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FM RECEIVER

DYNAMIC RESPONSE

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(This work was performed at the request of the Centaur RF Systems Group Dept 963-6 and funded by WAP K200594)

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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 presented. 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 application of the analog computer to the problem is outlined and the analog simulation of the receiver system described. Computer solutions 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 <u>natural behavior</u> 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 unusual 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.

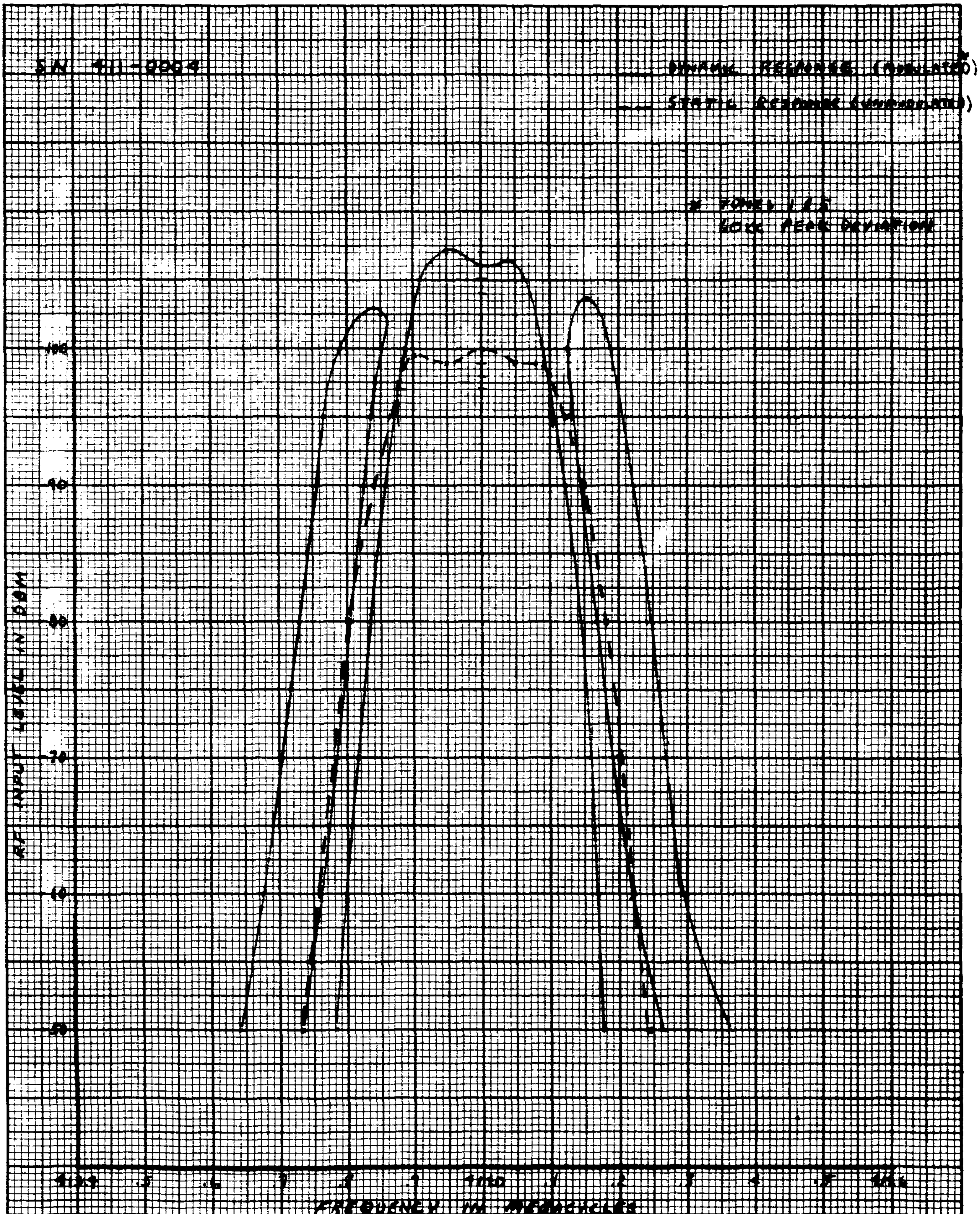


Figure 1. Typical Observed Response

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.

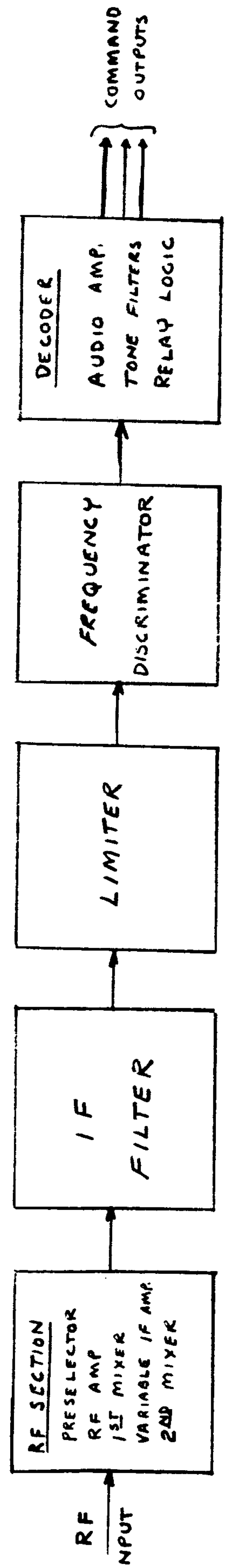


FIGURE 2. Simplified Block Diagram of the FM Command Receiver

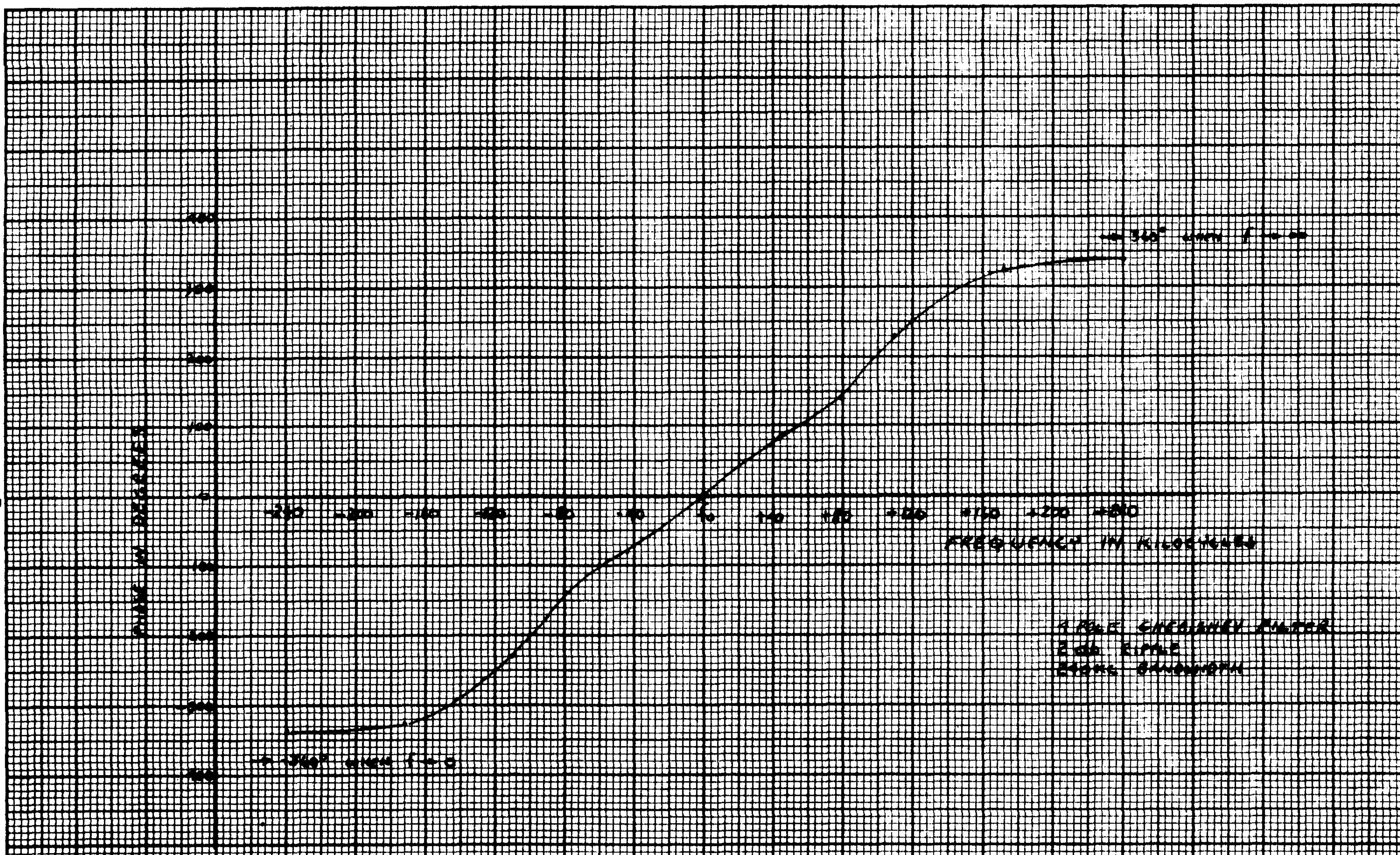
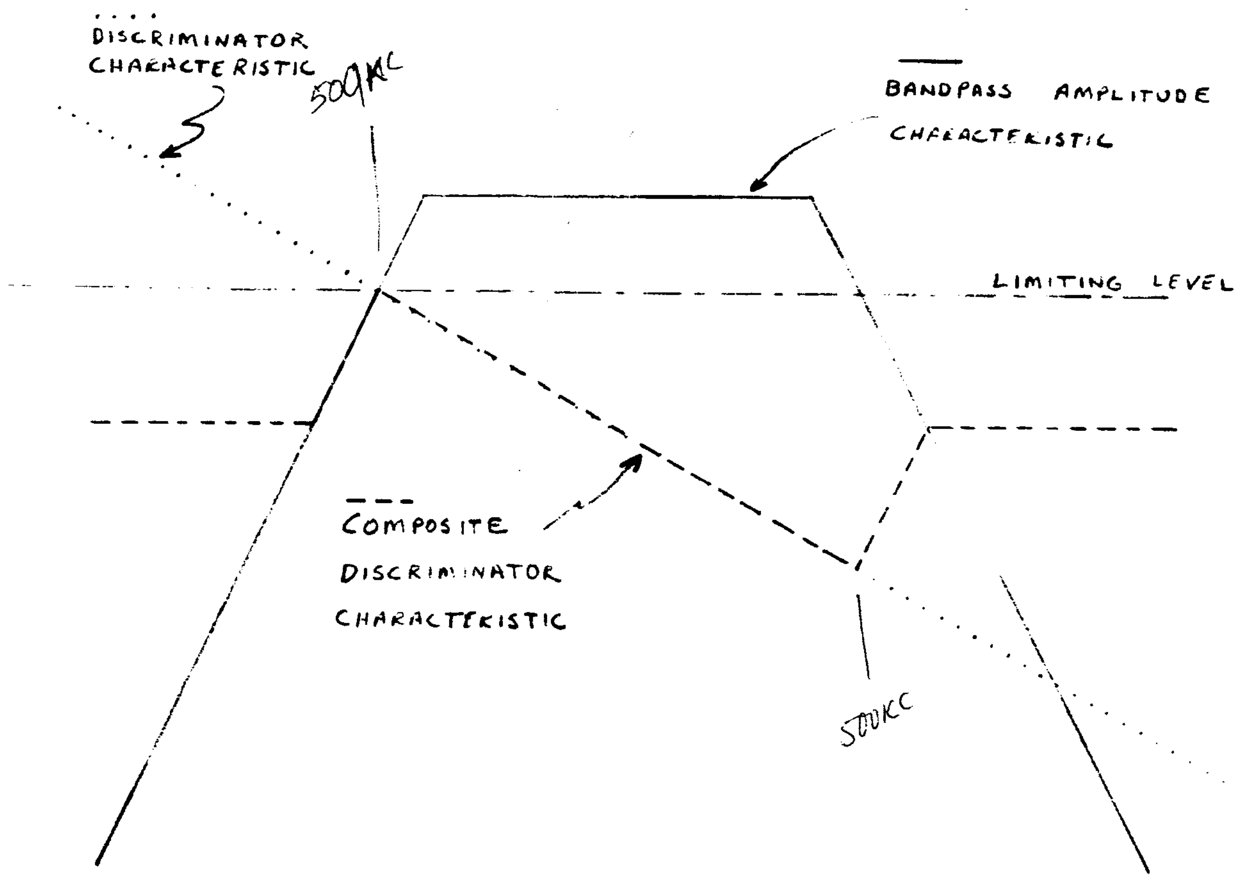


Figure 3. Chebishev IF Filter Phase Characteristic

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 composite 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 discrimination characteristic for a hypothetical receiver to aid discussion. 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



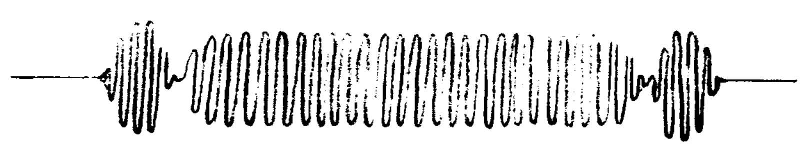


Figure 4. Frequency Characteristics

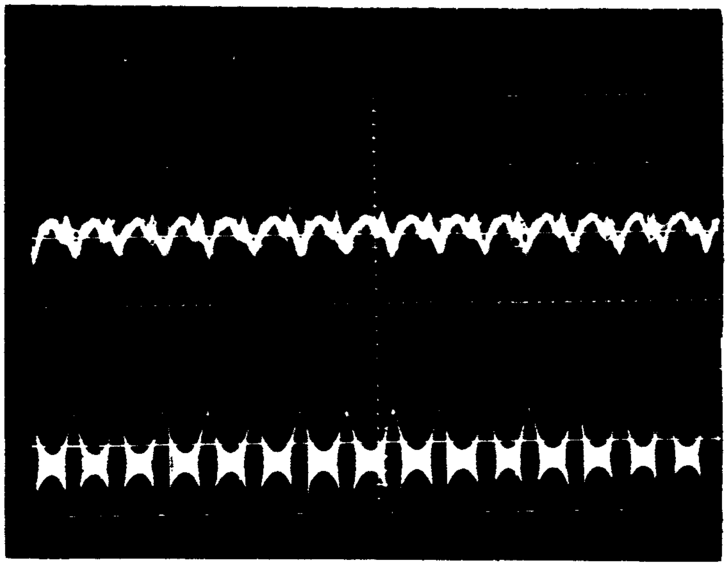
peaks where the demodulated output signal is cancelled and any resultant is a small amount of second harmonic. There are two positions, one on each side of the passband beyond the cancellation points, where the discriminator output signal again increases but is of opposite sign due to operating on the 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.

