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TECHNICAL REPORT ECOM-01711-3
**AUTOMATIC DETECTION AND DISPLAY EQUIPMENT
FOR
MOVING TARGET INDICATION
(ADDEM)**

THIRD QUARTERLY PROGRESS REPORT

BY

L.L. SCHOENFELD

NOVEMBER 1966

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CONTRACT DA 28-043 AMC-01711(E)
GOODYEAR AEROSPACE CORPORATION
ARIZONA DIVISION
LITCHFIELD PARK, ARIZONA

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TECHNICAL REPORT ECOM 01711-3

NOVEMBER 1966

AUTOMATIC DETECTION AND DISPLAY EQUIPMENT

FOR

MOVING TARGET INDICATION

(ADDEM)

THIRD QUARTERLY PROGRESS REPORT

1 MAY 1966 THROUGH 31 JULY 1966

REPORT NO. 3

CONTRACT NO. DA 28-043 AMC-01711(E)

DA TASK NR. 1P6-20901-A-188-03

PREPARED BY

L. L. SCHOENFELD

**GOODYEAR AEROSPACE CORPORATION
ARIZONA DIVISION
LITCHFIELD PARK, ARIZONA**

FOR

**U.S. ARMY ELECTRONICS COMMAND
FORT MONMOUTH, NEW JERSEY 07703**

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SECTION I - PURPOSE

The purpose of this program is to develop, fabricate, and test the Automatic Detection and Display Equipment for MTI. This equipment will be used in conjunction with battlefield surveillance radars such as the AN/TPS-25, AN/TPS-33, and AN/PPS-5 to detect moving targets and display them on a range versus azimuth display.

The more pertinent design requirements are:

1. Ten range-gated channels, each containing ten bandpass filters, detectors, integrators, threshold circuits, and indicators. The range gates will be either 50 or 100 meters in width and the range-gated interval will be adjustable in range between 0 and 9500 meters.
2. A range versus azimuth display to indicate the doppler frequency and range of a detected moving target. A total of 100 indicators will monitor 10 frequency channels in each of the 10 range gates
3. A range versus azimuth indicator to present the operator with a B-scope display. This indicator will have a persistence variable between 1 second and 30 minutes and will present only moving targets
4. A volume of less than 1 cubic foot, weight less than 25 pounds, and consumption of less than 25 watts of power
5. Operation under rigorous field conditions.

More detailed specifications can be obtained from Electronics Command Technical Requirement SCL-8029.

This program is being conducted under Contract Number DA28-043-AMC-01711(E) ending 30 April 1967, with the delivery of one fully tested detection and display unit.

B. G. Wood contributed the section on active filters, T. A. Hammond collected much of the experimental data, and C. C. Knight contributed the section on packaging.

SECTION II - ABSTRACT

This is the third quarterly report documenting the development of the Automatic Detection and Display Equipment for MTI.

The circuit design effort was approximately 80 percent completed at the end of this reporting period. The active filter was analyzed to minimize the effect of component tolerance and variation. A range-delay and range-gating circuit was also developed. The revised cathode-ray tube (crt) power supply, which includes provisions for dynamic focus, was completed and is ready for test.

Packaging studies and power estimates indicate that size and power consumption will be better than the specified values.

SECTION II - FACTUAL DATA

1. INTRODUCTION

This report describes the work accomplished, during the third quarter, toward development of the Automatic Detection and Display Equipment for MTI.

During this period, work continued on the final design of all circuits. At the end of the reporting period, approximately 80 percent of the circuit design has been completed. The value of the extended effort of circuit design is illustrated in the power estimate of 16.8 watts for the operating unit. Packaging studies were also conducted and a mockup constructed. It appears that the volume of unit will be approximately 0.75 cubic foot.

2. ACTIVE FILTERS

Since a configuration of the active pole had been previously selected which was capable of producing the high Q required, an extensive analysis of the effects of environmental changes upon each component in the circuit (Figure 1) was undertaken.

To meet the desired objectives of low power consumption, small size, and light weight, each component had to be selected with extreme care, weighing each of these factors against the required performance.

The largest components in each pole are the capacitors in the parallel T ; therefore, reducing their size is a method of reducing the over-all size of each filter channel. On the other hand, as the size of these capacitors is reduced, the impedance of the feedback network increases, thereby introducing instability in the amplifier. It was determined experimentally that the amplifier was stable with more than 300,000 ohms in the d-c feedback network. Therefore, this was used as the upper limit in the parallel T .

A computer program has been written which generates eight place values of capacitance, if the resistors are chosen, and one which generates eight place values of resistance for a given capacitor size. This makes it possible to generate a set of filters of any desired impedance. Another program was developed which analyzed these filters when each component was changed by any desired percent. This made it possible to verify the experimental data taken from environmentally testing each filter.

Figures 2 through 10 illustrate one of these programs. Figure 2 is the response of a single pole with all component values accurate to five significant figures. Figures 3 through 10 show the computed response with a selected component changed by ± 10 percent. These curves, therefore, indicate the effect on the circuit with any component in error or changing with environment. As may be seen, the shunting elements C_E and R_H have the greatest effect upon both amplitude and frequency response. Changing R_C or R_D and C_F or C_G also have a serious effect upon the response. It was determined, however, that some of these effects can be made to cancel one another. Figures 11 and 12 illustrate this result on a two-pole filter both from environmental data and the computer. Figure 11 shows the computed response of the circuit, when C_F , C_G , C_E , C_V , C_W , and

