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APPENDIX H

F.M. TRANSMITTER

Submitted as part of the Final Report

for RF Test Console on JPL

Contract No. 950144

NAS 7-100

DATE:

12 February, 1965

WESTINGHOUSE DEFENSE AND SPACE CENTER

SURFACE DIVISION

ADVANCED DEVELOPMENT ENGINEERING

MDE 4856 55

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I. F.M. Transmitter Requirements

This report is in response to Contract No. 950144 Statement of Work paragraph 8. Design parameters are established for linear frequency modulation of a locked oscillator in accordance with the simplified diagram of figure 1.

A. Conclusion

The F.M. Transmitter specifications are state-of-the-art. The designer is confronted with the problem of attempting to solve the problem with conventional techniques, whereby the required hardware is by necessity at or beyond the state-of-the-art. The alternative is to select a system that offers both basic advantages and unfortunately basic risks. The locked oscillator Frequency Modulator represents such a choice. The degree of risk is minimized by computation which prove that the system is feasible. In the final analysis, the system performance is dictated by the correlation between hardware performance and computation.

The system outlined in this report is feasible provided that 1) A VCO can be fabricated that exhibits the specified residual F.M. and linearity 2) a phase detector with a linear transfer of output voltage to input phase is used 3) a count down divider is used in the feedback channel to constrain the phase deviation within the linear limits of the phase detector. 4) The modulation index and the division of the feedback counter are matched such that the linear range of the phase detector is not exceeded 5) a realizable compensation network whose transfer function is the inverse of the transfer of frequency output to pre distort the baseband modulation.

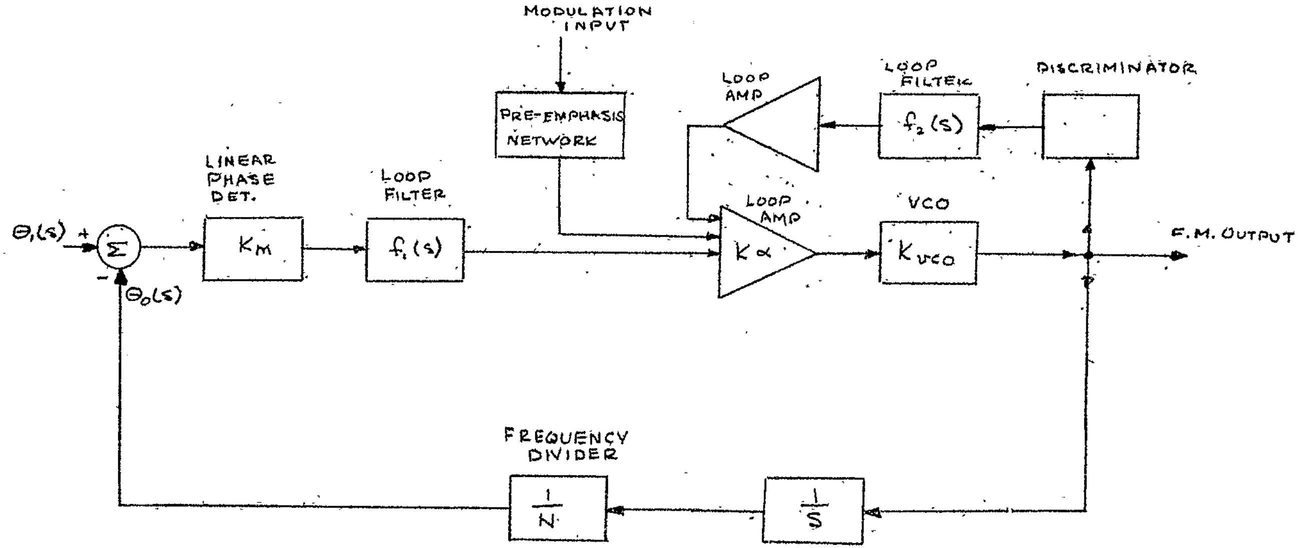


FIGURE 1. PHASE LOCK, AFC FM MODULATOR

B. Summary of Specifications

Reference JPL Spec. No. GPG-15062-DSW

1. FM Operational Modes. The RF test console shall contain a wideband FM transmitter capable of operating in either AFC or non-AFC modes, and both conventional and phase-lock FM demodulations.

2. Frequency Stability. The frequency stability of the FM transmitter-receiver pair shall be such as to cause less than 15 cps rms residual FM with the transmitter operating in the AFC mode, and 60 cps rms residual FM with the transmitter in the non-AFC mode, measured at the output of either the conventional or phase lock FM receiver in a 500 kc bandwidth.

3. Static Linearity. The transmitter-receiver pair shall exhibit a static linearity of ± 0.5 percent over the full-scale frequency deviation with the non-AFC transmitter and either receiver.

4. Dynamic Linearity. The transmitter-receiver pair shall exhibit a dynamic linearity of ± 1.0 percent over all combinations of modulating frequency and frequency deviation in both AFC and non-AFC transmitter modes and with either receiver.

5. Frequency. The transmitter center frequency shall be exactly 50 Mc and shall be continuously tunable ± 500 cps about this frequency by manual control.

6. Power Output. The transmitter power output shall be sufficient to drive the signal/noise mixer.

65-33G
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7. Frequency Response. The frequency response of the frequency modulator shall be constant within ± 0.1 db from 50 cps to 100 kc and ± 0.5 db from 3 cps to 50 cps and 100 kc to 500 kc.

8. Frequency Deviation. The modulator shall be capable of deviating the carrier ± 500 kc about its center frequency with a maximum modulation index of 512 in the AFC mode.

9. Deviation Linearity. The characteristics of the frequency modulator shall be sufficiently linear to meet the static and dynamic linearity requirements of par. 3.5.2.1.2 and 3.5.2.1.3, JPL Spec. 15062.

10. Incidental AM. Incidental amplitude modulation due to frequency modulation shall cause no more than ± 0.1 db envelope distortion under the most severe conditions.

11. AFC Operation. The FM transmitter shall be capable of operating either with or without automatic frequency control. In the AFC mode its modulation responses shall be from 3 cps to 500 kc, and in the non AFC mode, from dc to 500 kc.

12. Amplitude Modulator. The FM Transmitter shall contain a narrow-band amplitude modulator with characteristics identical to that in the PM transmitter as specified in 3.5.1.2.5. JPL Spec. 15062.

13. Bandwidth. The overall FM Transmitter bandwidth shall be consistent with the above performance specifications.

14. Transmitter Reference Output. The FM transmitter reference output shall be identical to that for the PM transmitter.

C. Design Objectives

1. Peak Frequency Deviation

The peak frequency deviation of the VCO and driving amplifier must be 0.5 mc.

2. Fidelity Allocations

The static and dynamic linearity allocations are given for the TX/RX pair as ± 0.5 percent and ± 1.0 percent respectively. The transmitter and receiver are designed separately. The allotted error will be equally divided between the TX and RX.

3. Phase Lock FM Transmitter Loop Characteristics

a. APC Loop Transfer Function

The transfer function of frequency output to modulation input will be of the form:

$$\frac{f_o}{V_m}(s) = \frac{N}{K_m}\left[\frac{s\left(\frac{s}{\omega_1}+1\right)}{\frac{s^2}{\omega_n^2}+\frac{2\varepsilon}{\omega_n}s+1}\right]$$

The baseband pre-distortion will be of the form:

$$f(s) = \frac{K_m}{N}\left[\frac{\frac{s^2}{\omega_n^2}+\frac{2\varepsilon}{\omega_n}s+1}{s^2}\right]$$

whereby the combined frequency response of the pre-distortion network and locked oscillator will be ± 0.1 db from 50 cps to 100 kc and ± 0.5 db from 3 cps to 50 cps and 100 kc to 500 kc.

b. __APC Transport Lag__

Not applicable to narrow band loop

c. __APC Loop Multiplier Reference__

$50.10^{+6}/512$ 97.5 kc (Maximum Reference Frequency)

d. APC Loop Rejection at 97.5 kc with respect to modulation.
−60 DB

e. __VCO__

Peak frequency deviation ± 0.5 mcs.

| | |
|---|---|
| Sensitivity | $2.10^{+6}$ $\frac{\text{cycles}}{\text{volts sec.}}$ |
| Linearity | 0.5% |
| Output carrier frequency | 50 Mcs |
| Q | 200 |

f. __Phase Detector__

Linearized (Sawtooth) characteristic Phase error Range ± $\pi$ Radians

| | |
|---|---|
| Linearity | 2% |
| Sensitivity | 1 volt/radian |

g. __D.C. Amplifier__

| | |
|---|---|
| Gain | 20 db |
| Linearity | 0.1% |
| Voltage Output Range | ± 10 volts |
| Equivalent Corner Frequency | 10MCS, |
| Drift | 100 $\mu$V/wk. |

h. APC Loop Filter Transfer Function

$$f(s) = \frac{\tau_2 s + 1}{\tau_1 s + 1}$$

i. Feedback Frequency Divider Ratio

N = 9 (Minimum)

j. AFC Loop Filter Transfer Function

$$F(s) = \frac{1}{\left(\frac{s}{\omega_1}+1\right)\left(\frac{s}{\omega_{n_0}}+1\right)^2}$$

k. AFC Open Loop Transfer Function

$$G(s) = \frac{K_P}{\left(\frac{s}{\omega_1}+1\right)\left(\frac{s}{\omega_{n_0}}+1\right)^2}$$

l. AFC Closed Loop Transfer Function

$$F(s) = \frac{1}{\left(\frac{1}{K_P}+1\right)\left(\frac{s}{\omega_n}+1\right)\left(\frac{s^2}{\omega_n^2}+\frac{2\varepsilon}{\omega_n}s+1\right)}$$

m. AFC Open Loop Gain

$$K_P = K_m \frac{\text{VOLTS}}{\text{MC}} \cdot K_\alpha \frac{\text{VOLTS}}{\text{VOLT}} \cdot K_{VCO} \frac{\text{MC}}{\text{VOLT}}$$

$$K_P = 0.25 \frac{\text{VOLTS}}{\text{MC}} \cdot 100 \frac{\text{VOLTS}}{\text{VOLT}} \cdot 2 \frac{\text{MC}}{\text{VOLT}} = 50$$

Note: The AFC amplifier gain $\frac{100 \text{ volts}}{\text{volt}}$ is attributed to two loop amplifiers. One amplifier is common to both the AFC and APC loop.

n. Discriminator

| | |
|---|---|
| Linearity | 0.3% |
| Gain | 0.25 volts/mcs |

## II. AFC System

The TX/RX pair specification states that the RMS residual F.M. be no larger than 15 cps measured in a 500 kc receiver bandwidth. The residual FM is related to the baseband noise measured at the receiver output as the ratio of $\frac{15\ \text{cps}}{500\ \text{kc}}$ or approximately -90 db. Transmitter VCO instability is one contributor to the system residual F.M. Specifically the drift rate of the transmitter VCO contributes to the noise spectrum measured at the RX output. The AFC system in the usual sense is intended to improve the stability characteristics of the deviable oscillator. Consider the system outlined in figure 2.

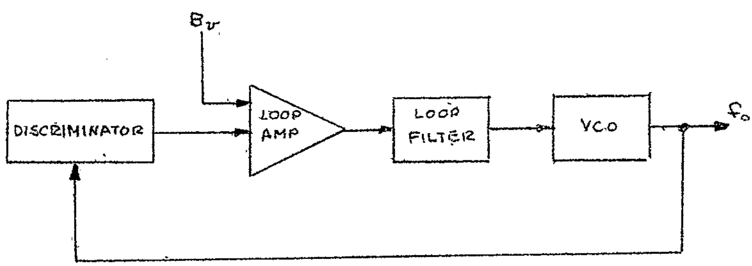



FIGURE 2. BASIC AFC SYSTEM

A simplified frequency model of this system is indicated in figure 3.

