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Tracking, Telemetry, and Command Studies of Ground and Satellite Subsystems

Volume IV: Airborne Transponder Configuration Study

DECEMBER 1963

Prepared by D. MATSON
Telecommunications and Tracking Department

Prepared for COMMANDER SPACE SYSTEMS DIVISION
UNITED STATES AIR FORCE
Inglewood, California

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FOREWORD

The staff of the Telecommunications and Tracking Department has for some time been engaged in supporting the command and control requirements of a variety of Air Force satellite programs. This support consists of analysis and design of the ground and airborne subsystems which perform the three main satellite support functions, i.e. tracking, telemetry, and command.

Because of the varied experience members of the department have gained from support of the many Air Force satellite programs, they have prepared a series of six volumes, entitled TDR-269(4110-01)-13, Tracking, Telemetry, and Command Studies of Ground and Satellite Subsystems, which is believed to present a new synthesis of the technical material in the field of advanced integrated tracking, telemetry, and command systems.

The six separate volumes of the report are titled as follows:

Volume I	Summary
Volume II	A Comparison of Several Ranging Systems Designed for Military Space Applications
Volume III	Acquisition of Space Vehicles by Ground Tracking Systems
Volume IV	Airborne Transponder Configuration Study
Volume V	Ground Station Antenna Survey
Volume VI	Parametric Analysis of Tracking, Telemetry, and Command Subsystems

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ABSTRACT

The report discusses the system requirements which reflect upon the design of CW watellite transponders. The Mariner prototype transponder and the Goddard Range and Range Rate Transponder are taken as currently operational examples of the state of the art. These two transponders are described in detail. Their primary advantages and disadvantages are noted.

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1.0 THE MARINER TRANSPONDER STUDY

The following section gives a brief description of the Mariner L-Band and S-Band communications systems. The second section provides a discussion of the system requirements of the S-Band communications system design and the last section deals with the details of the basic S-Band transponder.

1.1 Evolution of the Mariner Communications System

On the last Mariner flight, which was the Venus fly-by experiment, the communications system (see Figure 1) consisted of the L-Band transponder (890 mc/s up, 960 mc/s down) feeding either of two 3-watt power amplifiers. One 3-watt power amplifier feeds an omnidirectional antenna array and the other feeds a four-foot, high gain parabolic antenna. The signal from the ground is received through the omnidirectional command antenna and is fed directly to the L-Band transponder. The 25-watt power amplifier shown in Figure 1 was not carried on this mission.

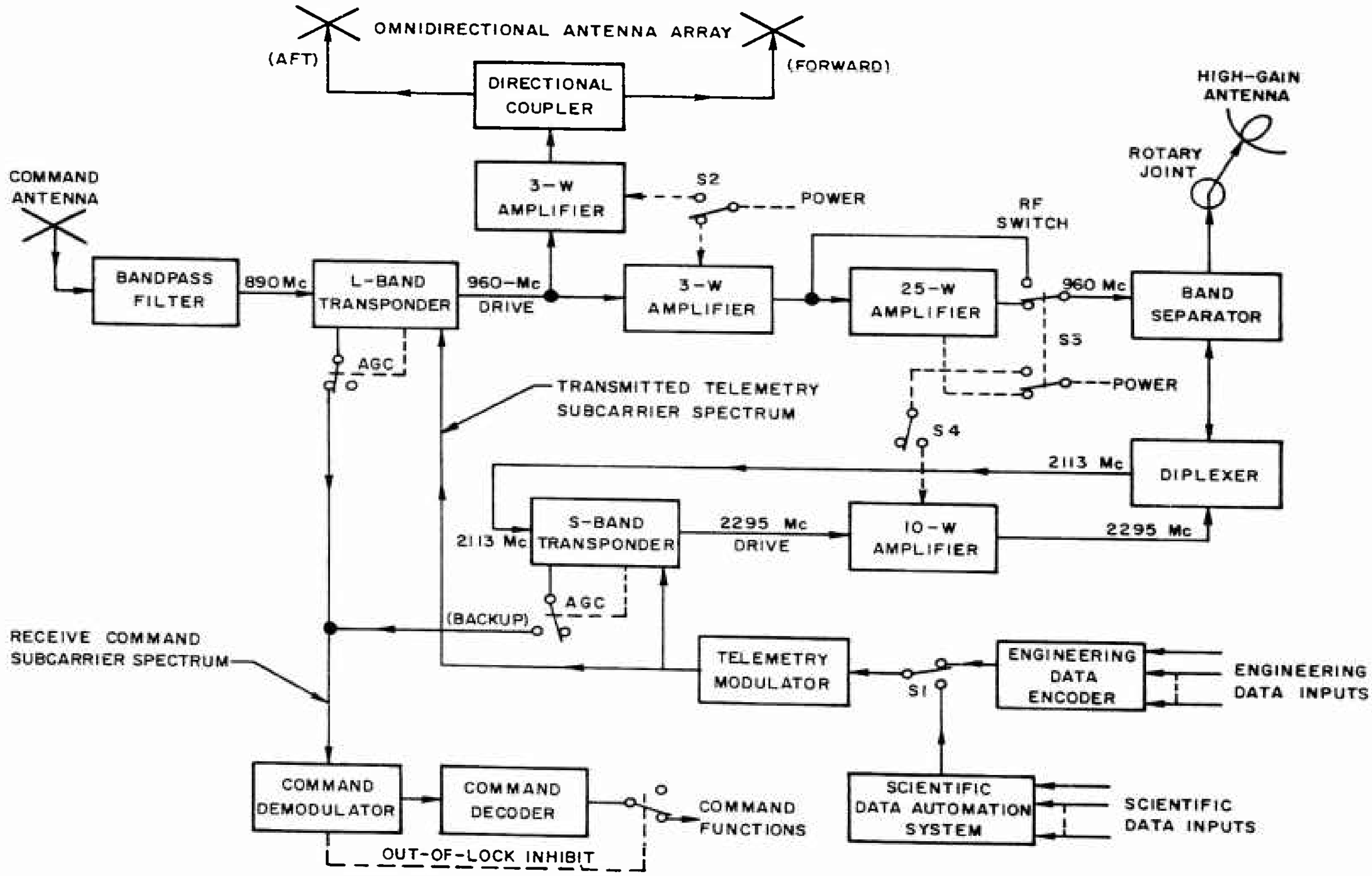

Figure 1. Mariner L-Band Transponder System Functional Block Diagram

An additional transponder, the preliminary engineering model of the present S-Band unit, was carried to obtain performance evaluation data. It was used only for short periods during the mission.

A prototype of this transponder is at present completing its development stage to replace the L-Band system. It is assumed that the command, ranging, and telemetry subsystem will also take on some modification before the next flight. The expected general configuration for the S-Band transponder is shown in Figure 2.

1.2 Communications System Design Constraints

This section will discuss the design constraints imposed upon the Mariner communications including comments about their influence upon the resulting subsystems and their application to the integration of the SSD tracking, telemetry and command systems.

1. A primary system design boundary imposed upon communications was that it must be completely compatible with the Deep Space Instrumentation Facility (DSIF). This implies that the carrier be either phase or amplitude modulated, and that the modulation spectrum occupy only those bands utilized by the DSIF receivers. Additionally, in order to perform automatic angle tracking and receive two way doppler information, the modulation must not completely suppress the carrier.

While economics plays some part in the realization of this constraint, the primary reason is to freeze the present ground environment in order to allow a more orderly development of the space systems and other aspects of the ground environment (namely reliability) while complying with a severe schedule. This comment appears quite pertinent relative to the SCF which must satisfy program requirements which dictate rapid change of the ground environment.

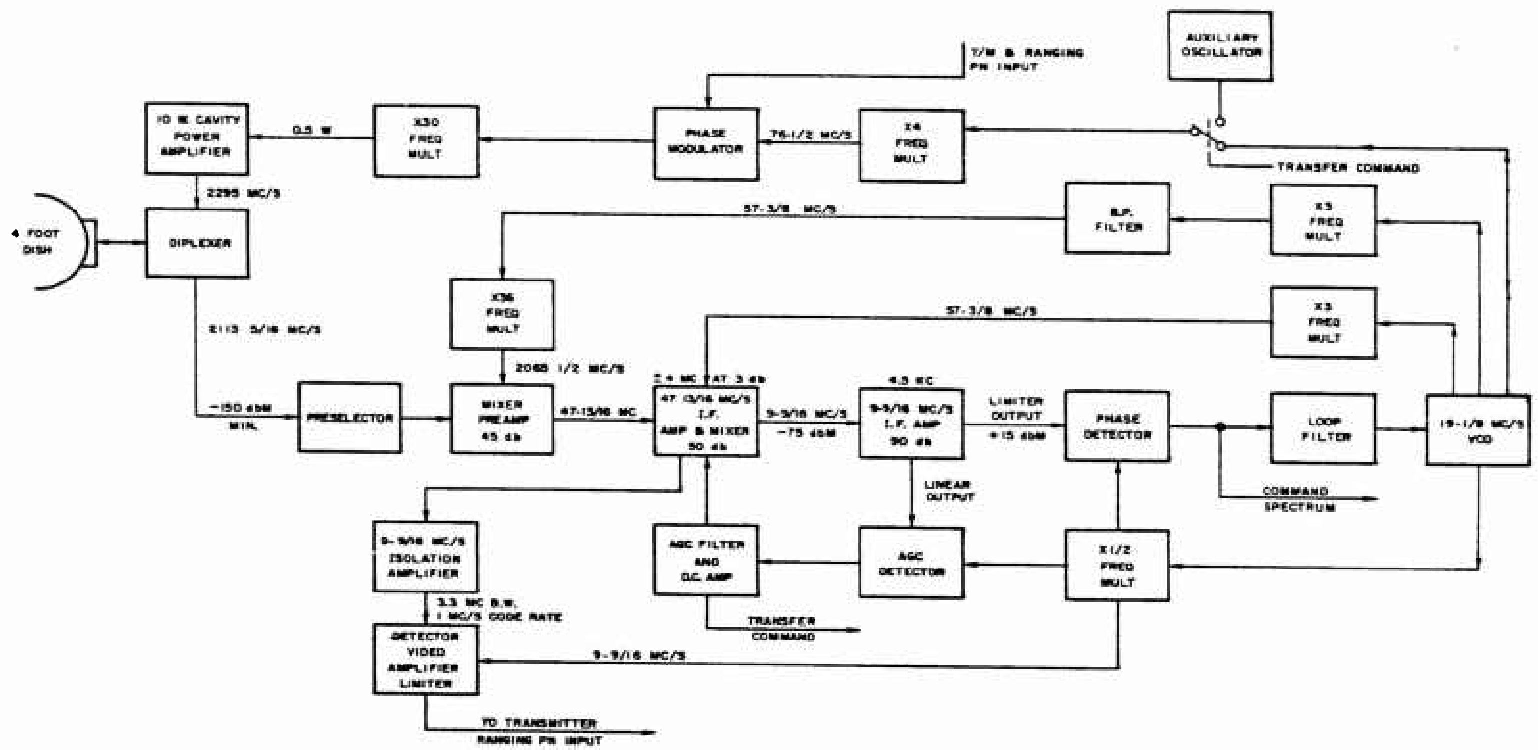

Figure 2. Mariner Lunar Ranging Transponder

2. It has been required of the DSIF that by 1 January 1963, it vacate the L-Band frequencies of 890 and 960 mc/s and commence operating at S-Band frequencies of 2113 mc/s and 2295 mc/s, up and down respectively. For this reason an S-Band experimental transponder was included in the Venus fly-by experiment.

3. Monetary costs cannot be ignored and therefore a definite limitation exists upon the elaborateness with which the system (ground and space) can be mechanized.

4. Studies of the available power generated by solar panels indicates that the output is a function of the trajectory. If the spacecraft is approaching the sun the longer communication range would be compensated by a higher solar panel power output (vice versa for a trajectory away from the sun). The Mariner type mission can assume continued illumination of their craft by the sun; however, this assumption is usually not true for SSD missions, especially for low altitude satellites. The trajectories encountered on Mariner Missions, in conjunction with the level of efficiencies of solar power systems, lead to a limitation on transmitter output power levels of about 25 watts at this point in the development of the state-of-the-art.

5. It appears that the deep space telemetry systems must have a variable data rate capability due to the great variation of communications range during a mission (up to 270 db one way loss at 2300 mc/s for a Venus or Mars mission). At any one of the data rates available, one would desire to match the transmitting power requirements as closely as possible to the solar panel output and at the same time minimize energy storage requirements, switching, and peak power transients. These considerations imply two results: continual telemetry transmission and the choice of phase modulation of the rf carrier.

The continual telemetry transmission can be shown to provide maximum information transfer to the ground for fixed energy output of the spacecraft solar panels. This mode of operation would be unsatisfactory for Air Force missions due to extended vehicle-earth-station eclipse time and the necessity for security.

The peak power in a phase modulated wave is constant regardless of variations in modulation index, type or rate of information being transmitted. For this reason it provides the most efficient use of the power available from solar panels.

In addition to the power matching requirement more efficient use can be made of the CW transmitter designs available when the peak and average power requirements are within a fraction of each other. Both continual transmission and phase modulation of the rf carrier are techniques which reduce the difference between peak and average power requirements.

6. After powered phases of a mission, such as launch or mid-course maneuver, it is highly probable that the vehicle high gain antenna will not be properly oriented to communicate with the ground control station. It is important that communications and control are maintained during these and other unstabilized conditions. Therefore, the maximum use must be made of vehicle omnidirectional antennas out to maximum range. These same constraints apply to earth satellites.

7. At some point in a space mission, depending on the range and vehicle power, communications must switch to a vehicle antenna with a directivity greater than isotropic. Use of a high gain antenna requires the proper operation of an antenna orientation mechanism.

It would seem probable that the reliability of this mechanism would not be commensurate with the command system bit error probability of 10^{-5}. Therefore, the command communications channel will be so designed as not to depend on the use of a high gain antenna. The command system will be designed to operate using only the omni antenna out to the maximum range that active control is required.

8. JPL favors the complete separation of vehicle diagnostic data handling from the scientific or payload data handling. The bandwidths of the various diagnostic data are known quite accurately before launch and would lend themselves easily to fixed formatting and sampling rates. Scientific experiments, however, may yield data much different from that anticipated and as a result much more flexibility and complication is needed to obtain the maximum amount of information. This argument is applicable to the SSD problem and might imply the greater standardization of diagnostic telemetry systems.

9. The system is confined to subcarrier modulation of the rf carrier due to the use of simple phase lock demodulators in the DSIF. In order to maintain stable lock in the rf carrier loop the instantaneous phase error must be limited to less than 1 radian (a somewhat arbitrary but practical definition). For this modulation index, about 50% of unmodulated carrier power remains in the carrier.

10. The Mariner communications will use only one rf carrier. This constraint derives from the desire for maximum efficiency of use of spacecraft power. While there may exist special cases of multiple signals yielding improved primary power utilization, such implementations have to the present time required increased hardware complexity. In the usual case, the use of a single carrier is dictated by a desire to maximize the ratio of sideband or information power to carrier power.

In maintaining that a single carrier can be implemented, one is assuming that a carrier can be placed outside the information bandwidth, detected, multiplied to the proper frequency and used to demodulate the information sidebands.

11. Digital techniques are chosen, over analog, for use in all JPL systems. This choice is justified by: a) the relative simplicity of the data processing and handling problem especially where large quantities of data must be contended with, b) the difficulty of somehow source coding analog data to render it more impervious to noise, and lastly, c) the use of digits allows the use of superposition of synchronization and data to obtain an optimum communication technique[1].

12. The choice of the modulation system for any future JPL system will be PSK at $\pm 90°$. This type of modulation complements the requirement of paragraph 11 for digital techniques. PSK at $\pm 90°$ also allows the simultaneous satisfaction of paragraph 9 (subcarrier modulation of the RF carrier) and paragraph 1 (two way doppler tracking).

