A Multiple-Rate Digital Command Detection System With Range Cleanup Capability

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A multiple-rate command transmission and detection scheme that utilizes a composite signal as a subcarrier is described. The composite subcarrier is constructed in such a way that it can be used for ranging. Furthermore, the ranging and command signals can be processed in a near-optimal fashion with the same detector system. The performance of the detector system is analyzed and found to be consistent with realistic mission requirements.

I. Introduction

Most space vehicles employ phase-locked receivers that track the unmodulated (or residual) uplink carrier signal. As a result, all information transmitted to the spacecraft must be such that, when modulated onto the carrier signal, the information contained in the modulation sidebands falls outside the receiver tracking loop bandwidth. Because of this, information signals containing a large amount of low-frequency energy (such as command signals) must first be modulated onto an intermediate carrier (or subcarrier) before finally being modulated onto the carrier.

The introduction of the subcarrier, however, is not a "cure-all," since in order to coherently demodulate the information signal one must now coherently track the subcarrier in addition to the carrier and data. The standard approach to all of the above is to provide independent tracking loops to "track out" all of the information contained in each of the carriers. It seems ironic that after going through all this trouble to eliminate the information contained in these carriers many spacecraft systems employ a second channel (subcarrier), the phase of which is used to gather range information.

In this article a command detector system is described that uses a composite signal as a subcarrier. This composite subcarrier is designed in such a way that it can also convey range information. As a result it is possible to perform essentially optimum detection of the command and ranging signals with the same detector. A preliminary analysis indicates that the performance of the detection system is adequate for most spacecraft missions.

Before discussing the system let us consider a few constraints within which the system must operate. First, we will make the command bit rate coherently related to the subcarrier rate. The reason for this is ease of generation of command signals on the ground as well as the reference signals in the spacecraft. Actually, we will consider only systems wherein the subcarrier and bit rates are related by a power of 2, although this constraint is not absolutely necessary. Second, we will consider systems having multiple command rate capabilities with the proviso that the lowest symbol rate is one symbol per second and that a particular symbol rate will be used only if the symbol energy-to-noise density ratio at that symbol rate exceeds 10 dB ($P_e \leq 10^{-5}$). Finally, we will allow the received signal at the spacecraft to undergo a doppler frequency shift. However, we will restrict the doppler magnitude (fractional shift δ in Hz/Hz) to less than 10^{-6}. Although many spacecraft missions actually experience absolute doppler shifts in excess of this amount the majority of this shift is predicted and compensated for at the ground transmitter. Thus, a bound of 10^{-6} appears to be reasonable for the doppler that the spacecraft actually experiences.

II. System Description

The system described here is motivated by the range cleanup loop discussed by Hurd (Ref. 1). This cleanup loop is actually a generalization of a phase-locked loop wherein the loop voltage-controlled oscillator (VCO) is replaced by a VCO/function generator combination. The function generator provides a synchronous version of a square wave signal modulated by the particular range code component that is being received. In addition the function generator also provides a second signal consisting of the same square wave signal modulated by the next lower range code component.

Also provided with the clean-up loop are two correlators. One of these correlates the received signal with the same composite function generator signal which is being used in the loop to derive the loop error signal. As long as this correlator output is sufficiently high we know that we are tracking with the proper range code reference. The second correlator compares the received signal with the other function generator (next range code) signal. If this correlator output becomes sufficiently high we know that the received signal has undergone a change in the range code component. Detection of this change causes the function generator to shift references enabling the loop to track the new composite signal. Likewise, the correlation detector references are changed so that they are looking for the present composite signal as well as the next lower composite signal.

If one thinks about the operation of the cleanup loop for a moment, it soon becomes apparent that the loop is no more than a phase-locked loop with a detector for sensing when the loop is in lock. Consequently, one would think that by adding an in phase data integrator leg (which already exists in the lock detector circuitry) and a mechanism whereby the effect of data modulation could be removed from the error channel, the cleanup loop could be converted to a subcarrier demodulator and bit (symbol) detector where the subcarrier is actually the composite range signal. A mechanism whereby this can be accomplished is shown in Fig. 1.

Before describing the operation of the system let us consider the structure of the composite signal we will use as a subcarrier. The subcarrier signal consists of a square wave of frequency f_{sc} phase modulated by one of its own subharmonics. For example if f_{sc} denotes a square wave of frequency f_{sc} then the ith possible composite subcarrier (also denoted c_i, the ith ranging code) is represented by

$$c_i = \overline{f_{sc}} \oplus \overline{\frac{f_{sc}}{2^i}}$$

where $\oplus$ indicates modulo 2 addition if the square wave levels are $\{0,1\}$ or multiplication if the levels are $\{+1,-1\}$. In order to use these signals as subcarriers the code components must be used sequentially. For example, we initially send component c_1. After a sufficient period of time we send c_2, and so on. (Note that c_1 is a square wave signal of frequency $f_{sc}/2$.)

With the aid of the flow chart of Fig. 2 we can now explain the operation of the demodulator/detector. Initially all the switches in Fig. 1 are in the acquisition (ACQ) position and the function generator is outputting components c_1 and c_2. As long as no signal is received by the demodulator, neither the c_1 nor the c_2 correlators in the lock detector will accumulate to sufficiently high value. As soon as component c_1 is received at the demodulator the signal (which is filtered and sampled at the Nyquist rate) is applied to the transition sample selector of the phase-locked (cleanup) loop. The tracking loop consists of an all-digital second-order phase-locked loop such as the one discussed by Holmes and Tegnelia (Ref. 2) with the VCO replaced by a cleanup loop function generator. A more detailed discussion of how the loop operates is given in Ref. 2.