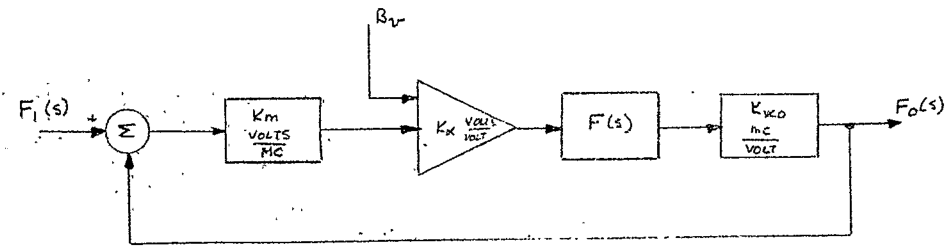



FIGURE 3. SIMPLIFIED MODEL, AFC LOOP

The transfer function $\frac{F_o}{B_v}$ is listed in equation 1

$$\frac{F_o}{B_v}(s) = \frac{K_\alpha K_{vco} F(s)}{1 + K_m K_\alpha K_{vco} F(s)} \qquad (1)$$

Let $K_p$ (position constant) = $K_m K_\alpha K_{vco}$

$$\frac{F_o}{B_v}(s) = \frac{1}{K_m}\left[\frac{1}{1 + \frac{1}{K_p F(s)}}\right] \qquad (2)$$

The TX/RX amplitude response is specified as ± 0.5 db at 500 kc. Assume that 0.1 db droop is allotted the FM Transmitter. Therefore, in equation (2) if $j\omega$ is substituted for s, the transfer $\left|\frac{E_o}{\beta_v}(j\omega)\right|$ must not droop more than 0.1 db as jw becomes $10^{+6}\pi$ $\frac{\text{rad}}{\text{sec}}$.

A. AFC Loop Synthesis

The closed loop Butterworth response synthesis developed during Phase I may be of value. The synthesis affords a convenient technique for selecting the open loop poles and gain such that the closed loop poles fall on a semi circle (cut off frequency) in the s plane. The open loop poles and gain are determined as a function of closed loop cut off frequency and damping factor. A three pole Butterworth closed loop response is assumed. The closed loop transfer function is of the form indicated by equation 3.

$$F(s) = \frac{1}{\left(\frac{s}{\omega_n}+1\right)\left(\frac{s^2}{\omega_n^2} + \frac{2\mathcal{E}}{\omega_n}s+1\right)} \tag{3}$$

The open loop transfer function is listed by equation 2. Let $F(s)$ of equation 2 become

$$F(s) = \frac{1}{s\left(\frac{s^2}{\omega_{n_o}^2} + \frac{2\mathcal{E}_o}{\omega_{n_o}}s+1\right)} \tag{4}$$

Let $K_m = 1$

$$\frac{F_o}{\beta_v}(s) = \frac{1}{\frac{s}{K_p}\left(\frac{s^2}{\omega_{n_o}^2} + \frac{2\mathcal{E}_o}{\omega_{n_o}}s+1\right)+1} \tag{5}$$

If equations (3) and (5) are expanded and equated, equation 6 results

$$\frac{s^3}{\omega_n^3} + \frac{1}{\omega_n^2}(2\xi+1)s^2 + \frac{1}{\omega_n}(2\xi+1)s + 1 = \frac{s^3}{K_p\omega_{n_o}^2} + \frac{2\xi_o}{K_p\omega_{n_o}}s^2 + \frac{s}{K_p} + 1 \qquad (6)$$

Equating the coefficients of like powers of s, the following is evident

$$K_p\,\omega_{n_o}^2 = \omega_n^3 \qquad (7)$$

$$\frac{K_p\,\omega_{n_o}}{2\xi_o} = \frac{\omega_n^2}{2\xi+1} \qquad (8)$$

$$K_p = \frac{\omega_n}{2\xi+1} \qquad (9)$$

The open loop parameters of interest are listed as follows:

$$K_p = \frac{\omega_n}{2\xi+1} \qquad (10)$$

$$\omega_{n_o} = \omega_n\sqrt{2\xi+1} \qquad (11)$$

$$\xi_o = \frac{\sqrt{2\xi+1}}{2} \qquad (12)$$

The closed loop resonant frequency (or closed loop filter cut off frequency) is specified by the allotted amplitude droop at 500 kc. A three pole Butterworth exhibits 0.1 db amplitude droop at 500 kc if fn is 1 mcs ( $\omega_n = 2\pi \cdot 10^{+6}$). Further, the closed loop damping, $\mathcal{E}$ of a three pole Butterworth is defined as 0.5. Therefore,

$$K_p = \frac{\omega_n}{2\mathcal{E}+1} = \pi \cdot 10^{+6} \tag{13}$$

$$\omega_{n_o} = \omega_n \sqrt{2\mathcal{E}+1} = \sqrt{2}\, 2\pi \cdot 10^{6} \ \frac{\text{rad}}{\text{sec}} \tag{14}$$

$$\mathcal{E}_o = \frac{\sqrt{2\mathcal{E}+1}}{2} = 0.707 \tag{15}$$

As indicated by equation (13) the open loop gain must be $\pi \cdot 10^{+6}$ Unfortunately, this is unrealizable. Therefore, a realistic synthesis must list a realistic loop gain as a constant and allow the other parameters to become dependent of the selected closed loop cut off frequency and loop gain.

A practical open loop gain is listed by equation 16.

$$\begin{aligned} K_p &= K_m \frac{\text{volt}}{\text{mc}} \cdot K_\alpha \frac{\text{volt}}{\text{volt}} \cdot K_{vco} \frac{\text{mc}}{\text{volt}} \\ &= 0.25 \cdot 100 \cdot 2 \\ K_p &= 50 \end{aligned} \tag{16}$$

The discriminator constant is based upon a linear delay line discriminator used previously. The open loop transfer function is of the form listed in equation 17.

$$G(s) = \frac{K_p}{\left(\frac{s}{\omega_1}+1\right)\left(\frac{s}{\omega_{n_o}}+1\right)^2} = \frac{1}{\frac{1}{K_p}\left(\frac{s}{\omega_1}+1\right)\left(\frac{s^2}{\omega_{n_o}^2}+\frac{2}{\omega_{n_o}}s+1\right)} \tag{17}$$

The loop filter $F(s)$ is comprised of real pole at $\omega_1$ and a double pole at $\omega_{n_o}$. The closed loop form is:

$$F(s) = \frac{G(s)}{1+G(s)} = \frac{1}{\frac{1}{K_p}\left(\frac{s}{\omega_1}+1\right)\left(\frac{s^2}{\omega_{n_o}^2}+\frac{2}{\omega_{n_o}}s+1\right)+1} \tag{18}$$

$$F(s) = \frac{1}{\frac{1}{K_p\omega_1\omega_{n_o}^2}s^3 + \frac{1}{K_p}\left(\frac{2}{\omega_1\omega_{n_o}}+\frac{1}{\omega_{n_o}^2}\right)s^2 + \frac{1}{K_p}\left(\frac{1}{\omega_1}+\frac{2}{\omega_{n_o}}\right)s + \frac{1}{K_p}+1} \tag{19}$$

The closed loop response required (three pole Butterworth) is of the form:

$$F(s) = \frac{1}{\left(\frac{1}{K_p}+1\right)\left(\frac{s}{\omega_n}+1\right)\left(\frac{s^2}{\omega_n^2}+\frac{2\epsilon}{\omega_n}s+1\right)} \tag{20}$$

$$F(s) = \frac{1}{\left(\frac{1}{K_p}+1\right)\frac{s^3}{\omega_n^3} + \left(\frac{1}{K_p}+1\right)\left(\frac{2\mathcal{E}+1}{\omega_n^2}\right)s^2 + \left(\frac{1}{K_p}+1\right)\left(\frac{2\mathcal{E}+1}{\omega_n}\right)s + \frac{1}{K_p}+1} \qquad (21)$$

Equating coefficients of like powers of s of equations 19 and 21 the following results:

$$(K_p+1)\,\omega_1\,\omega_{n_0}{}^2 = \omega_n{}^3 \qquad (22)$$

$$\left(\frac{2}{\omega_1\omega_{n_0}} + \frac{1}{\omega_{n_0}{}^2}\right) = (K_p+1)\left(\frac{2\mathcal{E}+1}{\omega_n{}^2}\right) \qquad (23)$$

$$\left(\frac{1}{\omega_1} + \frac{2}{\omega_{n_0}}\right) = (K_p+1)\left(\frac{2\mathcal{E}+1}{\omega_n}\right) \qquad (24)$$

The open loop gain, $K_p$ , is 50 as outlined earlier. The closed loop damping, $\mathcal{E}$, of a three pole Butterworth is 0.5. The open loop poles in terms of the closed loop cut off frequency $\omega_n$ were computed as follows:

$$\omega_1 = 2(\omega_n - \omega_{n_0}) \qquad (25)$$

$$\omega_{n_0} = 0.99\,\omega_n \qquad (26)$$

The cut off frequency, $\omega_n$, is selected as $2\pi \cdot 10^{+6}$ rad/sec as outlined earlier. The parameters of interest become:

$$\omega_n = 2\pi \cdot 10^{+6} \ \frac{rad}{sec} \tag{27}$$

$$\omega_1 = 1.25 \cdot 10^{+5} \ \frac{rad}{sec} \tag{28}$$

$$\omega_{nn} = 6.22 \cdot 10^{+6} \ \frac{rad}{sec} \tag{29}$$

$$K_p = 50 \tag{30}$$

$$\mathcal{E} = 0.5 \tag{31}$$

The open loop transfer function is

$$G(s) = \frac{50}{\left(\frac{s}{1.25 \cdot 10^5} + 1\right)\left(\frac{s}{6.22 \cdot 10^6} + 1\right)^2} \tag{32}$$

The closed loop transfer function becomes

$$F(s) = \frac{1}{\left(\frac{1}{50} + 1\right)\left(\frac{s}{2\pi \cdot 10^6} + 1\right)\left(\frac{s^2}{(2\pi \cdot 10^6)^2} + \frac{2(.5)}{2\pi \cdot 10^6} s + 1\right)} \tag{33}$$

The open and closed loop responses indicated by equations 32 and 33 are plotted in figure 4. The group delay of the closed loop transfer is plotted in figure 5. The Root locus plot of the system is shown in figure 6. Figure 5 indicates that the group delay variation to 500 kc is approximately 23 nanoseconds. If the cut off frequency is selected to achieve constant group delay rather than 0.1 db droop in amplitude response the cut off frequency must be extended from 1 mc to 2.5 mc. Figures 7 and 8 indicate the amplitude response and group delay for $\omega_H$ of $2\pi \cdot 2.5 \cdot 10^{+6} \frac{rad}{sec}$. The variation in group delay is decreased from 27 nanoseconds to 3 nanoseconds. The amplitude droop at 500 kc is decreased from 0.1 db to .0003 db. The attenuation to the 50 mc discriminator harmonic is -80 db.

