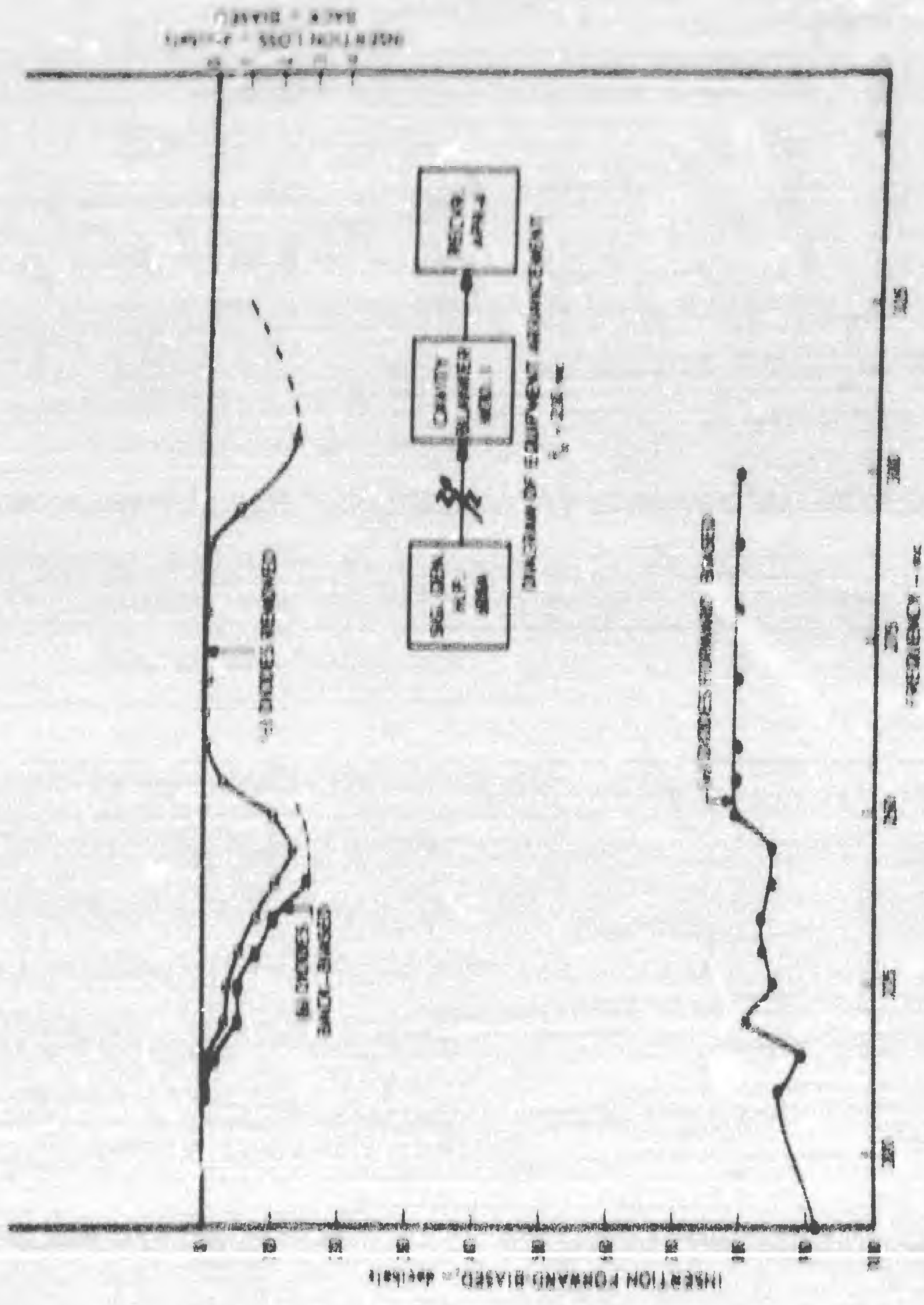

Figure 5- Insertion Loss Characteristics of Cavity Duplexer Model 1.

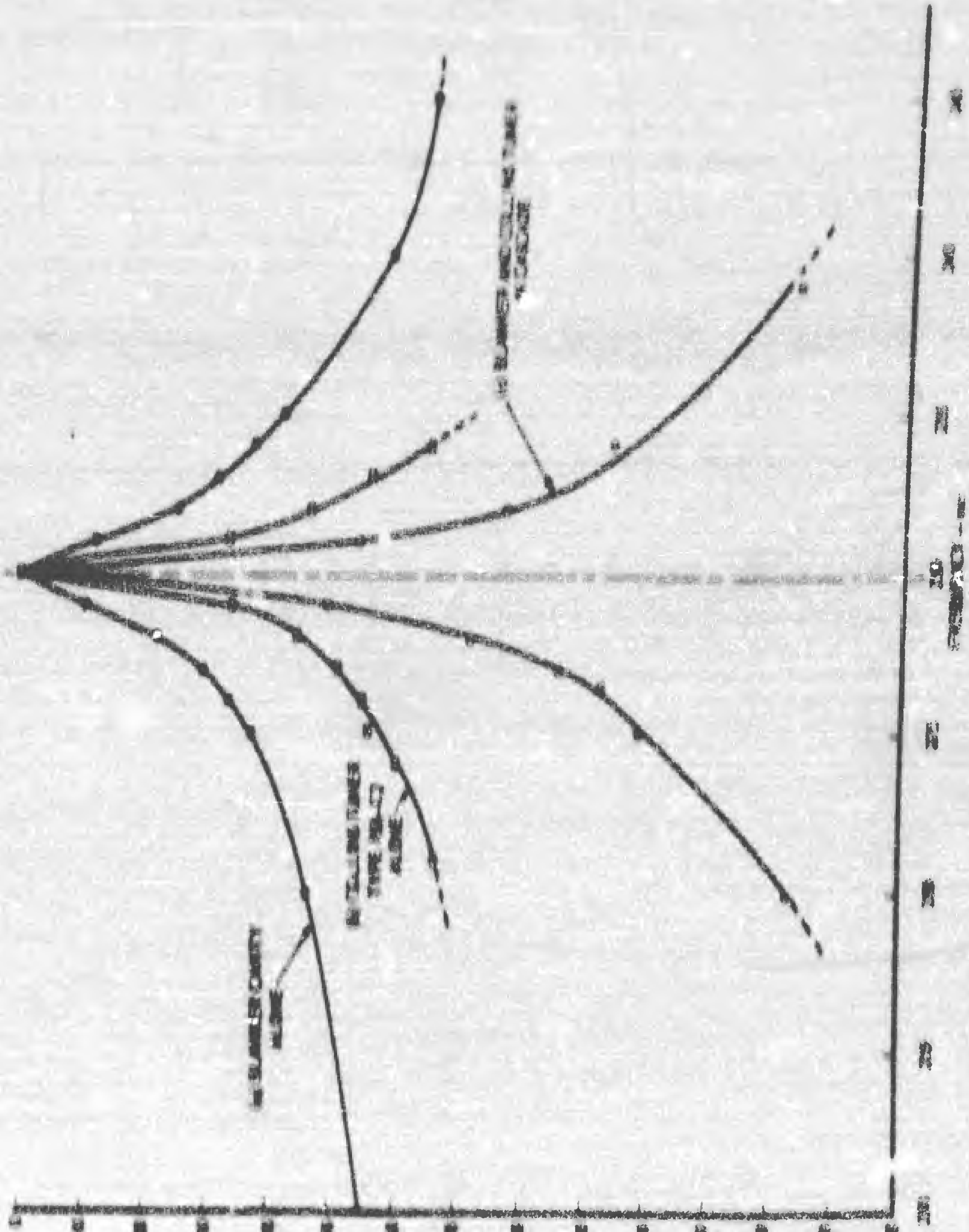

Figure 6.

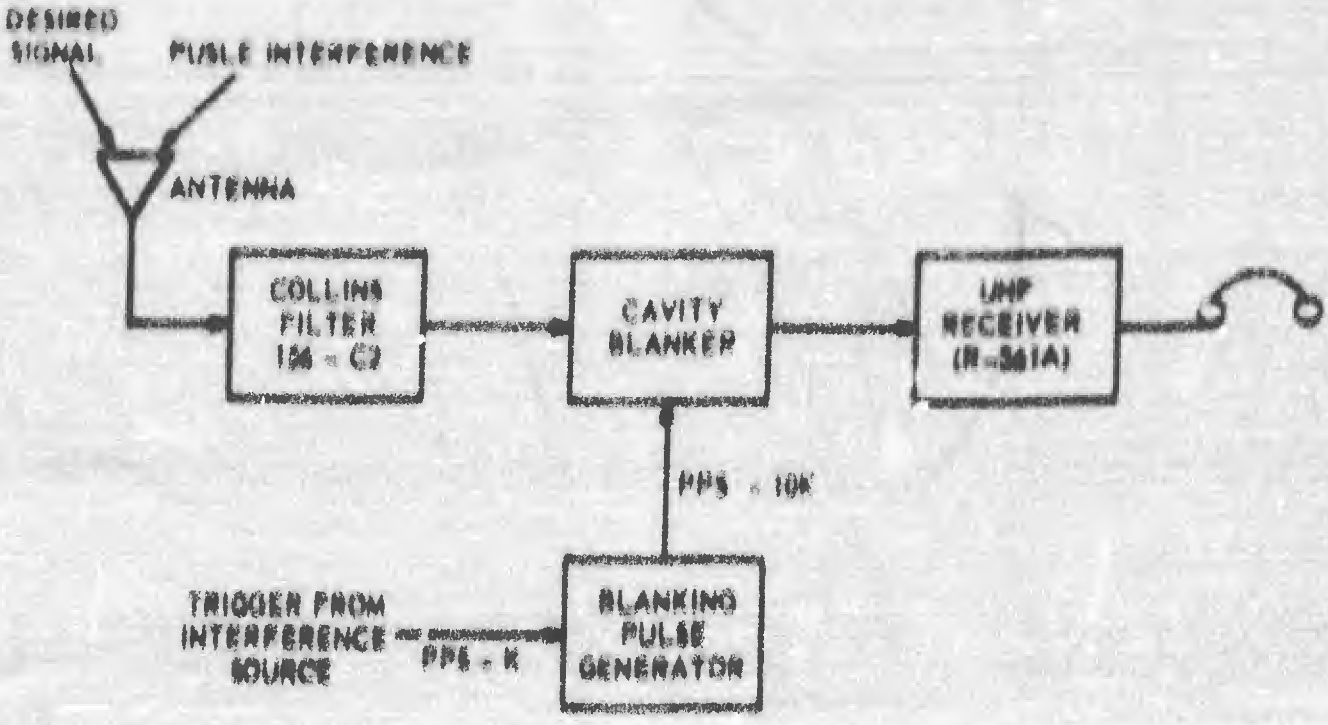

Figure 7. Equipment Arrangement Utilizing UHF Cavity Blanker.

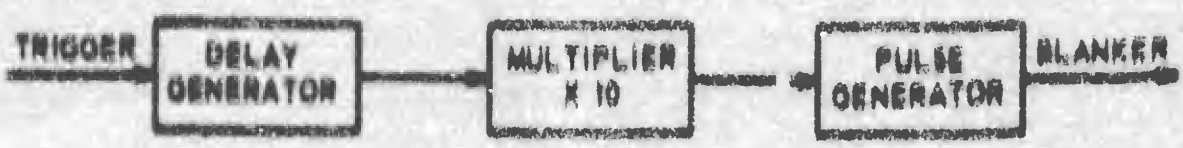

Figure 8. Functional Block Diagram of Blanking Pulse Generator.

tone was very much in evidence in the output. When the blanking pulse rate was increased to ten times the trigger pulse rate the output of the receiver was essentially free of any annoying tone, leaving the signal highly intelligible and the receiver noise level normal. The width of the blanking pulse in each test case was 20 microseconds, and the interfering pulse width was 10 microseconds.

Tests were also made to determine the intermodulation characteristics of the blanker. Qualitative observations showed that interference of this type was negligible when the diodes were either forward or reverse biased.

2.1.2 Broad Band Blanker. The narrow band blanker previously described must be retuned whenever operation at a new frequency is desired. To eliminate this tuning requirement, a blanker was developed which would operate over the entire UHF communication band without tuning controls. The schematic of Figure 9 illustrates the method of obtaining blanking action over a wide frequency range.