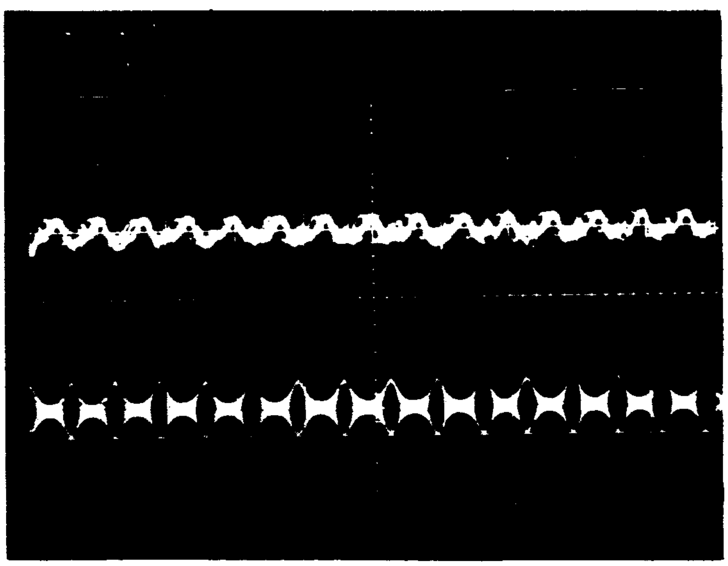
Low Frequency Null 214.668 MC



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 straightforward. 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.

$$f(t) = A \sin\left(\omega_0 t + \frac{\Delta\omega}{\omega_1} \sin \omega_1 + \frac{\Delta\omega}{\omega_2} \sin \omega_2 t\right) \tag{1}$$

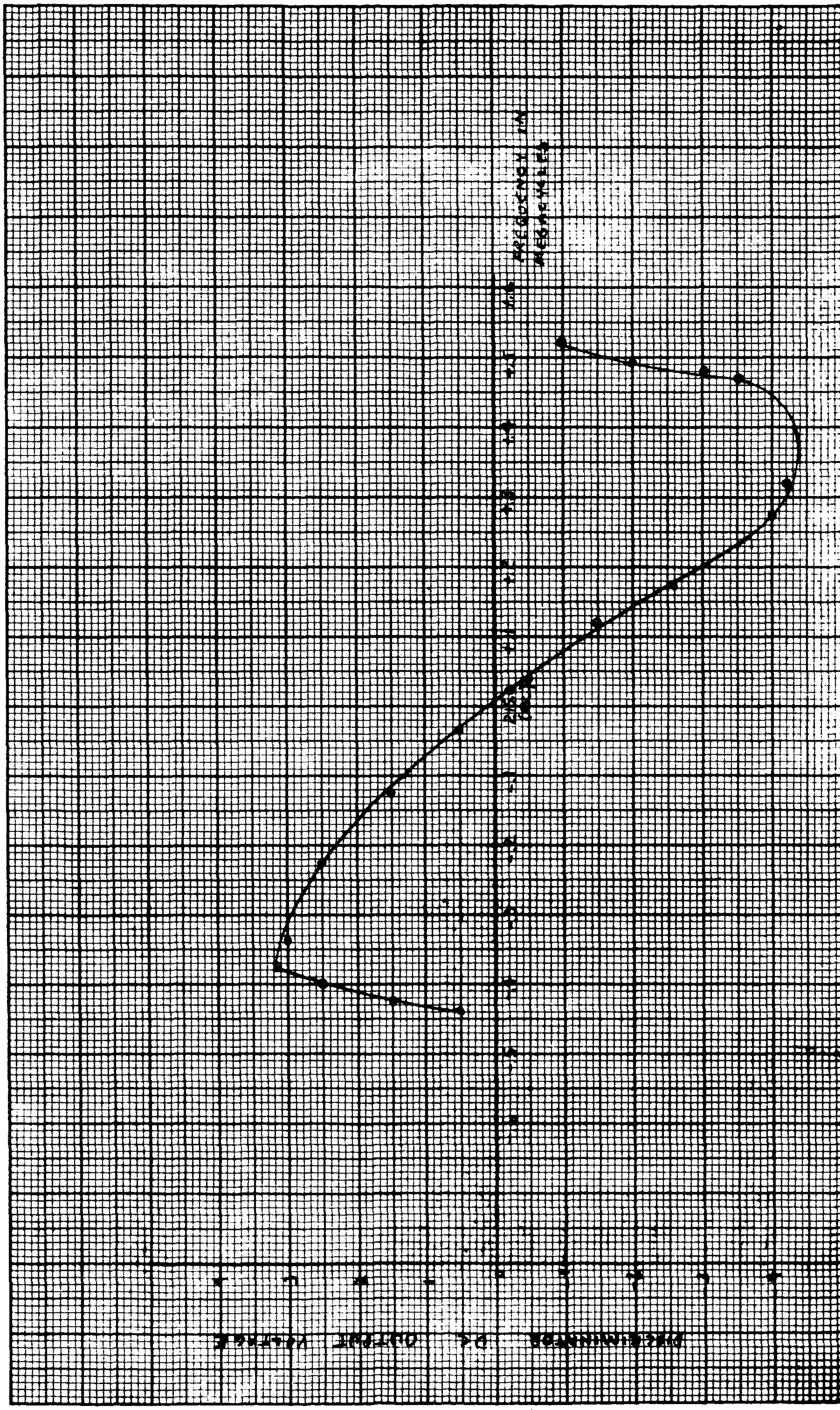
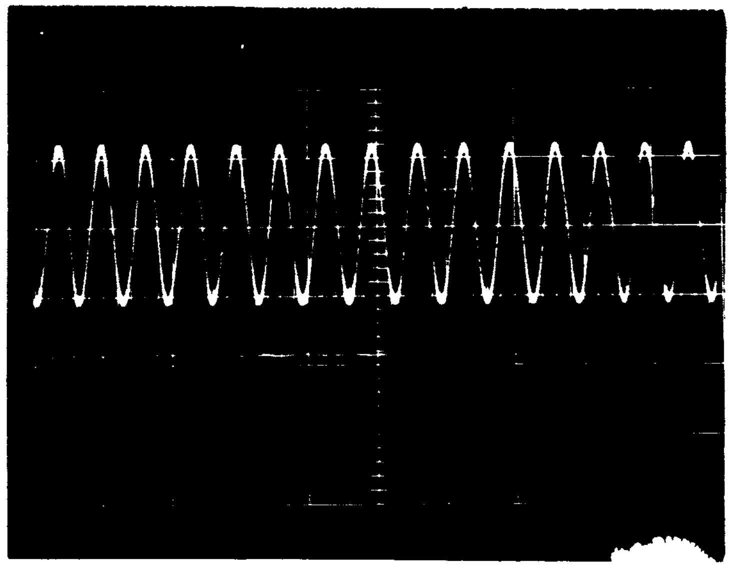


Figure 5. Nems-Clarke R1037A Receiver Composite Discriminator Characteristics



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

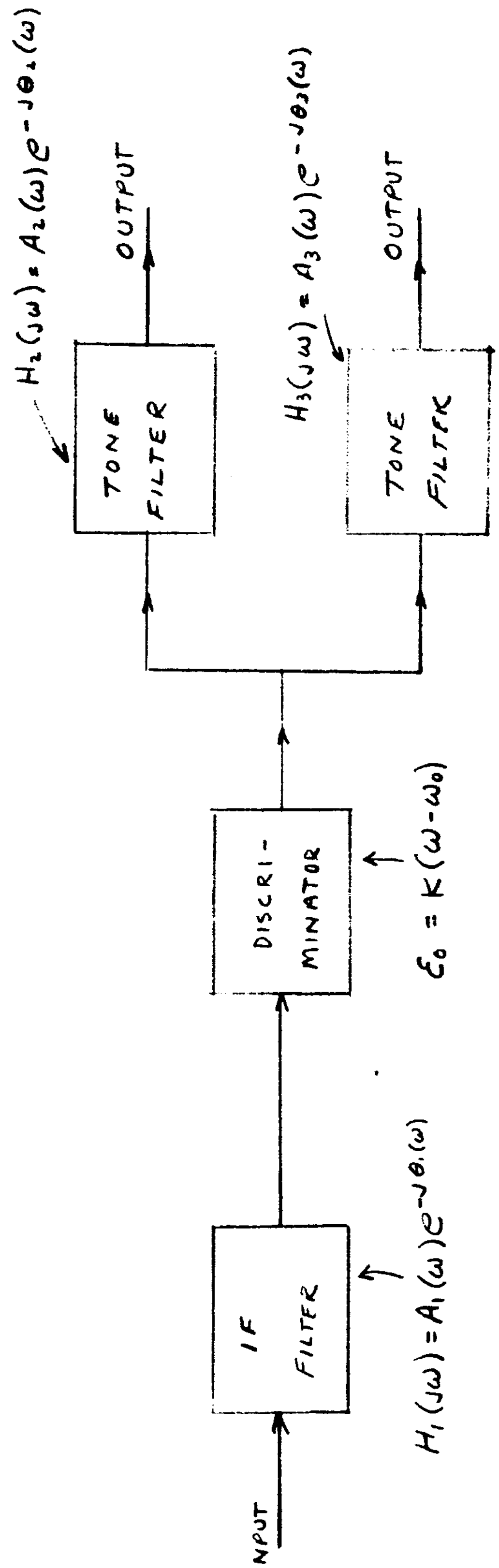


Figure 8. Mathematical Model of Receiver

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

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

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

$$G_1(j\omega) = F_1(j\omega)\, H_1(j\omega) \qquad (3)$$

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

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

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

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

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:

$$\cos\left(\frac{\Delta\omega}{\omega_1}\sin\omega_1 t\right) = J_0\left(\frac{\Delta\omega}{\omega_1}\right) + 2\sum_{1}^{\infty} J_{2m}\left(\frac{\Delta\omega}{\omega_1}\right)\cos 2m\,\omega_1 t$$

$$\sin\left(\frac{\Delta\omega}{\omega_1}\sin\omega_1 t\right) = 2\sum_{1}^{\infty} J_{2m-1}\left(\frac{\Delta\omega}{\omega_1}\right)\sin(2m-1)\,\omega_1 t \qquad (6)$$

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

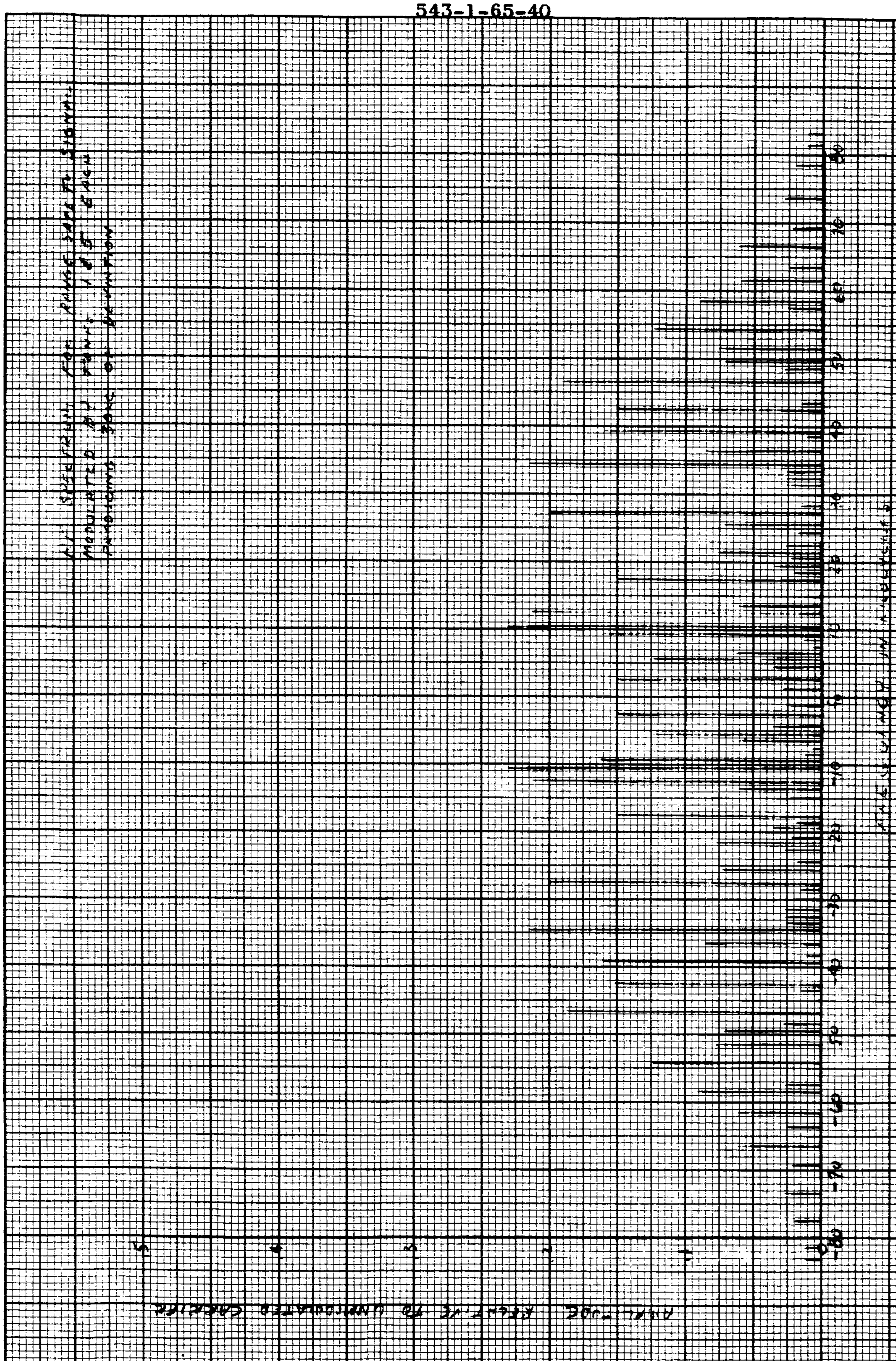


Figure 9. Two Tone RF Spectrum

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 frequency. 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[4,5] as early as 1961. 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 utilised; the simulation is described in the following section and the computer solutions are discussed in Section 7.

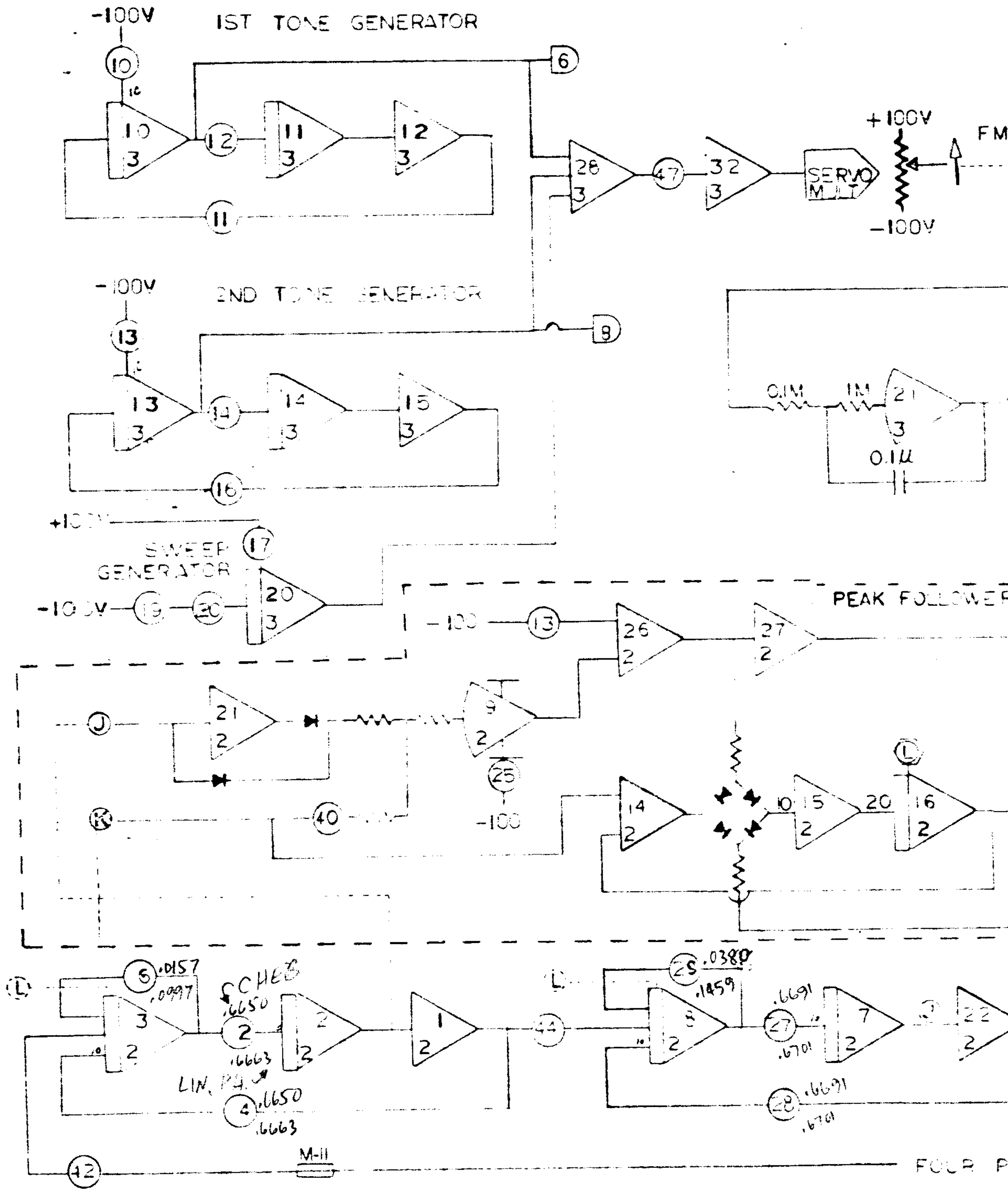
6. ANALOG COMPUTER SIMULATION

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.

<u>Unsuccessful Attempts to Use Scaling of 10^{-5}:</u> A 107 CPS constant amplitude oscillator capable of being frequency modulated was synthesized first.
6
This amplitude stabilized oscillator was later run at 1/10 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.


