RADC-TN-61-330

A PULSE INTERFERENCE BLANKER

By

W. B. Warren, Jr.

TECHNICAL NOTE NO. 1

Contract No. AF 30(602)-2366

Project No. A-525

Prepared for

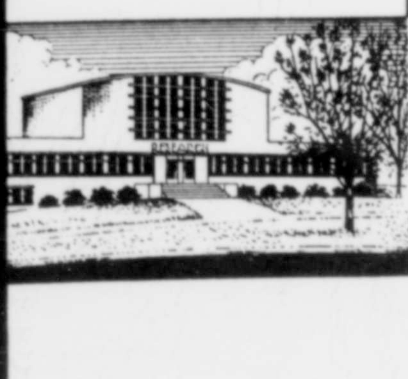
Rome Air Development Center

Air Research and Development Command

United States Air Force

Griffiss Air Force Base, New York

15 November 1961



Engineering Experiment Station

Georgia Institute of Technology

Atlanta, Georgia

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ABSTRACT

An interference blanker is described which is effective in overcoming the effects of pulse interference to narrow band communication receivers. This device employs both the techniques of "blanking" and "sampling" to obtain the desired interference suppression.

Explanations and mathematical justification for both modes of suppression are presented. A description is given of a completed device which permits the protection of communications equipment without the necessity of making internal modifications.

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1. INTRODUCTION

In many instances, it is necessary to co-locate narrow band communications equipment with high-powered, pulse-type equipment, such as radar. In this situation, it is not always possible to prevent a considerable amount of pulse energy from appearing at the input to the communications equipment. In some instances, this power level may be as high as several watts. Such power levels, incident upon conventional communications equipment, will cause serious overloading and desensitization and render impossible the reception of weak desired communications signals. This difficulty is due to the fact that most communication receivers do not provide sufficient frequency selectivity in their input stages, thereby permitting considerable amounts of the high-powered pulse interference to reach the grid circuits of the first few stages in the receiver. Overloading of these stages by the pulses causes grid current to be drawn with a consequent rapid charging of the AVC line and severe desensitization of the receiver because of the excessive AVC voltage which is developed. Consequently, pulse interference suppression techniques which are applied at a point in the receiver which follows these input stages are not effective in suppressing the effects of this interference, since the damage has already been done before the signal reaches the point where the suppression device is applied. Adequate suppression of the effects of this type of interference can be obtained only if the interfering pulse is rejected directly at the input terminals of the receiver.

It is the objective of this Note to describe a device which permits the suppression of the pulse interference directly at the input terminals to a communication receiver so as to avoid the difficulties mentioned previously. In addition, this device requires no internal modification to the receiver with which it is being used.

2. CONCLUSIONS

It is concluded that excellent protection of communications equipment from pulse interference can be obtained by the use of devices similar to the interference blanker described in this Technical Note. This device has been packaged in a form suitable for use in field testing, and its operation has been successfully demonstrated at RADC.

3. RECOMMENDATIONS

The results obtained in the application of the interference blanker to typical pulse interference situations indicate that this type of equipment may be successfully applied whenever problems of pulse interference to communications equipment are encountered.

It is recommended that this device be field tested in a variety of pulse interference situations to determine its applicability to a wide range of actual interference conditions.

4. DISCUSSION

In this device, called an "Interference Blanker", as is true with any other interference suppression device, it is necessary to recognize some essential difference between the desired and interfering signals and to make use of this difference to obtain the desired interference reduction. In the case of a narrow band desired signal, which is modulated with speech or other narrow band intelligence and which is being interfered with by a pulse-type signal, one essential difference lies in the fact that the interference is limited in time, while the desired signal is not. Thus, by turning the input to the narrow band receiver on or off at the proper instants of time, the interfering signal can be drastically reduced in amplitude or completely eliminated. On the other hand, the desired signal, not being of a time limited character, is sufficiently unaffected by this gating of the receiver input to permit the desired signal to be successfully recovered.

4.1. Blanking. The Interference Blanker applies the gating technique in either of two basic ways. The first of these is called "blanking". In the blanking technique, the receiver is turned off for the duration of each interfering pulse and turned on again between pulses, but the gaps in the desired signal due to this gating process do not seriously degrade the intelligibility of the desired signal. The blanking is accomplished by the insertion of a pulse controlled switch in series with the input to the communication receiver being protected, in the manner shown in Figure 1. The action of the pulse controlled switch in removing the interfering pulse signal is indicated by the wave forms of Figure 2, which illustrate the situation of a desired, sine wave modulated AM signal with a periodic pulse interference superimposed. Notice that this technique will be effective only if the blanking pulses are synchronous with the interfering pulse

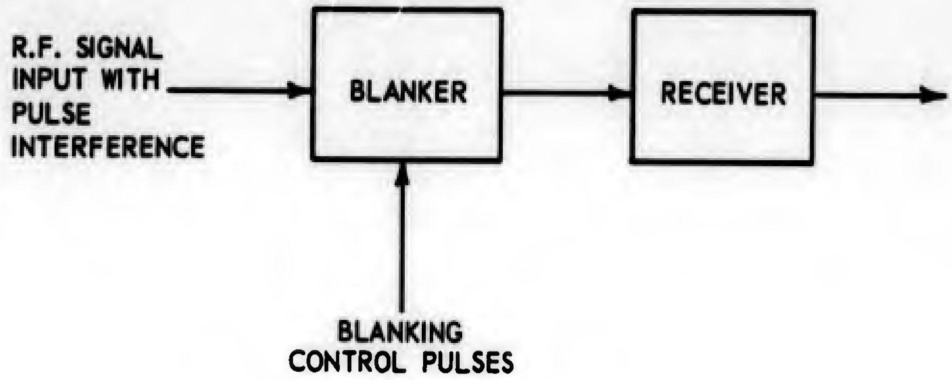


Figure 1. Interference Blanking

signal. In cases where it is possible, the best method to obtain this synchronization is to have a direct cable connection with the source of interference so that a pre-trigger for the blanking pulse can be obtained. The pre-trigger is necessary, since there is some delay in the generation of the blanking pulse and in the disconnect action of the blanking switch. As a result of this delay, it is not possible to blank a particular interference pulse by detecting that particular pulse and using this detected output to generate the necessary control pulse for the blanking switch unless an adequate delay in the main signal path can be obtained. In general, such a delay in the signal path is difficult to obtain and has the additional disadvantage that most devices which might be used for this purpose have an appreciable insertion loss so that the desired signal would be unduly attenuated in passing through them. However, if the proper pre-trigger pulse is available, the blanking switch may be operated by this pulse slightly before the arrival of the interference pulse at the receiver input and complete blanking of the interfering signal is obtained. If a direct cable connection to

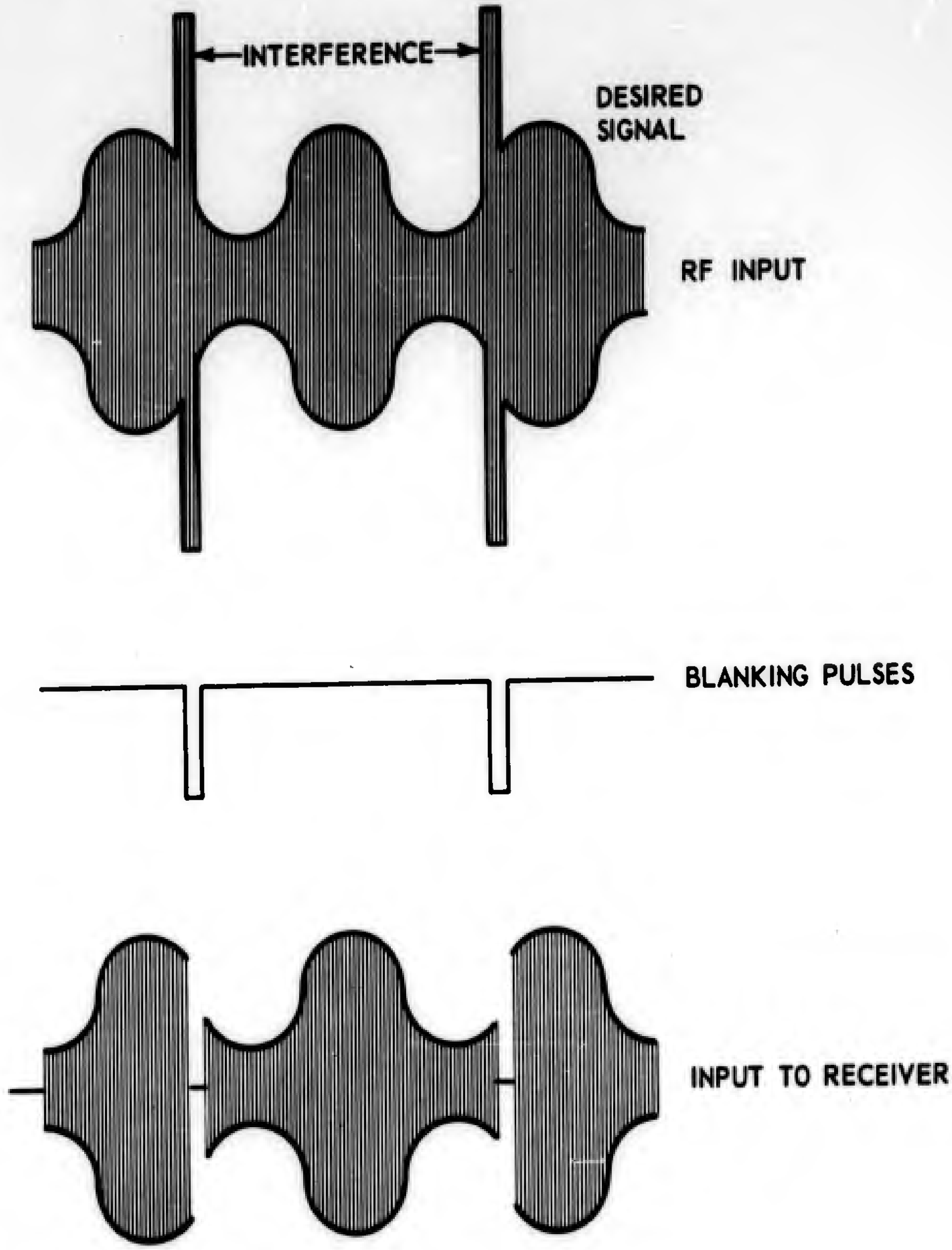


Figure 2. Blanking Waveforms.

the source of pulse interference is not available, an auxiliary receiver can be used to detect the interfering pulses and supply them to a blanking pulse generator as synchronizing information.

In the latter situation, it is not possible to have the proper pre-triggering information, due to the previously mentioned delay, which is associated with the switch and will also be associated with the auxiliary receiver. However, for periodic interference, the required pre-triggering information can be obtained by delaying the interfering pulses by an amount just short of one pulse interval. In this arrangement, the blanking pulse corresponding to a particular interfering pulse is generated by the previous interfering pulse delayed by almost one pulse interval. This technique is effective as long as the condition of periodicity is met.

