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THE MTI CONVERSION UNIT FOR SC-SK SERIES RADAR

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THE MTI CONVERSION UNIT FOR SC-SK SERIES RADAR

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March 2, 1948

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Glossary of Symbols

- c = Velocity of propagation of electromagnetic waves in a vacuum
- D = Target range
- F = Antenna height gain factor
- F_c = Height gain factor for clutter
- F_t = Height gain factor for target
- f = Frequency
- f_o = Transmitted frequency
- f_d = Doppler frequency
- f_r = Received frequency
- G = Antenna gain
- M = Pulses on target per scan
- N = Antenna rotation speed
- P_{tr} = Transmitter peak power
- P_r = Received echo peak power
- ΔP_r = Pulse-to-pulse change in received echo power
- (PRF) = Pulse repetition frequency
- R = Subclutter visibility - Numerical Power Ratio of clutter level to minimum detectable signal in that clutter
- R_{max} = Maximum permissible subclutter visibility
- S = Sensitivity
- S_r = Relative sensitivity
- T_d = Delay time of returning echo
- t = Time variable
- V_o = Velocity of propagation of wave motion under consideration
- $V_{max}^{(n)}$ = Speed for maximum response where n is the order of the maximum
- $V_{null}^{(n)}$ = Speed for no response where n is the order of the null
- ν = True velocity
- $\nu^{(r)}$ = Radial component of velocity
- ν_r = Receiver velocity
- ν_s = Ships velocity

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v_t = Target velocity

θ = Angle (Spatial)

θ_b = Beamwidth of radar antenna

θ_c = Angle between the target's course and the radius from ship to target

θ_r = Relative bearing of antenna beam

σ = Radar area

σ_c = Radar area of clutter

σ_t = Radar area of target

τ = Radar pulse length

ϕ = Electrical phase angle

ϕ_{tr} = Starting phase of radar transmitter

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ABSTRACT

A coherent pulse doppler MTI (Moving Target Indicator) system has been developed for application to the SC-SK series radars to eliminate echoes produced on the indicators by large areas of land clutter when attempting to detect and track attacking enemy planes. Basic theory, design, and parameters associated with the MTI system are fully discussed and are shown to depend on the elementary Doppler Shift equation. The important mathematical relations for the design of a coherent pulse doppler MTI are developed along with a theoretical discussion, and the equations are tabulated in an Appendix for convenient referencing.

The actual MTI conversion unit used for the SC-SK radar systems is presented and discussed in detail. It consists of two major components; one a new plate modulator for the transmitter, and the other the receiving and cancellation systems. The MTI system achieves a cancellation ratio of 30 db and a subclutter visibility of about 25 db. Maximum range is unaffected by the addition of the MTI. This MTI does, however, provide "solid" aircraft coverage in areas of clutter. System reliability has proven to be excellent.

Inasmuch as the main test of the system is covered in another report, the system performance is discussed only briefly.

PROBLEM STATUS

This is an interim report on this problem; work is continuing.

AUTHORIZATION

NRL Problem No. R02-24 (BuShips Problem S1055R-C).

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THE MTI CONVERSION UNIT FOR SC-SK SERIES RADAR

INTRODUCTION

Problem

During the closing months of the recent war, naval operations had to be carried out in the vicinity of large land masses in the Pacific. As a result, serious losses in the efficiency of radar in detecting attacking enemy planes occurred because large areas of land clutter echoes appeared on the indicators. To combat this loss in radar efficiency, an accelerated program of development of Moving Target Indicator equipment for use with existing radar systems was undertaken in the hope of getting usable results to the fleet by the time the main Japanese Islands were to be invaded. The early ending of the war obviated the need for any such equipment reaching the operational theatres and resulted in the substitution of a more orderly and systematic research program for the original crash program.

In connection with the early crash program, the Naval Research Laboratory was given the problem of developing an MTI conversion unit for the SC-SK series air search radars. A coherent pulse doppler type of system with a supersonic mercury-delay line cancellation system was chosen for this crash program because it fitted the basic radar system, it had been proven at Radiation Laboratory to be basically operable, and it was the one type of MTI capable of operating with PPI at normal scanning rates.

The first developmental model conversion unit, built under the crash program, was completed and put into operation in July just before the Japanese surrender. This unit, though it gave usable results, was quite critical to adjust and generally difficult to operate. Thus, with the ending of the war and the reduction in priority of the problem, it was decided to build a second improved model incorporating all the lessons learned in the construction of the first model. It is this second model which is described in this paper.

Attention should be called to the fact that although the SC-SK series radars are at least obsolescent (if not obsolete) the value of this work is not affected by the obsolescence of these radars. The MTI Conversion Unit can be applied with little change to the SR and SR-A series radars, and, with the inclusion of "velocity compensation by coho shift," and a change in the r-f head, the Conversion Unit can be adapted to other radars of the SR series.

References

NRL Report R-2480,¹ on anticlutter devices, led to this problem; Reports R-3065 and R-3138² describe the pressurized mercury delay line and the sea trial of the final conversion unit respectively.

¹ I. H. Page, NRL Report R-2480, "A Study of Radar Anti-Clutter Devices for Naval Use," 1 March 1945.

² T. H. Chambers, NRL Report R-3065, "A Pressurized Mercury Delay Line for Fleet Service," 28 April 1947; T. H. Chambers, NRL Report R-3138, "Sea Trials of the MTI Conversion Unit for SC-SK Series Radar," July 1947.

THEORETICAL CONSIDERATIONS

The Doppler Shift

A well-known phenomenon called the doppler shift results when the frequency of a wave from a stationary transmitter is measured from a moving platform. This shift in received frequency results from the fact that the number of cycles received in unit time is increased or decreased (depending on the direction of motion of the platform) by the number of wavelengths that the platform moves relative to the transmitter. The doppler frequency is then defined to be the difference between received and transmitted frequencies, or

$$f_d = f_r - f_o \quad (1)$$

where,

f_d = doppler frequency,
 f_r = received frequency,
 f_o = transmitted frequency.

In the case where the receiver is in motion, the following relation may be deduced:

$$f_{r_1} = \left[1 + \frac{\nu_r(r)}{V_o} \right] f_o \quad (2)$$

where,

$\nu_r(r)$ = radial speed of platform toward transmitter,
 V_o = velocity of propagation of waves.

Similarly, if the transmitter is set in motion and the receiving platform held stationary another shift will be observed. The resultant frequency is found from the equation:

$$f_{r_2} = \left[\frac{V_o}{V_o - \nu_t(r)} \right] f_o \quad (3)$$

where,

f_{r_2} = received frequency with transmitter in motion,
 $\nu_t(r)$ = radial speed of transmitter toward receiver.

Doppler in Radar Systems

In the case of a radar system, the target acts first as a receiver thus introducing a shift in apparent frequency according to equation (2). The apparent received frequency is then

$$f_{r_1} = f_o \left[1 + \frac{\nu_r(r)}{V_o} \right].$$

The target then reradiates this received frequency as a transmitter and introduces a second shift according to equation (3).

$$f_{r_2} = f_o \left[1 + \frac{\nu_r(r)}{V_o} \right] \left[\frac{V_o}{V_o - \nu_t(r)} \right]. \quad (4)$$

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This frequency is then the actual frequency that will be received by the radar from a moving target. If now we replace V_0 by c (the speed of electromagnetic waves),

$$f_{r_z} = f_0 \left[1 + \frac{\nu_r(r)}{c} \right] \left[\frac{c}{c - \nu_t(r)} \right]. \quad (4.1)$$

Furthermore, it will be noted that for the case of practical target velocities,

$$\nu_r(r) = \nu_t(r) \ll c.$$

Thus we may write,

$$f_{r_2} \cong f_0 \left[1 + \frac{2 \nu_t(r)}{c} \right]. \quad (4.2)$$

Substituting this in equation (1),

$$f_d \cong \frac{2 \nu_t(r)}{c} f_0. \quad (5)$$

Thus, the returning echo will differ in frequency from the transmitted signal by the amount f_d .

It is of interest to note the order of magnitude of this doppler frequency in a practical radar system. Assuming the radar to be one of the SC-SK series operating at 200 Mc and assuming a plane speed of 200 mph,

$$2 \times \frac{200}{3600} \times \frac{200 \times 10^6}{186,000} = 119 \text{ cps.}$$

Thus we see that the doppler frequencies will be very low, of the order of hundreds of cycles in a practical system. It is to be noted that doppler frequency is dependent only on system frequency and target velocity and is proportional to both of these. Thus, the doppler frequency for any system can be found with no knowledge of the system's characteristics other than its r-f frequency.

Detection of Doppler Shift

From the estimate made above of the amount of the doppler shift, it can be seen that the detection of this shift will not be an easy problem since it amounts to only about one part in one million. There are, however, several means which may be used to detect it. Perhaps the simplest of these is to beat the returning signal against the radiated signal as would be done in the case of c-w radar. The c-w radar is not, however, a good radar for general use because of its lack of range information. It is possible to secure range information from a c-w system by using modulation, but the required circuitry becomes complicated, the maximum scanning speed is very low, and the MTI characteristics of an unmodulated c-w radar are lost. For these reasons, and because of the accepted superiority of the pulsed radar for general military applications, only this latter type has been considered at this Laboratory to be suitable for MTI use.

When an attempt is made to detect the presence of a doppler shift within a given pulse, it is found that, assuming a 5-microsec pulse as in the SC-SK series, the only effect of the

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doppler is to introduce a relative phase rotation between the two signals of:

$$\phi = 119 \times 5 \times 10^{-6} \times 2\pi = .003 \text{ radians.}$$

Such a phase rotation would obviously be very difficult to detect. It appears best to detect the doppler shift by pulse-to-pulse comparison rather than by comparing from start to finish of one pulse.

To make this pulse-to-pulse comparison, it is necessary to establish a reference against which the phase of the returning echo may be compared. This may be done by means of a so-called coherent oscillator (coho). This oscillator is simply a stable r-f oscillator which operates at transmitter frequency and is pulled into phase with the transmitter each time a pulse is radiated. Hence, this coherent oscillator delivers essentially an extension of the transmitter pulse. Its output is therefore,

$$e_{\text{coho}} = \cos(2\pi f_0 t + \phi_{\text{tr}}) \quad (6)$$

during the interval,

$$t_0 \leq t < t_0 + \frac{1}{(\text{PRF})}.$$

The returning echo is, however, of the form,

$$e_{\text{sig}} = \cos[2\pi f_0(t - T_d) + \phi_{\text{tr}}], \quad (7)$$

where

$$\begin{aligned} f_0 &= \text{transmitter frequency,} \\ \phi_{\text{tr}} &= \text{transmitter starting phase,} \\ T_d &= \text{delay time of returning echo.} \end{aligned}$$

If now we add these two signals together during the interval,

$$T_d \leq t \leq T_d + \tau$$

where,

$$\tau = \text{Radar Pulse Length,}$$

we will get an output signal of the form,

$$e_{\text{coho}} + e_{\text{sig}} = 2 \cos(2\pi f_0 T_d) \cos(2\pi f_0 t + \phi). \quad (8)$$

The first cosine term is the amplitude factor, called A, and is the term of interest. (The second cosine term is removed by action of the detector). From this term, it is obvious that the amplitude of the output pulse will be dependent on T_d . But T_d is simply a measure of target range; thus,

$$T_d = \frac{2D}{c}, \quad (9)$$

where,

$$D = \text{target range.}$$

Now substituting for T_d in the amplitude factor of equation (8):

$$A = 2 \cos\left(\frac{4\pi f_0 D}{c}\right). \quad (10)$$

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From this it is obvious that for a target whose range is constant, the amplitude of the output pulse will remain constant. But for a target with radial motion, where D is changing, the amplitude will vary.

The variation of the amplitude factor with target speed may be shown as follows:

Assume that

$$D = D_1 + \Delta D = D_1 + \nu_t^{(r)} t,$$

where,

D_1 = an arbitrary starting point,

$\nu_t^{(r)}$ = target radial velocity.

Then,

$$A = 2 \cos \left[\frac{4\pi f_0 \nu_t^{(r)} t}{c} + \frac{4\pi f_0 D_1}{c} \right].$$

Now choose D_1 such that the second term in the brackets is an even multiple of π .

Then,

$$\begin{aligned} A &= 2 \cos \frac{4\pi f_0 \nu_t^{(r)} t}{c} \\ &= 2 \cos 2\pi \left[\frac{2\nu_t^{(r)}}{c} f_0 \right] t. \end{aligned} \quad (11)$$

Thus it is shown that the envelope of the returning pulse is nothing more than a cosine wave with a frequency,

$$f = \frac{2\nu_t^{(r)}}{c} f_0. \quad (12)$$

If we compare this with equation (5) we see that it is in actuality the doppler frequency showing up as the envelope of the resultant of the addition of the returning echo signal and the coho.

Further examination of this procedure shows that it is a sampling technique wherein the system has the same characteristics as a c-w system except that, instead of examining the envelope continuously, it is sampled at discrete time intervals. This sampling leads to some very important system characteristics, but before considering these there is an important consideration which should be taken up.

During the course of the above discussion the coho has been assumed to be at radio frequency. Suppose that it is put at the receiver intermediate frequency instead. Though the mathematics involved will not be given here, the important point is that the shifting of the relative phases of coho and signal, as shown by equations (6) and (7) is done at radio frequency, and once this phase shift is introduced, it is preserved unchanged through the frequency conversion. Hence the amplitude factor in equation (8) is the same regardless of where the coho is introduced.

The Consequences of the Sampling Procedure

The "sampling" technique used in examining the doppler introduces some important characteristics into the system. The amplitudes of successive pulses are constant for fixed targets but varying for moving targets. Hence, the separation of moving and fixed targets is made by comparing the amplitudes of these successive pulses and the sensitivity of the

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overall system is dependent on the rate of change in pulse amplitude. Although this sampling phenomenon could be investigated mathematically to determine sensitivity as a function of target speed, the mathematics becomes quite complicated and will not be gone into here. The results will simply be presented and discussed.

Investigation shows that the curve of sensitivity of the MTI system versus target speed is a rectified sine wave in which the nulls come at speeds where the doppler frequency is an integral multiple (including zero) of the system PRF. Thus the nulls come at,

$$\frac{2 V_{\text{null}}^{(n)}}{c} f_0 = n (\text{PRF}),$$

where,

$V_{\text{null}}^{(n)}$ = speed for zero response,

(PRF) = system pulse repetition frequency,

n = any integer (including zero),

or,

$$V_{\text{null}}^{(n)} = \frac{n (\text{PRF}) c}{2 f_0}. \quad (13)$$

The equation for response versus target speed is then obviously the absolute value of a sine wave which goes to zero when $\nu_t^{(r)}$ has the value V_{null} . Thus,

$$S_r = \left| \sin \frac{2 \pi \nu_t^{(r)} f_0}{(\text{PRF}) c} \right|, \quad (14)$$

where,

S_r = relative sensitivity.

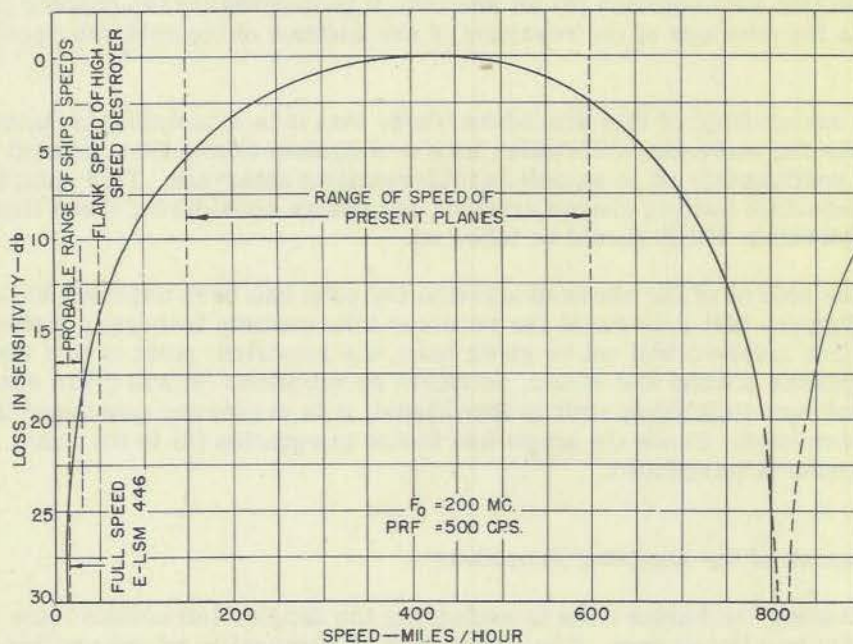


Fig. 1 - Graph Showing Loss in Sensitivity vs Speed
SC-SK MTI

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Such a curve is shown in Figure 1 with the relative response scale plotted in db. This curve is for the SC-SK MTI system. This system has the following characteristics:

$$\begin{aligned} \text{PRF} &= 500 \text{ cps} \\ f_0 &= 200 \text{ Mc.} \end{aligned}$$

Thus, for the first null,

$$V_{\text{null}}^{(1)} = \frac{500 \times 186,000 \times 3600}{2 \times 200 \times 10^6} = 837 \text{ mph.}$$

The response curve is shown dotted in the vicinity of this null because in the operation of an actual system, it will be found that the nulls will be "filled in" by propeller modulation and other effects which modulate the echo.

Velocity Compensation

In the last section it was pointed out that both radar itself and the target act both as transmitter and as receiver. Because of this symmetry of the path of the radio waves, it is obvious that there will be no difference between the effect of motion of the target and motion of the radar. (Note that for this to hold true, the reduction of equation (4.1) to (4.2) must be possible, thus v must be smaller than c ; as it is in the radar case.) Thus the only important motion is the relative motion of the target and the radar. Hence if the radar carrying vehicle is in motion all stationary targets except those dead ahead have relative motion and become, as the radar sees them, moving targets. Since this would defeat the purpose of the MTI, a method must be found to compensate for it.

There are two basic methods of compensating for the motion of the radar when used on shipboard. The first of these, and by far the simpler, relies on the wide difference between the speeds of ships and of planes. Ordinarily, planes will travel at speeds of several hundred miles per hour, whereas ships will travel at speeds of 40 miles per hour or less. Thus, if system parameters are selected so as to put the first peak in the response curve at 400 mph or higher, the response to ordinary ship speeds will be about 18 db down from peak response. This will allow a subclutter visibility of slightly less than 18 db in the system. Unfortunately, however, the highest possible ship speeds will give responses only 12 db down, and since this will severely limit the subclutter visibility, it will considerably reduce the usefulness of the MTI system.

The requirements which must be met by the system parameters to compensate in this manner are quite easily determined. The velocity at the first null in response is just twice the velocity at the first peak; thus, from equation (13) we get,

$$V_{\text{null}}^{(1)} = \frac{(\text{PRF}) c}{2 f_0} \gg 800 \text{ mph.}$$

Solving this for $f_0 / (\text{PRF})$, we get,

$$\frac{f_0}{(\text{PRF})} \ll 4 \times 10^5. \quad (15)$$

This condition is quite easily met for low frequency systems, but for high frequency systems the required (PRF) becomes so high as to restrict the range. Thus, in the present state of the art, this method of compensation cannot generally be used with higher frequency systems (L band and higher).

A second possible method of compensation is the use of special electronic circuits which introduce into the coho signal a phase rotation which is just equal in magnitude and direction to the phase rotation introduced into the returning echo signal by the ships motion. Then, when the comparison between the coho and returning echo is made, the phase rotation due to ships motion will be cancelled out and the effect of that motion will be eliminated.