-100V
1ST TONE GENERATOR
2ND TONE GENERATOR
SWEEP GENERATOR
SERVO MULT
+100V
-100V
FM
0.1M
1M
0.1μ
PEAK FOLLOWER
SCHEB
LIN. PH.
M-11
FOUR

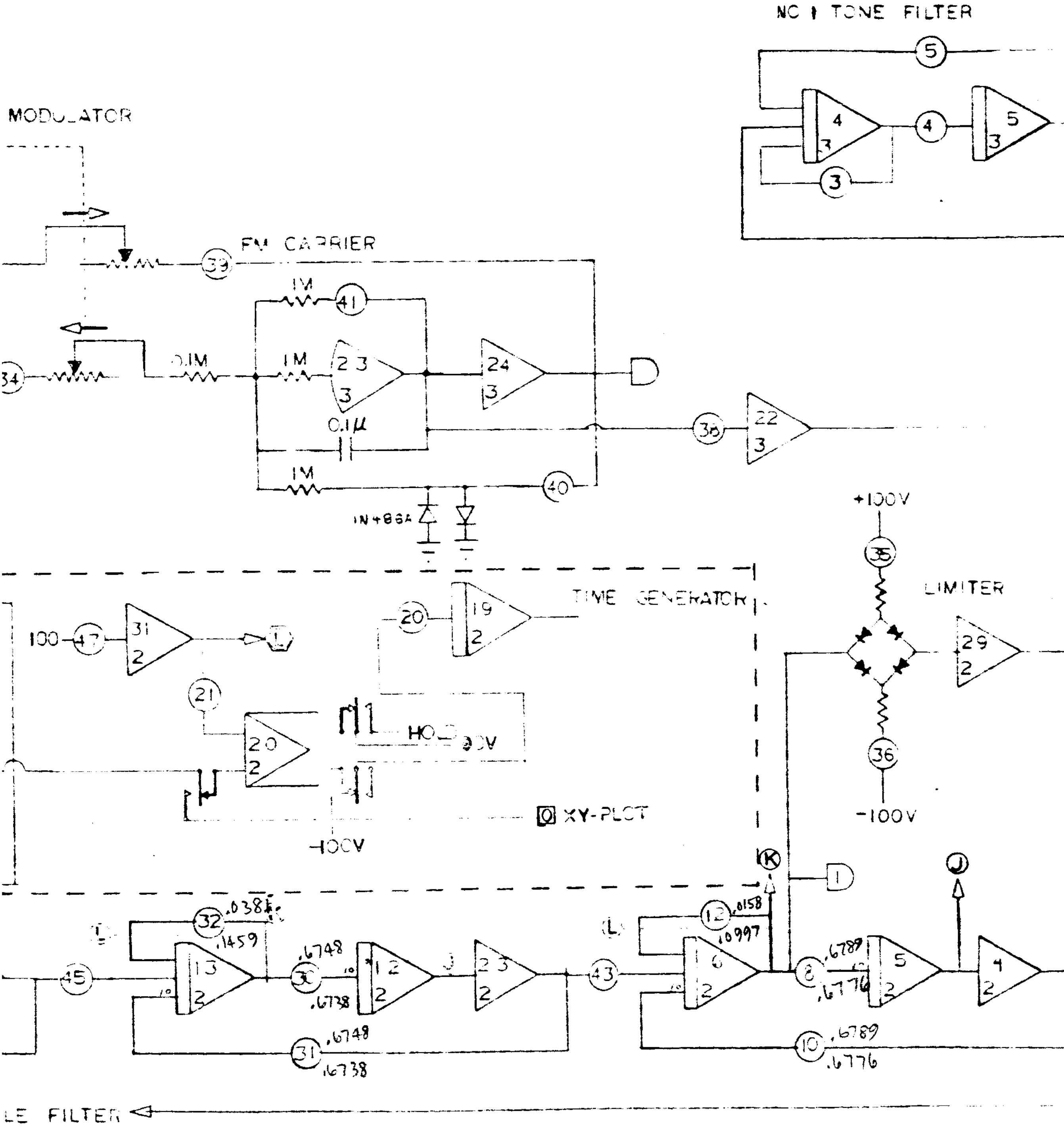
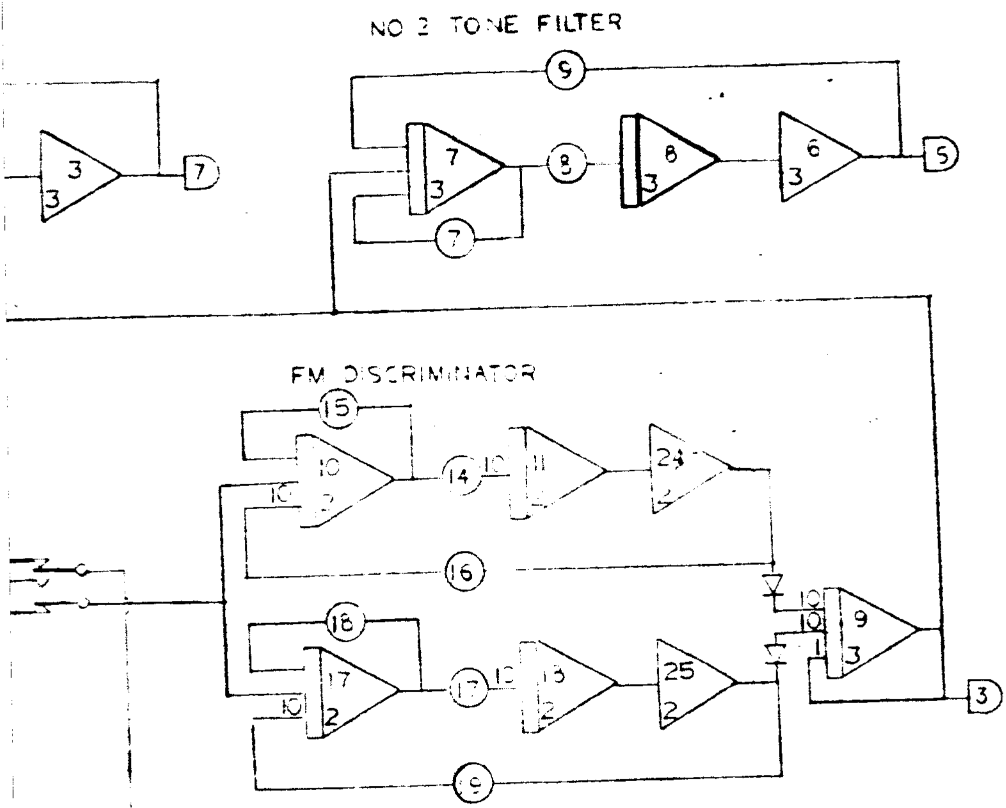


FIGURE 10 ANALOG COMPUTER SIMULATION



STATION 2

POT

2 - 6650	25 - 1000
4 - 6650	27 - 5691
6 - 0157	28 - 5691
8 - 5789	29 - 0380
10 - 5789	30 - 6748
12 - 0158	31 - 6748
13 - 0150	32 - 0331
14 - 6092	35 - 0050
15 - 0490	36 - 0050
16 - 6092	40 - 0020
17 - 7348	42 - 1400
18 - 0446	43 - 1400
19 - 7348	44 - 1400
20 - 0100	45 - 1400
21 - 5000	47 - 1000

STATION 3

3 - 0500	34 - 7728
4 - 0471	38 - 0500
5 - 0471	39 - 7728
7 - 0500	40 - 8000
8 - 0762	41 - 3000
9 - 0762	47 - TEXT
10 - 0300	
11 - 0471	
12 - 0471	
13 - 0300	
14 - 0762	
15 - 0762	
17 - 2000	
19 - 0010	
20 - 0005	

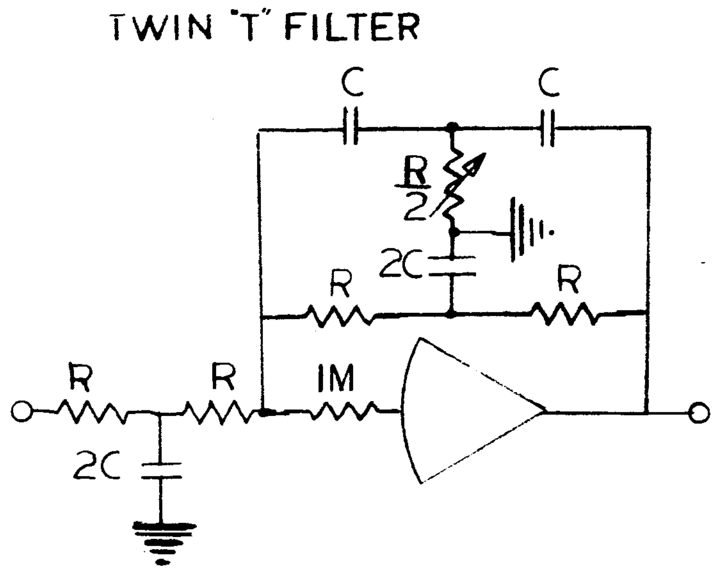


Figure 11. Twin "T" Filter

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 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

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

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.)

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 ω_o/Q = the pot setting. At the same time the Q is being tweeked in, the ω_o 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 calibration 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 computer 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 input 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 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 12 and 13 illustrate the results for the Chebishev and linear phase filters respectively. The mathematical description of the filter transfer functions are:

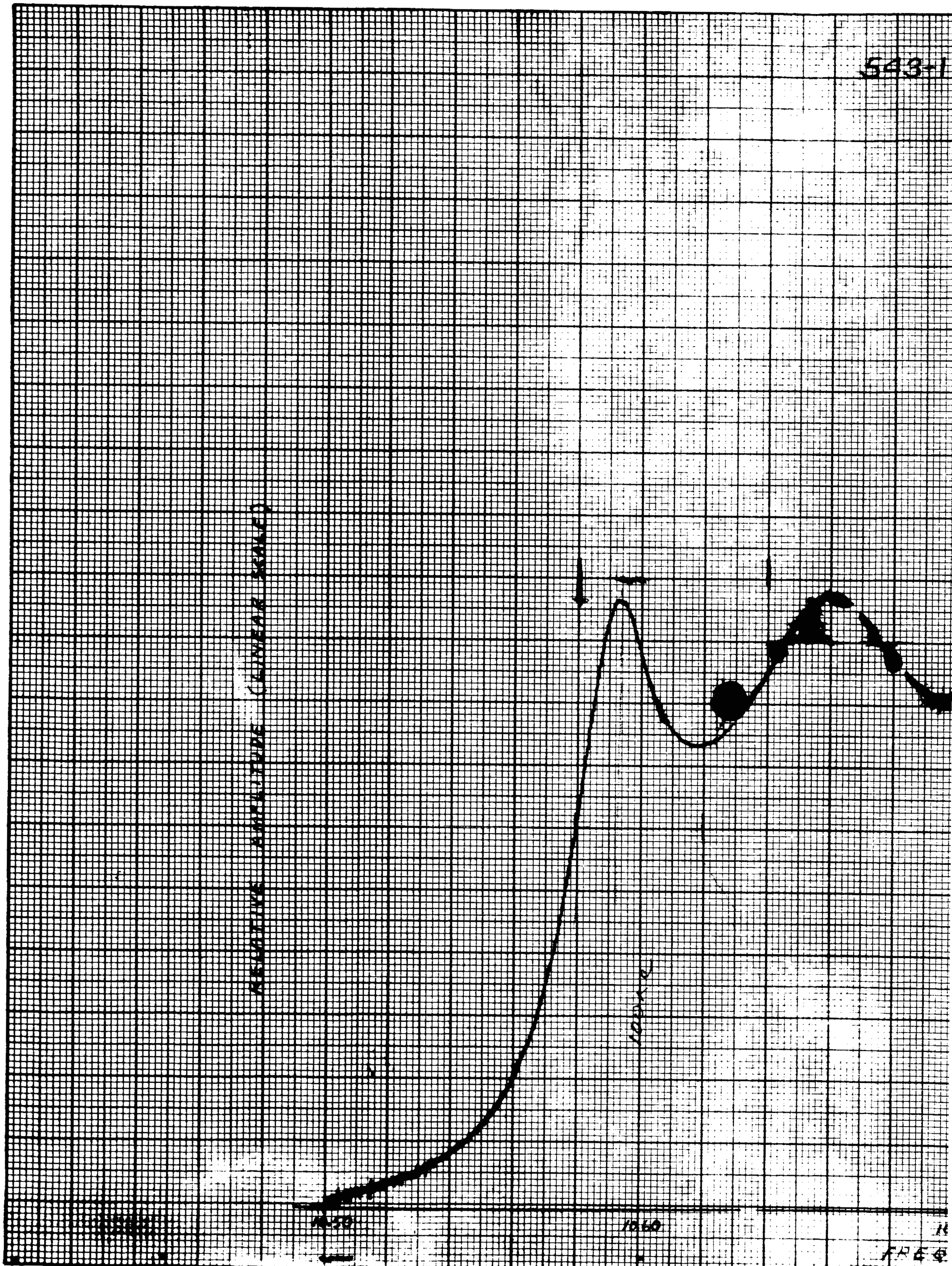
$$\frac{e_o}{e_{in}} = \frac{s}{[s-P_1][s-\bar{P}_1][s-P_2][s-\bar{P}_2][s-P_3][s-\bar{P}_3][s-P_4][s-\bar{P}_4]} \qquad (7)$$

Chebishev:

$P_1 = -.0788 + J66.50$
$\bar{P}_1 = -.0788 - J66.50$
$P_2 = -.1902 + J66.91$
$\bar{P}_2 = -.1902 - J66.91$
$P_3 = -.1902 + J67.48$
$\bar{P}_3 = -.1902 - J67.48$
$P_4 = -.0788 + J67.89$
$\bar{P}_4 = -.0788 - J67.89$

Linear Phase:

$P_1 = -.4984 + J66.63$
$\bar{P}_1 = -.4984 - J66.63$
$P_2 = -.7296 + J67.01$
$\bar{P}_2 = -.7296 - J67.01$
$P_3 = -.7296 + J67.38$
$\bar{P}_3 = -.7296 - J67.38$
$P_4 = -.4984 + J67.76$
$\bar{P}_4 = -.4984 - J67.76$



KEUFFEL & ESSER CO.