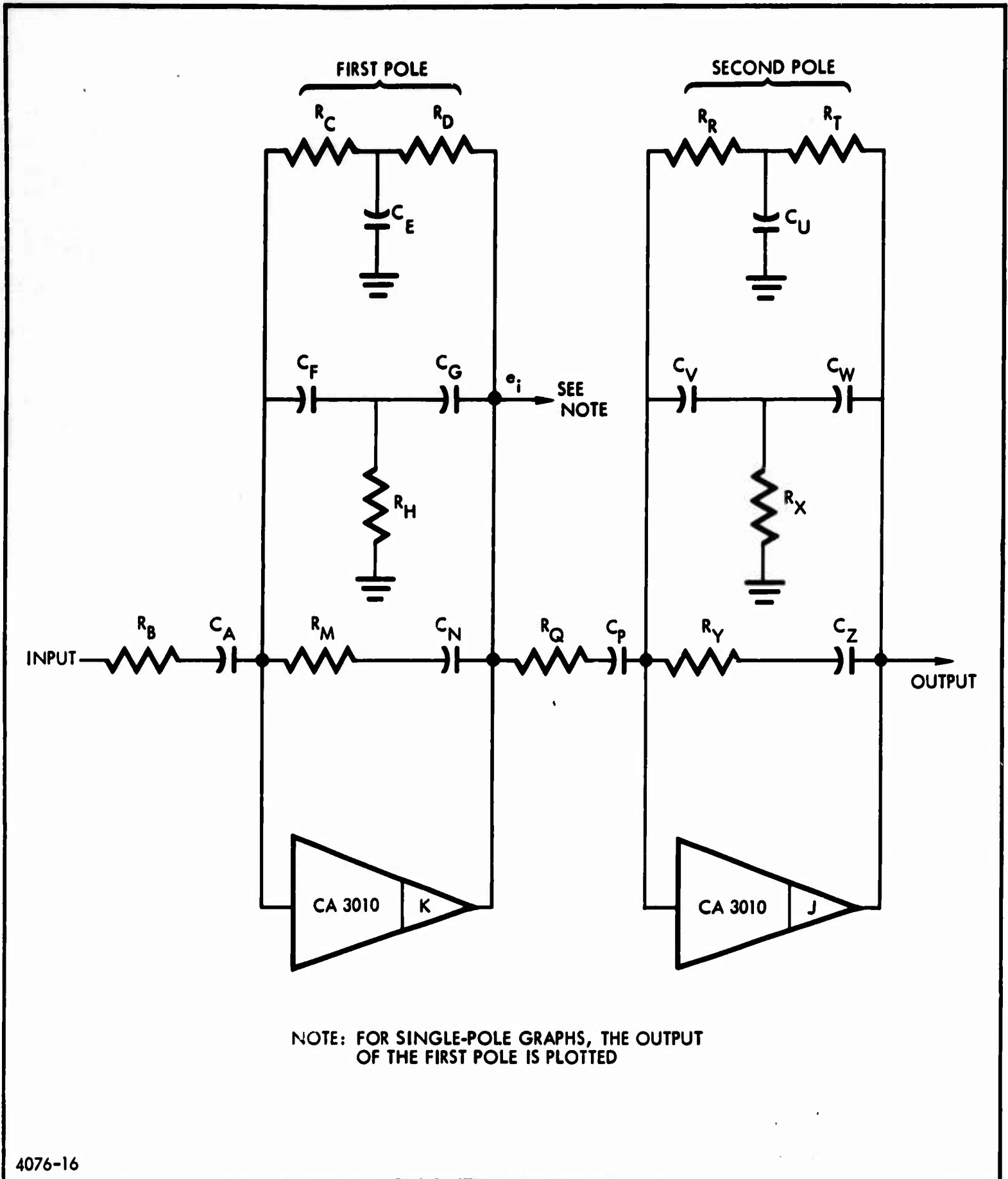


Figure 1 - Two-Pole Active Filter

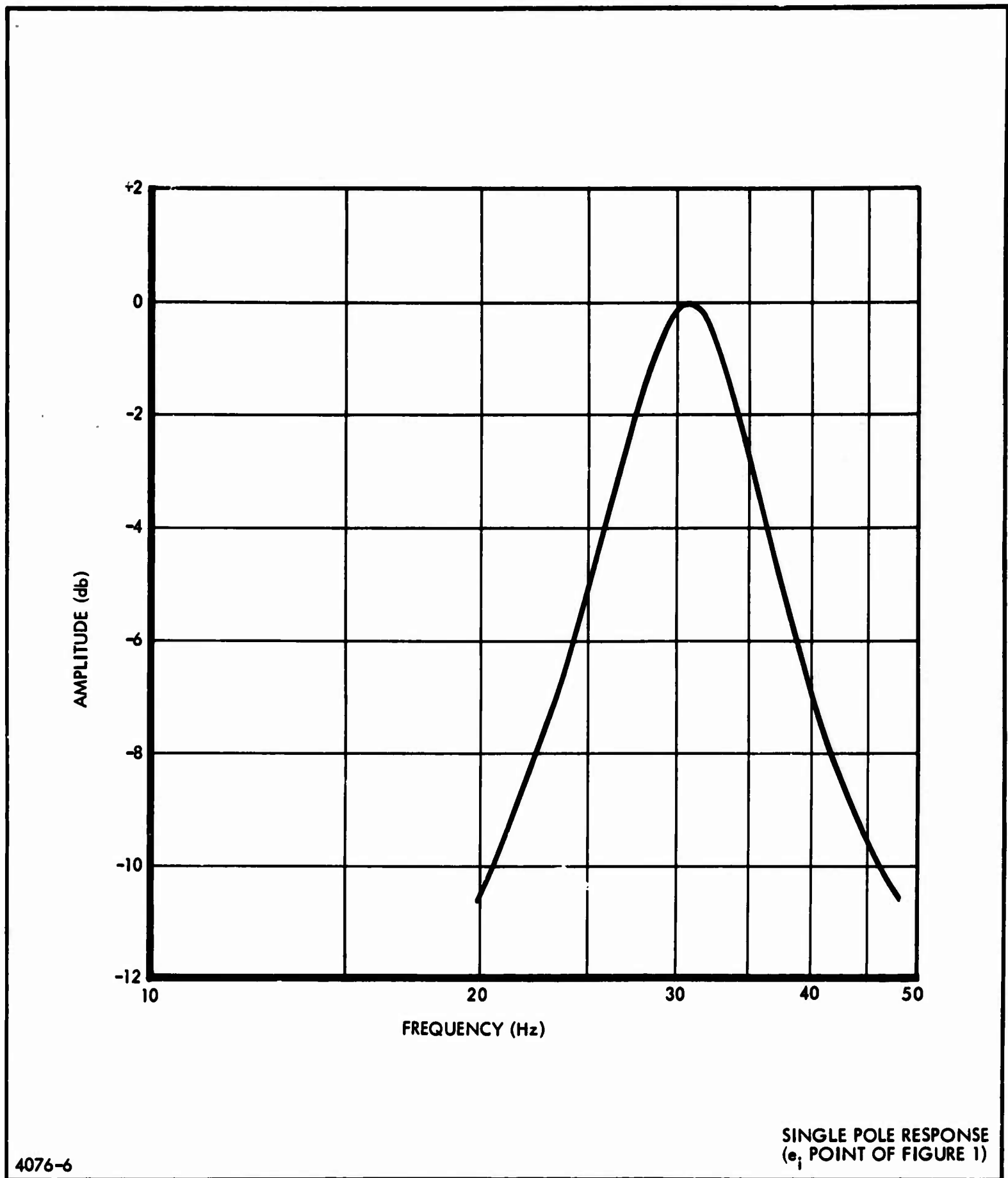
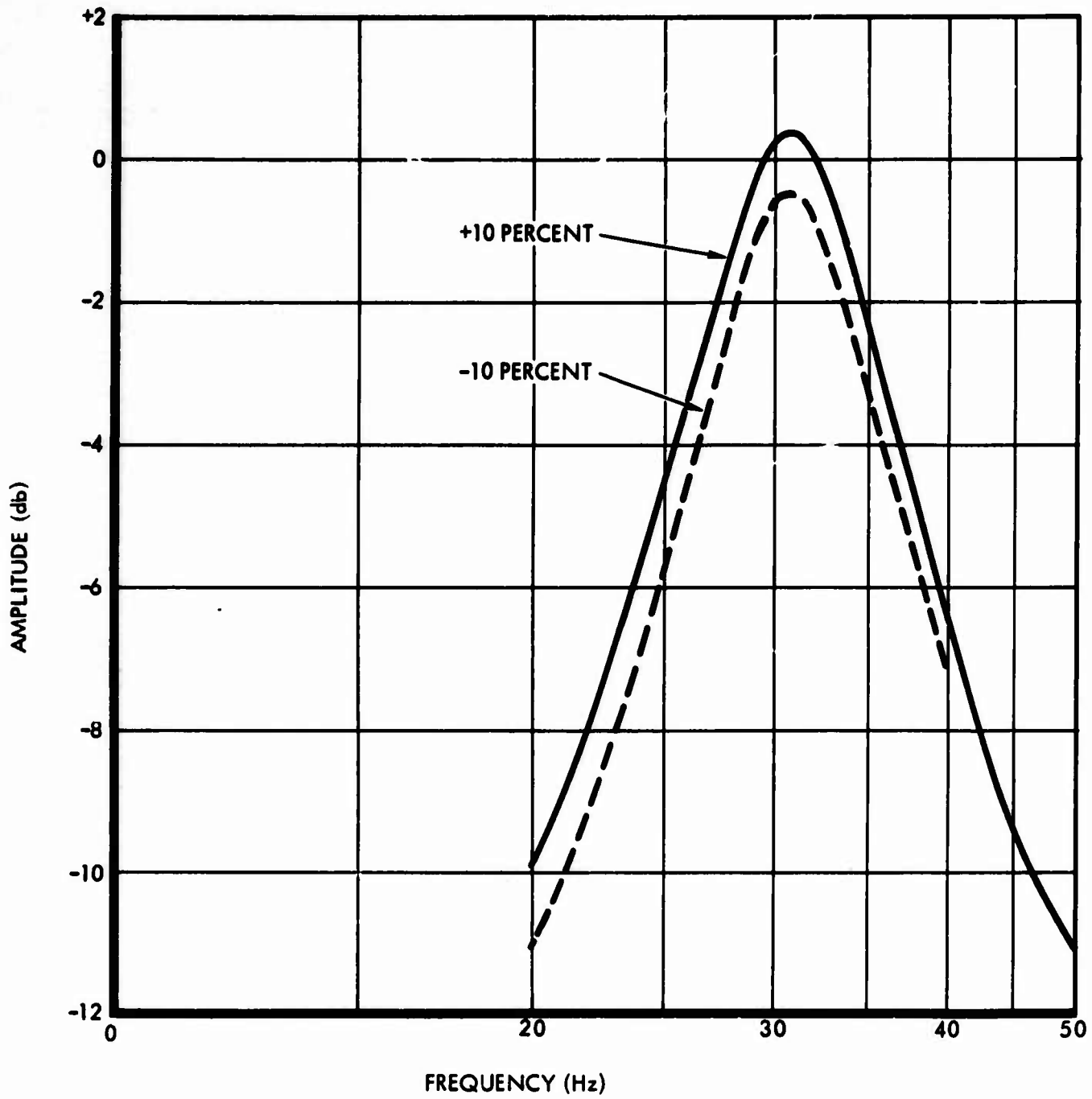


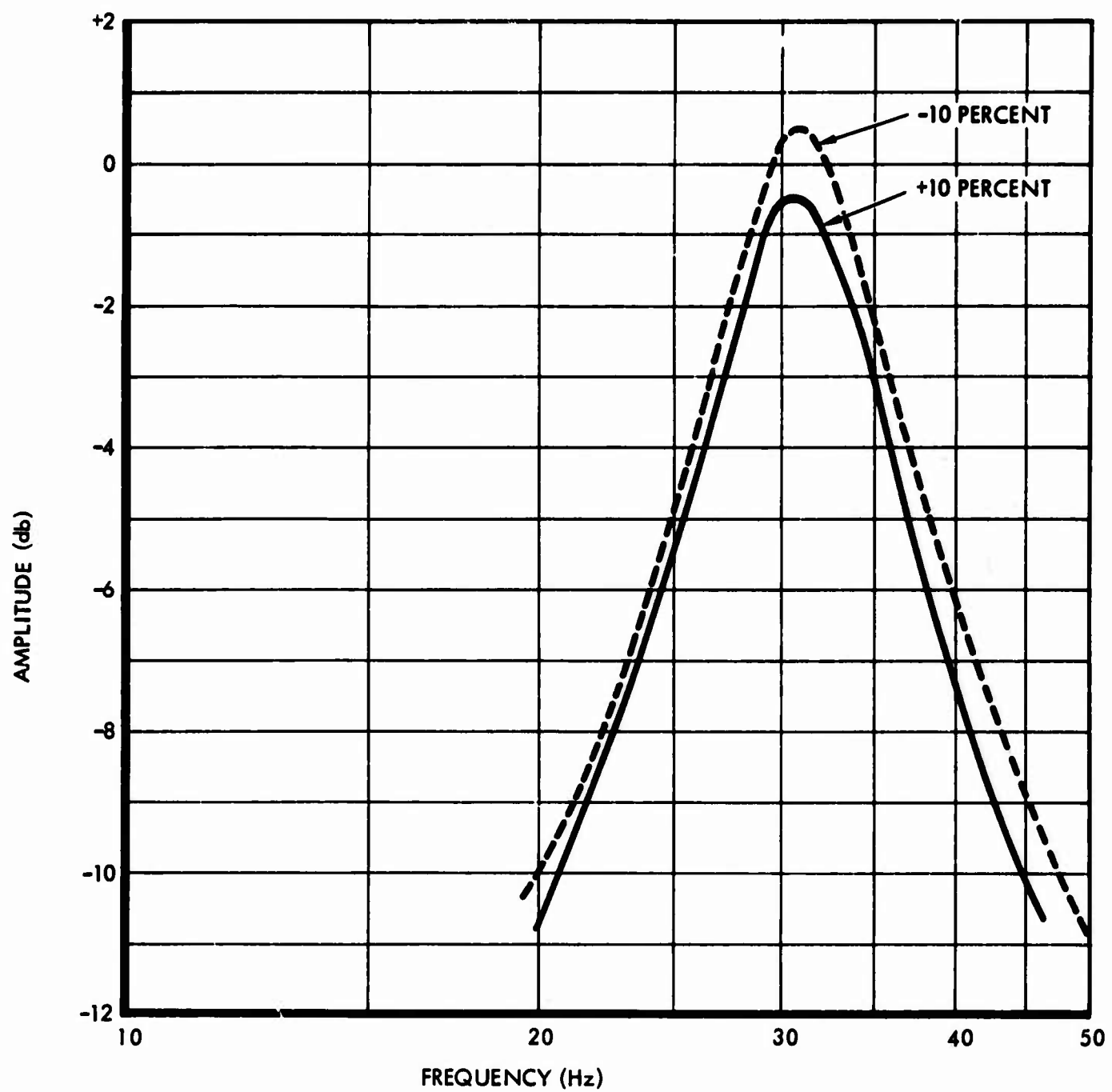
Figure 2 - Response of the Circuit with All Values Accurate to Five Significant Figures



SINGLE POLE RESPONSE
(e; POINT OF FIGURE 1)

4076-7

Figure 3 - Varying $C_A \pm 10$ Percent



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SINGLE POLE RESPONSE
(e_i POINT OF FIGURE 1)Figure 4 - Varying $R_B \pm 10$ Percent

