DESIGN STUDY
OF A HEAO-C SPREAD SPECTRUM TRANSPONDER TELEMETRY SYSTEM FOR USE WITH THE TDRSS SUBNET

FINAL REPORT
NASA GRANT NO. NSG-8013

Submitted To

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION
GEORGE C. MARSHALL SPACE FLIGHT CENTER

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HUNTSVILLE, ALABAMA

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SEPTEMBER 30, 1975

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NOTE: EACH SECTION HAS INDEPENDENT EQUATION NUMBERING

INTRODUCTION

This report gives the results of a design study of a spread spectrum transponder for use on the HEAO-C satellite. The transponder performs the functions of code turn-around for ground range and range-rate determination, ground command receiver, and telemetry data transmitter. The spacecraft transponder and associated communication system components will allow the HEAO-C satellite to utilize the Tracking and Data Relay Satellite System (TDRSS) subnet of the post 1978 STDN.

Use of the TDRSS by HEAO-C is being considered for the following reasons:

(1) The ground site subnet of the post 1978 STDN will include only six-to eight sites.

(2) Reduction in the HEAO-C tape recorder requirement to only 15 percent of an orbit recorded when out of view of the TDRSS.

(3) Allows high real-time data rate transmission.

(4) Near-continuous monitoring and near-instantaneous access leading to real-time command and control.

In TDRSS terminology, HEAO is a medium data rate user. As such, depending on mission requirements, it could operate as a multiple access or S-band single-access user. The transponder is designed to allow the use of either of these modes of operation under control of ground command. This will allow for the greatest freedom in TDRSS scheduling, allow for growth in the HEAO experimental package, and still guarantee the most economical use of the TDRSS subnet capabilities.

The transponder design includes an eleventh order gold code generator operating at .6 M CHIP/SEC (MA) or 6 M CHIP/SEC (SSA) along with carrier and code-delay-lock loops for code acquisition and coherent despreading. The return telemetry data is coded by a K=7, V=3 convolutional encoder. This allows a 6db coding-gain.

Associated communication systems components have been specified by a previous report, and will be only briefly described in this report.

Section 2 is a description of the present TDRSS subnet from the point of view of the user's interface. Section 3 is a general description of the HEAO-C-TDRSS system configuration. Sections 4, 5, 6, 7, and 8 describe components of the HEAO-C transponder including the gold-code generator, convolutional encoder, and carrier and code delay-lock loops.

Section 9 is a summary of the total spread spectrum transponder system, and section 10 is a list of reference literature.

2. TDRSS SUBNET DESCRIPTION

This section gives a description of the TDRSS subnet as it affects the HEAO-C as a system user. The transponder design allows ground command programming as a MA or SSA user, so each of these TDRSS support features will be described. This material is from the June 10, 1974 TDRSS Users' Guide (X-805-74-176).

The Tracking and Data Relay Satellite System (TDRSS) concept consists of two geosynchronous relay satellites, 130 degrees apart in longitude and a ground terminal centrally located in the continental United States. Additionally, the system includes two spare satellites: one in orbit, and one in configuration for a rapid replacement launch. The payload of each Tracking and Data Relay Satellite (TDRS) is the telecommunications service system which relays communication signals between low earth-orbiting user spacecraft and the TDRSS ground terminal. A "bent-pipe" concept is used in the design of the telecommunications service system (i.e., all communication signals received at the TDRS are translated in frequency and retransmitted).

The telecommunications link from the ground terminal to the TDRS to the user is called the forward link and will be used to carry user command data, tracking signals, and voice transmissions. The link from the user to the TDRS to the ground terminal is called the return link and will be used to carry user telemetry data, return tracking signals, and voice. Both the forward and return links consists of a space-to-space link between the TDRS and the user, and a space-to-ground link between the TDRS and the TDRSS ground terminal.

Each TDRS provides the following two types of space-to-space communication links:

a. Multiple-access System. One 10-element S-band phased array antenna system to support the forward link (command link) of 20 users (time shared), and one 30-element S-band phased array antenna to support the return link of 20 users simultaneously. The spacecraft supported by this system are called Multiple-access (MA) users.

b. Single-access System. Two 3.8 meter parabolic antennas, each operating at both S- and Ku-band. This configuration is called a single-access system because each antenna will normally support one user at a time. However, each antenna can support two users simultaneously (one at S-band and one at Ku-band) provided both users are within the beamwidth of the antenna. The user spacecraft supported by this system are called Single-access (SA) S- or Ku-band users.

The two-satellite TDRSS concept is illustrated in figure 2-1. The general TDRSS Frequency plan (TDRSS to user) is an follows:

FORWARD

(1) 2287.5 MHZ - MULTIPLE ACCESS
(2) 2200 TO 2300 MHZ SINGLE ACCESS
(3) 14.6 TO 15.25 GHZ SINGLE ACCESS

RETURN

(1) 2106.4 MHZ - MULTIPLE ACCESS
(2) 2025 TO 2120 MHZ SINGLE ACCESS
(3) 13.4 TO 14.05 GHZ SINGLE ACCESS

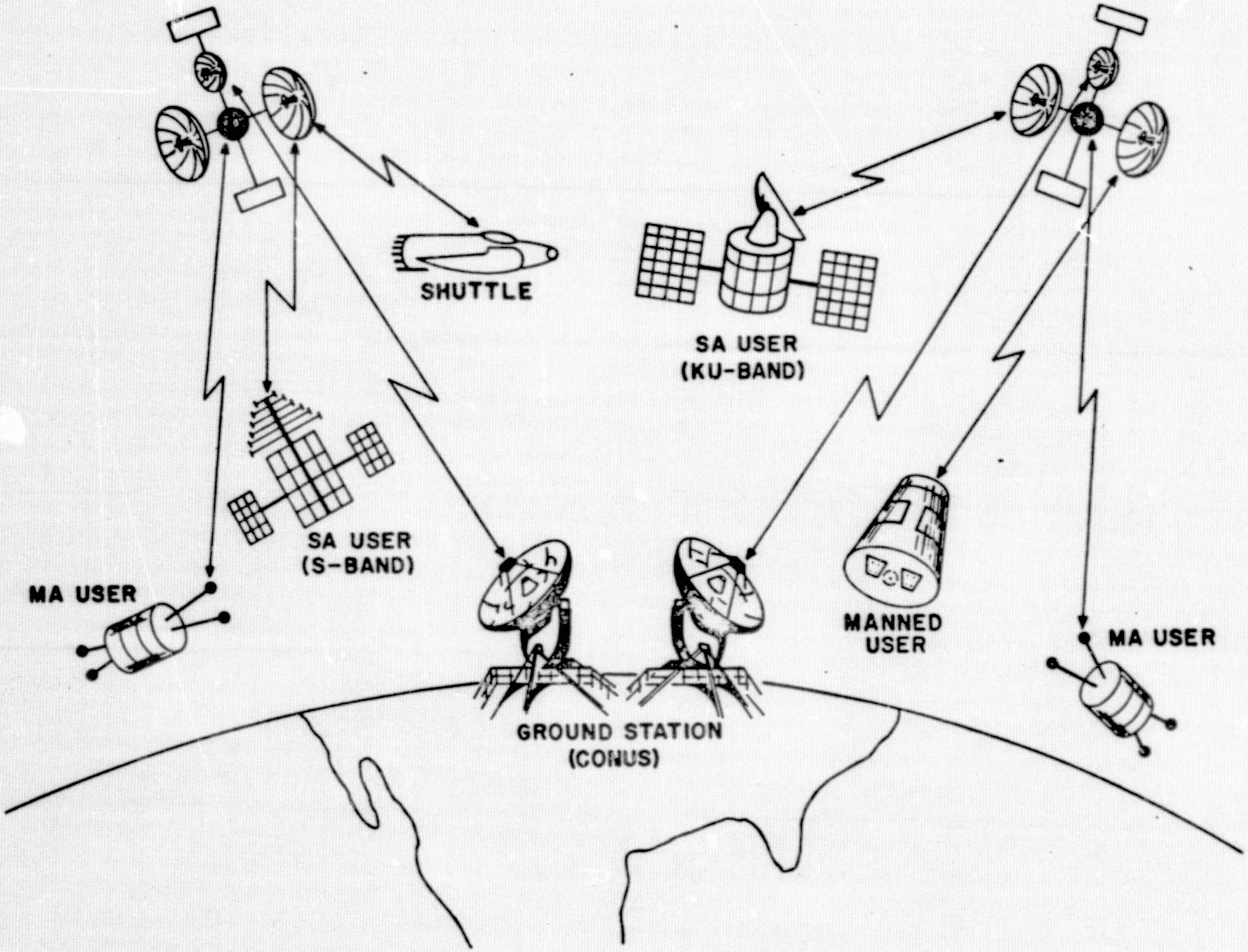


Figure 2-1 Two-satellite TDRSS Concept

The elements of the TDRSS described by support mode are as follows:

3.3 MULTIPLE-ACCESS SYSTEM

3.3.1 FORWARD (COMMAND) LINK

a.	Antenna	- 10-element phased array, 23-dB gain, single steered beam per TDRS.
b.	Frequency	- 2106.4 MHz, all users on same frequency.
c.	Bandwidth	- 5 MHz.
d.	TDRSS signal EIRP*	- 34 dBw peak.
e.	Duty factor	- Continuous.
f.	User command	- Time shared between users.
g.	Command rate	- 100 to 1000 b/s.
h.	Modulation	- PN spread spectrum. PSK ($\pm 90^0$), biphase.
i.	Operation	- All users on same command frequency, users separated by user unique codes, beam steered to desired user for duration of command and/or tracking sequence.
j.	Code type	- Gold, length to be defined ($\approx$ 2000 bits/code).

*EIRP in direction of user.