27-1

120KC

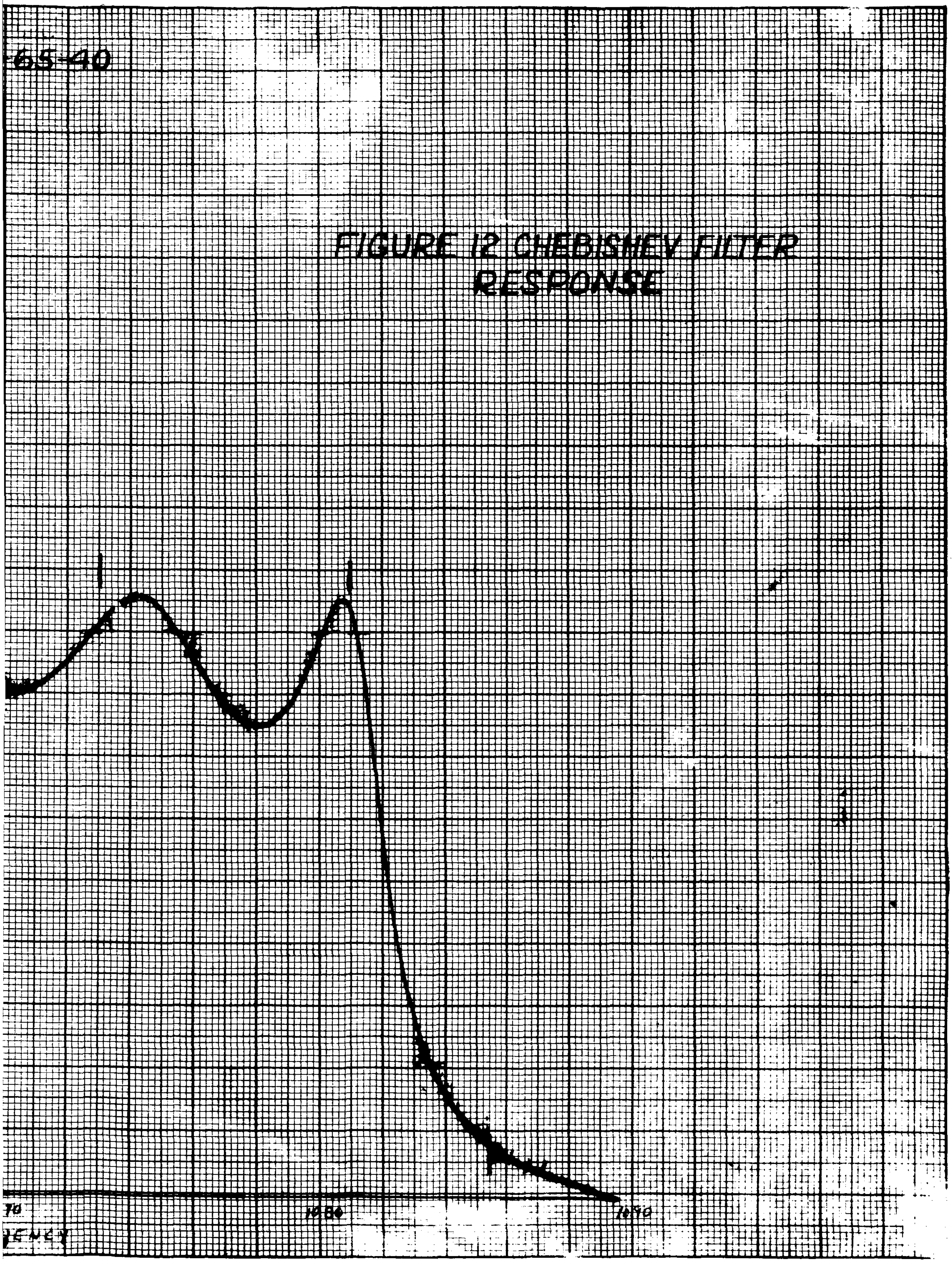


FIGURE 12 CHEBISHEV FILTER RESPONSE

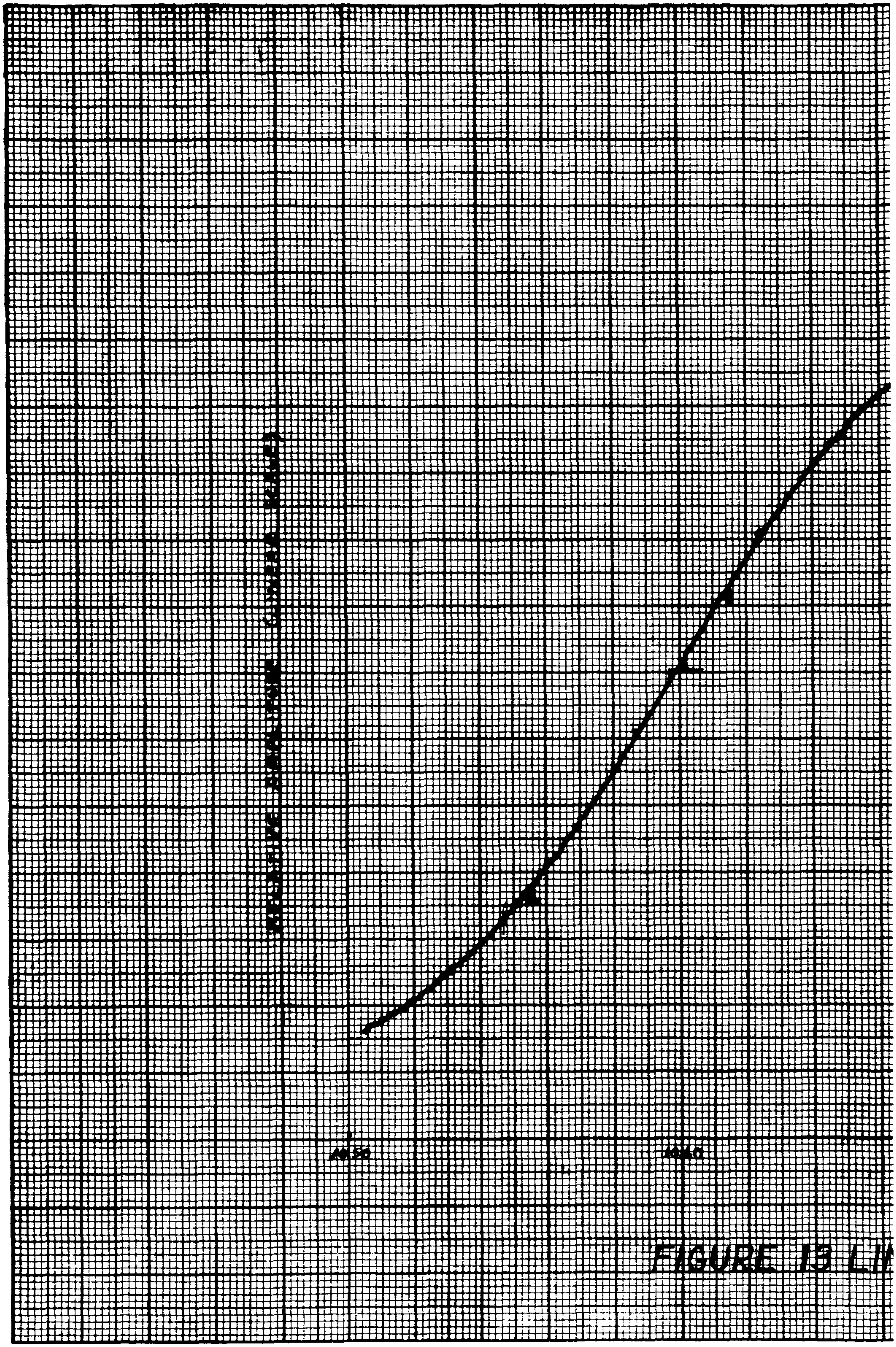
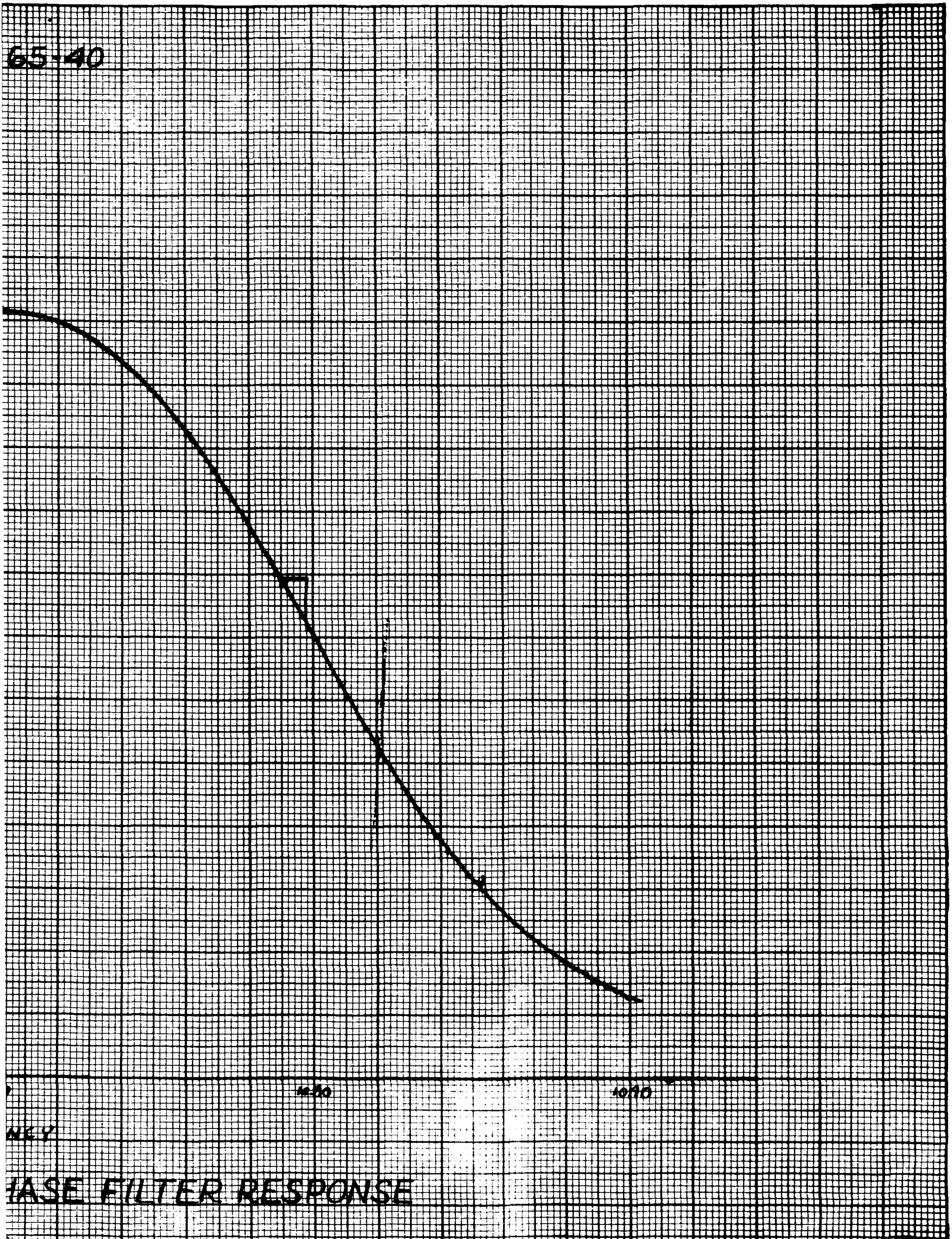


FIGURE 13 LI


65-40
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 vessus 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 continuous, the tone filter output is in phase with the tone oscillator. Null conditions occurred on

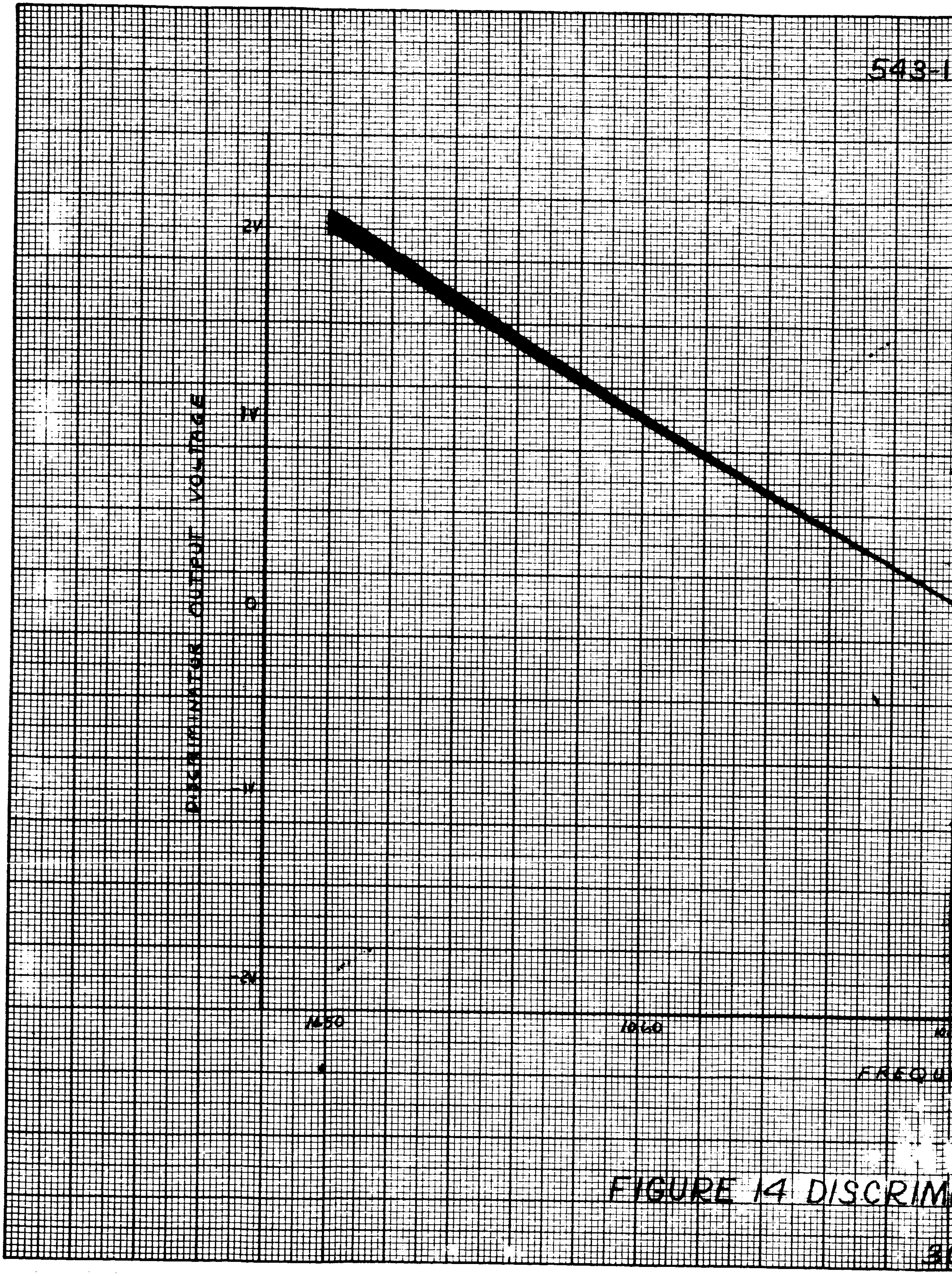
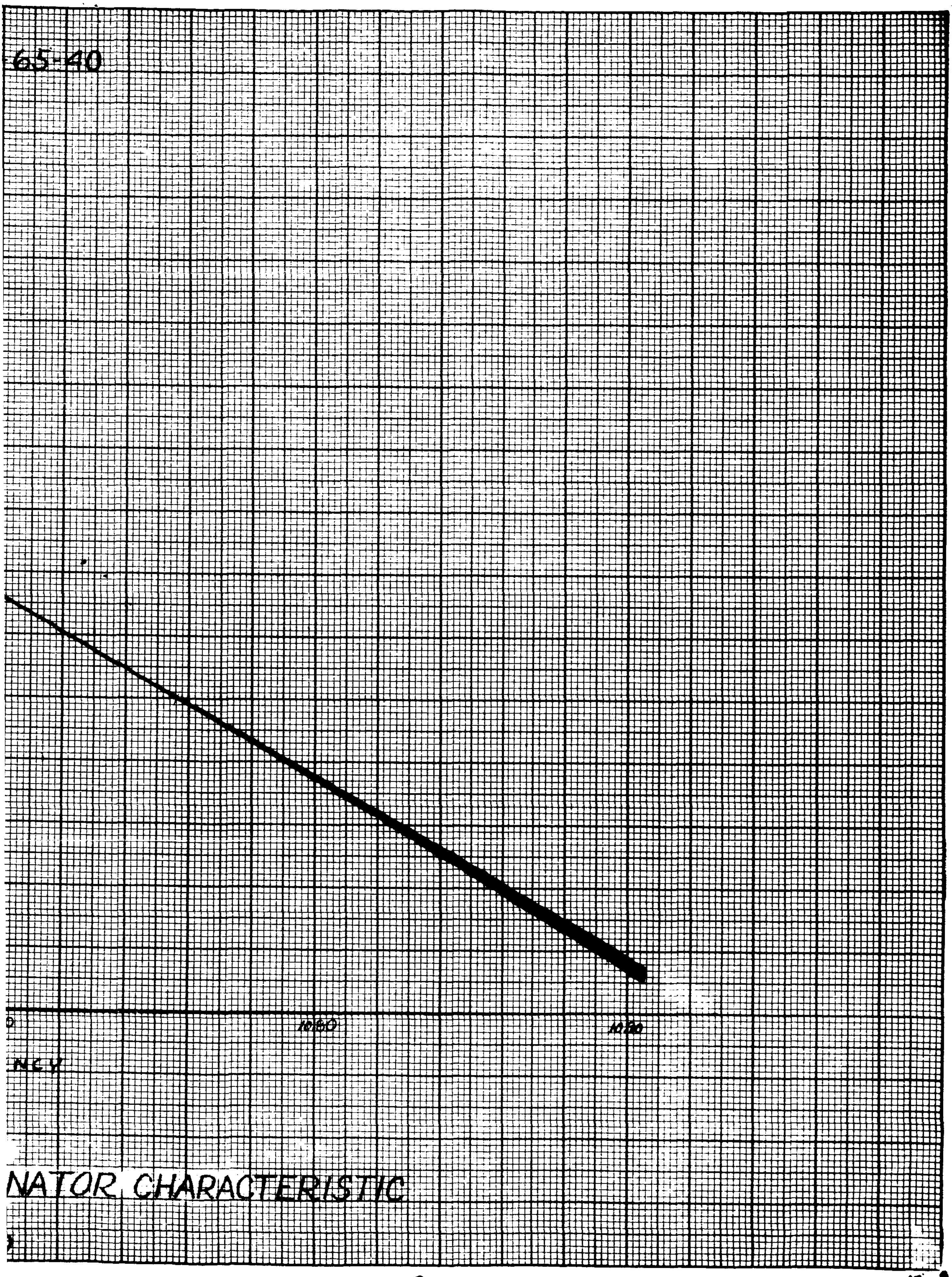
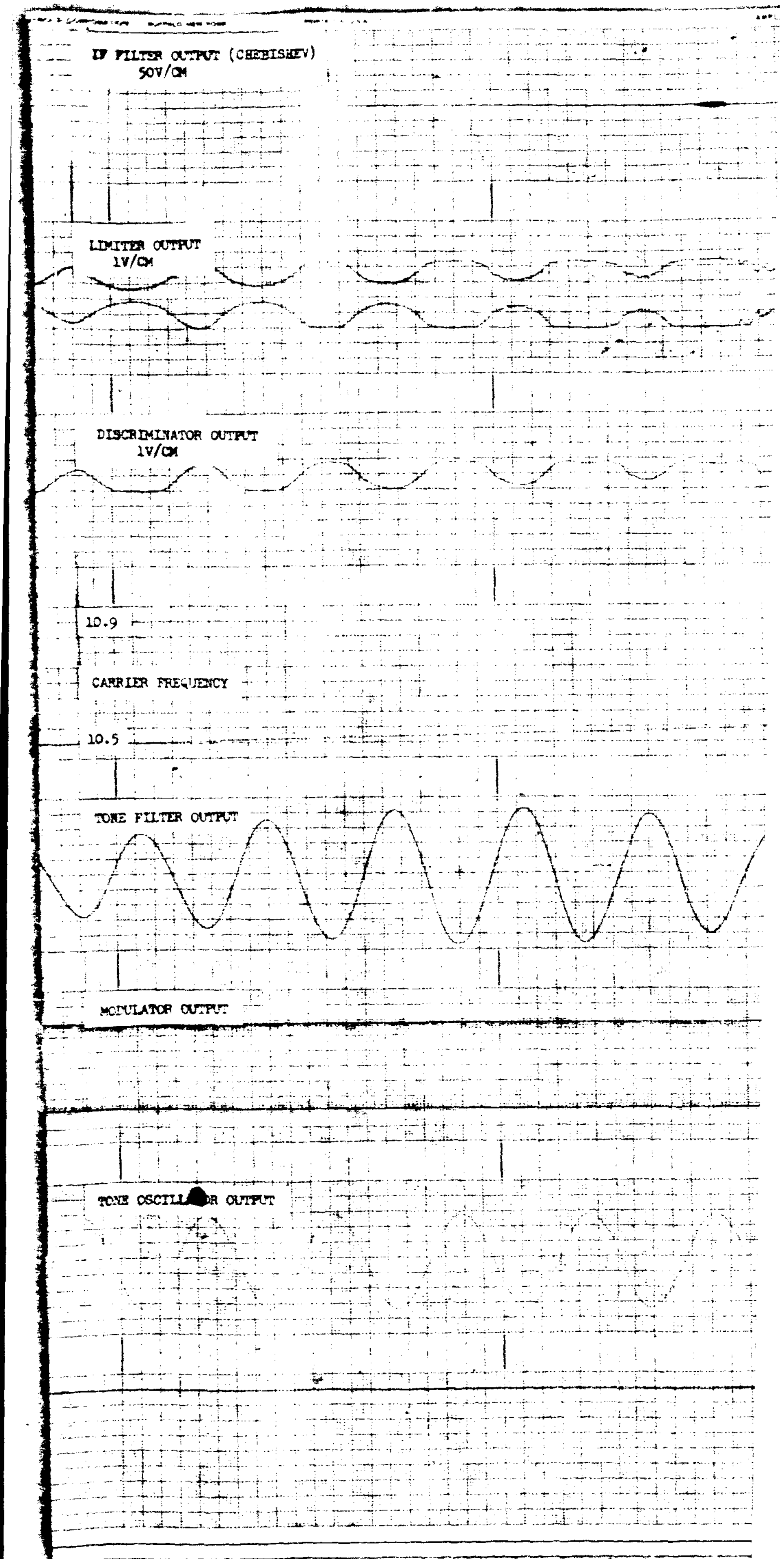


