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Third Quarterly Progress Report
for
COMMAND SYSTEM STUDY FOR THE
OPERATION AND CONTROL OF
UNMANNED SCIENTIFIC SATELLITES
TASK III
Command System Interference

1 January 1965 - 31 March 1965

Contract No. NAS 5-9705

Prepared for

Goddard Space Flight Center
Greenbelt, Maryland

Submitted by

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**ADVANCED COMMUNICATIONS • RESEARCH AND DEVELOPMENT**

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ABSTRACT

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This report covers progress during the reporting period of the Command System Study for the Control of Unmanned Scientific Satellites under Task III Command System Interference. The report contains a review of performance above threshold of AM, FM, and DSBSC modulation. An analysis of error rate degradation below threshold is presented for an FSK channel appearing as one of the subcarriers in an FDM/FM system. The effect of CW and modulated interference on the baseband of AM and FM demodulators is treated. An improved AGC circuit is described for overcoming the desired signal attenuation relative to its level without interference. Test procedures for evaluating the effects of CW interference are outlined. Specific conclusions are:

a. Discriminators are commonly used in place of phase demodulators in phase modulated command systems. It is shown that when the command information is carried by narrow subcarrier channels the loss of performance above threshold caused by the use of a discriminator for this purpose is insignificantly small.

b. CW interference in an AM command system employing narrow subcarrier channel is the least serious form for interference and has much less effect on the demodulated baseband than it does in a comparable angle modulated system. Unless the beat between the signal and interfering carriers falls within a subcarrier channel, a properly designed AM command receiver should not be seriously affected. Interference in an AM system from a narrowband angle modulated signal results in an effect very similar to CW interference. However, strong amplitude modulated interference in an AM command system may completely obliterate the command modulation through intermodulation distortion.

Author

# I INTRODUCTION

## 1.1 General

This report covers work during the first quarter on Task III of contract NAS 5-9705. A concurrent report on Task II entitled "Closed-Loop (Feedback) Verification Techniques" is being issued under separate cover. The objectives of Task III are:

To evaluate the effects of interference from earth and space emitters, spacecraft equipment, propagation anomalies, and thermal and sky noise upon the GSFC Standard Command Systems and the "Unified" system studied in task 1. Emphasis will be on the non-gaussian type noise/interference such as radar pulses, spurious CW emitters, and fading and multipath. Recommendations of appropriate measures to combat this interference will be included.

## 1.2 Summary of Work During Reporting Period

(a) The comparative performance above threshold in the presence of white noise interference was reviewed for amplitude modulation, angle modulation and double-sideband suppressed-carrier modulation.

(b) The performance in the region somewhat below threshold was analyzed for an FSK digital data channel comprising one of the subcarriers in an FDM/FM system.

(c) The influence of CW interference on AM and FM systems was studied. For AM, the effect of modulated interference was also investigated.

(d) An improvement in receiver design was evolved to combat one of the deleterious effects of interference.

(e) Procedures for testing interference effects on receivers were investigated, in particular, the details for CW interference tests.

# II DISCUSSION

## 2.1 System Performance in Gaussian Noise

### 2.1.1 PM, AM and DSBSC Above Threshold

#### 2.1.1.1 PM With Phase Demodulation

We will start our system comparison by analyzing the performance of a phase modulated system. A carrier phase modulated by a subcarrier can be represented by Eq. (1)

$$s(t) = \mathrm{Re}\left[A\, e^{j\beta \sin 2\pi f_m t}\, e^{j2\pi f_c t}\right] \tag{1}$$

A is the carrier amplitude, $\beta$ the peak deviation in radians and $f_m$ is the frequency of the subcarrier being considered. The signal described by Eq. (1) is typical of most angle modulated spacecraft command systems. The GRARR system is a good example of this.

Noise is added to the signal at the receiver input. We will here consider the case where the noise is gaussian and has a constant spectral density over the i-f passband of the receiver. This noise can be described by Eq. (2).

$$n(t) = \mathrm{Re}\left[\sum_{f_1}^{f_2} \sqrt{2 N_o \Delta f}\; e^{j\theta_n}\, e^{j2\pi f_n t}\right] \tag{2}$$

$N_o/2$ is the noise spectral density and $f_2 - f_1 = W_{i-f}$ is the passband of the i-f amplifier. The composite i-f signal resulting from signal and noise is given by Eq. (3).

$$v(t) = \mathrm{Re}\left[\left(A\, e^{j\beta \sin 2\pi f_m t} + \sum_{f_1'}^{f_2'} \sqrt{2 N_o \Delta f}\; e^{j(f_n' t + \theta_n)}\right) e^{j2\pi f_c t}\right] \tag{3}$$

where

$$f_n' = f_n - f_c \tag{4}$$

The carrier power is $C = A^2/2$. The noise power in the i-f bandwidth is given by Eq. (5).

$$\overline{n(t)^2} = N_o W_{i-f} \tag{5}$$

where

$$W_{i-f} = f_1' - f_2' = f_1 - f_2$$

As long as the i-f carrier-to-noise ratio exceeds 10 db the phase noise on the i-f signal can be expressed by Eq. (6)

$$\theta_n(t) = \frac{I_m\left[ \sum_{f_1'}^{f_2'} \sqrt{2 N_o \Delta f} \; e^{j\theta_n} \; e^{j2\pi f_n' t} \right]}{A} \tag{6}$$

The phase demodulator has an output voltage proportional to the phase angle of the i-f signal. The noise spectral density of the phase demodulator output is given by Eq. (7).

$$S_{\theta_n}(f) = \frac{N_o}{A^2} \quad , \quad -\frac{W_{i-f}}{2} < f < \frac{W_{i-f}}{2} \tag{7}$$

The signal component in the phase modulator output is:

$$\theta(t) = \beta \sin 2\pi f_m t \tag{8}$$

In this analysis we are interested in the signal-to-noise ratio in a particular subcarrier channel and not in the baseband signal-to-noise ratio in general. We will therefore center our analysis on the signal-to-noise ratio within the subcarrier channel bandwidth. We will call this bandwidth $W_{sc}$. The subchannel signal-to-noise ratio is:

$$(S/N)_{sc} = \beta^2 (C/N)_{i-f} \frac{W_{i-f}}{2W_{sc}} \tag{9}$$

Equation (9) is valid for a phase modulated system using a phase detector as a demodulator.

2.1.1.2 PM with Discriminator Demodulation

In most communication systems a frequency demodulator, usually a discriminator, is used in place of a phase demodulator for demodulation of phase modulated signals. This is usually done for the sake of convenience and system reliability. A true phase modulator requires a reference signal which is difficult to obtain. We shall see that the use of a discriminator for demodulation of phase modulated updata signals of the form given in Eq. (1) has little effect on the resulting subcarrier signals. The demodulator output is given by Eq. (10).

$$\dot{\theta}(t) = 2\pi f_m \beta \cos 2\pi f_m t \tag{10}$$

The average subcarrier power is

$$\overline{\dot{\theta}(t)^2} = 2\pi^2 f_m^{\ 2} \beta^2 \tag{11}$$

The noise spectral density at the discriminator output is:

$$S_{\dot{\theta}}(f) = 4\pi^2 \frac{N_o}{A^2} f^2, \quad -\frac{W_{i-f}}{2} < f < \frac{W_{i-f}}{2} \tag{12}$$

The noise within the subcarrier channel bandwidth is:

$$N_{sc} = \frac{8\pi^2 N_o}{A^2} \int_{f_1''}^{f_2''} f^2 \, df \tag{13}$$

$f_1''$ is the lower and $f_2''$ the upper edge of the subcarrier channel passband.

$$N_{sc} = \frac{8\pi^2 N_o}{3A^2} \left[ (f_2'')^3 - (f_1'')^3 \right] \tag{14}$$

Equation (14) can be simplified by introducing the subcarrier channel bandwidth, $W_{sc}$, and the fractional bandwidth $\eta = W_{sc}/f_m$. We will assume that $f_m$ is the center frequency of the subcarrier channel. By use of the fractional bandwidth, $\eta$, Eq. (14) can be simplified and becomes:

$$N_{sc} = \frac{8\pi^2 N_o W_{sc} f_m^2}{A^2} \left[ 1 + \frac{\eta^2}{12} \right] \qquad (15)$$

The subcarrier channel signal-to-noise ratio when a discriminator is used as the baseband demodulator for a phase modulated signal becomes:

$$(S/N)_{sc} = \frac{\beta^2 A^2}{4 W_{sc} N_o \left( 1 + \frac{\eta^2}{12} \right)} \qquad (16)$$

In most updata systems the fractional bandwidth of the subcarrier channel is small. When this is the case the last term in Eq. (15) can be dropped and Eq. (16) takes on the form:

$$(S/N)_{sc} = \beta^2 (C/N)_{i-f} \frac{W_{i-f}}{2W_{sc}} \qquad (17)$$

As we see Eq. (17) is the same expression for the subcarrier channel signal-to-noise ratio that we would have obtained had we used a true phase demodulator instead of a discriminator as the baseband demodulator in the updata receiver.

The loss in subcarrier channel signal-to-noise ratio caused by use of a discriminator to demodulate a PM updata signal is

$$\epsilon = 10 \log_{10} \left[ 1 + \frac{\eta^2}{12} \right] \text{ db} \qquad (18)$$

In Eq. (18), $\eta$ is the relative subcarrier channel bandwidth and $\epsilon$ the performance loss in db.

As a typical example we can take a subcarrier channel that has a center frequency of 10 kc and a bandwidth of 2 kc. This channel has therefore a relative bandwidth $\eta = 2$. The performance loss caused by use of a discriminator as the baseband demodulator can be evaluated from Eq. (18) and is found to be less than 0.1 db.