Equations 27 thru 33 must be modified as follows:

$$\omega_H = 5\pi \cdot 10^{+6} \frac{rad}{sec} \tag{34}$$

$$\omega_1 = 3.14 \cdot 10^{+5} \frac{rad}{sec} \tag{35}$$

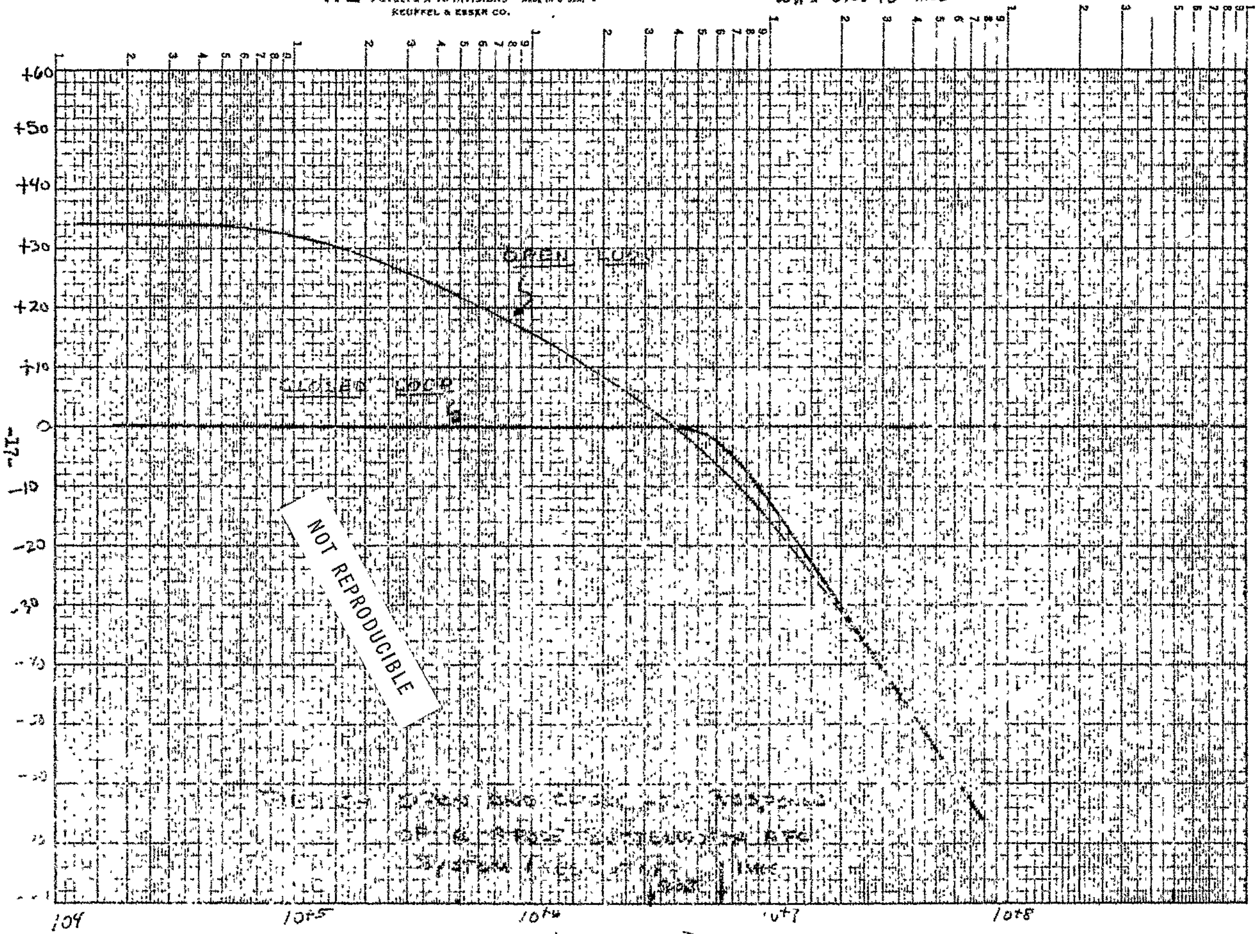
$$\omega_{n_o} = 15.54 \cdot 10^{+6} \frac{rad}{sec} \tag{36}$$

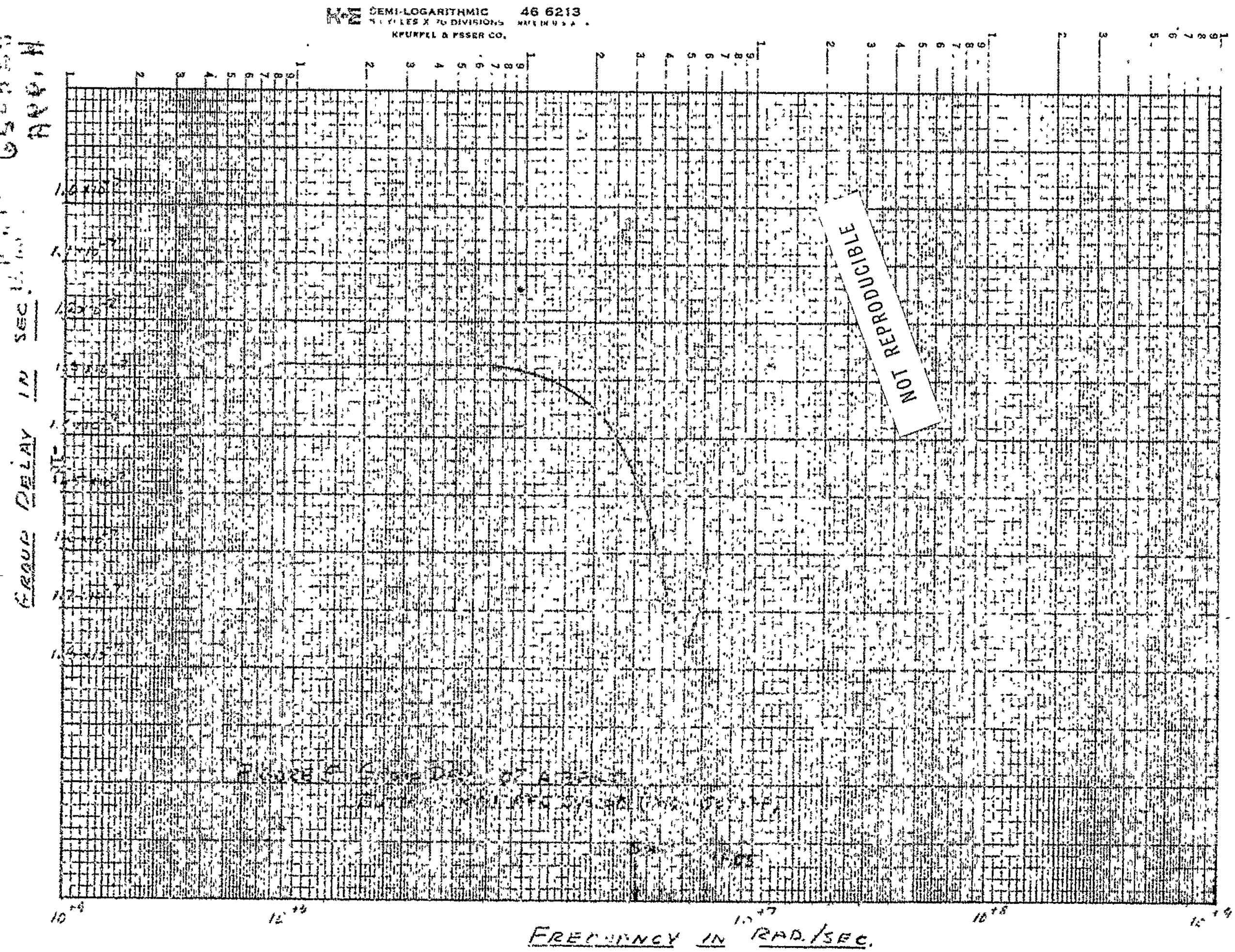
$$K_P = 50 \tag{37}$$

K+E SEMI-LOGARITHMIC 46 6213
5 CYCLES X 70 DIVISIONS MADE IN U.S.A.
KEUFFEL & ESSER CO.

$f_n = 10^{+6}$ mcs $\omega_{no} = 0.98\,\omega_n$

$\omega_n = 6.28 \cdot 10^{+6}$ mcs $\omega_1 = .04\omega_n$

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GROUP DELAY IN SEC
FREQUENCY IN RAD./SEC.

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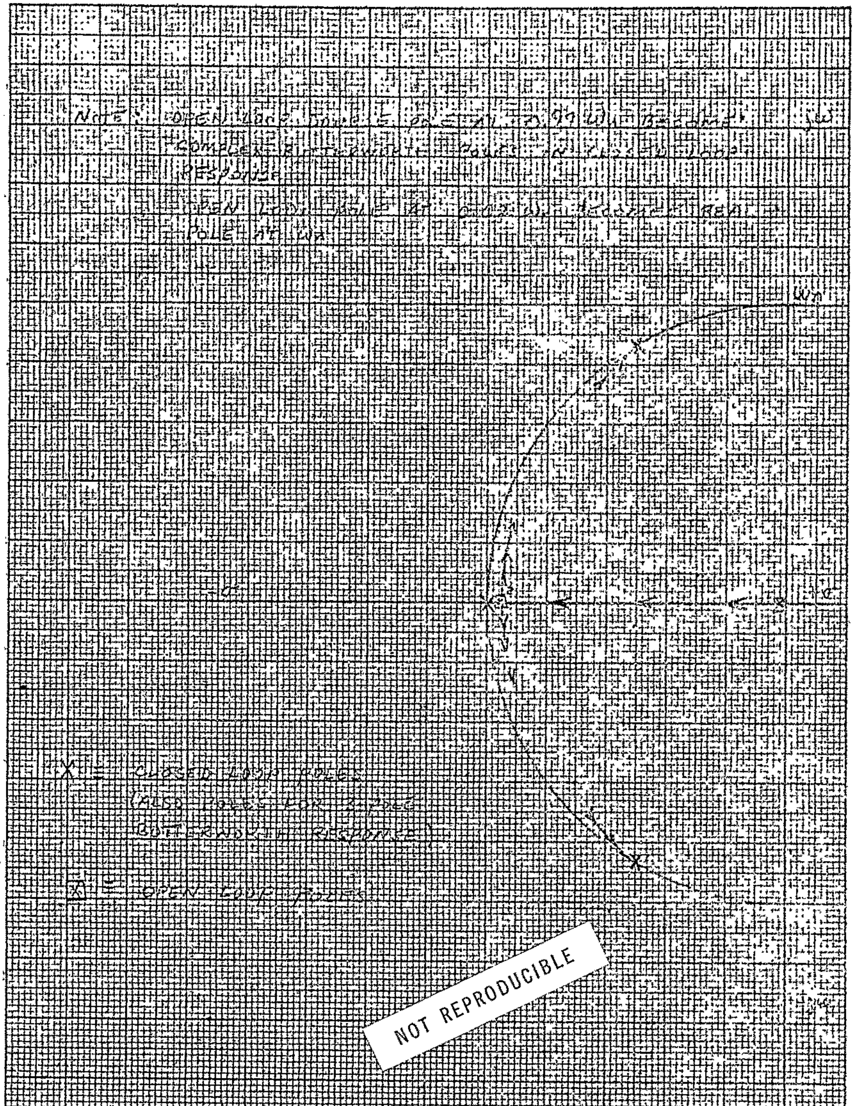
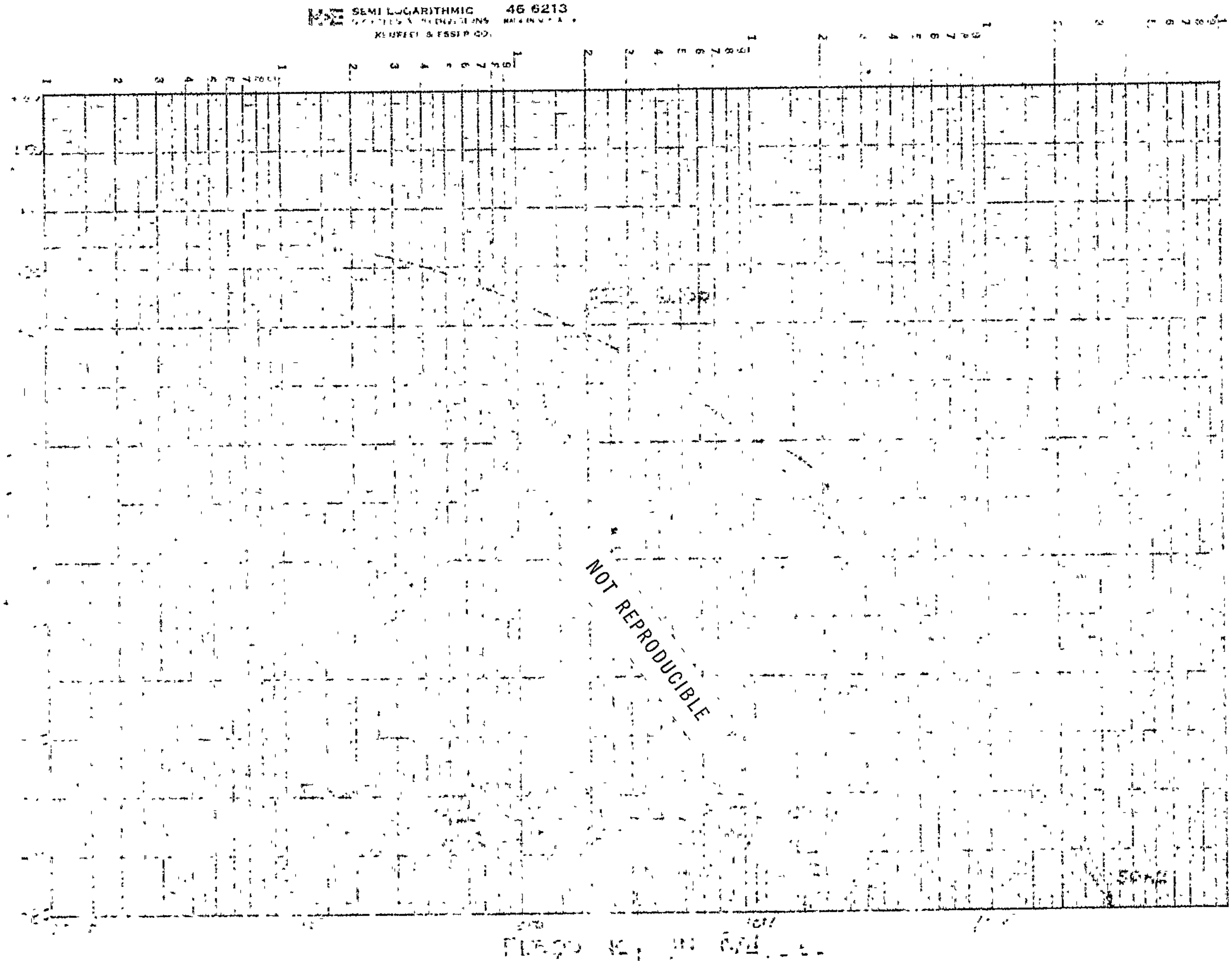



