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GENERAL DYNAMICS

~~ASTRONAUTICS~~

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A2136-1 (REV. 6-61)

FM RECEIVER

DYNAMIC RESPONSE

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SUMMARY

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This report describes a study into the dynamic response produced
by an FM receiver employing an amplitude sensitive discriminator.
A simple, graphical explanation of the anticipated response is pre-
sented. Empirical results obtained in the evaluation of an FM
telemetry receiver are included. Analytical techniques leading to a
mathematical proof of receiver behavior are discussed. The appli-
cation of the analog computer to the problem is outlined and the
analog simulation of the receiver system described. Computer solu-
tions for a number of conditions are included in the form of graphic
recordings. Finally, it is concluded that dynamic response such as
that exhibited by the AVCO range safety receiver is the natural be-
havior anticipated from such a receiver.

1. INTRODUCTION

A new receiver GD/C P/N 55-01233, manufactured by AVCO, is being
incorporated into the Centaur range safety system. The performance
requirements of the receiver are described by GD/C specification.[1]

Evaluation tests of the receiver at NASA test facilities and at GD/C
revealed a dynamic response characteristic that initially appeared un-
usual to test personnel. Range personnel were disturbed to the point
where they were reluctant to allow the receiver to be flown. The
dynamic test was performed by supplying an rf signal modulated by two
tones to the receiver. The rf signal carrier frequency was initially
set well outside the passband of the receiver (frequency lower than
receiver); then, with rf level held constant, the carrier frequency was
moved toward the receiver passband until a command actuation was
observed; this frequency was recorded. The carrier frequency movement
was then continued until the command deactivated and that frequency
recorded. This process was continued until the carrier frequency was
well beyond the passband on the high frequency side. The rf level was
then changed 10 db and another sweep made through the spectrum observing
again the carrier frequencies producing command "on-off" states. From
the resulting data a curve such as that of Figure 1 was generated. The
curve shows two side lobes and a wider center lobe where the command
is activated. Narrow command deactivate zones (approximately one-
half the peak deviation in width) separate the center lobe and sidelobes.
The center lobe is narrower than the unmodulated response (see Figure 1)
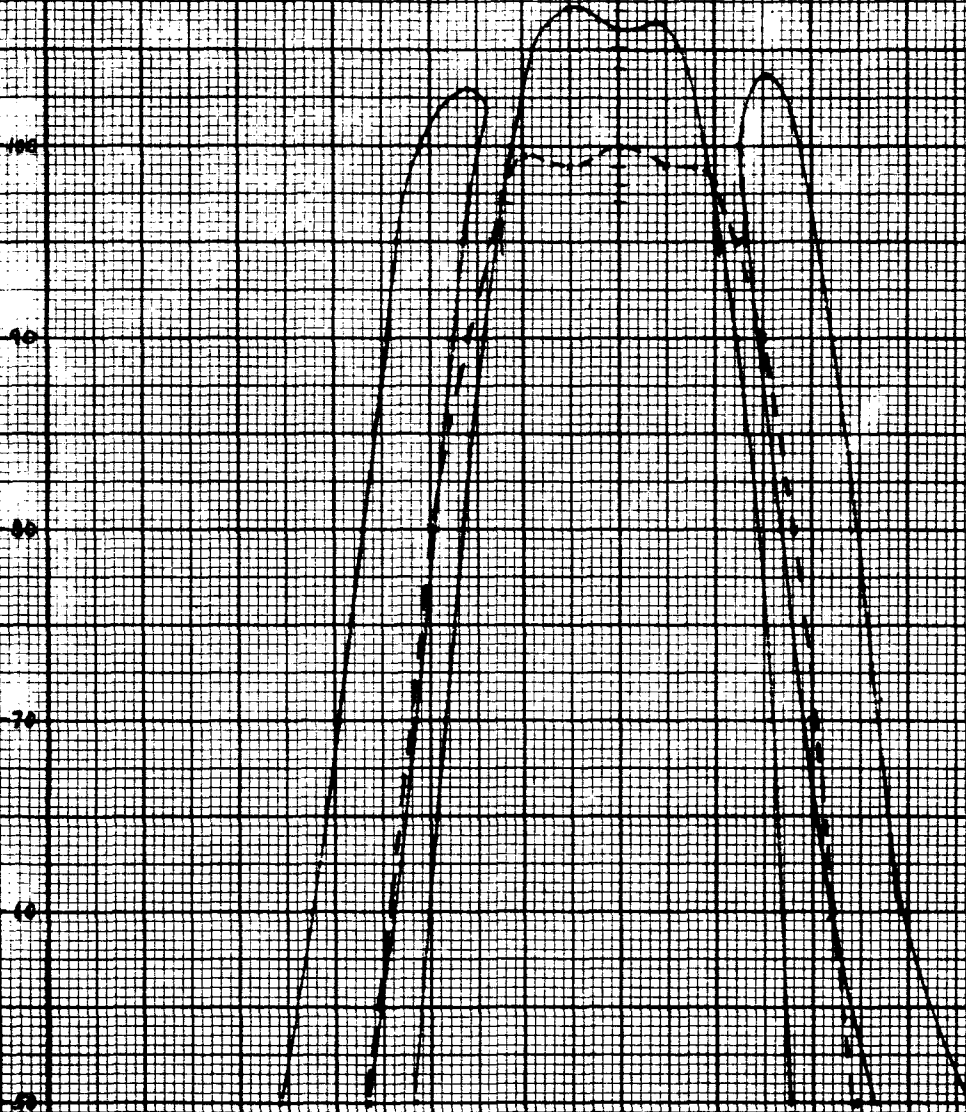by an amount approximately equal to the peak deviation.

Figure 1. Typical Observed Response

Typical Observed Response

The study described in this report was initiated to gain a better understanding of the dynamic response characteristic of the FM command receiver.

2. RECEIVER DESCRIPTION

A brief description of the receiver is included in this section for the reader's convenience. More detailed descriptive information is available in the receiver handbook.[2]

A simplified block diagram of the receiver is shown in Figure 2. The three center blocks of the diagram are of greatest interest for the discussion here for it is their characteristics that establish the dynamic response of the receiver.

The IF filter features a sharp selectivity with a low 60 db bandwidth to 3 db bandwidth ratio. The nominal filter transfer function is that of a 4 pole Chebishev filter having a bandwidth of about 240 kilocycles and 2 db passband ripple. The actual filter is a 4 crystal lattice network. This type of filter does not have a linear phase characteristic (a requirement for distortionless FM) in the passband and the phase response changes violently near the bandedges. Figure 3 is a graph of the anticipated phase response for the filter.

Limiting is accomplished in three amplifier stages that precede the discriminator. Each stage reaches its limiting level at a different rf input level to the receiver thereby providing a large overall dynamic limiting range.
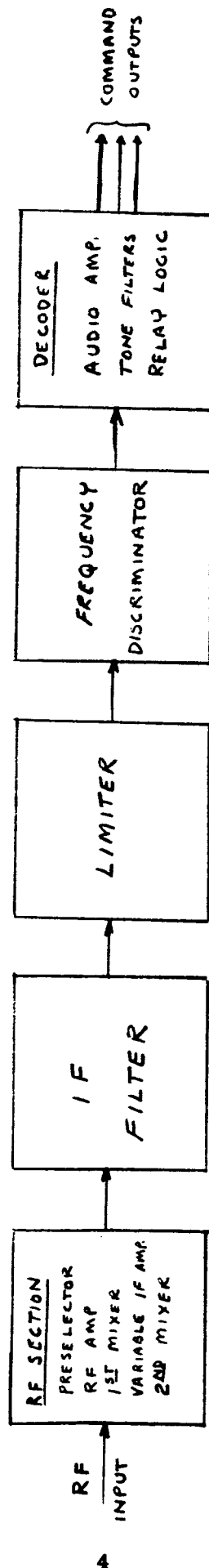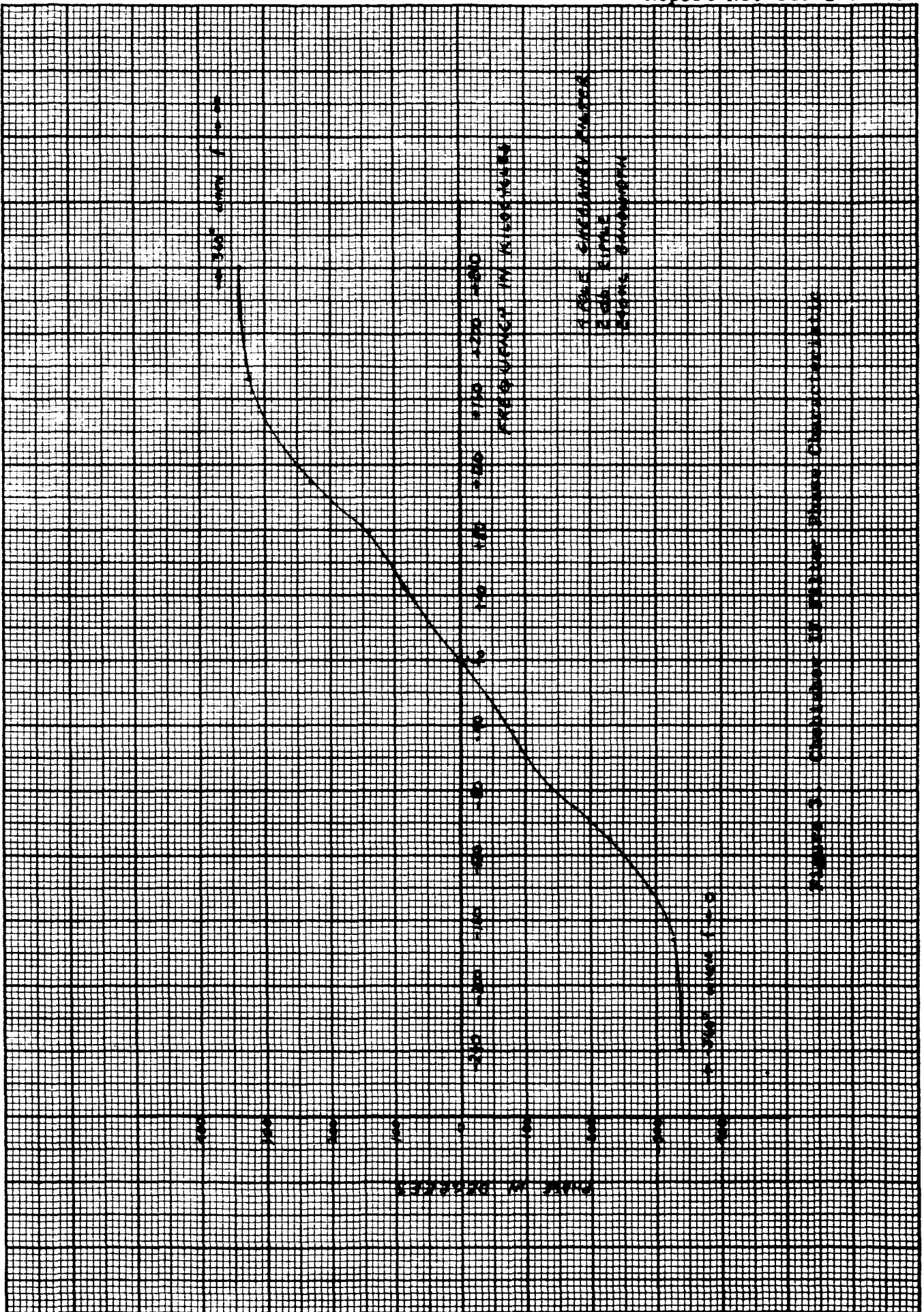
FIGURE 2. Simplified Block Diagram of the FM
Command Receiver

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Figure 3. Compilation of Phase-Shift Characteristic

The frequency discriminator is a conventional Foster-Seeley circuit.
The discriminator bandwidth (frequency separation between response
peaks) is much wider than the IF bandpass of the receiver to assure good
linearity over the region of interest.

3. ANTICIPATED DYNAMIC RESPONSE

The dynamic response exhibited by the range safety receiver (or any FM
receiver employing a Foster-Seeley discriminator) can be explained by
the "composite" frequency discrimination characteristic of the receiver.
The Foster-Seeley discriminator's output amplitude is proportional
to input voltage and the frequency deviation of the signal; for this
reason, the discriminator is always preceded by one or more limiter
stages. The selectivity of the receiver is established by the bandpass
filter located ahead of the limiter stages. Typically the IF bandwidth
is narrower than the frequency discriminator characteristic. The com-
posite frequency discrimination characteristics of the receiver can
be described by a superposition of the IF bandpass and FM discriminator
characteristics. Figure 4 illustrates such a composite frequency dis-
crimination characteristic for a hypothetical receiver to aid dis-
cussion. From Figure 4 it may be noted that the composite discriminator
output decreases when the input frequency is sufficiently offset from
band center because the IF bandpass filter attenuation drops the signal
below limiting level. Again, directly from Figure 4, it is seen that
the composite discriminator characteristic has slopes of opposite
sign at the bandedges compared to the band center slope. The result
is that there are points near the composite discriminator response
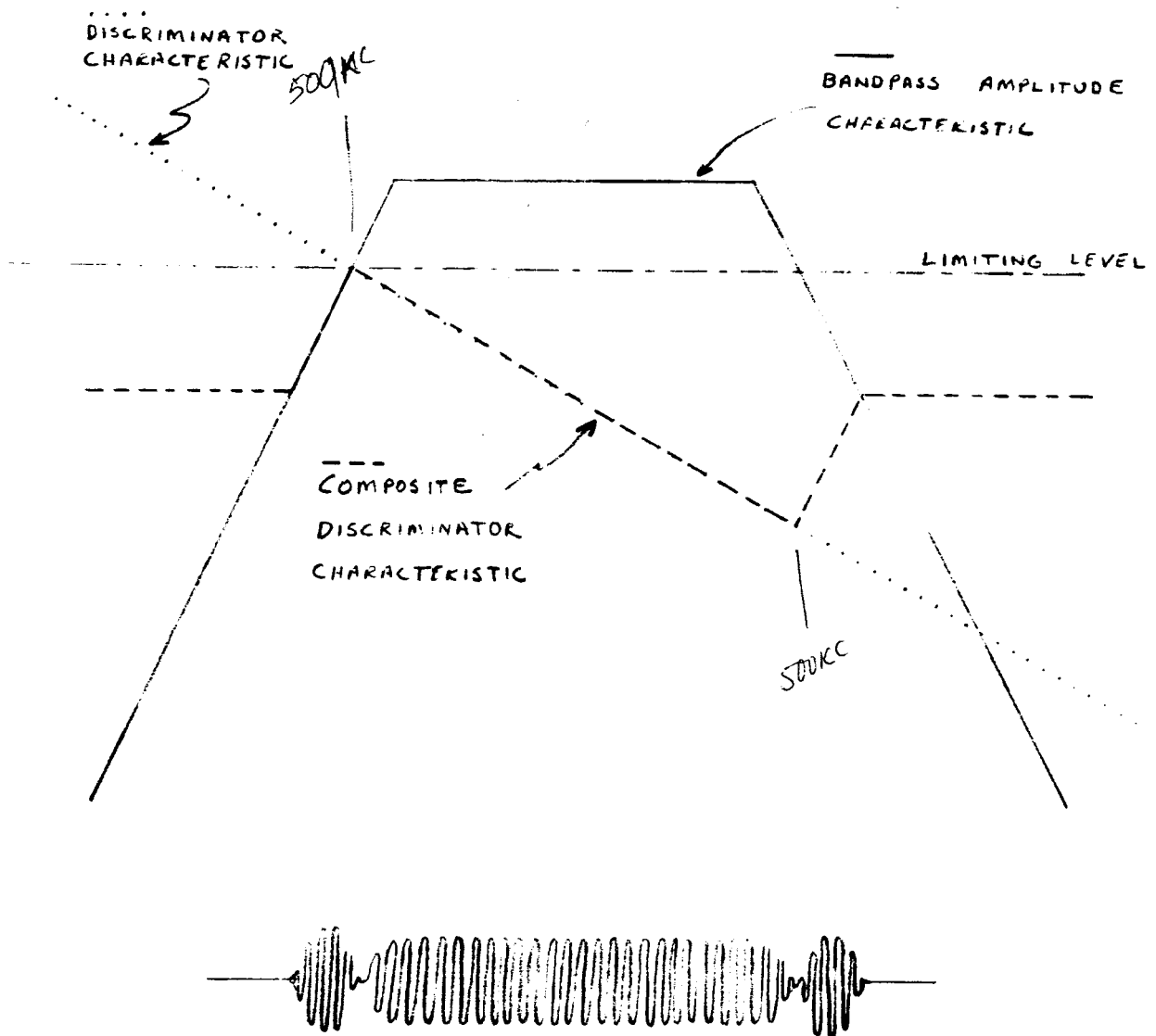
6