### 2.1.1.3 AM with Carrier

We will now investigate the performance of an AM command system. The AM modulated command signal is described by Eq. (19).

$$s(t) = \mathrm{Re}\left[ A(1 + a \sin 2\pi f_m t)\, e^{j2\pi f_c t} \right] \tag{19}$$

The noise can be expressed in the form of Eq. (2) which is repeated here for convenience.

$$n(t) = \mathrm{Re}\left[ \sum_{f_1}^{f_2} \sqrt{2N_o \Delta f}\; e^{j\theta_n} e^{j2\pi f_n t} \right]$$

$f_1$ is the lower and $f_2$ the upper edge of the i-f passband.

The resulting subcarrier channel signal-to-noise ratio can be evaluated from Eqs. (2) and (19). For a receiver employing an envelope demodulator operating with an i-f signal-to-noise ratio exceeding 6 db the subcarrier channel signal-to-noise ratio becomes:

$$(S/N)_{sc} = \frac{A^2 a^2}{4N_o W_{sc}} \tag{20}$$

Equation (20) can be put in the more conventional form:

$$(S/N)_{sc} = (C/N)_{i\text{-}f} \frac{a^2 W_{i\text{-}f}}{2W_{sc}} \tag{21}$$

### 2.1.1.4 Comparison of PM, AM and DSBSC

In this section we will compare the performance of PM, AM and double-sideband suppressed carrier (DSBSC) systems. In an AM system the carrier power, C, is not affected by modulation and can therefore be evaluated in the absence of modulation. Since the transmitter must be able to handle the total transmitted power it is more meaningful to use the average transmitted power, $P_a$ as a figure for comparison purposes. $P_a$ is the maximum average power the transmitter can handle. The relation of carrier power to $P_a$ is:

$$\text{PM}: \qquad C = P_a \tag{22}$$

$$100\%\ \text{AM}: \qquad C = 0.67\, P_a \tag{23}$$

In DSBSC transmission there is no carrier at all. $P_a$ is the total sideband power. The DSBSC signal can be expressed as shown in Eq. (24).

$$s(t) = \text{Re}\left[ A \sin 2\pi f_m t \; e^{j2\pi f_c t} \right] \tag{24}$$

The baseband noise in a DSBSC system is the same as in an AM system. The subchannel signal-to-noise ratio when a DSBSC system is used is:

$$\text{DSBSC}: \qquad (S/N)_{sc} = \left(\frac{P_a}{N}\right)_{i-f} \frac{W_{i-f}}{W_{sc}} \tag{25}$$

Using the same basis for comparison we obtain

$$\text{PM}: \qquad (S/N)_{sc} = \frac{\beta^2}{2} \left(\frac{P_a}{N}\right)_{i-f} \frac{W_{i-f}}{W_{sc}} \tag{26}$$

and

$$100\%\ \text{AM}: \qquad (S/N)_{sc} = 0.34 \left(\frac{P_a}{N}\right)_{i-f} \frac{W_{i-f}}{W_{sc}} \tag{27}$$

It has been assumed that the PM and AM systems operate with an i-f signal-to-noise ratio exceeding 10 or 6 db respectively. Eqs. (25), (26) and (27) are plotted in Fig. 1. The i-f bandwidth is assumed to be equal for all three systems.

## 2.2 System Performance in Interference

### 2.2.1 Interference in an AM Command System

An interfering signal can enter the i-f amplifier of the updata receiver in several different ways. The most obvious is for the interfering signal to fall within the i-f bandwidth of the carrier frequency of the desired signal. Another way is for the interference to be separated from the desired signal by twice the i-f frequency on the same side of the signal carrier as the local oscillator. A third way for the interfering signal to enter the i-f amplifier is for the interference to overload the front end of the receiver to such an extent as to cause crossmodulation with the desired signal.

The various detrimental effects of interference will be discussed in this section. As a preliminary to this analysis some of the effects of interference in a command system employing subcarrier channels will be considered.

It is a well known fact that nonlinear demodulators such as the envelope demodulator or the discriminator will suppress a weaker signal relative to a stronger one. If the interfering signal is weaker than the desired one, the interference will be suppressed. If, on the other hand, the interfering signal is stronger than the desired one, the desired signal will be suppressed and in many cases completely obliterated.

The suppression effect caused by the nonlinear action of the demodulator is a suppression relative to the interference and to the additive noise present.

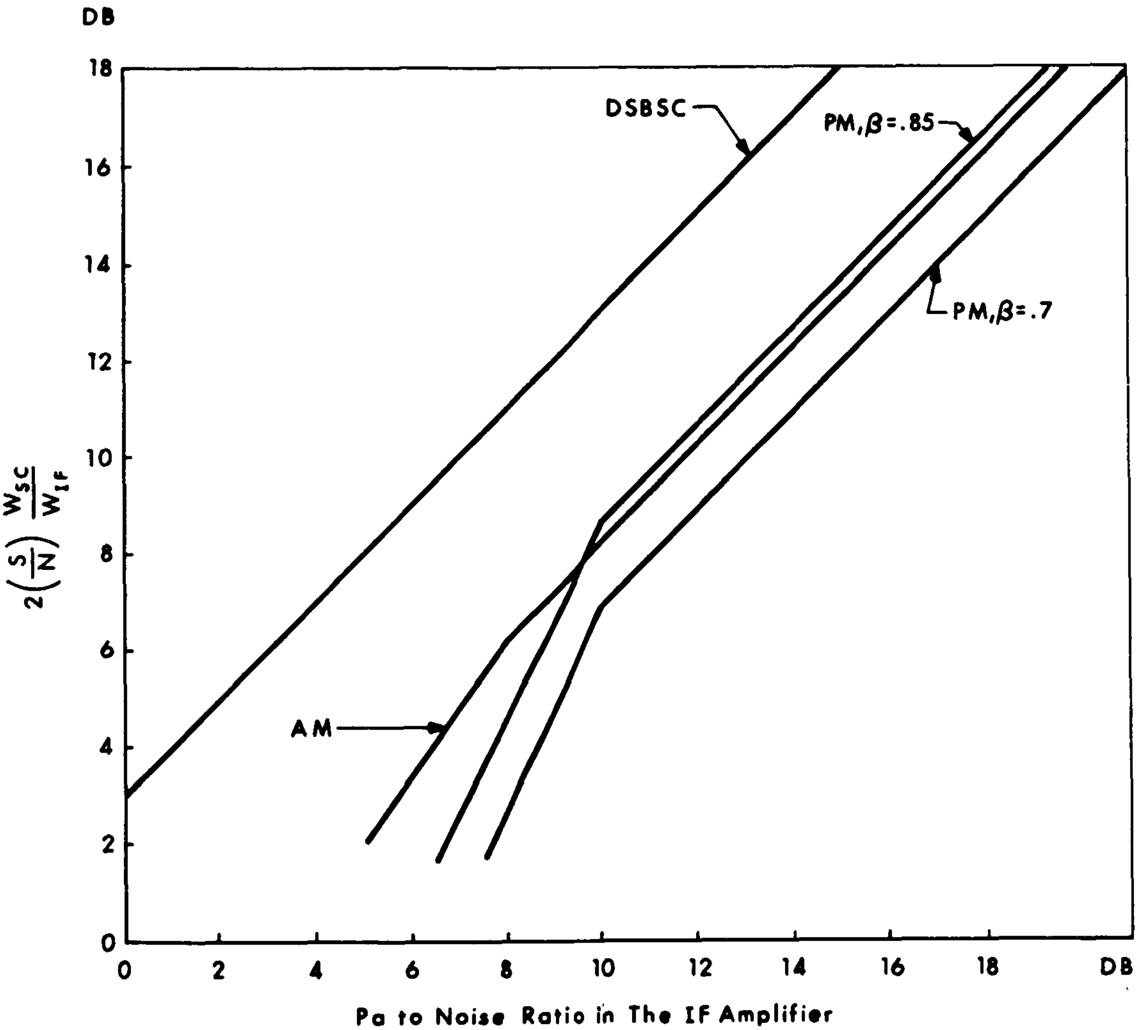


Fig. 1 Normalized subcarrier channel signal-to-noise ratio for three different modulation techniques. The comparison is based on uniform front-end noise spectral density, equal total transmitted power and equal i-f bandwidth.

Another effect that takes place in the receiver is the signal attenuation effect. This effect comes about because of the limiter or the AGC circuits. The purpose of these circuits is to keep the output level of the i-f amplifier reasonably constant subject to variations in input level. When interference is present the result is that the sum of signal and interference is kept fixed.

As long as the interference is much weaker than the signal, the sum of signal plus interference is approximately equal to the signal level and the attenuation is small. If, on the other hand, the interference is much stronger than the signal, the level of the i-f amplifier output is fixed mainly by the interference. Since the output level is fixed by the limiter or AGC circuits the result will be an attenuation of the signal at the input to the demodulator, relative to the signal level in the absence of interference.

The attenuation effect is different from the suppression effect in that it does not in itself cause a deterioration of the signal-to-noise or the signal-to-interference ratio. It will therefore cause difficulty only in cases where the dynamic range of the subsequent circuits is insufficient. This is the case where a fixed lower threshold is used in the subcarrier demodulator, as for instance in the noise immune filter. It will also be the case if there is too much noise in the subsequent stages so that the subcarrier channel signal-to-noise ratio deteriorates too far.

Besides the signal suppression and attenuation effects there are other causes of signal deterioration. First there is the possibility that the difference frequency between the signal and interfering carrier falls within a subcarrier channel. If a nonlinear demodulator is used there are, in addition to this component, several distortion terms. Strong interference in an angle-modulated system will lead to complete obliteration of the desired signal. The same is usually true when an AM system employing envelope demodulation is subjected to strongly modulated interference. Only when the AM system is subjected to CW interference, or when a synchronous demodulator is used, does the signal baseband survive unharmed even through it does contain interference and distortion products.