If the multiple interference sources are encountered and if the rates of these sources are not integrally related, it is not possible to gate out all the interference. In fact, if the blanking pulses are synchronized to one of the interfering sources, then every pulse of a non-related interfering source will enter the receiver, except those which overlap the pulse interval of the source to which the blanking pulses are synchronized.

4.2. Sampling. A second receiver gating technique which is used in the Interference Blanker to reduce or eliminate pulse interference is called "sampling". In this process, samples of the input signal are taken at a rate which is sufficiently high to reconstruct all the essential information in the desired signal from knowledge of the samples alone. Since the required information in the input signal is the envelope (in the case of an AM signal), it is not necessary to provide samples at a rate high in respect to the carrier frequency of the input signal, but rather it is sufficient to provide samples at a rate which is high with respect to the highest frequency components contained in the envelope of the desired signal. For the case of a speech modulated signal occupying the frequency range 200 to 4,000 cps, the required minimum sampling rate is 8,000 per second, in accordance with well known theory of sampled signals. However, in a practical situation, the perfect filters required by the theory are not available and a slightly higher sampling rate is necessary to permit "clean" recovery of the original signal from knowledge of the samples alone. For the case mentioned above, a practical sampling rate of the order of 10,000 samples per second is necessary. Since the sampling of the input signal is done at the RF input to the receiver, it is necessary to reconstruct the desired signal from the sampled signal by means of a band-pass filter, rather than by a low-pass filter, as is usually the case.

Figure 3 illustrates the technique of sampling an amplitude modulated signal and reconstructing the original signal by a band-pass filter. Since the output of the sampling switch is zero, except at the sampling instants, it is apparent that the output of the band-pass filter is dependent only on the values of the input AM signal at these sampling times, since the input to this filter is zero at all other times. Hence, any interfering pulse signals which might occur at times other than the sampling instants will not be present in the output of the sampling switch and, therefore, will not be present in the output of the band-pass filter.

In a practical situation, the bandwidth of the RF amplifier portion of a conventional receiver is not sufficiently narrow to act as a perfect band-pass filter, such as is assumed to be available in acquiring the wave forms of Figure 3. As a result, the input to the mixer and, hence, the input to the IF amplifier

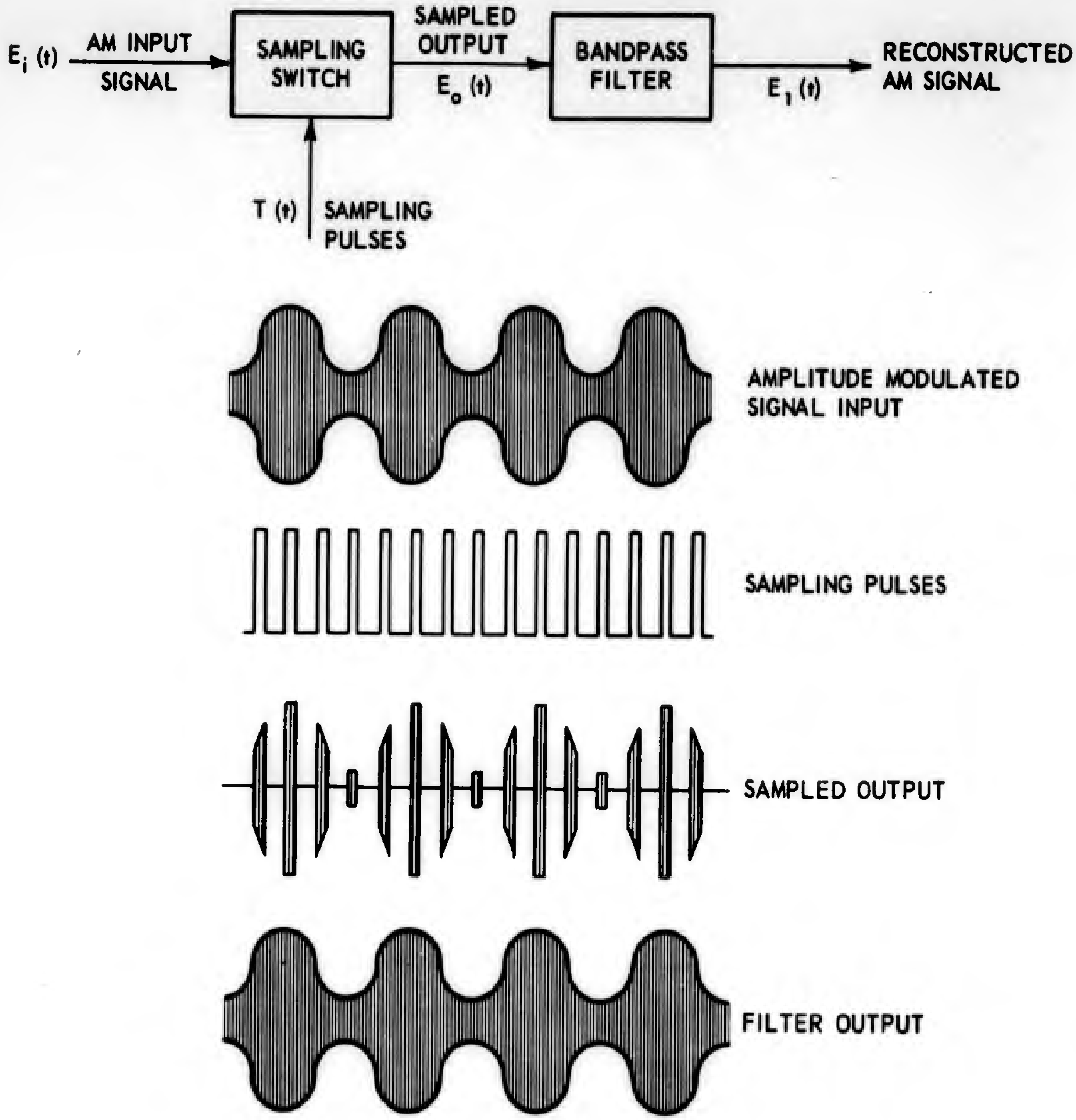


Figure 3. Sampling Technique.

is in the nature of a pulse amplitude modulated signal, similar to that labeled "sampled output" in Figure 3, with the exception that the carrier frequency of this pulse amplitude modulated signal is now the IF frequency of the receiver, rather than that of the original input signal. The narrow bandwidth of the IF amplifier then serves as the required band-pass filter so that its output will appear as the reconstructed AM signal and as such may be applied to an envelope detector to produce the desired envelope modulation at audio frequency.

4.2.1. Spurious Responses. A mathematical justification of the fact that this sampling technique does indeed preserve the envelope modulation on the desired signal is given in the Appendix. In this analysis, it is pointed out that the process of sampling constitutes a modulation of the input signal at the sampling rate, and hence produces sidebands which are distributed in frequency on both sides of the desired signal frequency. This is an undesirable situation, since it causes the receiver to have a spurious response at frequencies close to the frequency of the desired signal. The magnitude of these responses is governed by the magnitude of these sidebands, which in turn are functions of the shape of the sampling pulses. As a result, it is possible to reduce the level of these spurious responses considerably by careful attention to the shape of the sampling pulses. Nevertheless, this production of spurious responses represents a disadvantage in the use of the sampling technique. In addition, it should be noted that the desired signal source is connected to the receiver input for a small fraction of the total time; this fraction being equal to the duty cycle of the sample pulses. However, the receiver noise as generated in the input stages of the receiver is present all the time, with the result that the signal to noise ratio of the input signal is a function of the duty cycle of the sampling process. It is apparent, then, that for a constant sampling rate, the width of the sampling pulses should be made as wide as possible. On the other hand, the probability that interfering pulses will occur at one of the sampling instants is directly related to the width of the sampling pulses. Therefore, a compromise must be reached in which the maximum pulse width is used which is consistent with the desired degree of interference reduction.

4.2.2. Calculation of Interference Reduction. Some measure of the degree of interference reduction that may be obtained by the use of a sampling scheme such as is described here may be obtained by considering the probability that an interference pulse occurs during one of the sampling times. If there is synchronization between the sampling pulses and the interfering signal, then there is no matter of probability involved, since the existence or absence of interference depends on the phase relationship between the sampling pulses and the interfering signals and may be calculated explicitly. However, for those cases where the sampling pulses occur in a somewhat random fashion with respect to the interfering pulses, the question of the absence or existence of interference reduces to the determination of the probability that an interference occurs during the time of the sampling pulses.

At any instant of time, the probability that a sample pulse occurs is given by the duty cycle of the sample pulse train. Likewise, the probability of occurrence of an interfering pulse is given by the duty cycle of the interfering pulse train. If these two events are statistically independent, then the probability of simultaneous occurrence of these two events is given by the product of their probabilities. Thus, if

$$f_1 = \text{sampling pulse rate}$$

f_2 = interfering pulse rate

Δ_1 = sampling pulse width

Δ_2 = interfering pulse width

then the per cent of time that coincidence occurs is given by

$$\% \text{ coincidence} = (\Delta_1)(\Delta_2)(f_1)(f_2). \quad (1)$$

This particular mode of operation, then, is best suited to those situations where multiple, asynchronous, interference sources are encountered. Since the sampling pulses may be synchronized so that samples are taken immediately before or after the occurrence of the pulses of one of the sources, the number of the interfering pulses from other sources, which will enter the receiver, will then be reduced by the factor given in Equation (1).

4.3. Diode Switch. In the application of either the blanking or sampling techniques to the reduction of pulse interference, it is necessary to have a switch which can be controlled by the blanking pulses to effect a rapid disconnection of the receiver from its antenna so as to prevent interfering pulses from entering the receiver. At the same time, this switch must be capable of a very low minimum insertion loss when it is used to connect the receiver and the antenna together so as not to unduly attenuate low-level desired signals. The basic design of such a switch, capable of operation in the frequency range of 200 to 400 mc, has been developed by Georgia Tech under Contract No. AF 30(602)-1789. This switch is essentially a multi-section low-pass filter whose cut-off frequency is about 400 mc. Each section of this low-pass filter has a normally back-biased diode connected in shunt with it so that the effect of the diode is to add a small amount of shunt capacity to each section of the filter. This small added capacitance is taken into account in the design of the filter so that the effect of the back-biased diode is essentially negated and the attenuation in the pass band of the filter is quite small. However, the application of a forward biasing voltage to the shunt diodes causes them to exhibit a very low shunt resistance across each section of the filter, and the attenuation of the filter becomes very large. This large attenuation results from two mechanisms. The first is the considerable impedance mismatch encountered at the input to the filter when the diodes are forward biased, with the result that only a small fraction of the incident power actually flows into the switch proper; the larger portion of the power being reflected back to the source. That fraction of the power which does enter the filter is then attenuated in each successive section of the filter by an amount dependent upon the ratio of the shunt diode resistance to the series reactance of the inductive arm of that particular section.

