TDRSS TELECOMMUNICATIONS STUDY
PHASE II- FINAL REPORT

(NASA-CR-143689) TDRSS TELECOMMUNICATIONS STUDY, PHASE 2 Final Report, Aug. - Dec. 1974 (Magnavox Research Labs.) 59 p HC $4.25 CSCL 17B
N75-17545
Unclas
G3/32 12038

Advanced Products Division
The Magnavox Company
2829 Maricopa Street
Torrance, California 90503

1 December 1974
Final Report for Period August 1974 - December 1974

Prepared for

GODDARD SPACE FLIGHT CENTER
Greenbelt, Maryland 20771

1. Report No.	2. Government Accession No.	3. Recipient's Catalog No.
4. Title and Subtitle TDRSS Telecommunications Study Phase II - Final Report		5. Report Date December 1, 1974 6. Performing Organization Code
7. Author(s) C.R. Cahn R.S. Cnossen		8. Performing Organization Report No. R-4996
9. Performing Organization Name and Address The Magnavox Company Advanced Products Division 2829 Maricopa Street Torrance, California 90503		10. Work Unit No. 11. Contract or Grant No. NAS5-20047
12. Sponsoring Agency Name and Address National Aeronautics and Space Admin. Goddard Space Flight Center Greenbelt, Maryland 20771		13. Type of Report and Period Covered Type III (Final) Aug 1974 to Dec 1974 14. Sponsoring Agency Code Code 805
15. Supplementary Notes		

16. Abstract

Providing an extension to the Phase I report which performed a parametric analysis of the telecommunications support capability of the Tracking and Data Relay Satellite System (TDRSS), this Phase II report considers candidate modulation waveforms which could meet the Shuttle telecommunications requirements and also be compatible with the TDRSS Single Access S-band Service specified in Phase I. In addition, it considers the feasibility of modifying a Single Access S-band User Transponder for operation with conventional STDN signals emanating from remotely located ground stations.

17. Key Words (Selected by Author(s))	18. Distribution Statement
TDRS Satellite Link Analysis TDRS Transponder Design Modulation Tradeoffs Phased Arrays	

19. Security Classif. (of this report)	20. Security Classif. (of this page)	21. No. of Pages	22. Price*
Unclassified	Unclassified	59	

*For sale by the Clearinghouse for Federal Scientific and Technical Information, Springfield, Virginia 22151.

PREFACE

NASA has conducted several studies, over the past several years, of a Tracking and Data Relay Satellite System (TDRSS) to augment the current tracking and data network. The results of these studies, which have established that a TDRSS is both feasible and cost effective, led to awards for major definition studies in May 1971. The studies were directed primarily to the relay spacecraft elements of the system. At the same time, several Goddard Space Flight Center (GSFC) in-house studies were directed to the design of the ground elements and overall system operational aspects. The integrated results of all these studies were published in a "TDRSS Definition Phase Study Report", December 1973.

Subsequently, the Magnavox Company was awarded a study contract directed toward the telecommunications aspects of the system. A Phase I report dated September 1974, contained the results of a study to (1) determine the capability of the TDRS System as specified by the TDRSS Definition Phase Study Report, (2) determine potential configuration changes to satisfy all telecommunications performance objectives without major redesign, (3) perform analysis of significant system variables for the purpose of optimizing parameters and, (4) define the TDRSS telecommunications characteristics in greater detail than previously established.

This Phase II report, dated December 1974, contains the results of a study to (1) determine alternative TDRSS implementations that can provide superior telecommunications support capability within the fundamental launch vehicle, frequency allocation, weight, and power constraints, but without restriction to current system definition and (2) perform comparative performance analyses between the TDRSS defined in Phase I and alternative configurations.

Work on this report, entitled TDRSS Telecommunications Study, was accomplished by the Advanced Products Division of The Magnavox Company and complies with the requirements of Contract Number NAS5-20047. This report is the result of a study carried out by Dr. C.R. Cahn and Mr. R.S. Cnossen of the APD/Magnavox Research Laboratories. APD wishes to thank Mr. Leonard F. Deerkoski of NASA/GSFC, Greenbelt, Maryland, for his technical and administrative guidance during Phase II of this program.

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OBJECTIVE

The Tracking and Data Relay Satellite System (TDRSS) is designed to provide a relay capability between a ground station and low altitude user vehicles. The telecommunication services for S-band and Ku-band have been analyzed in detail during Phase I of the study. [1]

The Phase II goal of the study is:

a. Determine alternate TDRSS implementations that can provide superior telecommunication support capability within the fundamental launch vehicle, frequency allocations, weight, and power constraints; but without restriction to current system definition.

b. Perform comparative performance analyses between the TDRSS defined in Phase I and alternative configurations.

The following sections are in compliance with this Phase II goal.

SECTION II

TELECOMMUNICATION SERVICE TO SHUTTLE

This section discusses some of the special problems associated with TDRSS support of the Shuttle (Orbiter), in accordance with requirements defined by NASA-JSC. [2] The modulation approaches described here may differ from signal designs currently suggested by NASA-JSC.

2.1 SYNCHRONIZATION OF SPREAD SPECTRUM FORWARD LINK TO SHUTTLE

The single access forward link from TDRS to Shuttle on either S-band or Ku band is required to be spread spectrum, approximately 15 MHz bandwidth, so as to meet flux density limits on the TDRS transmissions. However, in contrast to other TDRS users, the Shuttle does not have a two-way ranging requirement. There is, instead, a one-way range rate measurement made with the aid of an on-board atomic frequency standard.

A demand for protection against the unlikely possibility of specular multipath is not imposed, and a relatively short PN code can be employed at a chip rate of approximately 12 Mbps. A code period of 2047 chips (or 2048 chips) is the minimum consistent with 4 kHz bandwidth for flux density (i.e., a discrete spectrum should have a spectral line spacing comparable to or less than 4 kHz).

The synchronization requirement for Shuttle is somewhat different than for other users in that digital voice (plus commands) is being transmitted. Useful (although noisy) voice can be received down to $E_b/N_o = 0$ dB, where the error rate is high. Rate 1/3 error correction coding is utilized, and we assume the data rate is 32 Kbps (thus, the symbol rate is 96 Kbps). The capability for synchronization at this minimum value of E_b/N_o is desired, although a longer acquisition time is acceptable than at the design point of $E_b/N_o = 3.9$ dB for 10^{-4} error rate.

The synchronization philosophy is that the transmission to the Shuttle is initiated with data already present, and acquisition should be completed in a reasonably short time on the order of 5 seconds, although 20 seconds or even longer would not be unacceptable for the initial acquisition. However, faster reacquisition should be possible after a dropout (due to antenna nulls) of a few seconds. The measure of acquisition time is the average time to sync, or, alternatively, the time to sync achieved on 90 percent of trials*.

2.1.1 SYNCHRONIZATION ANALYSIS

We adopt the synchronization process described in Appendix III of the Phase I - Final Report. This employs a sequential detection strategy to speed up the synchronization search. A single filter design is contemplated, with the structure shown in figure 2-1. Typical performance, as derived by computer simulations, is given in

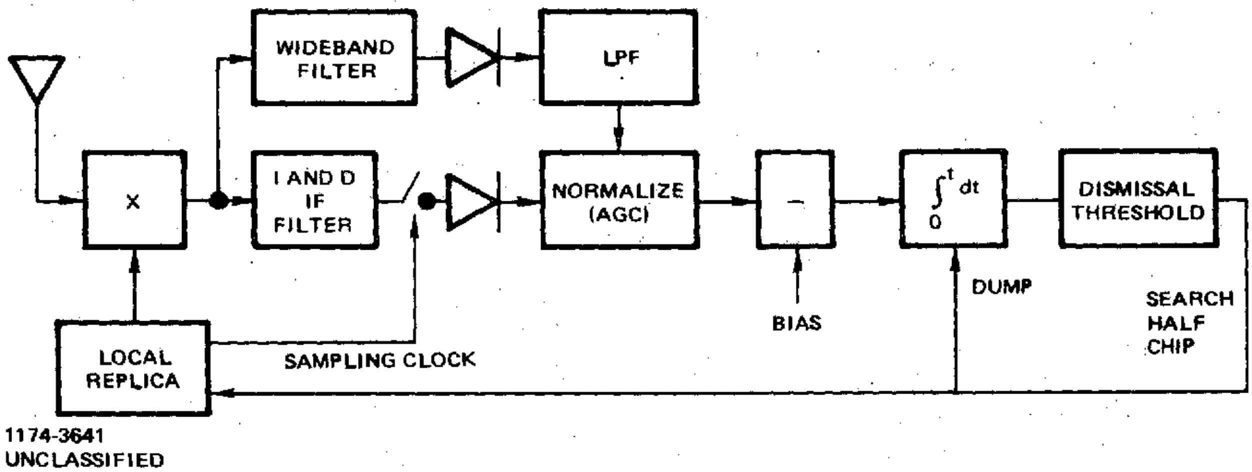

Figure 2-1. Synchronization Circuit Concept

figure 2-2. The bias is introduced to cause the integrated envelope to fall below a dismissal threshold when correlation does not exist in that particular search position. When correlation occurs in the correct search position, the envelope will tend to exceed the bias, and dismissal does not occur. Synchronization is declared by failure to dismiss after a designated truncation interval. The noise level is measured in a wide bandwidth to set receiver gain (i.e., a total power AGC is implied).

The average search rate enables the time uncertainty (one code period for initial acquisition) to be searched in a time T_s, defined by

$$T_s = \frac{\text{No. of PN chips of time uncertainty}}{\text{Average search rate in chips/sec}} \qquad (2\text{-}1)$$

*For other TDRSS users, we have adopted the criterion on achieving, say, 90 percent probability of acquisition in a specified maximum time.

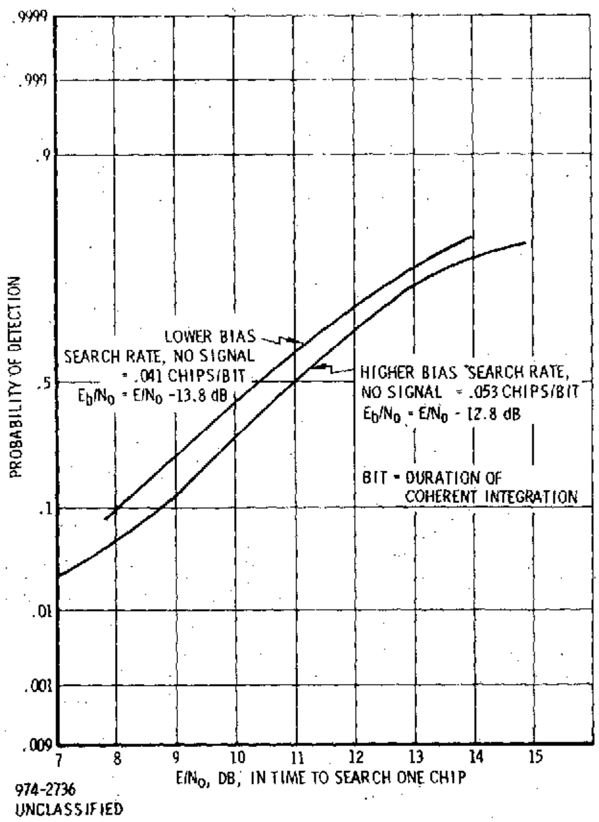

Figure 2-2. Synchronization Performance

figure 2-2 gives the probability P_d of detecting sync on a single pass through the time uncertainty, when the frequency error is zero. Then, the cumulative probability $P_d^{(M)}$ of detecting sync after M passes is

$$P_d^{(M)} = 1 - (1 - P_d)^M \tag{2-2}$$

since each pass is an independent chance to detect sync. The time $T_s^{(.9)}$ to detect sync which is not exceeded 90 percent is approximately given by solving (2-2) for M with $P_d^{(M)} = .9$ and assuming an average rate of search per pass, yielding

$$T_s^{(.9)} = MT_s = T_s \frac{\log(.1)}{\log(1 - P_d)} \tag{2-3}$$

figure 2-3 presents a plot of (2-3) as a function of P_d.

The average time to detect sync is computed by noting that the average number of passes required is $1/P_d$, obtained from the binomial distribution (i.e., "coin flipping"). Since detection occurs, on the average, halfway through the last pass,

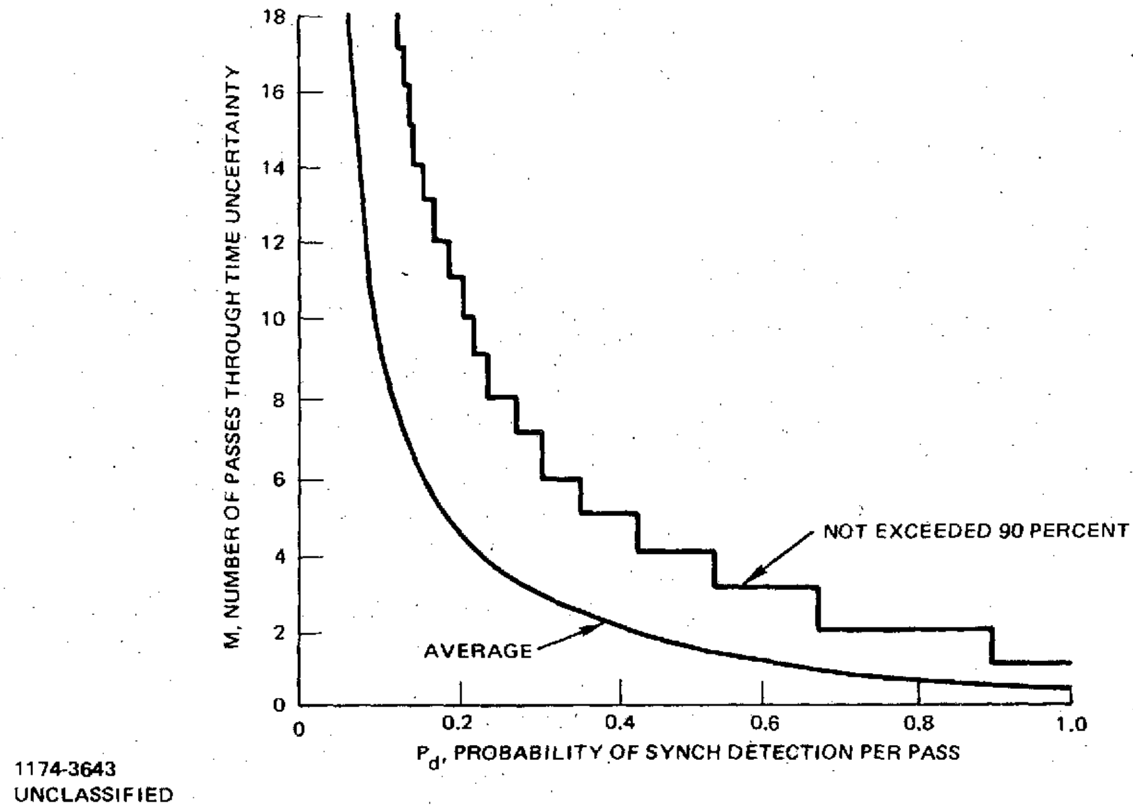

Figure 2-3. Number of Passes to Detect Sync

$$T_s^{(av)} = T_s\,(-.5 + 1/P_d) \tag{2-4}$$

Equation (2-4) is plotted in figure 2-3 as a function of P_d.

By combining figures 2-2 and 2-3, the average synchronization time and the time not exceeded 90 percent can be expressed as a function of received signal-to-noise ratio. However, it should be remembered that Doppler still must be taken into account.

2.1.2 NUMERICAL RESULTS FOR SHUTTLE AT S-BAND

For S-band operation to the Shuttle, assume a PN code period of 2047 chips and a symbol rate of 96 Kbps, the data rate being 32 Kbps biphase with rate 1/3 coding. Assume the same losses apply during synchronization as during data demodulation, and let the duration of coherent integration be one symbol, or 10.4 microseconds. Using the curve in figure 2-2 for lower bias, we find

$$\text{Average Search Rate} = .041 \times 96 \times 10^3 = 3900 \text{ chips/sec} \tag{2-5}$$

and for one pass through 2047 chips of uncertainty (one code period)

$$T_s = \frac{2047}{3900} = 0.5 \text{ second/pass} \tag{2-6}$$

As already mentioned, figure 2-2 can be used with figure 2-3 to yield the acquisition time as a function of E_b/N_o*. We have (using the curve for lower bias of figure 2-2)

$$\begin{aligned} E/N_o &= E_b/N_o - 4.8 \text{ dB} + 13.8 \text{ dB} \\ &= E_b/N_o + 9 \text{ dB} \end{aligned} \tag{2-7}$$

as the abscissa of figure 2-2 to find P_d.

We now bring in the effect of Doppler. Figure 2-2 is for zero Doppler, and with a carrier frequency error Δf, the correlated amplitude drops according to

$$\text{Amplitude} = \frac{\sin(\pi \Delta f T_{coh})}{(\pi \Delta f T_{coh})} \tag{2-8}$$

where T_{coh} is the duration of coherent integration (10.4 microseconds in the present case). Maximum Doppler at S-band (26 parts per million) is 55 kHz, for which (2-8) yields an amplitude of 0.54, or 5 dB loss. Because of this loss, we break the frequency uncertainty -55 kHz to 55 kHz into two regions, with a maximum loss now of 1.2 dB at maximum Doppler. The two frequency uncertainty regions are searched serially, doubling the acquisition time**.

*Note, figure 2-2 defines E_b/N_o for the duration of coherent integration, or one symbol. Here E_b/N_o is defined to apply to a data bit, and is 4.8 dB higher (rate -1/3).

**In addition to carrier Doppler, we sometimes have to consider code Doppler. For a PN chip rate of 12 Mbps, maximum code Doppler is 312 chips/sec over the full uncertainty, or 156 chips/sec with two uncertainty regions. Figure 2-2 is based on a truncation of about 200 coherent integration intervals, or 2 milliseconds. For a Doppler of 156 chips/sec, the relative drift is 0.3 chip during the truncation. Thus, we are about at the limit of tolerable code Doppler in the present design.

The alternative of offsetting the ground transmission by an estimate of Doppler is not allowed because of the one-way range rate measurement to be made on-board the Shuttle. Offsetting the receive frequency by an estimate of Doppler would be desirable, if such information is available on-board the Shuttle prior to the initial sync acquisition.

Figure 2-4 presents the computed acquisition performance results, with 1.2 dB of loss due to Doppler included, based on serial search over two frequency uncertainty regions.

2.1.3 REACQUISITION CONSIDERATIONS

If there is a temporary loss of signal in the Shuttle's receiver, such as could be caused by an antenna null, the receiver will stop tracking the PN, and there will be a drift of code phase in the receiver until the received signal level is restored to a usable value. Because Shuttle is making one-way Doppler measurements with a stable oscillator (say 10^{-9} accuracy), the drift is almost entirely due to unpredicted motion* relative to TDRS.

Let us assume dynamics due to orbiting only, which means a maximum possible acceleration of lg, or 9.8 m/sec^2, relative to TDRS. If the Shuttle makes no attempt to estimate the relative acceleration (a third-order carrier tracking loop would provide such an estimate) but uses the range rate measurement (which is essentially perfect), the range error builds up according to

$$\text{Maximum range error}\Big]_{\substack{\text{PN chips}\\ \text{at 12 Mbps}}} = 0.2t_{\text{sec}}^2 \qquad (2\text{-}9)$$

We see that an error of ±1024 PN chips can be reached after 72 seconds.

It is concluded that with a PN code period of 2047 chips, it is preferable to search about the estimated time position of correlation for a maximum of 72 seconds after loss of signal. The range uncertainty builds up as the square of elapsed time. After 72 seconds, the receiver should revert to the initial sync acquisition process. Note that a PN receiver implementation with a very narrow code tracking loop which is aided by scaling from the carrier tracking loop automatically extrapolates ranges on the basis of one-way Doppler information.

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*With 10^{-9} oscillator frequency accuracy, there is 0.1 microsecond error after 100 seconds, or one PN chip at 12 Mbps chip rate. The error of the one-way Doppler measurement itself is less than one-half cycle in 2 x 10^9 for one second averaging.