When an AM system is subjected to CW interference it may well be that because of the relatively narrow subcarrier channels even though interference and distortion products do appear in the demodulated baseband, none of these disturbing terms fall within the subcarrier channels. If this is the case, then the only effect of the interference is a reduction in signal-to-noise ratio because of signal suppression and an attenuation of the subcarrier level.

If, therefore, the subcarrier demodulator has sufficient dynamic range, or a suitable AGC circuit that counteracts the signal attenuation effect is used, considerable improvement in system performance can be achieved when the system is operating in a CW interference environment.

2.2.1.1 CW Interference in an AM System

We will here analyze the operation of an AM receiver, employing envelope demodulation, when the input is signal plus CW interference.

The desired signal, s(t), can be written as:

$$s(t) = A \operatorname{Re}\left\{[1 + am(t)]\, e^{j\omega_c t}\right\} \tag{28}$$

where

$$|am(t)| \leq 1$$

The CW interference can be expressed as:

$$i(t) = I \operatorname{Re}\left[ e^{j\omega_i t} \right] \tag{29}$$

where I is the amplitude of the interfering carrier. The composite output from the i-f amplifier is therefore:

$$u(t) = \mathrm{Re}[\{ I e^{j(\omega_i - \omega_c)t} + A[1 + am(t)] \} e^{j\omega_c t} ] \tag{30}$$

Equation (30) is valid as long as the i-f amplifier is operating within its linear range. When a conventional envelope demodulator is used the resulting baseband signal, v(t), is

$$v(t) = \frac{1}{2} \sqrt{ \{ I e^{j\omega_i' t} + A[1+am(t)]\} \{ I e^{-j\omega_i' t} + A[1 + am(t)]\} } \tag{31}$$

where

$$\omega_i' = \omega_i - \omega_c$$

Equation (31) can be simplified by introducing the signal-to-interference raio, $\rho = A/I$.

$$v(t) = \frac{1}{2} I \sqrt{ 1 + \rho^2 [1 + am(t)]^2 + 2\rho [1 + am(t)] \cos \omega_i' t } \tag{32}$$

Equation (32) gives evidence of three distinctly different regions of operation, $\rho \ll 1$, $\rho \approx 1$, and $\rho \gg 1$.

Weak Interference Region $\rho \gg 1$

In this region v(t) can be approximated by Eq. (33)

$$v(t) \approx \frac{1}{2} A [1 + am(t)] + \frac{A}{2\rho} \cos \omega_i' t \tag{33}$$

We see from Eq. (33) that the desired signal plus a small interference term are present in the output from the envelope demodulator.

The purpose of the AGC loop is to keep the average i-f level reasonably constant at the output of the i-f amplifier. This in turn means that the DC component of the demodulator output is reasonably constant. If the receiver has a perfect AGC loop, the carrier amplitude, A, will therefore be normalized at the demodulator output. Equation(33) will therefore take on the form:

$$v'(t) = 1 + am(t) + \frac{1}{\rho} \cos \omega_i' t \tag{34}$$

For a particular demodulator, v(t) may be multiplied by a constant which is independent of the AGC action. For convenience this constant has been set equal to two in obtaining Eq. (34).

## Medium Interference Region $0.1 < \rho < 10$

When the signal-to-interference ratio is close to unity any simple approximations for the square root in Eq. (32) fails. The physical significance of this is that the signal output from the demodulator may be seriously distorted in this region. There will still be some of the modulation remaining in the demodulated output but harmonic distortion will be high.

## Strong Interference Region $\rho \ll 1$

When $\rho \ll 1$ the output from the demodulator can be approximated by Eq. (35)

$$v'(t) = 1 + \frac{\rho^2}{2} \left[ 1 + 2\,am(t) + a^2 m^2(t) \right] + \rho \left[ 1 + am(t) \right] \cos \omega_i' t \tag{35}$$

From Eq. (35) we see that the modulation, m(t), has been attenuated by $\rho^2$ relative to its level in the absence of interference. Eq. (35) shows that the strongest interference term appears as DC component. Since this component is removed by the subcarrier channel filters, it is of no importance. The next strongest interference-caused component is $\rho[1 + am(t)] \cos \omega_i' t$. If this component is considered to be the actual interference we see that the baseband signal-to-interference ratio equals the i-f signal-to-interference ratio. It will be shown later that this is only the case when the interfering signal is a CW signal or when it is angle modulated.

Returning to the signal attenuation effect we see that if none of the interference or distortion terms fall within the subcarrier channels the main effect of the interference is to attenuate the desired signal amplitude by $\rho^2$ or the signal power by $\rho^4$. For a signal-to-interference ratio of -20 db the subcarrier power will be attenuated by 40 db relative to its standard level without interference. This will probably result in an unusable baseband signal even if none of the distortion or interference products should fall within the passband of the subcarrier filters.

2.2.1.2 AM Interference in an AM System

We will consider the interference situation that develops when an AM modulated signal interferes with the AM modulated command transmission.

Let the desired signal be described by Eq. (36)

$$s(t) = A \operatorname{Re}\{[1 + am(t)]\, e^{j\omega_c t}\} \tag{36}$$

Let the AM modulated interfering signal be described by Eq. (37)

$$i(t) = I \operatorname{Re}\{[1 + bn(t)]\, e^{j\omega_i t}\} \tag{37}$$

I is the amplitude, n(t) the modulation, and $f_i = \omega_i/2\pi$ the frequency of the interfering signal.

The resulting i-f signal can be described as:

$$u(t) = \operatorname{Re}[\{A[1 + am(t)] + I[1 + bn(t)]\, e^{j\omega_i' t}\}\, e^{j\omega_c t}] \tag{38}$$

where

$$\omega_i' = \omega_i - \omega_c$$

The output from the envelope demodulator is:

$$v(t) = \frac{1}{2} I \sqrt{\rho^2[1+am(t)]^2 + [1+bn(t)]^2 + 2\rho[1+am(t)][1+bn(t)] \cos \omega_i' t}$$

where $\rho = A/I$ is the signal-to-interference voltage ratio.

The square root in Eq. (39) is amenable to approximations for small and large values of $\rho$.

## Weak Interference Region $\rho >> 1$

When the interference is weak compared to the desired signal, Eq. (40) is a good approximation for Eq. (39).

$$v(t) = \frac{1}{2} I_\rho \left\{ [1 + am(t)] + \frac{1}{\rho} [1 + bn(t)] \cos \omega_i' t \right\} \tag{40}$$

In a conventional AM receiver the action of the AGC circuit will be to normalize the DC component in Eq. (40). When this is taken account of Eq. (40) can be rewritten as:

$$v'(t) = 1 + am(t) + \frac{1}{\rho}[1 + bn(t)] \cos \omega_i' t \tag{41}$$

By comparing Eq. (33) and Eq. (41) we see that in the case of weak interference, the only difference between the effects of AM modulated and CW interference is that the former results in AM modulation of the interference subcarrier. AM interference, even if it is weak, is therefore potentially more disturbing than CW interference.

## Strong Interference Region $\rho << 1$

When the interference is much stronger than the desired signal, Eq. (39) can be approximated by Eq. (42).

$$v'(t) = [1 + bn(t)] + \frac{\rho^2 [1 + am(t)]^2}{2[1 + bn(t)]} + \rho[1 + am(t)] \cos \omega_i' t \tag{42}$$

When we compare Eq. (35) and Eq. (42) we see that strong AM modulated interference can completely obliterate baseband signal components while in the case of CW interference it would only attenuate it.

To study this effect more closely we will expand the signal component in Eq. (42). We will assume that a = b = 1 and $m(t) = \cos\omega_1 t$, $n(t) = \cos\omega_n t$. In this case the signal component of Eq. (42) becomes:

$$\rho^2 \frac{\cos\omega_1 t}{1+\cos\omega_n t} = \rho^2 \frac{\cos\omega_1 t}{2\cos^2\frac{\omega_n}{2}t} = \frac{\rho^2}{2}\cos\omega_1 t\left(1+\tan^2\frac{\omega_n}{2}t\right) \tag{43}$$

As Eq. (43) shows with the type of modulation used in this example, some of the intended baseband signal does come through. It will, however, be severely disturbed by the presence of the $\tan^2$ function.

#### 2.2.1.3 Angle Modulated Interference in an AM System

Angle modulation is frequently used for ranging and command signals in spacecraft systems. Since the system under investigation is an AM system, the problem of interference into this system by angle modulated systems operating nearby is of great importance.

The amplitude modulated signal can be expressed as:

$$s(t) = A\,\mathrm{Re}\left\{[1+am(t)]\,e^{j\omega_c t}\right\} \tag{44}$$

The angle modulated interference can be expressed as :

$$i(t) = I\,\mathrm{Re}\left\{e^{j[\omega_i t+\beta f(t)]}\right\} \tag{45}$$

I is the amplitude, $\beta$ the modulation index and f(t) the modulation of the interfering signal. We will assume that the interference is such that the modulation does not cause the interfering signal to swing outside the edge of the i-f passband. If the interfering signal operates on the edge of the i-f passband I will be a function of f(t).

The composite signal at the output of the i-f amplifier is:

$$u(t) = \mathrm{Re}\left[\left\{A[1+am(t)] + I\,e^{j[\omega_i' t+\beta f(t)]}\right\}e^{j\omega_c t}\right] \tag{46}$$

where

$$\omega_i' = \omega_i - \omega_c$$

Equation (46) is valid as long as the i-f amplifier is operating within its linear range.

The resulting baseband signal at the output of the envelope demodulator is:

$$v(t) = \frac{1}{2} I \sqrt{1 + \rho^2 [1 + am(t)]^2 + 2\rho [1 + am(t)] \cos [\omega_i' t + \beta f(t)} \quad (47)$$

where $\rho = A/I$ is the signal-to-interference ratio. When we compare Eq. (47) with Eq. (32) we notice that interference from a narrowband angle modulated carrier or from a CW signal has very similar effects on an AM system. The main difference is that the interference term is now angle modulated with the difference frequency, $\omega_i'$, as a carrier. This will give rise to interference sidebands centered at $\omega_i'$. There will therefore be more interference terms present in the demodulated baseband spectrum. Except for this the results obtained for CW interference apply also when the interfering signal is narrowband angle modulated.