This mechanism may be visualized by referring to Figure 4 which shows, schematically, the construction of the switch. Two capacitors are used at the input and output of the filter to isolate the filter for DC; however, the reactance of these capacitors at frequencies in the 200 to 400 mc range is negligible. The control pulse is applied to the filter through an RF choke whose impedance is high with respect to the 50 ohm impedance level of the filter but does not present any appreciable reactance at frequencies contained in the control pulse. The details of the mechanical construction of this filter can be seen in the photograph of Figure 5. The series inductance is provided by short lengths of No. 10 wire, while the shunt capacitance is supplied by silver mica button capacitors. This particular type of capacitor was used because of the low inductance connection which can be made to it, and because it concentrates its capacitance at one point. These button capacitors also serve as feed through connections

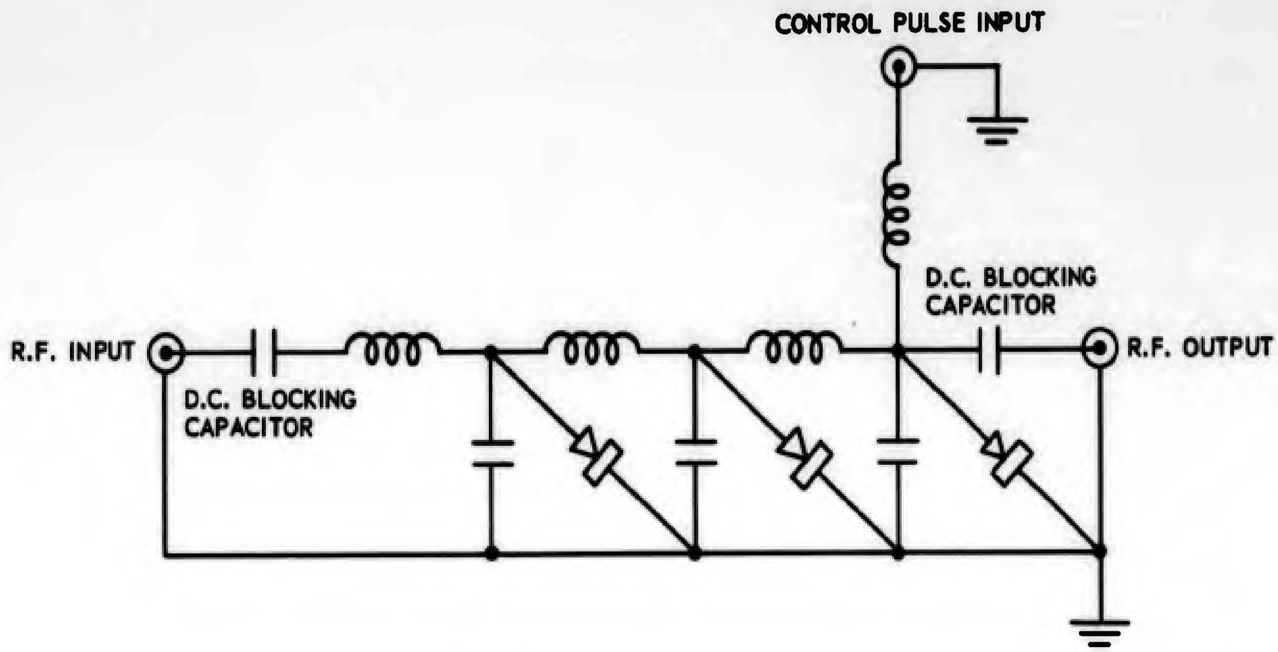


Figure 4. RF Switch Schematic

between adjacent sections of the filter. The diodes are connected between the center conductor and ground and are physically located at the point where the center conductor connects to the shunt capacitors. The entire assembly is constructed in a box approximately one inch in cross section and eight inches long. The 50-ohm impedance level of the filter is sufficiently low with respect to the impedance of the box viewed as a short length of transmission line so that the box introduces no appreciable effect on the response of the filter.

A set of typical frequency characteristics of a diode controlled switch of the type shown in Figure 5 is shown in Figure 6. The insertion loss with a 400 milliampere control current exceeds 56 db over the range of 200 to 400 mc, while the insertion loss with the diodes back-biased is less than 1 db over the same frequency range. This 56 db insertion loss is sufficient to reduce a 10 watt

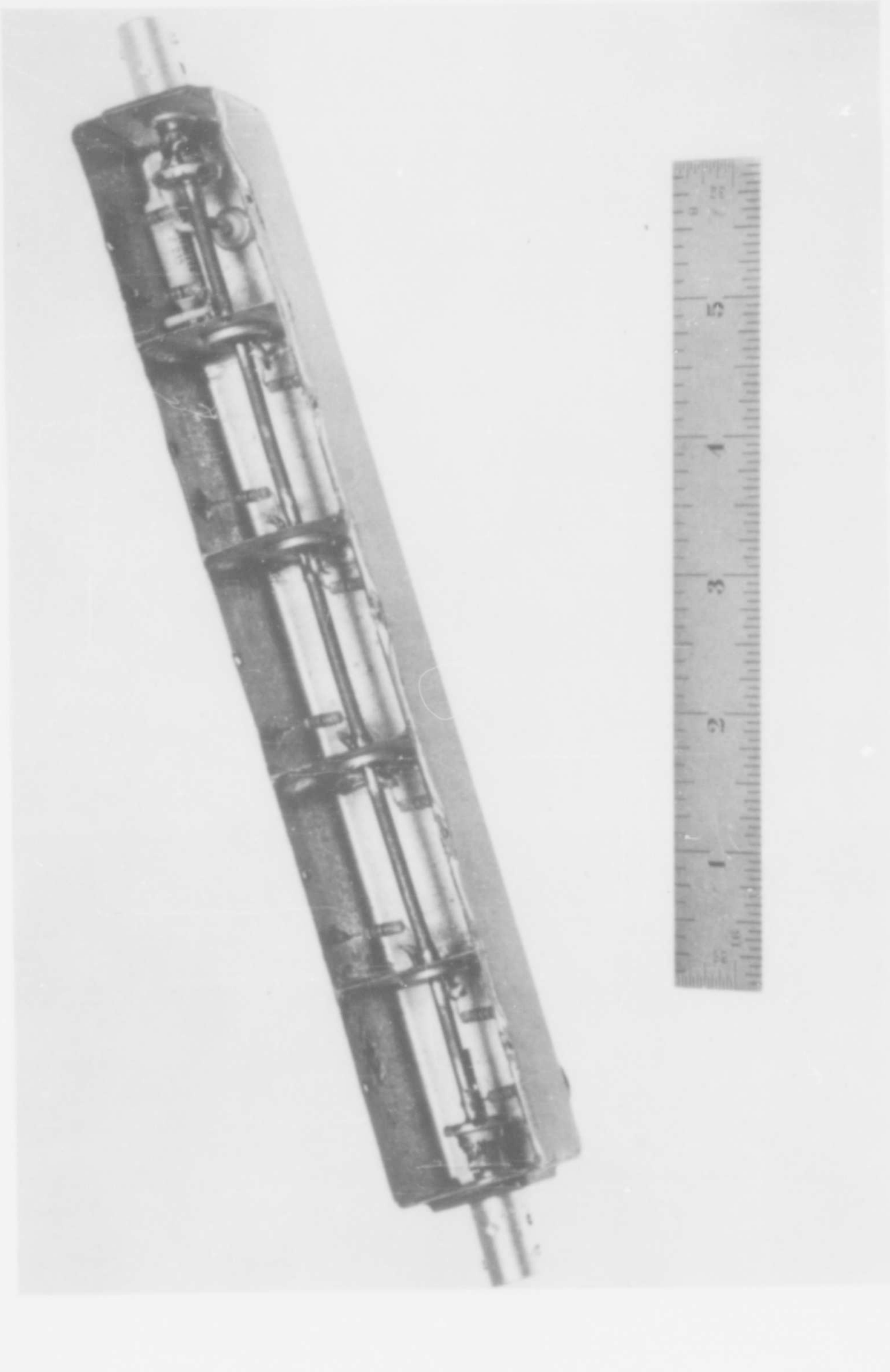


Figure 5. Photograph of Diode Switch.

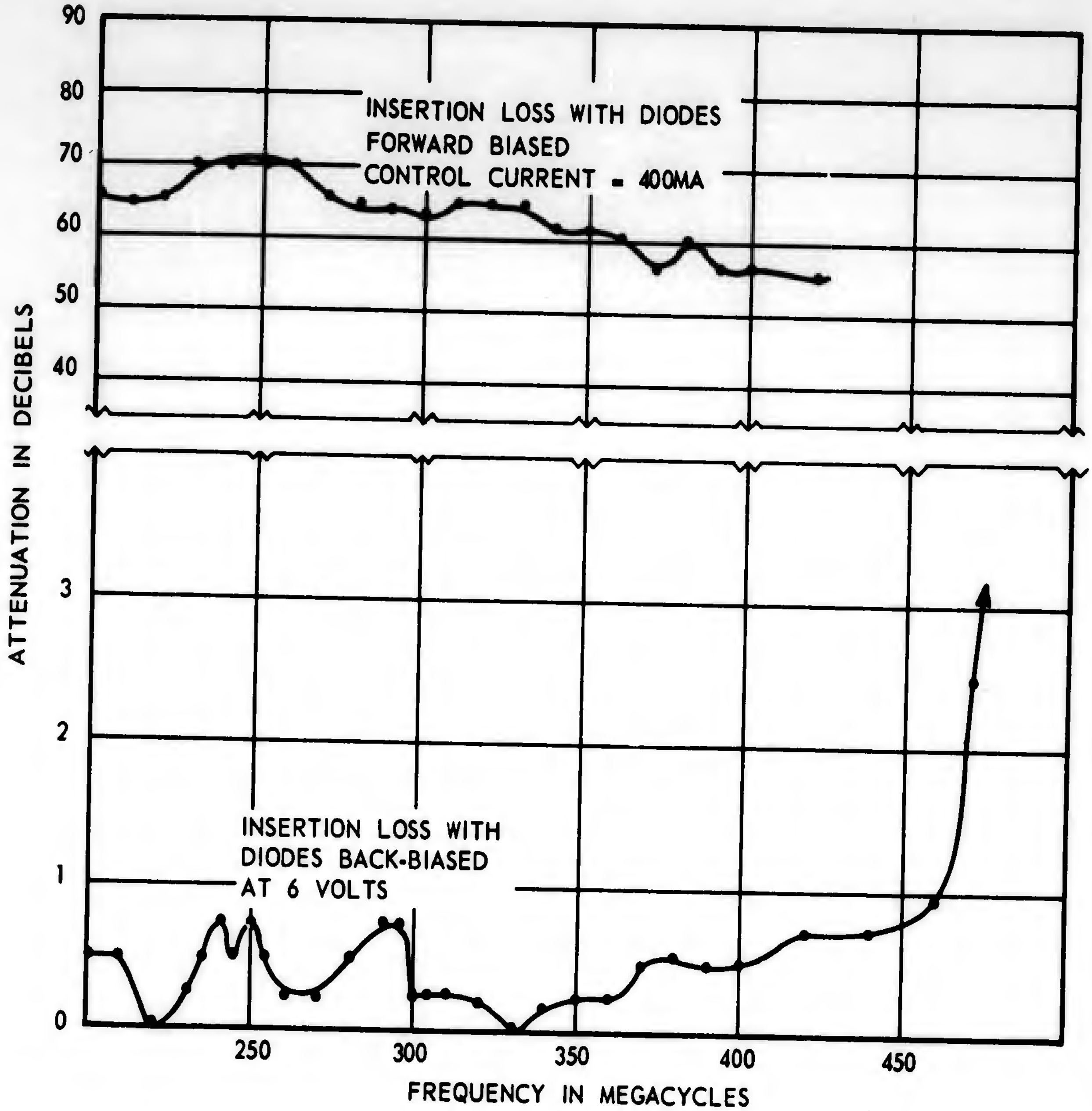


Figure 6. Switch Insertion Loss Characteristics.

interfering pulse appearing at the antenna terminals of the receiver to a level of 25×10^{-6} watts. Although this level is quite low and will not cause serious overloading in the front end of the receiver, it is still sufficiently large to cause considerable difficulty in the reception of desired signals close to the sensitivity threshold of the receiver. It is also possible that this remaining level is large enough to produce a considerable AVC voltage at the output of the AVC detector, especially in those receivers having conventional peak detecting AVC rectifiers. This excessive AVC voltage may cause sufficient desensitization of the receiver to prevent the reception of the very weak desired signals. In such an event, two of the diode switches may be connected in cascade so that an additional 56 db of attenuation of interfering pulse signals is obtained. This amount of attenuation is sufficiently large to reduce the level of the interfering pulses almost to the noise level of the receiver. Even with this arrangement, the maximum insertion loss in the back-biased condition is considerably less than 2 db over the entire range of interest from 200 to 400 mc. With regard to the cascade connection of two of these switches, care must be taken to connect the control pulse inputs of the two switches in a "back to back" manner. This is illustrated in Figure 7. This is necessary because the leakage of the signal out of one of the switches into the common control pulse lead can effectively by-pass the action of the second switch when the "back to back" connection is not used.

