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APP I

APPENDIX I

FM RECEIVER

Submitted as part of the Final Report

for RF Test Console on JPL

Contract No. 950144

NAS 7-100

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DATE: October 30, 1964

WESTINGHOUSE DEFENSE AND SPACE CENTER

SURFACE DIVISION

ADVANCED DEVELOPMENT ENGINEERING

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1.0 FM Receiver

1.1 Introduction

The FM receiver study was based on JPL Specification No. GPG-15062-DSN. The following list of the unit specifications are applicable to the FM receiver:

3.5.2.3 <u>FM Receivers.</u> The console shall contain a dual FM receiver containing both conventional and phase-lock FM detectors, shown functionally in Figure 6, with the following characteristics:

3.5.2.3.1 <u>Input Bandpass Filter.</u> The dual FM receiver shall contain three (3) easily interchangeable bandpass filters with the following characteristics:

a. <u>Bandwidths.</u> The input filters shall have half-power bandwidths within ± 2 percent of 1 Mc, [illegible], and [illegible] Kc. The filter noise bandwidths shall differ from the half-power bandwidths by no more than 10 [illegible].

b. <u>Amplitude Characteristics.</u> The input filters shall have maximally-flat amplitude characteristics with no more than 0.2 db peak-to-peak ripple across the bandwidth.

c. <u>Phase Linearity.</u> The phase characteristics of the bandpass filters shall be sufficiently linear and symmetrical so as to meet the static and dynamic linearity requirements of 3.5.2.1.2 and 3.5.2.1.3.

3.5.2.3.2 <u>Limiter.</u> The dual FM receiver shall contain a hard limiter following the input bandpass filter with the following characteristics:

a. <u>Dynamic range.</u> The limiter shall have a dynamic range consistent with the characteristics of the signal/noise mixer specified in 3.5.3.

b. <u>Output Waveform.</u> The limiter output waveform shall have delay, rise and fall-times, symmetry, and droop characteristics consistent with the requirements of the FM detectors which follow it.

3.5.2.3.3. <u>Conventional FM Detector.</u> The dual FM receiver shall contain a conventional FM detector with performance characteristics consistent with the frequency stability, and static and dynamic linearity requirements of 3.5.2.1.1, 3.5.2.1.2, and 3.5.2.1.3.

3.5.2.3.4 <u>Phase Lock FM Detector.</u> The dual FM receiver shall contain a modulation-tracking phase-lock FM detector with the following characteristics:

a. <u>Voltage-controlled Oscillator.</u> The voltage-controlled oscillator No. 3 shall have a center frequency of exactly 50 Mc, deviation capability consistent with the transmitter modulator characteristics specified in 3.5.2.2.3b, linearity consistent with the static and dynamic linearity specified in 3.5.2.1.2 and 3.5.2.1.3, and sufficient stability to satisfy the overall transmitter-receiver stability specified in 3.5.2.1.1.

b. <u>Phase Detector.</u> The phase detector bandwidth shall be consistent with the transmitter modulator frequency response specified in 3.5.2.2.3a. It shall be of sufficient fidelity to meet the dynamic linearity requirements of 3.5.2.1.3.

c. <u>Loop Gain.</u> The loop gain shall be sufficiently large such that under conditions of peak transmitter deviation the loop phase error due to finite loop gain will be less than 10° peak. The loop gain stability shall be ± 5 percent over all operating conditions.

d. Loop filter. The loop filter shall be of the passive configuration as shown in Figure 4b. Three (3) standard loop information bandwidths of 3, 30, and 300 kc shall be supplied, with component tolerances as specified in 3.5.1.3.8c. It shall be possible to simple and reliably change the loop filter components in order to operate with any information bandwidth from 100 cps to 1.0 Mc.

e. Compensation Filter. A single-pole RC low-pass filter shall be employed at the loop filter output such that the overall phase-lock FM detector transfer function is that of a "pure" loop. The component tolerances and interchangeability shall be the same as for the loop filter.

3.5.2.3.5 Output Lowpass Filter. The dual FM receiver shall contain a single low-pass output filter with the following characteristics:

a. Response Characteristics. The output filter shall have more than 3 poles and shall have either constant amplitude or linear phase response characteristics manually selectable.

b. Bandwidths. The output filter shall have three (3) standard bandwidths of 1, 10, and 100 kc. It shall be possible to simply and reliably change the bandwidth determining components in order to operate with any bandwidth from 10 cps to 500 kc.

3.5.2.3.6 Offset Frequency. The offset frequency, f_3, shall be exactly 55 Mc such that the output spectrum will be centered at 5 Mc. It shall be easy to replace this reference with an externally supplied frequency.

3.5.2.3.7 Balanced Modulator. The balanced modulator shall be identical to those specified in 3.5.1.3.3.

3.5.2.3.8 AC Isolation Amplifier. The ac isolation amplifier shall be identical to that specified in 3.5.1.3.3.

3.5.2.4 FM Predetection Playback. It shall be possible to easily modify the configuration of the FM receiver, as shown in Figure I, in order to demodulate FM signals previously predetection recorded.

1.2 Functional Block Diagram

A functional block diagram of the FM receiver is shown in Figure (1). The analysis of the various units that make up the FM receiver are listed in the following order: (1) Input BPF, (2) Limiter, (3) Phase lock detector, (4) Conventional FM detector, (5) Output LPF. Other supporting sections concerning the operational parameters of the receiver will then be discussed.

1.3 Input BPF

The required characteristics of the input BPF are stated in specifications 3.5.2.3.1 (a) (b) and (c). Three input RF filters centered at 50 mc whose bandwidths are 1 mc, 100 kc and 10 kc are required. The specifications state that the filter noise bandwidth will differ from the half-power bandwidth by no more than 10 percent, a 3 pole Butterworth filter meets this requirement. A plot of the amplitude and phase characteristics of this filter is shown in figures (2) and (3). The above characteristics are plotted as a function of radian frequency normalized to the radian frequency that corresponds to the half-power value of the amplitude response.

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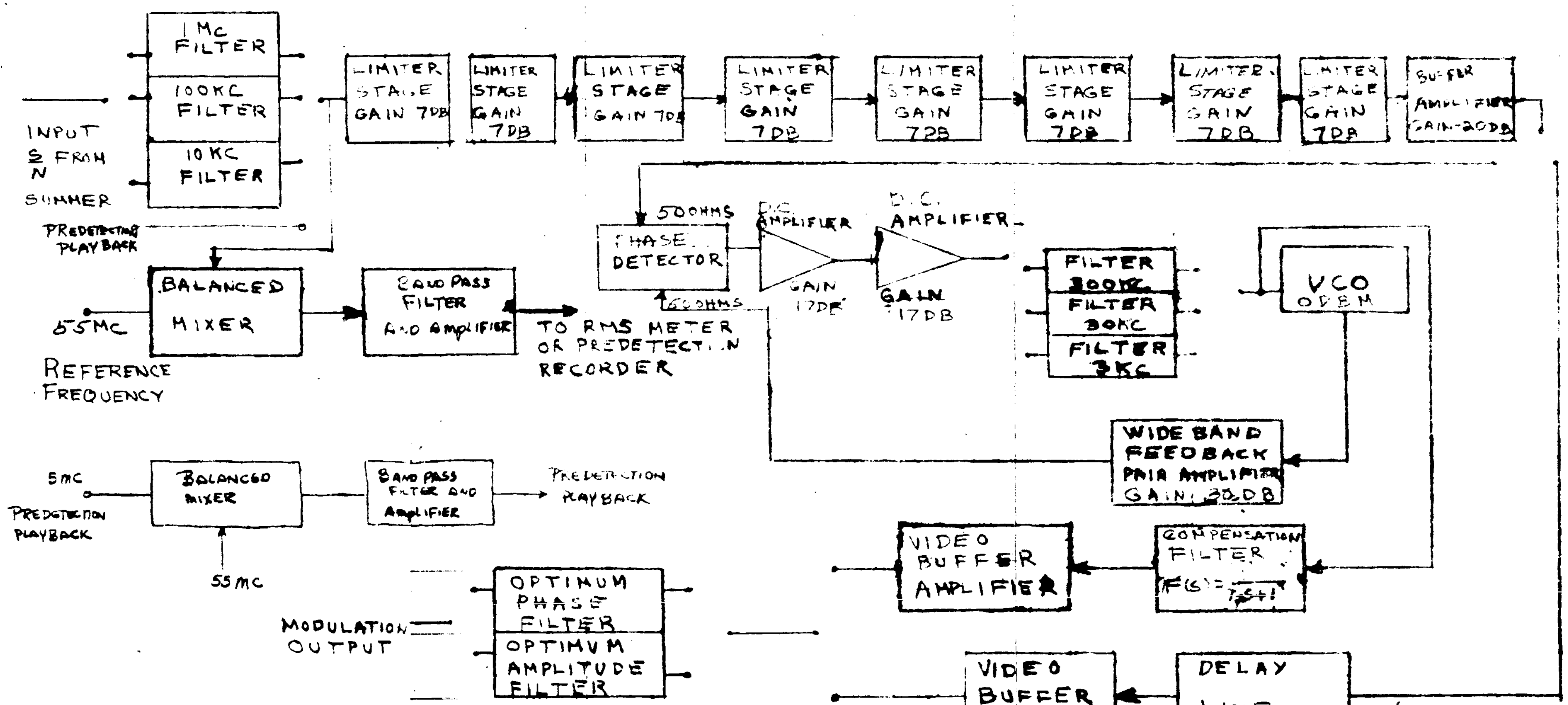
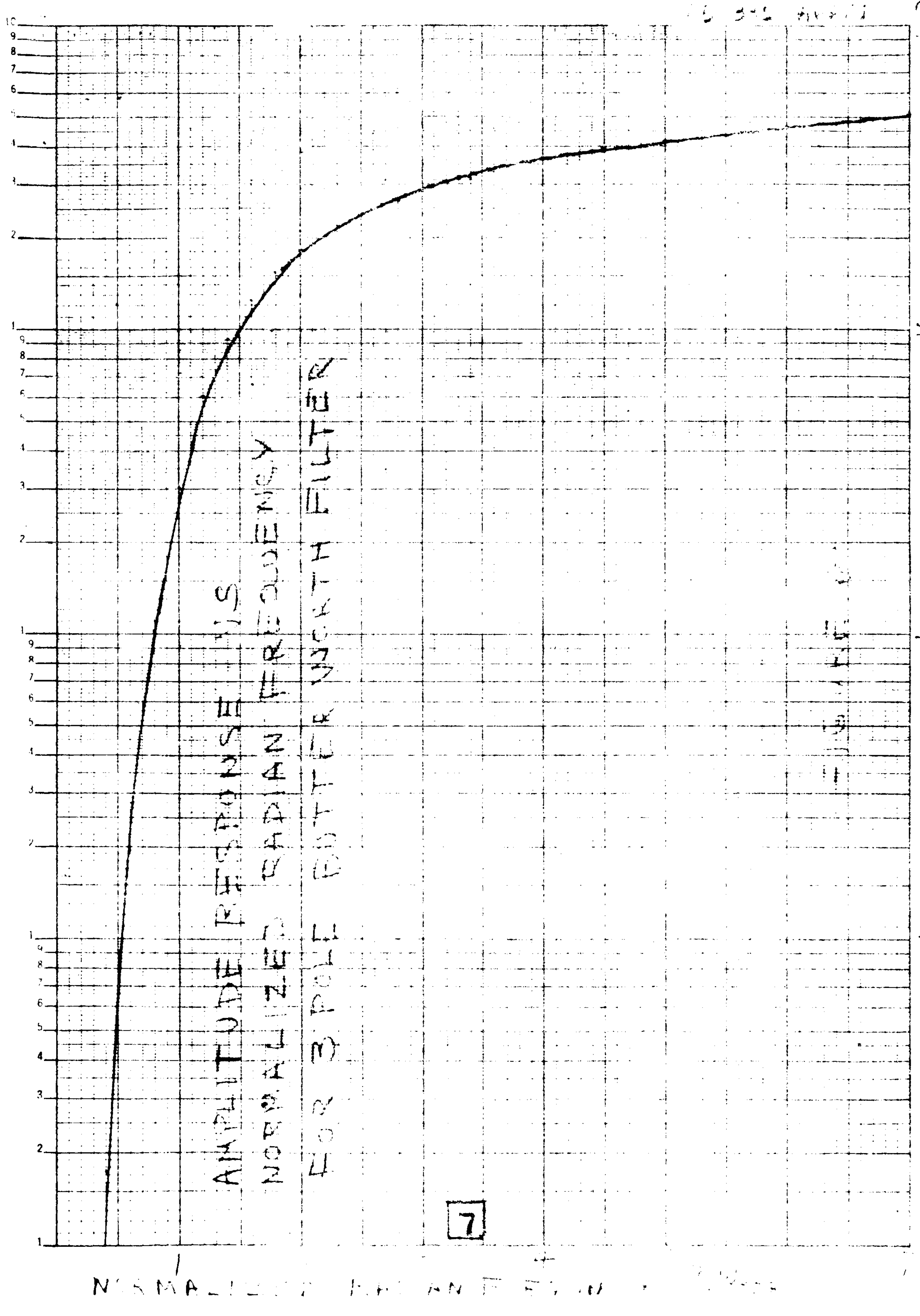
MINIMUM SIGNAL MAXIMUM NOISE 1Mc FILTER	-40DBM --4DBM	-40DB'M -15.8DBM	-33DBM -8.3DBM	-26DBM -1.8DBM	ODBM	ODBM	ODBM	OPBM	ODBM	-26.8DBM ~ODBM	+5.2DBM ~+20DBM
MAXIMUM SIGNAL MINIMUM NOISE 10KC FILTER	-20DBM -50DBM	-20DBM -81.8DBM	-13DBM -74.8DBM	-6TBM -67.8DBM	ODBM	ODBM	ODBM	ODBM	ODBM	ODBM -64.8DBM	+20DBM -44.8DBM
RMS POWER CAPABILITY LINEAR TO 36			ODBM	ODBM	ODBM	ODBM	ODBM	ODBM	ODBM	ODBM	+20DBM