If the frequency deviation of the interfering signal causes it to operate on the edge of the passband of the i-f amplifier, the angle modulation will be converted into amplitude modulation. The resulting amplitude modulated interference will have an effect similar to AM interference into an AM system. This leads, as we have seen, to much more serious interference problems.

We can therefore conclude that interference from narrowband angle modulated signals will have an effect very similar to CW interference as long as the interfering signal does not fall on or very near the edge of the i-f passband.

Interference from wider band angle modulated signals, or from narrowband signals falling close to the edge of the i-f passband will have an effect similar to that of AM interference because of the FM to AM conversion in the i-f amplifier.

To conclude this section we will analyze an example of angle modulated interference. Let the system and interference parameters be as follows:

$$\rho = 0.1 \therefore -20 \text{ db}, \quad m(t) = \cos 2\pi \cdot 14 \cdot 10^3 t$$

$$f_i' = 6 \text{ kc}, \quad a = 1$$

$$f(t) = \cos 2\pi \cdot 2 \cdot 10^3 t$$

$$\beta = 0.5$$

The demodulated baseband is given by Eq. (48).

$$v'(t) = \frac{1}{2}\left\{1 + 0.01\left[0.5 + m(t) + \frac{m^2(t)}{2}\right] + 0.1 \cos\left[2\pi \cdot 6 \cdot 10^3 t + 0.5 f(t)\right]\right\} \tag{48}$$

After approximations and rearrangement of terms we obtain:

$$\begin{aligned} v'(t) = \frac{1}{2}\{&1 + 0.01 \cos 2\pi \cdot 14 \cdot 10^3 t + 0.0025 \cos 2\pi \cdot 28 \cdot 10^3 t \\ &+ 0.1 \cos 2\pi \cdot 6 \cdot 10^3 t - 0.1[0.875 \cos 2\pi \cdot 2 \cdot 10^3 t \\ &- 0.042 \cos 2\pi \cdot 6 \cdot 10^3 t] \sin 2\pi \cdot 6 \cdot 10^3 t\} \end{aligned} \tag{49}$$

When Eq. (49) is written in terms of its individual frequency components, we obtain:

$$\begin{aligned} v'(t) = \frac{1}{2}\{&1 + 0.01 \cos 2\pi \cdot 14 \cdot 10^3 t + 0.0025 \cos 2\pi \cdot 28 \cdot 10^3 t \\ &+ 0.1 \cos 2\pi \cdot 6 \cdot 10^3 t + 0.044 \sin 2\pi \cdot 4 \cdot 10^3 t \\ &- 0.044 \sin 2\pi \cdot 8 \cdot 10^3 t + 0.0021 \sin 2\pi \cdot 12 \cdot 10^3 t\} \end{aligned} \tag{50}$$

The individual frequency components in the resulting baseband are clearly in evidence in Eq. (50). A sketch of the resulting power spectrum is given in Fig. 2.

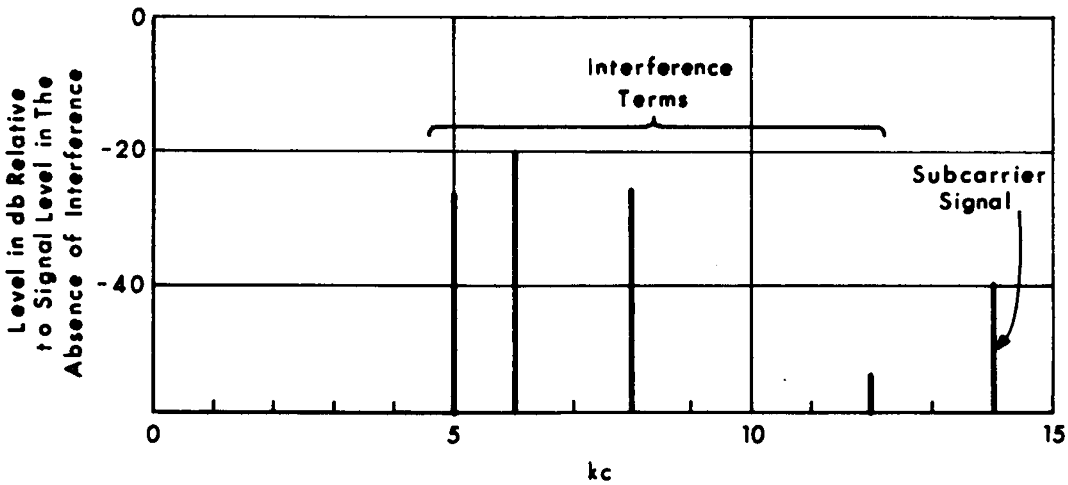


Fig. 2 Demodulated baseband resulting from a particular case of strong narrowband angle modulated interference in an AM system employing envelope demodulation.

## 2.2.2 Effect of CW Interference on a System Employing Angle Modulation

In this section we will study the performance of a command system employing angle modulation when it is subjected to CW interference.

The desired angle modulated signal can be described by Eq. (51).

$$s(t) = A \,\mathrm{Re}[e^{j\psi(t)} \; e^{j\omega_c t}] \tag{51}$$

where $\psi(t)$ is the modulating function.

The interfering carrier can be described by Eq. (52).

$$i(t) = I \,\mathrm{Re}[e^{j\theta} \; e^{j\omega_i t}] \tag{52}$$

I is the amplitude, $\theta(t)$ the modulation and $\omega_i$ the angular frequency of the interfering signal.

After limiting, the output from the i-f amplifier will be:

$$u(t) = \mathrm{Re}[\; e^{j\theta(t)} \quad e^{j\omega_c t} \;] \tag{53}$$

$\phi(t)$, the output from the phase demodulator is:

$$\phi(t) = \tan^{-1} \frac{\rho \sin\psi(t) + \sin[\omega_i' t + \theta]}{\rho \cos\psi(t) + \cos[\omega_i' t + \theta]} \tag{54}$$

where $\rho$ is the signal-to-interference voltage ratio and $\omega_i' = \omega_i - \omega_c$.

In order to simplify Eq. (54) we need to develop two trigonometric relationships.

Given the expression

$$\tan\gamma = \frac{\rho \sin\alpha + \sin\beta}{\rho\cos\alpha + \cos\beta} \tag{55}$$

We want to find two simple approximations for Eq. (55) valid for small and large values of $\rho$.

In the problem at hand the absolute values of the angles $\alpha, \beta$ and $\gamma$ are unimportant. Only the relative angles need be considered. We can therefore use any one of the three angles as reference for the two others. When we choose $\alpha$ as the reference angle we can rewrite Eq. (55) in the form:

$$\tan(\gamma - \alpha) = \frac{\rho \sin(\alpha - \alpha) + \sin(\beta - \alpha)}{\rho \cos(\alpha - \alpha) + \cos(\beta - \alpha)} \tag{56}$$

When $\rho \gg 1$, Eq. (56) is closely approximated by Eq. (57)

$$\tan(\gamma - \alpha) = \frac{1}{\rho} \sin(\beta - \alpha) \tag{57}$$

or

$$\gamma - \alpha = \frac{1}{\rho} \sin(\beta - \alpha) \tag{58}$$

When $\rho \ll 1$, it is more convenient to choose $\beta$ as the reference angle. With $\beta$ as the reference angle, Eq. (55) becomes:

$$\tan(\gamma - \beta) = \frac{\rho \sin(\alpha - \beta) + \sin(\beta - \beta)}{\rho \cos(\alpha - \beta) + \cos(\beta - \beta)} \tag{59}$$

since $\rho \ll 1$, Eq. (59) can be closely approximated by Eq. (60)

$$\tan(\gamma - \beta) = \rho \sin(\alpha - \beta) \tag{60}$$

or

$$\gamma - \beta = \rho \sin(\alpha - \beta) \tag{61}$$

Eq. (54) can now be simplified by use of the approximations (58) and (61) for appropriate values of the signal-to-interference ratio, $\rho$.

Suitable approximations for Eq. (54) can now be found for small and large values of $\rho$.

Weak Interference Region $\rho >> 1$

When the signal is much stronger than the interfering carrier the demodulated baseband voltage, $\phi(t)$, given by Eq. (54) is closely approximated by:

$$\phi(t) = \psi(t) + \frac{1}{\rho} \sin\left[\omega_i' t + \theta - \psi(t)\right] \tag{62}$$

From Eq. (62) we see that under conditions of weak CW interference, the desired modulating waveform is present in the demodulated baseband. It is disturbed by an angle modulated signal centered on the frequency difference between the signal and interfering carriers and modulated by the signal modulation. By comparing Eq. (34) and (62) we see that the strength of the interfering term is the same in an angle modulated system or an AM system employing envelope demodulation, as long as the interference is weak.

Strong Interference Region $\rho << 1$

When the interference is much stronger than the signal Eq. (54) can be approximated as shown in Eq. (63).

$$\phi(t) = \omega_i' t + \theta + \rho \sin\left[\psi(t) - \omega_i' t - \theta\right] \tag{63}$$

From Eq. (63) we see that when the interfering CW signal is much stronger than the desired carrier, the desired modulating waveform is completely removed from the demodulated baseband. The desired modulating waveform does appear as angle modulation on the interference subcarrier but there is no way in which it can be extracted by the baseband subcarrier filters.

## 2.2.3 Error Rates in FDM-FM Digital Data Transmission

### 2.2.3.1 Introduction

In this section we consider the problem of the evaluation of the error rate of binary information modulated upon a subcarrier in an FM system. The FM signal is assumed to be undistorted by the channel and received in additive gaussian noise flat over the frequency band occupied by the signal.

