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ABSTRACT

This document is the Fifth Quarterly Report on NASA Contract No. NAS 8-11050, Special AROD System Studies. It summarizes the results of the work performed on the Wideband Modulation Techniques task since its inception and the work performed on the Phase-Lock Loop effort during the quarter ending December 31, 1964.

Section 1 (Wideband Modulation Techniques) presents the results of a comparative analysis and evaluation of modulation techniques for AROD as they have been compiled up to the end of the quarter. Sidetone systems, pseudo-noise systems, and "hybrid" systems which combine periodic signals and pseudo-noise signals were analyzed. The principal conclusions that have evolved are:

a. Sidetone systems can approach the AROD accuracy objectives, but are too susceptible to CW interference.

b. A modulation system using only pseudo-noise signals cannot provide satisfactory tracking accuracy for AROD.

c. Hybrid systems can approach the AROD performance objectives and provide substantial interference rejection.

Section 2 (Phase-Locked Loop Advanced Circuit Investigations) reports on efforts to exploit memory, which characterizes Type II control systems, in minimizing reacquisition problems after signal fade.

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Section 1

WIDEBAND MODULATION TECHNIQUES

1.1 INTRODUCTION AND SUMMARY

This section presents the results of the work accomplished on the wideband modulation techniques task from its inception in October, 1964, through December, 1964. The material contained in this section is supplemented by four Appendixes in Volume 2 of this Quarterly Report.

The purposes of the wideband modulation techniques task are to analyze and evaluate modulation techniques for use in the AROD system and to recommend the technique(s) most suitable for implementation. Toward these ends, the study team conducted two initial parallel efforts. The first of these investigated systems employing only discrete sidetones to obtain the required range, range rate, and ambiguity resolution information; the second effort was directed toward systems employing pure pseudo-noise (PN) signals. At the conclusion of these two efforts, "hybrid" systems combining periodic signals and PN signals were analyzed. Block diagrams were generated for several promising hybrid systems, and competing approaches were evaluated; the recommendations resulting from the evaluation are contained in this section.

1.1.1 Summary

The investigation of systems employing only discrete sidetones concluded that these systems are capable of approaching the design objectives of the AROD system, but are very susceptible to CW interference. A detailed investigation of the inherent adaptivity of phase-locked loops indicates that the approach showing the most promise for achieving the AROD design objectives is linear amplification in the spacecraft receiver. Using this approach, phase-locked loop parameters were selected which provide well-balanced performance for the AROD signal parameters. These results are contained in (Tables 1-1 and 1-2), Sub-section 1.3.1.

The PN effort initially concentrated upon the use of a spectrum generated by modulating the rf carrier with a pure PN sequence. As indicated in Sub-section 1.3.2, the conclusion of this analysis is that the AROD accuracy objectives cannot be met with this type of spectrum. PN acquisition techniques and the associated acquisition times were also analyzed in this portion of the investigation.

The remainder of the modulation techniques investigation concentrated upon spectra which combine a pure PN sequence with a periodic signal from which more accurate ranging information can be derived. Several possible implementations for these hybrid systems are suggested in Sub-section 1.3.4. Advantages and disadvantages of each approach are discussed.

1.1.2 Conclusions and Recommendations

The conclusions that have been reached as a result of the modulation techniques investigation can be separated into two categories: (1) conclusions that are based upon detailed quantitative analyses, and (2) conclusions that are based upon qualitative assumptions concerning certain aspects of the AROD system. The latter conclusions are included at this time to indicate apparent trends, but it is strongly recommended that the modulation techniques investigation be continued in order to obtain the information required to substantiate the conclusions in these areas.

The strong conclusions which have resulted from the modulation investigations are:

a. A modulation system using only discrete sidetone components with a Fine Ranging Tone frequency of approximately 1 MHz can approach the AROD accuracy objectives, but such a system is too susceptible to CW interference.

b. A modulation system using only pseudo-noise (PN) sequences cannot provide satisfactory tracking accuracy for the AROD application; rather, a periodic component must be included in the modulation.

c. "Hybrid" systems combining periodic and PN signals can approach the AROD acquisition and tracking accuracy objectives and provide substantial interference rejection.

d. To approach the AROD objectives, the inherent adaptivity characteristics of phase-locked loops must be fully exploited; this requires essentially linear amplification (with no limiting) in the Vehicle Tracking Receiver (VTR). A set of loop parameters that has been balanced to provide a good tradeoff among the AROD design objectives is presented in Sub-section 1.3, Tables 1-1 and 1-2.

e. The slave PN generator in the VTR must be driven by the "clean" signal available at the output of the phase-locked loop tracking the periodic component in the hybrid spectrum.

f. In order to achieve good dynamic range, the PN sidelobes must be suppressed by filtering. Baseband (pre-modulation) filtering, combined with analog (angle) modulation, appears to be the preferable approach.

g. During initial acquisition, the first step following acquisition of the (doppler-wiped-out) carrier should be the acquisition of the PN sequence; only after this has been accomplished, should the doppler be restored to the signal.

h. The VTR should mix incoming signals with a local signal that essentially spreads an interfering sinusoid into a spectrum similar to that of pseudo-noise.

i. Proper selection of the harmonic relationship between the PN rate and the periodic component (FRT) can result in a power density spectrum that vanishes in the vicinity of the carrier.

Beyond the preceding conclusions, the principal recommendation that must be made, is for further analysis and experimentation to provide the supplementary information necessary for the selection of the final implementation. Of special importance are studies in the following areas:

a. Analyses of the spectra and errors resulting from filtering PN sequences when various modulation methods and parameters are employed.

b. Determination of the interfering signal environment and the multipath characterization applicable to an operational AROD system.

c. Optimization of the modulation parameters and allocation of power to the various spectral components; computer simulation of those aspects of the analysis which are not amenable to direct solution.

d. Evaluation of the utility of an on-board data processor for acquisition aiding and real-time signal dynamics compensation.

e. Optimization of the PN reacquisition techniques.

f. Determination of the suitability of the "unmatched index demodulator" and the Costas-type receiver for the AROD application.

g. Analysis of the effects of incorporating command data in the spectrum, and optimization of the command transmission and reception parameters.

h. Determination of the magnitude of the AM-to-PM conversion problem in the TWTs to be employed in AROD, and evaluation of the errors introduced by this conversion.

Several hybrid implementations appear to be satisfactory, with each being strong in some areas, and weak in others. The final selection of an implementation must reflect detailed equipment considerations and the results of analyses yet to be performed. Moreover, such factors as the magnitude of the potentially interfering signals and the severity of the multipath problem are not definitively known at this time. Consequently, a final decision regarding the implementation should be deferred as long as possible to permit the inclusion of additional information.

Since some design decisions might have to be made at this time, and since it would be useful to concentrate on a particular implementation for additional in-depth analysis and evaluation, it is suggested that the most promising implementation approach is the PN hybrid system discussed in Sub-sections 1.3.3.2 and 1.4.2. The basis for this suggestion are some relatively strong assumptions concerning the comparative importance of several factors in the overall AROD system environment. A version of the PN hybrid system which employs two (or three) phase-locked loops to track the (reconstituted) carrier and the periodic component (FRT) avoids the relatively unproven Costas-type and "unmatched index demodulator" receivers. The reacquisition difficulties with this system should be surmountable by employing sophisticated reacquisition procedures and probably will not be a significant detriment to system performance.

1.2 SCOPE OF THE STUDY

The purpose of the modulation techniques investigation is to analyze and evaluate wideband angle modulation techniques for use in the AROD system. The comparative evaluation is based on the following criteria:

a. Accuracy of the range and range rate measurements.

b. Interference resistance (including interchannel crosstalk, unintentional jamming, and spurious signals).*

c. Time required for signal acquisition and reacquisition.

d. Ease of implementation.

For each of the criteria, only those factors that influence, or are influenced by, the selection of the modulation technique, are to be considered. For example, the range and range-rate errors caused by the uncertainty of the vacuum velocity of light are not included in the evaluation process.

Most of the AROD system characteristics and specifications are included in the RFP for the AROD System Test Model Hardware.** The factors that are most important for the modulation techniques study (with a few modifications as directed by NASA's Contracting Officer's Technical Representative) are as follows:

Spacecraft System

Transmitted Frequency:	2276 MHz
Transmitter Output Power:	10 watts rms
Antenna Power Gain:	0 db
Receiver Noise Figure:	6 db
Frequency of Master Oscillator:	3.2 MHz
Down Link Bandwidth Allocation:	6-12 MHz

*An additional consideration is the resistance of each candidate modulation technique to multipath errors. However, the unavailability of quantitative information concerning the multipath parameters has permitted only a qualitative analysis to date.

**RFP No. 1-4-40-01283, NASA Marshall Space Flight Center, Huntsville, Alabama.

Ground Transponder System	
Transmitted Frequency:	1750 to 1850 MHz
Transmitter Output Power:	50 watts rms
Antenna Power Gain:	12 db
Receiver Noise Figure:	3 db
Up-Link Bandwidth Allocation:	14.4 MHz
Number of Channels:	4
Signal Dynamics	
Maximum Two-Way doppler*:	± 196 KHz
Maximum Two Way doppler Rate*:	± 6.3 KHz/sec
Dynamic Range:	70 db
Drop-out (for up to 2.5 sec):	90 db
Maximum Range:	4000 naut mi
Minimum Range:	90 naut mi
Performance Requirements	
Range Accuracy (3σ):	1.0 meter
Range Rate Accuracy (3σ):	0.03 m/sec
Acquisition Time:	2 seconds
Ambiguity Resolution:	2000 km

Several of these factors merit further discussion. For example, the bandwidth allocations indicated are to be considered as guidelines only. If the results of the modulation study indicate that a substantial advantage would result from utilizing wider bandwidths, such a recommendation should be made. The nominal arrangement for the four channels that are to be incorporated in the 14.4 MHz up-link bandwidth allocation assigns 3.2 MHz for each channel, three 400 KHz doppler guard bands between channels, and two 200 KHz guard bands at the edges of the total allocation. For the down-link bandwidth allocation, two alternatives are to be analyzed: 6 MHz and 12 MHz.

The maximum dynamic range of 70 db will only be encountered during launch; it allows for a range of 25 miles to one ground station, a range of approximately 250

*If a frequency of 2450 MHz were to be used.

miles to a second ground station, and a severe multipath cancellation effect at the second ground antenna. It is not necessary to maintain the full system accuracy in this situation; however, the system should be able to acquire the second ground station in the allotted time of 2 seconds, and must be able to track the weaker signals. The fundamental dynamic range due to range differences between two stations is approximately 20 db. As interference resistance is one of the evaluation criteria, it is necessary to determine the degradation in measurement accuracy that results from interchannel crosstalk when this maximum dynamic range and multipath degradation are considered.

The accuracy requirements stated are to be considered, for the purposes of the modulation study, as design objectives for spacecraft ranges of 90 to 4000 nautical miles. The ability of the candidate modulation techniques to achieve these objectives is, of course, one of the most important evaluation criteria.

The acquisition time requirement assumes that the VHF link is available to aid in the acquisition process and that the ground station has already been switched to the STANDBY condition. The 2 seconds is the time from the command to switch the ground station to the ON condition, until the new station has been acquired, and the Vehicle Tracking Receiver is tracking the station's signals with full accuracy. The ability of the vehicle receiver to rapidly reacquire the signals after the dropout of 90 db (during firing of a retro rocket, for example) must be considered in the evaluation process.

The selection of a modulation technique should also reflect the possibility that the system requirements may be changed at some time in the future. It would be desirable to implement a system which would be readily expandable to meet any increased requirement.

No satisfactory "model" for the interfering signals that might be encountered is available. The principal problem will be strong CW signals; radar pulses, however, may also be a serious problem.

1.3 MODULATION PARAMETER ANALYSIS

In the modulation study, the approach to the problem was to consider two different methods of range modulation for the tracking first. The two range modulations considered were: (1) sidetone ranging, and (2) pseudo-noise ranging.

The sidetone ranging is one in which harmonically related tones are combined, and coherently angle modulate the carrier. The received signal from the ground transponder is coherently tracked to obtain range and range rate. The received carrier is used to derive the range rate information; the fine range is obtained from the phase delay of the fine ranging tone. Lower harmonic tones are coherently tracked to resolve the range ambiguity of the fine ranging tone phase delay measurement, the pure pseudo-noise range modulation is one in which the output of the pseudo-noise generator angle modulates the carrier. The vehicle tracking receiver must coherently track the received carrier to extract the range rate information and must synchronize a slave pseudo-noise generator with the received pseudo-noise sequence. The fine range information is obtained from the clock driving the slave pseudo-noise generator.

After analyzing the sidetone modulation and the pure pseudo-noise modulation, two promising hybrid modulations were investigated. These two hybrid range modulations are referred to as the sidetone hybrid and the PN hybrid. The sidetone hybrid is one in which the range modulation consists of a discrete tone plus a pseudo-noise sequence modulo-two added with a clock. The pseudo-noise bit rate and the clock are a sub-harmonic of the tone. The pseudo-noise sequence is used for ambiguity resolution and the tone for fine ranging. In the PN hybrid, a pseudo-noise sequence is modulo-two added to a clock. The vehicle tracking receiver separates the clock and the pseudo-noise and then tracks the clock for fine ranging. The pseudo-noise is utilized for range ambiguity resolution.

The parameters were analyzed for each of the above range modulations to determine if the design goal objectives for the AROD system were met. For the purposes of the analysis, the pseudo-noise modulation was considered to bi-phase modulate the carrier. The reason for this is that the analysis is simplified. The use of baseband filtering or rf filtering will complicate the analysis and further study in this area is recommended.

1.3.1 Sidetone System

The sidetone range and range-rate system considered for possible AROD use was the typical sidetone system in which harmonically-related sidetones are combined and then angle modulate a carrier*. A typical block diagram of the sidetone system is shown in Figure 1-1. A typical receiver for the sidetone system is shown in Figure 1-2. The carrier tracking phase-locked loop is used to obtain the range rate information and the highest frequency tone phase-lock loop is used to measure fine range.** Lower frequency tones are used for ambiguity resolution.

In the transmitter of the vehicle, the sidetones are generated coherently, with respect to the master clock. The sidetone combiner weights each tone so that the power transmitted in the rf component for that tone is of the desired amount. The tones then angle modulate the S-band carrier. The modulator output drives the TWT power amplifier, providing an output of 10 watts at the vehicle. The ground station transponder acquires and tracks the received signal from the vehicle, translates the signal to a new frequency, and transmits the signal at a power of 50 watts. The signal received in the vehicle is coherently detected by use of phase-locked loops. The output of the oscillator of the carrier tracking phase-locked loop is used to determine the velocity by determining the doppler on the signal. The output of the fine range tone phase-locked loop is used to compare zero crossings with a reference oscillator to determine fine range. The ambiguity tones are used to determine coarse range.

Since the sidetone system relies on the phase-locked loop for the tracking information, the design of the carrier and fine-range loops were analyzed to determine the parameters needed to meet the AROD design goals, and to determine if a sidetone system could meet the design objectives. Also, since the pseudo noise tracking systems may utilize phase-lock loops, the information obtained can also be applied to the loops in the pseudo-noise tracking system.

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*RFP-1-4-40-01283, "AROD System Test Model Hardware Development," describes a system which may be a sidetone system. Another report describing a sidetone system is, "Goddard Range and Range Rate System Design Evaluation Report," Contract No. NAS 5-1926, Motorola, Inc., Report No. W 2719-2-1, 16 March 1962

**For an extensive bibliography on Phase-Lock Loops, consult Martin, B. D., "The Pioneer IV Lunar Probe: A Minimum Power FM/PM System Design," Technical report No. 32-215, Jet Propulsion Laboratory, March 15, 1962, NASA Contract No. NAS 7-100

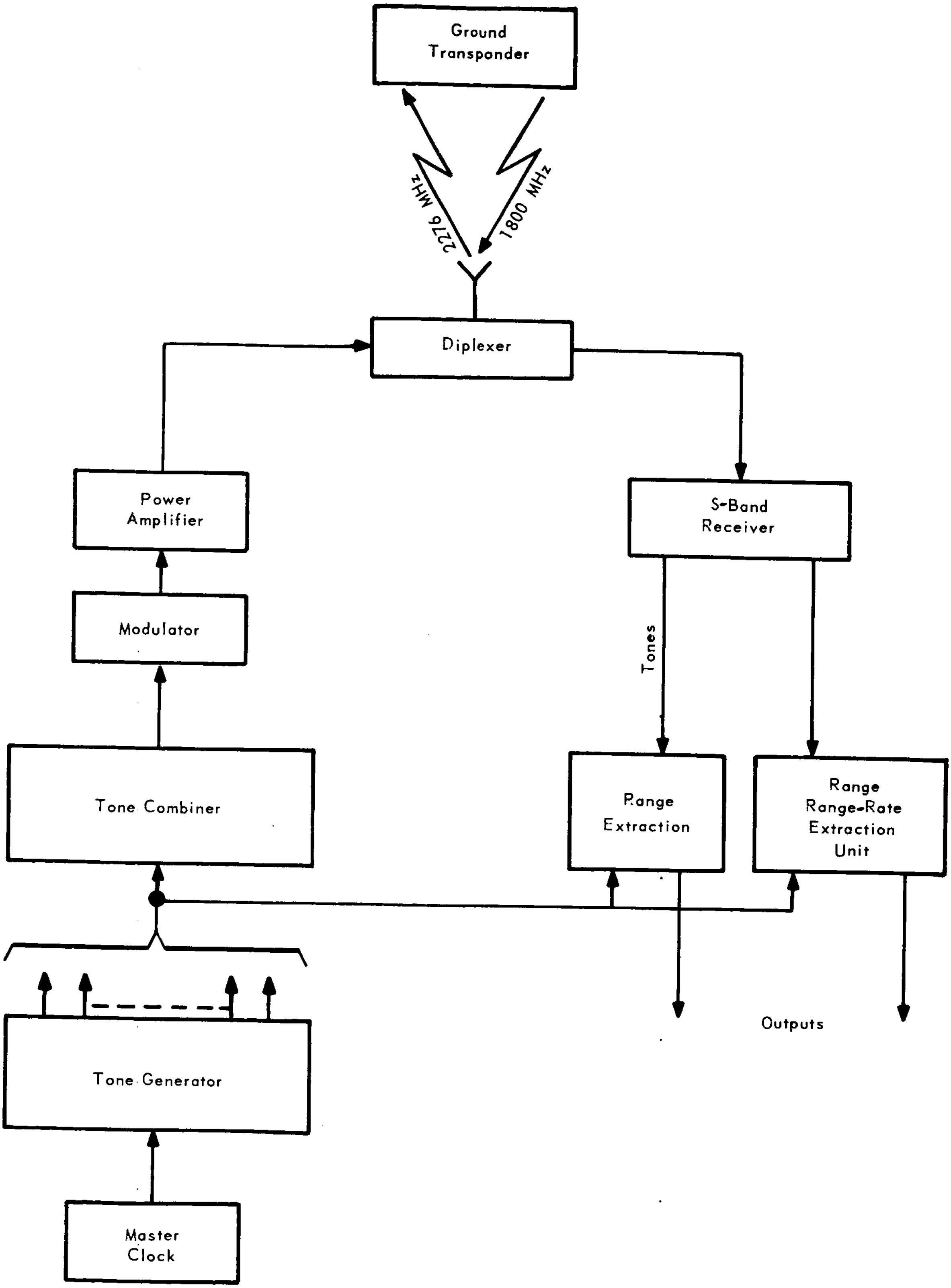

Figure 1-1. Sidetone Tracking Receiver

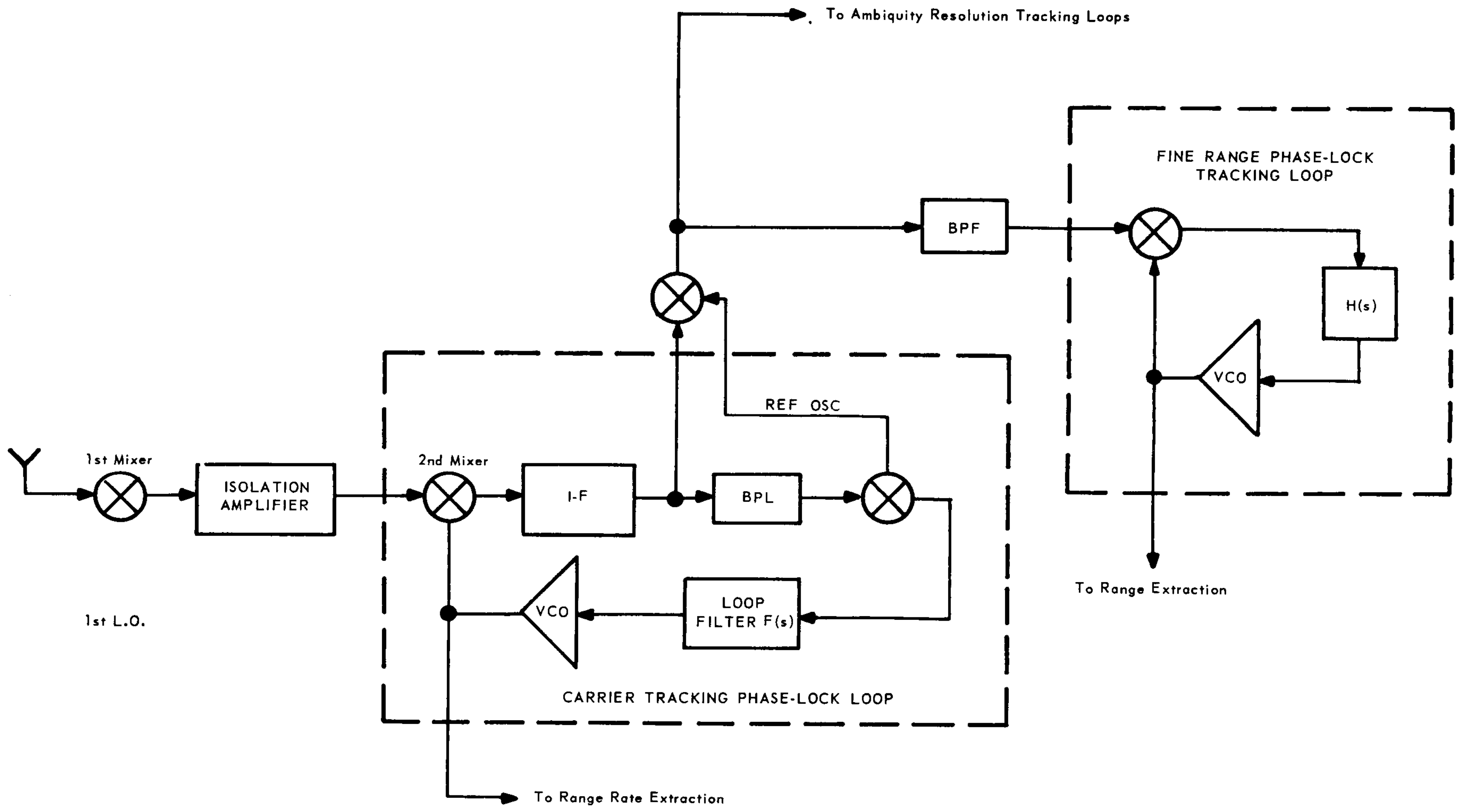

Figure 1-2. Block Diagram of Range and Range-Rate Tracking System

1.3.1.1 Thermal Noise Errors

The loop natural frequency of the carrier loop and tone loops must be chosen to meet the accuracy requirements at the minimum signal level. As shown in Appendix B (Volume 2), the phase variance at the output of the phase-locked loop which is preceded by a band pass limiter is given by:

$$\sigma_p^2 = \frac{\omega_{n_o} \Phi}{2 \zeta_o} \left[\frac{\alpha}{\alpha_o} 2 \zeta_o^2 + \frac{1}{2} \right]$$

where

ω_{n_o} = loop natural frequency at design point

ζ_o = loop damping factor at design point

α_o = limiter suppression factor at design point signal level*

α = limiter suppression factor above design point

Φ = noise power spectral density to signal power ratio

σ_p^2 = phase variance of VCO output

The loop damping factor, ζ_o is chosen equal to $1/\sqrt{2}$ for reasons of acquisition time. (Refer to Sub-section 1.3.1.4.)

