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ADVANCED TRACKING AND DATA RELAY EXPERIMENTS STUDY - Multimode Transponder Experiment Analysis Procedure

Magnavox Research Laboratories
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Torrance, California 90503

15 September 1973

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ABSTRACT/ CONCLUSIONS

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This report provides the procedures to analyze the data from the experiment described above and sets forth the criteria to establish the quality of signals.

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16. Abstract

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PREFACE

This report, dated 15 September 1973, is entitled "Multimode Transponder Experiment Analysis Procedure." It is the third in a series of three reports which contain the findings of a program titled "Advanced Tracking and Data Relay Experiments." The work was accomplished by the Magnavox Research Laboratories of Torrance, California and complies with the requirements of Contract Number NAS5-21824, Contract Data Item 6.

Plans and implementation concepts have been developed for a series of experiments utilizing a Multimode Transponder mounted in an aircraft working either through a spacecraft or directly with a ground station which would simulat a TDRSS user working through the TDRSS. The purpose of the experiments would be to determine the best modulation and encoding techniques for combating RFI and multipath propagation and to determine the characteristics of VHF and UHF RFI in discreet bands. The experiments would also determine the feasibility and accuracy of range and range rate measurements with the various modulation and encoding techniques.

This report provides the procedures to analyze the data from the experiment described above and sets forth the criteria to establish the quality of signals.

Magnavox wishes to acknowledge the assistance of Pat Mitchell, ATDRE technical officer and Keith Fellerman of the TDRSS program office, G.S.F.C.

This report was prepared by Messrs. R. Cnossen and J. Moses of MRL.

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ATDRE ANALYSIS PROCEDURE

The criteria for establishing the quality of the test results for those experiments outlined in paragraph 2.2 are developed in this section. The expected performance for the Multimode Transponder equipment is established for (1) data error rate, (2) range measurement and (3) range rate measurement. The results of these calculations are then presented as a family of curves so that the performance for many modes of operation can be easily determined and performance trends can be readily observed.

4.1 EVALUATION PROCEDURE

Basically, the test data derived from the experiment will be colated into a series of performance curves. In this format, the test data would be subjected to the performance criteria and expected results derived in this section. The performance curves for the many modes of operation would also be compared, to determine which operational configurations would be most suitable for the proposed TDRSS.

Five basic criteria have been selected from which the performance of the Multimode Transponder would be based:

- Signal Acquisition
- Data Error Rate
- Voice Quality
- Range Measurement Error
- Range Rate Measurement Error

The expected results for these performance criteria are presented in this section.

4.2 EXPECTED PERFORMANCE

The expected performance of the Multimode Transponder and associated ground equipment has been calculated and curves have been constructed to show the theoretical performance including implementation losses.

4.2.1 DATA ERROR RATE

4.2.1.1 Theoretical Performance

The basic method of transmitting data in a PN mode of operation is to modulo-2 add data with a PN code. As previously analyzed, the worst case processing gain for data using this technique is:

$$PG = 1/(T_c B_{data}) = B_{RF}\, T_b$$

where $B_{RF} = T_c^{-1}$ and $T_b^{-1} = B_{data}$, and the output noise is Gaussian. We can write for the tolerable interference to signal ratio:

$$\frac{N}{S} = \frac{PG}{E_b/N_o}$$

where E_b/N_o is a normalized performance parameter, the ratio of energy per bit to interference power per Hz.

Figure 4-1 displays theoretical probability of error versus E_b/N_o for white Gaussian noise. The binary modulation schemes are coherent PSK, differentially encoded ΔPSK, and differentially coherent DPSK. Note that coherent PSK is a practical impossibility since absolute phase is not available to the receiver. Differential encoding, where 1 denotes a phase change and 0 absence of a phase change, is a practical way to resolve the 0, π phase ambiguity of a Costas loop demodulator (shown in figure 4-2) for a biphase signal. Figure 4-1 gives the results of a computer simulation of ΔPSK. There is excellent agreement with theory, and the simulation shows that phase slips of the Costas loop can be made negligible.