After a sufficient length of time the tracking loop will lock on to the received component c_1, and the c_1 correlator in the lock detector will detect its presence. The system will continue in this mode indefinitely unless either the signal is lost or the c_2 correlator detects the presence of c_2. If c_2 is detected the function generator changes the signals c_1 and c_2 to c_2 and c_3 respectively and the system will continue indefinitely in this state until signal is lost or c_3 is detected, etc. Note that by supplying the signal labeled by c_i to the spacecraft transmitter the "cleaned up" range signal can be transmitted back to the ground.

Let us now associate with each ranging component c_i a particular command symbol rate. In particular let the symbol rate correspond to the frequency of the signal which when modulated onto f_{sc} produces the composite subcarrier. In other words when component c_i is sent, the corresponding symbol rate will be $f_{sc}/2^i$ symbols per second. Furthermore, since the symbol and subcarrier timing was specified as synchronous we see that detection of c_i corresponds to detection of the rate associated with c_i as well as the sequential determination of symbol timing at that rate. (Sequential symbol timing is determined by examining the sign of the c_{i+1} correlator at the time c_{i+1} is detected. This is the same process whereby range ambiguity is removed in the sequential ranging system.)

Recall that the switches in Fig. 1 were all set to the ACQ position. This is to allow efficient detection of the sequence of code components. However, once we reach the date rate at which we desire to send commands we must change all the switches to the DEMOD position. Since code components c_i and c_j, $i \neq j$ over an appropriate integration time are orthogonal, a convenient way to notify the detector that c_i corresponds to the desired command symbol rate is to follow the unmodulated transmission of $\overline{c_i}$ by a transmission of c_i (complement). Thus, the c_i correlator would serve not only as a c_i signal presence indicator but also as an end of acquisition detector. Once the end of acquisition sequence is detected, and the switches placed in the DEMOD position, the system is ready to demodulate and detect command data as well as track the c_i range code component. Note also that this demodulator does not have the characteristic phase ambiguity problem that most suppressed carrier tracking loops have, since we can define the transmission of $\overline{c_i}$ as actually the transmission of c_i with an all zeros command modulation signal.

Once the device is in the DEMOD mode the command symbol stream is integrated (accumulated) over the appropriate symbol time T_s and the symbol detected. The symbol sign is then applied to the lock detector to remove the effects of data modulation on the lock detector accumulators. Likewise, the symbol sign is injected into the loop to remove data modulation on the loop error signal. Finally, we note that a signal from the lock detector representing the c_i correlator most significant bit (MSB) is applied to the loop after the second loop accumulator. This signal normally is not needed by the loop. However, during the time that $\overline{c_i}$ is being received to indicate the end of the acquisition sequence, the loop error signal will be complemented and hence cause the tracking loop to correct in the wrong direction. Note also that the c_i correlator will produce a highly negative rather than a positive correlation during this time. Consequently if the total accumulation time of the two loop accumulators matches the accumulation time of the lock detector, the loop error signal can be inverted so that the loop retains lock.

III. System Design

The parameters associated with the demodulator/detector are as follows (Ref. 2):

Δ_2 = integral control phase bump size

Δ_1 = proportional control phase bump size

$n = \Delta_1/\Delta_2$

f_{sc} = subcarrier (basic) frequency

$T_{sc} = \dfrac{1}{f_{sc}}$

T_s = symbol time

L = phase bump exponent (i.e., $\Delta_2 = 2^{-L}$)

K = subcarrier frequency exponent (i.e., $f_{sc} = 2^K$)

M = number of subcarrier cycles per phase bump

W = bandwidth (one sided) of the pre-sampling filter $= \dfrac{8}{T_{sc}}$

$\alpha = \dfrac{1}{16}$ = portion of each subcarrier cycle used for linear signal transition (Ref. 2)

m = number of actual signal transitions occurring in M subcarrier cycles $\approx 2M$

A = received signal amplitude (outside the linear transition region)

N_0 = one-sided noise spectral density

Using these definitions we can specify the system constraints as

$$T_s \geq 1 \text{ second} \tag{1}$$

$$\frac{A^2 T_s}{N_0} \geq 10 \tag{2}$$

$$\delta \leq 10^{-6} \tag{3}$$

We also impose the additional constraints that

$$r \geq 2 \tag{4}$$

$$\sigma_L < 0.1 \text{ (radians)} \tag{5}$$

$$\frac{MT_{sc}}{\tau_2} \leq \frac{1}{3} \tag{6}$$

where r is the loop damping parameter, σ_L is the rms loop phase error and τ_2 is the loop secondary time constant. The constraint given by Eq. (6) is necessary to ensure that the loop will be stable in light of the fact that the loop accumulator produces a transport delay in the loop transfer function (Ref. 3, pp. 19–20). Using Eq. (6) and the fact that

$$\tau_2 = MT_{sc}\frac{\Delta_1}{\Delta_2} \tag{7}$$

results in the requirement

$$n \geq 3 \tag{8}$$

We will choose $n = 4$ for ease of implementation (although it may be necessary to increase n to 6 in order to retain stability at strong signal levels).