In the case in which there is no limiter or AGC preceding the loop, the amplitude of the input signal fed to the phase detector is directly proportional to the received signal amplitude, and the phase variance of the VCO output is given by

$$\sigma_p^2 = \frac{\omega_{n_o} \Phi}{2 \zeta_o} \left[\frac{A}{A_o} 2 \zeta_o^2 + \frac{1}{2} \right]$$

where A_o = signal amplitude at design point

A = signal amplitude above design point

*See Sub-section 1.3.2.3 for a discussion of suppression factor.

The design point (match point) is chosen as the point at which the signal level is at its minimum value. The phase variance of the VCO at the design point is the same for both cases and, for $\zeta_o = \frac{1}{\sqrt{2}}$, is given by

$$\sigma_p^2 = \frac{3\,\omega_{n_o}}{2\sqrt{2}}\,\Phi_o$$

where Φ_o = noise power spectral density-to-signal power at design point.

1.3.1.1.1 VTR Carrier Loop

From the system requirements given in Sub-section 1.2, the range rate accuracy objective is 0.01 m/sec standard deviation. The range rate accuracy is related to the phase accuracy of the VCO output by

$$\sigma_v^2 = \frac{\lambda_t^2\,\sigma_p^2}{(4\pi T)^2}$$

where T = averaging time of range rate extraction unit

σ_v^2 = variance of range rate measurement

λ_t = wavelength of transmitted wave.

Using the (normalized) value of Φ_o of $\frac{1}{2000}$ second, as calculated in Appendix A (Volume 2), for the noise power spectral density to signal power ratio, and assuming an averaging time of T = 0.1 second and and a 2000 GC carrier frequency, the constraint on the loop natural frequency, in order to meet the range rate accuracy objective at maximum range, is:

$$\omega_{n_o} \leq 13.3 \text{ rad/sec per watt of spacecraft transmitter power}$$

Thus, for 5 watts of transmitter power, a loop natural frequency of

$$\omega_{n_o} = 76.5 \text{ rad/sec}$$

will give a range rate accuracy of

$$\sigma_v = 0.01 \text{ m/sec}$$

A loop natural frequency of 76.5 rad/sec meets the range rate accuracy requirements; however, as will be seen later, when a limiter precedes the loop, it will not be capable of tracking the high Doppler rate encountered at minimum range.

1.3.1.1.2 VTR Fine Range Loop

The range accuracy objective, as given in Sub-section 1.2, is that the standard deviation be less than 1/3 meter. The range variance is related to the phase variance of the fine range tone VCO output by:

$$\sigma_R^2 = \left(\frac{\lambda}{4\pi}\right)^2 \sigma_p^2$$

where

λ = wavelength of fine range tone.

σ_R^2 = range variance

Assuming a fine range tone frequency of 1 MHz, it is found that to meet the range accuracy requirements at the design point (maximum range), the loop natural frequency must be constrained by:

$\omega_{n_o} \leq 0.8$ rad/sec per watt of spacecraft transmitter power

Assuming that 3 watts of power is allotted to the fine range tone, it is found that the pull-in time of the phase-lock loop will be approximately 82 seconds when the doppler offset uncertainty is reduced 95 percent by velocity aiding from the carrier loop. However, it will also be seen later, that to track the signal, a much higher loop natural frequency is needed at the minimum ranges. A tradeoff must be made of range accuracy due to thermal noise, range accuracy due to doppler rate, the pull in time, steady state phase errors, and lag corner time constant.

1.3.1.1.3 Ground Carrier Tracking Loop

The carrier tracking loop at the ground transponder must be capable of operating with accuracy equivalent (or higher) to the vehicle loop. The design of the ground loops is somewhat easier, however, since the dynamics of the signal are less than at the vehicle, and the signal to noise ratio is better.

The noise power spectral density to signal power ratio at the ground receiver (Appendix A, Volume 2) is $\Phi = \frac{1}{2600}$ seconds per watt^{-1} of transmitted power. The loop natural frequency of the ground carrier loop would have to meet the requirement:

$$\omega_{n_o} \leq 99.5 \text{ rad/sec}$$

for a 5-watt carrier power to have the phase variance equal the vehicle phase variance.

1.3.1.1.4 Ground Fine Range Loop

In order for the ground fine-range loop-phase variance to be equal to that in the vehicle, the ground loop natural frequency can be 1.3 times as large as that in the vehicle.

1.3.1.2 Steady-State Errors

The steady-state phase errors in the tracking loops can be divided into two types:

1. Static phase error—This error occurs in a phase-locked loop in which the loop fitler is a lag filter and is given by

$$\Phi_{ss} = \frac{\Omega}{K}$$

where

Ω = doppler offset from VCO center frequency in rad/sec

K = loop gain in sec^{-1}

2. Dynamic phase error—This error occurs in a phase-locked loop with a proportional plus integral control filter for the loop filter, for a doppler rate on the input signal. Since the lag filter approximates the proportional plus integral control filter, the same formula is used for both cases, and is given by:

$$\Phi_{ss} = \frac{D}{\omega_n^2}$$

where D = doppler rate rad/sec^2

ω_n = loop natural frequency in rad/sec

1.3.1.2.1 VTR Carrier Tracking Loop

The only requirement for steady state phase error in the carrier tracking loops is that the phase error be kept at a value such that the phase-locked loop operates in its linear range and remains in lock. The upper limit of phase error for the loop to remain locked is usually taken as 0.5 radian. If we assume that a maximum steady state error allowed is $\pi/8$ radians, then at maximum range, the gain constant of the phase-locked loop is constrained by

$$K \geq \frac{(2\pi)(160 \times 10^3)}{\pi/8}$$

$$K \geq 2.56 \times 10^6 \ \text{sec}^{-1}$$

However, at the maximum range (the design point), the doppler offset for a 2000 mile, circular orbit, will be approximately half the maximum offset, or 80 KHz. Thus, at the design point

$$K \geq 1.28 \times 10^6$$

This value of loop gain is easily attained; however, the lag corner time constant of the lag filter is directly proportional to the loop gain and inversely proportional to the square of the loop natural frequency. That is:

$$T_1 = \frac{K}{\omega_{n_o}^{2}}$$

where

T_1 = lag corner time constant

K = loop gain at design point

ω_{n_o} = loop natural frequency at design point.

The value of the lag corner time constant may be difficult, if not impossible to implement for certain loop bandwidths; thus, a tradeoff may be necessary between the steady state phase error and phase-noise errors in order to achieve a lag corner time constant which can be implemented.

The maximum dynamic steady state phase error occurs at the minimum range where the maximum doppler rate occurs. In order to keep the upper bound of this steady state error at $\pi/8$ radians, the loop natural frequency must meet the requirement

$$\omega_n^2 \geq \frac{D}{\phi_{ss}} = \frac{(2\pi)(5130)}{\pi/8}$$

$$\omega_n^2 \geq 101,000$$

$$\omega_n \geq 320 \text{ rad/sec}$$

1.3.1.2.2 Ground Carrier Tracking Loop

In the ground station, the same upper limit for phase errors are imposed as in the vehicle; however, the ground station signal dynamics are approximately one half of the vehicle signal dynamics. For an upper bound of $\pi/8$, for either the static phase error or dynamic phase error, it is found that the loop gain and loop natural frequency at the design point must meet the following requirements:

$$K \geq 0.728 \times 10^6$$

$$\omega_n \geq 208 \text{ rad/sec}$$

1.3.1.2.3 VTR Fine Range Tracking Loop

In the range loop, a steady state phase error will result in an error in the range measurement; therefore, the range loop has a stringent requirement than the carrier loop in that the phase error for range accuracy is much less than that required to maintain lock. The gain constant of the loop not only affects the static steady state phase error, but will also determine the lag corner for a particular loop natural frequency. Lag filters with large lag corner time constants become impossible to implement with passive discrete components due to circuit leakage resistances (eventually limited by capacitor quality). Even with an active lag filter utilizing operational amplifiers, a lag corner time constant of 200 seconds is difficult to attain. Assume a lag corner time constant of 200 seconds; the loop natural frequency and gain are constrained by:

$$\frac{K}{\omega_n^2} < T_1$$

where

K = loop gain, sec^{-1}

ω_n = loop natural frequency rad/sec

T_1 = lag corner time constant, sec.

Assuming a 1 MHz fine ranging tone and the doppler offset at maximum range of 40 cps, then to keep the static phase error at 1/3 m requires that

$$K > 20000$$

At the minimum range, maximum signal level, the static phase error will be negligible, since the doppler is approximately zero at this point and the gain of the loop has increased. (The variation of loop gain is discussed in Sub-section 1.3.1.3). However, the dynamic phase error will be a maximum at this point due to the doppler rate. In order to keep this error below 1/3 m for a 1 MHz tone, the phase error must be less than

$$\phi_{ss} \leq \frac{4\pi \Delta r}{\lambda} = (2\pi)\left(\frac{1}{3}\right)\left(\frac{1}{150}\right)$$

$$\phi_{ss} \leq 0.01397 \text{ rad.}$$

The maximum doppler rate for the 1 MHz tone will be $D = 2.57\ c/s^2$. The loop natural frequency of the tone loop must meet the requirement

$$\omega_n^2 \geq \frac{D}{\phi_{ss}} = \frac{(2\pi)(2.57)}{0.01397}$$

$$\omega_n^2 \geq 1160$$

$$\omega_n \geq 34.1 \text{ rad/sec}$$

1.3.1.2.4 Ground Fine Range Tracking Loop

Since the signal dynamics for the ground range loop are one-half that of the vehicle loop, and the signal power-to-noise power spectral density is somewhat higher, the problem is somewhat simpler. In Sub-section 1.3.1.1, it was seen that for the same phase-noise error in the ground loop as in the vehicle loop, the loop natural frequency of the ground could be 1.3 times as large as the vehicle loop natural frequency. This means that the gain constant of the ground loop could be 1.69 larger for a lag corner time constant of 200 seconds. Since the doppler at design point is half that in the spacecraft, the loop gain is constrained by

$$K > 10^4$$

If we require the same upper bound on dynamic steady state phase error on the ground as in the vehicle, then the ground loop natural frequency must meet the requirement

$$\omega_n \geq 24.1 \text{ rad/sec}$$

1.3.1.3 Adaptability

In Sub-sections 1.3.1.1 and 1.3.1.2, bounds on the loop natural frequencies were determined so that the phase-noise error and static phase errors were kept below the design objectives. For the vehicle carrier tracking loop, it was seen that for 5 watts of carrier power, the loop natural frequency must be ≤ 76.5 rad/sec in order to meet the range rate accuracy objective of 0.01 m/sec at the maximum range. It was also found in Sub-section 1.3.1.2, that the loop natural frequency must be ≥ 320 rad/sec at the minimum range for the loop to track. Fortunately, the apparent contradiction between these requirements can be reduced because of the inherent adaptivity of a phase-locked loop.

Due to the variation of the phase detector gain with the amplitude of the input signal, the gain of the phase-locked loop varies with the amplitude of the signal. If the input signal to the loop has an amplitude which is directly proportional to the received signal amplitude, then this will be referred to as the linear receiver. If the phase-locked loop contains a bandpass limiter preceding the phase detector, then this receiver will be referred to as the limiter receiver. In the limiter receiver, the gain of the loop will be directly proportional to the suppression factor of the limiter. The suppression factor (α) of the limiter is related to the noise power-to-signal power ratio at the output of the limiter by:

$$\alpha = \frac{1}{\sqrt{1 + \left(\frac{N}{S}\right)_{OUT}}}$$

where

α = suppression factor of limiter

$\left(\frac{N}{S}\right)_{OUT}$ = noise power to signal power ratio at output of limiter

Davenport[2] has shown that the noise-to-signal power ratio at the output of the limiter is $4/\pi$ times the noise-to-signal power at the input to the limiter. Thus, the suppression factor is:

$$\alpha = \frac{1}{\sqrt{1 + \frac{4}{\pi}\left(\frac{N}{S}\right)_{2B_N}}}$$

where

$2\ B_N$ = IF bandwidth.

The loop parameters are chosen for a particular value of the signal amplitude or limiter suppression factor. This point was called the design point or matched point. The value of the amplitude or suppression factor at the matched point will be denoted by A_o or α_o, respectively. The minimum signal level (maximum range) point will be chosen as the matched point. In Appendix B (Volume 2), it is shown that the loop natural frequency for the linear receiver is given by:

$$\omega_n = \sqrt{\frac{A}{A_o}}\ \omega_{n_o}$$

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[2]Davenport, W. B., "Signal-to-Noise Ratios in Bandpass Limiters" Journal of Applied Physics, Vol 24, pp 720-727, June 1963

where

ω_n = loop natural frequency

ω_{n_o} = loop natural frequency at match point

A = signal amplitude

A_o = signal amplitude at matched point

In the case of the limiter receiver, the loop natural frequency is given by:

$$\omega_n = \sqrt{\frac{\alpha}{\alpha_o}}\ \omega_{n_o}$$

In Figure 1-3, the amplitude ratio A/A_o, and the suppression factor ratio α/α_o for the carrier loops are plotted versus range. In calculating the suppression factor, α, it was assumed that the carrier power was 5 watts and that the bandpass limiter had a bandwidth of 0.5 MHz. The value of 0.5 MHz was chosen so that the carrier, with the maximum doppler rate, would pass through the filter; however an I-F carrier tracking loop would use a much narrower bandwidth. It is seen that the maximum value of A/A_o is 40, whereas the maximum value of α/α_o is 7.98. Thus, at minimum range (100 miles) the linear receiver will have a loop natural frequency 6.33 times the loop natural frequency at match point. The limiter receiver will have a loop natural frequency only 2.82 times that at match point. Thus, if we choose the loop natural frequency to be 76.5 rad/sec at the match point, the loop natural frequencies for the limiter receiver and linear receiver will be 216 rad/sec and 483 rad/sec, respectively. Hence, the limiter receiver would be unable to track the high doppler rates. A greater variation of suppression factor can be obtained by using a wider bandwidth. If we assume a 4 MHz bandwidth, then the signal power into the limiter is the full signal power from the 10 watt signal (except for high order sidebands). The largest value for $\sqrt{\alpha/\alpha_o}$ in this case is 3.92. The loop natural frequency at the minimum range would be 303 rad/sec (for 76.5 rad/sec at match point).

The better receiver or more optimum receiver for tracking range and range rate is the linear receiver. The optimum receiver derived by Jaffe and Rechtin[3] is

[3] Jaffe, R. and Rechtin, E. "Design and Performance of Phase-Lock Circuits Capable of Near-Optimum Performance over a Wide Range of Input Signal and Noise Levels", IRE Transactions - Information Theory, March 1955, pp 66-76

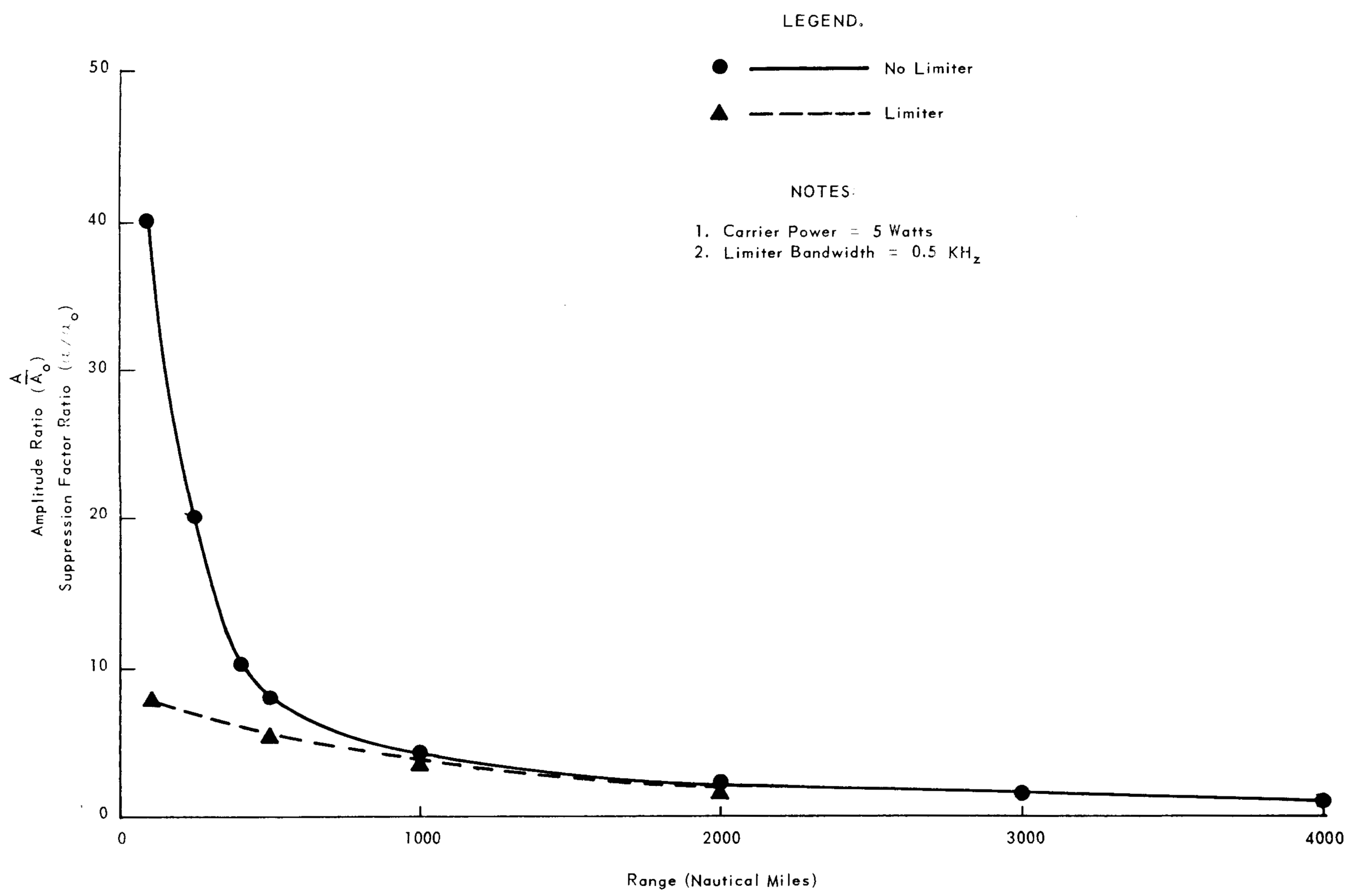

Figure 1-3. Amplitude Ratio and Suppression Factor Ratio Versus Range

optimized to meet a criterion for certain assumptions. It would be appropriate to review the assumptions and criteria for the Jaffe/Rechtin analysis. The assumptions for the Jaffe and Rechtin analysis are: (1) that the phase-lock loop is initially in a locked condition, (2) a frequency step is applied to the input to the phase-locked loop, and (3) the signal and noise are at a given power. Jaffe and Rechtin then found the loop filter that optimizes the loop in the sense that the phase noise output from the VCO plus the integral of the square transient error for the step input is minimized. The optimum filter was found to be the proportional, plus integral control filter. If a limiter is now placed in the input to the phase-locked loop, the loop will operate at near optimum for various signal and noise inputs. However, the optimum in this case is that the phase noise, plus integral square error, is minimized.

A receiver which minimizes transient error to a step in frequency is not optimum for a tracking application where steps in frequency are never applied. A more appropriate criterion for a tracking loop might be chosen as that loop which tracks the signal over the signal dynamics and has the greatest accuracy over the dynamics of the signal. Using this criterion, we see that the linear receiver is superior to the limiter receiver.

The implementation of a linear receiver for the AROD dynamic range of tracking from 4000 miles to 100 miles, 32 db does not present any problems. For larger signal levels than the level at 100 miles, a delayed AGC may be utilized.

1.3.1.4 Acquisition Time and Pull-In Ranges

Thus far, there has been no consideration given to the acquisition problem in the selection of the loop parameters. The selection of loop parameters has been made to meet accuracy requirements and to assure tracking over the dynamics of the signal. In addition to tracking within a prescribed accuracy, the total acquisition time of the spacecraft receiver must be less than 2 seconds from the time the "Standby to ON" command is given to the ground station. In order to meet the 2 second acquisition time requirement, the phase-lock loops must be able to acquire in a fraction of this time.

An approximate pull-in time has been given by Viterbi[4] to be

$$t_{pi} = \frac{\Omega^2}{2\zeta\omega_n^3}$$

where

Ω = doppler offset from VCO center frequency

ω_n = loop damping constant

ζ = loop natural frequency

1.3.1.4.1 VTR Carrier Loop

The doppler offset of the carrier at maximum range will be approximately 80 KHz. It is assumed that the ground station can reduce the doppler uncertainty to 1/500 of this value. At maximum range, for a pull-in time of 1 second for the carrier loop natural frequency must meet the requirement

$$\omega_n \geq \sqrt[3]{\frac{\Omega^2}{2\zeta t_{pi}}}$$

$$\omega_n > \sqrt[3]{\frac{(2\pi \cdot 160)^2}{\sqrt{2}}} > 90 \text{ rad/sec}$$

1.3.1.4.2 VTR Fine Range Loop

At maximum range, the doppler offset for a 1 MHz fine-range tone will be approximately 40 cps. For a pull-in time of one second, the loop natural frequency must meet the requirement:

$$\omega_n > \sqrt[3]{\frac{(80\pi)^2}{\sqrt{2}}} \geq 35.4 \text{ rad/sec}$$

[4] Viterbi, A. J. "Acquisition and Tracking Behavior of Phase-Locked Loops" Jet Propulsion Laboratory, External Publication No. 73, ASTIA No. AD-234-166

The carrier tracking loop can be used for doppler correction of the fine range loop. The carrier loop should correct the doppler offset of the fine range tone loop to less than 5 percent of the offset. Using the doppler correction, the loop natural frequency required for a 1 second acquisition time is

$$\omega_n > 3\sqrt{\frac{(4\pi)^2}{\sqrt{2}}} > 4.8 \text{ rps}$$

The maximum doppler offset which a phase-locked loop will pull in lock is given by:

$$\Omega \le 2\,\omega_n\sqrt{\frac{\zeta K}{\omega_n} + 1}$$

where

ω_n = loop natural frequency rad/sec
K = loop gain constant $\sec^{-1}$
ζ = loop damping factor
Ω = pull-in range

If the gain constant is large compared with ω_n, the pull-in range is given by

$$\Omega \le 2\sqrt{\zeta\omega_n K}$$

Thus, the requirement for $\omega_n K$ is that

$$\omega_n K \ge \frac{\Omega^2}{4\zeta}$$

The carrier tracking loop $\omega_n K$ product at match point must meet the requirement (assuming doppler wipe-out correction)

$$\omega_n K \ge \frac{(160 \cdot 2\pi)^2}{2\sqrt{2}} \ge 356 \times 10^3$$

The fine range tone-loop $\omega_n K$ product at match point must meet the following requirement if doppler correction from the carrier loop is not used:

$$\omega_n K \geq \frac{(80\pi)^2}{2\sqrt{2}} \geq 22,250$$

If doppler correction from the carrier loop is used, the $\omega_n K$ product requirement is

$$\omega_n K \geq \frac{(4\pi)^2}{2\sqrt{2}} \geq 556$$

In the acquisition time and pull-in time calculations, the value of the loop damping factors, ζ, was chosen as $1/\sqrt{2}$. The value of $\zeta = 0.5$ can be shown to result in a minimum loop two-sided noise bandwidth. The value of $\zeta = 0.707$ is the value found for the Jaffe and Rechtin optimum filter. Frazier and Page[5] found that, for best acquisition probability, the loop damping actor should be ≥ 0.5. Therefore, in order to allow for variation in circuit parameters and signal levels, a value of 0.707 was chosen to keep the damping factor above 0.5 for worst conditions.

