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NOSMO DOPPLER SIMULATOR

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REPORT

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RADIATION LABORATORY
MASSACHUSETTS INSTITUTE OF TECHNOLOGY
CAMBRIDGE · MASSACHUSETTS

DCR
Div. 14
OEMsr-262

Radiation Laboratory

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NOSMO DOPPLER SIMULATOR

Abstract

A unit to simulate Doppler beating phenomena for use in the airborne radar training program is discussed. The unit amplitude modulates returned signals with a random frequency spectrum whose upper frequency limit varies as the sine of the angle between the azimuth of the radiated beam and ground track of the aircraft. Practice in the determination of drift angle is therefore made possible.

W. Roth

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Title Page
25 Numbered Pages

NOSMO DOPPLER SIMULATOR

Introduction

As the design of radar equipment becomes more complex, the operator training program must be capable of duplicating the normal operating conditions to a more faithful degree. The recent development of methods for determining the drift angle of a moving aircraft by means of airborne radar has necessitated the design of a unit to be used with ground based airborne radar training equipment. Since the method is based on Doppler beating phenomena, the unit is required to simulate the Doppler beating which is observed in actual flight.

Theory

If the antenna of a search type of aircraft radar such as H2X is searchlighted, the returned signals as observed on an A scope will contain beat frequencies due to the motion of the aircraft with respect to the ground. In the design of radar trainers for such parent radar sets, it is necessary to simulate this type of Doppler beating so that radar attachments such as the Nosmo drift determination unit may be included in the training program. In order to fully understand the design problems encountered in the development of such a Doppler simulator, it will be found helpful if an elementary analysis of the formation of the Doppler beat frequencies is presented.

Figure 1 presents the geometry of the system under discussion which includes a moving aircraft and two point targets A and B located on the ground. In this discussion the earth is assumed to be flat. Both point targets are located at the same slant range R, but at different azimuth angles with respect to the ground track of the moving aircraft. It can be seen that the tilt angle of the radar spinner is the same for each target. Although the aircraft is assumed to be moving with a ground speed of v_o, it is convenient to consider the aircraft stationary and the ground targets moving under it with a ground speed of v_o.

From inspection of Figure 1, the velocity of point A with respect to the aircraft is given by:

$$v_A = v_o \cos \theta_A \cos \tau \qquad (1)$$

Similarly, the velocity of target B with respect to the aircraft is given by:

$$v_B = v_o \cos \theta_B \cos \tau \qquad (2)$$

Since relative motion exists between the aircraft and each respective target, the frequency of the received signal returned by reflection from the targets will not be the same as the transmitted frequency from the radar antenna. The relation between these two quantities is expressed by the usual Doppler formula:

$$f_r = f_t \frac{c}{c-v} \qquad (3)$$

where:

f_t = transmitted frequency

f_r = received frequency

v = relative velocity of aircraft and target

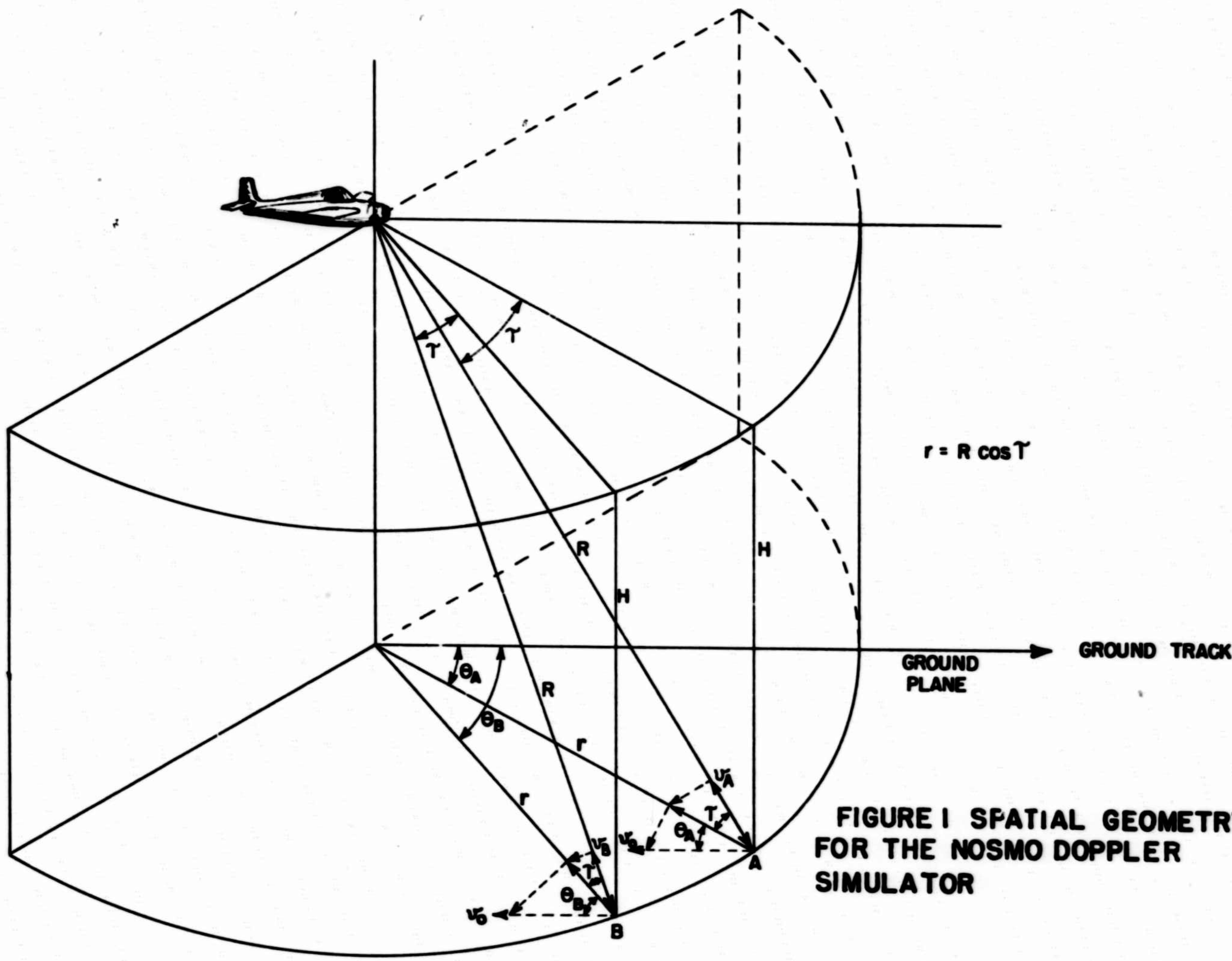

FIGURE I SPATIAL GEOMETRY FOR THE NOSMO DOPPLER SIMULATOR

c = velocity of propagation

λ_t = transmitted wave length

Since the frequencies of the reflected signals from the two targets A and B differ if the respective azimuth angles with the ground track are unequal, the difference frequency between these two received signals will appear as an amplitude modulation of the returned signal at any given range. This beat frequency Δf is developed with the use of equations (1), (2), and (3).