Figure 4. Frequency Characteristics

peaks where the demodulated output signal is cancelled and any resultant is a small amount of second harmonic. There are two positions, one on each side of the passband beyond the cancellation points, where the discriminator output signal again increases but is of opposite sign due to operating on the outside slopes of the composite discriminator characteristic. A frequency sweep of a modulated signal, therefore, produces a signal such as illustrated in the lower portion of Figure 4. (See also computer solutions in Section 7.) The center lobe is typically narrower than the width of the unmodulated passband because of the frequency deviation and the difference in the magnitude of the inter and outer discriminator slope magnitudes.

Pender & McIlwan[3] briefly discuss the dynamic response of FM receivers.

4. EMPIRICAL RESULTS

A Nems-Clarke R1037-A receiver was used to obtain data on FM receiver dynamic response. This receiver was used for the test to demonstrate that the dynamic response exhibited by the AVCO receiver is not peculiar to that receiver but typical of FM receivers employing the Foster-Seeley type discriminator. Additionally, it was convenient to use the Nems-Clarke receiver because of the availability of the limited IF, video and DC FM discriminator outputs at accessible terminals.

Photographs of Figure 5 show null outputs as they occurred at the points of slope reversal.
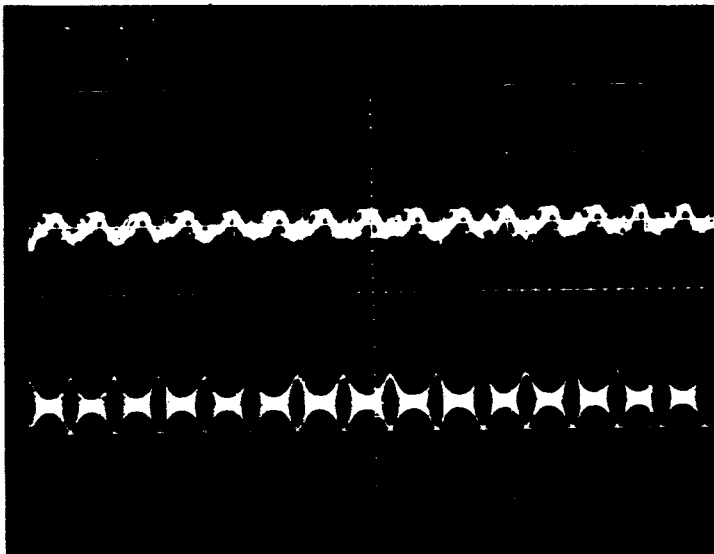
8

Low Frequency Null 214.668 MC
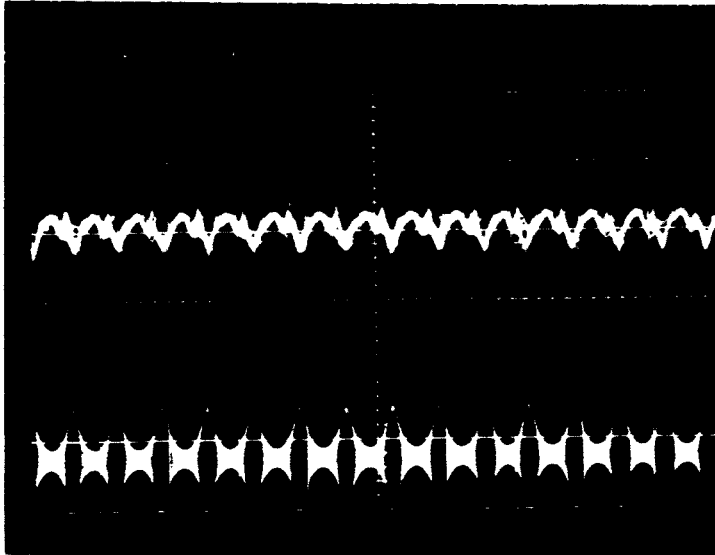
Receiver Video Output

Receiver Limiter Output

High Frequency Null 215.429 MC

Receiver Video Output

Receiver Limiter Output

Test Conditions (both photos)

Receiver: Nems-Clarke R1037-A S/N 355
          (300 KC IF Bandwidth)
Modulation: 7.5 KC Rate 30 KC Peak
RF Level: 3000   volts

Figure 5. Null Points Measured on the Nems-
Clarke Telemetry Receiver

The amplitude modulation which cancels the FM is clearly evident in the limited IF output (lower trace on the photos). The AM indicates operation below the limiting level and on the IF filter skirt. Figure 6 illustrates the measured composite discriminator characteristic, while, Figure 7 shows normal operation at band center.

5. ANALYTICAL TECHNIQUE

A mathematical analysis was undertaken in an attempt to improve the understanding of the dynamic response characteristics, for additional verification of experimental results, and to determine if any other factors (such as the Chebishev filter phase characteristic) played a significant role in the observed response.

The classical approach for the solution of a problem of this type is examined. While the actual mathematical solution of this problem is quite formidable, the approach is simple and straight-forward. A reasonable simplification of the receiver system considers only the IF filter, frequency discriminator and output tone filters (see Figure 8). The input to this system is easily described as a time function.
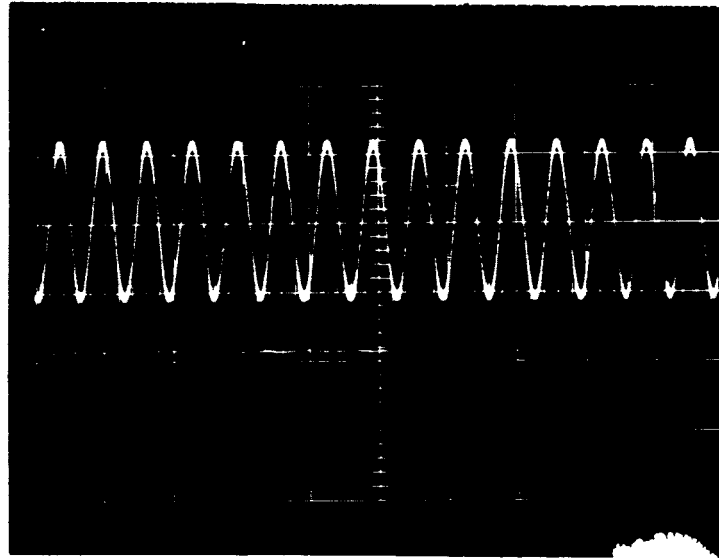
$$f(t) = A \sin \left(\omega_0 t + \frac{\Delta \omega}{\omega_1} \sin \omega_1 + \frac{\Delta \omega}{\omega_2} \sin \omega_2 t \right) \qquad (1)$$

Figure 5. Nomographs T10X7A Receiver Combined Discriminator Characteristics

Normal operation at band center   214.970 MC

Upper Trace: Receiver Video output

Lower Trace: Receiver Limiter output

Test Condition:

Receiver: Nems-Clarke R 1037-A   S/N 355
            (300 KC IF Bandwidth)

Modulation: 7.5 KC rate, 30 KC peak deviation
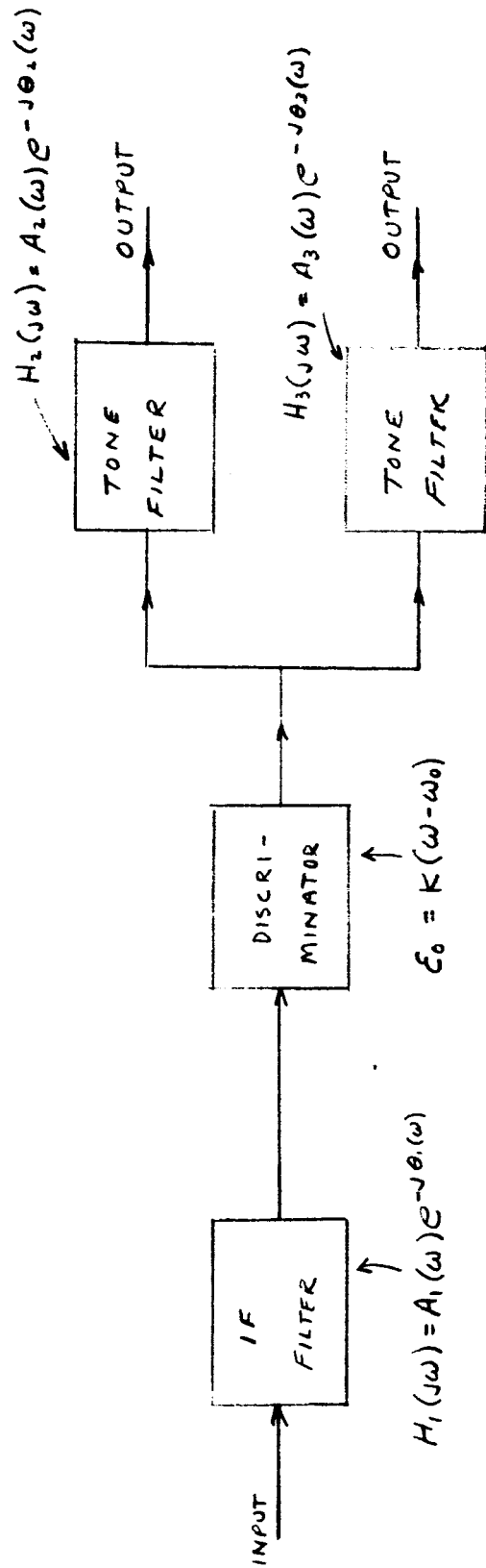
RF Level: 3000   volts

Figure 7.   Normal Operation

Figure 8. Mathematical Model of Receiver

By means of the Fourier integral the function (f (t)) is transformed to the frequency domain.

$$F_1 (j\omega) = \int_{-\infty}^{\infty} f(t) e^{-j\omega t} dt \tag{2}$$

If the IF filter frequency transfer function is $H_1(j\omega)$, then the filter output is

$$G_1 (j\omega) = F_1(j\omega) \, H_1(j\omega) \tag{3}$$

$G_1(j\omega)$ is transformed to the time domain by the inverse Fourier transform

$$g(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} G_1(j\omega) e^{j\omega t} d\omega \tag{4}$$

Examining the argument of the function g(t) the phase $\phi(t)$ may be deduced. The output $\mathcal{E}(t)$ of an ideal discriminator is simply the first derivative of $\phi(t)$ with respect to time