When the SNR (signal-to-noise ratio) at the frequency discriminator input is greater than around 1 or 2 db the baseband output noise has been shown by Rice[1] to be closely representable as the sum of an impulsive noise and a gaussian noise. The impulses occur essentially randomly (i.e., they have approximately Poisson distributed occurrence times) and their areas are approximately $\pm 1$ cps. Rice[1] gives expressions for the "instantaneous frequency modulation" and he also gives averaged impulse rate expressions for sinusoidal modulation and gaussian modulation. The gaussian noise component of the baseband output is assumed to be equal to the conventional noise output at high SNR which is proportional to the derivative of the component in quadrature to the signal vector.

When the input SNR drops below 1 db the above simple picture no longer obtains. For sufficiently small input SNR's, say less than -2 db, the output noise will be approximately that which would have existed in the absence of input signal. Blachman[2] has derived an integral that yields the noise spectral density at zero frequency assuming an unmodulated carrier. From the above results some useful information may be obtained about the output noise level for all input SNR's.

As has been proven by Rice[3] and Stumpers[4] the signal component of the ideal discriminator output is undistorted but is reduced in amplitude by a factor $(1 - e^{-\rho})$ where $\rho$ is the input SNR. Thus it is possible to determine output SNR's for a large range of input SNR's.

Strictly speaking the output SNR in a given bandwidth will be sufficient to determine error rate only if the noise is gaussian. However, if the bandwidth of the filter in the digital demodulator is sufficiently small so that many independent input noise samples are averaged, the noise at the filter output will be nearly gaussian. Thus for input SNR's greater than 1 db it is necessary that the average number of impulses per binary signaling element be much bigger than 1 if the detected noise due to the impulses is to be nearly gaussian. For input SNR's less than 1 db and detection bandwidths much smaller than the output noise bandwidth it appears reasonable to assume that the detected noise is approximately gaussian. Except for these special cases it is necessary to calculate error probabilities taking into account the non-gaussian nature of the output noise.

2.2.3.2 Summary of Some Known Results

In this section we summarize some known results on the baseband signal and noise output of an FM discriminator.

For $\rho$ > 1 db the output noise (in cps) is expressed as

$$h(t) = i(t) + y(t) \tag{64}$$

where i(t) is a Poisson distributed impulsive noise

$$i(t) = \sum \delta(t - t_k) - \sum \delta(t - t_\ell) \tag{65}$$

and y(t) is a gaussian noise with spectral density

$$P_y(f) = \frac{f^2}{2W\rho} \tag{66}$$

where W is the i-f bandwidth (cps) and $\rho$ is the input signal-to-noise (power ratio).

The number of impulses/sec is dependent upon the input modulation and thus is time variable. According to Rice[1],

$$N_+(t) = \frac{r}{2}\{\sqrt{1+\mu^2}\ \text{erfc}\sqrt{\rho}\sqrt{1+\mu^2} - \mu e^{-\rho}\,\text{erfc}\ \mu\sqrt{\rho}\} \tag{67}$$

$$N_-(t) = \frac{r}{2}\{\sqrt{1+\mu^2}\ \text{erfc}\sqrt{\rho}\sqrt{1+\mu^2} + \mu e^{-\rho}\,\text{erfc}[-\mu\sqrt{\rho}]\} \tag{68}$$

where $N_\pm(t)$ are the (instantaneous) number of positive and negative impulses/sec, r is the rms bandwidth of the r-f noise (assumed flat)

$$r = \frac{W}{\sqrt{12}} \tag{69}$$

$\mu$ is the input modulation, $\nu(t)$, (in cps) normalized to r, i.e.,

$$\mu = \frac{\nu(t)}{r} \tag{70}$$

and erfc( ) is the complementary error function

$$\text{erfc}\ x = \frac{2}{\sqrt{\pi}}\int_x^\infty e^{-t^2}\,dt = \frac{e^{-x^2}}{x\sqrt{\pi}}\left(1 - \frac{1}{2x^2} + \ldots\right) \tag{71}$$

For $\rho$ greater than 1 or 2 db the first terms in Eqs. (67) and (68) may be approximated by the first term in the asymptotic expansion (71), yielding

$$N_\pm(t) \approx \frac{r}{2}e^{-\rho}\left\{\frac{e^{-\rho\mu^2}}{\sqrt{\pi\rho}} \mp \mu\,\text{erfc}\,[\pm\mu\sqrt{\rho}\,]\right\} \tag{72}$$

The number of impulses/unit time may be averaged over the input modulation. In the case of a normally distributed modulation process (72) can be averaged to yield a simple closed-form expression for the average impulse rate. Thus one may show that

$$N_+(t) = N_-(t) = \frac{e^{-\rho}}{2}\sqrt{\frac{r^2 + 2\sigma^2\rho}{\pi\rho}} \tag{73}$$

where $\sigma^2$ is the mean squared value of the frequency deviation of the base-band signal (in cps)

$$\sigma^2 = \overline{[\nu(t)]^2} \tag{74}$$

As Stumpers[4] and Rice[3] have shown, the signal component of the discriminator output is given by

$$(1 - e^{-\rho})\, \nu(t) \tag{75}$$

This expression is valid for all input SNR.

2.2.3.3 Error Rate Calculation

In this section we shall consider the binary error rate for information modulated upon a subcarrier. It will be assumed that the average number of impulses per signaling element is sufficiently large so that the detected noise is essentially normally distributed. Also it will be assumed that the data subchannel occupies a bandwidth small enough compared to the width of the baseband so that the noise spectrum due to the y(t) term in (64) varies little over this subchannel.

In the case of matched filter reception in flat noise the error probability can be uniquely related to a signal-to-noise ratio

$$s = \frac{E}{2N} \tag{76}$$

where E is the energy of a signaling element and N is the (two-sided) spectral density. Thus Helstrom[5] shows that for coherent detection and antipodal signals (e. g., PSK with $\pm 180^{\circ}$) the error probability is given by

$$P = \frac{1}{2} \operatorname{erfc} [\sqrt{s}] \tag{77}$$

while for incoherent detection of orthogonal signals (e. g., FSK) the error probability is given by

$$P = \frac{1}{2} \exp [-\frac{s}{2}] \tag{78}$$

As is well known the (two-sided) spectral density of Poisson distributed impulse noise is numerically equal to the number of impulses per unit time. Thus the total power density due to the combined effects of the (positive and negative) impulsive noise i(t) and the continuous noise y(t) is given by

$$N = e^{-\rho} \sqrt{\frac{r^2 + 2\sigma^2 \rho}{\pi \rho}} + \frac{f_d^2}{2W\rho} \tag{79}$$

where $f_d$ is the frequency location of the data subcarrier.

If we assume that the typical signaling element is a sinusoidal burst of duration T and has a modulation index of $\mu$, then

$$E = \frac{1}{2} \mu^2 f_d^2 \, T \left(1 - e^{-\rho}\right)^2 \tag{80}$$

and the detected SNR is given by

$$s = \frac{\mu^2 f_d^2 \left(1 - e^{-\rho}\right)^2}{4e^{-\rho} \sqrt{\frac{r^2 + 2\sigma^2 \rho}{\pi \rho}} + \frac{2f_d^2}{W\rho}} \tag{81}$$

This expression may be inserted in Eq. (77) or (78) for computations of error rate.

## 2.3 An Improved AM Command System

### 2.3.1 An Improved AM Command Receiver

As the results of section 2.2 show, one of the detrimental effects of strong CW interference in a conventional AM data transmission system is a signal attenuation.

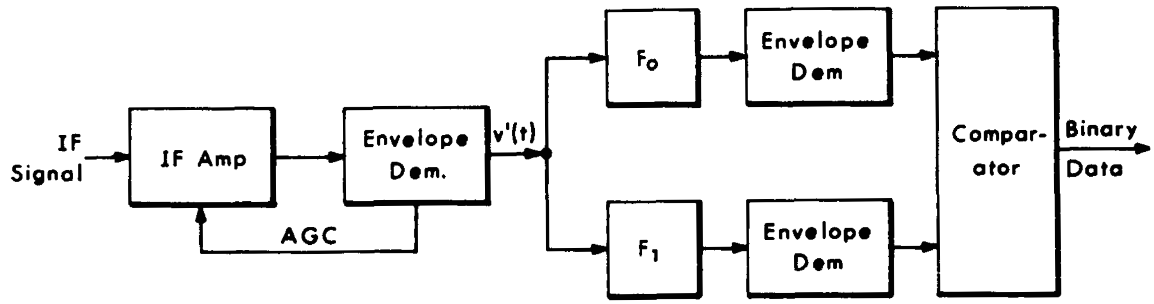


Fig. 3 AM receiver with FSK baseband demodulator.

Figure 3 shows the relevant portion of a conventional AM receiver with FSK baseband demodulation. The operation of this system was analyzed in Sec. 2.2. It was found that when a weak interfering CW signal is present, the input voltage to the baseband filters, v'(t), is given by:

$$v'(t) = 1 + am(t) + \rho \cos \omega_i' t \quad , \quad \rho \gg 1 \tag{82}$$

When the interference is strong, the input voltage to the baseband filters is:

$$v'(t) = 1 + \frac{\rho^2}{2} [1 + 2am(t) + a^2 m^2(t)] + \rho[1 + am(t)] \cos \omega_i' t, \quad \rho \ll 1 \tag{83}$$

$\rho = A/I$ = signal-to-interference voltage ratio

$am(t)$ = desired baseband signal

$\omega_i'$ = frequency difference between signal and interference.

From Eq. (83) we see that strong CW interference has two effects on the demodulated baseband. One is the creation of interference and distortion products within the passband of the baseband filter, the other is a substantial attenuation of the desired signal relative to its level in the absence of interference. If the signal-to-interference ratio is -10 db, we see from Eq. (83) that the baseband signal will be attenuated by 20 db relative to its level in the absence of interference.