$$\Delta f = f_t \left(\frac{c}{c - v_o \cos \theta_A \cos \tau} \right) - f_t \left(\frac{c}{c - v_o \cos \theta_B \cos \tau} \right) \quad (4)$$

$$\Delta f = f_t c \left(\frac{c - v_o \cos \theta_B \cos \tau - c + v_o \cos \theta_A \cos \tau}{c^2 - c\, v_o \cos \tau \{\cos \theta_A + \cos \theta_B\} + v_o^2 \cos \theta_A \cos \theta_B \cos^2 \tau} \right) \quad (5)$$

Since $c \gg v_o$, the last two terms of the denominator can be neglected.

$$\Delta f \cong \frac{f_t\, v_o \cos \tau}{c} (\cos \theta_A - \cos \theta_B) \quad (6)$$

Since $\lambda_t = c/f_t$

$$\Delta f \cong \frac{v_o \cos \tau}{\lambda_t} (\cos \theta_A - \cos \theta_B) \quad (7)$$

The characteristics of the Doppler beat frequencies obtained from the two point targets can be summarized by examination of equation (7). The beat frequency Δf is directly proportional to the ground speed of the aircraft, inversely proportional to the transmitted wavelength, directly proportional to the cosine of tilt angle and directly proportional to the difference of the cosine of the azimuth angles of the targets with respect to the ground track. For normal applications the aircraft speed, transmitted wavelength, and spinner tilt angle will not be varying to any great extent. Thus we can write the further equation (8) in which the above mentioned terms are considered to have negligible variation with respect to the magnitude of variation occurring in the bracketed term.

$$\Delta f = K (\cos \theta_A - \cos \theta_B) \qquad (8)$$

K = constant

To investigate further, the behavior of Δf will be examined for a fixed angular difference of θ_A minus θ_B as the absolute magnitude of these angles is varied from 0 to 90 degrees. For example, we can assume that the two point targets are displaced 5 degrees with respect to one another. If the targets are placed so that they straddle the ground track, then $\cos \theta_A - \cos \theta_B = 0$ and the Doppler beat frequency is absent. As the mean position of the two targets is shifted, the bracketed term increases reaching a maximum when the two targets straddle the 90 degree axis line. Hence, Δf varies as the sine of the mean azimuth angle of the two targets. In the usual radar application Δf varies from a minimum of two or three cycles per second when the targets are along the ground track to a maximum of the order of 400 cycles per second when the targets are perpendicular to the ground track.

In the above analysis it is shown that the one beat frequency obtained from two point targets varies as shown by equation (9).

$$\Delta f \cong K \sin \theta \quad \text{where} \quad \theta = \frac{\theta_A + \theta_B}{2} \qquad (9)$$

If we increase the number of point targets to three, the number of Doppler beat frequency components increases to three. In general, for a number N individual point targets the number of beat frequencies present equals the number of N things taken two at a time or $\frac{N(N-1)}{2}$

When flying over normal ground, the number of point targets within the radiated beam at a given range is very large. It is more likely for a large number of closely spaced point targets to be present within the beam than a large number of targets spaced widely apart. Thus, there are a greater number of low beat frequency components present than the higher frequency components and although the highest frequencies present vary as sin Θ, there will be low beat frequencies since point targets very closely spaced are always present. In summarizing the above discussion, the Doppler beat frequencies always contain the summation of a large number of frequencies with components extending from zero to an upper limit determined by the beam width of the radar set used. This upper beat frequency limit varies as the sine of the spinner angle as given by equation (9).

The above explanation has considered the phenomena taking place at a given range. However, a similar thing occurs at any other given range. It can be readily seen that the phases of the beat frequencies occurring at other ranges will be completely independent from the phase of the frequency in the above case since the distribution of targets need not be the same at different ranges. Therefore, any scheme for Doppler simulation should produce beat frequencies giving random phasing with range. Since the ground over which the aircraft is flying continually changes in character, the arrangement of point targets changes from any instant to the next. Hence, the phase and frequency components present at any given range should vary in a random manner with respect to time. Here again, the simulation scheme must be capable of this duplication.

In the development of equation (8), it was assumed that the pitch angle and hence cos Υ is a constant. This is certainly true for normal usage. However at short ranges where Υ becomes large, the rate of change of cos Υ with range also becomes large. Since the normal transmitted beam has a finite width, at short ranges the values of cos Υ for targets within the beam differ to a considerable degree. This causes modulation as the aircraft flies over uneven ground since K in equation (8) is no longer a constant. The effect at short ranges is to modulate the returned signals even though the spinner angle coincides with that of the ground track. Therefore, it is not advisable to observe the frequencies on the expanded sweeps. This feature should also be duplicated in any simulation scheme so that the student is trained to observe the Doppler beating on the proper portion of the correct range sweeps.

If the effects of the Doppler beat frequencies are to be accurately simulated the following features must be included: a) formation of a frequency spectrum from zero cycles per second and up whose components are completely random in phase; b) variation of the upper limit of this frequency spectrum in accordance with the sine of the angle between the spinner and the ground track; c) amplitude modulation of returned signals by the random spectrum in such a manner that the modulation is completely random with respect to range; d) modulation at short ranges even at zero spinner angle.

Circuit Discussion

The manner in which the above mentioned features have been incorporated into a simulator unit for trainer purposes can be seen from the block diagram Figure 2 and typical wave forms taken at various parts of the unit. A gas tube whose spectrum is shown in Figure (3a) is used as a source of random frequency noise. Here it is observed that the spectrum has a peak in the region of 750 kilocycles. Another feature of the spectrum at this particular frequency region is that the voltage output remains fairly constant with variation of tubes so that greater stability of noise output is obtained if this region is utilized. Since the desired noise spectrum must start at a frequency of zero cycles per second, a 750 kilocycle local oscillator is used to beat with the noise spectrum in order to produce the spectrum shown in Figure (3c). The spectrum obtained from the beating is passed through a lowpass filter whose band width can be varied in proportion to the angle between the actual ground track of