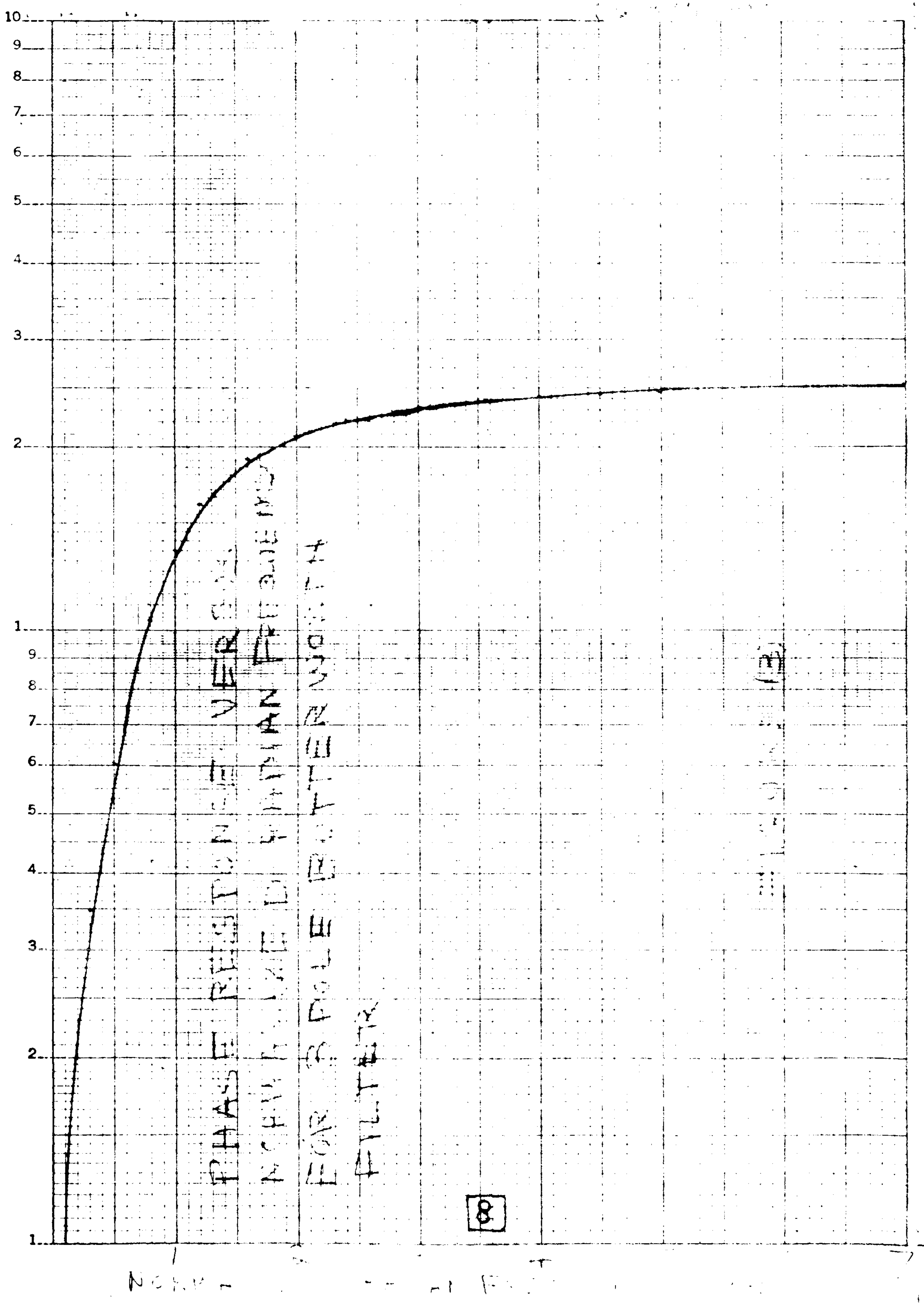


ALL AMPLIFIERS IN IF STRIP HAS BW = 10Mc.
DC AMPLIFIER IN LOOP - BANDWIDTH 10Mc
AC AMPLIFIER IN LOOP - BANDWIDTH - 10Mc
LIMITER STAGE - BANDWIDTH - 10Mc

GAIN OF VIDEO AMPLIFIERS WILL BE ADJUSTED TO OBTAIN A CONSTANT OUTPUT SIGNAL LEVEL FROM EACH OF THE DETECTORS.

Fig. 1 Functional Block Diagram, F.M. Receiver


AMPLITUDE RESPONSE VS
NORMALIZED RADIAN FREQUENCY
FOR 3 POLE BUTTER WORTH FILTER


PHASE RESPONSE VERSUS
NORMALIZED FREQUENCY
FOR 6 POLE BUTTERWORTH
FILTER
10
9
8
7
6
5
4
3
2
1
K+E SEMI-LOGARITHMIC 359-61
KEUFFEL & ESSER CO. MADE IN U.S.A.
2 CYCLES X 70 DIVISIONS

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The 3 pole Butterworth filter will satisfy the filter noise bandwidth requirement; however, the phase linearity of this unit is inferior to the Bessel filter. The stringent requirements of both phase and amplitude are incompatible from the standpoint of filter design. Hence, in order to obtain a maximally flat amplitude characteristic with steep skirts, the phase response of the filter must be degraded. Of course, the effect of this degradation can be minimized by only utilizing an FM wave that has significant sidebands well within the linear phase response of the filter.

The 3 pole Butterworth 1 mc filter can be instrumented by employing 3 helical type resonators. Its approximate size will be 1" x 1-3/4" x 3". Its expected insertion loss will be 2 db. The 100 kc and 10 kc filters will be instrumented by employing 3 or 4 single crystal lattice sections. These elements will give the required spurious suppression and amplitude characteristics. The overall dimensions for these units will be 1" x 1" x 2" and the upper limit on their insertion loss will be 3 db.

The bandpass filters described above can distort the impressed FM wave by either possessing a non-uniform amplitude response or a non-linear phase response. Intermodulation distortion is caused by the former factor because this effect will tend to attenuate some of the significant sidebands of the FM wave. The number and total frequency width of these sidebands are a strict function of the modulating frequency fm, and the peak frequency deviation, Δf, (Modulation index of FM wave, M is equal to $\Delta f/fm$). A general formula that is frequently employed to determine the required frequency width,[1] of the filter is as follows:

(1) $$B = 2\,fm\,(1 + M + \sqrt{M})$$

A more accurate estimation of bandwidth required to pass all sideband components exceeding 1% of the unmodulated carrier can be obtained from the FM spectrum for a given M and fm. A plot of the required value of B for the above condition is shown in figure (4). As shown, for a Δ f of 500 kc and an M of 1, the required value of B that is necessary to hold the distortion below the 1% value is 3 mc. Also, as M increases past the value of 6, the required value of B assymptotically approaches 2 Δ f. Specifications 3.5.2.2.3 (a) and (b) state that the transmitter-receiver pair must be capable of operating with a Δ f of 500 kc and an fm of 500 kc. However, if an FM wave with the above parameters are passed through the 1 mc filter, an output wave with 11% distortion will result. This value follows from the fact that the second sideband of the FM wave would be rejected by the filter. This sideband's amplitude with respect to the amplitude of the unmodulated carrier is .11.

As previously stated, a non-linear phase response of the filter will also cause distortion in the output FM wave. The distortion from this effect can be minimized by utilizing FM waves whose important sidebands fall within the linear phase response of the filter. The phase characteristics of a filter is usually given in terms of the group delay, To. Its relations with the actual phase shift and frequency of the FM wave can be stated as follows:

(2) $$B = T_o W + n \pi$$

where B is the phase shift

W is the radian frequency

n is a constant.

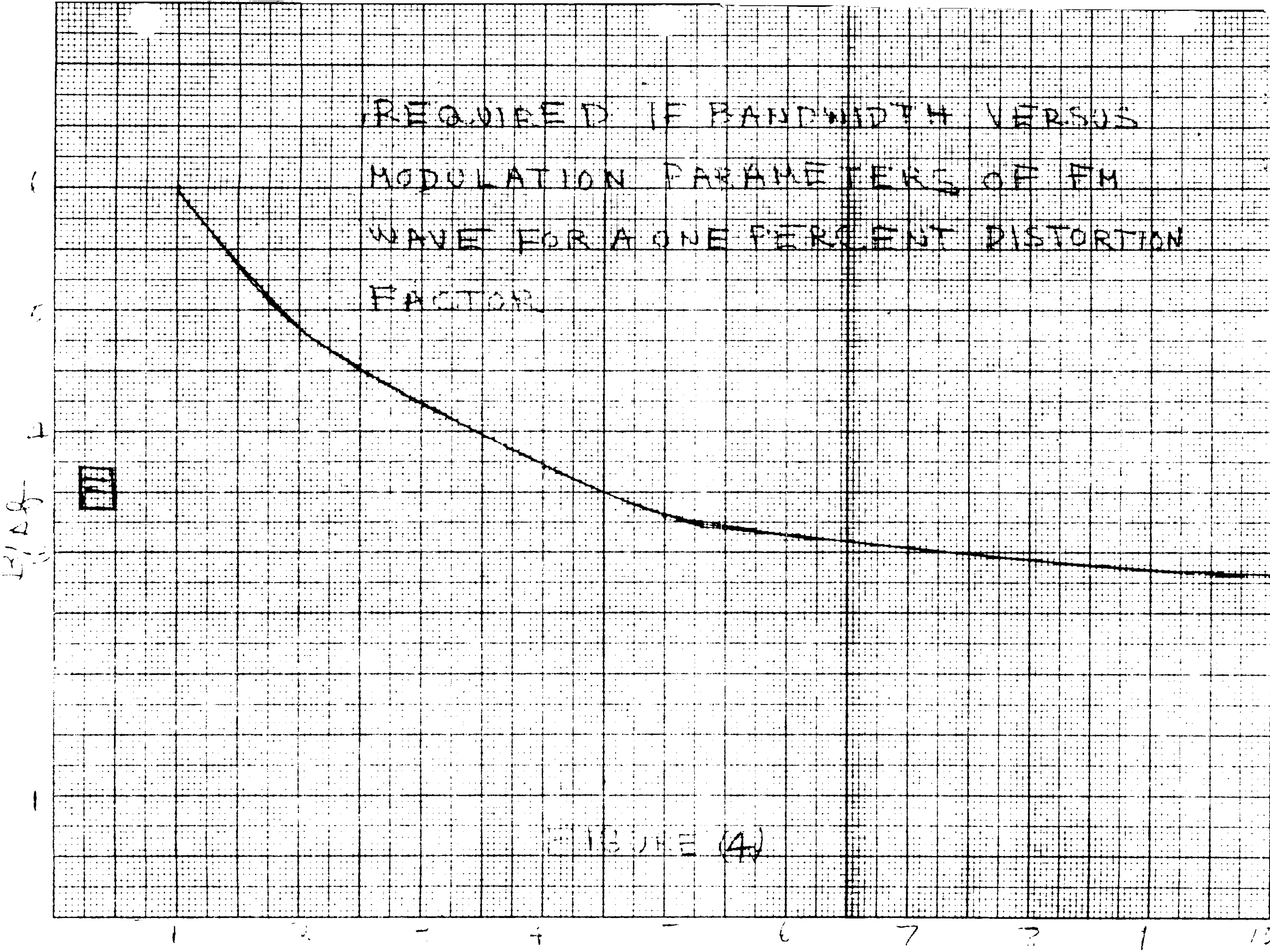
REQUIRED IF BANDWIDTH VERSUS
MODULATION PARAMETERS OF FM
WAVE FOR A ONE PERCENT DISTORTION
FACTOR

FIGURE (4)

From equation (2) it follows that $\frac{dB}{dW} = T_o$.

It follows that if dB/dW is constant across the frequency band of interest, then the phase shift B also varies linearly across this same band. Figure (5) is a plot of T_o vs. the normalized radian frequency ω/ω_{3DB} for a 3 pole Butterworth filter. As shown, T_o stays relatively constant over the frequency interval of $0 < \omega/\omega_{3DB} < .2$. Hence, from the phase standpoint the upper frequency limit on the three filter values are 10^5, 10^4 and 10^3 cps.

Signal fidelity degradations due to the filter's phase characteristic can be essentially eliminated by utilizing a passive equalizer at the output of the filter. This device has an all pass characteristic with respect to amplitude and a phase characteristic that is the reciprocal of the filter's phase response. This device will be required if the significant sidebands of the FM wave exceed the frequency limits stated above. Since specification 3.5.2.2.3 (a) and (b) only state the upper limits on the value of Δf and fm, no conclusions regarding the use of a phase equalizer can be made until intermediate values are chosen by JPL. In fact, the question as to whether the specification 3.5.2.1.3 concerning dynamic linearity can be fulfilled with the type of filters required to fulfill specification 3.5.2.3.1 (a) will mainly depend on the actual operational values of Δf and fm and not on the maximum values stated.

[illegible].4 Limiters

The required operating characteristics of the limiter unit is given in specification 3.5.2.3.2. In order to yield a wide operational signal range for the conventional and phase lock detectors, an input S/N range of -36db to +30 db was chosen from the S/N summer. The limiting

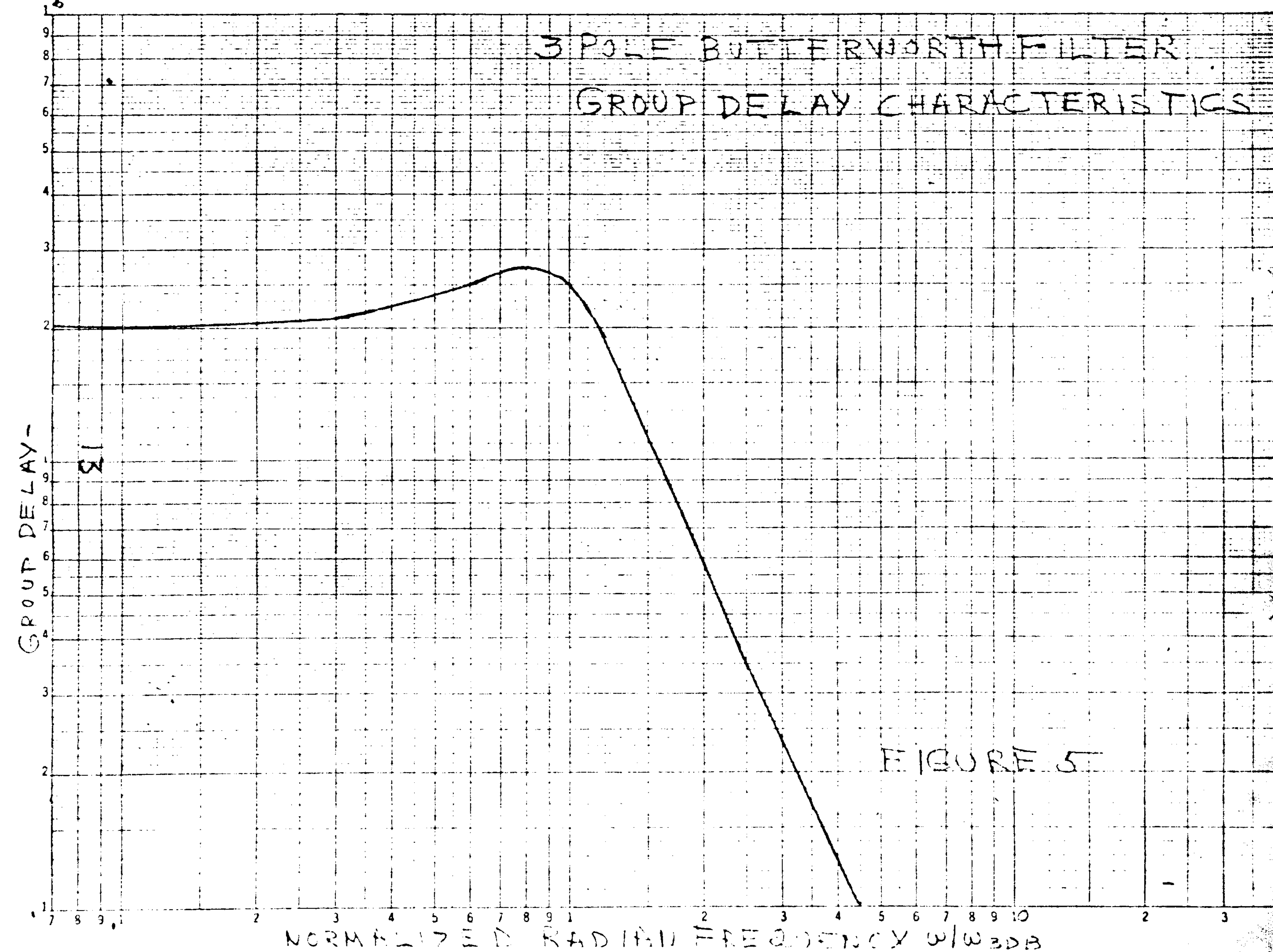

FIGURE 5

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factors for this range being the maximum attainable value of 170 db for the open loop gain of the phase lock detector. The above range will guarantee a minimum S/N input of -25.2DB to -4.7DB depending on the operational characteristics of the phase lock detector and the particular input filter that is employed. Also a maximum S/N range of +44.8 db to 64.8 db for the above conditions is also possible. With this wide possible operational range for the FM receiver, all practical cases for testing the detectors are present; hence, the need for the 100 db dynamic range for the limiters are given in specification 3.5.2.3.2 (a) isn't required.