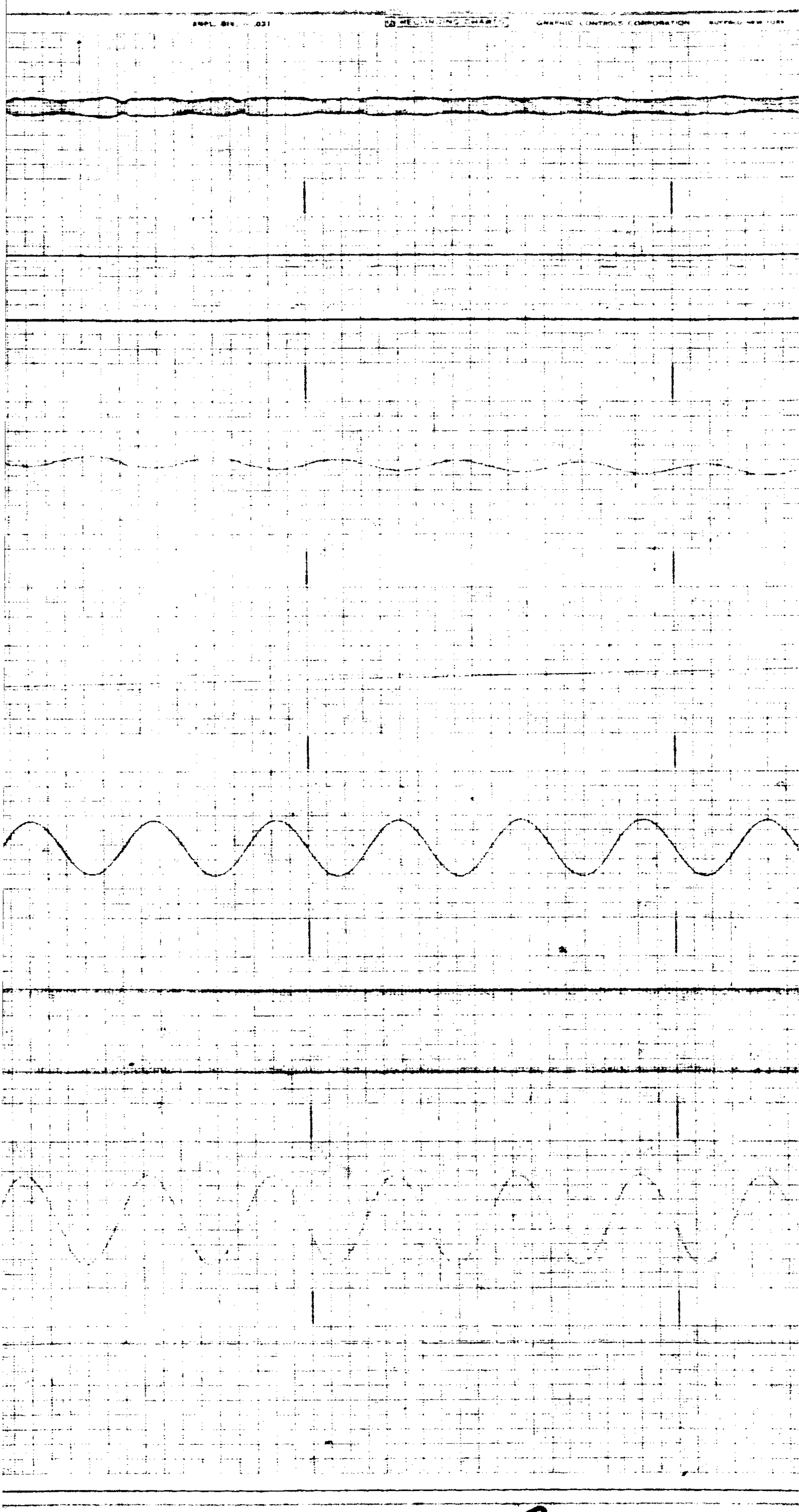
FIGURE 14 DISCRIM


65-40
1080
NCY
NATOR CHARACTERISTIC

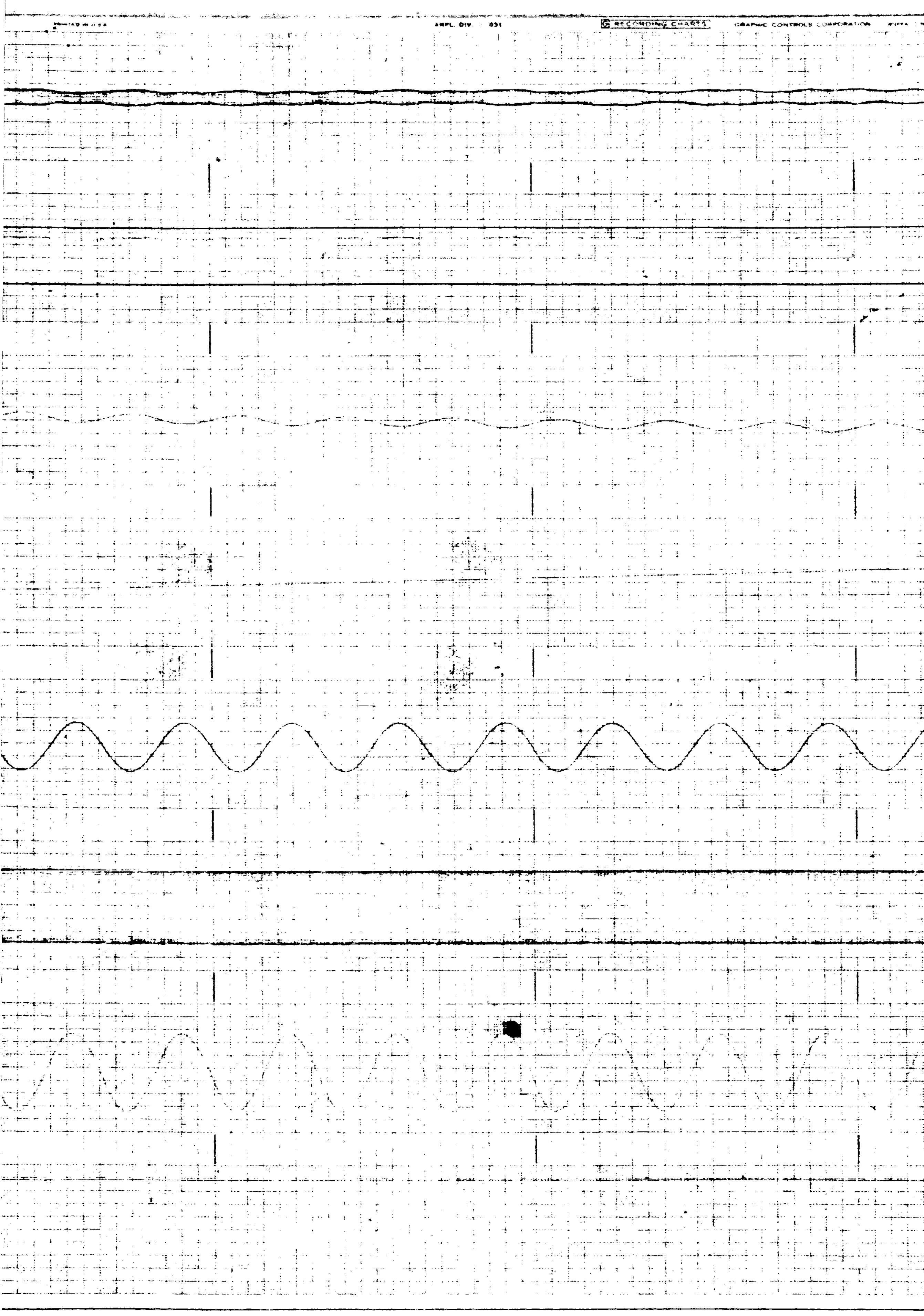

IF FILTER OUTPUT (CHEBISHEV)
50V/CM
LIMITER OUTPUT
1V/CM
DISCRIMINATOR OUTPUT
1V/CM
10.9
CARRIER FREQUENCY
10.5
TONE FILTER OUTPUT
MODULATOR OUTPUT
TONE OSCILLATOR OUTPUT

31-1

31-2



31-3



31-4

31-5

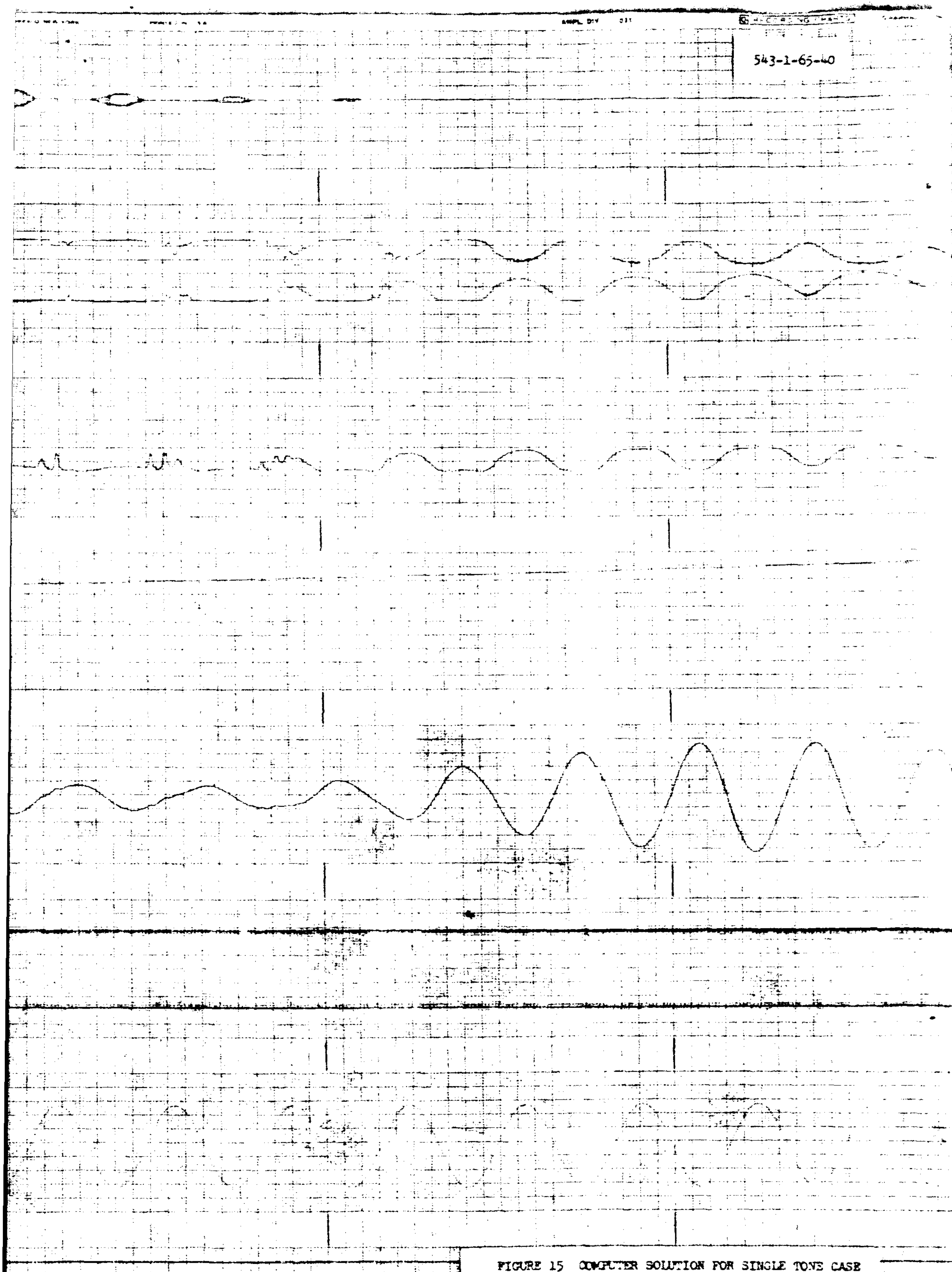
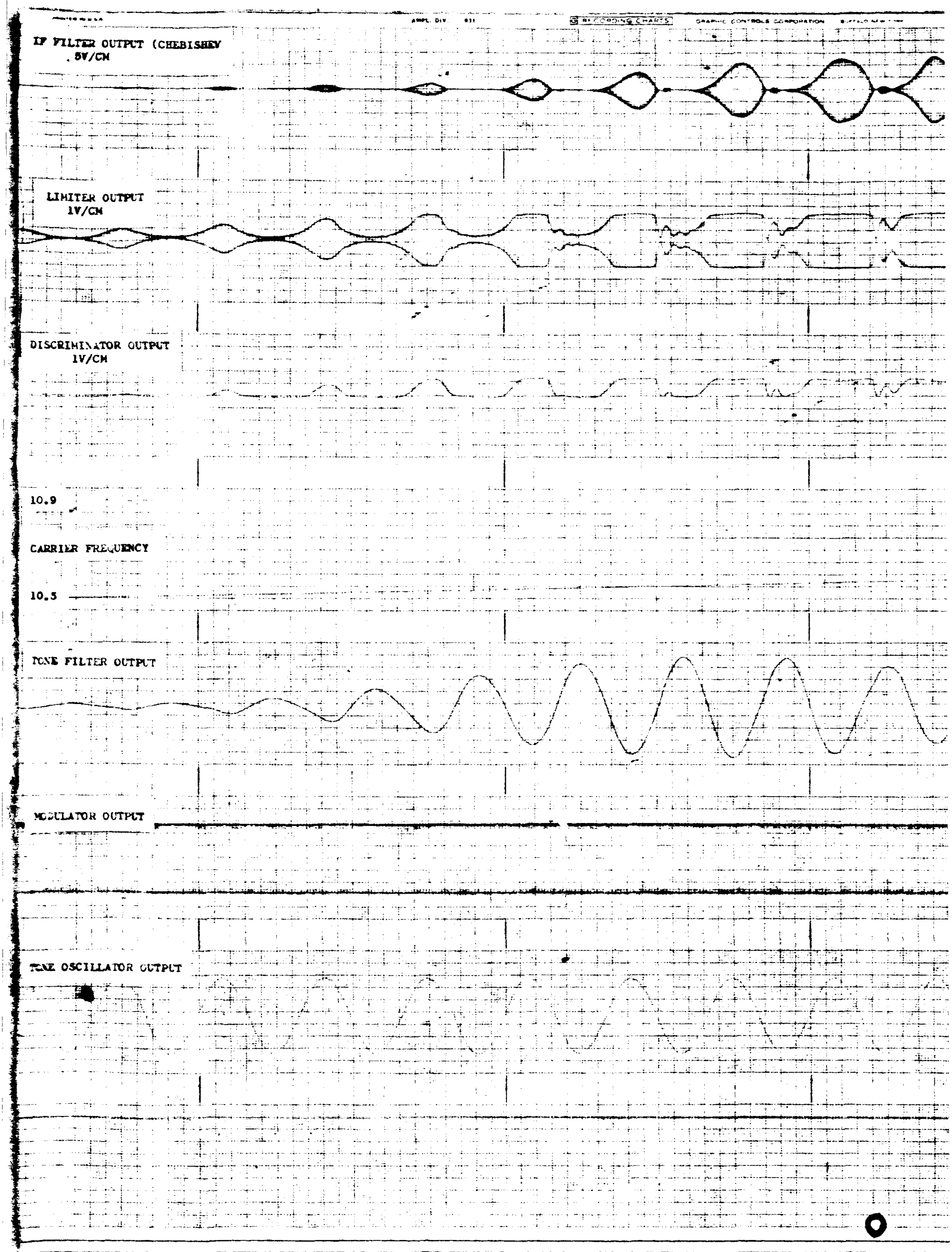


FIGURE 15 COMPUTER SOLUTION FOR SINGLE TONE CASE

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. Correcting for tone filter storage the null points are more accurately located at 10.565 cps and 10.835 cps. Another interesting observation 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.