There are therefore two ways in which CW interference can disable the AM up-data system. One way is for the interference and distortion products to fall within the passband of the subcarrier FSK filters. The other is for the subcarrier signal in the baseband to be attenuated to such an extent that it will no longer operate the FSK demodulator reliably.

This second failure mode can to a large extent be prevented by use of the modified AM receiver shown in Fig. 4.

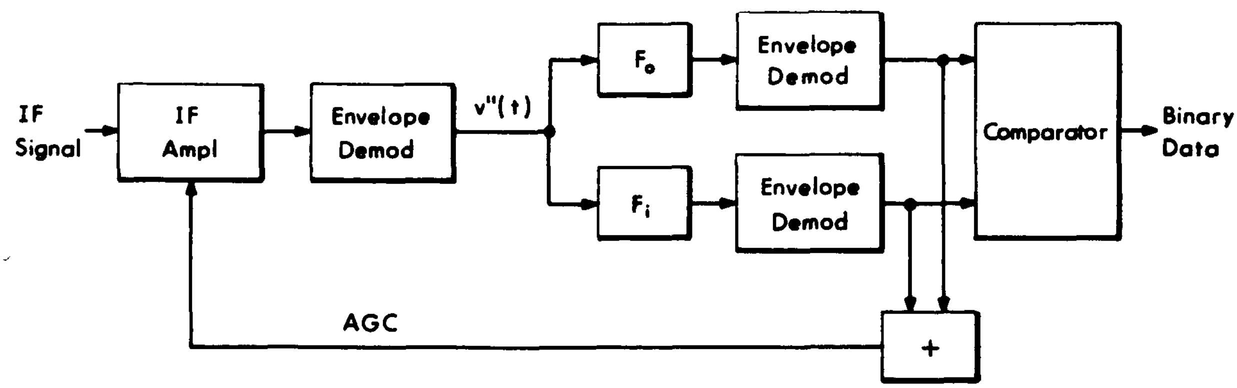


Fig. 4 AM receiver designed to minimize the effects of CW interference.

The performance of the receiver shown in Fig. 4 will, under conditions of severe CW interference, be significantly superior to the performance of the conventional AM receiver shown in Fig. 3. Since the AGC voltage is taken from the output of the subcarrier demodulator the signal attenuation effect is removed. The demodulator output voltage will therefore be:

$$v''(t) = 1 + am(t) + \rho \cos \omega_i' t \quad , \quad \rho >> 1 \tag{84}$$

and

$$v''(t) = \frac{1}{\rho^2} + \frac{1}{2} + am(t) + \frac{a^2 m^2(t)}{2} + \frac{1}{\rho}[1 + am(t)] \cos \omega_i' t \ , \quad \rho << 1 \tag{85}$$

Equations (84) and (85) are valid for the condition when no strong distortion or interference products fall within the passband of the subcarrier filters. It has also been assumed that the linear range of operation of the i-f amplifier is not exceeded by the total i-f signal.

When these conditions are fulfilled we see from Eq. (84) and (85) that the level of the signal component in the baseband is unaffected by CW interference. The signal attenuation effect has therefore been removed.

The proposed improved command receiver where the AGC control voltage is taken from the subcarrier demodulator output may fail under certain circumstances. If the i-f amplifier gets severely overloaded the nonlinearities thereby produced will reduce the AM modulated signal component severely. This will in turn reduce the AGC voltage and cause further overloading of the i-f amplifier. It is quite possible that this sequence of events can lead to blocking of the receiver.

In order to overcome the problem of driving the i-f amplifier into nonlinear operation, a double AGC loop can be employed. A receiver using this technique is shown in Fig. 5.

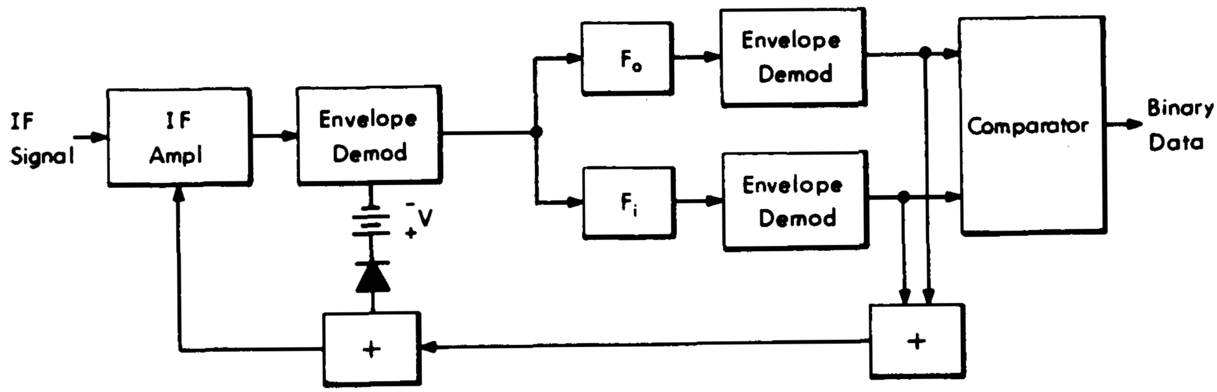


Fig. 5 AM receiver with double AGC loop.

The main AGC action in the receiver shown in Fig. 5 is caused by the output from the subcarrier demodulators.

The second AGC loop has delayed action. As long as the DC output from the baseband envelope demodulator is less than V, the diode is reverse biased and does not conduct. When the DC voltage from the baseband demodulator exceeds V, the diode conducts and a control voltage is applied to the AGC bus. The voltage, V, is selected large enough so that the baseband demodulator does not affect the AGC action under normal operating conditions and at the same time small enough so as to prevent severe overloading of the i-f amplifier. A suitable choice for V appears to a point slightly above the point where nonlinear action sets in.

The receiver shown in Fig. 5 should be capable of near optimum performance in a CW interference environment. This receiver is not subject to the detrimental effects of nonlinear operation of the i-f amplifier.

Since this receiver will perform as a conventional AM receiver after the inner AGC loop has taken over control it will suffer from the signal attenuation effect in this region of operation. It is therefore desirable to use an i-f amplifier capable of handling linearly as large output voltages as possible in order to obtain the maximum freedom from signal attenuation effects.

A graph comparing the performance of a conventional and a modified AM receiver is given in Fig. 6.

## 2.4 Interference Immunity Testing of the Command Receiver

### 2.4.1 Laboratory Test Methods

The two basic requirements placed upon an experimental setup to be used for studies of system degradation due to interference are that the conditions must be realistic and well controlled.

By realistic is meant that the actual operating conditions of the system must be approached as closely as practicable and that all important effects are considered. The requirement that the experimental conditions be well controlled is necessary from the point of view of data evaluation and repeatability of the experiments.

From these considerations it is clear that for the interference tests to be meaningful they should be performed on the complete space vehicle receiver including the r-f stage and the first converter. The inclusion of the receiver front end in this experiment is necessary since the frequency selective and nonlinear circuits found here can significantly effect the overall performance of the receiver. The next point to be considered is the level of the desired signal. Under actual operating conditions the signal level is a function of the distance between the spacecraft and the transmitter. The expected signal level can therefore be expected to vary over a wide range. The OGO command receiver shall operate without

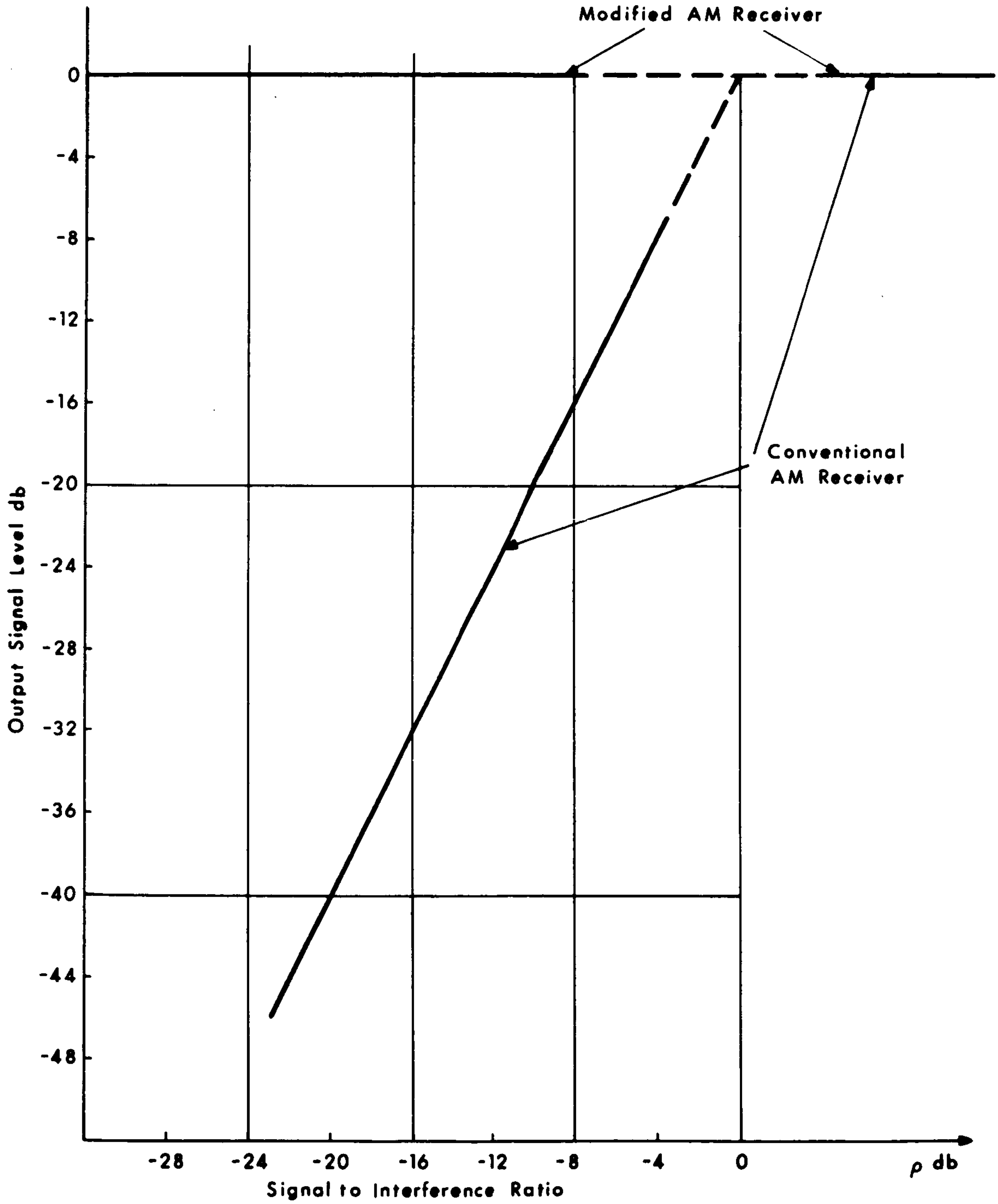


Fig. 6 Signal suppression in a conventional and a modified AM receiver subject to CW interference. No interference or distortion products fall within the passband of the baseband filters.

degradation over a range in input signal level of -110 to -40 db. Since the receiver possesses some essential nonlinearities, it is necessary for the tests to be run over the entire expected range of signal levels.