It was also shown in Ref. 2 that

$$\sigma_L^2 = \frac{\pi^3 N_0}{256\, nMT_{sc}A^2} + \frac{\pi^{5/2} n\sqrt{N_0}}{2^L\sqrt{MT_{sc}A^2}} \tag{9}$$

and

$$r = \frac{16\, n^2\sqrt{2MT_{sc}A^2}}{2^L\sqrt{N_0}} \tag{10}$$

Now, if the constraints Eqs. (4) and (5) are to be satisfied then we must require that

$$\xi \geq \max\left[\frac{1}{2}\log_2\left(\frac{25\pi^3}{64n}\right), \quad L + \log_2\left(\frac{\sqrt{\pi}}{8n^2\sqrt{2}}\right)\right] \tag{11}$$

and

$$L \geq \log_2\left[\frac{25\pi^{5/2} n\, 2^{\xi}}{\left(2^{2\xi} - \dfrac{25\pi^3}{64n}\right)}\right] \tag{12}$$

be simultaneously satisfied where

$$\xi = \frac{1}{2}\log_2\left(\frac{MT_{sc}A^2}{N_0}\right) \tag{13}$$

The equality portions of Eqs. (11) and (12) are shown in Fig. 3 for $n = 4$. Also shown in Fig. 3 is the line $L = 8$, which is the minimum value of L necessary to keep the phase bump quantizing error variance within reason. This results in an acceptable solution region as shown by the shaded region of Fig. 3.

It remains to determine the values of M and T_{sc}. Note that thus far all of the performance parameters depend on the product MT_{sc}. The two can be separated, however, by selecting M such that the variance of the phase error buildup due to doppler is less than 0.01 (rad)^2. Toward this end we have from Ref. 2:

$$\phi_d \triangleq 2\pi M\delta \leq \frac{2\sqrt{3}}{10} \tag{14}$$

from which we get

$$M \leq 5.45 \times 10^4 \tag{15}$$

For implementation reasons we would like M to be a power of 2. Thus, we consider candidate solutions of the form $M = 2^i$, $i = 1,2,\cdots,15$. Noting that large M corresponds to high subcarrier frequencies (which is desirable for ranging applications) we conservatively choose $M = 2^{14}$. Then, from Fig. 3, Eq. (13), and the design point condition

$$\frac{A^2 T_{s_{max}}}{N_0} = \frac{A^2}{N_0} = 10 \tag{16}$$

we find that $T_{sc_{min}} = 2^{-14}$. The significant loop design parameters then become

$$M = 2^{14}, \qquad T_{sc} = 2^{-14}$$
$$L = 8, \qquad n = 4 \tag{17}$$

from which all of the other parameters can be calculated.

Having chosen the loop parameters we now turn to the lock detector accumulators. Recall that Eq. (17) is sufficient to guarantee a symbol error probability of 10^{-5} whenever antipodal signals of mean value $\pm A$ are received and integrated for $T_{s_{max}} = 1$ sec. However, the lock detectors must distinguish between signals having mean values of $-A$, 0, $+A$ where the zero mean value results from correlating with one of the orthogonal signals. This means that the lock detectors are working at a 6 dB disadvantage relative to the symbol detector. If we desire the same error statistic (i.e., 10^{-5}) for miss and false alarm probabilities then we can simply increase the lock detector accumulation times by a factor of 4. This means that *a priori* (i.e., before symbol rate is known) we must accumulate over 4 seconds in the lock detectors. However, once the actual rate has been determined the lock detector accumulation times can be reduced to 4 T_s seconds.

Recall that in the last section we stated that the lock detector most significant bit signal could be used to correct the direction of the loop phase error when c_K was being received *if* the accumulation times of the loop accumulators and the lock detector matched. However, we have just required that the lock detector accumulate over four loop accumulation times. This presents no problem if we split the in-phase lock detector accumulator into two sections such that the first section accumulation time matches that of the loop. Figure 4 is a revised block diagram showing how this can be accomplished. (Note: the notation Σ/X means that the input is accumulated until the accumulation time, or the accumulation index equals X.)

IV. Loop Performance

From the equations of Ref. 2 one can easily find that the design point values for the present design are

$$\begin{aligned} w_L &= 0.44\,\mathrm{Hz} \\ r &= 2.52 \\ \sigma_L &= 5.67\,\mathrm{deg} \end{aligned} \tag{18}$$

These expressions are computed based on a linearized model of the loop phase detector. It turns out that the loop detector actually operates in a region where the linear theory does not apply. In order to improve our performance estimates we will use equivalent linearization about the particular average phase detector operating point. Toward this end let

$$A_{eq}(\phi) = \frac{\partial}{\partial \phi}\{E[\hat{\phi}|\phi]\} \tag{19}$$

be the equivalent phase gain when the loop phase error is ϕ. Since the loop is operating at a high loop signal-to-noise ratio (SNR) we can model the loop phase error as a gaussian process with mean zero and variance σ_L^2. In this case we can average Eq. (19) over this distribution to get

$$\overline{A_{eq}} = \frac{\sqrt{\frac{2\rho}{\pi}}}{\pi\alpha}\left(\frac{1}{1+\frac{\rho\,\sigma_L^2}{\pi^2\alpha^2}}\right)^{1/2} \tag{20}$$

where

$$\rho = \frac{mA^2}{N_0 W} \tag{21}$$

is the sample SNR. By using Eq. (20) in place of A_{eq} of Ref. 2 one obtains results that include the effects of limiter suppression. Note, however, that these equations require a recursive procedure for evaluation. Figures 5, 6 and 7 illustrate the behavior of r, w_L and σ_L as ρ varies. Note that the design point value of ρ is 4 dB. Also shown for comparison in Figs. 5 through 7 are the corresponding results using the linear theory.