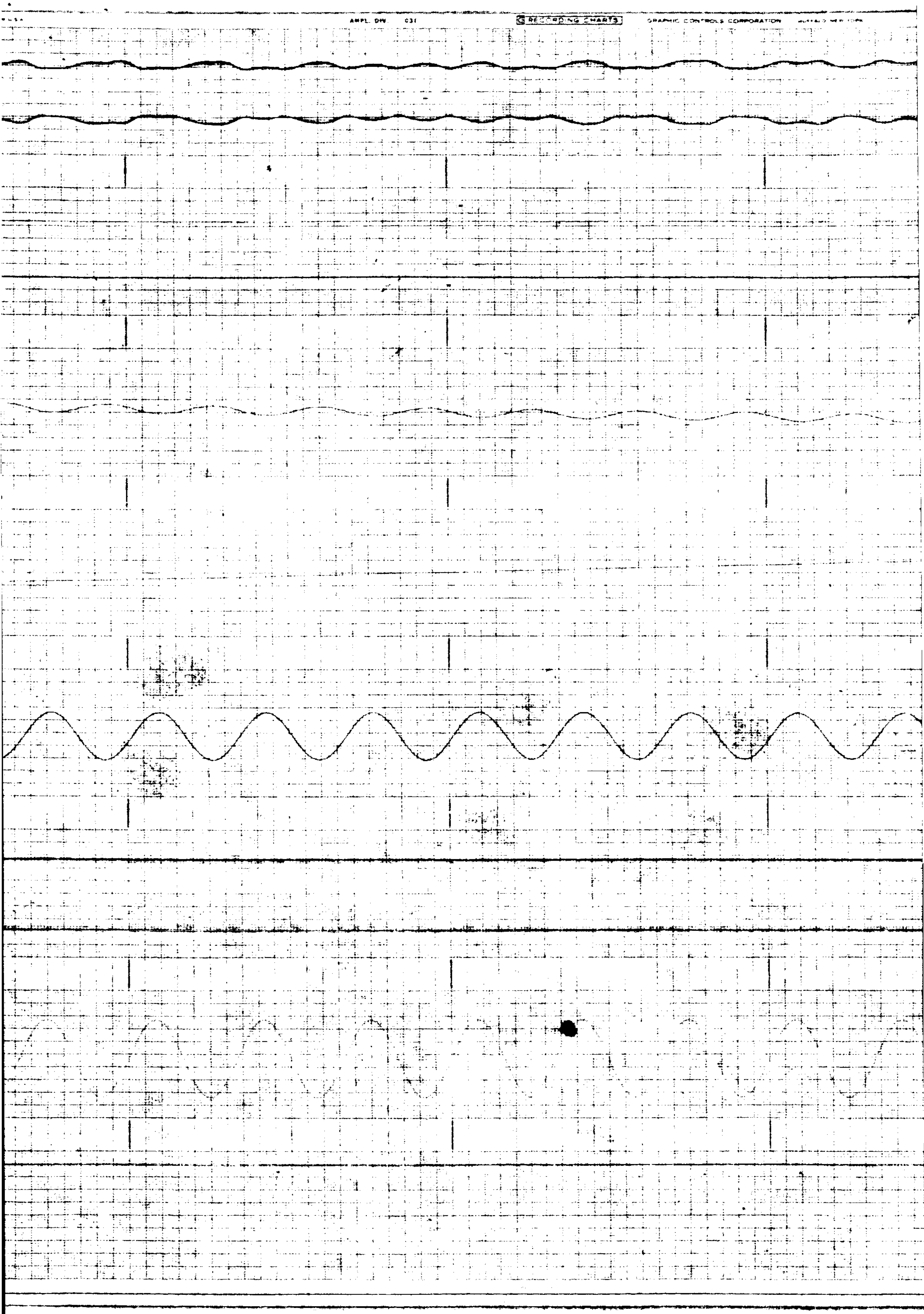
Data of Figure 16 is similar in all respects to the data of Figure except that the IF filter output (limiter input) is reduced 10 db. The effect is that the frequency range in which limiting is continuous is narrowed and the null points are closer together.

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


AMPL. DIV.
RECORDING CHARTS
GRAPHIC CONTROLS CORPORATION
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


AMPL. DIV. 031
RECORDING CHARTS
GRAPHIC CONTROLS CORPORATION

33-4


RECORDING CHARTS
GRAPHIC CONTROLS CORPORATION

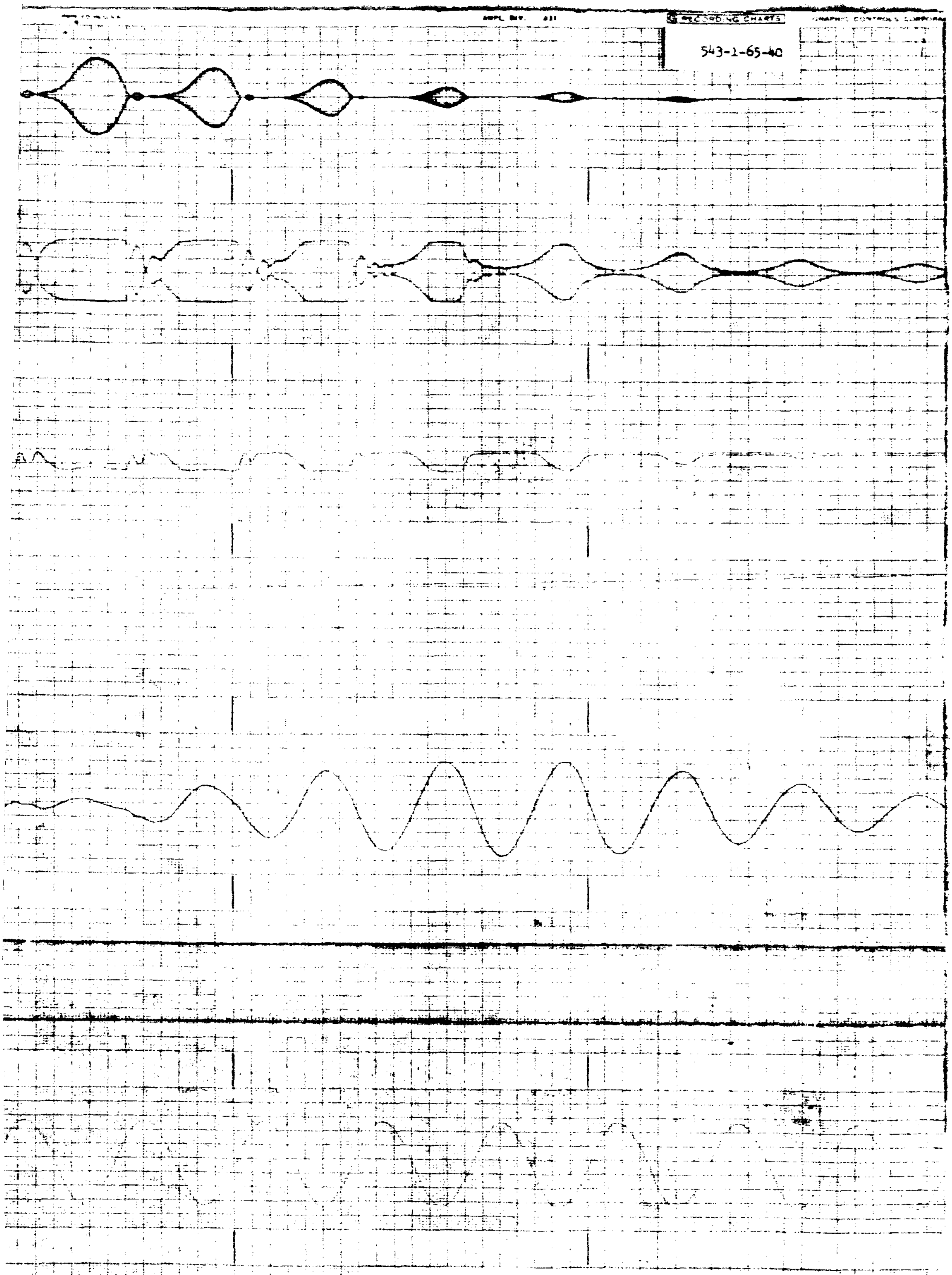
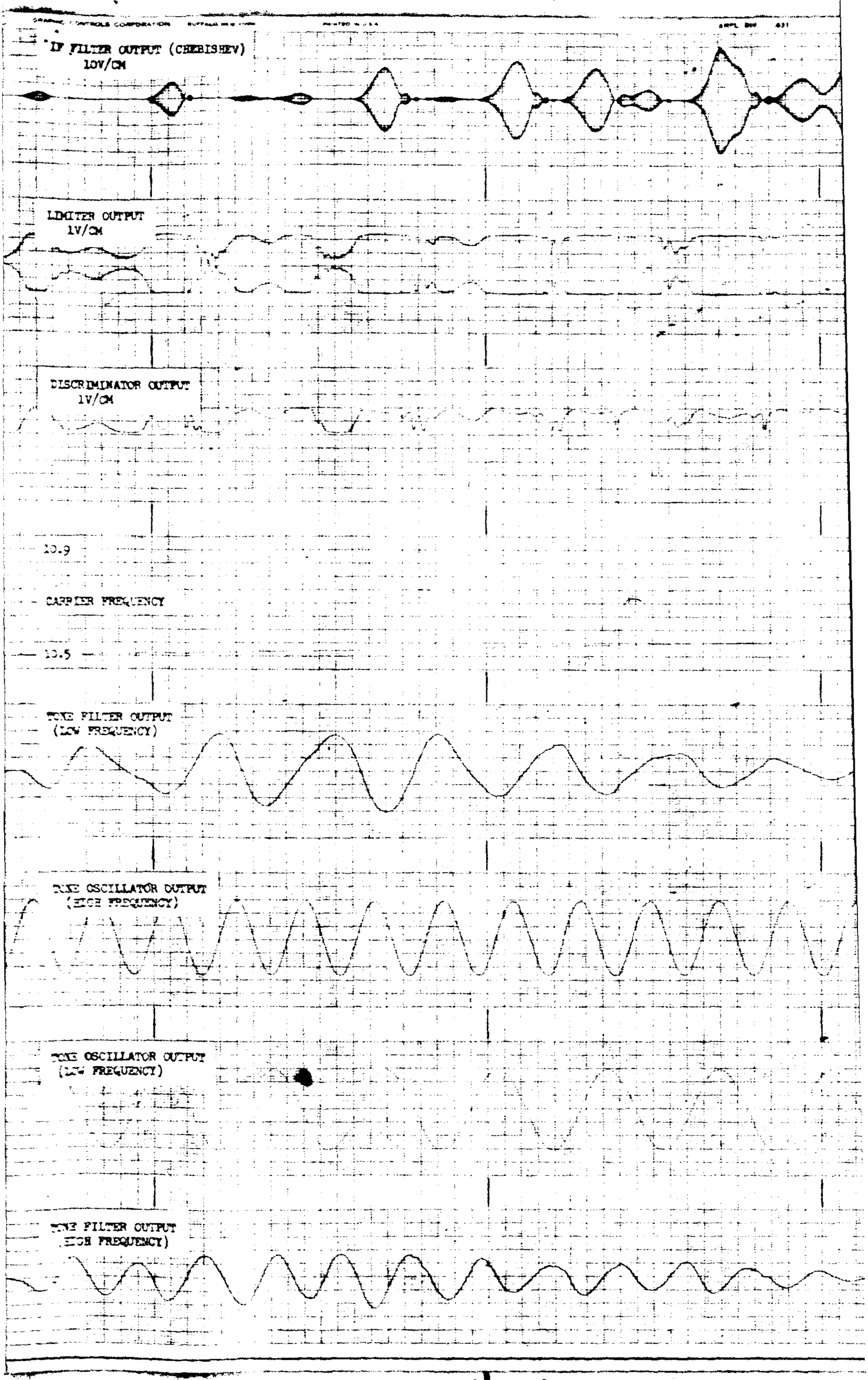
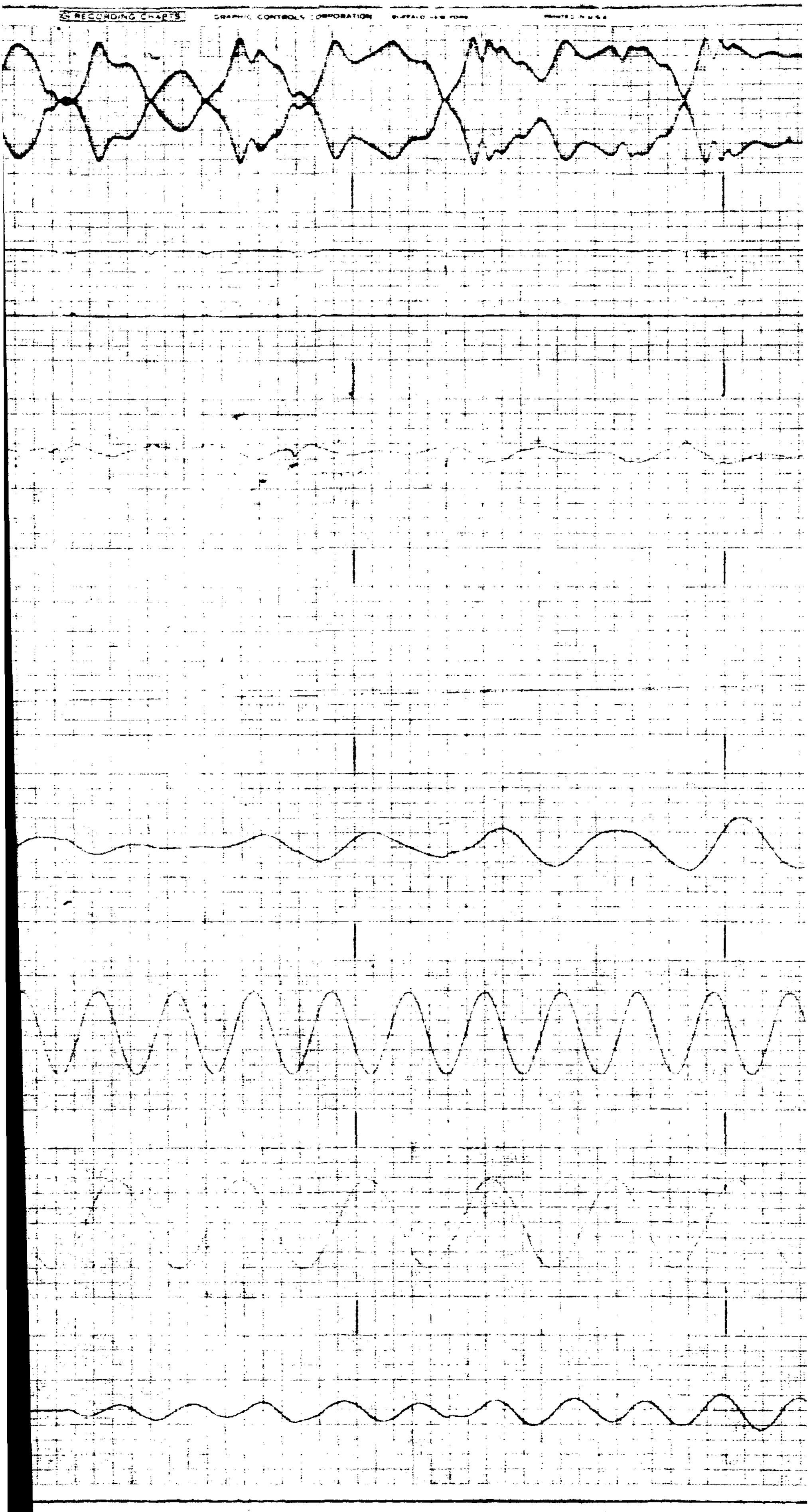
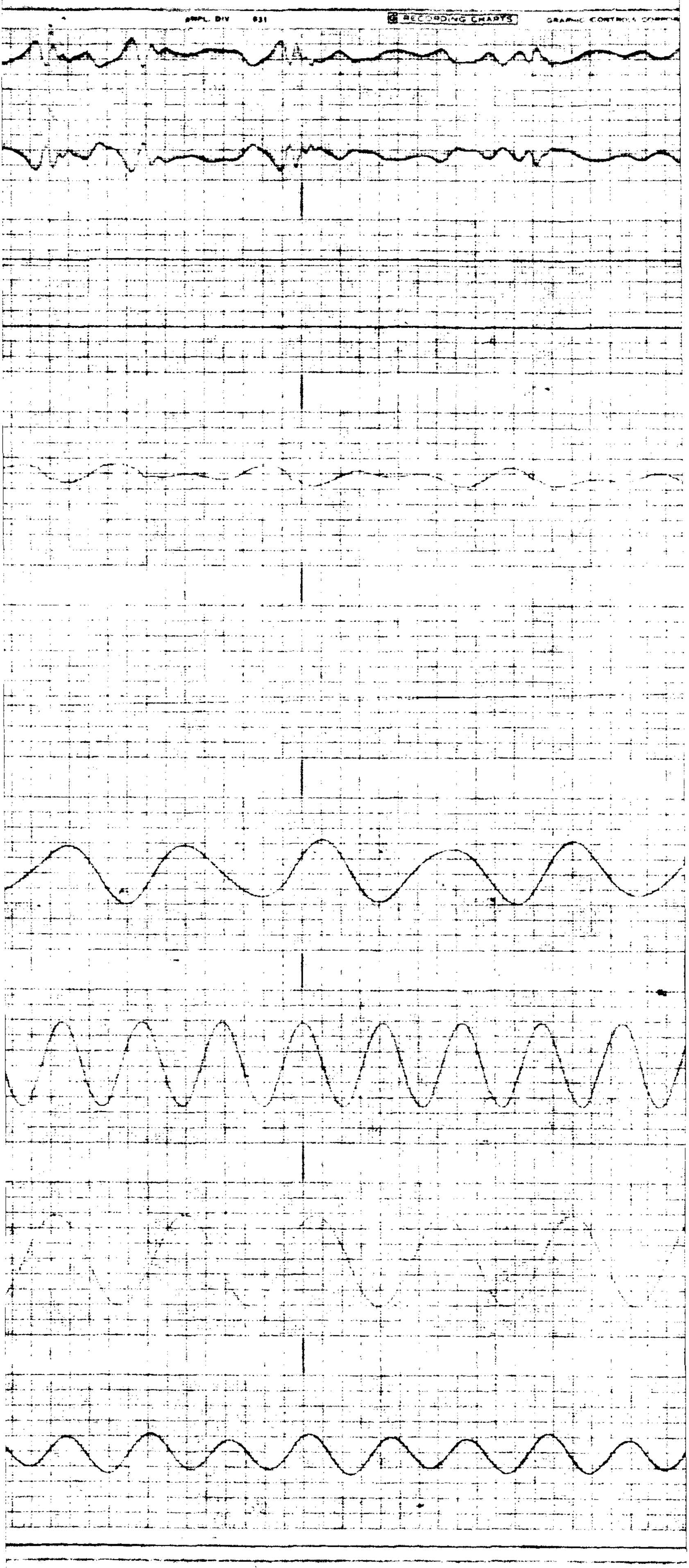