$$\mathcal{E}(t) = K \frac{d\phi(t)}{dt} \tag{5}$$

The discriminator output is then convolved with the time response of the tone filters to obtain the system output.
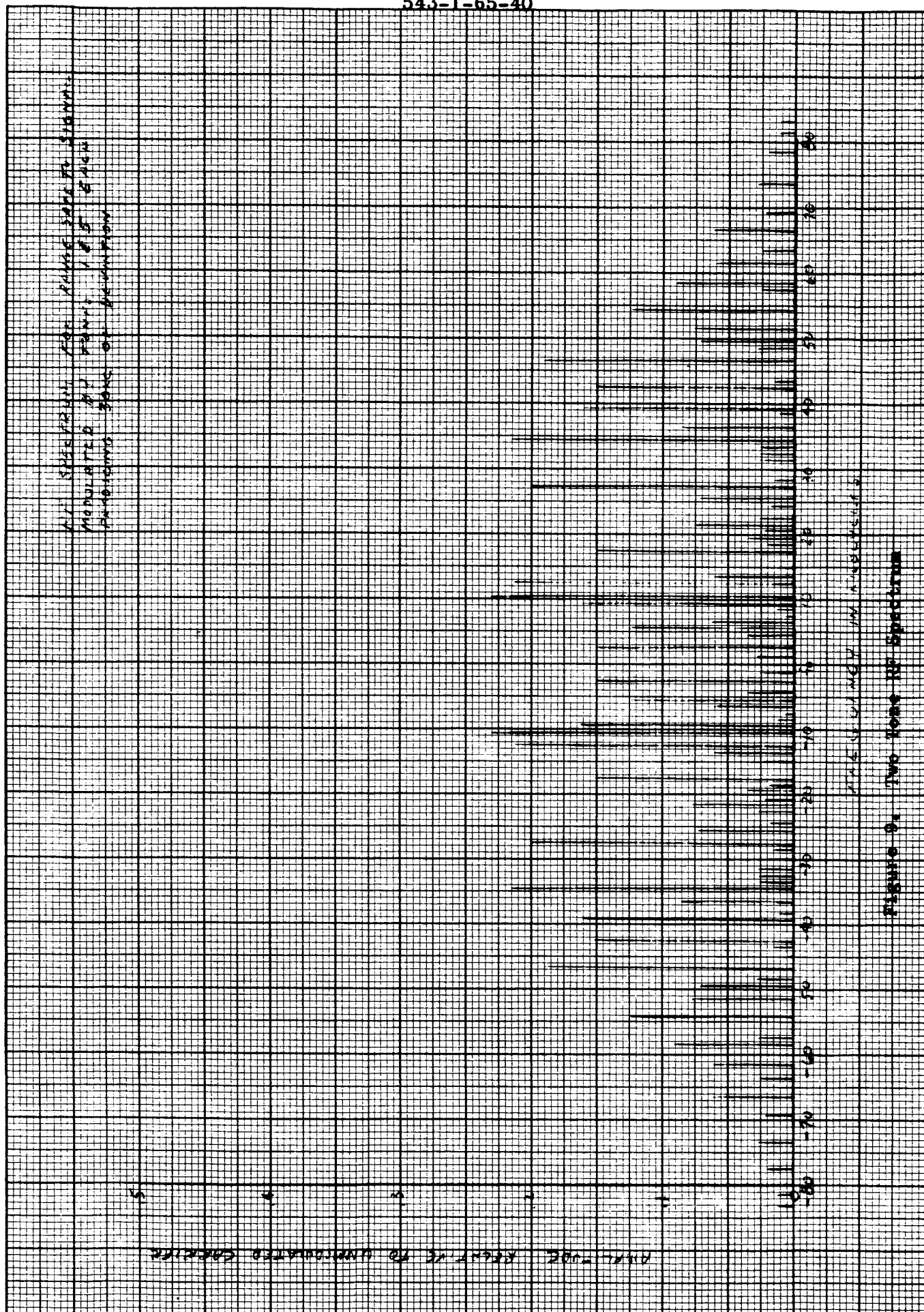
It was immediately apparent that the magnitude of the problem was such that a manual solution was not economically feasible. This

was true even for the simpler case of a single modulating tone. Computer support, therefore, was clearly in order if an analytical solution was to be obtained with a reasonable expenditure of time and money.

First consideration was given to the application of the digital computer for solution of this problem. "Canned" programs are available that accept the Laplace form of the filter transfer function. Getting the input drive function (see equation (1)) into a form appropriate for use as input data to the filter program presented a problem. It was necessary to expand the function into its component parts; this would have entailed a new program. The procedure is to expand the function of equation (1) by the sum of angles then apply the following identities:

$$\cos\left(\frac{\Delta\omega}{\omega_1}\sin\omega_1 t\right) = J_0\left(\frac{\Delta\omega}{\omega_1}\right) + 2\sum_1^\infty J_{2m}\left(\frac{\Delta\omega}{\omega_1}\right)\cos 2m\,\omega_1 t$$

$$\sin\left(\frac{\Delta\omega}{\omega_1}\sin\omega_1 t\right) = 2\sum_1^\infty J_{2m-1}\left(\frac{\Delta\omega}{\omega_1}\right)\sin(2m-1)\omega_1 t$$

(6)

The series are truncated arbitrarily at a point that includes all significant components. Because this would have required a new program and because its general solution was required only once, it appeared feasible to accomplish this by hand. This was actually accomplished and results are shown in Figure 9 where they are evaluated for tones 1 and 5. The discriminator function appeared as another problem for the digital computer. It was finally decided that a reasonable approach would be one that utilized the computer to detect

All frequencies for plane wave FT shown.
Modulated by two tones, 1 & 5 each.
Panoramic 30kc of deviation

frequencies in megacycles

Figure 9. Two Tone RF Spectrum

positive zero crossing of the function; the instantaneous frequency
is then defined as the reciprocal of the period between positive
zero crossings. The instantaneous deviation would then be the
difference between the instantaneous frequency and the carrier fre-
quency. If less than an ideal discriminator function were desired
(a requirement for the problem of interest) the limiting function
must also be included; this appeared relatively easy to accomplish.
Some rough estimations of the required computer time for a solution
indicated the digital approach might be expensive depending upon
the number of steady-state solutions required.

Consideration was then given to the application of the analog computer.
Examination of the requirements for the system simulation on the analog
computer revealed no unsurmountable problems. The solution of
communication problems by use of an analog computer is not new, such
techniques being described as early as 1961.[4,5] It appeared that use of
the analog computer for this problem would be less expensive and more
quickly accomplished than by digital, and it offered additional insight
in understanding the system because of "live" graphical output and on-
line adjustability. This was the computer approach finally utilized;
the simulation is described in the following section and the computer
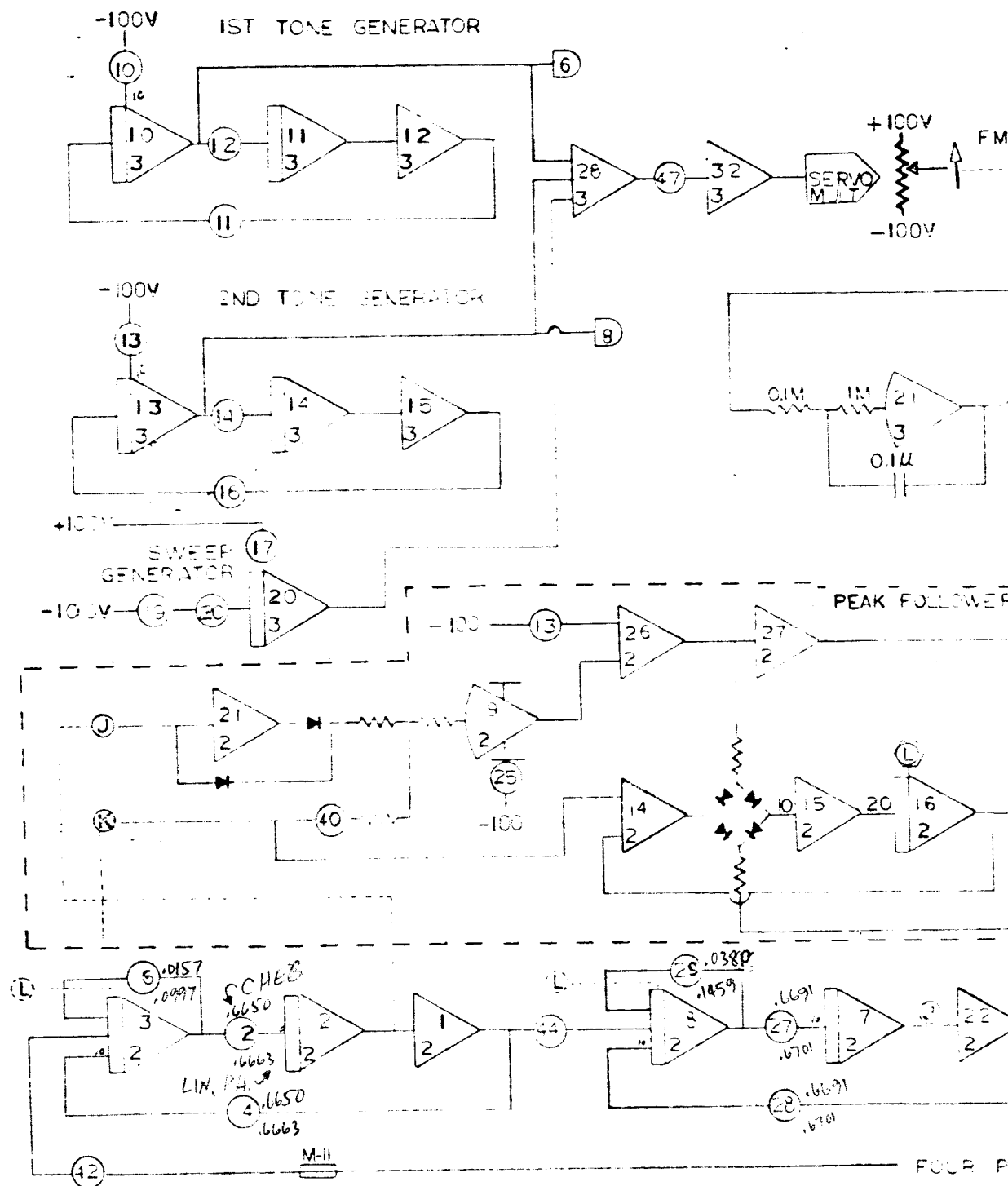solutions are discussed in Section 7.

6. ANALOG COMPUTER SIMULATION

This section describes (1) the method of solution by analog computer, (2) technical details of some of the circuits employed, and (3) instructions for calibration which will assist personnel at another laboratory in putting this simulation on their computer quickly. Figure 10 is a complete block diagram of the system, using conventional analog notation. Dotted lines indicate connections for calibration purposes, and solid lines indicate the basic FM system simulation.

Initial considerations indicated a frequency scaling factor of 10^{-5} or 10^{-6} for analog simulation of this FM receiver. Since frequency scaling by 10^{-5} would require less computer solution time (a factor of 10) this approach was investigated first.

Unsuccessful Attempts to Use Scaling of 10^{-5}: A 107 CPS constant amplitude oscillator capable of being frequency modulated was synthesized first. This amplitude stabilized oscillator was later run at $1/10^{6}$ the frequency and proved to be quite stable at either frequency. A second order system in the form of a three amplifier loop was tried for simulating each set of poles in a Chebishev filter, but the Q adjustments were unstable for large values.

The next attempt used a twin "T" circuit as a quadratic filter (see Figure 11) The general range of desired frequencies were easily obtained and Q's in excess of the maximum Q desired were attainable.
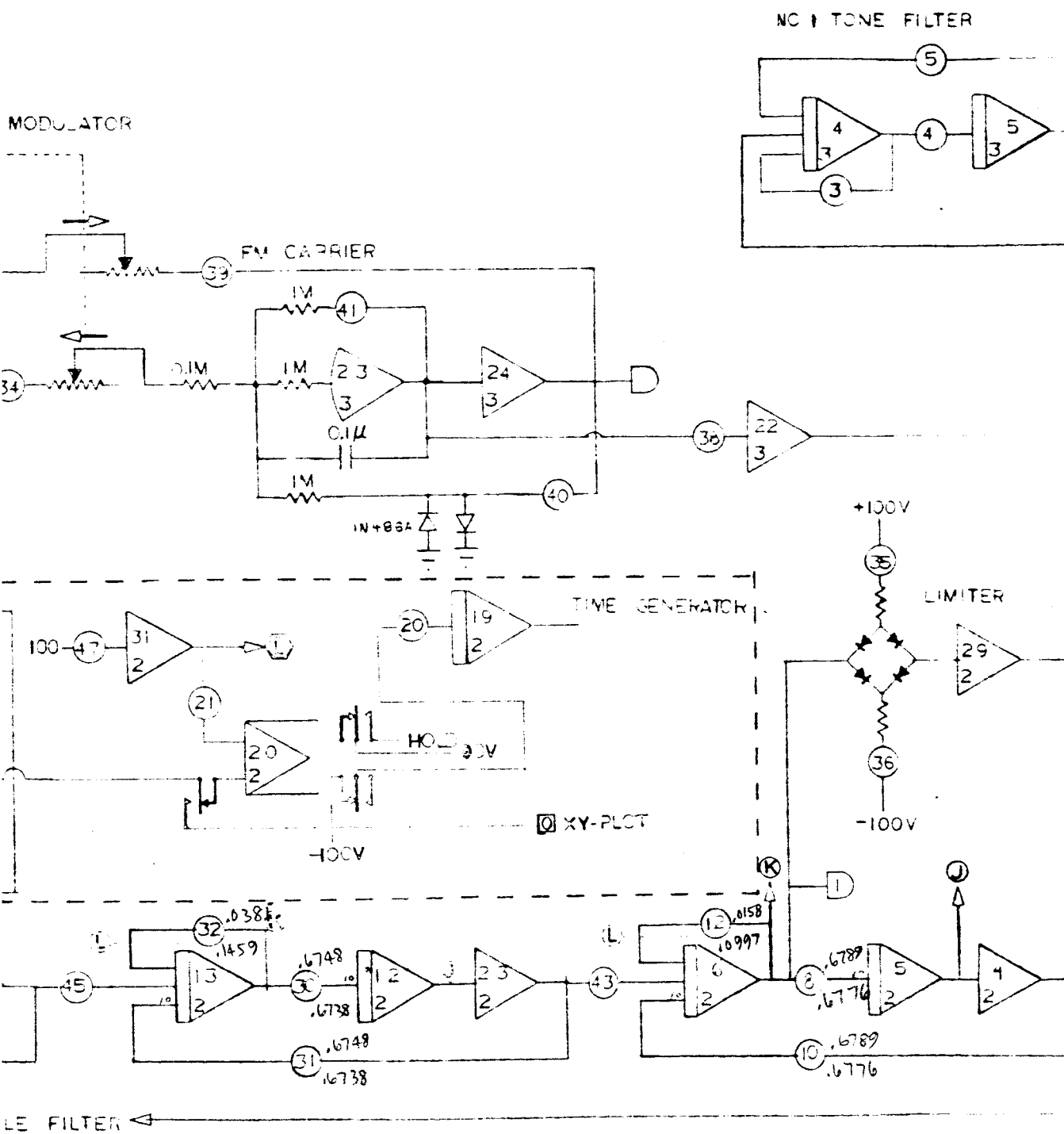
18

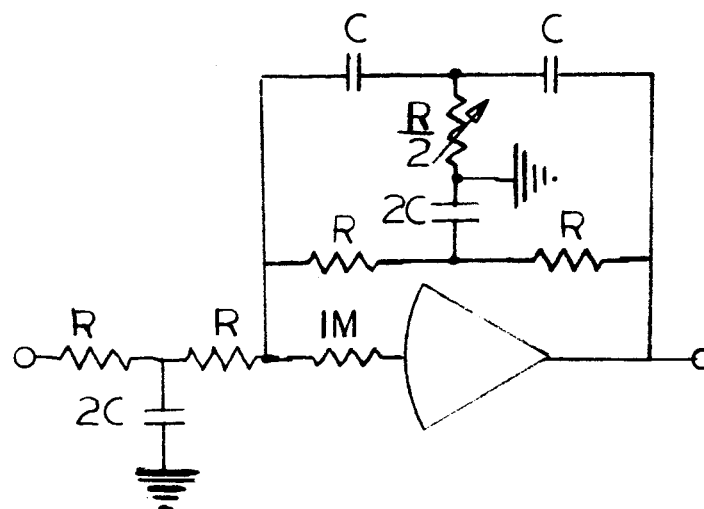19-1

FIGURE 10 ANALOG COMPUTER SIMULATION

NO 2 TONE FILTER

FM DISCRIMINATOR

STATION 2

STATION 3

POT

2 – 6650	25 – 1000	3 – 0500	34 – 7728
4 – 6650	27 – 6691	4 – 0471	38 – 0600
6 – 0157	28 – 6691	5 – 0471	39 – 7728
8 – 6789	29 – 0380	7 – 0500	40 – 8000
10 – 6789	30 – 6748	8 – 0762	41 – 3000
12 – 0159	31 – 6748	9 – 0762	47 – TEXT
13 – 0150	32 – 0331	10 – 0300	
14 – 6092	35 – 0050	11 – 0471	
15 – 0490	36 – 0050	12 – 0471	
16 – 6092	40 – 0020	13 – 0300	
17 – 7348	42 – 1400	14 – 0762	
18 – 0446	43 – 1400	15 – 0762	
19 – 7348	44 – 1400	17 – 2000	
20 – 0100	45 – 1400	19 – 0010	
21 – 5000	47 – 1000	20 – 0005	

19·3

TWIN "T" FILTER

Figure 11. Twin "T" Filter

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A test was conducted by frequency modulating the 107 cps carrier with a single tone generator and then demodulating the filtering to re-obtain the tone. Except for a DC drift out of the demodulator the results were very encouraging.