13. The system uses the PSK at $\pm 90°$ in a pseudo random coded form. The PR code is easily generated and has the properties of orthogonal codes. Appendix B gives theoretical and measured data which indicates that the system so implemented for Mariner approaches some sort of an optimum condition. Additional advantages accrued by the use of a PR scheme are:

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1 Viterbi, A. J., "On Coded Phase Coherent Communications", PG SET, Pg. 3-14, March 1961. See Appendix A

a) Ease of bit synchronization acquisition.

b) Bit and word sync as a by product.

c) Interference rejection inherent (13 db for the 50 bit/second command system).

d) Near optimum communication efficiency when used with PSK.

1.3 The Basic Vehicle Transponder

The ground transmitted signal at 2113 5/16 mc/s is received by the four foot spacecraft dish and fed through the diplexer to a receiver preselector (see Figure 2). The spacecraft basic transponder is a double conversion receiver with a carrier tracking phase lock loop. The pseudo-random (PR) ranging code is extracted after the second mixer and further processed by the ranging channel. The command PR code spectrum is demodulated by the carrier tracking phase lock loop and transferred to the command loop for further processing. A coherent frequency sample from the carrier loop VCO is multiplied and then modulated by the PR ranging and telemetry spectra before further multiplication to the transmitting frequency of 2295 mc/s.

1.3.1 System Pseudo Random Signal Spectra

The form of the spectra which the system is designed to handle is shown in Figure 3. The spectra are approximate in that the frequency component amplitudes have not been calculated. Note that the telemetry (down link) PR spectrum is relatively narrowband (87% of signal power within 2100 cps) with high power (45% of unmodulated carrier power) and conversely, the ranging PR code spectrum is relatively wide band (87% of signal power within 1 mc/s) with high power on the up link and low power on the down link. The command or telemetry signals (C/T signals) bi-phase modulate a subcarrier and this signal is linearly and incoherently added to the ranging code. The ranging code bit rate is one megabit per second. The relative amplitudes of the bi-phase signal and the ranging code signal are adjusted to yield the relative powers for the composite signal modulated on the carrier as indicated in Figure 3. Under this arrangement, the interference to the C/T signals due to ranging code spectra in the C/T band is negligible due to the narrow C/T bandwidth. Furthermore, the interference rejection due to C/T detection processing is an additional 13 db for a 50 bit per second

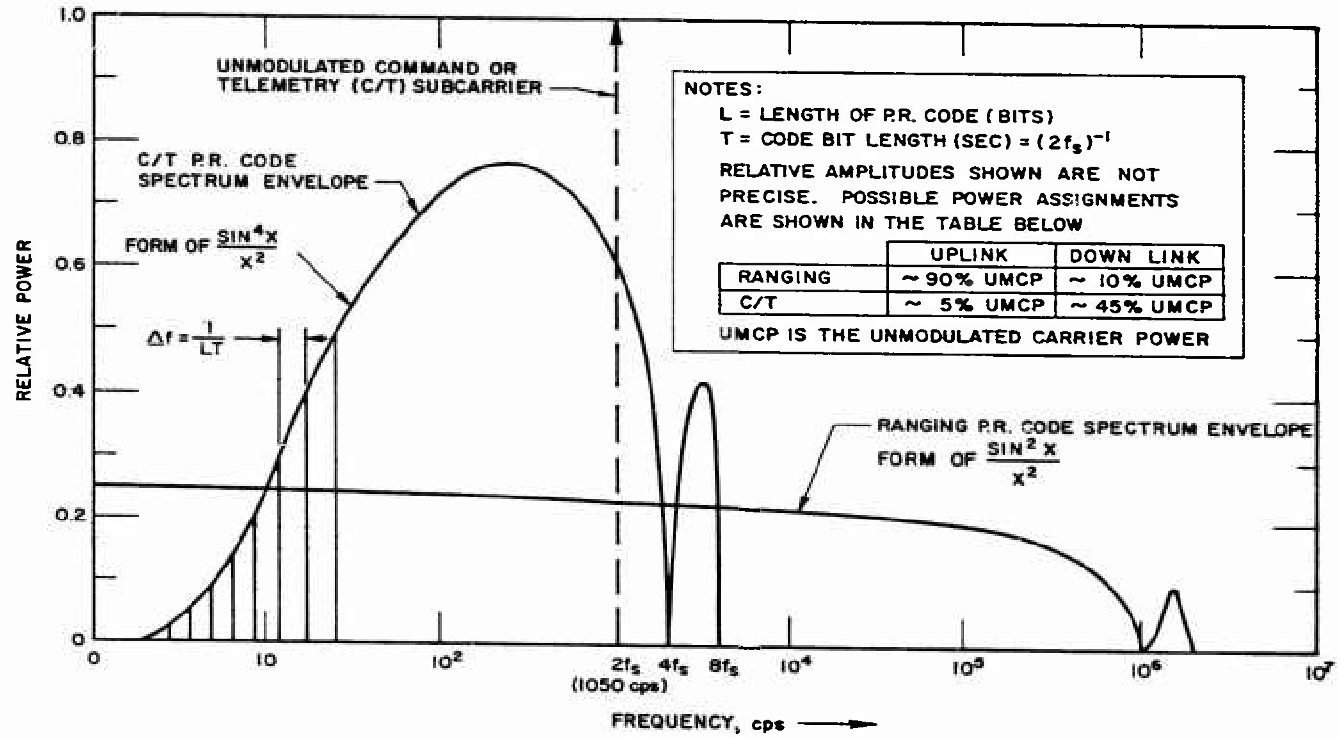

Figure 3. Approximate Spectra for Mariner Transponder

information rate and 1050 cps transmitted code rate. Conversely, the interference to the ranging signals by the C/T spectrum is eliminated by a high pass filter in the ranging video amplifier to attenuate the C/T power. The loss of ranging power in this bandwidth (2100 cps) is entirely negligible. Specifying the ranging processing gain is a problem when one attempts to assign a measure of information to the received code.

1.3.2 Receiver Thresholds

For the received command information the receiver carrier coherence threshold is defined, in this case, as carrier power equal to phase noise variance in the loop noise bandwidth, $2B_{Lo}$ or 20 cycles per second. At room temperature, the receiver input noise ($N = KTB$) is -174 dbm in a 1 cycle/second bandwidth or -161 dbm in the 20 cycle/second loop noise bandwidth at the loop threshold. With a system noise figure of 11 db maximum the receiver sensitivity is -150 dbm (see Appendix A). The command channel acquisition threshold has been determined empirically to be about 6 db higher than the coherence threshold (according to H. Donnelly of the JPL receiver group). These thresholds have been determined by experiment. In the case of these thresholds white noise and signal are applied to the system input at a controlled power ratio (S/N) and the initial frequency offset, pull in time or drop out time are measured.

Referring to Figure 2, it is noted that the ranging spectrum is extracted after the second mixer, detected to video, limited and remodulated on the down link carrier. The only threshold in the vehicle receiver for this channel is that due to the suppression of the signal by the limiter at low signal to noise ratios.

1.3.3 Coherent Frequency Multiplication and Mixing

No rf amplification is provided in the receiver; instead the composite signal from the preselector is mixed with a reference at 2065 mc/s and then amplified in the preamplifier (see Figure 2).

Originally the 2065 mc/s reference was obtained by multiplying the loop VCO frequency of 19 1/8 mc/s by 108 and then applying it to the mixer. It is known that the output of the frequency multiplication process contains strong subharmonic and harmonic content (see Figure 4), and these harmonics were causing spurious lock of the carrier phase lock loop. To prevent this situation, a +40 db signal to interference (S/I) margin at the input to the carrier loop phase detector has been found necessary. As a result of the frequency multiplication, the first mixer reference contains VCO harmonics every 19 1/8 mc/s. The most critical frequency of 2103 3/4 mc/s was only about 30 db below the desired reference at 2065 1/2 mc/s. This harmonic, when mixed with the incoming rf signal at 2113 5/16 mc/s produces a 9 9/16 mc/s component which feeds directly through the first if amplifier and mixer, is amplified in the pass band of the second if and causes spurious signal lock. Notice that the 2122 7/8 mc/s component also mixes with the received signal to form an output at 9 9/16 mc/s. Additionally, as the multiplier harmonics become appreciable an image frequency is generated every 19 1/8 mc/s across the preselector bandwidth. These images mix with the harmonics on the reference and produce 9 9/16 mc/s components which, though of third order, will add coherently with the second order interfering components. These problems are eliminated by additional filtering in the multiplier chain and this is accomplished by the band pass filter following the x3 multiplier. Spurious harmonics on the first mixer reference are now down at least 70 db from the desired frequency. This attenuation appears to be sufficient to prevent false lock from these effects.

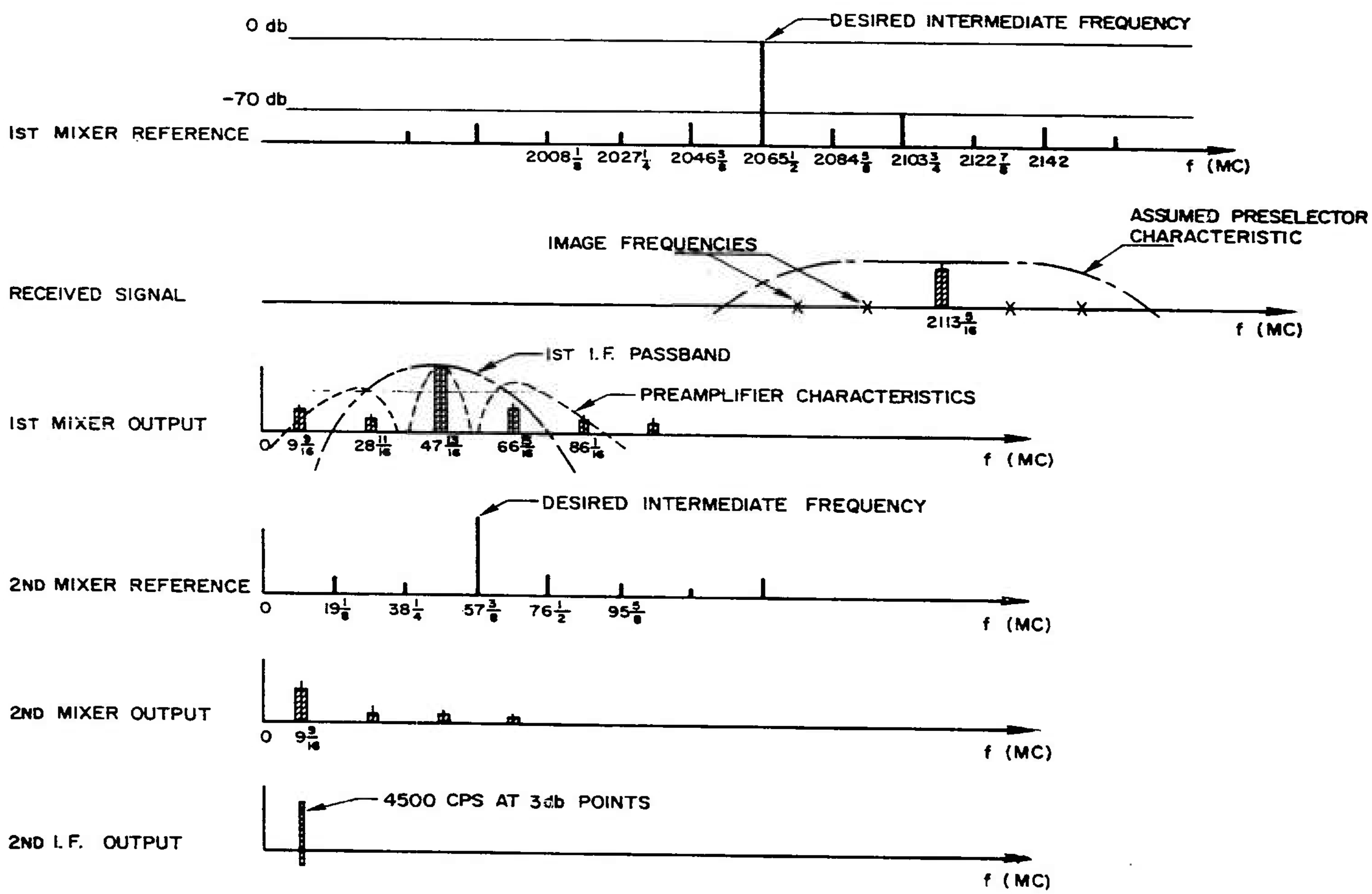

Figure 4. Internal Coherent Harmonic Generation

Another way in which the first mixer reference component at 2103 3/4 mc/s can cause false lock is probably more serious than that described above. Here the strong 2065 1/2 mc/s and 2103 3/4 mc/s combine in the first mixer to form a 38 1/4 mc/s component which is well within the pass band of the preamplifier and also within the skirts of the first if. After this amplification the component is mixed with the 47 13/16 mc/s (see Figure 4) desired spectrum to form a 9 9/16 component which is amplified in the second if providing a strong undesired coherent signal to the phase detector. To alleviate this problem, notch filters are included in the preamplifier providing about 35 db rejection at 38 1/4 and 57 3/8 mc/s (2122 7/8 mc/s mixes with 2065 1/2 mc/s to produce 57 3/8 mc/s) which in addition to the 30 db or so added attenuation in the multiplier chain is sufficient to eliminate false lock.

Apparently the desired signal at the output of the first if amplifier has been sufficiently filtered so that the harmonics of the second mixer reference cause no problems.

In brief, for a system in which all the frequencies are coherent and there exist many nonlinear elements which provide the opportunity for mixing of these frequencies, careful filtering becomes a major task.

1.3.4 Intermediate Frequency Amplifiers and AGC

The incoming 2113 5/16 mc/s is mixed with 2065 1/2 mc/s to form the first intermediate frequency of 47 13/16 mc/s (see Figure 2). The second mixing with 57 3/8 mc/s produces the second if at 9 9/16 mc/s. At this point the channel splits; the signal going to an isolation amplifier and on through the ranging channel, and alternatively through the carrier phase lock loop to the command detection system. The characteristics of the transponder amplifiers are given in Table 1.

TABLE 1

TRANSPONDER AMPLIFIER CHARACTERISTICS

Preamplifier		
Center Frequency		47 13/16 mc/s
3 db Bandwidth		7 mc/s
Gain		45 db
Filtering:	Notch filters at 38 1/4 and 57 3/8 providing 35 db of rejection to spurious mixer harmonics.	
First if Amplifier		
Center Frequency		47 13/16 mc/s
3 db Bandwidth	Spec.	8 mc/s
	Actual	10 mc/s
Gain		50 db
Dynamic Range (AGC)	Spec.	+50 to -70 db
	Actual	+50 to -80 db
Second if Amplifier		
Center Frequency		9 9/16 mc/s
3 db Bandwidth		4.5 kc/s
Gain		90 db
Filtering:	Input filter to limit bandwidth to 4.5 kc/s	
Ranging Channel Isolation Amplifier		
Center Frequency		9 9/16 mc/s
3 db Bandwidth		3.3 mc/s
Gain		0 db
Isolation		110 db

Figure 5 indicates the transponder gain profile. The preamplifier operates over -150 dbm to -50 dbm while supplying 45 db of gain. The first if amplifier incorporates AGC in its first three stages which controls the gain from +50 to -70 db. The second if amplifier at 9 9/16 mc/s accepts synchronous gain controlled inputs at -76 dbm and its output maintains a saturated state in the limiter. The input stage of this amplifier is a two pole crystal filter with a highly linear phase characteristic and a bandwidth of 4500 cps which sets the over-all predetection bandwidth. Originally a three pole filter was used here; however, spurious responses through the filter resulted in a severe degradation of the S/N ratio. The 4500 cycle bandwidth of this filter is determined by the requirements of the command system whose spectrum is shown in Figure 3. The succeeding stages of amplification providing about 90 db gain are wideband to introduce a minimum of additional phase/frequency distortion. The two outputs from the second if amplifier are:

a) A linear signal for the synchronous AGC and

b) A limited signal to be used as an input to the carrier PLL.

The limiter is followed by one stage of amplification whose output is about +15 dbm (signal plus noise) which is approximately the desired input to the loop phase detector.