FIGURE 16 COMPUTER SOLUTION FOR SINGLE TONE CASE (LOW LEVEL)

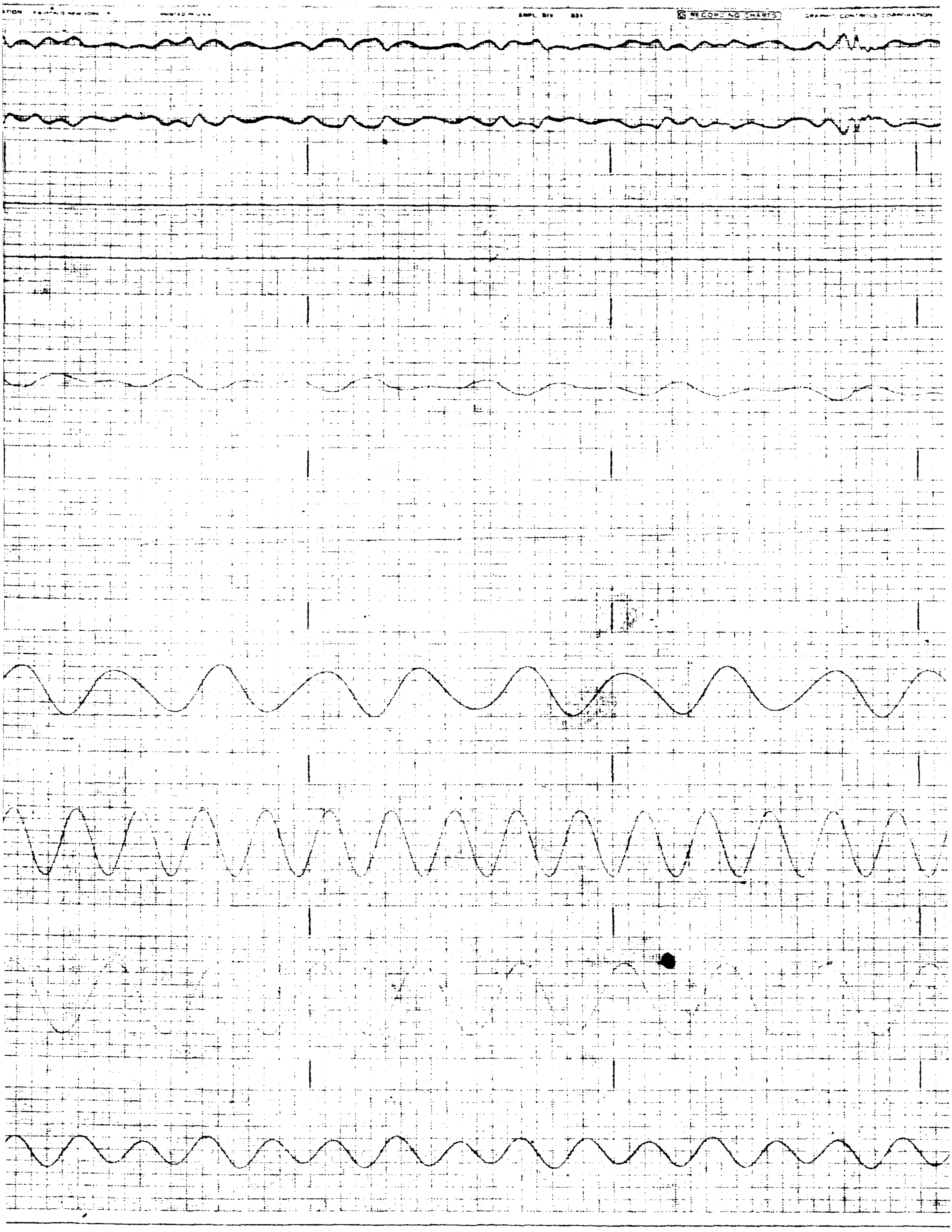

IF FILTER OUTPUT (CHEBISHEV)
10V/CM
LIMITER OUTPUT
1V/CM
DISCRIMINATOR OUTPUT
1V/CM
10.9
CARRIER FREQUENCY
10.5
TONE FILTER OUTPUT
(LOW FREQUENCY)
TONE OSCILLATOR OUTPUT
(HIGH FREQUENCY)
TONE OSCILLATOR OUTPUT
(LOW FREQUENCY)
TONE FILTER OUTPUT
(HIGH FREQUENCY)