The limiter preceding the phase lock detector is necessary in order to insure a relatively stable loop gain over the desired dynamic range of signal-to-noise ratio values ($\frac{S}{N}$) at the input to the limiter. Another advantage in the use of limiters is for the reduction of the impressed AM modulation on the input FM wave. This latter characteristic reduces the signal distortions in both the phase lock and conventional detectors.

The work of Davenport[2] on the limiter's effect on the S/N is useful in designing the phase lock detector. It was found that for input (S/N) into the limiters of less than -4 db, a reduction in the output (S/N) takes place. The amount of the reduction gradually increases as the input S/N decreases until it assymptotically approaches a limiting value of /4 for a (S/N) of -10 db. For a S/N input greater than -4 db a gain in the output S/N is obtained. This gain gradually increases until it assymptotically approaches a limit of 3 db at an input S/N of 10 db or greater. The above conclusions by Davenport were reached by assuming a single frequency input and an output filter that rejected all of the generated harmonics caused by the limiter. However, for a modulated input signal, the (S/N) out of the limiter will be higher than that predicted by Davenport since the main noise contribution resulting from the cross product of S x N will be further removed from the center frequency. Hence, this noise contribution will suffer more attenuation by the filter. In the unmodulated signal case, the SXN noise term will essentially fall within the passband of the filter since the signal spectrum is only one line. Also in computing the output S/N terms, Davenport included all of the noise terms that extended over infinite frequency limits. This was done for mathematical convenience. Actually, the output noise values are only those passed by the output filter; therefore, it follows that the actual (S/N) out will be higher than the theoretical value.

The transfer function of any active unit, in general, can be given by the expression $y = K x^{1/n}$, where y and x are the input signals and k is merely a constant. n is defined by the type of unit that is under consideration. For a perfect limiter, n = ∞. Of course, this performance isn't obtained in practice because there always exists a finite operational threshold region. Also, a practical limiter will only be effective over a given input power range of S + N values.

The required dynamic range over which the limiter must operate will essentially depend on the minimum desired value of S/N that is below the threshold level of the phase locked detector and the maximum value of S/N that can be obtained from the S/N summer. The threshold level of the phase lock detector is that value of S/N that is defined in the closed loop noise bandwidth, 2 BLC of the detector. The total possible variation in the output S/N from the S/N summer is +30 db to -70 db defined in a 15 mc bandwidth. The above range is obtained by a noise source that yields an output power range of 0 to -50 dbm. [illegible] the signal source produces a power output range of -20dbm to -[illegible] dbm.

As previously stated the S/N range that will be employed for the FM receiver is -35 db to + 30 db defined in the 15 mc noise bandwidth of the S/N summer. In order to obtain the above range, the signal power range will be taken between -20 dbm to -40 dbm. The noise power range will be taken as -4 dbm to -50 dbm. For the S/N limits defined above, figure (6) shows the various limits that can be obtained for the three standard input filters and the values of 2 BLo computed from the required loop information bandwidths given in specification [illegible](d). As showing the values of 2 BLo are 2 mc, 200 kc and 20 kc. The computation of these values will be discussed in the next section. Also it is noticed in figure (6) that in the case of BIF [illegible] 2BLo, the operational values of S/N depends only on the bandwidth ratio of 15 mc to the particular input filter bandwidth. Also, figure 7 is a plot of the corresponding values of the signal and noise powers chosen versus the output S/N for the summer.

At the input to the limiter, the lower S/N value as shown in figure (6) is -24.2 db. Referring to figure (7), it is seen that the corresponding signal and noise power levels are -40 dbm and -15.8 dbm. An upper limit for the input S/N to the limiter is 61.[illegible] db. The corresponding signal and noise powers for this case are -20 dbm and -81.8 dbm.

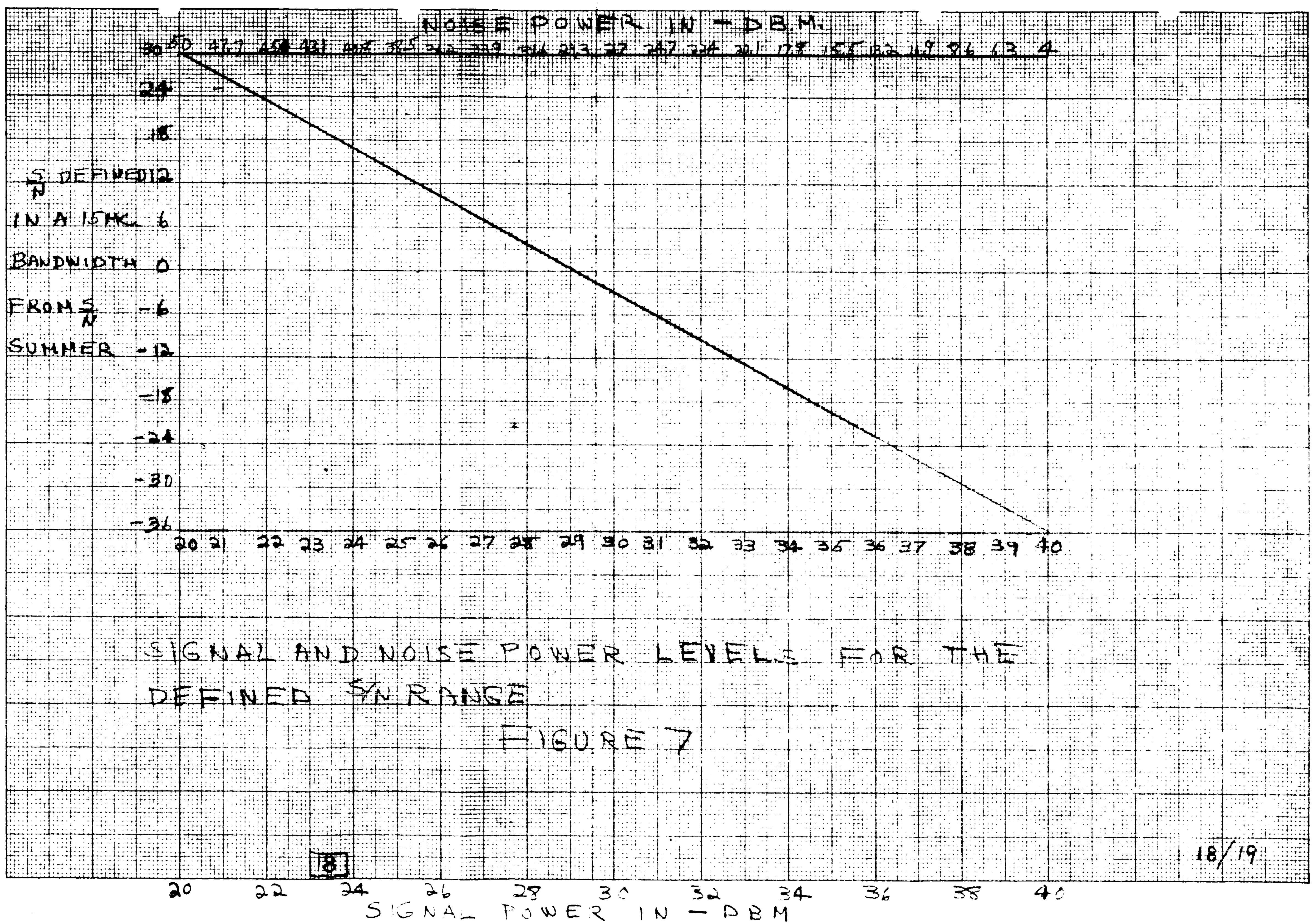
In order to obtain efficient limiting action on all possible variations of signal level, 10 db of band limiting for the lowest signal level should be obtained. Also, since the AM modulation will vary the signsl amplitude level by a 6 db value, it follows that a 1[illegible] db limiting level should be provided to cover all possible signal variations. Since the

OPERATIONAL RANGE FOR S/N VALUES

S/N IN 1MC	LIMITER FACTOR	S/N IN $2BL_0$=2MC	S/N IN $2BL_0$=[illegible]	S/N IN $2BL_0$ = 20KC
-24.2DB +41.7DB	-1DB +3DB	-25.2DB +44.8DB	-18.2DB +51.7DB	-8.2DB +61.7DB
S/N IN 100KC -14.2DB +51.8DB	-1DB +3DB	-1[illegible].2DB +[illegible]4.8DB	-15.[illegible]DB +[illegible]4.[illegible]DB	-7.2DB +61.7DB
S/N IN 10KC -4.2DB +61.7DB	-.5DB +[illegible]DB	-4.[illegible]DB +64.7DB	-4.7DB +64.8DB	-4.7DB +64.8DB

OVERALL S/N RANGE FROM SUMMER IS +30DB TO -36 DB DEFINED IN A 15Mc BANDWIDTH.

FIGURE 6



SIGNAL AND NOISE POWER LEVELS FOR THE DEFINED S/N RANGE

FIGURE 7

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minimum signal level is -40 dbm and 16 db of limiting is required for this value, it follows that 56 db of gain is required from the limiter unit.

For the above value an 8 stage tunnel diode limiter will be employed. This type of limiter has already been developed by Westinghouse. The output signal level of the limiter will be set at 0 dbm. Also each of the limiter stages will have a gain of 7 db. For an input signal level of -40 dbm, 16 db of limiting will occur in the last 3 stages of the limiter. At a maximum signal level of -20 dbm, 36 db of limiting will occur in the last 6 stages. The maximum noise power at the input to the limiter will occur at the S/N value of -24 db for the 1 mc input filter. Since the maximum noise power is -15.8 dbm, 40.2 db of limiting will result. In order to obtain good operational drive levels for the two FM detectors over all possible signal levels, a wideband feedback pair amplifier with a power gain of 20 db will be inserted at the output of the limiter. This unit is particularly necessary for the case of the phase lock detector when it is operating at the -40 dbm input signal level. In the case of the higher signal levels the 20 db gain factor could be reduced if it is desired.

In order to minimize signal distortion in the limiter, it is necessary that the bandwidth of each limiter stage be at least five to ten times the bandwidth of the signal. The above distortions occurs due to the deterring effect on the tuned circuits in the limiter cause by changes in the reactive and resistive impedance level of the limiting units. The variation in the resistive portion of the diode impedance tends to change ths shape of the response of the tuned circuit. The variation in the reactive impedance of the diode causes variations in the resonant frequency of the tuned circuit. It follows that if the bandwidth of the limiter stages are broad, the percentage change from the above effect will be low. A tunnel diode type of limiter will be employed, bandwidth values of 10 mc will be easily obtained. The 10 mc value is chosen from the highest input filter bandwidth value stated in specification 3.5.2.3.1 (a).

An upper limit on the value of the bandwidth is set by the requirement that the adjacent harmonic components of the limited signal be rejected. Since the operating frequency is 50 mc and the peak frequency deviation is 500 kc, a unit limiter bandwidth of 10 mc will satisfy the harmonic rejection requirements.

Instead of cascading limiters to equalize the 26 dbm input power differential, a combination AGC system and a smaller number of cascaded limiters could be employed. This latter technique is usually employed when the bandwidth of the input wave is high (about 10 to 20 mc). In this case, the AGC system is necessary because of the requirement that the unit limiter's bandwidth be about 10 times the value of the bandwidth of the input wave. Since the overall limiter section must produce a certain amount of gain in order to efficiently drive the detectors, a very large bandwidth would require a prohibitively large number of limiter stages. However, in the case under consideration, the 10 mc limiter stage bandwidth doesn't necessarily require the use of the AGC system. For the case of a relatively low bandwidth for the input signal, the method that is employed is optional.

1.5 Phase Lock Detector

As stated in specification 3.5.2.3.4, the FM receiver shall contain a modulation tracking phase-lock FM detector. The linearized version of the PLL shown in Figure (8) of the transmitter section of the report will be utilized to determine the required operating parameters of the loop. Specific values for the above parameters will be obtained from the JPL specification 3.5.2.3.4 (a) (b) (c) (d) and (e).

The three main parameters that must be defined for the phase lock detector are the required values of the loop information bandwidth, f_n, and the minimum value of the open loop gain G_o. As stated in specification 3.5.2.3.4 (d) the values of f_n are 3, 30 and 300 kc. The required value of G_o is determined by the static deviation value and the peak permissable phase error stated in specification 3.5.2.3.4 (e). Hence it follows that

$$G_o \qquad \frac{2\pi \Delta f}{.1745} \tag{3}$$

For a Δf of 500 kc it follows that G_o is 18×10^6 radians/sec.

The signal suppression factor α is proportional to the carrier voltage at the input of the limiter. The above factor is related to the input and output S/N of the limiter by the following expression:

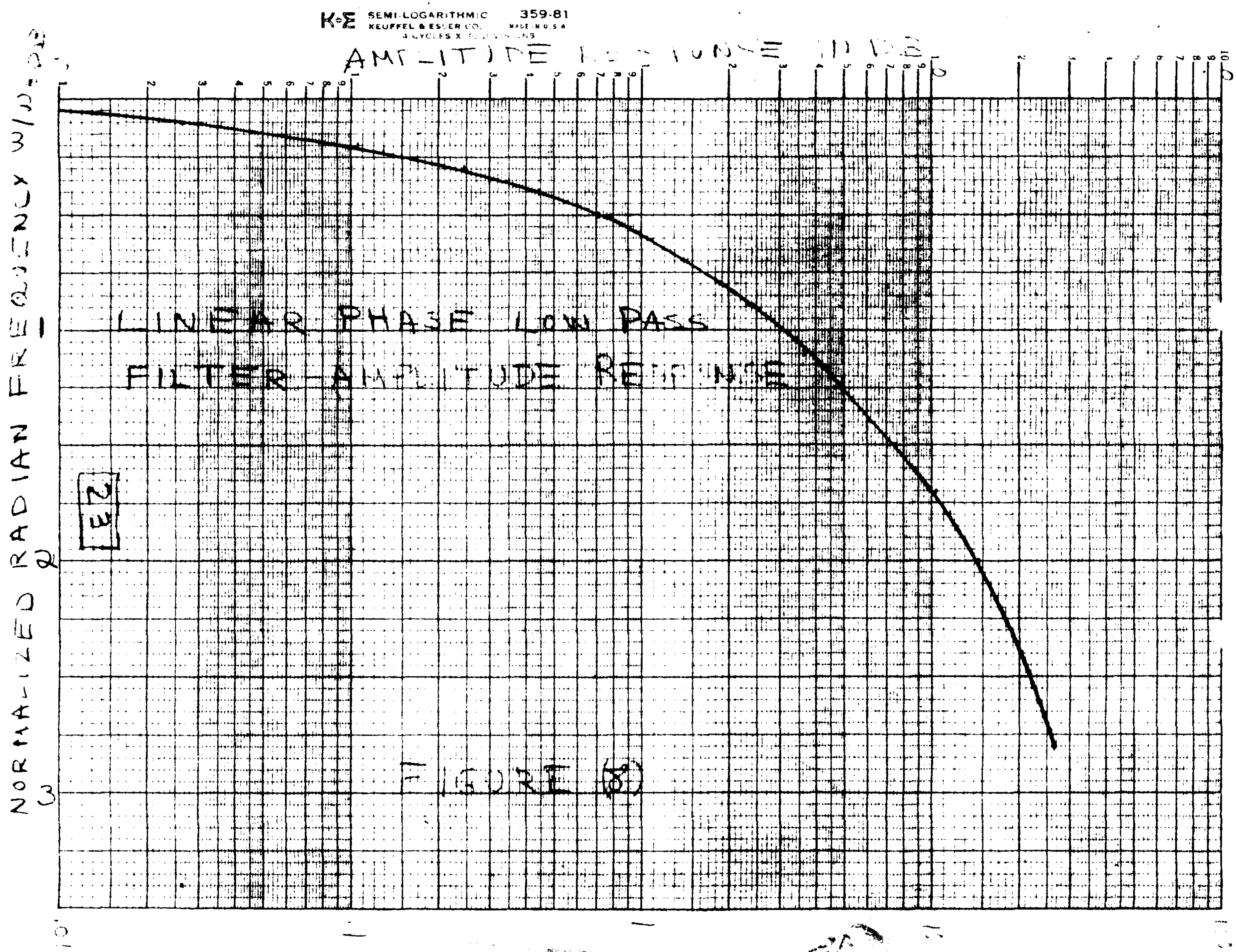
K+E SEMI-LOGARITHMIC 359-81
KEUFFEL & ESSER CO. MADE IN U.S.A.
AMPLITUDE
NORMALIZED RADIAN FREQUENCY
LINEAR PHASE LOW PASS
FILTER AMPLITUDE
23

FIGURE 8

$$\alpha^2 = \frac{1}{1+\left(\frac{N}{S}\right)_{OUT}} \sim \frac{1}{1+\frac{4}{\pi}\left(\frac{N}{S}\right)_{IN}} \qquad (4)$$

Since the input S/N values vary from -24.2 db to +61.8 db it follows from the above equation that α will very from .057 to 1. Hence, the open loop gain G will vary by a factor of 17.54 or 24.88 db. Since the maximum open loop gain G is stated to G_0 by the expression $G_0 = \alpha_0 G$ it follows that G is equal to 3.16×10^6 or 170 db.