A four pole Chebishev filter was synthesized using twin "T" networks. The FM carrier was driven with a sweep generator and the output of the filter was amplitude detected. The sweep generator drove the x axis of an x-y recorder and the amplitude detector drove the y axis. The output from this filter circuit was never satisfactory. The following problems caused the termination of this approach: (1) Q and frequency measurements were made with considerable difficulty. (2) Q and fo required too much time to adjust because of interaction between these two parameters. (3) Frequency pulling or shift in frequency was noted when the individual poles of the filter were connected in series. (4) Lack of sufficient isolation between filter stages was detected. (5) Major factor which caused this line of research to be abandoned was the apparent non-linearity which was detected with a change in gain; the cause was undetermined, but the d v/dt limitation of the operational amplifier was suspected.

Success in Using Scaling of 10^{-6}: The next alternative required frequency scaling by 10^{-6} and subsequent changes in the existing circuits. Two computing boards were used because of the number of amplifiers required and the availability of integrators with different RC time constants. Amplifiers in station two had RC time constants of 0.1 sec and station three had one second time constants. With the exception of the carrier

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oscillator all 10 cycle circuits were on station two and all tone circuits were on station three. The exact location may be found by noting the station number posted with the amplifier number on the main schematic in Figure 10.

The peak follower in Figure 10 serves two purposes: 1) calibration of the Q of the filter sections, and 2) peak reading of the filter output to the x, y plotter. It is not used to simulate any function of the FM receiver. The detailed operation of this circuit is given presently. Integrator 19-2 is the time generator.

This time generator should be checked with the EPUT Meter (combination Events per unit time and time interval meter EPUT and TIM). This will eliminate the possibility of an incorrect RC time constant from being perpetuated into other measurements.

The Qs of the individual filters which makes up the four pole filter are adjusted by using a peak follower, a comparator, and a time generator.

The technique is to excite the filter and measure the rate of decay of the sinusoid. The quantity actually measured is the time required for the exponential envelope to decay to one half of its initial value. This time and the circuit Q are related by the easily derived expression

$$t = \left(\frac{\ln 2}{\pi} \right) \frac{Q}{f} = 0.2206 \, \frac{Q}{f} \tag{7}$$

In the circuit the damping is tweeked until the t corresponds to the desired Q. (At the high Q values employed in these circuits the frequency "f" is essentially independent of Q.)

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Output of amp 3 1-2 is the initial condition for the pole being adjusted
and also the output of the peak follower. The comparator has 1/2 of the
initial condition as a bias. When the computer is set to "operate",
the peak follower tracks the output of the filter being adjusted to the
half voltage point where the computer is switched to hold. The output
of the time generator will agree with the computed time when the proper Q
has been set. An initial setting of the pot may be computed from the
relationship $\omega o/Q$ = the pot setting. At the same time the Q is being
tweeked in, the ωo pots may also receive final tweeking.

The TIM is connected to the output of the filter being tested for Q.
The initial condition behaves as if it were an impulse function causing
the filter to ring at the natural frequency. Both frequency and Q
may be adjusted simultaneously.

The peak follower is an instrumentation circuit which is used for cali-
bration and envelope detection.

The peak follower is a sample hold circuit that samples the f(t) term
when it is at a maximum. The circuit is included with the analog com-
puter FM receiver diagram (dashed enclosure). The input labeled "j"
is connected to the filter point whose value is k(f(t))/S. The 1/s
shifts the phase of f(t) by 90° so that when k(f(t)/s is equal to zero,
f(t) is a positive or negative extreme. Pot 40 is used to bias the in-
put to amplifier 9. Amplifier 9 has a cathode follower limiter around it
so that its output can only be negative. If f(t) is negative then
the input labeled "K" is negative and the output of amplifier 9 is

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blocked or zero. If f(t) is positive and k(f(t))/s is equal to zero the output of 9 will generate a pulse. The electronic switch between amplifiers 14 and 15 will close for the duration of the pulse.

Integrator 16 and amplifiers 14 and 15 have a R.C. time constant of 1/2 millisecond, but this value can be made smaller if desired. From the previous text it has been implied that the peak follower is polarity sensitive. This may be eliminated by appropriate use of absolute value circuits.

Pot 47 on station three depends on relationship of voltage to frequency. The servo pot varies the RC time constant of the amplifier by changing the input resistance. A way to determine this pot value requires the multiplier to be set with fixed voltages and the corresponding frequencies. The ratio of the change in voltage to the change in frequency can then be normalized to give .01 cycles per volt of modulation. This normalizing coefficient is the setting for pot 47 in Figure 10.

The following Table may be used to change the Chebishev filter into a linear phase filter.

Pots

2	- .6663	30	- .6738
4	- .6663	31	- .6738
6	- .0997	32	- .1459
27	- .6701	8	- .6776
28	- .6701	10	- .6776
29	- .1459	12	- .0997

The rest of the pot settings should remain the same.

Pots 42, 43, 44, and 45 were changed an equal amount to increase or decrease the gain of the filter. The change in pot value is approximately equal to the fourth root of the change in gain. A linear change in gain at the output can be made by pot 38.

7. COMPUTER SOLUTIONS

The analog computer simulation of the range safety receiver system was accomplished by scaling frequency by a factor of 10^{-6}. The receiver IF frequency is 10.7 megacycles and the computer simulated IF frequency 10.7 cycles.

Prior to making the actual computer "runs" it was necessary to verify that the computer did indeed include the proper transfer functions and that the accuracy of simulation was adequate. The static linearity of the modulator was measured first with the result that the maximum deviation from a straight line through the end points was 2 parts out of 464. The IF filter characteristics were then measured by driving the modulator with a ramp (at a very low sweep rate) and using the peak follower and an X-Y plotter to record the response. Figures 18 and 13 illustrate the results for the Chebishev and linear phase filters respectively. The mathematical description of the filter transfer functions are:

$$\frac{e_o}{e_{in}} = \frac{s}{[s-P_1][s-\bar{P_1}][s-P_2][s-\bar{P_2}][s-P_3][s-\bar{P_3}][s-P_4][s-\bar{P_4}]} \qquad (7)$$

Chebishev:		Linear Phase:	
$P_1 =$	$-.0788 + J66.50$	$P_1 =$	$-.4984 + J66.63$
$\bar{P_1} =$	$-.0788 - J66.50$	$\bar{P_1} =$	$-.4984 - J66.63$
$P_2 =$	$-.1902 + J66.91$	$P_2 =$	$-.7296 + J67.01$
$\bar{P_2} =$	$-.1902 - J66.91$	$\bar{P_2} =$	$-.7296 - J67.01$
$P_3 =$	$-.1902 + J67.48$	$P_3 =$	$-.7296 + J67.38$
$\bar{P_3} =$	$-.1902 - J67.48$	$\bar{P_3} =$	$-.7296 - J67.38$
$P_4 =$	$-.0788 + J67.89$	$P_4 =$	$-.4984 + J67.76$
$\bar{P_4} =$	$-.0788 - J67.89$	$\bar{P_4} =$	$-.4984 - J67.76$
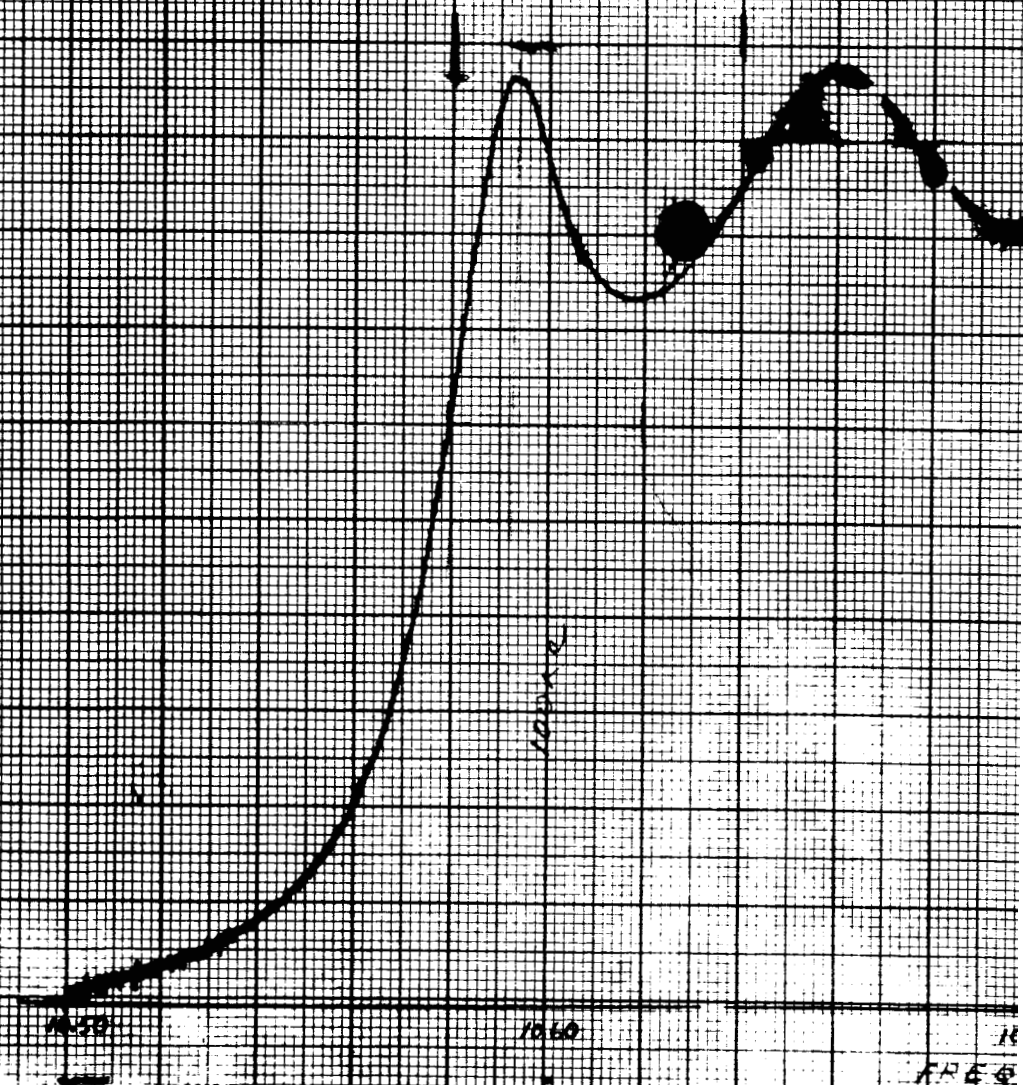
FIGURE 12 CHEBISHEV FILTER
RESPONSE