It should be noted that the sequential acquisition procedure described earlier requires that the tracking loop be free of cycle slipping. Thus, we would like the mean time to slip $T(2\pi)$ to be extremely large. Viterbi (Ref. 4) has shown that at high values of loop SNR ρ_L, the mean time to slip, is well approximated by

$$T(2\pi) \approx \frac{\pi}{2w_L}e^{2\rho_L}. \tag{22}$$

Since $\rho_L \approx 1/\sigma_L^2 \approx 100$ and $w_L \approx 1.0$ we see that cycle slipping while tracking a signal component is indeed an extremely rare event. Consequently, we can assume that cycle slipping can occur only during the time required to detect a code component change, since during this time the loop is operating with the wrong phase reference signal. Hurd (Ref. 5) has shown that if a second-order loop experiences a loss of signal at time $t = 0$ the phase error variance $\sigma_L^2(t')$ at time $t = t'$ is given by

$$\frac{\sigma_L^2(t')}{\sigma_L^2(0)} = 1 + \frac{4r^2}{r^2+1}w_L t' + \frac{4(2r^2+1)r^2}{(r^2+1)^3}(w_L t')^2 + \frac{16r^4}{3(r^2+1)^4}(w_L t')^3 \tag{23}$$

Using the design point values for the loop parameters and the lock detector time of 4 seconds for t' yields

$$\sigma_L^2(4\,\text{seconds}) = 0.141\ \text{rad}^2 \tag{24}$$

Finally, the application of Eq. (22) results in a mean time to slip when experiencing a code change of 4.52×10^6 seconds.

V. Lock Detector Performance

Recall that our model for the filtered subcarrier signal has a linear region of length $T_{sc}/16$ surrounding each subcarrier transition and that the received signal is sampled at 16 f_{sc}. Now, if the phase error is small then out of any 16 consecutive samples we will have seven with mean value $+A$, seven with mean value $-A$ and two samples which are zero mean. (This is strictly true when the square wave is unmodulated and is a pessimistic approximation whenever modulation is present.) If each sample is hard limited to ± 1 and multiplied by the in-phase correlator reference signal the mean value of the sampled input to the accumulator will be

$$\mu_s = \frac{7}{8}\,\text{erf}\left(\sqrt{\frac{R_s}{2^{K+4}}}\right) \tag{25}$$

where R_s is the symbol energy to noise ratio and 2^K represents the number of subcarrier cycles per symbol time T_s. Note also that the variance of the sampled values is $1 - \mu_s^2 \approx 1$. The accumulator sums these samples for 4 seconds, accumulating a total of 2^{20} samples, and compares the summation against a threshold value of T. If Σ represents the value of the accumulation then the probability that the in-phase signal will not be detected, given that it is being received, is given by

$$P_m \triangleq \Pr\left\{\sum < T \,\middle|\, \begin{matrix}\text{signal}\\ \text{present}\end{matrix}\right\} = \frac{1}{2}\,\text{erfc}\left[\frac{(7)2^{17}\,\text{erf}\sqrt{\frac{R_s}{2^{K+4}}} - T}{2^{10}\sqrt{2}}\right] \tag{26}$$

Likewise, the probability that the in-phase signal will be detected, given that it is not being received, is

$$P_f = \Pr\left\{\sum \geq T \,\middle|\, \begin{matrix}\text{signal}\\ \text{absent}\end{matrix}\right\} = \frac{1}{2}\,\text{erfc}\left(\frac{T}{2^{10}\sqrt{2}}\right) \tag{27}$$

Figure 8 illustrates the behavior of the false alarm and miss probabilities as the threshold varies. It should be noted that the probability of a miss is more critical in a sequential system, since the probability of not acquiring is the probability of the union of the miss events whereas the probability of false acquisition is the product of the constituent false alarm probabilities. Therefore, the threshold should be chosen in such a way as to favor the miss probability at the expense of the false alarm probability. We shall assume that $T = 2500$ is chosen which implies that $P_m \approx 10^{-6}$ and $P_f \approx 10^{-2}$ at the design point. Note that these are the component probabilities, not the overall system probabilities, which are discussed later.

A. Acquisition Sequence

We note first that the first signal component c_1 must be sent for a sufficiently long time for the tracking loop to acquire the signal. Based on the experimental evidence of Ref. 2 we see that after receiving the signal for a period of time equal to 60 loop update times (60 seconds in our case) the loop will have acquired with a probability of $1 - 10^{-5}$. Now, note that the time of arrival of the change from code c_i to c_{i+1} does not necessarily coincide with the start of one of the lock detector accumulator cycles. Thus, in general the code change will occur within an accumulator window making the detection of c_{i+1} at the end of this window very unreliable. Therefore, it is necessary that when a code change occurs the new code must be transmitted for a minimum of two lock accumulator intervals (8 seconds in our case). Actually, it may be necessary to transmit each new component for three lock detector intervals so that the loop phase error can stabilize during the third interval. We will, however, assume for now that two intervals are sufficient.