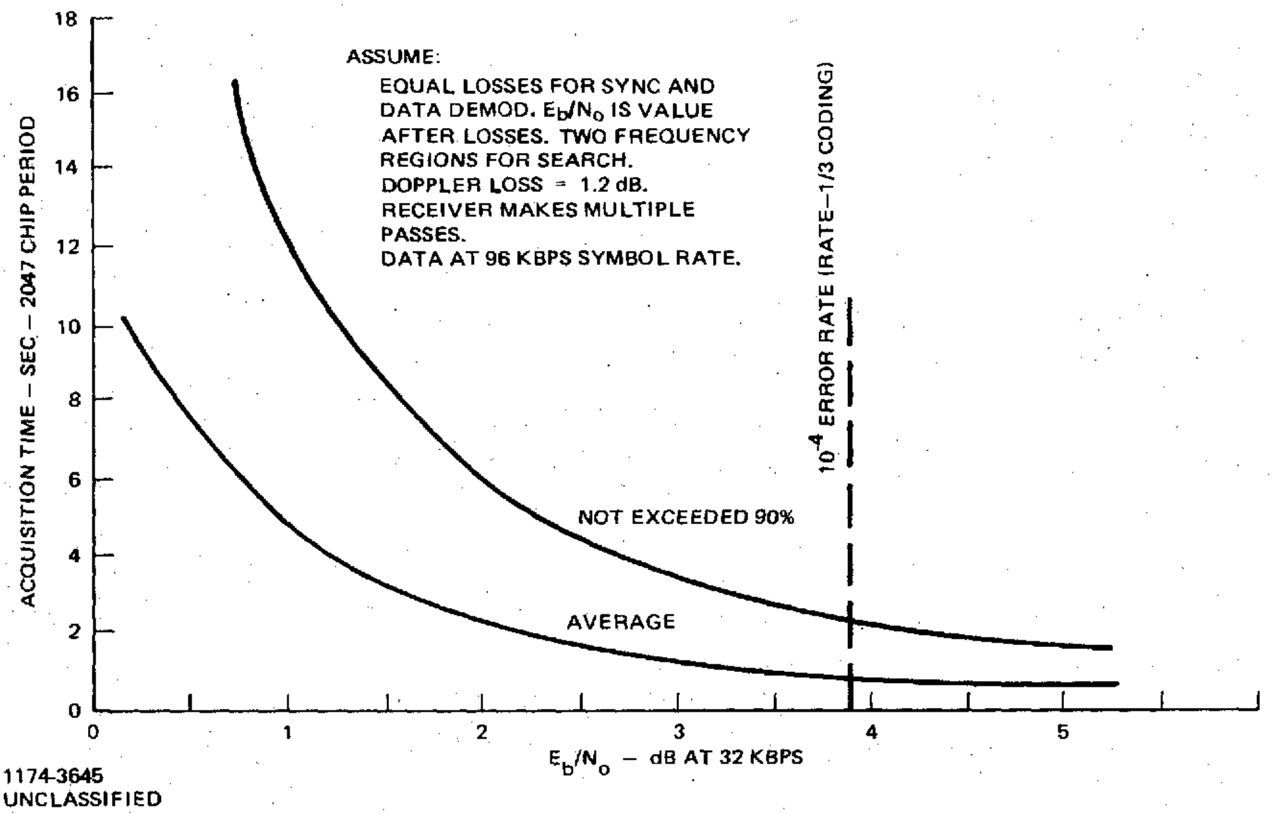


Figure 2-4. Acquisition Time as Function of Receiver S/N

2.1.4 CONCLUSIONS

An analysis of the TDRS-to-Shuttle link at S-band has been carried out for a data rate of 32 Kbps (the lowest data rate of concern). The receiver is presumed to search continually through the 2047 chip period of the PN code until synchronization is detected. As S/N in the receiver drops, the acquisition time increases because the receiver has to make several passes through the full period until sync is detected. Acceptable initial acquisition time (average of about 10 seconds) is achieved for a received signal-to-noise ratio as low as $E_b/N_o = 0$ dB (at 32 Kbps). The average initial acquisition time is less than 1 second at $E_b/N_o = 4$ dB. The receiver requires no knowledge of S/N for this acquisition concept, but sets the detection thresholds on the basis of total power. As E_b/N_o decreases, the probability of detection per pass decreases.

The acquisition time has been doubled and 1.2 dB of loss included to take into account a maximum Doppler of 55 kHz. If a coarse estimation of Doppler is possible on-board the Shuttle, improved acquisition performance can be realized.

If loss of signal occurs temporarily, the receiver should search about the code phase prior to loss updated by one-way Doppler. (A narrow PN delay lock loop aided by scaling from the carrier tracking loop automatically performs this updating.) The error due to acceleration in orbit of 1g is less than the code period until 72 seconds has elapsed. The search should be conducted over an aperture which is growing quadratically with elapsed time.

It should be pointed out that the implicit assumption has been made that the data is synchronous to the PN. The integrate-and-dump IF filter has its coherent integration extended over a data symbol, whose timing is fixed relative to the PN code repetition marker. The data symbols can be either in NRZ or Manchester format without affecting the analysis presented above. (The format is, of course, known.) With asynchronous data, the sync acquisition performance would be poorer, and a sampled IF would have to be replaced by a continuous filter. Also, data demodulation would be slightly degraded by the quantizing noise due to reclocking asynchronous data transitions. Thus, we recommend use of synchronous data on the TDRS-to-Shuttle link, where there is no operational difficulty involved. (Command bits are stored in a buffer anyway, and the voice digitizer can use a clock derived from the PN chip rate.)

As a final comment, the PN design has a code period of 170 microseconds and provides no protection against a possible multipath delay which is exactly a multiple of the code period. Delays as great as 33 milliseconds can be experienced at an altitude of 5000 Km.

2.2 RETURN LINK MODULATION SCHEME FOR SHUTTLE

There are two modes specified for the Shuttle/TDRS return link: [2]

Mode 1 - Simultaneous transmission of
2 Mbps digital data or 192 Kbps digital data,
50 Mbps wideband digital data.

Mode 2 - Simultaneous transmission of
4.2 MHz baseband video,
2 Mbps digital data,
192 Kbps digital data.

Simultaneous transmission in mode 1 of 2 Mbps data (playback from recorder) and 192 Kbps real time data is actually desired, along with the 50 Mbps wideband data. These three data streams have independent clocks, and the maximum rates are specified. The modulation should be of constant envelope type.

2.2.1 MODULATION - MODE 1

A solution for handling two independent data streams is by quadrature modulation as sketched in figure 2-5. After coherent demodulation in the receiver, there are two independent baseband outputs, and the respective data clocks are recovered by independent bit synchronizers. If the two channels have disparate data rates, they can be distinguished in the demodulator without ambiguity. In the modulator, the two channels ideally are given amplitude weighting proportional to the square root of their data rates. (A practical limit would be imposed on the amplitude ratio.) Note that the resultant modulated carrier has constant envelope even though the two channels do not have the same amplitude.

The above describes the technique for quadrature multiplexing of the 50 Mbps NRZ data stream with the 2 Mbps data stream. However, if the 50 Mbps is rate -1/2 error correction coded, the symbol rate is 100 Mbps. The quadrature multiplexing precludes handling this 100 Mbps stream as a staggered quadriphase transmission; however, the wideband Ku-band return link of TDRS has 225 MHz bandwidth, ample for 100 Mbps.

Because of the wide bandwidth needed to pass the 50 or 100 Mbps symbol rate, asynchronous TDM is feasible to multiplex 192 Kbps with the 2 Mbps data. The scheme reserves 10 percent of each 2 Mbps data bit for an independent low-rate asynchronous data channel, as suggested in figure 2-6. Bit synchronization of the 2

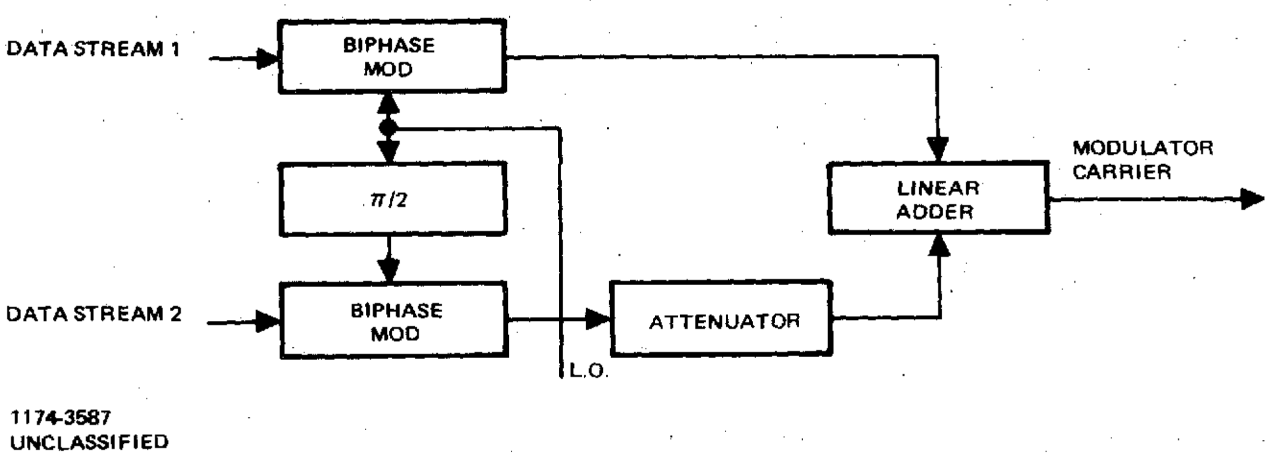


Figure 2-5. Quadrature Modulation with Two Data Streams

Mbps data is much more accurate than 10 percent, and the bandwidth is wide enough to resolve the 10 percent portion. Thus, we can transmit and receive the asynchronous 192 Kbps data by sampling the data waveform at the 2 Mbps data clock. An independently acting bit synchronizer to extract the asynchronous 192 Kbps clock is required.

There is a slight inefficiency associated with the asynchronous TDM but only on the 192 Kbps data. Hence, the inefficiency is inconsequential. The inefficiency arises because the 192 Kbps clock tends to move discretely by one sample as necessary to create the correct average rate; hence, an extra sample can occasionally be included incorrectly in a given bit (or taken away from a given bit). With 10 samples per bit, this effect is not of much concern.

2.2.2 MODULATION - MODE 2

In mode 2, we replace the 50 Mbps data with suppressed carrier PDM at a sampling rate in excess of twice the 4.2 MHz baseband to avoid aliasing. A sampling rate of about 10 Mbps can be employed. Figure 2-7 illustrates SCPDM[3], which generates a phase modulated square wave at half the sampling rate. Conditioning of the input may involve AGC, peak clipping, and preemphasis.

Theoretical analysis and computer simulation [4] shows that the output test-tone-to-noise ratio of unfiltered SCPDM with maximum-likelihood demodulation is approximated by

$$(S/N)_o \cong 0.5(E_s/N_o)^2 \qquad (2\text{-}9)$$

not including any further improvement from filtering to the baseband (if less than twice the sampling rate) and from preemphasis of the transmitted spectrum. In (2-9),

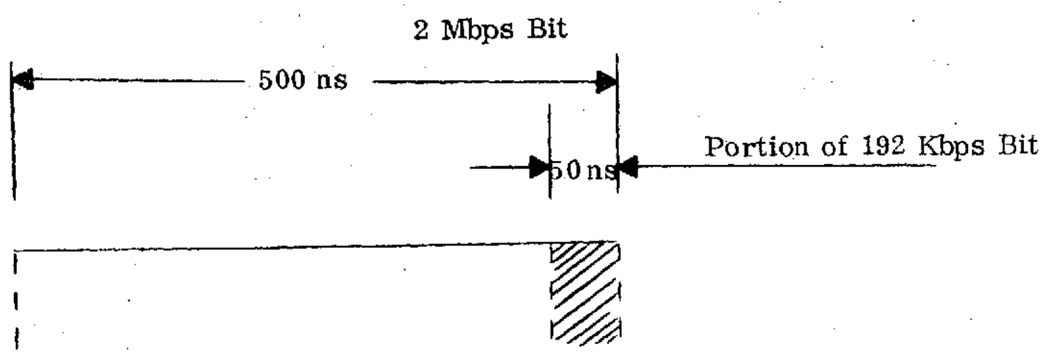
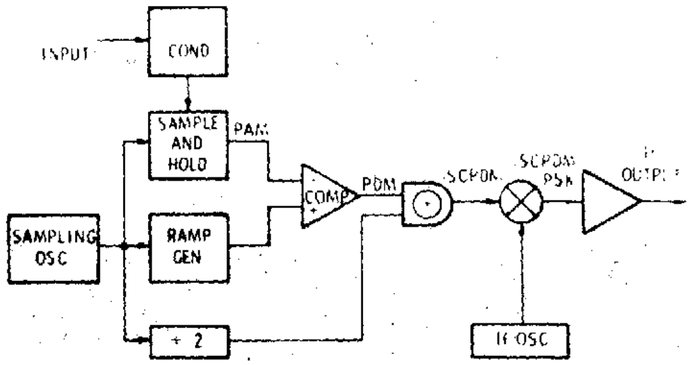
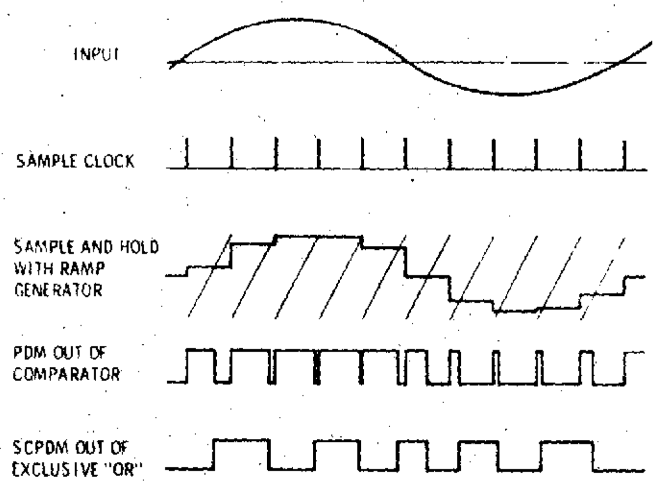


Figure 2-6. Asynchronous TDM Concept



(a) SCPDM Modulator



(b) SCPDM Modulator Waveforms

Figure 2-7. SCPDM

E_s/N_o is the energy/noise ratio over the sampling interval. Taking $(S/N)_o = 26$ dB as the requirement, (2-9) yields $E_s/N_o = 14.5$ dB. At the minimum theoretical sampling rate of 8.4 Mbps, $S/N_o = 83.7$ dB-Hz. The effect of filtering the SCPDM waveform is to create a finite switching time, and this ultimately causes (2-9) to be replaced by a linear input-output relation when E_s/N_o becomes sufficiently high. The 225 MHz bandwidth of the Ku-band return link does not degrade performance.

The SCPDM replaces the 50 Mbps data used in mode 1. Thus, the 192 Kbps data still is multiplexed with the 2 Mbps data by asynchronous TDM. The resultant is quadrature multiplexed with the SCPDM. (Note, S/N_o for 50 Mbps data, rate - 1/2 encoded, is 83 dB-Hz, based on $E_b/N_o = 6.0$ dB, and this is about the same as needed to handle 4.2 MHz baseband by SCPDM.)

The demodulator can unambigously identify which channel contains the SCPDM by recognizing the square wave component at half the sampling rate. This square wave is also used to regenerate the sampling clock. Consequently, the modulation baseband cannot extend all the way to DC, but must cutoff above the bandwidth of the SCPDM clock tracking loop.

An implementation of SCPDM quadrature multiplexed with digital data has been developed for an audio baseband (4 kHz audio and 2.4 Kbps data).[5] This modem does not attempt to perform maximum-likelihood demodulation since it emphasizes operation at low signal-to-noise ratios. Consequently, the demodulation is simply integrate-and-dump over the full sampling interval. The extension to a maximum-likelihood demodulation is conceptually straightforward, because the time of occurrence of the maximum integrated voltage during the sampling interval is the maximum-likelihood estimator of the switching instant.

SECTION III
SQPSK DEMODULATOR DESIGN PROBLEMS

A scheme for reconstituting the phase reference is required for coherent demodulation of staggered quadriphase (SQPSK). One alternative is, of course, to transmit a separate carrier component, unmodulated by the data, which shares the total power with the data modulated component. The phase reference is derived by tracking this carrier component, and if it is a spread spectrum signal, it serves for timing extraction also (e.g., for ranging)[6]. The amount of power to be allocated to the carrier component may be determined in a straightforward manner from the requirement that the signal-to-noise ratio in the bandwidth of the tracking loop exceed roughly 20 dB to produce an rms error of 0.1 radian (required value depends on desired error rate and error correcting code performance)[7].

We study here the use of suppressed carrier tracking, which for SQPSK is a generalization of the familiar Costas loop or squaring loop for BPSK. In particular, the problem of working at a low E_b/N_o is of concern. (Note, E_b/N_o is defined herein as the energy/noise density for a transmitted binary digit, rather than for an information bit. Thus, with a rate -1/2 error correcting code E_b/N_o = 2 dB when $E_{\text{information bit}}/N_o$ = 5 dB.) Implementation of QPSK demodulators has typically[8] not emphasized operation close to theoretical bounds at low E_b/N_o; however, this is essential for transmission of data which has been error correction coded or which conveys digital voice.

3.1 DECISION-FEEDBACK TRACKING FOR QPSK

Let us first assume QPSK instead of SQPSK. The QPSK waveform is

$$s(t) = a(t) \cos \omega_o t + b(t) \sin \omega_o t \tag{3-1}$$

where a and b denote the respective binary data streams, and the bit transitions on the two streams coincide. If the reconstituted phase reference has a shift θ, coherent demodulation of (3-1) with additive noise yields the two output channels

$$\begin{aligned} I &= a(t) \cos \theta + b(t) \sin \theta + x(t) \\ Q &= -a(t) \sin \theta + b(t) \cos \theta + y(t) \end{aligned} \tag{3-2}$$

where x and y are independent Gaussian noise voltages.

If good estimates of a and b are extracted by the demodulation, decision feedback yields an error voltage for tracking, according to

$$\begin{aligned} \epsilon &= \hat{a}(t)\,Q - \hat{b}(t)\,I \cong a(t)\,Q - b(t)I \\ &= -(a^2 + b^2)\sin\theta + a(t)\,x\,(t) - b(t)\,y(t) \end{aligned} \tag{3-3}$$

where $\hat{a}$ and $\hat{b}$ denote the estimates*. Note that terms involving $\cos\theta$ cancel out on the assumption that the estimates are correct. The form of (3-3) suggests deriving the error voltage by decision feedback as follows

$$\epsilon = \mathrm{sign}(I)\,Q - \mathrm{sign}(Q)I \tag{3-4}$$

where $\hat{a} = \mathrm{sign}(I)$ and $\hat{b} = \mathrm{sign}\,(Q)$. Of course, there are four stable tracking points spaced by multiples of $\pi/2$, where $\epsilon = 0$. Equation (3-4) is a generalization of the Costas loop for BPSK, which derives ϵ by one of the terms on the right side of (3-4).

If integrate-and-dump over the bit duration is performed on each channel, (3-4) is computed at the quadriphase symbol rate (equal to half the total bit rate). Alternatively, a single-pole low pass filter can be used on each channel to smooth I and Q prior to computing ϵ according to (3-4).

The performance of the suppressed-carrier tracking loop can be estimated from the linearized equivalent circuit defined by the average slope of ϵ as a function of θ and the noise density associated with the fluctuations of ϵ. These parameters can be measured by a computer simulation which measures mean and variance of ϵ. Results for QPSK with integrate-and-dump filtering are given in figure 3-1, with ϵ normalized by the signal amplitude. The variance of ϵ is indicated in terms of the equivalent noise density of white noise in the tracking loop, and the unit of bandwidth is the total bit rate**. The ideal curve for large E_b/N_o is $\epsilon = 2\sin\theta$, with the normalization employed, with $|\theta| < \pi/4$. The error characteristic is periodic, with period $= \pi/2$, corresponding to the four stable tracking points.