In the use of a switch of this nature, there is a change in the current in the inductance of the low-pass filter, when the diodes are switched from the forward to back-bias condition. Since a portion of this current flows in the output circuit of the switch, it is possible that extraneous signals may be presented to the input to the receiver even though no input signal is incident upon the

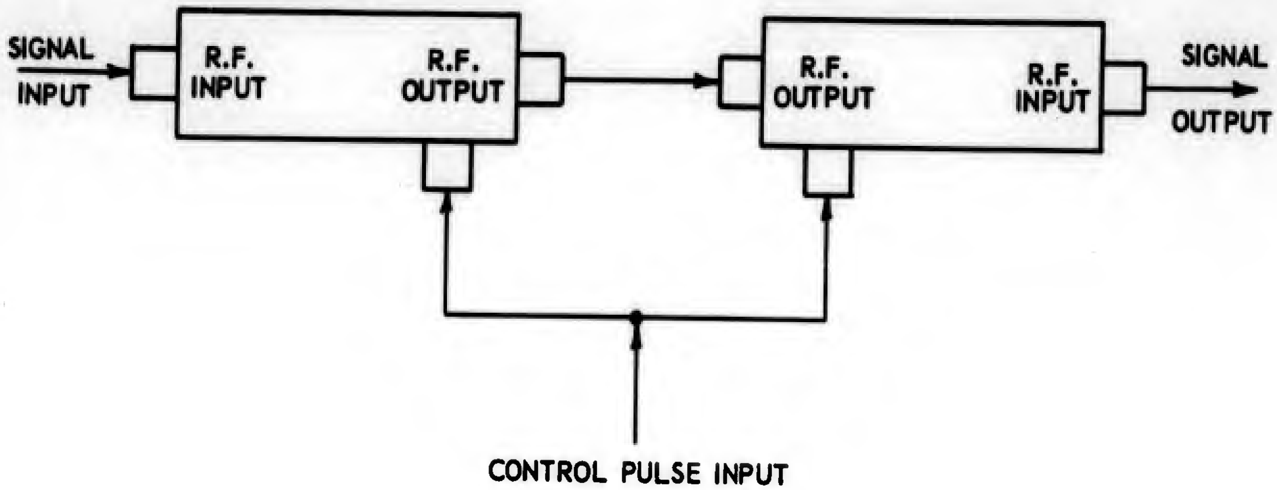


Figure 7. Proper Method of Cascading Two Diode Switches

antenna. Fortunately, this difficulty is not encountered in the use of this switch in the frequency range 200 to 400 mc, since the switching of the diodes from the forward to back-bias condition is slow enough that no components of this switching current pulse falling in the range 200 to 400 mc are of sufficient amplitude to exceed the receiver noise level. If the use of such a switch were contemplated at lower frequencies, it would be necessary to use some form of balanced switching circuitry so that the current in the inductance of the filters would not change when the condition of the diodes was reversed. One arrangement which meets this requirement is shown in Figure 8. In this case, the two components of the switching current in the output transformer are in the same direction and hence cause no net induced voltage in the transformer secondary. However, the desired signal is fed in a "push-pull" connection so that the two components in the output transformer are of opposite phase and, therefore, produce a secondary voltage which is proportional to the sum of the two primary components. Careful attention to the construction of the transformer and to the matching of the diode characteristics should permit reduction of the unbalanced components of control pulse current to the order of 1% or less.

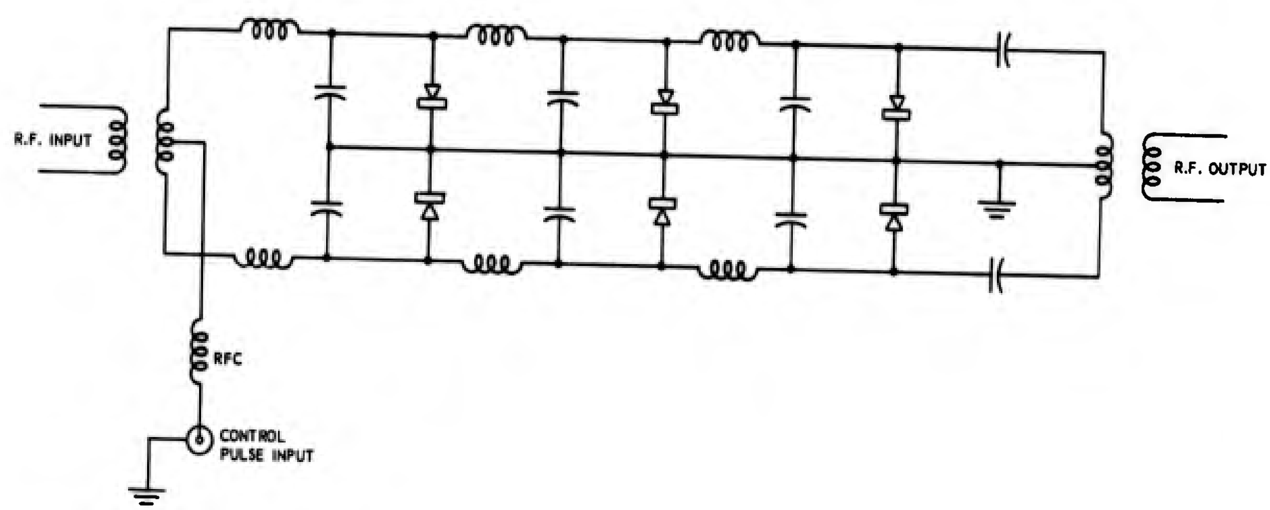


Figure 8. Balanced Switch

4.4 Equipment Design. The block diagram of Figure 9 illustrates the way in which the pulse control switch has been combined with the auxiliary receiver and suitable pulse delay and shaping circuitry to acquire proper synchronization of the pulse control switch for the suppression of periodic pulse interference. Referring to Figure 9, the desired signal and interference are superimposed at the

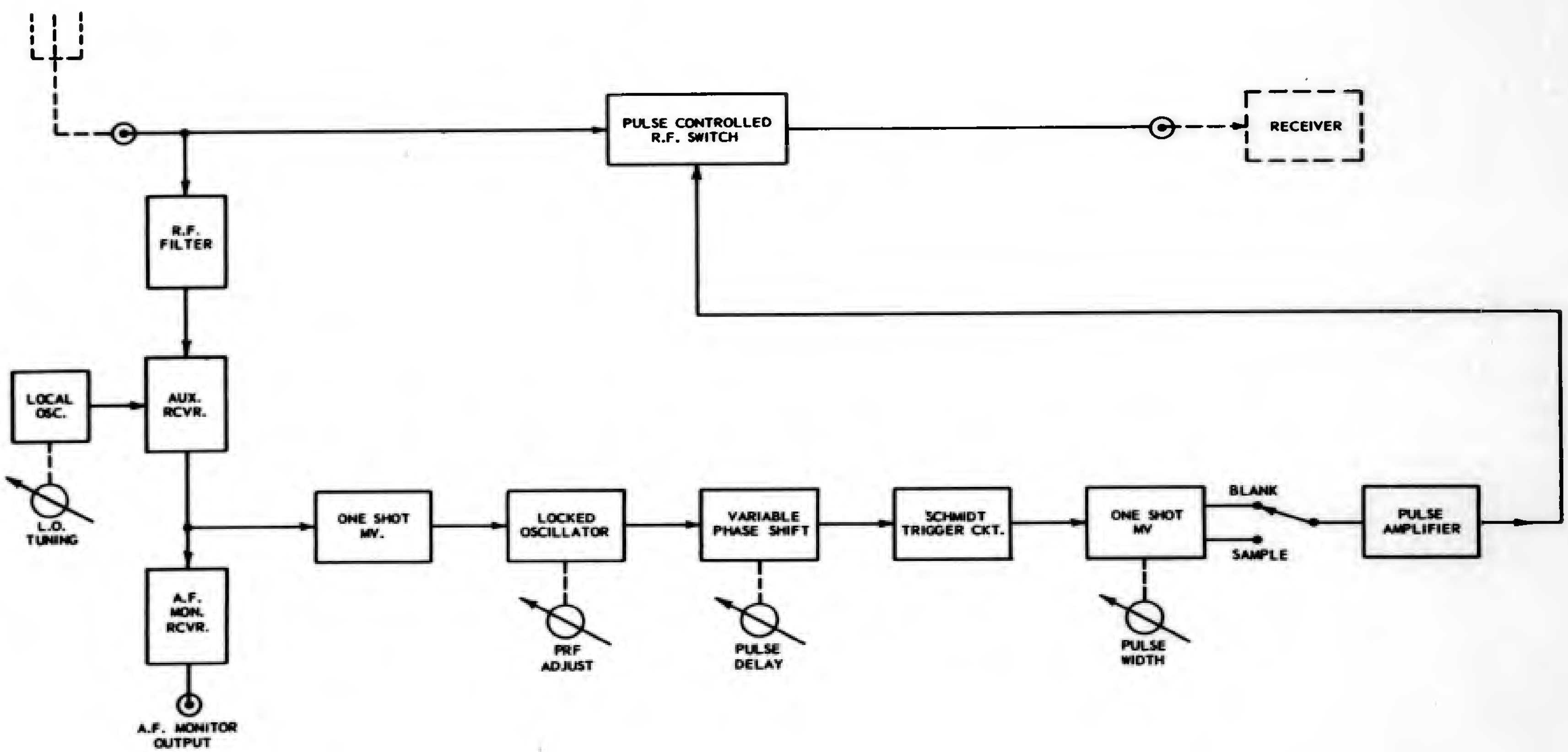


Figure 9. Block Diagram - Interference Blanker.

antenna and are applied both to the auxiliary receiver and to the pulse controlled switch. The auxiliary receiver supplies at its output one pulse for each pulse of the input interfering signal. Since the shape of this pulse depends on the shape of the transmitted interfering pulse, a one shot multivibrator is triggered by the output of the auxiliary receiver to produce a standard pulse shape which is independent of the shape of the input trigger pulse. This standard output pulse occurs at the same repetition rate as the input interfering pulse and is used to synchronize a locked oscillator at one of the harmonics of the repetition rate which lies in the frequency range 9 to 11 kc. The reason for this synchronization on one of the harmonics of the interfering repetition rate, rather than on the fundamental, follows. When the incoming desired signal is turned on and off by the blanking of the RF switch, a "hole" is created in the envelope of the desired signal. This "hole" constitutes an amplitude modulation of the desired signal at the blanking pulse rate. Since the repetition rate of most interfering pulse sources lies in the audio range, blanking the signal at the fundamental rate of the interfering pulse source would cause a corresponding amplitude modulation of the desired signal at an audio rate and this would result in an annoying tone interference in the output of the receiver; however, if the blanking pulse rate is at some harmonic of the input interfering pulse rate, the amplitude modulation of the desired signal will still occur but the lowest frequency component will be at the rate at which the blanking is being accomplished. For the case of the locked oscillator in question, this rate would be in the range 9 to 11 kc. As a result, low-pass filtering in the audio output of the receiver is effective in removing this tone, but does not affect the desired audio signal.