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[5] Frazier, J. P. and Page, J. "Phase-Lock loop Frequency Acquisition Study" IRE Transactions on Space Electronics and Telemetry, Sept, 1962, pp 210-227.

1.3.1.5 Tradeoffs and Conclusions

The conflicting requirements indicated in the previous sub-sections indicate the need for tradeoffs among the design objectives, and some compromises. In this sub-section, a representative set of parameters is suggested that will provide a system approaching the AROD design objectives.[6]

In Table 1-1, the design requirements dictated by the design objective are listed in the left-hand column and the compromise design parameters are listed in the right-hand column. It is assumed that the carrier frequency is 2 GC and that 5 watts of spacecraft transmitted power is allocated to the carrier. The carrier phase-lock loop parameter requirements dictated by the equations presented in the previous sub-sections are as follows: the phase noise error objective requires that the loop natural frequency at match point, ωn_o, $\leq$ 13.3 rad/sec per watt of transmitted power. The dynamic tracking requires a loop natural frequency be greater than 320 rad/sec at a 100 mile range. For dynamic tracking, it was assumed that the maximum allowable phase error would be $\pi/8$ radians. An acquisition time objective of one second requires $\omega n_o = 5620$ rad/sec for a doppler offset of 80 KHz. An offset of 80 KHz was assumed at the 4000 mile range rather than the maximum offset of 160 KHz, since calculations for a 2000 mile altitude circular orbit indicate that the doppler offset will only be 53 KHz at 5 degrees elevation. With a doppler wipeout correction on the ground it is reasonable to assume that the doppler offset uncertainty can be reduced to 1/500 of the 80 KHz value. The acquisition time objective of one second then, requires a $\omega n = 90$ rad/sec when the doppler wipeout is utilized. The pull-in objective with doppler wipeout requires that the product $\omega_n K < 356,000$. Implementation difficulties encountered with lag corner time constants greater than 200 seconds requires $\frac{K}{\omega_n^2} < 200$. The static tracking error objective requires that the loop gain at the match point be $K \geq 1.28 \times 10^6$.

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[6] Airborne Ranging and Orbit Determination Design Feasibility report, IBM Final Report, Vol 2, March 20, 1963, IBM Document No. TR-023-022 NASA Contract No. NAS 8-5098, p. D-10

Table 1-1

Carrier Tracking Loop, Parameter Tradeoff

(Carrier Frequency 2 Gc)

Requirements: Design Goal	Requirements: Constraint	Design Point Parameters
$\sigma_{\dot{r}} = 0.01$ m/sec $\phi_{ss_{\ddot{r}}} \leq \pi/8$ $\tau_{lag} \leq 200$	$\omega_{n_o} \leq 13.3$ rad/sec/watt $\omega_{n_{100}} > 320$ rad/sec $\frac{K}{\omega_n^2} \leq 200$	Let $\tau_{lag} = 200$ sec $\phi_{ss_{\dot{r}}} = \pi/8$ Carrier power = 5 watts $\omega_n = \sqrt{\frac{1.3 \times 10^6}{200}} = 80$ rad/sec $\sigma_{\dot{r}} \approx 0.01$ m/sec $\phi_{ss_{\ddot{r}}} = \frac{(2\pi)(5130)}{(80)^2\,(40)} = \pi/24$
$t_{pi} = 1$ sec	$\omega_{n_o} \geq \sqrt[3]{\frac{(2\pi.160)^2}{\sqrt{2}}}$ $\omega_{n_o} \geq 90$ rad/sec	$t_{pi} = \frac{(2\pi\,.\,160)^2}{\sqrt{2}\,(80)^3} = 1.5$ sec
Pull-in range $\Omega \geq 160$ cps	$\omega_n K \geq \frac{\Omega^2}{4\zeta}$ $\omega_n K \geq 356000$	Ω easily met
$\phi_{ss_{\dot{r}}} \leq \pi/8$	$K \geq \frac{(2\pi)(80)(10^3)}{\pi/8}$ $K \geq 1.28 \times 10^6$	$K = 1.3 \times 10^6$

As a reasonable compromise to these conflicting requirements, it is suggested that the loop gain at the match point be chosen as 1.3×10^6 sec^{-1}. The lag corner time constant restriction then requires that the loop natural frequency be 80 rad/sec. The phase-noise error at the match point will be approximately the design objective of 0.01 m/sec. The acquisition time is approximately 1.5 seconds (which may be satisfactory for the maximum range). Sweeping of the VCO could be used to reduce the acquisition time still more, if required. The pull-in range of the phase-lock loop is well above the doppler uncertainty. If a linear amplifier is used, the dynamic steady state phase error is $\pi/24$ at the range of 100 miles, and the loop is operating well within the linear region. The use of a hard limiter would result in a dynamic phase error of 0.6 radians at the range of 100 miles, and the loop would not operate in its linear region. It is therefore recommended that a linear receiver or soft limiting be used in the receiver.

If it is assumed that a fine ranging tone frequency of approximately 1 MHz is to be employed and 3 watts of spacecraft transmitter power are allocated to the fine ranging tone. The FRT phase-locked loop parameter requirements dictated by the equations presented in the previous sub-sections are as follows: The thermal noise error objective requires that the loop natural frequency at the match point be less than 2.4 rad/sec. The dynamic tracking error objective requires that the loop natural frequency at the minimum range of 100 miles be greater than 34 rad/sec. Allocating 1 second of acquisition time to the FRT phase-locked loop requires that the loop natural frequency at maximum range be greater than 34 rad/sec. The static tracking error objective requires that the loop gain be greater than 20,000 at maximum range. The pull-in range objective requires that the product of loop gain and loop natural frequency at maximum range be less than 22,250. Implementation difficulties encountered with a lag corner time constant greater than 200 seconds require that $\frac{K}{\omega_n^2}$ be less than 200 seconds.

As a reasonable compromise among these conflicting requirements, it is suggested that the loop natural frequency at maximum range be set equal to 4 rad/sec. This will result in a thermal noise error approximately 1.3 times that stated as the design objective; however, other sources of error (the uncertainty of the vacuum velocity of light, for example) will still dominate at maximum range. It should be noted that the phase-noise spectral density of the phase input to the fine ranging loop was taken to be $N_0/2S$ (noise power spectral density to twice the signal power ratio). This assumption is valid since the signal-to-noise ratio at the input to the fine ranging tone phase-lock loop will be 10 db, at maximum range, if a 600 cps bandpass filter is used at the input. With this value for loop natural frequency, the following system performance characteristics result. (See Table 1-2.) The loop natural frequency at minimum range, assuming linear amplification, is approximately 25 rad/sec; the resulting dynamic tracking error is slightly above the 0.3 meter objective, but calculations or calibrations can probably be employed to reduce this error if necessary. By employing doppler aiding in acquisition, to reduce the doppler uncertainty to 5 percent or less of its nominal value, the acquisition time is approximately 1.7 seconds (which may be satisfactory for maximum range) and the pull-in range requirement of approximately 40 cycles/second is easily met. By maintaining our lag corner time constant objective, we require that the loop gain at maximum range be equal to 3200; the steady state phase error resulting from this gain is $\frac{\pi}{40}$ radians, or 1.9 meters, which can be substantially reduced, if necessary, by the use of velocity aiding from the carrier loop or by calculations or corrections.

Table 1-2

Fine Range Tracking Loop, Parameter Tradeoff

(Fine Range Tone = 1 MHz)

Requirements		Design Point Parameters
Design Goal	Constraint	
$\sigma_r \leq 1/3$ m $\phi ss_{\ddot{r}} \leq \frac{1}{3}$ m $\tau_{lag} \leq 200$ sec	$\omega_{n_o} \leq 0.8$ rad/sec/watt $\omega_{n_{100}} \geq 34$ rad/sec $\frac{K}{\omega_n^2} \leq 200$	3 watt Tone power $\omega_{n_o} = 4$ rad/sec $\sigma_r = 0.43$ m $\phi_{ss_{\ddot{r}}}$ $\omega_{n_{100}} = 25$ rad/sec
$t_{pi} = 1$ sec	$\omega_n \geq \sqrt[3]{\frac{(80\pi)^2}{\sqrt{2}}}$ $\omega_n \geq 35.4$ rad/sec	$t_{pi} = 1.73$ sec when doppler aid is used.
Pull-in range $\Omega \geq 40$ cps	$\omega_n K > \frac{\Omega^2}{4\zeta} = \frac{(80\pi)^2}{2\sqrt{2}}$ $\omega_n K \geq 22250$	$\Omega =$ 30 cps, doppler aid will reduce uncertainty to 2 cps
$\phi_{ss_{\dot{r}}} \leq 0.3$ m	$K \geq \frac{(2\pi)(40)(300)}{(.3)(4\pi)}$ $K \geq 2.10^4$	$\phi_{ss_{\dot{r}}} = \frac{80\pi}{3200} = \frac{\pi}{40} = 1.87$ m $K = 200 \cdot 16 = 3200$

1.3.2 Pure PN Systems

The PN ranging systems discussed in this subsection use as a ranging signal, a carrier modulated by a maximal length sequence. This type of system will be referred to as a pure PN system as opposed to ranging systems in which the modulating signal will consist of a maximal length sequence combined with some periodic signal.

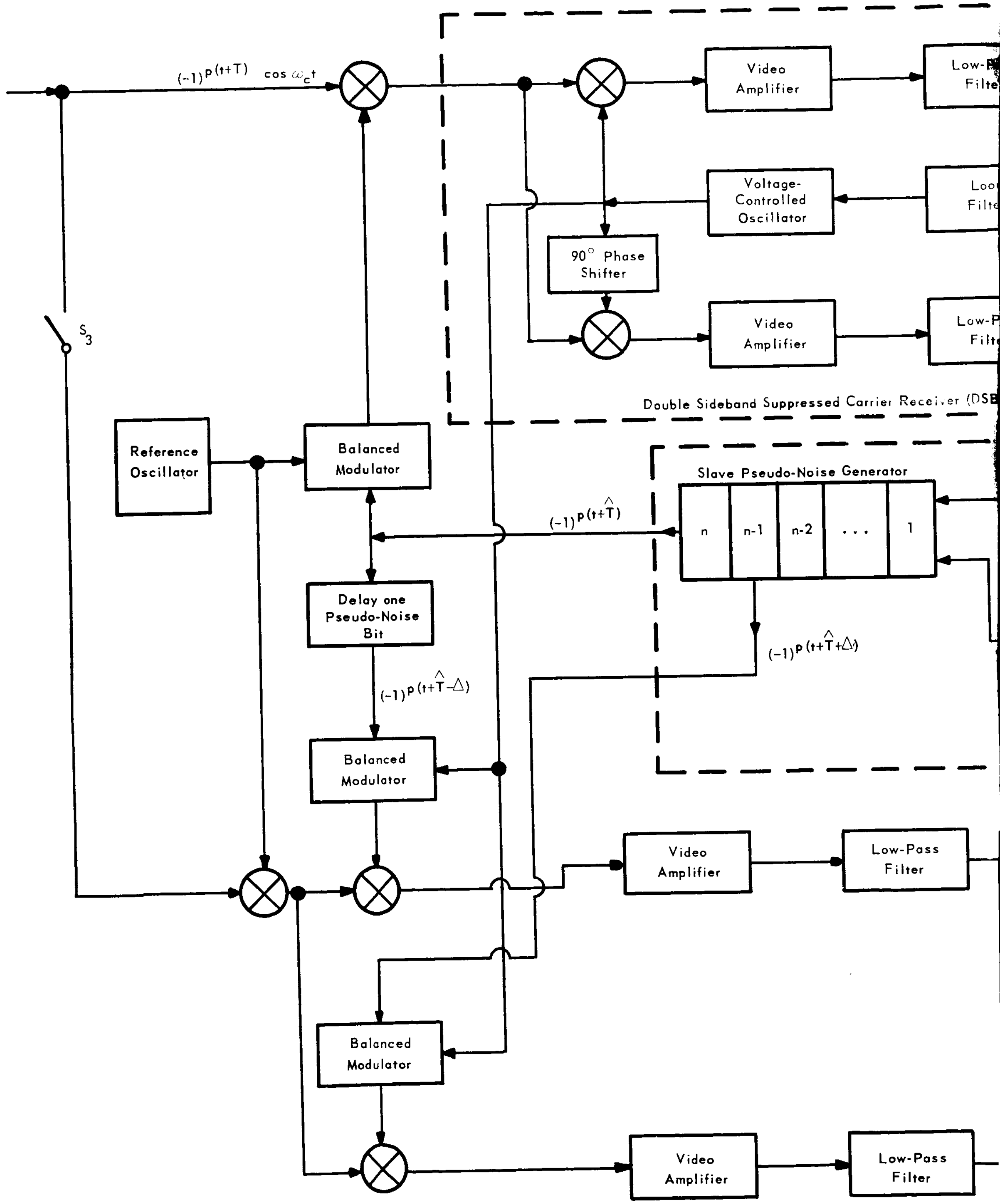
1.3.2.1 General System Description

A logical block diagram of a pseudo noise-tracking receiver is shown in Figure 1-4. The input signal is bi-phase modulated by a maximal length sequence sufficiently long, so there will be no repetition during the transit time over the ambiguity resolution range.

The DSBSC (double sideband suppressed carrier) receiver generates a VCO output coherent with the suppressed incoming carrier, and binary information detected at the output of the in-phase arm is used to aid the initial slave pseudo noise generator (PNG) acquisition process.

The acquisition program and control (APC) receives the event marker, which specifies the initial state of the slave PNG for acquisition, and sets up the slave PNG for acquisition. From this initial slave PNG state, the APC starts the process of bringing the slave PNG within a single bit of the incoming pseudo-noise modulation. At this stage of the acquisition process, the delay-lock discriminator (DLD) is switched in to maintain pseudo-noise (PN) bit synchronization of the slave PNG with the incoming PN modulation. There are alternatives to this approach, but the DLD is representative of methods to maintain bit synchronization.

The gross ranging function is performed by comparing the state of the reference PNG used to generate the original ranging transmission, and the slave PNG, which is synchronized to the PN modulation transponded from the ground. Fine ranging is obtained by comparing the phase relation of the clock used to drive the reference PNG, and the voltage controlled clock in the DLD. The mean-square delay error due to the input noise to the DLD determines the accuracy with which the fine ranging can be performed.



$(-1)^{p(t+T)}$ $\cos \omega_c t$
Video Amplifier
Voltage-Controlled Oscillator
$90°$ Phase Shifter
Video Amplifier
Double Sideband Suppressed Carrier Receiver
S_3
Reference Oscillator
Balanced Modulator
Slave Pseudo-Noise Generator
$(-1)^{p(t+\hat{T})}$
n
n-1
n-2
. . .
1
Delay one Pseudo-Noise Bit
$(-1)^{p(t+\hat{T}+\Delta)}$
$(-1)^{p(t+\hat{T}-\Delta)}$
Balanced Modulator
Video Amplifier
Low-Pass Filter
Balanced Modulator
Video Amplifier
Low-Pass Filter

1-33-1

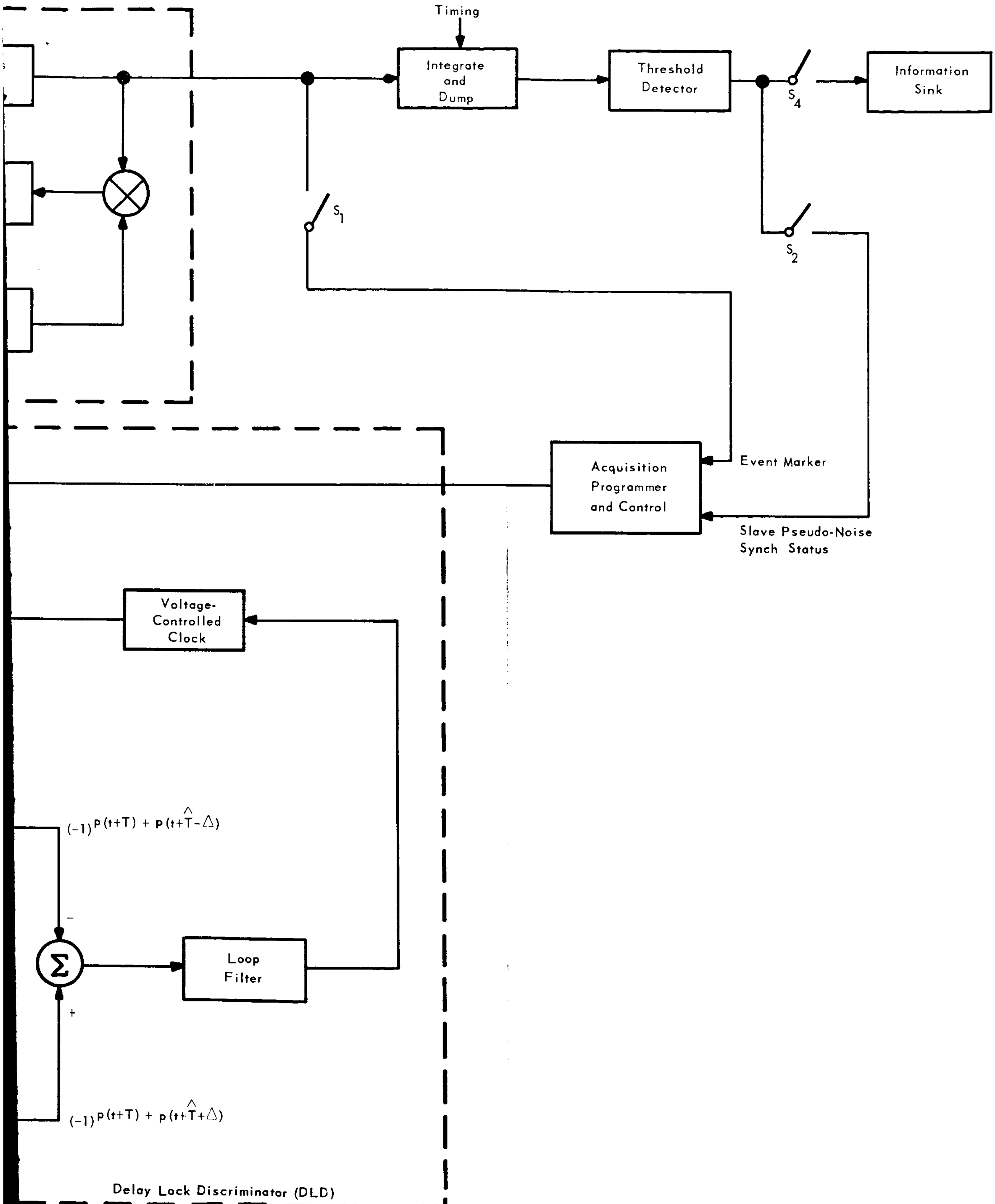


Figure 1-4. Pure Pseudo Noise-Ranging Receiver

The doppler rate is determined by comparing the DSBSC receiver VCO, which is generating a signal coherent with the incoming suppressed carrier, and the reference oscillator used to generate the initial ranging signal. The accuracy with which this operation can be performed depends upon the basic loop characteristics of the DSBSC receiver, and its tracking accuracy can be evaluated by using a linearized model of a phase-lock loop.

1.3.2.2 Acquisition Analysis

1.3.2.2.1 Procedural Analysis

The acquisition procedure involves the ground station and the vehicle as follows:

a. The vehicle receives a transmission, from the ground, in which the doppler offset has been estimated and removed before transmission. Switch S1 is closed, the others open, and an event marker bi-phase modulates the doppler wiped-out carrier to set the initial PNG state for the vehicle slave PNG acquisition.

b. The ground station inserts the bi-phase modulation on the doppler wiped-out carrier, using the PN derived from the PN ranging transmission from the vehicle. Switch S2 is closed, the others open, and the vehicle APC initiates the slave PNG acquisition procedure, which consists of generating a local replica sufficiently long in duration to see if the DSBSC receiver in-phase arm output exceeds a threshold level. If the slave PNG synchronization is not achieved, the PAC steps the slave PNG and determines again if slave PNG sync has been achieved, and continues this procedure until the slave PNG is brought into synchronization within a PN bit.

c. The next step is to transfer control of the slave PNG to the DLD. This is accomplished by closing switch S3, with all the others open.

d. The doppler offset is now removed by the ground station, and normal tracking functions ensue.

1.3.2.2.2 Estimate of Acquisition Time

The acquisition time can be estimated by evaluating the time required for each of the steps in the previous subsection. Major items to consider are the lock-up time of the DSBSC receiver for a given doppler frequency offset, the information bit duration for the event marker and the total number of bits, the number of ambiguous PN bits over which the slave PN generator performs its search, and the integration time associated with each PN ambiguity resolution process. These items will be discussed in order.

a. Acquisition time of DSBSC receiver for given doppler offset.— The DSBSC receiver generates the local replica of the carrier from the sidebands as shown in Appendix D (Volume 2). The error voltage into the VCO has the same form as the corresponding quantity in the case of the phase-lock loop. The dynamic performance of the DSBSC receiver can be similarly predicted from the phase-lock loop. The acquisition time for a given doppler frequency offset $\Delta\omega$ is given by,

$$\frac{\Delta\omega^2}{2\zeta\omega_n^3} \text{ sec},$$

where ω_n is the loop natural frequency in rad/sec, and ζ is the loop damping constant as discussed in Sub-section 1.3.1.5 of this report.