*Decision feedback, in general, introduces the problem of compensating for demodulation delay, but this is not a difficulty in the schemes described herein.
**The measured variance = Noise Density/(2 x integration interval).

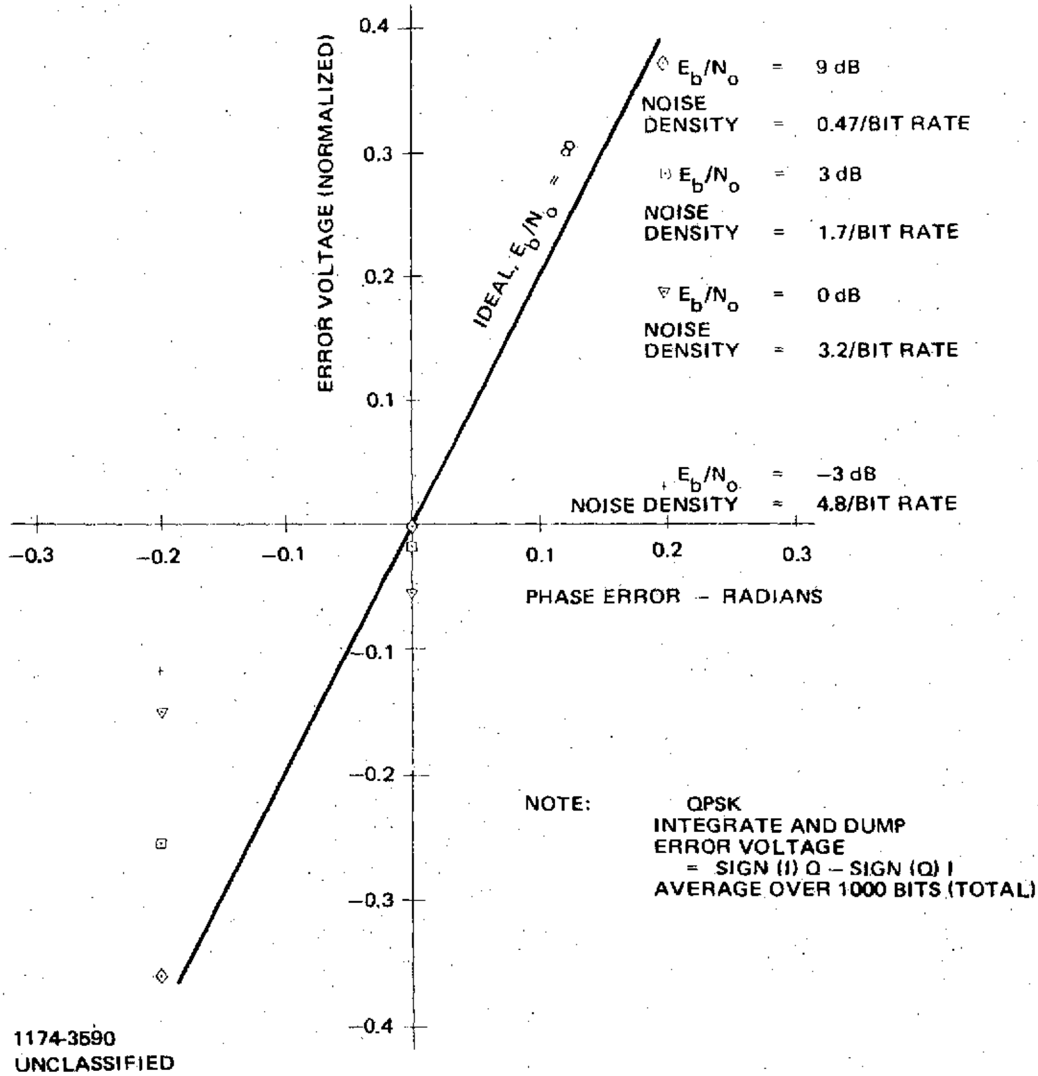


Figure 3-1. QPSK Error Characteristic

The use of the numerical information from figure 3-1 is now illustrated. For a loop of noise bandwidth B_L, the rms tracking error in radians is obtained from the linearized equivalent circuit to be

$$\sigma_\theta = (\text{Noise Density} \cdot B_L)^{.5}/\text{slope} \tag{3-5}$$

At $E_b/N_o = 0$ dB, the slope is approximately $.18/.2 = 0.9$. Since the Noise Density $= 3.2$/bit rate, (3-5) gives $\sigma_\theta = 2(B_L/\text{Bit Rate})^{.5}$. Then, if we require $\sigma_\theta = .1$ radian, B_L/Bit Rate = .0025.

There is a potential problem associated with suppressed carrier tracking, that a false lock condition can exist where the frequency is offset from the true frequency. For instance, with QPSK, a stable false lock can exist if there is a phase change of $\pi/2$ from one quadriphase symbol to the next. Thus, the offset frequency is 1/8 the total bit rate*. This false lock condition can be avoided provided that the maximum carrier frequency uncertainty is constrained to be less than the smallest false lock offset.

It is seen from figure 3-1 that a definite degradation occurs when E_b/N_o drops below roughly 0 dB, as indicated by the reduction in slope. This slope reduction is due to the high error rate in making the bit decisions at low E_b/N_o. Heuristically, at low signal-to-noise ratio, sign(I) is proportional to I and sign(Q) is proportional to Q; hence, ϵ in (3-4) approaches zero, explaining the threshold behavior. If single pole filters are used instead of integrate-and-dump, the wider noise bandwidth causes the threshold to occur at a higher E_b/N_o, as will be seen below when discussing SQPSK.

3.2 DECISION FEEDBACK TRACKING FOR SQPSK

We now assume SQPSK in which the bit transitions on the two channels are displaced by half the bit interval. Thus, we write

$$s(t) = a(t) \cos \omega_o t + b(t - T/2) \sin \omega_o t \tag{3-6}$$

to show the displacement explicitly. Now

$$\begin{aligned} I &= a(t) \cos \theta + b(t - T/2) \sin \theta + x(t) \\ Q &= -a(t) \sin \theta + b(t - T/2) \cos \theta + y(t) \end{aligned} \tag{3-7}$$

and after smoothing by single-pole filters, the error voltage can be computed according to (3-4). A typical design practice for BPSK is to set the 3 dB filter bandwidth equal to T^{-1}. Note that the noise bandwidth of the filter is 1.57 T^{-1}, while that of integrate-and-dump is 0.5 T^{-1}, a ratio of 5 dB. Because of the thresholding behavior of (3-4), a narrower single pole low pass filter may be desirable with SQPSK.

*With Costas loop tracking of BPSK, false lock can exist for a phase change of π, and the offset frequency for a false lock is half the bit rate.

Figure 3-2 gives the results of a computer simulation with SQPSK and single-pole filters with a 3 -dB bandwidth of T^{-1}. Because the filters are not memoryless, the noise density was obtained from the variance measured for ϵ averaged over an integration interval of 20 bits duration. It is expected that QPSK and SQPSK would behave similarly with I and Q smoothed by single-pole filters.

Figure 3-2 displays a sharp threshold behavior, and the slope has been reduced almost to zero at $E_b/N_o = 0$ dB. Reducing the low pass filter bandwidth should improve this threshold point, at the cost of degrading the slope at high E_b/N_o*.

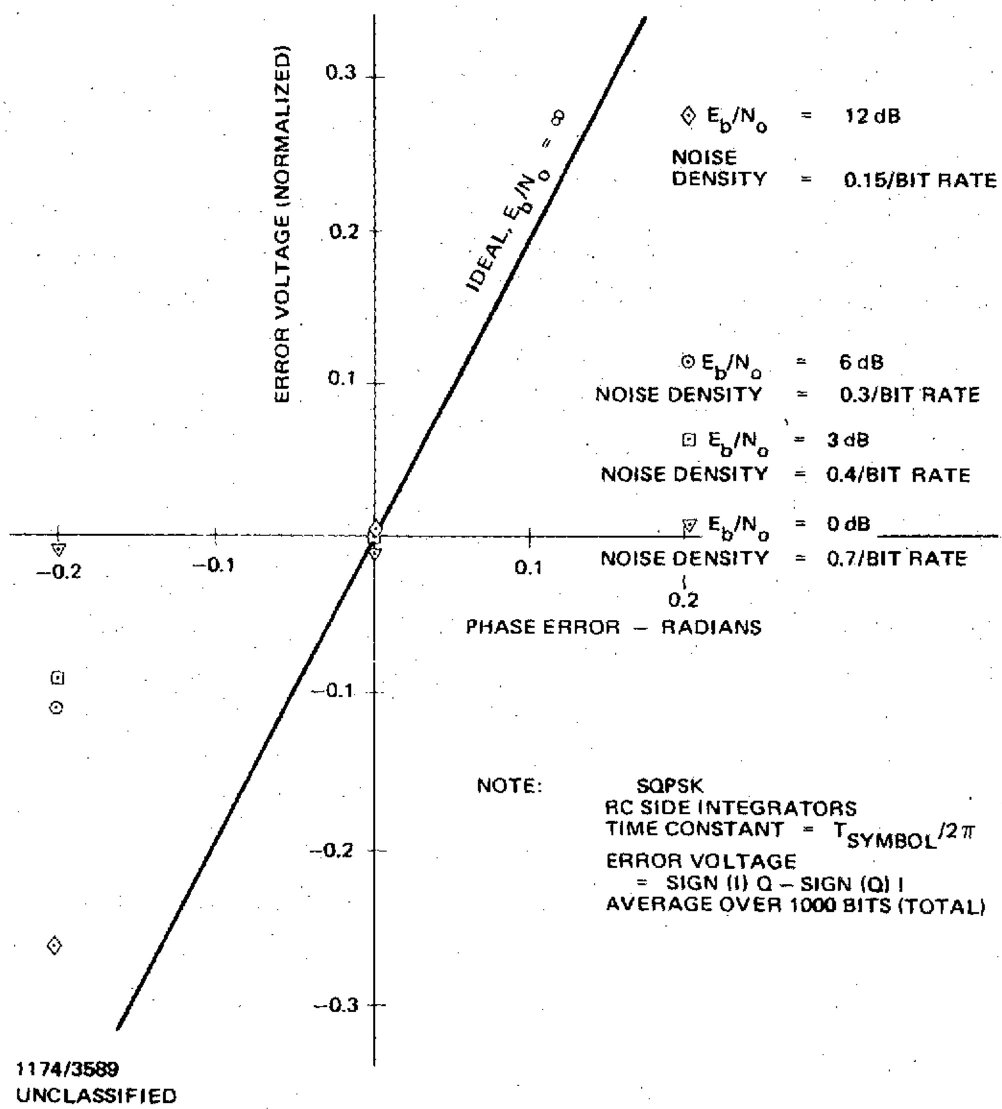


Figure 3-2. SQPSK Error Characteristic

*The transient due to a bit transition covers a larger portion of the bit duration, and this causes an effective reduction in signal amplitude.

Figure 3-3 shows simulation results with the single-pole filter 3 -dB bandwidth equal to 1/4T. As predicted, the slope is lower than in figure 3-2 when E_b/N_o = 12 dB (and much lower than the ideal slope), but substantially higher when E_b/N_o = 0 dB. (Several points are plotted for E_b/N_o = 0 dB to show the scatter of the 1000 bit average.)

In an attempt to apply integrate-and-dump filters to SQPSK, the difficulty is noted that the bit transitions occur at different times on I and Q. One possibility is to treat SQPSK as QPSK of twice the bit rate, but this obviously has a 3 dB disadvantage with respect to the degradation occuring at low E_b/N_o. A superior approach for integrate-and-dump filtering with SQPSK is now described. Referring to (3-3), the

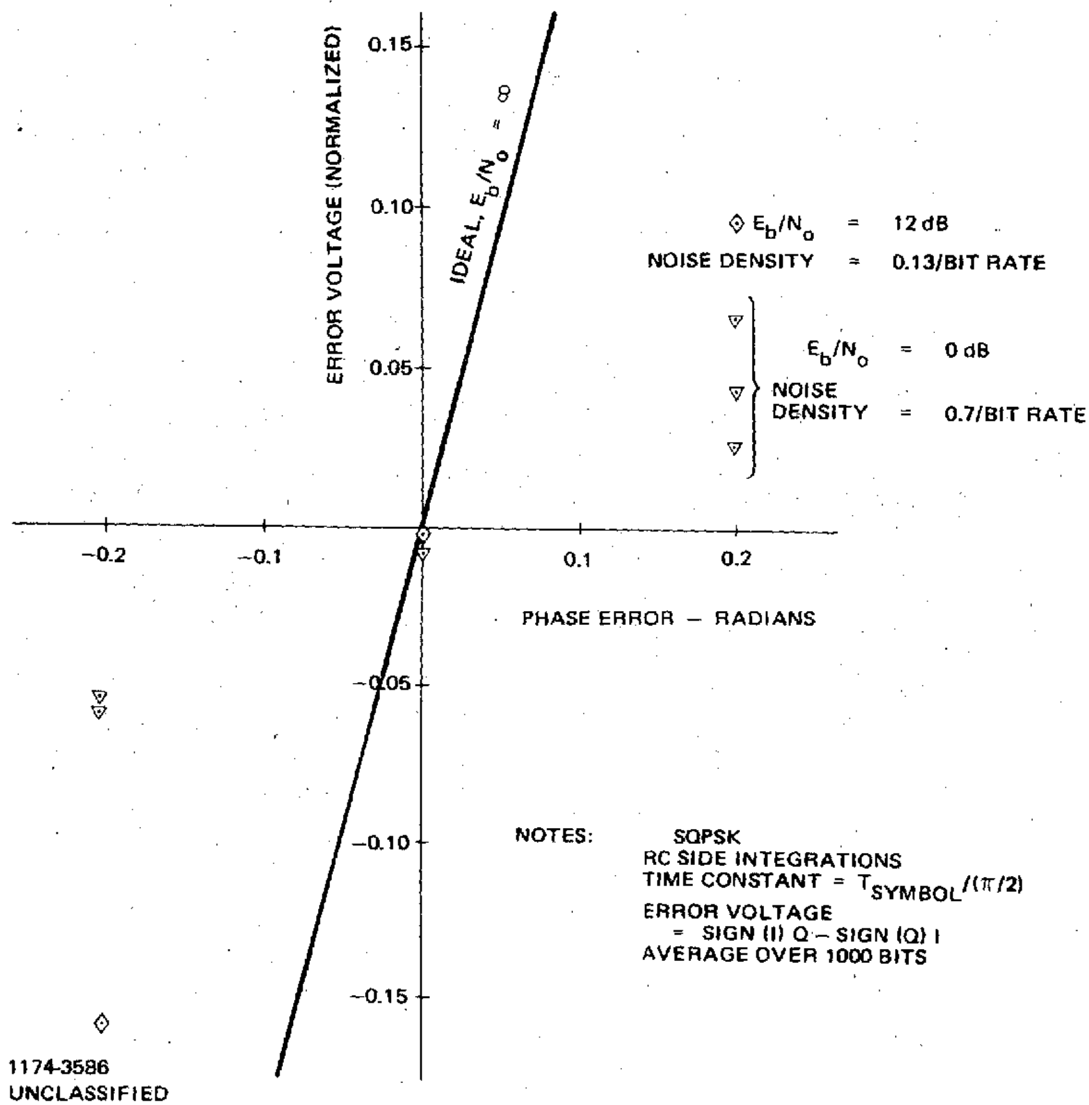


Figure 3-3. SQPSK Error Characteristic - Narrower Filters

observation is made that the estimates $\hat{a}$ and $\hat{b}$ are alternately available at the respective bit transition points. Thus, we write

$$\begin{aligned} \hat{a}\, Q &= \text{sign}(I)\, Q_I \\ -\hat{b}\, I &= -\text{sign}(Q)\, I_Q \end{aligned} \tag{3-8}$$

The notation is to convey that the integrate-and-dumps for I and Q_I are sampled at the I-bit transitions, while those for Q and I_Q are sampled at the Q-bit transitions.

Results of a computer simulation of SQPSK with integrate-and-dump as defined by (3-8) are given in figure 3-4. Compared to QPSK in figure 3-1, it appears that there is somewhat less degradation at low E_b/N_o.

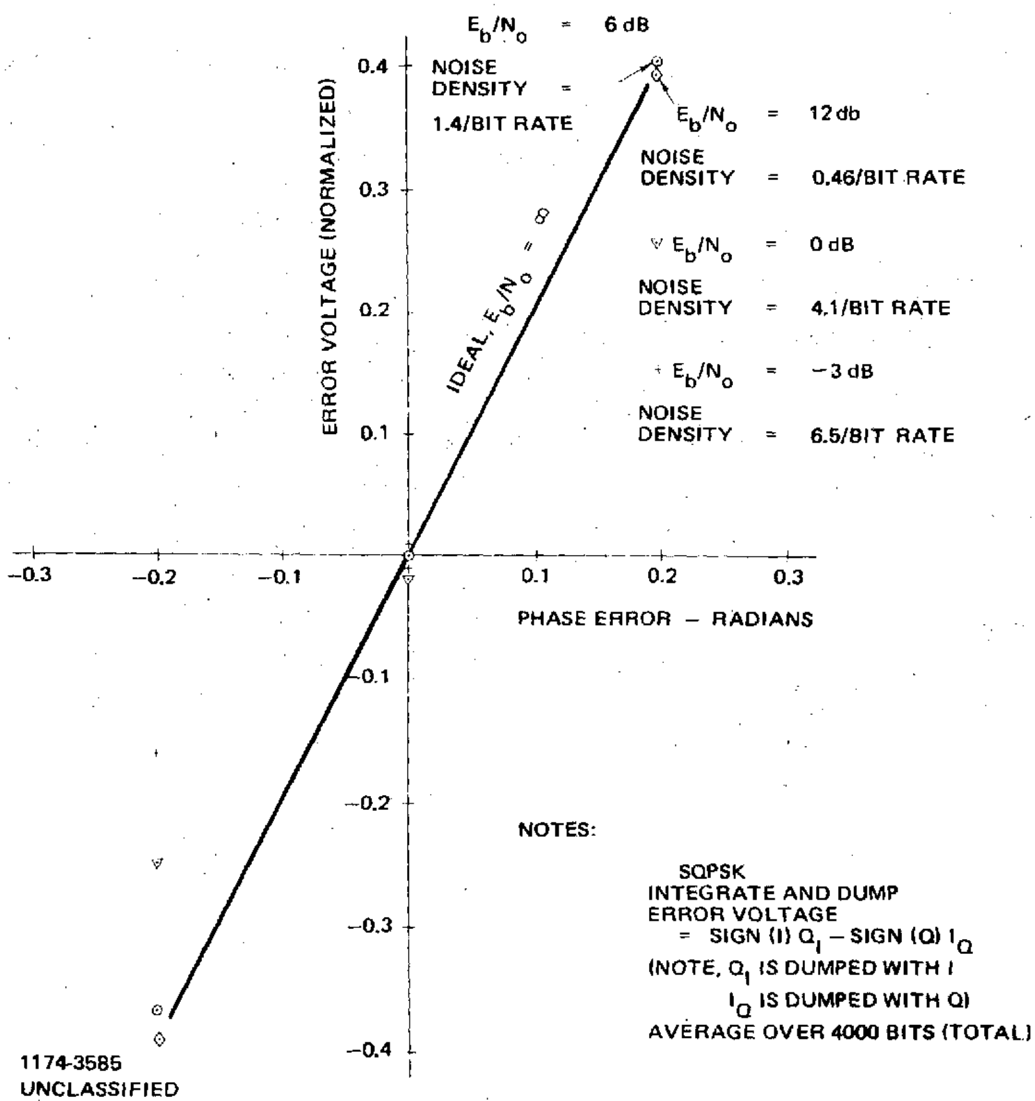


Figure 3-4. SQPSK Error Characteristic - Sampled Filters

3.3 DETECTION OF CARRIER SYNCHRONIZATION

To detect carrier synchronization in a decision-feedback type of tracking loop, an indicator is desired which becomes nonzero only for correct synchronization. The possibility of false lock exists with suppressed carrier tracking, with the phase reference changing by π radians per symbol for BPSK and $\pi/2$ radians per symbol (or $\pi/4$ per bit) for QPSK and SQPSK. Thus, a biphase Costas loop can false lock at a frequency offset of half the bit rate, and a quadriphase decision-feedback loop can lock at frequency offset of an eighth the total bit rate.