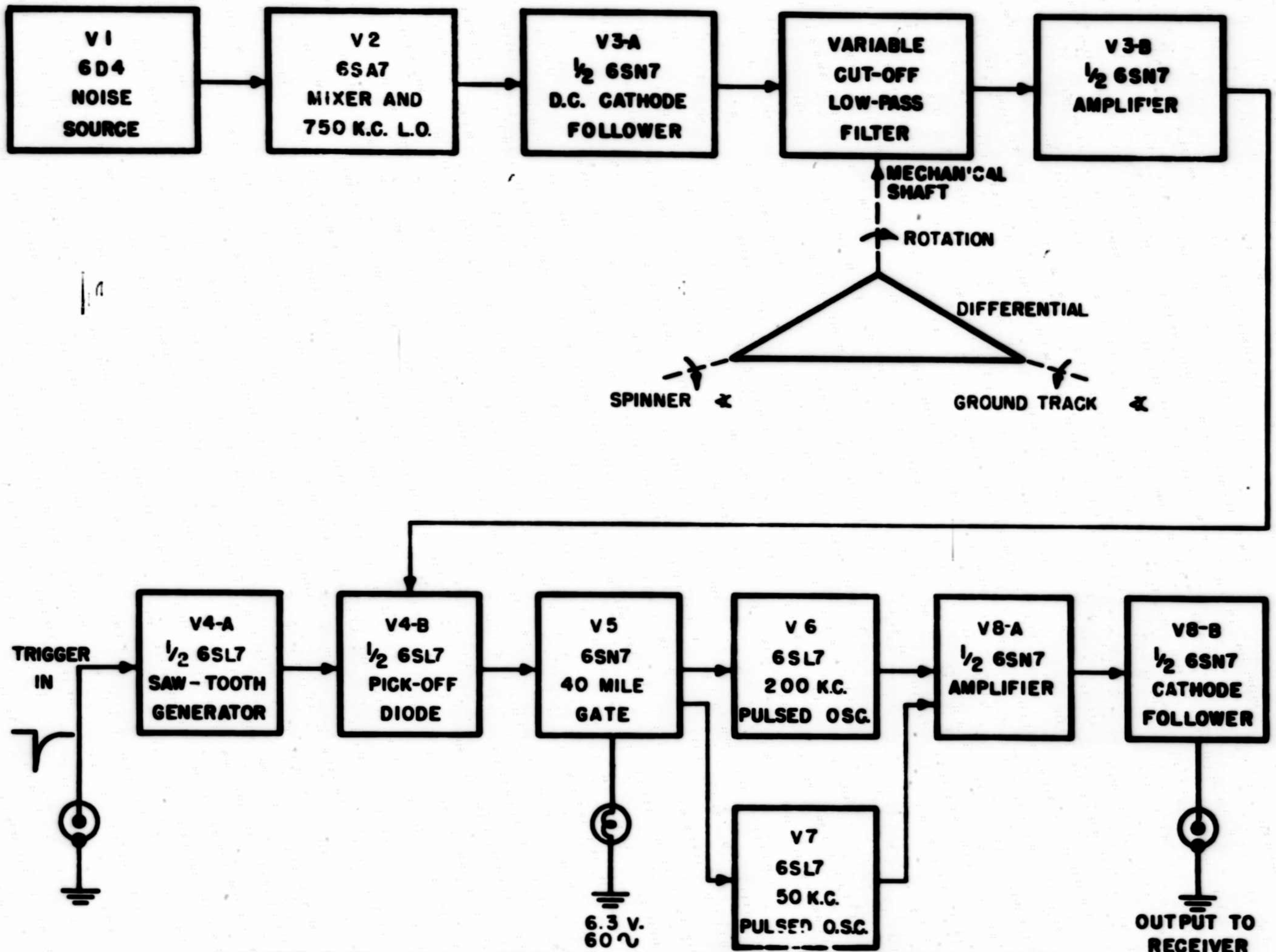

FIGURE 2 BLOCK DIAGRAM THE NOSMO DOPPLER SIMULATOR

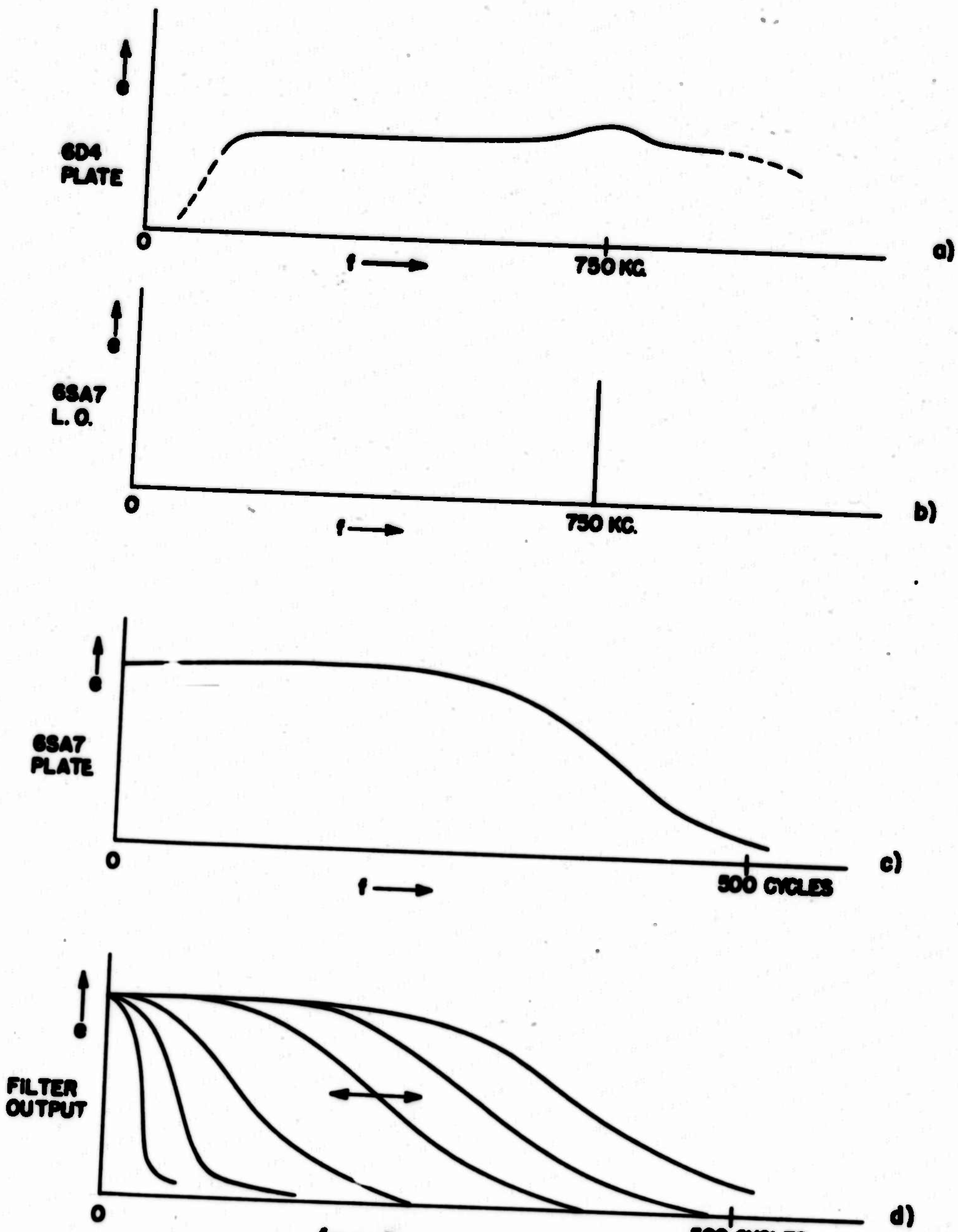

FIGURE 3 FREQUENCY SPECTRUM DEVELOPMENT

the simulated aircraft and the azimuth angle of the spinner. Hence, the spectrum of random frequencies contained in the output of the filter is a function of this angle as shown in Figure (3d). When the spinner angle coincides with that of the ground track, the upper frequency cutoff of the filter will be very low, that is, in the order of two cycles per second; whereas, if the spinner angle approaches 90 degrees to the ground track, the cutoff frequency of the filter will increase to approximately 400 cycles per second. Hence, we have obtained a source of random frequencies which is controlled by the angle between the ground track and the spinner heading in the desired manner. The method by which this noise spectrum is used will now be discussed.