The event marker will be a binary code, whose bit duration is of the order 1/5 of a millisecond. Coherently detected [7], this code can be distinguished from its complement with a probability of 2×10^{-3} for $\frac{E}{N_o} = 6$ db. If the code consists of 4 bits, the probability of at least one error occurring is 0.006 as shown in Appendix D (Volume 2). Hence, the probability of receiving all four bits correctly or with a single error is 0.994. This process requires 4/5 of a millisecond.

b. Slave PNG Acquisition — The APC controls clock pulses into the slave PNG during acquisition. The clock pulses can be generated using the vehicle reference oscillator, for with doppler wipe out, the effects of doppler on the bi-phase modulation of the input signal are insignificant. Hence, the slave PNG search is controlled by clock pulses generated from the vehicle reference oscillator. This method of slave PNG control does not necessarily provide a slave PNG output which is in PN

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[7] Lawton, J.G., Comparison of Binary Data Transmission Systems, Conference Proceeding of 2nd National Convention on Military Electronics, P. 54

bit sync with the incoming PN modulation, so the slave PNG will be shifted by 1/2 a PN bit before each comparison to establish slave PNG synchronization. In order to provide the required range ambiguity resolution, a 15 stage maximal length sequence is generated to provide a sequence whose period is 32,768 bits. At a megabit rate, it requires 32 milliseconds to recycle. Since the event marker considered here specifies a particular state in the maximal length sequence, a maximum of 128 milliseconds delay may be incurred if the reference state had just been passed when the need for an event marker arose. The event marker resolves the ambiguity of the incoming PN state to within 50 bits as shown in Appendix D (Volume 2), so the slave PNG searches over the ambiguous bits providing for two comparisons per ambiguous PN bit. If the PN bits are in bit synch, but not synchronized with respect to PN states, the cycle and add property of maximal length sequences [8] provides an input to the DSBSC receiver which is bi-phase modulated by a time shifted version of the incoming PN sequence. The power density spectrum [8] of this signal is given by,

$$\frac{p+1}{p^2 t_o^2}\left|H_1(f)\right|^2 \sum_{n=-\infty}^{\infty} \zeta\left(f-\frac{n}{pt_o}\right)-\frac{1}{pt_o^2}\left|H_1(f)\right|^2 \sum_{n=-\infty}^{\infty} \zeta\left(f-\frac{n}{t_o}\right)$$

where $p = 2^{15} - 1$, $t_o = 10^{-6}$ sec, and

$$H_1(f) = \frac{\sin(\omega t+\omega_c)\frac{t_o}{2}}{(\omega + \omega_c)} + \frac{\sin(\omega-\omega_c)\frac{to}{2}}{(\omega - \omega_c)}$$

where ω_c is the carrier frequency, and $\omega_c t_o = 2\pi n$, where n is some integer. Since the first zero occurs at $f = f_c \pm \frac{1}{t_o}$, the total signal power is spread out, with the first zeros of $\frac{\sin^2 x}{x^2}$ occurring at $\pm$ 1 MC about the carrier. It is possible for

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[8] Titsworth, R. C. and Welch, L. R., "Power Spectra of Signals Modulated by Random and Peudorandom Sequences". JPL Technical Report No. 32-140, Pages 22,24.

the DSBSC receiver to lose lock during this operation, so it is reasonable to estimate the acquisition time for the worst case with this possibility in mind. When the slave PNG is in synch with the incoming PN modulation, ideally, at least, the DSBSC receiver input is a sinusoid with a doppler frequency offset. The problem of acquisition, in this case, reduces to that of a phase-lock loop acquiring a doppler shifted signal. In this case, the constraint on the comparison time of the slave PNG operating in a particular state with the incoming PN modulation is that of allowing sufficient time for the phase-lock loop to acquire a doppler shifted signal. At acquisition, the doppler offset with the compensation is given by $\Delta\omega = 320 \times 2\pi$ rad/sec for low altitude orbits and $\Delta\omega = 160 \times 2\pi$ rad/sec for high altitude orbits. This assumes an average carrier frequency of 2 KMC, and a doppler wipeout factor of $\frac{1}{500}$. A loop-noise bandwidth of approximately $2 B_{Lo} = 1.06\ \omega_n = 330$ radians/sec is required to acquire the doppler offset of $\Delta\omega = 160 \times 2\pi$ rad/sec in 20 milliseconds. With 100 comparisons to make the permitted two-second acquisition, time is exhausted. In addition, the loop-noise bandwidth does not provide the tracking accuracy required. The results of the considerations, so far, indicate the possible need for a separate acquisition loop; however, the basic acquisition time requirement still presents a problem difficult to overcome. The critical item is to maintain lock of the DSBSC receiver during slave PNG acquisition, and this requires further investigation. A possibility is to use a low index PM modulation during the acquisition process in conjunction with a phase-lock loop receiver. Needless to say, once the DSBSC receiver has been acquired, additional time is required to remove the doppler. The fastest frequency change can be applied as long as the loop does not open.

1.3.2.3 Thermal Noise Ranging Errors

The delay lock discriminator (DLD) controls the voltage-controlled clock pulsing the slave PNG during the tracking operation. The variance of the estimated time $\hat{T}$, generated by the DLD, controls the system range accuracy[9]. This variance

[9] Spilker, J. J. "Delay Lock Tracking of Binary Signals." IEEE Transactions on Space Electronics & Telemetry, SET-9, March, 1963.

due to noise is:

$$\sigma^2_{\hat{T}} = 2.12\, t_o^2 \left(\frac{p}{p+1}\right)^2 \frac{N_o\, \omega_o}{P_s} \; \sec^2$$

ω_o is the DLD loop filter parameter, N_o (watts/cps) is input noise power density and P_s(watts) is the received signal power. Since $p = 2^{15} - 1$, $\frac{p}{p+1} \doteq 1$ and can be neglected. Using the value of $\frac{N_o}{P_s}$ for maximum range of 4000 nautical miles,

$$\sigma_{\hat{T}} = \sqrt{1.06\,\omega_o}\; 10^{-8} \sec$$

Converting this to the range variance

$$\sigma_R = 3\sqrt{1.06\,\omega_o} \; \text{m}$$

The function $1.06\,\omega_o$ is the loop noise bandwidth of the DLD. Even if this value is 1 rad/sec, the requirements are not met.

1.3.2.4 Common Channel Analysis

The interference rejection properties of the PN system are derived from the fact that the DSBSC receiver input and the DLD input are mixed with a locally generated PN sequence. This process spreads out the signal power of interfering signals, and collapses the spectrum of the signal of interest.

Apply this to the case in which the vehicle is performing PN ranging functions with several ground stations, simultaneously. When the geometry is such that the transit times are different, it is possible to consider the use of common channel ranging. With the assumption that the transit times are unique, it is possible that one or more stations will provide more signal power than the other stations. In particular, in terms of input signal power density at the frequency of interest, it is expected that the received signal power density can be 40 db above that at another frequency of interest. Here we are assuming co-channel operation, but some separation of the carrier frequencies in the common channel.

The fact that the mode of operation described above is not attractive in the presence of extreme signal power dynamic range can be seen from the consideration of separating the carrier frequencies in accordance with the 3.2 MC channel assignment. If the PN modulation is generated at the rate of 1 megabit, the power spill over into an adjacent channel is proportional to:

$$\frac{\pi}{2} - \int_{0}^{5.05 \text{ rad}} \frac{\sin^2 x}{x^2}\, dx = 0.1179,$$

or corresponds to 7.5 percent of the total power associated with the ranging signal.

We assume that the adjacent channel 2 is performing PN ranging functions, so each component of the discrete spectra spilling in from the other channel is mixed with the local replica and spread over into channel 2 and to its adjacent channels. For example, a spill-over component from channel 1 into channel 2, which is displaced $\Delta\omega \ \frac{\text{rad}}{\text{sec}}$ from the center frequency of channel 2 will have the form

$$\sin \left\{(\omega_2 - \Delta\omega)\, t + \theta_1\right\},$$

where θ_1 is some phase angle. This signal will combine with the ranging signal in channel 2, namely,

$$\sin (\omega_2 t + \theta_2(t)),$$

where $\theta_2(t)$ is the PN ranging modulation. The combined signal received will be

$$\sin (\omega_2 t + \theta_2 (t)) + \sin \left\{(\omega_2 - \Delta\omega)\, t + \theta_1\right\},$$

and will be mixed with the local replica for channel 2, namely

$$\cos (\omega_2 t + \omega_{IF} t + \theta_2 (t))$$

The low frequency output in this case will be,

$$\sin \omega_{IF} t + \sin \left\{(\omega_{IF} + \Delta\omega)\, t + \theta_2 (t) + \theta_1\right\}.$$

The second term is the interfering term, and has a spectrum similar to the ranging signal in channel 2, except that it is centered at $\omega_{IF} + \Delta\omega$.

In order to assess the effects of the spill-over from an adjacent channel, the total power from channel 1 must be determined by taking into consideration all the spill over components from channel 1, and the result of mixing with the replica used to receive the ranging signal in channel 2. For example, in order to assess the reduction in spill-over by filtering the baseband signal, consideration of this nature will be required. In order to obtain a worst-case estimate of the total effect of the energy spilling into channel 2, the total power from channel 1 is taken to be a sinusoid at the center of channel 2. If the geometry were such that channel 1 were receiving 40 db total power over channel 2, the 7.5 percent spill over will correspond to 39.5 db above the effective ranging signal power in channel 2. This interfering signal power will be spread out by the PN sequence in channel 2, so approximately 85 percent of the interfering power from channel 1 will be present in channel 2.

1.3.2.5 Tradeoffs and Conclusions

The acquisition time and tracking accuracy considerations make this approach not a very attractive one from the point of view of meeting performance objectives. The major difficulties encountered here can be remedied by the use of a hybrid system and are discussed further in Sub-section 1.3.3.2.

1.3.3 Hybrid Systems

1.3.3.1 Sidetone Hybrid

In the sidetone hybrid ranging system considered in this report, a fine ranging tone is added to a pseudo-noise sequence and the composite signal is used to phase (or frequency) modulate a carrier. The transmitted signal is given by:

$$y(t) = \cos(\omega_c t + \Delta \Theta \Phi(t) + \Delta \phi \cos \omega_r t)$$

where

$\Phi(t)$ = pseudo noise sequence
ω_r = ranging tone frequency rad/sec
$\Delta \phi$ = modulation index of tone
$\Delta \Theta$ = modulation index for pseudo noise sequence.

The pseudo-noise sequence is pure pseudo-noise sequence which has been modulo-two added to a clock. The pseudo-noise sequence is used for coarse ranging or ambiguity resolution. The pseudo noise bit rate, the clock rate, and the fine ranging tone frequency are chosen so as to be harmonically related. Thus, if we assume the pseudo-noise bit length is 2 microseconds, a 500 KC clock and a fine ranging tone of 1 MHz, the baseband spectral density envelope will have the shape as shown in Figure 1-5. The power spectral density is simply that of the pseudo noise modulo, two-clock pulse spectral density, $S\Phi(f)$, plus that of the tone, $S\omega_r(f)$. The rf spectral density as derived in Appendix C (Volume 2) will have the shape as shown in Figure 1-6.

The percentage of power in the portions of the rf spectrum, are found in Appendix C (Volume 2) to be:

Component	Percent
Carrier	$\cos^2 \Theta \, J_o^2 (\Delta\phi)$
Tone	$2 \cos^2 \Theta \, J_1^2 (\Delta\phi)$
$2 \omega_r$	$2 \cos^2 \Theta \, J_2^2 (\Delta\phi)$
⋮	⋮
PN	$\sin^2 \Theta$

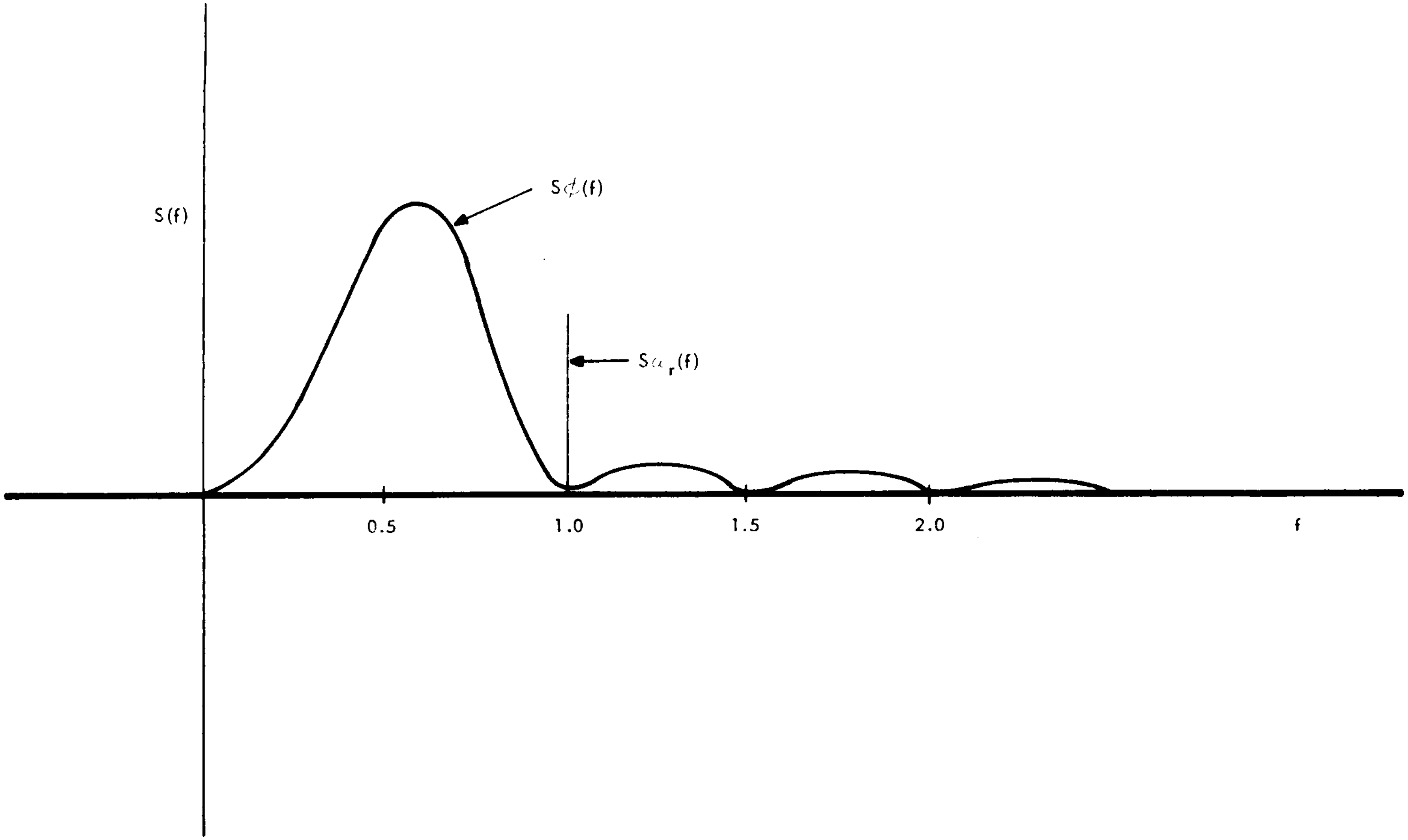


Figure 1-5. Baseband Power Spectral Density Envelope for PN + $2f_s$ + Tone

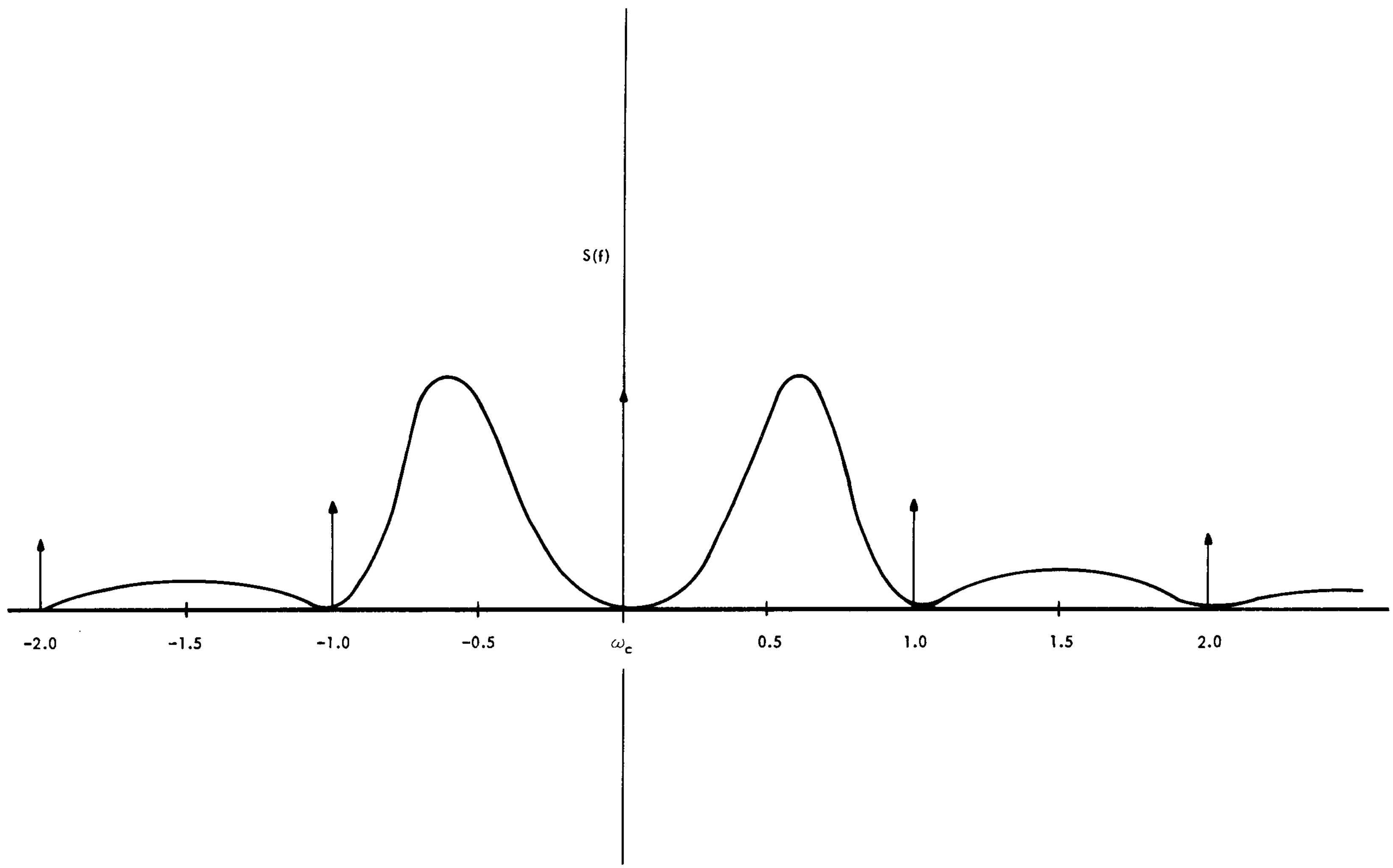


Figure 1-6. RF Spectral Density of PN + $2f_S$ + Tone

In the previous analysis of the sidetone system, Sub-section 1.3.1, it was seen that a 5 watt carrier power and a 3 watt tone power could give the required performance accuracy over a large volume of space. The pseudo-noise sequence is used only to resolve the course range ambiguity to within a half cycle of the fine ranging tone with a high probability. The selection of the carrier tracking loop parameters and fine ranging loop parameters will follow that as given for the sidetone system. If we assume the use of a delay lock discriminator (DLD) to track the pseudo noise, the time error in the DLD will be*:

$$\sigma_t^2 = 2.12\ \Delta^2 \left(\frac{M}{M+1}\right) \Phi\ \omega_{n_o}$$

where

Δ = PN bit length

M = number of bits in sequence

Φ = noise power spectral density to signal power ratio

ω_{n_o} = DLD loop natural frequency at the design point.

A 5 rad/sec loop natural frequency will be sufficient to allow tracking at the high doppler rates. Using a 5 rad/sec loop, the time error is found to be:

$$\sigma_t^2 = (2.12)\ (2 \times 10^{-6})^2\ \frac{5}{(2)\ (2000)} = 1.06 \times 10^{-4}$$

$$\sigma_t = 0.103\mu s$$

The period of the fine ranging tone is 1 microsecond; therefore, the probability of error is 0.0006 percent.

The advantage of the sidetone hybrid system, is that during acquisition and tracking of the PN sequence, a coherent clock pulse for driving the slave pseudo-noise generator is available from the fine range tracking loop. The output of the slave pseudo-noise generator is in clock synchronism with the incoming signal. Therefore, the problem of removing the clock from the PN $\oplus$ 2 f_s signal is obviated. The only problem remaining is to obtain bit synchronism. The output of the slave,

---

*The use of the delay lock discriminator is assumed for purposes of analysis and does not necessarily imply its use in the implementation of the system.

pseudo-noise generator can be cross-correlated with the received pseudo-noise sequence to obtain bit synchronization. The analysis given in section 1.3.3.2.2 for the PN hybrid is also applicable for this case.

Another advantage of this system is that interference rejection can be obtained by a feedback of the pseudo random sequence in the first mixer. A disadvantage of the system is that the rf spectrum may contain more power in the adjacent channels than for a PN hybrid. The reason for the power outside the channel can be seen when we consider the derivation of the rf spectrum. If we consider the carrier phase modulated only by the tone, then the rf spectrum contains the carrier, $J_o(\Delta\phi)$, and sidebands at $\omega_c \pm k\omega_r$, $J_k(\Delta\phi)$. When the pseudo noise is also applied, the resultant rf spectrum is the rf spectrum of the carrier with the tone modulation convolved with the pseudo-noise baseband spectrum.

1.3.3.2 PN Hybrid

The difficulties encountered in a system in which the carrier is directly bi-phase modulated by the output of a maximal length sequence generator were as follows:

a. The power density spectrum of the bi-phase modulated carrier has an envelope which does not vanish at the carrier frequency, and thus leads to the increased probability of the previously locked coherent receiver losing lock as a result of the decrease in signal power density resulting from the PN modulation. Experiments[10] with the DSBSC receiver have revealed that the reconstituted carrier becomes a noisy reference and results in a per digit error of 10^{-1}, approximately, when $\frac{E}{N_o}$ is 4 db, whereas the per digit error is approximately 8×10^{-4} when $\frac{E}{N_o} = 12$ db, where E is the received energy per bit of information.

b. The clock pulses driving the slave PNG during acquisition did not permit the slave PNG to operate bit synchronous with the incoming pseudo random modulation during the slave PNG synchronization procedure. In the system described in Sub-section 1.3.2, the slave PNG was shifted 1/2 a PN bit at a time, to compensate for the lack of a PN bit synchronized clock pulse to drive the slave PNG and thus, resulting in a longer acquisition time.