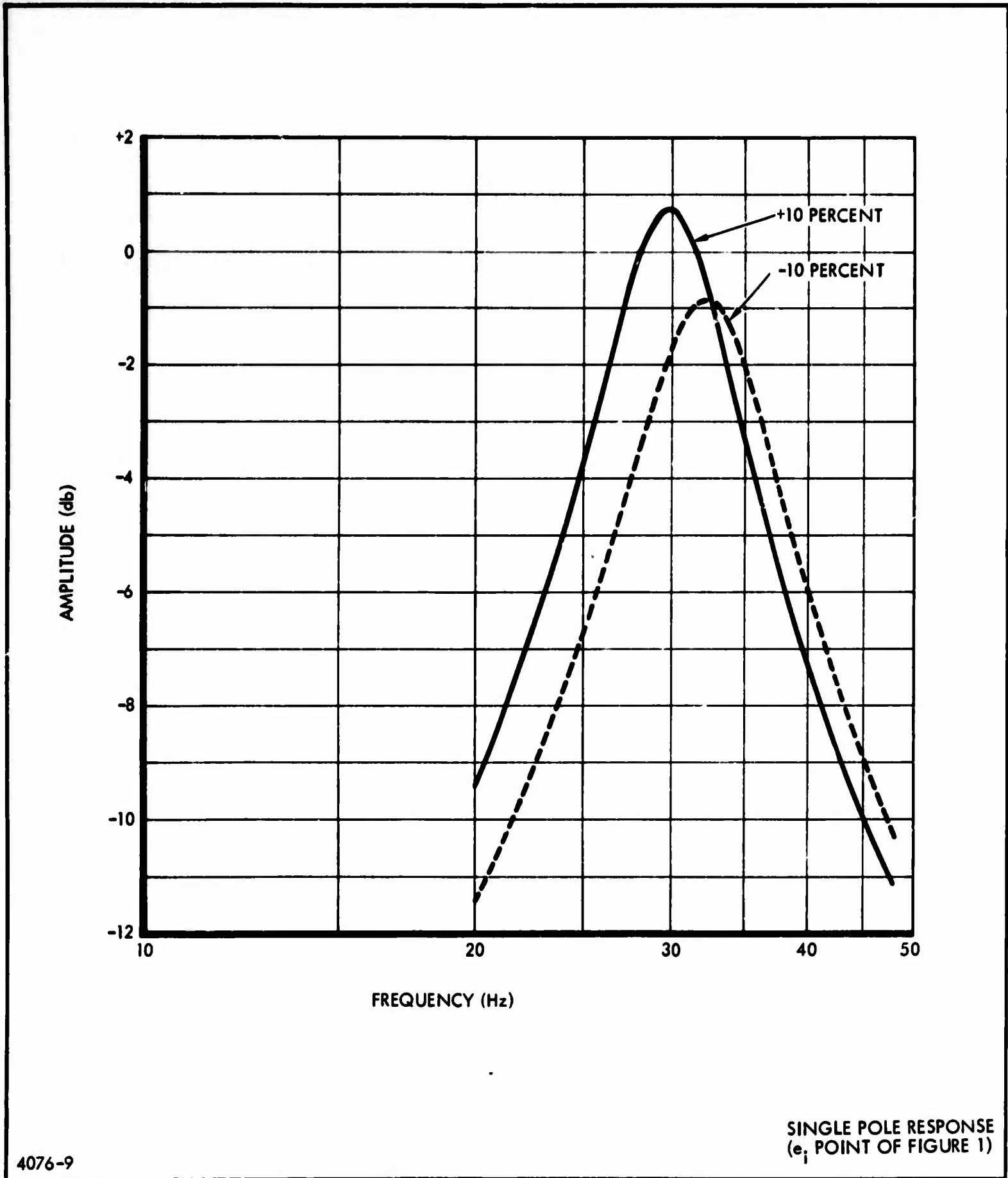
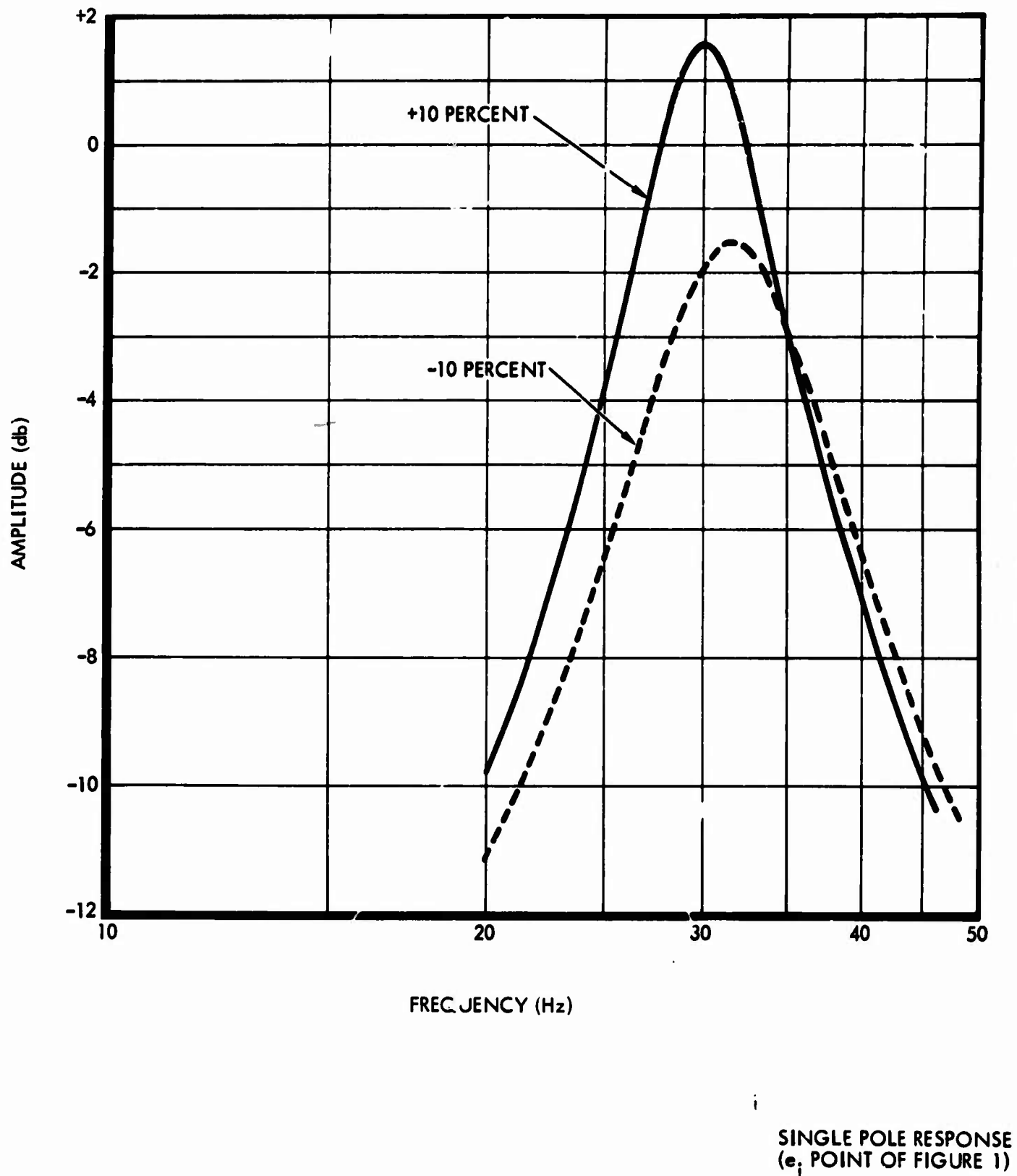
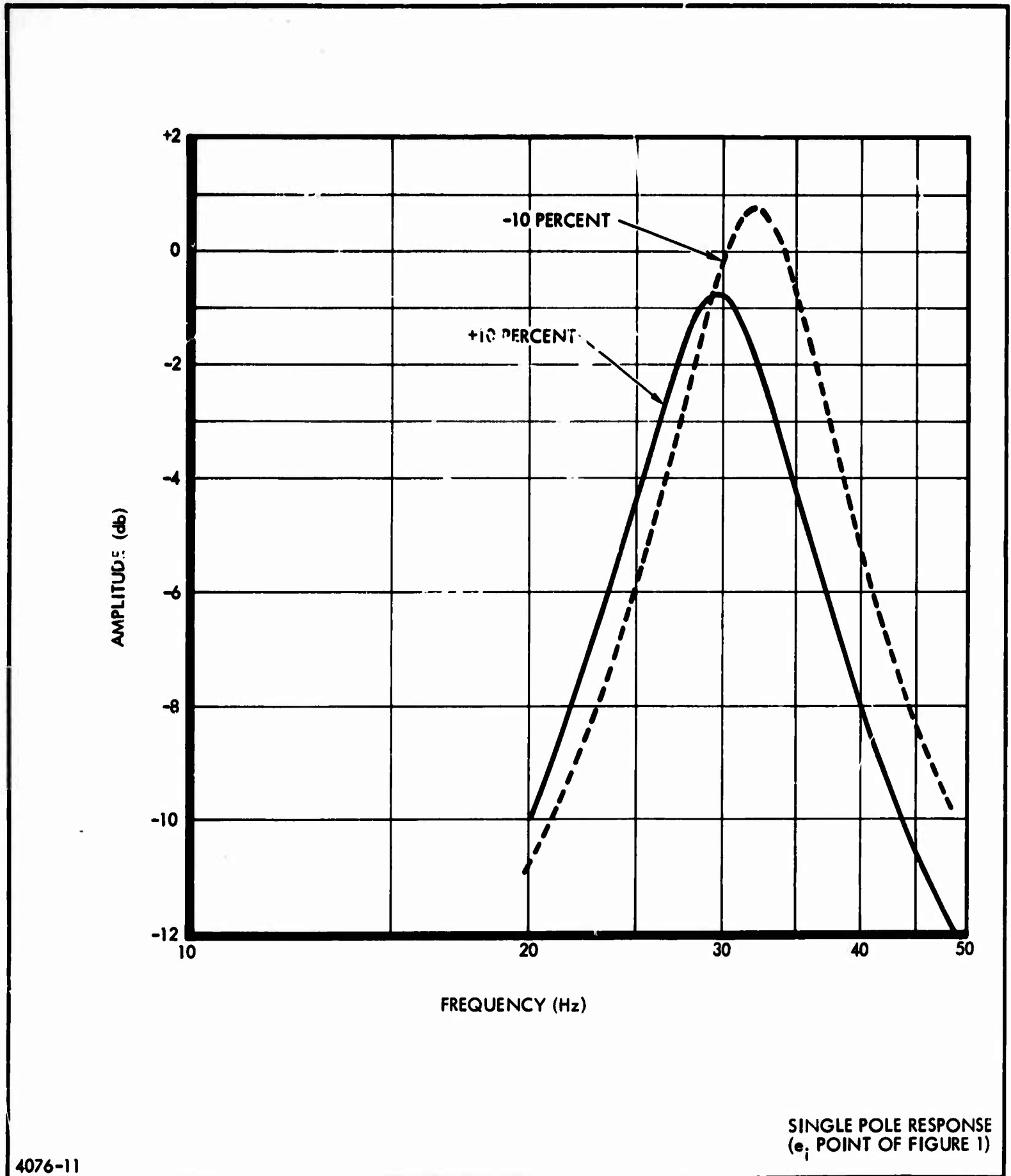


Figure 5 - Varying R_C or $R_D \pm 10$ Percent



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Figure 6 - Varying $C_E \pm 10$ Percent