Using the coherently derived reference from the PLL the AGC operates upon the linear signal plus noise output from the second if. The synchronous AGC has the advantage of not acting to decrease the signal power when the noise power increases (does not act as a limiter with its attendant suppression effect). The linear output of the AGC DC amplifier is fed back to the first stages of the first if amplifier. A second binary output provides a transfer command which switches the transmitter frequency drive from the VCO output to an

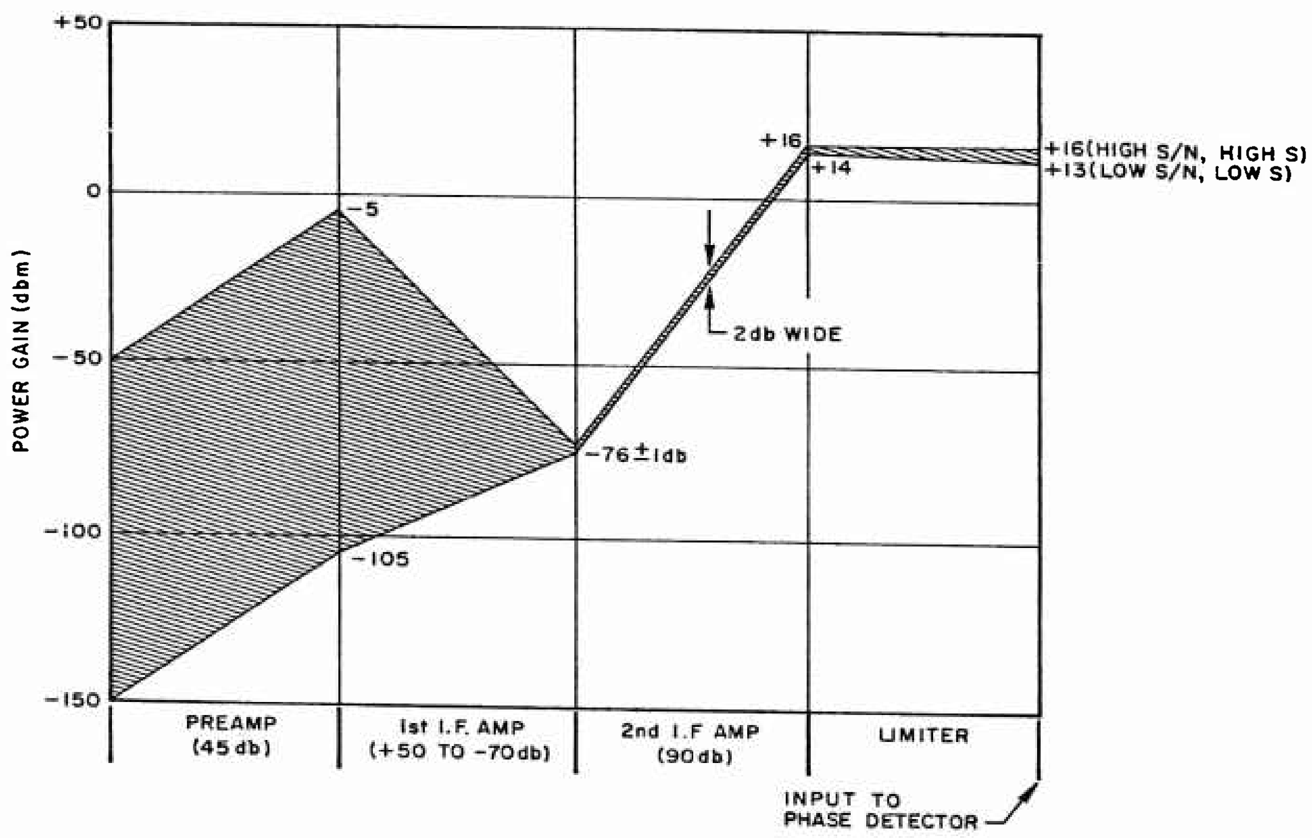

Figure 5. Carrier Power Gain Profile

auxiliary oscillator when the received carrier is below threshold. This allows the transmission of telemetry information to the ground independently of whether or not a signal is being received.

1.3.5. Carrier Phase Lock Loop

The carrier phase lock loop is a second order servo which tracks the incoming carrier, develops a coherent frequency for other receiver processing, and translates the command spectrum to baseband for detection by the command digital loop.

The sole variation in this circuit from the usual PLL is that the VCO frequency is double the if input frequency to the phase detector. This frequency choice is to prevent another, and probably the most serious of all modes of self lock. If the output of the VCO were 9 9/16 mc/s (see Figure 2), instead of 19 1/8 mc/s, the second mixer reference would be obtained by a x6 frequency multiplier. A 9 9/16 mc/s spurious frequency component possibly at -30 to -60 db from the VCO output level, will exist at the output of the multiplier. This component will feed directly through the second mixer with little further attenuation. With 90 db gain through the second if amplifier, it is not difficult to visualize conditions where this component will cause trouble. It is possible that the power of this amplified component can be so greatly in excess of any received signal power that the loop will remain permanently locked to this component and reject any received signal. In the present configuration a -40 db signal to interference margin is being required at the input to the phase detector. The only false lock phenomenon to be observed has been that due to transient ringing of the 4500 cps crystal filter in the second if amplifier due to excessively high input noise power. There has been no observation of false lock due to excessive time delay around the two mixer reference loops.

This loop has been optimized at a phase variance threshold of 1 radian and a threshold loop bandwidth of 20 cycles. Under conditions of strong signals the loop bandwidth will increase to a limit of approximately 450 cps and will afford increased capability to track higher doppler rates at close ranges.

In addition to the mixer references the VCO supplies, after division by 2, the 9 9/16 mc/s references for the loop phase detector, the synchronous AGC and the ranging spectrum detector. The phase detectors used in the phase lock loop and the synchronous automatic gain control (SAGC) are the same circuit. These phase detectors have two thresholds; one, (a), being the point at which the (S/N) is so low that the output is no longer a function of the signal input, the second threshold, (b), is that point of operation where the output is no longer a <u>linear</u> function of the input. These thresholds are stated in terms of input (S/N). The second threshold, (b), is the higher of the two at about (S/N) = -29 db and is the one of interest in this transponder.

The first threshold, (a), is the point of operation where the residual DC output from the detector due to the input signal and noise becomes much greater than the output due to the input signal. This DC residual null output is apparently a function of the input noise power and diode mismatch.

The second threshold, (b), is a function of the phase detector balance drift over the operating temperature range and will exist for any diode selection since they can be matched at one temperature only.

With a receiver threshold (S/N) of -23.5 db (measured in a 4500 cps bandwidth), the nonlinearity threshold, (b), is 6 db further down and should be no problem with this transponder. This threshold does set a definite upper limit on the if bandwidth realizable with the given PLL bandwidth.

1.3.6 Phase Locked Receiver Tracking

It is of interest to estimate the receiver tracking errors and obtain an idea of the receiver stability mechanism. A simplified diagram of the frequency linearized receiver is shown in Figure 6, where modulators are replaced by frequency adders and only the second if amplifier is of interest.

Since there is no fixed stable reference oscillator in the receiver it might appear, after cursory inspection, that difficulty would be encountered in preventing the receiver second if from stabilizing on the skirts of the narrow input filter. This condition would result in distortion and attenuation of the desired signals.

The initial carrier frequency acquisition is accomplished by sweeping the ground transmitted carrier over the range of uncertainties at the vehicle receiver. At some point in this range of sweep the PLL locks to and follows the received carrier variations. It is assumed that the ground transmitted frequency returns to the initial point after the acquisition routine. At this point, the receiver is momentarily locked to the incoming signal, f_1. It is desired to determine whether a) the second if frequency, f_3, will fall within the 4500 cps passband and b) whether the receiver frequencies, f_2 and f_3, will drift to a stable frequency outside the 4500 cps if pass band.

At any instant the second if, f_3, is related to the received carrier, f_1, and the VCO output, f_2, by

$$f_3 = 6f_2 - (f_1 - 216\ f_2) \qquad (1)$$

At some point in the acquisition f_2 became equal to f_3, then

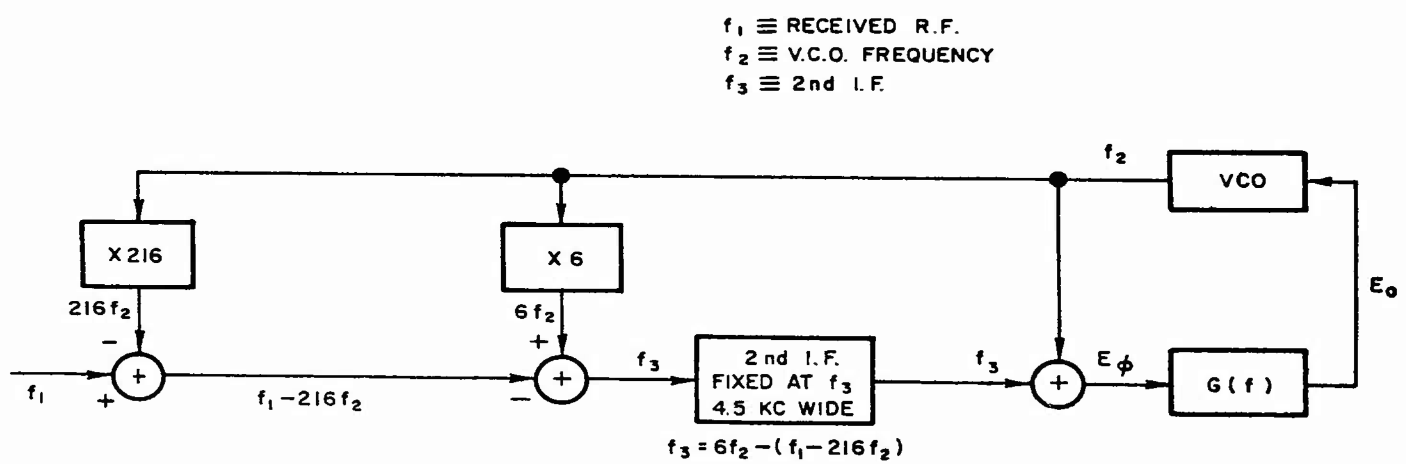

Figure 6. Linearized Tracking Receiver

$$f_3 = 6f_3 - (f_1 - 216f_3) \text{ or } f_3 = \frac{f_1}{221} \quad (2)$$

The if offset, Δf_3, due to any error in the received frequency (due to doppler, oscillator instability, etc.) is,

$$\Delta f_3 = \frac{\Delta f_1}{221} \quad (3)$$

In other words, received errors are reduced by a factor of 221 in the second if. This amounts to about a 450 cps uncertainty due to a 100 kc/s doppler uncertainty. Consequently the received signal is being tracked to within ± 450 cps in the 4500 cps second if pass band.

Note that if the VCO were replaced by a fixed oscillator with no PLL, the Δf_3 at if equals Δf_1 at rf and there is no effective division of frequency shifts to guarantee that f_3 stays in the if pass band. Additionally, if a fixed reference oscillator is used with a VCO in the receiver, as is done in some systems, there is no effective division of Δf_1; however, now the input signal is translated perfectly to the selected if. The system utilizing the fixed reference oscillator is more complex than necessary for the Mariner application.

To estimate whether or not the receiver drifts out of the if pass band after acquisition, assume the VCO frequency, f_2, is perturbed by a small noise, Δf_2. The second if is now,

$$f_3 + \Delta f_3 = 6(f_2 + \Delta f_2) - \left[f_1 - 216(f_2 + \Delta f_2)\right] \quad (4)$$

Combining eq. (1) and (4) a restoring signal at the if output is,

$$\Delta f_3 = 222\, \Delta f_2 \quad (5)$$

from which the high receiver frequency gain in the reverse direction is apparent. The loop phase detector output, E_ϕ, can then be approximated by,

$$E_\phi = K_\phi \left[\Delta\phi_3(s) - \Delta\phi_2(s) \right] \tag{6}$$

where $\Delta\phi_3(s) - \Delta\phi_2(s)$ is the input phase difference, and $K\phi$ is the detector gain constant. Since

$$\Delta\phi_3(t) - \Delta\phi_2(t) = \int \left[\Delta f_3(t) - \Delta f_2(t) \right] dt \tag{7}$$

then
$$\Delta\phi_3(s) - \Delta\phi_2(s) = \frac{\Delta f_3(s) - \Delta f_2(s)}{s} \cong \frac{\Delta f_3 - \Delta f_2}{\omega_2} \tag{8}$$

Equation (6) is now, with $2\pi f_2 = \omega_2$,

$$E_\phi = \frac{K_\phi \; (\Delta f_3 - \Delta f_2)}{\omega_2} \tag{9}$$

The input voltage to the loop VCO, E_o, is

$$E_o = E_\phi \, G(\omega) = E_\phi G(2\pi\Delta f_3 - 2\pi\Delta f_2) = E_\phi \, G(\Delta\omega)$$

$$= \frac{K_\phi \; (\Delta f_3 - \Delta f_2) \; G(\Delta\omega)}{\omega_2} \tag{10}$$

where G(s) is the loop filter transfer function. Combining Equation (5) with (10),

$$E_o = \frac{221K_\phi \, G(\Delta\omega) \, \Delta f_2}{\omega_2} \tag{11}$$

E_o is applied to the VCO in such polarity as to make Δf_2 approach zero. In the above, the receiver is assumed to be dynamically stable (no poles in the right half plane).

1.3.7 The Transponder Ranging Channel

At present JPL is concerned with three types of ranging:

a. Skin tracking or passive reflectors.

b. Turnaround transponders where the signal is basically amplified and placed on a different down link carrier.

c. Recoder transponder where the received ranging code is detected and a different code is transmitted to the ground.

The Mariner S-Band transponder falls into the b. category.

The total received spectrum is taken from the output of the second mixer (see Figure 2) at 9 9/16 mc/s and fed through an isolation amplifier. Reverse isolation of 110 db is necessary at this point to prevent the 9 9/16 mc/s reference, used in the ranging detection, from feeding back into the second if amplifier band pass and causing receiver self lock as described in the previous section. The isolation amplifier characteristics are tabulated in Table 1.

To avoid double phase modulation of the down link carrier and attendant difficulties of detection on the ground, the ranging spectrum output from the isolation amplifier is phase detected to baseband. In order to control the modulation index of the ranging spectrum upon the down link carrier the detector output is amplified and limited.

The binary output of the limiter is linearly added to the telemetry PR spectrum and applied to the phase modulator. The relative amplitudes of the signals are controlled to give about the same relative power in each signal as shown in Figure 3; namely: 45% of unmodulated carrier power in the telemetry signal and about 10% in the ranging signal.

It is pertinent to note the difference between the ranging channel and the command channel and their relation to the carrier tracking phase lock loop. Both the ranging and command detectors in the vehicle process binary symbols and therefore phase nonlinearity and distortion do not degrade the signals so long as phase polarity information is retained with an unchanged probability of error. By contrast, the carrier tracking channel (first and second if, PLL, etc.) maintains carrier frequency error control and to do so must maintain phase linearity. These facts imply that the amplification and detection circuits for the ranging and command channels can be far simpler than the circuits needed to linearly amplify and detect carrier frequency variations over large changes in vehicle range. For this reason the wideband ($\pm$1.8 mc/s) ranging spectrum is fed through the separate channel described above. The command channel is carried along with the carrier tracking channel since the command bandwidth is fairly narrow ($\pm$2.1 kc/s) and it is convenient to recover the demodulated command baseband at the output of the PLL phase detector. Furthermore, it is apparently no design problem to open the carrier tracking channel bandwidth from a bandwidth required by carrier doppler rate to one which accommodates the command spectrum.

If elimination of the separate ranging channel (isolation amplifier, detector, video amplifier, limiter) is attempted by widening the carrier tracking channel to about 4 mc/s, the receiver performance is degraded. As the bandwidth is increased the carrier to noise ratio in the if, $(C/N)_{if}$ is decreased by the ratio of the initial to the final if bandwidth.

This is about 30 db. The input (C/N) to the phase detector is now about -53.5 db which is far below the required -29 db linearity limit noted previously. Additionally, the total (S+N) power is increased by 30 db and, since the second if incorporates 30 db of linearity margin, the second if is at saturation. If the gain-bandwidth product of the second if is somewhat constant the if gain will be decreased approximately proportional to the increase in bandwidth. Any loss of this type will dissipate the 20 db receiver gain margin and if in excess of 20 db will raise the threshold sensitivity of the receiver. However, in this receiver, the gain-bandwidth product is probably not constant since the input crystal filter sets the bandwidth of the second if while all other stages are wideband.