With a Costas loop for BPSK, an indicator of sync detection is given by $I^2 - Q^2$, provided I and Q have been smoothed by non-sampled filters*. With false lock at an offset $\Delta\omega$,

$$\begin{aligned} I &= \sqrt{2}\, A \cos(\Delta\omega t + \theta) \\ Q &= \sqrt{2}\, A \sin(\Delta\omega t + \theta) \end{aligned} \tag{3-9}$$

and I^2 and Q^2 both have the same time average**. This would not be true, however, for integrate-and-dump filtering of I and Q.

With QPSK and SQPSK, the sync indicator is generalized from that for biphase by defining

$$\text{sync indicator} = (I^2 + Q^2) - 2.74\,\epsilon^2 \tag{3-10}$$

where ϵ is given by (3-4). In the absence of a carrier, I and Q are independent Gaussian processes with variance σ^2, and

$$\begin{aligned} \text{Av}\left\{I^2 + Q^2\right\} &= 2\sigma^2 \\ \text{Av}\left\{\epsilon^2\right\} &= \text{Av}\left\{I^2 + Q^2 - 2\,|I|\,|Q|\right\} = .73\,\sigma^2 \end{aligned} \tag{3-11}$$

sync detector averages to 0. For a false lock with I and Q given by (3-4), we find

$$\begin{aligned} \text{Av}\left\{I^2 + Q^2\right\} &= 2\,A^2 \\ \text{Av}\left\{\epsilon^2\right\} &= \text{Av}\left\{2\,A^2 - 2\,A^2\,|\sin 2\Delta\omega t|\right\} = .73\,A^2 \end{aligned} \tag{3-12}$$

The sync indicator averages to 0, provided that I and Q are non-sampled filters. However, for a true lock $\epsilon \to 0$, and a positive sync indication results, on the average.

*Q is driven to zero and $|I|$ to a maximum by the tracking loop for BPSK.

**A narrow tracking loop bandwidth, compared to the offset frequency, is assumed, so that θ varies negligibly over one cycle of $\Delta\omega$.

3.4 IMPROVED THRESHOLD BY SIMULTANEOUS CARRIER PHASE AND BIT TRACKING

We see from the above that it is difficult to maintain carrier phase synchronization when demodulating staggered quadriphase at low E_b/N_o, such as for operation with error correction coding*. We take as the objective that $E_b/N_o \cong 0$ dB for the transmitted bits.

One approach to derive an error characteristic for phase tracking is to extract I and Q baseband channels by product detection and filter them with a single pole low pass filter prior to computing

$$\epsilon_\theta = \text{sign(I)Q} - \text{sign(Q)I} \tag{3-13}$$

This has a null at zero phase error. After carrier synchronization is achieved, the bit timing can be extracted by an independent bit synchronizer which locks to the bit transitions in the conventional way. In this approach, SQPSK is treated the same as QPSK, and there is an ambiguity of multiples of $\pi/2$ in carrier phase tracking. With SQPSK, a $\pi/2$ slip in carrier phase must be detected since the bit transitions on the two channels are displaced. At low E_b/N_o, this approach has the disadvantage of a degraded carrier tracking threshold due to use of single pole filters rather than I and D. The advantage is that carrier synchronization is not dependent on having bit synchronization.

To improve the threshold, I and D filters matched to the bit duration are needed. Now, the problem is how to derive bit timing simultaneously with achieving carrier phase tracking**. We discuss this simultaneous synchronization problem for SQPSK only.

To begin the discussion, assume that the SQPSK demodulator has good estimates of carrier phase and bit timing. Figure 3-5 shows the bit streams in the respective channels and I and D with timing matched to the I channel with an error τ. That is, the SQPSK signal is

$$s(t) = a(t) \cos \omega_o t + b(t) \sin \omega_o t \tag{3-14}$$

*In this discussion, E_b/N_o is the energy/noise density per transmitted bit; hence, redundancy of error correction is not included.

**If bit timing where known from another source, this problem would be alleviated.

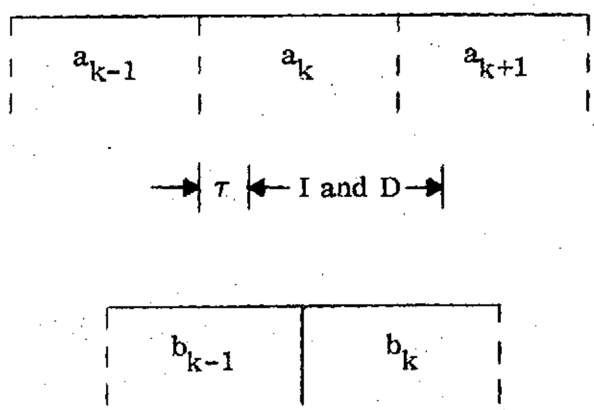


Figure 3-5. I and D Timing With τ Error

and with perfect phase synchronization, $I = a(t)$ and $Q = b(t)$. More generally, there is a phase error θ, and

$$\begin{aligned} I &= a(t) \cos \theta + b(t) \sin\theta \\ Q &= -a(t) \sin \theta + b(t) \cos \theta \end{aligned} \tag{3-15}$$

From figure 3-5, we see that for $0 < \tau < .5$ the I and D outputs are

$$\begin{aligned} I_D &= [a_k (1 - \tau) + a_{k+1} \tau]\cos \theta + [b_{k-1} (.5 - \tau) + b_k(.5 + \tau)] \sin\theta \\ Q_D &= -[a_k (1 - \tau) + a_{k+1} \tau]\sin \theta + [b_{k-1}(.5 - \tau) + b_k(.5 - \tau)]\cos\theta \end{aligned} \tag{3-16}$$

The expressions for $-.5 < \tau < 0$ can be similarly written.

For $\tau = 0$ and $\theta < \pi/4$, we find the error voltage to be

$$\epsilon_\theta\Big]_{\tau=0} = \text{sign}(I_D)Q_D = -a_k^2 \sin \theta + .5\, a_k[b_{k-1} + b_k]\cos\theta \tag{3-17}$$

and there is an average restoring force proportional to $\sin\theta$. A similar derivation with timing matched to the Q channel leads to

$$\epsilon_\theta\Big]_{\tau=0} = -\text{sign}(Q_D)I_D = -b_k^2 \sin\theta - .5\, b_k[a_k + a_{k+1}]\cos\theta \tag{3-18}$$

and the cross-product terms in (3-17) and (3-18) will cancel when summed over the sequence of bits. Thus, if the timing error is small, carrier phase synchronization can occur, since a restoring force exists proportional to $\sin\theta$.

Now, let us assume $\theta = 0$ and $0 < \tau < .5$. A conventional bit synchronizer derives the error voltage from an I and D which straddles the bit transitions, and corrects the sign by observing the polarity of the bit transition. Thus, with timing matched to the I channel, Q_D straddles the transitions in the Q channel, and

$$Q_D\Big]_{\theta=0} = b_{k-1}(.5 - \tau) + b_k(.5 + \tau) \tag{3-19}$$

The bit transition polarity is determined from the polarities of

$$\begin{aligned} Q_D\Big]_{\substack{\theta=0 \\ .5 \text{ bit early}}} &= b_{k-1}(1-\tau) + b_k \tau \\ Q_D\Big]_{\substack{\theta=0 \\ .5 \text{ bit late}}} &= b_k(1-\tau) + b_{k+1} \tau \end{aligned} \tag{3-20}$$

With timing matched to the Q channel, a similar computation derives an error from I_D.

3.5 BLOCK DIAGRAM OF SQPSK TRACKER

Figure 3-6 gives a block diagram of the SQPSK demodulator implemented in accordance with the above concept. There are two I and Ds on each channel, dumped, respectively, at the I transitions and the Q transitions. The I and D outputs are used, as required by (3-17) and (3-18), to compute the error voltage from the two channels for carrier phase tracking, and also to derive the error voltage for bit tracking. The bit transitions in the output data streams are detected by appropriate transition detection logic and used to correct the error voltage for bit tracking bit. Note that the second I and D in the I channel, which is dumped by Q timing, contains the bit tracking information for the I bit transitions, and vice versa in the Q channel.

It is assumed that the carrier frequency error is initially small so that a false lock does not happen. A feature of the loop is that a $\pi/2$ phase slip is not stable as long as the bit timing does not change very much during the slip[9].

3.6 COMPUTER SIMULATION RESULTS

The question now is whether pull-in from an arbitrary θ, τ initial state will take place. To understand how pull-in can take place, note that the maximum initial time error is 0.5, where unity denotes the symbol duration on either channel. The maximum initial phase error is $\pi/4$, since initial acquisition does not distinguish between the channels, and four stable tracking positions can result from the acquisition process. Alternatively, we can view the maximum initial time error as 0.25, with the maximum initial phase error being $\pi/2$.

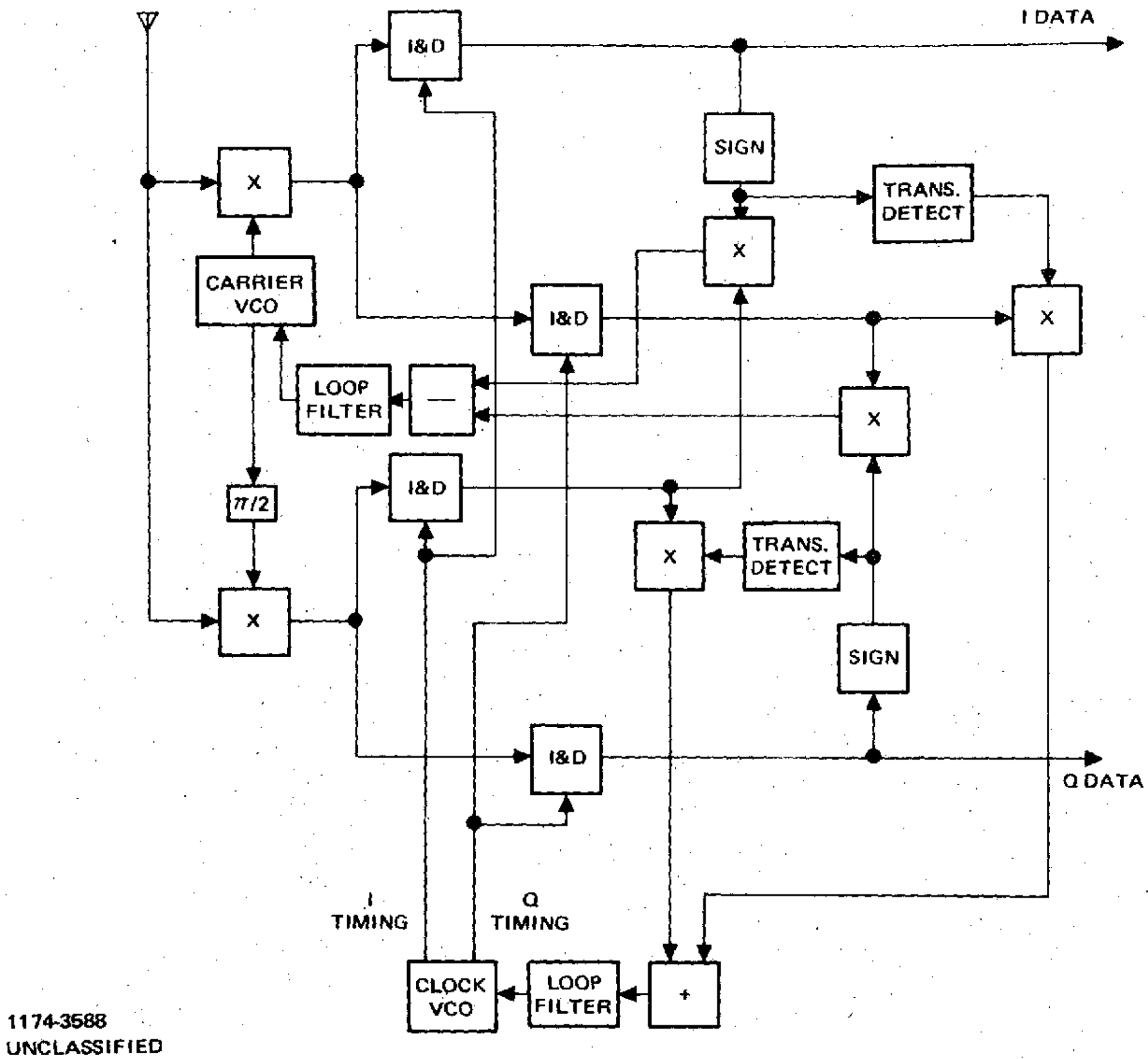


Figure 3-6. Block Diagram of SQPSK Demodulator

A computer simulation was performed by programming (3-16) to be valid over the range $-.5 < \tau < .5$. The error voltage for carrier tracking was derived from (3-17) and (3-18), and the error voltage for bit tracking was derived from I_D and Q_D by detecting apparent bit transitions. Noise of variance $\sqrt{N_o/2E_b}$ was introduced on each channel with the bit amplitudes normalized to unity. Each iteration moves time by 0.5; hence, the noise on each channel is represented by the sum of two Gaussian variables, the second of which is reused on the next iteration.

It should be noted that the simulation is valid only if θ varies slowly with respect to the bit rate, but this is typically the case for the tracking bandwidth of interest. A second-order carrier loop and a first-order bit loop were used in the simulation.

Figure 3-7 shows typical acquisition results when the time error and phase error are taken with respect to one of the two channels (arbitrarily designated). The acquisition limit is where pull-in usually went to the designated channel, rather than the other channel. The dotted line is drawn to best fit the observed pull-in behavior, while including the total uncertainty region (area equal to 0.5 bit multiplied by $\pi/4$ radians). The figure clearly shows that pull-in will always take place even at $E_b/N_o = 0$ dB, for the indicated loop bandwidths, $B_{phase} = .001$ and $B_{\cancel{phase}\ timing} = .001$. ~~an~~

With $B_{phase} = .01$, a phase slip was observed after a few bits at $E_b/N_o = 0$ dB; hence, this bandwidth is too wide. With $B_{timing} = .0001$ and $B_{phase} = .001$, an initial phase error of $\pi/2$ with $\tau = 0$ was reduced to zero while the timing error remained small, verifying that $\pi/2$ slips are unstable as long as the timing is disturbed only slightly. However, if B_{timing} is widened to .001, the timing was pulled to $\tau = .5$. Thus, the timing loop should be made narrow compared to the phase tracking loop, once acquisition has taken place, in order to eliminate $\pi/2$ phase slips which would cause loss of bit integrity.

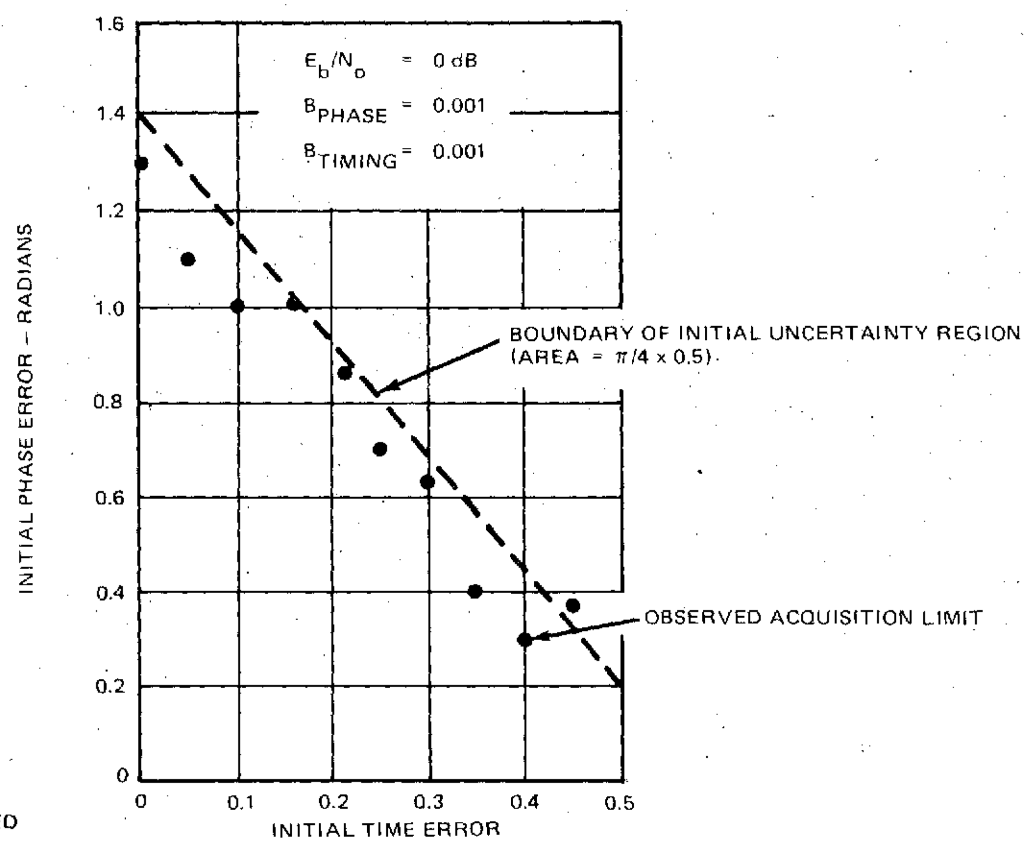


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Figure 3-7. Acquisition Limit to Designated Channel

3.7 CONCLUSIONS

A scheme for acquiring and maintaining SQPSK carrier phase and bit synchronization simultaneously has been described, and will function at $E_b/N_o \cong 0$ dB. Integrate and dump filters are utilized, and they perform the three functions of demodulating data, deriving an error voltage for carrier phase tracking, and deriving an error voltage for bit tracking. The double, interacting loops will pull in from an arbitrary θ, τ initial condition, provided that the frequency error is negligible. After acquisition, the bit tracking loop should be made tighter than the carrier phase tracking loop, so that $\pi/2$ slips are unstable.

SECTION IV

TDRS USER TRANSPONDER OPERATION WITH REMOTE GROUND STATIONS

From an operational standpoint, it is desirable to extend the capability of the S-band single access transponder to include operation with a remotely located ground station. Three separate approaches for implementing this capability were considered:

a. Addition of a TDRS Receiver-Transmitter.

b. Addition of a simplified TDRS R-T equipment at each remote ground station and a modified transponder with time shared search mechanisms for acquiring either a FH TDRS signal or a PN remote ground station signal.

c. Modification of a user transponder to detect, acquire and transpond STDN signals from an unmodified remote ground station.

In summary, the first two approaches were ruled out in favor of the third approach for (1) economic reasons and (2) the relative ease of modifying the user transponder design for compatibility with a STDN signal. The three approaches are briefly described in the following paragraphs.

4.1 REMOTE GROUND STATION EQUIPMENT REQUIREMENTS FOR COMPATIBILITY WITH TDRS USER TRANSPONDERS

Basic functions of a remote ground station receiver-transmitter compatible with a S/A user transponder are shown in figure 4-1. The receiver portion consists of a code tracking loop for demodulating the spread spectrum portion of the signal and a carrier tracking loop for demodulating the data from the signal. The transmitter portion consists of a frequency hop synthesizer for generating a FH preamble during initial acquisition and a PN sequence generator for modulating the transmit carrier during track.

Note that all frequency synthesis is referenced to a station frequency reference (5 MHz). Transmit carrier offsets are synthesized by a "digital VCO" which is a rate multiplier operating in conjunction with an incremental phase modulator. Receiver center frequency offsets (to compensate for the anticipated return link doppler) are generated from the same type of mechanism.