In order to determine the characteristics of the loop filter, the various terms for the PLL defined in the transmitter section of the report will be employed. From equation (3) of this section it is shown that

$$H(s) = \frac{G_0 F(s)}{s + G_0 F(s)} \qquad (5)$$

The type of loop filter that will be employed is stated in specification 3.5.2.3.4 (d). It has a transfer function F(S) of $\tau_2 s + 1/\tau_1 s + 1$. Hence in inserting this value in equation (5) it follows that

$$H(s) = \frac{\tau_2 s + 1}{\frac{\tau_1}{G_0} s^2 + \left(\frac{1}{G_0} + \tau_2\right) s + 1} \qquad (6)$$

In relating equation (6) to the general expression for a second order servo it follows that

$$\tau_1 = \frac{G_0}{\omega_n^2} \qquad (7) \qquad\qquad \tau_2 = \frac{2\xi}{\omega_n} - \frac{1}{G_0} \qquad (8)$$

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Utilizing the above equations with the given values of f_n of 3×10^5 cps 3×10^4 cps and 3×10^3 cps with a G_o of 18×10^6 rad/sec. it is possible to compute the required values of τ_1 and τ_2 for an ξ of .7. Table I shows the result of these computations. In addition the value of f_{τ_2} is given to show its proximity to f_n for loop stability purposes.

TABLE (I)

G_o (rad/sec)	f_n (cps)	τ_2 (μ sec)	τ_1 (μ sec)	f_{τ_2} (cps)
18×10^6	3×10^5	.638	5.07	2.31×10^5
18×10^6	3×10^4	7.37	507	2.116×10^4
18×10^6	3×10^3	74.24	50700	2.14×10^3

As stated in the transmitter section of the report, the maximum open loop gain G is equal to $2 \pi K_m K_{vco} K_\alpha$. The required value of G is 170 db or 3.16×10^8 radians/sec. For an input S/N to the limiter of +10 db or greater the value of K_m will be 2 volts/radian. The value of K_{vco} is determined by the slope characteristics of the VCO. For the unit chosen, it will be 5×10^5 cps/volt. Therefore, it follows that the required value of K_α is 50.3 or 34 db. In order to realize this value, 2 d.c. amplifiers each with a gain of 17 db each will be employed.

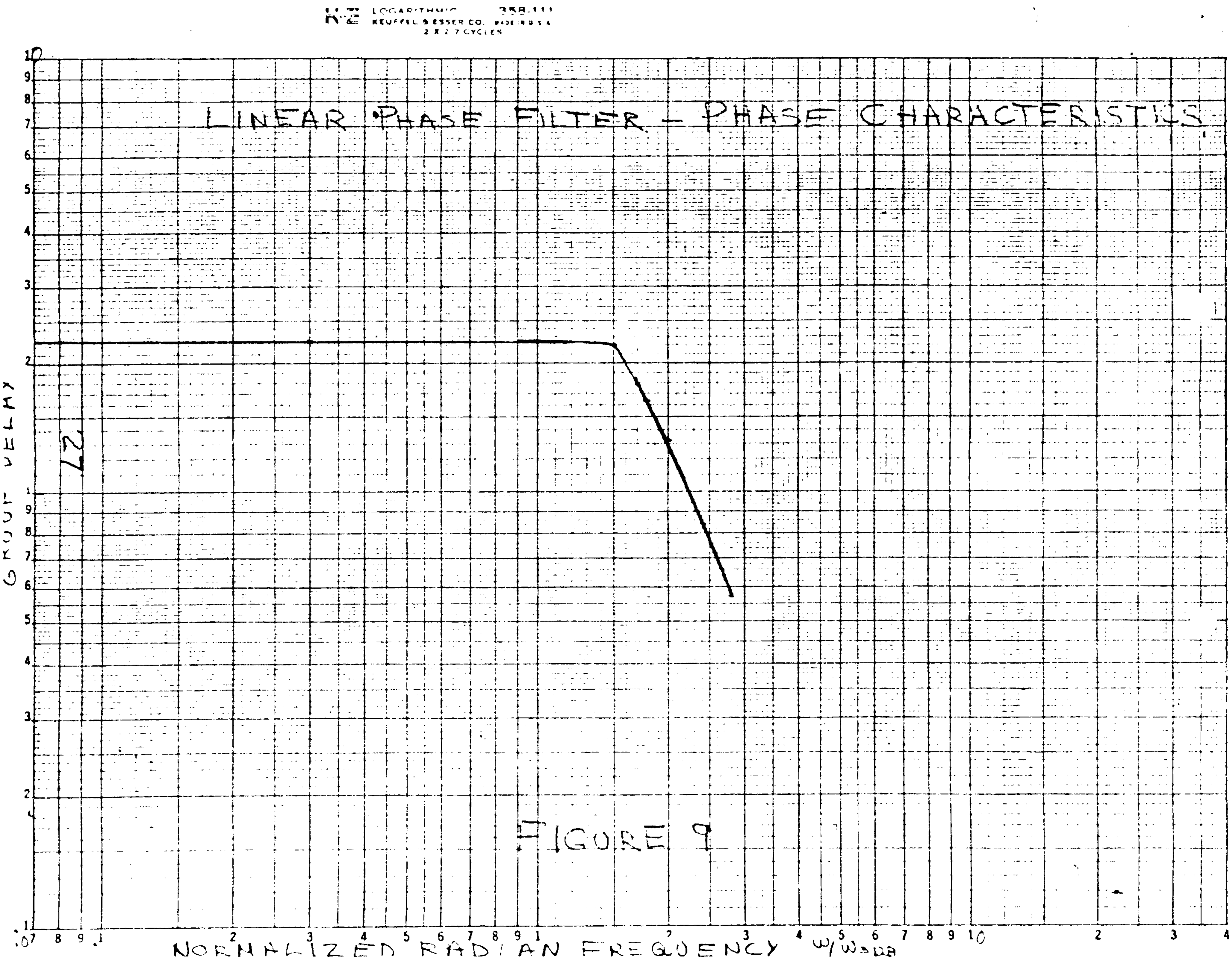
As the input S/N to the limiter decreases the corresponding value of K_m will also decrease. Hence for an input S/N of -24.2 db the value of K_m will be .114 volts/radian. For a high input S/N to the limiter the input signal level to the phase detector will be +20 dbm. This corresponds to a input voltage level of 2.24 volts rms for an input impedance level of 50 ohms. Since the suppression minimum input signal voltage level that the detector must handle is .128 volts rms.

cation 3.5.2.3.4.). For this case, the filter : ' have a transfer function of the form $1/\tau_2 s + 1$. The time constant, τ_2, is identical to the values computed for the specified loop filter. Hence, as the loop information bandwidth, fn, is varied the values of τ_2 must also be changed to comply with the desired operating conditions.

1.6 Output Low Pass Filters

The requirements for the low pass filters are stated in specification 3.5.2.3.5 (a) and (b). Linear four pole filters with bandwidths of 1, 10 and 100 kc will be employed to satisfy the above requirements. In the case of the linear phase requirements, a four pole equiripple phase approximation will be employed. For the constant amplitude requirement, a four pole Chebyshev magnitude approximation will be employed. The amplitude and group delay factors for the optimum phase filter is shown in figures (8) and (9). For this case, a .05° phase variation is encountered up to the normalized frequency value of 1.5 W/W3DB. As the frequency increases past this value, the phase characteristics of the filter deteriorates quite rapidly.

For the optimum amplitude filter, a plot of the group delay versus the normalized radian frequency is shown in figure (10). The amplitude characteristics for this filter is given in Table II.



LINEAR PHASE FILTER - PHASE CHARACTERISTICS
GROUP DELAY
NORMALIZED RADIAN FREQUENCY ω/ω3dB
K&E LOGARITHMIC 358-111
KEUFFEL & ESSER CO. MADE IN U.S.A.
2 X 2 7 CYCLES

FIGURE 9

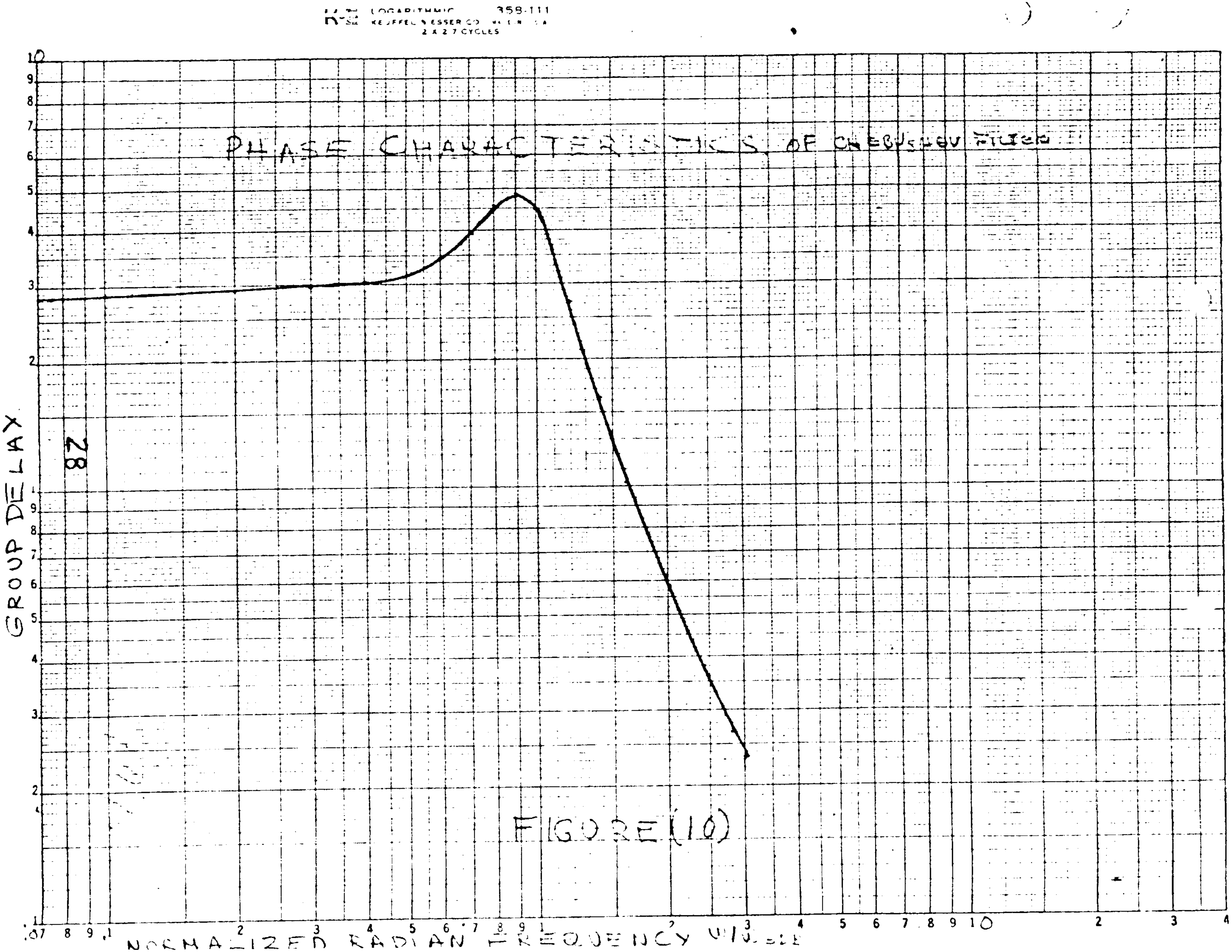

LOGARITHMIC 358-111
KEUFFEL & ESSER CO.
2 X 2 7 CYCLES
PHASE CHARACTERISTICS OF CHEBYSHEV FILTER
GROUP DELAY
NORMALIZED RADIAN FREQUENCY
82

FIGURE (10)

Frequency W/W_{3DB}	Amplitude (-DB)
.02	.00986
.06	.0088
.08	.0079
.1	.0069
.2	.0014
.3	.0062
.4	.0065
.5	.0098
.6	.0016
.7	.0021
.8	.258
.9	1.13
1	3.01
1.2	8.74
1.4	14.57
1.6	19.68
1.8	24.12
2	28.03
2.4	34.7
2.8	40.25

TABLE II Output Filter Response

As shown in the above table, the amplitude response has a ripple factor of .01 db within the frequency limit of W/W_{3DB} of .7. Past the above frequency value, the amplitude factor deteriorates quite rapidly.

3.7 Conventional FM Detector

The requirement for a conventional FM detector is stated in specification 3.5.2.3.3. The various operating characteristics that must be considered in choosing the type of discriminator to employ are as follows:

a. Required static and dynamic linearity factors

b. Operating requirements of the driving unit.

c. Distortion due to imbalance in the discriminator

d. Instrumentation factors.

Taking the above factors into consideration, the choice of the type of unit to employ essentially narrows down to a high or low slope discriminator. The advantages of the former type is essentially that it requires less input drive power and the AM rejection requirements on the limiter unit is less severe. Also, the effects of imbalance from the standpoint of output distortion is not as severe for this type of unit. The advantages of the latter type is that it has better linearity characteristics and the components required to instrument the unit are more adaptable to the required 50 mc. operating frequency. Since stringent linearity requirements are imposed on this unit by specifications 3.5.2.1.2 and 3.5.2.1.3, a low slope delay line frequency discriminator will be employed. This type of discriminator has been built and tested by Westinghouse. A static linearity factor, in the region of the center operating frequency of .05% has been obtained. A circuit diagram of this unit is shown in figure (11). The phase relationships at the various points along the delay line is shown in figure 12.