FIGURE 6. ROOT LOCUS OF THREE POLE BUTTERWORTH AGC SYSTEM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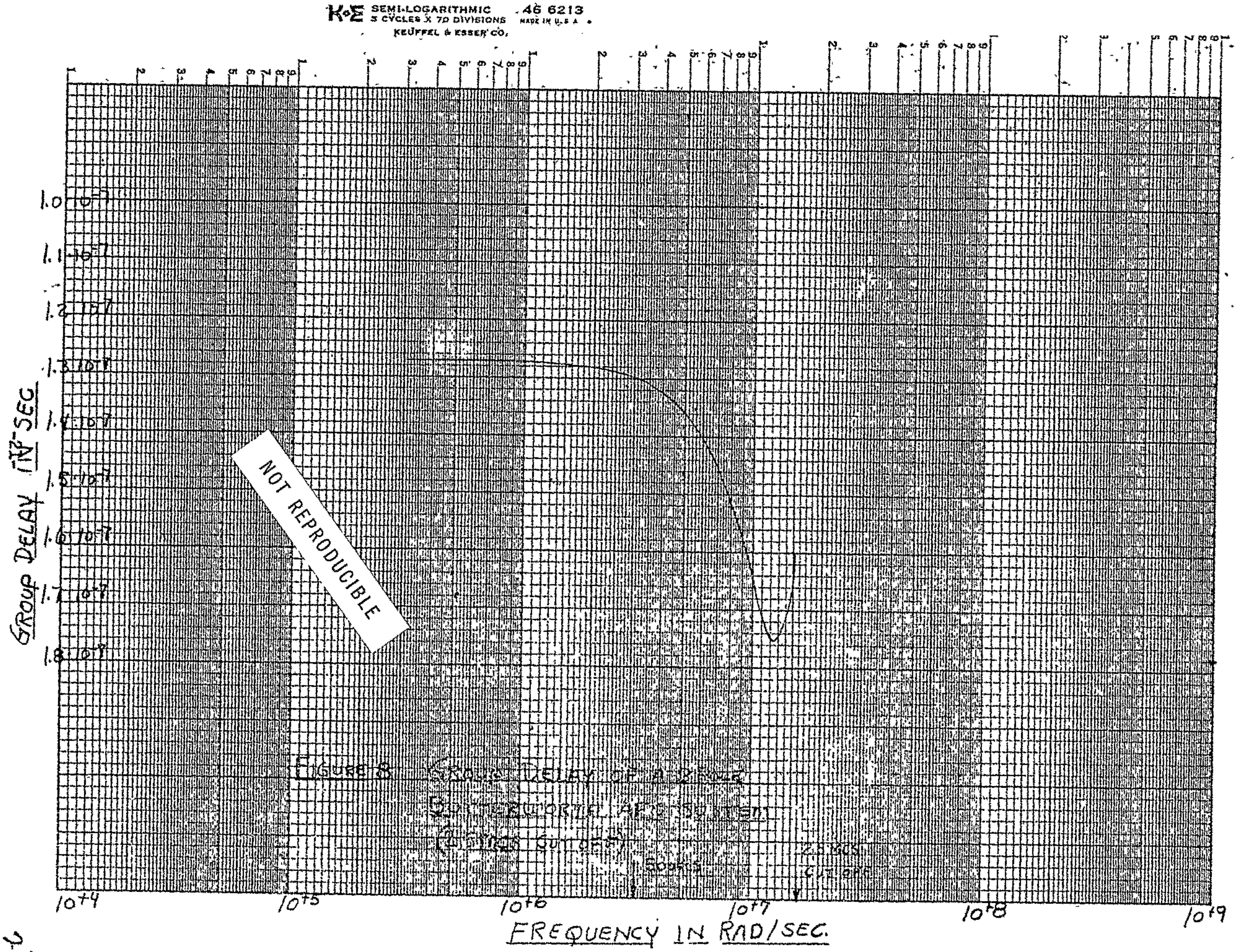


Figure 8. Group Delay of a 2-Pole Butterworth APS System (2.5 MC cutoff)

$$\xi = 0.5 \tag{38}$$

$$G(s) = \frac{50}{\left(\frac{s}{3.14\ 10^{5}} + 1\right)\left(\frac{s}{15.59\ 10^{6}} + 1\right)^{2}} \tag{39}$$

$$F(s) = \frac{1}{\left(\frac{1}{50} + 1\right)\left(\frac{s}{5\pi\ 10^{4}} + 1\right)\left(\frac{s^{2}}{(5\pi\ 10^{4})^{2}} + \frac{2(.5)}{5\pi\ 10^{4}} + 1\right)} \tag{40}$$

B. <u>AFC Loop Linearity Considerations</u>

The simplified model of the AFC loop shown in figure 3 is repeated in figure 4. The principal loop non linearity is attributed to the VCO. The linearity of the delay line discriminator considered has been measured as 0.3%. Therefore, the VCO non linearity is considered as a loop input ( $\Phi_{vco}(s)$ ) and the transfer $\frac{F_o}{\Phi_{vco}}(s)$ examined as shown by equation 41.

$$\frac{F_o}{\Phi_{vco}}(s) = \frac{K_{vco}}{1 + K_{vco} K_m K_\alpha F(s)} \tag{41}$$

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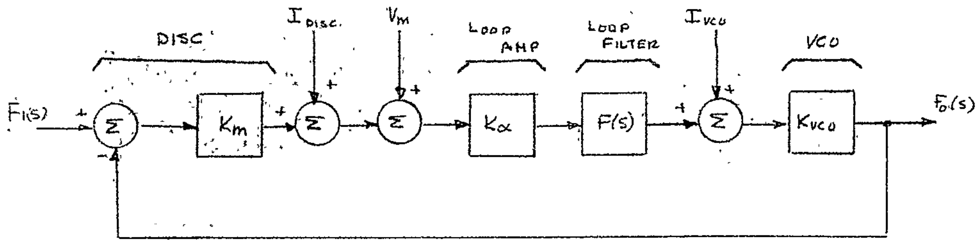



FIGURE 9. LINEAR FM MODULATOR WITH SIMULATED INTER-MODULATION SOURCES

Let $K_\rho = K_{vco} K_m K_\alpha$

$$\frac{F_o}{I_{vco}}(s) = \frac{1}{K_m K_\alpha}\left[\frac{K_\rho}{K_\rho + F(s)}\right] \tag{42}$$

As outlined earlier $F(s) = \dfrac{1}{\left(\frac{s}{\omega_1}+1\right)\left(\frac{s}{\omega_{n_2}}+1\right)^2}$

$$\frac{F_o}{I_{vco}}(s) = \frac{1}{K_m K_\alpha}\left[\frac{\left(\frac{s}{\omega_1}+1\right)\left(\frac{s}{\omega_{n_2}}+1\right)^2}{\left(\frac{s}{\omega_n}+1\right)\left(\frac{s^2}{\omega_n^2}+\frac{2\zeta s}{\omega_n}+1\right)}\right] \tag{43}$$

The VCO intermodulation distortion is attenuated by the loop as indicated by equation 43 and shown in figure 10.

Similarly, the intermodulation distortion introduced by the discriminator, $I_{DISC}$, is attenuated by the gain of the discriminator. Therefore, in the AFC mode the VCO non-linearity is attenuated by the gain as shown in figure 10. The loop performs little correction on the discriminator non-linearity. Therefore, a successful FM modulator, organized as indicated, requires a linear discriminator. The delay line discriminator outlined on pages 30 to 38 in the FM Receiver report (Appendix I) references a unit that exhibits 0.3% distortion at 20 percent bandwidth. This discriminator is applicable or at least provides a basis for initial design.