The sinusoidal output of the synchronized oscillator then is supplied through a continuously variable phase shifter to actuate the Schmidt trigger circuit which produces an output pulse at the zero crossing of each of the cycles of the phase shifter output. A variation of the phase shift causes a corresponding variation in the zero crossings of the phase shifter output and a resulting shift in the time position of the output pulse of the Schmidt trigger circuit. This output pulse is used to trigger a one shot multivibrator whose output pulse width is controllable. This output pulse, of the proper width to blank out the interfering signal, is passed through a pulse amplifier to control the RF switch in series with the input to the receiver. In operation, then, the interfering pulse is picked up by the auxiliary receiver whose output is used to synchronize pulses in the frequency range 9 to 11 kc, whose width and time of occurrence are manually adjustable. These pulses are used in the proper polarity to gate the input to the receiver off or on as the circumstances may require. A more detailed discussion of the individual components of this interference blanker is given in the following paragraphs.

4.4.1 Auxiliary Receiver. The requirements of an auxiliary receiver to detect the presence of interfering pulses and supply synchronization information to the locked oscillator depend upon the expected levels of the interfering pulse signals. For those situations where only large amplitude pulse interference will be encountered, a simple crystal-video receiver has sufficient sensitivity. A typical receiver of this type is illustrated in Figure 10. The principal limitation on the sensitivity of receivers of this type is the lack of diode nonlinearity in the region near zero bias. Although this can be remedied to some extent by the application of a small forward bias to the diode, generally, rectification efficiency at signal levels below several millivolts is quite poor. It is necessary to provide some means of RF selectivity in front of the diode detector in order that pulse signals whose frequencies are far removed from the frequency range in

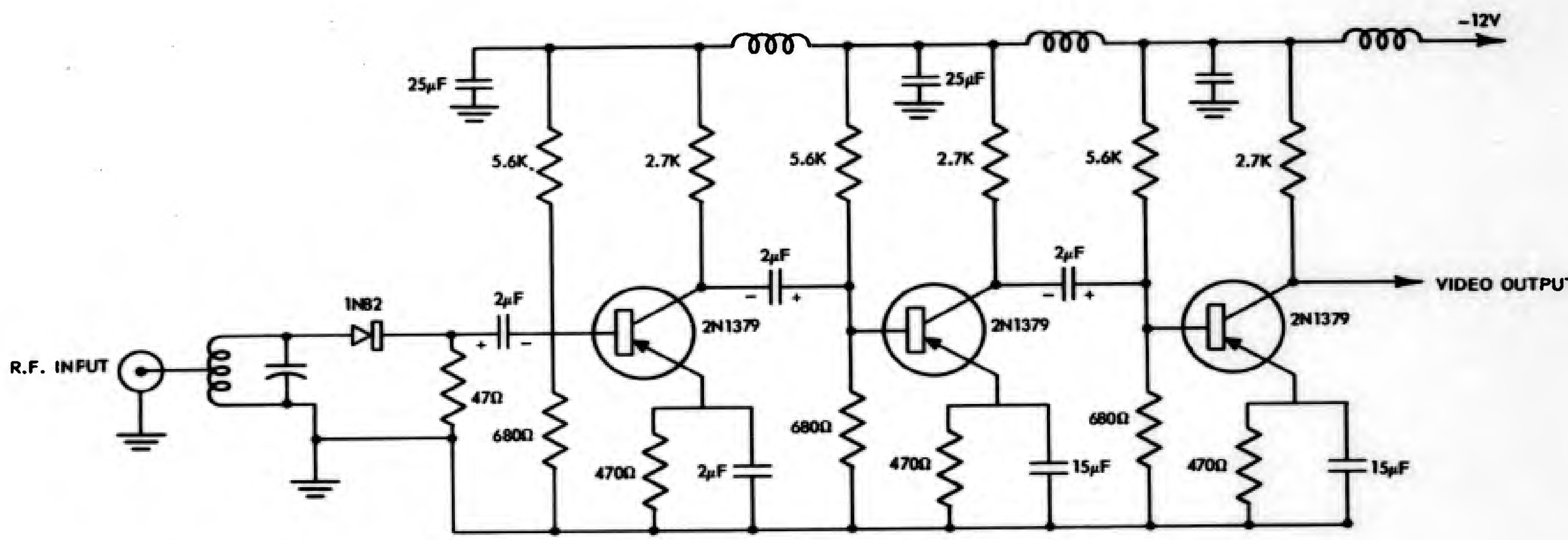


Figure 10. Crystal-Video Receiver.

which operation is contemplated do not produce false triggering of the blanking circuitry. Since no stringent selectivity requirement need be placed on this filter, adequate protection against this false triggering may be acquired with a simple single tuned RF filter. One such filter which has been successfully applied to this application is shown in Figure 11. This filter has essentially 50-ohm input and output impedances and has an effective "Q" of several hundred. The short section of two-wire transmission line is effectively short circuited by the ground plane to which it is attached. Since the length of this transmission line is considerably less than one-quarter wave length over the frequency range 200 to 400 mc, the reactance presented at the open end of this line section is essentially inductive. This inductance is resonated by a variable capacitor placed directly across the open end of the transmission line. One major disadvantage in the use of this particular configuration lies in the fact that both the rotor and stator of the variable capacitor are at RF potential with respect to the ground plane. This difficulty was overcome by mounting the capacitor in a polystyrene block and mechanically coupling to the rotor with an insulated shaft. The entire assembly was then enclosed in a shield can with adjustment of the resonant frequency being made by means of an insulated shaft. The input and output connections are made by means of coaxial cables which connect to the BNC connectors on either side.

Listening tests have demonstrated that when pulse signals of one-half millivolt or larger are superimposed on desired signals of the level of 2 to 4 microvolts, an appreciable interference occurs. For this reason, it is necessary to provide blanking action at the input to the receiver whenever the level of interfering pulses exceeds one-half millivolt. Since adequate sensitivity could not be obtained with a simple crystal-video receiver to operate at these low signal levels, it was necessary to construct a superheterodyne receiver to acquire the desired sensitivity.

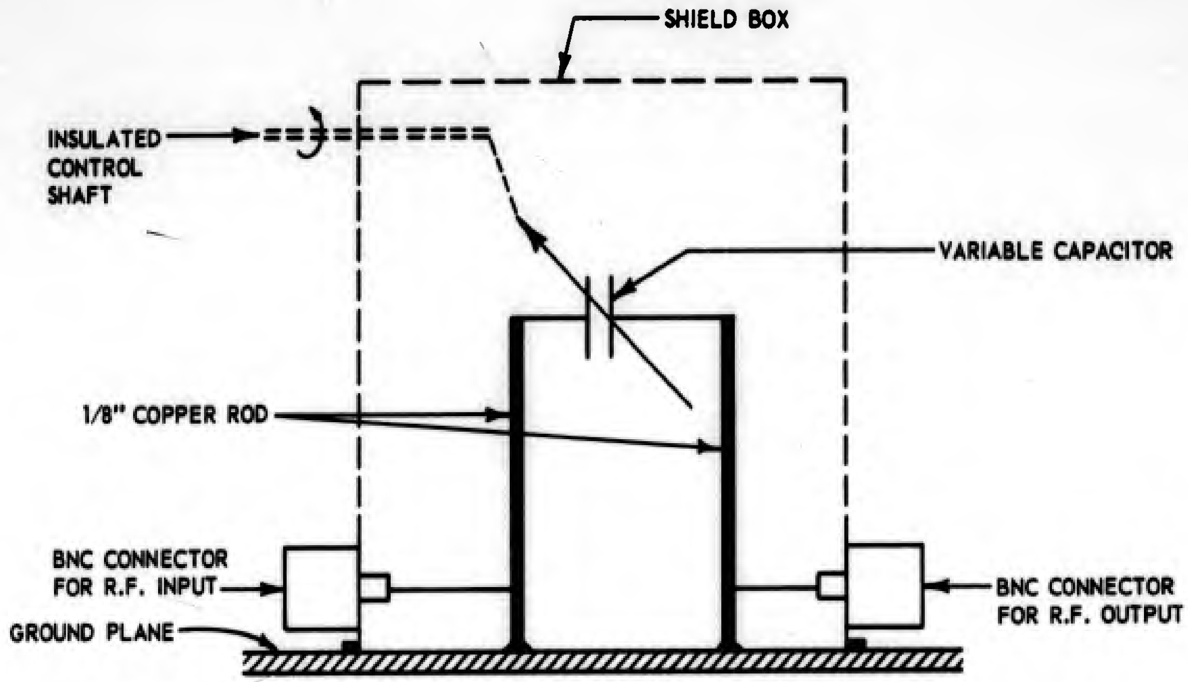


Figure 11. Tunable RF Filter

The physical construction of the receiver is illustrated in the photograph of Figure 12 and a schematic is given in the Appendix. A simple crystal mixer is used to heterodyne the output of the RF filter with the local oscillator to produce a 40 mc IF signal. The 40 mc IF frequency was chosen to obtain the necessary bandwidth for undistorted reproduction of the interference pulse. Three stages of IF amplification are provided to raise the signal level to a value sufficiently large to obtain efficient rectification in the diode detector. The output pulse from the detector is then used to drive a three-stage video amplifier which raises the level of the detected pulse to a value sufficiently large to insure reliable triggering of the one-shot multivibrator. An additional one-stage audio amplifier is connected to the video output to provide an audio monitoring output for use as an aid in receiver tuning. In order to simplify the