The interference to be used in this test should be as close an approximation to the actual expected interference as possible. On the other hand the interference should be as simple as possible to generate so that the experimental difficulties encountered do not get out of hand. Relatively simple interference will also in general lead to more generally applicable results.

As a reasonable compromise between these two conflicting requirements we will choose three representative types of interference. They are CW, pulsed carrier and modulated carrier.

The CW tests are the simplest to specify and perform. They will probably also give the most useful information. For this reason they will form the nucleus of the testing procedure.

The CW interference tests require only control over the frequency and power of the interfering carrier.

The pulsed carrier tests require in addition control over pulse length and timing since synchronized interference will in general have a different effect than when the pulses are randomly timed. The modulated carrier tests have the added complexity of modulating function. Both amplitude and angle modulation should be used.

It should be clear that the problem of specifying a meaningful testing procedure for interference immunity is a very complex task. Because of the large number of variables involved the number of readings required will be very large. In order to maximize the amount of useful information obtained with a minimum of effort, careful planning is required.

In the following section a suitable test setup is introduced. This is followed by a discussion of the various tests and the ranges of the parameters involved. The section will conclude with a tabular representation of the various experiments presenting the ranges of the variables involved as well as the data that must be recorded and a suggestion as to convenient graphical representation of the results.

### 2.4.2 Test Setup for Interference Immunity Tests

A setup suitable for complete interference immunity testing of the OGO command receiver is shown in Fig. 7. The pseudo random bit generator (PRBG) is explained in detail in Sec. 4.2.1.

The interference immunity of the receiver should be tested over its entire rated range of signal levels. The rated range for the OGO command receiver is -40 to -110 dbm. This range can be lowered when the signal generator is adjusted to give a level of -40 dbm at the spacecraft receiver input when the attenuator $A_s$ is set to zero. The signal power at the receiver is thus:

$$P_s = -40 - A_s \quad \text{dbm}$$

It is desirable to test the interference immunity of the command receiver over a signal-to-interference ratio range of -40 to +20 db. This is easily accomplished when the interference generator is adjusted to give a level of zero dbm at the input of the spacecraft receiver when $A_i$ is set to zero. The interference level at the receiver input is thus:

$$P_i = -A_i \quad \text{dbm}$$

The signal-to-interference power ratio, $\rho^2$, is now easily evaluated.

$$\rho^2 = -40 - A_s + A_i \quad \text{db}$$

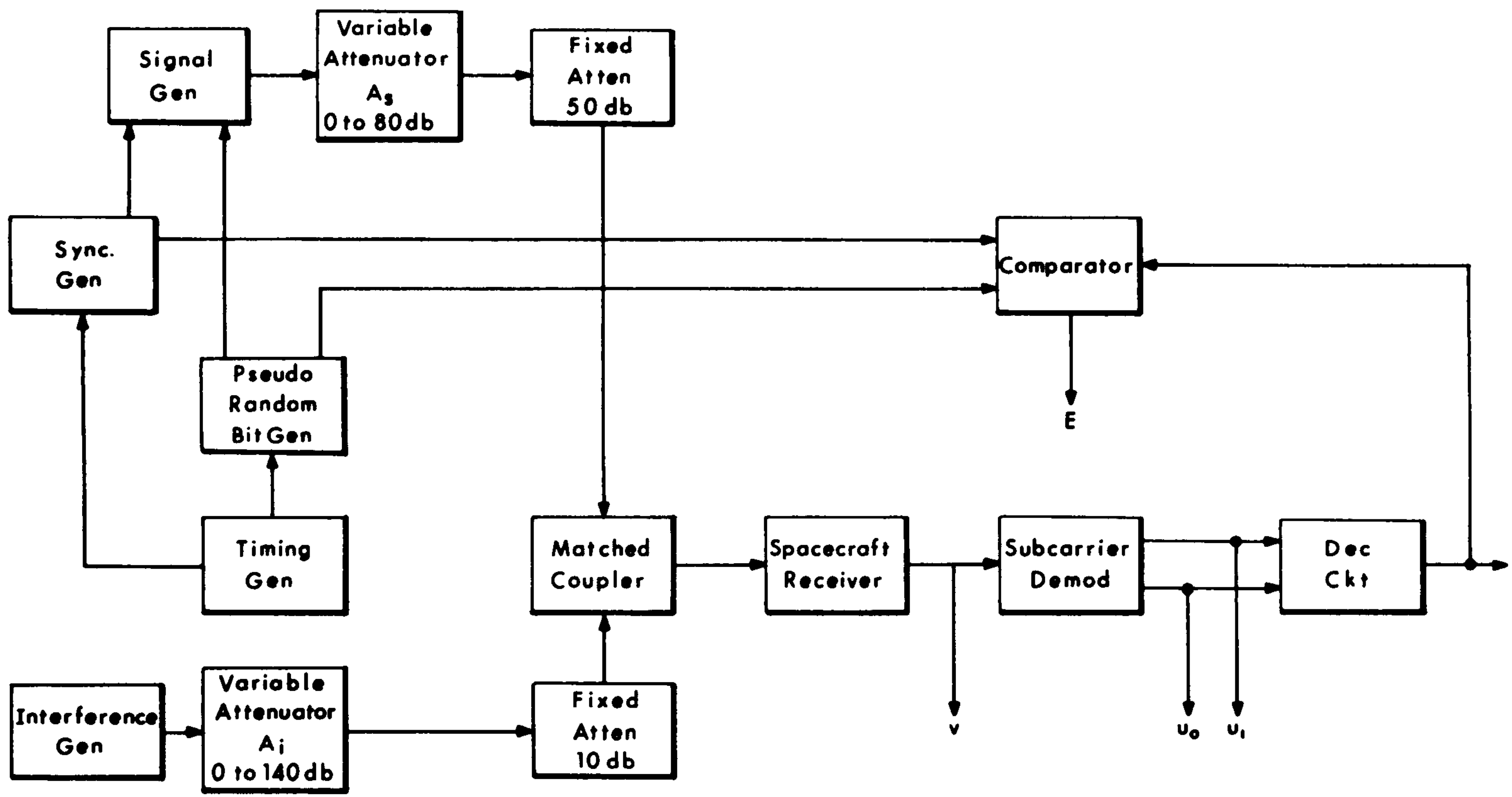


Fig. 7 Test setup for interference immunity tests.

The signal-to-interference ratio range of most interest is therefore $0 < A_i - A_s < 60$ db.

The frequency of the interfering carrier must be varied over a range sufficient for the interference to sweep well over the i-f response of the receiver. The i-f bandwidth of the OGO command receiver is 36, +8, -0 kc at the 3 db points. It may therefore be necessary for the interfering carrier to deviate $\pm$ 40 kc from the signal carrier frequency.

The output voltages v, $u_0$, and $u_1$ are the baseband and subcarrier signals respectively at the outputs of their corresponding envelope demodulators. It is important that v is extracted before its DC level has been removed.

2.4.2.1 The Pseudo Random Bit Generator (PRBG)

During part of the interference immunity testing, the signal generator should be modulated by a simulated command signal. One of the simplest ways to accomplish this is by use of a pseudo random sequence generated by a shift register with suitable feedback connections.

The pseudo random sequence must be substantially longer than the memory of the command receiver and decoder. From a study of information available about these system components, it appears unlikely that the carryover will be more than a few bits. It therefore appears that a pseudo random sequence of length 15 bits is sufficient. Such a sequence can be generated by use of a four-stage shift register with appropriate feedback connections. A suitable circuit is shown in Fig. 8.

The shift register shown in Fig. 8 consists of four flip-flops and one exclusive or circuit. The shifts are commanded by the clock pulses.

Initially all the flip-flops except the first one must be set to zero. As clock pulses are applied the output, which can be taken from any of the points A, B, C or D will go through all possible binary sequences of four digits except the four zeros. This one is excluded since zeros stored in all four shift registers would not allow the process to start again by itself.