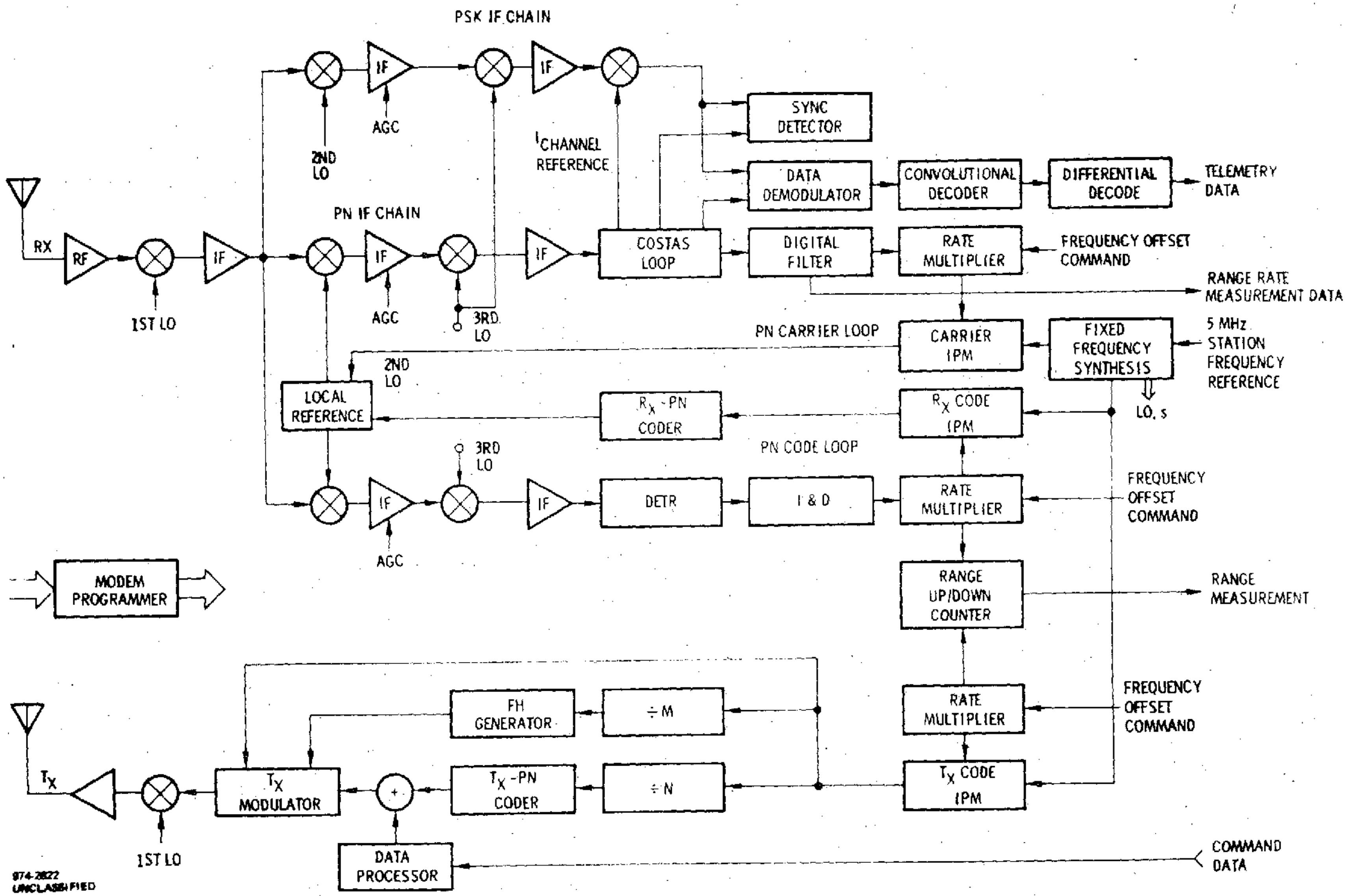

Figure 4-1. Ground Receiver-Transmitter Block Diagram

Range rate data are extracted from the digital filter of the carrier loop. Basically, this data is gathered by accumulating the carrier loop error signals (digital words at a fixed iteration rate) over a period of 1 or 10 seconds. The accumulated word is then scaled so that the digital-output word is in terms of meters/seconds. Range measurement is accomplished by counting the accumulated difference between receiver and transmitter code clock increment commands ($\pm 1/96$ chip steps). This accumulated difference is then scaled so that the digital word output is in terms of meters.

For users whose telemetry data is modulo-two added with a PN sequence, data is demodulated with a Costas demodulator after the pseudorandom sequence has been stripped from the signal. For users employing a clear mode for data along with a minimum power PN ranging signal, data is downconverted in a separate PSK IF chain and synchronously detected using a coherent I channel reference from the PN carrier tracking loop. For users employing a clear mode for data in the return link without a ranging signal, the PN local reference of the PN carrier loop is gated "off" and the data is extracted in the Costas demodulation. If quadriphase shift key modulation is employed in the return link, a separate QPSK demodulator must be used.

4.1.1 CONCLUSION

Technically, there is no reason why a remote ground station could not be modified to operate with a TDRS user transponder. The R-T equipment would be almost identical to the proposed TDRS ground R-T. There is little doubt that the signal interfaces of this additional equipment could be integrated into existing ground stations and that operational procedures could be established for acquiring and tracking ing a TDRS user satellite. User transponder tracking bandwidth could be increased, after acquisition, to track the increased doppler rate of change at perigee, since the received signal power tends to be maximum at this point. The directive ground antenna would also make interference from other TDRS signals negligible. The major disadvantage of this approach is the high cost of modifying existing ground stations and the maintenance of additional equipment.

4.2 TIME DIVISION MULTIPLEXING FOR REMOTE STATION OPERATION

The R-T equipment required for remote ground station operation (as described in section 4.1) could be simplified. The frequency hop preamble required for 20 second acquisition in the TDRS link could be eliminated from the waveform of a remote station link due to the 60 dB S/N advantage offered by STDN type stations (compared to TDRS satellite S/N ratios).

The impact of this approach would be a modified user transponder with a time shared search mechanism for acquiring a FH TDRS signal or a PN remote ground station signal. The transponder could be programmed to perform a normal FH search for 11 seconds (the probability of acquiring a TDRS signal in 11 seconds is 0.97) and then switch to a very fast PN search for 1 second (a 300 kilochip 1 second search rate would allow a complete scan of an 18 stage PN code in 1 second).

4.2.1 CONCLUSION

For this approach, no substantial implementation problems are anticipated. A 10% code slewing rate is not difficult. Doppler uncertainties fall well within the sync decision bandwidths. The performance impact on the TDRS forward link would be acquisition performance. Since the FH acquisition strategy allows for several passes through the FH preamble period of 11 seconds, an occasional 1 second signal loss for a remote ground station signal search would result in a slight decrease in the probability of FH signal detection. The major disadvantage to this approval is the high cost of modifying existing ground stations and maintenance of the additional equipment. In addition, the user transponder would have to be substantially modified to accommodate (1) the wide variation in signal strength, (2) wide range of PN search rates and (3) broad range of PN tracking bandwidths.

4.3 MODIFIED TRANSPONDER FOR USE WITH STDN GROUND STATIONS

A third approach to remote ground station operation is the functional augmentation of the user transponder to enable it to work with a ground station of the Spaceflight Tracking and Data Network (STDN). Since this network tracks satellites via the Goddard Range and Range Rate System, which is a tone ranging system, the PN TDRSS transponder must be modified for a compatible mode.

4.3.1 RECOGNITION OF STDN SIGNAL

The basic requirement is to incorporate a STDN recognition function which can command the transponder to stop searching for synchronization to a TDRSS forward link signal. The strong uplink signal of STDN will make this recognition function relatively straightforward, and also eliminate any concern for Doppler and Doppler rate of change (particularly the latter when a low altitude satellite goes by directly overhead).

Table 4-1 reviews the uplink propagation parameters from STDN to the satellite. It is clear that the power received from STDN is tremendous relative to that from TDRS, being about 60 dB greater for equal user antenna gains. Thus a signal from TDRS will negligibly affect S/N for receiving the STDN signal. Also, it is impossible to receive a signal from TDRS when the STDN signal is present.

The STDN uplink signal has a 500 kHz subcarrier continuously phase modulated on the carrier for S-band operation. The peak phase deviations are chosen so that the carrier component is reduced by about 4 dB from the unmodulated case. The 500 kHz subcarrier tone has a deviation of about one radian.[10] Thus a second-order phase lock loop in the transponder of relatively wide bandwidth, say 2 kHz to be within the 4 kHz sideband of the lowest tone, can track the carrier and demodulate the 500 kHz subcarrier with a loop signal-to-noise ratio of 94.4 dB-Hz - 33 dB - 4 dB $\cong$ 57 dB.

Table 4-1. Uplink From STDN

STDN Antenna Gain	43 dB
STDN Tx Power	40 dBw
RF Losses	-2 dB
Pointing Loss	0 dB
EIRP	81 dBw
Space Loss (4, 000 miles)	-176 dB
User Antenna Gain (Worst)	-10 dB
P_s Out of User Antenna	-105 dBw
T_s (Antenna Output)	824° K
KT_s	-199.4 dBw/Hz
P_s/KT_s	94.4 dB-Hz
Transponder Bandwidth B	4.5 MHz
P_s/KT_sB	28 dB

Without Doppler compensation on the transmit signal, the offset due to Doppler can be ± 55 kHz. The acquisition time for pull-in of a second order loop from an offset Δf is approximately[11]

$$t_{\text{pull-in}} = 4.1\, \Delta f^2/B_L{}^3 \tag{4-1}$$

For $\Delta f = 55$ kHz and $B_L = 2$ kHz, $t_{\text{pull-in}} = 1.6$ seconds.

After acquisition, the phase lock loop acts as a demodulator to extract the 500 kHz subcarrier so as to detect the presence of the STDN signal. At maximum Doppler, the offset on the subcarrier is ± 19 Hz; hence, the presence of the subcarrier can be detected in a filter of minimum bandwidth equal to 40 Hz. For a peak deviation of 1 radian, (rms of .7 radian) the S/N in this filter is 94.4 dB-Hz -16 dB -3 dB $\cong$ 75 dB, which enables exceedingly reliable detection of the STDN signal when it impinges on the user. The detection filter can be much wider than 40 Hz if desired for implementation convenience.

Finally, we note that the maximum carrier Doppler rate of change $\dot{f}$ which can be tracked by a second order loop is given by

$$2\pi\dot{f}/(1.89\, B_L)^2 \cong 1 \text{ radian} \tag{4-2}$$

For B_L = 2 kHz (4-2) yields $\dot{f} = 2.3 \times 10^6$ Hz/sec, which corresponds to $\ddot{R} = 3 \times 10^5$ m/sec^2. However, [12] maximum $\ddot{R}$ is 352 m/sec^2.

4.3.2 GENERAL DESCRIPTION OF USER EQUIPMENT MODIFICATIONS

The single access S-band user transponder presented in Chapter 3 of the Phase I - Final Report can be modified to handle STDN signals with the addition of two modules for detecting the presence of a STDN signal and subsequently transponding it in a coherent manner. The additional circuitry and its interface with a S/A user transponder is shown in figure 4-2.

Assuming that the center frequencies of a Goddard Range and Range Rate signal (GRARR) are compatible with the Rx and Tx center frequencies of the selected user transponder, common RF receiver, transmitter and L.O. synthesis sections can be used for both signals in the transponder. Separate phase locked loops (PLLs) for TDRS and STDN forward link signals are recommended because of the substantial difference in waveform structure. To receive STDN signals, the TDRS Costas loop would have to be converted to a PLL, tracking and data bandwidths would have to be extended and loop gain would have to be corrected for proper operation. Finally, by using separate PLLs, reliable performance is enhanced and a STDN mode can also be added to an "off-the-shelf" TDRS user transponder with minimal interface impact.

Incoming STDN signals are routed to a STDN signal detection module from a signal splitter at the output of the 1st IF (prior to the FH/PN correlator). In the detector module, STDN signals are amplified by a 2nd IF (a 3rd IF is not required due to the high signal level with respect to TDRS signals). A PLL is used to acquire and track the GRARR carrier and, subsequently, provide a coherent frequency reference for retransmitting the signal tone package. The presence of a STDN signal is indicated by detection of the 500 kHz subcarrier which is demodulated along with the other tones by the PLL.

In the STDN modulator module, the baseband tones modulate a 1.2 MHz subcarrier and this composite signal in turn linearly modulates a Tx carrier which is coherently derived from the STDN PLL. To provide a STDN mode of operation, the controller module must: (1) disable the PN or PSK modulator, (2) apply the STDN modulator and, (3) select a frequency reference from the STDN PLL instead of the TDRS PLL for RF synthesis. Accidental selection of a STDN mode is not likely because a STDN mode command requires both a locked PLL and detection of a 500 kHz subcarrier. In addition an established TDRS return link mode will override a STDN mode command.

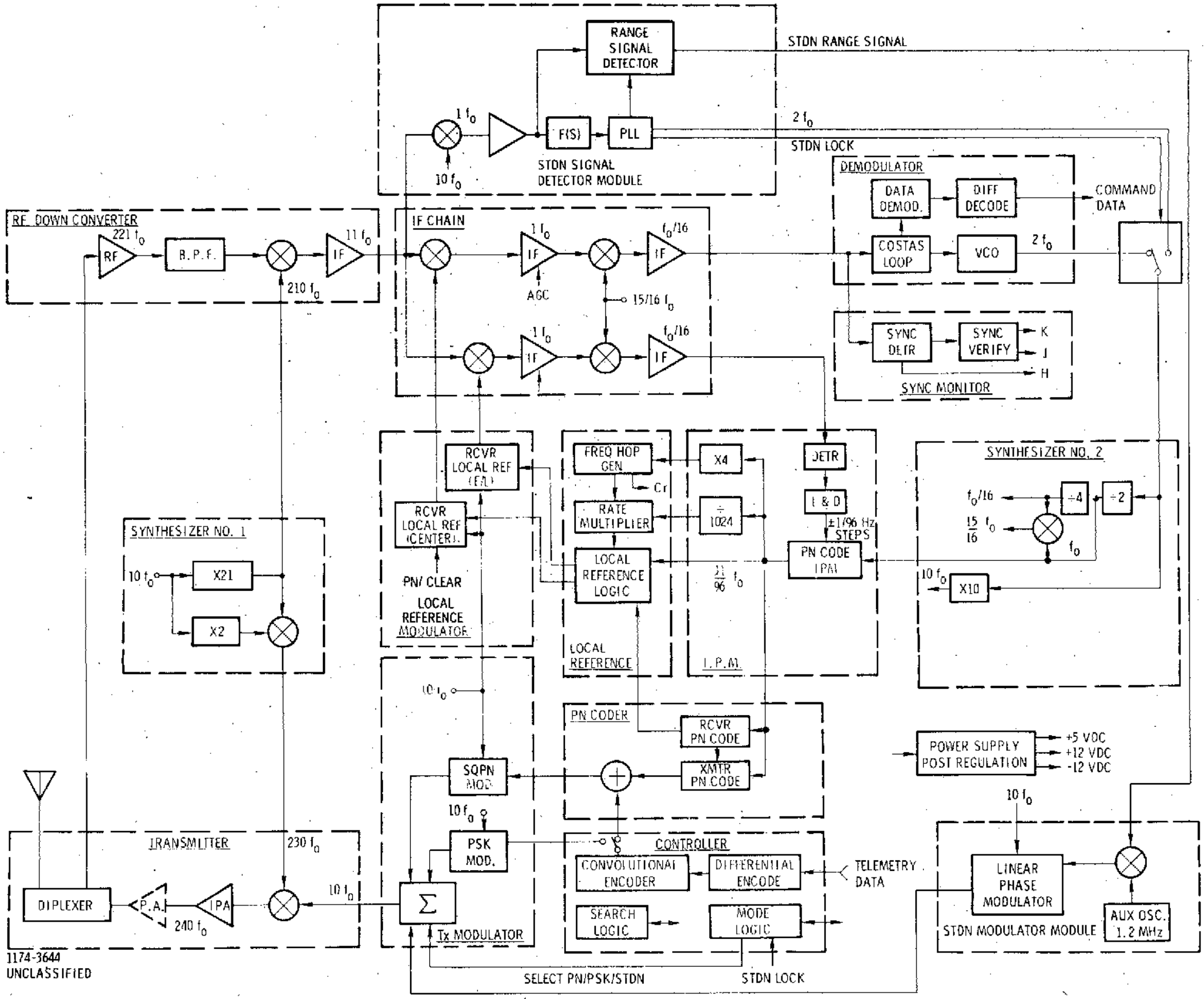

Figure 4-2. Multiple Access User Transponder Modified for STDN Signals

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Details of the STDN signal detection module are presented in figure 4-3. A crystal filter is used to isolate the signal carrier. This carrier is amplitude limited prior to phase detection in the PLL. Range tones are preconditioned in a wideband AGC amplifier prior to baseband detection. After detection, they are reamplified and filtered. A separate Q channel in the PPL is used to detect the 500 kHz subcarrier. This subcarrier is subsequently bandpass filtered (BW ≈ 40 to 100Hz) and used to trigger a threshold device which provides a very good indication of the presence of a STDN signal. Detection of the 500 kHz subcarrier was used for STDN signal detection, rather than PLL lock, to prevent false detection from spurious carrier signals and unintentional jammers.

4.3.3 SIZE, WEIGHT, AND POWER

The single access S-band transponder, modified to handle STDN signals from remote ground stations, contains two additional modules:

a. STDN Signal Detection Module
b. STDN Modulator Module

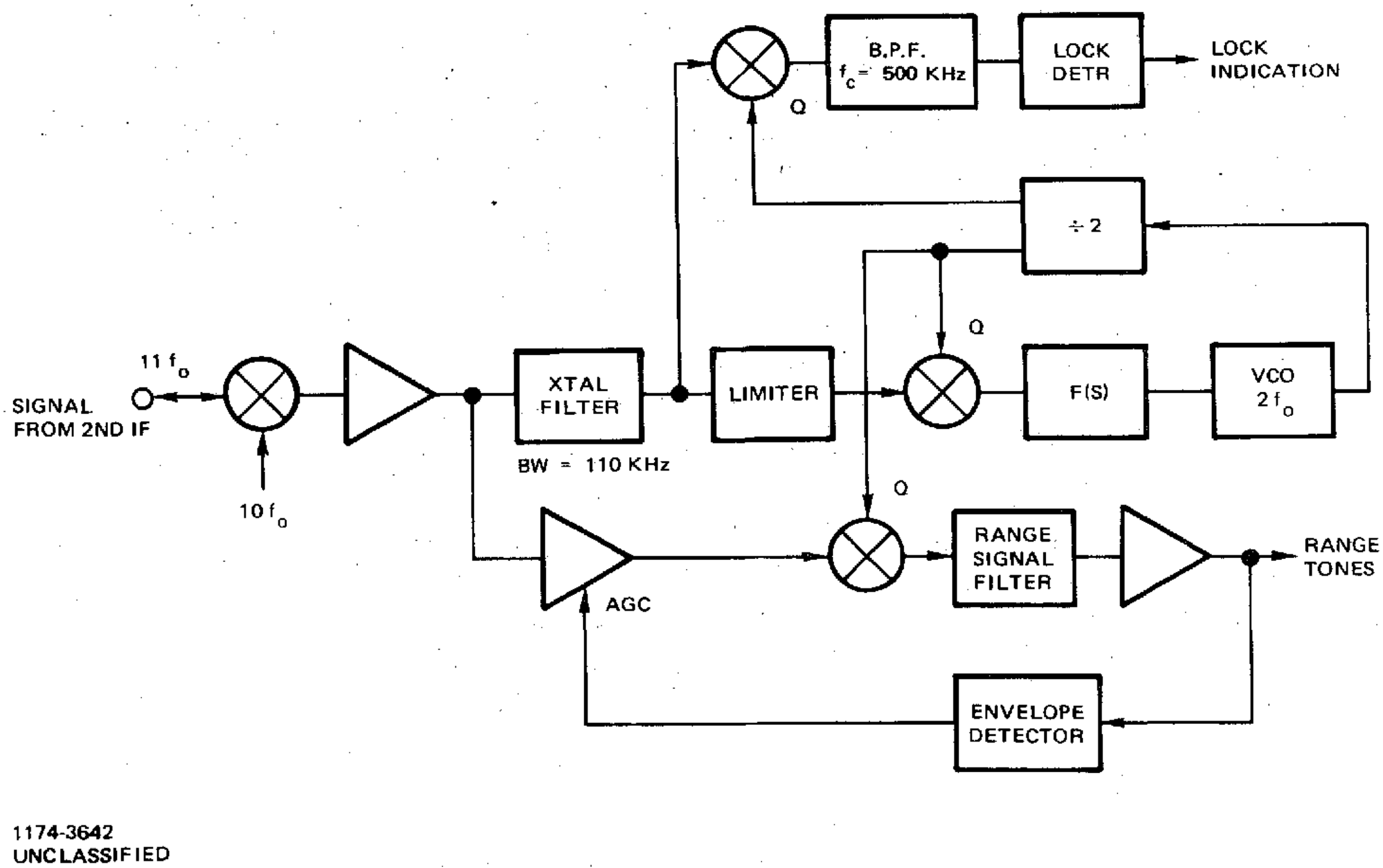

Figure 4-3. STDN Signal Detector Module

The power, weight and size requirements for this transponder are summarized in table 4-2.