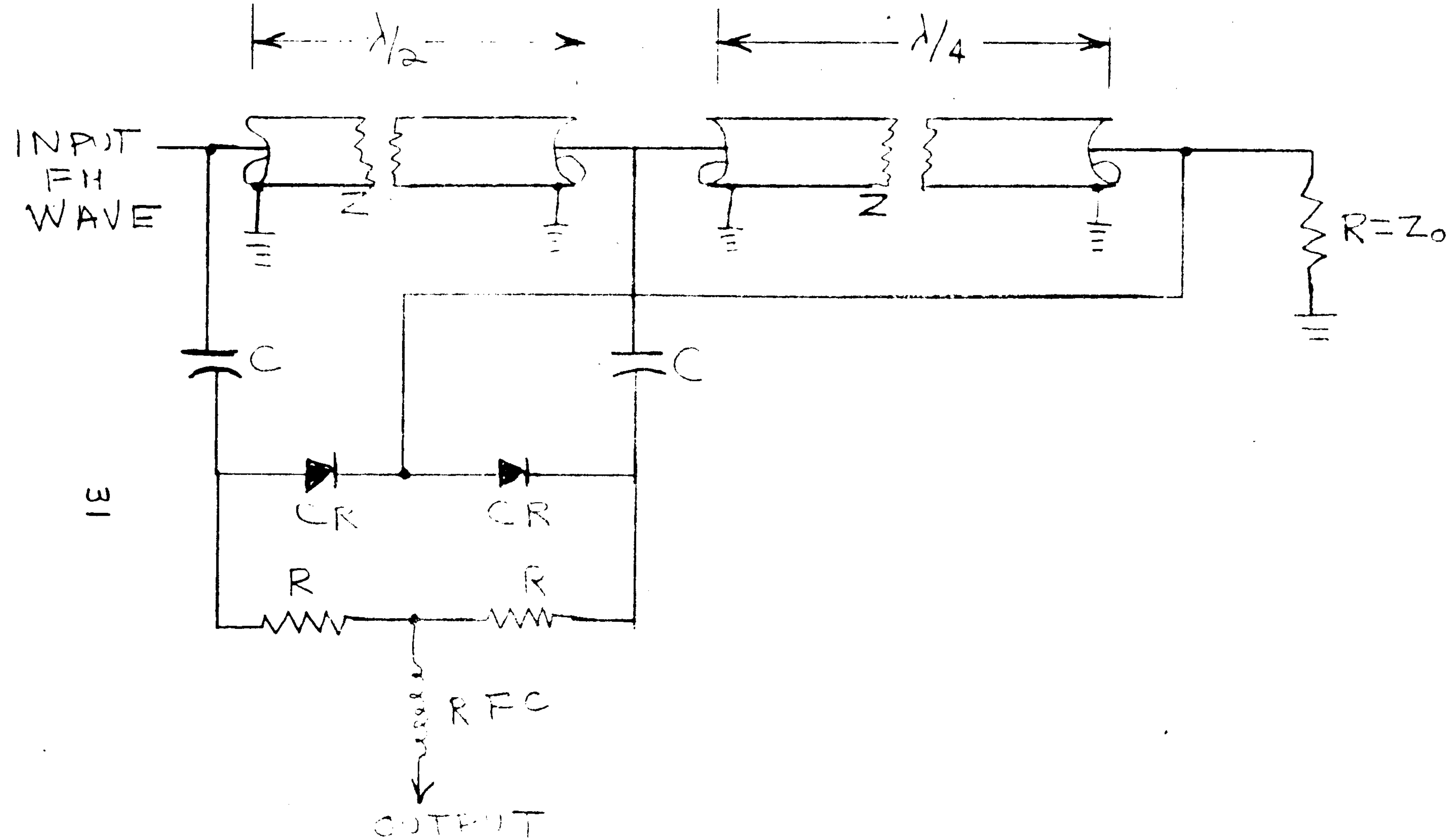


FIGURE (11)

DELAY LINE F [illegible]

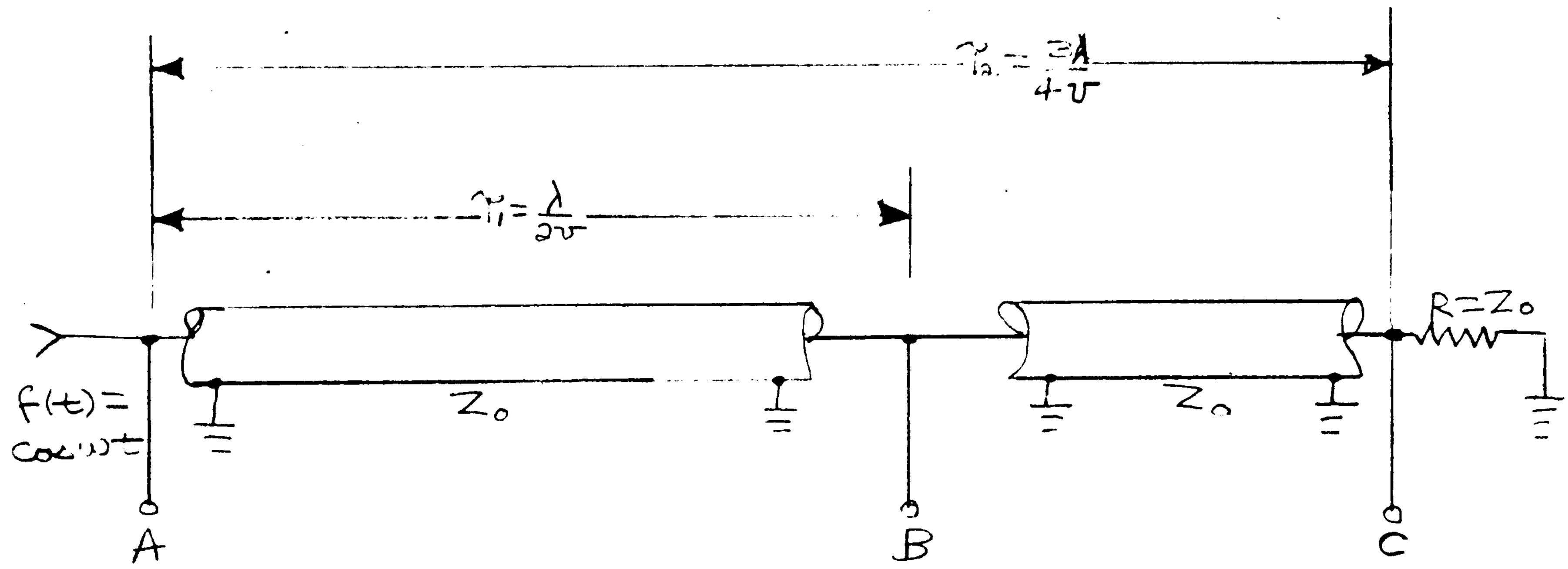


FIGURE (12)

DELAY LINE phase

As shown in figures 11 and 12 both phase splitting and the development of quadrature voltage is accomplished by means of a delay line. The $\lambda/2$ section of line provides out of phase "sampling" voltages while the $\lambda/4$ section provides the quadrature voltage to be sampled. The peak detecting or sampling diodes are connected in a well known phase detector fashion.

The discriminator characteristic is computed for sine wave inputs using the information of figure 12. The phase shift at points B & C are:

$$(12) \quad \phi_B = \pi f/f_0 \qquad \phi_C = \frac{3}{2}\pi f/f_0$$

moreover, the voltages at A, B, & C may be written

$$(13) \quad V_A = \cos \omega t$$

$$(14) \quad V_B = \cos(\omega t - \phi_B)$$

$$(15) \quad V_C = \cos(\omega t - \phi_C)$$

The first peak detector will have an output proportional to the difference between V_A and V_C.

$$(16) \quad V_{A-C} = \cos(\omega t - \phi_B) - \cos(\omega t - \phi_C)$$

While the second peak detector will have an output proportional to the difference between V_B and V_C

$$(17) \quad V_{B-C} = \cos(\omega t - \phi_B) - \cos(\omega t - \phi_C).$$

Equations (16) and (17) may be reduced to the forms

(18) $$V_{A-C} = -2 \sin\left(\omega t - \frac{\Phi_C}{2}\right) \sin \frac{\Phi_C}{2}$$

(19) $$V_{B-C} = -2 \sin\left(\omega t - \frac{\Phi_B}{2} - \frac{\Phi_C}{2}\right) \sin\left(\frac{\Phi_C}{2} - \frac{\Phi_B}{2}\right)$$

Thus one detector will have an output nearly equal the peak value of V_{A-C} (V_1) and the other a value near to the peak value of V_{B-C} (V_2).

(20) $$V_1 = \sin \Phi_C / 2$$

(21) $$V_2 = \sin\left(\frac{\Phi_C}{2} - \frac{\Phi_B}{2}\right)$$

The detector outputs are of opposite polarity being summed with equal weight giving an output proportional to $V_1 - V_2$.

(22) $$V_1 - V_2 = K\left[\sin \frac{\Phi_C}{2} - \sin\left(\frac{\Phi_C}{2} - \frac{\Phi_B}{2}\right)\right]$$

Using the fact that $\phi_C = \frac{3}{2}\phi_B$, $V_1 - V_2$ becomes

(23) $$V_1 - V_2 = 2K \sin \frac{\Phi_B}{4} \cos \frac{\Phi_B}{2}$$

Using the relation between phase and frequency $B = 180\, f/f_0$ equation (23) becomes

(24) $$V_1 - V_2 = \sin 45\, f/f_0 \cos 90\, f/f_0$$

Table III shows computations of output vs normalized frequency to evaluate the theoretical possibilities of the discriminator. Consider the region from

(25) $$f/f_0 = \frac{52}{45}$$

TABLE III. Output Characteristic of Transmission Line Discriminator

(Computations By Means of 5 Place Logarithms - Accuracy to 4 Places)

$(45\ f/f_o)^o$	$(90\ f/f_o)^o$	sin $45\ f/f_o$	cos $90\ f/f_o$	sin $45\ f/f_o$ x cos $80\ f/f_o$	DIFFERENCES
40	80	.6428	.1737	.1116	.0416
42	84	.6691	.1046	.06999	.0457
44	88	.6947	.0349	.02425	.0494
46	92	.7193	-.0349	-.02510	.0526
48	96	.7431	-.1045	-.07765	.0556
50	100	.7660	-.1737	-.1331	.0575
52	104	.7880	-.2419	-.1906	.0594
54	108	.8090	-.3090	-.2500	
55	110	.8192	-.3420	-.2802	.0605
56	112	.8290	-.3746	-.3105	.0613
58	116	.8481	-.4384	-.3718 ctr	.0612
60	120	.8660	-.5000	-.4330	
61	122	.8746	-.5299	-.4635	.0608
62	124	.8830	-.5592	-.4938	.0596
64	128	.8988	-.6157	-.5534	.0579
66	132	.9136	-.6691	-.6113	.0556
68	136	.9272	-.7193	-.6669	.0529
70	140	.9397	-.7660	-.7198	

to

(26) $f/f_o = \frac{64}{45}$

The deviation from linearity at the center of the selected region is

(27) $$\frac{\frac{.5534 + .1906}{2} - .3718}{.5534 - .1906} \times 100 = 05\%$$

The deviation from linearity at the center of the lower half of the region is

(28) $$\frac{\frac{.3718 + .1906}{2} - .2802}{.5534 - .1906} \times 100 = + .275\%$$

While for the upper half the result is

(29) $$\frac{\frac{.5534 + .3718}{2} - .4635}{.5534 - .1906} \times 100 = - .248\%$$

This accuracy is obtained over a bandwidth of

$$\frac{\frac{64}{45} - \frac{52}{45}}{\frac{58}{45}} \times 100 = 20.7\%$$

Note that the region of extreme linearity is displaced from the balance point $f/f_o = 1$. Measurements have shown linear operation of this discriminator, which was confirmed by the above analytical development.

As shown in the above table, the optimum linear operational point occurs at an f/f_o value of 1.2688. Hence, the lengths of the various sections of the delay line must be cut so that for an f of [illegible] mc, the above condition is satisfied. For the above values an fo of 3[illegible].79 mc is required. When the carrier is deviated over the maximum frequency excursion of ± 500 kc, the corresponding values of f/f_o are 1.30[illegible] and 1.[illegible]. Hence, in utilizing the equations shown above, the static linearity value for the above peak frequency excursions is .3[illegible]%. As the peak frequency deviation is taken over a wider interval, the actual static linearity value will improve when computed on a percentage basis for the whole frequency interval.

From the above discussion, it follows that the delay line discriminator has good linear characteristics over a 20% bandwidth value. Since no tuned circuits are employed in its construction, output signal degradation due to frequency drift or misalignment will be minimized. This is an important point since one of the major disadvantages of a low slope tuned circuit discriminator is the increased signal degradation due to [illegible] over a high slope unit. The type of units required to instrument the discriminator all have broad band characteristics; hence, relatively large operational bandwidths are possible with this type of discriminator. Finally, at the [illegible] mc operating frequency this overall unit can be more easily instrumented when compared to a tuned circuit type.

The output S/N of an FM system will vary in a linear manner with respect to the input S/N when the system operates above its threshold value. Below this point, the output S/N decreases very rapidly as the input S/N decreases. The above characteristic is due to the basic properties of frequency modulation. It has been found that the magnitude of the threshold value $(S/N)_T$ is a function of the modulation index, M. For values of $M > 5$, this threshold point is quite abrupt and can easily be determined in a plot of output S/N versus input S/N. For lower values of M, the curve at the expected threshold point becomes more rounded; hence, it is more difficult to determine. This latter condition makes it difficult to obtain an analytical expression that relates (S/N_T) to the FM system parameters. However, an empirical expression that approximates the required value of $(S/N)_T$ has been attained. It is as follows

$$\left(\frac{S}{N}\right)_T = 6.3\left[\left(\frac{B_N}{2f_m} - 1\right)\right]^{1/2} \tag{30}$$

B_N is the noise bandwidth of the IF filter

f_m is the bandwidth of the post detection filter

$(S/N)_T$ is defined in the above noise bandwidth. For the IF filter, the approximate value of B_N will be very close to the specified 3 db bandwidth. Therefore, for a B_N of 1 mc and an f_m of 100 kc, the required value of $(S/N)_T$ is 12.6 or 11 db. In the above computation the plot in Figure 4 was employed to compute the maximum permissable value of f_m for an M of 2. For an $(S/N)_T$ of 11 db in a 1 mc bandwidth, the corresponding values of S and N at the output of the filter would be -29.4 dbm -40.4 dbm. For the above signal value, the limiting action will take place in the last two stages of the limiter.

Since digital data will be employed with the FM system, the values of certain operational parameters should be chosen so as to insure optimum operation. The parameters of interest are the IF bandwidth, bit rate and the peak frequency deviation Δf. For a given FM system, the maximum bit rate that can be employed would be a value equal to the IF bandwidth. In fixing the above two parameters, it follows that the magnitude of the S/N at the output of IF filter can be maximized by correctly choosing the value of Δf. This follows from the fact that the signal energy of the FM spectrum within given frequency limits is a function of Δf and the bit rate. Since the bit rate and the frequency limits have been chosen, a maximization of the signal energy can occur for a certain value of Δf. It has been found that in order to obtain a minimum probability of error, the optimum value of Δf, for the above conditions, is .715 times the bit rate.