4.2.1.2 Practical Performance

There are two major sources of E_b/N_o degradation from the theoretical for DCPSK detection. The first, which is less significant, is the imperfect carrier tracking due to front-end noise. The second, which is the major part of the total E_b/N_o degradation, is the loss due to RC data filtering in the demodulator.

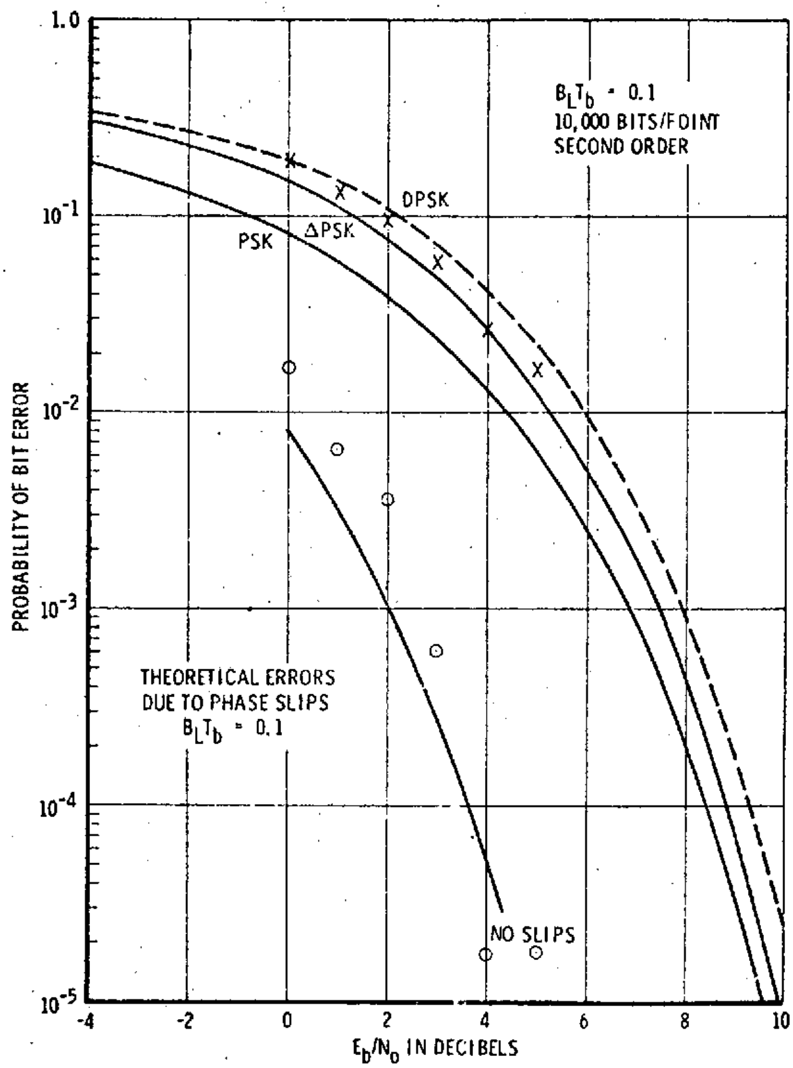

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Figure 4-1. ΔPSK Performance