Figure 7 - Varying C_F or $C_G \pm 10$ Percent

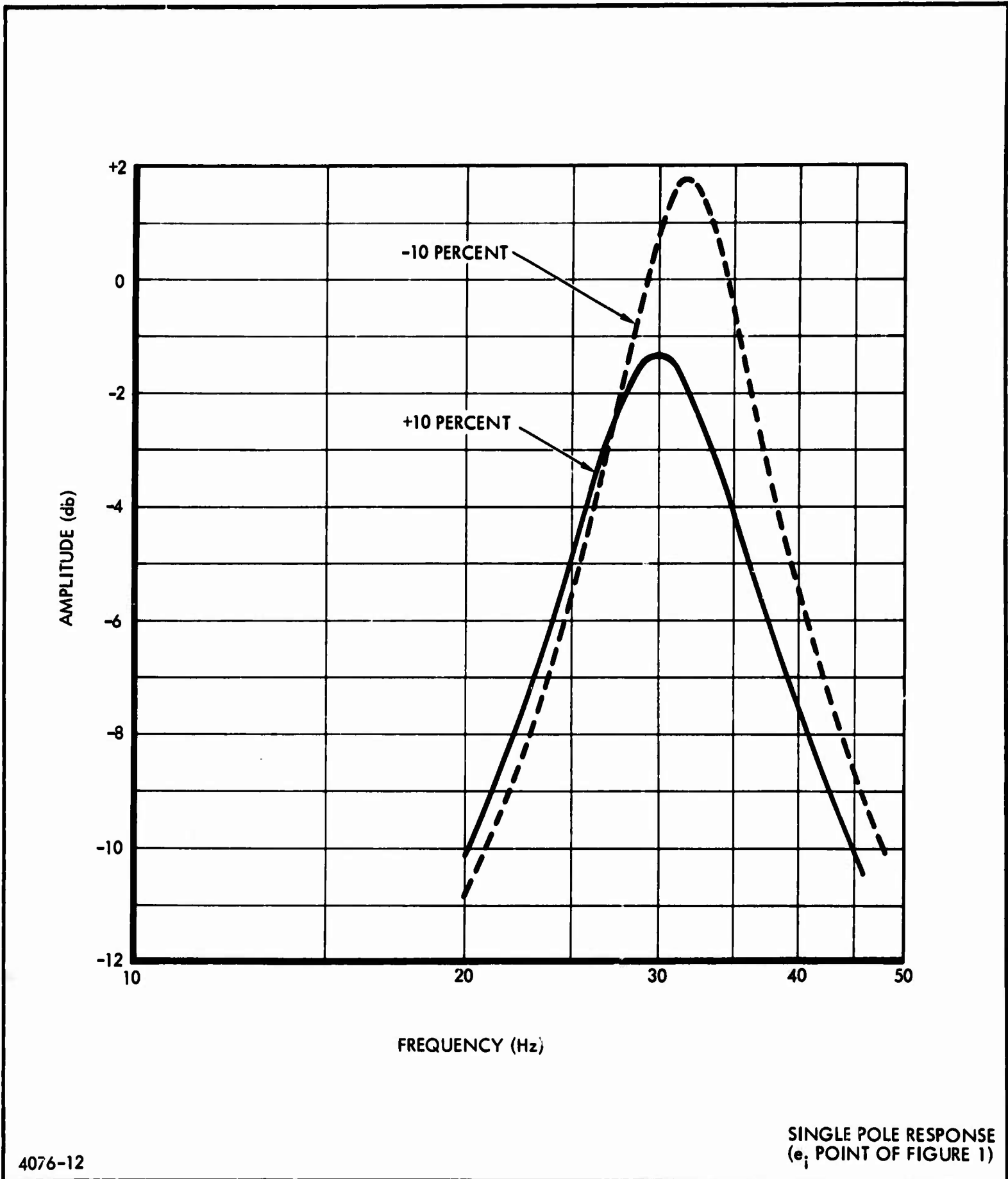


Figure 8 - Varying $R_H \pm 10$ Percent

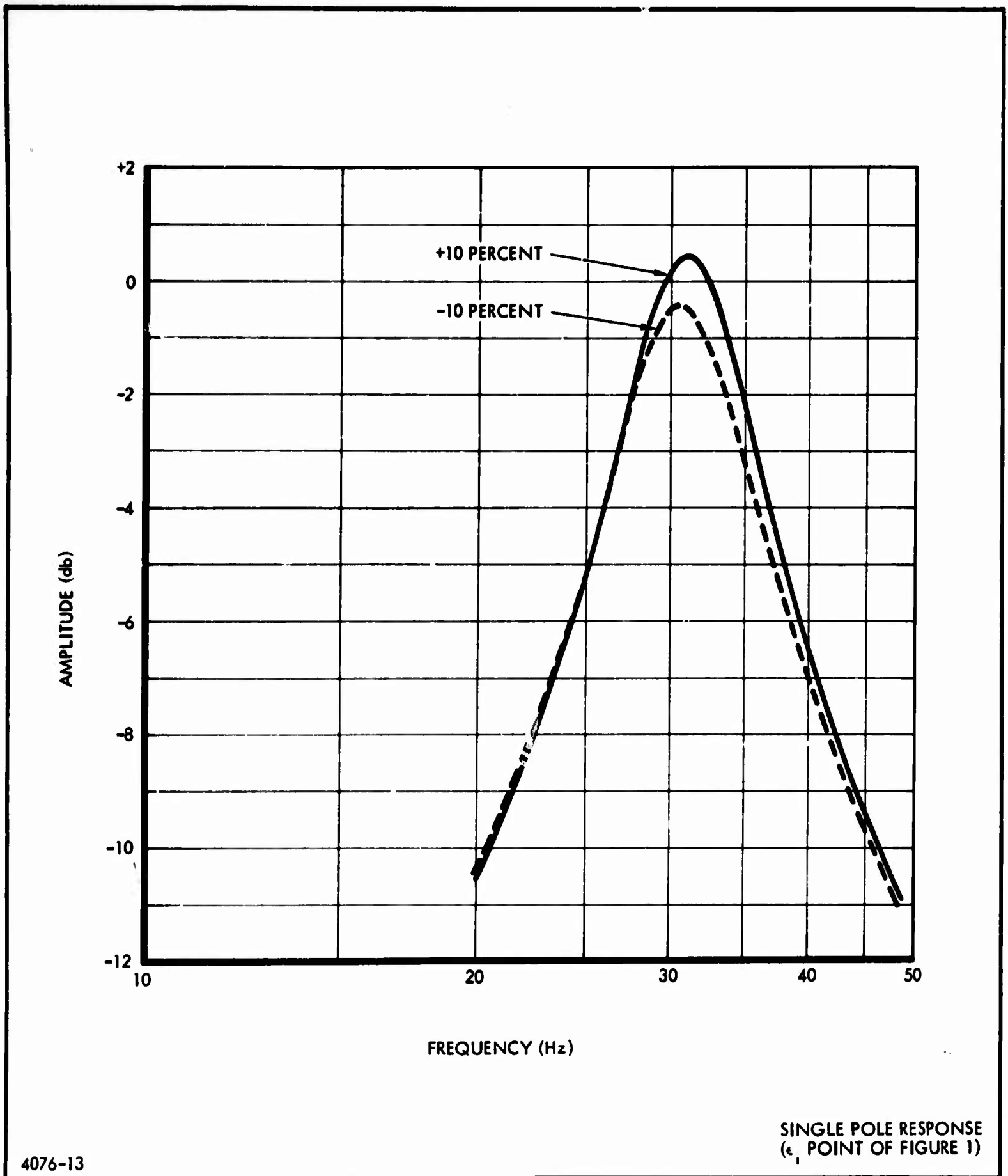
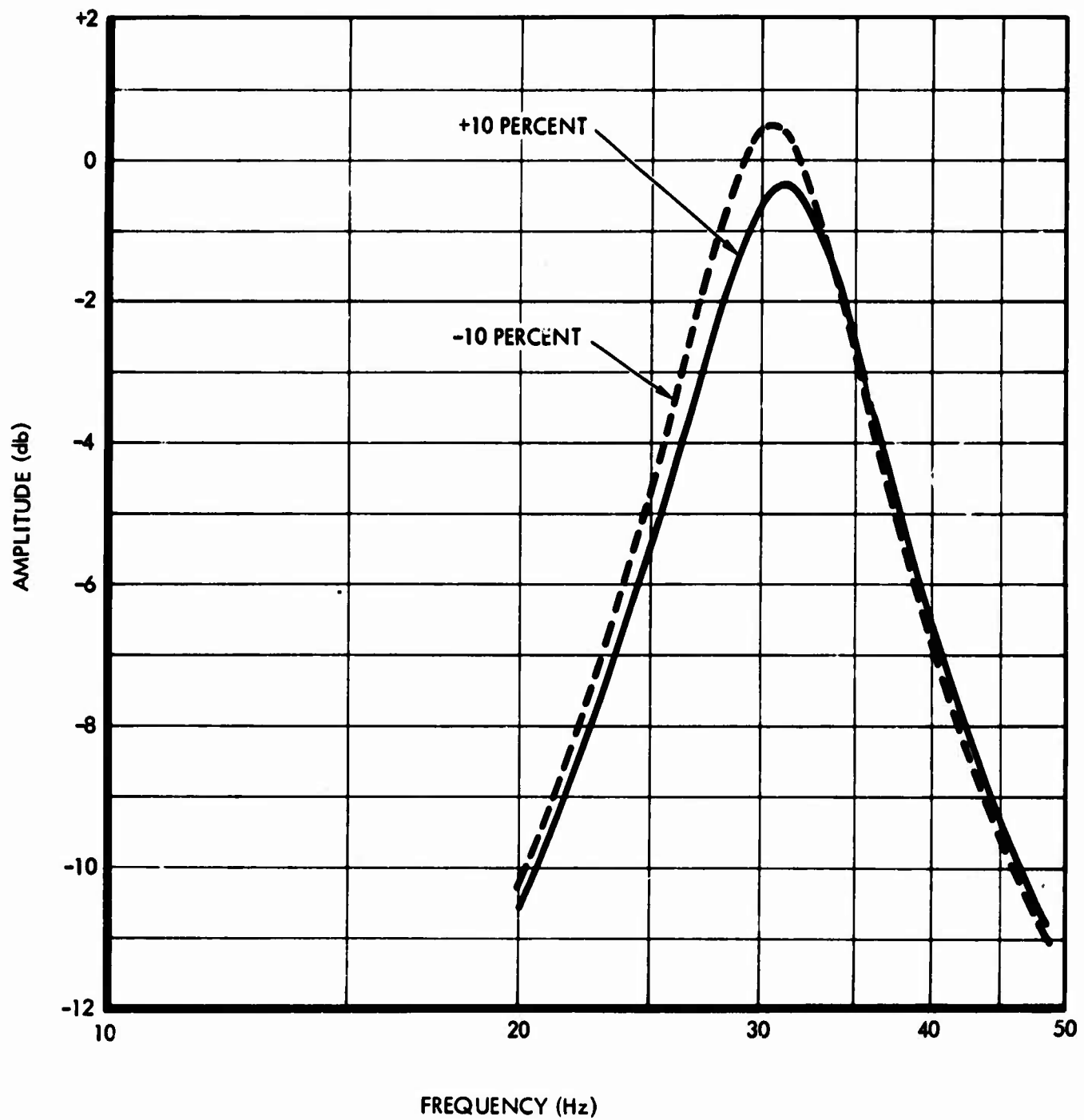


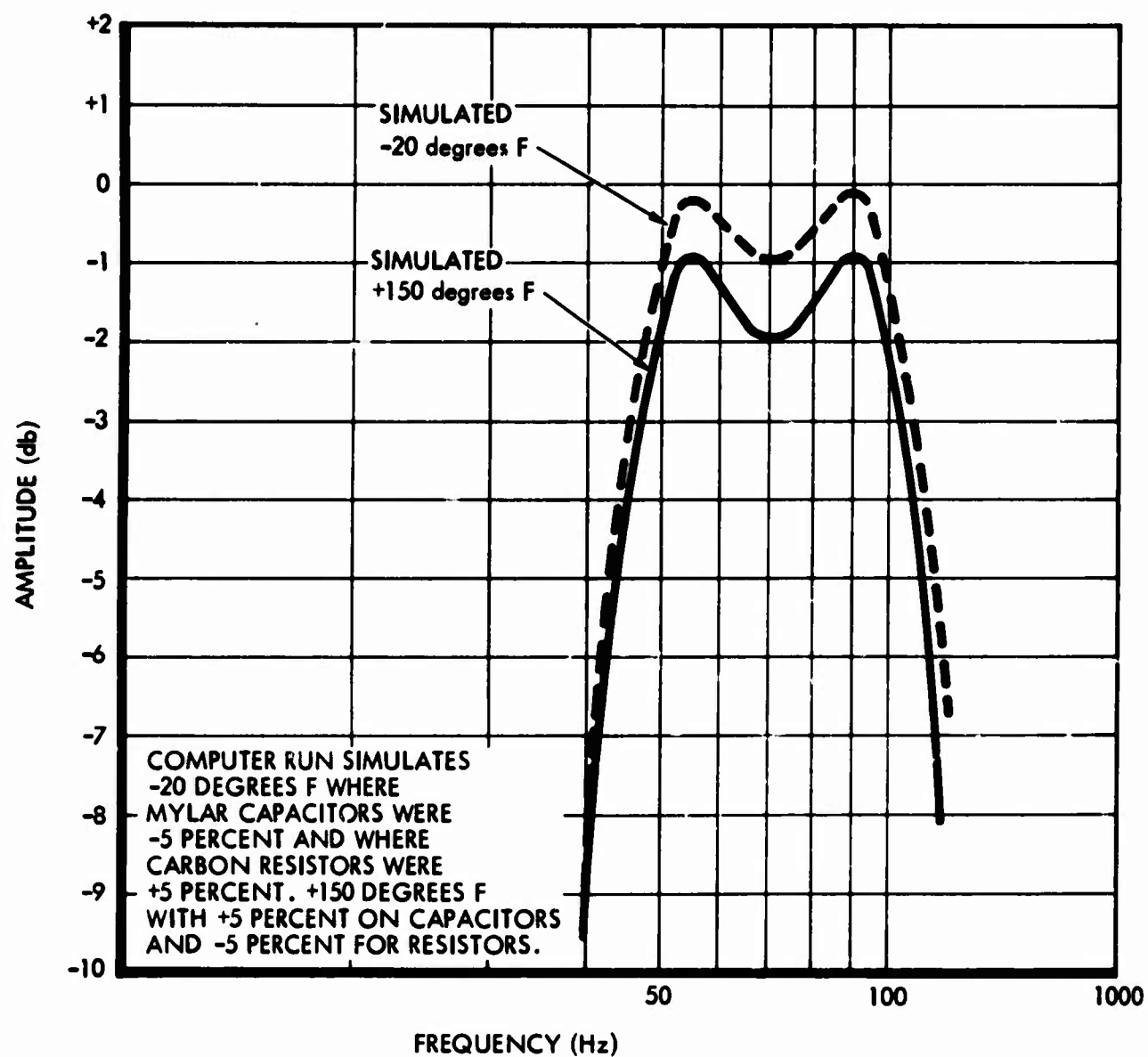
Figure 9 - Varying $R_M \pm 10$ Percent



SINGLE POLE RESPONSE
(e_i POINT OF FIGURE 1)

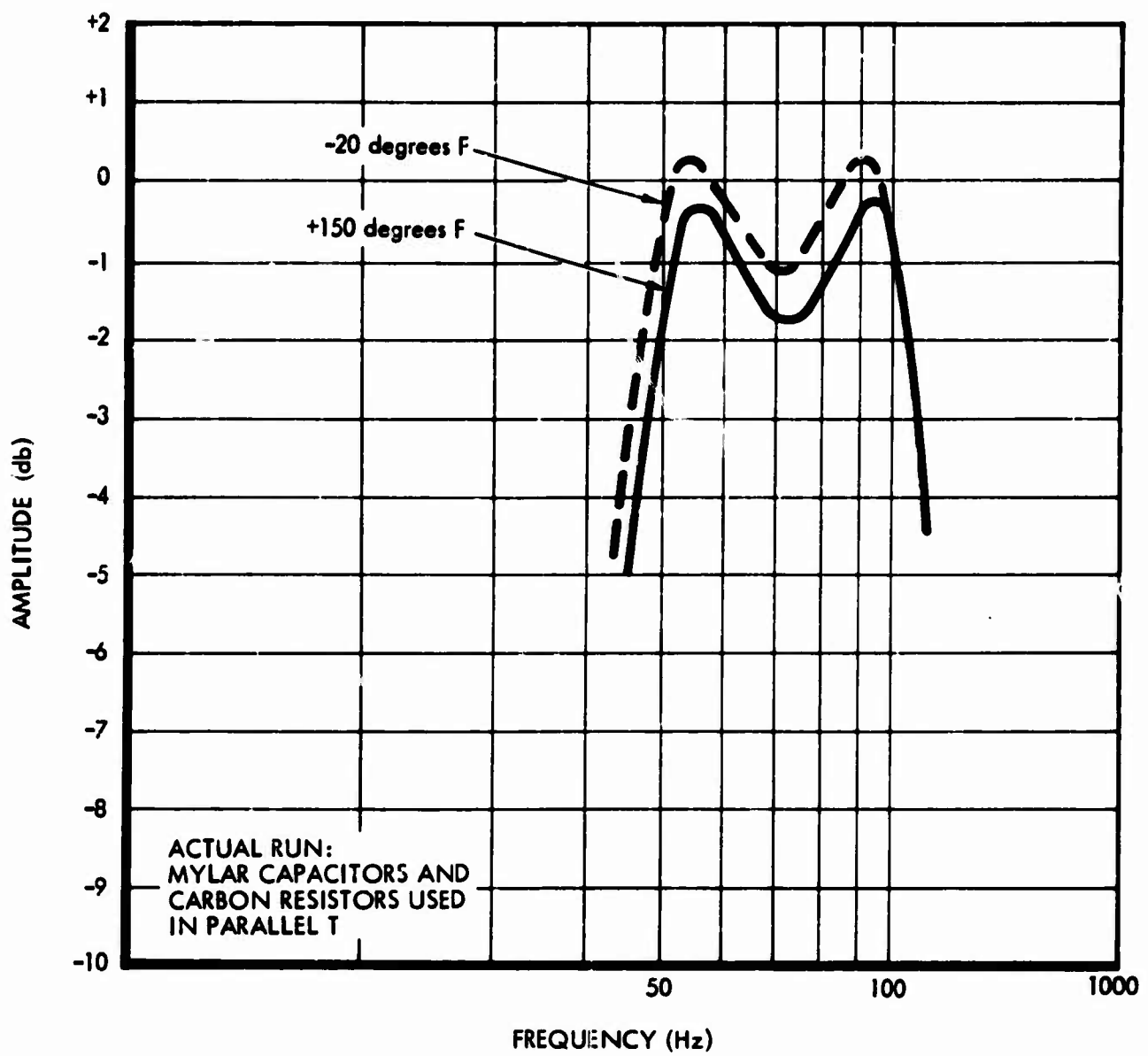
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Figure 10 - Varying C_N ±10 Percent



4076-17

Figure 11 - 50-100 Hz Filter Response Computer Results



4076-18

Figure 12 - 50-100 Hz Filter Response Environmental Results

C_U are changed -5 percent, while R_C , R_D , R_H , R_R , R_T , and R_X are changed +5 percent. As can be seen, the resulting response change is relatively minor, considering the amount the component values changed. Figure 12 verifies this experimentally. Mylar capacitors were found to have a positive temperature coefficient over the 170 F temperature range which is required of the system, while 1/4-watt carbon resistors have an equal negative temperature coefficient over this same range. Carbon resistors used in complementary positions with Mylar capacitors therefore cancel the radical effects on gain and frequency shift encountered when using either with a stable component. Figure 12, an actual environmental run, agrees with the computer evaluation and offers acceptable performance. This is considered a major breakthrough and means that the design of the filters is essentially complete except for possible minor refinements.