C. <u>AFC Carrier Tracking Loop</u>

The previous sections indicated one form of F.M. Modulator whereby the amplitude response and group delay of the output as a function of modulation is forced to assume a three pole Butterworth characteristic. Further, the loop attenuation of the system's principal source of non-linearity (VCO) was examined. One may conclude, that if the system can be fabricated as outlined, the results would be compatible with the spec. Unfortunately, the system has one serious shortcoming. The discriminator (figure 9) provides the loop reference frequency $F_i(s)$

The discriminator linearity and slope sensitivity are inversly proportional to bandwidth. The discriminator reference stability ( $F_i(s)$ ) is directly proportional to the bandwidth. As outlined earlier, a linear

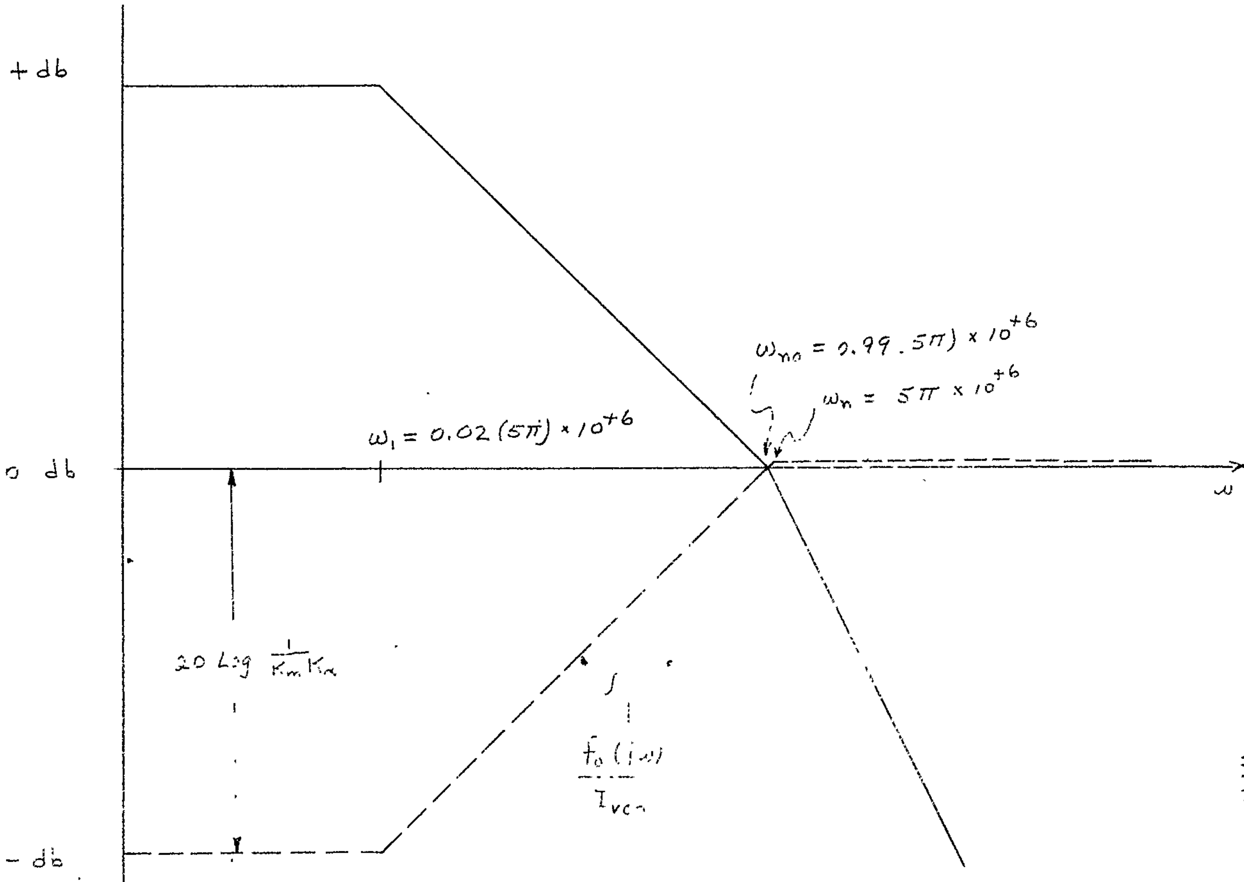



FIGURE 10 AFC LOOP RESIDUAL VCO NONLINEARITY

discriminator is essential, the reference instability must be accepted. However, if the reference frequency is allowed to vary, the VCO output, $F_o$, will follow. The rate of change of $F_o$ will be detected in the receiver as residual F.M. Therefore, the system must provide a means of stabilizing the VCO output, aside from the AFC loop.

Two systems were considered. The first consists of a narrow band auxiliary AFC Loop whereby the discriminator characteristics include narrow bandwidth, high slope sensitivity and stable reference (i.e. Crystal Discriminator). The block diagram of this system is indicated in figure 11.

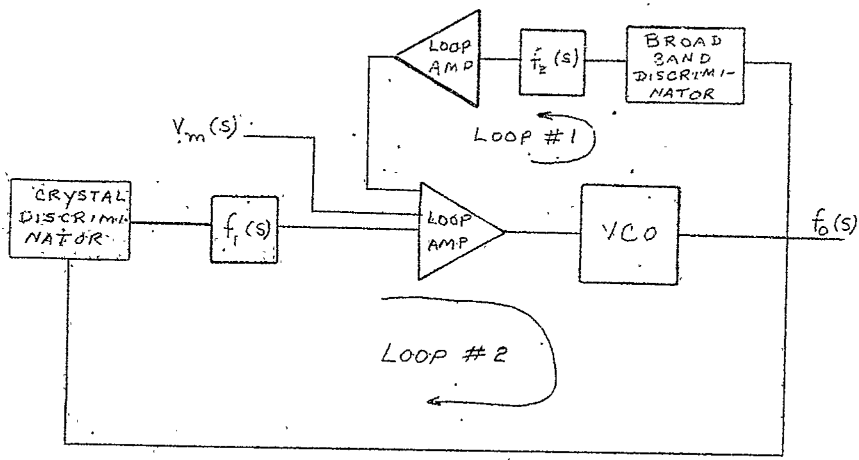



FIGURE 11. SIMPLIFIED DIAGRAM TWO DISCRIMINATOR AFC SYSTEM

In general terms, loop #2 can be replaced with an equivalent linear VCO and loop #1 analyzed for its influence on the average center frequency. The drift and drift rate of the output center frequency of this system is determined primarily by the crystal discriminator stability and the characteristics of loop #1. The worst case crystal discriminator reference frequency stability is approximately ± .01 % or ± 5 kc at 50 mcs based on the characteristics of units previously designed. The residual FM is specified as 15 cps. Therefore, the system shown in figure 11 is not considered applicable.

A second system considered to stabilize the VCO center frequency is outlined in figure 12.

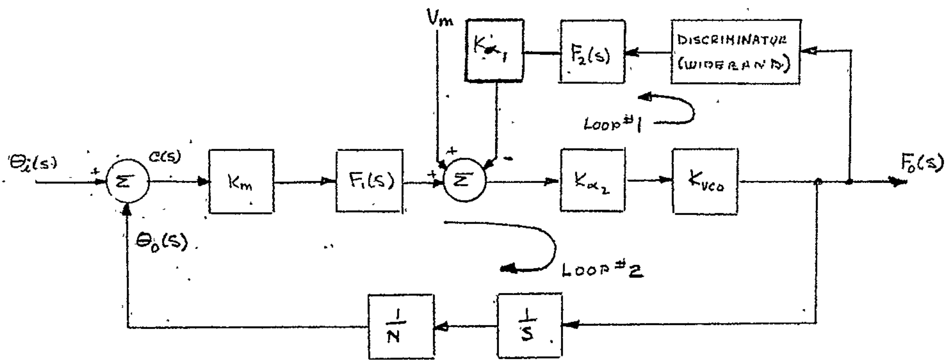



FIGURE 12. LINEAR LOCKED OSCILLATOR FREQUENCY MODULATOR
WITH AFC LINEARIZATION OF VCO

It is assumed that loop 1 can be replaced with a linear VCO yielding the model of Figure 13.

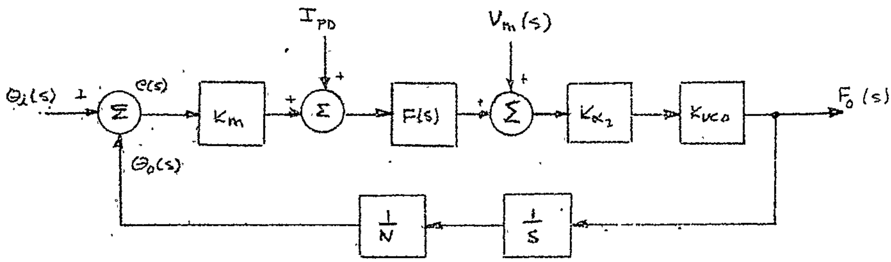



FIGURE 13. SIMPLIFIED LOCKED OSCILLATOR FREQUENCY MODULATOR

The transfer function of interest, $\frac{F_o}{V_m}(s)$, is expressed as follows:

$$\frac{F_o}{V_m}(s) = \frac{K_{\alpha_2} K_{vco}}{1 + \frac{K_{\alpha_2} K_{vco} K_m F(s)}{N s}} \tag{44}$$

Let $K_v = \dfrac{K_{\alpha_2} K_m K_{vco}}{N}$

$$\frac{F_o}{V_m}(s) = \frac{N}{K_m}\left[\frac{K_v}{1+\frac{K_v F(s)}{s}}\right] = \frac{N}{K_m}\left[\frac{1}{\frac{1}{K_v}+\frac{F(s)}{s}}\right] \tag{45}$$

Assume the loop filter is of the form

$$F(s) = \frac{\tau_2 s + 1}{\tau_1 s + 1} \tag{46}$$

$$\frac{F_o}{V_m}(s) = \frac{N}{K_m}\left[\frac{1}{\frac{1}{K_v}+\frac{\frac{\tau_2 s+1}{\tau_1 s+1}}{s}}\right] = \frac{N}{K_m}\left[\frac{1}{\frac{s(\tau_1 s+1)+K_v(\tau_2 s+1)}{K_v\, s(\tau_1 s+1)}}\right] \tag{47}$$

$$\frac{F_o}{V_m}(s) = \frac{N}{K_m}\left[\frac{s(\tau_1 s+1)}{\frac{\tau_1 s^2}{K_v}+\left(\tau_2+\frac{1}{K_v}\right)s+1}\right]_{\tau_2+\frac{1}{K_v} \ll 1} \tag{48}$$

$$= \frac{N}{K_m}\left[\frac{s(\tau_1 s+1)}{\frac{\tau_1 s^2}{K_v}+\tau_2 s+1}\right]$$

Let $\omega_n = \sqrt{\frac{K_v}{\tau_1}} \qquad \tau_2 = \frac{2\epsilon}{\omega_n} \qquad \tau_1 = \frac{1}{\omega_1}$

Equation 48 is of the form

$$\frac{F_o}{V_m}(s) = \frac{N}{K_m}\left[\frac{s\left(\frac{s}{\omega_1}+1\right)}{\left(\frac{s^2}{\omega_n^2}+\frac{2\epsilon}{\omega_n}s+1\right)}\right] \tag{49}$$

Let $\epsilon = 0.707$ and $s = j\omega$

The amplitude vs frequency response of equation 49 is

$$\left|\frac{F_o}{V_m}\right| = \frac{N}{K_m}\left[G(s) \cdot G(-s)\right]^{1/2} \tag{50}$$

$$\left|\frac{F_o}{V_m}\right| = \frac{N}{K_m}\left\{\frac{\omega\sqrt{\frac{\omega^2}{\omega_1^2}+1}}{\sqrt{\frac{\omega^4}{\omega_1^4}+1}}\right\} \tag{51}$$

$$\left| \frac{F_o}{V_m} \right| = 20 \log \frac{N}{K_m} + 20 \log \omega + 10 \log \left( \frac{\omega^2}{\omega_1^2} + 1 \right) - 10 \log \left( \frac{\omega^4}{\omega_n^4} + 1 \right) \qquad (52)$$