In actual operation the introduction of this phase rotation becomes, because of its continuous additive nature (a constant amount is added between adjacent radar pulses), an actual frequency shift. This frequency shift is equivalent to the doppler shift introduced by the motion of the ship. Thus, it will be given by equation (5) and will be,

$$f_d \cong \frac{2\nu_s^{(r)}}{c} f_0, \quad (16)$$

where,

$$\begin{aligned} f_d &= \text{necessary frequency shift,} \\ \nu_s^{(r)} &= \text{component of ships velocity in direction of target,} \\ c &= \text{velocity of light,} \\ f_0 &= \text{radar frequency.} \end{aligned}$$

But,

$$\nu_s^{(r)} = \nu_s \cos \theta_r, \quad (17)$$

where,

$$\begin{aligned} \nu_s &= \text{ships forward speed,} \\ \theta_r &= \text{relative bearing of antenna.} \end{aligned}$$

Substituting for $\nu_s^{(r)}$, equation (16) becomes:

$$f_d \cong \frac{2\nu_s \cos \theta_r}{c} f_0. \quad (16.1)$$

Thus, this frequency must be added to the coho to compensate for the ships motion.

Unfortunately, this system does not give perfect velocity compensation. In the above discussion, θ_r has been taken as the antenna bearing, but it is immediately evident that, because of the finite beamwidth of the antenna, not all targets will be on that bearing; thus, the ships apparent radial speed $\nu_s^{(r)}$ will not be the same for all targets in the beam at a given time. This effect puts a limitation on the degree of compensation attainable and therefore limits the possible subclutter visibility. Since it is a fundamental limitation on the system, it will be discussed in the next section of this paper.

SYSTEM LIMITATIONS - GENERAL

Radial Velocity

Throughout the previous part of this paper, it was assumed that the target was moving radially with respect to the radar system. Since the system is sensitive to radial velocity only, the component of target velocity which is radial to the radar will be the only one which produces indications on the MTI. Thus, as far as a doppler system is concerned, the target will have a radial speed of:

$$\nu_t^{(r)} = \nu_t \cos \theta_c, \quad (18)$$

where,

$$\begin{aligned}\nu_t^{(r)} &= \text{radial speed,} \\ \nu_t &= \text{true speed,} \\ \theta_c &= \text{angle between the target's course and the radius} \\ &\quad \text{from ship to target.}\end{aligned}$$

This effect could be taken into account in Figure 1 (thus extending this figure to cover motion in any direction) simply by dividing the speed scale by the factor,

$$|\cos \theta_c|.$$

Or, it may be taken into account in equation (14), thus obtaining the equation (14.1) for the sensitivity of the system in terms of plane speed in any direction,

$$S_r = \left| \sin \frac{2\pi \nu_t \cos \theta_c}{(\text{PRF}) c} f_0 \right|. \quad (14.1)$$

From the above discussion two important facts can be seen. First, there will be an apparent general shifting of the true speed scale in Figure 1 to higher speeds for motion which is not truly radial, and second, there will be certain combinations of speed and angle of motion which will give response nulls. Unfortunately these two factors must, in the case of an MTI, be added to the already long list of factors which must be taken into account in selecting appropriate parameters for the system. Fortunately, however, they do not introduce serious limitations. In the case of the first factor (the general shifting of the curve to higher true speeds) the shift is not great over the range of angles that is important. In military operations, radar targets which are not headed toward the radar ship cannot, in general, cause damage, and they therefore become of lesser importance. This cannot be construed as meaning that they can be neglected, since ships which are not detected until they are within gun range are an exception to this rule. However, it is reasonably safe to neglect targets which have an angle of more than 60 degrees to the radial. This 60 degree angle amounts to a shift in the true speed scale of 2 to 1. Thus if the system is being designed to have a relatively high speed of maximum response, care must be taken to see that it is not set so high that this factor opens an avenue of approach to an attacking enemy.

In the case of the second factor (the nulls at various speeds and angles) the situation again is not too serious. Fortunately, there are many features which tend to modulate the returned echo from a target (notable among these is the propeller modulation in the case of an aeroplane) and, while this modulation is not due to doppler frequencies, it is still effective in providing an echo visible through the cancellation system. While these nulls would theoretically go to zero, in actual use it has been reported that they cause a loss in sensitivity of 6 to 20 db and if subclutter visibility is higher than this, the target can still be seen.

Subclutter Visibility

The subclutter visibility attainable will in general determine whether MTI will or will not be useful in a given system.

Consider the equation for radar performance,

$$P_r = \frac{G^2 P_t \sigma \lambda^2 F^2}{(4\pi)^3 D^4}, \quad (19)$$

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where,

P_r = power received at the receiver terminals for the given target,
 G = antenna gain,
 P_t = transmitted power,
 σ = radar area of the given target,
 λ = wavelength,
 D = target range,
 F = height gain factor.

If now we substitute the appropriate factors and find the ratio of clutter strength to target echo strength we get,

$$\frac{P_{rc}}{P_{rt}} = \frac{G^2 P_t \lambda \sigma_c F_c^2}{(4\pi)^3 D^4} \div \frac{G^2 P_t \lambda \sigma_t F_t^2}{(4\pi)^3 D^4} \quad (20)$$

where the subscript t indicates factors concerned with the moving target and the subscript c indicates factors concerned with the clutter. Since all factors except P_r , σ , and F are obviously the same in both numerator and denominator, equation (20) reduces to:

$$R = \frac{P_{rc}}{P_{rt}} = \frac{\sigma_c F_c^2}{\sigma_t F_t^2} \quad (20.1)$$

where,

R = ratio of clutter strength to strength of desired echo.

Thus, the clutter return will exceed the return from the desired signal by the ratio R and, if the signal is to be seen, the subclutter visibility, expressed as a ratio, must quite obviously exceed this value of R .

A study of equation (20.1) shows that there are four factors which determine the value of R . The echoing areas and the height gain factors for both the target and the clutter. Unfortunately, sufficient work has not been done to make it possible to evaluate quickly these four factors on paper in any given case. Rather it will generally be necessary to run actual tests to evaluate either these factors, or the ratio R directly. There are, however, certain cases where useful simplification can be made.

In cases where the altitude of the target and the clutter are the same, (or approximately so) such as the case with low flying planes, or submarine periscopes, the factors F_t and F_c become the same and equation (20.1) reduces to:

$$R = \frac{\sigma_c}{\sigma_t} \quad (21)$$

Thus, it is necessary to know only the relative echoing areas of the target and the clutter.

In any event, regardless of how the value of R is found, if the MTI is to have any value in the system under consideration, the subclutter visibility must exceed this ratio R . Thus, before plans are made to incorporate MTI in a radar system, the value of R for the types of clutter and the types of targets expected must be determined. If then the expected subclutter visibility, as limited by the cancelling system, etc., does not exceed R by several db, the MTI will not render the desired targets visible, and its inclusion in the system may not be warranted.

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The first limitation on subclutter visibility is the cancellation ratio of the cancelling system. Quite obviously subclutter visibility cannot exceed cancellation ratio, since such a condition would require discrimination between very small uncanceled moving targets and larger uncanceled residual from stationary targets.

A second fundamental limitation on subclutter visibility is the antenna scanning rate. To a first approximation we may assume that the increase in antenna gain is linear between the 3-db (one way pattern) point and the maximum. This being the case, it is then quite obvious that the scanning of the antenna will cause a variation in echo power of

$$\frac{\Delta P_r}{P_o} = \frac{.75}{.625 M} \times 2 = \frac{2.4}{M},$$

where,

ΔP_r = pulse-to-pulse change in receiver echo power,
 P_o = mean echo power level,
 M = pulses on target per scan.

Thus, the scanning of the antenna introduces a modulation of all echoes, and if this modulation is not to render all targets visible as moving targets, the system subclutter visibility must be:

$$R_{\max} < \frac{P_o}{\Delta P_r},$$

where,

$R_{\max}$ = maximum permissible subclutter visibility.

Then, since

$$R_{\max} < \frac{M}{2.4}$$

and since,

$$M = \frac{60 (\text{PRF}) \theta_b}{360 N},$$

where,

(PRF) = pulse repetition frequency,
 θ_b = antenna beamwidth to the 1/2 power points
 (one way pattern) in degrees,
 N = antenna rotation speed in rpm,

we may write:

$$R_{\max} < \frac{(\text{PRF}) \theta_b}{15 N}. \quad (23.1)$$

This scanning limitation is considered in more detail by A. G. Emslie.³ It is interesting to note that, using a Gaussian antenna pattern, he obtains,

$$\frac{\Delta E_r}{E_o} = \frac{1.66}{M}$$

and therefore,

$$\frac{\Delta P_r}{P_o} = \frac{2.75}{M}$$

Thus, the error introduced by the simplifying assumptions used above is less than 15 percent.

³ A. G. Emslie, MIT Rad Lab Report 1080, "Moving Target Indication on MEW," 19 February 1946

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Another important limitation on subclutter visibility arises from the compensation of the system for the motion of the radar-carrying vehicle. This limitation was mentioned earlier but discussion postponed to the present section.

Quite evidently, the two different methods of velocity compensation will lead to different limiting values for the subclutter visibility, so they will be discussed separately. The first one to be discussed will be velocity compensation by the choice of parameters (in accordance with equation (15)).

Consider first equation (14) for the relative sensitivity to moving targets of various speeds. If we substitute $\nu_s^{(r)}$ for $\nu_t^{(r)}$, we get the system sensitivity to the ship's speed; thus,

$$S_r = \left| \sin \frac{2\pi \nu_s^{(r)} f_0}{(\text{PRF}) c} \right|. \quad (24)$$

Since this is the relative response to fixed targets due to a ship's motion and since the subclutter visibility must be less than this if the fixed targets are not to appear as moving targets,

$$R_{\max} \ll \left| \sin \frac{2\pi \nu_s f_0}{(\text{PRF}) c} \right|^{-1}$$

Or, written in terms of $V_{\text{null}}^{(l)}$,

$$R_{\max} \ll \left| \sin \frac{\pi \nu_s}{V_{\text{null}}^{(l)}} \right|^{-1}, \quad (25)$$

where ν_s has been substituted for $\nu_s^{(r)}$ since it is the maximum possible value of $\nu_s^{(r)}$. If the parameters are set according to equation (15) with f_0/PRF being only slightly less than 4×10^5 , then the curve of Figure 1 will be found to apply and the subclutter visibility will be found to be limited about as follows:

ν_s	$R_{\max}$
0	No limit
12 knots	23 db
25 knots	18 db
40 knots	12 db

Turning now to the other method of velocity compensation, the coho-phase-rotation method, consider three targets off the beam of a moving ship. One of these targets is dead abeam on, say, 090° relative while the other two are on $090^\circ \pm \theta_b/2$ relative. The antenna is pointed on 090° relative. Thus, one target is in the center of the beam and the other two are at the two half-power points. It will be obvious that, since the ship's motion is exactly perpendicular to the radius line to the target on 090° relative, there will be no motion with respect to this target and it will be cancelled as a stationary target. However, since both the other targets are on bearings $\theta_b/2$ degrees off the normal to the ship's course, the ship will have a relative velocity with respect to them (toward one and away from the other). This relative velocity can be shown to be:

$$\nu_s^{(r)} = \nu_s \sin \frac{\theta_b}{2}, \quad (26)$$

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where,

$$\begin{aligned}\nu_s^{(r)} &= \text{ships relative velocity,} \\ \nu_s &= \text{ships true speed,} \\ \theta_b &= \text{antenna beamwidth.}\end{aligned}$$

Quite obviously this velocity cannot be compensated since it is of opposite sign on the two sides of the beam. If now we refer to equation (14), and substitute this value of ν in it in place of $\nu_t^{(r)}$ we get,

$$S_r = \left| \sin \frac{2\pi f_o \nu_s \sin \theta_b / 2}{(\text{PRF}) c} \right|,$$

or,

$$S_r = \left| \sin \frac{\pi \nu_s \sin \theta_b / 2}{V_{\text{null}}^{(1)}} \right|. \quad (27)$$

Thus there will be a doppler response of this (voltage) level to the clutter at the edges of the beam even though the target at the center of the beam is stationary. This condition, where the antenna is headed broadside, is the worst condition, so if subclutter visibility is restricted to a value such that it will not detect this signal, it will be satisfactory for all antenna bearings. Thus subclutter visibility must be limited to:

$$R_{\text{max}} \leq \left| \sin \frac{\pi \nu_s \sin \theta_b / 2}{V_{\text{null}}^{(1)}} \right|^{-1}, \quad (28)$$

where,

$$\begin{aligned}\nu_s &= \text{maximum speed of ship,} \\ V_{\text{null}}^{(1)} &= \text{speed at which the first null in the system response} \\ &\quad \text{occurs (eq. 13),} \\ \theta_b &= \text{antenna beamwidth.}\end{aligned}$$

The other limitations on subclutter visibility are mainly system stabilities. They will not be discussed here since they are covered in the next section, in connection with the SC-SK Series Radars.

SYSTEM LIMITATION - MTI FOR SC-SK SERIES RADAR

The delay line cancellation unit built for this MTI system has been found to provide a reliable cancellation ratio of 30 db. Thus, this source will limit the subclutter visibility of the system to slightly less than this value.

In the SC-SK system, the antenna scanning rate is 5 rpm, the antenna beamwidth is about 20° and the (PRF) is 500 cps. Hence there are about 330 pulses on the target per scan and the pulse-to-pulse variation in echo power, due to this cause, is down to about 10^{-5} . Since this is 50 db down it will obviously not be a limitation.

Velocity compensation is accomplished by parameter selection in accordance with equation (15). The first null in the response curve comes at 837 mph, so the limiting values for subclutter visibility given on page 12 apply to this system.

Up to the present time it has been assumed that everything was perfectly stable and had no effect on system performance. It is now necessary to investigate the stability requirements which must be met by the various units. Before such an investigation can be made, however, it is necessary to assume some degree of desired cancellation. Since a cancellation ratio of 25 db or better can be used at speeds of 12 knots or less, and since the delay-line system is capable of providing a 30-db cancellation ratio, let us assume that the system is to have a normal cancellation ratio 30 db.

If a 30-db cancellation ratio is to be maintained, it is evident that the aggregate of all effects which will cause variations in pulse amplitude, either with or without coho, must be 30 db below peak pulse amplitude. Thus they must be limited to a total of 3 percent (in voltage). The effects which contribute to this total will be transmitter amplitude and frequency stability, coho phasing and frequency stability, and local oscillator frequency stability. Fortunately, transmitter frequency and amplitude stability are so related that if the amplitude stability requirement is met, the frequency stability requirement will be much more than met and can be neglected. Thus we have four effects and if we limit each of these to a variation of 1 percent, the statistical average variation will be less than the allowable 3 percent.

The requirement for transmitter pulse amplitude simply means that the voltage pulse from the modulator must change less than 1 percent from pulse to pulse. Thus the filter on the main modulator high voltage supply must be heavy enough to discharge less than 1 percent per pulse between charging cycles of the 60 cycle power source.

Again, the problem of coho frequency stability is simply one of constant supply voltage and freedom from microphonics. The former can be solved by a voltage regulator and the latter by proper design. The problem of phasing stability is somewhat more complicated, being dependent entirely on the method used to phase the coho. Assume, however, that coho phasing is done by injecting a radio frequency pulse into the coho while it is running normally. If then the residual signal in the coho is to be 30 db below the resultant, it is evident that

$$E_{\text{phasing}} > 30 E_{\text{coho}} \quad (29)$$

Thus, the phasing tube which drives the coho must deliver a voltage 30 times the normal coho level. It is also of interest to note the magnitude of the frequency stability required of the coho. If it is to introduce less than 1 percent change in pulse voltage amplitude when beating with two adjacent similarly phased pulses, its phase must vary less than .01 radians from pulse to pulse. Thus, in the present system, its stability must be:

$$\frac{df}{dt} = \frac{.01}{2} \times 500^2 = 400 \text{ cycles/sec/sec.} \quad (30)$$

This seems, on first consideration, difficult to obtain, but in actual practice it presents no trouble if good design practice is followed.

The frequency stability requirements of the local oscillator are the same as those given by equation (30) for the coho. Here again the problem is one of voltage regulation and microphonics. Because of the higher frequency at which this oscillator operates, the requirements are harder to meet than for the coho but again with careful design and especially, with careful choice of tube type, no special problems should be encountered for S-band and lower frequencies.

SC-SK SYSTEMS CONSIDERATIONS

System Parameters

Quite logically, the first thing which must be considered in the application of MTI to any existing radar system is the question of circuit parameters and whether the parameters of the system are suitable for MTI. If they are not, the questions then arise as to what changes must be made to adapt them to MTI system use. Like any other problem in radar, the answer to this question must be a compromise. The changes to the system must be limited to avoid the pitfall of designing a complete new system, and at the same time the revision must be complete enough to make the MTI practical and useful.

The r-f frequency of the system is, of course, fixed and a change of this parameter cannot even be considered. Indeed, it is mentioned here only because a knowledge of its value is necessary in the paragraphs that follow. The SC-SK series radars use an r-f frequency of $200 \text{ Mc} \pm 25 \text{ Mc}$ and thus the MTI system must be designed for this r-f frequency.

The (PRF) of the SC-SK system was originally 60 cps. This is not satisfactory for an MTI using a mercury delay line since it would require a line 16,000 microseconds long (about 80 ft). Furthermore, at 60 cps the speed for peak response would be about 50 mph and this is too low for a shipborne air search system unless special velocity compensation is used. Fortunately, both of these characteristics indicate that the (PRF) should be increased. In considering this increase in repetition rate there are several factors which must be considered.

Since velocity compensation in this system is to be by selection of parameters, equation (15) will be used to find the (PRF). Transmitter frequency f_0 is 200×10^6 , therefore,

$$(\text{PRF}) > \frac{200 \times 10^6}{4 \times 10^5} = 500 \text{ cps.}$$

Thus, the (PRF) must be 500 cps or higher. This (PRF) requires a delay line 2000 microseconds long and allows a range sweep of about 150 miles. Both these are satisfactory. It does, however, increase the duty cycle of the transmitter more than 8 to 1. It has, though, proven practical to reduce the duty cycle more than 2:1 by shortening the pulse length and then operating the transmitter at a somewhat greater than normal average power. Thus, 500 cycles was chosen as the (PRF) of the modified system.

As mentioned in the preceding paragraph, the pulse length of the system was shortened to reduce the amount that the duty cycle of the transmitter had to be increased in connection with the increase to 500-cycle (PRF). The pulse length selected was 2 microseconds. This allowed the use of a 500-cycle (PRF) with a duty cycle increase of slightly less than 4:1. Extensive tests were run with the transmitter plate-keyed at this new (PRF) and pulse length and it was found that tube life did not suffer.

In the original SC-SK transmitter, the peak pulse power was limited by peak cathode emission and by arcing rather than by plate dissipation in the tubes themselves. Because of this, and because of the increase in efficiency when plate keying was used, it proved practical to maintain the peak power of the transmitter unchanged in spite of the increase in duty cycle.

The MTI System

Having decided on these two changes in system parameters it is possible to proceed with the design of the system in general. The technical details and circuitry will not be

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considered at this point but will be left to later sections where each unit will be discussed in detail.

Diagram 1⁴ shows a block diagram of the system (less power supplies, antenna control, and indicators). The transmitter and duplexer are the normal SC-SK series units except that the power for the transmitter is now supplied by a new modulator (unit No. 7) instead of by the built-in power supplies. The receiver (unit No. 0) is a complete new receiver supplying two separate output signals. One of these is a normal video signal of negative polarity and about 2-volt level to the video distribution system. This normal presentation is protected against jamming by a-j protection built into the receiver. The second output is a 15 Mc i-f signal fed at a level of about 1 millivolt to unit No. 6, the MTI i-f and coho unit.

The MTI i-f and coho unit receives this signal from the receiver and amplifies it up to a level of several volts, limits it, and then feeds it into a phase detector. This unit also contains the coherent oscillator and the circuits which lock it in correct phase with the transmitter. Output from this coherent oscillator is fed to the coho mixer where it is added to the i-f signal and the resultant signal is detected in the phase detector to give the coherent video signal. This coherent video signal is then fed to the remodulator.