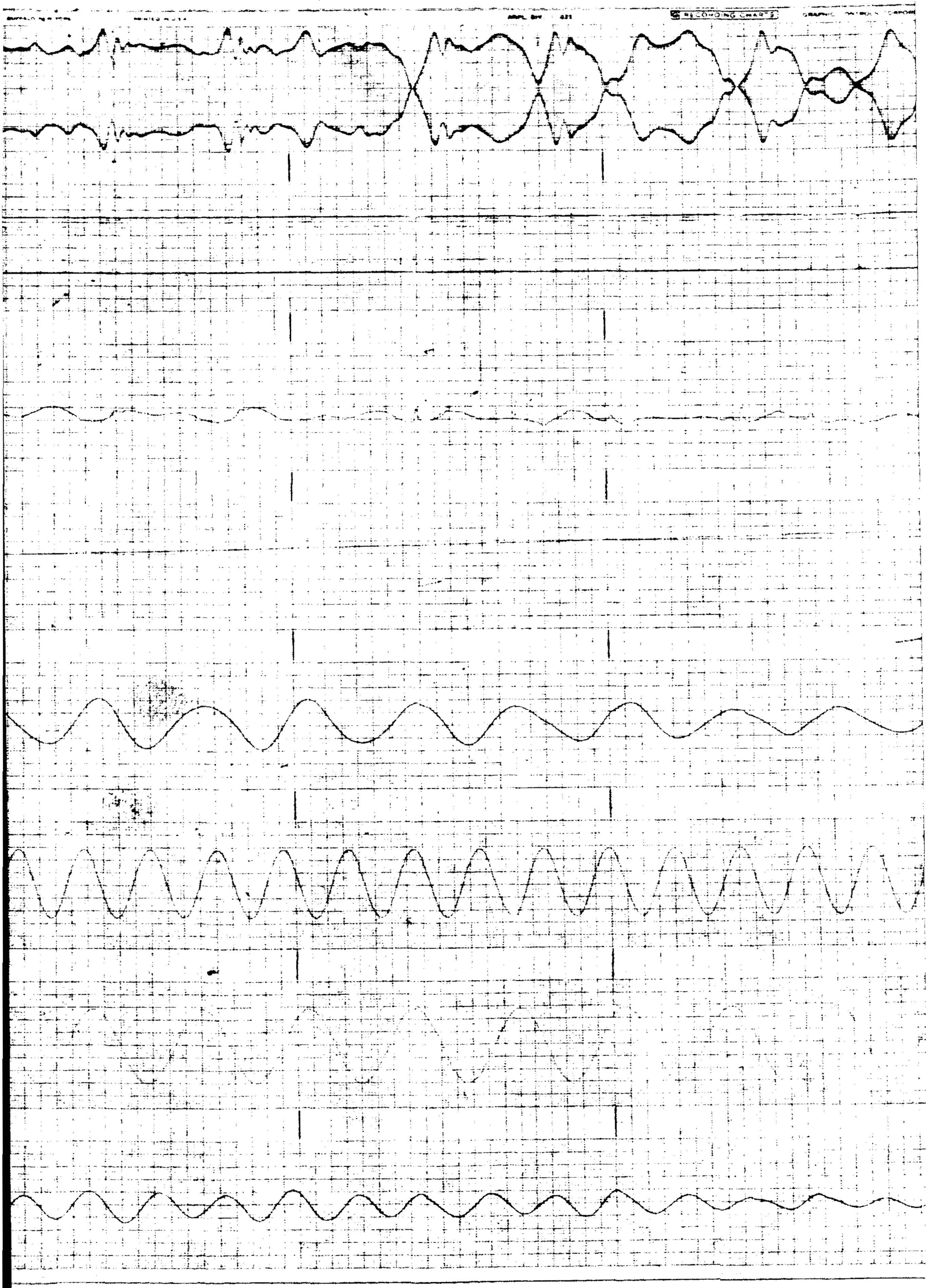
34.2



34-3



34-4



34-5

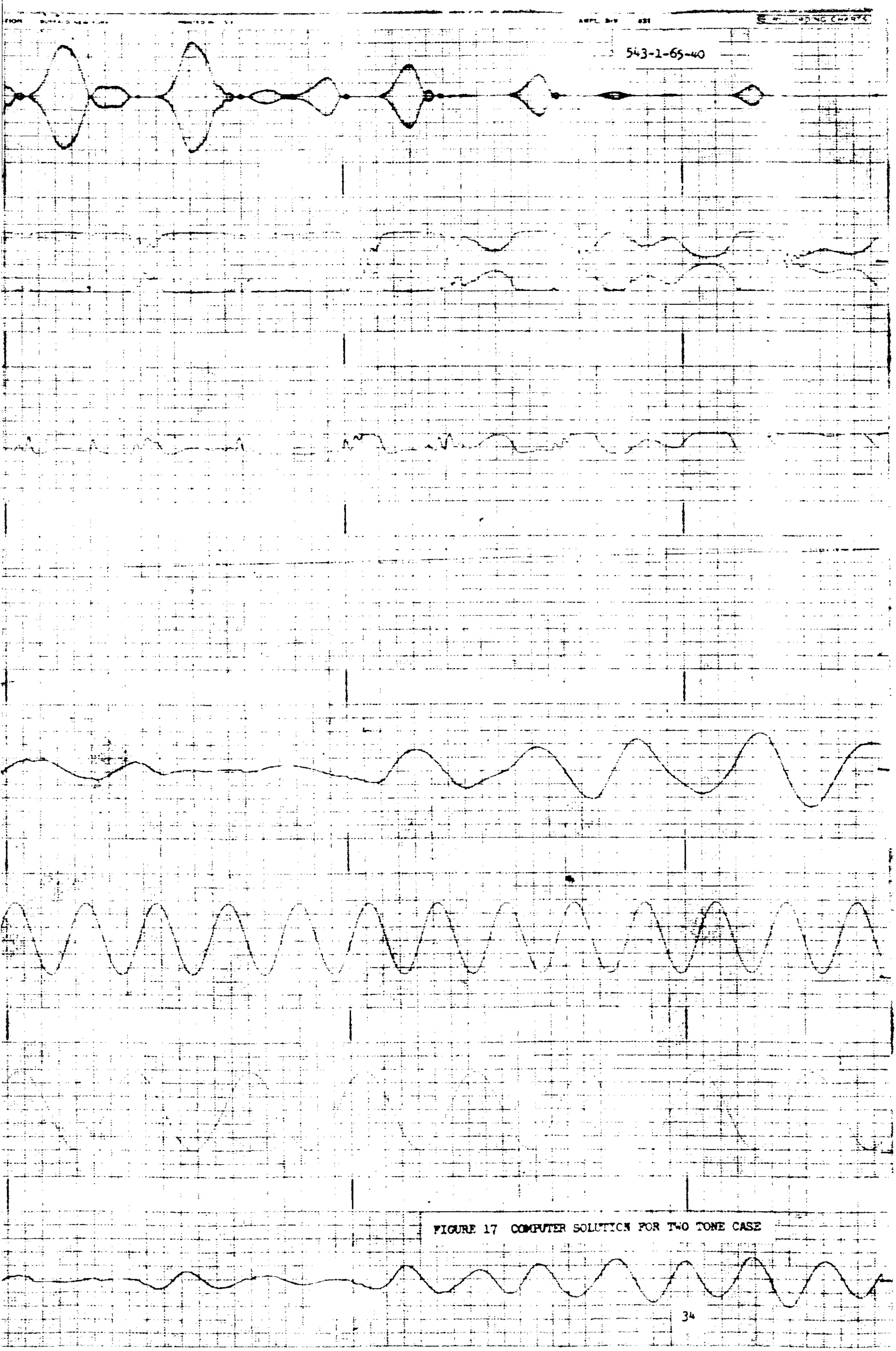
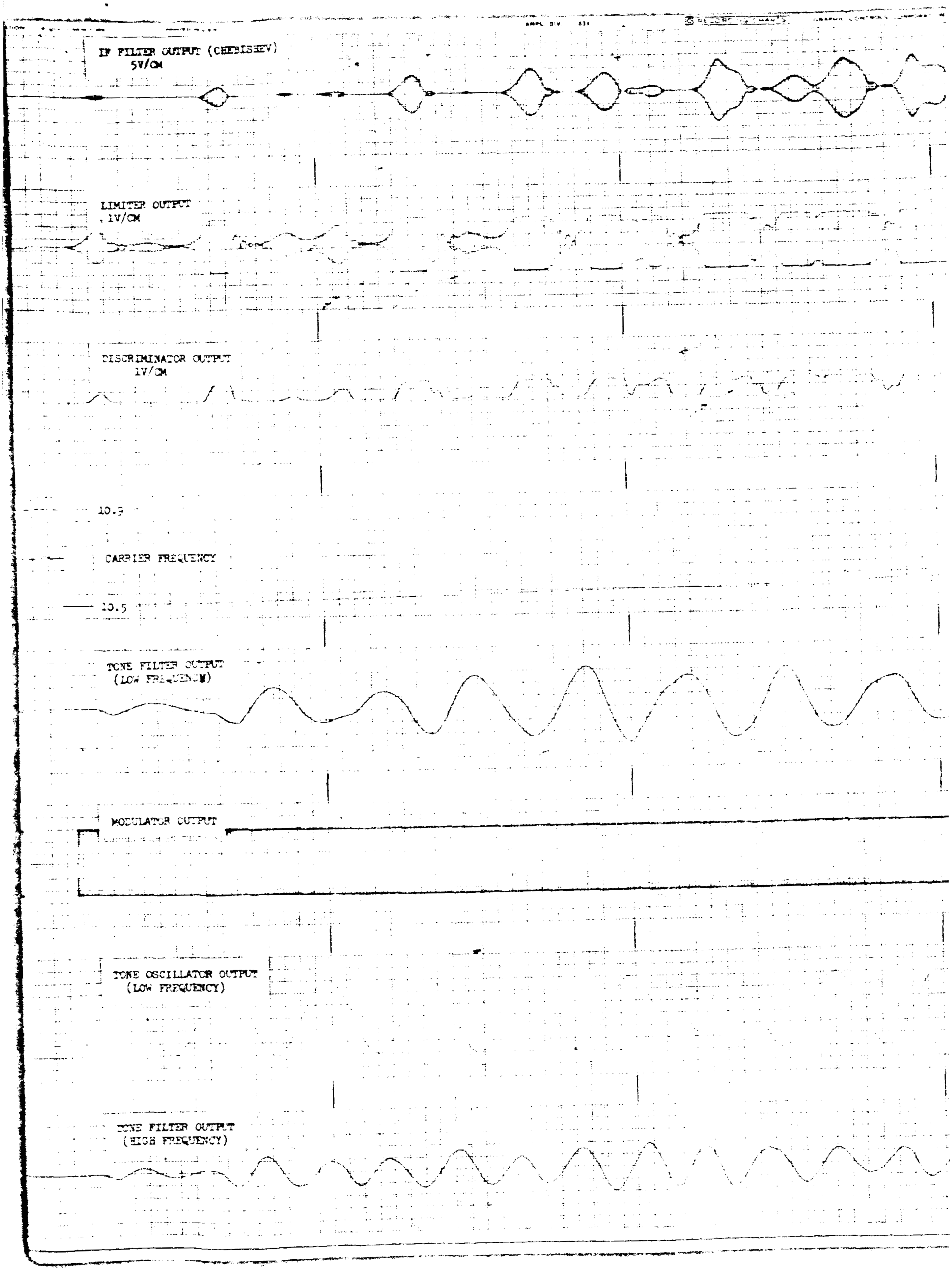
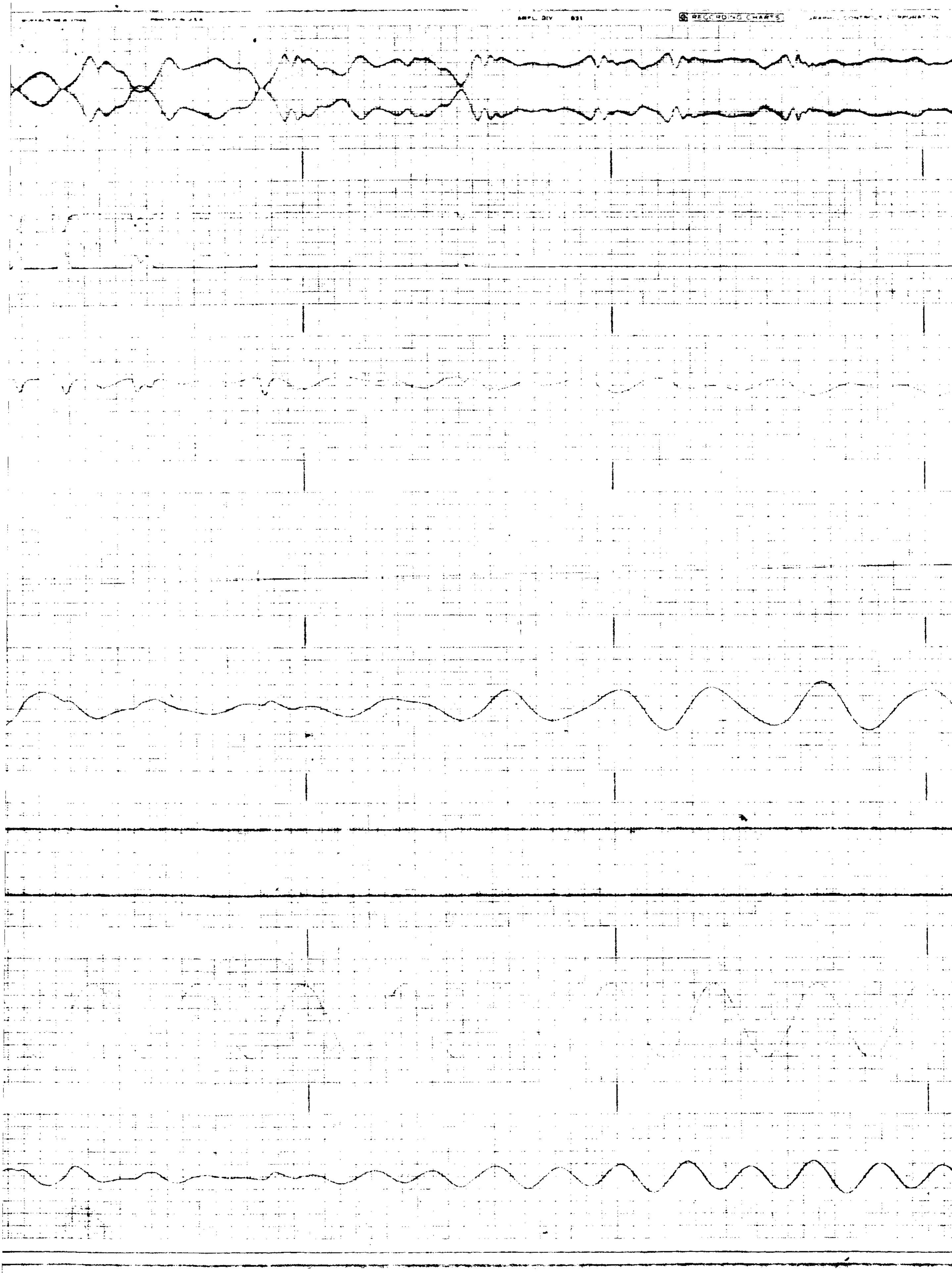


FIGURE 17 COMPUTER SOLUTION FOR TWO TONE CASE


IF FILTER OUTPUT (CHEBISHEV)
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)



35-2

35-3

35-4

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

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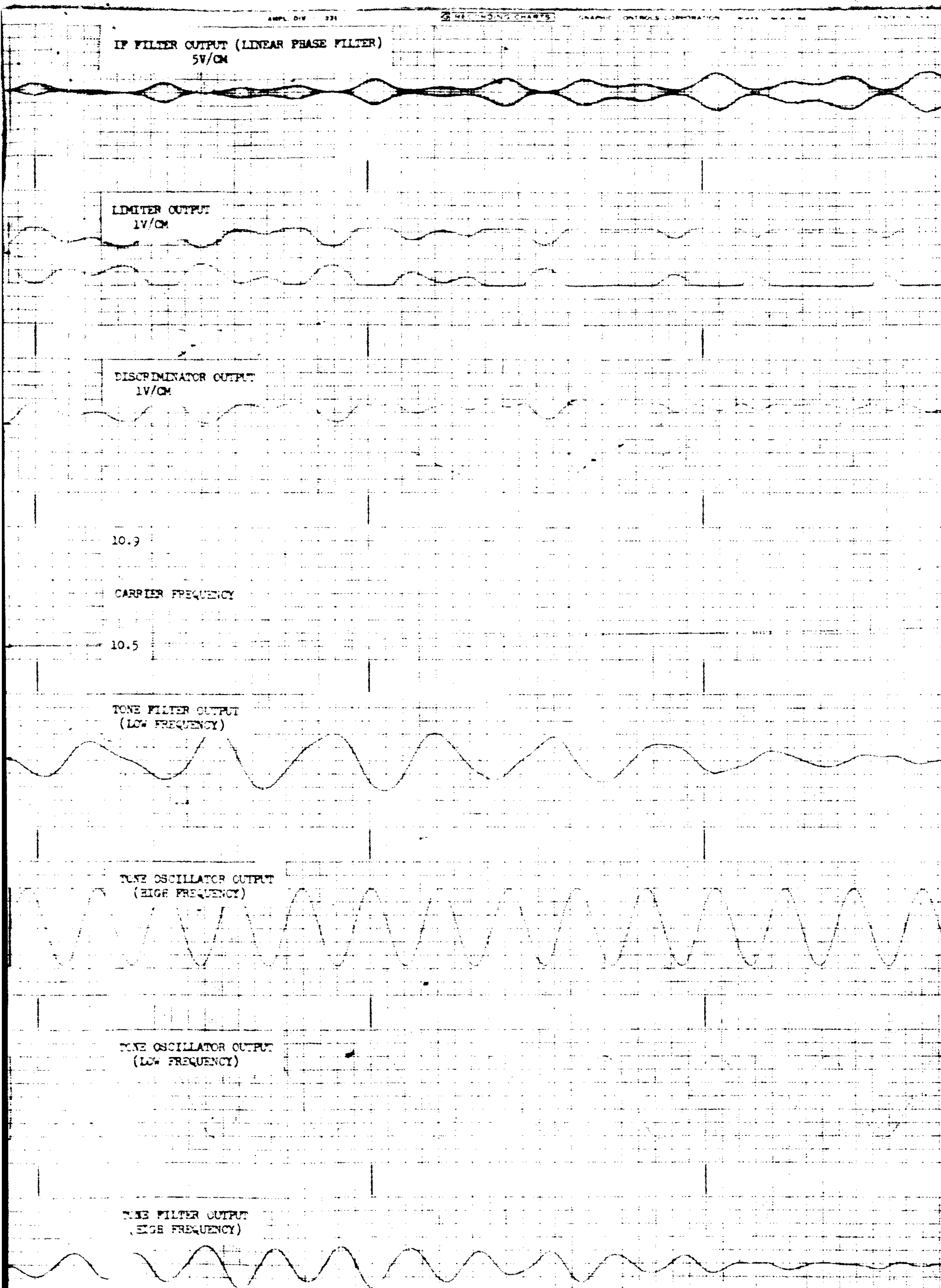
and the high frequency tone oscillator output omitted. Basic data characteristics are the same as those for the single tone case. The tone filter outputs are out of phase with the tone oscillator outputs when operation is below limiting level and in phase in the center portion of the band where limiting 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.


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)

37-2

37-3

37-4

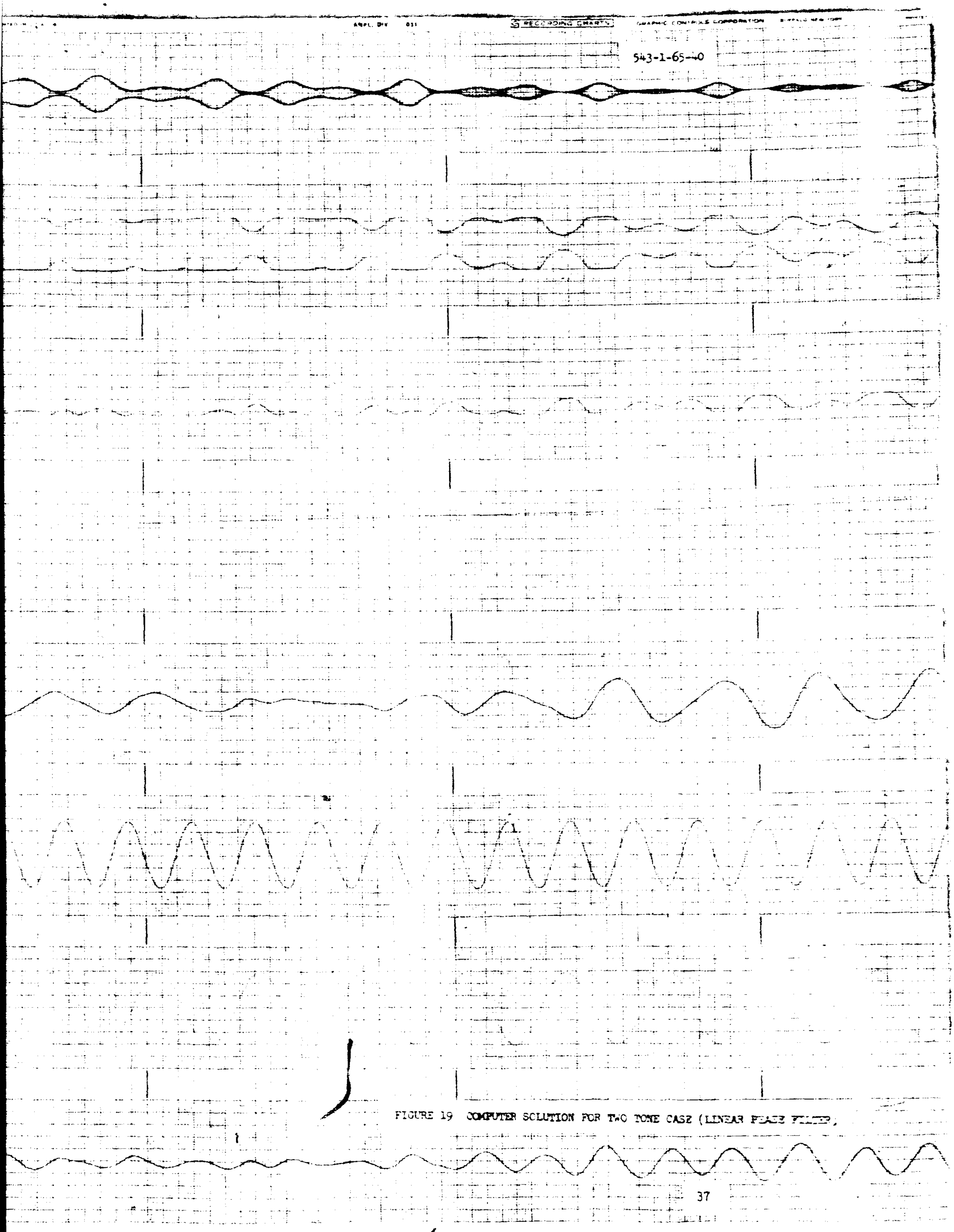


FIGURE 19 COMPUTER SOLUTION FOR TWO TONE CASE (LINEAR PHASE FILTER)

37-5

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

The contributions of Mr. George Cooper, Mr. John Rhamy and Dr. Allan Wilson are acknowledged. Mr. George Cooper set up the analog computer simulation and prepared the material of Section 6 of this report. Mr. John Rhamy contributed the explanation for the natural behavior of FM receivers (with amplitude sensitive discriminators) as described in Section 3. Dr. Allan Wilson assisted in resolving simulation problems and developed the "proof" of FM simulation as included in the Appendix.

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

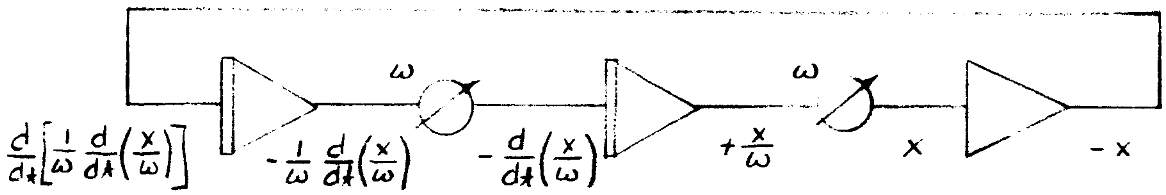
5. Missiles and Space December 1961 by John M.Lambert "GE's Electronic System Evaluator."

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 $\omega = \omega_0 + \Delta\omega \cos pt$.

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 $X = A \sin\left(\omega_0 t + \frac{\Delta\omega}{p} \sin pt\right)$ satisfies the differential equation solved by the analog loop.

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

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

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

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

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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