A negative trigger which synchronizes the range sweeps in the radar set must be brought into the simulator unit. A positive sawtooth is developed from these triggers as shown in Figure 4, a and b. The noise spectrum described above is applied to the cathode of a pickoff diode so that the output from this diode is varied in the manner shown in Figure 4c. Since the instantaneous value of the cathode potential varies in a random manner with the frequency components present in the noise spectrum, the time at which the sawtooth obtained from the pickoff diode begins will also vary in a random manner and with frequency components corresponding to those in the spectrum of the filter output. The output from the pickoff diode is used to trigger a 40-mile gate circuit. Since the diode output is jittered in time, the 40-mile gate will also be jittered in time so that the gate will start at a different time with respect to the start of each sweep of the parent set. Positive and negative 40-mile gates are obtained from the same gate circuit. The negative gate is employed to turn on a 200 kilocycle pulsed oscillator and the positive gate is used to turn off a 50 kilocycle pulsed oscillator.

Since the 200 kilocycle pulsed oscillator is started by the 40-mile negative gate, its oscillations will be phase modulated with respect to the start of the radar sweep in accordance with the information contained in the noise spectrum previously described. This is shown in Figure 4f. The trailing edge

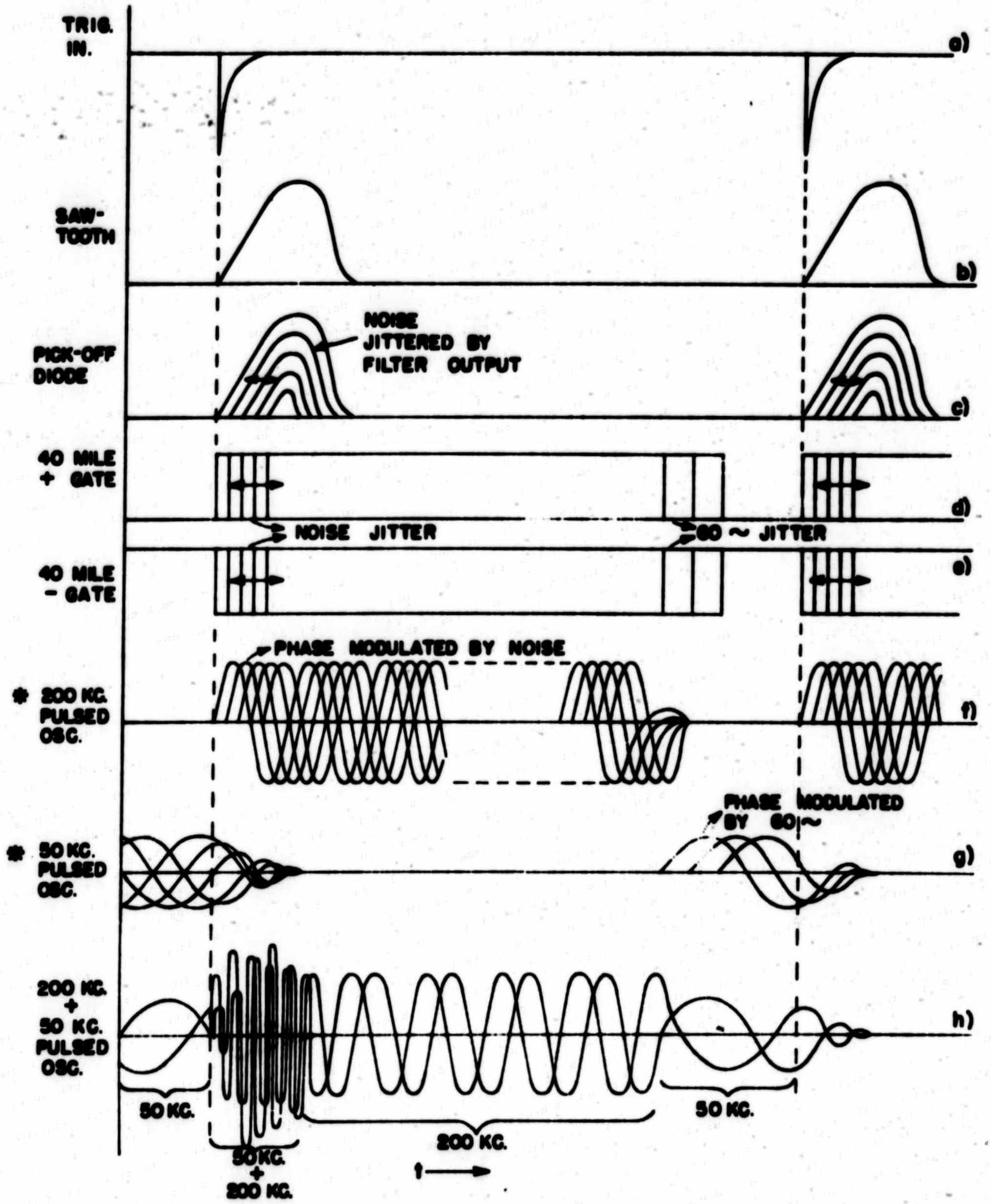

FIGURE 4 VOLTAGE WAVE FORMS FOR THE NOSMO-DOPPLER STIMULATOR

of the 40-mile gate is jittered in time by a 60 cycle voltage injected into the gate circuit. This jittering is constant and independent of spinner position. Since the positive gate is employed to turn off the 50 kilocycle pulsed oscillator, the time at which the 50 kilocycle oscillations begin is determined by the trailing edge of the 40-mile gate. This is jittered by the 60 cycle voltage so that the 50 kilocycle oscillations are phase modulated at the constant 60 cycle rate. As has previously been mentioned, this action is independent of the spinner positions. Since the 60 cycle jitter on the trailing edge of the gate is superimposed upon the jitter present in the leading edge of the gate, the 50 kilocycle pulsed oscillations are jittered by the summation of the noise jitter, which is a function of spinner position, and the 60 cycle jitter, which is constant. Hence, even though the antenna may be along the ground track which results in little noise jitter, the 60 cycle jitter is sufficient to phase modulate the 50 kilocycle pulsed oscillations by a considerable degree. This is shown in Figure 4g.

The output of the simulator comprises the summation of the 200 and 50 kilocycle pulsed oscillations which gives the resultant wave form as shown in Figure 4h. The pulsed oscillators are designed so that the effect of the superposition of the 50 kilocycle oscillation dies out at a time equivalent to approximately eight miles from the start of the sweep of the parent radar set. The 200 kilocycle oscillations are then present until the end of the 40 mile gate whereupon the 50 kilocycle oscillations reappear. Hence, the summation only appears in the overlapping interval from 0-8 miles. Since the 50 kilocycle oscillations are phase modulated even though the spinner angle may be zero, the echoes appearing along the sweep within the 0-8 mile region will be amplitude modulated so that the student will be taught not to use this portion of the sweep.