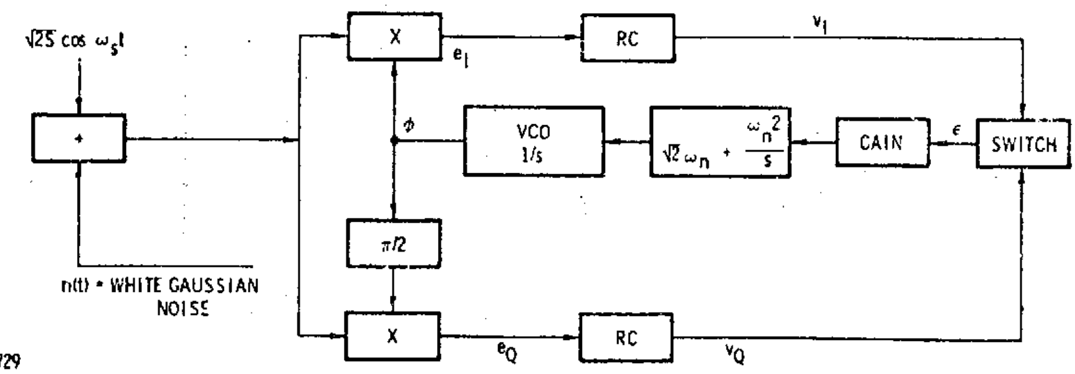

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Figure 4-2. Costas Loop Tracking for ΔPSK Demodulation

4.2.1.2.1 Imperfect Carrier Tracking

Imperfect carrier tracking due to front-end noise is interpretable in terms of a phase error in the reference. As illustrated in figure 4-3, a phase error ϕ in the reference changes the in-phase channel voltage by cos ϕ with a noise contribution of sin ϕ. Thus, the bit energy to noise density varies as $(\cos \phi + \sin \phi)^2$.

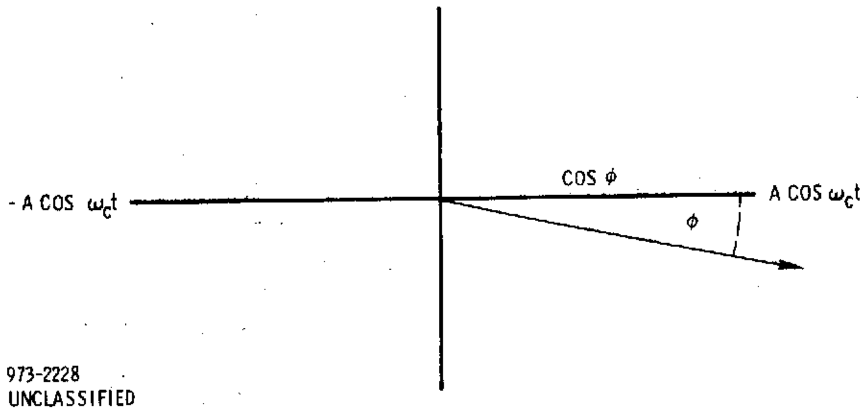

Figure 4-3. Effect of Reference Angle Error ϕ on Signal Vector

For small phase errors the above is approximated by $(1 + \phi)^2$. The probability of bit error conditional on the reference angle error ϕ is given by:

$$P_e(\phi) = \frac{\exp\left[-E_b/N_o (1 + \phi)^2\right]}{2 \quad 2\, E_b/N_o\, (1 + \phi)}$$

The probability of bit error is found from the integral given by:

$$P_e = \int_{-\infty}^{\infty} P_e(\phi)\, P(\phi)\, d\phi$$

Upon integration, with the appropriate substitution for the approximately Gaussian variable $P(\phi)$, we get P_e [exp $- (2\, E_b/N_o + 1)^2 \frac{\sigma 2}{2}$]. Where σ^2 is the mean square value of the noise in the carrier loop and P_e is the probability of error for a DCPSK system with a perfect reference. σ^2 will not be evaluated. On a low pass equivalent basis, since B_L (loop noise bandwidth) is a low pass one sided bandwidth, there is an effective processing gain in the loop equal to 10 log $\frac{\text{Data Rate}}{B_L}$. Thus, the receiver front end noise, white gaussian noise, after filtering by the carrier loop is $-(E_b/N_o + 10 \log \frac{DR}{BL})$ relative to unit signal power, thus this is σ^2. The E_b/N_o is obtained from the theoretical curve at the particular point where the degradation is required to be known. Using the equation for P_e and the relationship above for σ^2, the E_b/N_o degradation due to imperfect carrier tracking is obtained for a data rate of 100 bps, 300 bps and 1000 bps. This is shown in figure 4-4.