The component values listed in Table I are the one-percent tolerance values selected for the entire filter set. Some evaluation remains in the area where a choice exists between the larger but more stable Polystyrene and the more compact but less stable Mylar. Evaluation quantities have been ordered in these cases.

These particular types of components were chosen for different reasons. The Polystyrene capacitors in the parallel T will consist of two capacitors in parallel to get extremely close to the five place computed values. The input and feedback capacitors will be one percent molded Polystyrene. The Mylar capacitors will be one percent and bridged to match each other. All of the resistors, except those used with Mylar capacitors, will be epoxy coated metal film one-percent resistors. The carbon resistors used will be bridged from five-percent values. Approximately 80 percent of the parts for the filters have been ordered, including all of the operational amplifiers.

The reason for the choice of these components, with this undesirable type of coefficient, is the desire to reduce the volume of each filter. The lower frequency channels require capacitors of 0.1 μ i. These capacitors, being Mylar type and not electrolytic, are fairly large but are smaller than more stable types, such as Polystyrene. Their size, being the largest in the complex of components, is the width limit for the printed circuit boards. The use of Polystyrene capacitors in the higher frequency channels affords excellent stability for temperatures below 170 F.

The cross-over point, where the Polystyrene capacitors are small enough to fit the maximum width of the printed circuit cards, is 200-400 Hz. On this filter, the size of the Polystyrene capacitors is the approximate size of the largest Mylar capacitor.

The curves previously discussed (Figures 11 and 12) give some indication of the tolerance which may be expected from the filters.

For this test, all values of components were within one percent of the computed values, except for R_M and R_Y . They were changed to give 0 db of gain at the resonance frequencies. Figure 12 shows that the Q of each of the poles was then too high, giving excessively steep slope to the bandpass characteristic and too much ripple in the center of the bandpass. To compensate for this, it may be necessary to tailor R_B and R_Q as well as R_M and R_Y on some filters.

TABLE I - ACTIVE FILTER COMPONENTS

Frequency		FIRST POLE							
		R_C and R_D (k Ω)	R_H (k Ω)	R_B (k Ω)	R_M (k Ω)	C_F and C_G	C_E	C_A	C_N
30	30.7	<u>100</u>	<u>49.9</u>	715	715	<u>0.0523</u>	<u>0.1036</u>	0.00715	0.00715
50	51.5	<u>100</u>	<u>49.9</u>	536	536	<u>0.0309</u>	<u>0.0619</u>	0.00576	0.00576
100	103.0	<u>100</u>	<u>49.9</u>	536	563	<u>0.0154</u>	<u>0.0309</u>	0.00301	0.00301
200	206.1	76.8	38.3	412	412	0.01	0.02	0.00182	0.00182
400	408.0	39.2	19.6	348	348	0.01	0.02	0.0011	0.0011
600	608.8	26.7	13.0	332	332	0.01	0.02	0.787 pf	787 pf
800	809.2	19.6	9.76	324	324	0.01	0.02	604 pf	604 pf
1000	1009.5	31.6	15.8	634	634	0.005	0.01	249 pf	249 pf
1200	1209.6	26.1	13.0	619	619	0.005	0.01	210 pf	210 pf
1400	1409.8	22.6	11.3	619	619	0.005	0.01	182 pf	182 pf

SECOND POLE

Frequency		SECOND POLE							
		R_R and R_T (k Ω)	R_X (k Ω)	R_Q (k Ω)	R_Y (k Ω)	C_V and C_W	C_V	C_P	C_Z
50	48.8	<u>100</u>	<u>49.9</u>	215	715	<u>0.0324</u>	<u>0.0649</u>	0.0147	0.00453
100	97.1	<u>100</u>	<u>49.9</u>	162	536	<u>0.0162</u>	<u>0.0324</u>	0.01	0.00309
200	194.1	<u>49.9</u>	<u>24.9</u>	80.6	267	<u>0.0165</u>	<u>0.0324</u>	0.01	0.00309
400	388.2	41.2	20.5	64.9	221	0.01	0.02	0.00619	0.00182
600	588.2	26.7	13.3	73.2	243	0.01	0.02	0.00374	0.0011
800	788.4	20	10	76.8	255	0.01	0.02	0.00261	787 pf
1000	988.6	16.2	8.06	78.7	216	0.01	0.02	0.0020	604 pf
1200	1188.7	26.7	13.3	162	536	0.005	0.01	825 pf	249 pf
1400	1388.8	22.6	11.5	162	549	0.005	0.01	698 pf	210 pf
1600	1588.9	20	10	187	549	0.005	0.01	604 pf	182 pf

- Notes: 1. Underlined values denote carbon resistors and mylar capacitors. Otherwise all resistors are metal film and capacitors are polystyrene.
2. Values of components shown in Figure 1.

From these and similar results, it is expected that the over-all response of all of the filter channels may be held to no greater variation than ± 1 db with all component tolerances and amplifier gain variations.

3. RANGE DELAY AND GATED OSCILLATOR

Signal Corps Technical Requirement SCL-8029 specifies 10 range gates of either 50 or 100 meters, incrementally delayable in 20 steps of 5 range gates each. Furthermore, the maximum rms jitter between radar trigger and the range gates is to be less than 1.5 nanoseconds.

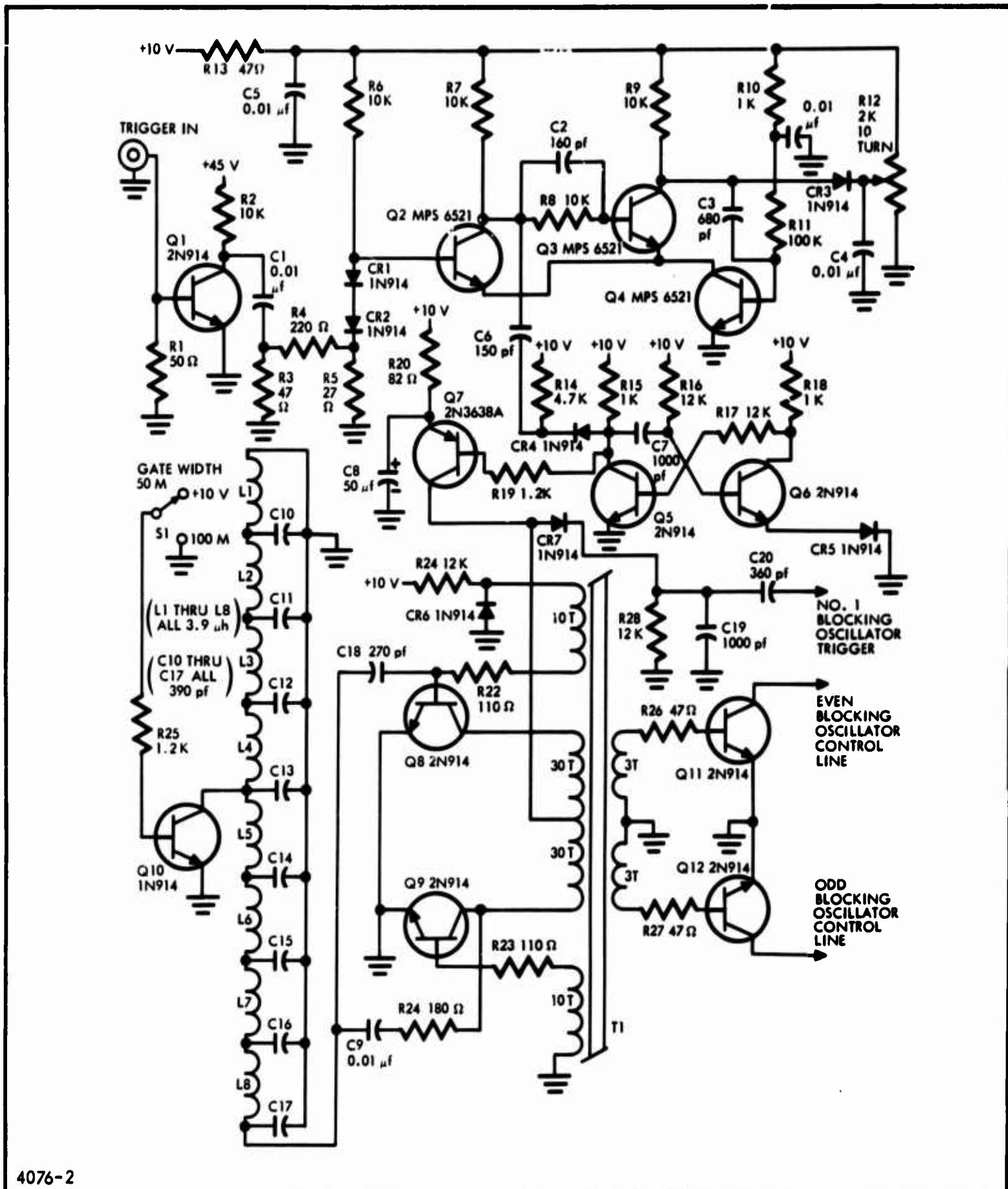
The initial concept for circuits to accomplish the delay and range gating employed digital techniques. However, further investigation revealed that a rather complex circuit would be required and, while use of integrated circuits would reduce size and weight, power consumption and cost would be greater than desired.

To overcome this problem, other delay techniques were considered. Previous work with a transistorized analogy of the phantastron delay circuit had indicated extremely low jitter. Therefore, after discussion with the project office, jitter tests on this circuit were conducted. The measured jitter was approximately that allowed and remained at this magnitude during changes in voltage and temperature. As a result of these tests, it was decided to utilize this circuit instead of the digital delay technique. The advantages of the phantastron circuit are:

1. Less complex circuits
2. Extremely low power consumption
3. A continuously variable range delay from 0 to 10,000 meters, controllable with a knob calibrated directly in meters
4. No change in range delay when switching between gate widths.

The range-delay circuit schematic diagram is shown in Figure 13. Q1 is a trigger-standardizing circuit, providing a constant output pulse over a wide variation of input pulses. This operation is achieved by operating the transistor in the avalanche mode. Any input pulse which exceeds the emitter-base threshold places Q1 in avalanche, and C1 is discharged through Q1 and R3, producing a pulse across R3 with extremely fast rise time and constant shape.

Transistors Q2, Q3, and Q4 comprise the phantastron delay circuit. During the inter-pulse period, Q2 and Q4 conduct and Q3 is cut off. Also, C3 charges toward +10 volts through R10 until caught by CR3 at a voltage determined by the setting of R12. The negative triggering pulse developed across R3 is coupled through R4, CR1, and CR2 and cuts Q2 off, beginning the timing state. In this condition, Q3 saturates and Q2 is held off. Q4 now operates as an integrator and discharges C3 at a constant rate. Since the base voltage of Q4 remains approximately constant, the collector of Q3 exhibits a linear



4076-2

Figure 13 - Range Delay and Gated Oscillator Schematic Diagram

run-down toward ground. This run-down will continue until the emitter of Q2 becomes less positive than its base. At this point, Q2 saturates and Q3 is cut off, terminating the delay cycle, and C3 is allowed to recharge for the next cycle. The length of the delay period is controlled by the initial voltage across C3, and this in turn is controlled by the setting of R12. Because the discharge of C3 is linear, the delay control, R12, can be linearly calibrated directly in range. A further advantage is that delay is independent of supply voltage for first order effects.

At the termination of delay, a negative transition is produced at the collector of Q2. This transition is coupled through C6 and triggers the oscillator gate generator, a monostable multivibrator (Q5 and Q6). A negative pulse is generated at the collector of Q5 which saturates Q7, through R19, and provides power for the gated oscillator (Q8, Q9, and T₁). The duration of this pulse is greater than 6.6 μ sec to allow ten 100-meter range gates to be generated.

The gated oscillator is a standard magnetic multivibrator circuit (Q8, Q9, and T₁) operating at approximately 500 kHz. This configuration was chosen over the standard LC ringing circuit because of its low output impedance capabilities. However, to make it usable for this application, it must always start with the same phase, and its frequency of oscillation must be controlled by a more stable element than the saturation characteristics of T₁. Proper starting phase is ensured by returning Q9 base winding directly to ground while Q8 base winding is returned to a positive voltage through R24. Thus, when power is applied to the circuit by Q7, Q8 always starts in the saturated condition.