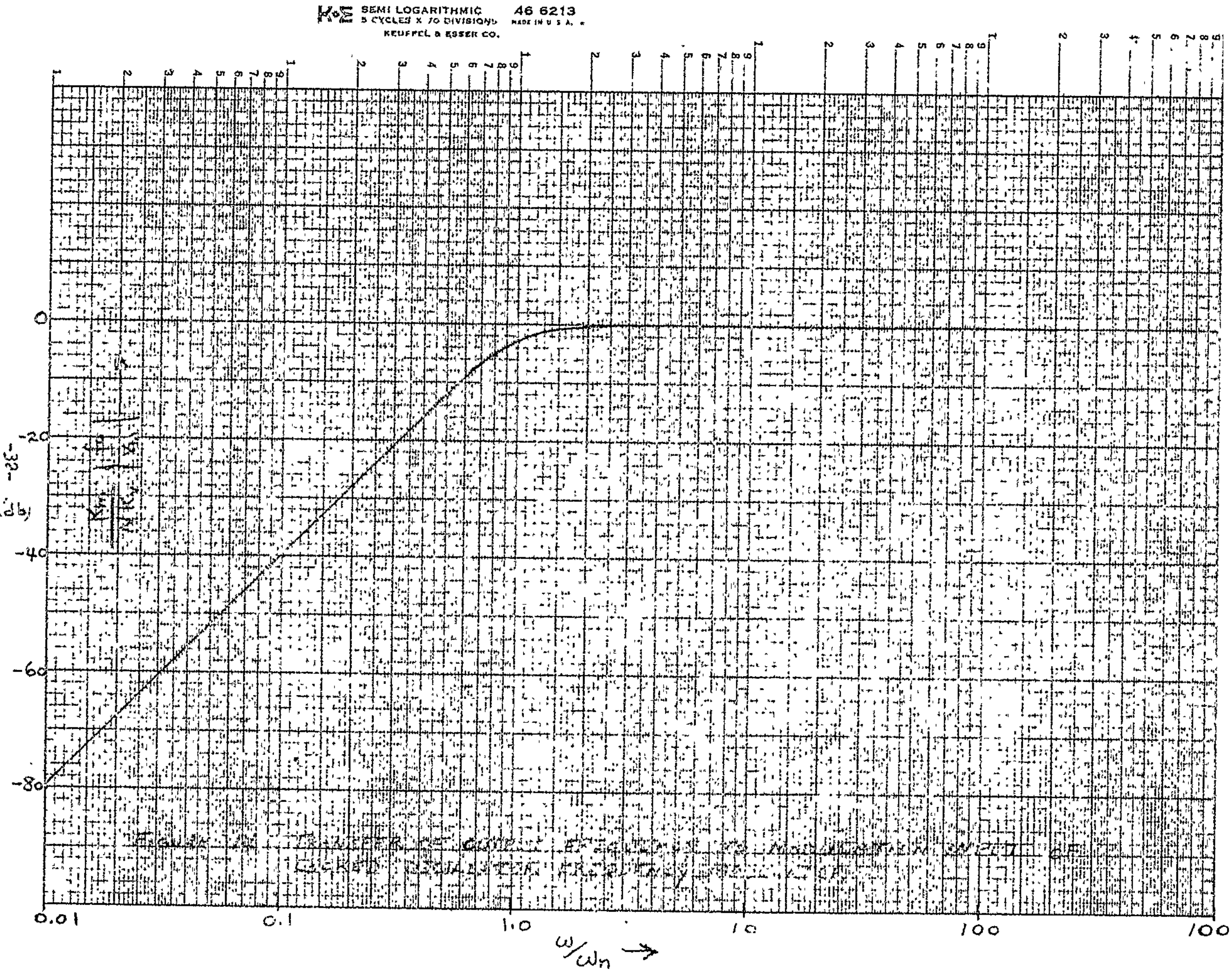
Figure 14 indicates the plot of equation 52.

The high pass characteristic of $\frac{F_o}{V_m}(s)$ is not compatible with the amplitude response specification ($\pm$ 0.1 db from 50 cps to 100 kcs and $\pm$ 0.5 db from 3 cps to 50 cps and 100 kc to 500 kc), if the lower bound of the baseband approaches $\omega_n$. Although a numerical value of $\omega_n$ has not been established it appears practical to pre-emphasize the modulation baseband and thereby extend the low frequency portion of the baseband. The pre-emphasis network transfer function must be the inverse of equation 49 as indicated by equation 53.

$$F(s) = \frac{K_m}{N} \left[ \frac{\frac{s^2}{\omega_n^2} + \frac{2\varepsilon}{\omega_n} s + 1}{s \left( \frac{s}{\omega_1} + 1 \right)} \right] \qquad (53)$$

If $\omega_1 << 1$ equation 53 becomes

$$F(s) = \frac{K_m}{N} \left[ \frac{\frac{s^2}{\omega_n^2} + \frac{2\varepsilon}{\omega_n} s + 1}{s^2} \right] \qquad (54)$$


K+E SEMI LOGARITHMIC 46 6213
5 CYCLES X 70 DIVISIONS MADE IN U.S.A.
KEUFFEL & ESSER CO.
db
0
-20
-40
-60
-80
0.01
0.1
1.0
10
100
100
ω/ωn →

App H

Equation 54 is of the form

$$\frac{e_o}{e_{in}}(s) = K\left[\frac{As^2+Bs+1}{s^2}\right] \tag{55}$$

$$e_o(s) = K\left[Ae_{in}(s) + e_{in}(s)\frac{B}{s} + \frac{e_{in}(s)}{s^2}\right] \tag{56}$$

Equation 56 may by synthesized with either active or passive components. The passive synthesis is not without problems, as the required component values are unrealistic. The active synthesis is a standard analog computer problem. Figure 15 shows one means of mechanizing equation 56.

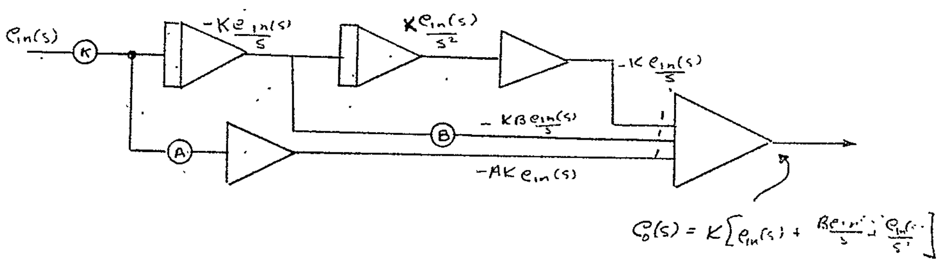



FIGURE 15. <u>ACTIVE PRE-EMPHASIS MECHANIZATION</u>



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Aside from the transfer of $\frac{F_o}{V_m}(s)$ and the form of the pre-emphasis network, the influence of the phase detector harmonics and the phase detector reference feed through on the output is of interest. Refer to figure 13. The transfer function $\frac{F_o}{V_m}(s)$ is repeated by equation 57 (A linear system is assumed and superposition applied)

$$\frac{F_o}{V_m}(s) = \frac{K_\alpha K_{VCO}}{1 + \frac{K_m K_\alpha K_{VCO} F(s)}{N s}} \tag{57}$$

The transfer $\frac{F_o}{I_{PD}}(s)$ is indicated by equation 58.

$$\frac{F_o}{I_{PD}}(s) = \frac{K_\alpha K_{VCO} F(s)}{1 + \frac{K_m K_\alpha K_{VCO} F(s)}{N s}} = F(s)\left[\frac{K_\alpha K_{VCO}}{1 + \frac{K_m K_\alpha K_{VCO} F(s)}{N s}}\right] \tag{58}$$

Equation 58 divided by equation 57 yields

$$\frac{V_m}{I_{PD}}(s) = F(s) \tag{59}$$

Equation 59 indicates that the relative magnitude of $V_m$ to $I_{PD}$ is determined by the APC loop filter $F(s)$. The loop filter is of the form $\frac{\tau_2 s+1}{\tau_1 s+1}$, a simple lag lead configuration. If the lower frequency component of $I_{PD}$ (the phase feedthrough) is greater than $\frac{1}{\tau_2}$ and if $\tau_1$ equals $1000\,\tau_2$, $I_{PD}(j\omega)$ becomes -60 db with respect to $V_m$. Numerical values will be assigned later in the report.

D. APC Loop Modulation Considerations

In implementing this phase locked AFC unit, the most critical problem is developing a system that will operate with all of the variations in the modulation parameters that will be impressed on the desired carrier frequency. Specification 3.5.2.2.4 defines the modulating frequency range of 1 cps to 500 kc. The lower limit of the above range was raised to 3 cps after discussing this point with JPL personnel. Also, the peak frequency deviation value given in specification 3.5.2.2.3 (b) is 500 kc. Hence the above modulation parameters are the values that will be employed in the following design considerations for the AFC unit.

The phase lock AFC system is a carrier tracking loop. Hence for proper operation of the loop two basic conditions must be fulfilled: First, a spectral line at the center frequency must exist that is either displaced from the sidebands generated by the modulation process or if the sidebands are close to the carrier frequency they must be at least 20 db down relative to the strength of the carrier frequency line. Second, a given minimum amount of power must always be present in the carrier frequency component in order to insure proper operation of the phase detector. The question of whether the above two operating conditions can be met will mainly depend on the magnitudes and characteristics of the modulation parameters of the FM wave.

The ideal assumption for the above parameters is that the carrier frequency is modulated with a single tone frequency. In this case, a spectral line will be present at the desired center frequency; hence, the first condition can be fulfilled. However, the second condition can't necessarily be met since the energy content of the center frequency can vary from a high value for $M < 1$ to a very low value for $M \sim 2.4$

A plot of the energy content of the carrier versus M approximates a $\sin x / x$ function. Hence, no conclusions regarding the discrete tone case can be made until some operational range of values for M have been stated as a definite requirement. Actually, the assumption of a discrete tone modulating signal isn't realistic because this type of signal doesn't contain any useful information at the receiver. In practice, a signal that contains useful information is actually made up of a large number of discrete tones. Hence, this type of signal will be assumed for the design of the AFC loop.

Due to the fact that a multitone signal can't be readily represented in a concise mathematical form, a simplifying assumption concerning its characteristics must be made. It is assumed that the modulating multitone signal has the properties of white Gaussian noise. Hence, the problem reduces to determining the resulting power spectrum of a carrier frequency that has been frequency modulated with white Gaussian noise.

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App. H

For the case of a noiselike signal FM modulating the carrier, the important parameters to be considered are the minimum and maximum frequencies ($f_1$ and $f_N$) of the modulating spectrum, which in this case is assumed to be uniform. Also the rms deviation of the carrier, D, and not the peak deviation is the important parameter. For a very good approximation, the peak deviation of the carrier can be taken as four times the rms deviation. Of course, there is a finite probability that the peak value of the carrier will exceed 4 times the rms value if the observation time is long enough; however, from a practical standpoint, the magnitude of the above factor is sufficient.

From a practical standpoint, it can be assumed that $f_1 << f_N$. Hence, the effective rms modulation index m can be given as $D/f_n$. It can be shown that the resulting FM spectrum, for the case under consideration, for an $m \geq 2$ is given by the expression,

$$f_N \cdot w_1(x) = \frac{1}{\sqrt{2\pi}\, M} \cdot exp^{-x^2/2m^2} \qquad (60)$$

where $x = \frac{f}{f_n}$ - the normalized relative frequency.

The total input carrier power is taken as unity for convenience. An important factor to note in this case is that the resulting FM spectrum is independent of the shape of the modulation spectrum and also of the values $f_1$ and $f_N$ and is only dependent on the rms carrier deviation, D.