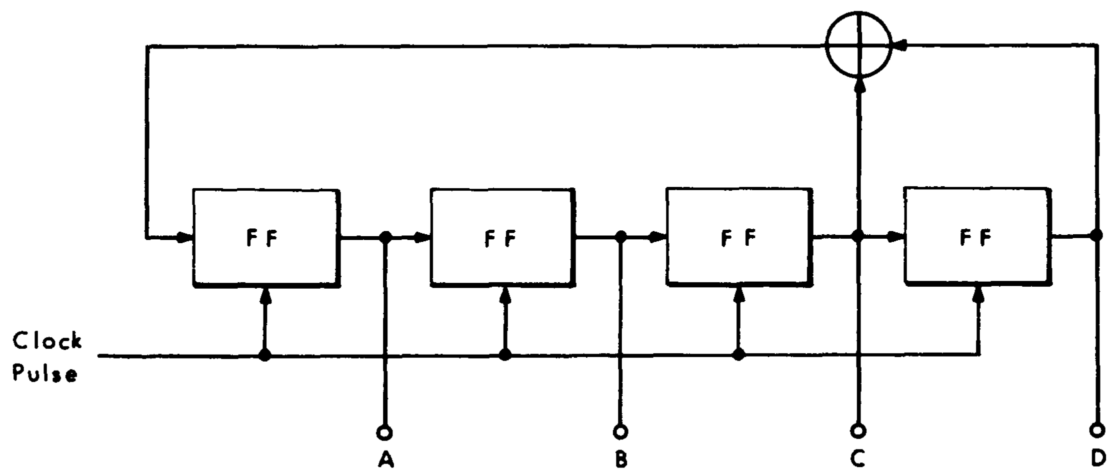


Fig. 8 Shift register with feedback connections for generation of a sequence of length 15.

The longest string of consecutive zeros that can appear at the output is therefore three. If a longer string of consecutive zeros turn out to be desirable one must use a shift register with more stages.

From Fig. 7 we see that one of the outputs from the PRBG feeds the signal generator modulator. This signal is in turn fed through the receiver and data demodulator. The other output from the pseudo random bit generator is fed directly to the comparison circuit. Since the two resulting outputs must be compared with reference to the same bit, the time delays along both paths must be within one bit interval of each other. Since the path through the modulator, receiver and data demodulator has more delay than the other path, this signal must come out of the PRBG at an earlier time. This is, fortunately, not very difficult to accomplish. When the signal to the comparator is taken of lead D in Fig. 8 the output to the data modulator can be taken of leads A, B, or C, whichever results in the proper timing.

### 2.4.3 CW Interference Test

This test can logically be divided into two parts, Test I and Test II.

#### Test I

The purpose of Test I is to assess the signal attenuation and the baseband distortion interference caused by an interfering carrier. The transmitted signal will therefore be continuously modulated by one of the two baseband subcarriers during this test.

The recorded data will consist of signal strength, interference strength, the frequencies of the carrier and interfering signal, baseband demodulator DC and peak levels, and the output voltages from the two subcarrier demodulators.

This test is of fundamental importance in evaluating the interference immunity of the command receiver. A good picture of the performance of the receiver can be obtained by plotting the output levels of the two subcarrier demodulators, $u_o$ and $u_1$, as functions of the signal-to-interference power ratio, $\rho^2$, with the frequency difference between the signal and interfering carriers as a parameter.

We will assume that the active subcarrier corresponds to $u_o$. In this case the ideal situation would be that $u_o$ is constant while $u_1$ is zero.

The plot of $u_o$ will evidence the amount of signal attenuation while the plot of $u_1$ will show the disturbing distortion and interference products.

Test II

The purpose of Test II is to assess how CW interference effects the error rate of the system.

During Test II the transmitter will be modulated by the two data subcarriers keyed by a pseudo random stream of mark and space signals.

The recorded data will consist of the level of the signal and interference, the frequencies of the carrier and of the interfering carrier, the baseband demodulator output voltage, the number of bits transmitted during the test, and the number of error indications that occurred during the test.

### 2.4.4 Testing Procedures

#### 2.4.4.1 Determination of the Linear Range of the IF Amplifier

The purpose of this test is to obtain data on the maximum signal level the i-f amplifier will handle in a linear manner. When carrier only is applied to the receiver there exists a linear relationship between the DC component of the baseband demodulator output and the i-f signal level.

The test procedure will be as follows:

The test setup shown in Fig. 7 will be employed. The signal generator is unmodulated and the interference generator is turned off. The signal generator is adjusted to give a level of -40 dbm at the receiver input when $A_s = 0$. The frequency of the signal generator is set to the carrier frequency of the spacecraft receiver. $A_s$ is set to its maximum value. The AGC loop of the spacecraft receiver is disabled. The setting of the variable attenuator, $A_s$, and the DC component of v, the baseband demodulator output voltage, will be recorded. Decrease the setting of the variable attenuator to zero in steps of 2 db. Plot $\bar{v}$ as a function of $A_s$.

At the point where this curve starts to saturate, the i-f amplifier is overloaded. At this point $\bar{v} = V_x \cdot V_x$ can be used for adjusting the delayed loop of a double loop AGC system as well as for interpretation of the data from the interference immunity tests.

2.4.4.2 <u>CW Interference Testing Procedures</u>

<u>Test I</u>

The purpose of this test is to obtain data on signal suppression and baseband interference and distortion.

The test setup shown in Fig. 7 will be used. The random bit generator and the comparator are not used during Test I.

The transmitted signal will be continuously modulated by the subcarrier used to indicate a binary zero. The modulation percentage will be the same as will be used during actual updata transmission. Set the frequency of the signal generator to the carrier frequency of the updata receiver. Set the signal generator output level so that the signal level is -40 dbm into the receiver when $A_s = 0$. Set the interference generator so that the interference level is zero dbm into the receiver with $A_i = 0$.

During the measurements, $A_s$ shall be changed in steps of 10 db over its range except from 65 to 75 db where the steps shall be 5 db. The receiver may fail to operate for $A_s > 75$ db. $A_i$ shall be changed in steps of 5 db over the range $A_s < A_i < A_s + 60$ db except in the range $A_s + 30 < A_i < A_s + 50$ db where the steps shall be 2 db.

The frequency of the interfering carrier, $f_i$, shall be varied over the range $f_c \pm 40$ kc in steps of 2 kc except for the range $|f_c - f_i| \leq f_1 + 1$ kc where steps shall be 500 cps. $f_1$ is the frequency of the highest frequency command tone.

In particular the frequencies

$$|f_c - f_i| = f_o$$

and

$$|f_c - f_i| = f_1$$

shall be tested even if they do not fall exactly into the regular pattern of test frequencies.

The recorded data shall be the settings of the two attenuators, $A_s$ and $A_i$, the frequencies of the signal and of the interfering carrier, and the voltages v, $u_o$ and $u_1$.

Test II

The purpose of this test is to obtain data on how CW interference affects the error rate of the OGO command link.

The test setup shown in Fig. 7 will be used during test II.

The transmitted signal will be modulated by the mark and space subcarriers. These subcarriers will be keyed by the output from the PRBG. The two outputs from the PRBG must be selected so as to compensate for the delay in the transmitter-receiver loop. Set the frequency of the signal generator to the carrier frequency of the updata receiver. Set the signal generator output level so that the signal level is -40 dbm into the receiver when $A_s = 0$. Set the interference generator so that the interference level is zero dbm into the receiver with $A_i = 0$.

During the measurements, $A_s$ shall be changed in steps of 10 db over its range except from 65 to 75 db where the steps shall be

5 db. The receiver may fail to operate for $A_s > 75$ db. $A_i$ shall be changed in steps of 5 db over the range $A_s < A_i < A_s + 60$ db except in the range $A_s + 30 < A_i < A_s + 50$ db where the steps shall be 2 db.

The frequency of the interfering carrier, $f_i$, shall be varied over the range $f_c \pm 40$ kc in steps of 2 kc except for the range $|f_c - f_i| \leq f_1 + 1$ kc where steps shall be 500 cps. $f_1$ is the frequency of the highest frequency command tone.

In particular the frequencies

$$|f_c - f_i| = f_o$$

and

$$|f_c - f_i| = f_1$$

shall be tested even if they do not fall exactly into the regular pattern of test frequencies.

The recorded data shall be the settings of the two attenuators, $A_s$ and $A_i$, the frequencies of the signal and interfering carrier, the voltage v, the number of digits transmitted during each run and the number of errors occurring during each run.

Special attention shall be paid to catastrophic failures such as loss of sync or severe receiver overloading. Comments shall be made in the experimental log at points where they occur.

## III. PROGRAM FOR THE NEXT REPORTING INTERVAL

a. Analysis of interference effects in AM and FM basebands will be continued. Impulse interference will be covered.

b. Data demodulation will be studied to determine error rates in the presence of interference.

c. Additional test procedures for laboratory measurements of performance under various interference conditions will be developed.

d. RFI analysis techniques suitable for spacecraft will be investigated.

# IV CONCLUSIONS AND RECOMMENDATIONS

a. Discriminators are commonly used in place of phase demodulators in phase modulated command systems. It is shown that when the command information is carried by narrow subcarrier channels the loss of performance above threshold caused by the use of a discriminator for this purpose is insignificantly small. Below threshold, baseband impulses begin to appear and performance degrades in accordance with the results of Sec. 2.1.2.

b. CW interference in an AM command system employing narrow subcarrier channels is the least serious form for interference and has much less effect on the demodulated baseband than it does in a comparable angle modulated system. Unless the beat between the signal and interfering carriers fall within a subcarrier channel a properly designed AM command receiver should not be seriously affected. Interference in an AM system from a narrowband angle modulated signal results in an effect very similar to CW interference. However, strong amplitude modulated interference in an AM command system may completely obliterate the command modulation through intermodulation distortion.

c. One of the effects of strong CW interference on an AM system is to cause a strong attenuation of the desired signal relative to its level in the absence of interference. Unless the subcarrier demodulator has sufficient dynamic range to cope with the variations in signal levels, the system may fail even though no interference or distortion products fall within the passbands of the subcarrier filters. This failure mode may be partly removed by use of a modified AGC circuit where part of the AGC voltage is taken from the subcarrier demodulators.

d. A program for interference immunity testing of the command receivers requires careful planning because of the large number of variables involved. It appears that the CW interference tests will give the most immediately useful data. They should, therefore, be undertaken as a first step in the testing program.

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