Table 4-2. Modified S-band Single Access Transponder
Size, Weight and Power

Modules	Power (watts)	Weight (oz.)	(Size (in^3)
Narrowband Transponder			
RF Down Converter	1	16	24
IF Chain	1.5	12	24
Synthesizer No. 1	2	10	24
Synthesizer No. 2	1.5	10	24
Demodulator	2	10	24
Sync Monitor	.5	6	12
Controller	1.5	6	12
Modulator	1	16	24
Transmitter (100 mW)	2	24	36
Post Regulator	3	24	36
Chassis	-	72	78
Subtotal	16	206	318
Pseudonoise Assemblies			
IPM	2	10	24
PN Coder	2	6	12
Local Reference	1.5	6	12
Local Reference Modulator	1.5	12	24
Subtotal	7	34	72
STDN Assemblies			
Detector	1	10	18
Modulator	1	6	12
Subtotal	2	16	30
TOTAL	25	256	420

In summary, the Single Access S-band Transponder will require approximately 25 watts of power (assuming a 100 mW transmitter), weigh on the order of 16 lbs., and occupy 420 cubic inches in a 5 in. x 6 in. x 14 in. configuration.

4.3.3.1 Size

The length and width dimensions of each assembly are 4.5 in. x 6 in. with a useful circuit area of 24 square inches. The height dimension varies from 0.5 in. for logic boards to 1 in. for analog and RF assemblies to 1.5 in. for the transmitter and power supply assemblies.

The physical configuration of the transponder is envisioned as a tray of fixed-mount assemblies supported by a pair of rigid walls. Estimated dimensions for the transponder are 5 in. x 6 in. x 14 in. with 318 cubic inches of the 420 cubic inches apportioned to the narrowband functions , 72 cubic inches to the PN functions and 30 cubic inches to the STDN functions.

4.3.3.2 Weight

The estimated weight for the various assemblies is itemized in table 4-2. In general, logic assemblies are lighter than analog assemblies and much lighter than "canned" RF assemblies. The heaviest item in the transponder is the chassis at 4.5 lbs. The second heaviest items are the power supply and transmitter modules at 1.5 lbs each. In summary, the transponder weight is estimated at 16 lbs.

4.3.3.3 Power

A power estimate for the transponder is 25 watts. The largest variation in this value for potential users lies in the transmitter, since the EIRP requirements vary over a range of 30 dB. For purposes of an estimate, a 100 mW transmitter requirement was assumed.

Variation in power supply requirements from satellite to satellite also poses somewhat of a problem in estimating power requirements for the transponder. For this estimate, power supply regulation was assumed for all receiver and PN functions with an average operating efficiency of 80%.

4.3.4 MODULE SPECIFICATIONS

Detector Module

IFs	
Bandwidth	
1st	10 MHz
2nd	5 MHz
3rd	5 MHz
PPL	
Type	2nd order loop
B_L	2 kHz nominal
Predetection Bandwidth	
Offset at Tx	100 kHz nominal
No offset at Tx	17 kHz
Subcarrier Detector	
Bandwidth	100 Hz nominal
P_d	>.99

Modulator Module

Type	Linear
Subcarrier Frequency	1.2 MHz nominal
Input	
Sensitivity	1 volt/radian
Impedance	1000 ohm
Output	
Level	-10 dBm RMS
Impedance	50 ohm resistive

4.3.5 OPERATIONAL PROCEDURES

Assuming that an S-band S/A user transponder has been modified to operate with TDRS or STDN signals, the following operational procedures should be followed to establish a STDN transpond mode of operation from a remote STDN ground station:

a. Select the appropriate RF frequencies compatible with the forward and return links of the desired user transponder.

b. Offset the transmit frequency to accommodate the Doppler offset and point the ground transmitter antenna.

c. Begin a STDN uplink transmission with a signal which has a 500 kHz subcarrier continuously phase modulated on the carrier (other tones may also be present).

d. User transponder acquires STDN signal within 2 seconds (if a normal TDRS forward link is established, it will be disrupted due to the relative signal strength of the STDN signal).

e. User transponder begins a STDN transpond mode. (If a normal TDRS return link has been previously established, a STDN transpond mode will be disallowed until the TDRS return link transmission is completed.)

f. At the STDN ground station, acquire and track the return STDN signal from the user transponder.

g. The user transponder will automatically revert back to a normal TDRS mode of operation when the STDN forward link signal is discontinued or disrupted.

SECTION V

TABULATION OF CODES FOR MULTIPLE ACCESS AND S-BAND SINGLE ACCESS USERS

This section provides a listing of frequency hop and PN codes for the multiple access and S-band single access services. Parameters and generation techniques are as recommended in the Phase I - Final Report.

All codes have a period which is a power of 2. Since a maximal linear code sequence generator has a period of the form $K = 2^k-1$, an extra chip must be inserted when this generation technique is utilized. This can be done conveniently by recognizing the unique string of k-1 0's in the period and inserting an extra 0 at this point in the period. The code generator is simply inhibited from shifting for one clock when the string of k-1 0's occurs.

5.1 FH CODE FOR MULTIPLE ACCESS FORWARD LINK

The frequency hopping code for the forward link of the multiple access source has a period of 256 chips and a hopping rate of 3.006 kHz. A primitive root generation technique is utilized, so that the sequence of frequencies (represented by numbers in the range from 0 to 255) for the nth code is computed according to

$$f_i^{(n)} = [(3^i \bmod 257) - (27^i \bmod 257)] \bmod 256 \qquad (1)$$

The mod p function means to add or subtract p until the result is in the range 0 to p-1.

Table 5-1 is a computer printout of 30 different codes generated according to (1). The sequence of frequency numbers is read horizontally across the table.

5.2 PN CODES FOR MULTIPLE ACCESS

The forward and return links of the multiple access service use staggered quadriphase PN, generated by two different maximal binary codes. The code pair biphase modulates carriers phase shifted by 90°, and the clocks are displaced by a half chip. The period is $2^{18} = 262143$ chips, and the chip rate is 3.078 MHz.

Table 5-1. Printout of FH Codes for Multiple Access

FREQUENCY HOP CODE NUMBER= 1

44	114	61	254	58	103	190	213	200	4	30	62	200	236	81	195	60	140	110	213	255	111	100	96	176	104	91	216	166	26	103	212
103	228	69	225	162	204	143	21	98	210	190	203	158	38	164	42	84	94	76	12	34	48	227	124	187	140	53	191	181	187	150	66
110	10	127	203	91	164	149	100	182	216	167	55	150	250	182	35	185	174	206	2	254	34	183	115	110	182	97	165	206	216	23	74
129	137	35	151	183	116	251	138	205	60	49	156	145	143	203	110	140	173	18	148	155	114	147	234	26	248	223	171	122	38	171	26
212	142	195	2	198	153	66	43	56	252	226	194	56	20	175	61	196	116	146	43	1	145	156	160	80	152	165	40	90	230	153	44
153	28	187	31	94	52	113	235	158	46	66	53	98	218	92	214	172	162	180	244	222	208	29	132	69	116	203	65	75	69	106	190
146	246	129	53	165	92	107	156	74	40	89	201	106	6	74	221	71	82	50	254	2	222	73	141	146	74	159	91	50	40	233	182
127	119	221	105	73	140	5	118	51	196	207	100	111	113	53	146	116	83	238	108	101	142	109	22	230	8	33	85	134	218	85	230

FREQUENCY HOP CODE NUMBER= 2

108	43	200	152	131	18	208	185	215	152	170	14	189	197	29	76	188	254	131	10	144	198	136	40	206	143	116	120	144	203	255	231
102	201	111	74	197	122	214	166	157	32	240	12	114	135	212	80	82	40	160	223	104	139	114	157	50	40	152	65	93	125	247	142
106	158	39	113	230	90	180	165	165	15	110	58	231	125	121	186	177	215	29	79	20	141	245	246	76	36	239	171	112	221	157	121
113	219	197	49	228	74	118	145	137	23	78	168	212	153	216	204	108	80	77	199	246	29	138	248	146	174	25	195	0	57	197	214
148	213	56	104	125	238	48	71	41	104	86	242	67	59	227	180	68	2	125	246	112	58	120	216	50	113	140	136	112	53	1	25
154	55	145	182	59	134	42	90	99	224	16	244	142	121	44	176	174	216	96	33	152	117	142	99	206	216	104	191	163	131	9	114
150	98	217	143	26	166	76	91	91	241	146	198	25	131	135	70	79	41	227	177	236	115	11	10	180	220	17	85	144	35	99	135
143	37	64	207	28	182	138	111	119	233	178	88	44	103	40	52	148	176	179	57	10	227	118	8	110	82	231	61	0	199	59	42

FREQUENCY HOP CODE NUMBER= 3

37	182	98	225	46	36	180	200	107	36	122	3	150	145	166	204	46	19	184	155	231	234	80	70	245	168	20	98	65	99	18	230
75	243	216	109	115	193	103	225	235	82	49	224	211	183	250	78	28	124	115	37	195	26	147	20	206	139	26	233	31	222	67	138
254	70	205	252	156	121	245	148	220	214	113	139	106	64	16	178	218	38	106	101	127	203	120	212	186	178	245	77	117	99	204	105
195	120	90	94	186	197	125	77	100	52	90	235	222	166	54	172	15	139	128	34	161	20	152	112	72	232	49	73	19	83	129	150
219	74	158	31	210	220	76	56	149	220	134	253	106	111	90	52	210	237	72	101	25	22	176	186	11	88	236	158	191	157	238	26
181	13	40	147	141	63	153	31	21	174	207	32	45	73	6	178	228	132	141	219	61	230	109	236	50	117	230	23	225	34	189	118
2	186	51	4	100	135	11	108	36	42	143	117	150	192	240	78	38	218	150	155	129	53	136	44	70	78	11	179	139	157	52	151
61	136	166	162	70	59	131	179	156	204	166	21	34	90	202	84	241	117	128	222	95	236	104	144	184	24	207	183	237	173	127	106

FREQUENCY HOP CODE NUMBER= 4

176	80	171	140	64	8	195	92	247	244	111	220	98	26	38	62	67	72	73	242	11	178	110	109	14	72	254	19	217	118	17	203
117	92	251	27	186	82	162	47	29	147	5	65	3	221	248	24	112	79	185	128	82	59	10	176	49	13	194	171	128	42	63	30
166	236	88	178	187	186	228	203	163	217	194	14	45	215	8	219	41	115	128	208	189	78	86	66	72	184	151	82	251	146	188	187
96	18	135	52	53	204	57	40	129	64	157	245	235	4	22	79	74	190	219	205	152	34	16	38	130	0	183	92	45	15	65	221
80	176	85	116	192	248	61	164	9	12	145	36	158	230	218	194	189	184	183	14	245	78	146	147	242	184	2	237	39	138	239	53
139	164	5	229	70	174	94	209	227	109	251	191	253	35	8	232	144	177	71	128	174	197	246	80	207	243	62	85	128	214	193	226
90	20	168	78	69	70	28	53	93	39	62	242	211	41	248	37	215	141	128	48	67	178	170	190	184	72	105	174	5	110	68	69
160	238	121	204	203	52	199	216	127	192	99	11	21	252	234	177	182	66	37	51	104	222	240	218	126	0	73	164	211	241	191	35

FREQUENCY HOP CODE NUMBER= 5

74	153	86	158	36	23	87	232	199	233	72	168	235	154	152	83	120	217	160	22	211	208	149	134	174	50	175	171	236	117	246	245
222	127	169	98	75	141	240	97	94	103	102	113	41	219	194	108	67	149	20	15	115	178	166	19	179	181	132	12	204	38	211	198
76	119	14	209	252	169	27	146	166	42	69	209	196	207	49	42	118	137	235	14	64	44	196	208	78	90	156	216	42	130	14	88
250	63	93	175	60	136	20	69	141	131	167	2	73	228	185	138	125	25	134	196	166	154	198	96	154	134	202	118	233	207	136	82
192	103	170	98	220	233	169	24	57	23	184	88	21	102	104	173	136	39	96	234	45	48	107	122	82	206	81	85	20	139	10	11
34	129	87	158	181	115	16	159	162	153	154	143	215	37	62	148	189	107	236	241	141	78	90	237	77	75	124	244	52	218	45	58
180	137	247	47	4	87	229	110	90	214	187	47	60	49	207	214	138	119	21	242	192	212	60	48	178	166	100	40	214	126	242	168
6	193	163	81	196	120	236	187	115	125	89	254	183	28	71	118	131	231	122	60	90	102	58	160	102	122	54	138	23	49	120	174

FREQUENCY HOP CODE NUMBER= 6

147	68	104	130	51	171	227	184	188	194	20	49	107	12	173	136	9	48	196	222	241	247	174	38	152	227	71	190	235	90	32	94
1	45	240	243	134	219	34	162	50	200	150	151	39	165	22	63	137	240	163	48	234	78	9	149	91	119	229	88	200	186	123	108
215	45	45	18	235	224	226	149	247	173	8	104	188	248	128	119	140	244	41	145	30	154	82	214	240	95	34	7	26	212	171	242
39	21	216	182	248	99	49	81	208	141	180	96	41	135	244	189	216	196	125	210	30	80	0	120	32	153	228	50	169	22	253	184
109	188	152	126	205	85	29	72	68	62	236	207	149	244	83	120	247	208	60	34	15	9	82	218	104	29	185	66	21	166	224	162
255	211	16	13	122	37	222	94	206	56	106	105	217	91	234	193	119	16	93	208	22	178	247	107	165	137	27	168	56	70	133	148
41	211	211	238	21	32	30	107	9	83	248	152	68	8	128	137	116	12	215	111	226	102	174	42	16	161	222	249	230	44	85	14
217	235	40	74	8	157	207	175	48	115	76	160	215	121	12	67	40	60	131	46	226	176	0	136	224	103	28	206	87	234	3	72

Table 5-1. Printout of FH Codes for Multiple Access (Continued)

FREQUENCY HOP CODE NUMBER= 7

62	86	76	145	199	55	179	173	149	142	157	177	221	33	226	25	96	84	140	252	24	16	78	16	73	123	90	189	208	132	137	129
175	116	129	46	212	13	99	118	147	248	138	149	241	249	233	133	228	127	196	167	134	177	139	61	29	216	49	84	92	98	33	247
141	76	110	1	34	167	229	230	122	112	159	96	229	71	205	141	247	50	172	111	140	40	88	120	245	229	81	247	108	113	69	31
253	144	223	114	211	128	61	148	218	154	18	64	204	194	39	24	131	187	139	74	212	138	24	254	51	179	160	242	240	139	99	111
194	170	180	111	57	201	77	83	107	114	99	79	35	223	30	231	160	172	116	4	232	240	178	240	183	133	166	67	48	124	119	127
81	140	127	210	44	243	157	138	109	8	68	107	15	7	23	123	28	129	60	89	122	79	117	195	227	40	207	172	164	158	223	9
115	180	146	255	222	89	27	26	134	144	97	160	27	185	51	115	9	206	84	145	116	216	168	136	11	27	175	9	148	143	187	225
3	112	33	142	45	128	195	108	38	102	238	192	52	62	217	232	125	69	117	182	44	118	232	2	205	77	96	14	16	117	157	145

FREQUENCY HOP CODE NUMBER= 8

80	58	91	37	83	7	168	134	97	23	29	35	242	86	115	112	132	28	170	35	49	176	56	193	225	142	89	162	250	237	172	47
246	5	188	124	6	78	55	215	195	30	186	95	69	204	47	224	115	160	59	67	233	51	51	255	126	36	45	232	4	8	172	173
172	141	93	56	233	170	54	105	61	7	151	137	52	148	227	248	53	181	138	221	26	46	250	125	123	20	65	73	9	11	114	245
120	151	155	77	240	140	128	158	231	248	242	227	7	245	130	195	122	201	3	0	14	162	158	17	77	111	96	57	101	241	26	196
176	198	165	219	173	249	88	122	159	233	227	221	14	170	141	144	124	228	86	221	207	80	200	63	31	114	167	94	6	19	84	209
10	251	68	132	250	178	201	41	61	226	70	161	187	52	209	32	141	96	197	189	23	205	205	1	130	220	211	24	252	248	84	83
84	115	163	200	23	86	202	151	195	249	105	119	204	108	29	8	203	75	118	35	230	210	6	131	133	236	191	183	247	245	142	11
136	105	101	179	16	116	128	98	25	8	14	29	249	11	126	61	134	55	253	0	242	94	98	239	179	145	160	199	155	15	230	60

FREQUENCY HOP CODE NUMBER= 9

52	73	239	177	35	252	129	82	234	151	143	56	39	231	202	148	76	58	209	60	209	154	233	89	244	141	62	204	99	16	90	118
135	64	10	174	71	34	152	7	233	28	132	179	24	18	138	111	148	23	215	166	107	219	245	96	202	32	193	144	170	147	98	204
237	124	148	255	236	251	185	44	212	255	192	216	129	170	78	54	184	147	248	107	32	208	255	3	170	4	147	230	163	56	72	112
127	83	118	106	252	207	138	171	69	216	149	30	58	80	45	186	136	65	185	58	38	40	177	43	9	47	167	174	203	168	111	178
204	183	17	79	221	4	127	174	22	105	113	200	217	25	54	108	180	198	47	196	47	102	23	167	12	115	194	52	157	240	166	138
171	192	246	82	185	222	104	249	23	228	124	77	232	238	118	145	108	233	41	90	149	37	11	160	54	224	63	112	86	109	158	52
19	132	108	1	20	5	71	212	44	1	64	40	127	86	178	202	72	109	8	149	224	48	1	253	86	252	109	26	93	200	184	144
129	173	138	150	4	49	118	85	187	40	107	226	198	176	211	70	120	191	71	198	218	216	79	213	247	209	89	82	53	88	145	78

FREQUENCY HOP CODE NUMBER= 10

67	221	123	129	24	213	77	219	106	9	164	109	184	62	238	92	106	97	234	220	187	75	129	108	243	114	104	53	134	190	161	7
194	142	60	239	27	131	200	45	231	230	216	134	94	109	25	144	11	179	58	40	19	157	86	172	198	180	105	54	53	73	129	13
220	179	91	2	61	126	124	195	204	40	15	37	151	21	140	185	150	1	134	113	194	213	133	50	154	86	48	128	208	14	195	119
59	46	147	118	63	217	151	9	37	123	208	81	149	251	36	200	0	247	243	82	172	59	203	231	201	118	28	20	130	253	93	206
189	35	133	127	232	43	179	37	150	247	92	147	72	194	18	164	150	159	22	36	69	181	127	148	13	142	152	203	122	66	95	249
62	114	196	17	229	125	56	211	25	26	40	122	162	147	231	112	245	77	198	216	237	99	170	84	58	76	151	202	203	183	127	243
36	77	165	254	195	130	132	61	52	216	241	219	105	235	116	71	106	255	122	143	62	43	123	206	102	170	208	128	48	242	61	137
197	210	109	138	193	39	105	247	219	133	48	175	107	5	220	56	0	9	13	174	84	197	53	25	55	138	228	236	126	3	163	50