[10] Matson, D., "Tracking Telemetry, and Command Studies of Ground and Satellite Subsystem", Vol IV, Aerospace Corp., AD 438 031.

c. The DLD in the system of Sub-section 1.3.2 did not provide the range tracking accuracy required. These shortcomings can be alleviated by mixing clock pulses with the PN modulation. Using bi-phase modulation of the carrier frequency ω_c, the ranging signal will be,

$$(-1)^{p(t)}(-1)^{c(t)} \cos(\omega_c t + \theta),$$

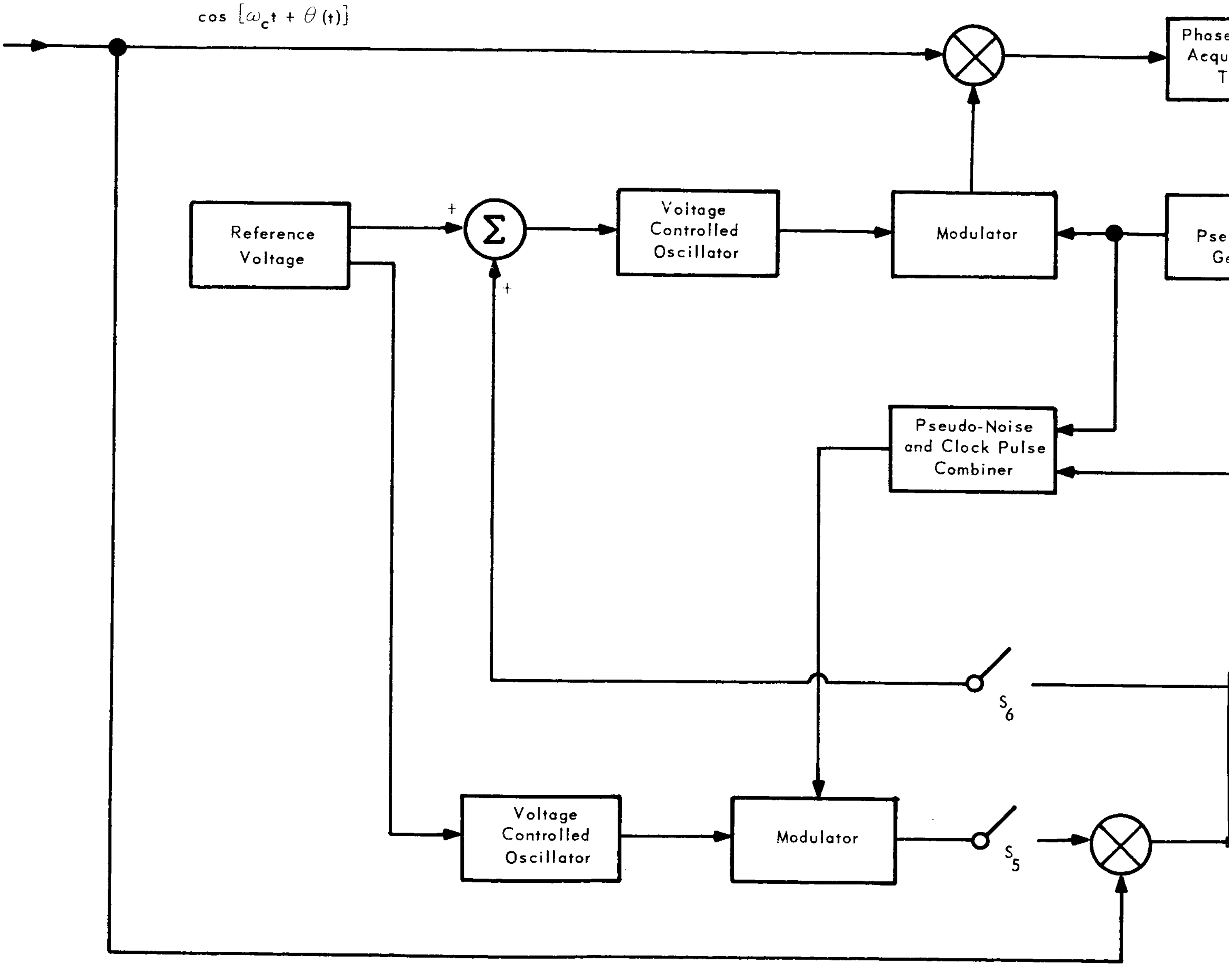
where p(t) and c(t) have values 0 and 1, and are the outputs of the PNG and clock pulse generator (CPG) respectively; θ is an arbitrary phase angle. A logical block diagram of the system is shown in Figure 1-7.

The result of multiplying the clock pulses with the maximal length sequence generator changes the power density spectrum of the ranging signal. The spectrum depends upon the relation between the PN bit duration, t_o, and the period of the clock pulses. The carrier modulated by the maximal length sequence has a line spectra discussed in section 1.3.2.2.2. The line spectra resulting from the combination of the maximal length sequence and the clock pulses will differ from that of the maximal length sequence alone. This can be determined from the resulting changes in the probability distributions associated with the states of the combined sequence. The more important item is the envelope of the ranging signal spectrum resulting from modulating the carrier with the sequence combined with the clock pulses. This has been obtained and it can be seen that

$$H_1(f) = \frac{-\omega + e^{j\omega to} \; (\omega \cos \omega_c t_o - j\omega c \sin \omega_c t_o)}{j(\omega^2 - \omega_c^2)}$$

when the clock pulse period is $2t_o$. Here, $t_o = \frac{2\pi n_o}{\omega_c}$ and n is an integer. When the clock period is t_o, then

$$H_1(f) = -\frac{1}{4j} \; \frac{(\omega - \omega_c)\left(e^{j(\omega+\omega_c)\frac{t_o}{2}} - 1\right)^2 + (\omega + \omega_c)\left(e^{j(\omega-\omega_c)\frac{t_o}{2}} - 1\right)^2}{\omega^2 - \omega_c^2}$$



cos [ωct + θ(t)]
Phase
Acqu
T
Reference Voltage
Σ
+
+
Voltage Controlled Oscillator
Modulator
Pse
Ge
Pseudo-Noise and Clock Pulse Combiner
S6
Voltage Controlled Oscillator
Modulator
S5

1-47-1

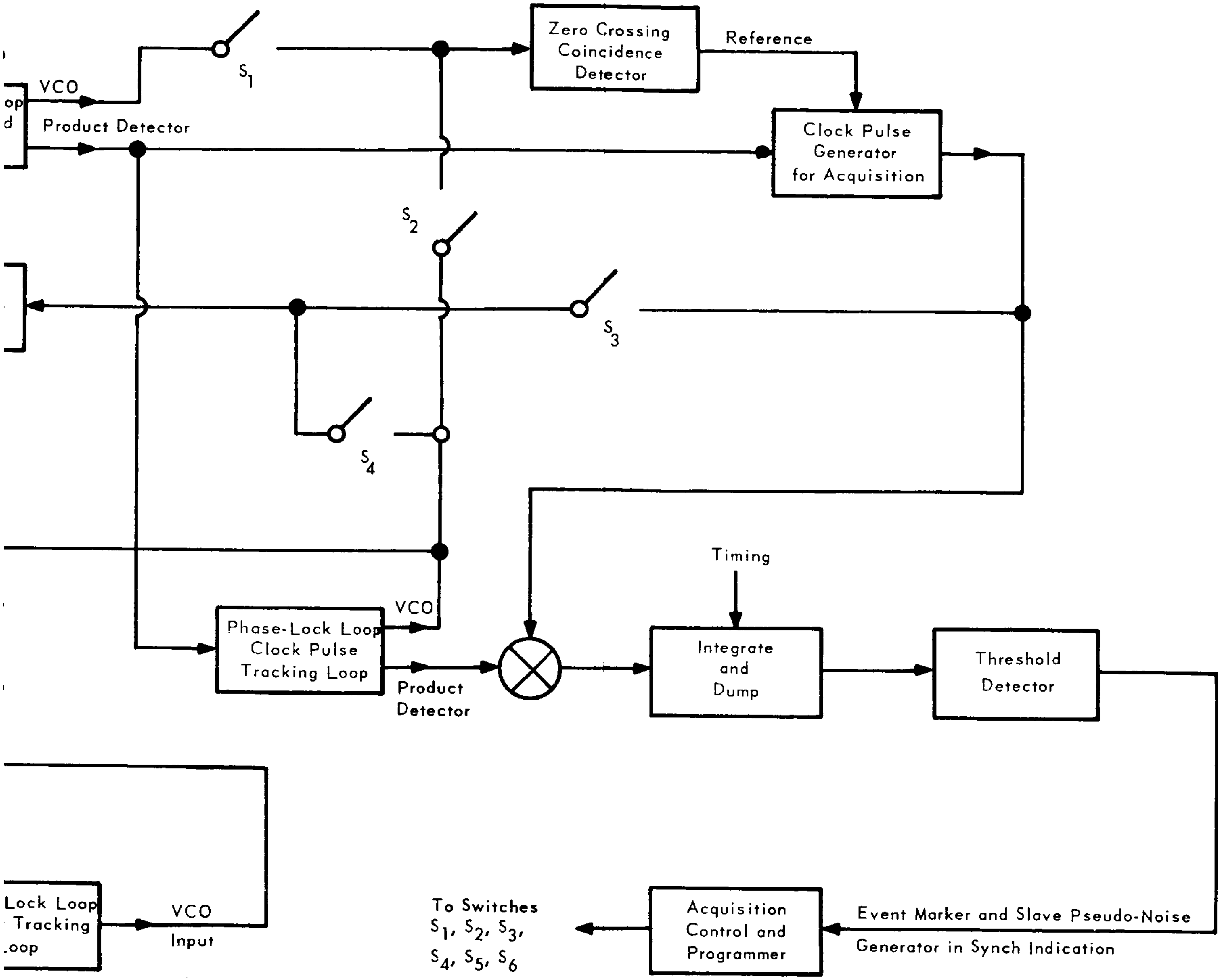


Figure 1-7. PN Hybrid System Receiver

and by applying L'Hospital's Rule:

$$H_1(f)\Big|_{2\pi f\ =\ 2\pi f_c} = 0$$

and

$$H_1(f)\Big|_{2\pi f\ =\ 2\pi f_c + \frac{4\pi}{t_o}} = 0$$

This serves to show that the effects of PN modulation can be eliminated in the vicinity of the carrier frequency.

The logical block diagram of this ranging system consists of phase-locked loop receivers in lieu of the DSBSC receiver previously considered. The system consists of three phase-locked loops, of which one is used to acquire the doppler wiped-out carrier and, at a later stage, to provide an input to the phase-lock loop tracking the clock pulses. In addition to the phase-locked loop tracking the clock pulses, there is another phase-locked loop provided to track the carrier frequency. Once the tracking mode is established, the carrier-tracking loop tracks a sinusoidal signal resulting from the removal of the PN and clock pulses. The input to the carrier tracking loop VCO is fed to the acquisition and tracking phase-lock loop to aid the tracking process. The acquisition and tracking phase-lock loop receives the ranging signal with the PN removed and corrected in doppler frequency shift. The clock pulse tracking loop tracks the clock pulses from the output of the product detector in the acquisition and tracking phase-lock loop. The slave PNG is controlled during tracking by the VCO output from the clock pulse tracking loop.

1.3.3.2.1 Acquisition Procedural Analysis

a. The carrier will be acquired with the doppler wiped out. The VCO output of the acquisition and tracking phase-lock loop will be coherent with the incoming doppler wiped-out carrier.

b. Clock pulses whose period corresponds to a single PN bit duration are inserted on the doppler wiped-out carrier by means of phase modulation. An index less than 1 will be used in order to maintain a carrier component to aid the tracking functions of the acquisition and tracking, phase-lock loop. When a clock pulse which is considered to be a squarewave, and has a period T, phase modulates a carrier with a phase deviation Φ, the modulated signal is

$$A_c \cos \left(\omega_c t + \Phi \sum_{\substack{n=1 \\ n\,odd}}^{\infty} \frac{4}{n\pi} \sin\left(\frac{2\pi nt}{T}\right)\right)$$

Using Jacobi's expansion, the preceding can be rewritten as,

$$A_c R_e \left(e^{i\omega_c t} \prod_{\substack{n=1 \\ n\,odd}}^{\infty} e^{i\Phi \frac{4}{n\pi} \sin\left(\frac{2\pi nt}{T}\right)} \right)$$

The carrier component has the coefficient,

$$A_c J_o \left(\Phi \frac{4}{\pi}\right) J_o \left(\Phi \frac{4}{3\pi}\right) J_o \left(\Phi \frac{4}{5\pi}\right) \cdots\cdots$$

With a deviation of $\Phi = \frac{\pi}{4}$,

$$A_c J_o \left(\Phi \frac{4}{\pi}\right) J_o \left(\Phi \frac{4}{3\pi}\right) J_o \left(\Phi \frac{4}{5\pi}\right) \cdots \doteq .72\ A_c$$

Perhaps this is not the optimum deviation to use, but this will be determined from the loop requirements and other concerning factors to be considered during acquisition, which are discussed later. The acquisition and tracking loop's product detector output is tracked by the clock pulse tracking loop. The VCO output of the acquisition and tracking loop is phase coherent with the incoming signal, so this VCO output and the VCO output of the clock pulse tracking loop are fed into a zero crossing coincidence detector to establish a coherent clock pulse to drive the slave PNG during acquisition.

c. The event marker is transmitted by means of performing a bi-phase modulation of the clock pulses, or equivalently using the clock pulse as a subcarrier. The event marker is detected by linear demodulation of the modulated clock pulses, using as a reference, the clock pulses obtained from the VCO tracking the input doppler wiped-out signal. Having detected the event marker, the slave PNG is ready to be stepped for the purpose of synchronizing it to the incoming PN modulation.

d. Item 1: the doppler wiped-out carrier which is being modulated by the clock pulses, is now modulated by a combination of the clock pulse and the PN sequence. The bit duration of the PN is equal to one period of the clock pulse. The power density spectrum of the carrier bi-phase modulated by this type of signal was discussed in Sub- section 1.3.3.2. For the purposes of acquisition, a narrow-deviation phase modulation will be continued to be used during this stage of acquisition. Maintaining lock of the acquisition and tracking phase-lock loop can be eased by providing a carrier component for the phase-lock loop receiver to track. The slave PNG under the control of the acquisition clock pulse generator modulates a sinusoid and mixes it with the incoming signal, which in the bi-phase case is

$$(-1)^{p(t)}(-1)^{c(t)}\cos(\omega_c t + \theta),$$

as discussed in Sub-section 1.3.3.2. In this case, the local replica will be

$$(-1)^{p_e(t)} \cos(\omega_c t + \omega_{IF} t)$$

where $p_e(t)$ will be the local slave PNG estimating the state of the incoming PN modulation based on the event marker. In this case, when $p_e(t)$ is synchronized with $p(t)$, the input to the acquisition and tracking phase-lock loop will be

$$(-1)^{c(t)}\cos(\omega_{IF} t + \theta).$$

Earlier, it was found that a low index phase modulation is desirable. Also it was found that it is possible to combine the PN sequence with clock pulses having a period equal to the PN bit duration to keep the spectrum due to the modulating signal, away from the carrier frequency. In this case the local replica for acquisition, as well as the input signal, are slightly different. Assuming that the clock pulse is a two valued squarewave with amplitudes $\pm \frac{\pi}{8}$, and the PN waveform also takes on the same two values. The ranging signal in this case will be

$$\cos(\omega_c t + \theta(t)),$$

where $\theta(t)$ is determined by the clock pulse and the PN. When they disagree, $\theta(t) = 0$, and when they agree, one or the other of the values $\pm\frac{\pi}{4}$ is the modulating signal. In this case, the local replica is

$$\cos(\omega_c t + \omega_{I-F} t + \theta_e(t))$$

and the low frequency output of the multiplier is,

$$\cos(\omega_{I-F} t + \theta_e(t) - \theta(t))$$

When the slave PNG is synchronized to the state of the PN sequence which determined the value of $\theta(t)$, $\theta_e(t) - \theta(t)$ is a clock pulse train taking on the values $\pm\frac{\pi}{8}$. When this is not the case $\theta_e(t) - \theta(t)$ is a random waveform taking on values $\pm\frac{3\pi}{8}$ and $\pm\frac{\pi}{8}$. The fact that the slave PNG has achieved synchronization will be detected by the same process that detected the event marker. Item 2: at this stage of acquisition, it is also necessary to discuss the number of PN bits on which to perform the comparison before making the decision whether the slave PNG is generating the same incoming PN states. Assuming that the slave PNG is in PN bit step, but not synchronized in PN state, the comparison process ultimately results in producing a signal in the integrate and dump circuitry which is other than a positive constant; ideally speaking, during the comparison time. The process then, in principal, is that of detecting a constant level signal as opposed to some other signal which is not constant. The latter signal is determined primarily by the correlation coefficient, ρ, between the portion of the incoming PN states and the slave PNG output states being compared. Hence, ideally, it is desirable to choose the maximal length sequence used for ranging on the basis of trying to maintain ρ close to -1 over the range of ambiguous PN bits, over which the final brute force search must be performed to synchronize the slave PNG, and a value of $\rho = 1$ when synchronization is achieved. In other words, when the incoming PN states are compared with the output of the slave PNG, there will be a corresponding correlation coefficient ρ_i associated with each comparison. The requirement on the set of all ρ_i, namely, $\{\rho_i, 1 \leq i \leq n\}$, where n is the number of comparisons performed, should be close to -1, other than at the synchronized state. This probably can be achieved to various degrees, but it is reasonable to choose the ranging sequence based on this type of criteria and use the event marker to operate in the portion of the sequence that has this desirable characteristic. In the ideal case, when $\rho = -1$, a comparison time of 1 millisecond gives rise to $\frac{E}{N_o} = 13$ db, and an error probability better than 10^{-5}. When the correlation coefficient $\rho = 0$, the error probability is still better than 10^{-5}. For $\rho = \frac{1}{2}$, the error probability is 5×10^{-4}.

e. Once the slave PNG is acquired, the control of the slave PNG is transferred to the VCO output of the clock pulse tracking phase-lock loop. The clock pulse and slave PNG are combined to remove the corresponding modulation from the input to the carrier tracking loop. Once the carrier tracking loop is locked, the VCO input from this loop can be inserted to aid the tracking of the acquisition and tracking phase-lock loop. However, there are several considerations of importance. For acquisition, the clock pulse period was equal to the PN bit duration in order to keep the frequency range in the vicinity of the carrier frequency free from the spectral components due to the PN modulation. Once the slave PNG has been acquired, this is no longer a necessity, and it is advantageous to distribute a high percentage of the signal power in the channel being used. This can be achieved by reducing the clock period to twice the duration of a PN bit. In this case, the power density spectrum reduces to that discussed in Sub-section 1.3.3.2. At the same time, for tracking purposes, there is no need to operate at a low deviation as in the case of acquisition. These considerations become important when interference rejection during tracking is of importance. The local replica is narrowing the ranging spectrum at the inputs to the carrier tracking, and the acquisition and tracking phase-lock loops, and is effectively spreading out interfering signals as discussed in section 1.3.2.4. However, the one big disadvantage of adjusting the phase deviation and clock pulse rate is that, if, for some reason, one of the phase lock receivers loses lock and re-acquisition is required, it may be necessary to repeat the acquisition procedure.

f. The last stage of acquisition is to remove the doppler wipe-out and perform tracking functions.

1.3.3.2.2 - System Errors

The system errors in this case are similar to what is obtained in the systems discussed in previous sub-sections, once the ranging modulation is removed. In addition to the phase errors in the phase-lock loop due to thermal noise, there is also the possibility of practical limitations in removing the PN modulation and the errors resulting from the residual PN. Before this item is discussed, the signal power into the various loops will be determined.

Considering the channel bandwidth to be 3.2 MC, and the carrier being bi-phase modulated by the combined PN and clock pulses, whose period is twice the PN bit duration at a 1 MC bit rate, the total power in the channel has been estimated to be 92.3% of the total power, with 7.7% of the power falling outside of the channel. Upon removal of the PN modulation, an IF signal, bi-phase modulated by clock pulses, is obtained. Theoretically, this signal should contain 92.3% of the total power the range equation predicts will be received. In the clock pulse tracking loop, the first harmonic of the clock pulse is tracked. The first harmonic has approximately 83% of the total clock-pulse power, so 76.5% of the total power predicted by the radar equation is tracked by the clock pulse phase-lock loop. The carrier tracking loop receives 92.3% of the total power calculated as a result of removing the PN and clock pulse modulation. This estimate is based on the received carrier frequency being located at the channel center, so it is optimistic and should be considered, with this in mind.

1.3.3.2.2.1 - Errors Due to Residual PN Modulation.

In the tracking mode, the slave PNG is controlled by the clock pulse phase-lock loop. With bi-phase modulation, the incoming signal is

$$(-1)^{p(t)} (-1)^{c(t)} \cos \omega_c t,$$

as discussed before neglecting doppler. The local replica is,

$$(-1)^{p(t + \Delta(t))} \cos \tilde{\omega}_c t$$

in order to generate the input to the acquisition and tracking phase-lock loop. The low frequency component of the actual input is

$$(-1)^{p(t) + p(t + \Delta(t)) + c(t)} \cos (\omega_c - \tilde{\omega}_c) t.$$

The function, $\Delta(t)$, is a random variable with some distribution, and in order to meet the range accuracy requirement of 1/3 meter standard deviation, $\Delta(t)$ must have a standard deviation of $1/9 \times 10^{-8}$ seconds. Since the loop noise bandwidth of the clock pulse phase-lock loop is of the order of several radions at most, $\Delta(t)$ will vary slowly. With this assumption

$$p\,(t + \Delta(t)) \fallingdotseq p(t) + \Delta(t)\, p'(t)$$

Hence, the input to the acquisition and tracking loop during tracking will be

$$(-1)^{\Delta(t)\, p'(t)}\, (-1)^{c(t)} \cos (\omega_c - \tilde{\omega}_c)\, t.$$

In order to assess the effects of the disturbance term $(-1)^{\Delta(t)\, p'(t)}$, consider the properties of the maximal length sequence. A detailed discussion of its statistical properties are found in [8]. The frequency of occurrence of each PN state and the joint occurrence of the states separated by a given number of PN bit durations can be specified. The disturbance effects the input to the acquisition and tracking loop only when a PN state change occurs. During a run in which no PN state transition takes place, the disturbance has no effect. When a PN state change occurs $p'(t)$ will be a positive or negative going pulse. The clock pulse phase-lock loop will have an extremely narrow noise bandwidth so the disturbance will be attenuated, and the acquisition and tracking phase-lock loop noise bandwidth must be designed to reduce the loop phase error due to this disturbance.

1.3.3.2.2.2 - Tracking Accuracy

1.3.3.2.2.2.1 - Input Bandwidths to Phase-Lock Loops

The input bandwidth to the acquisition and tracking phase-lock loop. BW_1, is determined by the maximum doppler frequency offset at maximum range and the capability of the ground doppler wipe-out system to remove the doppler uncertainty during initial acquisition. The initial doppler offset into the acquisition and tracking

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[8] Titsworth, R. C. and Welch, L. R., op. cit, (Sub-section 1.3.2.2.2)

loop has been estimated to be $160 \times 2\pi \frac{\text{rad}}{\text{sec}}$. Hence, $B_{W_1} = 200$ cps is a reasonable estimate.

The input bandwidth, B_{W_2}, to the carrier tracking loop depends upon the specific time during acquisition at which it is locked, and whether the tracking information from the acquisition and tracking phase - lock loop is utilized. With this in mind,

$$B_{W_2} \leq B_{W_1},$$

The particular system considered in Figure 1-7.1 permits B_{W_2} to be used as a parameter to tradeoff in achieving the desired operation in terms of range rate accuracies and pull-in time.