The remodulator (unit No. 4) consists of a 15-Mc oscillator driving a pair of 6AG7's in parallel. The coherent video signal is applied to the grids of the 6AG7's, along with the trigger pulse, to modulate this 15-Mc carrier which is then fed directly from this unit to the transmitting crystal in the mercury delay line. Also on this chassis is the through channel. This channel takes a fraction of a volt of the output of the 6AG7's through an attenuator, amplifies it through a single stage (with a gain control for controlling its level), and detects it to give the undelayed video signal.

The delay line (which receives the output of the remodulator) is a five channel folded mercury delay line. Diagram 10 is an assembly drawing of this line and shows its general structure. It is made of stainless steel and weighs about 80 pounds when filled. It is discussed in greater detail in NRL Report R-3065.⁵ The output of the receiving crystal is fed to a single tube preamplifier built into the delay line, and then to the delayed channel trigger amplifier. Although this amplifier is built into this same unit as the delay line (unit No. 5), it is functionally part of the trigger channel and will be discussed in connection with the trigger system. The delayed channel signal is also taken off the cathode of the first tube in this amplifier and fed up to unit No. 3, the delay video and cancellation unit.

In the delay video and cancellation unit, the delayed carrier from the delay line pre-amplifier is amplified and detected to give the delayed video signal. This signal is then delayed another 2 microseconds by a video delay line. This further delay is necessary in connection with the trigger circuit used and will be discussed in that connection. The output from this video delay line is then fed to the resistance cancellation network where it is added, in opposite polarity, to the through video signal fed in from unit No. 4. The output from this cancellation network is then a video signal in which all fixed targets have been cancelled out and all moving targets appear as pips of either positive or negative polarity. Because of this double sided character this video signal must be processed further before going to the indicators. It is first amplified and then fed to a phase inverter whose output is rectified by a full-wave rectifier to give a single-sided video. This single-sided MTI video signal is then fed to the video distribution system. This video distribution

⁴ All diagrams are placed at the end of the text.

⁵ T. H. Chambers, op. cit.

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system contains the necessary gain controls, clippers, amplifiers and cathode followers to feed two 75-ohm outputs to PPI indicators and two 75-ohm outputs to "A" scopes (one each with normal video, and one each with MTI video).

The other unit in the system is the trigger unit (unit No. 1). This unit generates a pulse which simultaneously keys the transmitter and is fed to the remodulator where it is modulated on the carrier at extremely high level. This high level pulse then travels down the delay line, along with the video signal, and is picked up by the receiving crystal and amplified by the delay line preamplifier. It then goes to a separate amplifier and detector and clippers where it is raised to working level detected and the video signal clipped off, leaving only the delayed trigger pulse. This pulse then keys a multivibrator with a pulse length of 1.6 microseconds variable ± 0.5 microseconds. The trailing edge of this pulse then initiates a new cycle of operation. Thus the total delay of the trigger system is the delay-line time plus 1.6 ± 0.5 microseconds plus 0.7 microseconds stray delay time for a total of delay line time of $+2.3 \pm 0.5$ microseconds. But the signal delay time is delay line time plus 2.0 microseconds added video delay time plus 0.3 microseconds stray delay giving a total of delay line time $+2.3$ microseconds. Thus, not only is the system (PRF) automatically slave to the delay line time, but a very fine adjustment of ± 0.5 microsecond is provided for precise balance.

Equipment Changes

Because of the changes necessary in the system itself, certain of the components, while not necessarily new additions to the system, will have to be either modified or replaced.

In the original SC-SK, the r-f oscillator was of the self-blocking type and was synchronized to the 60-cycle supply for timing control. Since the keying must now be controlled externally and the (PRF) and pulse length must be changed, the blocking oscillator function of the transmitter was removed completely and the grids returned to ground through a suitable bias source. Keying is then done by plate modulating the transmitter. Unfortunately, the power supply built into the transmitter was not sufficiently heavy to meet the strict stability requirements at the higher repetition rate so it was disconnected and a new high voltage supply built into the modulator.

The CG46ACQ replacement receiver for the SC-SK series radar systems was replaced with one of more modern design.

Since both the "A" scope and the PPI for this series radar were designed for use at a (PRF) of 60 cycles, both are unsatisfactory for use at the 500-cycle (PRF) of the MTI system. Although it required a very complete revision, the "A" scope was changed to operate at the new frequency. The PPI, on the other hand, would require such complete revision (and would still be an obsolete type) that instead of attempting to use it, it was replaced by one of the newer remote PPI's capable of operation at the higher repetition rate and with the narrower pulses.

Two new units were added to the system in conversion to MTI operation. The first of these is the new modulator for the transmitter and the second is the unit containing the receiver, coherent, cancellation, and trigger circuitry. For the sake of brevity, this second unit is called "The MTI Conversion Unit."

As mentioned previously, plate modulation was applied to the transmitter. This modulator unit (Figure 2) uses d-c resonant charging of a pulse-forming line which is then discharged into the transmitter by a hydrogen thyratron. The modulator unit is 21 inches

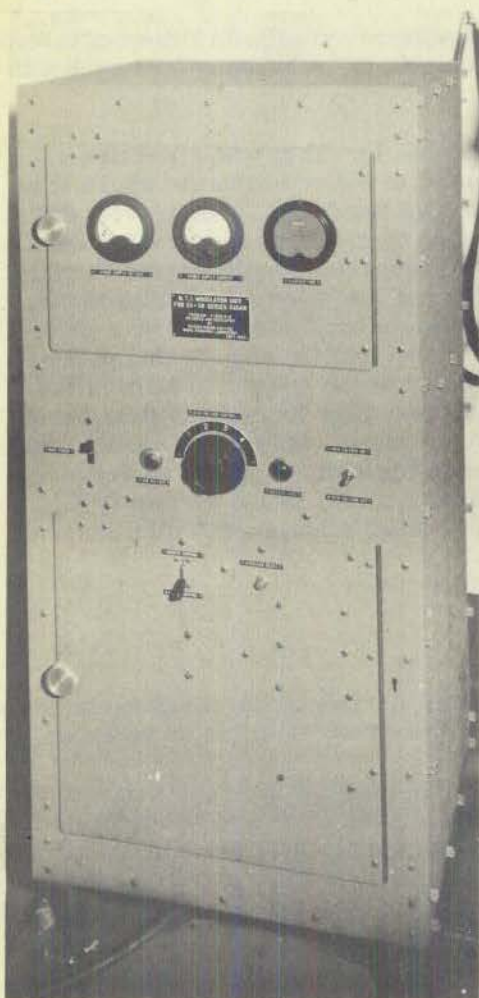


Fig. 2 - The Modulator Unit -
Unit No. 7

wide by 40-1/4 inches high (exclusive of mounting channels) by 19-1/2 inches deep and weighs about 175 pounds. It is described in detail in a later part of this report.

The MTI Conversion Unit (Figures 3 and 4) is the second new unit added to the SC-SK series radar. It is built in the form of six separate chassis which fit into a single cabinet intended to occupy the space at present occupied by the original PPI unit. This unit is 41 inches wide by 29 inches high (exclusive of shock mounts) by 34-1/2 inches deep and weighs 755 pounds. The six units contained in it are the receiver unit, the MTI i-f and coho unit, the re-modulator unit, the delay line unit, the delay i-f and video unit, and the trigger unit. It performs all functions necessary to MTI operation, except for the necessary revisions to the transmitter. It, also, is described in detail in a later part of this report.

Although there are several changes to be made in the transmitter, these changes are quite simple and quickly made. The following is a list of these modifications:

- a. The high-voltage rectifier tubes and their plate leads are removed.
- b. Power to the high-voltage plate transformer is disconnected.
- c. The plate-voltage supply to the r-f oscillator is disconnected.
- d. The pulse-input-power fitting is installed and connected.
- e. The grid return for the r-f oscillator is disconnected and a 250 ohm non-inductive resistor for grid leak is installed.
- f. The r-f coupling loop (for the coho circuit) is installed in the antenna feed line.

THE MODULATOR

The modulator circuit is shown in Diagram 2. It will be noted that the modulator uses the familiar type of pulse-forming network with d-c resonant charging and hydrogen thyratron keying.

The main high-voltage supply uses two type 100 R rectifiers in a full-wave circuit, together with a single pie section filter to develop a maximum of 4000 v. The regulation and filtering of this supply is very good and the filter is heavy enough so that output voltage

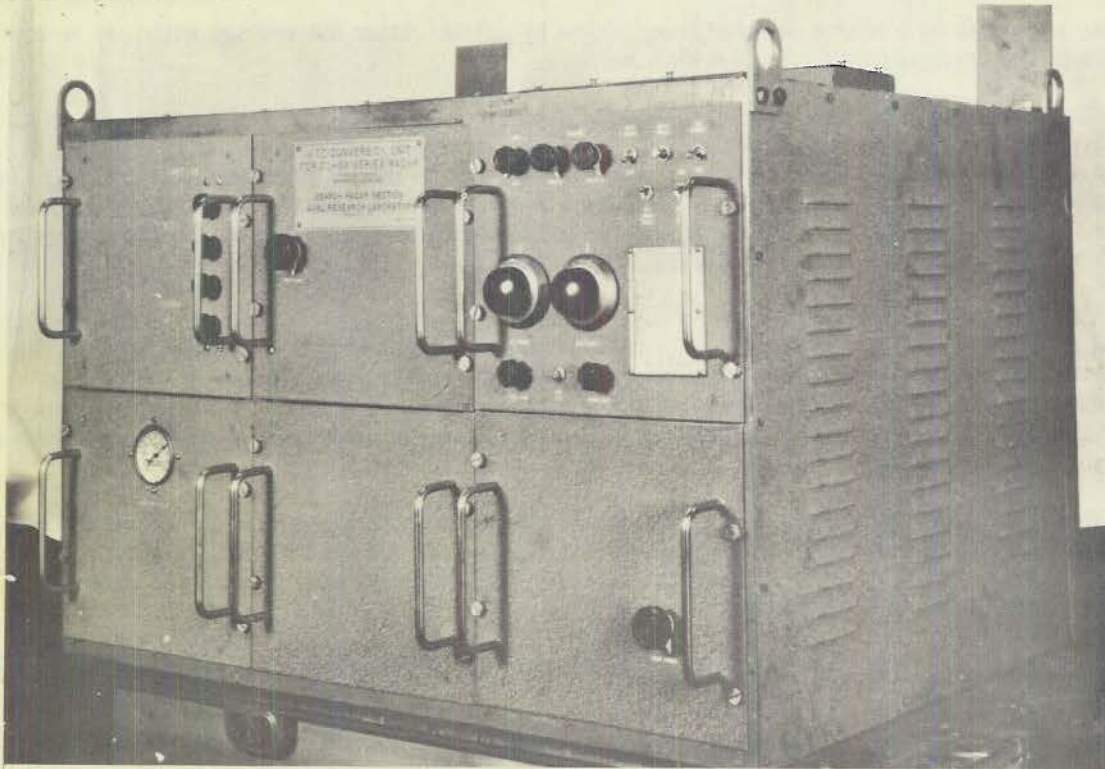


Fig. 3 - The MTI Conversion Unit for SC-SK Series Radar (Oblique View)

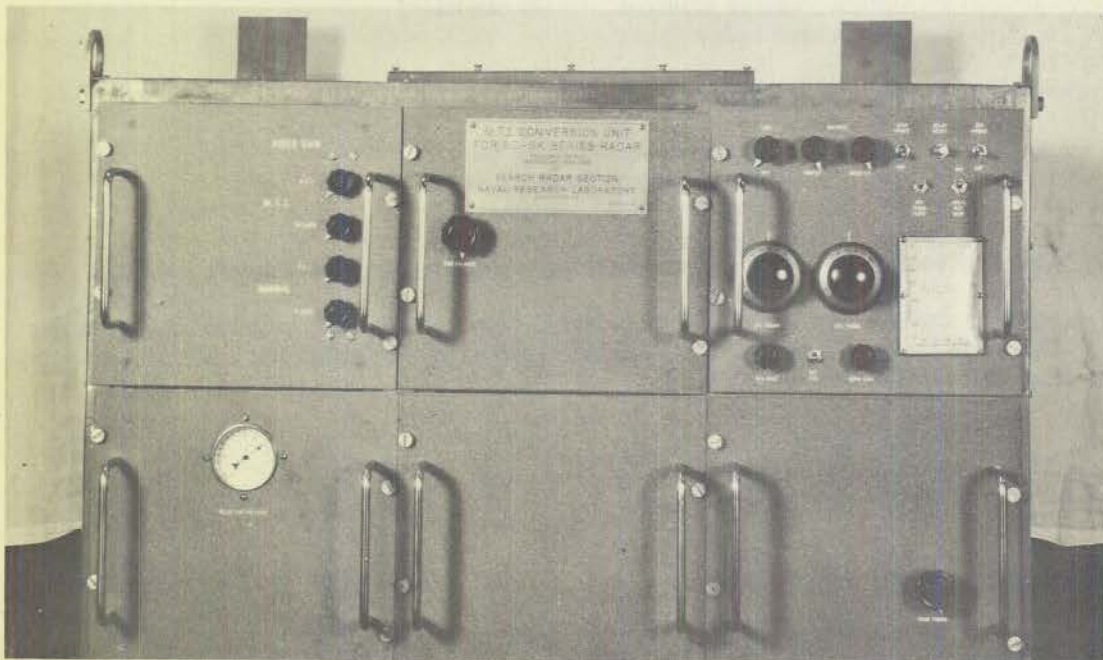


Fig. 4 - The MTI Conversion Unit for SC-SK Series Radar (Front View)

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varies less than 1/2 of one percent from pulse to pulse. Thus the voltage stability requirements of the transmitter are more than met.

The pulse-forming network is a Sprague type 15-E7-2-240-50P series 1897. It is a 7 section 2 microseconds network of 50-ohm characteristic impedance. It is charged to about a 7.9 kv resonant peak through a heavy choke and is then discharged through a pulse transformer with an impedance stepup ratio of 9.6:1 to match into the 480-ohm load presented by the transmitter. The discharge of the network is controlled by a 5C22 hydrogen thyatron which is in turn controlled by the keying unit.

The keying unit consists of three tubes which deliver a high-level pulse to the hydrogen thyatron. This pulse may be either locally or remotely generated. For local keying, the first two tubes are connected in a multivibrator circuit to deliver a pulse of 400-cycle repetition rate. The last tube then acts as a cathode follower to lower the impedance of the pulsing circuit as seen by the hydrogen thyatron. When "normal" operation is desired this local keying is used.

When the MTI is in operation, and a remotely-generated trigger pulse is used to trigger the unit, a relay applies cut-off bias to one side of the multivibrator, thus rendering it inoperative; and the other side is used as a clipper-amplifier to feed the trigger pulse from the input amplifier to the cathode follower.

The modulator may be controlled either locally from its front panel or remotely from the conversion unit.

For local control it is provided with the following controls:

Main Power	on-off
High Voltage	on-off
High Voltage Control	(4 position)
Overload relay reset	

It is also provided with the following meters:

High Voltage Voltmeter
High Voltage Current Meter
Elapsed Time Meter

A switch located behind the panel is provided for selecting local or remote trigger.

For remote control of the modulator, a local remote switch is provided on the front panel. When this switch is thrown to the remote position, the following functions are controlled from the conversion unit:

High Voltage	on-off
Overload Relay Reset	
Local or Remote Trigger	

THE MTI CONVERSION UNIT

The Receiver

As pointed out earlier, the CG46ACQ receiver was satisfactory electrically and could have been used in the MTI system for the normal channel and also to supply i-f to the MTI

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channel with only slight change. Its mechanical layout did not, however, fit into the plans for the MTI. For this reason a complete new receiver was built into the MTI conversion unit. This new receiver is shown as unit No. 0 in the block diagram (Diagram no. 1). The schematic diagram of the receiver is shown in Diagram no. 3 and photographs in Figures 5, 6, and 7.

Referring to the circuit of Diagram no. 3, it will be noted that a new r-f system incorporating grounded-grid triodes has been used in this receiver. The use of these grounded-grid triodes gives a noise factor about 2 db better than the lighthouse tube preamplifier used in the CG46ACQ receiver. The local oscillator used is a 2C43 lighthouse tube. Although several different oscillators using different miniature tubes were tried, they all were found to be too microphonic to be satisfactory in this application. On the other hand, the lighthouse tube was found to be stable and satisfactory. It is used in a double re-entrant tank circuit which is capacity-loaded to shorten the tank and is capacity-tuned by means of a double-spaced condenser. Output for the signal converter and for the coho converter (in unit No. 6) is taken off by means of two small coupling loops in the side of the tank and is fed to the converters over coaxial lines. The signal converter deserves some mention because of the unique method of coupling in the local oscillator signal. This converter is essentially a "grounded" grid stage with the r-f signal applied to the cathode. It differs, however, from the standard grounded-grid amplifier in that only pins 5 and 6 of the type 6J4 tube used in it are grounded. Pin 1 is tied to the coaxial line coming in from the local oscillator. Thus, the reactances of the grid leads within the tube are used as a voltage divider by means of which a portion of the output from the local oscillator is applied to the converter grid without introducing excess reactance between grid and ground. Efficient conversion by the tube is assured by applying sufficient bias to insure operation on the nonlinear portion of its characteristic.

The i-f system can conveniently be divided into two parts, the preamplifier (tubes V-005, V-006, and V-007) and the normal channel i-f amplifier (tubes V-008 to V-011 inclusive). The preamplifier consists of two ordinary i-f amplifier stages using 6AC7's. The main system gain control is applied to these two stages using a variable negative potential applied to the grids. The third stage, also using a type 6AC7, serves both as an amplifier in the normal channel and as a cathode follower to drive the line to the MTI I-F and coho unit (unit No. 6). The normal channel i-f amplifier consists of three stages of 6AC7's and a final stage of 6AG7 with non-amplified back-bias applied to all stages. This gives good a-j protection against jam levels up to greater than 50 db j/s ratio.

The detector is a 6AL5 diode (V-012) operated at a level high enough to insure linearity throughout its operating range. It gives an output signal of positive polarity.

The video system consists of three tubes. The first of these is a permanently-connected 6AL5 diode (V-013) D. C. reinserter. Although not normally necessary, this diode serves as third detector for off-frequency jamming, and, when the switchable FTC (of the RL type) is in use, it serves to remove overshoot. The second tube is an amplifier serving to raise the level into the 6AG7 cathode follower output tube. Output is of negative polarity and at a level of about 2 volts.

All power for this unit is supplied by a built-in power supply providing 300 volts positive at 150 ma for all plate power in this unit only. This supply also delivers 75 volts negative at 20 ma, regulated by a VR tube, for bias to this and all other units of the conversion unit.