The action of the phase modulated oscillators when the simulator output is employed to amplitude modulate the radar receiver can be explained with the use of Figure (5). Figure (5a)

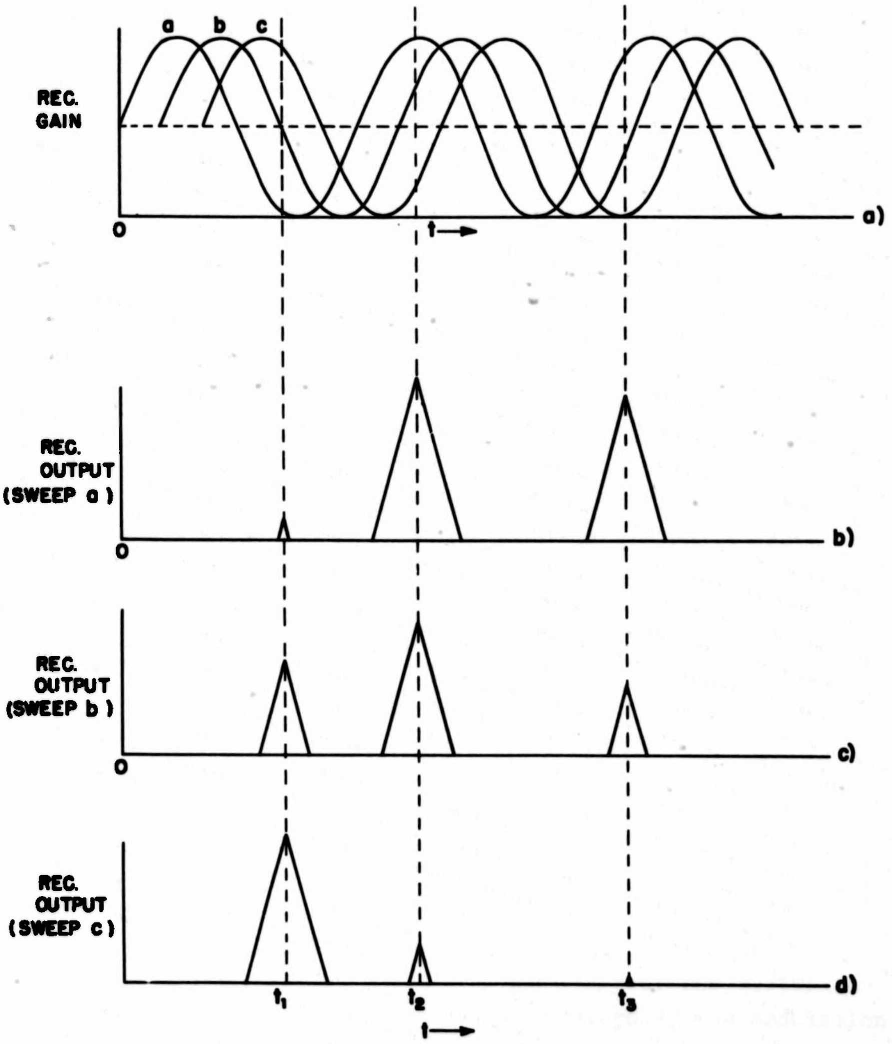

FIGURE 5 MODULATION EFFECT ON SIGNALS.
THE NOSMO DOPPLER SIMULATOR

shows the instantaneous gain of the receiver as a function of time for the three different sweeps, a, b and c. The horizontal dotted line indicates the gain of the receiver with the simulator unit turned off while the sinusoids a, b and c drawn for three different sweeps are drawn arbitrarily to indicate 100% modulation. The effect of such a phase modulated sinusoidal modulation signal is shown by Figure 5b, c and d. In these diagrams three targets are shown at ranges corresponding to time delays after the synchronizing trigger of t_1, t_2 and t_3. During sweep a, the gain of the receiver is varied in time by the sine wave labeled as shown in Figure 5a. Hence, the relative height of the three targets shown in Figure 5b can be obtained from this former curve. The height of the targets during sweep b will be controlled by the instantaneous value of the modulating sinusoid during sweep b. Throughout sweep c the relative height of the targets will be controlled by sinusoid c. By comparison of curves b, c, and d of Figure 5, it is seen that the height of targets at different ranges will vary in a manner determined by the phase modulation on the modulating signal. Since the modulating signals are phase modulated by the random noise voltage, the amplitude modulation of the signals at any given range is also of a random nature. The frequency components in the random modulating signal are determined by the position of the lowpass filter cutoff. Since this is controlled by the angle between the spinner and aircraft ground track, the amplitude modulation of signals is also controlled by this angle as is desired.

One of the stated conditions for proper simulation was that the amplitude modulation must be random with range. Strictly speaking, the simulation scheme does not provide this since the height of each target at any given range is determined by the phase of the modulating signal at the start of the particular sweep in question. So far as the eye is concerned, the modulation with range appears to be random in nature because the rate of change of gain per degree phase modulation of the modulating signal is variable with range. This is due to the fact that the slope of the sinusoid is not a constant. For instance, some targets appearing at a time corresponding to the top of the sine wave will not have much change in gain from sweep to sweep whereas targets appearing at a time corresponding to that at which the sinusoid passes through its axis will have the maximum change in gain from sweep to sweep. Therefore, some targets will appear to be changing slowly in amplitude whereas others will be changing rapidly. This can also be seen by examination of the signals of b, c, and d of Figure 5. Thus, the original criteria have been fulfilled by the method outlined above.

Description of Actual Circuits

The schematic diagram of the simulator unit is shown in Figure 6. The noise source employed is a 6D4 miniature gas thyratron with a permanent magnet surrounding the tube. The magnet is used to suppress inherent oscillations within the gas plasma which develops a non-uniform frequency apectrum from tube to tube with respect to the absolute noise level at any given frequency. Since a very narrow band of frequencies is actually required (a band of only 500 cycles in width) a tuned circuit comprising L1 and C2 is used to remove the undesired frequencies from appearing on the grid of the 6SA7 converter stage V2. This feature prevents overloading of the tube. The 750 kilocycle local oscillator is also incorporated in this stage by a standard oscillator circuit. The output from the plate of V2 consists of the random frequency spectrum from zero to approxiamtely 500 cycles. Capacitor C6 is employed to eliminate the undesired frequencies above 500 C.P.S. which would ordinarily appear at the output of the converter stage. The first half of tube V3 is used as a direct coupled cathode follower in order that the low-pass filter be driven from a low impedance source. A detailed discussion of this low-pass filter follows.

$$\frac{E_2}{E_1} = \frac{1}{\left[1 - (\omega T)^2\right] + 3j\omega T} \quad \text{where } T = RC \qquad (10)$$