3.3.2 RETURN (TELEMETRY) LINK

a.	Antenna	- 30-element phased array (gain 28 dB).
b.	Frequency	- 2287.5 MHz (all users on same frequency).
c.	Bandwidth	- 5 MHz.
d.	Array beam forming	- All element combining/beam forming performed at ground terminal. Separate array beam formed for each user simultaneously.

e. Return link signal characteristics - Code division multiplex/PRN spread spectrum modulation (tentative value 3.0 Mch/s). PSK ($\pm 90^{\circ}$), biphase.

f. Maximum single user telemetry rate - 48 kb/s.

g. Average user telemetry rate≈ 10 kb/s - Each user's supportable data rate is a function of the number, EIRP, and data rates of the other simultaneously-supported users.

h. Support duration/user - Continuous when in view of either TDRS (at least 85 percent of each low earth orbit).

i. Data handling - Data returned to user in real time.

j. Code type - Gold, length to be determined (≈ 2000 bits).

SINGLE-ACCESS SERVICE

GENERAL

Each single-access system can operate at S-band (command and telemetry), Ku-band (command and telemetry), or both simultaneously. There are two single-access systems per TDRS.

S-BAND SINGLE-ACCESS SERVICE

Forward (Command) Link

a. Antenna - 3.8-meter diameter parabolic reflector.

b. Antenna gain - 35.4 dB.

c. Frequency - 2025 to 2120 MHz. Each user at a separate frequency.

d. TDRS signal EIRP - 43.4 dBw peak normal; 46.0 dBw peak high power.

e. Bandwidth - 20 MHz narrowband, tunable over 100-MHz band.

f. Duty factor - Scheduled as required on a continuous basis.

g. Modulation - PN spread spectrum. PSK (±90°), biphase.

h. Desired user I.D. - By beam pointing and frequency.

Return (Telemetry) Link

a. Antenna - 3.8-meter diameter parabolic reflector.

b. Antenna gain - 36 dB.

c. Frequency - 2300 MHz, users separated by frequency.

d. Bandwidth - 10 MHz.

e. Telemetry data rate - Up to 5 Mb/s.

f. Spectrum spreading - Not required by TDRSS.

g. Modulation - PSK (±90°) biphase (other modulation schemes available because TDRS is a bent pipe and IF outputs from the receiver are available at the ground station).

The required forward link PN spectrum spreading is as follows:

	EIRP	FLUX DENSITY
MA	34 dbw	- 154 dbw/M^2/4KHZ
SSA	43.4 dbw	- 154 dbw/M^2/4KHZ

The implied required chip rates (minimum) are:

MA - .6M CHIPS/SEC
SSA- 6.0 M CHIPS/SEC

3. TDRSS-HEAO-C SYSTEM CONFIGURATION

The telecommunication requirements of the HEAO-C satellite for the two TDRS system are assumed as follows:

	LOW RATE MODE	2 TDRS
Forward Link:	1 kbps	Command Channel
Return Link:	6.4 kbps	Real time telemetry
	3.2 kbps	Recorded data
	9.6 kbps	TOTAL DATA RATE

	HIGH RATE MODE	
Forward Link:	1 kbps	Command Channel
Return Link:	128 kbps	Real time experimental data

Power link margin is specified as 6db for a telemetry BER of 1 part in 10^5.

The low rate mode represents minimum requirements and could be serviced by the MA mode of the TDRS. The high rate mode would require the SSA mode of the TDRS, and allows for growth in the HEAO-C experiment package.

The forward TDRS-HEAO-C link (command channel) is selected to have no error control coding. This avoids the implementation of a decoding algorithm in the spacecraft. Power link margin is specified as 10db for a telementry BER of 1 part in 10^6.

Tables 3-1 and 3-2 are the forward link power budgets for the TDRS-HEAO-C link. They are for the multiple-access S-band case and the single-access S-band case.

The forward acquisition sequence is divided into two subsequences. They are:

(1) Acquisition on low gain antenna (HEAO-C) and reception of high gain antenna pointing commands.

(2) Acquisition on high gain antenna (HEAO-C) and reception of spacecraft commands.

Table 3-1. Calculation for Multiple-access Forward Link, S-band

BER	10^{-6}
TDRS Antenna Gain (dB)	23.0
TDRS Transmit Power (dBw)	13.0
RF Transmit Loss (dB)	-1.0
Transmitted EIRP (dBw) Peak (S+N)	35.0
TDRS Transponder Loss (db)	-1.0
Peak Signal EIRP (dBw)	34.0
Antenna Pointing Loss (dB)	0.0
Signal EIRP (dBw)	34.0
Space Loss (db)	-191.6
User Antenna Gain (dB)	G_u
Polarization Loss (dB)	-0.5
P_s - Signal Power Out of User (dBw)	$-158.1 + G_u$
T_s (Antenna Output) (^oK)	824
T_s (dB)	29.2
KT_s (dBw/Hz)	-199.4
P_s/KT_s (dB-Hz)	$41.3 + G_u$
Demodulation/Bit Sync Loss (dB)	-1.5
Demodulation Loss (PN) (dB)	-1.0
Residual Carrier Loss (dB)	0.0
Required E_b/N_o (dB-Hz) (ΔPSK)	10.8
System Margin (dB)	-M
Achievable Data Rate (dB)	$28.0-M+G_u$
Coding Gain	G_c
Achievable Data Rate (dB)	$28.0-M+G_c+G_u$

Table 3-2. Calculation for Single-access Forward Link, S-band

BER	10^{-6}
TDRS Antenna Gain (dB)	35.4
TDRS Transmit Power (dBw)	11.5
RF Transmit Loss (dB)	-2.0
Transmitted EIRP (dBw) Peak (S+N)	44.9
TDRS Transponder Loss (dB)	-1.0
Peak Signal EIRP (dBw)	43.9
Antenna Pointing Loss (dB)	-0.5
Signal EIRP (dBw)	43.4*
Space Loss (dB)	-191.6
User Antenna Gain (dB)	G_u
Polarization Loss (dB)	-0.5
Ps - Signal Power Out of User (dBw)	$-148.7 + G_u$
T_s (Antenna Output) (^oK)	824
T_s (dB)	29.2
KT_s (dBw/Hz)	-199.4
P_s/KT_s (dB-Hz)	$50.7 + G_u$
Demod/Bit Sync Loss (dB)	-1.5
Modulation Loss (PN) (dB)	-1.0
Residual Carrier Loss (dB)	0.0
Required E_b/N_o (dB-Hz) (PSK)	10.8
System Margin (dB)	M
Achievable Data Rate (dB)	34.4-M+Gu
Theoretical FEC Gain R=3, K=7 (dB)	Gc
Achievable Data Rate (dB)	34.4-M+Gc+Gu

The achievable data rates during the two acquisition phases are summarized by table 3-3

MA HEAO-C (FORWARD)

	MARGIN	Gc	Gu	DATE RATE
PHASE 1	8	0	0	100 BPS
PHASE 2	10	0	20	6.3 KBPS

SSA HEAO-C (FORWARD)

MARGIN	Gc	Gu	DATA RATE
8	0	0	436 BPS
10	0	20	27.5 KBPS

Table 3-3 Achievable Data Rates by Acquisition Phase, MA and SSA:

The data rates during the two stage acquisition for both the MA and SSA cases are selected as:

PHASE 1 - 100 BPS ANTENNA COMMANDS
PHASE 2 - 1KBPS SPACECRAFT COMMANDS

Tables 3-4 and 3-5 are the return link power budgets for the TDRS-HEAO-C link. They are for the multiple-access S-band case and the single-access S-band case. The return link (telemetry data) is selected to have error control coding. The code selected is a K=7, V=3 convolutional en-coding/soft-decision viterbi decoding. Power link margin is specified or +6db for a telemetry BER of 1 part in 10^5.

Table 3-4. Calculation for Multiple-access Return Link, S-band

BER	10^{-5}
User EIRP (dBW)	EIRP
Space Loss (dB)	-192.2
Polarization Loss (dB)	-1.0
TDRS Antenna Gain @ $\pm 13^{o}$(dB)	28.0
P_s at Output of Antenna (dBW)	-165.2 + EIRP
T_i (antenna output terminals) (^{o}K)	824
T (due to direct other user interference)	255
$K(T_s + T_i)$ (dBW)	-198.3
$P_s/K(T_s + T_i)$	+33.1 + EIRP
Transponder Loss (dB)	-2.0
Demodulation Loss (dB)	-1.5
PN Loss (dB)	-1.0
AGIPA Loss (dB)	-0.5
System Margin (dB)	-M
Required E_b/N_o (10^{-5}BER), ΔPSK	-9.9
Achievable Data Rate (dB)	18.2-M+EIRP
FEC Gain, R = 3, K = 7 (dB)	6.0
Achievable Data Rate (dB)	24.2-M+EIRP

Table 3-5. Calculation for Single-access Return Link, S-band

BER	10^{-5}
User EIRP	EIRP
Space Loss (dB)	-192.2
Pointing Loss (dB)	-0.5
Pol. Loss (dB)	-0.5
P_s at Output of Antenna (dBW)	-157.2 + EIRP
TDRS Antenna Gain (dB)	36.0 (50%)
T_i (because of direct other user interference) (^oK)	----
T_s (Antenna Output Terminals) (^oK)	824
KT_s at Output of Antenna	-199.4
P_s/KT_s	42.2 + EIRP
Transponder Loss (dB)	-2.0
Demodulation Loss (dB)	-1.5
PN Loss (dB)	0.0
Residual Carrier Loss (dB)	0.0
AGIPA Loss (dB)	0.0
System Margin (dB)	-M
Required E_b/N_0, ΔPSK	-9.9
Achievable Data Rate (dB)	25.8 + EIRP -M
FEC Gain, R = 2, K = 7 (dB)	6.0
Achievable Data Rate (dB)	31.0 + EIRP -M

The HEAO-C return link EIRP is selected to be 24.8 dbw with the antenna system selected previously (FINAL REPORT NGR-01-001-021). The achievable data rates are summarized in table 3-6.

MA HEAO-C (RETURN)

MARGIN	EIRP	DATA RATE
6 db	24.8 dbw	19.9 KBPS

SSA HEAO-C (RETURN)

MARGIN	EIRP	DATA RATE
6	24.8 dbw	229 KBPS

The power link budgets shows that the multiple-access mode is sufficient for the low-rate HEAO-C with a growth factor of two. The S-band single access mode is required for the high rate HEAO-C with a growth factor of two.