1.3.8 <u>The Transponder Transmitter</u>

The composite spectrum from the phase modulator is frequency multiplied by 30 to give a down carrier of 2295 mc/s at .5 watt and applied to the vehicle power amplifier. In past flights this amplifier has been a 3 watt unit. In future flights either a cavity amplifier supplying 10 watts at rf (developed by Res Del, Pasadena) or an "Amplitron" supplying 20 watts at rf (developed by Raytheon, Mass.) will be incorporated into the transmitter.

2.0 THE GODDARD TRANSPONDER STUDY

2.1 System Performance and Objectives

The missions which the Goddard RARR system is designed to support are the near earth highly elliptical orbiting satellites. The applicable orbit parameters extend to 100,000 kilometers apogee with 100 kilometers perigee, circular or elliptical. Under these conditions the vehicle velocities and accelerations for both long and short range operations present fairly broad and exacting requirements to the tracking system.

The objectives of the tracking system are:

a) Transportable ground equipment to perform ranging simultaneously from three accurately positioned ground stations.

b) Simple and foolproof in concept.

c) Uncomplicated operation and calibration.

d) Dependable performance.

The measurements to be performed on the satellite in the orbits noted above are:

a) Range, using 7 range tones on each of three channels, with a precision of ±15 meters.

b) Range rate to a precision of ±.1 meter per second using 2 way coherent carrier doppler.

c) Angle tracking, using phase and amplitude monopulse error signals, to ±.1 degree.

d) Time resolution to within 10 μsec of WWV.

Short acquisition times are required for compatibility with low altitude satellites. In general, the system must provide instrumental accuracies which are comparable in magnitude to present knowledge of the propagating medium and the velocity of light.

According to the Motorola Engineering Analysis Report No. W2719-2-1, Revision 1, dated 23 November 1962, the system presents an integrated approach to the tracking, command and data communication problems associated with space vehicles; however, the report only mentions the command and telemetry problem. All analysis in the report is directed toward tracking considerations. Motorola has performed a minor study of the problem (Motorola TM 2719-6-5) in a manner which is not partiularly applicable to the SSD integrated approach to TT&C systems.

2.2 System Description [2]

A simplified block diagram of the over-all R and RR system is shown in Figure 7. The system consists of a VHF ranging subsystem and a separate S-Band ranging subsystem. The two are virtually independent.

The S-Band system provides precise tracking while avoiding the propagation anomalies of lower frequencies. The VHF is an optional aid to S-Band antenna acquisition since the use of VHF is required on future NASA vehicles. Use of VHF frequencies also allows smaller transponders. The S-Band subsystem is capable of independent operation and therefore the following discussion will be concerned solely with the S-Band ground system and transponder.

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2. Extensive discussion of the over-all system may be found in the references cited at the end of the section.

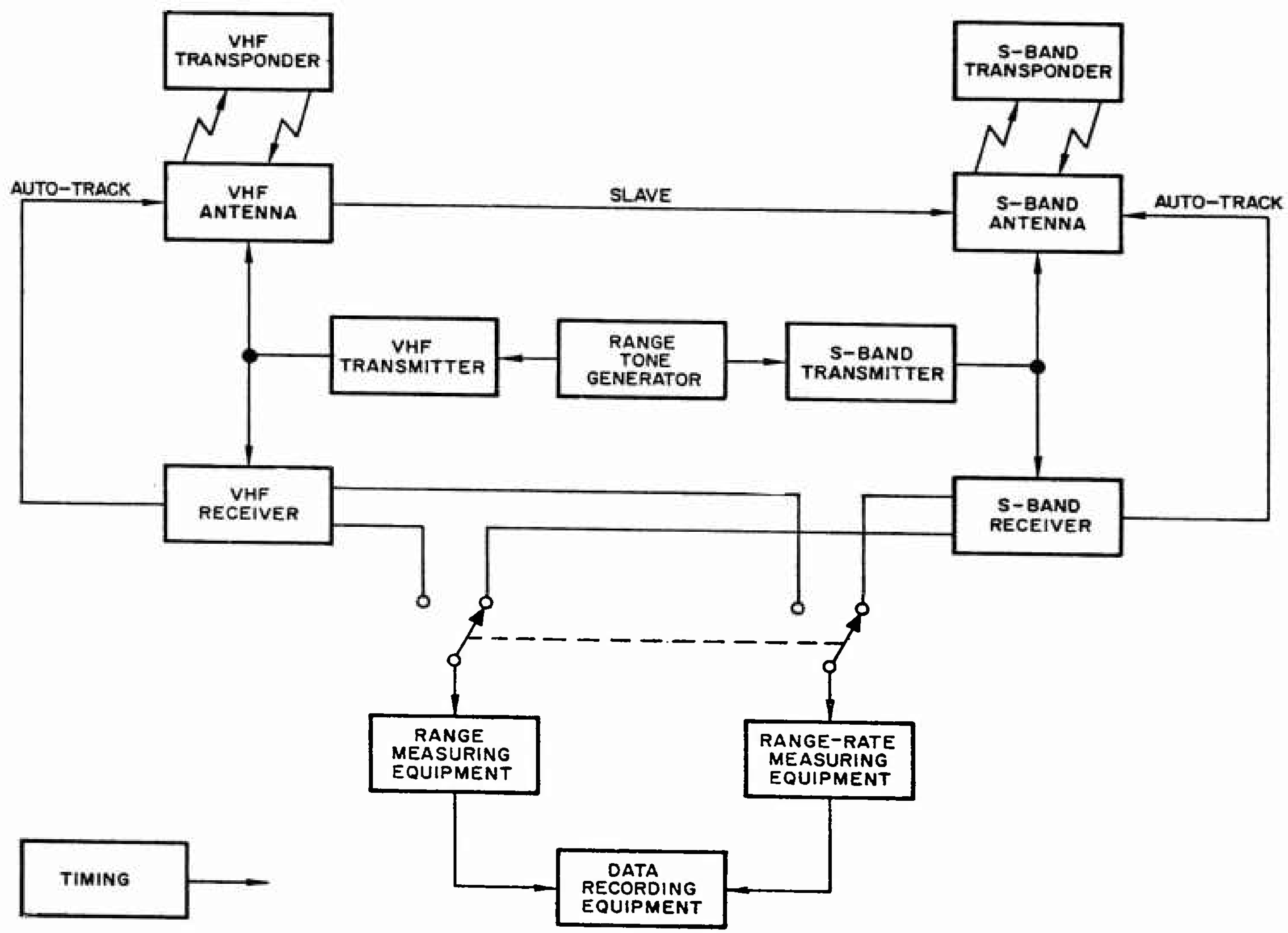


Figure 7. Simplified System Block Diagram

The S-Band transponder, as presently fabricated, includes the capability for simultaneous operation with two ground stations out to a certain limited range. Originally, three separate ranging channels were to be provided which would allow an instantaneous fix by trilateration. Since there appears to be no application for such a technique in the foreseeable future, one of the three channels was eliminated. This leaves the transponder as a 2-channel redundant unit. Table 2 shows the various transponder operating modes. Though the first models of the transponder will have only two channels, the system design constraints are based upon a three channel transponder. For this reason three channels will be referred to from time to time. Each mode effectively uses the total transponder bandwidth.

The range tones are phase modulated on the up-link carrier. Phase modulation is used here to a) avoid having to "pull" a very stable crystal oscillator to a large frequency deviation, and b) to obtain better transmitter power efficiency with constant peak power signal. The up-link frequency is nominal 2270 mc/s. Low indices of modulation are used to restrict the signal to approximately a one megacycle bandwidth. The system spectra are shown in Figure 8.

The S-Band ground antennas are hydraulically driven on XY pedestals. The XY pedestals are used to obtain accurate angular data around the zenith without the undue complexity associated with other mounts. Apparently NASA believes that only XY mounts will be used in the future. The system consists of two 14-foot Cassegrainian parabolic reflectors mounted side by side on the same pedestal. One is used for transmitting and one for receiving with gains of 35 and 33 db respectively (Motorola maintains the transmitting antenna attains a 39 db gain). The separate antennas allow a lower receiver noise temperature due to elimination of transmitter interference and lossy elements.

TABLE 2

TRANSPONDER TRACKING MODES

No. of Channels	No. of Range Tones	Highest Tone Freq.	Maximum Range (KM)	Transponder Tx Power	Transponder Ant. Gain
2	7	100 kc/s	32,000	1 watt	0 db
1	7	100 kc/s	100,000	1 watt	0 db
1	8	500 kc/s	5,000	1 watt	0 db
1	500 kc/s PR code	Broad	Lunar	1 watt	10 db

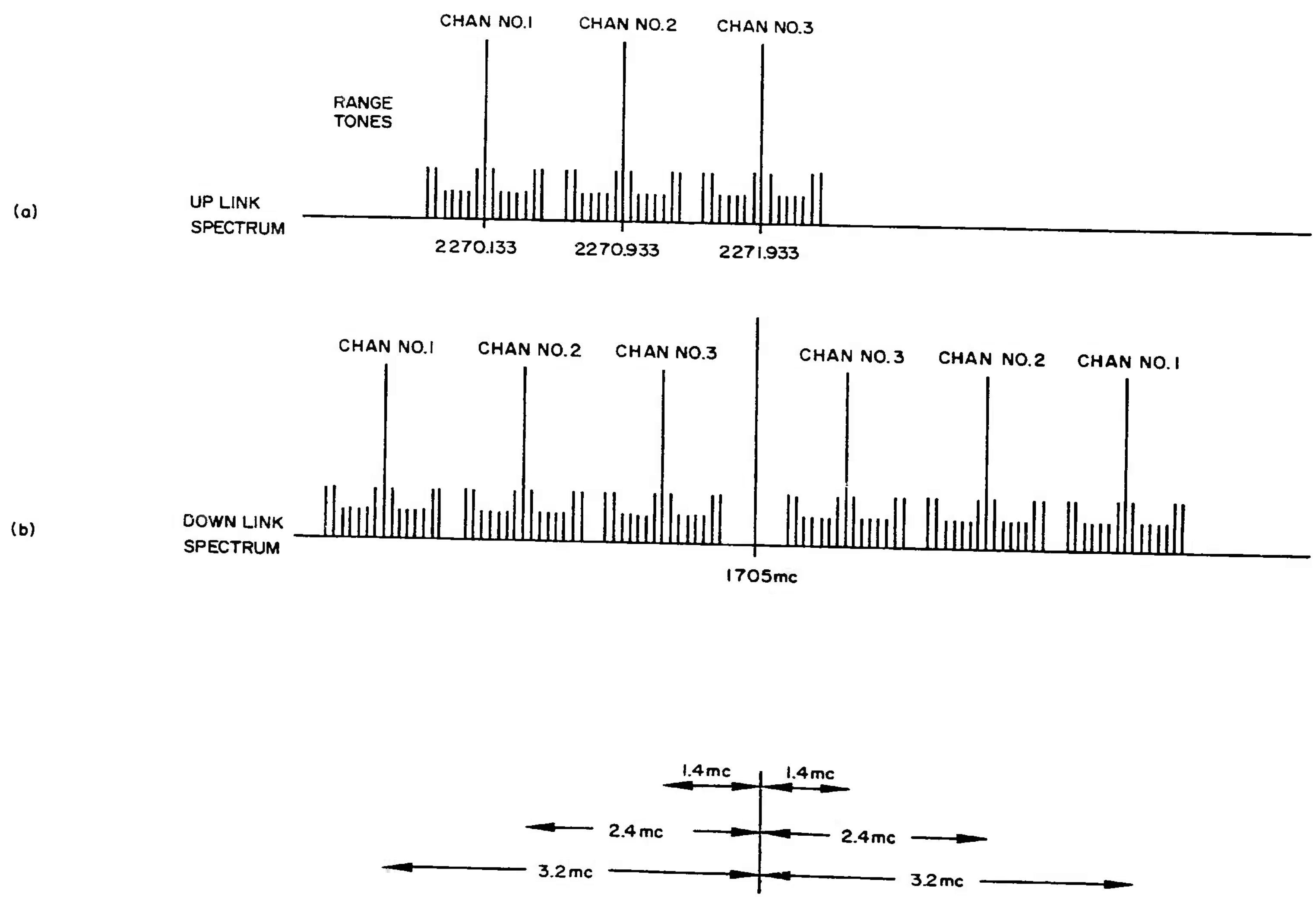


Figure 8. Spectrum of S-Band Transponder

The ground transmitter power is 10 Kw and can be switched to 1 Kw service for short range passes.

The transponder receiver antenna is usually 0 db, transmit and receive. As noted in Table 2, the transponder will radiate 1 watt on the down link. Increase in range over 100,000 kilometers is attained by increased antenna gain and PR coding.

The range tones received in the transponder are heterodyned, with a non-coherent local oscillator, down to a low if for linear addition to the other two channel signals (see Figure 8a). The composite 3 channel signal is phase modulated on the down link carrier (1705 mc/s) with the resulting spectrum as shown in Figure 8b. The modulation system then is PM/PM. The 1705 mc/s down frequency was chosen as the down link because this is a designated NASA down link frequency. No reason for the 2270 choice is apparent, except allocation. Apparently it is easier to buy microwave components at these rather than other frequencies; for instance, rather than 1400 mc/s.

On the up link the three rf carriers are spaced by .8 mc/s between the lowest and center, and 1.0 mc/s between the highest and center. Since each up carrier becomes a subcarrier phase modulating the down link, the sum and difference frequencies of these down link subcarriers and their associated range tones can cause coherent interference. The above up link carrier spacing was chosen as a result of a computer study which incorporated the following restrictions; a) the sums and differences of the subcarrier frequencies do not fall within a ±200 kc/s band about a desired frequency; b) the highest subcarrier frequency is limited to 3.5 mc/s because of transponder bandwidth limitations; c) the subcarrier frequencies must be sufficiently separated to permit filter separation in the transponder in the presence of sidetone modulation and doppler. Apparently there would be no problem if discriminator type demodulators were used on the ground; however, in the product type detector these sums and differences of all the harmonics can act as interference.

The modulation indices on the up and down links are respectively .7 and .2 for major and minor side tones and .5 for each sub-carrier. Motorola analysis indicates that this choice of indices produces the best compromise between signal to noise and signal to interference.

2.2.1 Ranging

The eight ranging side tones are 500 kc/s, 100 kc/s, 20 kc/s, 4800 cycles, 4160 cycles, 4032 cycles, 4008 cycles, and 4 kc/s. All of the tones may be used singly or in combination. Used singly, the tones provide greater precision due to the increase in transmitted power (both ground and space) and the resulting increase in S/N, which yields smaller phase errors.

When using more than one transponder channel at one time, the highest permissible side tone is 100 kc/s. Use of a higher range tone frequency for multiple channel operation would require widening the transponder channel filters. If the over-all transponder bandwidth was not to increase then the interchannel crosstalk would add intolerable errors as would the added phase/frequency nonlinearity. For these reasons, the 500 kc/s range tone is provided a separate broadband (approximately 3.3 mc/s) channel which overlaps all three other channels. The higher frequency tones determine the system precision and the low frequencies set the maximum unambiguous range. The 5 to 1 frequency ratio is a conservative value which allows higher confidence of ambiguity resolution and is determined by the actual accuracy of the system phase detectors.

Though the lowest frequency noted above is 4 kc/s (which corresponds to about 37.5 km resolution), the information as to the maximum unambiguous range lies in the difference between 4 kc/s and 4.008 kc/s. This "complementing" of the low frequency range tones is used to remove energy from within the passband of the ground carrier tracking phase lock loop and allows more accurate and reliable loop operation.

It is a requirement then, upon all frequency selections, that under various conditions of doppler shift they do not fall within the loop noise bandwidths of other frequencies.

2.2.2 Range Rate

Range rate information may be obtained from either the range tones or the carrier doppler. Normally, doppler tracking will be utilized for this purpose. The transponder incorporates an incoherent local oscillator whose drift must be cancelled out on the ground if accurate two way doppler is to be obtained. By a fairly complicated process* of manipulating the received and reference frequencies the two way component plus bias is separated from the other frequencies.