1.3.3.2.2.2.2 - Constraints On Phase-Lock Loop Parameters Due to System Accuracy Requirements

a. Range Rate Requirement—In Sub-section 1.3.1.1, the relation between the variance of the range rate measurement and the variance of the loop-phase error is given for an averaging time of T seconds. Taking $\frac{1}{\tau_{ave}} = \frac{1}{\tau_d} + \frac{1}{\tau_u}$,

where τ_d is the wavelength of the vehicle to ground signal in meters, and τ_u is the wavelength of the corresponding transponded signal in meters. Imposing the requirement that the range rate be $10^{-2} \frac{m}{sec}$ when $\tau_{ave} = 0.15$ m, $T = 0.1$ second and $\Phi = \frac{1}{2 \times 10^3}$, the constraint on ω_{n_o}, the loop natural frequency at maximum range is,

$$\omega_{n_o} \leq 13.2 \frac{\text{rad}}{\text{sec}}$$

per watt transmitted.

For the system under discussion, 92.3% of the transmitted signal power enters the carrier tracking loop, so Φ should be modified accordingly. In this case,

$$\Phi = \frac{1}{1848}$$

With this value

$$\omega_{n_o} \leq 12.25 \frac{\text{rad}}{\text{sec}}$$

per watt transmitted.

Since 10 watts are transmitted, the loop noise bandwidth at maximum range is

$$\omega_{n_o} = 122.5 \ \frac{\text{rad}}{\text{sec}} \cdot$$

With this loop-noise bandwidth, at minimum range, where the doppler rate of change is,

$$\dot{\omega} = \pm 6.3 \text{ k cps}^2$$

the steady state phase error is given by

$$\frac{\dot{\omega}}{\omega_{n_o}^{\ 2}} = 2.64 \text{ rad.}$$

In order to maintain lock at minimum range, the loop steady state phase error must be maintained at approximately 0.523 radians. To meet this requirement, a bandpass limiter is used to adapt the loop noise bandwidth, which is expressed in terms of the suppression factor as discussed in Sub-section 1.3.1.3,

$$\omega_n = \sqrt{\frac{\alpha}{\alpha_o}} \ \omega_{n_o}.$$

When

$$\frac{\alpha}{\alpha_o} = 5.04 \text{ or} \sqrt{\frac{\alpha}{\alpha_o}} = 2.25$$

the range rate tracking accuracy can be met at minimum range.

b. Range Tracking Accuracy—The range accuracy obtained in the clock pulse tracking loop is discussed in Sub-section 1.3.1.1. The first harmonic of the clock pulse is tracked by the clock pulse phase-lock loop, and to obtain a range standard deviation accuracy of 1/3 meter

$$\omega_{n_o} \leq 0.8 \ \frac{\text{rad}}{\text{sec.}}$$

per watt power transmitted.

In this case the power transmitted is 10 watts, so

$$\omega_{n_o} = 6.0 \ \frac{\text{rad}}{\text{sec}}$$

In obtaining this parameter, it was assumed that 76.5% of the calculated power received centers the clock pulse phase-lock loop.

There will be no discussion concerning steady state phase error; this has already been considered in Sub-section 1.3.1.2.

1.3.4.2.3 - Acquisition Time Estimate

The acquisition procedure was discussed in Sub-section 1.3.4.2.1, and the time required for each of the stages can now be estimated.

a. The acquisition time of the acquisition and tracking phase-lock loop is determined by,

$$\frac{(\Delta\omega)^2}{2\zeta\omega_n^3}$$

If the desired pull-in time is 50 milliseconds during acquisition, when the doppler wipe-out gives an initial frequency offset of

$\Delta\omega = 160$ cps,

then $\omega_n = 250$ rad/second. When K is taken to be 2.56×10^6 sec^{-1} in accordance with section 1.3.1.2, then

$T_1 = 41$ sec

$T_2 = 5.66 \times 10^{-3}$ sec.

This choice of ω_n does not require that conformance be made with the steady state phase error due to doppler rate at minimum range, considered in Sub-section 1.3.1.2, for the acquisition and tracking loop receives tracking aid from the carrier tracking loop.

b. The lockup time of the clock pulse tracking loop can be considered to be negligible regardless of the small loop noise bandwidth, for the input doppler offset is removed by the acquisition and tracking loop and can be considered practically zero.

c. The event marker process will require 4/5 of a millisecond, as discussed in Sub-section 1.3.2.2.2.

d. The slave PNG ambiguity resolution will require 50 milliseconds as per discussion in Sub-section 1.3.2.2.2.

e. The lockup time of the carrier tracking loop can also be reduced practically to zero by utilizing the VCO output of the acquisition and tracking phase-lock loop to reduce the frequency offset into the carrier tracking loop practically to zero. It is estimated that the functions described can be performed within 1 second. The remaining time is allocated to the coordination of operations between the vehicle and ground and the removal of doppler on the ground after the vehicle has acquired the doppler wiped-out signal. The time required for the latter operation is determined by the maximum rate the doppler can be restored without opening up the phase-lock loops at the vehicle. This can be performed more rapidly than at rates Frazier and

Page[11] used to search the VCO for the purposes of locking-up the PLL to a signal with unknown frequency.

1.3.4.2.4 - Tradeoffs and Conclusions

The method considered has the following advantages:

a. Efficient utilization of the power available to perform ranging operations.

b. Relaxes the constraints on the phase-lock loop parameters by separating the carrier tracking function and the clock pulse tracking functions.

c. Possesses good interference rejection capability during acquisition, as well as tracking.

The major disadvantage is the involved acquisition process. An estimate of the acquisition time indicates a good chance of meeting the acquisition time requirement, provided the coordination process between vehicle and ground can be performed within a second. If reacquisition is required, the initial acquisition process must be repeated.

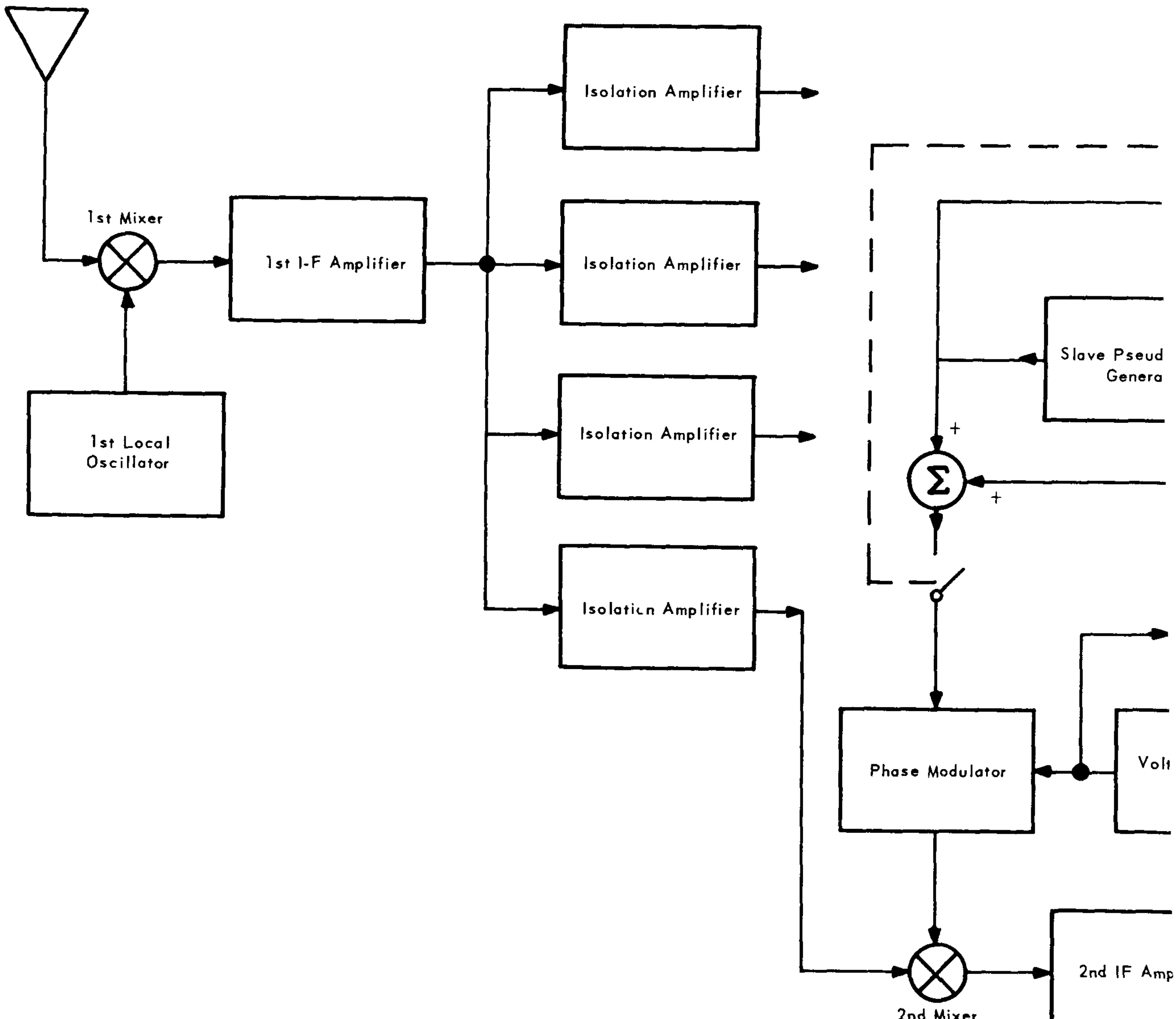
1.4 SYSTEM DESIGNS

1.4.1 Sidetone Hybrid

1.4.1.1 System Diagram

The receiver system diagram of the sidetone hybrid system is shown in Figure 1-8. The sidetone hybrid is one in which a sidetone for fine ranging is added to a pseudo random sequence to angle modulate the carrier. The pseudo random sequence is a pseudo sequence which has been modulo-two added to a clock. The pseudo random sequence bit rate, clock rate, and tone frequency are selected so that the pseudo random sequence spectrum has a null at the carrier and at the tone, as shown in Subsection 1.3.3.1. The modulation index of the pseudo random sequence and the

[11] Frazier, J. P. and Page, J., "Phase-Lock Loop Frequency Acquisition Study", IRE Transactions on Space Electronics and Telemetry, Sept. 1962, pp 210-227



1st Mixer
1st I-F Amplifier
1st Local Oscillator
Isolation Amplifier
Isolation Amplifier
Isolation Amplifier
Isolation Amplifier
Slave Pseud
Genera
+
+
Phase Modulator
Volt
2nd Mixer
2nd IF Amp

1-59-1

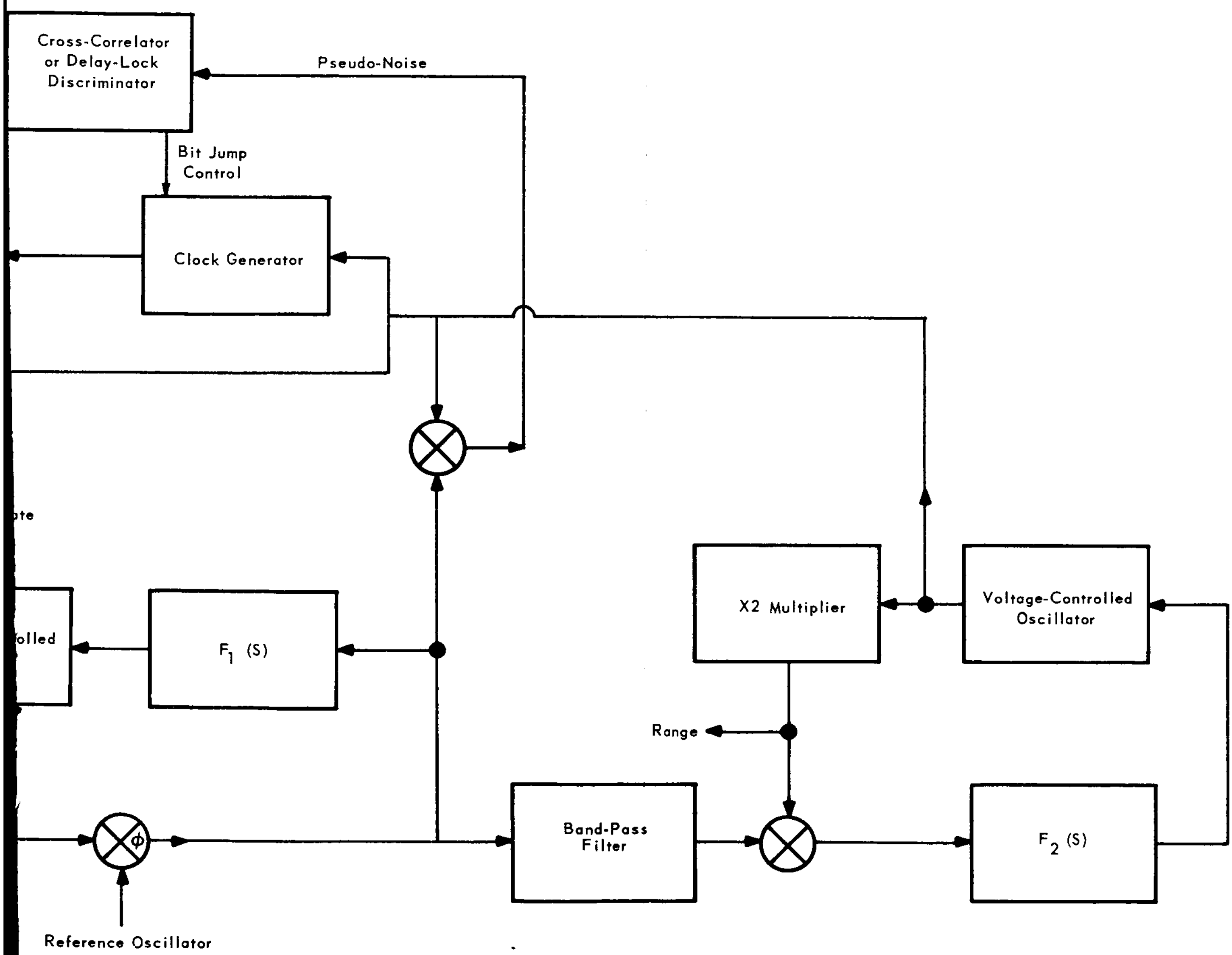


Figure 1-8. Sidetone Hybrid System Receiver

modulation index of the sidetone will determine the amount of power in the received carrier component, the sidetone, and the pseudo random sequence.

The receiver system consists of three main portions: (1) the carrier tracking loop, (2) the fine range tracking loop, and (3) the ambiguity resolution circuit.

1.4.1.2 Acquisition

In acquisition, the ground station first transmits the carrier with the fine range tone modulation on it, and with doppler wiped out. The carrier tracking loop first locks on the carrier and then the fine range tracking loop locks on the fine ranging tone. After locking on to both thc carrier and fine ranging tone, a command is generated to the ground station to apply the pseudo-noise ambiguity modulation to the carrier along with the fine ranging tone. It should be noted that the transmission of the carrier with the fine ranging tone offers some interference protection, since the receiver must lock on to both the carrier and tone before the vehicle considers the signal which it has locked on to as being the proper signal. In order to lock on to a false signal, it would be necessary to have an interfering signal with a modulation at the tone frequency, or two interfering signals which differ in frequency by the tone frequency. After the ground receives the command from the vehicle to apply the PN ambiguity signal, the ground station first transmits a coded word on the S-band as the event marker. The event marker starts the spacecraft slave PNG from some initial position. The ground station then begins to transmit the signal with both pseudo-noise ambiguity and range-tone phase modulating the carrier. During acquisition of the pseudo-random sequence, the modulation index of the pseudo-random sequence can be chosen at the ground station to place more power in the pseudo-random sequence portion of the rf spectrum than is used in the tracking. Since the carrier tracking loop and fine range loop are well above the threshold of dropout, they will continue to track the signal. After acquisition, the modulation index will be changed to place the normal tracking power in the carrier and tone frequencies.

The pseudo-noise ambiguity is a PN sequence which is modulo-two added to a sub-harmonic of the fine range tone. Since the sub-harmonic is coherent with the fine range tone, the output of the fine range tone tracking loop can be divided down to obtain a coherent replica of this sub-harmonic. This can be used to provide the slave PNG with an accurate timing pulse and also to mix with the PN modulo-two signal to obtain the PN signal. This PN signal and the output of the slave PNG are then correlated to obtain an indication when the two are in synchronization. After the slave PNG is locked in, the output of the slave PNG is modulo-two added to the sub-harmonic. This signal is then used to phase modulate the output of the carrier tracking VCO at the input to the second mixer.

After the vehicle receiver is locked on, the carrier, tone, and pseudo-noise ambiguity, a command is generated and sent to the ground station to remove doppler. The ground station then removes the doppoer wipe-out in a ramp fashion, and changes the modulation index of the pseudo-random sequence to the tracking mode index.

1.4.1.3 Tracking

Since the incoming signal is mixed with the local pseudo-random sequence modulo-two added to a clock, PN $\oplus$ $2f_s$, the output spectrum of the mixer will be determined by the modulation index of the incoming pseudo-random sequence and the modulation index of the pseudo-random sequence which is fed back to the mixer.

During acquisition, the modulation index of the pseudo random sequence might be chosen as $\pi/4$. The transmitted signal from the ground will be

$$x(t) = \cos\left(\omega_c T + \frac{\pi}{4}\Phi(t) + \Delta\phi \sin\omega_r T\right)$$

The percentage of received carrier power will be $1/2\, J_o^2(\Delta\phi)$. The percentage power in the fine ranging tone will be $J_1^2(\Delta\phi)$ The total percentage power in the pseudo random sequence spectrum will be 50%.

After the spacecraft slave pseudo-noise generator is synchronized with the incoming signal, the switch is closed, causing the VCO output to be modulated by the pseudo-noise modulo-two added to a clock. The modulation index of the pseudo noise fed back to the second mixer is chosen as $\pi/2$. The process of feeding back a PN signal with a modulation index of $\pi/2$ which is greater than the incoming signal will be referred to as "unmatched index demodulation". The output of the mixer will not change if the pseudo-noise sequence is synchronized. If the local signal fed to the mixer is

$$y(t) = \sin \left[(\omega_c - \omega_{if})t + \pi/2\ \Phi\ (t) \right]$$

the output of the mixer is, neglecting higher frequency terms, given by

$$z(t) = x(t)y(t) = \frac{1}{2} \sin \left[-\omega_{if}t + \frac{\pi}{2} \Phi(t) - \frac{\pi}{4} \Phi(t) - \Delta \phi \sin \omega_r t \right]$$

$$= -\frac{1}{2} \sin \left[\omega_{if}t + \frac{\pi}{4} \Phi^*(t) + \Delta\phi \sin \omega_r t \right]$$

where

$$\Phi^*(t) = -\Phi(t)$$

If the local signal fed back is

$$y(t) = \sin \left[(\omega_c + \omega_{if}) t + \frac{\pi}{2} \Phi(t) \right]$$

then the mixer output is given by

$$z(t) = x(t)y(t) = \frac{1}{2} \sin \left[\omega_{if}t + \frac{\pi}{4} \Phi(t) - \Delta\phi \sin \omega_r t \right]$$

In either case, the power spectral density is the same about the i-f center frequency.

After the spacecraft slave pseudo-noise generator is synchronized and unmatched index demodulation is in use, the ground station removes the doppler wipe-out and changes the modulation index of the pseudo-random sequence to the tracking mode value.

The removal of doppler wipe-out after unmatched index demodulation is in use will afford interference rejection to interfering sine waves. Assume an interfering sine wave is given by

$$x_i(t) = B \cos (\omega_i t + \Psi)$$

The local signal into the second mixer is

$$y(t) = A \sin \left[(\omega_c - \omega_{if})t + \frac{\pi}{2} \Phi(t) \right]$$

The output of the mixer will be:

$$z(t) = \frac{AB}{2} \sin \left[(\omega_c - \omega_i - \omega_{if}) t + \pi/2 \ \Phi(t) + \psi \right]$$

The interfering sine wave is thus spread out by the pseudo-random sequence.

In the tracking mode, the modulation index of the transmitted pseudo-random sequence is chosen as ϵ, then the modulation index of the pseudo-random sequence at the output of the mixer will be $\pi/2 - \epsilon$. Thus, ϵ can be chosen along with $\Delta\phi$, the tone modulation index, to distribute the proper amount of power in the carrier and tone components for tracking accuracy. Since the clock of the slave PNG is being derived from the fine range tracking loop, there need only be enough power in the pseudo-random sequence to keep the probability of false alarms in the cross correlator to a prescribed value.

1.4.1.4 Reacquisition

In the tracking mode, with an 80 rad/sec loop natural frequency, and 5 watts for the carrier power, the signal-to-noise ratio in the loop two-sided noise bandwidth at maximum range will be 21 db. The signal-to-noise ratio in the fine-range tracking loop with a 4 rad/sec loop and a 3 watt tone power will be 31.7 db. The absolute threshold for tracking in a phase-lock loop is 0 db; therefore, both loops are well above the threshold of loss of lock. In the case of the fine ranging tracking loop skipping a cycle due to high impulse noise, the slave pseudo-noise generator will no longer be in bit synchronization with the incoming sequence. The PN cross correlator will indicate loss of lock and stop the unmatched index demodulation. The modulation index of the pseudo-random sequence at the output of the mixer will be the same as the received signal, that is, ϵ. This modulation index will be such that most of the power is placed in the pseudo-noise portion of the spectrum which will aid in the reacquisition of the pseudo-random sequence bit synchronization.

In the event of complete signal dropout due to retrofiring, all loops will lose lock. In this case, the unmatched index demodulation is again removed. If the signal is lost for several seconds, it will be necessary to have phase-lock loops that not only have a doppler offset memory, but also have a doppler rate memory (with integration) to keep the VCO center frequency at approximately that of the received signal when the flame attenuation subsides. Since the pull-in range of the carrier tracking loop was found in to be 22,600 rad/sec (Sub-section 1.3.1.5) a doppler rate of 1440 KHz/sec with a 2.5 second duration of loss of signal will cause the received signal to be outside the pull-in range of the VCO; therefore, a doppler memory will be needed. If, after a period of time, the signal has not been acquired, the reacquisition procedure would be the same as the initial acquisition procedure.

1.4.1.5 Ground Transponder Equipment

In the modulation study, the greater emphasis was placed on the spacecraft. Since at the ground equipment, the signal power-to-noise power spectral density is greater than the spacecraft, the expected interference is much less than the spacecraft, and there is no adjacent channel interference problem; this emphasis is justified.

The tracking loop parameters have been discussed (Sub-section 1.3.1) and will also apply for the PN, sidetone hybrid. The acquisition problem at the ground station is different from that of the spacecraft, since the ground station must acquire the signal with both the pseudo-random sequence and sidetone modulation on it. However, the use of the VHF doppler estimate and event marker on the VHF will enable the ground station to acquire even though there is modulation by both the pseudo-random sequence and tone. Since the carrier and tone are at nulls of the pseudo-random sequence in the rf spectrum, the VHF doppler estimate can place the carrier VCO near the carrier and there will be negligible power from the pseudo-random sequence components entering the carrier phase-lock loop noise bandwidth. A sweeping of the carrier VCO may also be used to reduce the acquisition time. Since the accuracy of the VHF doppler estimate is not available, no calculations of the acquisition time for the ground carrier loop and tone loop was made. Since the interference problem is not serious, the use of unmatched index demodulation does not seem warranted in the ground station.