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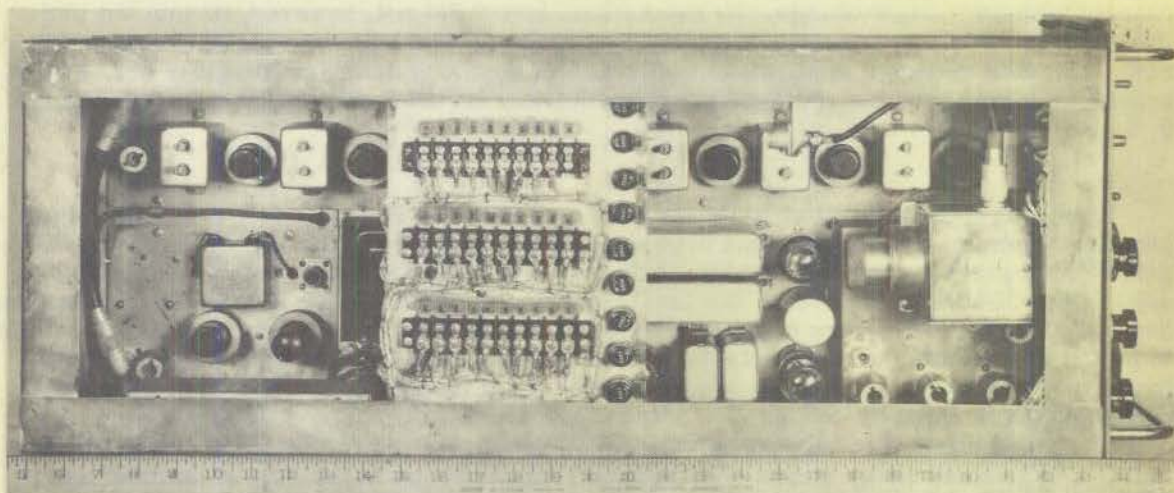


Fig. 5 - The Receiver Unit - Unit No. 0 (Top View)

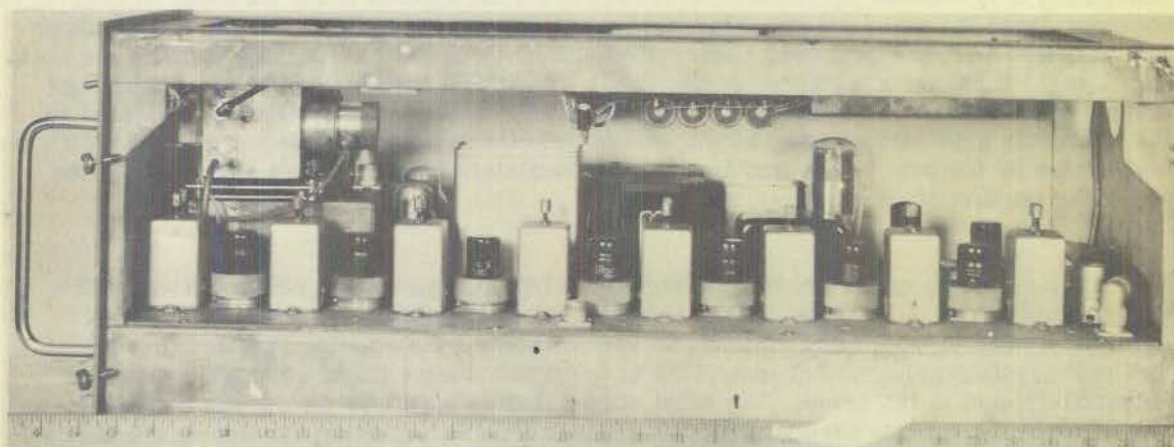


Fig. 6 - The Receiver Unit (Side View)

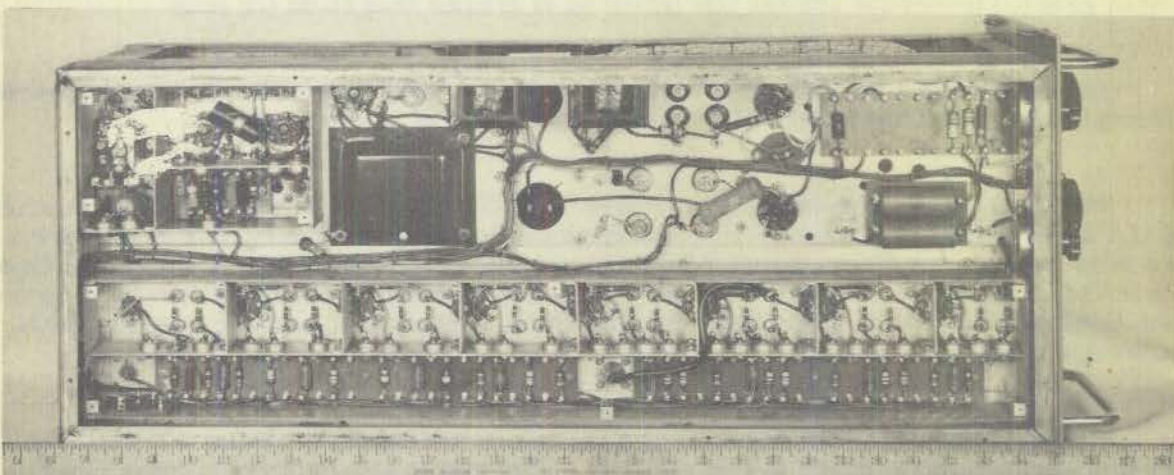


Fig. 7 - The Receiver Unit (Bottom View)

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The Control System. All controls for the conversion unit, with the exception of coho tuning, time balance, and video gain controls are brought out to the panel of this unit. They are as follows:

- Receiver Controls
 - Main Tuning
 - Osc. Tuning
 - Main I-F Gain
- Normal Channel Controls
 - Normal Channel I-F Gain
 - FTC On-Off Switch
- MTI Channel Controls
 - MTI Channel I-F Gain
 - Delayed Channel Gain
 - Through Channel Gain
- Power Controls
 - Main Power On-Off
 - MTI Power On-Off
- Transmitter Controls
 - Transmitter Plate Power On-Off
 - Repetition Rate MTI-400 cy
 - Overload Relay Reset

The MTI I-F and Coho Unit

The MTI I-F and coho unit (Figures 8 and 9) is shown in the block diagram (Diagram no. 1) as unit No. 6. The schematic diagram is shown in Diagram no. 4.

Referring to Diagram no. 4, it will be seen that this unit consists of two major parts: the MTI I-F system consisting of tubes V-601 to V-608 inclusive and their associated circuits, and the coherent oscillator system consisting of tubes V-617 to V-622 inclusive and their associated circuits. The MTI I-F system will be described first.

Signal from the last stage of the I-F preamplifier in the main receiver (unit No. 0) is fed into a 3-stage amplifier at input connector 30. This amplifier uses three type 6AC7 tubes connected in normal fashion and using single-tuned interstage coupling networks. Gain control is applied to the second and third stages. This amplifier feeds, again through a single-tuned coupling network, to a 6AC7 limiter. This limiter is of the plate current saturation type with regulated screen supply. Its output is fed, again through a single-tuned network, to the coho injector. This coho injector deserves some mention as it has some rather unique features. It is a type 815 tube connected for balanced operation from echo signals and for single-ended operation from the coho. A type 815 tube is used to insure adequate dynamic range to handle the combined output from the echo signal limiter and the coho. Signal is supplied to the grids from a balanced single-tuned network consisting of L_2C_2 and the tube capacities. Coho signal is then applied to the common cathode by means of the impedance matching pie network consisting of L_3 , C_4 and C_5 . Thus, in-phase coho voltages are combined in the two halves of the tube with out-of-phase echo-signal voltages.

The output of this tube is fed to the balanced phase-detector by means of circuits which are again single-tuned, but which are so arranged as to give independent control of the resonant circuits for signal and coho outputs. Coil L_4 and the tube capacities form the resonant circuit for the push-pull echo-signal output, while L_5 , the tube capacities, and

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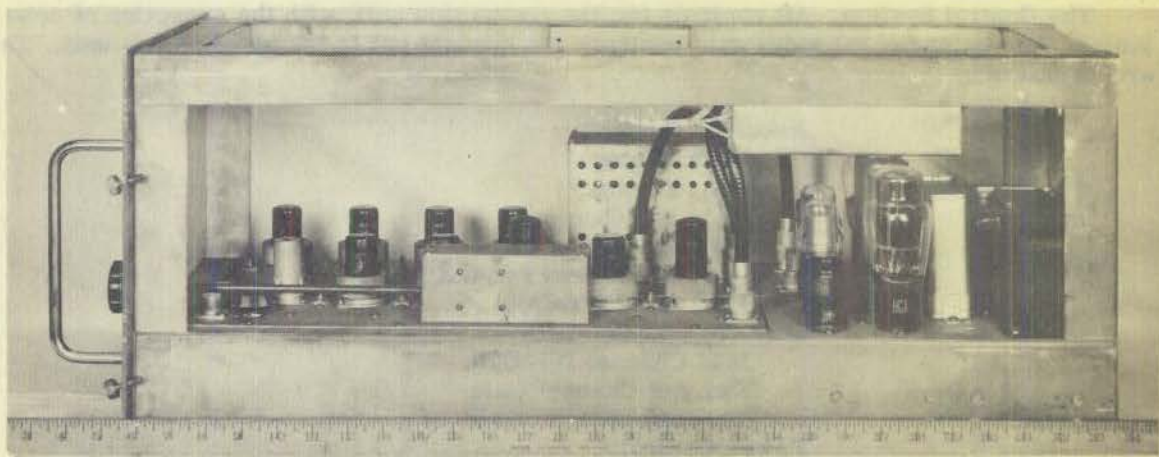


Fig. 8 - The MTI IF and Coho Unit - Unit No. 6 (Side View)

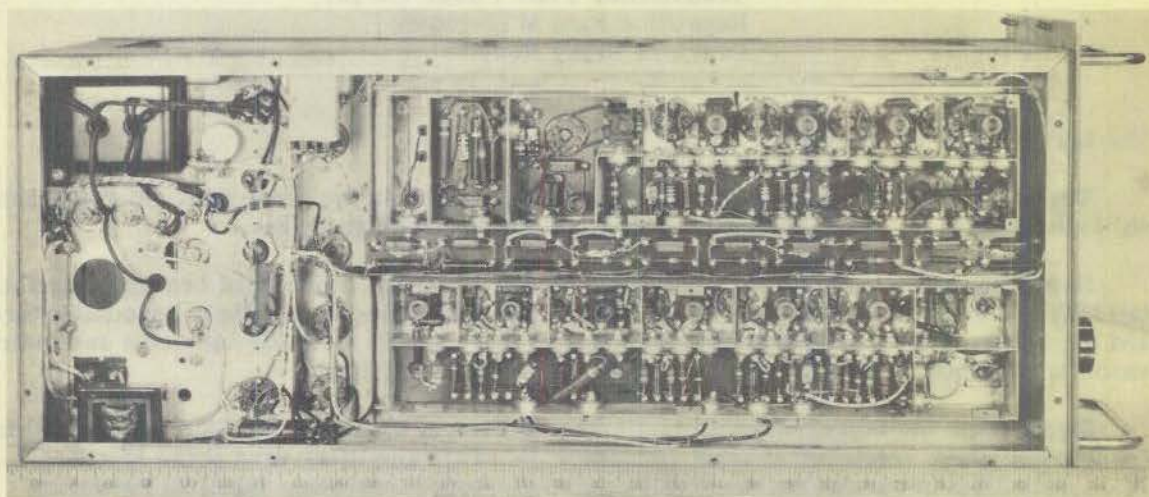


Fig. 9 - The MTI IF and Coho Unit (Bottom View)

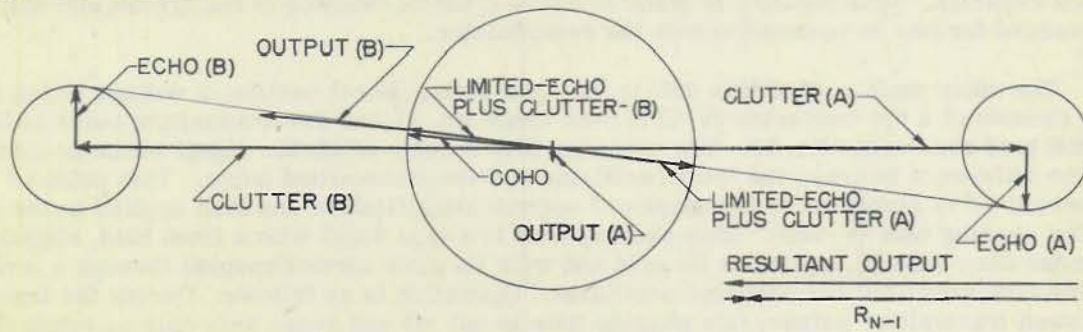
the capacity of L_4 to ground form the resonant circuit for the parallel coho signal. This system has been found to be very easy to adjust and has given excellent performance.

The phase detector consists of two type 6AL5 double diodes connected to give signals of opposite polarity. The outputs of these tubes are then connected together through individual load resistors to a common i-f filter and load resistor. Operation of this phase detector is best shown by the vector diagram of Figure 10. It is shown for the case of clutter above limit level, small echo signal, and with the coho almost in phase with the clutter, since this is the situation when the advantage of the phase detector is most apparent.

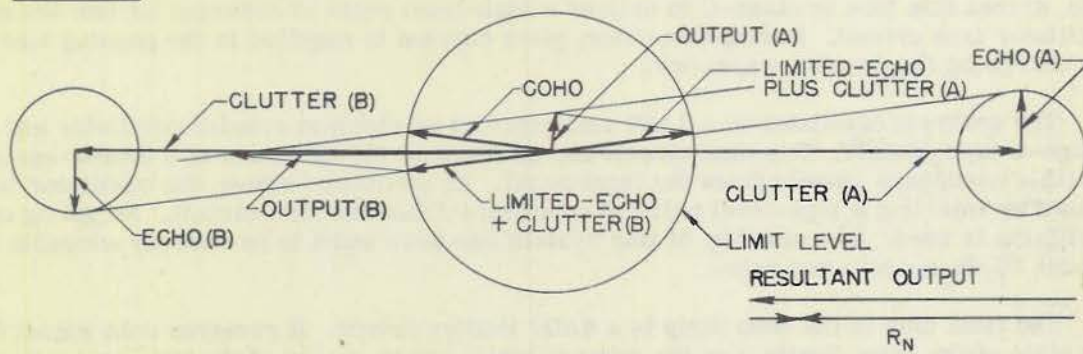
The upper diagram of Figure 10 shows the vector relations for the $(N-1)$ th pulse and shows how the resultant R_{N-1} is obtained by balancing the rectified outputs (A) and (B). The lower diagram shows the relations for the N th pulse (where the target is assumed to

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(N-1)TH PULSE



NTH PULSE



CHANGE IN VIDEO OUTPUT

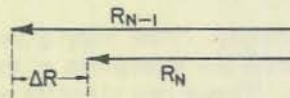


Fig. 10 - Operation of the Balanced Phase Detector

be traveling at a velocity $V_{\max}^{(n)}$ and how the resultant R_N is obtained. Comparing these two resultants then gives the pulse-to-pulse change in echo amplitude, R .

A thorough study of the possible relations between echo signal, clutter signal, and coho amplitudes and phases will not be given here; instead reference is made to Radiation Laboratory Report No. 1010⁶ which considers the relative merits of the various types of MTI receivers. The limiter and balanced phase-detector combination have essentially the same characteristic as the "lin-log" receiver except that the null in response when the plane is traveling at $V_{\max}^{(n)}$ and when the clutter happens to be in the same phase as the coho, is much narrower, and hence the reduction in coverage is less.

The coherent-video output from the phase detector is fed to a single-stage amplifier and then out of the unit over coaxial line number 61. It should be mentioned that this coaxial

⁶ W. Selove, MTI Rad Lab Report 1010, "Notes on MTI Receivers," 25 March 1946.

line is not matched. Instead, it is driven by a high impedance source and it acts as a lumped shunt capacity. This capacity is made to serve a useful function in the circuit and will be discussed further in connection with the remodulator.

The other main part of this unit is the coho strip. Local oscillator voltage is fed into the cathode of a 6J4 converter (V-617) over cable No. 21 and the transmitter pulse is fed to the grid over cable No. 22. The output is then a pulse of 15-Mc signal which is coherent to the difference between the local oscillator and the transmitted pulse. This pulse of coherent i-f is amplified in two stages of normal amplification and then applied to the grid of the phasing tube (V-620). This phasing tube is a type 6AG7 with a fixed bias, slightly greater than cut-off, applied to its grid and with its plate circuit coupled through a condenser to the tank circuit of the coherent oscillator. Operation is as follows: During the time between transmitter pulses, this phasing tube is cut off and hence acts only as small fixed capacity across the oscillator tank circuit. Also during this time the plate load resistor for this phasing tube (which must be made high enough not to load excessively the oscillator tank circuit with which it is essentially in parallel) charges the coupling condenser to power supply voltage. Then when the transmitter fires, the pulse of coherent i-f, which has been built up to high level by the two stages of gain and which is applied to the grid of the phasing tube, drives this tube in class-C to deliver a high-level pulse of coherent i-f into the coherent-oscillator tank circuit. During this pulse, plate current is supplied to the phasing tube from the charge on the coupling capacitor.

The coherent oscillator is a 6AC7 connected as an electron coupled oscillator and using a high-C tank circuit. This tank circuit can be tuned ± 1 Mc by means of a double-spaced variable condenser tunable from the front panel. As mentioned above, the oscillator is phased by injecting a high-level pulse of coherent i-f into the tank circuit. No gating of the oscillator is used. The stability of this system has been found to be entirely adequate to secure 30-db cancellation ratio.

The final tube in the coho strip is a 6AG7 limiter driver. It receives coho signal from the plate of the coho, limits it to the correct level, and by means of the two pie-section impedance-matching networks and interconnecting cable No. 39, drives the cathode of the coho-injection tube.

All power for this unit is supplied by a built-in power supply. This supply uses electronic regulation of the main supply voltage to all circuits with the exception of the echo-signal limiter and coherent-oscillator screen-grid circuits. These two circuits are supplied from sources regulated (in addition to the main electronic regulator) by separate gas tubes. This supply also supplies 6.3 volts ac to all heaters. Two a-c voltages are fed to this unit from the receiver unit (unit No. 0). These are the gain control voltage for the MTI i-f strip (0 to 20 volts negative) and a negative voltage for the gain control on the coho-strip i-f amplifiers. This voltage is negative 75 volts and is reduced by a network (adjustable from the top of the chassis by means of a screwdriver) and applied to the grids of the amplifier stages.

This unit has one front panel control; the coherent oscillator tuning control.

The Remodulator and Through Video Signal Channel

The remodulator and through video signal channel (Figures 11 and 12) is shown in Diagram no. 1 as unit No. 4. A schematic diagram of this unit is shown in Diagram no. 5.