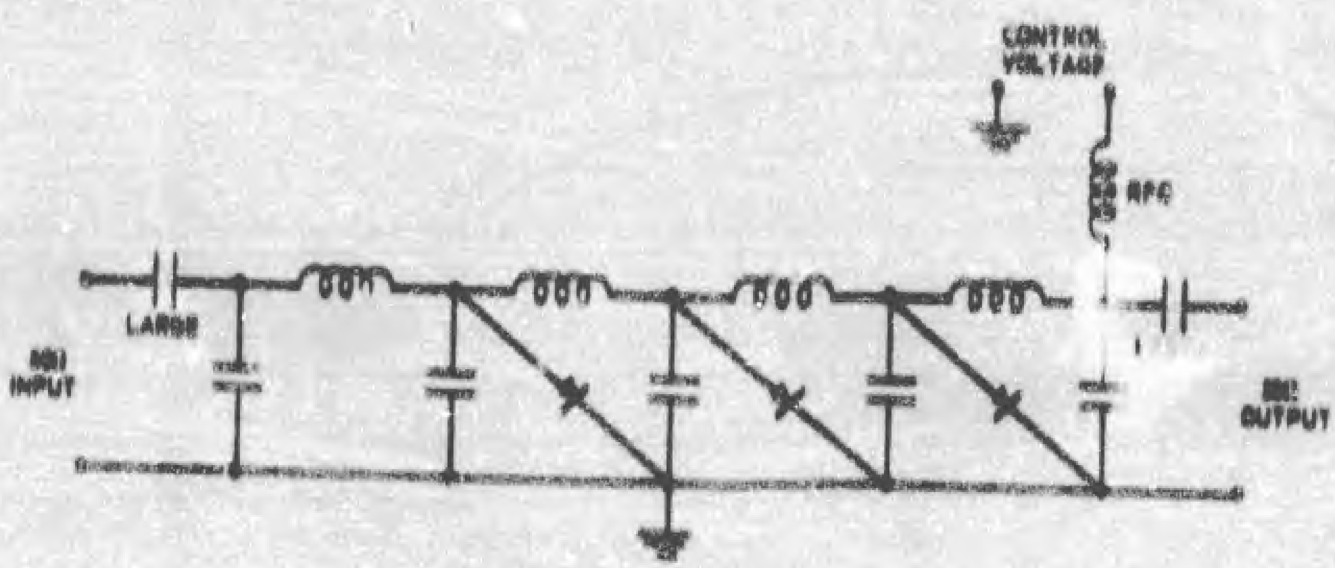

Figure 9. Broad Band Blanker

The blanker consists of a multisection, low-pass filter whose cutoff frequency is above 400 mc/sec. Each filter section has a normally back-biased diode in shunt with it, so that the transfer ratio of the filter at frequencies in the pass band is essentially unity. If the diodes are forward biased by the application of a suitable control voltage, each diode presents a low shunting resistance on the output of each filter section, and a large insertion loss is obtained. Hence application of the proper control voltage allows the insertion loss of the filter to be switched from a very low value to a very high value. The large capacitors in series with the input and output terminals serve to isolate the control voltage from the source and the load without attenuation of the r-f signal.

The curves of Figure 10 indicate the magnitude of the minimum and maximum insertion loss obtained with a laboratory model of the blanker under the conditions indicated in the figure. These curves show that a very low insertion loss can be obtained while maintaining a blanking insertion loss of greater than 50 decibels over the range 200 - 400 mc/sec. A higher loss of about 60 decibels is obtained in the lower part of the range where it is anticipated that the greatest use of this device would be made. Additional development effort could possibly raise the maximum attenuation to 70 or 80 decibels.

2.1.3 Local Oscillator Gating. In many pulse interference situations, the level of the interfering pulses is not sufficient to overdrive the r-f amplifiers and mixer of the receiver, but is still of sufficiently high level to seriously impair the receiver's performance in receiving a desired signal. In these instances blanking obtained by switching off the local oscillator of the receiver in synchronism with the interfering pulses will greatly reduce the effects these pulses have on the reception of a desired signal. In those time intervals during which the local oscillator signal is absent, no i-f voltage appears at the mixer output and blanking of the interfering pulse is accomplished. In order that the blanking action not generate interference of its own due to transients produced by rapidly starting and stopping the local oscillator, the blanking should be initiated a few microseconds before the occurrence of an interfering pulse, and should not be completely removed for an additional few microseconds after the interfering pulse has terminated.

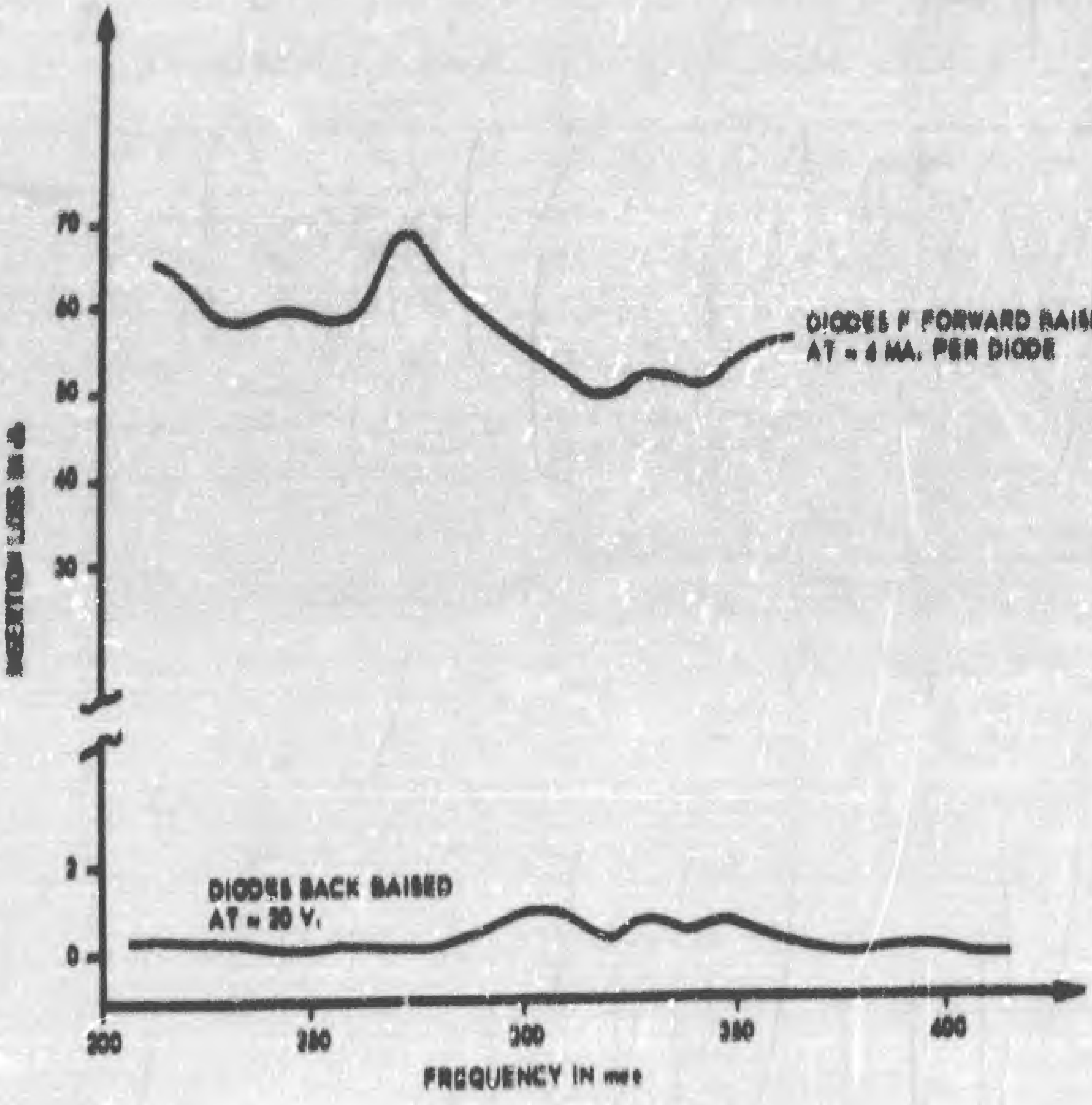

Figure 10. Blanker Insertion Loss Characteristics.

Operation in this manner allows the local oscillator to be switched on and off with a slowly rising and falling waveform, thus avoiding the sharp transients which might be generated with rapid on-off action.

The use of a slow rise and fall in the keying of the local oscillator also reduces the width of the band over which the receiver has spurious responses due to the multiple frequency components of the local oscillator voltage. The following analysis indicates the manner in which these responses arise.

When the local oscillator is being switched on and off in synchronism with a periodic interfering pulse, its output can be written as:

$$e_{LO} = \sum_{N} a_N \cos\left[(\omega_{LO} \pm N\omega_s)\right] \qquad (1)$$

where ω_{LO} = frequency of local oscillator

ω_s = frequency of local oscillator switching voltage.

If an input signal of frequency ω_i is applied to the mixer,

$$e_{input} = A \cos \omega_i t \, . \qquad (2)$$

Then, ignoring any harmonic mixing, the significant terms in the mixer output are:

$$e_{\substack{mixer\\out}} = (A \cos \omega_i t) \sum_{N} b_N \cos\left[(\omega_{LO} \pm N\omega_s)t\right] \qquad (3)$$

where the b_N's are functions of the a_N's and the i-f load impedance presented to the mixer.

Expanding and neglecting sum frequencies gives:

$$e_{\substack{mixer\\out}} = \sum_{N} \frac{Ab_N}{2} \cos\left[(\omega_{LO} \pm N\omega_s - \omega_i)t\right] \, . \qquad (4)$$

The desired output occurs when $\omega_i = \omega_0$, the frequency of the desired signal, giving an i-f output at:

$$\omega_{i-f} = \omega_{LO} - \omega_0 . \qquad (5)$$

However, the values of ω_i given by

$$\omega_i = \omega_0 \pm n\omega_s \qquad (6)$$

also produce components in the mixer output at the i-f frequency, $\omega_{LO} - \omega_0$; this can be verified by substituting these values for ω_i in the expression for the mixer output. Fortunately, the magnitude of these responses falls off rapidly for large n if a slow rise and fall is used in switching the local oscillator, thus eliminating these responses for all practical purposes. An example serves to demonstrate the type of difficulty that might arise if fast local oscillator switching were used.

Assume Switching rate = 1000 cps

Frequency of the undesired signal = ω_0 + 1 mc.

Amplitude of this undesired signal = 80 db above amplitude of the desired signal.

Switching duty cycle = 0.99 .

For a rectangular switching waveform the component of local oscillator voltage at ω_{LO} + 1 mc would be 80 db below the component at ω_{LO}, and the i-f output due to the undesired signal would be of the same amplitude as the desired signal i-f component. At higher switching rates the interference would be even larger.

2.2 Guard-Band Receiver

Pulse interference generally covers a wider band of frequencies than the desired signal, speech for example, and because of this fact a suppression technique can be devised which derives its action from information found in the channels which straddle the desired signal channel. This device has been called a guard-band receiver; it is an auxiliary receiver having a response characteristic which is notched for minimum transfer in the desired signal channel, but has maximum signal transfer in the upper and lower adjacent channels as shown in Figure 11.

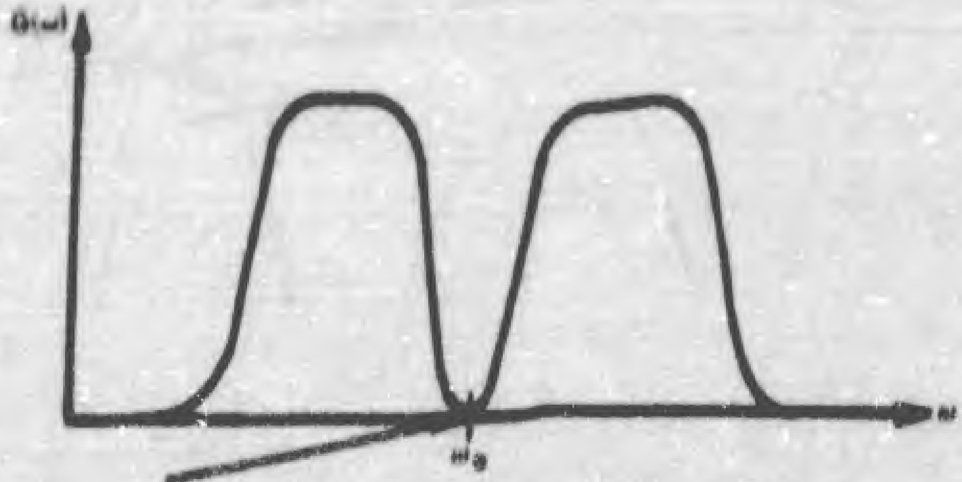

Figure 11. Guard-Band Receiver Response.

This type of characteristic permits the guard-band device to respond only to the interference which exists in the adjacent channels either because this interference is centered on the adjacent channel frequency or is the result of "spill-over from the desired channel. Such response may be utilized to derive gating control voltages that are applied to the desired signal channel.

One manner in which this technique might be applied to reduce the effects of pulse interference is shown in Figure 12.

In this system, both the pulse and desired signals appear at the input to the rejection filter which has a null of transmission to the desired signal. The interfering pulses are not rejected and are passed on to the interference amplifier. This amplifier must have a wide bandwidth to insure that the pulse shape is not distorted. The pulse envelope is detected by the rectifier, and used to switch the gated amplifier to the off condition, so that no signal is transmitted to the second detector during those times when an interfering pulse is present. Thus the signal supplied to the receiver i-f amplifier will be free of the pulse interference.