FIGURE 13 LIN

NCY
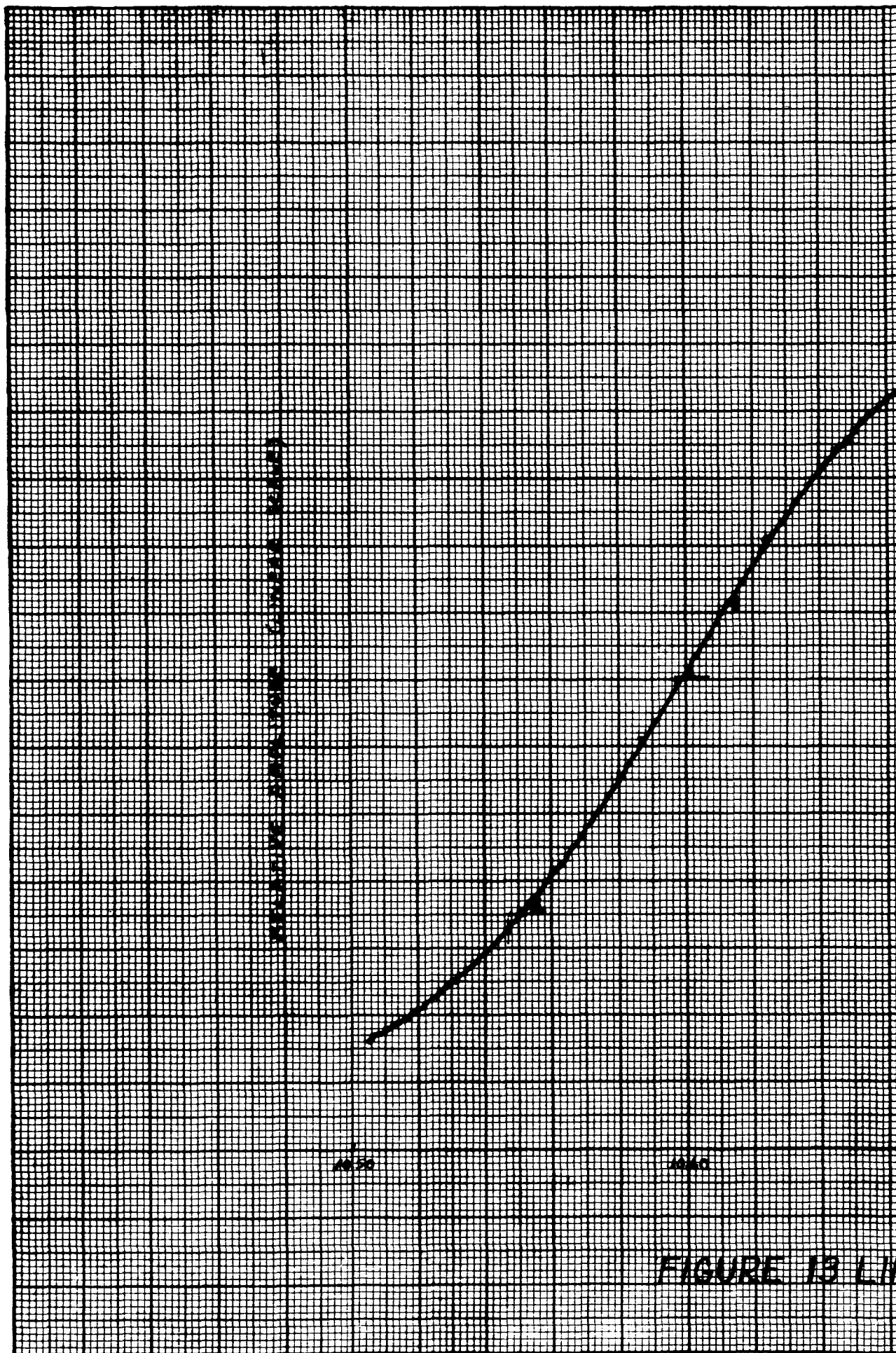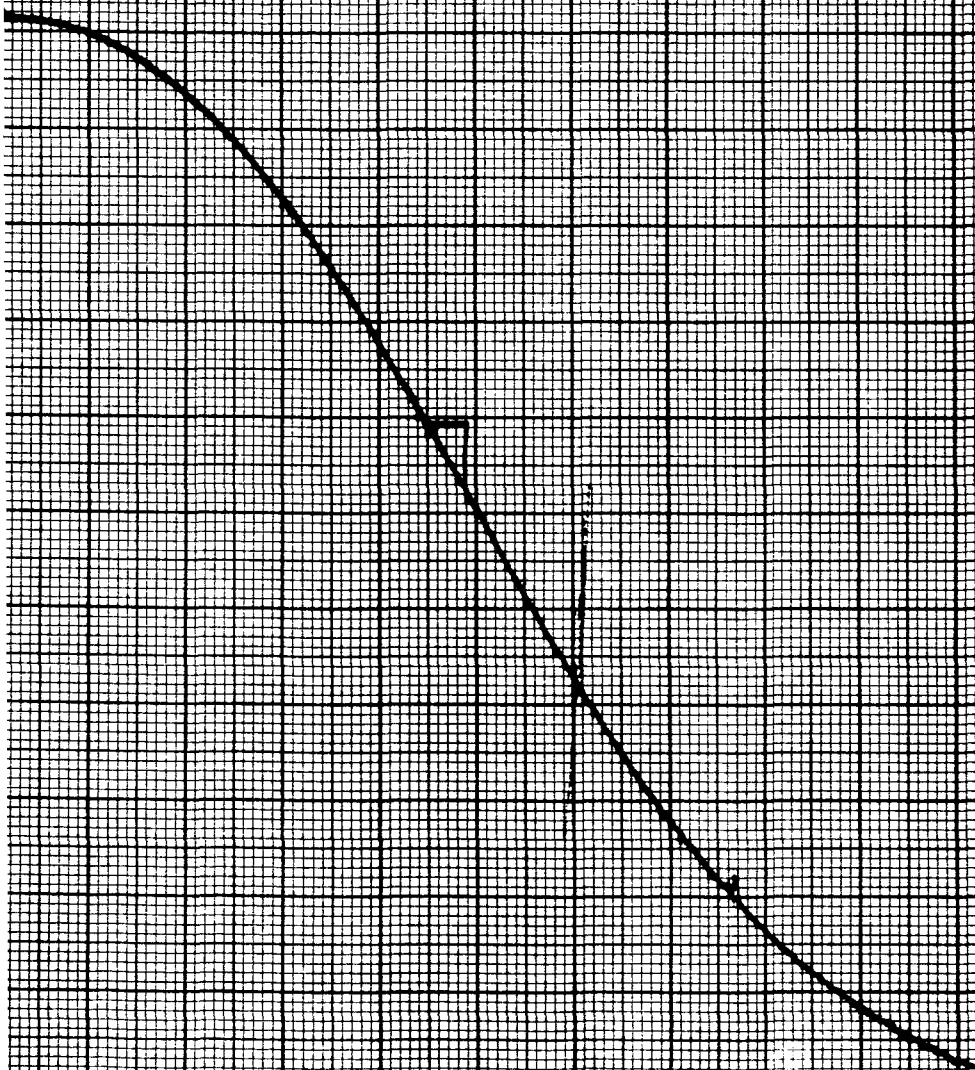
HASE FILTER RESPONSE

The frequency discriminator characteristic was determined also by a sweep of the modulator over the frequency range of interest. Figure 14 is an X-Y slot of the discriminator voltage versus frequency characteristic.

The first case evaluated on the computer was a single tone case. This case is equivalent to the unscaled condition where the signal is modulated by a 7.5 kilocycle sine wave producing a peak deviation of 30 kilocycles. A ramp voltage, summed with the modulation, drove the modulator slowly over the frequency range of interest (each run required approximately two hours). The sweep range was chosen such that the limiter preceding the discriminator dropped below limiting level at both ends of the sweep. Figure 15 is a recording of run results. The output from the IF filters (Chebishev transfer function) is recorded as are the outputs of the limiter, discriminator, ramp generator, tone filter, modulator, and tone oscillator. The modulator output was included in the recording for the purpose of showing that it was free of amplitude modulation. The tone oscillator was included to provide a phase reference for comparing tone filter output. A fairly high level signal was provided for this run (limiter input was 30 volts peak-to-peak at bandcenter) and limiting level was reached quickly. It may be noted from Figure 15 that the tone filter output is large and approximately 180° out of phase with the tone oscillator output when operation is below the limiter level. When the carrier frequency is moved into the pass band where limiting is con-tinuous, the tone filter output is in phase with the tone oscillator. Null conditions occurred on

FIGURE 14 DISCRIM

65-40

1000   1500

NCV

NATOR CHARACTERISTIC

30-2

IF FILTER OUTPUT (CHEBISHEV)
50V/CM

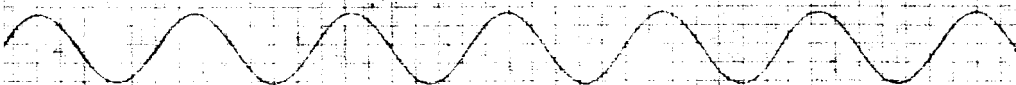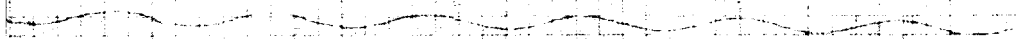LIMITER OUTPUT
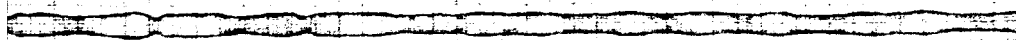1V/CM

DISCRIMINATOR OUTPUT
1V/CM

10.9

CARRIER FREQUENCY

10.5

TONE FILTER OUTPUT

MODULATOR OUTPUT

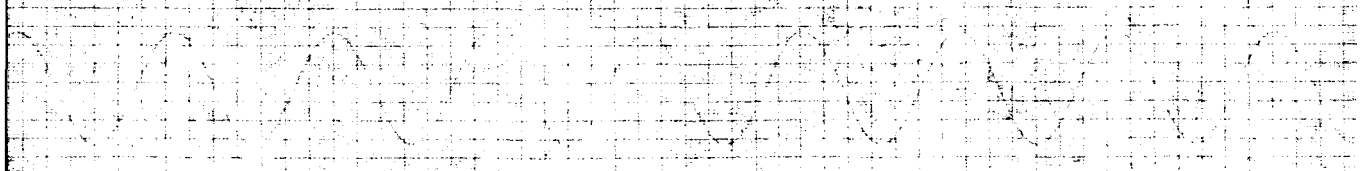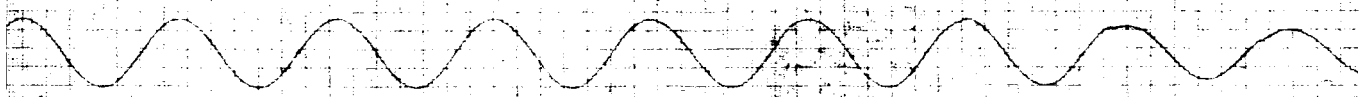TONE OSCILLATOR OUTPUT

IF FILTER OUTPUT (CHEBISHEV)

1-1

31-2

31-3

31-4

31-5
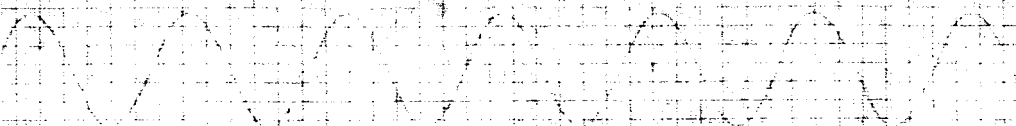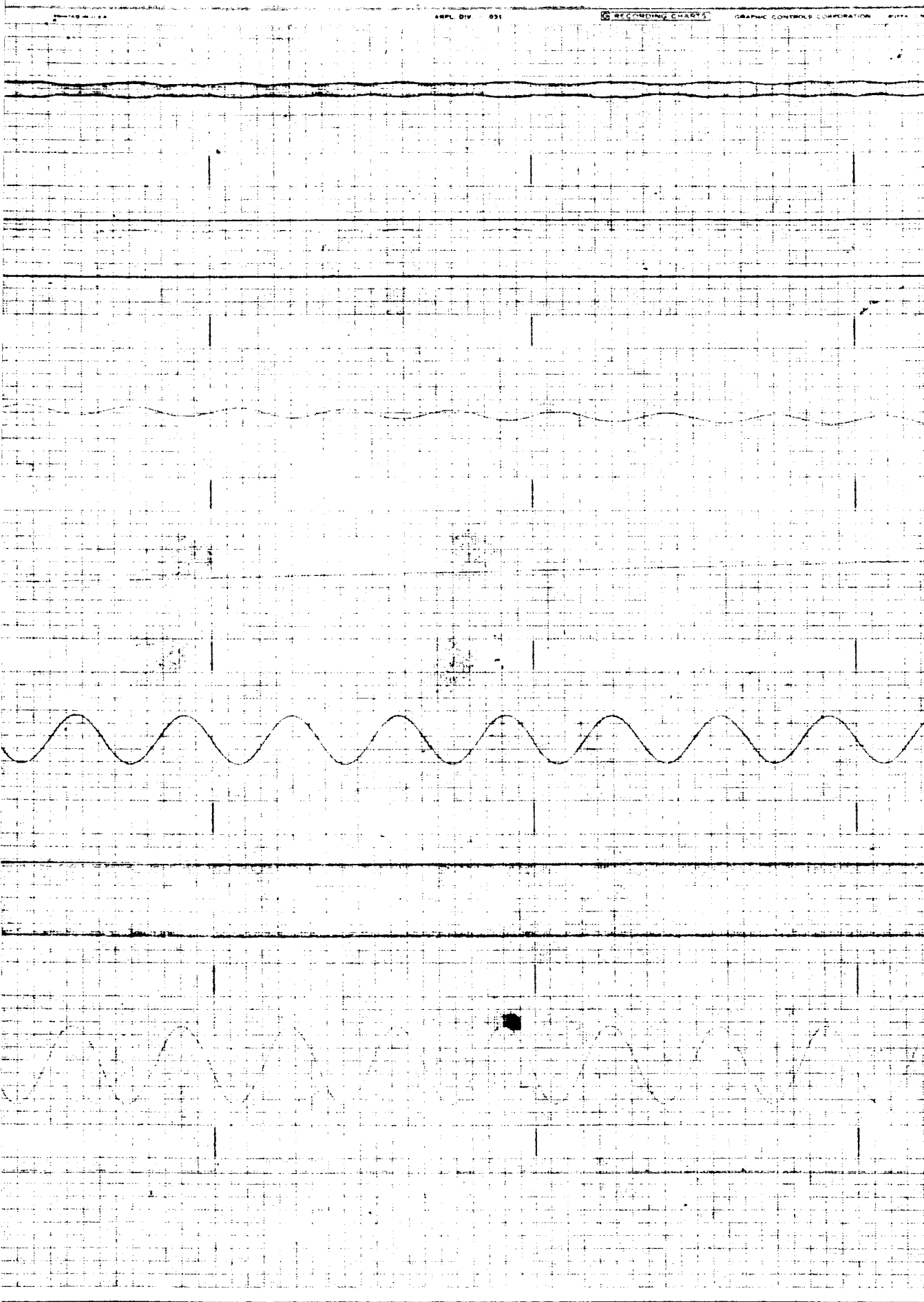
FIGURE 15   COMPUTER SOLUTION FOR SINGLE TONE CASE

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the skirts of the passband and are measured at frequencies of
10.58 cps and 10.86 cps for this case. For proper evaluation one
must consider the storage associated with the tone filters; for
the circuit Q's used, a storage of two cycles is reasonable. Correct-
ing for tone filter storage the null points are more accurately
located at 10.565 cps and 10.835 cps. Another interesting observa-
tion one can make is to note the large signal distortion present
in the discriminator output when operating near the bandedges of the
Chebishev IF filter but at a point where limiting is continuous.
This distortion is attributed to the non-linear phase characteristic
of the filter.

Data of Figure 16 is similar in all respects to the data of Figure
except that the IF filter output (limiter input) is reduced 10 db.
The effect is that the frequency range in which limiting is con-
tinuous is narrowed and the null points are closer together.