1.4.2 PN Hybrid

Many items pertaining to system design with regards to the PN Hybrid system have been discussed. (Refer to Sub-section 1.3.3.2.)

1.4.2.1 System Diagrams

a. The generation of clock pulses in bit synchronization with the incoming PN bits to operate the slave PNG during acquisition is the important means of reducing the acquisition time in this system. The input to the acquisition and tracking phase-lock loop is a carrier phase modulated by a square wave (or the clock pulses). The VCO output of this loop provides a signal coherent with the incoming signal. The clock pulse loop VCO provides a signal which is coherent with the clock pulses modulating the incoming doppler-wiped-out signal. The period of the clock pulse is an integer multiple of the input carrier period, so the coincidence of a positive going pulse out of the differentiators provide the reference signal.

b. During the slave PNG synchronization process, the combined clock pulse and PN modulate the incoming carrier, so the VCO output of the acquisition and tracking phase-lock loop is used to obtain the pulses to drive the slave PNG. The VCO output is coherent with the incoming carrier, but not necessarily at the frequency at which the slave PNG is to be operated. The i-f frequency out of the VCO will be divided down to the operating frequency of the slave PNG, and the reference signal obtained from the acquisition reference generator of Figure 1-9 provides the information required to generate the required clock pulses.

A critical fact assumed here is that doppler is not affecting the time scale of the signal modulation. If $\cos(\omega_c t + \theta(t)$ is the modulated signal of the modulator output on the ground, the received signal will be[12]

$$\cos(\omega_c Xt + \theta(Xt))$$

where

$$X = \frac{\omega_c \pm \text{(residual doppler frequency)}}{\omega_c}$$

For this case,

$$X = 1 \pm 8\text{x}10^{-8},$$

and this means that a unit of time will be scaled by 8 parts in 10^8.

[12] Wainstein and Zubakov, Extraction of Signals from Noise, Prentice Hall, p 370

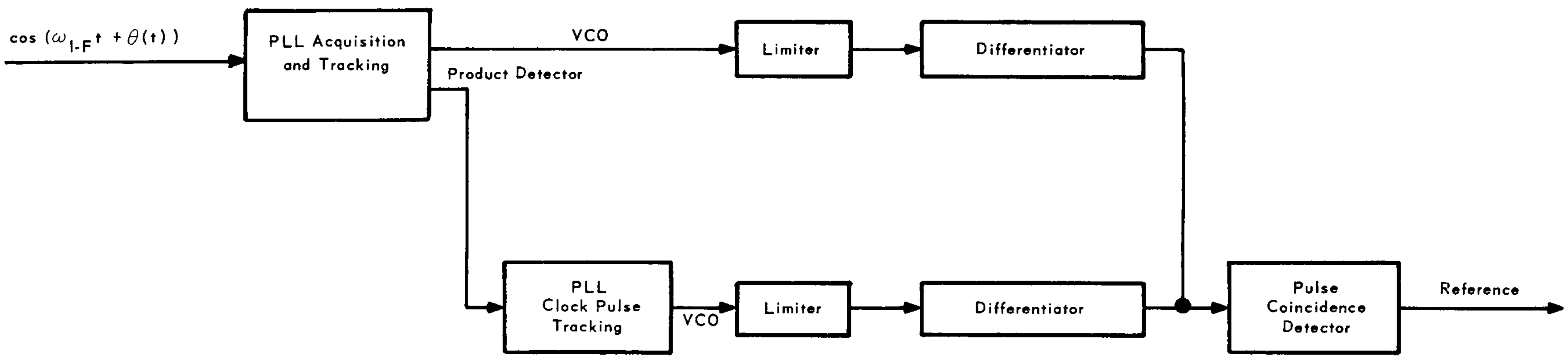


Figure 1-9. Acquisition Reference Generator

1.4.2.2 Ground System

A brief description of the ground system required to operate in conjunction with the vehicle ranging system is given. The ground system functions are: (1) cooperation with vehicle during acquisition, and (2) performance of normal tracking functions with the vehicle.

The VHF link is used during the acquisition for the purpose of information transfer, to coordinate the acquisition functions to be performed by the ground system. Upon receipt of a command from the vehicle, the ground receiver acquires the S-band incoming ranging signal. The phase-lock loop is locked to the carrier of the ranging signal, a slave PNG on the ground is synchronized to the incoming PN modulation, and a clock pulse phase-lock loop is locked to the clock pulse. The clock pulse period is equal to twice the PN bit duration for normal ranging, and this is the ranging signal from the vehicle to which the ground system locks.

The acquisition procedure can be programmed and executed automatically, allowing sufficient time to enhance the probability that the vehicle has completed the operation or it can be done sequentially upon command from the vehicle. The acquisition steps are:

a. Estimate the two-way doppler shift incurred by a signal originating at the vehicle and received at the vehicle. The carrier translated to the assigned ranging channel incorporating the estimated doppler shift at the vehicle is the doppler wiped-out carrier frequency. The doppler wiped-out carrier is phase coherent with the carrier frequency of the ranging signal received at the ground station.

b. The carrier with doppler wiped-out is modulated by the clock pulses and transmitted to the vehicle. The clock pulse period is equal to the basic PN bit duration.

c. Combined PN and clock pulse modulate the doppler wiped-out carrier. The clock pulse period is equal to the PN bit duration as in item b.

d. Combined PN and clock pulse modulate the doppler wiped-out carrier. The clock pulse period is twice the PN bit duration.

e. Using the same modulation as in item d, remove the doppler offset.

Section 2

PHASE-LOCKED LOOP ADVANCED CIRCUIT INVESTIGATIONS

2.1 DESIGN CONSIDERATION FOR TYPE II LOOPS

The Fourth Quarterly Report* on this contract described the use of state-controlled resonator elements to implement loop variable frequency oscillators (VFOs) which could develop a second ideal integration in the loop phase transfer function. The major problem area at that report time was the inability of available ferro-electric capacitor elements to deliver useful or stable amounts of state-control action. Subsequent contact with potential component suppliers, as well as investigation of current research efforts, indicated little industry interest in improving the properties of these materials for applications such as the AROD tracking filters. Although high-quality single crystal elements developed for research investigations demonstrate many desirable properties for this application, the single crystal types were not available in the parameter ranges required.

The breadboard loop developed at the time of the Fourth Quarterly Report was essentially a Type I loop with memory. This memory was implemented with special ferrite multi-aperture core elements which demonstrated adequate state-control resonator properties. At that report time, the two-perfect-integration dynamic stability problem caused temporary abandonment of pure Type II loop action in order to concentrate on the technology of implementing state-controlled resonator elements. Attempts to stabilize the loop, when pure state-control action was actually supplying

*Fourth Quarterly Report, Special AROD System Studies, 7 August 1964, IBM Document Number TR-74-179.

the desired second loop integration, with series network phase-lead equalization techniques did not succeed largely because of gain change difficulties. The state-control action was suppressed for normal loop tracking operation while the memory characteristic of state-control action was present only upon signal drop-out. Successful unaided acquisition for steady (i.e., zero doppler rate) loop inputs was achieved. However, the loop parameters were not optimized at that time since such a course required refinement of the multi-aperture core characteristics. Subsequent efforts again proved disappointing in that materials and device improvement stimulated little supplier interest.

During the time since the last report, the use of minor loop compensation rather than series network compensation was investigated as a powerful method of loop gain variation reduction as well as for the achievment of dynamically stable operation. This technique would linearize and gain-stabilize the overall transfer function of a rather "imperfect" stat e-control element such as the multi- aperture core, if sufficient local loop gain were achieved, and the feedback element was a perfect differentiator (of phase or frequency--depending on where the 1/s term resides in the loop phase transfer block diagram). The burgeoning complexity which appeared to accompany any effective use of state-control elements led to a re-examination of implementation goals from AROD system and new technology viewpoints. New developments in semiconductor devices permitting vastly improved analog storage, coupled with realization that the signal doppler rate (D) has a far greater impact on acquisition and reacquistiion than doppler offset have led to new directions for study.

A study of the AROD system characteristics reveals that if a Type II phase-locked loop is to be operationally useful during signal fading, a perfect memory of doppler offset at drop-out alone is insufficient. Since it is most likely that during a "burn," possibly accompanied by signal fading, a high doppler rate may be encountered, a loop with a perfect memory, remaining at the last known signal position, will be considerably off-frequency when the signal re-appears. Unaided re-acquisition against a still existing doppler rate will then be difficult at best and may require an intolerably high loop natural frequency and loop bandwidth, with a long pull-in time.

This can be further evaluated by means of the analysis for a Type II phase-locked loop given by Viterbi.* The maximum pull-in range, for a signal with a constant doppler rate (D) is given by

$$\Omega_{max} = \zeta \frac{\omega_n^3}{D} \tag{1}$$

with

Ω_{max} = maximum pull-in range rad/sec

ζ = loop damping factor

ω_n = loop natural frequency rad/sec

D = doppler rate ($\dot{A}$) rad/sec^2

In order for the loop to reacquire this maximum pull-in range must be larger than the offset incurred during the signal drop-out, which is

$$\Omega_{max} = DT \tag{2}$$

with T = drop-out time interval in seconds

Combining (1) and (2) gives:

$$\zeta \omega_n^3 = D^2 T \tag{3}$$

This condition for the loop natural frequency is shown in Figure 2-1 where an arbitrary value of $\zeta = 0.5$ is chosen. In this figure the absolute minimum ω_n required for reacquisition is plotted versus the doppler rate, for various drop-out time intervals. In practice one would select a value of ω_n at least a factor of two higher. As a typical example consider a drop-out time of 2.5 seconds while a constant doppler rate of 1000 Hz/sec is present during the drop-out and the reacquisition cycle. The figure shows that a ω_n of at least 600 radians per second is required and thus one may select a value of 1200 radians per second to ensure a stable and relatively fast pull-in.

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*"Acquisition and Tracking Behavior of Phase Locked Loops," A. J. Viterbi, JPL Ext. Pub. No. 673.

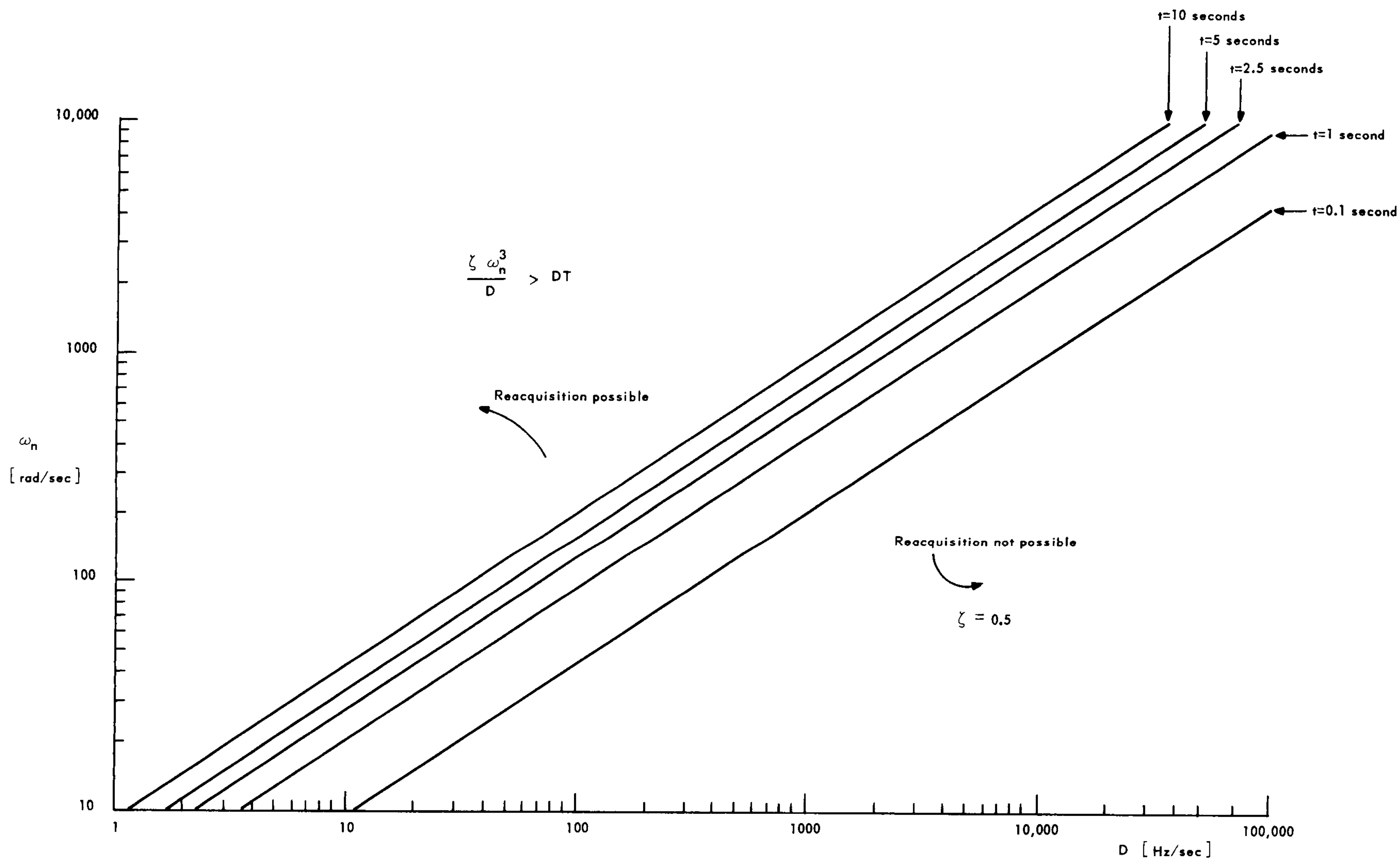


Figure 2-1. Maximum Permissible Doppler Rate vs Loop Natural Frequency

The required value of the loop natural frequency may be far in excess of that necessary to maintain satisfactory tracking performance. This can be particularly true, since with no doppler rate information available, the loop natural frequency has to be increased to allow for the highest anticipated doppler rate in the AROD system.

It is clear that some form of loop velocity (or rate) aiding is highly desirable. This can be accomplished by means of a prediction or estimate of the signal characteristics during the signal drop-out and reacquisition stages. Although this information could be obtained in an accurate form from the associated AROD computation capability, it appears more advantageous to design the phase-locked loop with an additional capability of doppler rate memory, as well as doppler offset memory, to automatically update the loop. This will avoid dependence on external aiding.

If the doppler rate is memorized by analog storage and used to continuously update the loop during the signal drop-out phase only, then the unavoidable error in the doppler rate prediction, and thus in loop VCO frequency, will still require a fairly large loop natural frequency.

A factor of n times reduction in offset at the time of signal reappearance will only reduce the required ω_n by a factor of $n^{1/3}$. If however, the estimated doppler rate information is also used during the reacquisition process, as well as during drop-out, the effective doppler rate incurred is reduced by the factor n and the required loop natural frequency is reduced by a further factor of $n^{1/3}$. Thus, to obtain a factor of 10 improvement in required ω_n, the accuracy of doppler offset and doppler rate estimation is $10^{-3/2}$ or 3 percent. It is anticipated that this accuracy can be obtained even with allowance for imperfect memories with the phase-lock loop design presently under development and described in Sub-section 2.2. It uses signal sampling techniques and a first-order hold between "samples", which for the case of phase-lock loops is between the last lock and the moment of re-lock. This method will obtain a memory of last doppler, a memory of its derivative and a continuous updating of the loop until the next "sample."

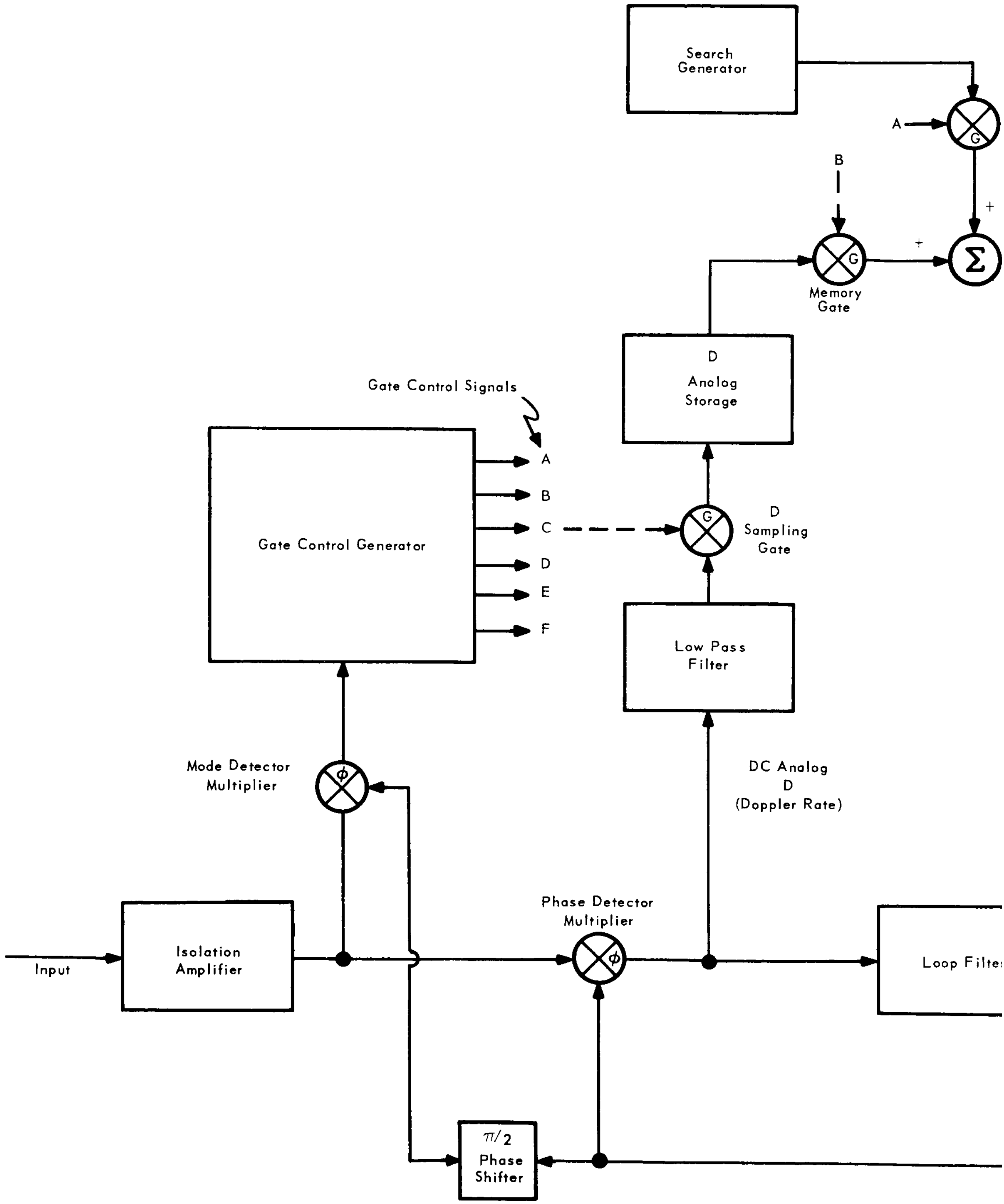
Examination of the AROD system parameters indicates that with the memory and self compensation characteristics, the parameters of the phase-lock loop do not need to be changed between the tracking and reacquisition phases, if the loop is self adapting for signal-to-noise ratio variations.

2.2 DESIGN APPROACH

In order to increase the loop's ability to reacquire rapidly after signal fading, it is necessary to memorize both the last doppler and doppler rate and to continuously update the loop during drop-out and reacquisition. This necessity, outlined in the previous sub-section, can be implemented by two storage devices which are supplied separately with doppler and doppler rate information during the tracking phase of the loop. Upon loss of signal, normal loop operation is discontinued and the voltage controlled oscillator (the output of the phase-locked loop) is supplied with the stored last data of doppler, further corrected by the integral of the last doppler rate. This control, representing an estimate of the signal position, is continued upon reappearance of the signal until phase-lock has been achieved, but is discontinued after a reasonable wait-time indicates no signal reappearance. In the latter event normal acquisition procedure is initiated.

This operation is shown in a block diagram, Figure 2-2, and is based on both the possibility of separating the two required data elements and on the ability to achieve analog storage functions of high quality with newly available state-of-the-art circuit technology. The doppler information is available in analog form as the voltage controlling the loop oscillator. The doppler rate is present in analog form as the loop phase detector (multiplier) output voltage, providing that the loop gain is high enough to make the doppler error negligible. This indicates the necessity of a very low first corner frequency in the loop filter. This can be achieved with the same technology used for implementing the storage function. In addition to the loop and its supporting circuitry, a control function is needed to sense the condition of the loop referenced to the signal and to gate the loop into different modes of operation. A comparison of the loop input with the VCO output in a 90 degree phase-shifted external phase comparator (multiplier) supplies information regarding the loop's condition, i.e., tracking a signal, a signal within range or no signal. The latter condition will initiate, after an appropriate delay, a search cycle which is simply implemented with this approach.

Referring to the block diagram of Figure 2-2, the typical internally referenced phase-locked loop configuration of a phase detector, a filter, a DC amplifier and a VCO is supplemented with a loop gate and a summer preceding the VCO. This gate discontinues the data from the DC amplifier and replaces it with the memory output.