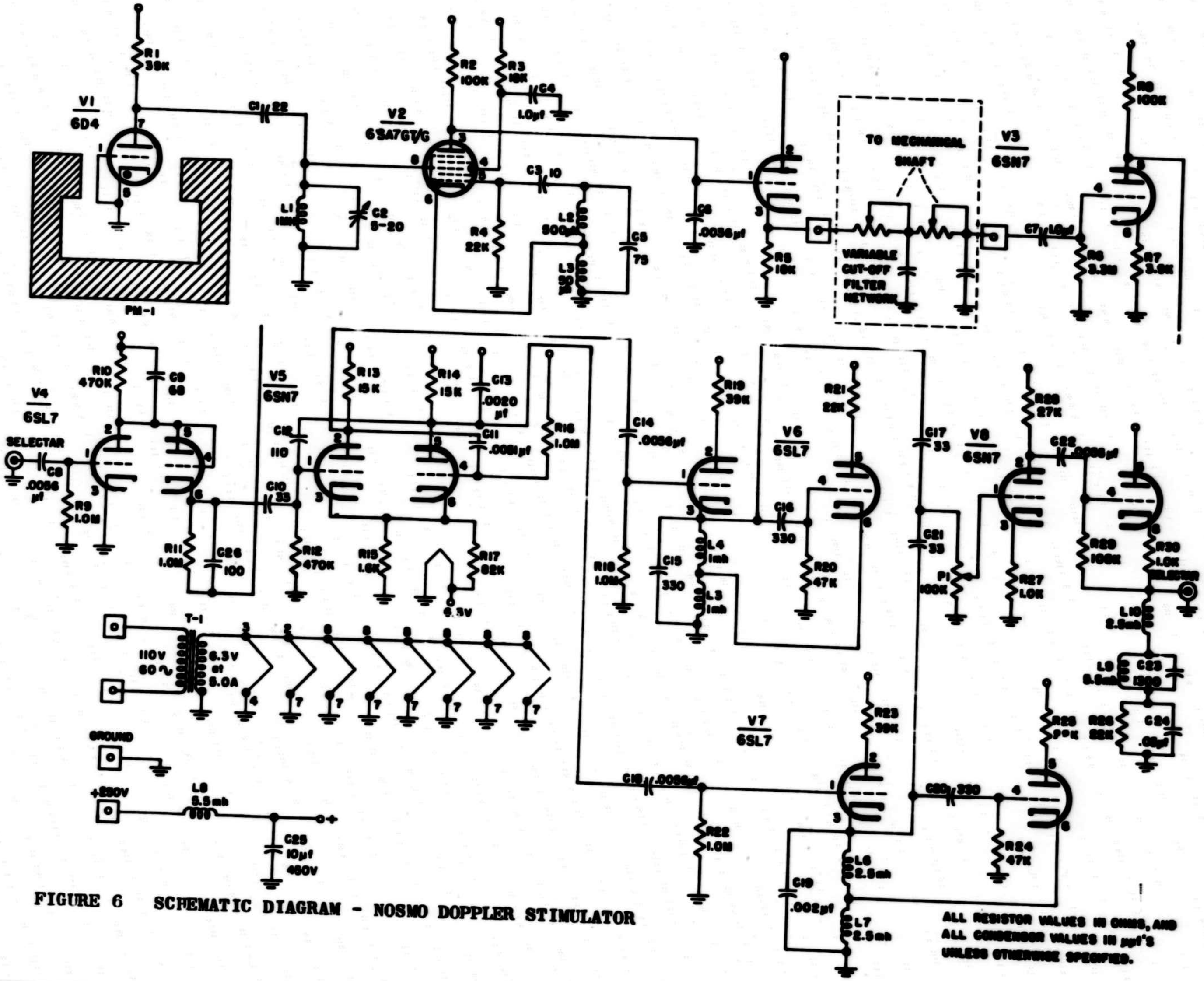

FIGURE 6 SCHEMATIC DIAGRAM - NOSMO DOPPLER STIMULATOR

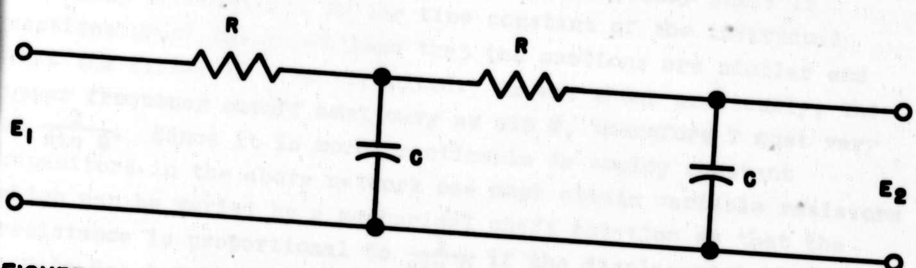

FIGURE 7 R-C FILTER NETWORK USED IN THE NOSMO DOPPLER SIMULATOR.

A plot of the absolute value of the transfer characteristic versus frequency is shown in Figure 8.

If one defines the upper frequency cutoff in the usual manner as that frequency for which the transfer characteristic equals .707, it is given by:

$$f_{c.o.} \cong \frac{.05}{T} \qquad (11)$$

Here it is seen that the upper frequency limit is inversely proportional to the time constant of the individual sections under the conditions that the sections are similar and that the filter is open circuited. As was shown previously, the upper frequency cutoff must vary as sin θ, therefore T must vary as $\frac{1}{\sin \theta}$. Since it is more practicable to employ constant capacitors in the above network one must obtain variable resistors which can be varied by a mechanical shaft rotation so that the resistance is proportional to $\frac{1}{\sin \theta}$ if the displacement shaft equals θ. A potentiometer approximating the desired curve is commercially feasible and at the present writing such potentiometers are being developed. In the actual installation, the potentiometers in the two respective sections are ganged and the mechanical shaft rotation varies the upper frequency cutoff in the desired manner to give the proper variation in filter bandpass with spinner angle.

The output from the low pass filter is amplified in the second half of the tube V3 (see Figure 6) after being a-c coupled through the network comprising C7 and R6. Since the time constant of these two elements is 3.3 seconds only the frequency components very close to zero c.p.s. are discriminated against. The output from this section of V3 consists of the amplified low frequency spectrum whose upper cutoff frequency is controlled by the angle the spinner makes with the ground track of the aircraft.

A negative trigger is brought into the simulator unit and is used to cut off the first half of V4. Since capacity is

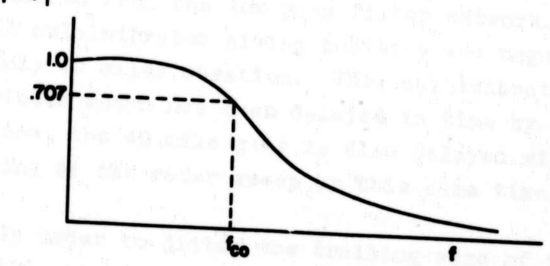

FIGURE 8 TRANSFER FUNCTION FOR FILTER USED IN THE NOSMO DOPPLER SIMULATOR

present in the plate circuit of this tube, an approximate sawtooth will be generated at the plate. It is applied to the plate of the second half of tube V4 which is connected as a diode. The previously obtained low frequency spectrum voltage is applied to the cathode through R11 and controls the instantaneous bias. The diode conducts only when the plate potential exceeds that of the cathode, hence, the sawtooth voltage appears across resistor R11 when the plate voltage exceeds the instantaneous potential of the noise spectrum. In this manner, the leading edge of the sawtooth appearing at the cathode of the diode is delayed in time proportional to the instantaneous value of the voltage obtained from the low pass filter network. Tube V5 is a biased multivibrator giving positive and negative gates of approximately 40 miles duration. This multivibrator is triggered by the sawtooth which has been delayed in time by the pickoff diode. Hence, the 40 mile gate is also delayed with respect to the beginning of the radar sweep by this same time.

In order to jitter the trailing edge of the 40 mile gate, a small amount of 60 cycle voltage is injected across the cathode resistor of tube V5. Since the trigger initiating the multivibrator action is large, the 60 cycle voltage applied to the cathode does not effect the start of the gate. However, since the application of the 60 cycles effectively varies the grid biases of tube V5, the time at which the multivibrator again assumes its stable state (trailing edge of gate) will be jittered at a 60 cycle rate. The amplitude of the jitter is determined by the amplitude of the 60 cycle voltage injected into the cathode circuit. In every other feature, the delay multivibrator is standard.