4. GOLD-CODE GENERATOR

The HEAO-C transponder will require an eleventh order gold code generator. This is required for the multiple-access mode and is selected for the single-access mode to provide range and range-rate data to the ground HEAO control.

This section gives the gold-code selection procedure, the results for synthesis of eleventh order codes, and a design of the generator.

Gold codes are a particular type of a larger group of sequences called non-maximum length. A sequence generating structure is described by its characteristic polynomial, and characteristic polynomials can be divided into subgroups as shown in figure 4-1.

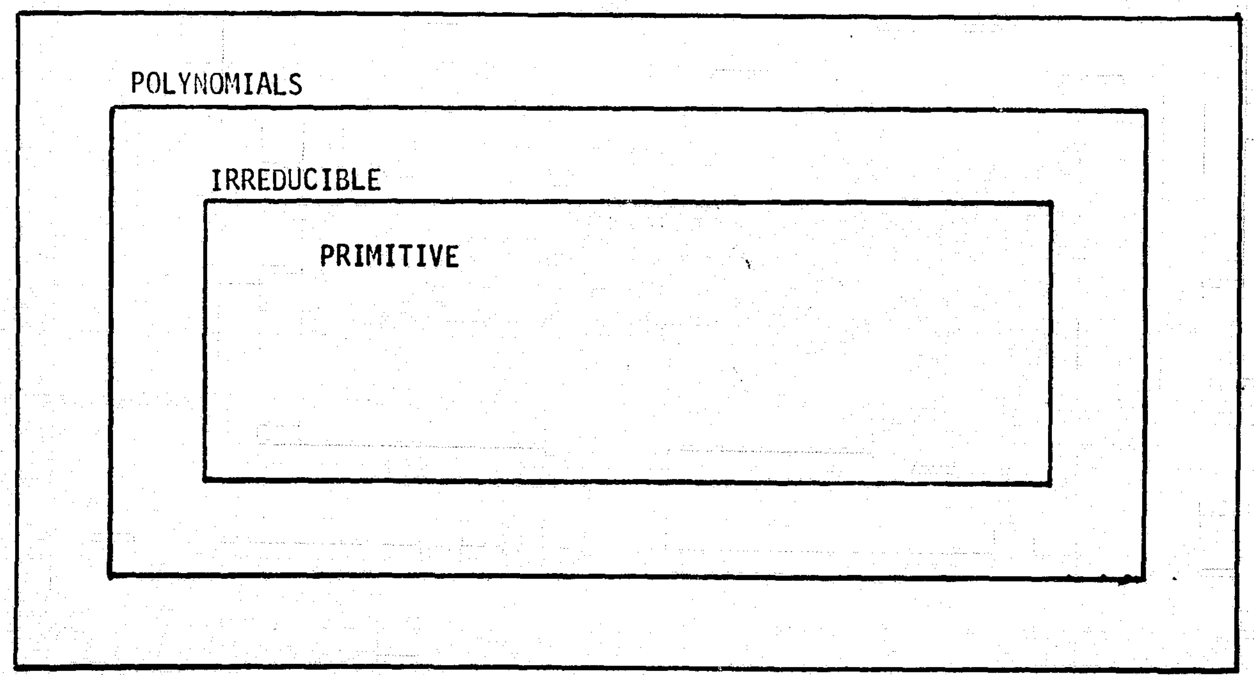


Figure 4-1 Classification of Polynomials

If the polynomial (describing a generating structure) is factorable, then the sequence depends on initial conditions and in general the sequences produced (Non ML) depends on initial conditions, and the sequences have different lengths.

If the polynomial is irreducable then all sequences out are of the same length. Example: $1+x+x^2+x^3+x^4$ gives three sequences of period 5.

If the polynomial is prime (irreducable) and primitive (maximal) then the sequences generated are maximum length.

Irreducible polynomials are tabulated in several coding references including Peterson's book on error correcting codes.

If the polynomial factors into two primitive irreducible polynomials of same order, n, then it gives 2^n+1 codes of length 2^n-1 and is a candidate gold code generator, also it gives codes of length $2(2^n-1)$.

If the polynomial factors into two primitive irreducible polynomials whose code lengths are relatively prime, then it gives 1 code of length $(2^{n_1}-1)(2^{n_2}-1)$ and is a hybrid-sum sequence.

If the polynomial factors into primitive irreducible polynomials or irreducible polynomials whose code lengths are not relatively prime, then it gives non-ML sequences of different lengths with initial condition dependence.

Examples of the different types of polynomials are as follows:

PRIMITIVE: x^4+x^3+1

IRREDUCIBLE (NOT PRIMITIVE): $x^4+x^3+x^2+x+1$

NOT IRREDUCIBLE: $x^6+x^3+x^2+x+1=(x^4+x^3+1)\ (x^2+x+1)$ NON-MAXIMAL

$x^7+x^5+x^4+x^2+1=(x^4+x^3+1)\ (x^3+x^2+1)$ HYBRID-SUM CODE

$x^{10}+x^9+x^7+x+1=(x^5+x^4+x^3+x+1)\ (x^5+x^3+1)$ GOLD CODE

As an example of the listing of irreducible polynomials, the irreducible polynomials of order 6 from Peterson are:

Polynomial (OCTAL)		Binary	Polynomial
103	*	110000100	$1+x+x^6$
127		111010100	$1+x+x^2+x^4+x^6$
147	*	111001100	$1+x+x^2+x^5+x^6$
111		100100100	$1+x^3+x^6$
015		101100000	$1+x^2+x3$
155	*	101101100	$1+x2+x^3+x^5+x^6$
007		111000000	$1+x+x^2$

*PRIMITIVE

There also exist the reverse code polynomials (not shown by Peterson).

Polynomial (OCTAL)		Binary	Polynomial
141	*	100001100	$1+x^5+x^6$
165		101011100	$1+x^2+x^4+x^5+x^6$
163	*	110011100	$1+x+x^4+x^5+x^6$
111		100100100	$1+x^3+x^6$
013		110100000	$1+x+x^3$
133	*	110110100	$1+x+x^3+x^4+x6$
007		111000000	$1+x+x^2$

Gold codes are useful in communications systems with multiple users on the same channel. With gold codes, user seperation can be achieved with code division multiplexing. Figure 4-2 illustrates a multi-station communication system as is the case in the multiple access mode of the TDRSS.

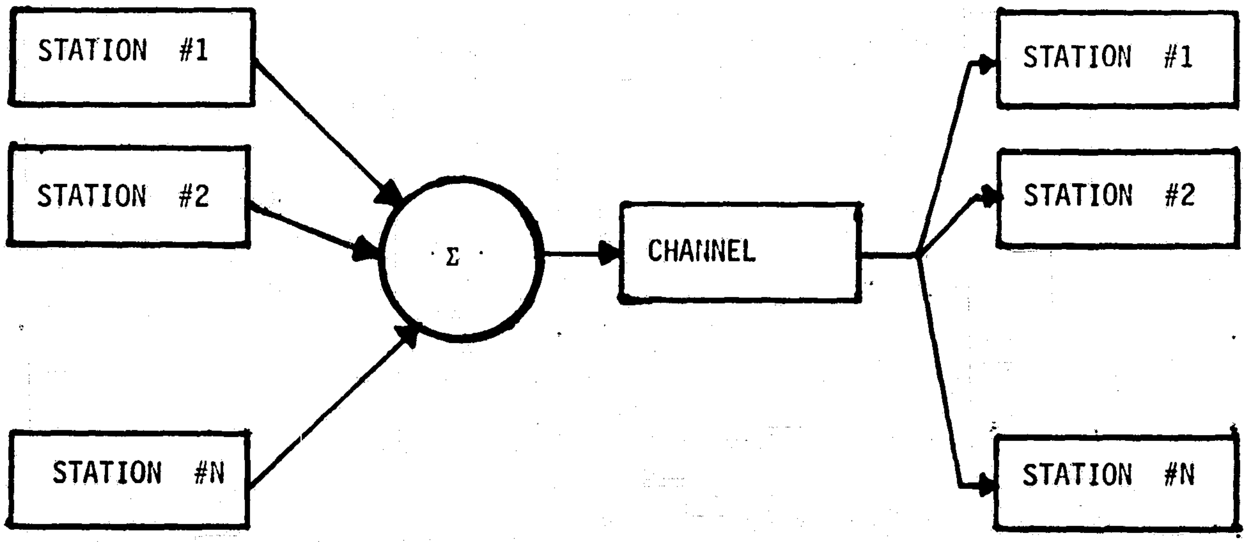


FIGURE 4-2 MULTI-STATION COMMUNICATIONS

The use of gold codes (Pseudo-Orthogonal Codes) allows effective code division multiplexing by minimizing code cross-correlation.