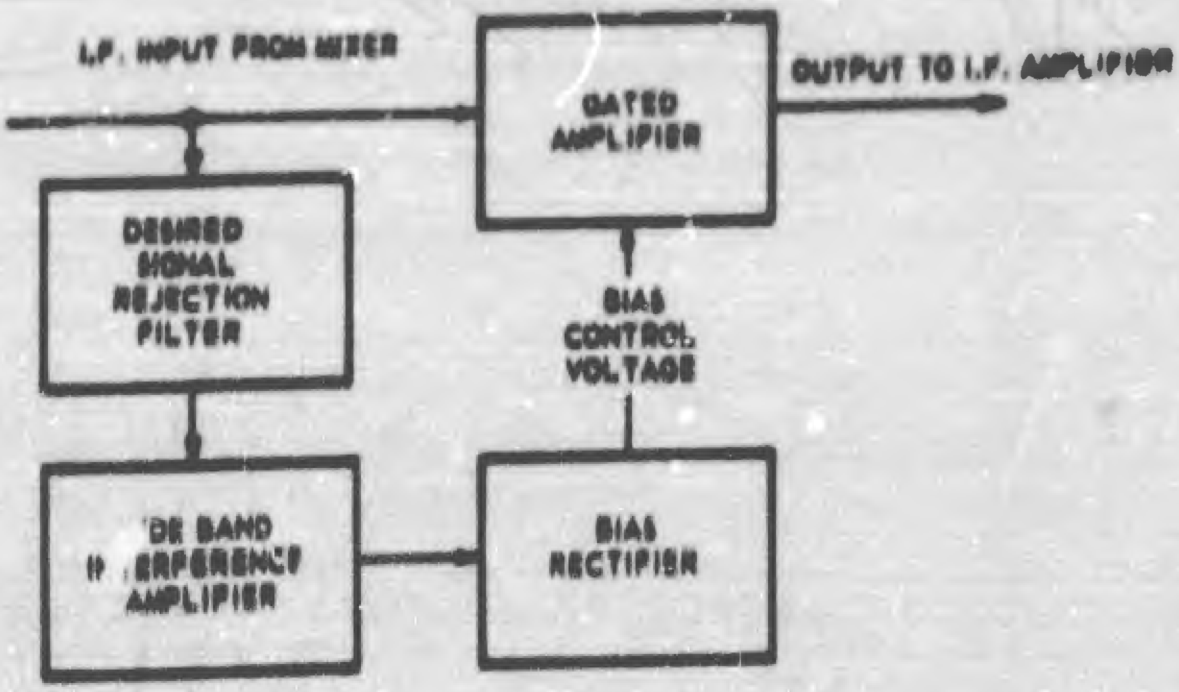

Figure 12. Application of Guard-Band Technique.

An unclassified report describing a device of this type is Reference 2.

2.5 Limiters

In order for an interfering pulse to cause serious interference in a conventional AM communications receiver, the pulse amplitude must be considerably larger than that of the desired AM signal. This wide difference in amplitude between the desired signal and the interference may be exploited to provide considerable protection against pulse type interference. An amplitude limiter that operates at a threshold above the peak amplitude of the desired signal will reduce the interference to signal ratio, thus providing the desired protection. Ideally, it would be desirable to limit the interfering pulse so that the signal to interference ratio would be unity and the desired signal remain undistorted by the limiter action. Various techniques based on this principle have been utilized to combat different types of impulsive noise in communication receivers. These techniques fall generally into two categories: (1) broad band limiting (r-f and i-f), and (2) narrow band limiting (audio channel).

2.3.1 Broad Band Limiters. Recent use has been made of a broad band limiter technique to effect an improvement in the detection of signals in the presence of high level pulse interference. Representative of this technique are the Hoffman Incremental Limiter (4) and other devices. These devices employ wide-band limiting of the input signal. The use of a wide-band signal path before the limiter is essential in preserving the time limited character of the pulse interference. The degree of improvement obtained with the limiting technique depends on the duty cycle and amplitude of the interfering pulses, the largest improvement in signal-to-noise ratio being obtained when the pulse is of low duty cycle and has a large peak amplitude. When the level of the pulse is comparable to that of the desired signal, no appreciable improvement is obtained, but pulse interference of this level does not seriously degrade the reception of the desired signal.

2.3.1.1 Other Devices.

See Volume III.

x.x.x.x Hoffman Incremental Limiter. The Hoffman Incremental limiter[5] is a device which is used with narrow band communication receivers to reduce the stretching of interfering pulses by narrow band i-f circuitry. The way in which this pulse stretching arises can be seen from examining the impulse response of a narrow band tuned circuit such as shown in Figure 14.

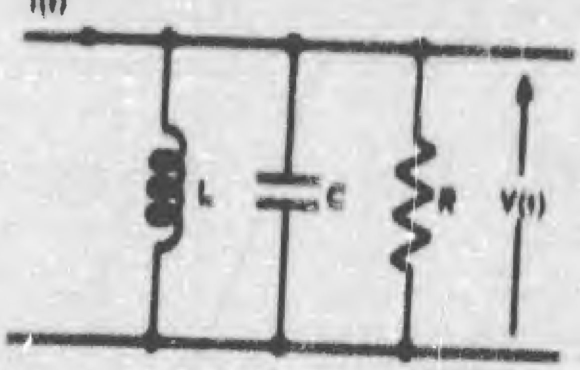

Figure 14. Typical Tuned Circuit.

Although the actual interfering pulses are not true impulses, they are sufficiently short so that the impulse response of the circuit closely describes the response to an actual pulse.

In Laplace Transform notation, the voltage across the circuit is given by:

$$V(s) = \frac{I(s)}{Y(s)} = \frac{1}{C}\left(\frac{s}{s^2 + \frac{1}{RC}s + \frac{1}{LC}}\right), \qquad (21)$$

Then: $v(t) = L^{-1}\{V(s)\} =$

$$\frac{1}{C}\left[e^{-t/RC}\right]\left[\cos\left(\sqrt{\frac{1}{LC} - \frac{1}{4R^2C^2}}\right)t + \frac{1}{RC}\sin\left(\sqrt{\frac{1}{LC} - \frac{1}{4R^2C^2}}\right)t\right]. \qquad (22)$$

A sketch of this function is shown in Figure 15. From this figure it is apparent that the effect of the pulse lasts for a considerable time after the pulse has disappeared. If a weak desired signal is also present, it will be masked for a much larger fraction of time by the stretched pulse than it was masked by the original pulse, and if the repetition rate of the pulse is high or the damping factor, $e^{-t/RC}$, is quite small, then masking may take place for one hundred per cent of the time, and the desired signal will be completely suppressed.

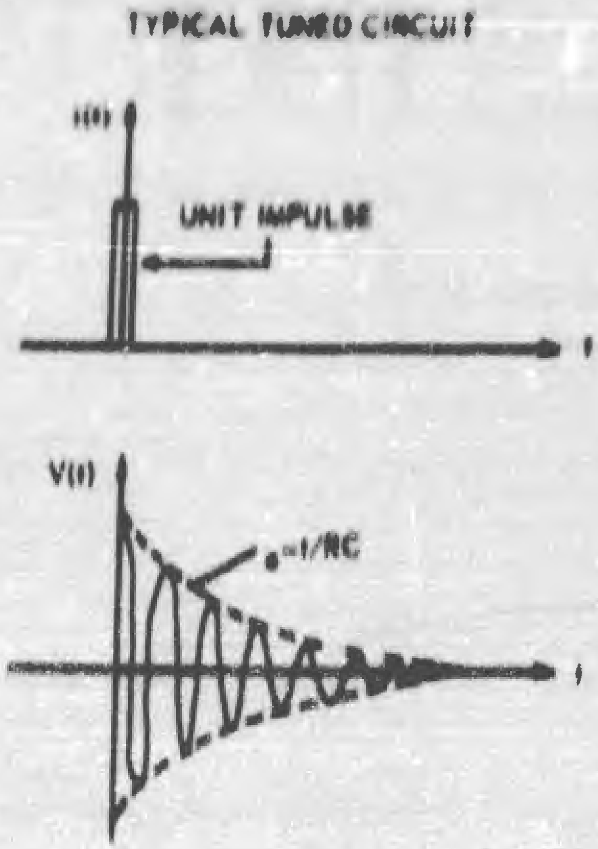

Figure 15. Pulse Stretching Waveforms.

In the Huffman Limiter this pulse stretching is drastically reduced by paralleling the resistance, R, of the tuned circuit by a non-linear resistor as shown in Figure 16.

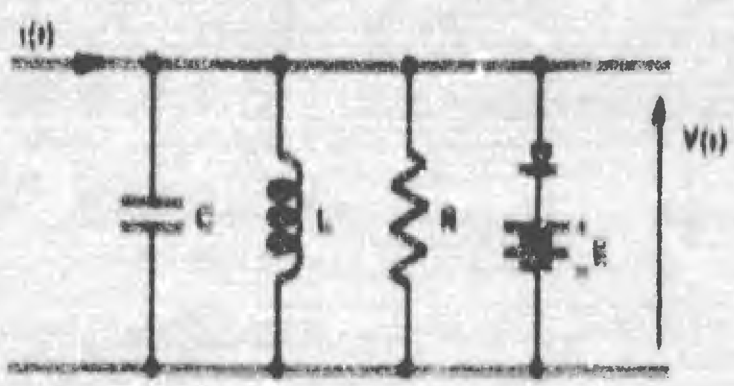

Figure 16. Addition of Non-Linear Resistor to Tuned Circuit.

The value of the voltage E is selected to be equal to the peak desired signal so that the diode does not clip out the modulation on the desired signal. When a large pulse of interference appears across R, the diode conducts, and shunts a low resistance across the tuned circuit for the duration of the interfering pulse. This causes the damping factor, $e^{-t/RC}$, to become very large, and the voltage across the tuned circuit falls rapidly to zero at the termination of the interfering pulse, so that pulse stretching is avoided. The shunting diode also acts as a limiter, thus reducing the peak amplitude of the interfering pulse. If several stages of a narrow band amplifier are modified by the addition of shunt diodes to the several tuned circuits associated with these stages, then by proper choice of the biasing voltages on the shunt diodes, the amplitude of the interference can be reduced to the level of the desired signal, thus providing a considerable increase in the original signal to interference ratio.

2.2.1.3 Comments. In the application of the broad band limiter technique to UHF communications receivers, a fixed limiting level should not be used, since for weak desired signals the clipping might be too high to provide any appreciable improvement in the signal to noise ratio, while for high levels of the desired signal the modulation may be clipped out. Consequently, it is essential that some form of control be provided for the level at which limiting occurs. This could be a front panel adjustment of the bias level of the limiter which is manually set for optimum reception in a given interference situation. If a manual control is not desired, then the AGC voltage might be used to control the limiting threshold. If this type of operation is employed, care must be taken to provide an AGC system which responds to the true average value of the input signal. The conventional diode peak detector is not satisfactory in this respect.

In a conventional UHF communication receiver the limiting should take place immediately following the first mixer, so that pulse stretching in the subsequent narrow band i-f amplifier is not encountered. The r-f portion of the receiver then serves as the wide band signal path preceding the required narrow band path. This is illustrated in Figure 17.

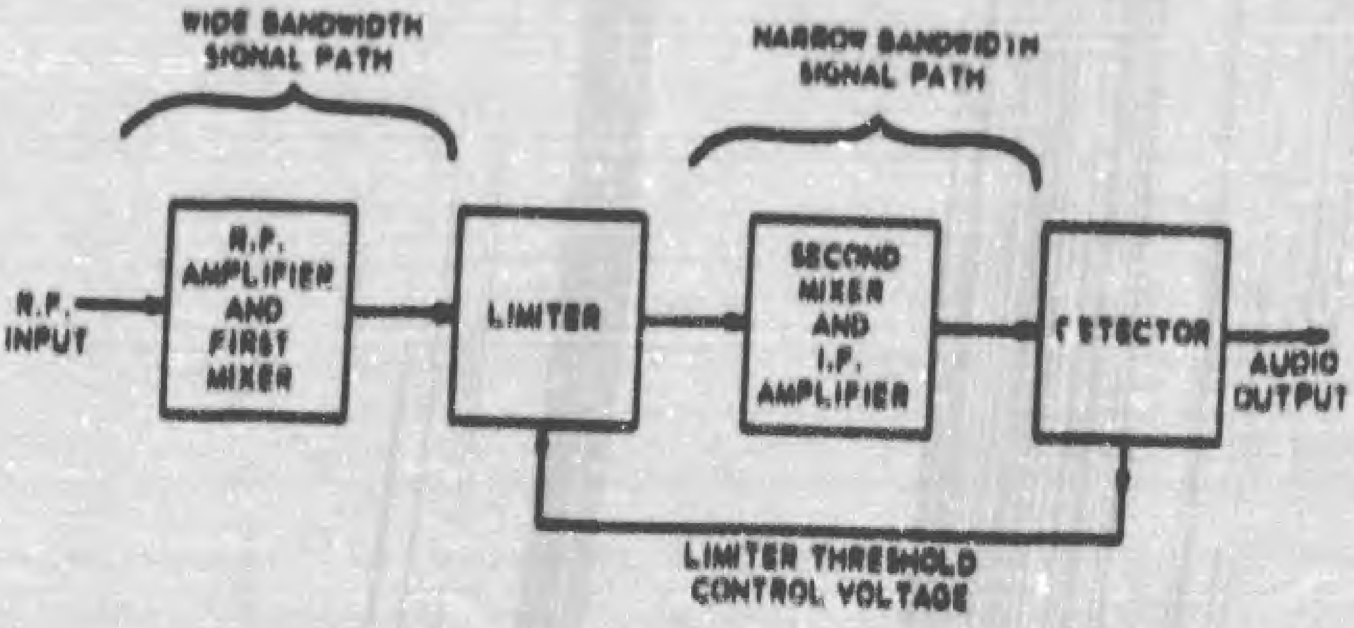

Figure 17. Application of Limiter to UHF Communication Receiver.

2.3.2 Narrow Band Limiters.

Narrow band limiters are distinguished from broad band limiters, described above, by their association with narrow band signal channels, principally, the second detector and audio portions of a receiver. Although there are some schemes which are associated with i-f amplifiers, these operate as auxiliary devices, and posses narrow band characteristics. Many of these limiter techniques have been employed to alleviate random impulse noise or periodic low frequency impulse noise (ignition, commutator) in commercial and military type communication receivers.

Noise limiters in general fall into two groups. One group reduces the amplitude of the signal coincident with the pulse, so that the receiver output is momentarily reduced. The other group functions by limiting the maximum excursion of a signal voltage produced by the pulse to a value which is not appreciably greater than some predetermined level, and are usually called output limiters.

Each of these groups usually operate with a delay bias that keeps the threshold of operation slightly greater than the peaks of the desired signal modulation. When changes in signal level are encountered, adjustment of the limiter threshold is necessary for most effective action of its characteristic. This adjustment may be accomplished by either manual or automatic means.