4.2.1.2.2 RC Data Filtering

The degradation due to RC data filtering is essentially obtained from the work done by J. Jay Jones [1]. He provides a relationship of probability of error vs E_b/N_o for a single pole RC filter for cpsk with the 3 dB RF bandwidth and data rate as variable parameters. This is shown in figure 4-5.

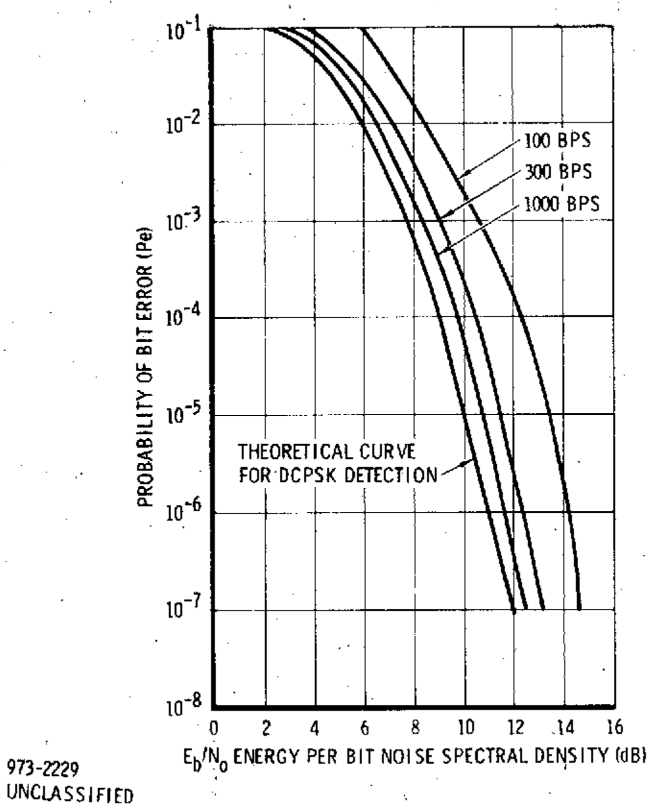

Figure 4-4. P_e vs E_b/N_o for dcpsk with imperfect carrier tracking for the specified data rates

[1] Jones, Jay, J., "Filter Distortion and Intersymbol Interference effects on psk signals", IEEE Transactions on Communication Technology, Vol. COM-19 No. 2, April, 1971.