Frequency stability is controlled by a shorted delay line connected to Q9 collector through C9. The two-way path delay of the line determines the half cycle pulse width. C18 provides cross-coupling speedup to ensure proper operation. Range-gate width is changed by Q10 which, when saturated, places a short on the delay line at the halfway point. With Q10 cut off, the oscillator provides a square wave output with a half cycle width equivalent to a 100-meter radar range. When Q10 is saturated, the half cycle width is equivalent to 50 meters.

The output of the gated oscillator drives Q11 and Q12. These transistors alternately short the even and odd blocking oscillator control lines to ground. The number 1 blocking oscillator trigger is obtained from the leading edge of the oscillator gate. CR7, C19, and R28 provide a pulse stretching circuit to remove hash from the trigger line.

4. RANGE GATES

The range-gate timing consists of 10 blocking oscillators connected so that the termination of one triggers the succeeding one. The duration of each blocking oscillator is

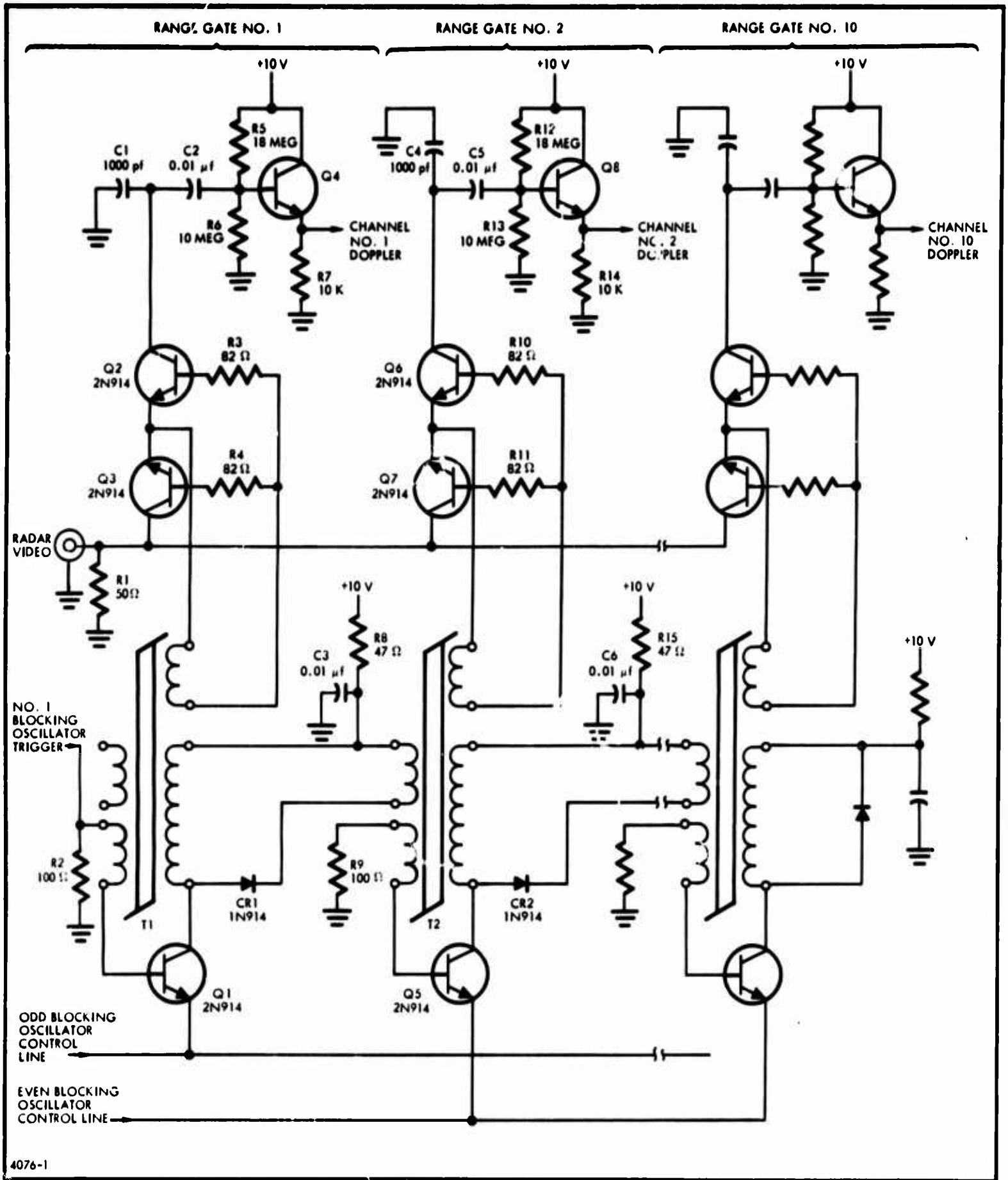
determined by the gated oscillator previously discussed. Blocking oscillators were chosen over digital circuits for several reasons:

1. Low power consumptions, since they draw power only during their active interval
2. Low output impedance for driving the range gate
3. Isolated output to allow use of a floating type range gate.

Figure 14 is the schematic diagram of the range gate circuit. At the start of the range gate, number 1 blocking oscillator trigger arrives across R2, turning Q1 on. At the same time, the first half cycle of the gated oscillator is under way and the odd blocking oscillator line is grounded. Under these conditions, Q1 and T1 act as a normal blocking oscillator and an output pulse is applied to Q2 and Q3, the actual range gate. Q1 remains in conduction until the first half cycle of the oscillator has terminated. At this time, the ground is removed from the odd blocking oscillator line and applied to the even blocking oscillator line. Q5 ceases conduction at this point, and the pulse is removed from the range gate. The collapse of the field of T1 causes an inductive kick above +10 volts which is caught by CR1 and dissipated across the trigger winding of T2. This pulse places Q5 into conduction, and the number 2 range gate, Q6 and Q7, is energized. Q5 will conduct until the second half cycle of the gated oscillator terminates. The inductive kick of T2 triggers the following blocking oscillator. The action of the series triggering and the blocking oscillator control lines sequentially energize all 10 blocking oscillators. Since the natural period of each blocking oscillator is approximately 1 μ sec, either 50- or 100-meter range gates can be utilized by merely changing the half cycle period of the gated oscillator.

The video is sampled by a two-transistor gate, Q2 and Q3, and stored on capacitor C1. Q2 and Q3 are normally cut off and present a high impedance between the video line and the storage capacitor. When the blocking oscillator fires, a positive pulse is applied between the base-emitter junction of Q2 and Q3 through R3 and R4 and places both these transistors hard into saturation. As long as the emitter-base current is greater than the emitter-collector current, the gate is bidirectional and gives an extremely low series impedance. With a 50-ohm video impedance, C1 can be fully charged to the video level during the 50-meter gate width.

To preserve the charge on the storage capacitor during the interpulse period, Q4, a field effect transistor connected as a source follower, is provided. With this circuit, C1 discharges less than five percent at a prf of 1600 pps, when driving a 10-k Ω load.



4076-1

Figure 14 - Range Gate Schematic Diagram

5. LOW-VOLTAGE POWER SUPPLY

Figure 15 is the schematic diagram of the power inverter and time delay portions of the low-voltage power supply (Q1, Q2, and T1 comprise the 10-kHz power inverter). This circuit is of standard design and needs no descriptions, except it should be noted that PNP transistors are used so that the output swings between +10 volts and -10 volts. Q3 is a unijunction oscillator used as a starter circuit for the inverter. A secondary winding and two rectifiers, CR13 and CR14, provide 6.3 volts to power the flood gun filament of the storage display tube.

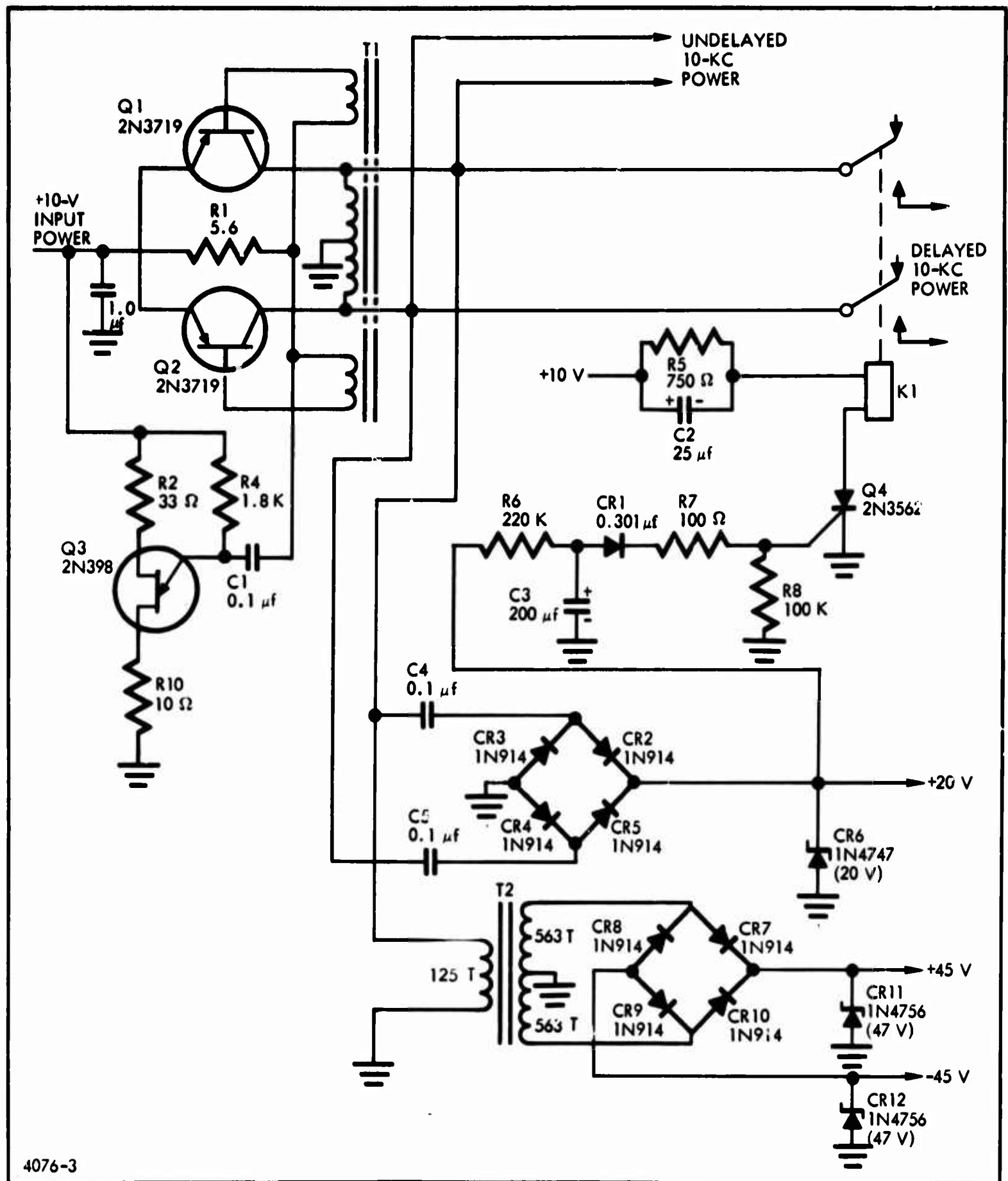
When cold, the two filaments of the storage display tube reflect a low resistance back to the inverter. The 10-kv power supply also reflects a heavy load until the capacitors are charged. If these two loads occur simultaneously, the inverter becomes unreliable. For this reason, and also to protect the crt during warmup, a 30-second time delay has been provided. When power is first turned on, relay K1 and the SCR, Q4, are open. C1 starts to charge, through R6, toward the +20 volts supplied from the rectifier CR2 through CR5. After approximately 30 seconds, C3 has charged to +10 volts and, at this point, CR1, a four layer diode, conducts, discharging C3 into the gate of Q4. Q4 is thus turned on and will remain on until the +10-volt supply is turned off. To reduce holding current, R5 is placed in series with K1 coil. C2 provides heavy initial current so that K1 will close.

T2 and the rectifiers, CR7 through CR10, supply plus and minus 45 volts. The -45 volts provide grid bias for the crt flood gun. Also, the undelayed 10 kHz powers the write gun filament and bias circuit. All of the remaining supplies are powered from the delayed 10-kHz power bus. CR11 and CR12 are overshoot clippers and not regulators.

Figure 16 shows the remaining portion of the low-voltage power supply. The basic technique used is series connected rectifiers with parallel connected inputs. This arrangement provides high efficiency, low-output impedance, and stable voltages, and, because of the high operating frequency and low voltage per step, the transformers can be small (approximately $1/2 \times 7/8 \times 1/4$ inch) and light. The zener diodes CR5, CR8, CR17, CR19, and CR24 are again used to clip the overshoot rather than regulate the output.