Many types of noise and output limiters are found[2], but the series and shunt type limiters predominate in use. For their simplicity these are highly effective devices in reducing the annoyance of impulse noise, and are widely used in communication receivers.

A typical second detector circuit with series and shunt type noise limiters is shown in Figure 10. The second detector, V1, supplies audio to the series limiter, V2, which is continuously conducting under average signal conditions. A large peak of interference drives the plate of the limiter negative with respect to the cathode, and thus cuts off the tube for the duration of the interference pulse. The threshold of the clipping may be determined by adjustment of the cathode voltage, and is governed by the modulation percentage of the signal.

There is some attenuation introduced by the series limiter circuit on the audio signal out of the second detector, and this means that greater gain in the audio amplifier section is required to restore the output to normal level.

The operation of the series limiter is enhanced by the addition of a circuit called a "noise cancellation detector." Noise energy remaining in the output of the series limiter diode can be further reduced by the introduction of noise pulses of about equal amplitude, but of opposite polarity to produce noise cancellation. These cancellation pulses are generated by the noise cancellation detector shown as V4 in Figure 10. The circuit is

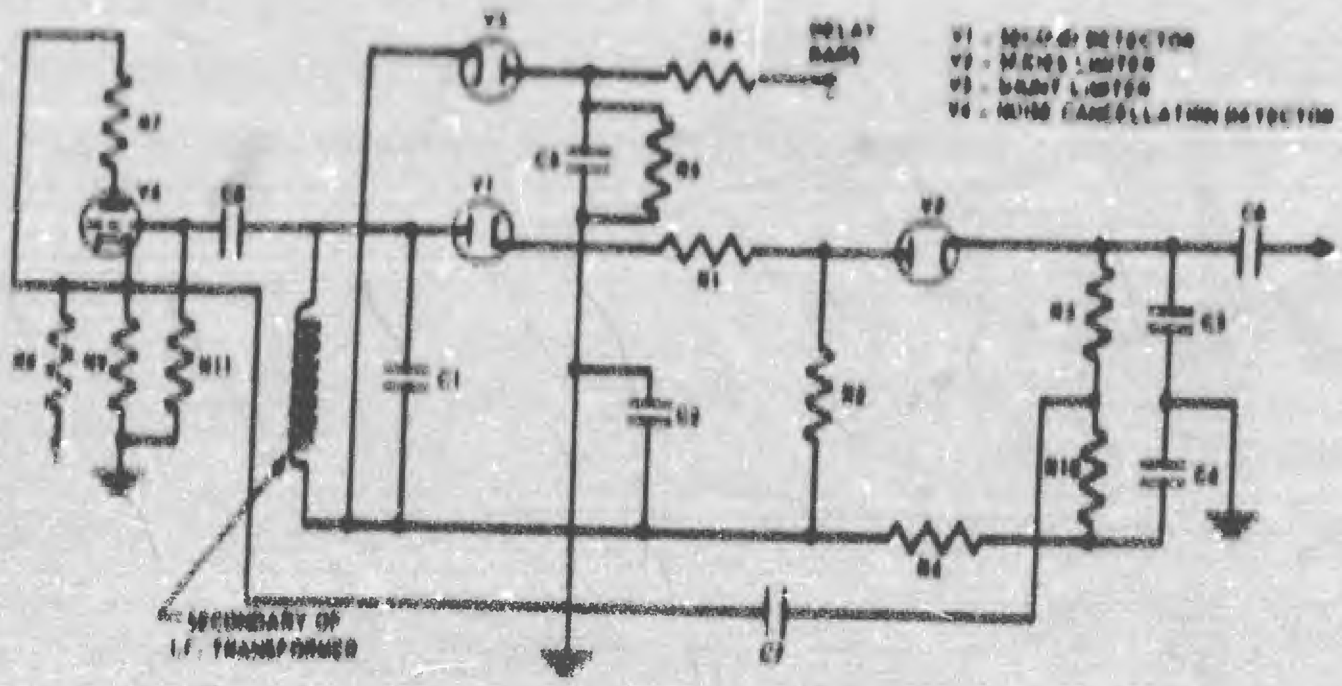

Figure 18. Second Detector with Series and Shunt Diode Noise Limiter and Cancellation Detector

an infinite impedance detector in which the detected signal appears at the cathode. This signal is coupled back to the output of the series limiter of the detector when the i-f voltage is below some predetermined value.

The shunt diode limiter V3 is connected so that a large negative excursion of i-f voltage will cause conduction and place a low impedance across the terminals of the i-f transformer, so that the i-f signal output to the second detector is momentarily reduced by heavy loading of the diode. The shunt limiter is designed primarily to prevent operation of the AVC and squelch circuits by interference pulses.

Another technique which has been used to some extent in suppression of pulse interference is one which operates from the i-f voltage picked off at the output of the mixer. This is the principle used by the Lamb noise silencer circuit[6] that is popular among amateur radio operators. A typical circuit is shown in Figure 19.

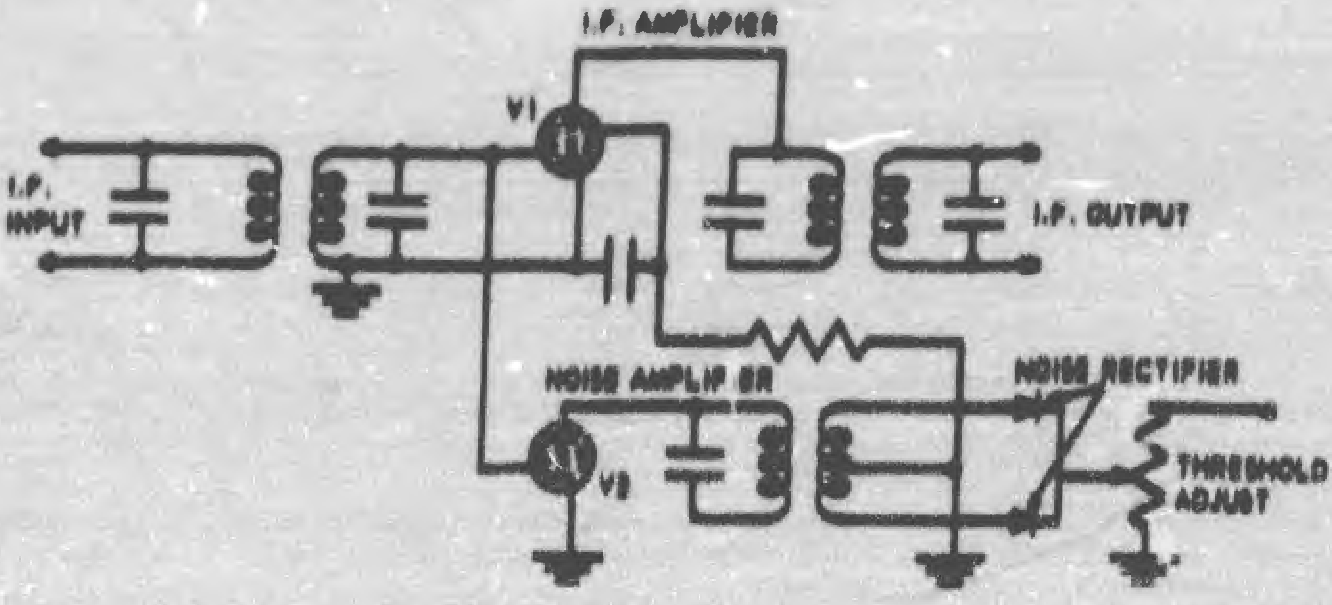

Figure 19. Basic Circuit of I-F Impulse Noise Suppressor.

The noise amplifier, V2, picks off a voltage at the input to the first i-f amplifier, V1. The output of the noise amplifier is rectified by a full wave circuit, and this negative dc voltage is applied as bias to the suppressor grid of the i-f amplifier. If the rectifier cathodes are biased such that normal signals will not produce an output, then the i-f amplifier will operate with normal gain. When a pulse signal of appreciable amplitude appears, a negative pulse is produced on the suppressor grid of V1, reducing its gain. This negative pulse may be applied to several such stages to effect a greater signal handling capability, which is necessary for extremely large pulse amplitudes.

By applying the i-f noise limiter control voltage to early i-f stages, the resonant circuit ring down (pulse stretching) caused by steep front pulses is reduced, thus minimizing signal masking in the receiver output.

Since the i-f noise limiter reduces the gain of the i-f amplifier during the pulse incidence, a discontinuity appears in the envelope of the signal which amounts to actual modulation of the signal by the pulse. This is heard in the output as a tone, but at a level comparable to the signal. This tone may or may not annoy the operator in understanding the desired signal, but in case it is desired to eliminate the tone, an audio filter such as the one described in Section [illegible].4 may be used.

Experiments were conducted on two receiver equipments designed for operation in the UHF range. These were the R-361 (GRR-7) and the R278-B (GRC-27). In these tests the objective was to determine the effectiveness of the noise limiter in the specific equipment on hand and under various signal conditions. Signal sources used were a voice modulated Hewlett Packard model 608C UHF generator and a laboratory constructed 10 watt UHF pulsed oscillator. The pulse repetition rate was 1,000 cycles per second, and the pulse width was 10 μ sec. The desired signal level to the receivers was fixed at 10 μ volts, and the pulse signal was adjusted to a value for each receiver determined by the overload characteristic of the early r-f and i-f stages.

The R361-A uses a series noise limiter, but does not have the shunt limiter circuit. Provision is made in this receiver to switch the series limiter in or out of the circuit according to the operating requirement. The R278-B receiver has both shunt and series noise limiters in addition to a noise cancellation detector. This receiver has no manual controls that directly affect the operation of any of these circuits which means they are active under all operating conditions.

Initial tests were conducted on each receiver to determine the effectiveness of the series limiter circuit in reducing the effect of pulse interference on the desired signal received in the output. Two cases of desired signals were investigated, one using a 400 cycle per second tone modulated CW and the other using voice modulated CW. Only the voice modulated tests are reported, since data on tests using tone modulation were not reliable or conclusive.

With limiter out the interference level was adjusted until approximately 50 per cent of the message was understood. The level of interference was noted to be one volt peak (.02 watts across 50 ohms). Switching the limiter back into the circuit permitted the operator to understand approximately 70 per cent of the message. Adjustment of the clipping level made no significant improvement in the intelligibility of the message. Going to higher input levels, a point was observed at which the signal was barely understood (10 per cent or less articulation score). At this point it was noted that the limiter gave no improvement, and it was found that the receiver was being blocked by the high signal level. Shorting out the AGC line at the point where it enters the r-f chassis permitted the signal to be detected, and its intelligibility to increase to about 50 per cent. Switching in the limiter increased this figure 10 per cent. These results appeared to be significant enough to demonstrate the amount of improvement obtained by the limiter. Articulation Tests would be required for quantitative data.

2.4 Pulse Suppression with Audio Filter

Several techniques have been discussed which are effective in reducing or eliminating the effects of pulse interference. In some cases, especially when the level of interference is very high, there remains some residual interference in the audio output of the receiver. In these situations a properly designed audio filter can greatly reduce the annoyance arising from residual tone, and considerably increase the readability of the desired signal. The desired transfer function of the audio filter is shown in Figure 20. The success of this method of filtering depends on the fact that the interference is periodic with period $\frac{2\pi}{\omega_0}$. In this case, all the frequency components of the pulse fall at the nulls of the filter response, and no pulse appears in the output. The narrow rejection bands have no appreciable effect on the desired speech signal. The width of the nulls are made broad enough to accommodate small variations in the repetition rate of the pulses, so that for a given nominal repetition rate, there is no need to readjust the filter.

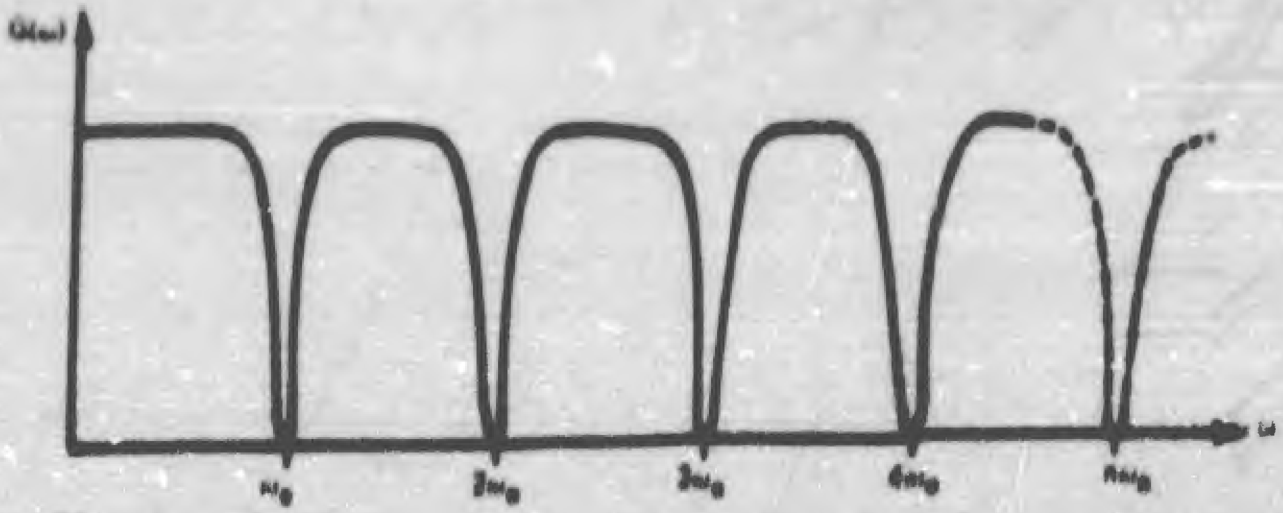

Figure 20. Audio Filter Response.