(1) "Power Spectrum of a Carrier Modulated in Phase or Frequency" Electronic and Radio Engineering, July 1957, p. 246

Also for this case the resulting spectrum is continuous; hence, the frequency spread at the important interval of the center frequency is a continuum. For this condition, the PLL unit won't operate in the correct manner since it requires a distinct spectral line.

In order to obtain a distinct spectral line at the carrier frequency, the rms modulation index of the input FM wave must be less than unity. Further restrictions are also imposed in the form of the factor $m^2/x^2$ or $D^2/f_1 f_n$. From an analytical standpoint, the resulting FM spectrum can only be solved for the condition of $D^2/f_1 f_n >> 1$ and $D^2/f_1 f_n << 1$. Also the resulting FM spectrum for the case of $m < 1$ is considerably influenced by the value of $f_1$. For the case when $D^2/f_1 f_n >> 1$, the FM spectrum is given by the following expression:

$$f_n \cdot w_2(x) = \frac{2m^2}{\pi^2 m^4 + 4x^2} \qquad (61)$$



In this case, the FM spectrum is continuous; hence, a frequency smear will exist in the region of the center frequency. This condition is similar to the case of $m > 1$. For the condition of $D^2/f_1 f_n << 1$, the overall power is divided into a residual carrier factor which is given by the expression:

$$W_D = \exp\left(-D^2/f_1 f_n\right) \qquad (62)$$



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and a continuous function which is given by

$$f_N \cdot W_2(x) = \frac{m^2}{2x^2(1-x_1)}. \qquad (63)$$

Hence, the magnitude of the power given by both expressions must be summed to unity. It can be shown that if $D^2/f_1 f_N = .1$, the power in the carrier of the resulting FM wave will have .9 of the total power. Hence, the remaining 10% of the power will exist in the sidebands of the FM wave. Therefore, it follows that a distinct spectral line will exist at the center carrier frequency if the above condition on D, $f_1$, and $f_N$ is fulfilled. Stated in another form, since $f_1$ and $f_N$ are usually taken as independent variables, the allowable value of D will be $.315 \sqrt{f_1 f_N}$.

For the specified values of $f_1$ and $f_N$ of 3 cps and 500 kc the allowable rms carrier deviation would be only 387 cycles/sec. This marginal value would clearly not be acceptable in a practical system. Hence, in order to increase the allowable value of D and still obtain the desired distinct spectral line, a frequency divider will be inserted between the FM modulation and the input to the phase detector. In this case, if the FM wave is divided by a factor N, then the allowable value of D will be $.315N \sqrt{f_1 f_N}$. From specification 3.5.2.2.3 it follows that the required value of D will be 125 kc. Hence, the required value of N is 325.5. From a design standpoint, the value of N must be some multiple of the base two. Therefore, the division factor of 512 which is equal to $2^9$ will be employed.

The need for the above division approach stems from the fact that given a continuous power spectrum at the frequency of interest, no other method could be found to obtain the desired spectral line that is required for operation of the APC loop. Since the sideband frequencies are infinitesimally close to the center frequency no loop filter characteristic can be employed to obtain the desired attenuation of the sideband frequencies. Hence, the only alternative is to vary the modulation parameters of the FM wave to obtain the correct operating condition. The advantage of this division technique is that if the value of N is chosen to obtain the desired value of $D^2/f_1 f_n$ for the maximum rms modulation index, m, stated, any smaller value of m will only put more power in the carrier, thus increasing the desired operating conditions for the AFC loop.

E. APC Loop Parameters

The APC loop design is based on the procedure outlined by JPL. The following relationships are repeated.

$$G = \frac{2\pi K_m K_\alpha K_{vco}}{N} \tag{64}$$



$$F(s) = \frac{\tau_2 s + 1}{\tau_1 s} \tag{65}$$

$$\tau_1 = R_1 C = \frac{G}{\omega_n^2} \tag{66}$$

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$$\tau_2 = R_2 C = \frac{\sqrt{2}}{\omega_n} \qquad (67)$$

$$H(s) = \frac{\frac{\sqrt{2}}{\omega_n} s + 1}{\frac{s^2}{\omega_n^2} + \frac{\sqrt{2}}{\omega_n} s + 1} \qquad (68)$$

$$\frac{e(s)}{\theta_i(s)} = 1 - H(s) = \frac{s^2}{s^2 + \sqrt{2}\,\omega_n s + \omega_n^2} \qquad (69)$$

The APC loop design is determined primarily by the loop natural resonant frequency. The transfer of $\frac{F_o}{V_m}(s)$ is high pass as outlined earlier. Therefore, to extend the lower frequency range of $\frac{F_o}{V_m}(s)$, (aside from pre-emphasis) $\omega_n$ must be small as possible. Conversely, if $\omega_n$ (or loop gain) is too small the loop error may become excessive if either the reference or VCO frequency change rapidly. It is specified that the transmitter center frequency be variable ± 500 cps from 50 mc. The 50 mc reference built during Phase I exhibited a drift rate ($\dot{\omega}$) of $521 \cdot 10^{-4}$ rad/sec$^2$ as a function of temperature variations. However, the drift rate of the crystal reference frequency is not the limiting factor. The drift rate of the VCO or the detuning rate of the reference frequency are far more significant. The VCO will be an L/C oscillator. A typical drift rate of a precision L/C VCO is 1 part in $10^{+6}$ per sec. or 50 cycles/sec$^2$ at

50 mcs. Conversely, if the same detuning rate of the reference frequency is allotted, the APC loop design will be based upon loop parameters sufficient to constraint the loop phase error in the presence of a frequency ramp of 50 $\frac{\text{cycle}}{\text{secs}^2}$.

The loop error attributed to a frequency ramp is outlined as follows:

$$f_i(s) = \frac{\dot{\omega}}{s^2} \quad ; \quad \theta_i(s) = \frac{\dot{\omega}}{s^3} \tag{70}$$

$$e(s) = \frac{\dot{\omega}}{s^3} \cdot \frac{s^2}{s^2 + \sqrt{2}\,\omega_n s + \omega_n^2} \tag{71}$$

$$= \dot{\omega} \, \frac{1}{(s+r)\left[(s+\alpha)^2 + \beta^2\right]} \tag{72}$$

$$= \dot{\omega} \, \frac{1}{(s+r)\left[s^2 + 2\alpha s + \alpha^2 + \beta^2\right]} \tag{73}$$

where $r = 0, \; \alpha = \varepsilon \omega_n, \; \alpha^2 + \beta^2 = \omega_n^2, \; \beta = \omega_n \sqrt{1-\varepsilon^2}$ (74)

The inverse transform yields the following transient error

$$e(t) = \frac{\dot{\omega}}{\omega_n^2} \, e^{-\frac{\omega_n}{\sqrt{2}} t} \sin\left(\frac{\omega_n}{\sqrt{2}} t - \psi\right) \tag{75}$$

The steady state error is:

$$e(t)_{ss} = \frac{\dot{\omega}}{\omega_n^2} = \Delta\Theta_{ss}(rad) \qquad (76)$$

The maximum phase error becomes

$$e(t)\,max = \frac{1.043\dot{\omega}}{\omega_n^2} \text{ at } t = \frac{\pi\sqrt{2}}{\omega_n} \qquad (77)$$

If we choose to constrain e(t) max to 5.7° ($\frac{1}{10}$ rad) then

$$\omega_n = \sqrt{\frac{1.043\dot{\omega}}{0.1}} = \sqrt{32.8 \times 10^2} = 57.3\ rad/sec \qquad (78)$$

The loop gain is

$$G = \frac{2\pi K_m K_\alpha K_{vco}}{N}$$

$$G = \frac{6.28 \cdot 1 \cdot 10 \cdot 2 \cdot 10^{+6}}{5/2} = 2.45 \cdot 10^{+5} = 107.8\ dB \qquad (79)$$

$$\tau_1 = R_1 C = \frac{G}{\omega_n^2} = 74.6\ sec \qquad (80)$$

$$\tau_2 = R_1 C = \frac{\sqrt{2}}{\omega_n} = 0.0247 \text{ sec} \tag{81}$$

## III. Implementation of Circuits

### A. VCO

The transmitter VCO center frequency is stabilized by the carrier tracking APC loop. However, the APC loop performs virtually no correction on the VCO short term instability which is manifest by residual FM in the TX/RX pair. The TX/RX pair residual FM is specified over the band DC to 500 kc in the non-AFC mode. Therefore, VCO disturbances, introduced by the outside world, with frequency components from DC to 500 kc must be minimized. Further, the magnitude of these disturbances is defined by the ratio of the design goal residual FM to the maximum frequency deviation (20 log $\frac{15 \text{ cps}}{500 \text{ kcs}}$ or -90.5 db AFC and 20 log $\frac{60 \text{ cps}}{500 \text{ kcs}}$ or -78.4 db non AFC). If half the residual FM is allotted the transmitter and half the receiver, the transmitter residual FM becomes -96.5 db in the AFC mode. The VCO contribution tothe system residual FM is determined by the following:

1) Power Supply Ripple
2) Vibration
3) Sustaining Circuit Noise
4) Tank Circuit Noise

The power supply ripple must be less than -100 db with respect to the nominal voltage. Assume a 15 volt power supply. The ripple must be less than 150 μ volts. The power supply regulator must be a low noise unit. A cascaded emitter follower buffer between the power supply and VCO will provide additional filtering aside from the power supply regulator.

Cabinet blower motors, building vibrations etc. can excite mechanical resonances in the VCO tank circuit which in turn produce sidebands displaced from the carrier by the resonant frequency. This problem is more pronounced in airborne equipment; however, it must also be considered as applied to the R.F. Test Console. Several precautionary measures include the following: 1) The tank coil will be wound on a mechanically stable, grooved teflon bobbin 2) The VCO assembly will be encapsulated in polyurethane foam. The usual approach to vibration problems is to change the mechanical resonant frequency such that the resultant sidebands fall outside the band of interest. This procedure is not available to the designer as the band of interest extends from DC to 500 kc.

The oscillator sustaining circuit noise, unfortunately, prevents practical realization of the theoretical short term oscillator stability. The oscillator sustaining circuit noise will be minimized by using low noise transistors and quartz capacitors. Conversely, the VCO oscillation level must be relatively large to provide as large a signal to noise ratio as practical within the constraints of long term stability and linear operation of the oscillator.

The tank circuit design presents two conflicting choices. The tank circuit Q should be high as possible for good short term stability.

However, the VCO frequency response must extend to 500 kc with only $\pm$ 0.1 db variation. The VCOphase transfer function is $\frac{1}{S\left(\frac{S}{\omega_1}+1\right)}$

The VCO 3 db response occurs at $W_1$ whereby $W_1$ is defined as $\frac{\pi f_o}{Q}$

Conversely the VCO short term stability is inversly proportional to the oscillator Q as indicated by equation 82.

$$\frac{\Delta f}{f} = \frac{2\pi}{\tau}\sqrt{\frac{4KT}{PQf_o}} \qquad (82)$$

$\tau$ = integration time

Q = crystal storage factor

K = Boltzman's constant

P = signal power dissipated in tank circuit series resistance

T = absolute temperature

$f_o$ = oscillator center frequency

Clearly, the VCO cannot simultaneously exhibit the required short term stability and frequency response. The short term stability takes priority. The oscillator loaded Q will be a minimum of 200, about the limit for an L/C unit. The resulting oscillator 3 db corner occurs at $250 \times 10^{+3}$ rad/sec. The VCO driving amplifier must exhibit a zero at $250 \times 10^{+3}$ rad/sec $\left[K\left(S/\omega_1+1\right)\right]$.