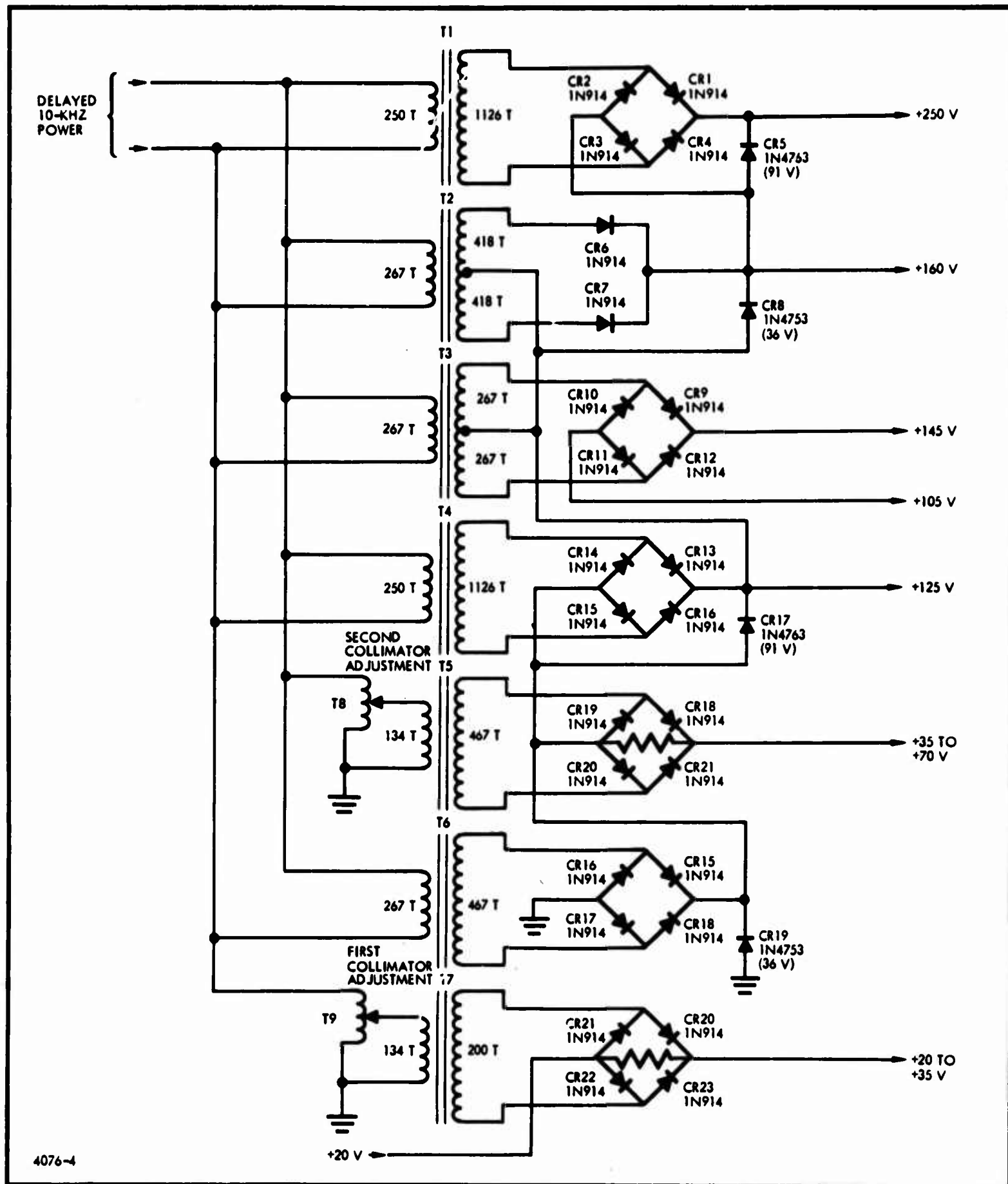
6. HIGH-VOLTAGE POWER SUPPLY

The high-voltage power supply is shown in Figure 17. The undelayed 10-kHz power is applied to T3 and rectifiers CR49 through CR52 and energizes the filament for the write gun. Ten-kHz power is also supplied to T4 and rectifiers CR53 through CR56. This second supply develops 90 volts negative, with respect to -2 kv, and ensures cutoff of the write gun. An optically coupled semiconductor device, Q2, is used to couple video



4076-3

Figure 15 - Power Inverter and Time Delay Circuit Schematic Diagram



4076-4

Figure 16 - Low-Voltage Power Supply Schematic Diagram

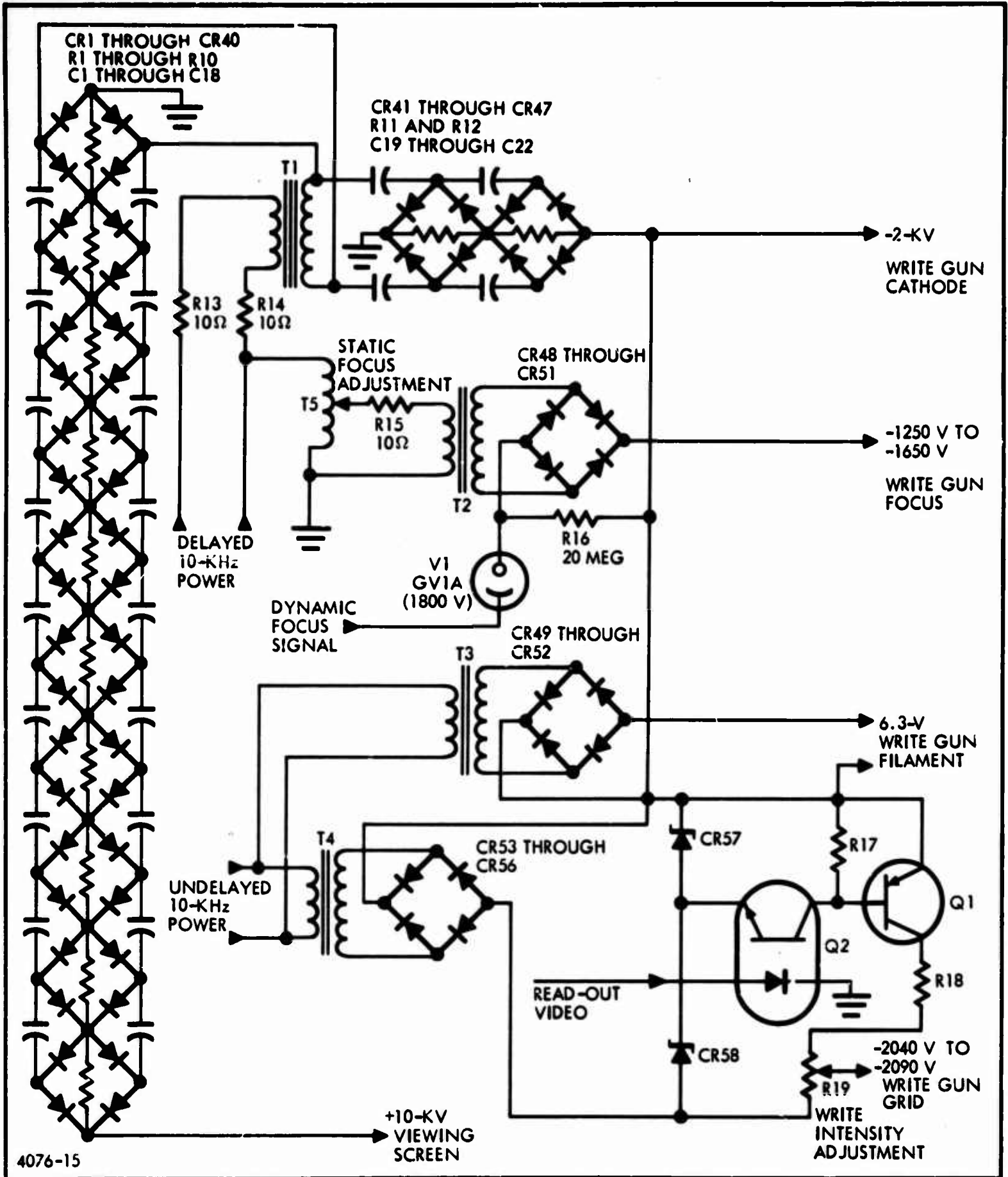


Figure 17 - High-Voltage Power Supply Schematic Diagram

from ground potential to the -2 kv. The transistor portion of Q2 is normally cut off and thus Q1 is held cutoff. In this condition, the write gun grid is held at -90 volts, with respect to its cathode, regardless of the setting of R19. When video is applied, the diode portion of Q2 emits sufficient light to cause the transistor portions of Q2 to conduct. Because the signal is coupled between the diode and transistor of Q2 by light, they can be insulated for high voltage. When the transistor of Q2 conducts, Q1 is placed in saturation, connecting R18 and R19 directly across the 90-volt supply. In this condition, the grid bias for the write gun is controlled by the setting of R19.

The delayed 10 kHz supplies power for the +10 kv, -2 kv, and focus supplies. The voltage multiplying rectifiers have been discussed in previous reports and will not be described here. The focus electrode is supplied from rectifiers CR48 through CR51 and transformer T2. To allow adjustment of this voltage, a variable transformer, T5, is provided. To meet resolution requirements, it has been found necessary to accomplish dynamic focusing. A straight line approximation of a negative going parabolic waveform is generated in the horizontal deflection circuit. This signal is coupled to the negative source through the corona type regulator tube V1, with R16 as the load resistor. The use of V1 allows the dynamic focus waveform to be coupled to the negative voltage without loss of waveform amplitude.

7. ESTIMATE OF POWER CONSUMPTION

The estimated power consumption of the entire unit is shown in Table II for both a 10-channel model and a 50-channel model. It will be noted that for the 10-channel model the power consumed is extremely moderate, and the 50-channel model is almost within the specified 35 watts.

The power required is, to an extent, determined by the number of range versus frequency lights that are illuminated. For purposes of this estimate, it has been assumed that 20 percent of the total number are energized at one time. It is believed that the value of 20 percent is a realistic value for typical field operation.

The 22 watts consumed by the filters in the 50-channel model could be reduced to a value of 13 watts with some circuit redesign. The extra effort was not considered necessary on the 10-channel model since only 1.8 watts would be saved.

8. PACKAGING STUDIES

A mockup of the Automatic Detection and Display Equipment for MTI (Figures 18 and 19) has been constructed as an aid in packaging studies. The high-voltage power supply can be identified as the potted module at the base of the crt. The low-voltage power supply, deflection amplifiers, and crt control circuits will be contained on the printed circuit

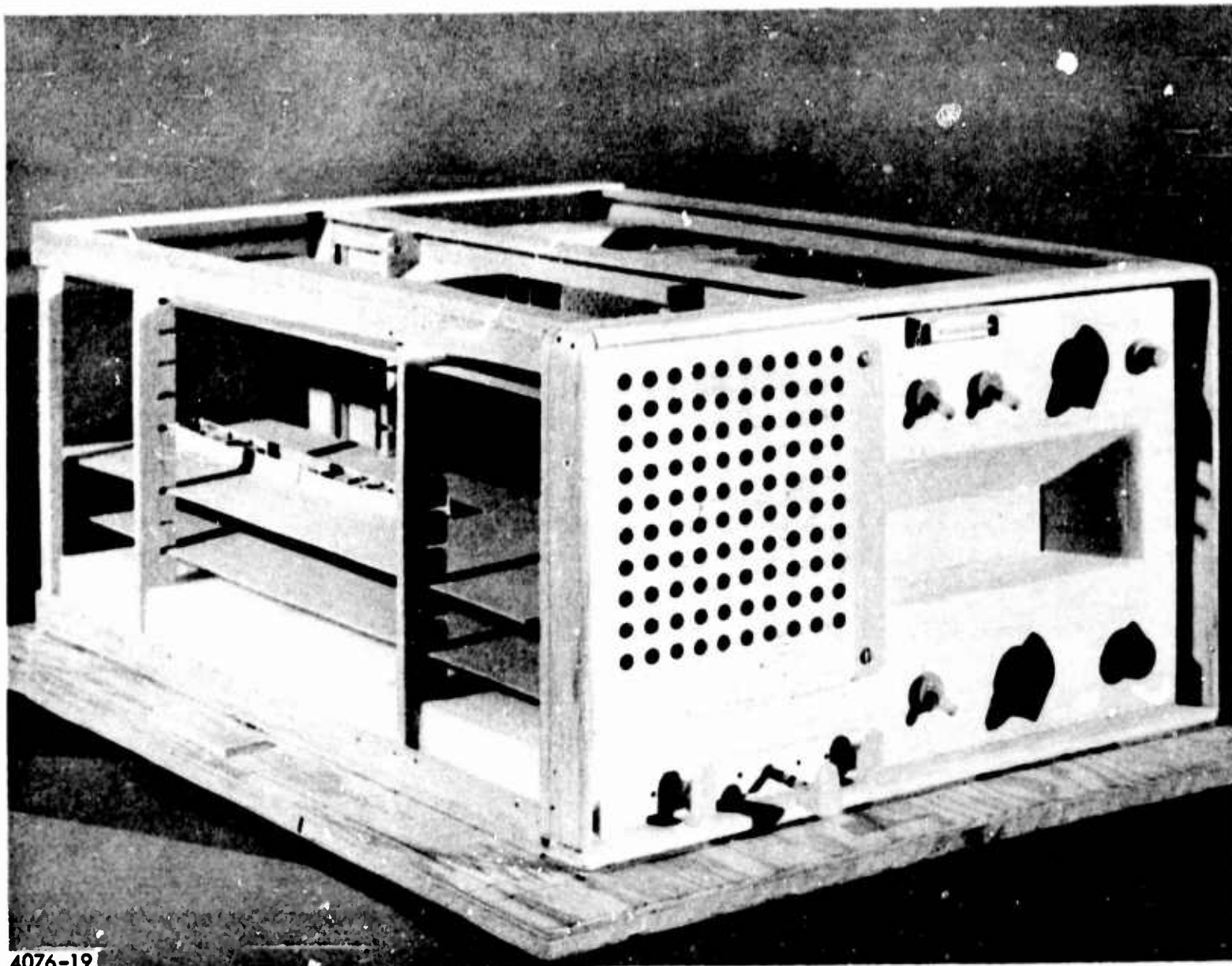
**TABLE II - ESTIMATED POWER CONSUMPTION OF
AUTOMATIC DETECTION AND DISPLAY EQUIPMENT FOR MTI**