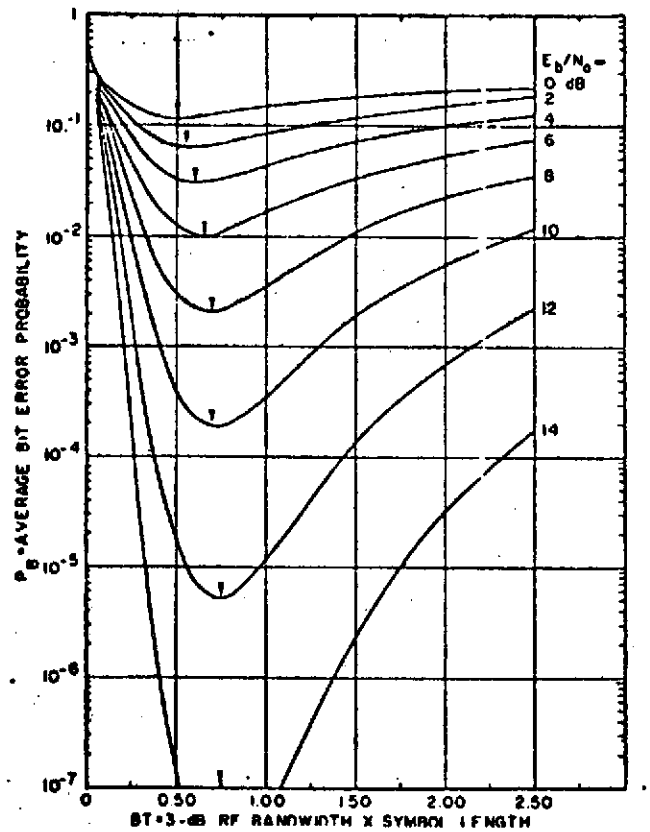

Figure 4-5. Single pole RC data filter detection for cpsk

For the design criteria of the 3 dB bandwidth being equal to the data rate, a relationship of P_e vs E_b/N_o can be obtained for the single pole RC filter. The relationship for P_e vs E_b/N_o for a single pole RC filter for dcpsk is shown in figure 4-6. Combining the degradation from data filtering and imperfect carrier tracking provides the required C/N_o for the specified P_e at the required data rates. This is shown in figure 4-7.

It should be pointed out that data demodulation performance improves by approximately 3 dB when both diversity receivers are tracking and their data outputs are combined. Also during a transpond mode of operation, the return link performance will be reduced somewhat if the signal source (MMT transmitter) is noisy.

For the PN modes of operation, a 1 dB loss in performance can be expected due to imperfect code tracking. Also, since the data is reclocked with the PN code clock, quantization losses can be expected on the order of 1 dB for code/data clock ratios of 10/1 and 2 dB for code/data clock ratios of 3/1.

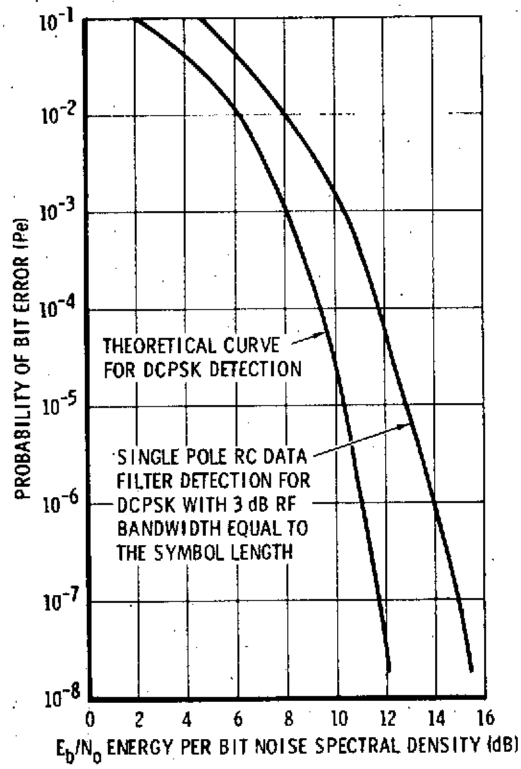

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Figure 4-6. P_e vs E_b/N_o with RC data filtering for dcpsk

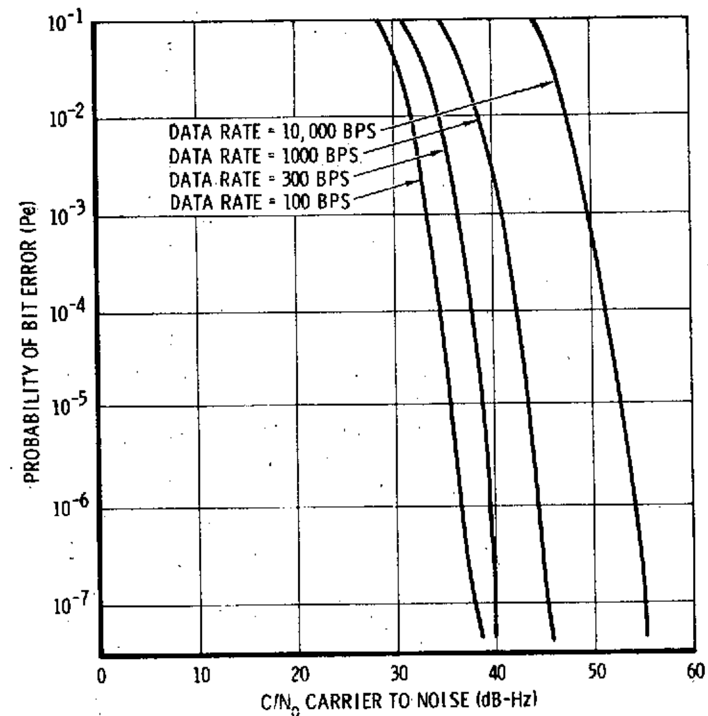

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Figure 4-7. Expected Multimode Transponder Data Recovery Performance

4.2.2 RANGE MEASUREMENT

The expected two-way ranging performance for the Multimode Transponder has been calculated. Bandlimiting and tracking S/N losses have been considered.