Since the doppler processing is a short time count it is important that the reference oscillator have good short term stability. Apparently the long term stability of a crystal can be sacrificed to obtain better short term stability by the particular cut made upon the crystal used in the reference oscillators.

2.2.3 Command

Motorola maintains that a digital command system such as that used in the Gemini Program, can be incorporated into the system simultaneously with ranging. A subcarrier would be chosen between 5 kc/s and 20 kc/s or 20 kc/s and 100 kc/s such that the command spectrum

*See Section 4.8.3, Reference 1.

did not overlap the range tones. This will require addition of a discriminator, filter and decoder in the vehicle. The addition of energy within the ranging bandwidth will degrade the ranging performance but not significantly, so maintains Motorola.

2.2.4 Telemetry

The spacecraft data to be transmitted to the ground would be modulated on a subcarrier of frequency equal to that which is not being transmitted from the ground. Then, as any other received subcarrier, it is modulated upon the down carrier. 2400 bit per second telemetry service is available to a 10,000 mile range when ranging is not being used.

2.3 S-Band Ranging Transponder

The over-all block diagram of the S-Band transponder is shown in Figure 9. The diagram shows the 4-channel system as planned. Table 3 gives pertinent parameters of the unit.

The major characteristic of the transponder is the absence of any signal detection circuitry. The received rf carriers are mixed down to subcarrier frequencies and then remodulated by a carrier for transmission to the ground station. This process seems reasonable since a) a range signal is never taken to baseband in order to preserve the doppler polarity information and b) a very simple transponder equipment results which minimizes vehicle weight, power and complexity; confining most of the equipment to the ground. The primary gain is the minimization of ranging and doppler errors which might be added for other more complex schemes. There is also one less frequency acquisition which is a minor consideration. The major problem associated with the transponder is minimizing the phase/frequency nonlinearity which contributes directly to the range

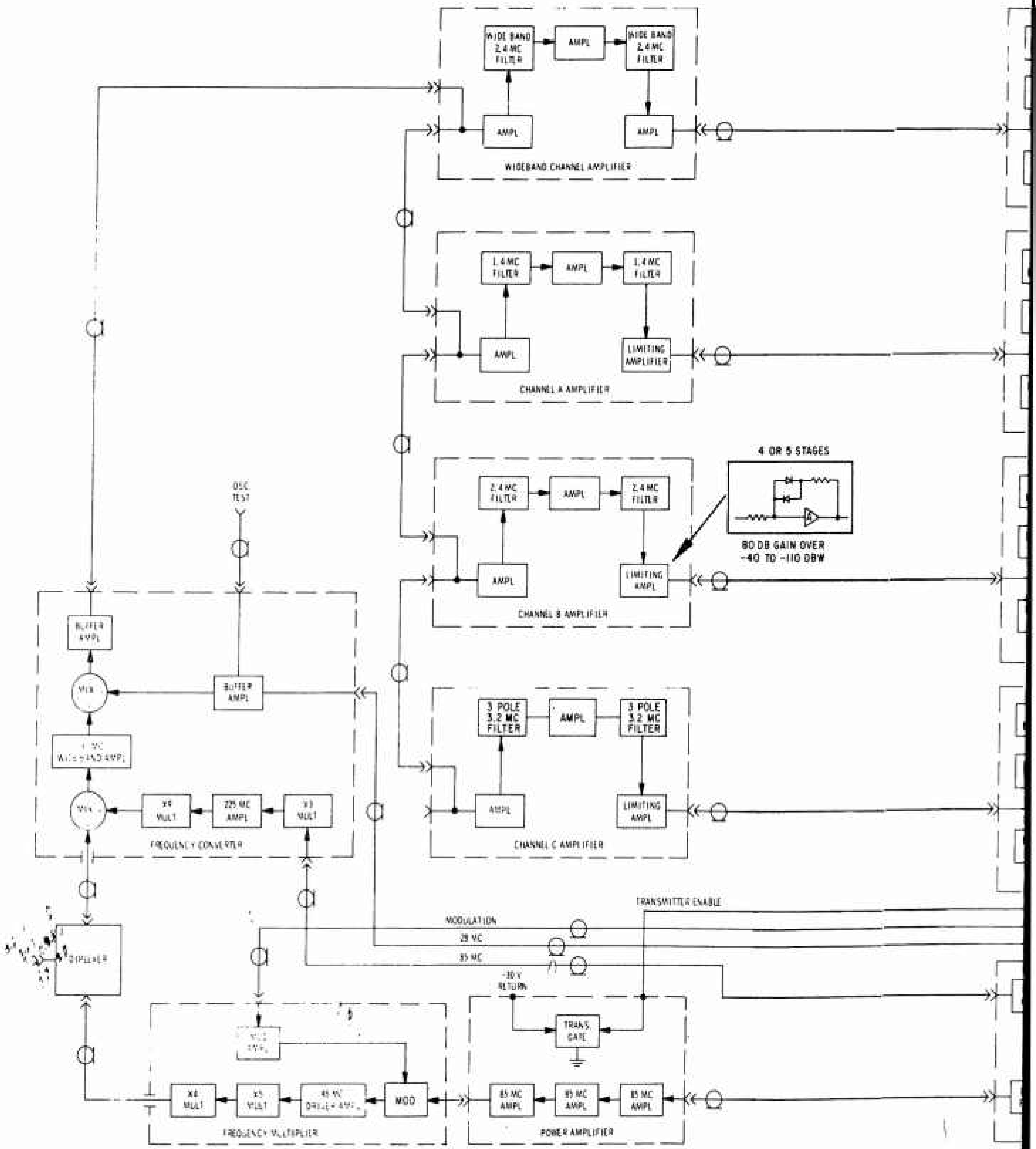


Figure 9. Range and Range-Rate Trans

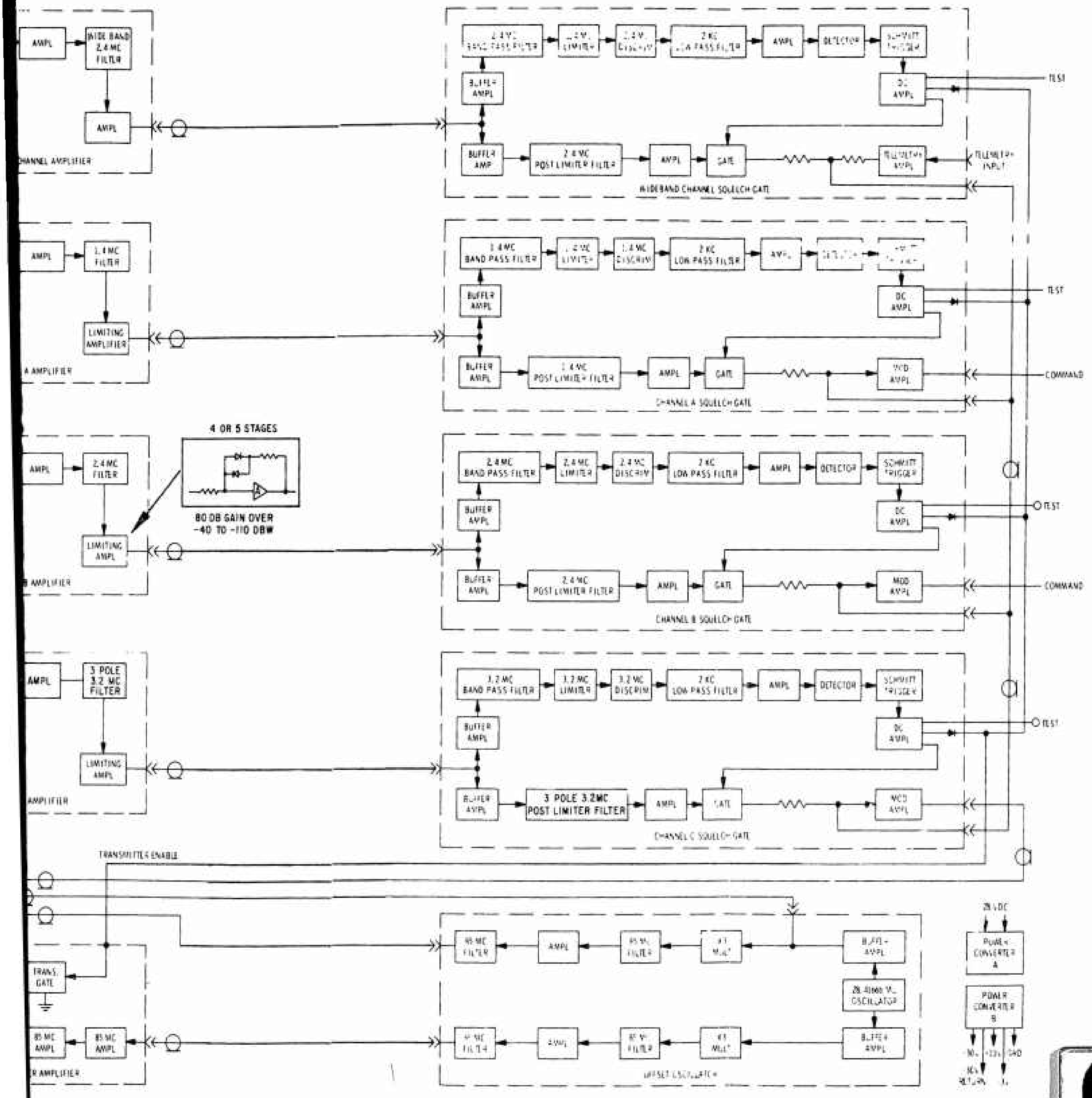


Figure 9. Range and Range-Rate Transponder

TABLE 3

TRANSPONDER PARAMETERS

Input Signal Level	-40 dbm to -110 dbm (each channel)
Received Frequencies	2270.1328 mc/s
	2270.9328 mc/s
	2271.9328 mc/s
Subcarrier Channel Frequencies	3.2, 2.4, 1.4 mc/s
Channel Power Difference	0 to 32 db (any channel high)
Multiplication Ratio	60 (transmit)/80 (receive)
Channel Range Tone Bandwidth	200 kc/s
Wideband Channel Bandwidth	3.3 mc/s
One Way Doppler Shift	±85 kc/s maximum
Transmitted Carrier Frequency	1705 mc/s
Transmitter Power	1 watt
Transponder Uncalibrated Group Delay	±50 nanoseconds maximum

uncertainty. It turns out that this error is about equal to the errors contributed by noise and oscillator instability. These three factors account for about 80% of the total range error.

2.3.1 Frequency Conversion

The receiver frequency converter uses a double frequency conversion. The incoming frequencies around 2271 mc/s are mixed with a high reference at approximately 2302 mc/s, resulting in a first if of approximately 31 mc/s. The second mixer reference is 28.4 mc/s and results in the channel center frequencies of 1.4, 2.4, and 3.2 mc/s. The reasons for double conversion are: a) the desire to attain maximum amplification, and for stability reasons this is best done at different frequencies; and b) to allow elimination of images. Apparently the results of the design were such that, according to Motorola, single conversion would have been satisfactory, since the required if gain is not too excessive and the undesired images were not effectively eliminated by the double conversion.

The adjustment of power ratios of the various signals used in the mixer is apparently a problem. The design of the first mixer and if amplifier are determined by the requirement for as low a noise figure as possible, which in this case turns out to be 10 db. The second mixer is designed to reduce the higher order mixer products to 45 db below the first order products.

2.3.2 Non-Coherent Local Oscillator and Frequency Multiplexers

The spacecraft reference frequencies are supplied from an oscillator using a third overtone crystal operating at 28.4166 mc/s, which is expected to yield acceptable long term and short term stability for the uncontrolled temperature environment. The expected temperature stability is ± 5 ppm from 0° to +50°C and the expected noise contribution, due to short term effects, is less than 2° rms phase error in a 50 cps noise bandwidth at 2271 mc/s.

The oscillator output is multiplied by 81 to provide the reference for the first mixer. The multipliers are a combination transistor and varactor. The last multiplier provides 2 mw to the first mixer.

The frequency multiplier chain in the transponder transmitter consists of bandpass type varactor multipliers. Maintaining frequency linearity through these circuits is a problem and as a result the selection of the point at which to apply modulation has been open to experiment.

2.3.3 Channel Amplifiers and Limiters

Essentially identical amplifiers are used in each of the three subcarrier channels at 1.4, 2.3 and 3.2 mc/s. These channel amplifiers perform the primary function of equalizing the output level of each channel, independently of the input signal levels, in order to maintain an equal transmitter output power allocation to each channel. It is apparent that these channel filters must be as narrow as possible to attain maximum attenuation of noise before the limiters and yet they must be wide enough to attain a very linear phase/frequency characteristic.

Since each channel frequency can have one way doppler shift, the bandwidths must be greater than 400 kc/s ($\pm$ 100 kc/s range tones plus approximately $\pm$ 100 kc/s one way doppler). Additionally, the linearity of the phase slope must be maintained over this bandwidth so that the uncalibrated group delay is less than 50 nanoseconds while maintaining 60 db rejection to adjacent channel signals (60 db at about 650 kc/s for a three section filter).

In the over-all transponder design, the most difficult problem was the synthesis of these relatively narrow filters. Imperfections in the phase/frequency characteristic add directly to the ranging errors making up the allotted 50 nanoseconds. The resulting filters take the form of three isolated three pole filters which were computer synthesized to minimize the group delay variation. Figure 10 shows the amplitude characteristic. It is not unusual that the calculated values agree with the measured values when precisely synthesizing a linear phase/frequency characteristic.

A measured plot of the group delay through the filter system shows the characteristic of Figure 11. The approximate curves show a skewness which is probably the result of misalignment of center frequencies and causes a delay variation of about ±70 nanoseconds. The correct characteristic is shown by the dotted line and approximates the required ± 50 nanoseconds. The group delay errors cannot be calibrated because the doppler shift is not known. To apply a frequency dependent correction would require a complicated iteration process to determine the final doppler or range figure and this iteration itself may add inseparable errors.

A limiting amplifier is placed between the second and third sections of the bandpass filters. This amplifier consists of four or five stages of nonlinear feedback amplifiers. These amplifiers supply 80 db gain over an input dynamic range of 70 db (over -40 to -110 dbm for each channel) with a time delay of less than ± 8 nanoseconds. Careful attention in the design should be given to minimizing the A.M. to P.M. conversion effect and also to minimizing the change in time delay through the limiters as a function of signal level.