This unit consists of two main parts. The first of these is the remodulator. This remodulator is, in reality, a small transmitter which supplies an r-f carrier, modulated

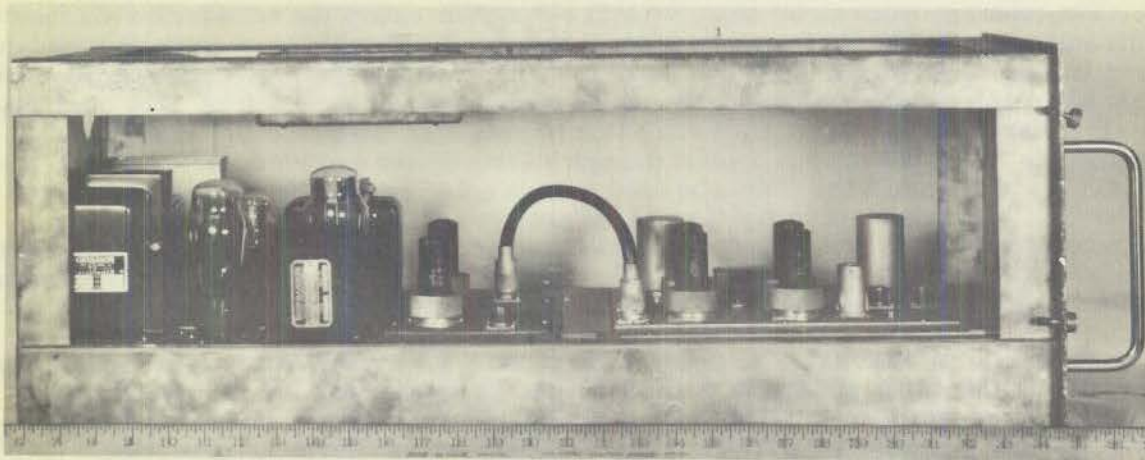


Fig. 11 - The Remodulator Unit - Unit No. 4 (Side View)

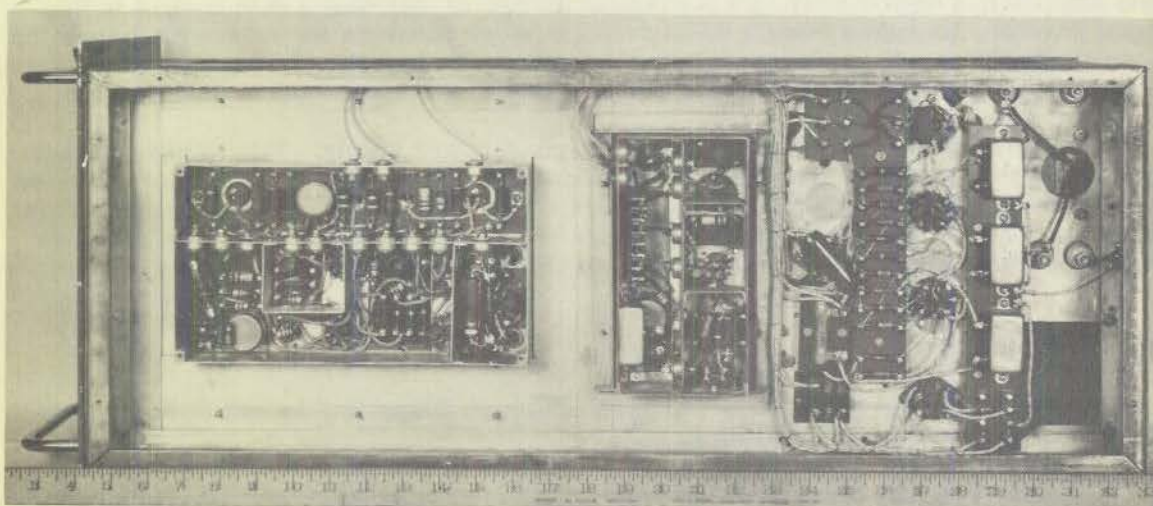


Fig. 12 - The Remodulator Unit (Bottom View)

by the coherent-video signal, to the transmitting crystal in the mercury delay-line. The second part of this unit is the through video-signal channel. This is essentially a monitor which takes a small part of the remodulator output and rectifies it to obtain an undelayed video signal for use in the cancellation unit. This undelayed video signal is obtained by the use of this monitor (rather than by taking a portion of the incoming coherent video signal) so that the through and delay channels will both be subjected to the same distortion in the remodulator and thus this distortion will not become a limiting factor on system cancellation.

The remodulator employs five tubes. The first of these is a 6AG7 carrier oscillator (V-401). This oscillator is electron-coupled and uses a high-C tank circuit to obtain stability. It feeds the grids of a pair of 6AG7's connected in parallel. These 6AG7's are the final amplifier and are coupled by a pie-section network into the 75-ohm cable which goes to the delay line (cable No. 51). Modulation is applied to this final amplifier in its grid circuit.

The coherent-video signal from the MTI I-F system enters the remodulator via cable 61. This cable acts as a lumped capacity since it is not matched in its characteristic impedance. This lumped capacity, which has a value of $120 \mu\mu\text{f}$ is used as the input condenser of a low pass filter which limits the band pass into the remodulator to about 400 kc. The other elements of this filter are in the input end of the remodulator video circuit. After passing through the filter, the video is applied through a screwdriver-adjusted level control to the grid of the modulator tube. This modulator tube is a 6AG7 connected as cathode follower. It develops the modulation voltage across the grid-return condenser for the final stage. Also connected to this grid return of the final stage is a 6AL5 trigger-pulse injector. This trigger-pulse injector serves to apply to the final amplifier a high-amplitude positive pulse which is necessary for operation of the trigger circuit (to be described later).

The remodulator delivers a carrier level of about 8 volts into the 75-ohm cable to the delay line. This, when stepped up to the 2700-ohm impedance level of the crystal, gives about a 50-volt level across the crystal. Modulation is at a level of ± 70 percent peak amplitude for the double-sided coherent video, and $+125$ percent for the trigger pulse.

The through video-signal channel consists of three tubes. The first of these is a 6AG7 carrier amplifier which receives about 0.3 volt of modulated carrier from the remodulator. This amplifier is used simply as a convenient method of including a gain control in this signal channel. Its output feeds a 6AL5 detector which recovers the video signal from the modulated carrier. The video signal then is fed out over line No. 62 by a cathode follower. Since the detector is connected to give positive output, this line output is also positive.

All power for the unit is supplied by an internal power supply. This supply delivers 500 volts electronically-regulated to the remodulator output stage and 300 volts regulated in a two-stage electronic regulator to all other tubes. One d-c voltage is brought in from unit No. 0. This is a variable negative voltage (0 to -20 volts) for the gain control of the through video-signal channel. The unit has no front panel controls.

The Delay Line Unit

The delay line unit (Figures 13 and 14) is shown in Diagram 1 as unit No. 5. A schematic diagram is shown in Diagram 6.

The delay line unit consists of two major parts, the delay line itself and the trigger channel carrier amplifier. Only the electrical characteristics of the delay line will be mentioned here, since its construction and mechanical features are covered completely in NRL Report R-3065.

The high level modulated carrier signal comes into this unit from the remodulator over cable number 51. This cable goes directly to the transmitting crystal compartment. Here a pie-network steps it back up to an impedance level of 2700 ohms and applies it to the crystal. This pie network, together with the pie network in the remodulator output circuit and the interconnecting cable form, effectively, a double-tuned transformer with slightly greater than critical coupling. Thus a double-peaked response, about 1 Mc broad, is obtained.

The receiving crystal of the delay line is matched into the grid of a 6J6 cathode-coupled preamplifier by a pie-section coupling network. This coupling network is about 1 Mc broad, and since it gives a single tuned response curve, it corrects the double-peaked response of the transmitting crystal circuits to give essentially flat overall response throughout the pass band. This preamplifier is built into the crystal compartment and its output is coupled

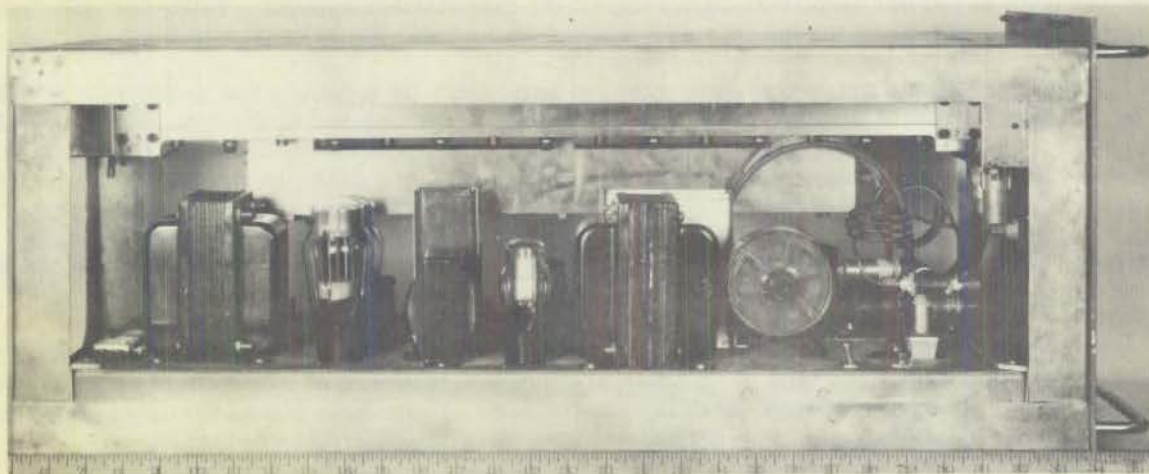


Fig. 13 - The Delay Line Unit - Unit No. 5 (Top View)

by means of a pair of pie sections and interconnecting cable No. 54 to the second main part of this unit; the trigger-channel carrier-amplifier.

The electrical characteristics of the delay line are about as follows: It has a delay time of 2000 microseconds, and a loss of about 80 db without the preamplifier and 65 db with the preamplifier. It has a bandwidth of about 1 Mc with its associated circuits. All its spurious responses are down more than 40 db below the main signal.

The second main part of this unit is the delayed-trigger carrier-amplifier. This amplifier receives the output of the delay-line preamplifier, builds it up to a level of several volts and then detects it. The only unusual feature of this amplifier is the signal take-off on the first stage. This first stage acts not only as an amplifier but also as a cathode follower to feed signal to cable 53 whence it goes to the delayed-video carrier-amplifier (in unit No. 3). For the sake of clarity, attention should be called at this point to the difference between these two amplifiers. In the delayed-trigger carrier-amplifier (the one in this unit) the trigger pulse will be clipped off its output, hence its gain must be maintained constant to secure proper operation of the trigger circuit. In the delayed-video carrier-amplifier (in unit No. 3) the video signal will be used, and gain must

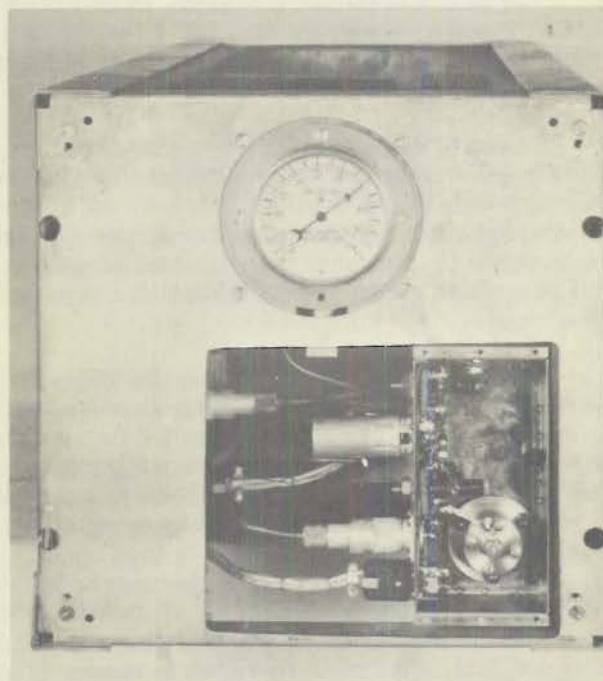


Fig. 14 - The Delay Line Unit (Front View) with Panel Removed

be variable to permit amplitude balancing of the cancellation system. Thus, separate amplifiers are used, even though they are similar and handle the same signal.

Returning to the delayed trigger carrier amplifier; its output is fed to detector V-505. This detector gives a positive output signal which is then applied to the non-linear amplifier V-506. This amplifier is operated with high bias so that the video signal is compressed in the low-gain portion of its characteristic, and the trigger pulse is strongly accentuated on the high-gain portion of its characteristic. This suppression of the video is in preparation for the clipping which will be done later (in unit No. 1) to completely separate the trigger pulse from the composite signal. Output from this amplifier is fed to cable No. 81 and thence to unit No. 1.

All power for this unit is supplied by a built-in electronically regulated power supply. Also contained within the unit is an automatic pressurization unit which supplies 35 psi of compressed air to pressurize the delay line. There are no controls on the front panel but there is one indicating meter; a pressure meter showing delay line pressure.

The MTI Cancellation and Video-Distribution Unit

The MTI cancellation system, shown in Diagram 1 as unit No. 3, and the Video Distribution System, shown as unit No. 2, together comprise unit No. 2-3 of the MTI Conversion Unit. A schematic diagram of this unit is shown in Diagram 7 and photographs in Figures 15 and 16.

The MTI delayed-carrier signal is fed into this unit (from unit No. 5) by cable No. 53. This cable brings the delayed carrier in at low level from the cathode of V-501 (in unit No. 5) and delivers it to the delayed-video carrier-amplifier where it is built up to a level of several volts and detected to recover the delayed-video signal. This is a conventional four-stage 15-Mc amplifier. It uses single-tuned circuits and has a bandwidth of about 1.5 Mc. Gain control is by means of a negative voltage applied to the grids of the first two stages.

The detector used to recover the delayed video signal is again a 6AL5 diode operated at a level high enough to be linear throughout its operating range. The output of this detector is fed through a lumped-constant LC delay network (whose function will be discussed later in connection with the trigger circuit), having a 2-microsecond delay time, to a cathode follower. The output of this cathode follower is then the delayed coherent-video signal of a polarity such that an increase in modulated carrier level causes an increase in the positive direction. This signal is then added, in a resistance-cancellation network consisting of R_1 , R_2 , R_3 and R_4 , to the undelayed video signal of opposite polarity, and fed in over cable No. 62 (from unit No. 4) to secure cancellation of the fixed targets. The output from this cancellation network is then a bipolar video signal representing only the moving targets, and, after it is rectified to render it single sided, it will become the MTI video signal.

The rectification of this video signal is accomplished in the following way: The output of the cancellation network is fed through a single-stage amplifier to a low-pass filter which removes the high-frequency uncancelled components and limits the overall system bandwidth to the desired 400 kc. The output of this filter is then fed to a degenerative cathode follower type of phase inverter, the two outputs of which are fed to the two grids of a 6J6 double triode. The plates and cathodes of this double triode are connected together and it is operated at approximately cut-off bias so that rectification is secured. The resultant signal, single sided and of positive polarity, is taken off the cathodes and fed to a cathode follower whence it leaves the unit over cable No. 65.

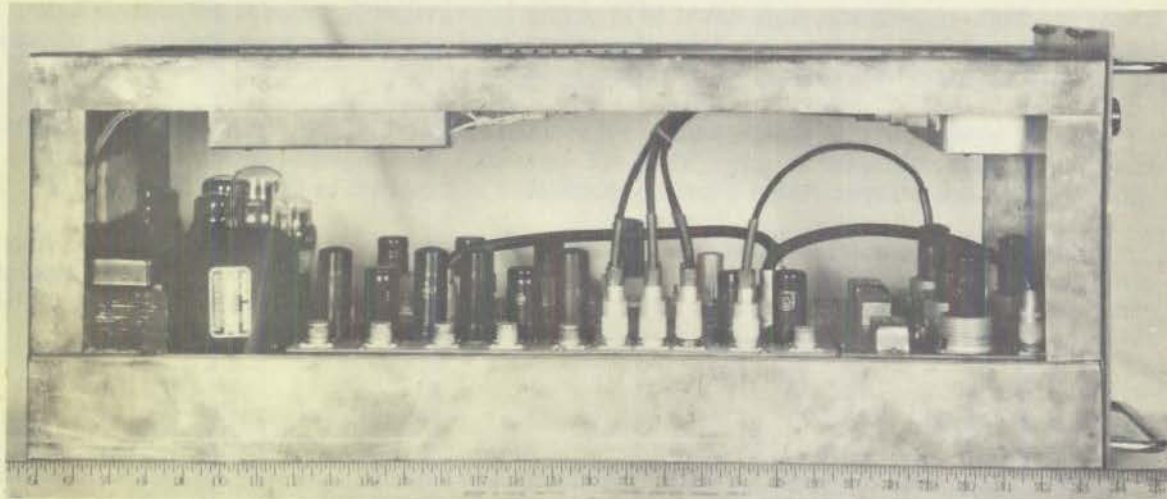


Fig. 15 - The Cancellation and Video Distribution Unit
Units No. 2 and 3 (Side View)

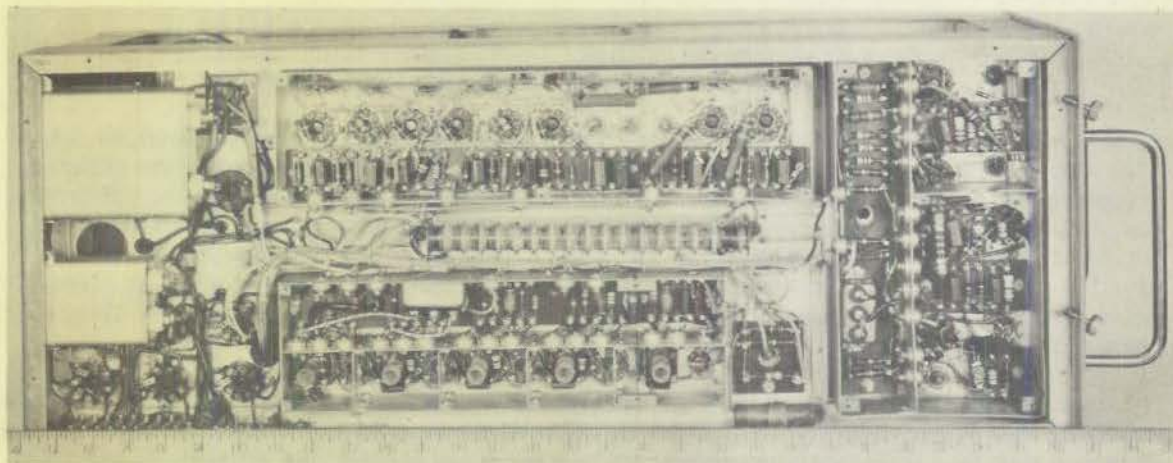


Fig. 16 - The Cancellation and Video Distribution Unit (Bottom View)

Unit No. 2, the video distribution unit, receives MTI video signal over cable No. 65 from unit No. 3 and normal video signal over cable No. 40 from unit No. 0. Its outputs are as follows:

Cable No.	Signal	Polarity	Level	Impedance
70	MTI	Neg.	50V	3000 ohms
71	Normal	Neg.	2V	75 ohms
72	Normal	Pos.	4V	75 ohms
73	MTI	Neg.	2V	75 ohms
74	MTI	Pos.	4V	75 ohms

Thus, one high-impedance, high-level MTI output is provided for direct connection to a local monitor scope and four low-impedance low-level outputs are provided for connection to the ship's radar-information distribution system.

All the power for this unit, with the exception of -75 volts bias, is supplied by a built-in electronically regulated supply. This -75 volts bias, and a gain-control voltage of 0 to -20 volts (for the delayed-video carrier-amplifier) is supplied to this unit from unit No. 0.

This unit has four front panel controls. These are the four gain controls for the four low-impedance outputs. (Attention should be called to the fact that, although the PPI channel controls are in reality gain controls, they do not change the peak output level. Their action, since they precede the limiters, is more nearly the action of a contrast control.)

The Trigger Unit

The trigger unit (Figures 17 and 18) is shown in Diagram 1 as unit No. 1. A schematic diagram of the trigger unit is shown in Diagram 8. In discussing this unit, reference will also be made to diagrams for the timing and cancellation circuits (Diagram 9).

The trigger unit receives, over cable No. 81, the delayed video and trigger signal shown as wave form No. 12 on Diagram 9. This signal is delayed from the start of the previous trigger pulse by a time, $T_L + a$

where,

$$\begin{aligned} T_L &= \text{delay line delay time,} \\ a &= 0.3 \text{ micro-sec (delay in detector output filter).} \end{aligned}$$

This signal is partially differentiated and amplified in V-101 to give wave form No. 13. This signal is in turn clipped, at the levels shown, by tube V-102 to recover the delayed trigger pulse as shown in wave form No. 14. This delayed trigger pulse is then differentiated in a very short time constant (wave form No. 15) and amplified to give wave form No. 16. This pulse is again clipped at the level shown in plate 2, by tube No. V-104 to give the pulse shown as wave form No. 17. This last differentiation and clipping gives a pulse of about 0.4 microsecond duration. This pulse is used to synchronize the delay-equalizing multivibrator.