The case of two modulating tones was run next. This condition is
equivalent in the unscaled situation to sinusoidal modulating tones
of 7.5 kilocycles and 12.14 kilocycles each producing 30 kilocycles
of deviation. Figure 17 includes data for a high limiter input
level while Figure 18 data has a limiter input 10 db lower. The
modulator output was not included in the run shown in Figure 17
because of the lack of recorder capacity. The higher frequency
tone oscillator output was recorded to allow phase comparisons. To
demonstrate that the modulator was free of amplitude modulations
under two tone modulation it is again recorded in Figure 18 data

32

IF FILTER OUTPUT (CHEBISHEV
.5V/CM

LIMITER OUTPUT
1V/CM

DISCRIMINATOR OUTPUT
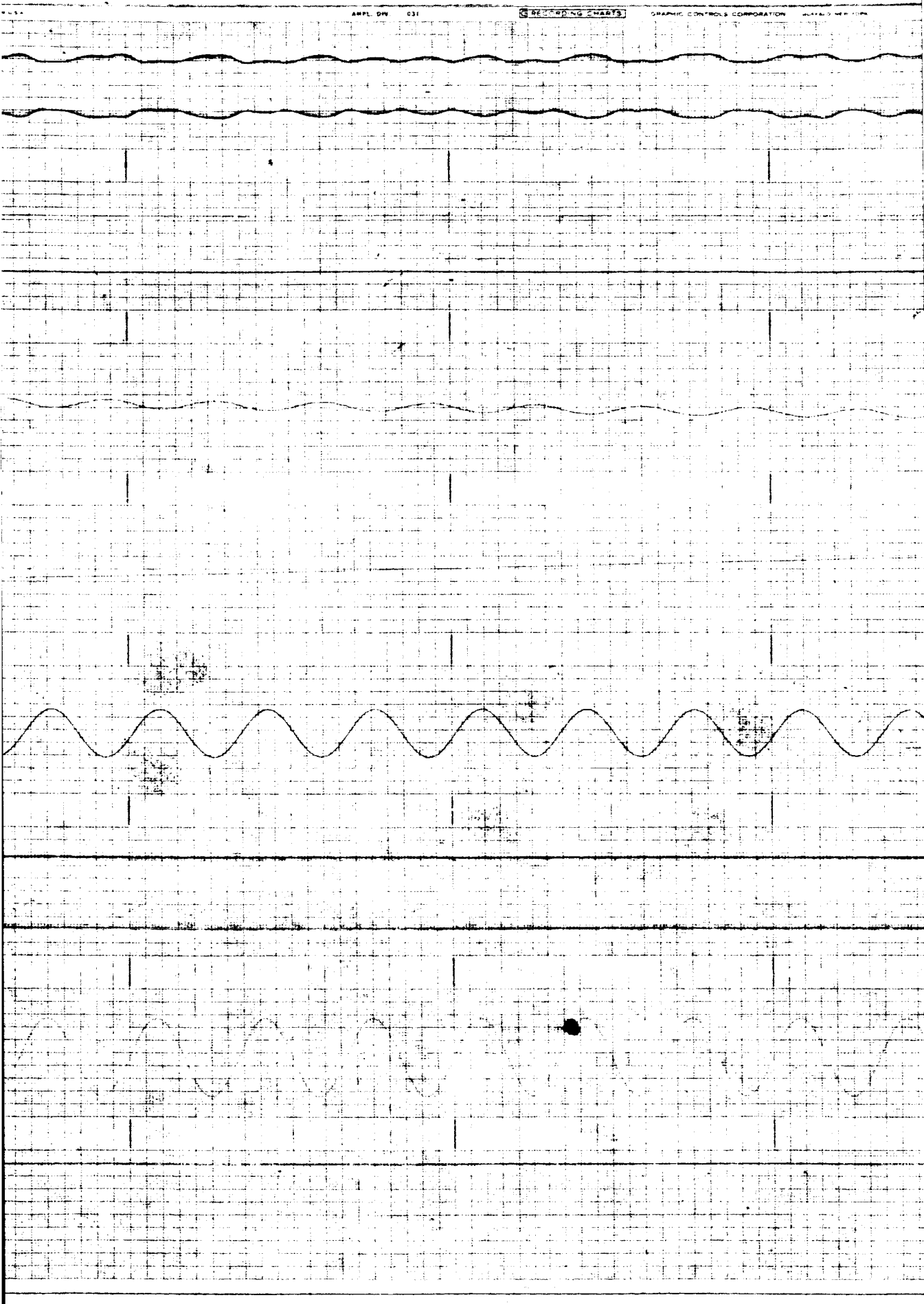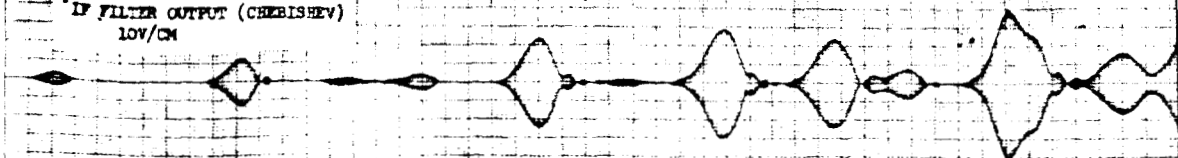1V/CM

10.9

CARRIER FREQUENCY

10.5

TONE FILTER OUTPUT

MODULATOR OUTPUT

TONE OSCILLATOR OUTPUT

33-1

33-2

335

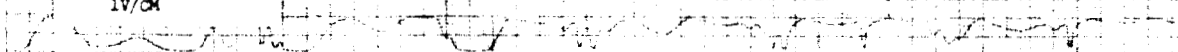FIGURE 16   COMPUTER SOLUTION FOR SINGLE TONE CASE (LOW LEVEL)

33-6

IF FILTER OUTPUT (CHEBISHEV)
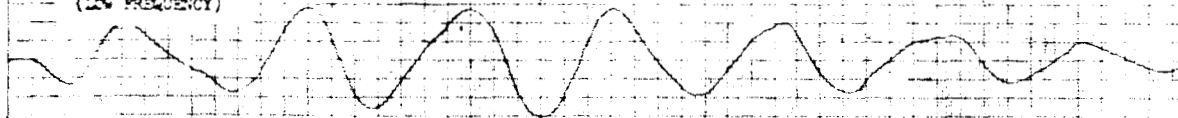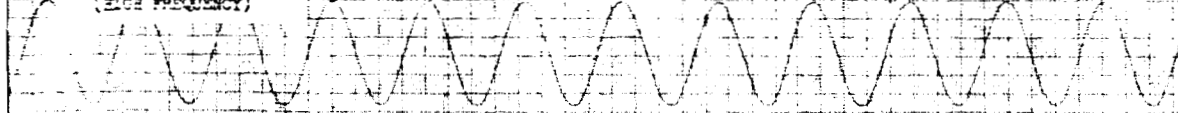10V/CM

LIMITER OUTPUT
1V/CM

DISCRIMINATOR OUTPUT
1V/CM

10.9

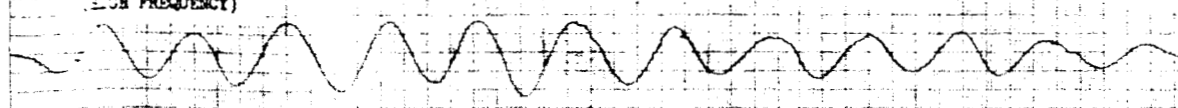CARRIER FREQUENCY

10.5

TONE FILTER OUTPUT
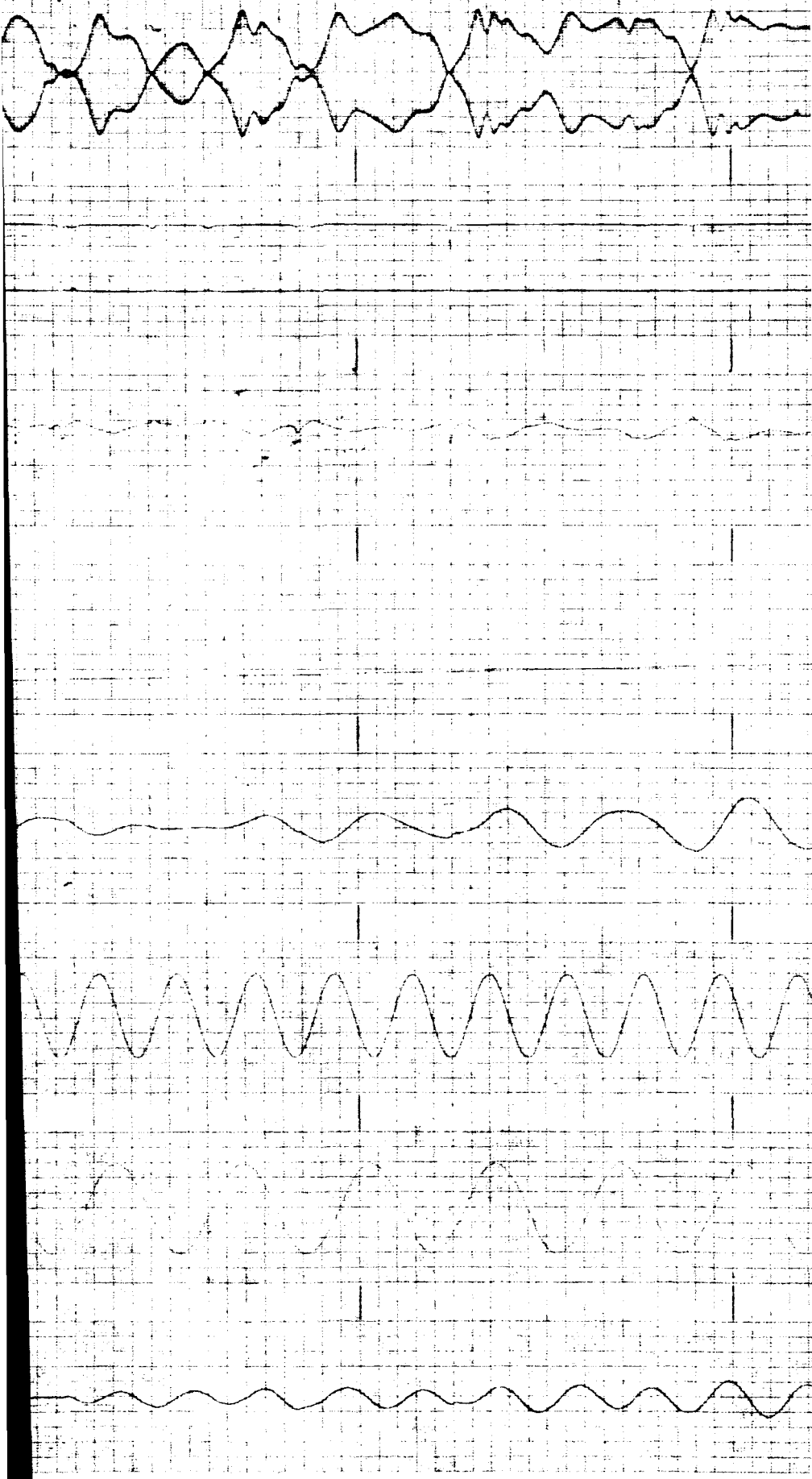(LOW FREQUENCY)

TONE OSCILLATOR OUTPUT
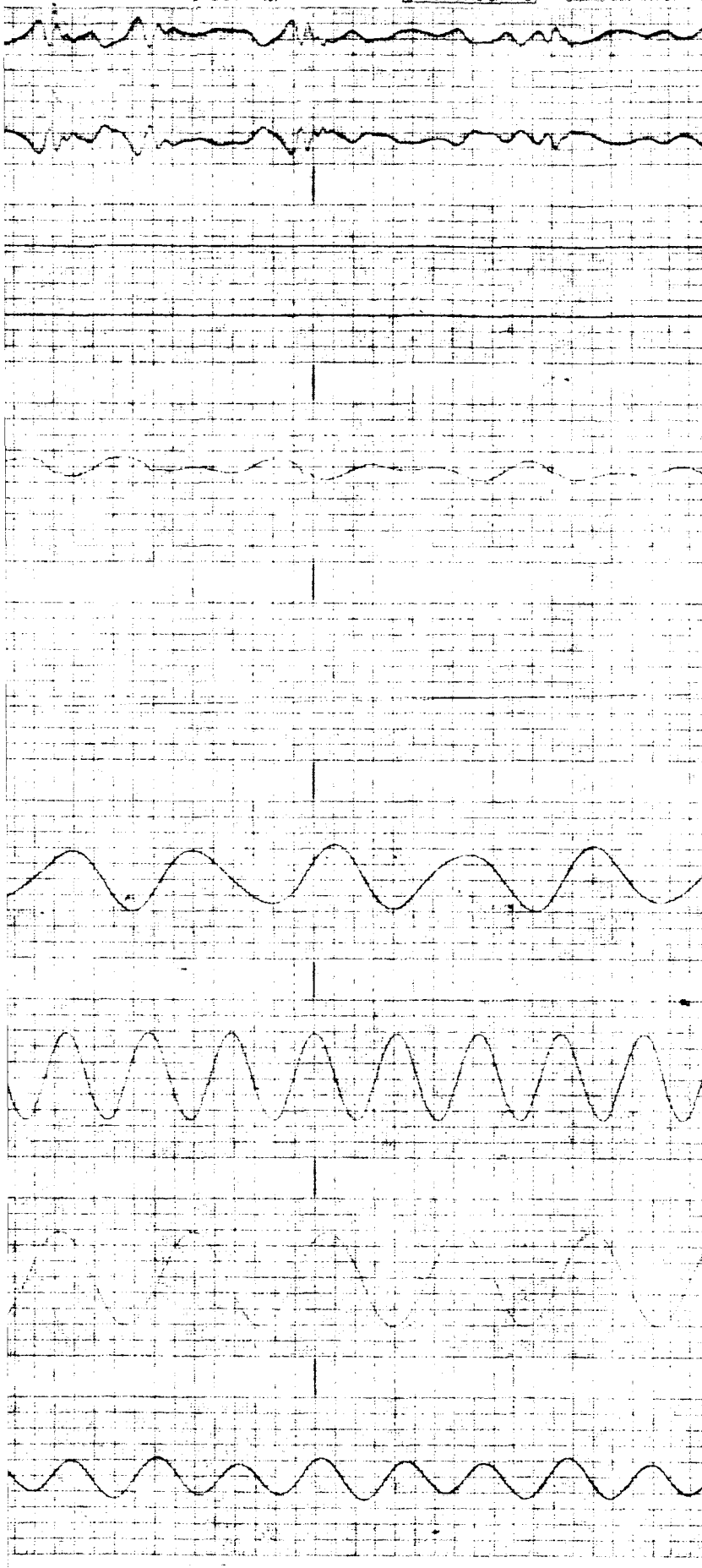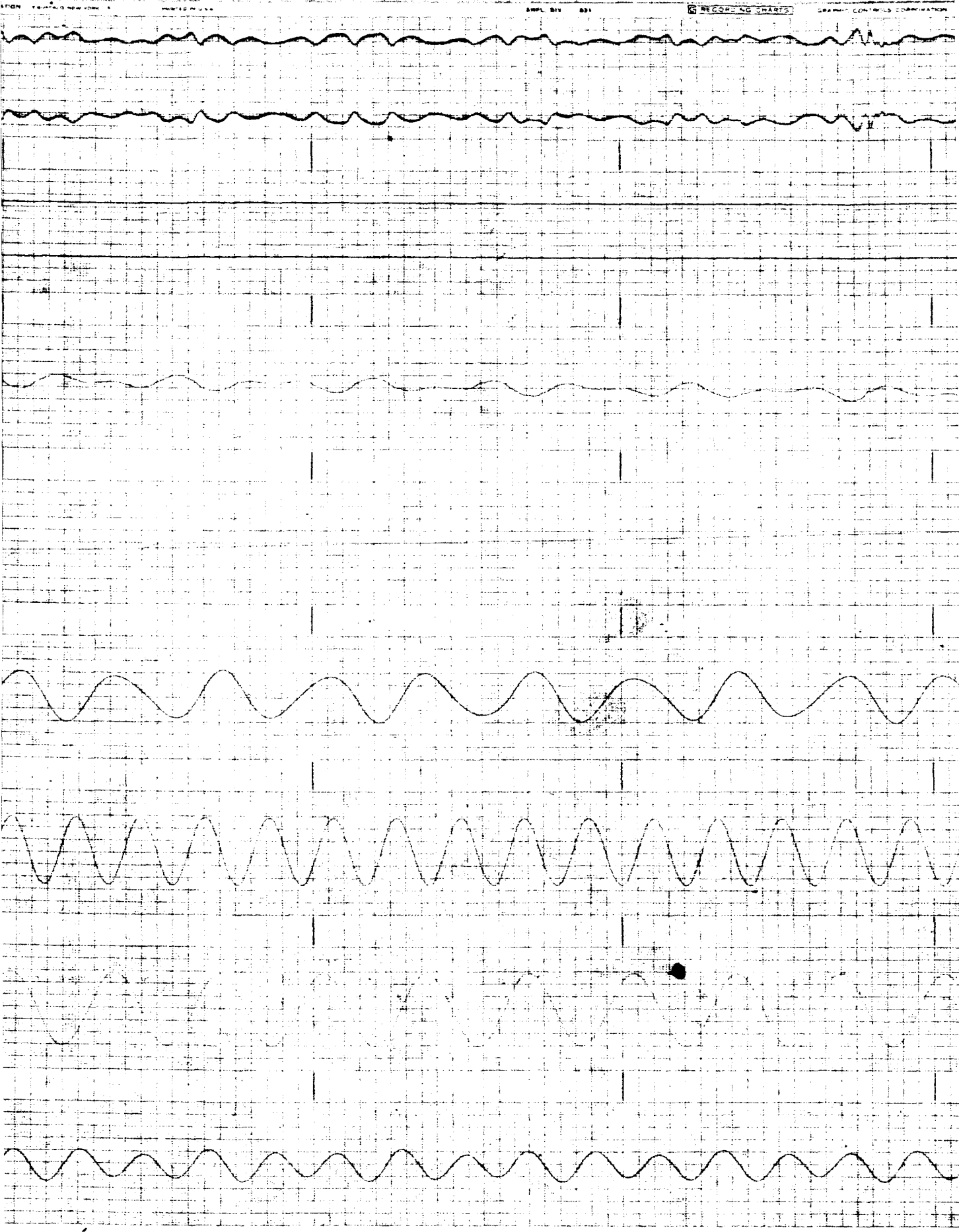(HIGH FREQUENCY)