As a review of correlation of codes consider the two sequences:

(a) 1 1 1 0 1 0 0
(b) 1 0 0 1 0 1 1

Define correlation as $\theta ab\ (\ell) = Na - Nd$ (4-1)

where Na: No. of agreements

Nd: No. of disagreements

ℓ : Phase shift

For the codes shown, the cross-correlation as a function of code phase difference is:

$\theta ab\ (0) = -5$
$\theta ab\ (1) = +3$
$\theta ab\ (2) = +3$
$\theta ab\ (3) = -1$
$\theta ab\ (4) = +3$
$\theta ab\ (5) = -1$
$\theta ab\ (6) = -1$

The following transformation in the "logical 1" and "logical 0" digits of the code can be made:

$1 \longleftrightarrow -1$
$0 \longleftrightarrow +1$

With this transformation, the cross-correlation can be expressed.

$$\theta ab\ (\ell) = \sum_{k=0}^{L-1} a(k)b\ (k+\ell) \qquad (4\text{-}2)$$

For the example given the cross-correlation function is shown in figure 4-3:

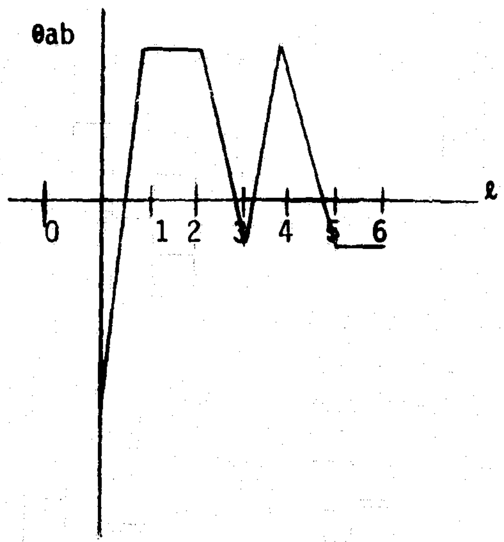


Figure 4-3 Example code cross-correlation

The code cross-correlation unbalance is calculated by integrating over all possible code cross phase positions.

$$\sum_{L = c}^{L - 1} \theta ab\ (\ell) = 1 \qquad (4\text{-}3)$$

Now a phase coded spread spectrum signal can be represented as:

$$s(\tau) + a_0 f_0(\tau) + a_1 f_0(\tau-\Delta) + a_2 f_0(\tau- 2\Delta) + \qquad (4\text{-}4)$$

$$= \sum_{i=-\infty}^{\infty} a_i f_0(\tau - i\Delta)$$

Where f is the general representation of the carrier waveform and the "a" terms are the code digits. A matched filter receiver for this particular waveform can be formed as illustrated in figure 4-4.

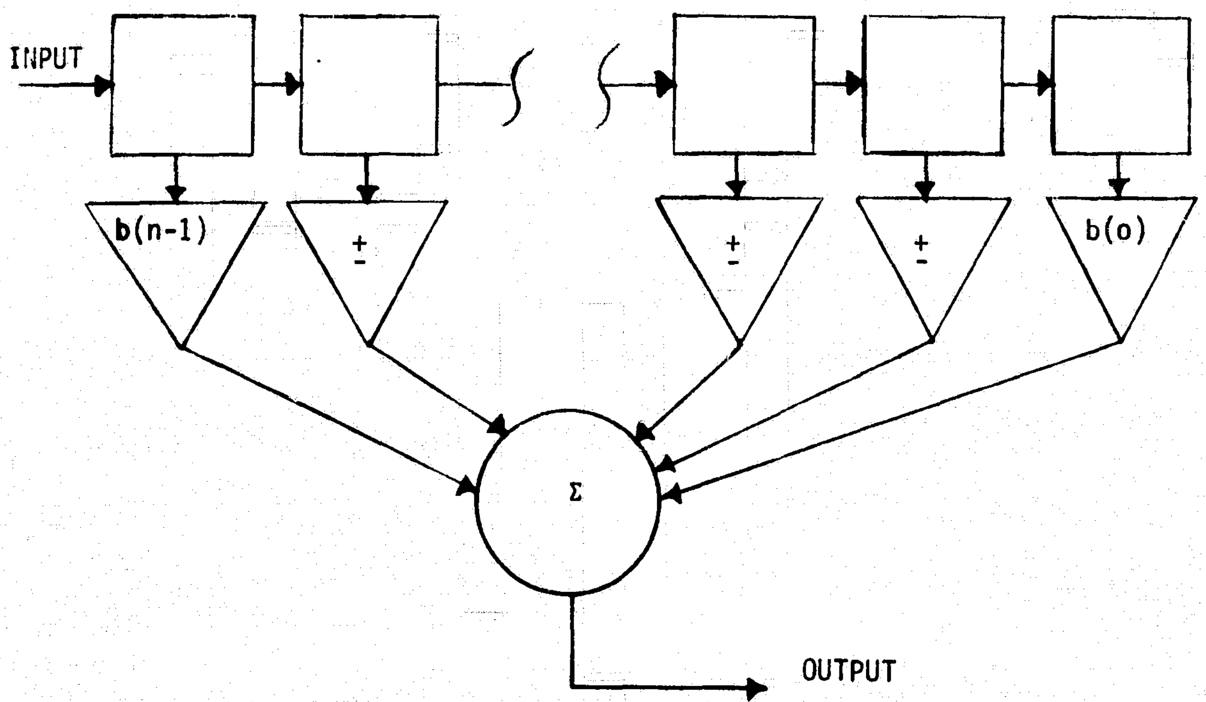


Figure 4-4 Matched filter receiver for spread spectrum waveform:

The receiver has impulse response

$$h(\tau) = b(n-1) \quad \delta(\tau-\Delta) + b(n-2)\,\delta(\tau-2\Delta) +$$

$$h(\tau) = \sum_{k=0}^{n-1} b(n-1-k)\ \delta(\tau-\Delta\{k+1\}) \qquad (4\text{-}5)$$

The receiver output for a general signal **input** $(\Delta\{\tau\})$ is

$$0(\tau) = \int_{\infty}^{\alpha} s(T)h\,(\tau-T)\ dT \qquad (4\text{-}6)$$

Now substitute $h(\tau-T)$ and $s(T)$ into (4-6)

$$0(\tau) = \int_{-\infty}^{\infty}\Big\{\sum_{i=-\alpha}^{\infty} a_i\ f_0\ (T-i\Delta) \sum_{k=0}^{n-1} b(n-1-k)\ \delta(\tau-T-\Delta\{k+1\})\quad dT \qquad (4\text{-}7)$$

Now performing the integration over T, solving for $\delta(0)$ condition,

$$T=\tau-\Delta(k+1), \qquad (4\text{-}8)$$

and obtain

$$0\ (\tau) = \sum_{i=-\alpha}^{\alpha} \sum_{k=0}^{n-1} a_i\ f_0\ (\tau-\Delta\{k+i+1\}\ )\ \ b(n-1-k) \qquad (4\text{-}9)$$

let

$$\xi = k+i+1 \qquad (4\text{-}10)$$

then

$$0(\tau) = \sum_{\zeta=-\infty}^{\infty} \sum_{k=0}^{n-1} a(\xi-k-1)\ f_o\ (\tau-\Delta\xi)\ \ b(n-1-k) \qquad (4\text{-}11)$$

now let

$$\zeta = n-1-k \qquad (4\text{-}12)$$

then

$$0(\tau) = \sum_{\zeta=-\infty}^{\alpha} \sum_{\zeta=0}^{n-1} a(\xi+\zeta-n)\ b(\zeta)\ f_0\ (\tau-\Delta\xi) \qquad (4\text{-}13)$$

Now, the period of the code is n, so the waveform $s(\tau)$ is cyclic over n,

$$0(\tau) = \sum_{\zeta=-\infty}^{\infty} \sum_{\zeta=0}^{n-1} a\xi+\zeta)\ b(\zeta)\ f_0\ (\tau-\Delta\xi) \qquad (4\text{-}14)$$

now

$$\sum_{J=0}^{n-1} a(\xi+\zeta)\ b(\zeta) = \theta ab\ (\xi) \qquad (4\text{-}15)$$

which is the cross-correlation function.

then

$$O(\tau) = \sum_{\zeta=-\infty}^{\alpha} \theta ab(\xi)\ f_0\ (\tau-\Delta\xi) \qquad (4\text{-}16)$$

and the output of the matched filter receiver depends on the cross-correlation between codes "a" and "b".

If "a" and "b" are the same ML sequence, the output is small except when $\xi=0$, or ξ is an integer multiple of n where

$$n=2^n -1 \qquad (4\text{-}17)$$

and where N is the order of the code. For example if N=12, N=4095 and

$$O_{MAX}\ (\tau) = 4095\ f_0\ (\tau-\Delta\xi) \qquad (4\text{-}18)$$

now if "a" and "b" are not the same ML sequence $\theta ab\ (\xi)$ can be large. For the case N = 12 for "a" and "b", but "a" and "b" not both same sequence, $\theta ab\ (\xi)$ can be as large as 1400. This is a large cross correlation for codes that are to be used in code division multiplexing. Now consider codes generated from multiple polynomials as shown in figure 4-5.

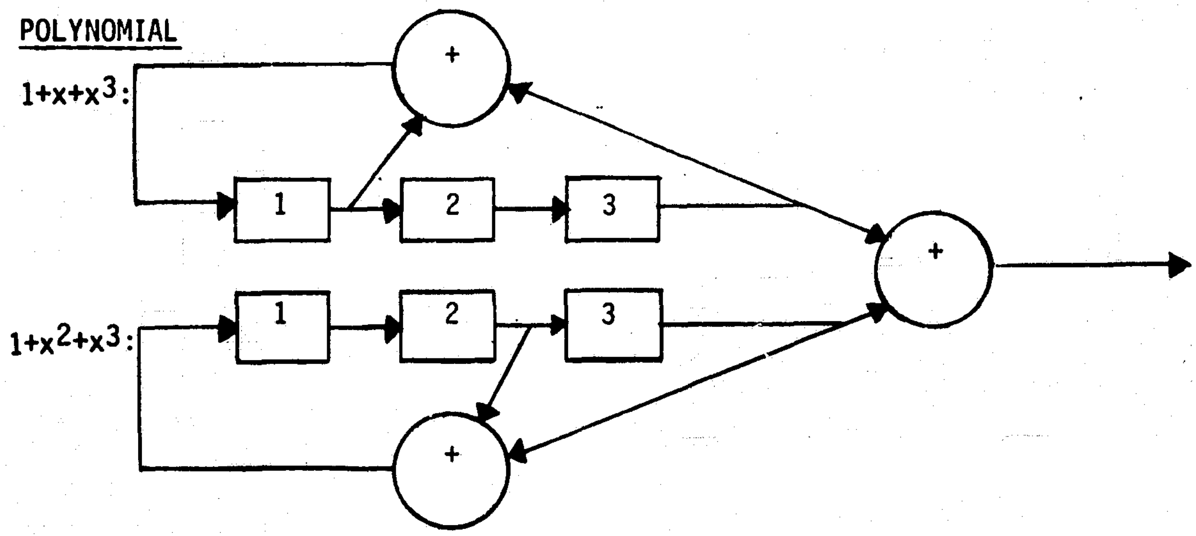
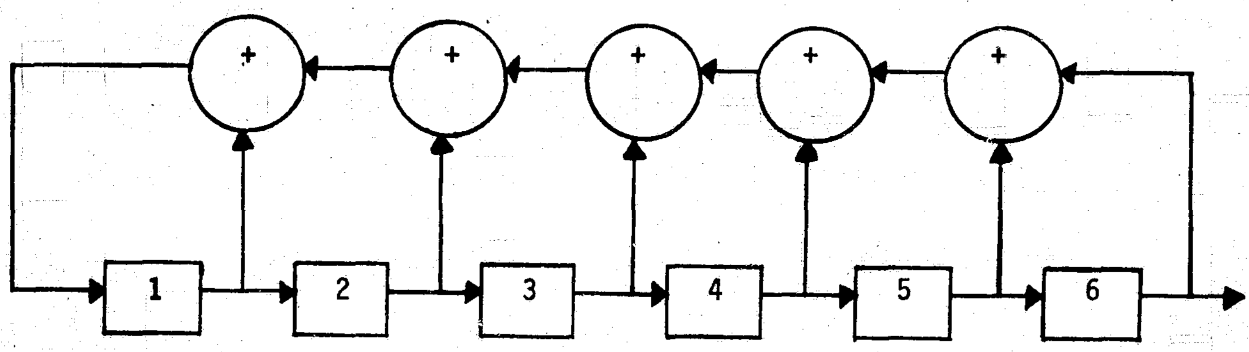