The delay lock loop rms range tracking error may be written as:

$$\sigma\Delta\tau \text{ rms} = \frac{1}{\sqrt{\frac{S}{2N_o B_L}}} \; \frac{R(0) - R(2\tau d)}{2R'(\tau_d)}$$

Where:

B_L = one sided loop noise bandwidth
N_o = one sided noise power density
S = average signal power
τ_d = jitter time displacement in bits

The bandlimited autocorrelation junction used above can be shown to be given by:

$$R(\tau_d) = \frac{1}{\pi}\left[\frac{2}{B}\cos B\tau_d\,(\cos B - 1) - 2\tau_\tau\ \mathrm{Si}\,(B\tau_d) + (1 + \tau_d)\,\mathrm{Si}\,[B\,(1 + \tau_d)] + (1 - \tau_d)\,\mathrm{Si}\,[B\,(1 - \tau_d)]\right]$$

Where:

$B = 2\pi f_r T$
f_r = one sided bandwidth of the receiver filter
T = PN code bit width

The correlation function and the slope of the correlation function are shown in figure 4-8 and figure 4-9 as a function of jitter time displacement τ_d and frequency-time product B.

From figures 4-8 and 4-9, the rms range tracking error vs the carrier to noise ratio for a loop B_L of 4 Hz, a f_r = 1.5 MHz and for two way ranging is shown in figure 4-10 for a time jitter displacement = .5 bits.

The chip rates of 34.133 kcps, 102.4 kcps and 1024 kcps are considered for both the uplink and downlink. The rms range error is also shown for an uplink of 102.4 kcps and a downlink of 1024 kcps.

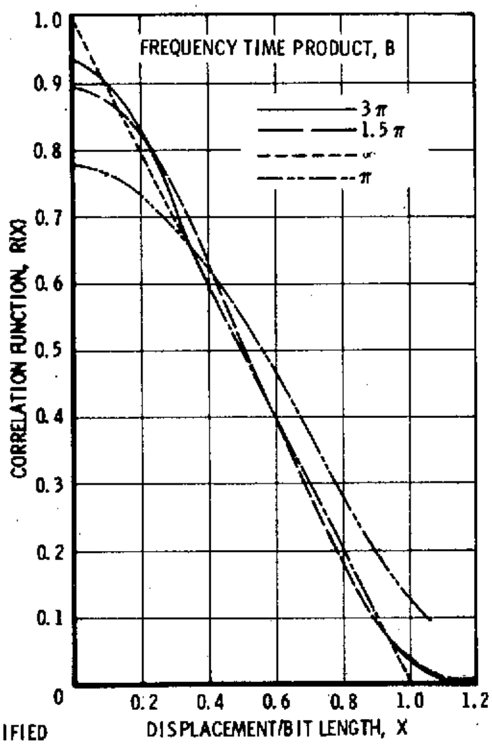

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Figure 4-8. Correlation Function R (τ_d)