The delay equalizing multivibrator consists of tubes V-105 and V-106. It is a one-shot multivibrator generating a pulse 1.3 ± 0.5 microseconds long as shown by wave form No. 18. This pulse is then differentiated, as shown in wave form No. 19, and clipped, by tube No. V-107 at the level indicated. Since this differentiation and clipping recovers the trailing edge of the multivibrator pulse, the resultant pulse lags the preceding synchronizing pulse by,

$$T_L + (0.3 + 1.3 \pm 0.5) \text{ microseconds}$$

or,

$$T_L + (1.6 \pm 0.5) \text{ microseconds.}$$

This pulse is then used to initiate a blocking oscillator (V-108) which develops the actual trigger pulse. This final trigger pulse is fed through cathode followers V-109 and V-110, through isolating filters, to the modulator (unit No. 7) and to the remodulator (unit No. 4) for reinjection into the delay line carrier. The isolation filters add an additional delay of 0.4 microseconds so the total delay around the trigger loop is,

$$T_L + (2.0 \pm 0.5) \text{ microseconds.}$$

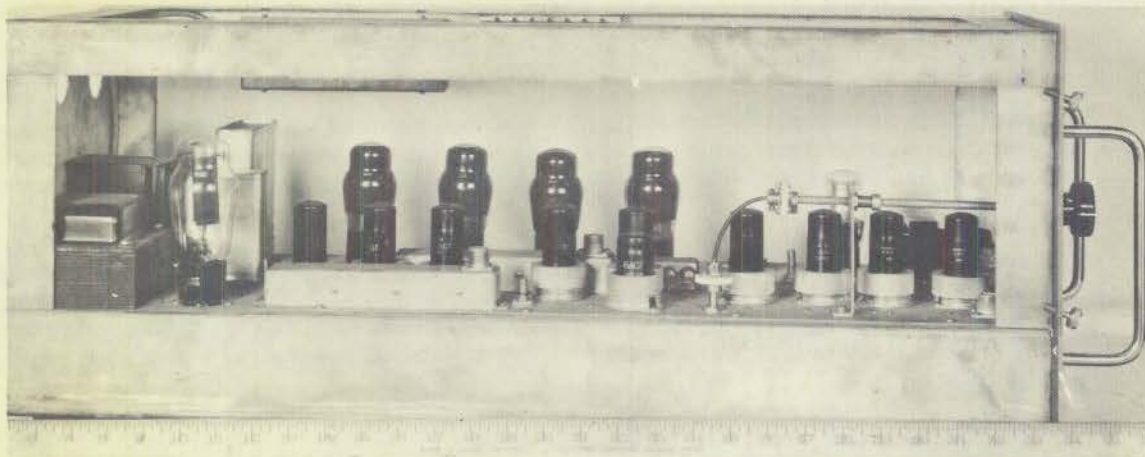


Fig. 17 - The Trigger Unit - Unit No. 1 (Side View)

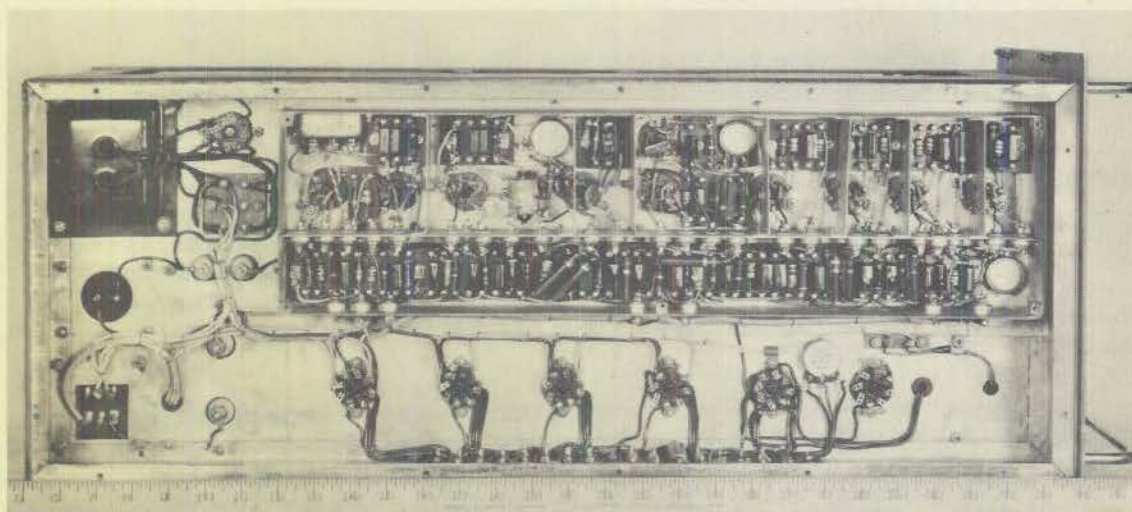


Fig. 18 - The Trigger Unit (Bottom View)

Consider the cancellation of a given echo. Obviously for cancellation to take place the delay in the delayed channel must be $T_L + 2.0$ microseconds. (The ± 0.5 microseconds is used to balance the system exactly.) Thus the two-microsecond LC delay line mentioned earlier is added in the delayed video channel.

All power for this unit with the exception of -75 volt bias is supplied by a built-in electronically regulated power supply. (The -75 volt bias is fed in from unit No. 0). This unit has one front panel control; the multivibrator adjustment which acts as the time balance adjustment.

SYSTEM PERFORMANCE

Cancellation Ratio

The cancellation ratio of the cancelling system was checked by feeding a 2 microsecond pulse into the remodulator over cable No. 60. This pulse gave a somewhat pessimistic measurement, since it had very steep sides and thus was much richer in high-frequency components than a radar echo.

The actual measurement was made by adjusting the input pulse to the desired polarity and level and then adjusting the cancelling system to best balance. Residual pulse energy was then measured. The delayed channel was then inactivated and the input pulse reduced in level to give the same output pulse energy. This reduction in level was then equal to the cancellation ratio.

The cancellation ratio was found to be as follows:

Level	Positive Pulse	Negative Pulse
5 volts	31 db	29 db
10 volts	32 db	28 db
20 volts	32 db	28 db

Thus, the average cancellation ratio is 30 db.

Attention is called to the fact that this is a true cancellation ratio. Limiting was not used to give an apparent cancellation ratio much higher than the true cancellation ratio.

Subclutter Visibility

Subclutter visibility measurements were made by injecting a noncoherent pulse of r.f. into the system while it was in normal operation over an area of considerable clutter. This pulse was supplied by a signal generator, and, although it does not seriously affect the results, the measurement had to be made with a 4 microsecond pulse because of the minimum pulse width of the signal generator.

The measurement was made by first measuring clutter strength by comparison with the signal generator output using the normal channel. The signal-generator pulse was then moved into the clutter and reduced to the minimum level visible on MTI. The ratio of clutter strength to this minimum visible echo level then gave the subclutter visibility of the system.

Subclutter visibility was found to be approximately equal to clutter level up to a clutter level of 27 db. At higher clutter levels, the subclutter visibility remained approximately constant at 27 db.

Sea Trials

No extensive discussion of the sea trials of the system will be included here as they are covered in a separate report.⁷ It would, however, be in order to mention the results obtained in these trials.

⁷ T. H. Chambers, NRL Report R-3138, "Sea Trials of the MTI Conversion Unit for SC-SK Series Radar," July 1947.

The first part of the sea trials was an investigation of the effect of the ship's speed on MTI performance. The investigation could not be carried to speeds higher than 12 knots as this was the ship's "full speed," however, at this speed, the results were found to be in agreement with the predictions.

The second part of the test was an investigation of the effect which the addition of MTI to the system had on the maximum detection range. It was found that the sensitivity of the MTI system or aircraft was essentially the same as the sensitivity of the normal system.

The third part of the test concerned the ability of the system to track low-flying planes through land clutter. In 24 runs at altitudes from 250 to 1500 feet, 80 percent hits were made using the PPI indicator. This is generally considered "solid" coverage.

The last part of the test was concerned with delay line performance under conditions of heavy weather. Although the ship rolled through a double amplitude of 30 degrees and pitched 8 to 10 degrees, there was no effect whatsoever on system performance.

System Reliability

Throughout the many hours of system operation at this Laboratory and during the 60 hours of operation afloat, the reliability of the system has been found to be excellent. All units have proven capable of operation for extended periods of time without adjustments or servicing. Stability has proven such that good balance is maintained, without adjustment, for many hours of operation.

The reliability of the delay line used in this system indicates that mercury delay lines can be designed which will operate satisfactorily aboard ship. Although it has been in service for almost a year, it has had no repair or adjustment except to be bled of an air bubble formed during the shock and vibration testing it was given shortly after its completion.

* * *

APPENDIX I

Equations Useful in the Design of an MTI System

Definition of Doppler frequency,

$$f_d = f_r - f_o \quad (1)$$

Received frequency in the radar case, (radar assumed to be stationary)

$$f_{r_2} \cong f_o \left[1 + \frac{2\nu_t^{(r)}}{c} \right] \quad (4.2)$$

Doppler frequency in the radar case, (radar assumed to be stationary)

$$f_d \cong \frac{2\nu_t^{(r)}}{c} f_o \quad (5)$$

Speed for zero response,

$$V_{\text{null}}^{(n)} = \frac{n (\text{PRF}) c}{2 f_o} \quad (13)$$

System sensitivity as a function of target radial speed

$$S_r = \left| \sin \frac{2\pi \nu_t^{(r)} f_o}{(\text{PRF}) c} \right| \quad (14)$$

System sensitivity as a function of target velocity and course

$$S_r = \left| \sin \frac{2\pi \nu_t f_o \cos \theta_c}{(\text{PRF}) c} \right| \quad (14.1)$$

Condition for velocity compensation by parameter selection

$$\frac{f_o}{(\text{PRF})} \ll 4 \times 10^5 \quad (15)$$

Coho frequency shift necessary for velocity compensation by coho shift

$$f_d \cong \frac{2\nu_s \cos \theta_r}{c} f_o \quad (16.1)$$

Required subclutter visibility - General case

$$R = \frac{\sigma_c F_c^2}{\sigma_t F_t^2} \quad (20.1)$$

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Required subclutter visibility case where clutter and target are at the same elevation

$$R = \frac{\sigma_c}{\sigma_t} \quad (21)$$

Limit on subclutter visibility due to antenna scanning

$$R_{\max} \ll \frac{(\text{PRF}) \theta_b}{15 N} \quad (23.1)$$

Limit on subclutter visibility due to velocity compensation by choice of parameters

$$R_{\max} \ll \left| \sin \frac{\pi \nu_s}{V_{\text{null}}^{(1)}} \right|^{-1} \quad (25)$$

Limit on subclutter visibility due to velocity compensation by coho shift

$$R_{\max} \ll \left| \sin \frac{\pi \nu_s \sin \theta_b / 2}{V_{\text{null}}^{(1)}} \right|^{-1} \quad (28)$$

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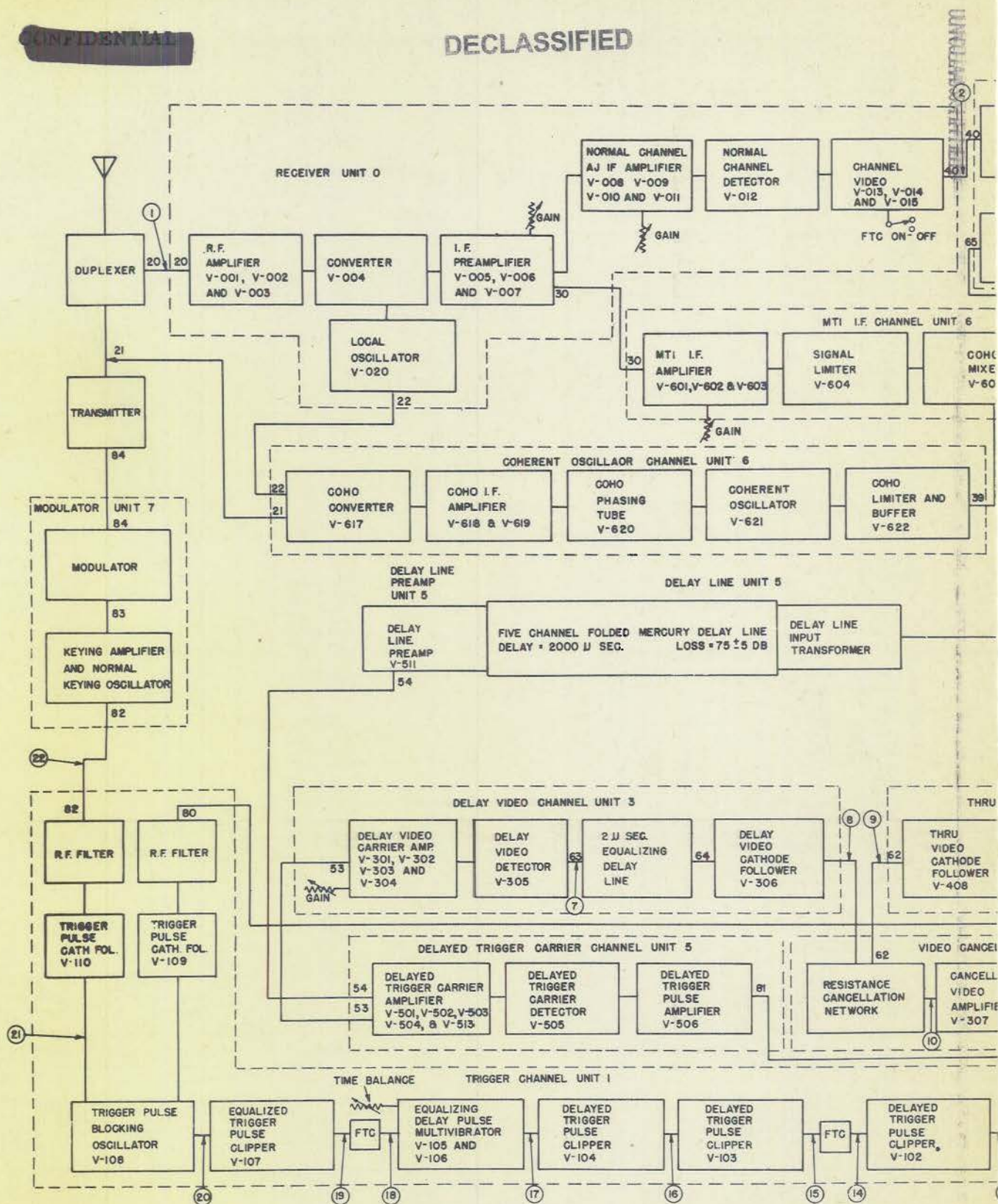
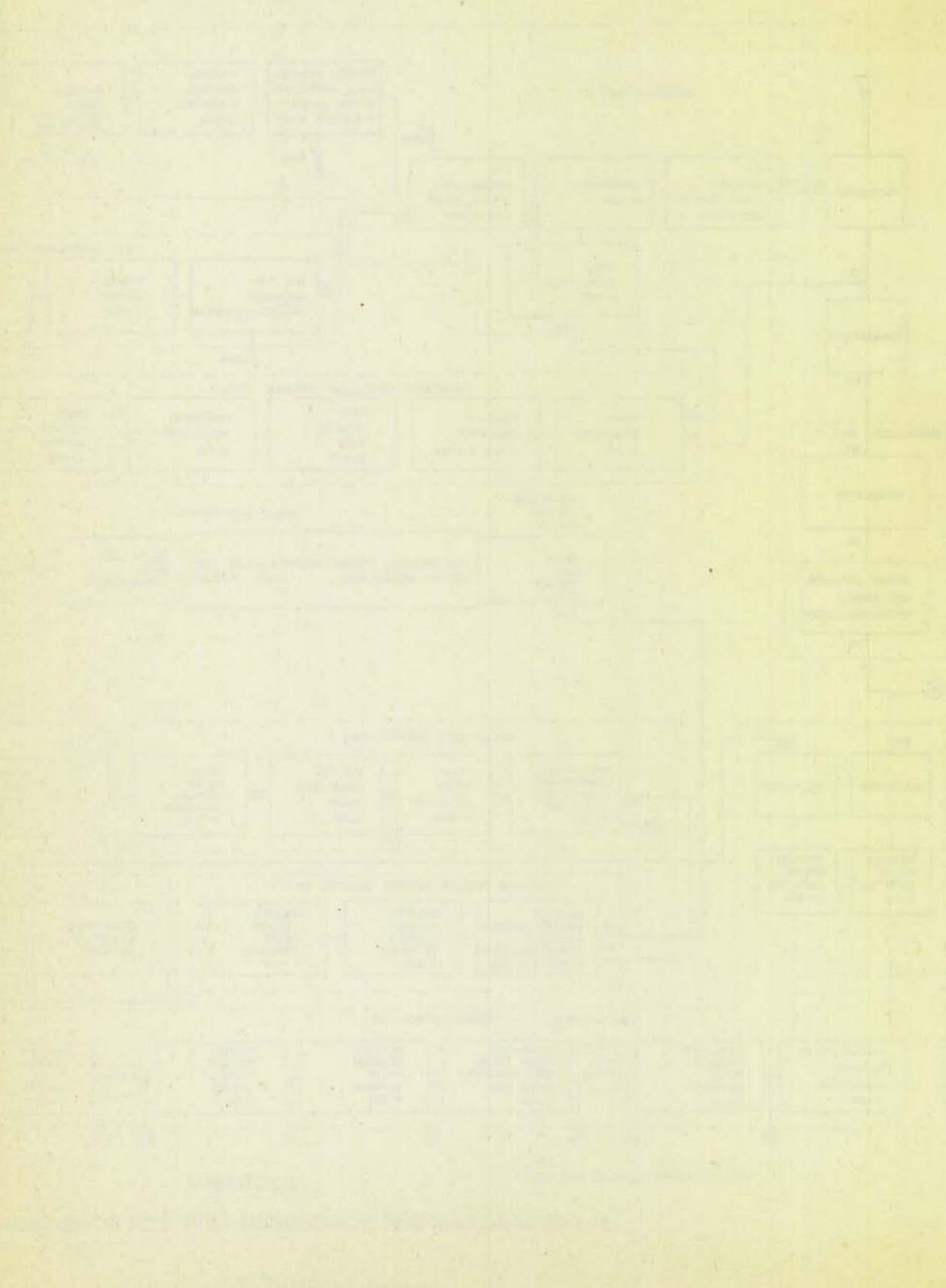
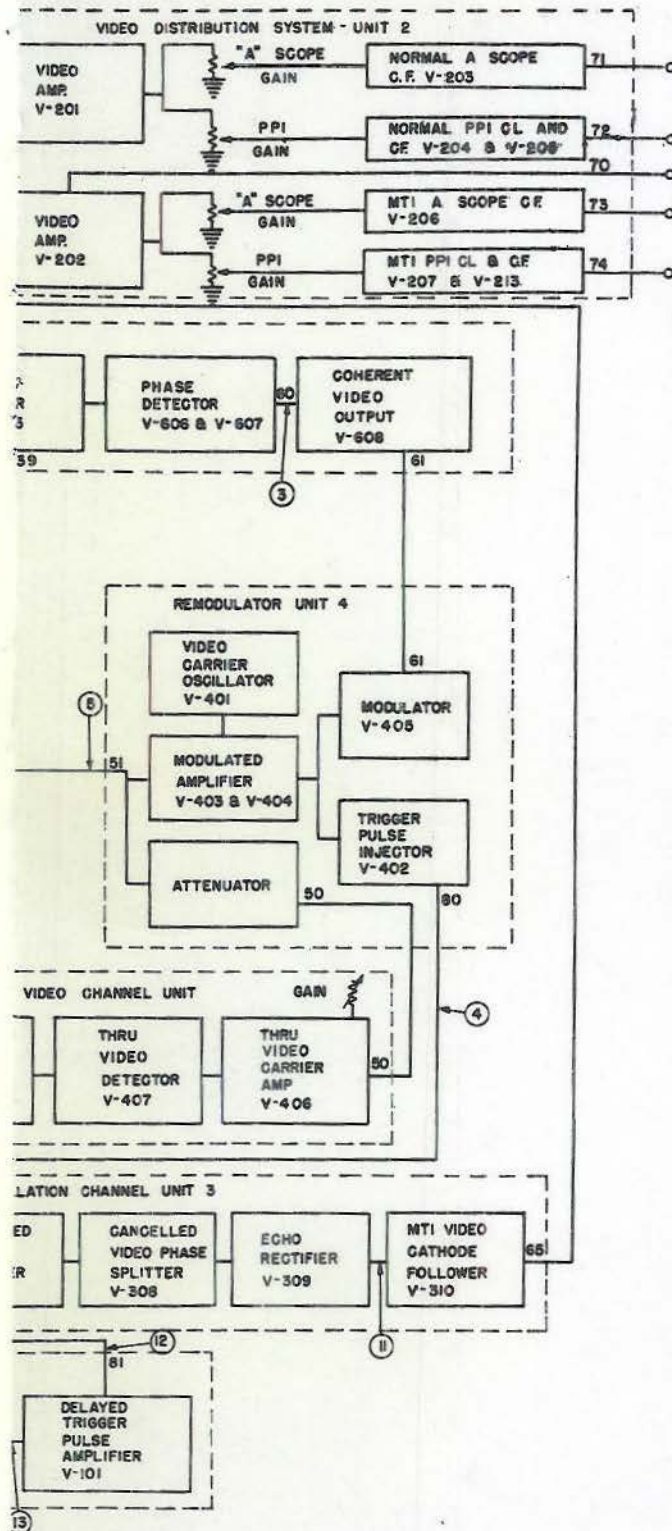