Figure 4-5 Multiple polynomials generating structure. An equivalent generating structure can be found as shown in figure 4-6.

$$(1+x+x^3)\ (1+x^2+x^3) = 1+x+x^2+x^3+x^4+x^5+x^6 \qquad (4\text{-}19)$$



The output sequence depends on initial conditions. Figure 4-6 equivalent generating structure. Now consider the example shown in figure 4-7.

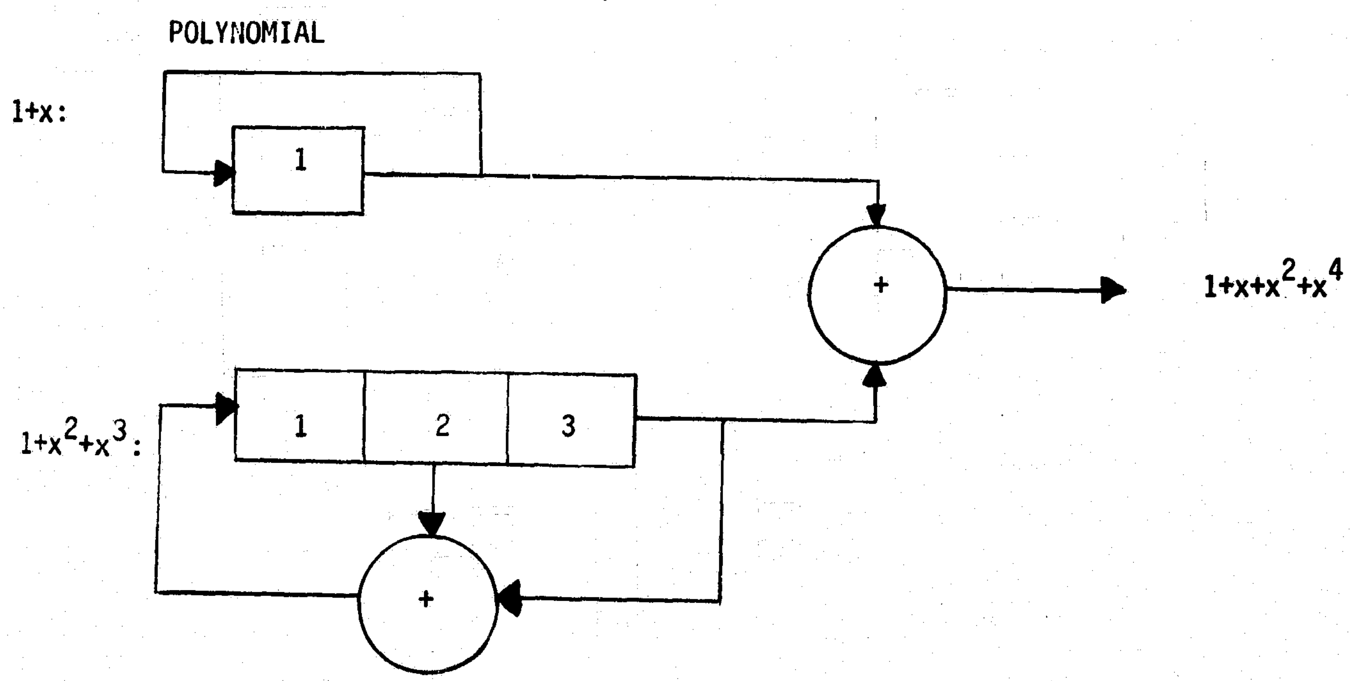


FIGURE 4-7 CODE GENERATING STRUCTURE

If the shift register polynomial has (1+x) as a factor, then if it generates sequence "a" it will also generate $\bar{a}$ (the compliment of "a"). This is obvious because (1+x) will generate a "1" or "0" all the time depending on initial conditions.

If sequences "a" and "b" can be generated in a shift register, then $a \oplus b$ can also be generated.

$Ia \rightarrow a$ (initial conditions Ia generates a)
$Ib \rightarrow b$ (initial conditions Ib generates b)

Then $Ia \oplus Ib \rightarrow a \oplus b$

Now further consider the cross correlation between two sequences of period n:

$$\theta ab\ (\ell) = Na-Nd$$
$$n = Na+Nd$$

where $n=2^N-1$ <u>or</u> (4-20)

$\theta ab\ (\ell) = n-2Nd$ (4-21)

where Nd is the "Hamming distance."

The question in optimizing a multiple access code division mutiples system is "How do you pick pairs of ML codes with minimum θab (ℓ)"? Consider the basic theorem from error correcting code study.

<u>Theorem</u>: Let α be any primitive 2^N-1 root of unity. Let f_i be the minimal polynomial of α.

Let

$$g_k(x) + \frac{1 + X^{(2^N-1)}}{f_1(x) \times f_2(x) \times f_3(x) + \cdots \times f_k(x)} \tag{4-22}$$

where there are no repeats in the f terms then

$a, b \in V(g_k) \Rightarrow \|a+b\| > k$

where

$V(f)$: The set of all sequences $\|a+b\| = Nd$: The hamming distance

α^i is a coset group, i is the lable from the table $f_i(x)$ is the polynomial with lable i. The table below gives an example of coset groups and equivalent lables (5th degree)

1	2	4	8	16	
3	6	12	24	17	
5	10	20	9	18	MOD 31
15	30	29	27	23	
7	14	28	25	19	
11	22	13	26	21	

As an example of the use of the theorem for 5th degree codes select k=5 (arbitrary). Then

$$\frac{1+x^{31}}{P_1P_2P_3P_4P_5} = \frac{1+x^{31}}{P_1P_3P_5} \tag{4-23}$$

Where repeats from the coset group table have been eliminated.

IF $a,b \in V\left(\frac{1+x^{31}}{P_1P_3P_5}\right)$ $\|a+b\| = Nd>5$

The hamming distance is greater than 5. As a second example select $k=30$. Then

$$g_{30}(x) = \frac{1+x^{31}}{f_1 f_2 \quad f_{30}} = \frac{1+x^{31}}{f_1 f_3 f_5 f_{11} f_{15} f_7} \tag{4-24}$$

Where all coset repeats are eliminated (4-25)

$$g_{30}(x) = 1+x$$

because

$$(f_1 f_3 f_5 f_{11} f_{15} f_7)(1+x) = 1+x^{31} \tag{4-26}$$

$a,b \varepsilon\ (1+x) \qquad \| a+b \| > 30$

Since a: sequence of 31 "1"
b: sequence of 31 "0"

it is seen that $Nd = 31>30$ (4-27)

The maximum value of the code cross-correlation function can be bounded by use of the following theorem:

Theorem: If $a,b \varepsilon\ V(g_k) \qquad |\theta ab| < 2^N-1-2k$

Proof: If $a,b \varepsilon\ V(g_k)$
then $a+b \varepsilon\ V(g_k)$ (because $I_a \to a$, $Ib \to b$, $I_a + Ib \to a+b$)
and $\overline{a+b} \varepsilon\ V(g_k)$ (because $(1+x)$ is a factor of $g_k(x)$)

from the first theorem

$\| a+b \| > k \qquad\qquad\qquad \| a+b \| > k$
or

$\| a+b \| = (\{2^N-1\} - \| a+b \|) > k$

$a+b \quad < (2^N-1) - k \qquad\qquad\qquad \| a+b \| > k$

$k < \| a+b \| < (2^N-1) - k$

$-2k > -2 \| a+b \| > -2(2^N-1) + 2k$

$-\{(2^N-1) - 2k\} < (2^N-1) - 2 \| a+b \| < (2^N-1) - 2k\}$

OR

$|(2^N-1) - 2 \| a+b \| | < 2^N-1-2k$

OR

$$|\theta ab| \quad 2^N-1-2k \tag{4-28}$$

This theorem gives the method for selecting g_k which generates sequences a, b, ··· such that the cross-correlation function is bounded.