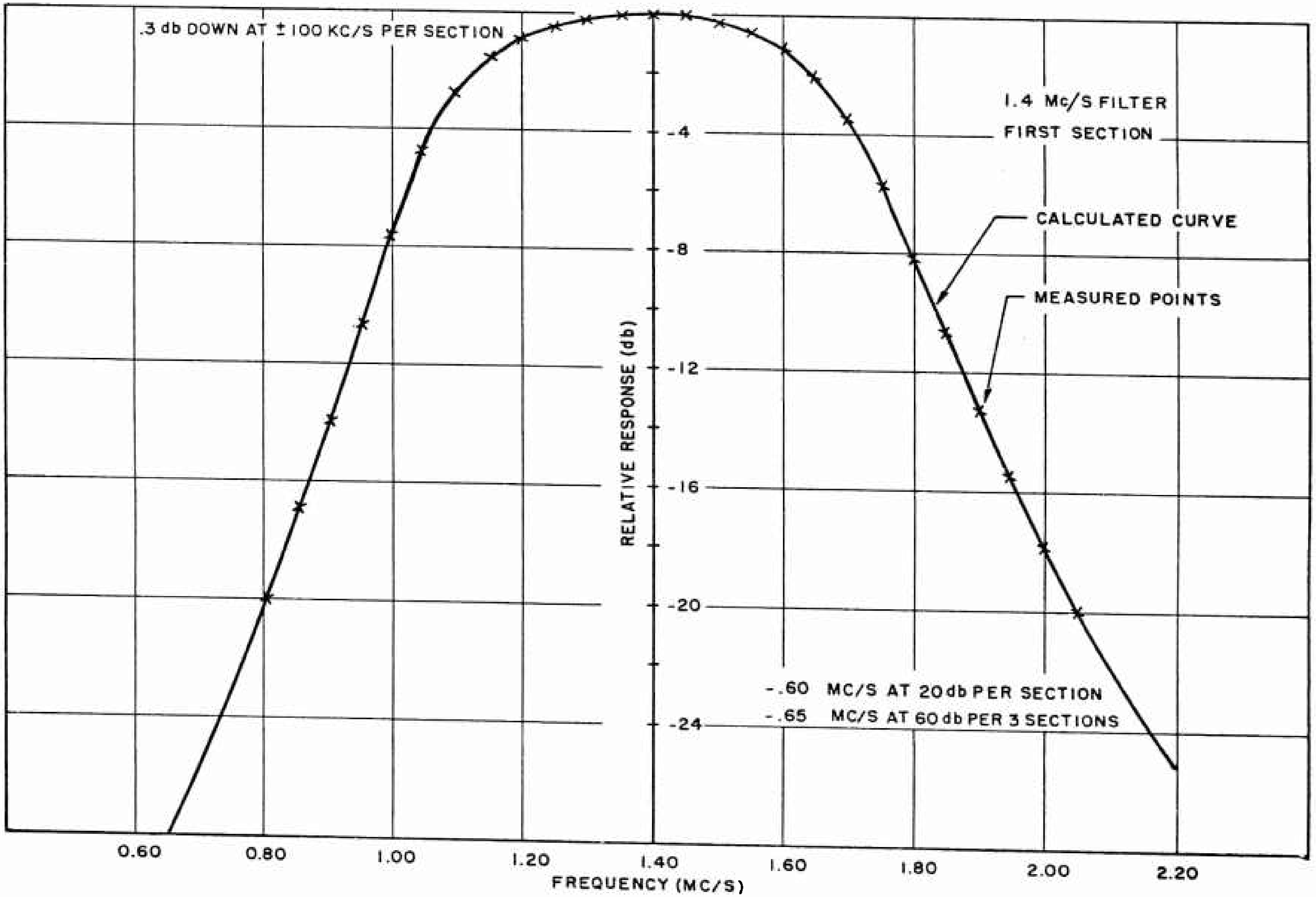


Figure 10. Transponder Channel Filter

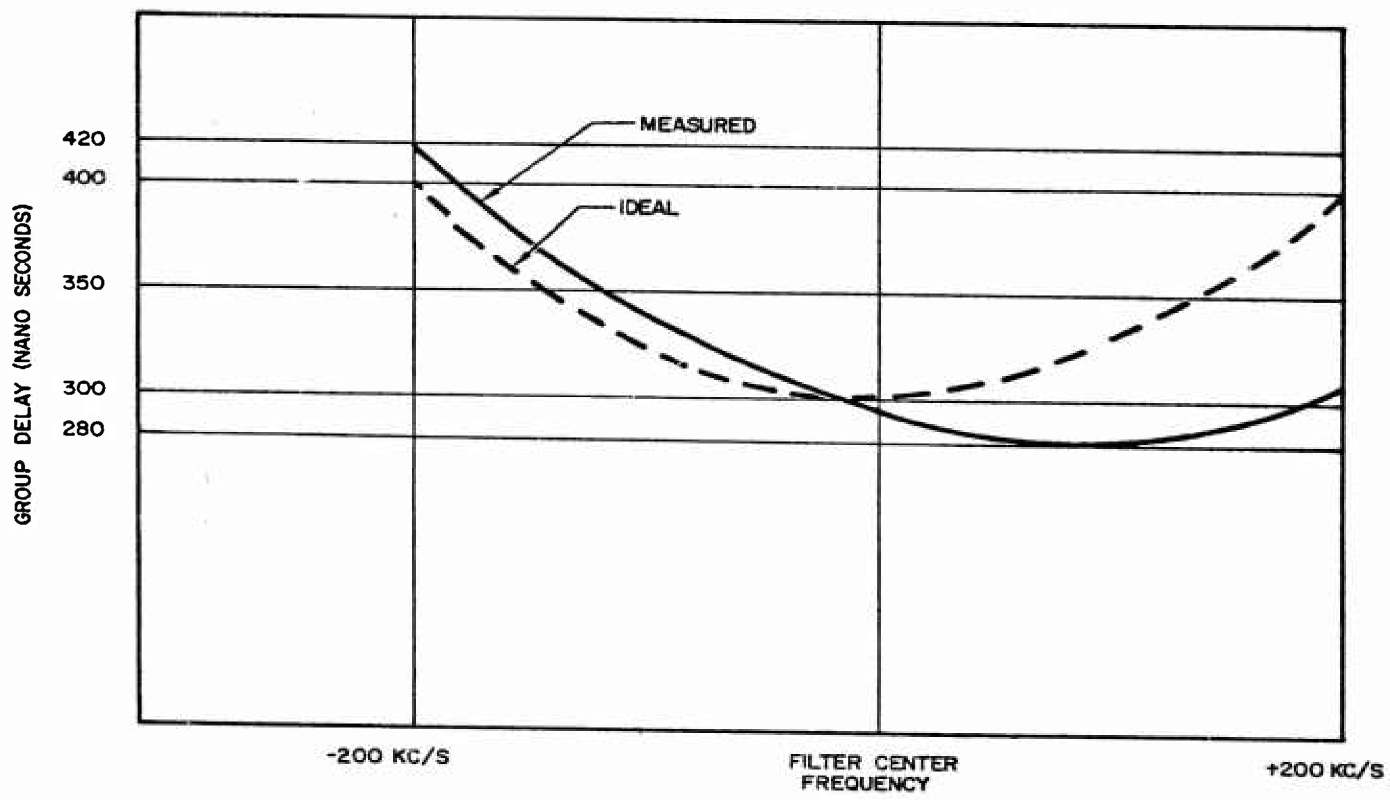


Figure 11. Channel Filter Group Delay Characteristic for Three Sections

NASA indicates that the phase linearity and isolation of the channel filters did not come up to the original requirements; however, the result was considered satisfactory for the application.

Aside from these channel amplifiers, all other amplifiers in the transponder are designed to be as wide band as possible to minimize the differential phase delay.

2.3.4 Squelch Circuits

The squelch circuitry functions to prevent transmission of noise when a channel signal is not present. If no signal is present in any channel, the transmitter is disabled to reduce non operating power consumption.

The squelch circuit utilizes the noise quieting effects in a limiter discriminator circuit. The discriminator is followed by a 2 kc/s low pass filter to reject the 4 kc/s and higher tones. Increasing signal power will decrease the noise power in this bandwidth, thus permitting the recognition of the presence of a signal. This circuit actuates with a (S/N) ratio in the channel bandwidths of approximately -6 db or an input signal level of -110 dbm. This corresponds to a range at which the S/N in the ground receiver carrier loop is approximately zero db when using a one watt transponder power output on single channel operation.

The output of the squelch amplifier drives a schmidt trigger which drives the channel signal gate and enables the transmitter circuitry.

Referring to Figure 9, a question arises when considering the operation of the transponder in the wideband mode. During this mode of operation it would appear that both the wideband and channel B squelch gates would be activated (due to identical center frequencies) thereby adding the output from channel B to the output of the wideband channel. Distortion or interference would result due to feeding the same signal through filters with different phase characteristics and then summing the two outputs.

2.3.5 Phase Modulator and Transmitter

Approximately 8 watts of power at 85.25 mc/s are developed in parallel transistor amplifiers which receive their signal from the first tripler. The output of this power amplifier is then phase modulated by the summed subcarriers. A x5 multiplier and a x4 multiplier generate the modulated rf signal at 1705 mc/s and the output power level, at the diplexer, of 1 watt. As noted above, the point at which to apply the phase modulation has been open to experiment due to difficulty in controlling frequency non-linearities of the multipliers which, of course, in a ranging system is disastrous. There also exists the problem of varactor parameter variations with temperature.

2.4 References for Goddard R and RR System

1. Goddard Range and Range Rate System Design Evaluation Report, No. W2719-2-1, Revision 1, 23 November 1962, Contract Number NAS 5-1926, prepared for NASA by Motorola, Inc. Military Electronics Division.
2. AD/COM High-Accuracy Satellite Tracking Systems, First Quarterly Report 30 June 1961, Including Task 1 Report.
3. AD/COM High-Accuracy Satellite Tracking Systems, Second Quarterly Report 30 September 1961.
4. AD/COM High-Accuracy Satellite Tracking Systems, Third Quarterly Report 31 December 1961.
5. AD/COM High-Accuracy Satellite Tracking Systems, Fourth Quarterly Report 31 March 1962.
6. AD/COM High-Accuracy Satellite Tracking Systems, Progress Report, 30 June 1962
7. AD/COM High-Accuracy Satellite Tracking Systems, Sixth Quarterly Report 30 September 1962.
8. AD/COM High-Accuracy Satellite Tracking Systems, Seventh Quarterly Report 31 December 1962.
9. Command, Telemetry and Communication Capabilities of the Goddard Range and Range Rate Transponder, Motorola T.M. No. 2719-6-5, 22 August 1962.
10. The System Generated Intermodulation of the Goddard Range and Range Rate System, Motorola T.M. No. 2719-2-4.
11. Summary of RARR Specifications, Error Sources and Effects, Motorola TM.
12. Up-Link Spectrum and Modulation Efficiency for the Goddard RARR System, Motorola T.M.; Ken Wood.
13. Error Analysis of S-Band System in Goddard Range and Range Rate System, Motorola T.M. 2719-7-8, 23 August 1962.
14. Error Analysis of VHF System in Goddard Range and Range Rate System, Motorola T.M. 2719-7-7, 15 August 1962.
15. S-Band and VHF Receiver Bandwidth Selection and Acquisition Time for Goddard RARR System, Motorola T.M. 2719-7-11, 14 December 1962.

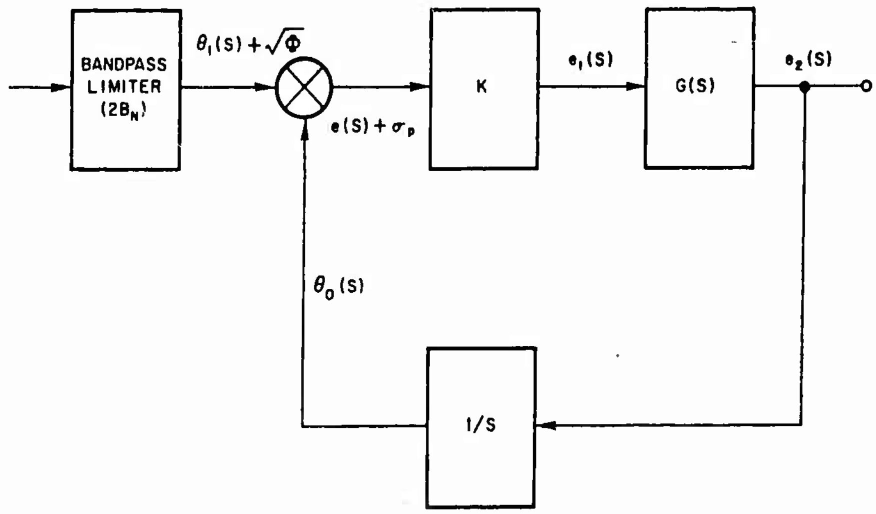
APPENDIX A

THE LINEARIZED SECOND ORDER PHASE LOCK LOOP INCLUDING THE BANDPASS LIMITER SUPPRESSION FACTOR

In the design of a communications receiver system, there exists a condition of signal power in relation to noise (or interference) power at which, regardless of the available system gain, the signal is no longer detectable. This is known as the receiver threshold signal to noise ratio. This threshold is usually due to the receiver's inability to efficiently process lower and lower (or higher and higher) signal power and exists because of some nonlinearities in the system. In other terms, if the system were linear for all input values it would only require additional system gain to satisfactorily process weaker signals. In the present case we are interested in examining the threshold behavior of a phase lock loop used to track only an rf carrier and translate incoming modulation. The receiver is the JPL Mariner S-Band transponder. Specifically, we would like to know the minimum received signal to noise power ratio to satisfy some threshold criteria within the loop. In this type of receiver the threshold effect arises due to the nonlinearity of the input multiplier or phase detector in which the output is proportional to cosine θ, where θ is the phase difference between the input and a reference signal.

The linear second order servo loop is shown in Figure A-1. The input to the bandpass limiter is a sinusoidal carrier with doppler rate superimposed along with a phase noise spectral density, Φ_N. The bandpass limiter characteristic of interest is its signal suppression factor, α (see Equation A1-3). This factor is derived from the limiter properties a) of Equation A1-1 which indicates that for the signal to noise power ratios existing in the last if bandwidth, $2B_N$,[1] the S/N through the limiter is only modified by a factor of $\frac{\pi}{4}$ and b) that the limiter (total S+N)

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



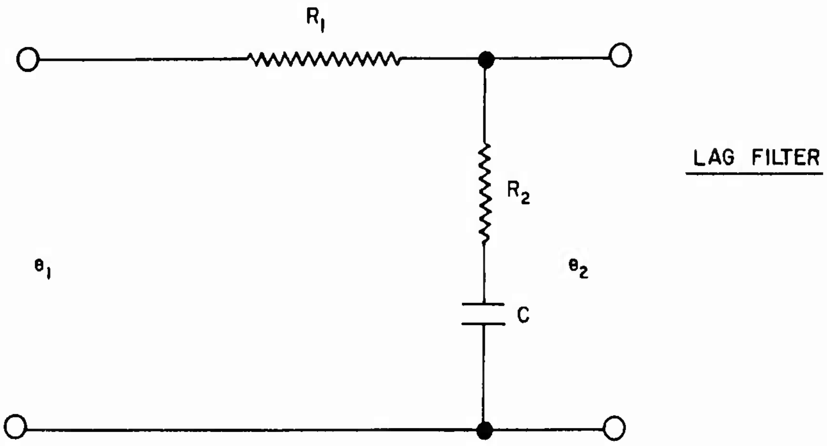
$$\left(\frac{S}{N}\right)_0 = \frac{\pi}{4}\left(\frac{S}{N}\right)_i, \text{ FOR } \left(\frac{S}{N}\right)_i \rightarrow 0 \qquad \text{(AI-1)}$$

$$K = a K_m K_{vco} K_A \qquad \text{(AI-2)}$$

$$a = \frac{1}{\sqrt{1 + \frac{4}{\pi}\left(\frac{N}{S}\right)_{2B_N}}} \qquad \text{(AI-3)}$$

$$G(S) = \frac{1 + T_2 S}{T_1 S} \qquad \text{(AI-4)}$$

Figure A-1. Linear Second Order Servo with a Bandpass Limiter Input



$$\frac{e_2}{e_1} = \frac{1+T_2S}{1+(T_1+T_2)S} \qquad \text{(A2-1)}$$

$$G(S) = \frac{1+T_2S}{T_1S} \text{ FOR } T_1 \gg T_2 \text{ AND } T_1 \gg 1 \qquad \text{(A2-2)}$$

Figure A-2. Practical Loop Filter Which Can Be Optimized by Choice of Parameters output power is essentially constant regardless of total input power. The limiter suppression factor, α, is important because it acts as a variable signal gain factor within the loop and consequently as the last if $(S/N)_{2B_N}$ varies, the loop noise bandwidth, $2B_L$, the loop signal to noise power ratio, $(S/N)_{2B_L}$ and the loop phase variance, σ_p^2, all vary with resultant change in the tracking performance. Due to α being included as a variable loop parameter the closed loop transfer function can only be optimized at one value of $(S/N)_{2B_N}$. Usually this transfer function is optimized at some defined threshold value and is therefore non-optimum at other values of $(S/N)_{2B_N}$.

The servo loop filter of the form of Equation A2-2 has been shown, by Jaffe and Rechtin[2] to be optimum in the sense that it minimizes VCO phase-noise jitter, σ_p, for specific values of transient error to a specified input signal (Weiner technique). In other terms, if we choose the closed loop damping ratio, ζ, to be $1/\sqrt{2}$ then the transient phase error, E_T, to a step frequency input has been minimized and then also the phase jitter, σ_p.

Additionally, the choice of filter time constants should be such as to satisfy $T_1 >> T_2$ and $T_1 >> 1$ which poses no design problem.