1.8 Distortion Considerations

Output distortion components from the transmitter-receiver pair will in general be caused by the following factors:

(1) Frequency instability in the FM transmitter

(2) Phase and frequency variations due to power supply ripple and circuit capacitances

(3) Noise

(4) Imbalance in the FM detectors.

Carrier frequency instabilities and the power supply ripple will both add directly to the resulting residual frequency deviations in the detectors. These deviations, will, in turn, degrade the dynamic linearity characteristics of the above pair. Hence, the frequency stability requirements and the dynamic linearity are related. Variations in circuit capacitances will cause phase modulation of the carrier. The residual frequency deviations due to this effect can be relatively minimized by employing wide band circuits through the transmitter receiver pair. Since this design factor will be employed, the effects of this degradation will be relatively unimportant. The imbalance factor will be negligible for the conventional FM detector due to the fact that no-tuned circuits are employed in this unit. The balance point is merely a function of delay line length; hence, this passive factor has a stable operating condition. The imbalance factor for the phase lock detector will mainly be determined by the stability properties of the loop VCO. Since, in this case, active elements are involved, larger relative drifts in the operating characteristics of the unit will occur.

Thermal noise has a detrimental effect on the frequency stability of an FM system due to the fact that it phase modulates the carrier. Hence, it does, in turn, cause a frequency modulation. Due to the fact that noise directly causes phase modulation, the output noise spectrum from an FM detector will have a triangular shape. Therefore, it follows that the resulting residual frequency excursion will be a strict function of the frequency difference between the center frequency, F_o, and the frequency F_i of the noise spike causing the phase modulation. Mathematically, the residual frequency value due to thermal noise can be expressed as follows:

(31) $$\Delta f_n = \frac{n}{c}\left(F_o - F_i\right)$$

where c is the peak carrier voltage

n is the peak noise voltage

For a given frequency interval, $F_o - F_i$, the value of Δ fn will be determined by n/c. Since the maximum value of Δ fn is specified the above factor will essentially set a limit on the lowest operational n/c value that can be tolerated. It follows that the most detrimental effects on the value of fn will occur when the receiver is operating with the 1 mc input filter.

1.9 Specification Exceptions

(1) Specification 3. .2.3.1

The above specification puts stringent requirements on both the amplitude and phase response of the input filters. In order to comply with the requirements stated in parts (a) and (b) of the specification, a non-uniform phase response past the frequency [illegible] will result. Hence, in order to comply with the dynamic [illegible] specification 3.5.2.1.3, the modulation characteristics of the input [illegible] wave should have its significant sidebands within the above frequency limits. Also as pointed out in the report, an FM wave with the modulation characteristics of f = 50[illegible] and fm = 500 [illegible] can't be passed through the [illegible] without greatly exceeding the distortion spec. set on the transmitter-receiver pair.

(2) Specification 3.5.2.3.2

For this case, an exception is taken for the [illegible] db dynamic range requirement of the limiter. The actual dynamic range is determined by the maximum signal-to-noise ratio that can be obtained from the signal-to-noise summer and the minimum signal-to-noise ratio that is consistent with the desired value of $\mathfrak{S}_o$ and the maximum permissable value of [illegible]. As shown in the report, the [illegible] value of S/N for the S/N summer is [illegible]. Since $\mathfrak{S}_o$ is set at [illegible] radians/sec and the maximum value of [illegible] is [illegible].16 x 1[illegible] radians/sec [illegible] that the lowest permissable value of [illegible] = [illegible]7. The [illegible] to the limiter that corresponds to this suppression factor is -2[illegible]. Hence for the 1 mc input filter the lowest S/N from the [illegible] of interest is -36 db. The above value corresponds to a [illegible] signal [illegible] range of -40 dbm to -20 dbm. Lowering the signal value below the [illegible] magnitude would be unrealistic from the standpoint of the minimum input signal level that the phase detector can successfully handle. Hence the S/N dynamic range that will be employed for the FM receiver will vary from -36 db to

+30 db for a 76 db range.

2.1 Circuit Mechanization

The following sections include a brief discussion of the circuit mechanization considered. In many instances particular circuits developed on former projects provide a basis for further development if not the final solution. Most of the circuits referenced in succeeding sections require additional development and test data especially with regard to phase linearity.

2.1.1 Limiter

The gain, dynamic range, limit level and linear power capability of the FM Receiver is listed in figure 1 as 35DB, 33DB, ODBM and ODBM respectively. Two basic tunnel diode limiters have been developed by Westinghouse (figure 14 and 15) that provide a basis for comparison if not a finalized circuit.

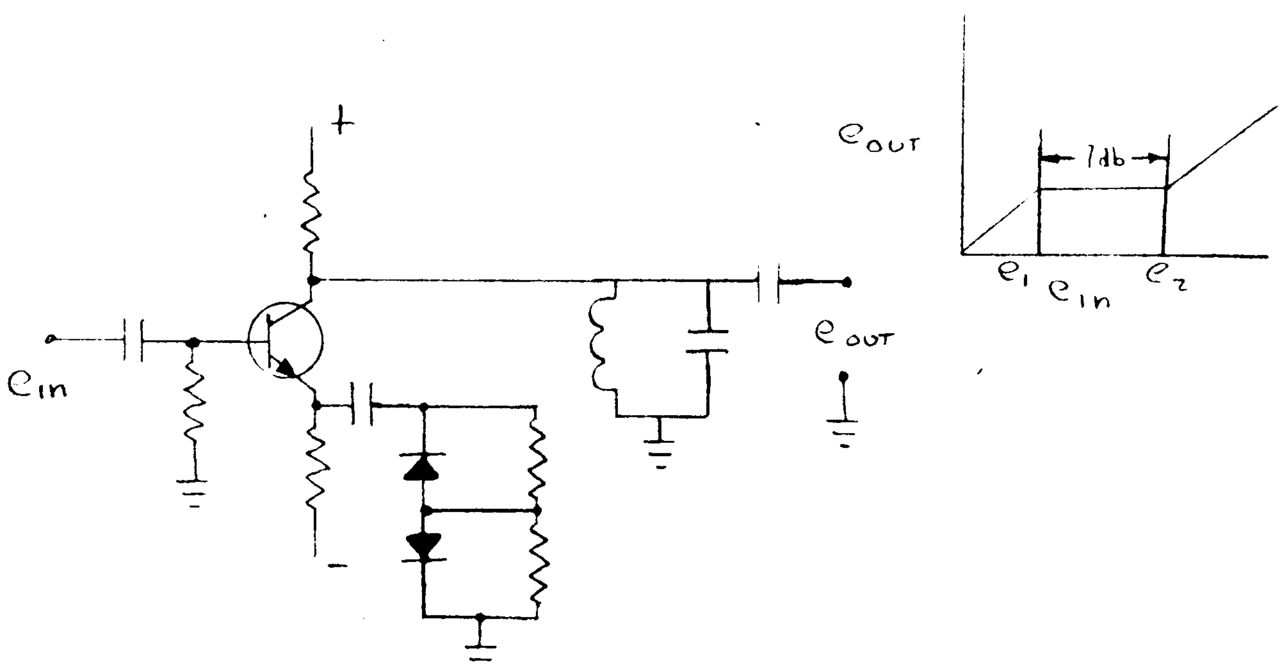


Figure 13. Tunnel Diode Limiter CRT 1

2.1.2 Buffer Amplifier

As shown in figure 1 the buffer amplifier must provide +20 DBM of gain between the last limiter stage and the phase detector. The buffer amplifier must exhibit a linear power capability to +30 DBM. Further, the power gain must be stable as a function of temperature and the bandwidth sufficient not to degrade the TX/RX pair frequency response. The broadband feedback pair illustrated in figure 14 exhibits all these characteristics except the linear power capability. The power capability will be achieved by a similar arrangement of power transistors.

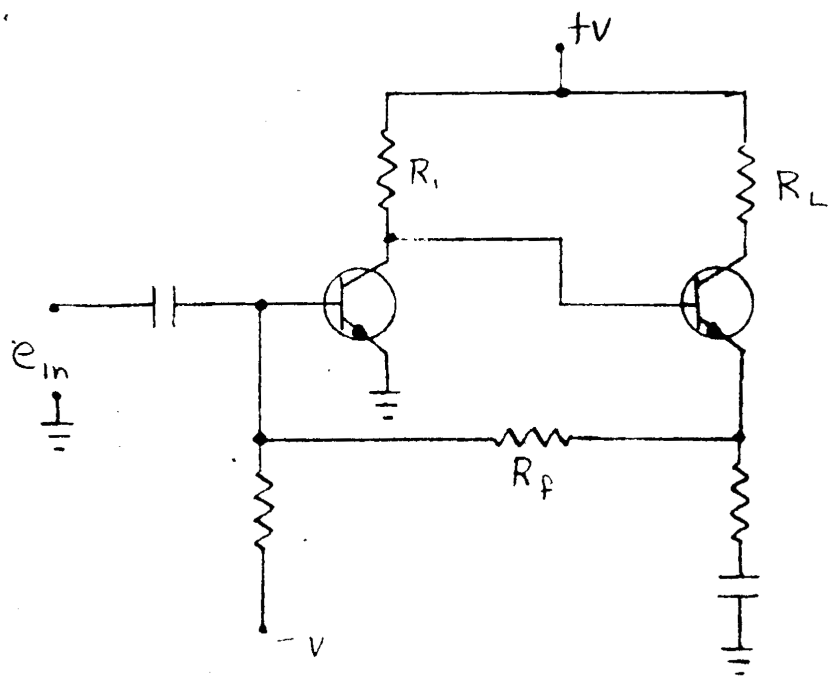


Figure 14. Feedback Pair

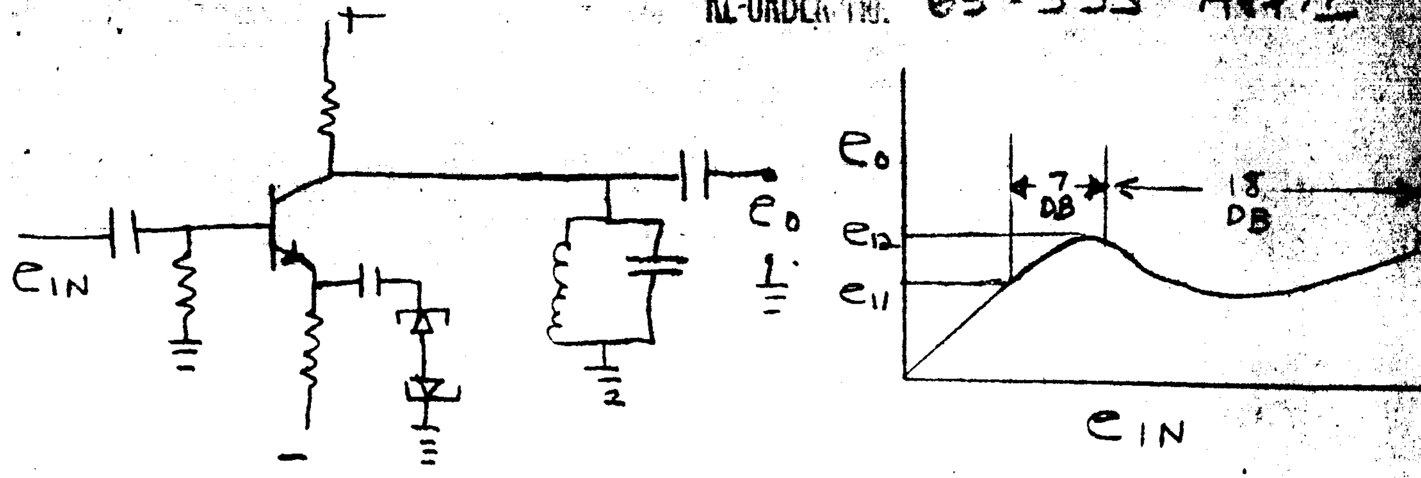


Figure 15. Tunnel Diode Limiter CKT 2

If the limiter illustrated in figure 14 is preceded by the limiter of figure 15 the overall characteristic of figure 16 results

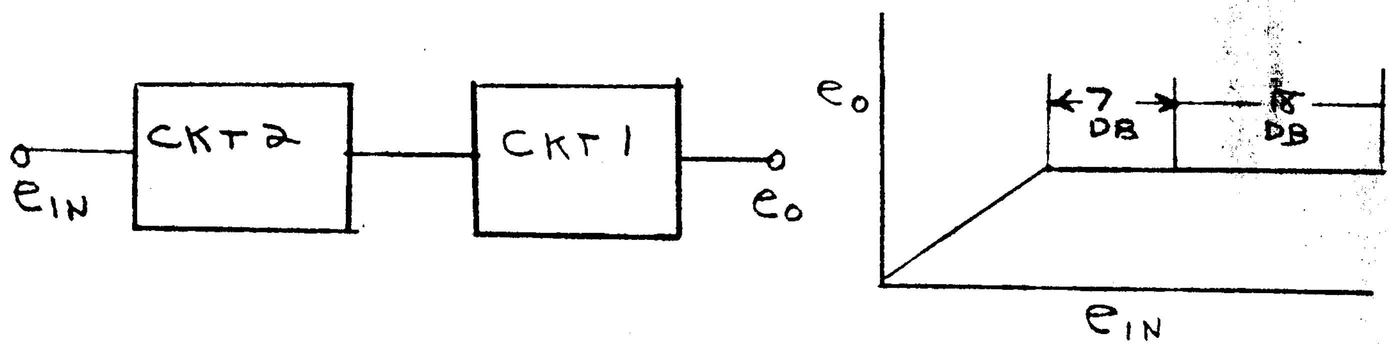


Figure 16. Cascaded Tunnel Diode Limiter

Cascading additional stages of CKT simply adds 18DB of dynamic range and 7DB of gain per stage. Therefore, a limiter chain consisting of four stages of CKT2 terminated by one stage of CKT_1 will yield 35DB of gain and 79DB of dynamic range. An eight stage tunnel diode wideband limiter was built using this technique. The characteristics are listed as follows:

1. Gain	60 DB
2. Dynamic Range	60 DB
3. Limit Level	200 mv (± 0.25% variations over 60DB input range)

4. Frequency Tested — up to 100 mc
5. Bandwidth — 30 mc
6. Power Dissipation — 1/4 watt

A comparable limiter patterned after the work of Ruthroff (1) was fabricated. A typical amplifier limiter stage is shown in figure 17. Four stages were cascaded and the following characteristics achieved.

1. Gain	60 DB
2. Input Dynamic Range	60 DB Min
3. Limit Level	200 mv (negligible Variation with 60DB input change
4. Frequency Tested	Up to 100 MC
5. Bandwidth	20 MC
6. Power Dissipation	3 watts

The comparative phase shift of both types of limiter was measured at 30 MC over 50 DB change of input level. The tunnel diode limiter exhibited 17 degrees of phase shift. Ruthroff's limiter shifted the phase of a 30 mc unmodulated carrier 35 degrees over the same dynamic range of input. A comparable test for shift in limit level as a function of temperature variation was conducted. The tunnel diode limiter level shifted 5.4% from a 200 mv limit level over a temperature range of $-30^{\circ}C$ to $+70^{\circ}C$. The conventional limiter limit level shifted 20% from 200 mv over the same temperature level. The tunnel diode limiter is less temperature sensitive and yields less phase shift as a function of drive level; however, the limit level variation as a function of drive level is inferior to the conventional limiter. The tunnel diode limiter approach will be applied to the FM Receiver.

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1. "Amplitude Modulation Suppression in FM Systems" by C T Ruthroff, The Bell System Technical Journal, July 1958.