As an example consider the case N=5, $2^N-1=31$, k=4, for this case

$|\theta ab| < 31-8=23$

and

$$g^4 = \frac{1-x^{31}}{P_1P_2P_3P_4} = \frac{1+x^{31}}{P_1(x)\ P_3(x)}$$

$$= (1+x)\ \frac{P_1P_3P_5P_{15}P_7P_{11}}{P_1P_3} = (1+x)\ P_5P_{15}P_7P_{11}$$

$$= (1+x)\ (1+x+x^2+x^4+x^5)\ (1+x^3+x^5)\ (1+x+x^2+x^3+x^5)\ (1+x+x^3+x^4+x^5)$$

A polynomial of degree 21, as a second example consider the case N=5, k=6:

$|\theta ab| < 31-12+19$

$$g_6 = \frac{1+x^{31}}{P_1P_3P_5} = \frac{(1+x)\ P_1P_3P_5P_{15}P_7P_{11}}{P_1P_3P_5} = (1+x)\ P_{15}P_7P_{11}$$

A 16th order polynomial, as a third example consider the code N=5, k=10:

$|\theta ab| < 31-20+11$

$$g_{10} = \frac{(1+x)\ P_1P_3P_5P_7P_{11}P_{15}}{P_1P_2P_3P_4P_5P_6P_7P_8P_9P_{10}} = (1+x)\ P_{11}P_{15}$$

where coset repeats in the denomination have been eliminated. Actually $|\theta ab| \leq 9$ for this case because $|\theta ab|$ cannot be an even number for odd code lengths.

We now have a procedure for selecting polynomials with bounded cross-correlation, that is:

(1) Form

$$g_k = \frac{(1+x)\ P_1P_3P_5P_{11}P_{15}P_7}{P_1P_2 \cdots P_k}$$

the (1+x) Factor just produces compliment sequences

(2) Take k as large as possible leaving one pair in the numerator

$g_{10} = (1+x)\ P_{11}P_{15}$

(3) Then $|\theta ab|$ is optimally bounded. For example

$|\theta ab| < 11$ (actually $|\theta ab| \leq 9$ since θab is not even)

If we let

a: P_{11}

b: P_{15}

Then $|\theta ab| \leq 11$

The value of k such that one pair is letf gives minimum cross correlation. The pair of remaining polynomials are the prefered pair.

This is the basis of gold codes, developed by Robert Gold. The following theorem is the form usually seen in discussions of gold codes.

<u>Theorem</u> Let f_1 (x) be a primitive polynomial of degree N

Let x^1 (The 1 is the lable) be a root of f_1 (x)

Let f_2 (x) be the irreducible polynomial such that

$2^{\frac{n-1}{2}}+1$ is the root of f_2 (x) for N-odd

$2^{\frac{n-2}{2}}+1$ is the root of f_2 (x) for N-even

Then if a and b are sequences such that

a: generated by $f_1(x)$

b: generated by f_2 (x)

Then

$$|\theta ab| < \begin{cases} 2^{\frac{N+1}{2}} + 1 & N\text{ - Odd} \\ 2^{\frac{N+2}{2}} + 1 & N\text{ - Even} \end{cases}$$

For example consider the polynomial

$$f_1\ (x) = 1+x^2+x^5$$

$$2^{\frac{N-1}{2}}+1 = 5$$

$$f_5\ (x) = 1+x+x^2+x^4+x^5$$

If a: $f_1\ (x)$

b: $f_5\ (x)$

Then $|\theta ab| \leq 9$

If it were required to use label 7 from the 5th degree coset group that could be done as follows:

$7\ (1,5) + (7,35) \rightarrow (7,4) \rightarrow (7,1)$

and polynomials with 5th degree coset lables 1 and 7 become the preferred pair. Other possible preferred pairs of this degree are:

$(1,5) \rightarrow (3,15)$

$(1,5) \rightarrow (11,55) \rightarrow (11,24) \rightarrow (11,3)$

$(1,5) \rightarrow (15,75) \rightarrow (15,13) \rightarrow (15,11)$

There are 2^N+1 different codes in the psendo orthogonal code group for a preferred pair of polynomi ls of degree N.

To illustrate gold code cross correlation for the 5th degree preferred pair $f_{11}f_{15}$ a demonstration system was constructed. Now

$$f_{11}f_{15} = (x^5+x^4+x^3+x+1)\ (x^5+x^3+1) = x^{10}+ x^9+x^7+x+1$$

and the generating structure is shown in figure 4-8.

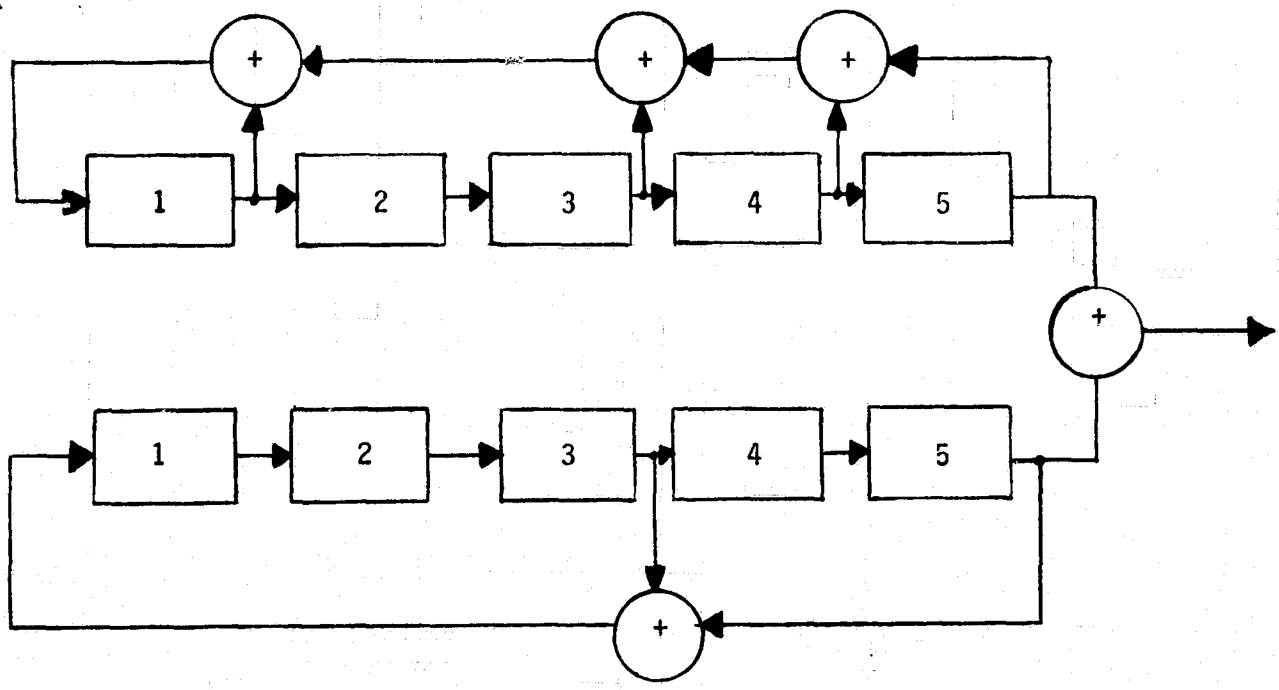


Figure 4-8 Generating structure for 5th degree preferred pair with coset lables 11 and 15

Two of the generators of the form shown in figure 4-8 were contructed along with a cross-correlator. Each generator wasdriven by a separate clock signal in the system shown in figure 4-9.

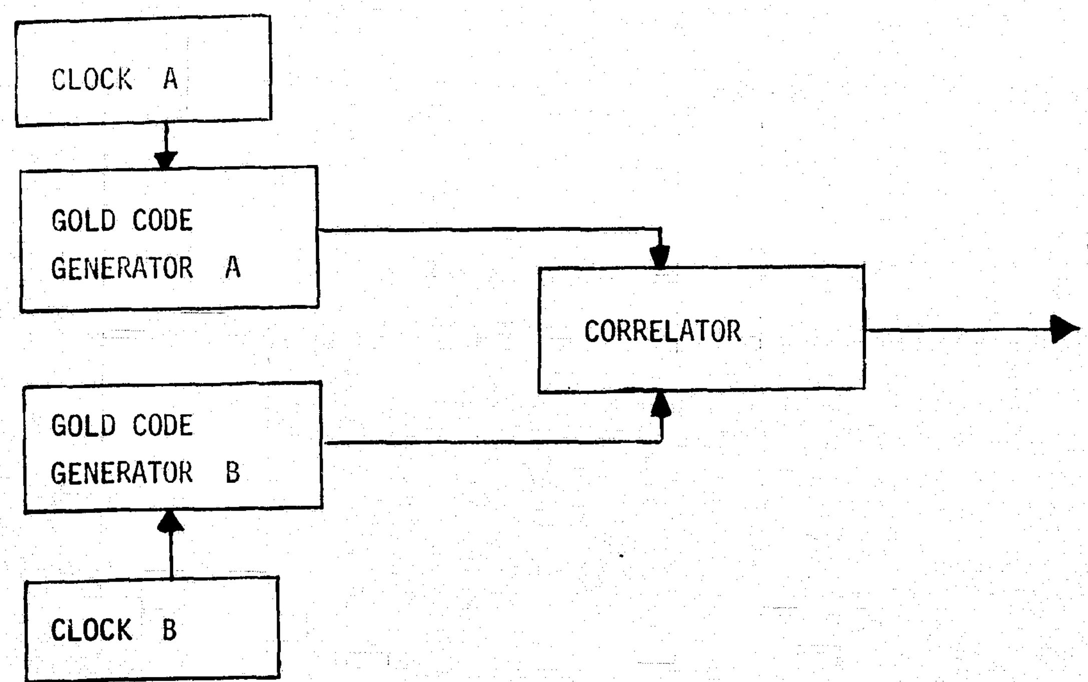


Figure 4-9 Gold Code generators and correlator

Figure 4-10 is a photograph of an oscilloscope trace of the output of the correlator, and shows an example cross-correlation trace (lower) with a code auto-correlation trace (upper). The auto-correlation function has peak value of 31 (degree 5 code). The gold cross-correlation is bounded by

$$|\theta ab| \leq 9 \quad \text{and} \quad \frac{|\theta aa|MAX}{|\theta ab|MAX} = 10.7 \text{ db}$$

and, as seen in the trace, actually takes an the values 7,-1, and -9.