Component	10 Channel (watts)	50 Channel (watts)
Crt, deflection, and power supplies	10.00	10.00
Filters	4.40	22.00
Integrators, threshold, and indicators	0.28	1.40
Delay and range gates	0.38	0.50
Read-out shift register	0.10	0.50
Read-out timing	<u>0.75</u>	<u>0.75</u>
Total	15.91	35.15
Assuming 95-percent efficiency of input voltage regulator	16.80	37.00

boards mounted around the neck of the tube in front of the high-voltage power supply. Ten identical range bin printed circuit boards will be mounted horizontally, to the left of the crt.

Packaging design of the high-voltage power supply is nearing completion, and fabrication will begin in the near future. This unit will consist of two printed circuit boards on which all components are mounted. These two boards and the crt socket will be potted in a clear flexible compound to provide insulation, shock, vibration isolation, and flexible restraint for the rear end of the crt. The entire potted assembly will be placed in a metal can with suitable radio frequency interference shields at all openings.

To reduce size and weight, a modified cordwood construction technique will be used for the 10 range bin boards. All passive components will be mounted between long, narrow printed circuit boards to form a cordwood module. All transistors and integrated circuits will be mounted on a large "mother board." The cordwood modules will also be mounted to the "mother boards" with components horizontal and interconnecting boards



4076-19

Figure 18 - Automatic Detection and Display Equipment for MTI Mockup (Front View)

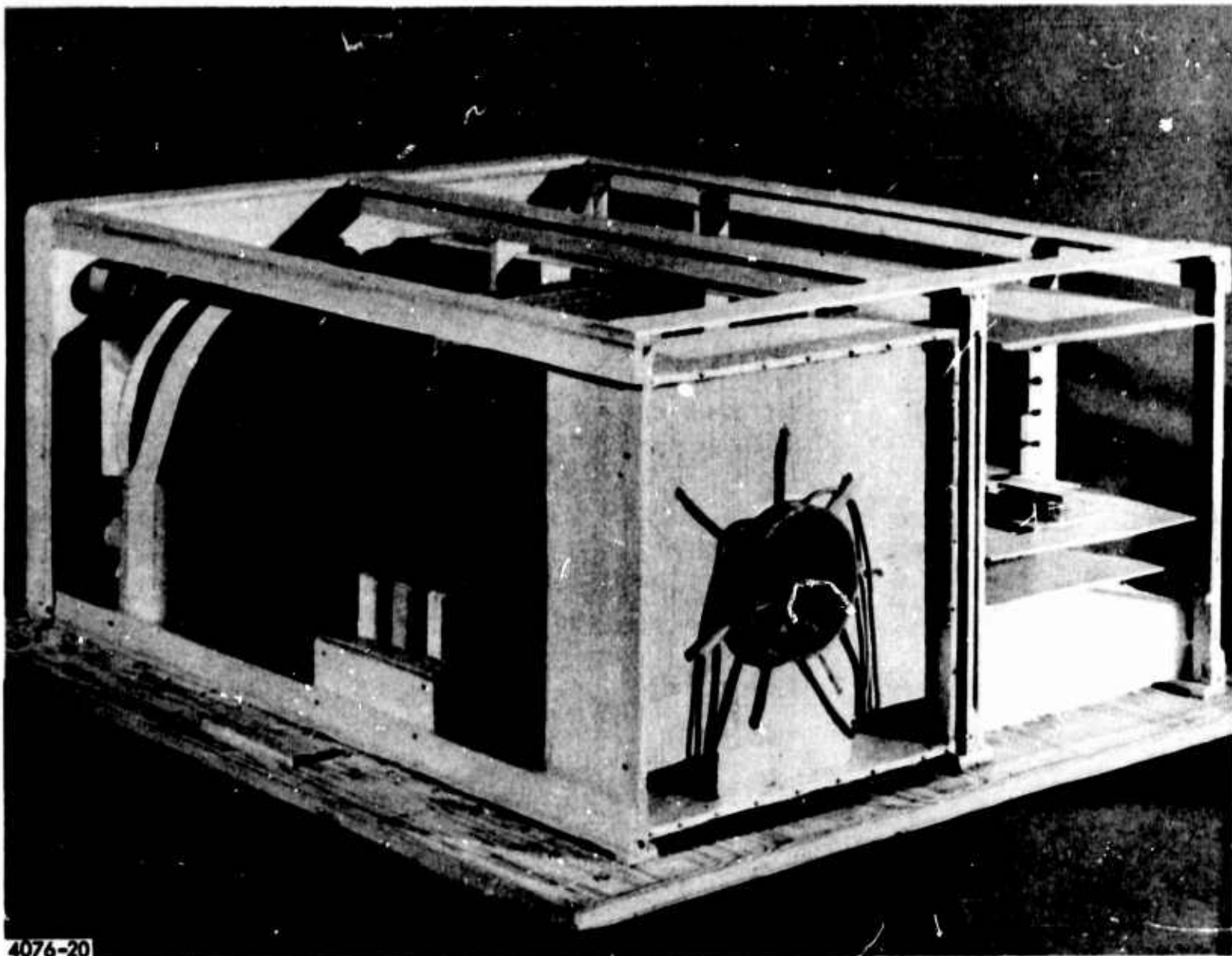


Figure 19 - Automatic Detection and Display Equipment for MTI Mockup (Rear View)

vertical and running parallel to the long dimension of the "mother board." Thus, the interconnecting boards of the modules will not only support components but also stiffen the "mother board." Additional stiffeners will be placed across the short dimension of the "mother board." Figure 20 shows a mockup of this packaging technique.

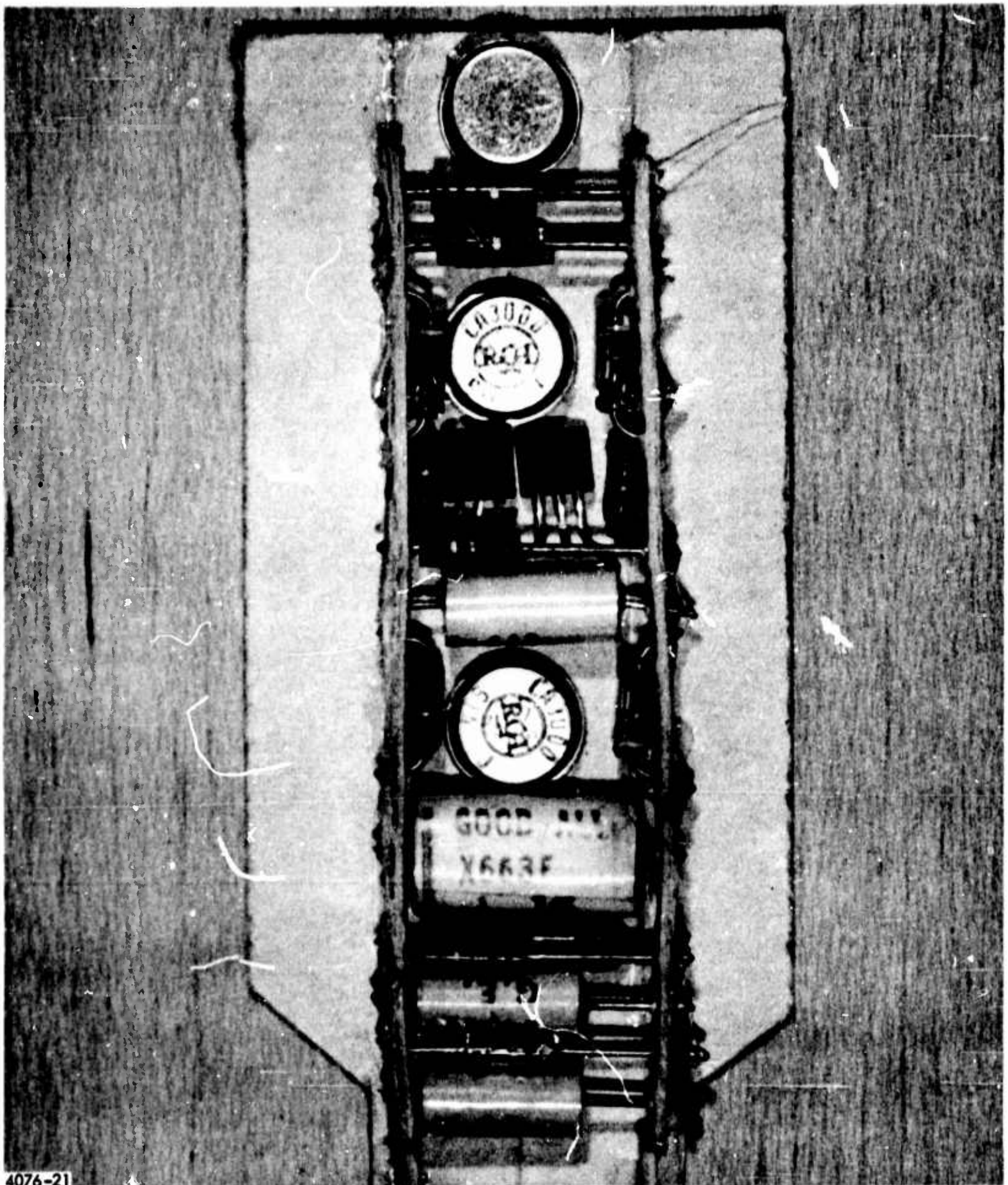


Figure 20 - Mockup of Modified Cordwood Packaging Technique

**SUPPLEMENT TO REPORT NO. 3, DATED 1 MAY THROUGH 31 JULY 1966. FOR SERVICES
UNDER CONTRACT NO. A-28-143 AMC-0711(E)**

1. IDENTIFICATION OF KEY TECHNICAL PERSONNEL

The following is a list of key technical personnel and the manhours performed by each on the Automatic Detection and Display Equipment for MTI program from 1 May 1966 through 31 July 1966.

<u>Name</u>	<u>Title</u>	<u>Manhours</u>
Knight, C. C.	Design Engineer	232
Miller, C. R.	Development Engineer	327
O'Herren, D. H.	Development Engineer	136
Roehrman, K. E.	Development Engineer	256
Schoenfeld, L. L.	Project Engineer	435
Wood, B. G.	Development Engineer	443

No new personnel were assigned to the program during this reporting period.

2. CONFERENCES

One conference was held during the reporting period.

Date: 23 June 1966

Place: Goodyear Aerospace Corporation
Litchfield Park, Arizona

Organizations Represented:

Goodyear Aerospace
(Arizona)

CS/TA Laboratory
Fort Monmouth, New Jersey

J. W. Cusick

S. Graveline

L. C. Graham

O. E. Rittenbach

S. D. Robertson

L. L. Schoenfeld

B. G. Wood

3. PROGRAM FOR THE NEXT INTERVAL

The remaining portions of the circuit design will be completed, but the main effort will be directed toward packaging the equipment.

DOCUMENT CONTROL DATA - R&D

(Security classification of title, body of abstract and indexing annotation must be entered when the overall report is classified)

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		2b. GROUP	
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13. ABSTRACT This is the third quarterly report, documenting the development of the Automatic Detection and Display Equipment for MTI. The circuit design effort was approximately 80 percent completed at the end of this reporting period. The active filter was analyzed to minimize the effect of component tolerance and variation. A range-delay and range-gating circuit was also developed. The revised cathode-ray tube (crt) power supply, which includes provisions for dynamic focus, was completed and is ready for test. Packaging studies and power estimates indicate that size and power consumption will be better than the specified values.			

14. KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
Video Processor Range Gated Filters B Display Doppler Display Active Filters Combat Surveillance Radars Operational Amplifiers Doppler Simulator						

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