The transmitter/receiver pair is allotted 1% dynamic linearity. If 0.5% linearity is assigned the transmitter the VCO must exhibit approximately 0.3% linearity. Two push pull VCO's that track will



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exhibit a linearity that is comprised of the difference in linearities of the two VCO's. However, the auxiliary AFC loop discussed earlier, will transfer the discriminator linearity (0.3%) to the VCO within the constraints of the AFC loop. In the Non AFC mode the transmitter linearity is determined primarily by the VCO.

B. Discriminator

The delay line discriminator outlined on pages 30 thru 38 of Appendix I, (F.M. Receiver), will be implemented.

C. Loop Amplifiers

The two operational amplifiers required for the AFC and APC loops will be the same configuration outlined in Appendix D (Locked oscillator Phase Modulator) page 108 thru 110.

D. Phase Detector

The phase detector inputs consist of the 50 mcs reference and 50 mcs VCO output divided by the feedback divider and the reference dividers respectively. The divider count (N) is a function of the modulation index and phase detector dynamic range. The minimum value of N of 512 has been suggested (N can be a variable if required). Assume $N = 512$, the phase detector inputs become $\frac{50.10^{+6}}{512}$ or 97.6 kc. The phase detector will be a digital unit as described on pages 126 to 133 of Appendix D, Locked Oscillator Phase Modulator.

E. Frequency Dividers

The frequency dividers must have a clock frequency of 50 mc and provide a minimum binary count $(N = 9)$ of 512. The value of N can be made larger and the modulation index increased accordingly. It is apparent however that as N becomes larger, the phase detector harmonics become lower in frequency. In the limit, the phase detector harmonic attenuation starts to diminish when the phase detector reference frequency is equal to $W_2$ (the APC loop filter pole). The first five binary counters reduce the unmodulated carrier from 50 mc to 50mc/32. The Thin Film flip flops outlined on pages 125 to 126 of Appendix D, Locked Oscillator Phase Modulator, are suitable. The clock speed of the remaining counters (approximately 1.5 mc) is relatively low and a standard transistor flip flop is suitable.

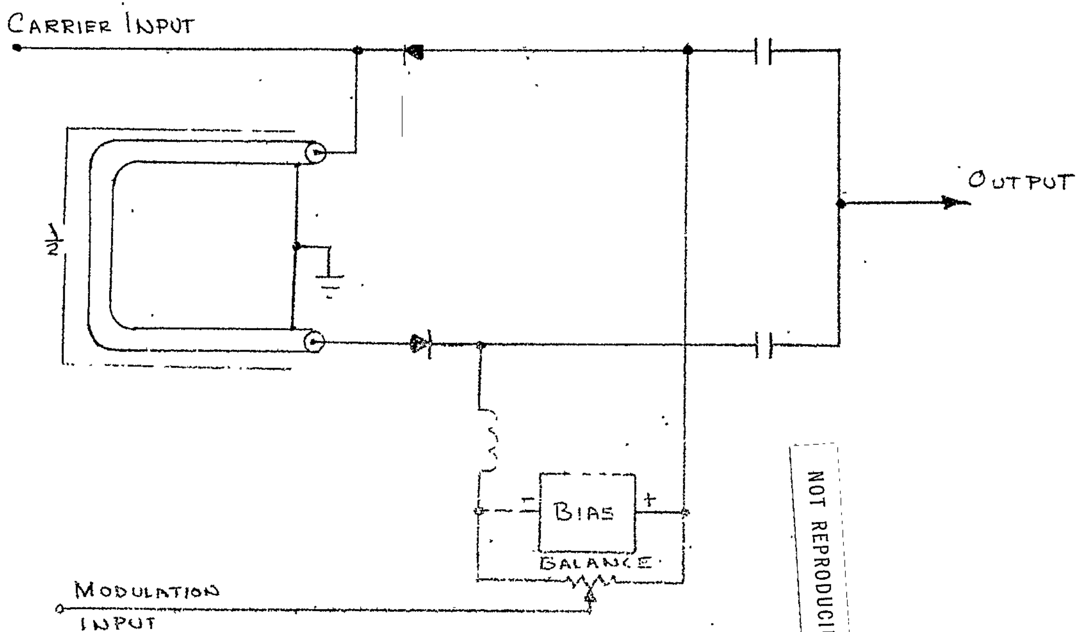
F. AM Modulator

In order to meet the requirements for the AM modulator stated, in paragraph 3.5.1.2.5 of the specification, the above unit must have a wide bandwidth. This follows from the fact that incidental FM modulation is produced from a variation in the circuit capacitances of the unit. In order to realize this requirement, it is best to employ an AM modulator that doesn't contain any band limiting transformers or active elements. Hence, it was concluded that an AM modulator that has good linear characteristics over the 0 to 5 modulation index range, can be operated with a modulating frequency range of 0 to 5 kc and has a bandwidth on the order of 20 mc would be optimum.

One type of AM modulator that was advertised to have the above characteristics was tested for use in the FM transmitter. It is a balanced type unit manufactured by the General Radio Company. An elementary schematic diagram of this unit is shown in Figure 16 . A unit was borrowed from the above company and tested with an FM signal generator. Due to the limitations of the test equipment, a peak frequency deviation of 240 kc and a modulating frequency of 60 kc was employed. The AM modulating signal was a 5 kc sine wave.

As a result of the tests, it was found that the AM modulation index, m, is a function of not only the amplitude of the 5 kc tone but also of the bias and balance settings. Hence, with these three variable adjustments, an m of 0 to 1 was easily obtained.

In determining the value of m as a function of the amplitude of the test tone to check the linearity of the modulator, a method proposed by General Radio was employed. This method entailed the use of a trapezoidal pattern that was displayed on a scope whose sides are denoted by A and B. It can be shown that the value of the modulation index, m, is equal to $B-A/_{B+A}$. It follows that as m approached zero, B approached A; therefore, the difference measurement B-A became very difficult to measure accurately. To within the accuracy of determining the magnitudes of B and A from the scope; it was found that perfect linearity between m and the input signal level of the test tone was obtained for values of $m > .3$. The procedure entails varying the amplitudes of the test tone in increments of 1 db, and then adjusting the bias and balance controls to obtain the desired value of m. Below the above value of m, the magnitude of the actual value of m was difficult to determine because of the small value of B-A. Hence, the use of this unit as the AM modulator will depend on whether an accurate measuring procedure for low values of m can be devised.

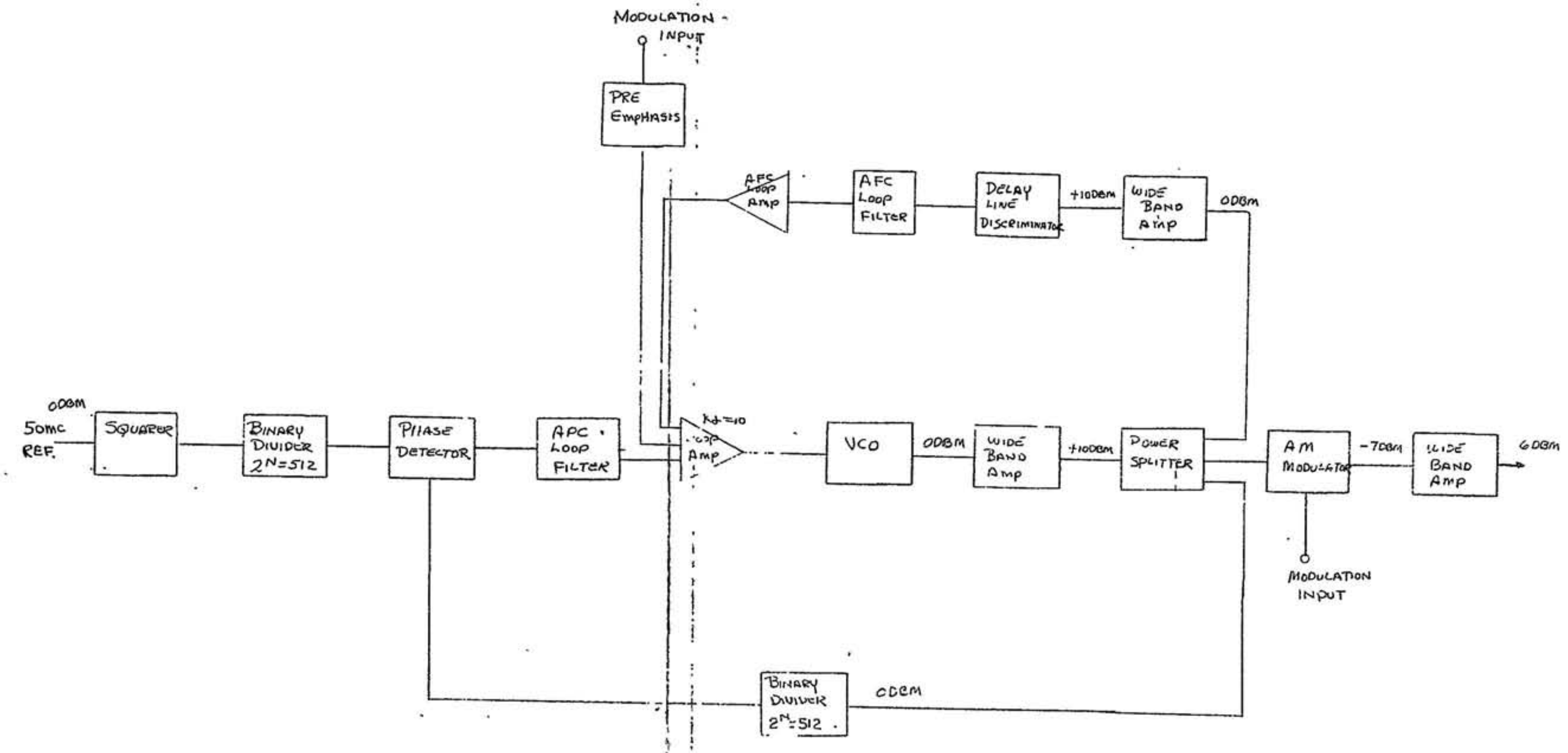



BALANCED MODULATOR

FIGURE 26

## IV. F.M. Transmitter Level Diagram

Figure 17 indicates the principal power levels around the APC and AFC loops.








ENGINEERING SKETCH

ENG.
APPROVED
CHARGE

TITLE
FIGURE 17
F.M. TRANSMITTER
LEVEL DIAGRAM

WESTINGHOUSE ELECTRIC CORPORATION
ELECTRONICS DIVISION
BALTIMORE, MD., U. S. A.

SK B

SHEET OF SHEETS

RECEIPT TYPE & FORMAT ☐ MAGNETIC

☐ LOAN ☒ PC ☐ 35 MM ☐ CARDS

☒ RETAIN ☒ MF ☐ 16 MM ☐ OTHER

STOCK RECEIVED FOR SALE

PC 0 MF

LOAN DOCUMENT RETURNED

DUE

OUT

TRANSACTION

NEW ITEM ☒ DUPE ☐ SUPERSEDES ☐ PRIOR NUMBER ☐

VOL 6826 ☐ TAB ☒ UNANN ☐ NO 1 03

14 REPRODUCTION INSTRUCTIONS

BLOWBACK → PRINT ↓

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| 1UP | 2 | 5 | 8 |
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27 PUBLIC RELEASABILITY ET

8 REPORT NUMBERS (XREF) ☒

NASA-CR-63908

9 RELATED DOCUMENT ☐

16 REMARKS

10 CONTRACTING OFFICE — BILLING CODE

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28 FORM & PRICE

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