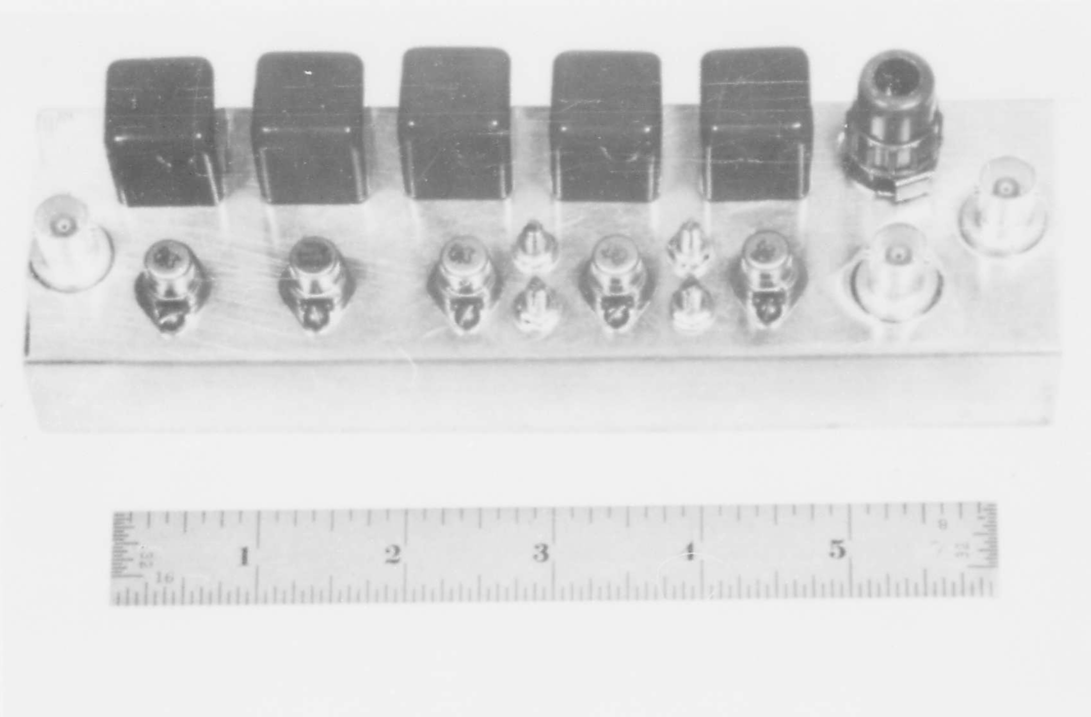


Figure 12. Auxiliary Receiver

receiver construction and to avoid tracking problems, the RF filter and the locked oscillator are separately tuned; the proper setting being obtained by maximizing the output of the audio monitoring amplifier on the interfering pulse signal.

4.4.2. Locked Oscillator. Figure 13 illustrates schematically the construction of the locked oscillator, along with the one-shot multivibrator which is used to standardize the synchronizing pulse for the locked oscillator. The

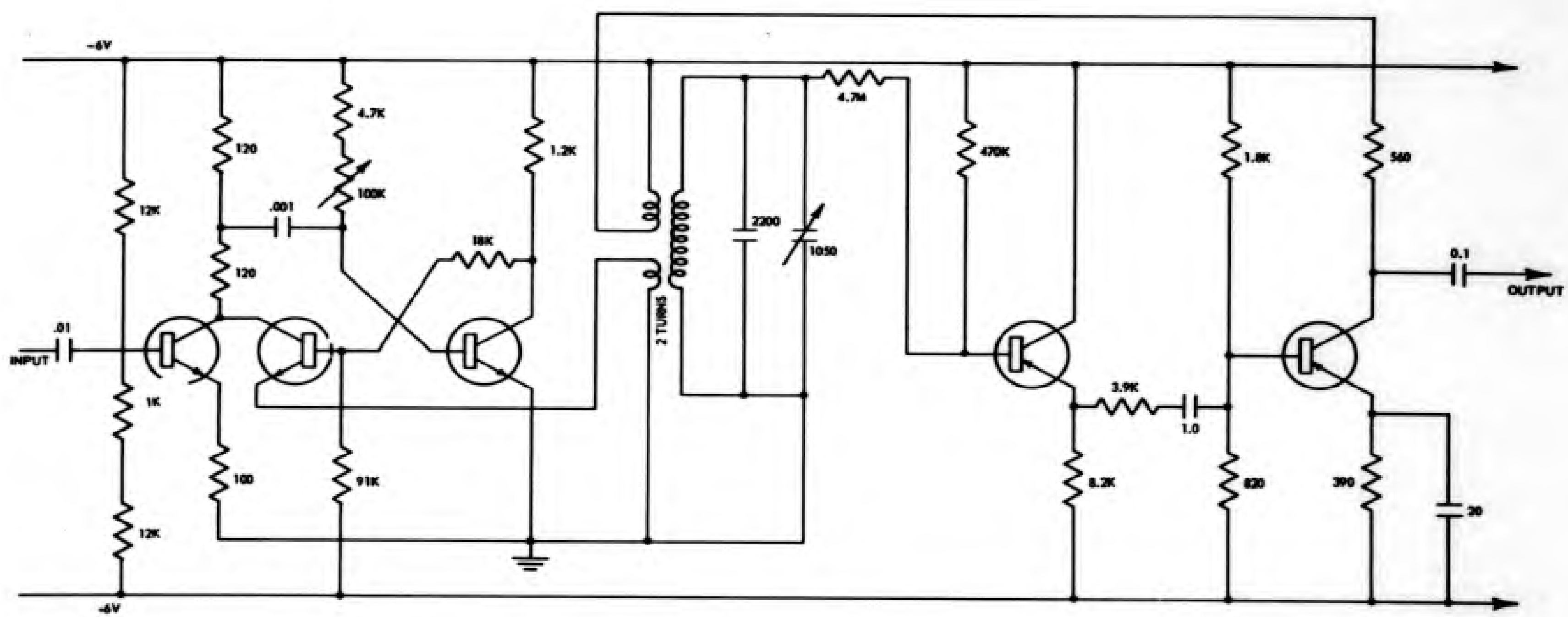


Figure 13. Locked Oscillator.

first transistor in the multivibrator is used simply as a means to parallel trigger the last two transistors, which form the multivibrator proper. Timing of the output pulse width is provided by proper selection of the coupling capacitor and is set for approximately 35 microsecond pulse width. The output pulse is taken as a current sample in the emitter of one of the transistors in the multivibrator. This current pulse is passed through a one-turn winding on the resonant circuit which controls the frequency of the locked oscillator. Due to the high "Q" of the oscillator coil, no appreciable circulating current flows in the oscillator tank circuit at frequencies only slightly removed from the resonant frequency of this circuit. Consequently, when the natural frequency of the oscillator is made nearly equal to one of the harmonic frequencies contained in the current pulse from the multivibrator, the only frequency component acting in the oscillator tank circuit to produce synchronization of the oscillator is that component which exists in the neighborhood of the resonant frequency of the oscillator circuit. For an interfering pulse rate in the neighborhood of 1,000 cycles, which is a typical value, the tenth harmonic of the pulse in the output of the multivibrator will produce across the tuning capacitor of the oscillator a voltage which is approximately 20 per cent of the voltage existing at this point due to the natural oscillation of the circuit. Since the "Q" of the oscillator coil is approximately 100, the frequency range over which synchronization can be effected may be calculated by the following relation:

$$BW_{(sync)} = \frac{\omega_o}{2Q_o}\left(\frac{V_s}{V_o}\right) \qquad (2)$$

where
$$\begin{cases} \omega_o = \text{natural frequency of oscillation} \\ Q_o = \text{Q of oscillator resonant circuit} \\ V_s = \text{amplitude of the synchronizing signal} \\ V_o = \text{amplitude of the natural oscillation} \end{cases}$$

From this relationship, it is seen that a synchronizing bandwidth for the tenth harmonic of approximately 60 cycles is obtained. This is sufficiently wide to accommodate small drifts in the repetition rate of the interfering pulse source without the need for manual readjustment of the oscillator frequency. As a practical matter, tests were made on the synchronizing action of the oscillator in the range 9 to 11 kc when the repetition rate of the interfering pulse signal was varied. It was found that satisfactory synchronization could be maintained when the repetition rate was as low as 50 cps. The oscillator proper consists of a two-stage transistor feedback amplifier in which the feedback is coupled to the tuned circuit by means of a small two-turn inductive link on the oscillator tank circuit. The input of the transistor amplifier is isolated from the tuned circuit by means of a 4.7 megohm resistor which serves to make impedance variations at the input of the amplifier have negligible effect on the tuned circuit. In addition, it permits the maintenance of a low amplitude of natural oscillation in the tuned circuit so as to provide the maximum synchronization bandwidth. The output is taken from the collector of the output amplifier in the oscillator circuit.

4.4.3. Variable Phase Shifter. The output of the locked oscillator is supplied to a three-stage variable phase shifter which positions the point of zero crossing of the sine wave output of the locked oscillator so the blanking pulses which are generated at these points of zero crossing may have their time position of occurrence varied. The construction of the phase shifter as illustrated in the schematic in the Appendix is a familiar RC network whose phase shift is variable from 0 to 180° but whose gain is independent of frequency. Since the time position of the output pulses must be varied over a range of almost one full period, it is necessary to provide 360° of phase shift to insure that the range is adequately covered. Since the 0 and 180° points in the phase shifter occur for the extreme values of RC = 0 and RC = ∞, it is necessary to provide three sections of phase shift to insure adequate coverage of the desired range without the necessity of operating at either of the two extremes. In the construction of the phase shifter, the phase in each section is controlled by one gang of a three-gang potentiometer so that a single control will give simultaneous variation of the phase shift in all three sections. Biasing for the three split loaded phase inverters is provided by a return of the base bias resistors to the collectors. This connection provides a degree of DC stability due to the negative feedback while at the same time maintaining a relatively high input impedance due to the large value of base biasing resistor used. The unbypassed emitter circuit also serves to raise the input impedance. Though this connection does not represent a necessarily optimum connection for obtaining high input impedance, it does supply the necessary unloading of the phase shift network and simultaneously gives the required phase inversion for driving the next succeeding phase shift network. Emitter followers are used at both the input and the output of the phase shifter in order to provide isolation of these terminals from external circuitry.

The output of the phase shifter is used to drive a Schmidt trigger circuit which fires at the zero crossing of the output of the phase shifter. This Schmidt trigger circuit in turn triggers a one-shot multivibrator; the proper trigger shape for this operation being obtained by means of a small differentiating transformer connected in the output of the Schmidt trigger. A diode is placed across the secondary of this transformer to remove the positive going portion of the wave form which might otherwise cause difficulty in triggering the one shot multivibrator.

The one shot multivibrator pulse width is adjustable from 5 to 35 microseconds so that variable width blanking pulses or sampling pulses may be generated. These output pulses are used to drive a pulse amplifier whose output stage is a power transistor capable of supplying the necessary 400 milliampere pulse to actuate the RF switch. Operation in either the blanking or the sampling mode is made possible by the simple expedient of reversing the connection of the output of the one shot multivibrator from one collector to the other.

4.4.4. Equipment Adjustment. The completed equipment is pictured in the photographs of Figures 14 and 15, which show the internal construction and parts layout as well as arrangement of controls on the front panel. With the exception of the auxiliary receiver and its local oscillator, the circuitry is constructed on phenolic boards which are arranged for simple removal from a central mounting structure in the center of the chassis. Each board is equipped with an easily removable plug which contains all connections to that particular board.