Since each of the codes must be transmitted in sequence, we have that the minimum acquisition time for the demodulator/detector when operating at a symbol rate such that

$$T_s = 2^K T_{sc} \tag{28}$$

is

$$T_{ACQ_{\min}}(K) = 8K + 60\,\text{seconds} \tag{29}$$

At the lowest symbol rate this corresponds to a minimum acquisition time of 172 seconds.

B. Acquisition Statistics

The two acquisition statistics of interest are the probability of not acquiring, given that the preamble was transmitted, and the probability of falsely acquiring, given that the preamble was not sent. If we define A_K as the event that the transmission of c_K was missed (not detected) then the probability $P_{NA}(K)$ of not acquiring at the rate for which $T_s = 2^K T_{sc}$ is upper bounded by

$$P_{NA}(K) \leq \sum_{k=1}^{K} \Pr\{A_k\} + \Pr\{A_{\bar{K}}\} \tag{30}$$

where $\Pr\{A_{\bar{K}}\}$ is the probability of missing when $\bar{c}_K$ is transmitted. Recalling that $\Pr\{A_1\} = 10^{-5}$, $\Pr\{A_j\} = 10^{-6}$ for $j \neq 1$, and $P_{NA}(K) \leq P_{NA}(14)$ yields

$$P_{NA}(K) \leq 10^{-5} + 1.4 \times 10^{-5} = 2.4 \times 10^{-5} \tag{31}$$

For false acquisition at rate K we have that

$$P_{FA}(K) = \left[\prod_{k=1}^{K} (P_{f_k}) \right] P_{f_{\bar{K}}} \tag{32}$$

where P_{f_i} is the probability of falsely detecting the presence of component c_i. If we assume that 64 symbols per second is the highest symbol rate for the unit then

$$P_{FA}(K) \leq P_{FA}(K = 8) = 10^{-18} \tag{33}$$

It appears that these statistics are compatible with most mission design values.

VI. Conclusions

Described herein is a method for transmitting command information on a subcarrier signal which can also be used for ranging. Next, a detector system was described that was capable of detecting the command modulation without destroying the range information. The detector was then analyzed and found to have false acquisition and failure to acquire probabilities which are compatible with realistic mission requirements. However, seldom does one obtain something in one area without giving up something in another, and this is no exception. In order to use the system described herein one must be willing to accept the increased acquisition time associated with sequential detection. The minimum acquisition time appears to be approximately twice that achievable in more conventional systems. Methods whereby the acquisition time can be reduced as well as experimental verification of these preliminary findings will be subjects of a future article.

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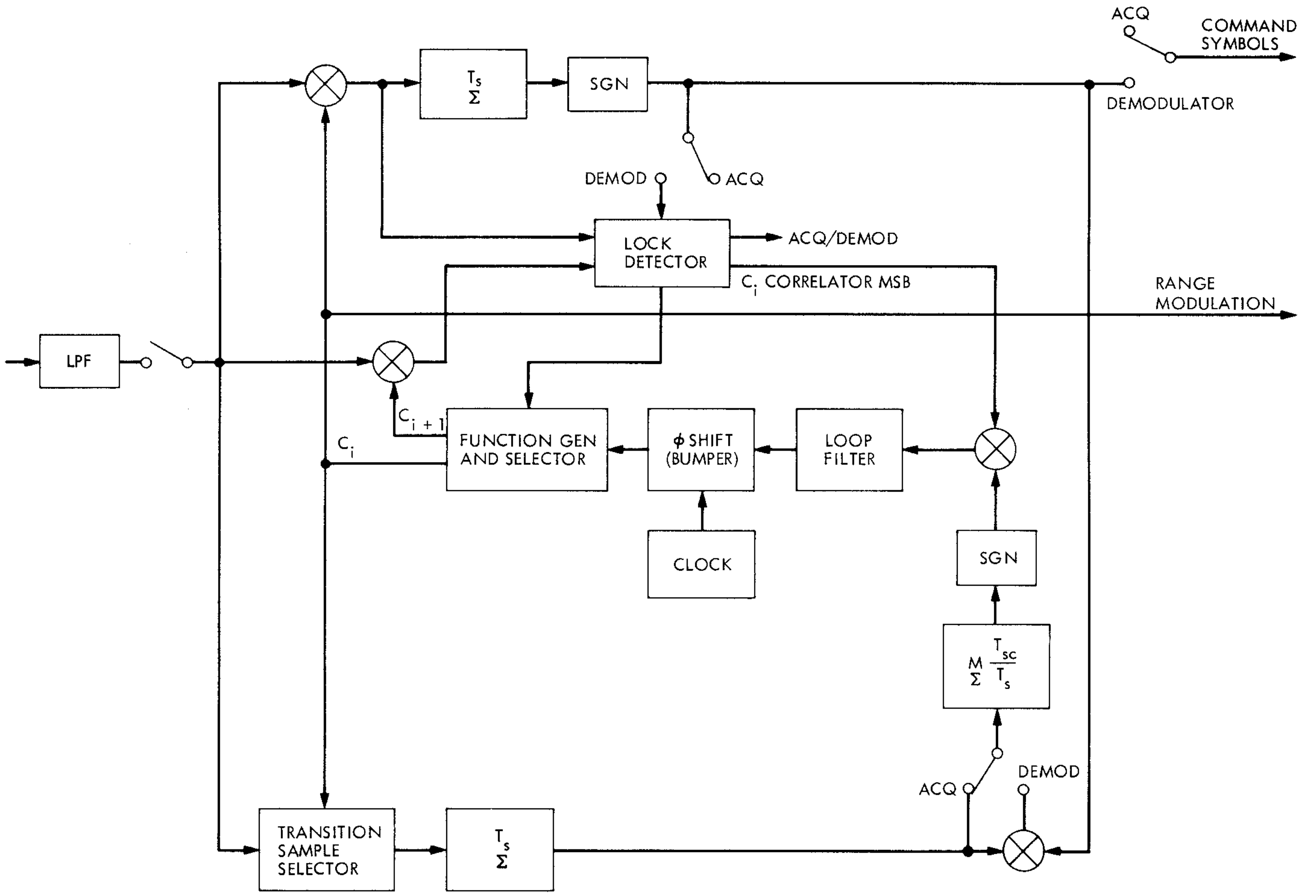