Such a filter was constructed using a series of bridged T filters tuned to the fundamental, and several harmonics of a 1,000 cps pulse interference. The filter passed no frequencies above 4,000 cps, and hence had a sufficient bandwidth to pass the desired audio signal. Using this arrangement, an increase of about 20 decibels in the tolerable pulse interference level was obtained. If many different pulse rates are encountered, it becomes necessary to make the filter tunable. This requirement complicates the filter design considerably, but such a filter can be constructed. One approach might be to build a series of bridged T or twin T null networks whose tuning is tracked, but this would be difficult to accomplish. Another method of constructing the desired transfer function would be to use a feedback amplifier whose feedback network contains a delay line. This line would have a variable delay which would be set equal to the period of the pulse interference. It is shown in Appendix II that such an arrangement has the desired transfer function.

It should be reiterated that an audio filter alone does not constitute a solution to the problem of pulse interference, since overloading effects ahead of the audio system may seriously degrade the desired signal response. Rather, the audio filter serves as a useful adjunct to other techniques in that it reduces the interference remaining after other techniques have been applied.

4.3 Desensitization Reduction

Receiver desensitization is a condition produced by strong undesired signals appearing in the receiver signal channel and capturing the AGC. The result is that the sensitivity of the receiver to the desired signal is greatly reduced.

Domination of the AGC operation by the undesired signal comes from the response of the AGC detector to excessive carrier level of this signal and from grid circuit rectification in the early r-f and i-f stages that cause charging of the AGC line. Additional effects that produce an apparent desensitization are signal clipping that occurs when the desired signal rides the crest of the large undesired signal and is forced into the grid cutoff, or plate-saturation region of the amplifier tube characteristic. This condition would be noticeable where the undesired signal consists of a strong CW carrier or modulation sidebands lying somewhat above the audio region, but within the channel bandwidth of the i-f stages. However, this condition is rather rare and was not investigated in the desensitization problem of the receiver. The principal types of undesired signals which produce receiver desensitization are CW carrier (cowed), unmodulated FM carrier, intelligence-bearing sidebands, and pulsed carriers. Desensitization resulting from pulsed type signals was the principal situation studied in the work reported here. These signals are encountered in communication installations in radar sites where strong fields are created by high power r-f radiation or modulator emissions. The nature of a radar pulse signal is that it is of short duration and of high peak power. Generally the duty cycle of these signals is very low (around 1 per cent) and therefore the average power is quite small in comparison to the peak power. The ideal AGC circuit should respond only to the average power of the signal and be insensitive to the peak power. It was the desirability of producing this kind of AGC response that guided the selection of remedial techniques investigated in the project. These techniques attempt: (1) to alleviate the condition of AGC line charging by rectification, and (2) to obtain an AGC characteristic required for proper receiver performance.

[illegible] AGC Line. A common condition found in most communication receivers is that capacitors associated with circuitry in the signal path may, with the incidence of strong signals, charge rapidly by grid rectification through a low resistance (short time constant) path and discharge slowly because of a high resistance (long time constant) discharge path. Thus in the case of AGC circuitry associated with r-f amplifiers, strong signals cause grid rectification and charge the AGC line, but since this line is normally designed to have a very long time constant discharge characteristic (low pass filter) a large voltage is developed across the grid capacitors and biases the tube into the region of cutoff and remains in that condition until the capacitor charge is reduced. Several stages in tandem having these characteristics can produce complete blocking of the receiver, and reduce the desired signal to zero in the output.

UHF communications receivers generally utilize a grounded grid amplifier configuration for the r-f front end, at least in the initial stages. Grid rectification occurs when the cathode is driven in the negative direction by the signal. Figure 21 shows a typical circuit of that type of amplifier improved by the addition of a crystal diode connected across its input terminals.

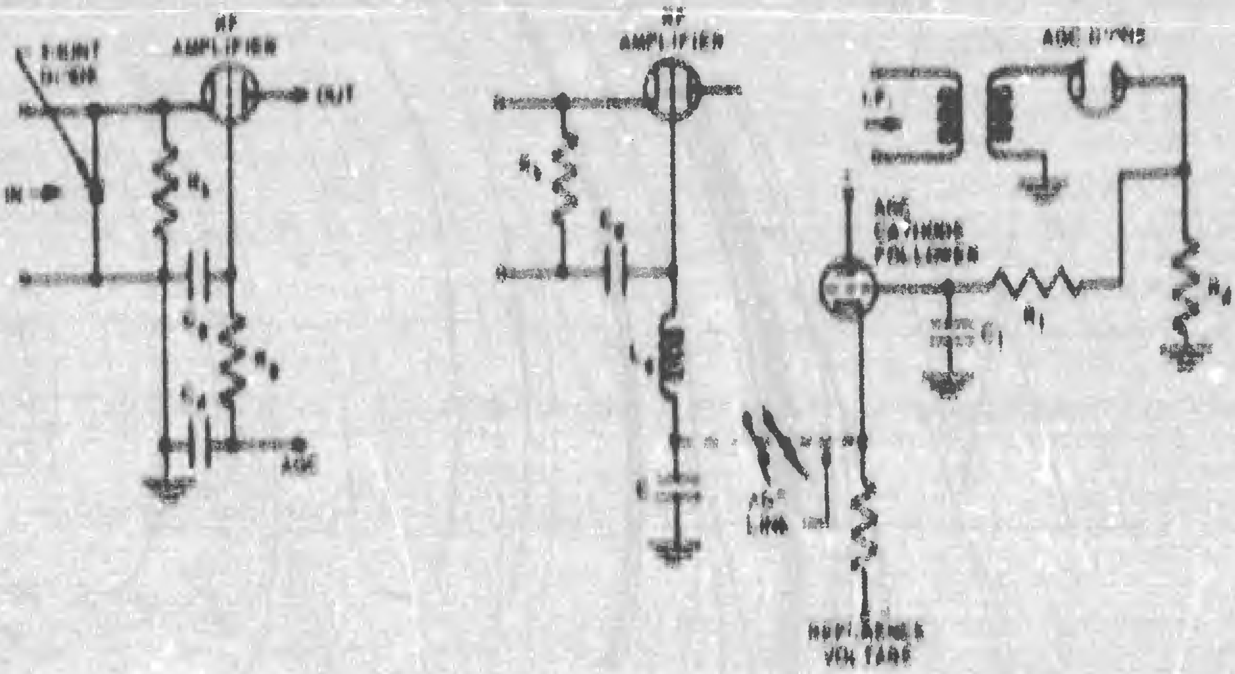

Figure 21. AGC Schemes for R-F Amplifier Circuits.

The crystal is connected so that a large negative excursion of the signal will cause it to conduct and essentially short circuit the input for the duration of the excursion. This prevents the signal from driving the cathode below ground potential, and hence becoming negative with respect to the grid. The normal operating cathode bias of the amplifier prevents the crystal from conducting under normal signal conditions. This technique does not prevent the cathode from being driven highly positive by the positive portion of the signal, but this condition only causes the tube to be driven into cutoff and results in the signal being clipped in the amplifier output and the AGC line is not disturbed. Type 1N34 crystal diodes were tested across the terminals of the grounded grid amplifiers of the r-f amplifier section of the R301-A. Applying a high level pulse signal to the input before the diodes were connected produced a voltage across the grid terminals of the second stage of about 3-1/2 volts negative. Adding the diodes to the amplifier inputs dropped the voltage to about .1 volt negative, indicating that some grid rectification was occurring. The pulse amplitude at the input to the second stage was about 10 volts peak and 10 microseconds long.

To reduce the time constant for discharge of the grid capacitor, C_g in Figure 21 , a low resistance to ground must be provided. This may be accomplished by using a low impedance driving source for the AGC line, such as a cathode follower. Isolation between the several stages that are connected to the line may be provided by using inductance elements which have a high ac impedance but low dc resistance. The desired time constant characteristics for the AGC circuit are provided by $R_1 - C_1$ in Figure 21 . No tests were made to evaluate this technique. It would require extensive modification of the AGC circuit, but might be well worth-while.

[illegible] AGC Detector. A typical means of generating the AGC voltage in the usual communications receiver is by rectifying the i-f output with an envelope detector. A voltage is desired that is proportional to the average value of the desired signal carrier. Therefore, an ideal AGC voltage generator is one which produces a true average of all signal carriers impressed upon it independent of the modulation envelope. However the envelope detector represents a peak type detector, since the storage of charge on the filter input capacitor develops a voltage that is proportional to the peak of the i-f voltage. The time constant of discharge of this capacitor is determined by the resistance of the RC filter, and in a typical receiver this may amount to a hundred microseconds or more. The R361-A, for instance, has a time constant of about 500 microseconds, and to obtain a true average of a signal carrier the modulation should be no higher than 400 cycles per second. Signals containing modulation components appreciably higher than this would generate AGC voltages proportional to the peak value of these components.

Figure 22 shows the envelope of a one kc sine wave modulation on the i-f carrier, and a 1,000 cps pulse with a duration much less than the period of the sine wave applied to a typical detector circuit.

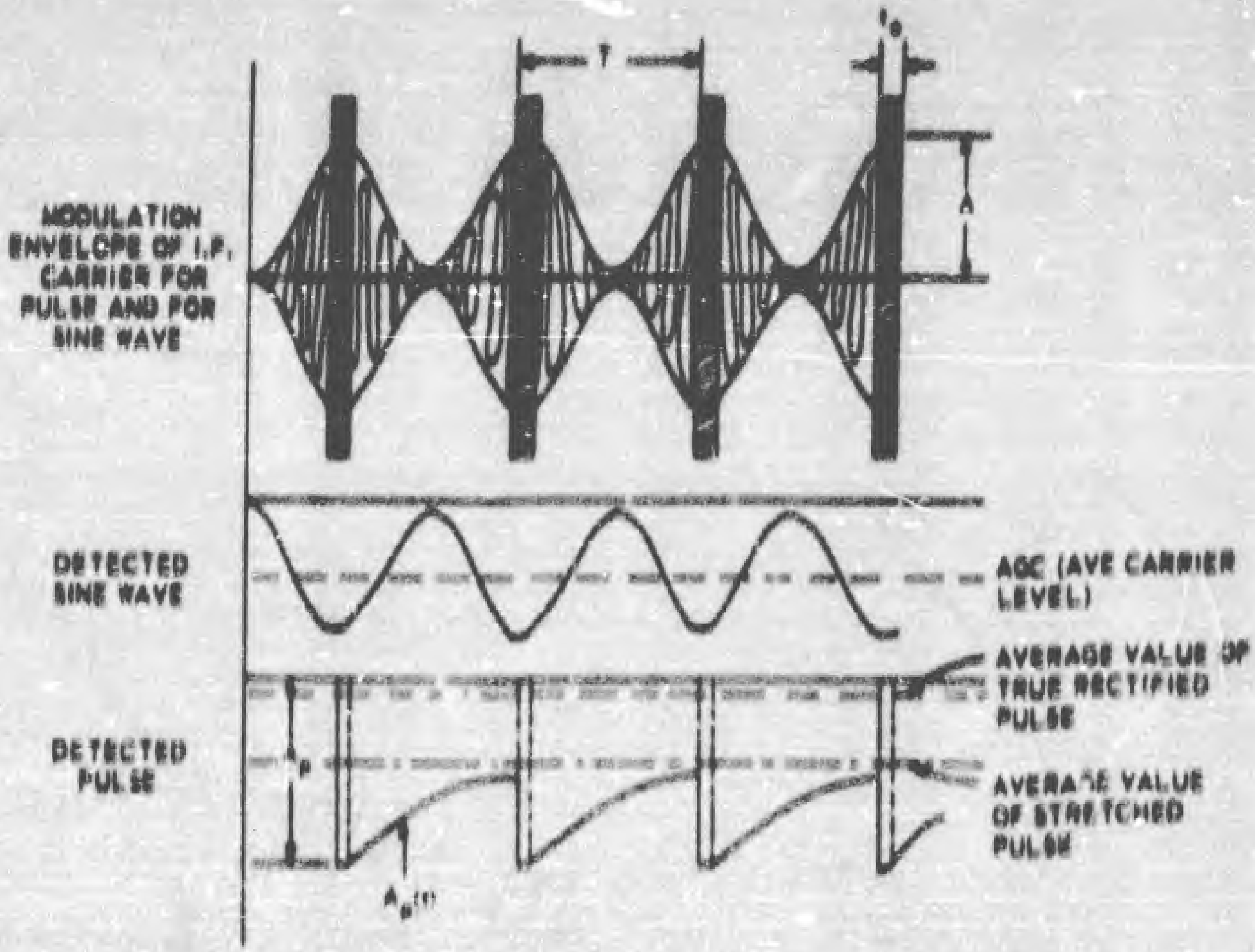

Figure 22. Output Characteristics of Typical AGC Detector for Two Types of Signals

The sequence of detector operation is shown for both cases. For the sine wave it can be seen that an average value has been established corresponding to the average carrier level. In the case of the pulse signal, detection results in faithful response to the leading edge, but the long time constant of the RC filter stretches the trailing edge and the average voltage that is established is greater than the average that would be obtained from consideration of the true waveform of the pulse alone.

From Figure 22 for the sine wave case, E_{AGC} is A/2 of the envelope for 100 per cent modulation and for the pulse E_{AGC} is ideally, $A\frac{\tau}{T}$. If, however, the time constant of the discharge path is considered, then the average value of the detected pulse is

$$\bar{E}_{ave} = \frac{1}{T}\int_0^{t_1} A_p(t)dt \ .$$

It is significant to compare the AGC voltage generated by a stretched pulse with that generated by a true pulse. Figure 23 shows the detected pulses sketched in Figure 22, but on an expanded scale.