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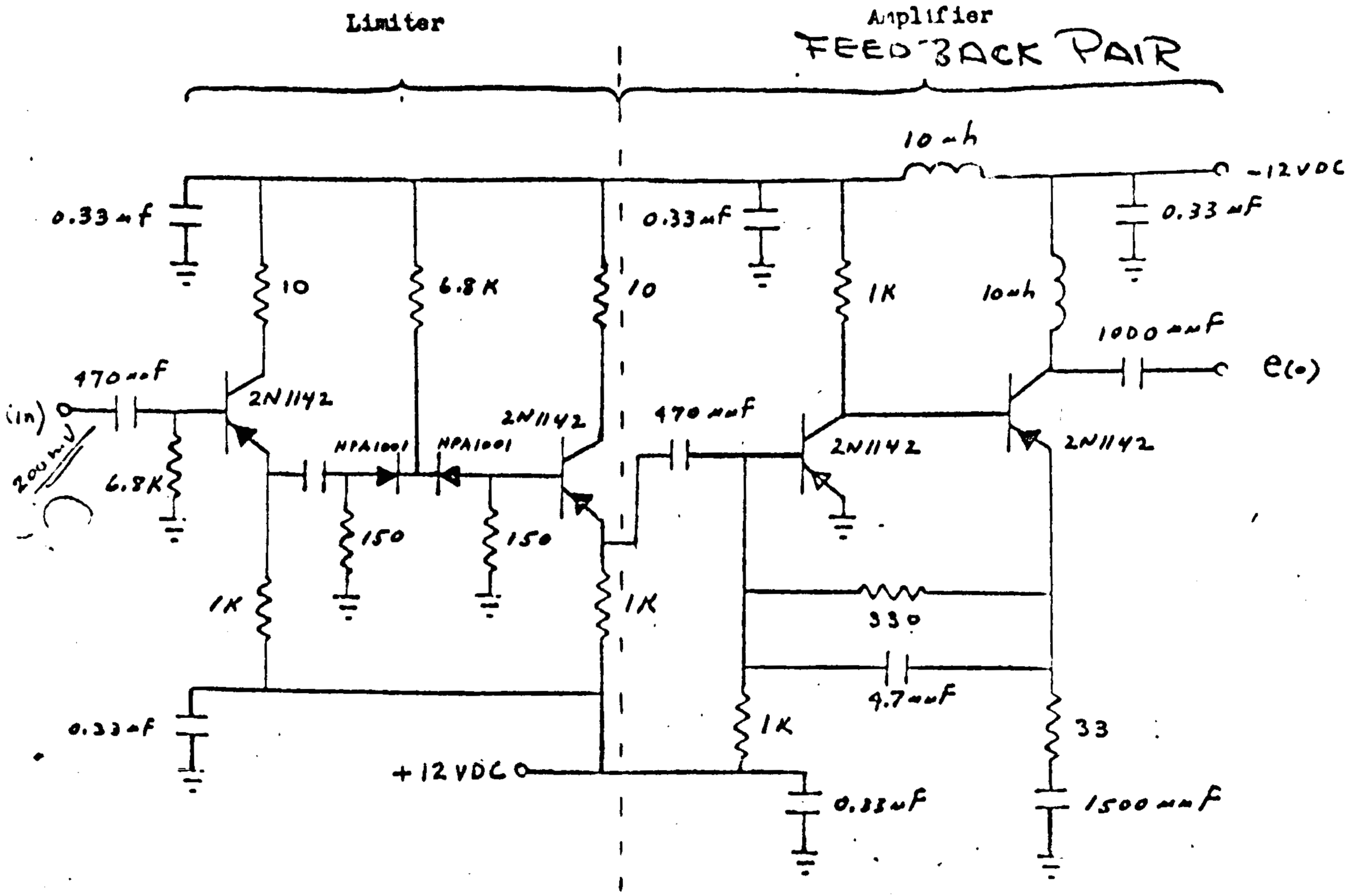

FIGURE 17 (Basic conventional limiter-amplifier circuit)

65-23 AVI

2.1.3 Phase Detector

The phase detector constant (Km $\frac{\text{radians}}{\text{volt}}$) was referenced earlier in this report as 2 volts/radian. This value is based on the characteristic of the phase detector illustrated in figure 18, whereby the scale factor of 2 volts/radian corresponds to a signal drive level of +10DBM. The referenced phase detector was fabricated and tested during the Phase I program and the characteristic is referenced here as a basis for the Phase Lock Detector design goal. The video response is indicated in figure 19.

2.1.4 Loop Amplifier

The loop DC amplifier gain (K_{ω}) was listed in section 3.5 as $50.3\frac{\text{volts}}{\text{volt}}$. Ideally, the loop amplifier gain is independent of frequency such that the loop transfer function is established primarily by the loop filter. However, as shown by Table I section 3.5 the maximum information bandwidth is listed as $3\text{x}10^{5}$cps. The loop amplifiers closed loop 3DB corner should be at least ten times greater than the largest loop information bandwidth to prevent excessive influence on the loop transfer function. For example, computer investigation of the 3 pole Butterworth closed loop response included in the PM Transmitter study indicated that an additional pole five times higher in frequency than the loop cut off frequency yielded 0.74DB peaking. An additional pole ten times the cut off frequency resulted in 0.34 DB peaking. Several commercially available operational amplifiers were considered. The Philbrick SP -456 (the chopper stabilized version of the Philbrick P-45) exhibits a 3DB closed loop bandwidth of 2.5 mc at a closed loop gain of 8. However, the overall 3DB bandwidth is extended as outlined below:

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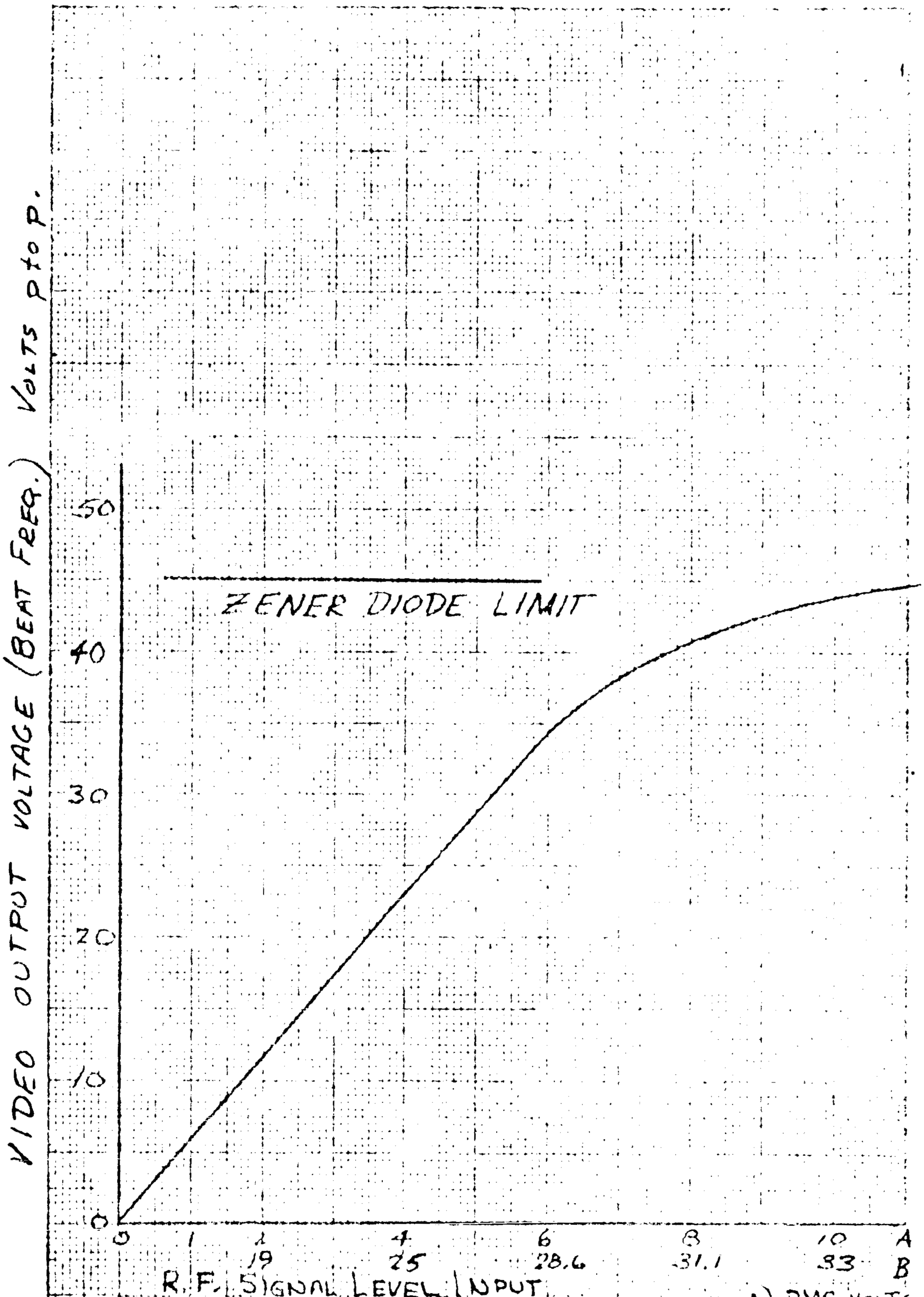

FIGURE 18 LINEARITY OF PHASE DETECTOR VOLTAGE TRANSFER

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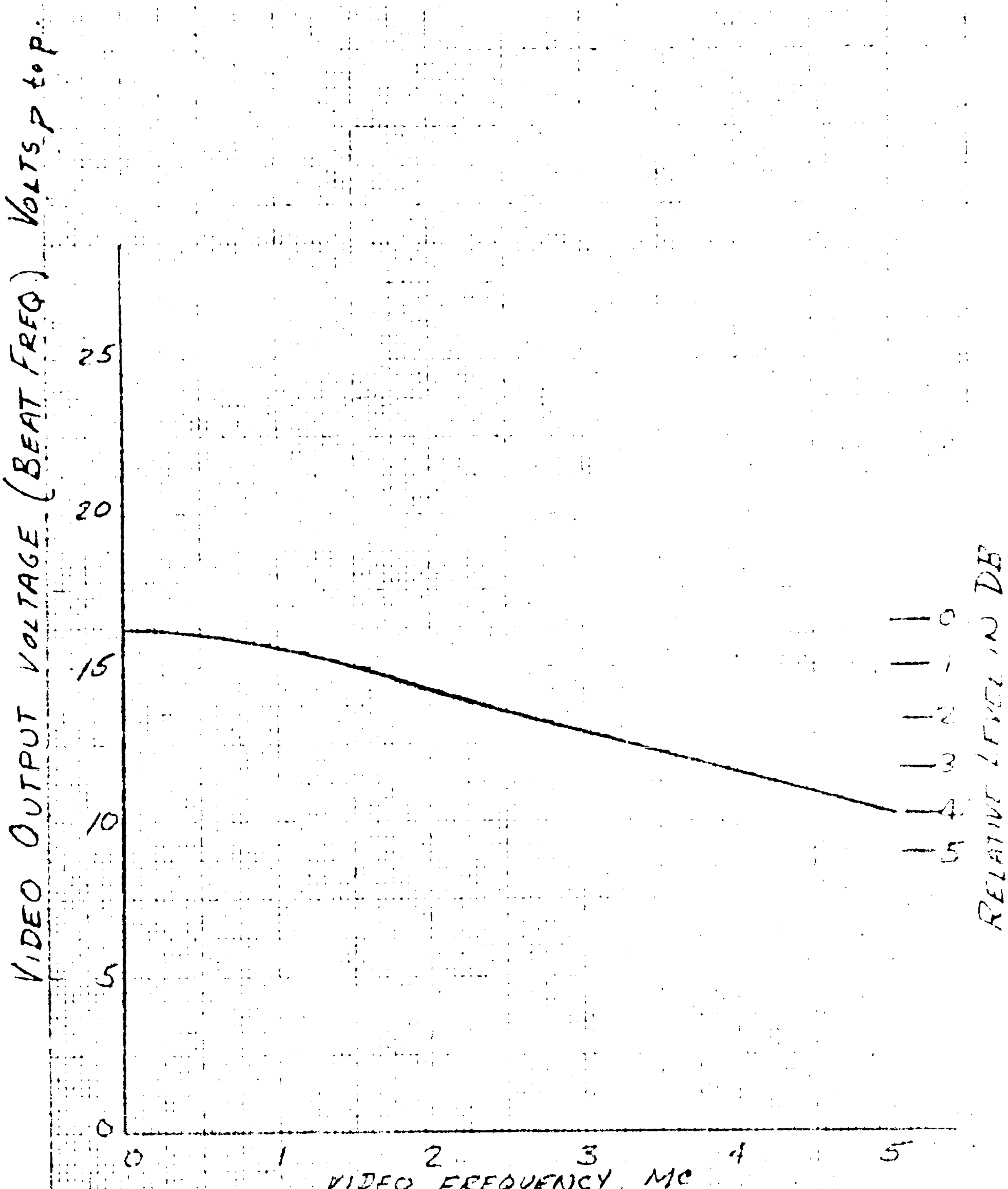

FIGURE 19 PHASE DETECTOR VIDEO BANDWIDTH CHARACTERISTICS

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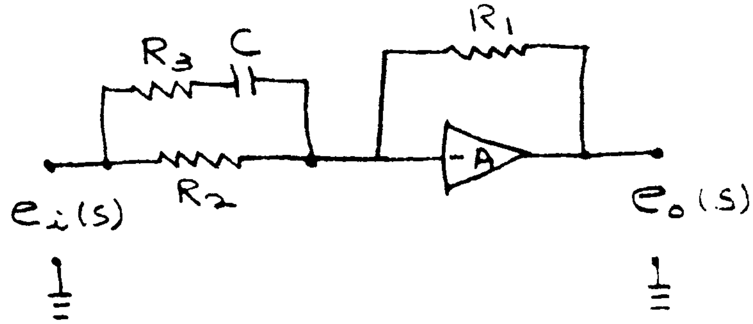

Figure 20. Compensated Operational Amplifier

$$\frac{e_o(s)}{e_i(s)} = \frac{R_1}{R_2}\left[\frac{1 + s(R_2+R_3)C}{1 + sR_3C}\right] = \frac{R_1}{R_2} f_1(s) \qquad (34)$$

$$f_1(s) = \frac{s/\omega_1 + 1}{s/\omega_2 + 1} \qquad \omega_1 = \frac{1}{(R_2+R_3)C} = 2\pi \times 2.5 \times 10^6 \frac{rad}{sec} \qquad (35)$$

$$\omega_2 = \frac{1}{R_3C} = 2\pi \times 15 \times 10^6 \frac{rad}{sec}$$

Figure 21 indicates the compensated amplifier response.