Fig. 1. Block diagram of demodulator/detector system

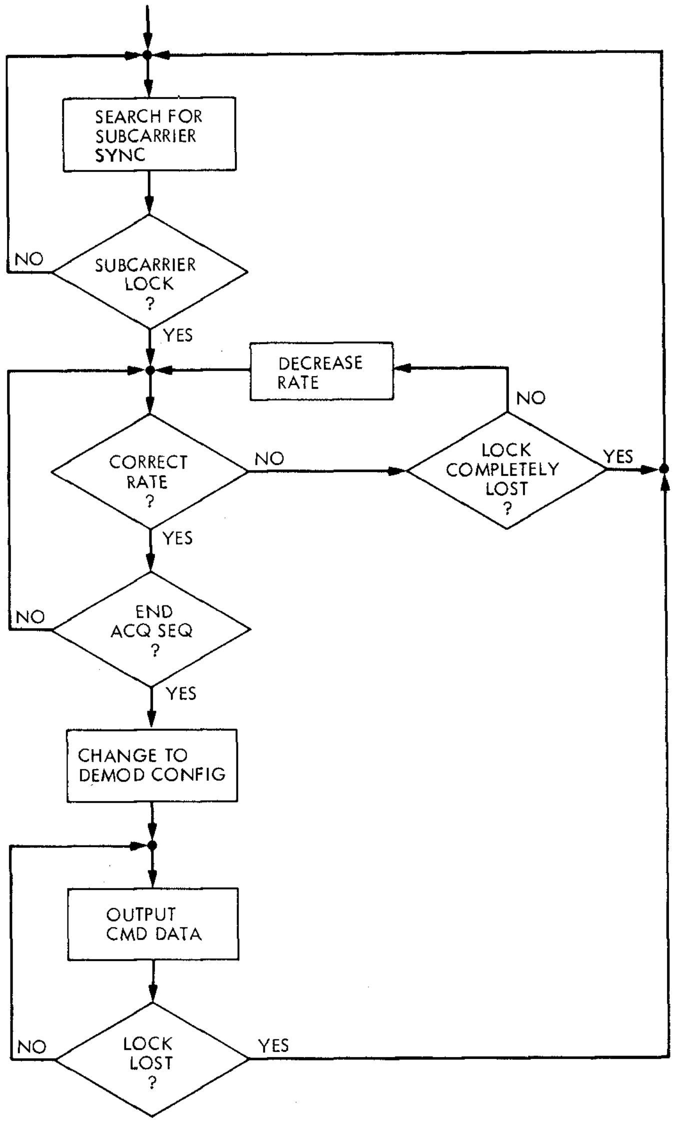

Fig. 2. Flow chart of demodulator/detector operation

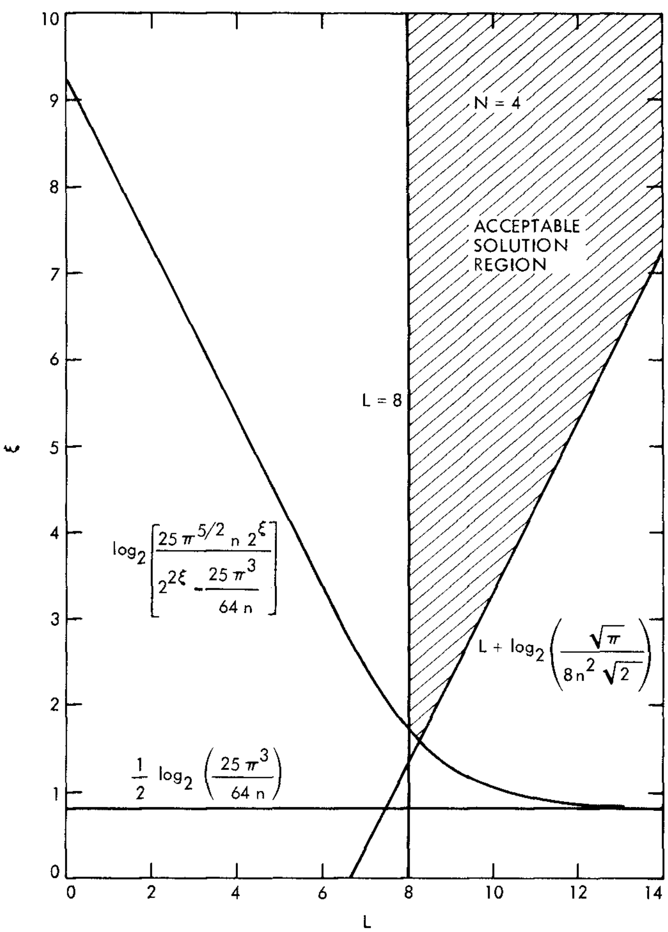

Fig. 3. Parametric representation of ξ and L