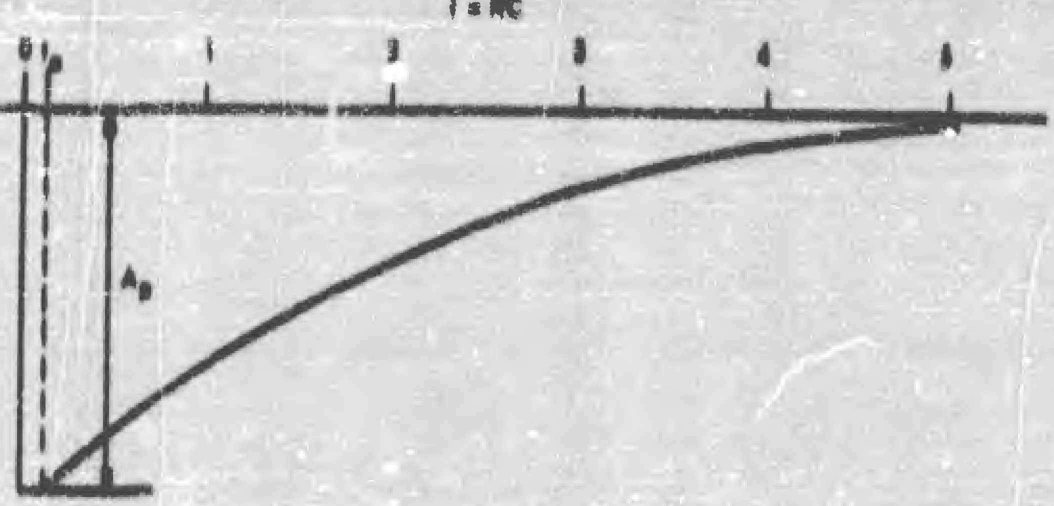

Figure 23. Characteristic Response of RC Filter for AGC Rectifier when Pulse Is Applied.

For the stretched pulse the average value,

$$\bar{E}_{sp} = \frac{1}{T}\int_0^{t_1} A_p(t)dt \qquad (24)$$

where

$$A_p(t) = A_p e^{-t/RC} \ ,$$

Substituting

$$\bar{X}_{sp} = \frac{1}{T}\int_0^{t_1} A_p e^{-t/RC} = \frac{A_p}{T}\left(\frac{e^{-t/RC}}{-1/RC}\right)\Bigg|_0^{t_1} . \tag{25}$$

If t_1 = 5RC then,

$$\bar{X}_{sp} \approx \frac{A_p RC}{T} . \tag{26}$$

The average value of a rectangular pulse is,

$$\bar{X}_p = A_p \frac{t_p}{T} . \tag{27}$$

The ratio of the two average values shows the relation between the AGC voltages produced in each case:

$$\frac{\bar{X}_{sp}}{\bar{X}_p} = \frac{\dfrac{A_p RC}{T}}{\dfrac{A_p t_p}{T}} = \frac{RC}{t_p} , \tag{28}$$

Since RC represents the time of the discharge path, a practical example can be given to show how much the AGC voltage would be reduced from the stretched pulse condition if the average of the true pulse were considered. Using the RC values found in the R361-A receiver, (R = 100 K ohms; C = 5,000 mmf),

$$\bar{X}_p = \frac{t_p}{RC}\bar{X}_{sp} = \frac{5}{500}\bar{X}_{sp} , \tag{29}$$

which indicates that $\bar{X}_p$ would produce an AGC voltage only one per cent of the stretched pulse value. The technique suggested here is to make the AGC detector a true averaging rectifier. Figure 24 gives a typical circuit which would accomplish this operation.

It consists of an AGC detector with load R_d connected to a lowpass filter, $R_f C_f$, and AGC cathode follower. The fact that the voltage produced is proportional only to the average value of the i-f carrier instead of the peak value may supply less voltage for AGC control than is required, and therefore an AGC amplifier before or after the AGC detector will be required. The time constant of the AGC filter utilized in this circuit is the same as in any typical AGC design. To provide an AGC output that is zero when the signal input is zero, the cathode load resistor is returned to a negative reference voltage.

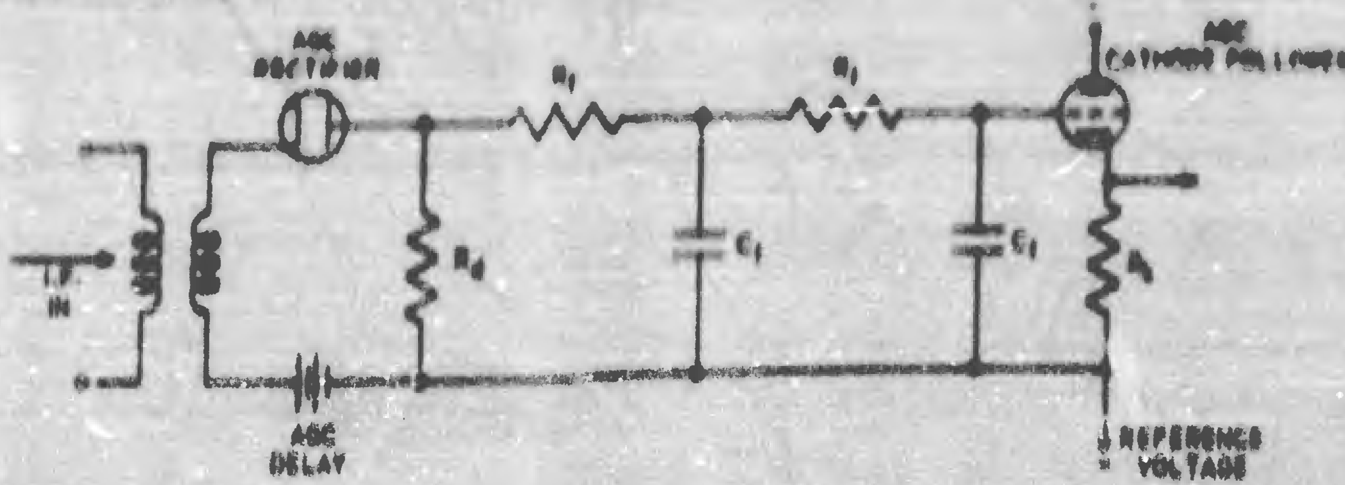

Figure 24. Modified AGC Circuit.

An investigation of the feasibility of this technique was made on the R361-A UHF communication receiver, since tests and experience showed it to be particularly susceptible to desensitization by pulse signals. Figure 25 shows a circuit of the modification made in the AGC detector section. A bridge rectifier was utilized in order to provide a higher AGC voltage than would be obtained in a half wave rectifier configuration. The same load resistance value was used as was used in the regular AGC circuit, but the filter capacitor, C-344, was omitted.

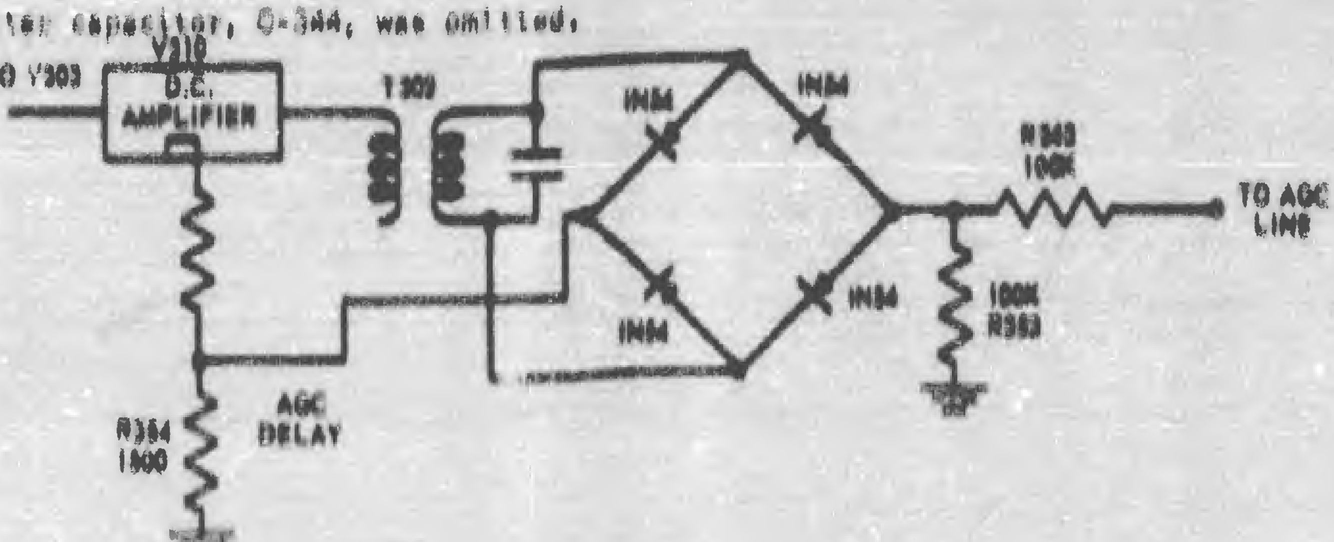

Figure 25. Modification of AGC in R361-A Receiver.

The conditions of the test were set up first to establish, with unmodified AGC, the value of the interference at the input which would suppress the desired signal until it was no longer understandable, and could hardly be dete[illegible]ed in the output. The desired signal input was 6 microvolts. Making the modifications permitted the interference to be increased from 15 to 25 decibels before the signal was again suppressed. It was felt that this degree of improvement was significant in demonstrating the effectiveness of an average value rectifier in reducing the desensitization caused by pulse signals.

A study of the R-278-B receiver showed that the AGC detector circuit in this unit already had a sufficiently short time constant.

2.6 Miscellaneous Techniques

This section includes a brief discussion of several suppression schemes which were conceived during the project, but which were given only a very preliminary investigation.

2.6.1 Voltage Sensitive Capacitors. The characteristic of the capacity across the junction of a semi-conductor diode is represented by the function in Figure 26. A typical Q value for these devices is about 30 at

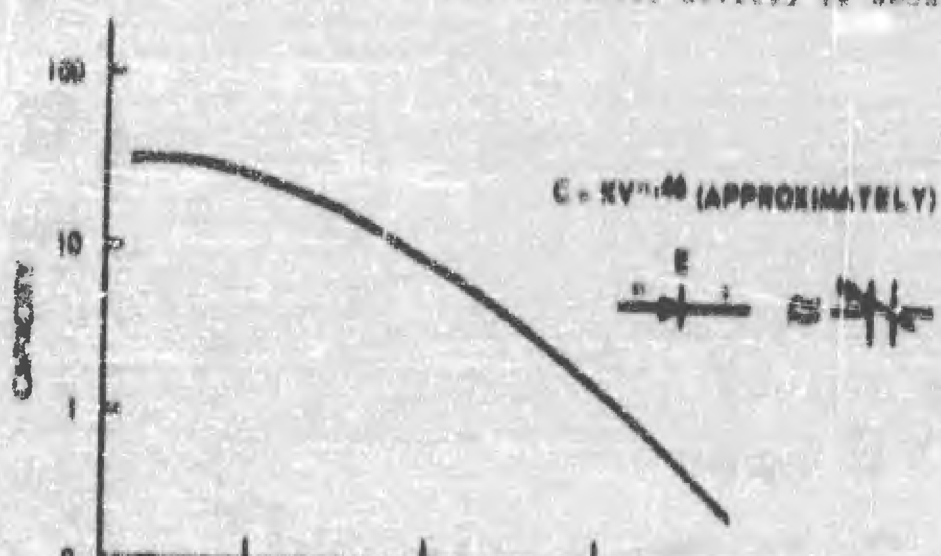

Figure 26. Typical Characteristic of a Crystal Diode Capacitor

50 mc/sec and is adequate for i-f circuit application. Such a device suggests a few techniques that may be useful in combating interference effects.

Consider the i-f amplifier circuit in Figure 27 with voltage sensitive capacitors, C_{v1} and C_{v2}, placed in the plate and grid circuits respectively. The tuned frequency of each circuit may be affected by the application of a step voltage to each diode capacitor.

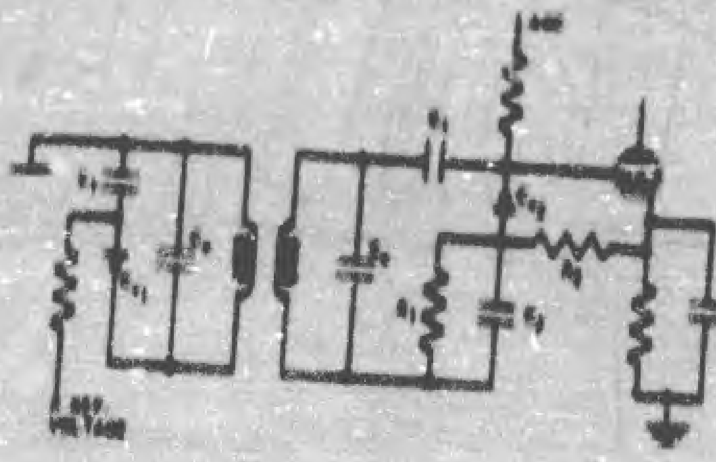

Figure 27. Application of Voltage Sensitive Crystal Diode Capacitors in I-F Amplifiers.