Search Generator
A
G
B
+
+
G
Σ
Memory Gate
D
Analog Storage
Gate Control Signals
A
B
C
D
E
F
Gate Control Generator
G
D
Sampling Gate
Low Pass Filter
DC Analog
D
(Doppler Rate)
Mode Detector Multiplier
φ
Phase Detector Multiplier
φ
Input
Isolation Amplifier
Loop Filte
π/2
Phase Shifter

2-7-1

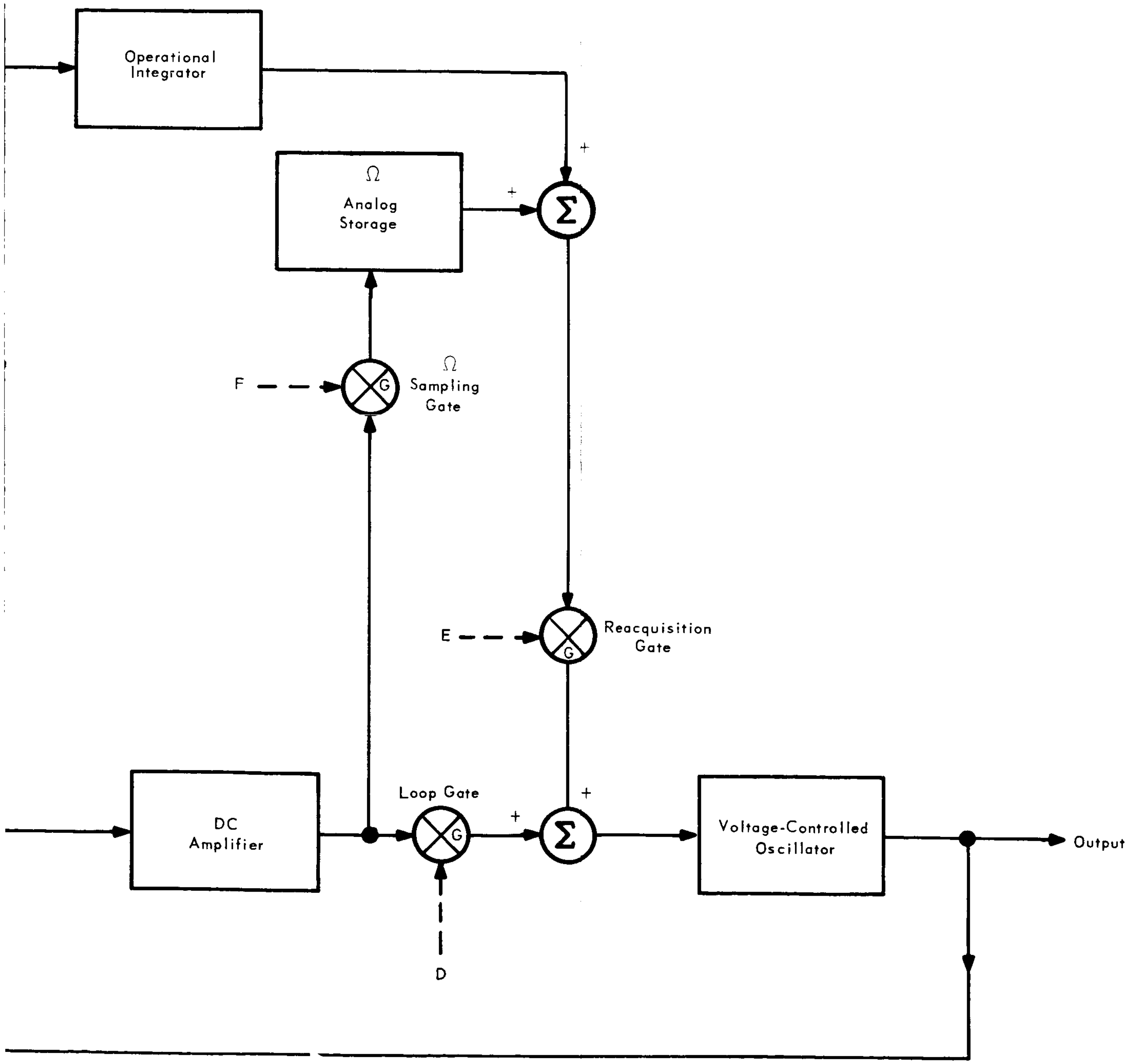


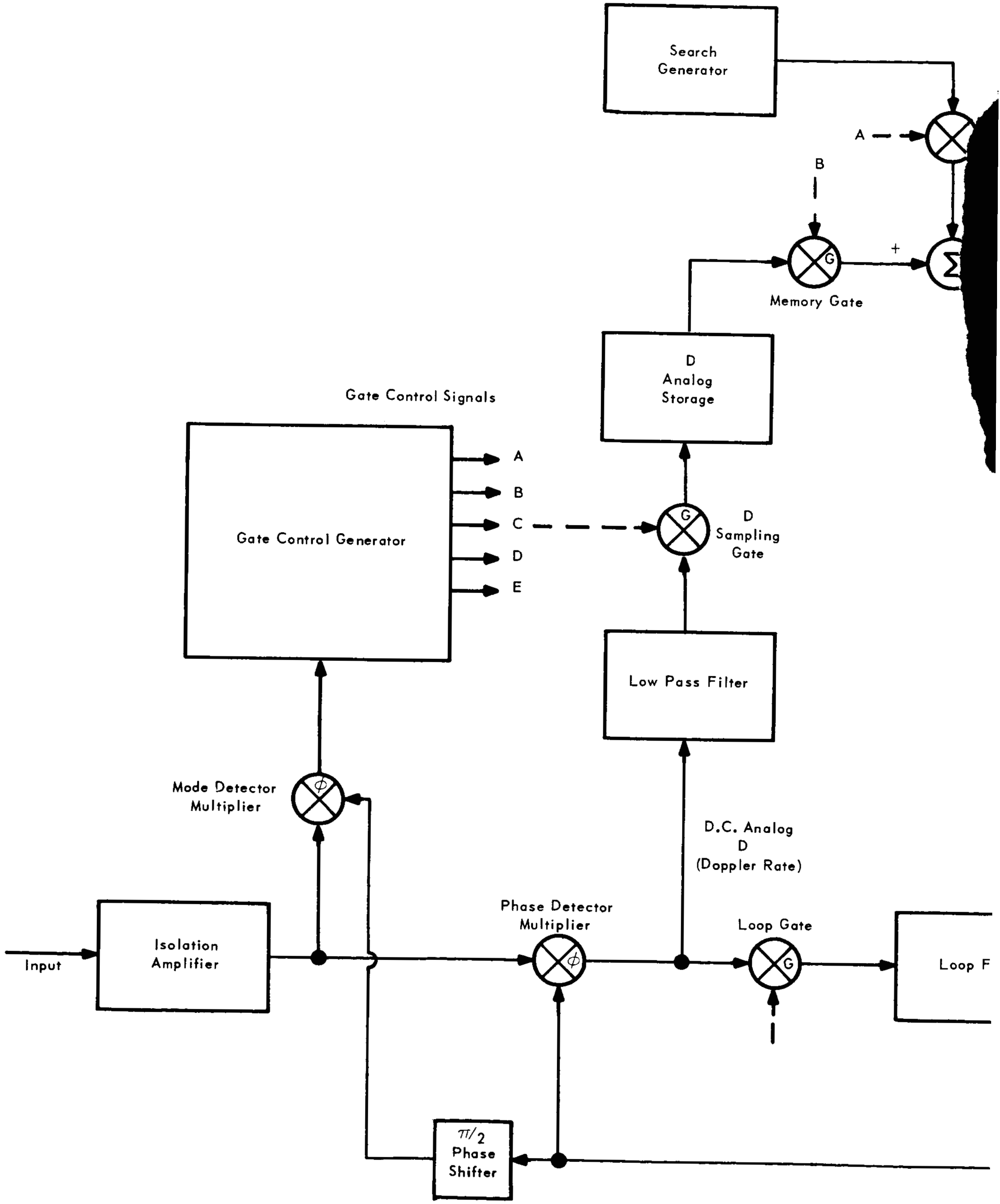
Figure 2-2. Memory Aided Phase-Locked Loop

A sense of the condition of the loop is available by a phase comparison of the input signal to the loop with a 90 degree phase shifted loop oscillator output. The comparator output is supplied to a Gate Control Generator which generates signals to establish the different modes of operation of the loop.

The output of the DC amplifier is connected during the track mode to an analog storage which continuously supplies doppler information to a summing device. The output of the loop phase detector is filtered and applied to an analog doppler rate storage during the track mode. This data is integrated during memory and reacquisition and added to the doppler data in the summing device, which at that time feeds the VCO. Integration is obtained by an operational integrator circuit. Search can be added at the input of this integrator to obtain a linear search and can be reversed in direction by applying a square wave to the integrator.

A shortcoming of this "brute force" method is immediately apparent. If after the sequence of loss-of-signal, memory and reappearance of signal the loop finally reacquires, the majority of the data controlling the VCO is then supplied by the memory devices and only a correction is supplied by the loop itself. Disconnecting the memory circuits at this time would result in a sudden and very sharp shift in the frequency of the VCO. This would cause an immediate un-lock with no chance of a recovery in a reasonable time of the lost doppler offset and thus of the signal. Instead of an abrupt discontinuation of memory-aiding, the data (charge on the storage capacitor) present in the memory must be slowly transferred to the integrator (filter) of the loop. The rate of transfer is not only limited by the device itself, but to a certain extent by the maximum doppler rate which the locked loop can track. This transfer procedure is essentially identical to an additional internally generated doppler rate. This causes a limitation in the loop characteristics in that a second signal fading cannot be tolerated until after some settling time.

To circumvent this difficulty, and to simplify the circuitry one can make use of the storage device already available in the loop. Particularly, since the doppler offset information is stored in the loop filter, a separate doppler memory circuit is not necessary. Thus the generalized system diagram of Figure 2-2 simplifies to that shown in Figure 2-3 where now upon signal drop-out the normal loop is broken at a point preceding the loop filter. This filter essentially serves two purposes. During



Search Generator
A
B
+
Memory Gate
D Analog Storage
Gate Control Signals
Gate Control Generator
A
B
C
D
E
D Sampling Gate
Low Pass Filter
Mode Detector Multiplier
D.C. Analog D (Doppler Rate)
Phase Detector Multiplier
Loop Gate
Input
Isolation Amplifier
Loop F
π/2 Phase Shifter

2-9-1

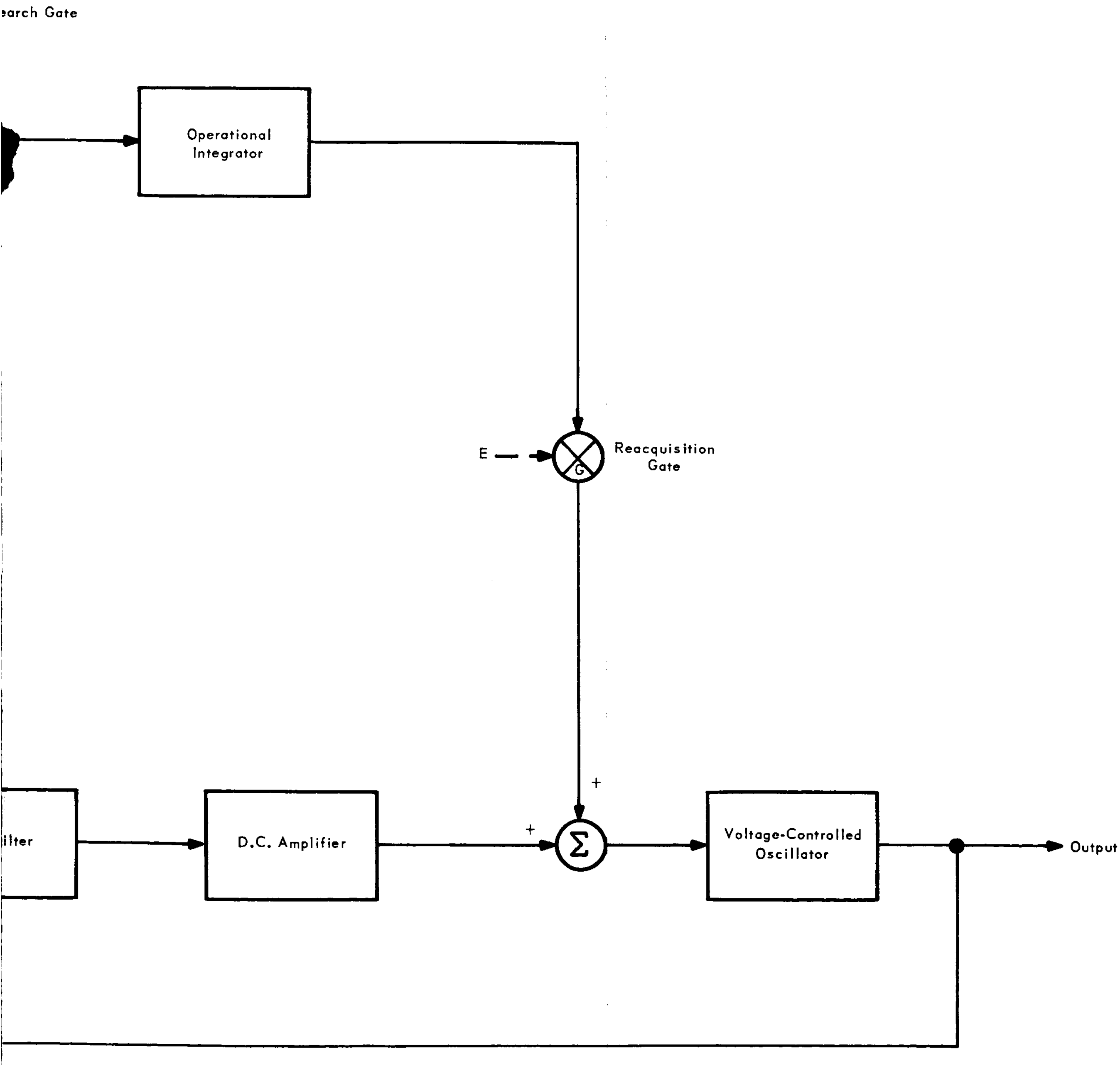


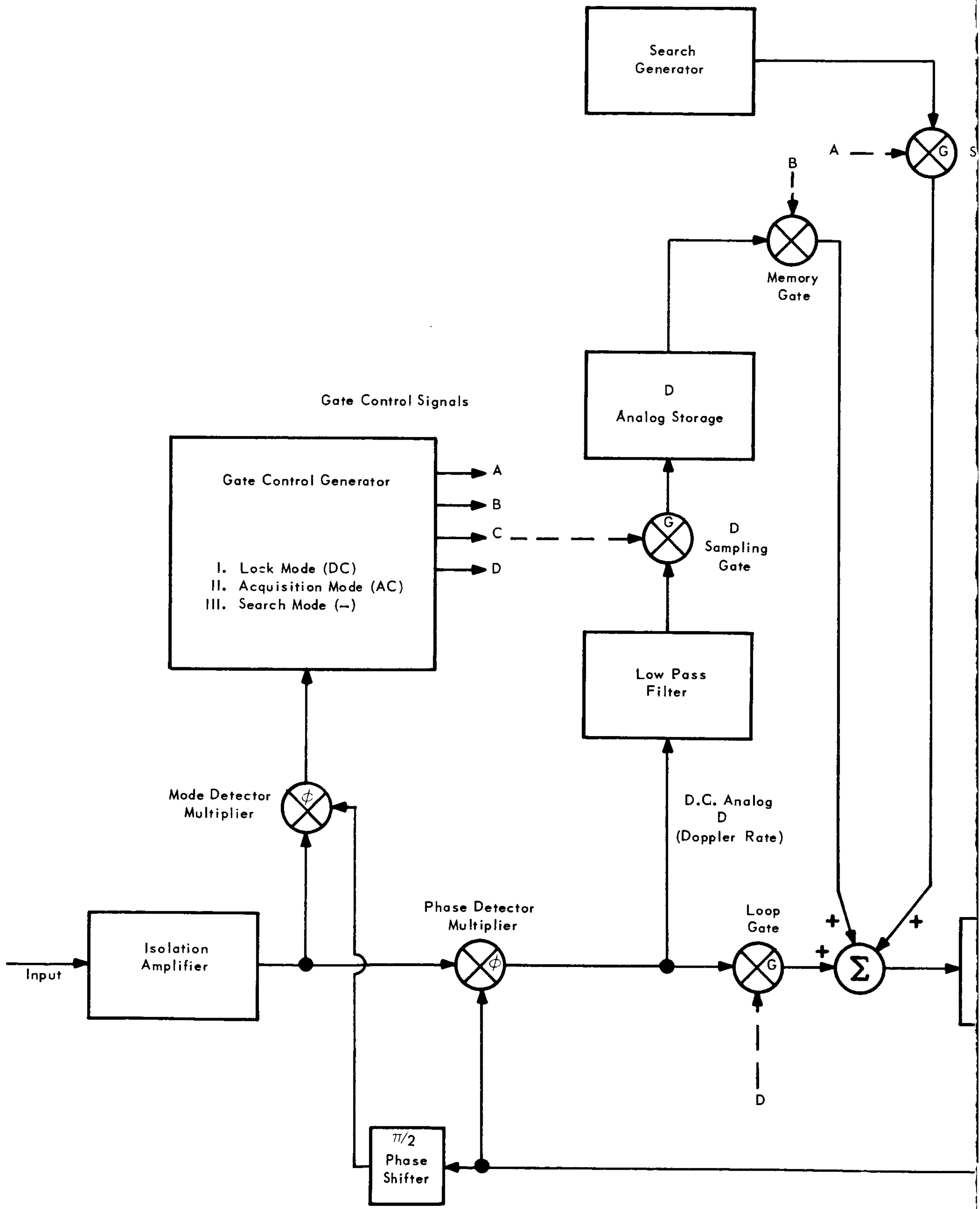
Figure 2-3. Simplified Memory Aided Phase-Lock Loop

the track mode, it determines the characteristics of the loop behavior, while during the memory and reacquisition modes it serves as a memory device for storing the last known doppler offset.

Although the problem of transferring the data of a large, last known doppler offset has been eliminated, a considerable doppler correction can build up during the signal drop-out period, which when applied to the VCO input will still have to be transferred. This leads to consideration of a further simplification by assigning an additional function to the loop filter, i.e., that of the operational integrator used to correct the loop's position with a doppler rate derived signal change estimate. This is shown in Figure 2-4 which represents the selected design approach. The stored doppler rate data is now gated into the loop filter during the memory and reacquisition mode and can be removed safely after lock-on is achieved, since it will be immediately replaced by a shift in output of the phase detector, generated by a phase skip of the VCO frequency.

This type of phase-locked loop, which uses a multi-function integrator and which could be called a "reflex phase-locked loop", has a clear advantage over typical loops. This is because it can remember both the last doppler offset and the last doppler rate. It also automatically self-corrects its output to reflect estimated signal changes during signal drop-out. It contains the apparatus to sense the loop's condition, lends itself to any required search cycle and has flexibility in the selection of the basic loop parameters.

The philosophy used to acquire a signal after a signal fade by a compensation of both doppler and doppler rate and based on the inability of the loop to pull in and lock against high doppler rates, can be considered for the search mode where no a priori data is available. Rather than sweeping the loop in a sawtooth or linear-flyback fashion, which can prevent signal lock-on in some applications, the loop can be swept by searching for both the doppler offset and the doppler rate. This is similar to techniques used for search radars. It can be achieved by superimposing upon a linear (or sawtooth) sweep a voltage which simulates the range of possible positive and negative doppler rates. A superimposed sine wave (or sawtooth) voltage is a sufficient doppler rate range simulation. This would enable the loop to acquire signals of any unknown parameters or dynamics as used for spacecraft maneuvering.


Search Generator
A
G
S
B
Memory Gate
D
Analog Storage
Gate Control Signals
Gate Control Generator
I. Lock Mode (DC)
II. Acquisition Mode (AC)
III. Search Mode (–)
A
B
C
D
G
D Sampling Gate
Low Pass Filter
D.C. Analog
D
(Doppler Rate)
Mode Detector Multiplier
Isolation Amplifier
Input
Phase Detector Multiplier
Loop Gate
+
+
+
Σ
D
π/2 Phase Shifter

2-11-1

earch Gate

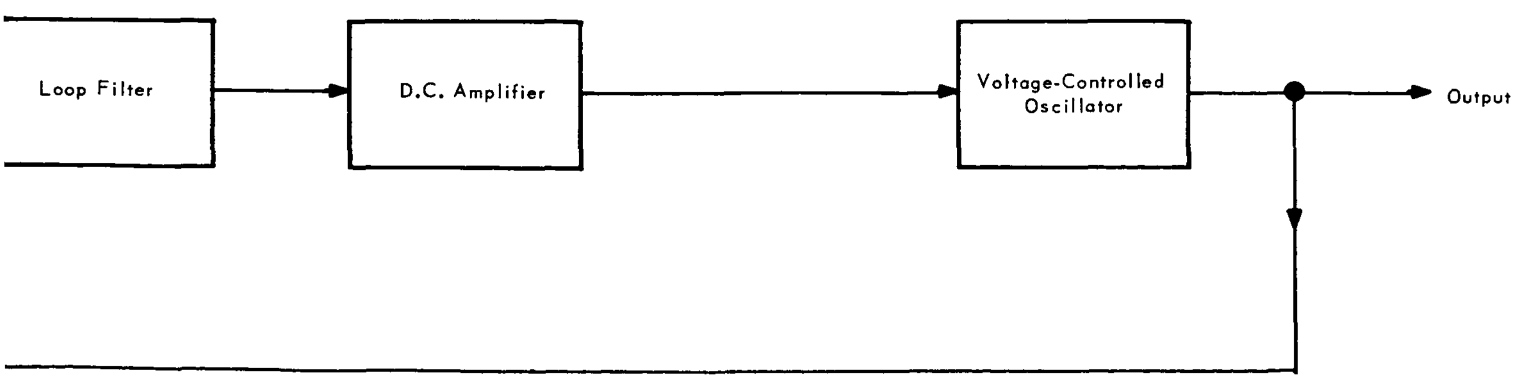


Figure 2-4. Reflex Phase-Lock Loop

For the loop under development, however, a simple sawtooth sweep will be used to better evaluate the main function of the phase-locked loop.

A description of the essential circuitry used to implement the required functions can be given. The limiter, phase shifter and the two multipliers are previously developed designs requiring a stable balance of the loop multiplier as a primary goal. The four gate circuits, driven by the gate generator, are connected to memory devices and therefore need a high isolation in the closed position. Use of metal-oxide-silicon field effect transistors (MOS-FETS) fulfills the isolation and gating requirement. The gate control generator is supplied with the output of the mode detector. A DC voltage indicates a locked condition. A low frequency beat indicates the presence of a signal in proximity. No detector output indicates absence of a signal within a reasonable passband. This passband, obtained by the use of a low pass filter, is selected as a balance between maximum noise rejection and filter delay time when searching. The four output control voltages are derived simply with several OR circuits using conventional computer logic techniques, while a time delay is incorporated to prevent an immediate search on loss-of-signal. The loop filter is combined with the DC amplifier to increase the effective value of a storage capacitor. A MOS-FET device is incorporated at the input of the DC amplifier to obtain a high isolation between input and output terminals. Operational amplifier techniques, using integrated circuits, make the loop natural frequency independent of this amplifier gain since the loop time constant changes simultaneously with this gain.

2.3 STORAGE IMPLEMENTATION

The primary implementation problem in achieving the reacquisition (and also acquisition) capability of the "Reflex" phase-lock loop configuration outlined in Subsection 2.2 is related to analog storage implementation with MOS-FET devices. It should be noted that the storage could easily be implemented in terms of digital storage elements of suitable bit precision (seven bits would provide better than 1 percent resolution) followed by D/A conversion. While such an approach would permit inclusion of the on-board AROD computation capability to aid both acquisition and reacquisition operations of the tracking loops, unaided operation would require the implementation of an A/D conversion operation, with penalties assignable to the loops themselves. Furthermore, implementation of both A/D and D/A converters makes wide use of analog gate elements, as required in the selected design approach, a similar necessity.

The analog storage approach avoids all the complexities of digital storage, while still permitting computer-aiding through a single time-shared (by all tracking loops) D/A converter by exploiting the extremely low leakage of MOS-FET elements used as analog multiplexer gates. The primary advantage of MOS-FET devices in implementing precise, long-term analog storage is a consequence of their effective gate leakage resistances of the order of 10^9 megohms. This permits use of capacitors with similarly low leakage resistances which are relatively small in capacity. The capacities would be of the order of 0.1 ufd. A further advantage is that field effect devices such as MOS-FET inherently have zero-offset voltages in contrast to conventional bipolar junction transistors. This property is of great value in simplifying the required analog signal gating functions indicated for the design approach of Sub-section 2.2.

Primary effort will be directed at implementing the analog storage circuits as required by the "Reflex" loop configuration. This technique will also meet the special requirements of the loop filter inherent in this approach. These elements, along with a functionally simplified Mode Detection and Gate Control Generator, will be imposed upon an existing Type I loop (with increased gain) in order to study reacquisition capability.