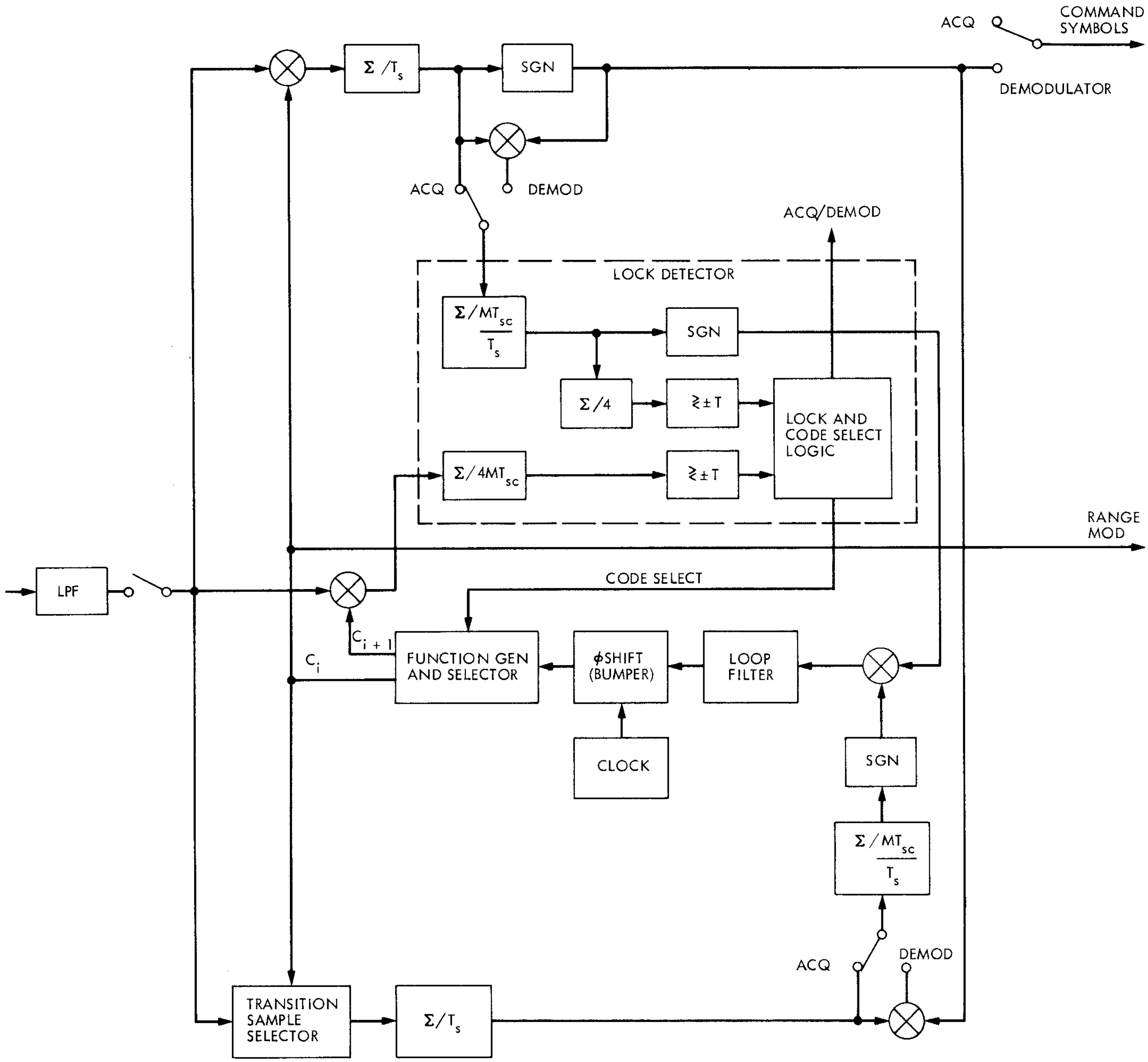

Fig. 4. Revised block diagram of demodulator/detector

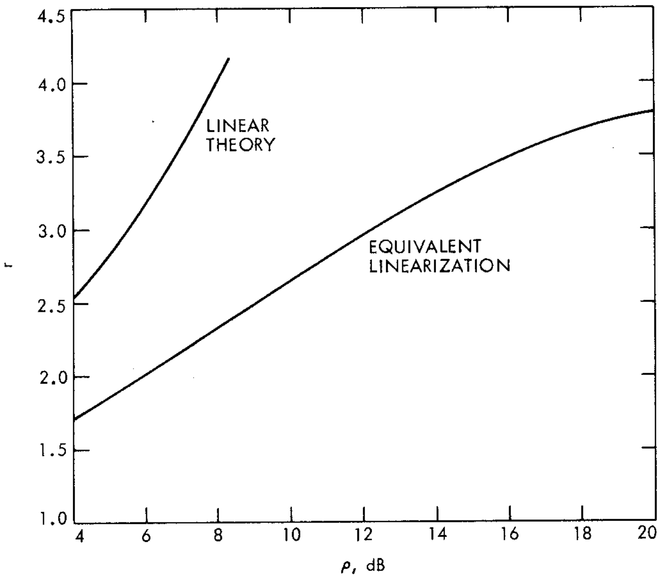

Fig. 5. Variation of damping parameter *r* with sample SNR

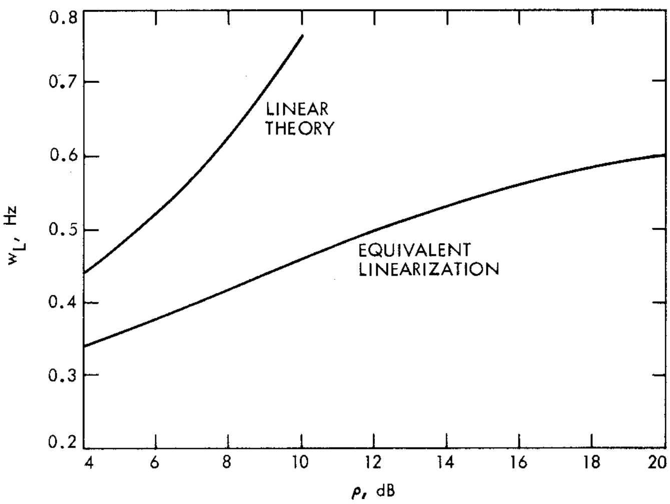