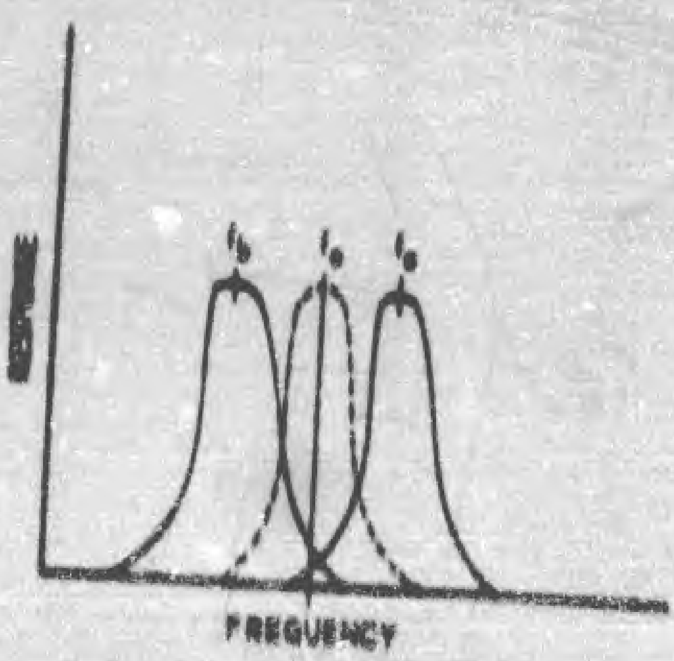

Figure 28. Theoretical Response Characteristics of Tuned Circuits Using Voltage Variable Capacitor Diodes.

Thus, if Figure 28 represents the normal response of the amplifier interstage network at f_o, then the application of a voltage to the capacitors will cause a shift in the tuned frequency depending upon the polarity of the applied control voltage. The new response may be above or below as desired. By applying a control voltage of one polarity to the plate circuit diode and a voltage of opposite polarity to the grid circuit diode, an opposite shift of one response with respect to the other can result so that the net response is one which has a deep slot at the desired frequency f_o. This situation is more desirable, since synchronous shifting of the response in one direction may place it on an adjacent channel or other interference. In circuits where large ac voltage excursions may occur, there would result an automatic detuning action that could be used to good advantage.

Application of capacitor diodes at UHF has not been thoroughly explored, nor do the existing market types necessarily represent what may appear in the next few months, since this device is relatively recent in appearance.

<u>2.6.2 Pulse Interference Reduction by Sampling</u>. Any device which reduces the effects of an interfering signal must depend for its operation upon recognition of some basic difference in the character of the desired and interfering signals. In the case of a pulse train interfering with an amplitude modulated carrier, one such difference that may be exploited is the fact that the pulse is time limited, while the desired signal is not. One means of utilizing this difference in the character of the two signals is indicated in Figure 29.

Since the desired signal is not time limited, a sample of the desired signal appears at the sampler output for every sample pulse, and if the rate of these samples is sufficiently high, the filter output will be a replica of the desired signal. However, due to the time limited character of the pulse interference, a sample of the interference appears in the sampler output only when the interfering pulse is in time coincidence with one of the sample pulses. Hence there may be fewer interference pulses per unit time in the output of the sampler than at the input to the sampler, and a reduction in the amount of interference is obtained. If the time of occurrence of the

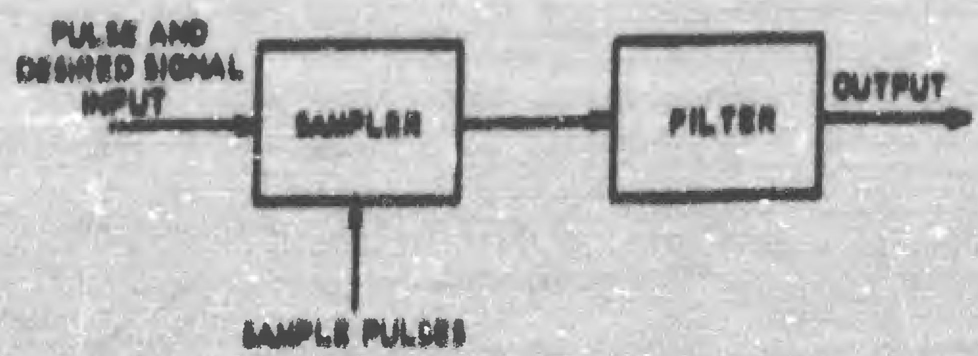

Figure 29. Sampling Filter.

sample pulses is made statistically independent of the interfering pulses, then the average rate of interfering pulses in the sampler output may be computed from the probabilities involved.

The improvement obtained with such a sampling filter is twofold. First, a signal to interference ratio improvement is obtained; and second, a reduction in prf of the interference in the outp. over that of the input. The lower rate results in less effect on the intelligibility of the desired signal, even signal to interference ratio improvement had been obtained.

The ne insuring that the time of occurrence of the sample pulses be stati ally independent of that of the interfering pulses arises from the fact that, if a constant sampling rate were used, the samples might fall in step with the interfering pulses, and no interference reduction would take place. The necessary statistical independence of the sample pulses can be obtained by jittering the sample pulse rate in accordance with some random function, such as the output of a noise generator.

The use of a sampling filter in a communication receiver is illustrated in Figure 30.

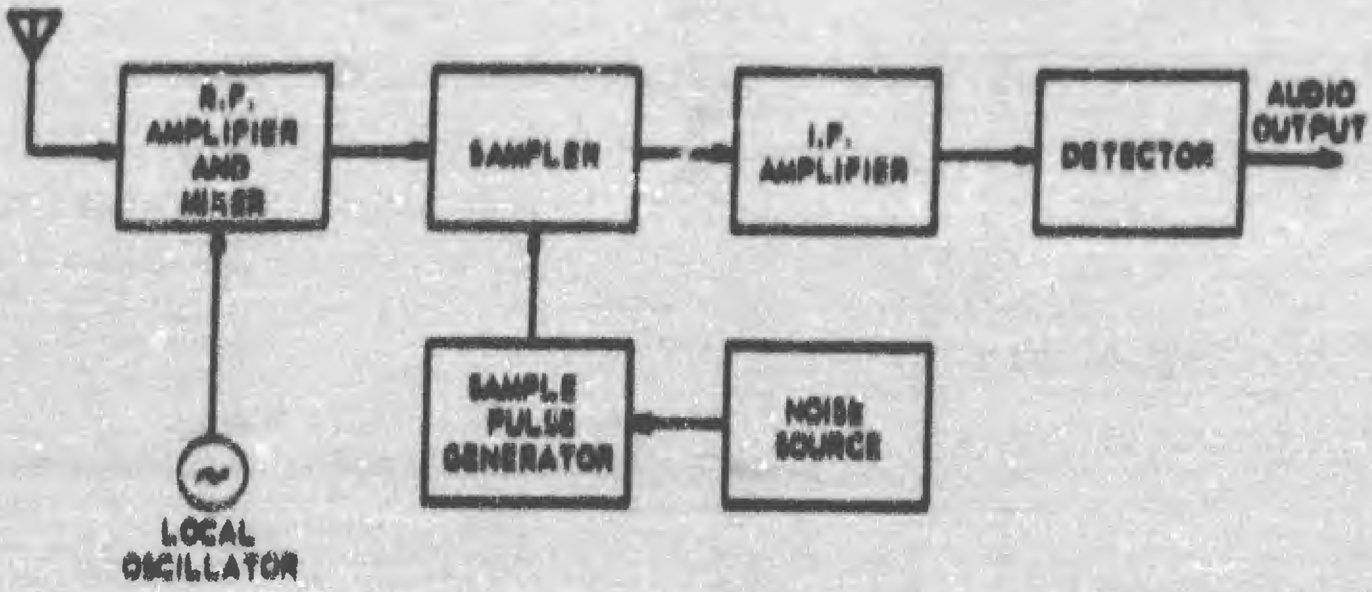

Figure 20. Application of Sampling Filter.

The sampler is placed ahead of the i-f amplifier to avoid stretching of the interfering pulses before they are sampled. The i-f amplifier with its band pass selectivity characteristics serves as the filter to reconstruct the desired signal from the sampler output. For the reception of speech modulated signals, an average sampling rate of 25 kc jittered from 20 kc to 30 kc might be used. The use of a sample width of 0.5 microseconds would give a sampling pulse duty cycle of 0.01, so that for most pulse interference situations a reduction in interfering pulse rate by a factor of 50 to 100 would be obtained.

[illegible]. Conclusions

Pulse interference can be grouped into two levels of severity, high level and low level. High level pulse interference will overload early stages of the receiver and result in blocking of its operation. Low level pulse interference does not overload early stages, but may produce desensitization and an annoying audio tone in the output of the receiver.

Pre-receiver blanking devices and limiters in the r-f front end are the techniques which are effective in combating high, as well as low, level pulse interference. The most effective one of these two schemes is the pre-receiver blanker (either narrow band or broad band). However, the operation of the blanker is dependent upon synchronizing triggers being available from the interfering source. If more than one source is troublesome, then synchronizing voltages must be obtained from each one. Auxiliary receivers may be used to detect and generate synchronizing voltages, but this method causes some deterioration of the blanking operation because of signal delay in the receiver circuitry. Limiting devices in the r-f front end of the receiver prevent overloading of these stages, and reduce the severity of overloading in succeeding stages, but are not as effective as blankers.

Low level pulse interference may be alleviated by techniques which are attached to, or built into, the internal circuitry of the receiver. These techniques include broad-band and narrow band limiters for i-f circuitry; i-f sampling, audio limiters, and AGC techniques. The most effective one (or class) of these techniques is the broad band limiter. It not only restricts the amplitude of the pulse interference, but also minimizes pulse stretching that has adverse effects on AGC operation. The sampling filter technique may be a good competitor for the broad band limiting technique on the basis of preliminary test and theoretical prediction. Further evaluation is necessary before a definite conclusion on this point can be reached. Improvement of the AGC circuitry is perhaps next in the line of effective pulse suppression devices. Desensitization effects can be reduced by fairly simple modifications in AGC design. Series and shunt audio limiters do very little to relieve the desensitization problem, but are effective in reducing the masking of the desired signal intelligence by pulse interference (when the receiver is not desensitized).

In each of the above techniques, whether dealing with high level or low level interference, the use of an audio filter may be desirable to suppress the audio tone that still exists either as residual interference or as a by-product of the suppression technique utilized earlier in the receiver. [Where high speed blanking or sampling rates are used, the by-products are above the audio range (or out of the audio bandwidth of the receiver) and no filter is required.]

4. Bibliography

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(7) Sylvania Electric Products, Incorporated, Anti-Jamming Techniques Study Final Report, Contract No. AF 30(602)-1593, October 30, 1957, ASTIA Document 140671; (SECRET).

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[illegible] Appendices

APPENDIX I

SCHEMATICS OF THE BLANKING PULSE GENERATOR AND ITS POWER SUPPLY FOR THE UHF CAVITY BLANKER

This Appendix contains the schematics of the Blanking Pulse Generator (Figure I-1) and its Power Supply (Figure I-2) for the UHF Cavity Blanker.

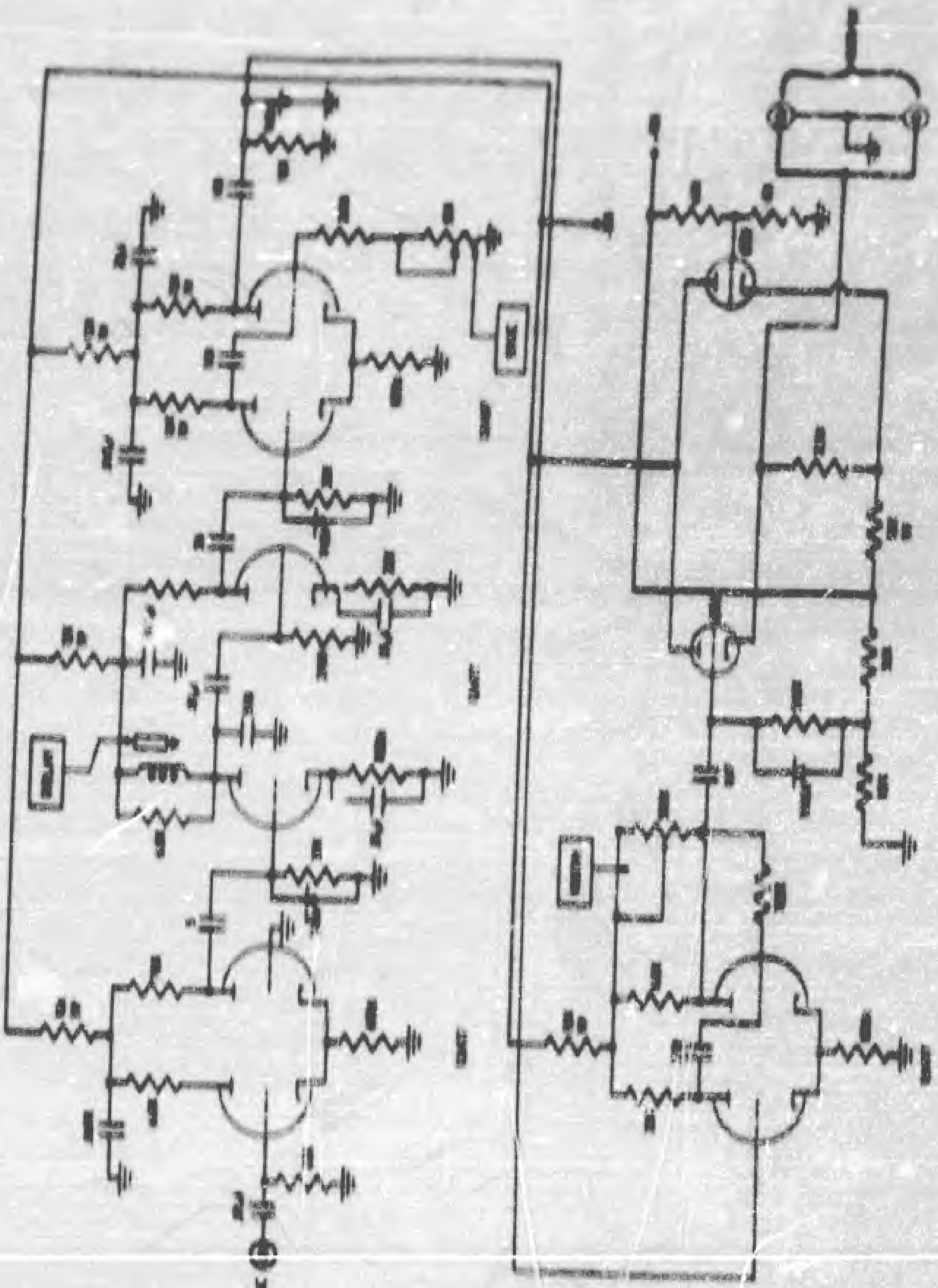

Figure 11. Marking Pulse Generator.