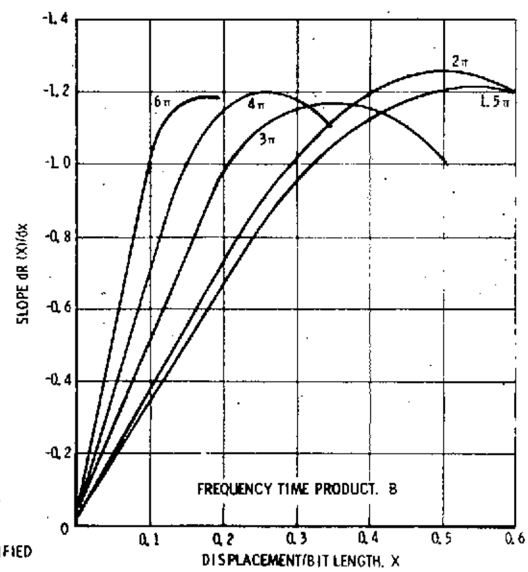

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Figure 4-9. Slope of Correlation R' (τ_d)

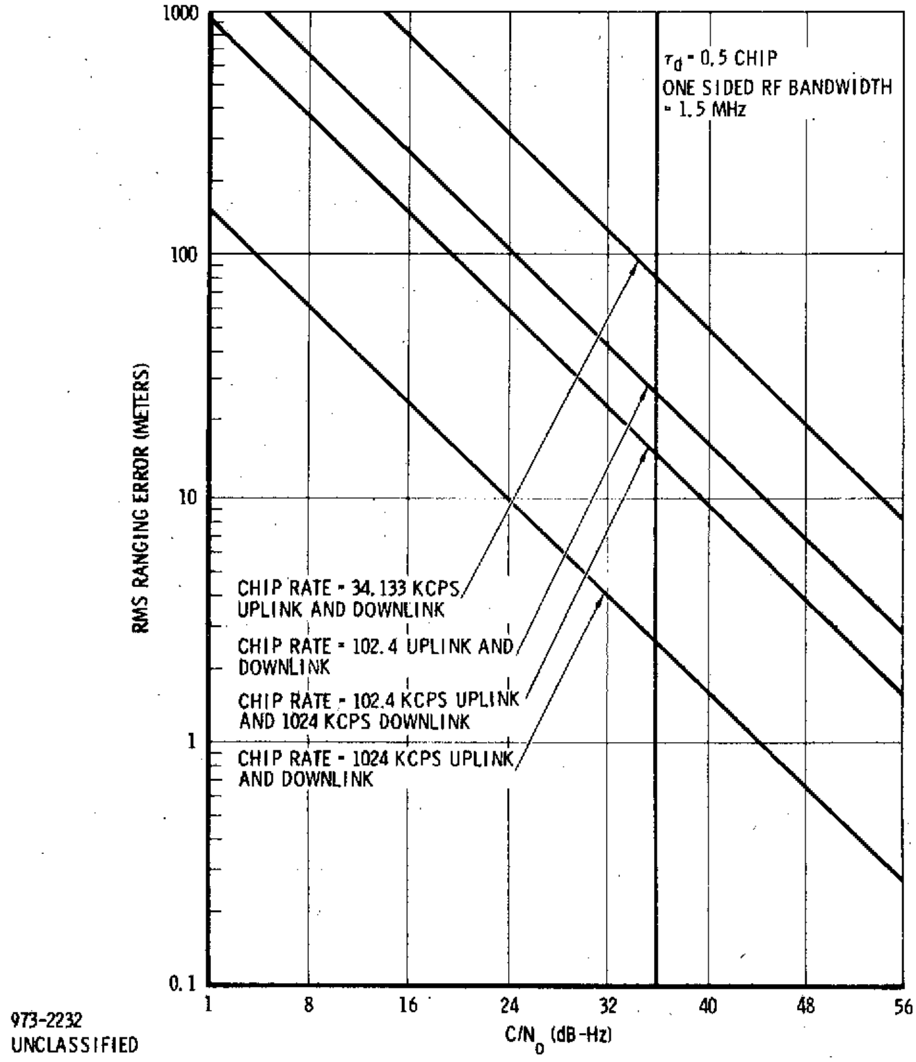

Figure 4-10. Delay Lock Loop rms range tracking error vs carrier to noise ratio for two way ranging