TONE OSCILLATOR OUTPUT
(LOW FREQUENCY)
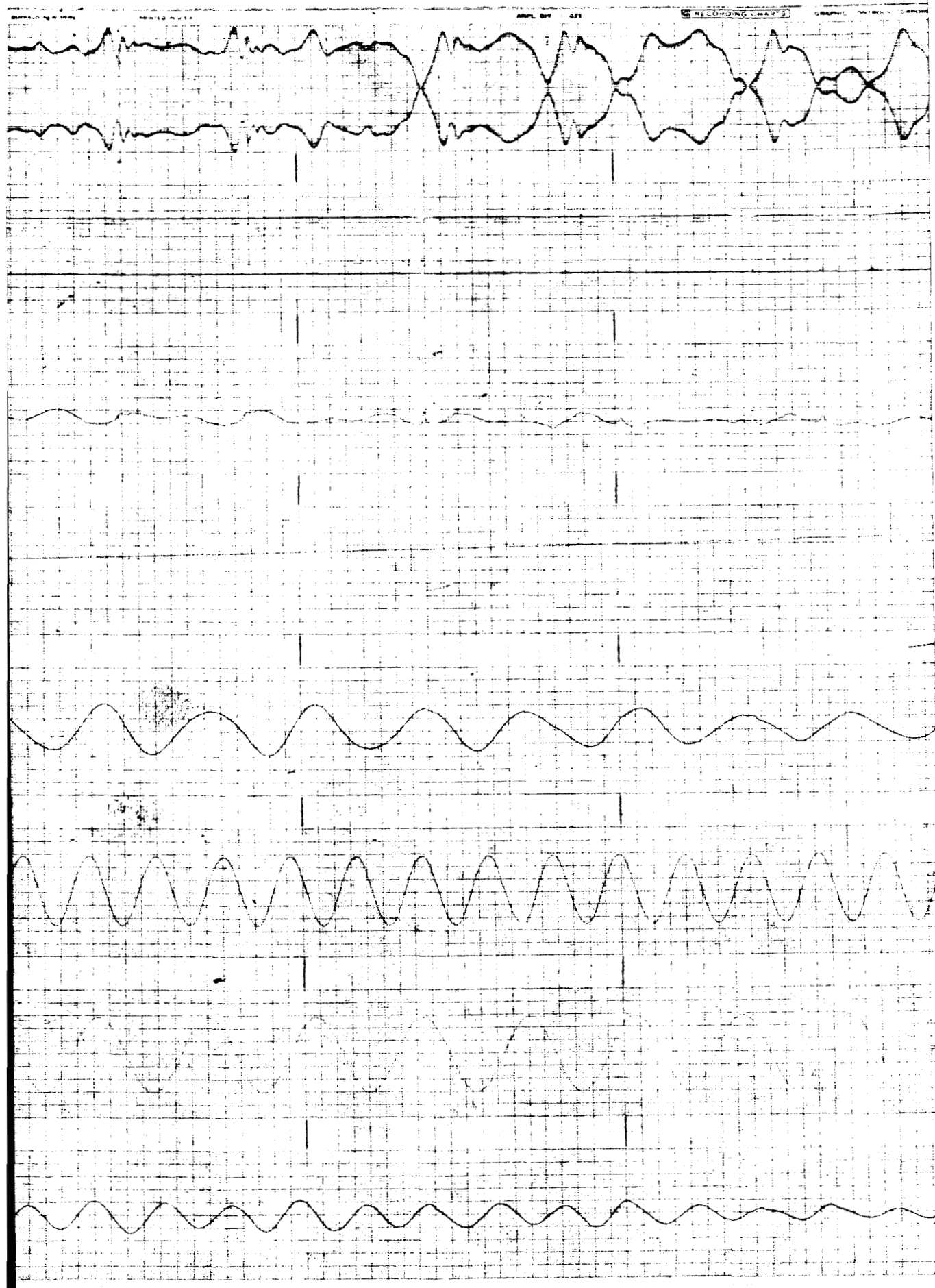
TONE FILTER OUTPUT
(HIGH FREQUENCY)

34-1

34.2

34-3

34-4

34-5

FIGURE 17 COMPUTER SOLUTION FOR TWO TONE CASE

543-1-65-40

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IF FILTER OUTPUT (CEEBISHEV)
5V/CM

LIMITER OUTPUT
.1V/CM

DISCRIMINATOR OUTPUT
1V/CM

10.9

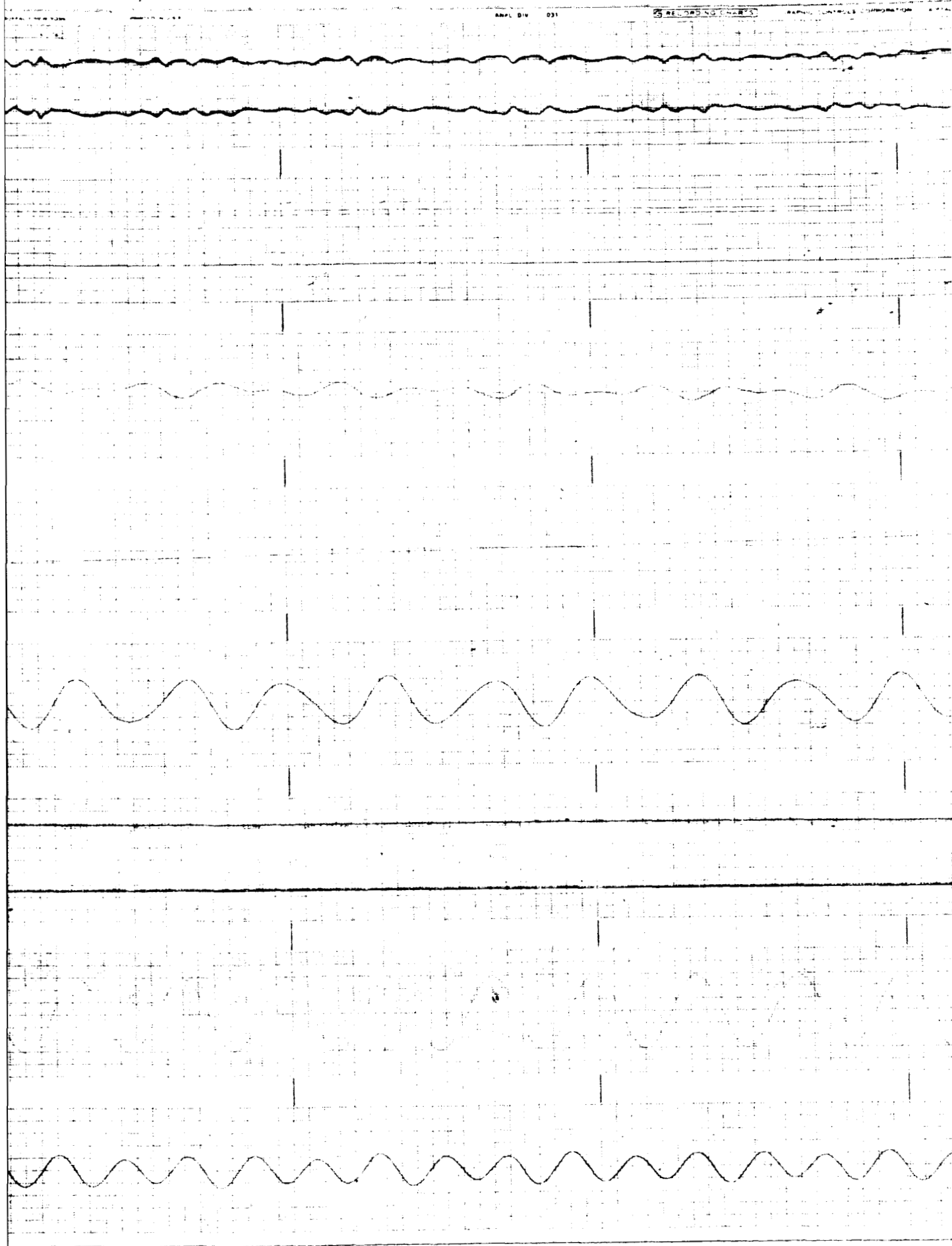CARRIER FREQUENCY

10.5

TONE FILTER OUTPUT
(LOW FREQUENCY)

MODULATOR OUTPUT

TONE OSCILLATOR OUTPUT
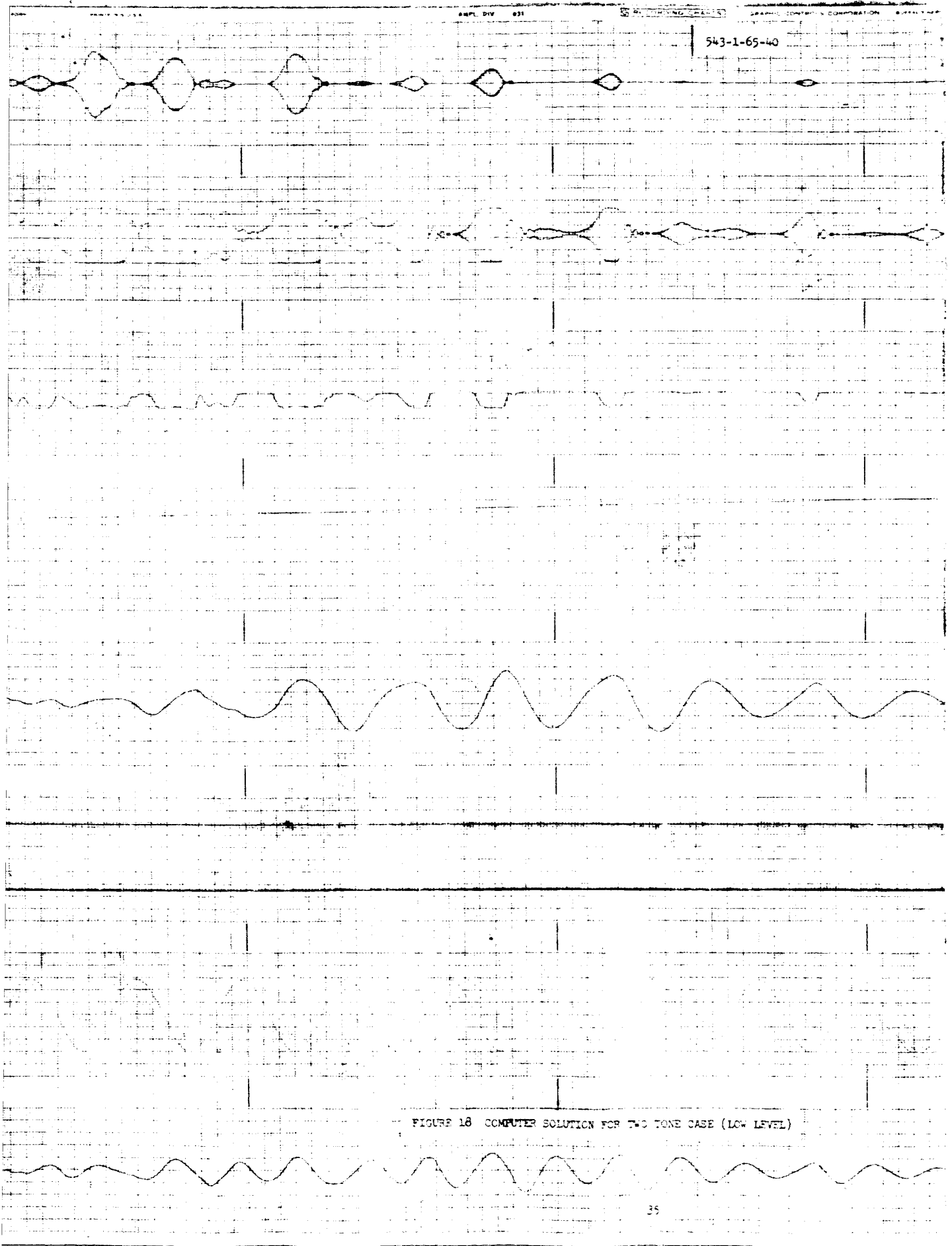(LOW FREQUENCY)

TONE FILTER OUTPUT
(HIGH FREQUENCY)

35-1

35-2

35-3

35-4

543-1-65-40

FIGURE 18 COMPUTER SOLUTION FOR TWO TONE CASE (LOW LEVEL)

35

356

and the high frequency tone oscillator output omitted. Basic data
characteristics are the same as those for the single tone case.
The tone filter outputs are out of phase with the tone oscillator
outputs when operation is below limiting level and in phase in
the center portion of the band where limiting is continuous. The
apparent distortion in the tone filter outputs is the result of cross
talk between channels and is due to the low Q's of the tone band pass
filters which limited separation. These tone filter Q's were chosen
as a compromise to provide reasonable separation yet allow a computer
run in the two hour time period.