Tube V6 generates the 200 kilocycle pulsed oscillations in the following manner: In the stable condition, the first half of tube V6 is conducting so that current flows through the tank circuit of the Hartley oscillator comprising the second half of tube V6 and the associated elements. This flow of current through the tank circuit lowers the Q and prevents oscillation. The application of the 40 mile negative gate cuts off the first half of tube V6 and the Hartley oscillator is free to oscillate until the negative gate is completed. Tube V7 operates in an exactly similar manner with the exception that the circuit constants are designed to give 50 kilocycle oscillations. Here the positive 40 mile gate is impressed across the first section of tube V7 so that the oscillator is on in the steady state condition but is turned off by the application of the positive gate. It may seem strange that one oscillator is normally off while the other is normally on, inasmuch as both have essentially the same circuit constants. This occurs since the duty cycle of the gates is large, in the order of 70%. The average d-c potentials on the control grids (Pin 1) of both V6 and V7 must be zero while the instantaneous potentials before the gates arrive are determined by the instantaneous potentials of the gates with respect to their average values. The instantaneous potential of the negative gate waveform before the gate arrives is positive while that of the positive gate is negative. Therefore, the first half of V6 is conducting heavily before the arrival of the negative gate, while that of V7 is cut-off in the same period. This prevents V6 from oscillating while permitting V7 to oscillate. Capacitor C13 shapes the leading edge of the positive gate so that the 50 kilocycle oscillations damp out exponentially, decreasing to a negligible value at a range corresponding to about 8 miles.

The outputs from the two oscillators are added through the capacitor adding network, C17 and C21. A gain control potentiometer P1 is provided so that the output level can be controlled to provide a suitable proportion of modulation for different receivers that may be used. The first half of tube V8 amplifies the sum from the two pulsed oscillators and applies the amplified signal to the second half of tube V8 which serves as a cathode follower designed to drive a capacity of approximately 300 micromicrofarads.

The network used in the cathode circuit of the output cathode follower prevents serious frequency discrimination which otherwise would be encountered since two frequencies differing by a ratio of 4 to 1 are to be passed. Unless the cathode circuit of a cathode follower is resistive for the frequencies to be passed, a phase shift occurs between the cathode voltage and grid voltage. When this occurs, the cathode degeneration voltage and input signal are not 180° out of phase and if the applied signal is large the tube will be found to cut off during portions of the cycle and hence will create distortion in the output wave form. The network shown in the cathode circuit of the output cathode follower is designed to create a resistive cathode impedance at the two frequencies of 50 kilocycles and 200 kilocycles when the total capacity of the load is approximately 300 micromicrofarads. This latter figure includes the capacity of any cable used to connect the simulator unit to the receiver which is to be modulated. Since capacitor C24 is essentially a short circuit for the frequencies in question, the network which presents a resistive impedance at the two named frequencies is as shown in Figure 9.

Cathode Follower Load Network

Capacitor C1 represents the capacity of the load being driven plus the capacity of the cabling. The set of equations below are the defining relations which can be used to find the values of the elements for any arbitrary values of C1, ω_1 and ω_2, in order that the input impedance of the network be resistive at the two frequencies.

$$L_2 = \frac{\left(\omega_2^2 C_1 L_1 - 1\right)\left(1 - \omega_1^2 C_1 L_1\right)}{\left(\omega_1^2 \omega_2^2 C_1^2 L_1\right)} \qquad (12)$$

$$C_2 = \frac{C_1}{\left(\omega_2^2 C_1 L_1 - 1\right)\left(1 - \omega_1^2 C_1 L_1\right)} \qquad (13)$$

$$\omega_1^2 \omega_2^2 = \frac{1}{L_1 C_1 L_2 C_2} \qquad (14)$$

By employing this cathode network, distortion and

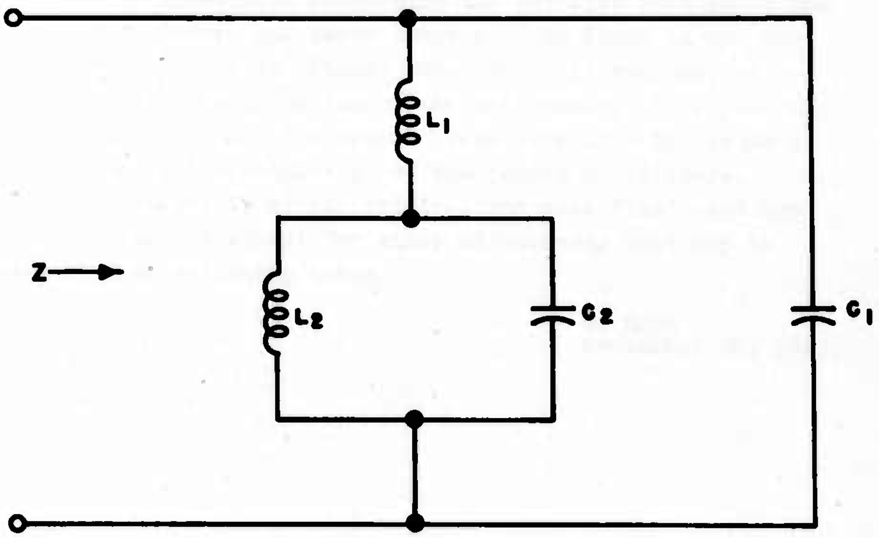

FIGURE 9 CATHODE FOLLOWER LOAD NETWORK
THE NOSMO DOPPLER SIMULATOR

and discrimination of the two frequency components present is prevented and large voltage output is obtained across the load capacitor without overdriving the cathode follower. For the AN/APQ-T1 radar trainer, the output from the Doppler simulator unit is applied to the screen of one of the pre-amplifier stages to modulate the gain of the receiver. Since the screen has an R-F bypass capacitor, the cathode follower load impedance network described above is required.

Pertinent data concerning the voltages throughout the circuit, wave forms, and power input will be found in the included wave form charts (Figure 10). The unit requires no adjustment with the exception of the gain control to adjust the percent modulation and the trimmer capacitor C2 which adjusts the amount of phase modulation of the pulsed oscillators. Neither adjustment is at all critical and once fixed need not be touched further except for minor adjustments that may be necessary when replacing tubes.

W. Roth
September 28, 1945

	TUBE NO. V1	TUBE NO. V2	TUBE NO. V3-1	TUBE NO. V3-2	TUBE NO. V4-1	TUBE NO. V4-1	TUBE NO. V5-1
GRID — DC VOLTAGE / WAVE SHAPE POS PEAK, NEG PEAK	DC 0 t=0 PIN NO. 1	DC 0 +4ᵛ / -4ᵛ t=0 NOISE PIN NO. 8	t=0 SAME AS PIN 3 OF V2 PIN NO. 1	DC 0 0 t=0 PIN NO. 4	DC -6 +6 / -100 t=0 1000µs PIN NO. 1	t=0 SAME AS PIN 2 V4-1 PIN NO. 4	DC 0 +25ᵛ / -50ᵛ t=0 450 µs PIN NO. 1
PLATE — DC VOLTAGE / WAVE SHAPE POS PEAK, NEG PEAK	DC +15ᵛ +8ᵛ / -8ᵛ t=0 NOISE PIN NO. 7	DC 50-100 +15ᵛ / -15ᵛ t=0 NOISE PIN NO. 3	DC +250 t=0 PIN NO. 2	DC 120 0 t=0 PIN NO. 5	DC 65ᵛ +80ᵛ / -17ᵛ t=0 PIN NO. 2	t=0 SAME AS PIN 2 V4-1 PIN NO. 5	DC 210ᵛ +35ᵛ / -90ᵛ t=0 PIN NO. 2
CATHODE — DC VOLTAGE / WAVE SHAPE POS PEAK, NEG PEAK	DC 0 t=0 PIN NO. 5	DC 0 +10ᵛ / -10ᵛ t=0 730-770KC PIN NO. 6	DC 50-100 t=0 SAME AS PIN 3 OF V2 PIN NO. 3	DC 5ᵛ 0 t=0 PIN NO. 6	DC 0 t=0 PIN NO. 3	DC 115ᵛ +40ᵛ / -8ᵛ t=0 400 µs PIN NO. 6	DC 15ᵛ +28ᵛ / -8ᵛ t=0 PIN NO. 3
SCREEN	t=0	DC 75 t=0 PIN NO. 4	t=0	t=0	t=0	t=0	t=0
REMARKS	RANDOM NOISE OBSERVED ON PLATE (60~ SWEEP)	G2 TUNED FOR MAX SIGNAL ON PIN 3 OF V3-1 LO FREQUENCY TO BE 730-770 KC MEASURED WITH RECEIVER (LOOSE COUPLING)	60~ SWEEP	WITH CLIP LEAD FROM PIN 4 TO GROUND (ZERO SIGNAL INPUT)	TRIG INPUT TO V4-1 100ᵛ 10 µsec		