2.1.5 VCO

The VCO contribution to the loop transfer function is $\frac{K_{VCO}}{s(s+\omega_1)}$. The corner W_1 is related to the VCO center frequency and Q as indicated by equations 36 and 37.

$$BW = f_0/Q \qquad (36)$$

$$\omega_1 = \frac{BW}{2} \qquad (37)$$

There are two choices available for the application of W_1 to the overall loop transfer function. The corner W_1 can be made one of the principal contributions to the loop transfer function or W_1 can be made very large with respect to the largest loop information bandwidth such that its influence (peaking)

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K+E SEMI-LOGARITHMIC 359-81 KEUFFEL & ESSER CO. MADE IN U.S.A. 4 CYCLES X 70 DIVISIONS

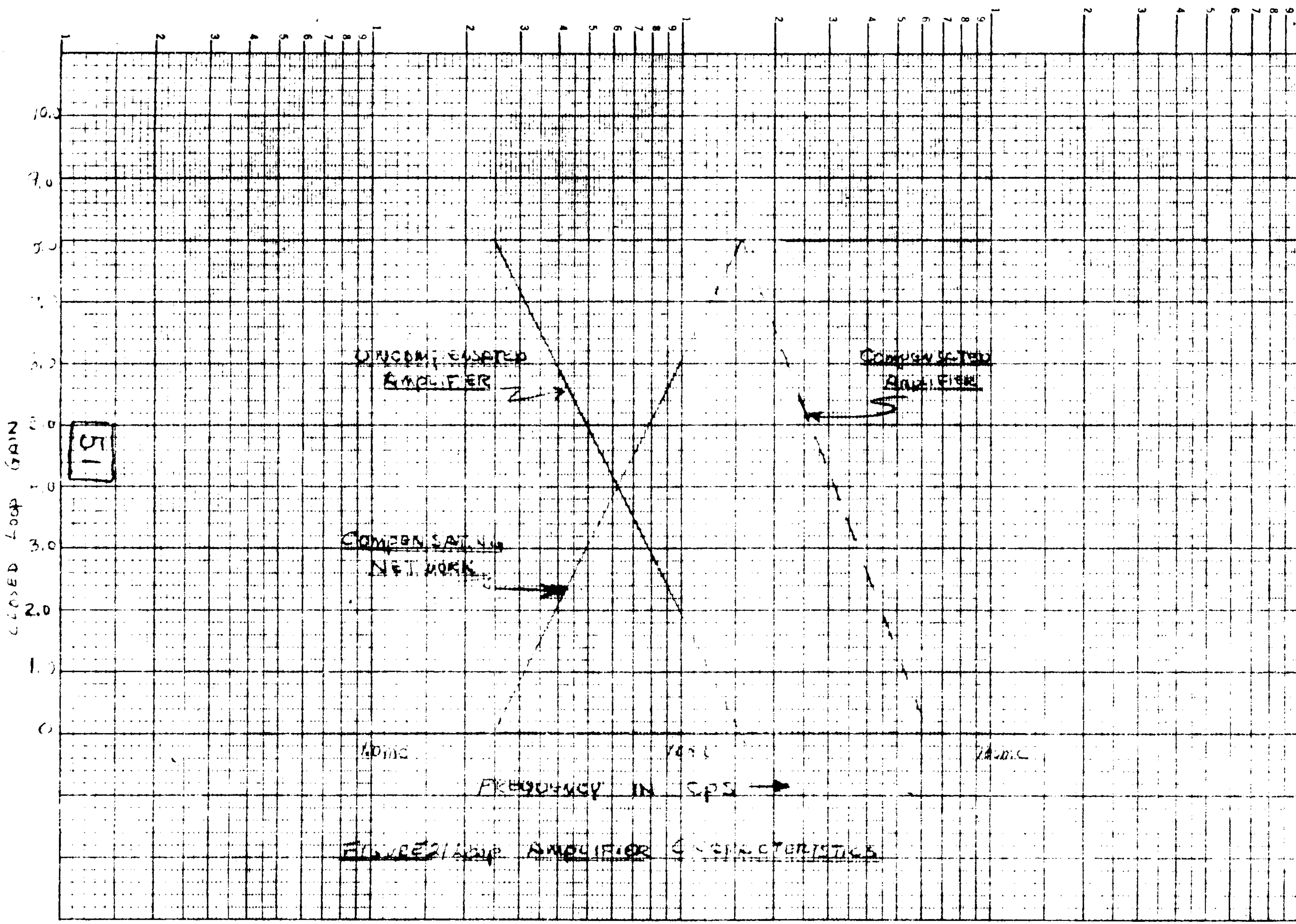
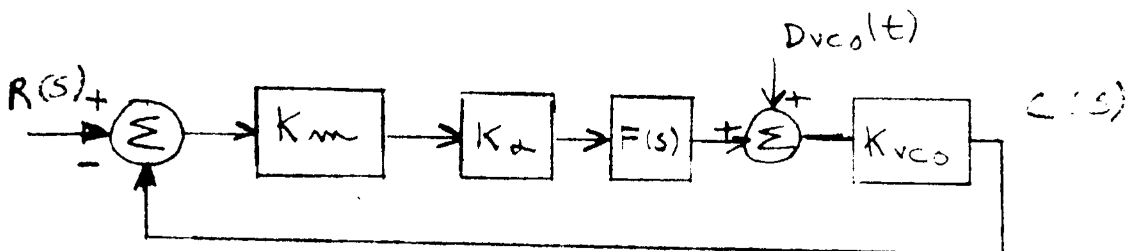


FIGURE 21 LOOP AMPLIFIER CHARACTERISTICS

is ignored. Since several loop information bandwidths are specified, the former choice is not attractive. The latter choice (a low Q VCO) has the disadvantage that the inherent short term stability of the oscillator is relatively poor and is dependent on the loop response time. At low values of loop information bandwidth the loop is restricted in the amount of oscillator short term stability it can correct. However, the latter choice will be used whereby a loaded Q of 5 is the design goal.

The non-linearity of the oscillator is connected to some extent by the gain of the phase detector and loop amplifier. Consider the following.



$$C(S) \doteq \frac{K_{vco} D_{vco}(t)}{1 + K_m K_d K_{vco} F(S)} = \frac{1}{1/K_{vco} + K_m K_d F(S)} \qquad (38)$$

If $D_{vco}(t)$ is considered a second loop input equivalent to the distortion contributed by the VCO it is obvious that the VCO distortion is attenuated by the product of the gain of the loop amplifier and phase detector. Therefore, it behooves the designer to assign as much loop gain as practical to the loop amplifier and minimum gain to the VCO within the constraints of the loop amplifier dynamic range, drift and linearity.

The 50 mc VCO will be based on the Clapp oscillator design. The oscillator design equations are included are shown in equations 40 thru 45 of the PM Receiver Study. A typical schematic diagram and transfer of E_o/E_{IN} are shown in figure 22 and 23 for a 70 mc oscillator patterned after the Clapp design. Although not shown an auxiliary AGC loop that regulates the oscillator power

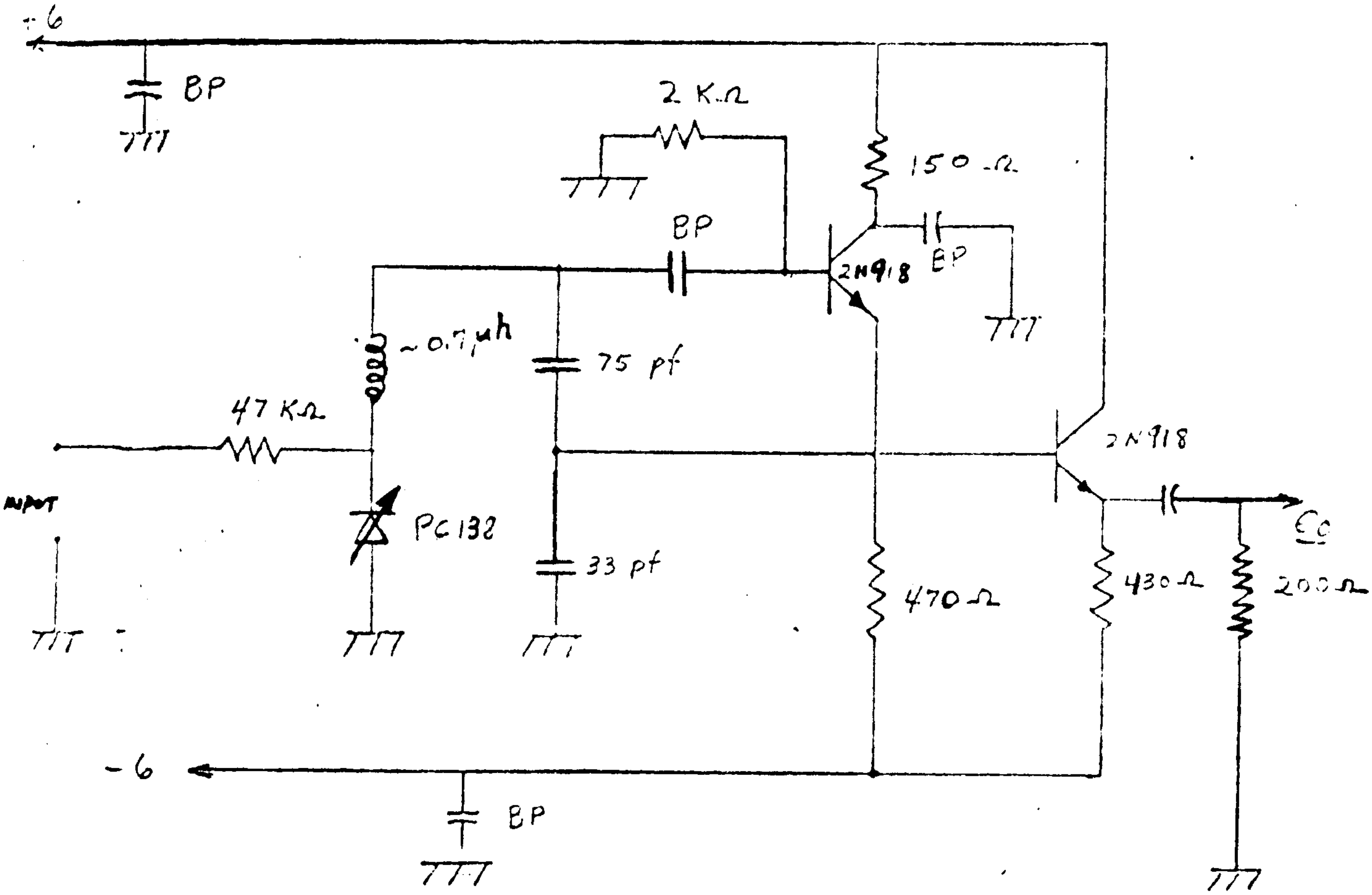


FIGURE 22 70 Mc/s VCO CIRCUIT DIAGRAM

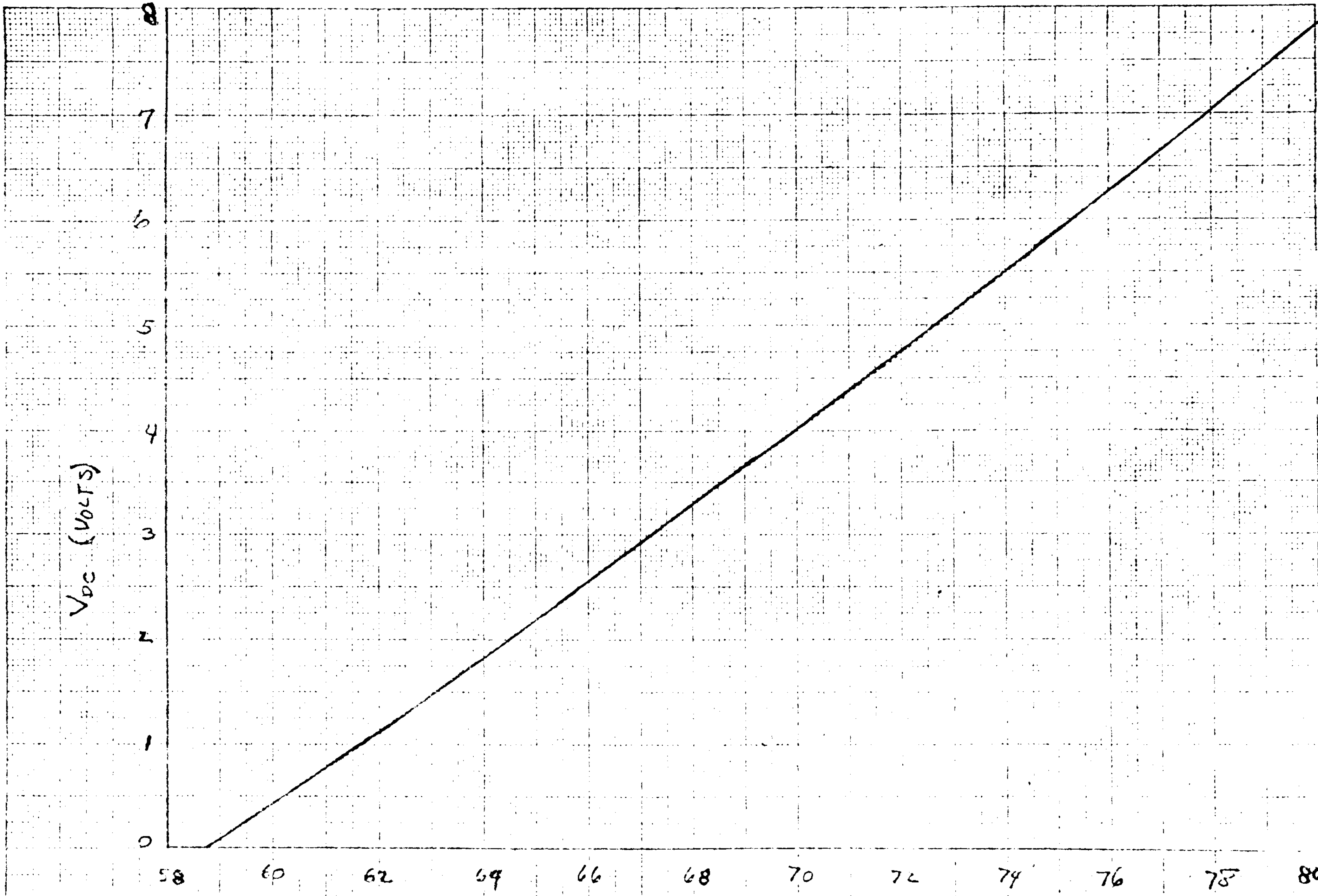


FIGURE 23
70 MCS VCO TRANSFER

and maintains class A operation as a function of frequency changes will be included.

2.1.6 Loop AC Amplifier

The phase detector reference power level was listed as 30DBM (10DB above RMS noise input) in section 3.5 of this report. There are two choices of supplying the reference power level. The oscillator can be built to deliver 30DBM directly (power oscillator) or a wide band medium power feedback pair with approximately 15DB of gain between a low power VCO and phase detector constitutes the second choice. The second choice suffers the disadvantage of contributing an additional loop corner and additional transport lag. However, the VCO linearity and Q requirements referenced earlier are enough burden for this unit aside from the relatively high power level. The medium power feedback pair is essentially the same circuit referenced in figure 13 except the output transistor must be capable of delivering 30 DBM.

2.1.7 Balanced Mixer

The FM Receiver requires two balanced mixers. One mixer down converts the 50 MC frequency modulated carrier to 5 MC by multiplying the modulated carrier with a 55MC reference. The second mixer reverses the process or up converts the FM spectrum from 5 MC to 50 MC. Only one stage of conversion is required as no overlap of the spectrum occurs as in the case of the PM Receiver. Each mixer is followed by a band pass filter. The filter phase response shall be linear (Bessel). The conversion processes are straight forward except in the case of the up converter the 55 MC reference is at the edge of the modulation spectrum centered on 50 MC. The Bessel filter will provide very little attenuation to the 55 MC reference.

The mixer balance to the reference will be the principal source of reference rejection. The balanced mixer is discussed in section 4.9 of the FM Receiver Study.

2.1.8 Output Amplifiers

The Video Buffer Amplifiers shown in figure 1 will be Philbrick SP-456 operational amplifiers.

2.1.9 Discriminator

The Discriminator characteristics and mechanization was discussed in section 3.7 of this study.

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