In order to place the equipment into operation, the antenna connection to the receiver is removed and instead is connected to the antenna output jack in the back of the chassis of the interference blanker. An output cable is then run from the output jack at the rear of the interference blanking chassis to the input jack of the receiver. Using the audio monitor output as an indication of proper tuning, the auxiliary receiver is tuned to the interfering pulse signal. Then, with the blanking pulse width set to its widest position, the PRF control which tunes the locked oscillator is varied until synchronization of the locked oscillator with the interfering pulse signal is obtained. Then by narrowing the

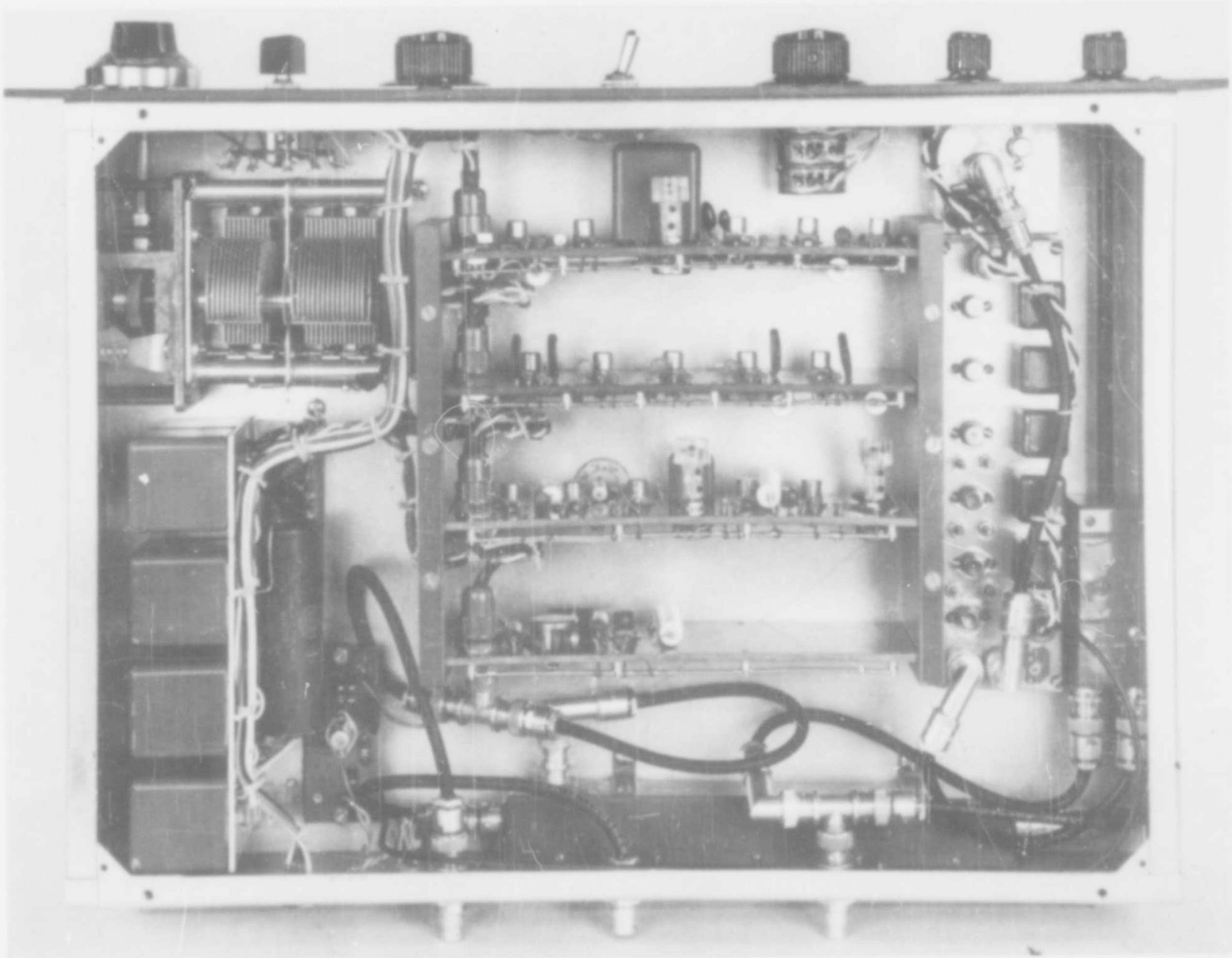


Figure 14. Internal Construction - Interference Blanker.

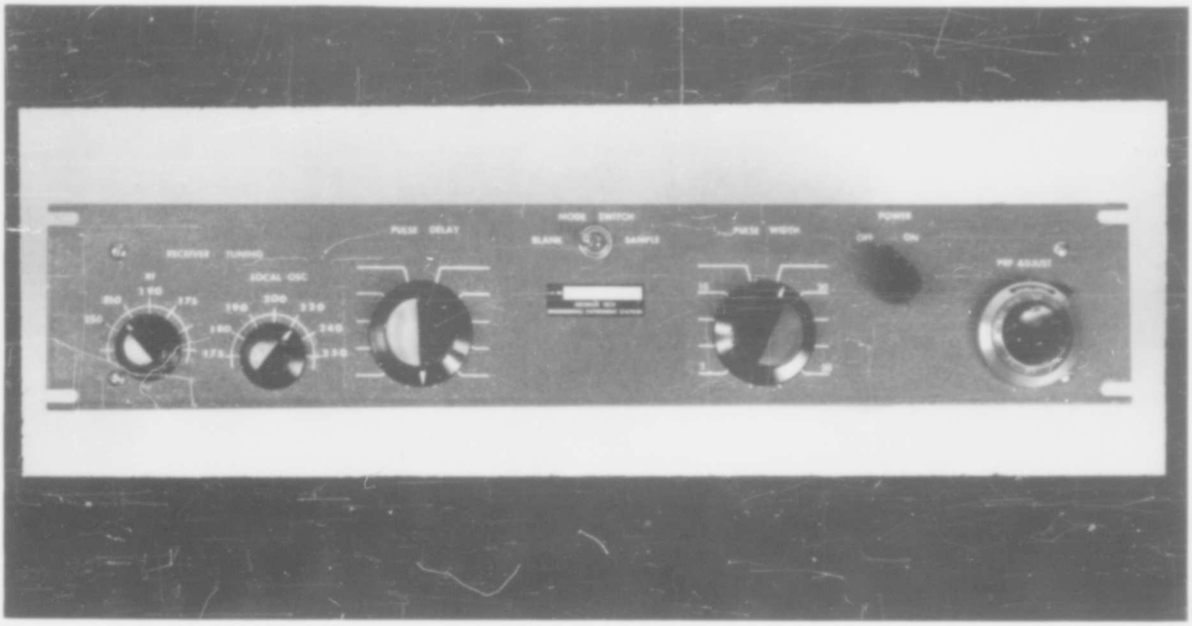


Figure 15. Front Panel - Interference Blanker

width of the blanking pulse and by adjustment of the pulse delay, proper position of the blanking pulse with respect to interfering pulse may be obtained. In this manner, interfering signals may be easily removed from the input to the receiver and excellent reproduction of desired signals at levels of a few microvolts can be obtained.

5. ACKNOWLEDGEMENTS

This program was conducted under the general supervision of W. B. Wrigley, who is head of the Communications Branch. P. A. Kresge also made significant contributions to the work described in this report.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

M. W. Long, Chief
Electronics Division

6. APPENDIX

6.1. Analysis of the Sampling Technique. The following analysis points out the relationship between the input and reconstructed AM signals in more detail. Referring to Figure 3, the output of the sampling switch is given by

$$E_o(t) = E_i(t) \quad T(t), \tag{3}$$

Where $E_i(t)$ is the input signal to the switching device and $T(t)$ is the switch transfer function. Since the switch is always on or off, the transfer function $T(t)$ must be of a form similar to Figure 16.

Here the switch is operating periodically at the frequency f_s. Because of this periodicity, $T(t)$ may be expanded in a Fourier series,

$$T(t) = \frac{A_0}{2} + \sum_{n=1}^{\infty} A_n \cos (n2\pi f_s t), \tag{4}$$

which, upon substitution in (1), gives

$$E_o(t) = E_i(t) \left\{ \frac{A_0}{2} + \sum_{n=1}^{\infty} A_n \cos (n2\pi f_s t) \right\}. \tag{5}$$

If $E_i(t)$ is a narrow band signal such as the AM signal of Equation (6),

$$E_i(t) = \left[1 + m \cos (2\pi f_m t)\right] \cos 2\pi f_o t. \tag{6}$$

Then $$E_o(t) = \left\{\left[1 + m \cos (2\pi f_m t)\right] \cos 2\pi f_o t\right\} \left\{ \frac{A_0}{2} + \sum_{n=1}^{\infty} A_n \cos (2\pi n f_s t)\right\} \tag{7}$$

or

$$E_o(t) = \frac{A_0}{2} E_i(t) + \sum_{n=1}^{\infty} \frac{A_n}{2} \cos (2\pi) (f_o \pm nf_s)t$$

$$+ \sum_{n=1}^{\infty} \frac{mA_n}{4} \cos (2\pi) (f_o + f_m \pm nf_s)t \tag{8}$$

$$+ \sum_{n=1}^{\infty} \frac{mA_n}{4} \cos (2\pi) (f_o - f_m \pm nf_s)t.$$

The band-pass filter removes all except the first term so that the output signal is

$$E_1(t) = \frac{A_0}{2} E_i(t). \tag{9}$$

From Equation (9), it is seen that the desired signal gain is

$$\text{Gain} = \frac{A_0}{2} \tag{10}$$

which is simply the average value of the transfer function of the switch. For the rectangular pulses of Figure 16,

$$\frac{A_0}{2} = \text{average value of } T(t) = \frac{\Delta}{T} = \text{duty cycle of the switching signal.} \tag{11}$$

6.2. Spurious Responses. In sampling or blanking the front end of a receiver in order to reduce the effects of pulse interference, several undesirable responses arise which are due to the switching process. These responses can be eliminated only if sufficient selectivity is placed ahead of the switching element to reject signals whose frequencies lie on one of these response frequencies.