4.2.3 RANGE RATE TRACKING

The theoretical range rate accuracy by carrier doppler is dependent on the loop signal-to-noise ratio neglecting dynamics. From a known carrier frequency, the doppler measurement is made by taking the ratio of the total phase increment (relative to a reference oscillator) divided by the time interval τ. The phase is, therefore, measured first at the start of the interval and then at the end. If $B_L \cdot \tau >> 1$, the two phase measurements have independent errors and the rms range rate tracking error can be written as:

$$\sigma \Delta v = \frac{C}{2\pi f_o} \sqrt{\frac{2\, N_o B_L}{S}}; \quad B_L \tau >> 1$$

Where:

C = velocity of propagation

f_o = carrier frequency

τ = interval time

For a carrier frequency of 137 MHz and an interval time of 1 sec, a relationship of rms error vs carrier-to-noise ratio can be obtained for a loop B_L of 40 Hz. This is shown in figure 4-11.

Note that this is one-way range-rate error. The actual two-way performance would have to account for the cumulative carrier noise filter for both the forward and return links divided by two.

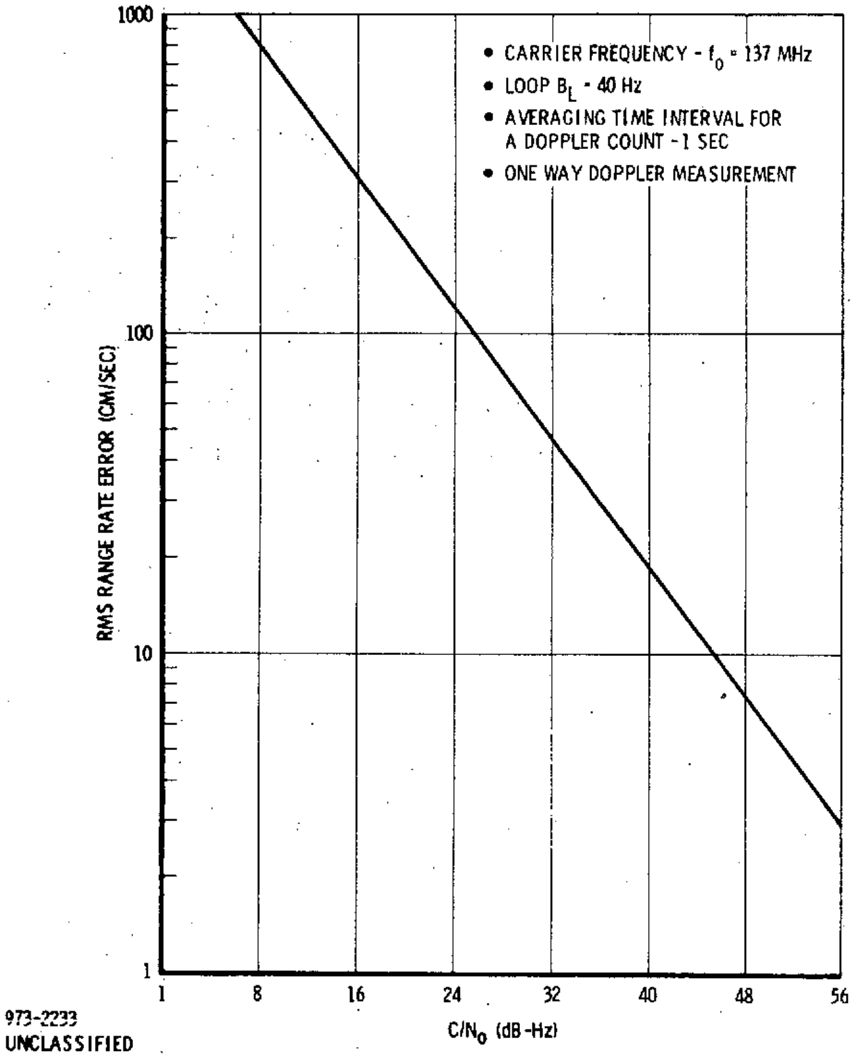

Figure 4-11. RMS range rate error for carrier doppler vs carrier to noise ratio