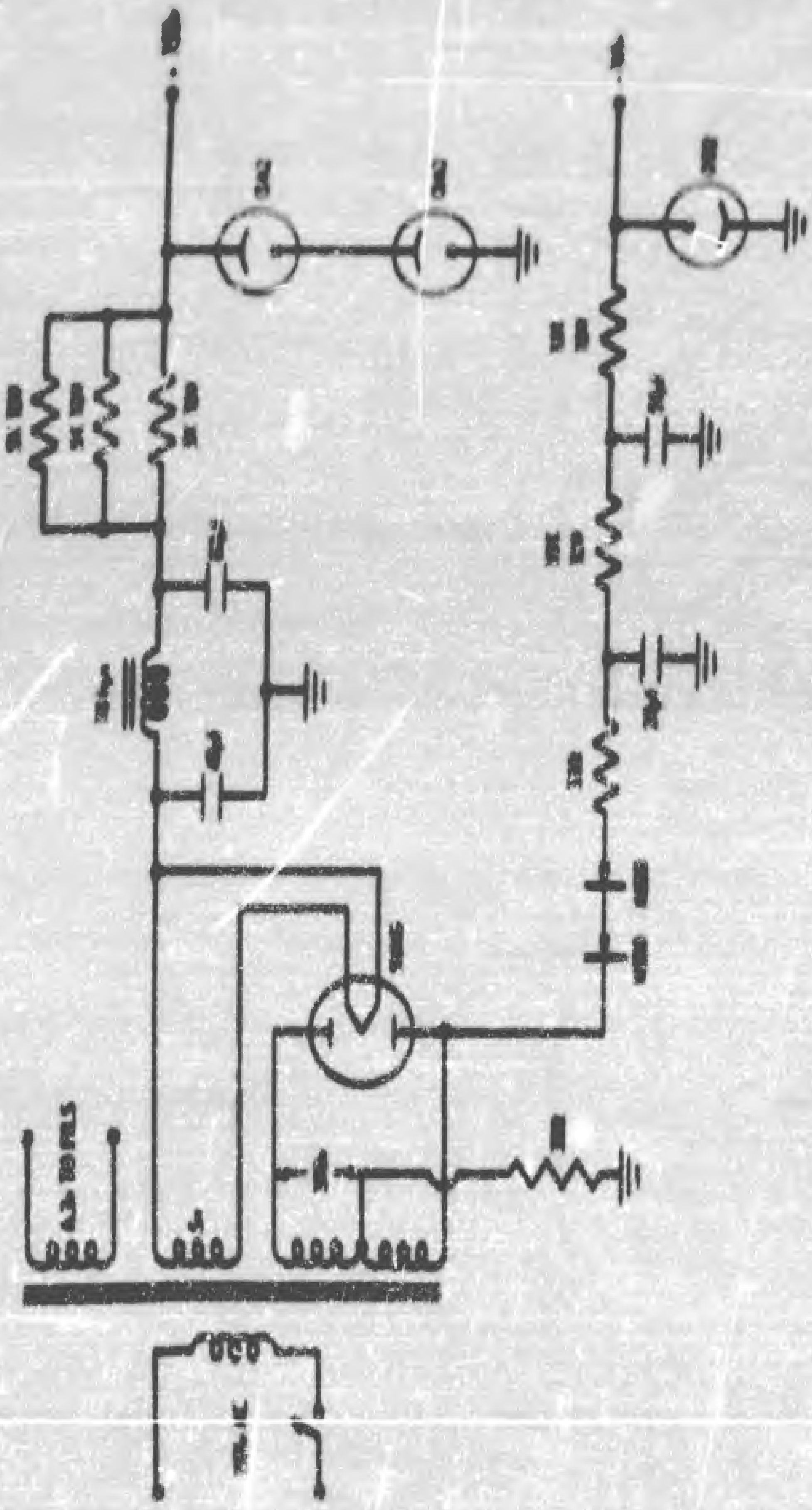

Figure 12. Power Supply for [illegible] Pulse Generator.

APPENDIX II
AUDIO FILTER ANALYSIS

The audio filter is a feedback amplifier in which the response of the feedback network is the inverse of the desired overall transfer function. Figure II-1 shows the desired response of the audio filter.

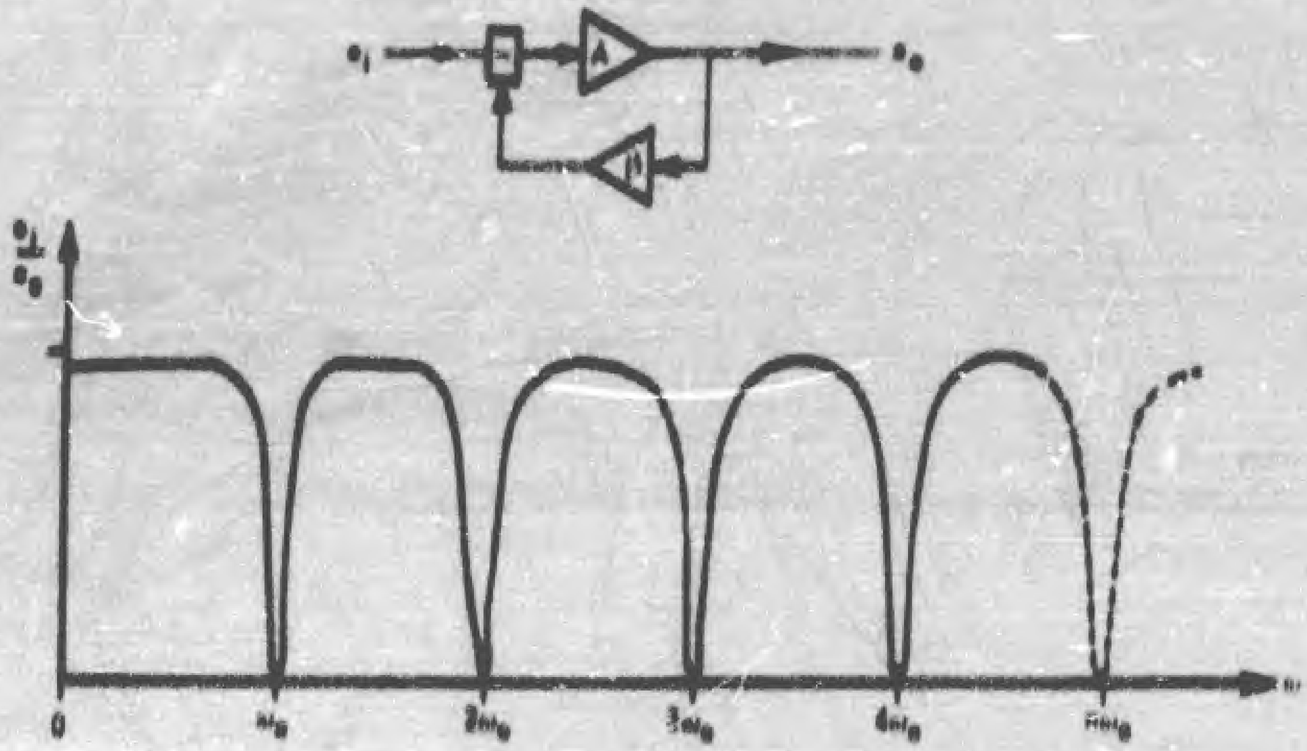

Figure II-1. Response of Audio Filter

To obtain this function the β network itself is a feedback amplifier as shown in Figure II-2. From this Figure:

$$E_o(t) = A[E_i(t) + E_o(t - T)]$$

Taking Laplace transform of both sides:

$$E_o(s) = A\,E_i(s) + Ae^{-sT}\,E_o(s)$$

$$E_o(s)\,[1 - Ae^{-sT}] = A\,E_i(s)$$

$$\beta(s) = \frac{E_o(s)}{E_i(s)} = \frac{A}{1 - Ae^{-sT}}$$

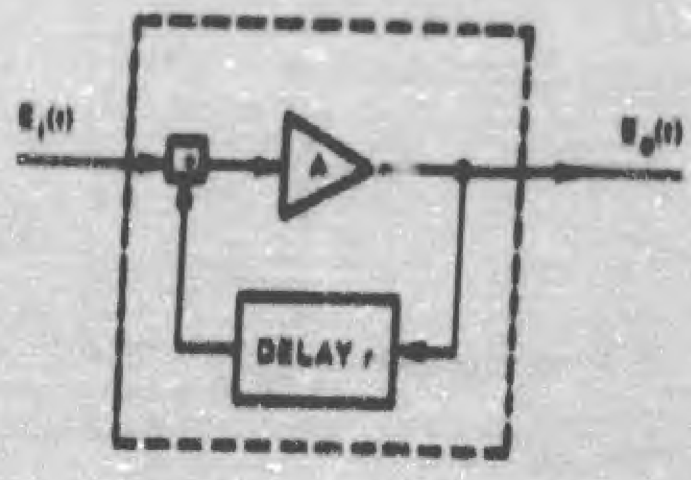

Figure II-2. Beta Network

The steady state response is given by $s = j\omega$

$$\beta(j\omega) = \frac{A}{1 - Ae^{-j\omega\tau}}$$

If ω_0 is taken to be: $\omega_0 = \frac{1}{\tau_0} = 2\pi f$, then:

$$\beta(j\omega) = \frac{A}{1 - Ae^{\left(-j\frac{2\pi\omega}{\omega_0}\right)}},$$

This function is shown in Figure II-3. If this $\beta(j\omega)$ is used as a feedback network around an amplifier of large forward gain, then the overall transfer function of the combination is given by:

$$G(j\omega) \approx \frac{1}{\beta(j\omega)}$$

so that the transfer would be the reciprocal of Figure II-3. This is quite like to the desired transfer function shown in Figure II-1. The relative attenuation at the frequencies $\omega = n\omega_0$ is:

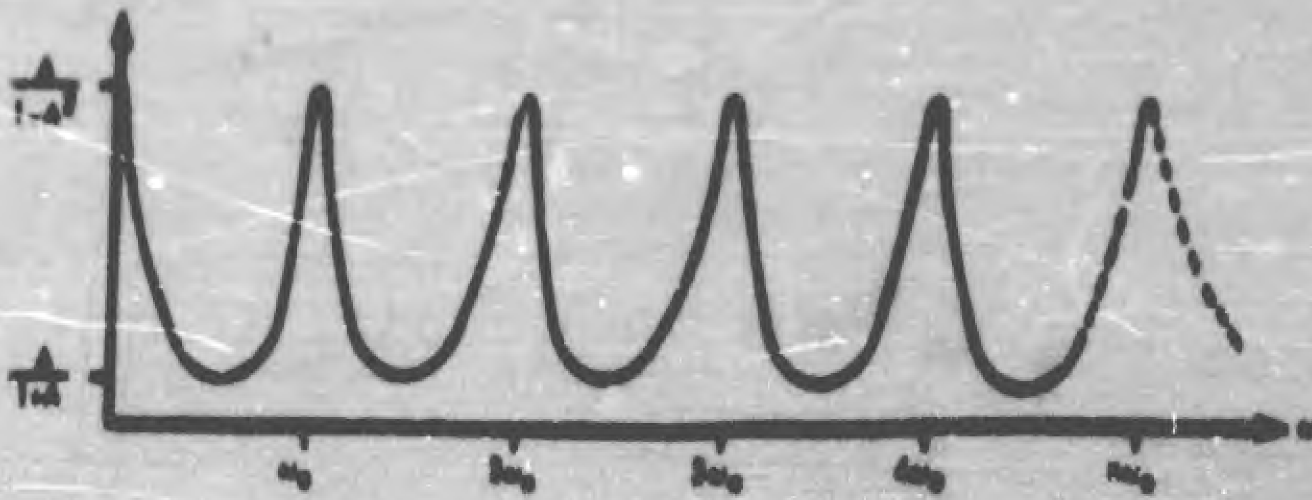

Figure II-3. Response of Nota Network.

$$\text{Gain}_{\text{relative}} = \frac{\left(\frac{A}{1+A}\right)}{\left(\frac{A}{1-A}\right)} \approx 1 - A$$

For $A = 0.99$, $\quad \text{Gain}_{\text{relative}} = .01 = -40 \text{ db}$;

so that the components of a periodic pulse signal of period $\frac{2\pi}{\omega_0}$ suffer an attenuation of 40 decibels. The use o. a variable delay line allows the null frequencies of the audio filter to be adjusted to the period of the interfering pulse signal.

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