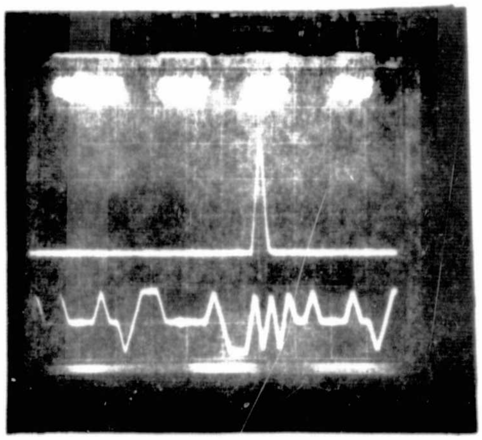
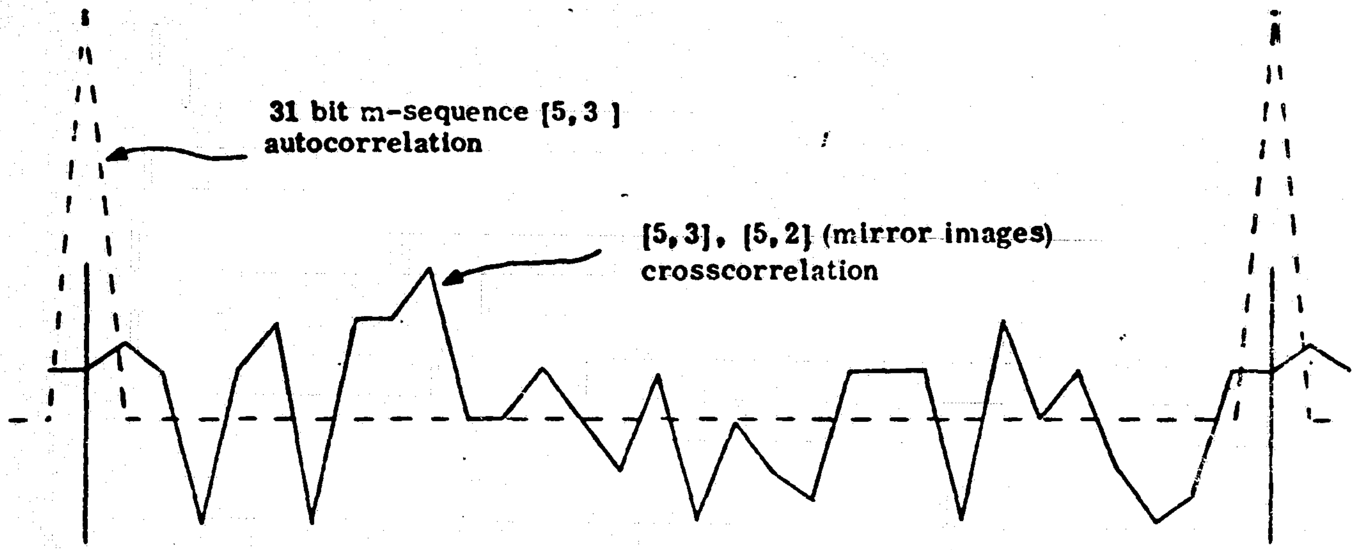


Figure 4-10 gold code auto and cross correlation waveforms.

The results shown in figure 4-10 for the cross-correlation between 5th order gold codes can be compared to the cross-correlation between maximum length 5th order sequences. Figure 4-10B shows the calculated cross-correlation for the codes (5,3,0) and (5,2,0). These are mirror image m-sequences. The cross-correlation advantage of the Gold code is 1.7db. More pronounced cross-correlation or code interference advantages are found for higher order gold codes. For example for 12th order codes $|\theta ab|MAX$ can be as large 1400 for "a" and "b" being ML codes. The equivalent gold family has $|\theta ab|MAX < 129$ which gives a cross-correlation advantage of 20.7db.

FUGURE 4-10B • **Comparative autocorrelation and crosscorrelation for 31 bit mirror image m-sequences.**



Shift		Agreements		Disagreements		A-D	
0	17	17	15	14	16	3	-1
1	18	18	13	13	18	5	-5
2	19	17	12	14	19	3	-7
3	20	11	17	20	14	-9	3
4	21	17	17	14	14	3	3
5	22	19	17	12	14	7	3
6	23	11	11	20	20	-9	-9
7	24	19	19	12	12	7	7
8	25	19	15	12	16	7	-1
9	26	21	17	10	14	11	3
10	27	15	13	16	18	-1	-5
11	28	15	11	16	20	-1	-9
12	29	17	12	14	19	3	-7
13	30	15	17	16	14	-1	3
14	31	13	17	18	14	-5	3
15		17		14		3	
16		11		20		-9	

TDRS user guidelines have specifed that multiple access users will share the TDRS MA channel by code division multiplexing, and that SSA channels will be PN spread spectrum at least on the forward link. In the case of the MA channel the code family for code division multiplexing has been selected as a Gold code group. Each MA user will be assigned a unique member of this family. TDRS user guidelines suggest that this code will be approximately 2000 bits in length. For the purpose of this design study, an 11th order gold code generator was selected. This generating structure is capable of producing a family of 2049 pseudo-orthogonal codes of length 2047 bits. The gold codes in the family will have cross correlation limited by $|\theta ab| < 2^{\frac{N+1}{2}} +1 =65$ and a jamming immunity to other MA channel user of 20 log (2047/65)= 30db.

There are 176 primitive eleventh degree polynomials. The size of the coset table would be 11x176 members. The lables for the preferred pair of polynomials can be calculated by using the previous theorem. A preferred pair would be primitive polynomials with labes 1 and $2^{\frac{N+1}{2}} +1$, or lables 1 and 33. The primitive polynomials for these two coset lables are:

1 : $x^{11}+x^2+1$ (4005 OCTAL)

33: $x^{11}+x^{10}+x^9+x^7+x^6+x^4+x^3+x^2+ 1$ (7335 OCTAL)

The characteristic polynomial for the code generating structure is

$$x^{22}+x^{21}+x^{20}+x^{18}+x^{17}+x^{15}+x^{14}+x^{12}+x^{11}+x^{10}+x^8+x^7+x^5+x^3+ 1$$

and the particular gold code generated would depend on initial loading of the generating register.

The gold code generator design is composed of the following parts:

(1) Generating register
(2) Code feedback logic
(3) Initial loader
(4) Code generation monitor

The generating register includes twenty-two storage stages, and the code feedback logic is designed to implement the given characteristic polynomial. Since the gold code generated depends on initial loading of the register, and since a unique gold code will be assigned in the MA user configuration, an initial loader will load a word into the register to insure the generation of the proper code. A code generation monitor will track the code being generated and make sure the proper gold code is being generated during operation of the transponder.

Figure 4-11 is an overall block diagram of the gold code generation. The operation of the code initial loading and reload logic is as follows:

<u>Initial Load</u>

1. SET ÷ 2047 TO ALL ZERO WORD
2. SET GENERATING REGISTER TO INITIAL CODE WORD
3. START GENERATOR

<u>Reload</u>

1. HALT GENERATOR
2. SET ÷ 2047 TO ALL ZERO WORD
3. SET GENERATING REGISTER TO INITIAL CODE WORD
4. START GENERATOR

The reload sequence is initialized by the occurrence of two ÷ 2047 count pulses in the sequence with no code word correlation pulse occuring during this period.

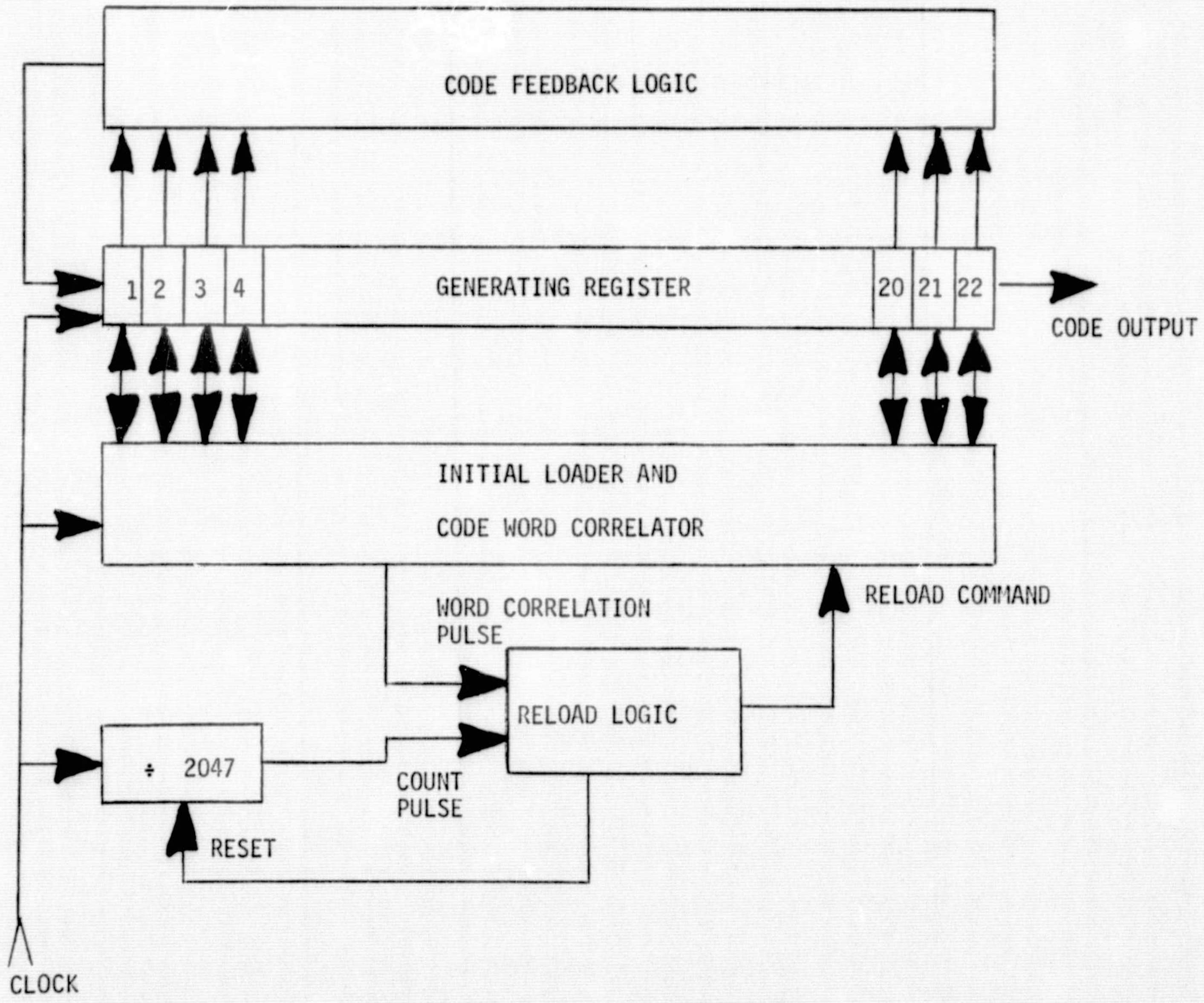


FIGURE 4-11 GOLD-CODE GENERATOR

Figure 4-12 A and 4-12 B are the electrical schematics of the code feedback logic and the generating register. The unit uses internal feedback which limits the gate delay problem that would exists if the characteristic polynomial $x^{22}+x^{21}+x^{20}+x^{18}+x^{15}+x^{14}+x^{12}+x^{11}+x^{10}+x^{7}+x^{5}+x^{3}+1$ were implemented with external configuration. This would result in eleven (11) gate delays in the feedback logic.