Returning to our original goal, we wish simply to determine the relation between the signal to noise ratios in the loop noise bandwidth $(S/N)_{2B_L}$, and the receiver input, $(S/N)_{2B_N}$. Knowing this relation and choosing a threshold criteria we can then determine the minimum required receiver input signal power. The open loop transfer function, given G(s) of A1-4 is,

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2. 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", PGIT, March 1955, pp 66-76.

$$H_o(s) = \frac{KG(s)}{s} = \frac{K(1 + T_2 s)}{T_1 s^2} \qquad (A3)$$

The closed loop transfer function is,

$$\frac{\theta_o(s)}{\theta_i(s)} = H(s) = \frac{H_o(s)}{1 + H_o(s)} = \frac{1 + T_2 s}{1 + T_2 s + \frac{T_1}{K} s^2} \qquad (A4)$$

comparing the R.H.S. of (A4) with the characteristic equation of the second order loop we have:

$$\frac{1 + T_2 s}{1 + T_2 s + \frac{T_1}{K} s^2} = \frac{F(s)}{1 + \frac{2\zeta}{\omega_n} s + \frac{s^2}{\omega_n^2}} \quad \text{and by A1-2}$$

$$\omega_n = \sqrt{\frac{K}{T_1}} = \sqrt{\frac{\alpha K_1}{T_1}} \qquad (A5)$$

$$\zeta = \frac{T_2}{2}\sqrt{\frac{K}{T_1}} = \frac{T_2}{2}\sqrt{\frac{\alpha K_1}{T_1}} \qquad (A6)$$

Examining (A6), we note that T_1, T_2 and K_1 are constants. If α varies then it is impossible for ζ to equal $1/\sqrt{2}$ except at one value of α which is usually chosen as the value at threshold, α_o. Under this choice then the loop is only optimum at the defined threshold.

In the loop linearization, J. and R.[3] show that the phase noise spectral density, $\Phi_N(s)$ is, for flat amplitide spectral distribution,

$$\frac{\Phi_N(s)}{p} = \frac{\Phi_N(o)}{p} = \frac{1}{2B_N} (N/S)_{2B_N} \tag{A7}$$

and, if we take the same liberties as Martin[4], the normalized amplitude noise spectral density, Φ_A, is then,

$$\Phi_N(o) = \Phi_A = \frac{1}{2B_N} (N/S)_{2B_N} = \frac{\Phi_{KT} + \Phi_{NF}}{p} \tag{A8}$$

where Φ_{KT} and Φ_{NF} are respectively the system noise spectral densities due to received thermal noise at room temperature, KT, and the receiver noise figure (which is about 11 db maximum for this receiver).

The loop phase variance, σ_p^2, and the loop bandwidth are given by the contour integrals,

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3. op. cit.

4. Martin, B. D., "A Coherent Minimum Power Lunar Probe Telemetry", JPL-EP No. 610, 12 May 1959.

$$\sigma_p^2 = \frac{1}{2\pi j}\int_{-j\infty}^{+j\infty} \left| H(s) \right|^2 \Phi_N(o)\ ds \tag{A9a}$$

$$2B_L = \frac{1}{2\pi j}\int_{-j\infty}^{+j\infty} \left| H(s) \right|^2\ ds \tag{A9b}$$

Usually σ_p^2 is taken to equal the loop signal to noise ratio, $(S/N)_{2B_L}$. Using H(s) as in (A4) and since $\Phi_N = \Phi_A$ is constant with s, the variable of integration, the loop phase variance is[5],

$$\sigma_p^2 = (N/S)_{2B_L} = \frac{\Phi_A}{2\pi j}\int_{-j\infty}^{+j\infty} \left| \frac{1 + T_2 s}{1 + T_2 s + \frac{T_1}{\alpha K_1} s^2} \right|^2 ds$$

$$= \left[\frac{\alpha K_1 T_2}{2T_1} + \frac{1}{2T_2} \right] \Phi_A \tag{A10}$$

---

5. The contour integral is tabulated in "Theory of Servomechanisms" Rad. Lab Series, No. 25, pp 369-370

At this point we wish to fix our filter gain and time constants. We will agree to let the damping ratio, ζ, equal the optimum value, $1/\sqrt{2}$, at the loop threshold and let the loop threshold natural frequency, ω_n, equal B_o. These substitutions give, from (A5) and (A6),

$$T_2 = \frac{\sqrt{2}}{B_o} \quad , \quad \frac{K_1}{T_1} = \frac{B_o^{\,2}}{\alpha_o} \tag{A11}$$

and from A1-3, the limiter threshold suppression factor is,

$$\alpha_o = \frac{1}{\sqrt{1 + \frac{4}{\pi}\,(N/S)_{2B_{No}}}} \tag{A12}$$

where $(N/S)_{2B_{No}}$ is the last if signal to noise power ratio at the loop threshold. Equation (A10) is now,

$$\sigma_p^{\,2} = (N/S)_{2B_L} = \frac{B_o\,\Phi_A}{\sqrt{2}}\left[\frac{\alpha}{\alpha_o} + \frac{1}{2}\right] = 2B_L\,\Phi_A \tag{A13}$$

The R.H.S. of (A13) comes from (A9b). To evaluate α_o, let $\alpha = \alpha_o$ in (A13), then the loop bandwidth at threshold is,

$$\frac{\sigma_{po}^2}{\Phi_{Ao}} = 2B_{Lo} = 1.06\ B_o \approx B_o \tag{A14}$$

which checks with the currently accepted value of threshold loop bandwidth, $2B_{Lo}$. If we use (A8) in (A14) we obtain the if threshold signal to noise ratio as a function of the system bandwidths and the chosen threshold loop noise value, σ_{po}^2,

$$(S/N)_{2B_{No}} = \left(\frac{2B_{Lo}}{2B_N}\right) \frac{1}{\sigma_{po}^2} \tag{A15}$$

Now, from (A12)

$$\alpha_o = \frac{1}{\sqrt{1 + \frac{4\sigma_{po}^2}{\pi}\left(\frac{2B_N}{2B_{Lo}}\right)}} \tag{A16}$$

or

$$\frac{\alpha}{\alpha_o} = \sqrt{\frac{1 + \frac{4\sigma_{po}^2}{\pi}\left(\frac{2B_N}{2B_{Lo}}\right)}{1 + \frac{4}{\pi}\ (N/S)_{2B_N}}} \tag{A17}$$

using (A8) again in (A13), and using the approximations in (A14), we get,

$$\sigma_p^{\,2} = (N/S)_{2B_L} = \frac{1}{\sqrt{2}}\left(\frac{2B_{Lo}}{2B_N}\right)(N/S)_{2B_N}\left\{\left[\frac{1+\frac{4\sigma_{po}^{\,2}}{\pi}\left(\frac{2B_N}{2B_{Lo}}\right)}{1+\frac{4}{\pi}(N/S)_{2B_N}}\right]^{\frac{1}{2}}+\frac{1}{2}\right\} \qquad (A18)$$

Equation (A18) gives the loop signal to noise ratio as a function of the last if signal to noise ratio. Various criteria can be used to choose the parameters $2B_N$, $2B_{Lo}$ and $\sigma_{po}^{\,2}$.

The last if bandwidth, $2B_N$, is determined by the information to be translated by the loop under consideration. The threshold loop noise bandwidth might be chosen to accommodate the maximum system doppler rate, since $2B_{Lo}$ is the narrowest loop bandwidth during coherent operation. In the case of the Mariner S-Band transponder $2B_{Lo} = 20$ cps and the last if bandwidth, $2B_N$ is 4500 cps.

Often the current literature has taken the loop threshold to be that condition where the rms loop phase error, σ_p, is equal to $\pi/2$. $\pi/2$ is chosen since at this value of phase error the multiplier gain, $K_{m(actual)}$, and consequently the loop gain, K, is zero and the received signal and loop reference coherence is definitely lost,

$$K = \alpha\, K_{m(actual)}\, K_{vco} K_A = \alpha K_m^{\,'} \cos\theta\, K_{vco} K_A = \alpha K_m^{\,'} \cos\sigma_p\, K_{vco} K_A \qquad (A19)$$

This equation is true for no input modulation or Doppler rate on the tracked carrier. Under this assumption the loop should track down to a threshold signal to noise ratio, $(S/N)_{2B_{Lo}}$, of

$$\sigma_p^{\ 2} \ = \ (\frac{\pi}{2})^2 = 2.46 \text{ or } (S/N)_{2B_{Lo}} \ = \ -3.9 \text{ db} \qquad \text{(A20)}$$

σ_p equaling $^{\pi}/2$ means that 32% of the time the instantaneous phase error will be in excess of $^{\pi}/2$ which may be interpreted as the amount of time that no phase coherence exists. A loop operating at this point would appear to be below any practical threshold and this is borne out in practice. Experiment has shown that a reliable and repeatable threshold is about 0 db $(S/N)_{2B_{Lo}}$ which corresponds to a loop phase error of about 1 radian rms. This is somewhat more palatable since one might expect a true coherence threshold somewhere less than $\sigma_p = {}^{\pi}/2$ (loop gain, K = 0) but more than $\sigma_p = 0$ $(S/N \rightarrow \infty)$. The obvious trouble with these analyses is that the linearized system can give no direct information on the actual system nonlinearities and the resulting actual thresholds.

Using 1 radian as the "one sigma" phase noise at threshold and the other Mariner parameters stated above, from (A15) we get,

$$(S/N)_{2B_{No}} \ = \ \frac{20_{cps}}{4500_{cps}} \ \frac{1}{1 \text{ rad}^2} = \ -23.5 \text{ db} \qquad \text{(A21)}$$

Using (A8) and changing to db, the receiver signal threshold is,

$$p_o \ = \ 2B_N + \Phi_{KT} \ + \ \Phi_{NF} \ + \ (S/N)_{2B_{No}}$$

$$= \ + 36.5 \text{ db} \ -174 \text{ dbm} + 11 \text{ db} \ -23.5 \text{ db} \ = \ -150 \text{ dbm} \qquad \text{(A22)}$$

Referring to (A18), we note that as the if input S/N increases, the noise power approaches zero in the limit. It is further interesting to note that for high if input S/N, the loop bandwidth B_L approaches an asymptotic limit. From (A13) and (A18)

$$2B_L = \frac{2B_{Lo}}{\sqrt{2}} \left[\frac{16.9}{\sqrt{1 + \frac{4}{\pi} (N/S)_{2B_N}}} + \frac{1}{2} \right] \tag{A23}$$

$$\lim_{\frac{N}{S} \to 0} 2B_L = 24.7\, B_{Lo} \tag{A24}$$

The loop bandwidth, $2B_L$, for high S/N will then have a better capability to handle the higher doppler rates at closer ranges. For the S-Band Mariner transponder this limit is about 495 cycles per second for a 20 cps threshold value.

NOMENCLATURE FOR APPENDIX A

B_o	rad/sec	loop natural frequency
E_T	rad	transient phase error
f_o	cycles/sec	loop information bandwidth
$G(s)$		loop filter transfer function, excluding K_A
$H_o(s)$		open loop transfer function
$H(s)$		closed loop transfer function
K	volts/radian	combined loop gain function including limiter suppression factor
K_A		loop amplifier constant
K_m	volts/radian	instantaneous phase detector gain, a function of loop error θ
K'	volts/radian	phase detector gain constant
K_{vco}	volts/cps	loop oscillator gain
K_1		loop gain excluding suppression factor
o	subscript	indicates the value at threshold
s		complex variable
$(S/N)_i$		signal to noise power ratio at input of limiter
$(S/N)_o$		signal to noise power ratio at output of limiter
$(S/N)_{2B_L}$		signal to noise power ratio in 2 sided loop bandwidth
$(S/N)_{2B_N}$		signal to noise power ratio in 2 sided if bandwidth
T_1, T_2	per second	loop filter time constants
$2B_L$	cycles/sec	two sided effective loop noise bandwidth
$2B_N$	cycles/sec	two sided effective if noise bandwidth
p	dbm	receiver signal power

α		limiter suppression factor
Φ_A	sec/cycle	normalized noise amplitude spectral density
Φ_N	sec/cycle	normalized noise phase spectral density
Φ_{KT}	watts/cps	thermal noise density at 290°K
Φ_{NF}	watts/cps	receiver noise density above 290°K
σ_p	rad	rms loop phase error
σ_p^2	rad^2	loop phase variance
θ	rad	phase difference between phase detector input and reference; loop error, $\theta_i - \theta_o$
θ_i	rad	input phase, absolute
θ_o	rad	output phase, absolute
ω_n	rad/sec	loop undamped natural frequency
ζ		closed loop damping ratio

APPENDIX B

MARINER PR/PSK COMMAND AND TELEMETRY SYSTEMS

JPL has used (and is still using) what they call a "two channel PSK system" (see Figure B-1) as the standard technique for their digital communications systems. Essentially, this system transmits pseudo random coded synchronizing information by phase shift keying at $\pm 90°$ and then phase modulating this signal upon a carrier. The spectrum of this signal is shown in Figure B-2 and is similar to the command or telemetry (C/T) spectrum shown in Figure 3. The data subcarrier at the PR code null at $4f_s$, is modulated by the data and added to the sync information. Note that the unmodulated sync subcarrier is at f_s.

Apparently, during the breadboard testing of these systems, the self noise effects within the detector proved to be more seriously degrading than originally anticipated. The phase noise on the VCO output, f_s, was about .15 radians rms for a 6 db (S/N) in 2 B_{LO}. Since the data channel reference is derived, by multiplication to $4f_s$, from this VCO the phase jitter on this reference is, according to Springett, about .6 radians. The effective S/N in the data channel is theoretically degraded by 1.6 db[1]. Measured losses on a breadboard system ran over 4 db for the command system. Attempted solutions to these problems were to increase the power in the sync channel to twice the data channel power where, neglecting the phase noise, the data channel should theoretically have four times the power in the sync signal. The data subcarrier was also moved to $2f_s$ to cut the phase jitter by half.

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1. SPS No. 37-16, Vol IV, JPL, 31 August 1962

None of the attempted solutions satisfactorily solved the reference phase jitter problem. The two channel system will eventually be discarded in favor of the more efficient single channel PSK system.

These problems, encountered by JPL, point up to importance of accurate sync information in a coherent system. According to Stockett's[2] report, the poor performance of General Electric's Costas receiver was due to degradation of the quality of the sync information in that receiver. It is interesting to note that this system is very similar to JPL's single channel system.

In the single channel system (see Figure B-3) the data modulates the synchronization code rather than the separate data subcarrier. (The spectrum is shown in Figure 3). One advantage of this arrangement is that all the power is simultaneously in both the data and synchronization. A major advantage is the elimination of the requirement for one noisy coherent reference. It is important to note that this modification does not change the form or content of the data modulation: it has not been PR coded. The data detection at the receiver will still be a coherent PSK bit by bit detection. Only the synchronization signal is PR coded. It should be noted that, except for the sync signal, the integrate and discharge filter would work just as efficiently if the PR coding was completely removed from the system. These considerations become important when making a comparison with a standard. In his discussion of the system, Springett[3] compares

2. Stockett, T. "Comparison of PSK and PSK/PM Receivers", Technical Memo 1953.5-050, T&T Department, Aerospace, February 1963.

3. Springett, J. "Pseudo Random Coding for Bit and Word Synchronization of PSK Data Transmission Systems", to be given at the International Telemetry Conference, London, 1963.