DIAGRAM 1

BLOCK DIAGRAM MTI CONVERSION UNIT FOR SC-SK SERIES R

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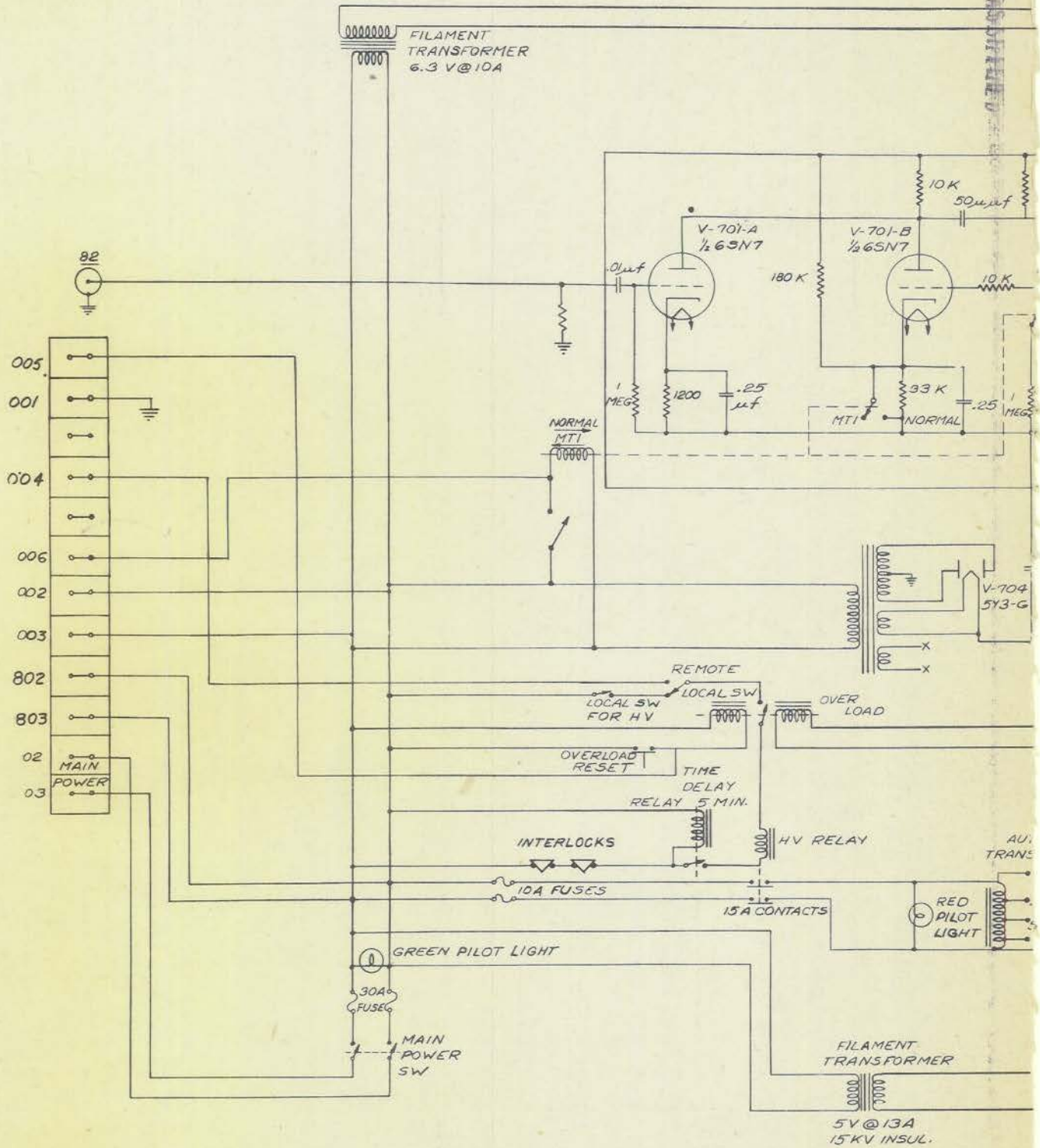




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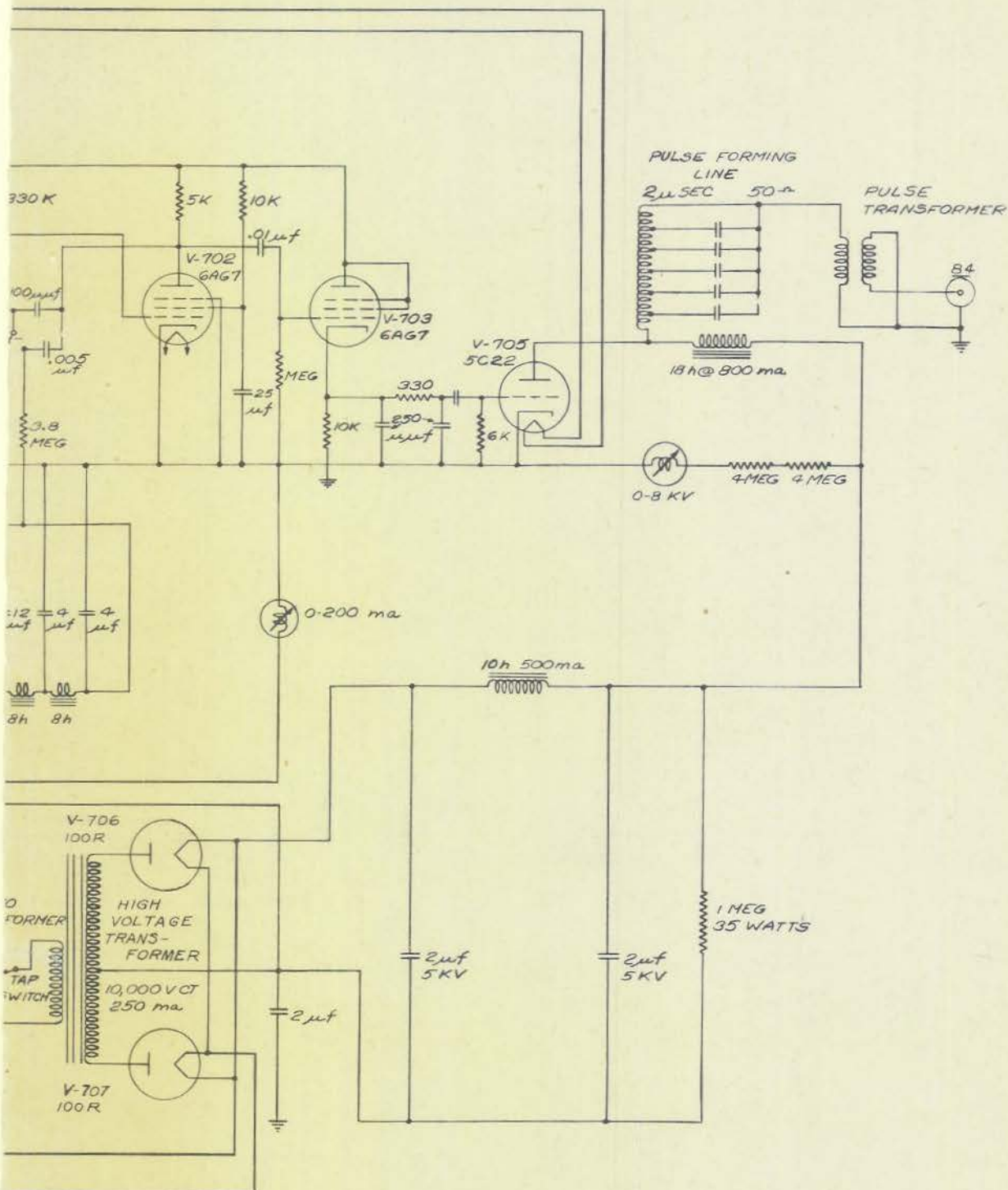
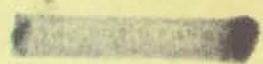


DIAGRAM 2
MODULATOR
UNIT #7
MTI CONVERSION UNIT
FOR SC-SK SERIES RADAR

43

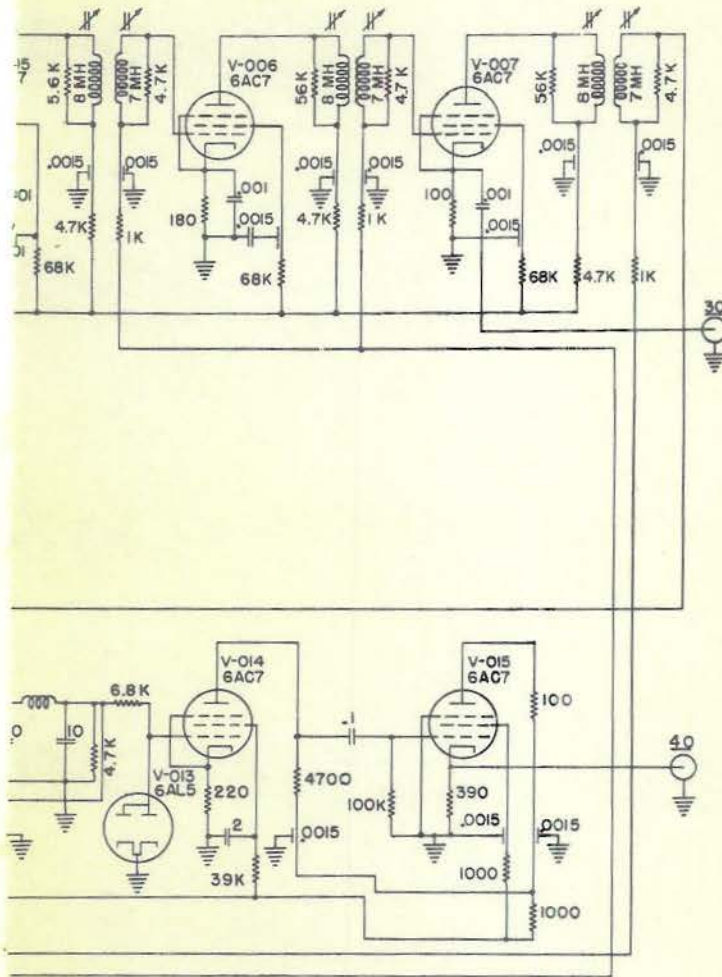


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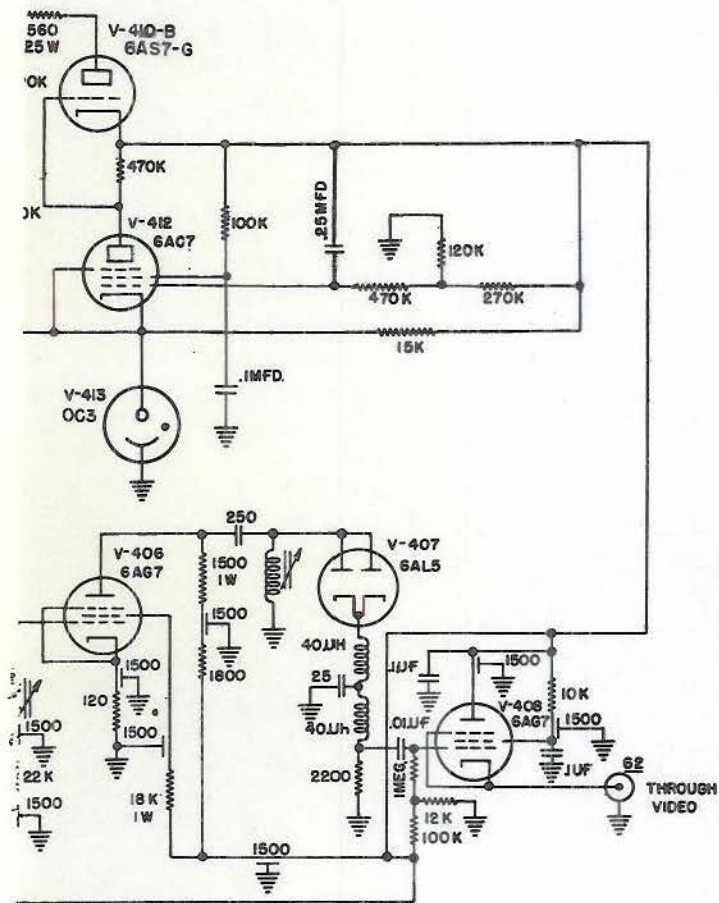
U. S. DEPARTMENT OF AGRICULTURE

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ANNEL UNIT #4
RADAR

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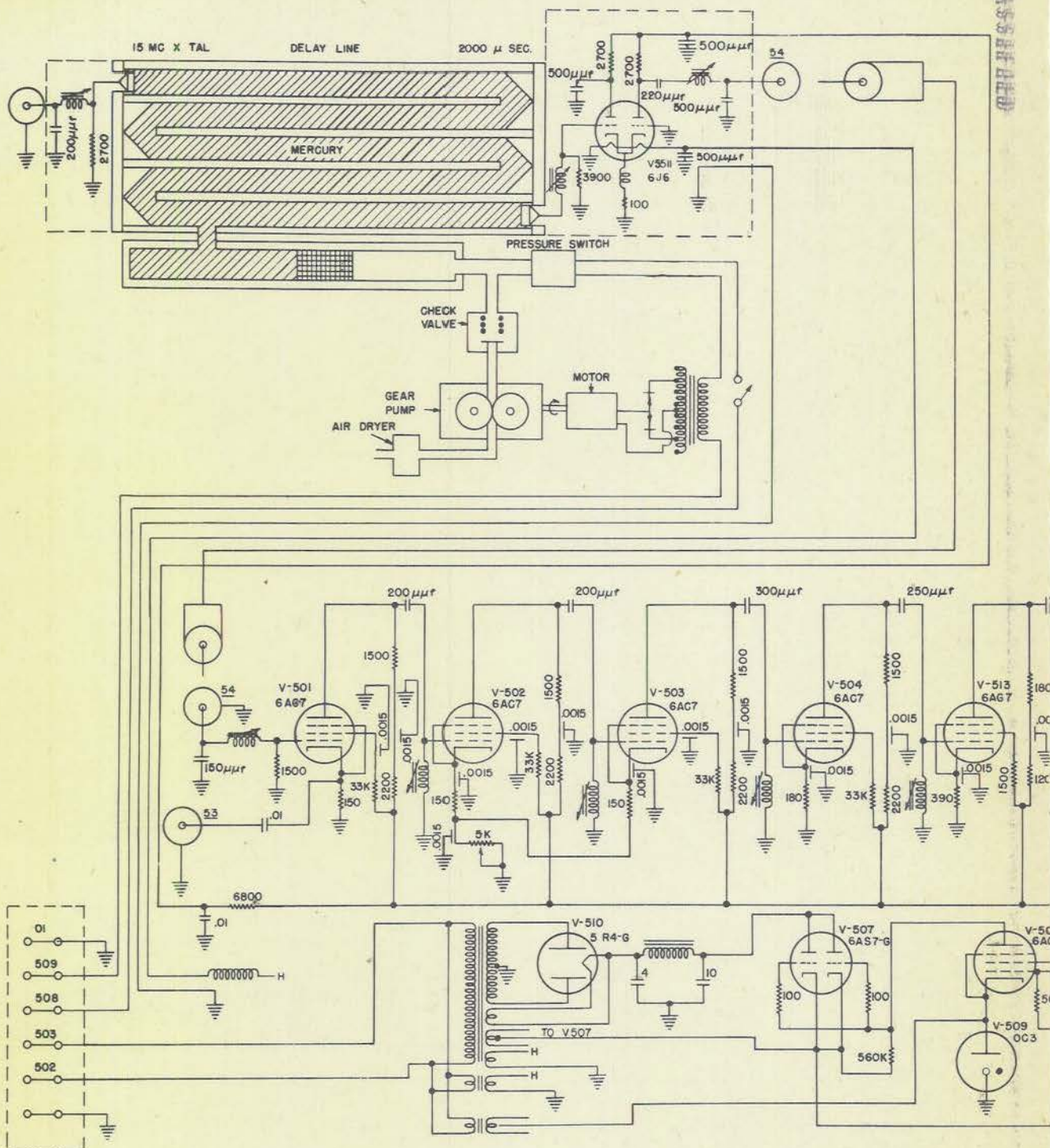
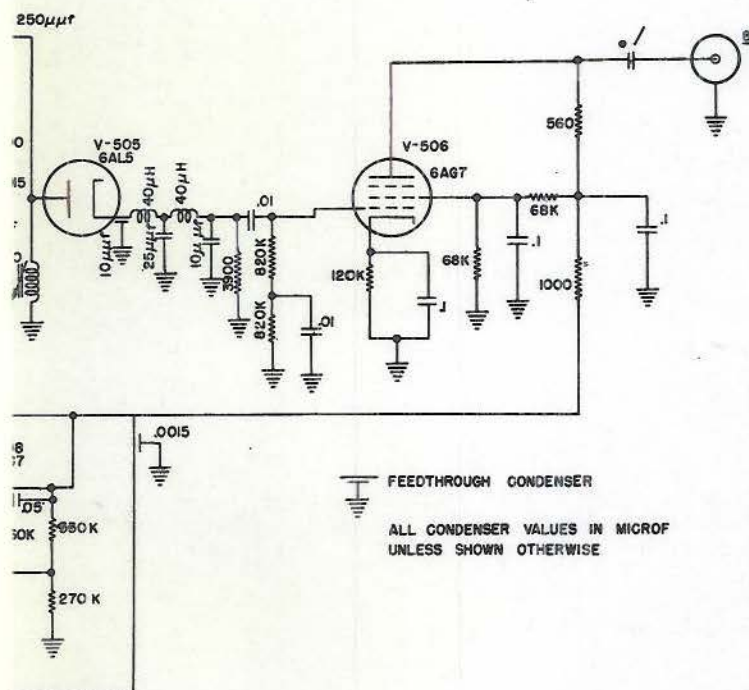