Fig. 6. Variation of loop bandwidth with sample SNR

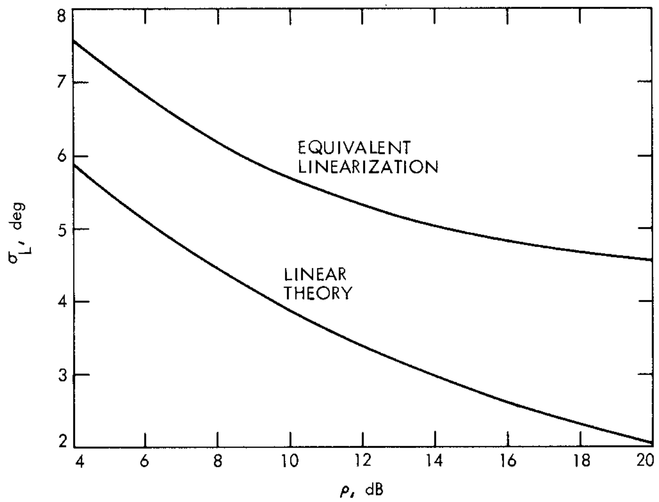

Fig. 7. Variation of rms loop phase error with sample SNR

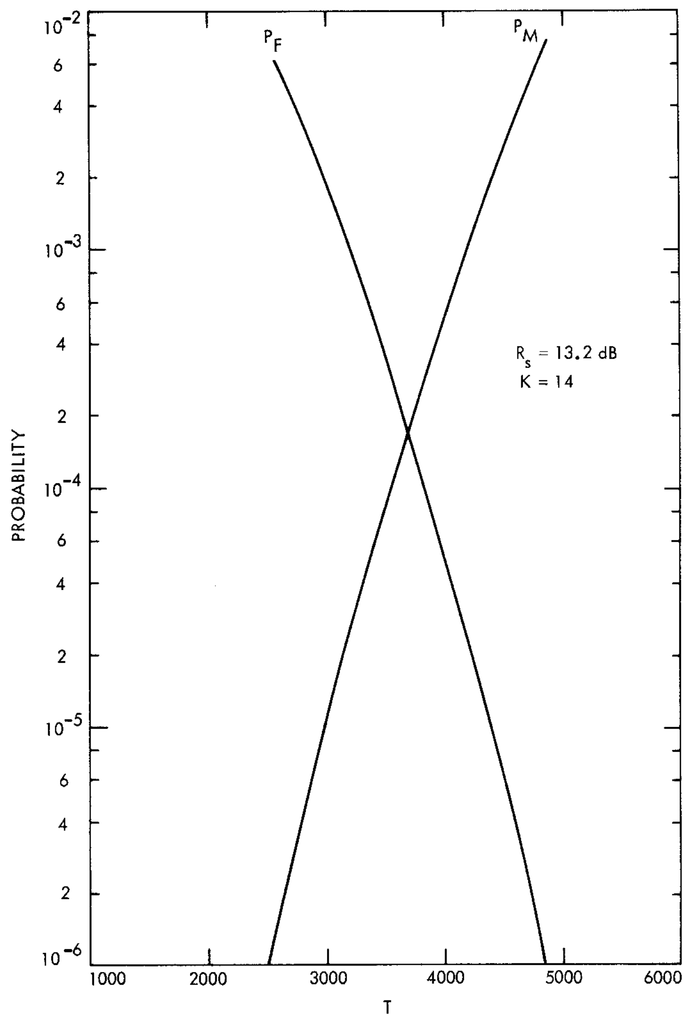

Fig. 8. Comparison of false alarm and miss probabilities vs threshold