A single run was made using the linear phase IF filter characteristic.
This data is included in Figure 19 . The two tone condition was
utilized for this run and filter output was at low level. The
purpose of using the linear phase characteristic was to demonstrate
that the dynamic response observed was not peculiar to the Chebishev
filter. It is readily observed that the tone filter outputs experience
the same phase reversals observed with the Chebishev and null points
are clearly evident. The spacing between null points has increased
slightly because the linear phase filter's skirt slope is not as
steep immediately beyond the pass band. Comparison of the discriminator
output using the Chebishev filter indicates that distortion is not
significant near the bandedges with the linear phase filter.

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IF FILTER OUTPUT (LINEAR PHASE FILTER)
5V/CM

LIMITER OUTPUT
1V/CM

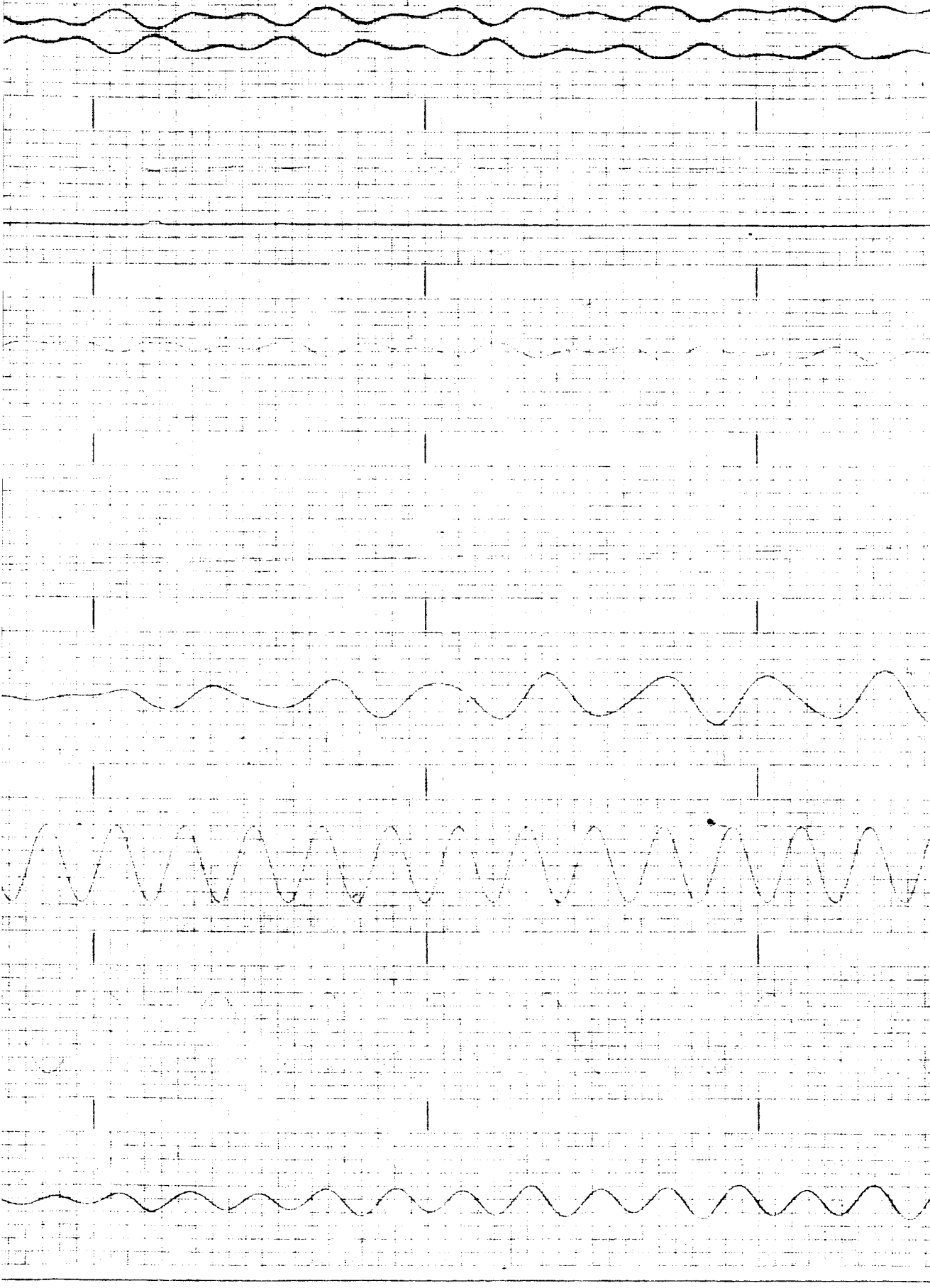DISCRIMINATOR OUTPUT
1V/CM

10.9

CARRIER FREQUENCY

10.5

TONE FILTER OUTPUT
(LOW FREQUENCY)

TONE OSCILLATOR OUTPUT
(HIGH FREQUENCY)

TONE OSCILLATOR OUTPUT
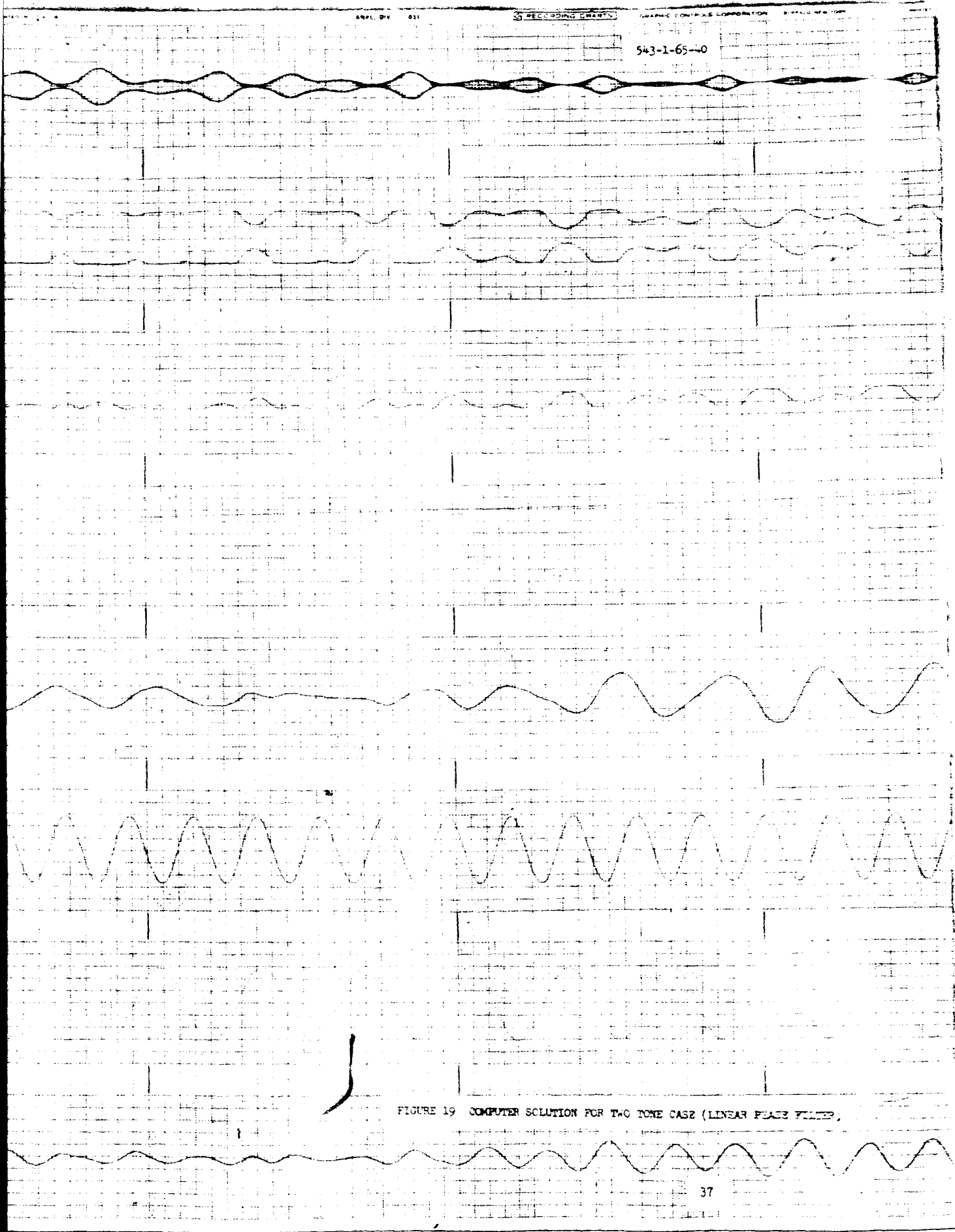(LOW FREQUENCY)

TONE FILTER OUTPUT
(HIGH FREQUENCY)

37-1

37-2

37-4

FIGURE 19   COMPUTER SOLUTION FOR TWO TONE CASE (LINEAR PHASE FILTER)

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CONCLUSIONS

As a result of the arguments and computer results in this report, the following conclusions are reached.

(1) The dynamic response characteristic exhibited by the AVCO range safety receiver is the natural response anticipated from a receiver utilizing a Foster-Seeley discriminator.

(2) The IF filter phase characteristics do not have any significant influence in the generation of the dynamic response observed.

(3) The Chebishev filter introduces significant distortion into the FM system when signal frequency is near the filter band-edges.

ACKNOWLEDGEMENTS

REFERENCES

1. Command Receiver Decoder, Vehicleborne Specification 55-01233-1

2. 80045 XRW 10/964 Handbook of Operating and Maintenance Instructions
    Command Receiver Decoder, Vehicleborne Part No. 185010-1 thru -151
    AVCO Corp. Electronics Division, Cincinnati, Ohio 45241 10 September 1964

3. Electrical Engineers Handbook Fourth Edition P8-29 Pender and
    McIlwan

4. Proceedings of the Western Joint Computer Conference May 9-11,1961
    p. 490

5. Missiles and Space December 1961 by John M.Lambert "GE's Electronic
    System Evaluator."

6. Simulation January 1964, p. 31

APPENDIX I

Simulation of Frequency Modulation by Use of an Analog Computer

(Dr. A. N. Wilson)
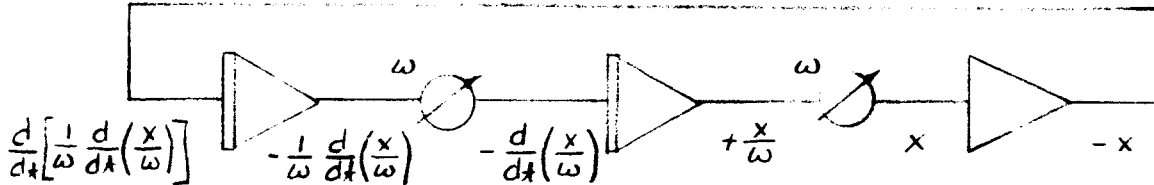
Assuming carrier frequency ω_0, modulation frequency p, and the desired deviation $\Delta\omega$, the necessary instantaneous frequency of the resonant element is $\omega = \omega_c + \Delta\omega \cos$ pt.

The analog loop gain above is ω^2 (amplifier gains are assured to be unity, for simplicity). The instantaneous oscillator frequency is thus $\sqrt{\omega^2} = \omega$, which may be identified by analogy with $1/\sqrt{LC}$ of the conventional tank circuit used in a frequency modulator. This analogy should constitute sufficient "proof" of proper simulation of FM on the analog computer.

An alternate "proof" consists in showing $X = A \sin (\omega_0 t + \frac{\Delta\omega}{p} \sin pt)$ satisfies the differential equation solved by the analog loop.

$$\frac{d}{dt}\left(\frac{x}{\omega}\right) = \frac{d}{dt}\left[\frac{A \sin (\omega_0 t + \frac{\Delta\omega}{p} \sin pt)}{\omega_a + \Delta\omega \cos pt}\right]$$

$$= A \left[\cos (\omega_0 t + \frac{\Delta\omega}{p} \sin pt) + \frac{p\Delta\omega \sin pt \sin (\omega_0 t + \frac{\Delta\omega}{p} \sin pt)}{(\omega_0 + \Delta\omega \cos pt)^2}\right]$$

Since $\frac{\Delta\omega}{\omega_c} << 1$ and $\frac{P\Delta\omega}{\omega_o^2} << 1$, the second term may be neglected, Thus

$$\frac{d}{dt}\left(\frac{x}{\omega}\right) \approx A\cos\left(\omega_o t + \frac{\Delta\omega}{P}\sin pt\right)$$

Similarly we can show

$$\frac{d}{dt}\left[\frac{1}{\omega}\frac{d}{dt}\left(\frac{x}{\omega}\right)\right] \approx -A\sin\left(\omega_o t + \frac{\Delta\omega}{P}\sin pt\right) = -x \qquad Q.E.D.$$

The approximation employed here is exactly the same as used in showing
a conventional modulator produces FM as defined above. One makes a
binomial expansion approximation on $\sqrt{C} = \sqrt{C_o + \Delta C}$.
(See Seely "Electron Tube Circuits" page 377, or Terman, or numerous
other references.)

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