FREQUENCY HOP CODE NUMBER= 11

215	105	75	118	241	161	214	91	220	30	217	254	15	98	182	122	145	122	138	198	108	227	148	107	216	156	209	88	52	5	50	66
16	192	125	195	124	179	238	43	177	58	171	204	185	252	58	7	167	22	188	208	213	254	162	168	90	92	15	193	235	104	194	252
19	122	94	83	192	65	19	187	245	119	92	59	2	83	15	151	4	143	140	19	199	91	180	34	236	243	202	173	166	137	202	51
22	75	159	185	73	230	245	233	200	182	3	172	64	242	50	64	182	49	11	216	191	85	135	167	16	235	130	203	215	235	121	191
41	151	181	138	15	95	42	165	36	226	39	2	241	158	74	134	111	134	118	58	148	29	108	149	40	100	47	168	204	251	206	190
240	64	131	61	132	77	18	213	79	198	85	52	71	4	198	249	89	234	68	48	43	2	94	88	166	164	241	63	21	152	62	4
237	134	162	173	64	191	237	69	11	137	164	197	254	173	241	105	252	113	116	237	57	165	76	222	20	13	54	83	90	119	54	205
234	181	97	71	183	26	11	23	56	74	253	84	192	14	206	192	74	207	245	40	65	171	121	89	240	21	126	53	41	21	135	65

FREQUENCY HOP CODE NUMBER= 12

99	57	64	79	189	42	86	205	241	83	106	85	51	42	212	161	170	26	116	119	4	246	147	80	2	5	244	6	123	150	109	144
66	1	81	36	172	217	236	245	5	13	241	39	72	29	177	163	10	152	100	146	54	74	158	60	2	2	154	119	10	169	177	51
218	125	175	214	131	216	11	228	68	196	114	166	64	214	237	5	146	149	46	24	77	138	164	116	137	141	247	131	33	144	134	14
51	87	226	195	86	68	213	140	3	233	94	87	55	0	170	246	240	73	145	235	217	17	71	238	133	81	57	32	197	7	106	43
157	199	192	177	67	214	170	51	15	173	150	171	205	214	44	95	86	230	140	137	252	10	109	176	254	251	12	250	133	106	147	112
190	255	175	220	84	39	20	11	251	243	15	217	184	227	79	93	246	104	156	110	202	182	98	196	254	254	102	137	246	87	79	205
38	131	81	42	125	40	245	28	188	60	142	90	192	42	19	251	110	107	210	232	179	118	92	140	119	115	9	125	223	112	122	242
205	169	30	61	170	188	43	116	253	23	162	169	201	0	86	10	16	183	111	21	39	239	185	18	123	175	199	224	59	249	150	213

Table 5-1. Printout of FH Codes for Multiple Access (Continued)

FREQUENCY HOP CODE NUMBER= 13

51	46	25	27	70	170	200	226	38	228	193	121	251	72	251	186	74	4	37	15	23	245	120	122	107	40	162	77	12	209	187	194
131	213	178	84	210	215	182	73	216	83	76	182	105	148	77	6	140	64	38	243	130	70	50	228	168	141	80	150	75	152	232	250
221	206	50	153	26	208	52	51	145	218	221	228	195	180	91	147	152	55	51	158	124	122	246	17	35	186	205	254	40	76	97	43
63	154	236	208	180	36	120	199	54	68	9	78	69	120	96	48	8	207	164	5	149	209	142	99	235	8	142	14	225	248	214	159
205	210	231	229	186	96	56	30	218	28	63	135	5	184	5	70	182	252	219	241	233	11	136	134	149	216	94	179	244	47	69	62
125	43	78	172	46	41	74	183	40	173	180	74	151	108	179	250	116	192	218	13	126	186	206	28	88	115	176	106	181	104	24	6
35	50	206	103	230	48	204	205	111	38	35	28	61	76	165	109	104	201	205	98	132	134	10	239	221	70	51	2	216	180	159	213
193	102	20	48	76	220	136	57	202	188	247	178	187	136	160	208	248	49	92	251	107	47	114	157	21	248	114	242	31	8	42	97

FREQUENCY HOP CODE NUMBER= 14

40	7	229	164	198	28	221	23	183	59	229	65	25	111	20	90	52	181	189	34	22	218	162	227	142	214	233	222	71	31	237	3
87	54	226	122	209	161	10	28	30	174	219	215	224	48	176	136	52	2	135	63	126	218	218	138	51	67	111	215	58	207	175	253
46	81	245	48	18	249	131	128	167	69	27	103	161	34	233	153	58	60	185	205	108	204	147	171	80	144	72	5	228	39	126	55
130	164	249	46	148	199	179	250	145	239	0	92	189	46	154	72	142	226	190	193	85	24	3	201	162	93	124	42	210	100	74	207
216	249	27	92	58	228	35	233	73	197	27	191	231	145	236	166	204	75	67	222	234	38	94	29	114	42	23	34	185	225	19	253
169	202	30	134	48	95	246	228	226	82	37	41	32	208	80	120	204	254	121	193	130	38	38	118	205	189	145	41	198	49	81	3
210	175	11	208	238	7	125	128	89	187	229	153	95	222	23	103	198	196	71	51	148	52	109	85	176	112	184	251	28	217	130	201
126	92	7	210	108	57	77	6	111	17	0	164	67	210	102	184	114	30	66	63	171	232	253	55	94	163	132	214	46	156	182	49

FREQUENCY HOP CODE NUMBER= 15

1	211	110	36	56	49	18	168	14	95	173	95	64	136	180	68	229	77	208	33	251	4	11	6	60	29	122	25	149	81	46	215
184	102	8	120	154	245	221	98	121	61	252	78	124	147	50	48	246	99	211	59	18	130	128	21	233	98	176	198	113	150	178	78
177	20	140	40	59	72	208	150	18	131	158	69	15	176	239	59	63	194	232	189	190	105	45	216	38	11	79	193	191	68	138	122
140	177	87	14	55	2	230	85	60	230	14	212	115	104	178	206	161	252	122	129	156	141	105	128	247	75	152	27	62	216	122	218
255	45	146	220	200	207	238	88	242	161	83	161	192	120	76	188	27	179	48	223	5	252	245	250	196	227	134	231	107	175	210	41
72	154	248	136	102	11	35	158	135	195	4	178	132	109	206	208	10	157	45	197	238	126	128	235	23	158	80	58	143	106	78	178
79	236	116	216	197	184	48	106	238	125	98	187	241	80	17	197	193	62	24	67	66	151	211	40	218	245	177	63	65	188	118	134
116	79	169	242	201	254	26	171	196	26	242	44	141	152	78	50	95	4	134	127	100	115	151	128	9	181	104	229	194	40	134	38

FREQUENCY HOP CODE NUMBER= 16

205	92	238	150	77	102	163	255	50	39	203	134	89	40	158	245	125	96	207	6	37	109	46	180	131	174	181	103	199	146	2	56
232	140	6	66	238	200	35	189	8	94	115	234	223	21	218	242	87	175	207	207	186	40	11	203	8	163	159	253	56	153	3	209
116	171	132	81	138	149	230	1	80	6	124	179	157	182	145	64	197	241	216	15	91	3	90	174	161	18	11	156	220	80	205	132
153	15	55	177	114	53	65	0	51	244	134	138	173	128	56	225	187	184	58	200	17	243	32	213	229	103	137	135	178	8	133	1
51	164	18	106	179	154	93	1	206	217	53	122	167	216	98	11	131	160	49	250	219	147	210	76	125	82	75	153	57	110	254	200
24	116	250	190	18	56	221	67	248	162	141	22	33	235	38	14	169	81	49	49	70	216	245	53	248	93	97	3	200	103	253	47
140	85	124	175	118	107	26	255	176	250	132	77	99	74	111	192	59	15	40	241	165	253	166	82	95	238	245	100	36	176	51	124
103	241	201	79	142	203	191	0	205	12	122	118	83	128	200	31	69	72	198	56	239	13	224	43	27	153	119	121	78	248	123	255

FREQUENCY HOP CODE NUMBER= 17

86	220	96	171	130	247	250	35	250	69	242	159	249	18	79	141	144	95	180	48	142	144	220	251	20	233	3	153	8	102	99	104
14	138	208	150	193	14	126	76	41	213	15	77	97	189	156	83	163	171	99	119	96	179	193	234	73	146	214	196	59	234	134	148
11	163	173	160	215	171	81	63	211	228	234	65	163	88	150	198	244	225	42	172	245	48	48	41	168	206	230	185	232	147	215	145
247	239	218	236	165	144	236	247	65	108	60	196	197	6	75	251	119	120	129	61	119	170	117	195	1	88	245	251	226	19	172	53
170	36	160	85	126	9	6	221	6	187	14	97	7	238	177	115	112	161	76	208	114	112	36	5	236	23	253	103	248	154	157	152
242	118	48	106	63	242	130	180	215	43	241	179	159	67	100	173	93	85	157	137	160	77	63	22	183	110	42	60	197	22	122	108
245	93	83	96	41	85	175	193	45	28	22	191	93	168	106	58	12	31	214	84	11	208	208	215	88	50	26	71	24	109	41	111
9	17	38	20	91	112	20	9	191	148	196	60	59	250	181	5	137	136	127	195	137	86	139	61	255	168	11	5	30	237	84	203

FREQUENCY HOP CODE NUMBER= 18

214	78	117	224	19	78	30	235	24	108	11	63	227	195	231	160	143	68	222	153	177	62	35	140	79	55	53	218	220	199	147	142
12	84	36	105	7	105	13	109	160	113	114	207	9	127	253	159	159	63	11	29	235	105	224	43	56	201	157	199	140	109	73	43
3	204	252	237	237	22	143	194	177	82	120	71	69	93	28	245	228	51	199	70	34	6	171	48	100	169	3	197	43	157	228	239
215	146	21	31	0	59	227	5	185	34	118	220	75	25	101	183	55	191	246	163	46	255	99	223	242	196	105	43	237	58	224	172
42	178	139	32	237	178	226	21	232	148	245	193	29	61	25	96	113	188	34	103	79	194	221	116	177	201	203	38	36	57	109	114
244	172	220	151	249	151	243	147	96	143	142	49	247	129	3	97	97	193	245	227	21	151	32	213	200	55	99	57	116	147	183	213
253	52	4	19	19	234	113	62	79	174	136	185	187	163	228	11	28	205	57	186	222	250	85	208	156	87	253	59	213	99	28	17
41	110	235	275	0	197	29	251	71	222	138	36	181	231	155	73	201	65	10	93	210	1	157	33	14	60	151	213	19	198	32	84

Table 5-1. Printout of FH Codes for Multiple Access (Continued)

FREQUENCY HOP CODE NUMBER= 19

72	99	170	113	106	114	230	9	63	133	171	41	148	91	250	159	116	110	71	188	95	133	180	199	157	105	118	174	61	247	185	140
214	168	247	175	98	248	46	228	60	212	244	119	203	224	73	155	51	231	177	168	161	136	33	26	111	144	160	24	15	48	224	35
44	27	73	3	88	84	18	160	31	224	126	233	74	227	75	229	54	208	97	115	248	129	178	236	63	198	15	8	53	170	66	207
122	205	72	122	171	50	241	125	111	92	142	98	94	51	33	119	126	52	92	90	131	237	127	208	94	56	153	54	20	110	87	44
184	157	86	143	150	142	26	247	193	123	85	215	108	165	6	97	140	146	185	68	161	123	76	57	99	151	138	82	195	9	71	116
42	88	9	81	158	8	210	28	196	44	12	137	53	32	183	101	205	25	79	88	95	120	223	230	145	112	96	232	241	208	32	221
212	229	183	253	168	172	238	96	225	32	130	23	182	29	181	27	202	48	159	141	8	127	78	20	193	58	241	248	203	86	190	49
134	51	184	134	85	206	15	131	145	164	114	158	162	205	223	137	130	204	164	166	125	19	129	48	162	200	103	202	236	146	169	212

FREQUENCY HOP CODE NUMBER= 20

93	152	59	200	142	58	4	48	88	37	149	218	44	110	249	132	158	215	106	106	166	22	239	21	207	170	74	15	109	29	183	86
42	123	61	10	241	25	165	128	159	86	156	57	44	44	69	47	219	141	60	94	192	201	16	81	54	147	241	155	210	199	216	76
123	104	95	110	150	215	240	14	173	230	32	238	208	18	59	55	211	106	142	73	115	136	110	199	92	210	82	18	66	8	34	114
181	0	163	37	162	64	105	51	169	116	20	117	120	239	225	190	243	154	19	175	113	9	112	60	210	104	164	93	72	229	215	186
163	104	197	56	114	198	252	208	168	219	107	38	212	146	7	124	98	41	150	150	90	234	17	235	49	86	182	241	147	227	73	170
214	133	195	246	15	231	91	128	97	170	100	199	212	212	187	209	37	115	196	162	64	55	240	175	202	109	15	101	46	57	40	180
133	152	161	146	106	41	16	242	83	26	224	18	48	238	197	201	45	150	114	183	141	120	146	57	164	46	174	238	190	248	222	142
75	0	93	219	94	192	151	205	87	140	236	139	136	17	31	66	13	102	237	81	143	247	144	196	46	152	92	163	184	27	41	70

FREQUENCY HOP CODE NUMBER= 21

146	41	146	236	86	88	43	73	248	15	70	114	63	109	222	174	7	250	24	177	55	81	61	71	16	126	171	63	147	27	129	170
253	193	152	153	18	144	65	227	33	254	94	154	120	40	217	215	129	24	242	125	1	184	71	24	57	228	116	94	105	191	1	155
200	126	202	172	25	181	94	156	179	136	37	116	255	2	141	212	109	151	100	196	122	68	73	228	104	21	92	31	160	232	197	173
232	91	78	28	176	184	31	109	193	250	39	143	52	175	40	51	89	81	104	157	141	250	220	176	2	115	203	145	191	101	101	165
110	215	110	20	170	168	213	183	8	241	186	142	193	147	34	82	249	6	232	79	201	175	195	185	240	130	85	193	109	229	127	86
3	63	104	103	238	112	191	29	223	2	162	102	136	216	39	41	127	232	14	131	255	72	185	232	199	28	140	162	151	65	255	101
56	130	54	84	231	75	162	100	77	120	219	140	1	254	115	44	147	105	156	60	134	188	183	28	152	235	164	225	96	24	59	83
24	165	178	228	80	72	225	147	63	6	217	113	204	81	216	205	167	175	152	99	115	6	36	80	254	141	53	111	65	155	155	91

FREQUENCY HOP CODE NUMBER= 22

35	128	182	180	116	127	68	233	226	192	222	133	62	82	8	23	42	168	95	66	114	159	111	136	228	223	219	101	145	229	213	125
67	28	39	186	137	44	164	101	201	192	191	230	116	188	129	125	12	206	17	190	240	239	14	27	138	103	55	245	97	232	80	232
222	233	8	47	247	35	236	162	85	141	171	163	239	84	42	110	154	109	223	203	54	31	102	240	171	31	105	125	128	139	0	224
67	6	69	42	40	110	89	133	71	13	65	75	244	246	157	153	16	166	86	185	126	102	80	224	13	154	255	8	63	243	80	112
221	128	74	76	140	129	188	23	30	64	34	123	194	174	248	233	214	88	161	190	142	97	145	120	28	33	37	155	111	27	43	131
189	228	217	70	119	212	92	155	55	64	65	26	140	68	127	131	244	50	239	66	16	17	242	229	118	153	201	11	159	24	176	24
34	23	248	209	9	221	20	94	171	115	85	93	17	172	214	146	102	147	33	53	202	225	154	16	85	225	151	131	128	117	0	32
189	250	187	214	216	146	167	123	185	243	191	181	12	10	99	103	240	90	170	71	130	154	176	32	243	102	1	248	193	13	176	144

FREQUENCY HOP CODE NUMBER= 23

122	164	126	210	155	152	228	211	147	88	241	132	35	124	113	58	216	239	240	125	192	209	176	92	69	15	1	99	91	57	168	195
158	171	72	49	37	143	38	13	139	33	11	226	8	100	39	8	194	237	82	173	39	182	17	108	13	42	206	237	138	55	157	254
73	39	139	13	101	177	242	68	90	19	218	147	65	241	196	155	112	232	230	135	17	60	114	51	181	44	199	93	35	198	51	59
238	253	83	162	222	168	113	11	90	39	253	11	59	107	3	80	101	148	114	170	234	218	128	235	52	206	118	136	205	222	27	223
134	92	130	46	101	104	28	45	109	168	15	124	221	132	143	198	40	17	16	131	64	47	80	164	187	241	255	157	165	199	88	61
98	85	184	207	219	113	218	243	117	223	245	30	248	156	217	248	62	19	174	83	217	74	239	148	243	214	50	19	118	201	99	2
183	217	117	243	155	79	14	188	166	237	38	109	191	15	60	101	144	24	26	121	239	196	142	205	75	212	57	163	221	58	205	197
18	3	173	94	34	88	143	245	166	217	3	245	197	149	253	176	155	108	142	86	22	38	128	21	204	50	138	120	51	34	229	33

FREQUENCY HOP CODE NUMBER= 24

158	108	156	249	180	56	206	132	43	107	240	105	77	229	148	232	31	128	43	203	242	18	132	189	117	53	255	45	175	12	238	30
45	204	191	205	136	17	206	207	236	109	7	118	176	10	178	190	225	46	65	228	238	185	98	239	208	193	198	22	217	132	179	105
135	170	105	123	243	183	148	73	224	66	202	229	222	139	241	113	235	239	162	98	46	72	181	61	194	138	167	0	94	249	142	230
229	11	203	88	24	192	247	30	116	227	189	82	176	209	186	165	83	176	99	22	94	10	139	18	104	69	246	22	184	169	138	136
98	148	100	7	76	200	50	124	213	149	16	151	179	27	108	24	225	128	213	53	14	238	124	67	139	203	1	211	81	244	18	226
211	52	65	51	120	239	50	49	20	147	249	138	80	246	78	66	31	210	191	28	18	71	158	17	48	63	58	234	39	124	77	151
121	86	151	133	13	73	108	183	32	190	54	27	34	117	15	143	21	17	94	158	210	184	75	195	62	118	89	0	162	7	114	26
27	245	53	168	232	64	9	226	140	29	67	174	80	47	70	91	173	80	157	234	162	246	117	238	152	187	10	234	72	87	118	120

Table 5-1. Printout of FH Codes for Multiple Access (Continued)

FREQUENCY HOP CODE NUMBER= 25

102	138	195	18	84	34	127	28	62	106	213	147	182	8	66	47	176	187	121	253	51	230	229	237	155	51	201	129	130	82	73	173
78	67	91	48	10	185	144	48	56	105	155	30	86	149	104	221	34	29	120	171	241	10	229	178	103	185	239	101	38	154	30	167
10	136	215	9	249	89	153	207	15	50	28	130	120	184	199	236	242	171	125	127	58	139	191	74	32	106	74	59	145	84	57	221
243	131	129	146	48	70	10	56	48	163	4	199	22	136	15	147	111	161	207	138	142	21	178	70	223	197	132	1	131	24	51	100
154	118	61	238	172	222	129	228	194	150	43	109	74	248	190	209	80	69	135	3	205	26	27	19	101	205	55	127	126	174	183	83
178	189	165	208	246	71	112	208	200	151	101	226	170	107	152	35	222	227	136	85	15	246	27	78	153	71	17	155	218	102	226	89
246	120	41	247	7	167	103	49	241	206	228	126	136	72	57	20	14	85	131	129	198	117	65	182	224	150	182	197	111	172	199	35
13	125	127	110	208	186	246	200	208	93	252	57	234	120	241	109	145	95	49	118	114	235	78	186	33	59	124	255	125	232	205	156