	TUBE NO. V5-2	TUBE NO. V6-1	TUBE NO. V6-2	TUBE NO. V7-1	TUBE NO. V7-2	TUBE NO. V8-1	TUBE NO. V8-2
GRID — DC VOLTAGE / WAVE SHAPE POS PEAK, NEG PEAK	DC -7ᵛ +45ᵛ / -100ᵛ t=0 PIN NO. 4	DC -30ᵛ t=0 SAME AS PIN 2 V5-1 PIN NO. 1	DC -2ᵛ +30ᵛ / -30ᵛ t=0 PIN NO. 4	DC -80ᵛ t=0 SAME AS PIN 5 V5-2 PIN NO. 1	DC -2ᵛ +20ᵛ / -20ᵛ t=0 PIN NO. 4	DC 0 +4ᵛ / -4ᵛ t=0 PIN NO. 1	DC 85ᵛ t=0 SAME AS PIN 2 V8-1 PIN NO. 4
PLATE — DC VOLTAGE / WAVE SHAPE POS PEAK, NEG PEAK	DC 150ᵛ +85ᵛ / -70ᵛ t=0 PIN NO. 5	DC +185ᵛ +50ᵛ / -55ᵛ t=0 PIN NO. 2	DC +180ᵛ +50ᵛ / -60ᵛ t=0 PIN NO. 5	DC +200ᵛ +40ᵛ / -100ᵛ t=0 PIN NO. 2	DC +200ᵛ +40ᵛ / -75ᵛ t=0 PIN NO. 5	DC 130ᵛ +60ᵛ / -60ᵛ t=0 PIN NO. 2	DC 250ᵛ t=0 PIN NO. 5
CATHODE — DC VOLTAGE / WAVE SHAPE POS PEAK, NEG PEAK	t=0 SAME AS PIN 3 OF V5-1 PIN NO. 6	DC 0 +30ᵛ / -30ᵛ t=0 PIN NO. 3	DC 0 +20ᵛ / -20ᵛ t=0 PIN NO. 6	DC 0 +20ᵛ / -20ᵛ t=0 PIN NO. 3	DC 0 +10ᵛ / -10ᵛ t=0 PIN NO. 6	DC 45ᵛ +2ᵛ / -2ᵛ t=0 PIN NO. 3	DC 105ᵛ t=0 SAME AS PIN 2 V8-1 PIN NO. 6
	t=0	t=0	t=0	t=0	t=0	t=0	t=0
REMARKS		OSCILLATION = 200 KC		OSCILLATION = 54 KC		GAIN POT ADJUSTED FOR 80ᵛ PEAK-PEAK OUTPUT AT 20 MILES RANGE	USE OUTPUT LOAD OF 300 µµf TO GROUND IF NOT CONNECTED TO PRE-AMP

D.C VOLTAGE MEASURED WITH 20,000 Ω/V METER
WAVE SHAPE DETERMINED WITH DUMONT TYPE 208 + MODEL G SYNCHRONIZER
WAVE SHAPE VOLTAGE MEASURED WITH DUMONT TYPE 208
WAVE SHAPE WIDTHS MEASURED IN MICROSECONDS WITH MODEL G SYNCHRONIZER
REP RATE OF TRIGGERS = 1000 CPS

POWER REQUIREMENTS	
VOLTAGE	CURRENT (AMP)
+250 DC?	92 MA?
[illegible]	
[illegible]	[illegible]

[illegible]

FIGURE 10 TEST DATA ON TWINKLE SIM. UNIT # 1,2,3,

REEL – C

487

A.T.I.

13732

TITLE: Nosmo Doppler Simulator	ATI- 13732
	REVISION (None)
AUTHOR(S): Roth, W.	ORIG. AGENCY NO. R-857
ORIGINATING AGENCY: Massachusetts Institute of Technology, Cambridge, Mass.	
PUBLISHED BY: Office of Scientific Research and Development, NDRC, Washington, D. C.	PUBLISHING AGENCY NO. (None)

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Sept '45	Unclass.	U.S.	Eng.	26	diagrs, graphs

ABSTRACT:

A unit to simulate Doppler beating phenomena is described for use in air-borne radar training program. Unit amplitude modulates returned signals with a random frequency spectrum whose upper limit varies as the sine of the angle between the azimuth of the radiated beam and ground track of the aircraft. Practice in determination of drift angle is herewith made possible.

DISTRIBUTION: Copies of this report obtainable from Air Documents Division; Attn: MCIDXD

DIVISION: Electronics (3)	SUBJECT HEADINGS: Doppler effect, Radio and radar (30940);
SECTION: Radar (2)	Simulators, Flight (86730)
ATI SHEET NO.: R-3-2-51	

TITLE: Nosmo Doppler Simulator	ATI- 24454 (25)
AUTHOR(S): Roth, W.	REVISION (None)
ORIGINATING AGENCY: Massachusetts Inst. of Technology, Radiation Lab., Cambridge	ORIG. AGENCY NO. R-857
PUBLISHED BY: Office of Scientific Research and Development, NDRC, Div 14	PUBLISHING AGENCY NO. (None)

DATE	DOC. CLASS.	COUNTRY	LANGUAGE	PAGES	ILLUSTRATIONS
Sept '45	Unclass.	U.S.	Eng.	26	diagrs

ABSTRACT: 23) * Radar P 17/9

A unit to simulate Doppler beating phenomena for use in the airborne radar training program is discussed. The unit amplitude modulates returned signals with a random frequency spectrum whose upper frequency limit varies as the sine of the angle between the azimuth of the radiated beam and ground track of the aircraft. Practice in the determination of drift angle is therefore made possible. The noise source employed is a miniature gas thyrathron with a permanent magnet surrounding the tube to suppress inherent oscillations within the gas plasma. Pertinent data concerning the voltages throughout the circuit, wave forms, and power input are included in wave form charts.

DISTRIBUTION: Copies of this report obtainable from Air Documents Division; Attn: MCIDXD

DIVISION: Electronics (~~8~~) 6	SUBJECT HEADINGS:
SECTION: Radar (~~2~~) 6	Training, Radar - Devices (95105.3)
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