DIAGRAM 6
 THE DELAY LINE UNIT
 UNIT # 5
 MTI CONVERSION UNIT
 FOR SC-SK SERIES RADAR



51

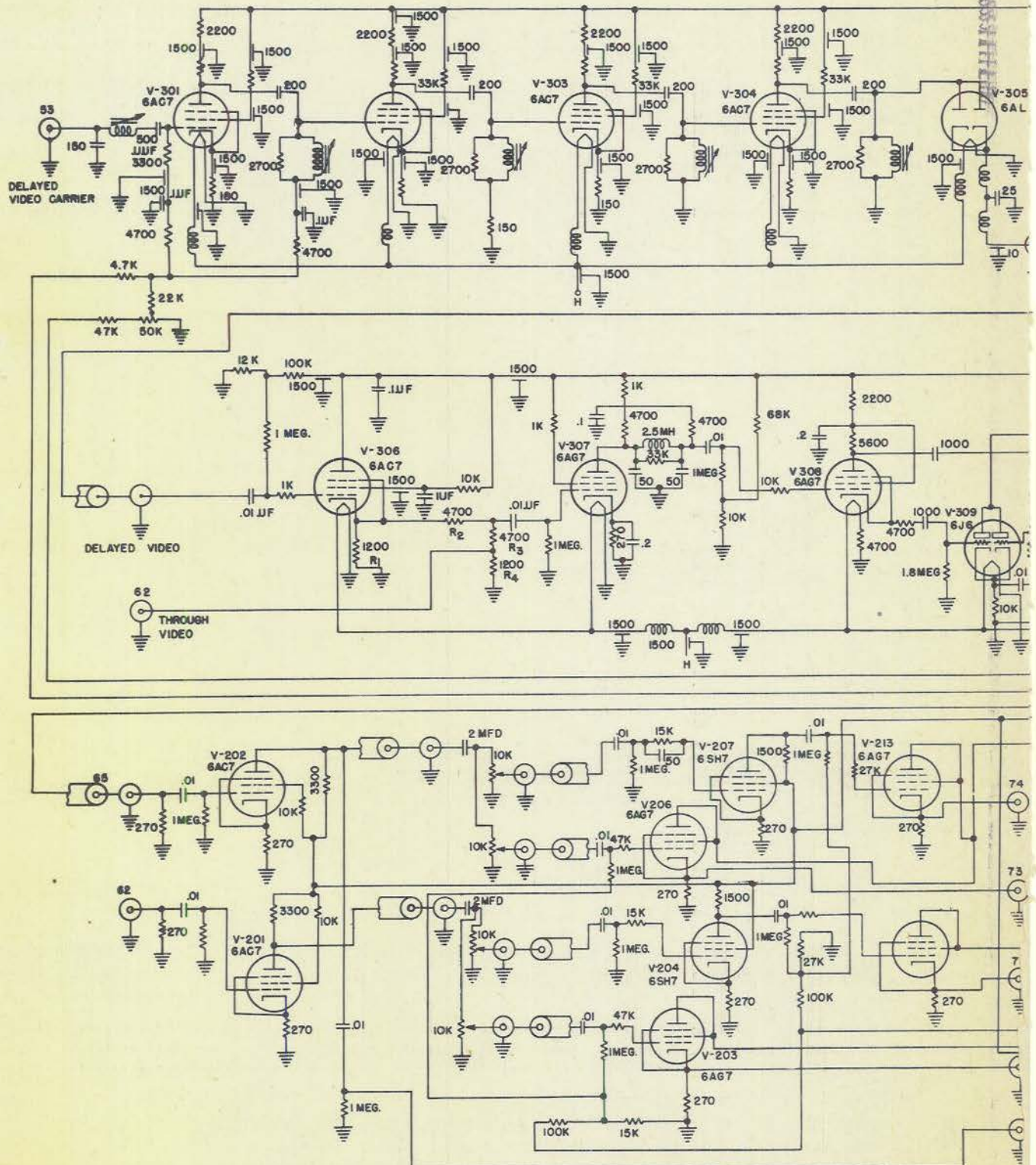


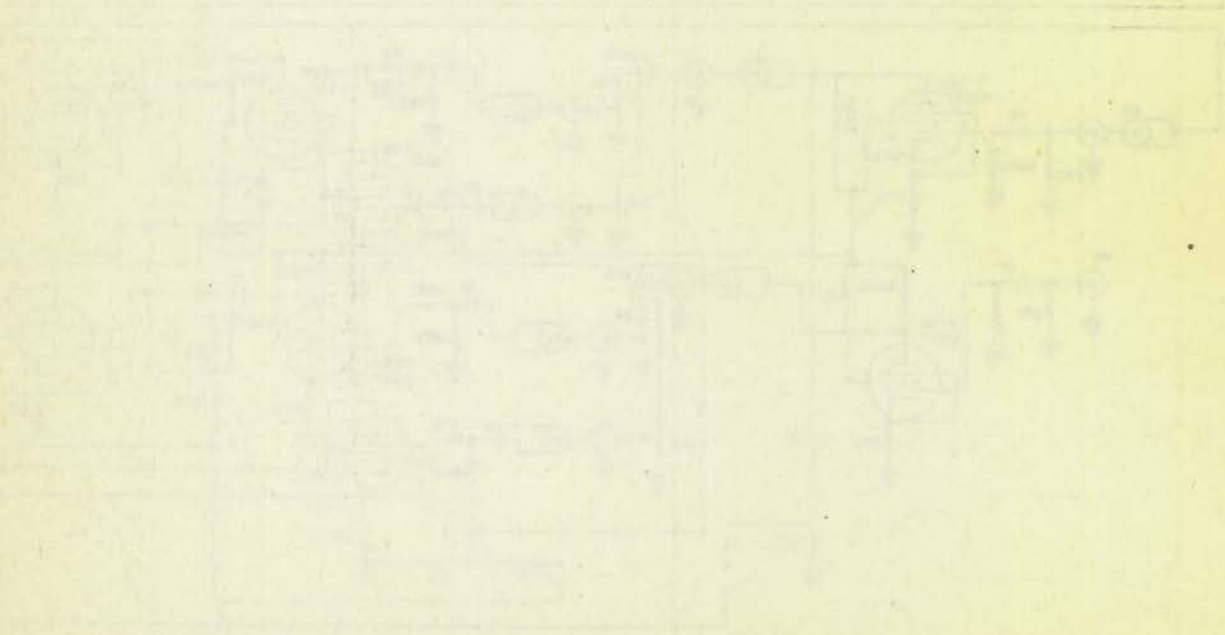
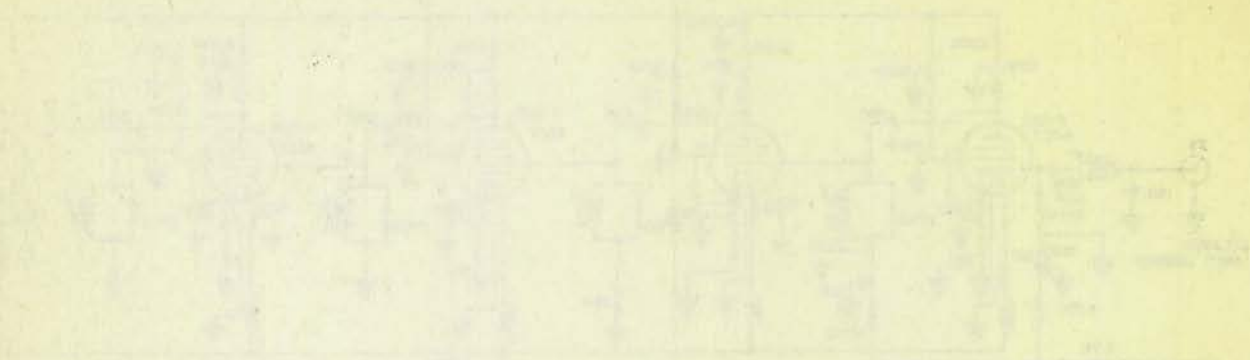
DIAGRAM 7

THE CANCELLATION AND VIDEO DISTRIBUTION UNIT
MTI CONVERSION UNIT FOR SC-SK SERIES RADA

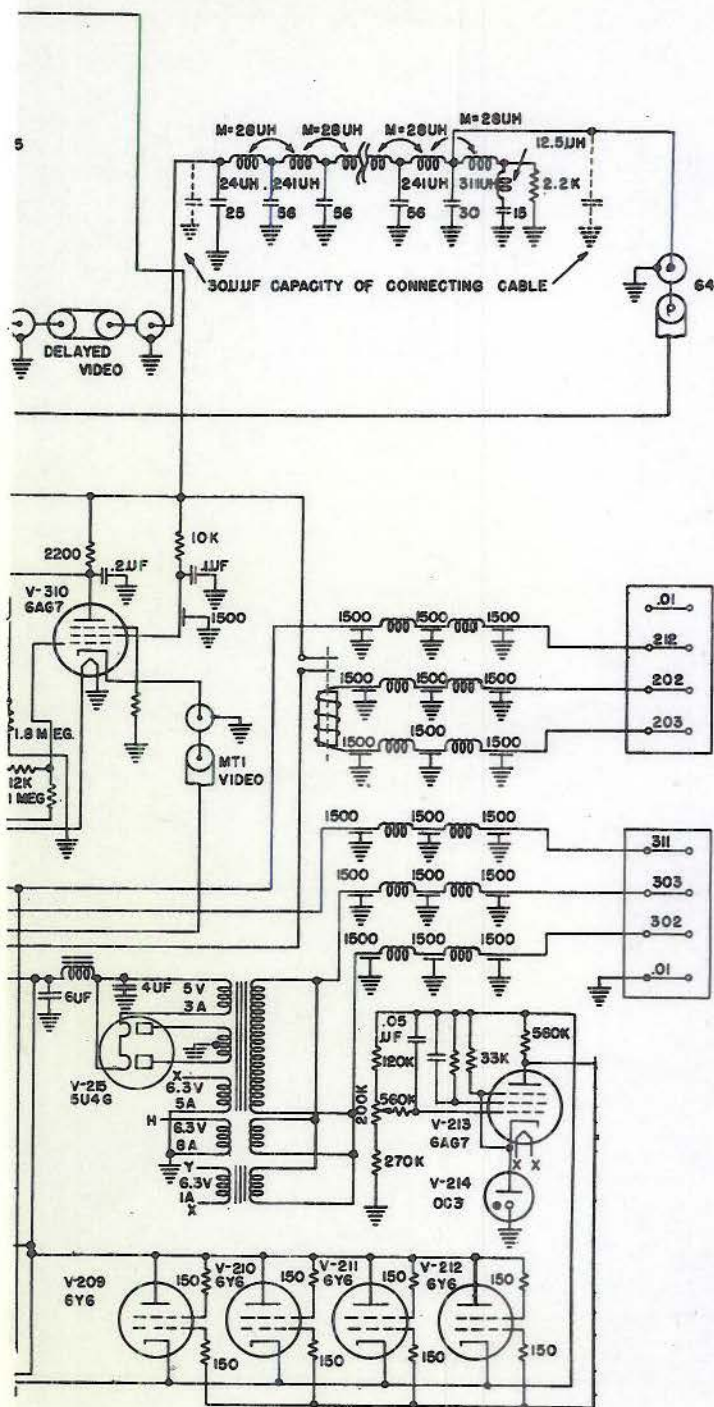
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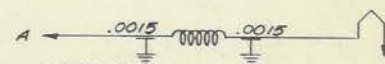


70 "A" SCOPE
VIDEO OUTPUT

2-3

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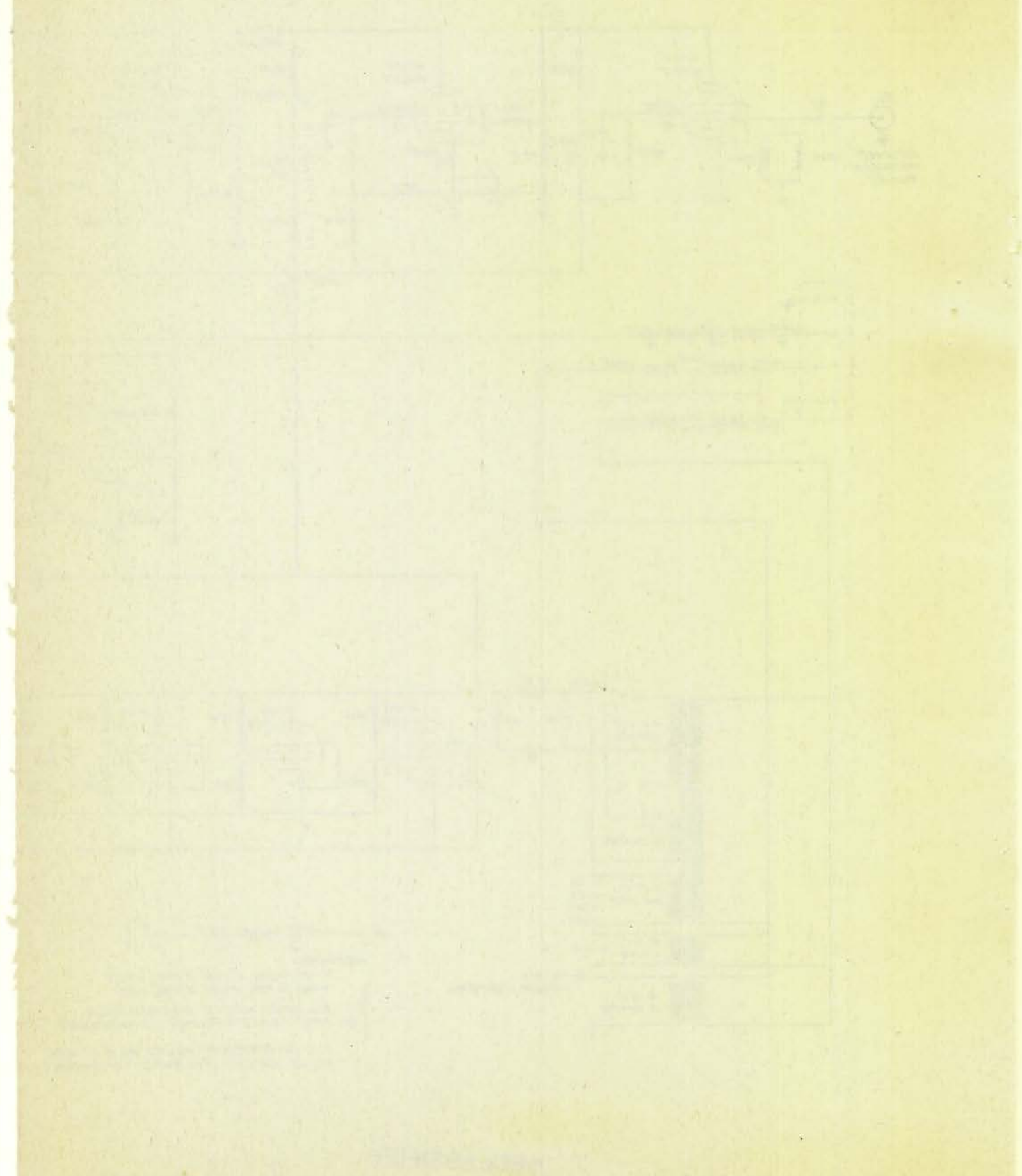
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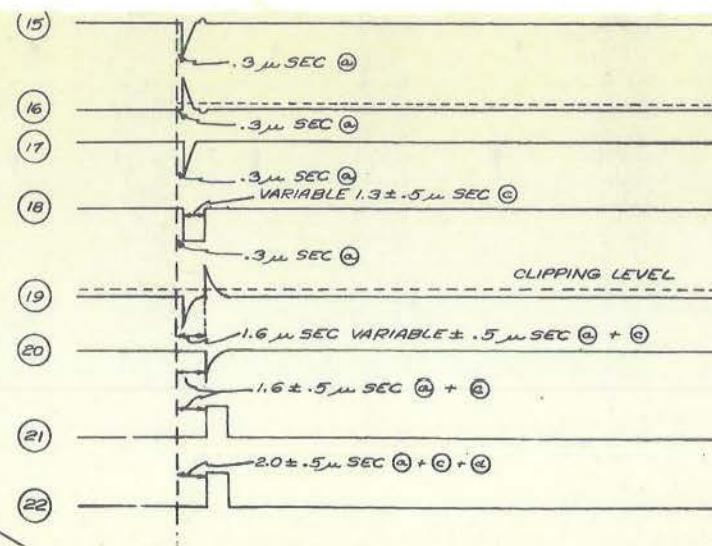
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TIMING AND CANCELLATION
OF THE Nth PULSE
MTI CONVERSION UNIT
FOR SC-SK SERIES RADAR

TRIGGER CIRCUIT WAVE FORM



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~~CONFIDENTIAL~~

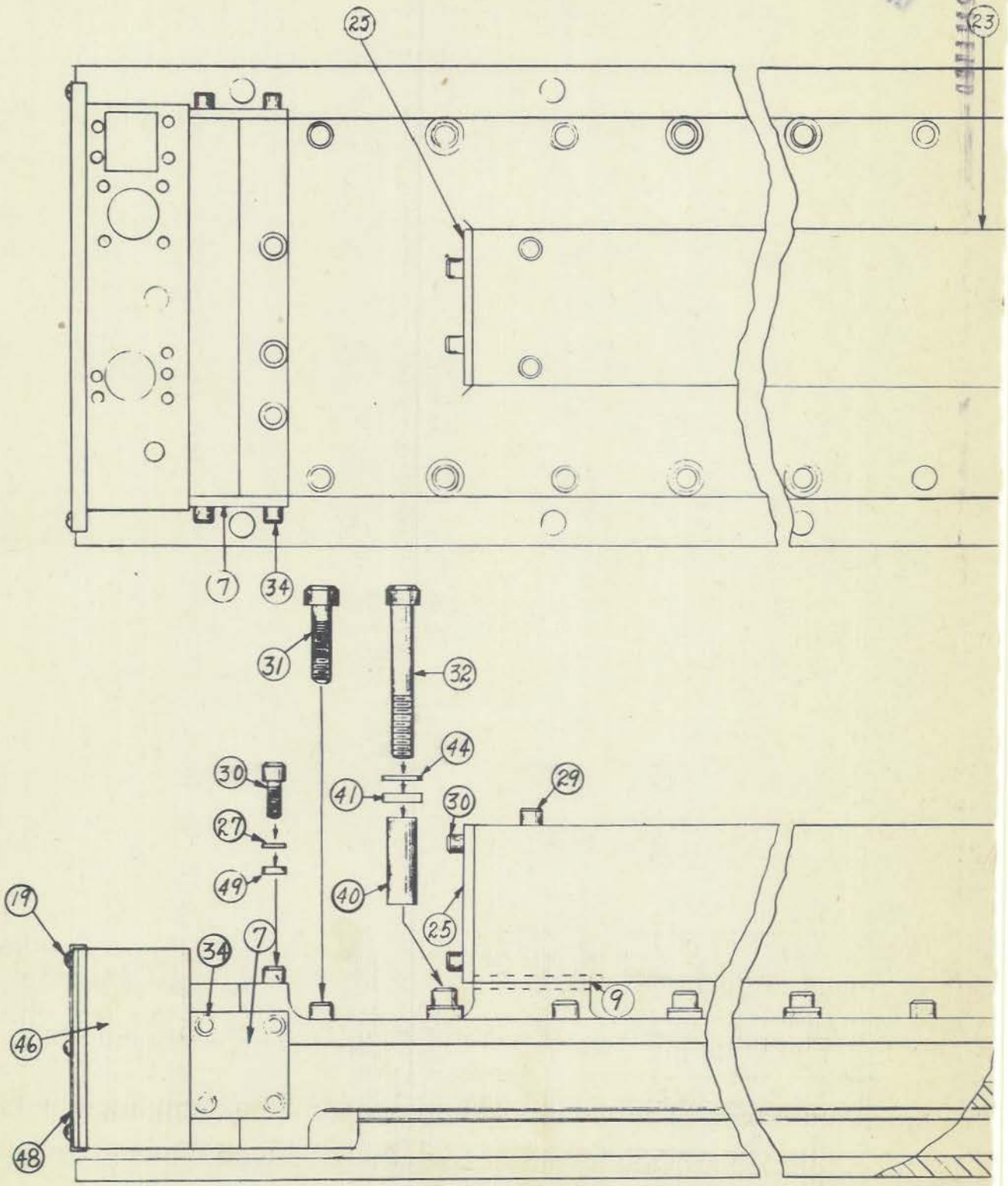


DIAGRAM 10
ASSEMBLY DRAWING - PRESSURIZED MERCURY

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