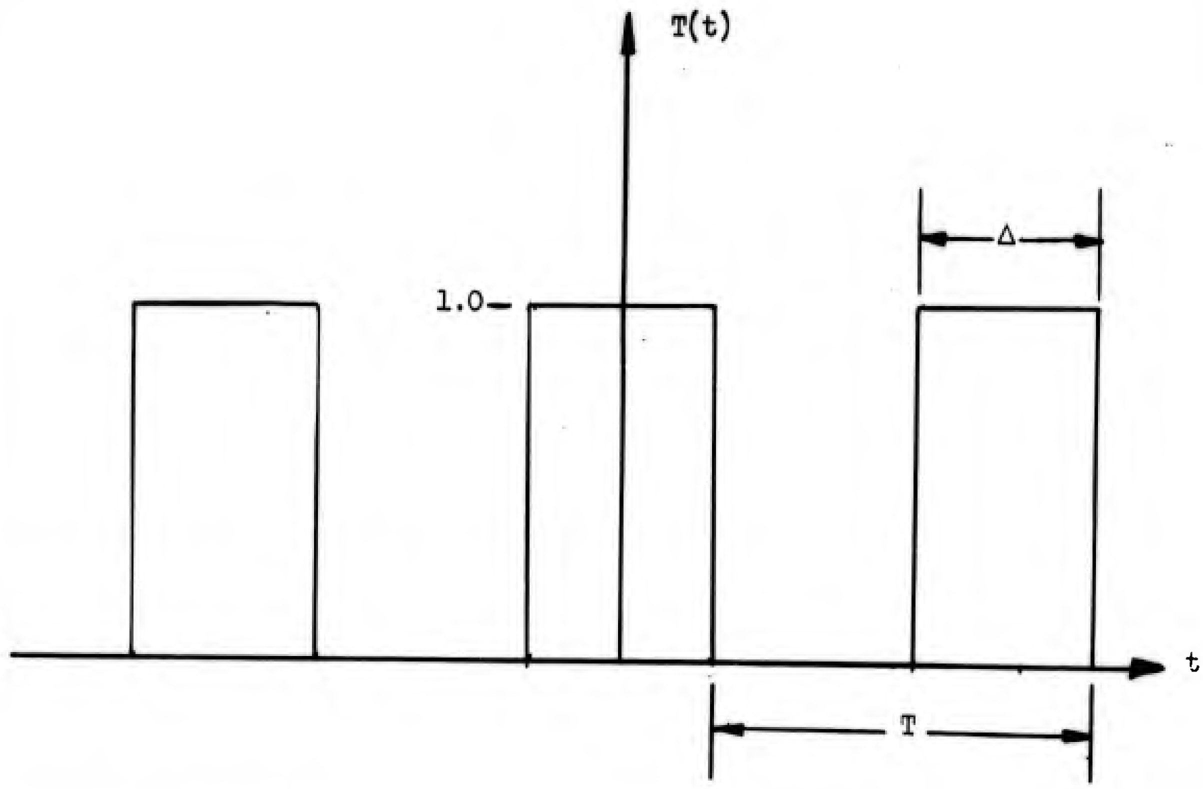


Figure 16. Sampling Switch Transfer Function

These responses arise because of the switching at the blanking or sampling rate. The following analysis points out the location and relative magnitudes of these responses:

If, referring to Equation (5),

$E_i(t)$ is a single tone signal of frequency f_i, i.e. (12)

$E_i(t) = B \cos (2\pi f_i t)$,

then

$$E_o(t) = B \cos(2\pi f_i t) \left\{ \frac{A_0}{2} + \sum_{n=1}^{\infty} A_n \cos(2\pi n f_s t) \right\} \tag{13}$$

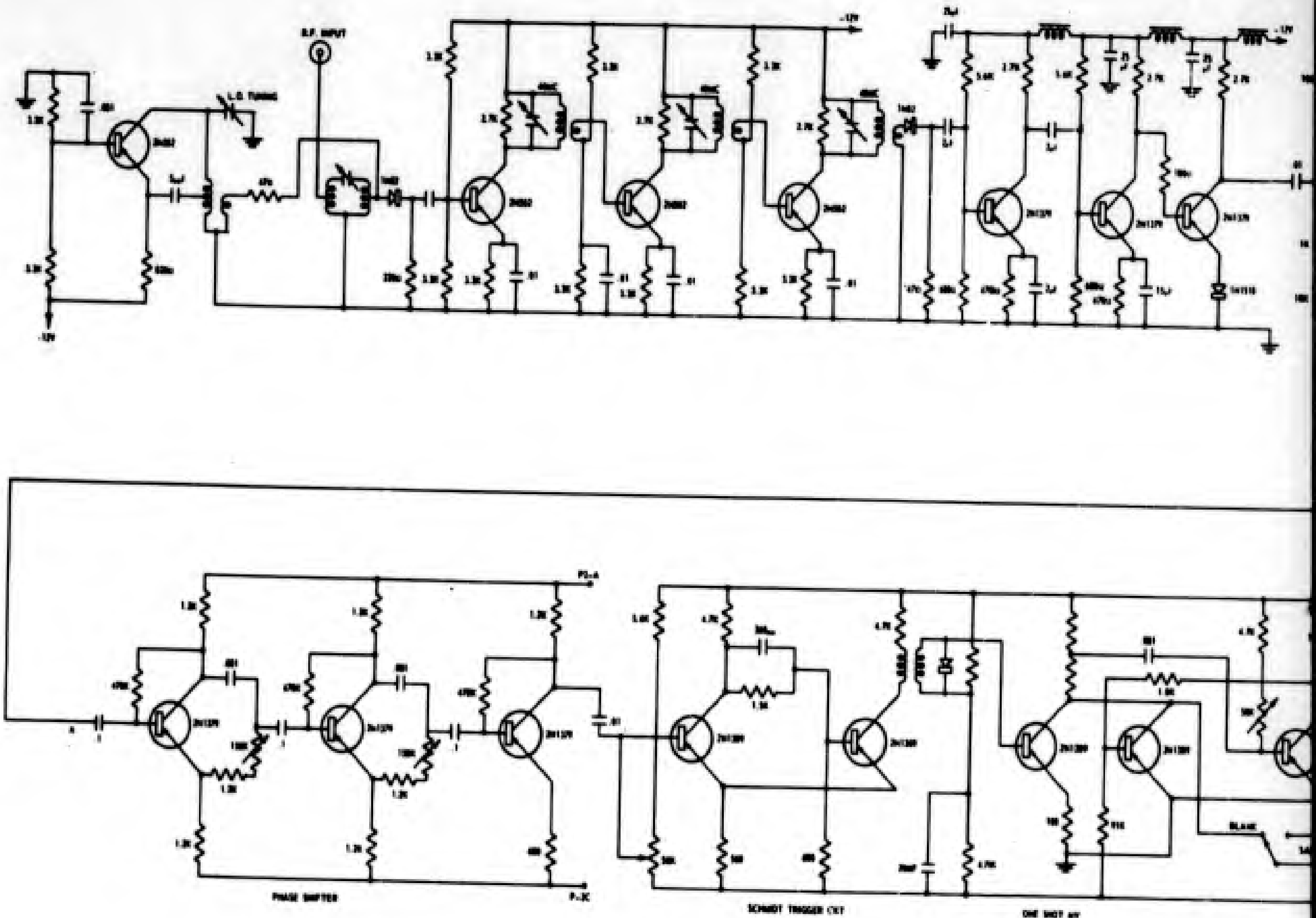
or

$$E_o(t) = \frac{A_0}{2} B \cos(2\pi f_i t) + \sum_{n=1}^{\infty} \frac{A_n B}{2} \cos 2\pi(f_i \pm n f_s)t. \tag{14}$$

Now the frequencies, f_i, are determined so that $E_o(t)$ has the components at the tuned frequency of the receiver, f_o. There are two conditions which satisfy this requirement. They are:

(1) $f_i = f_o$ — This is the output of the switching device due to a signal at the operating frequency, f_o.

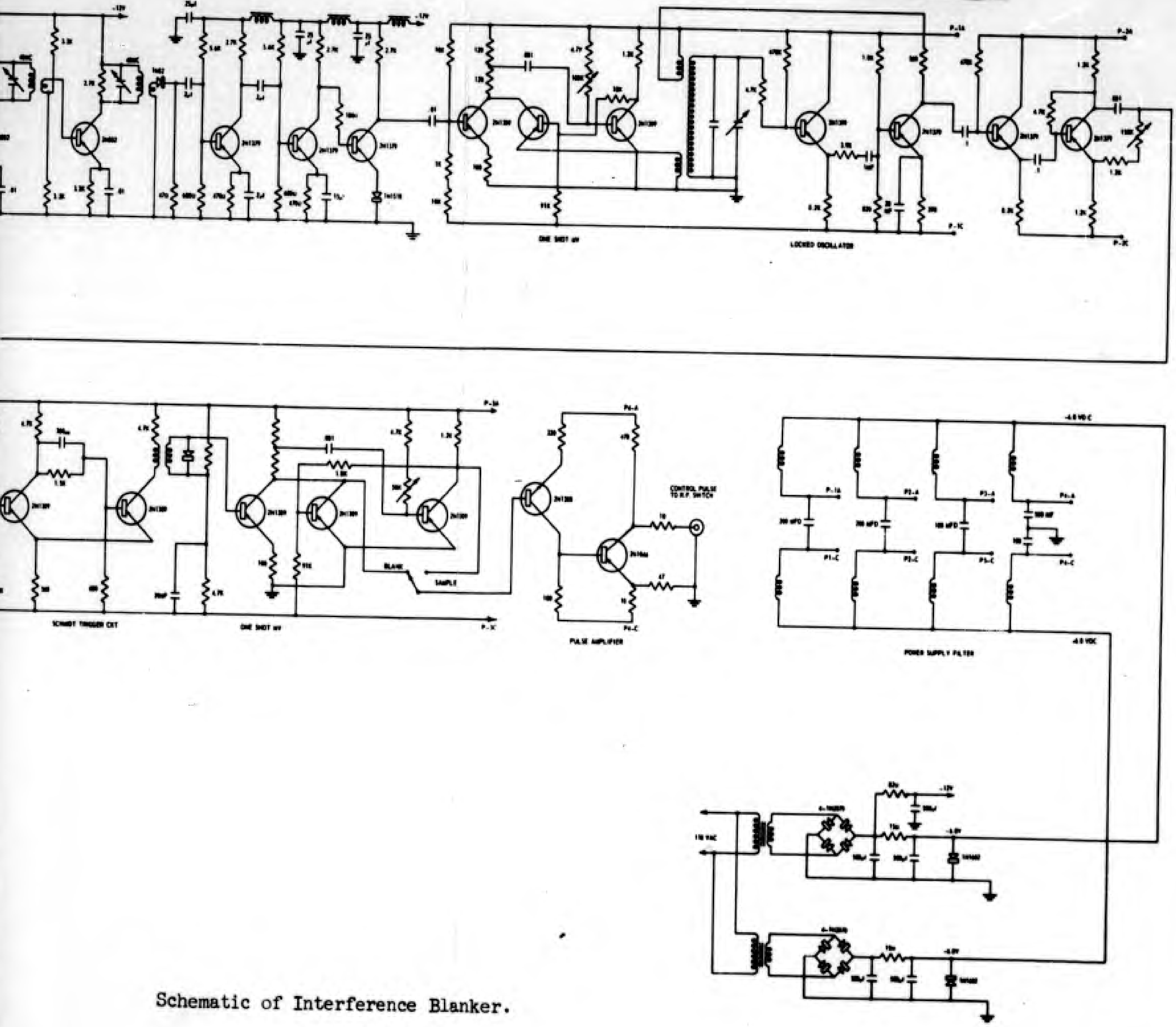
(2) $f_i = f_o \pm nf_s$ — These values of f_i give the frequency positions of the spurious responses, the relative amplitudes being determined by the Fourier Coefficients, A_n, in Equation (4).



1

Schematic of Interference B

2



Schematic of Interference Blanker.

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