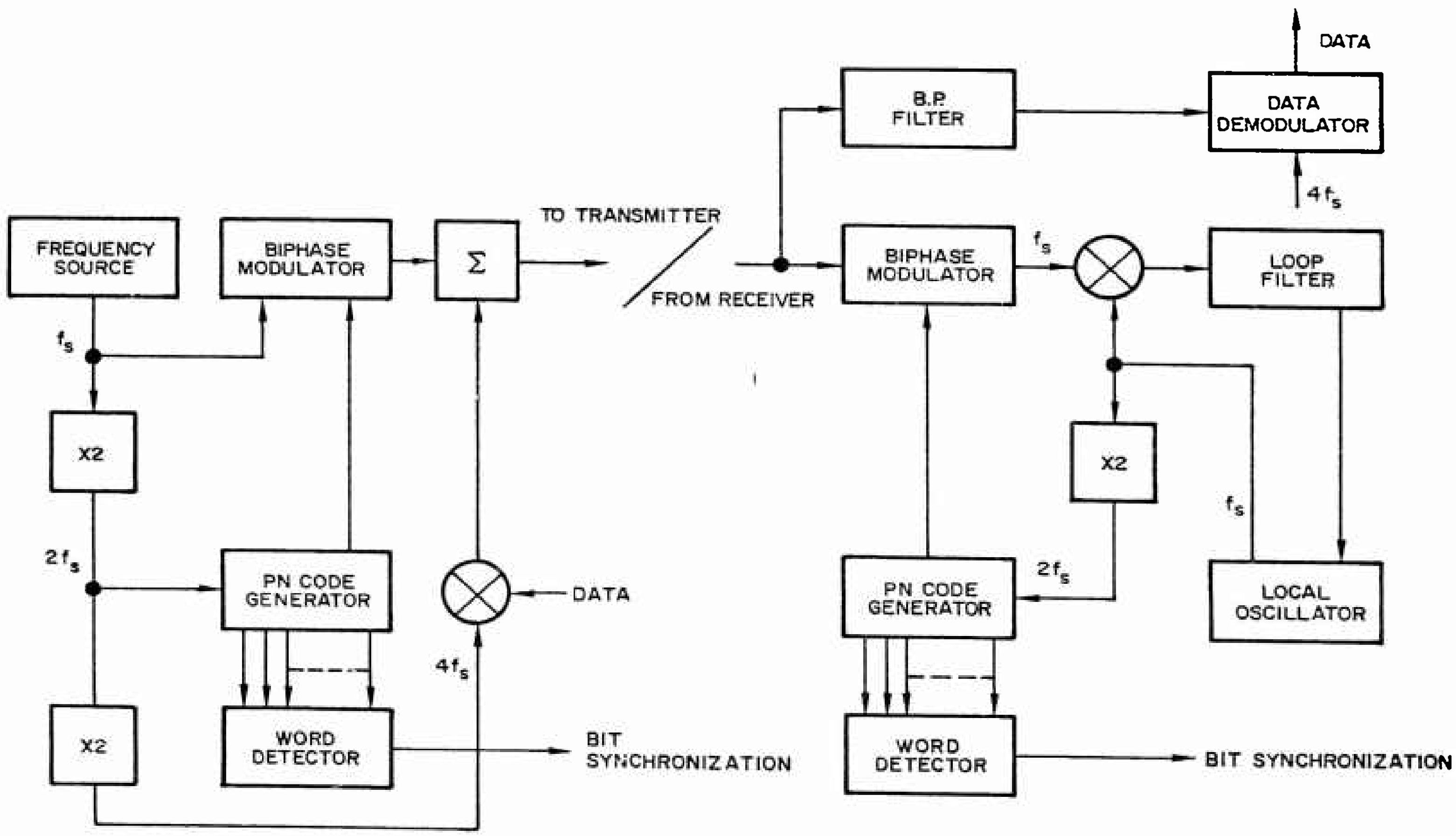

Figure B-1. Two-Channel PSK System

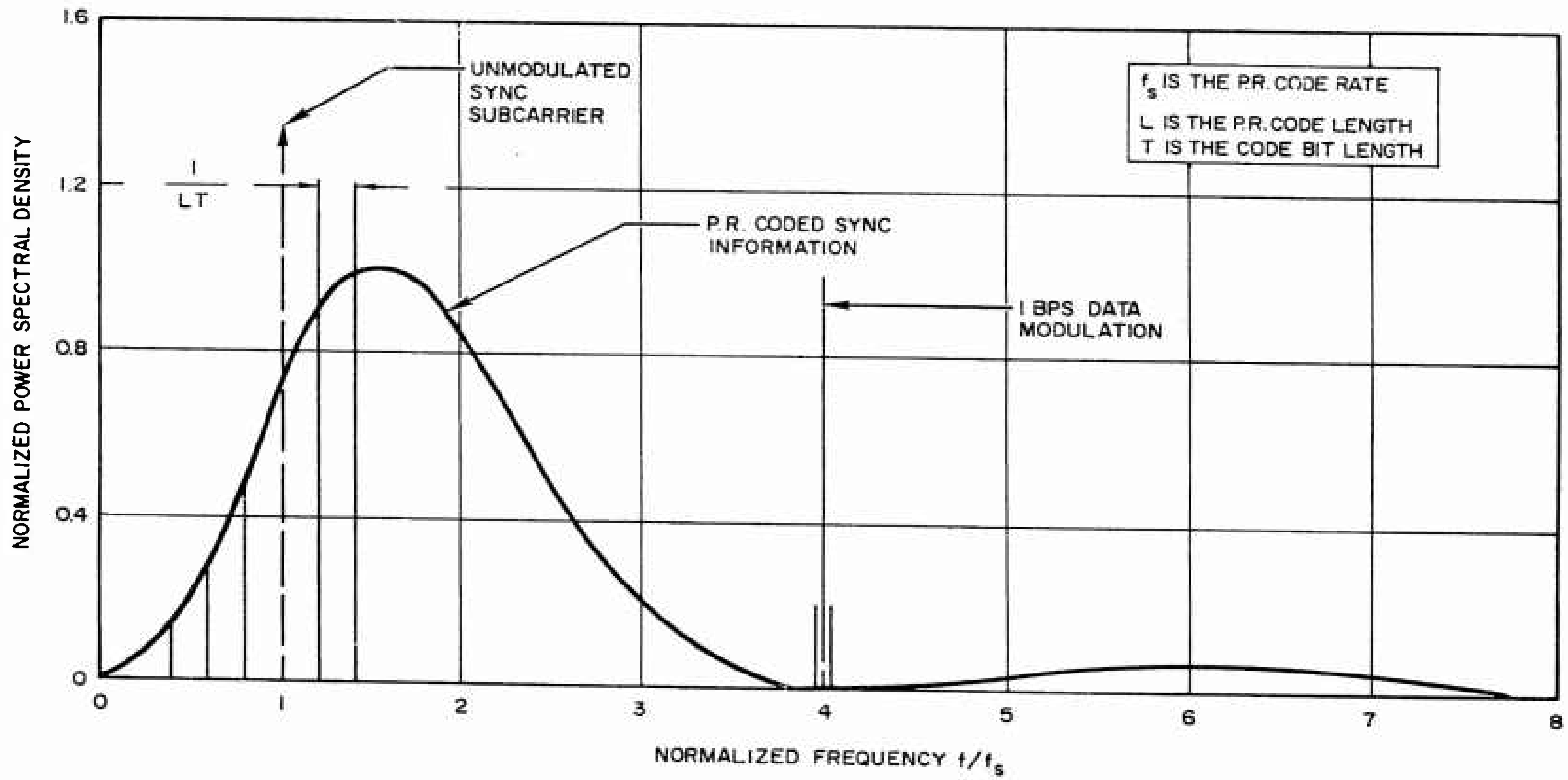

Figure B-2. Two-Channel PSK Spectrum

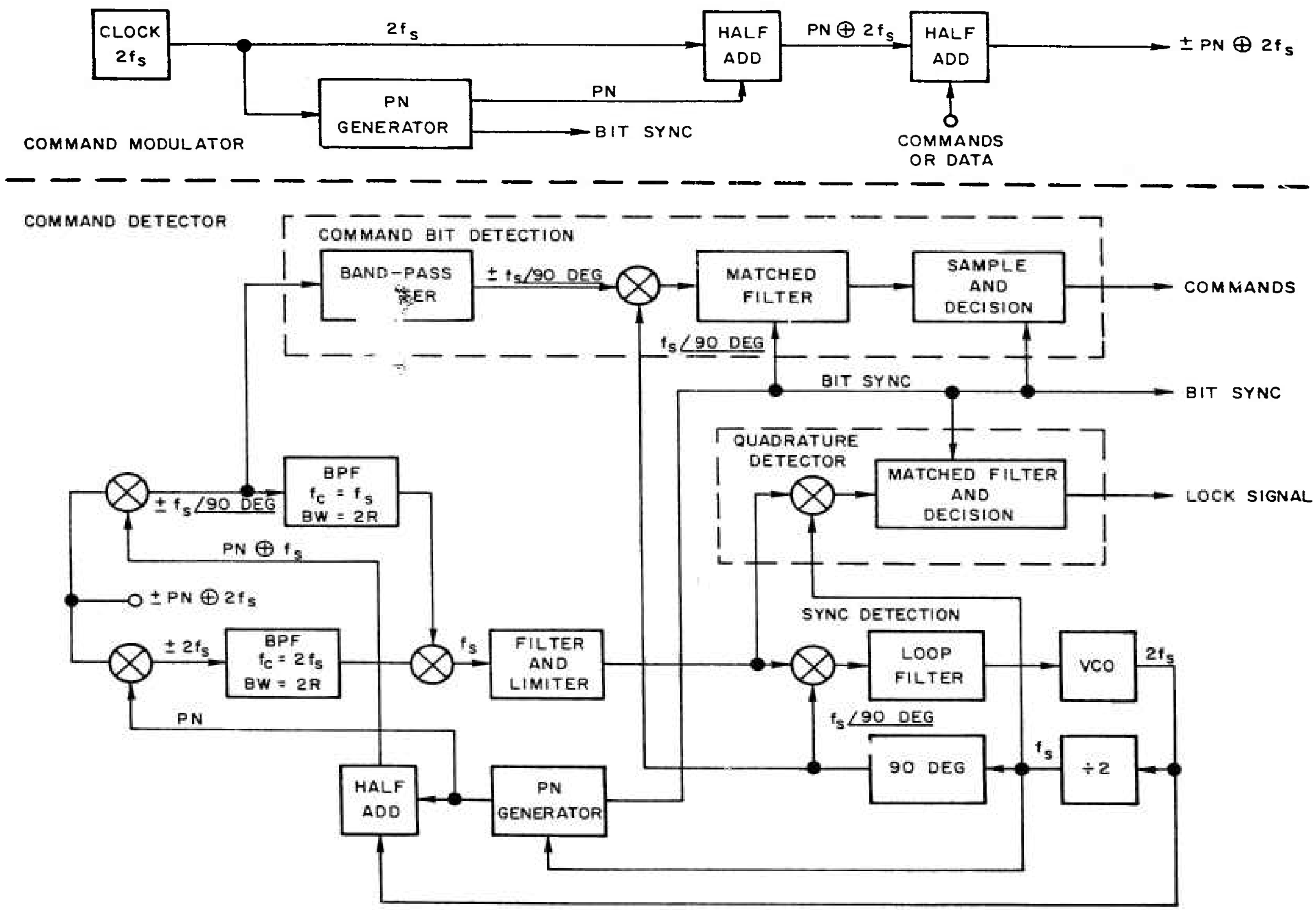

Figure B-3. Single-Channel Command Modulator and Detector the measured performance of the single channel system with the theoretical curve for uncoded coherent PSK bit by bit detection. The curves are shown in FigureB-4,where curve 3 is the "ideal" to which the "single channel" system (curve 1) is compared. The average measured performance of the "single channel" system approaches to within 1/3 db of curve 3. It is tempting to make the mistake of requiring curve 1 to compare against curve 4 for coded coherent PSK bits. This is the case where sequences of information bits are coded into words and is not what occurs in the "single channel" system. Other theoretical curves are shown for further comparison. Curve 2 shows the measured performance of the GE Costas receiver which is quite similar to the single channel receiver in that it performs an in-phase and quadrature multiplication to eliminate modulation from the signal from which the references are derived. Note that the single channel performance is about 4.5 db better than the Costas receiver.

Upon first examining the "single channel" system it would appear that PR coding has little effect upon the data detection (except for synchronization). Due to the particular manner in which the synchronization bits are modulated by the PR code, the data signal energy is translated away from the carrier allowing wider loop filter bandwidths. This then allows a very long data bit time, or high energy per bit, and a long integration time.

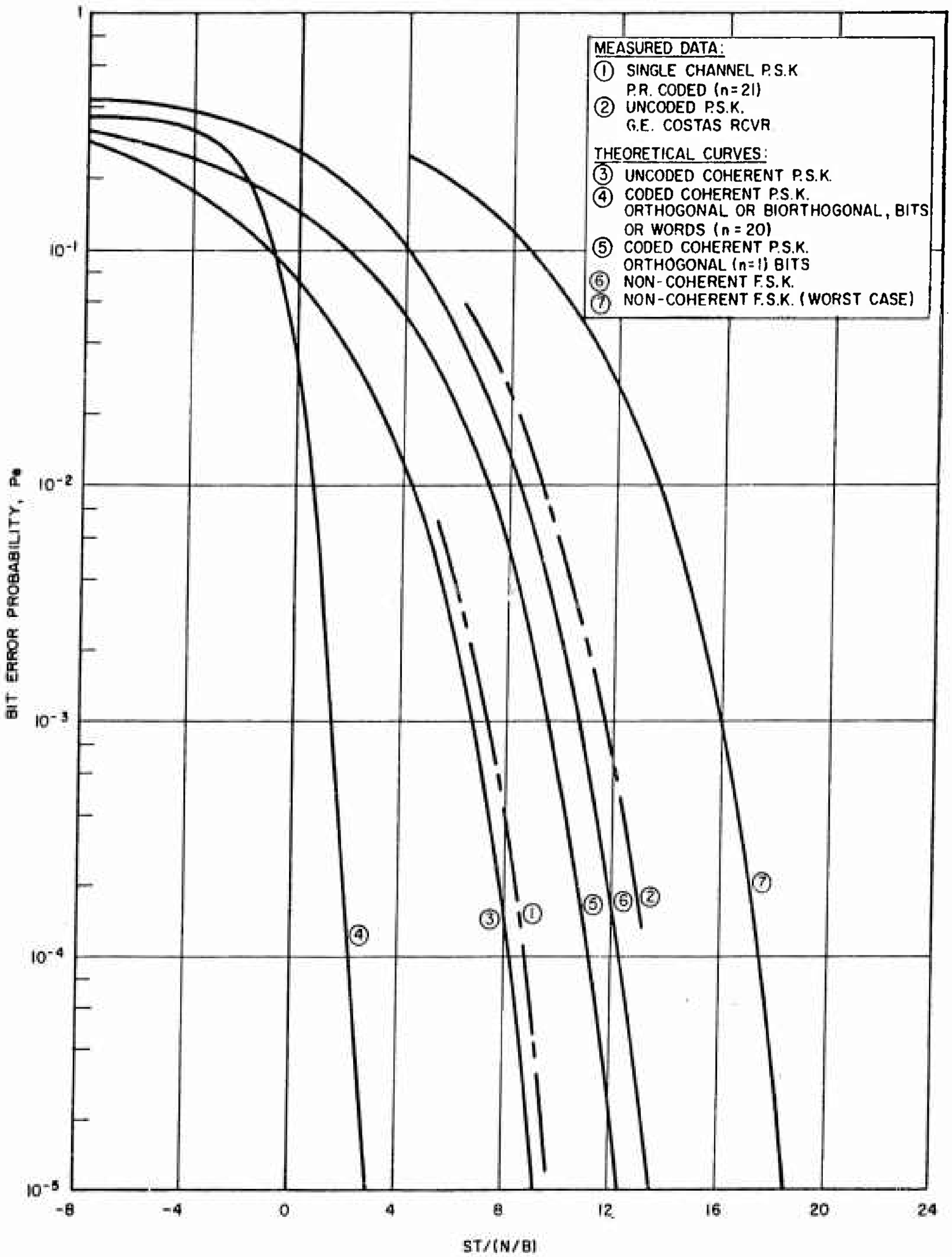

Figure B-4. Comparison of Various Coded and Uncoded Modulation Processes

References for Figure B-4:

Curve 1. Springett, J., "P.R. Coding for Bit and Word Synchronization of PSK Data Transmission Systems", Nat. Telemetry Convention, May 1963.

Curve 2. Stockett, T., "Comparison of PSK and PSK/PM Receivers", Technical Memo 1953.5-050, T&T Department, Aerospace, February 1963.

Curve 3. Cahn, C. R., "Performance of Digital Phase Modulation Systems", PGCS, May 1959.

Curve 4 and Curve 5. Viterbi, Al, "On Coded Phase Coherent Communications Systems", T.R. 32-67, JPL, 31 March 1961

Curve 6 and Curve 7. Springett, I., "Command Techniques for the Remote Control of Interplanetary Spacecraft", T.M. 33-88, JPL, 21 May 1962.

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