FREQUENCY HOP CODE NUMBER= 26

132	177	220	178	62	211	23	47	61	79	255	252	217	182	137	192	235	9	171	62	7	71	21	19	153	253	29	84	200	173	216	206
197	223	190	178	178	123	241	124	52	253	67	196	225	75	135	30	17	84	63	174	66	141	168	73	95	226	62	178	60	5	92	42
232	246	101	15	155	94	31	254	255	132	185	28	165	142	66	243	174	134	154	139	125	149	204	168	0	13	133	110	236	255	48	235
107	57	187	170	182	89	36	244	240	234	121	45	205	221	253	175	96	13	67	186	153	60	230	189	95	83	111	204	242	193	15	156
124	79	36	78	194	45	233	209	195	177	1	4	39	74	119	64	21	247	85	194	249	185	235	237	103	3	227	172	56	83	40	50
59	33	66	78	78	133	15	132	204	3	189	60	31	181	121	226	239	172	193	82	190	115	88	183	161	30	194	78	196	251	164	214
24	10	155	241	101	162	225	2	1	124	71	228	91	114	190	13	82	122	102	117	131	107	52	88	0	243	123	146	20	1	208	21
149	199	69	86	74	167	220	12	16	22	135	211	51	35	3	81	160	243	189	70	103	196	26	67	161	173	145	52	14	63	241	100

FREQUENCY HOP CODE NUMBER= 27

171	202	124	156	239	107	42	46	34	121	104	31	135	253	26	251	57	59	236	18	104	119	59	17	99	81	240	154	35	60	249	69
97	66	64	90	116	220	61	120	200	165	233	79	151	106	200	13	72	27	66	255	197	80	63	65	136	49	139	200	167	67	223	8
86	132	107	177	160	228	78	238	81	33	83	73	123	9	73	175	137	163	166	206	135	162	42	136	163	72	184	201	151	246	62	99
33	115	211	48	201	115	224	180	55	95	223	228	34	203	25	160	204	129	115	197	192	112	93	61	237	62	58	59	155	157	71	126
85	54	132	100	17	149	214	210	222	135	152	225	121	3	230	5	199	197	20	238	152	137	197	239	157	175	16	102	221	196	7	187
159	190	192	166	140	36	195	136	56	91	23	177	105	150	56	243	184	229	190	1	59	176	193	191	120	207	117	56	89	189	33	248
170	124	149	79	96	28	178	18	175	223	173	183	133	247	183	81	119	93	90	50	121	94	214	120	93	184	72	55	105	10	194	157
223	141	45	208	55	141	32	76	201	161	33	28	222	53	231	96	52	127	141	59	64	144	163	195	19	194	198	197	101	99	185	130

FREQUENCY HOP CODE NUMBER= 28

196	106	102	77	135	126	41	19	76	226	139	205	206	142	85	73	107	124	192	115	152	157	57	219	183	36	54	245	178	93	112	225
196	196	232	28	213	40	57	12	112	75	116	5	182	171	183	68	15	30	147	130	136	231	55	106	215	126	161	51	229	198	189	118
228	138	13	182	38	19	62	64	238	187	128	31	246	16	5	138	166	175	233	216	148	0	10	43	222	123	19	116	142	4	182	25
91	139	89	67	227	47	160	251	172	197	150	57	16	231	10	12	64	177	126	236	244	231	221	203	216	9	169	228	119	213	41	87
60	150	154	179	121	130	215	237	180	30	117	51	50	114	171	183	149	132	64	141	104	99	199	37	73	220	202	11	78	163	144	31
60	60	24	228	43	216	199	244	144	181	140	251	74	85	73	188	241	226	109	126	120	25	201	150	41	130	95	205	27	58	67	138
28	118	243	74	218	237	194	192	18	69	128	225	10	240	251	118	90	81	23	40	108	0	246	213	34	133	237	140	114	252	74	231
165	117	167	189	29	209	96	5	84	59	106	199	240	25	246	244	192	79	130	20	12	25	35	53	40	247	87	28	137	43	215	169

FREQUENCY HOP CODE NUMBER= 29

100	84	23	229	154	125	14	61	181	5	57	20	95	201	163	123	172	80	33	163	190	155	3	47	138	106	145	132	211	212	12	68
70	108	170	125	33	36	205	180	22	214	42	36	247	154	238	11	18	111	22	69	31	223	96	185	36	148	12	113	104	164	43	4
234	44	18	60	85	3	144	221	136	232	86	154	253	204	224	167	178	242	243	229	242	224	173	102	17	214	190	107	156	124	108	83
115	17	108	93	159	239	231	112	18	124	235	39	44	216	118	128	112	188	165	32	107	103	107	182	163	120	82	192	175	183	2	62
156	172	233	27	102	131	242	195	75	251	199	236	161	55	93	133	84	176	223	93	66	101	253	209	118	150	111	124	45	44	244	188
186	148	86	131	223	220	51	76	234	42	214	220	9	102	18	245	238	145	234	187	225	33	160	71	220	108	244	143	152	92	213	252
22	212	238	196	171	253	112	35	120	24	170	102	3	52	32	89	78	14	13	27	14	32	83	154	239	42	66	149	100	132	148	173
141	239	148	163	97	17	25	144	238	132	21	217	212	40	138	128	144	68	91	224	149	153	149	74	93	136	174	64	81	73	254	194

FREQUENCY HOP CODE NUMBER= 30

78	5	175	248	153	98	56	166	216	179	128	165	154	23	213	188	128	177	81	201	188	101	87	2	208	197	32	165	74	112	111	198
238	46	11	201	29	184	117	90	161	140	73	101	230	209	181	14	99	242	217	220	23	8	175	6	58	255	74	244	70	18	185	10
140	49	152	107	69	85	45	119	181	190	209	161	185	167	253	179	245	252	0	67	210	131	232	153	108	129	181	121	20	50	166	107
249	36	134	25	95	54	92	214	201	209	217	67	29	68	234	176	123	227	217	151	235	245	86	129	18	33	46	248	145	144	233	158
178	251	81	8	103	158	200	90	40	77	128	91	102	233	43	68	128	79	175	55	68	155	169	254	48	59	224	91	182	144	145	58
18	210	245	55	227	72	139	166	95	116	183	155	26	47	75	242	157	14	39	36	233	248	81	250	198	1	182	12	186	238	71	246
116	207	104	149	187	171	211	137	75	66	47	95	71	89	3	77	11	4	0	189	46	125	24	103	148	127	75	135	236	206	90	149
7	220	122	231	161	202	164	42	55	47	39	189	227	188	22	80	133	29	39	105	21	11	170	127	238	223	210	8	111	112	23	98

Table 5-2 is a listing of 30 pairs of maximal code sequence generators for the forward link. The codes are represented by a polynomial expressed in conventional octal notation. The corresponding binary number specifies the nonzero coefficients of the polynomial defining the tap connections; e.g.,

octal
1 0 0 0 4 7

binary
1 0 0 0 0 0 0 0 0 0 0 0 0 1 0 0 1 1 1

This example feeds back the mod-2 sum of the 14th, 17th, 18th, and 19th stages to the first stage of the shift register generator. Alternatively, as described in the Phase I - Final Report, a modular generator configuration can be used which has mod-2 adders between the register stages. Then, the binary number specifies where mod-2 adders must be located. (The leftmost 1 specifies feedback to the first stage. The rightmost 1 specifies feedback from the last stage. The remaining 1's denote where mod-2 adders must be inserted.)

Table 5-3 is a listing of 30 different code pairs for the multiple access return link.

There is a return-only mode which utilizes staggered quadriphase PN codes of a shorter period. A listing of 30 code pairs of period $2^{11} = 2048$, $2^{13} = 8192$, or $2^{15} = 32768$ is given in table 5-4.

It should be noted that one code of the pair can be arbitrarily shifted with respect to the other. For convenience, both can be started together with the same register contents, say, all 1's.

5.3 FH CODE FOR SINGLE ACCESS

The frequency hopping code for the S-band single access service has a period of 512 chips and a hopping rate of 6.012 kHz. The FH sequence is generated by a maximal binary sequence generator. At each successive shift, the shift register contains a number, in binary form, over the range from 0 to 511, and this specifies the frequency number.

Table 5-5 lists 20 polynomials to be used for FH code generation.

5.4 PN CODES FOR SINGLE ACCESS

The PN codes for single access are staggered quadriphase with a period of 2^{19} = 524287 chips and a chip rate of 6.156 MHz. Table 5-5 lists 20 pairs to be used for PN code generation on the forward link. Table 5-6 lists 20 additional pairs for the return link.

Table 5-2. MA 18-Stage Forward PN Pairs (octal polynomials)

1 0 0 0 0 4 7	1 0 0 1 7 2 7	Augment period by one chip
1 0 0 0 0 7 7	1 0 0 1 7 4 1	
1 0 0 0 1 1 5	1 0 0 2 0 3 1	
1 0 0 0 1 7 3	1 0 0 2 0 6 1	
1 0 0 0 2 0 1	1 0 0 2 0 7 5	
1 0 0 0 3 3 3	1 0 0 2 1 3 3	
1 0 0 0 3 4 7	1 0 0 2 1 7 1	
1 0 0 0 3 5 5	1 0 0 2 2 1 1	
1 0 0 0 4 0 7	1 0 0 2 2 4 1	
1 0 0 0 5 1 7	1 0 0 2 4 4 1	
1 0 0 0 6 2 1	1 0 0 2 6 2 3	
1 0 0 0 7 4 3	1 0 0 2 6 3 7	
1 0 0 0 7 5 1	1 0 0 2 7 0 5	
1 0 0 0 7 5 7	1 0 0 2 7 3 3	
1 0 0 1 0 1 3	1 0 0 2 7 4 1	
1 0 0 1 0 2 3	1 0 0 2 7 7 7	
1 0 0 1 1 4 1	1 0 0 3 0 1 1	
1 0 0 1 1 6 5	1 0 0 3 0 3 5	
1 0 0 1 2 5 3	1 0 0 3 0 5 3	
1 0 0 1 3 6 1	1 0 0 3 1 7 5	
1 0 0 1 4 2 7	1 0 0 3 2 1 5	
1 0 0 1 4 5 3	1 0 0 3 4 5 1	
1 0 0 1 4 5 5	1 0 0 3 4 6 1	
1 0 0 1 5 6 7	1 0 0 3 5 2 1	
1 0 0 1 6 0 7	1 0 0 3 6 7 7	
1 0 0 1 6 2 5	1 0 0 3 7 1 5	
1 0 0 1 6 3 1	1 0 0 3 7 6 1	
1 0 0 1 6 5 1	1 0 0 4 0 7 3	
1 0 0 1 6 6 1	1 0 0 4 1 6 3	
1 0 0 1 7 0 5	1 0 0 4 2 0 5	

Table 5-3. MA 18-Stage Return PN Pairs (octal polynomials)

1 0 0 7 1 2 1	1 0 1 6 5 6 1	Augment period by one chip
1 0 0 7 1 6 5	1 0 1 6 5 6 7	
1 0 0 7 1 7 1	1 0 1 6 7 0 5	
1 0 0 7 2 6 3	1 0 1 7 0 2 1	
1 0 0 7 3 1 5	1 0 1 7 0 7 1	
1 0 0 7 4 1 7	1 0 1 7 1 6 1	
1 0 0 7 5 0 1	1 0 1 7 2 1 5	
1 0 0 7 5 4 3	1 0 1 7 2 4 3	
1 0 0 7 6 3 7	1 0 1 7 2 6 1	
1 0 0 7 7 0 5	1 0 1 7 3 1 1	
1 0 0 7 7 7 7	1 0 1 7 3 2 7	
1 0 1 0 0 4 5	1 0 1 7 3 3 3	
1 0 1 0 0 5 1	1 0 1 7 5 0 3	
1 0 1 0 1 3 3	1 0 1 7 5 1 1	
1 0 1 0 1 6 3	1 0 1 7 5 5 5	
1 0 1 0 2 1 1	1 0 1 7 5 6 5	
1 0 1 0 3 1 3	1 0 1 7 6 1 1	
1 0 1 0 4 6 3	1 0 1 7 7 2 3	
1 0 1 0 5 5 1	1 0 1 7 7 2 3	
1 0 1 0 6 1 3	1 0 2 0 0 0 7	
1 0 1 0 6 1 5	1 0 2 0 0 6 1	
1 0 1 0 7 2 7	1 0 2 0 1 2 1	
1 0 1 0 7 4 1	1 0 2 0 1 7 1	
1 0 1 1 0 4 1	1 0 2 0 2 4 1	
1 0 1 1 0 5 5	1 0 2 0 2 5 5	
1 0 1 1 2 4 5	1 0 2 0 2 7 7	
1 0 1 1 2 6 1	1 0 2 0 3 1 5	
1 0 1 1 2 6 7	1 0 2 0 4 7 1	
1 0 1 1 3 3 3	1 0 2 0 5 0 7	
1 0 1 1 3 4 7	1 0 2 0 5 1 3	

Table 5-4. MA Return Link Only (octal polynomials)

11-Stage Pairs		13-Stage Pairs		15-Stage Pairs	
4 0 0 5	6 7 1 1	2 0 0 3 3	2 5 5 1 1	1 0 0 0 0 3	1 3 1 4 2 7
4 4 4 5	7 7 1 5	2 3 2 6 1	2 6 5 3 3	1 0 2 0 4 3	1 6 1 6 1 5
4 2 1 5	6 3 4 3	2 4 6 2 3	3 3 3 4 3	1 1 0 0 1 3	1 1 5 1 5 5
4 0 5 5	6 2 2 7	2 3 5 1 7	3 3 7 2 7	1 0 2 0 6 7	1 2 3 0 6 7
6 0 1 5	6 2 6 3	3 0 7 4 1	2 7 2 7 1	1 0 4 3 0 7	1 0 2 5 6 1
7 4 1 3	5 2 3 5	2 1 6 4 3	2 5 0 1 7	1 0 0 3 1 7	1 7 0 0 5 7
4 1 4 3	7 4 3 1	3 0 1 7 1	2 6 0 4 1	1 7 7 7 7 5	1 2 5 2 3 5
4 5 6 3	6 4 5 5	2 1 2 7 7	2 1 1 0 3	1 0 3 4 5 1	1 7 3 1 1 7
4 0 5 3	5 2 4 7	2 7 7 7 7	2 7 2 6 3	1 1 0 0 7 5	1 3 4 5 3 1
5 0 2 3	5 2 6 5	3 5 0 5 1	2 4 5 1 3	1 0 2 0 6 1	1 2 5 5 0 7
5 6 2 3	4 7 6 7	3 4 7 2 3	3 2 3 1 1	1 1 4 7 2 5	1 7 1 7 3 7
4 5 7 7	5 6 0 7	3 4 0 4 7	3 1 7 4 3	1 0 3 2 5 1	1 5 2 4 1 7
6 2 3 3	4 6 0 3	3 2 5 3 5	2 4 0 3 7	1 6 3 0 0 5	1 4 2 3 0 5
6 6 7 3	6 5 6 1	3 1 4 2 5	3 0 7 1 1	1 1 2 6 1 1	1 2 0 0 4 3
7 2 3 7	7 1 0 7	3 7 5 0 5	3 2 6 4 1	1 2 0 2 6 5	1 3 6 1 7 3
7 3 3 5	7 0 4 1	3 6 5 1 5	2 4 6 5 7	1 1 7 4 2 3	1 2 2 2 3 1
4 5 0 5	4 2 5 1	2 6 0 7 7	3 2 4 3 7	1 0 6 3 4 1	1 6 4 7 0 5
5 3 3 7	5 6 7 5	3 5 6 7 3	2 0 2 1 3	1 6 1 0 0 7	1 7 7 7 5 7
5 2 6 3	4 1 7 3	2 0 6 3 5	2 5 6 3 3	1 7 4 0 0 3	1 4 6 6 3 7
5 3 6 1	4 7 0 7	3 3 7 6 3	3 1 3 0 3	1 1 3 3 3 7	1 7 7 5 3 5
5 1 7 1	5 4 6 3	2 5 7 4 5	2 2 5 2 5	1 2 6 0 0 7	1 0 2 6 4 3
6 6 3 7	5 7 5 5	3 6 5 7 5	3 4 6 2 7	1 0 5 2 5 7	1 0 3 1 4 5
7 1 7 3	6 6 7 5	2 6 6 5 3	2 5 7 7 5	1 1 4 4 6 7	1 1 2 7 5 1
5 7 1 1	7 6 5 5	2 1 1 3 3	2 1 6 0 7	1 7 7 2 0 7	1 5 1 5 3 7
5 2 2 1	5 5 3 1	2 2 4 4 1	2 5 3 6 3	1 4 7 0 4 7	1 1 5 1 3 5
6 3 0 7	7 2 4 3	3 0 4 1 7	2 7 2 1 7	1 1 1 5 1 1	1 3 7 0 6 7
6 2 1 1	7 6 2 1	3 2 5 1 7	3 3 7 4 1	1 1 4 6 3 3	1 7 4 4 4 3
5 7 4 7	7 1 6 1	3 7 3 3 5	3 7 6 1 1	1 3 3 6 6 3	1 0 0 5 4 1
4 5 3 3	4 7 3 1	2 5 3 2 7	2 3 0 7 7	1 0 2 1 7 1	1 1 2 2 7 3
4 3 4 1	4 4 5 1	2 3 2 3 1	2 1 2 6 3	1 7 0 4 6 5	1 4 5 5 7 3

Augment period by one chip.

Table 5-5. S-Band SA Forward Link (octal polynomials)

9-Stage for FH	19-Stage Forward PN Pair		
1021	2000047	2000677	Augment period by one chip
1131	2000077	2000707	
1461	2000107	2000715	
1423	2000123	2000737	
1055	2000131	2001013	
1167	2000143	2001023	
1541	2000157	2001031	
1333	2000175	2001045	
1605	2000223	2001061	
1751	2000257	2001117	
1743	2000341	2001121	
1617	2000503	2001153	
1553	2000541	2001155	
1157	2000553	2001203	
1715	2000565	2001205	
1563	2000605	2001227	
1713	2000635	2001241	
1175	2000641	2001441	
1725	2000655	2001507	
1225	2000663	2001551	

Table 5-6. S-Band SA Return PN Pair (octal polynomials)

19-Stage Pairs		
2 0 0 1 5 5 7	2 0 0 2 4 5 3	Augment period by one chip
2 0 0 1 6 3 7	2 0 0 2 4 7 1	
2 0 0 1 6 4 5	2 0 0 2 5 0 1	
2 0 0 1 6 5 1	2 0 0 2 5 2 3	
2 0 0 1 7 1 1	2 0 0 2 5 3 7	
2 0 0 1 7 2 7	2 0 0 2 5 6 1	
2 0 0 1 7 3 3	2 0 0 2 5 6 7	
2 0 0 1 7 3 5	2 0 0 2 6 5 1	
2 0 0 1 7 6 3	2 0 0 2 6 7 3	
2 0 0 2 0 6 7	2 0 0 2 7 2 1	
2 0 0 2 1 2 1	2 0 0 2 7 2 7	
2 0 0 2 1 6 5	2 0 0 2 7 3 5	
2 0 0 2 1 7 7	2 0 0 2 7 5 3	
2 0 0 2 2 1 1	2 0 0 2 7 6 5	
2 0 0 2 2 2 1	2 0 0 2 7 7 1	
2 0 0 2 2 6 5	2 0 0 3 0 1 1	
2 0 0 2 2 7 1	2 0 0 3 0 2 1	
2 0 0 2 3 2 3	2 0 0 3 0 6 3	
2 0 0 2 3 3 1	2 0 0 3 1 0 1	
2 0 0 2 4 3 5	2 0 0 3 1 1 3	

SECTION VI
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