Figure 4-13 is the code word correlator and initial loader. Figure 4-14 is the ÷ 2047 network and the reload logic. The following symbols used in the drawings are defined as follows:

F - CLOCK LINE

C - GENERATOR RUN COMMAND LINE

D - GENERATOR RELOAD COMMAND LINE

L - GENERATOR RELOAD LINES TO BE CONNECTED TO T OR U LOAD LINES DEPENDING ON WORD TO BE LOADED

B - INTERNAL FEEDBACK LINE

P - WORD CORRELATION PULSE LINE

Y - 2047 COUNT PULSE LINE

R - RELOAD COMMAND LINE

The C-line is connected to the U-line and the D-line is connected to the T-line.

The reload and internal feedback feature make this generator design safe for the high spread transponder for spectrum spreading and range and range-rate tracking.

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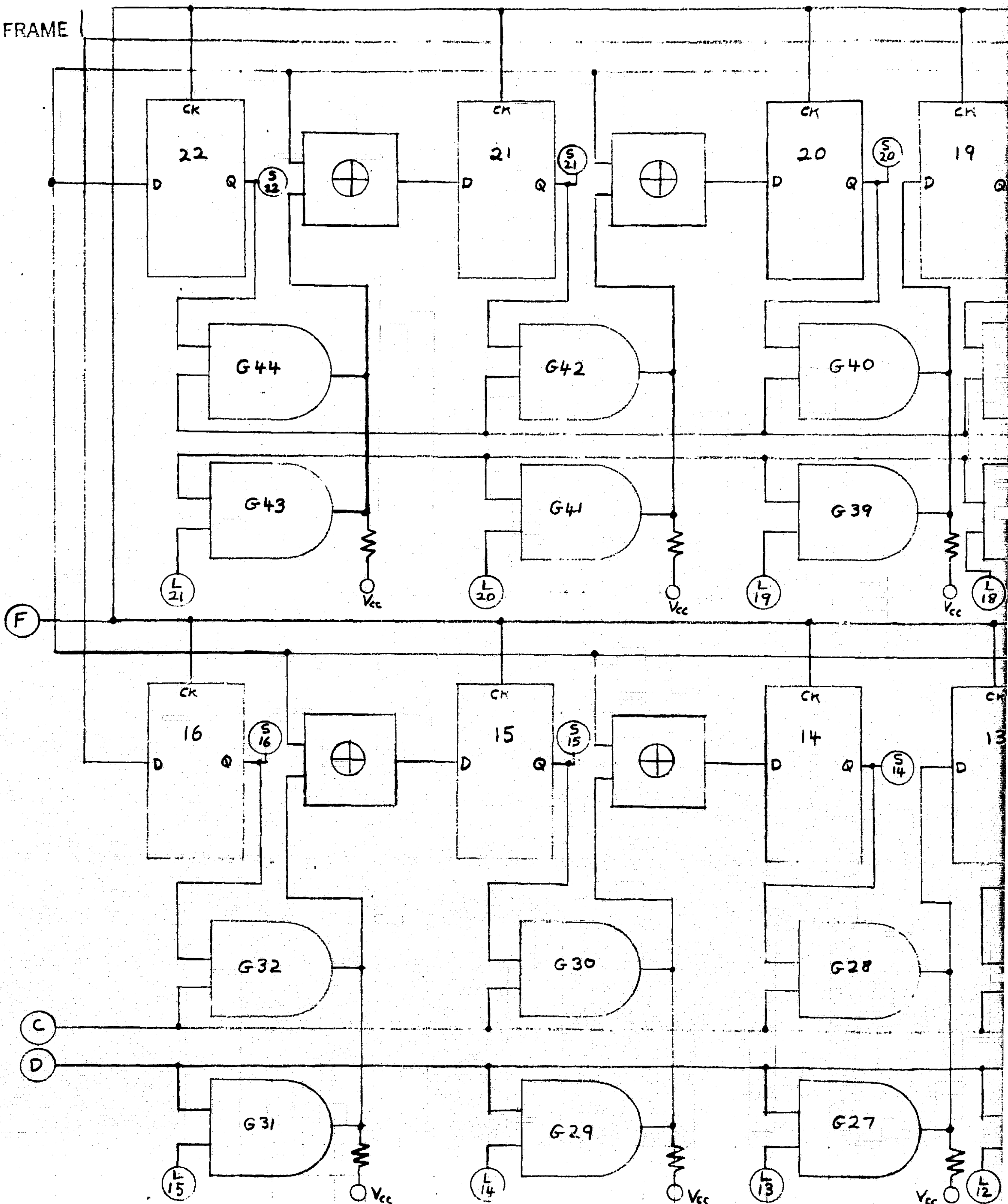

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Fig. 4-12A Code Feedback Logic and Generating Register

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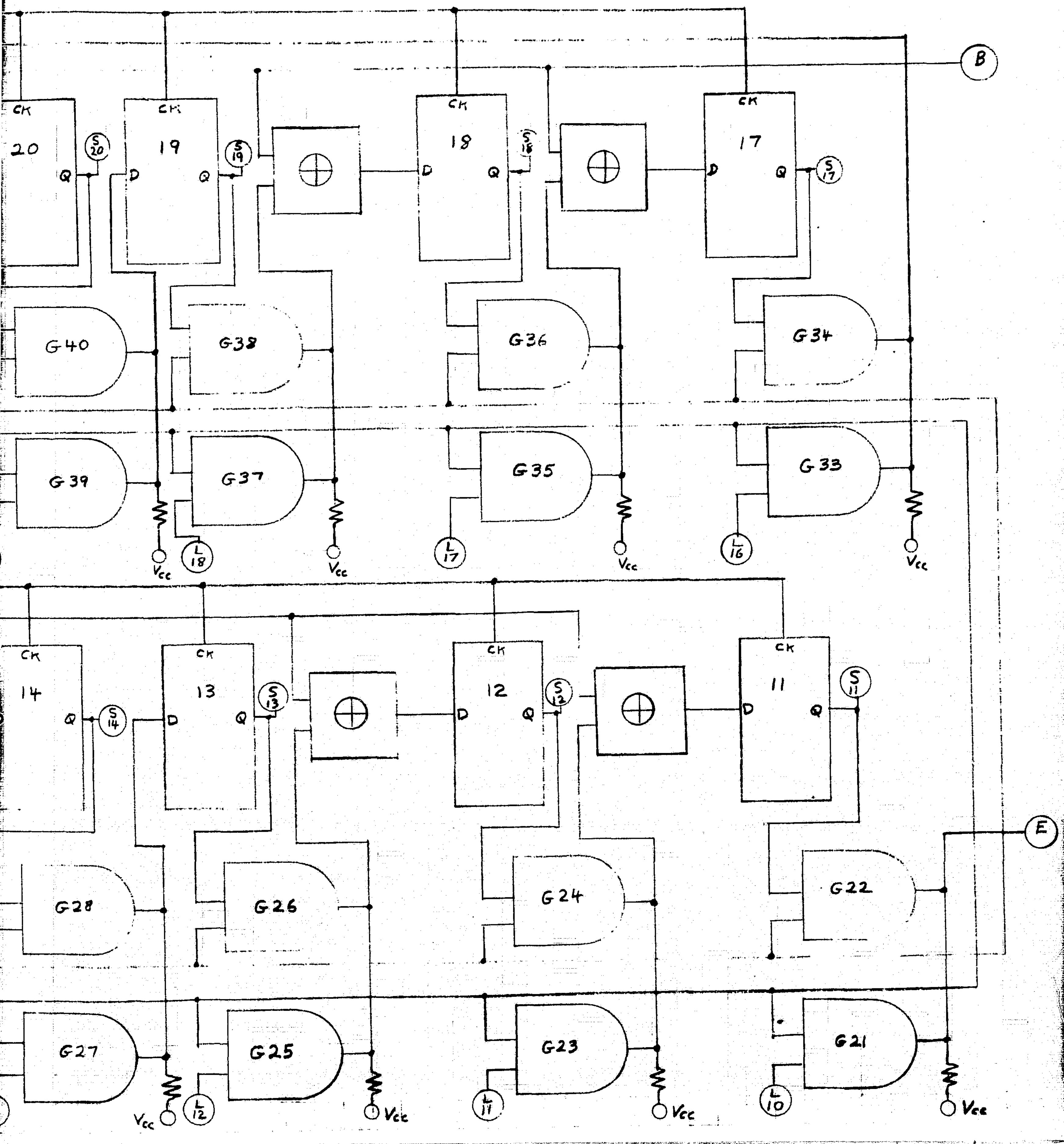

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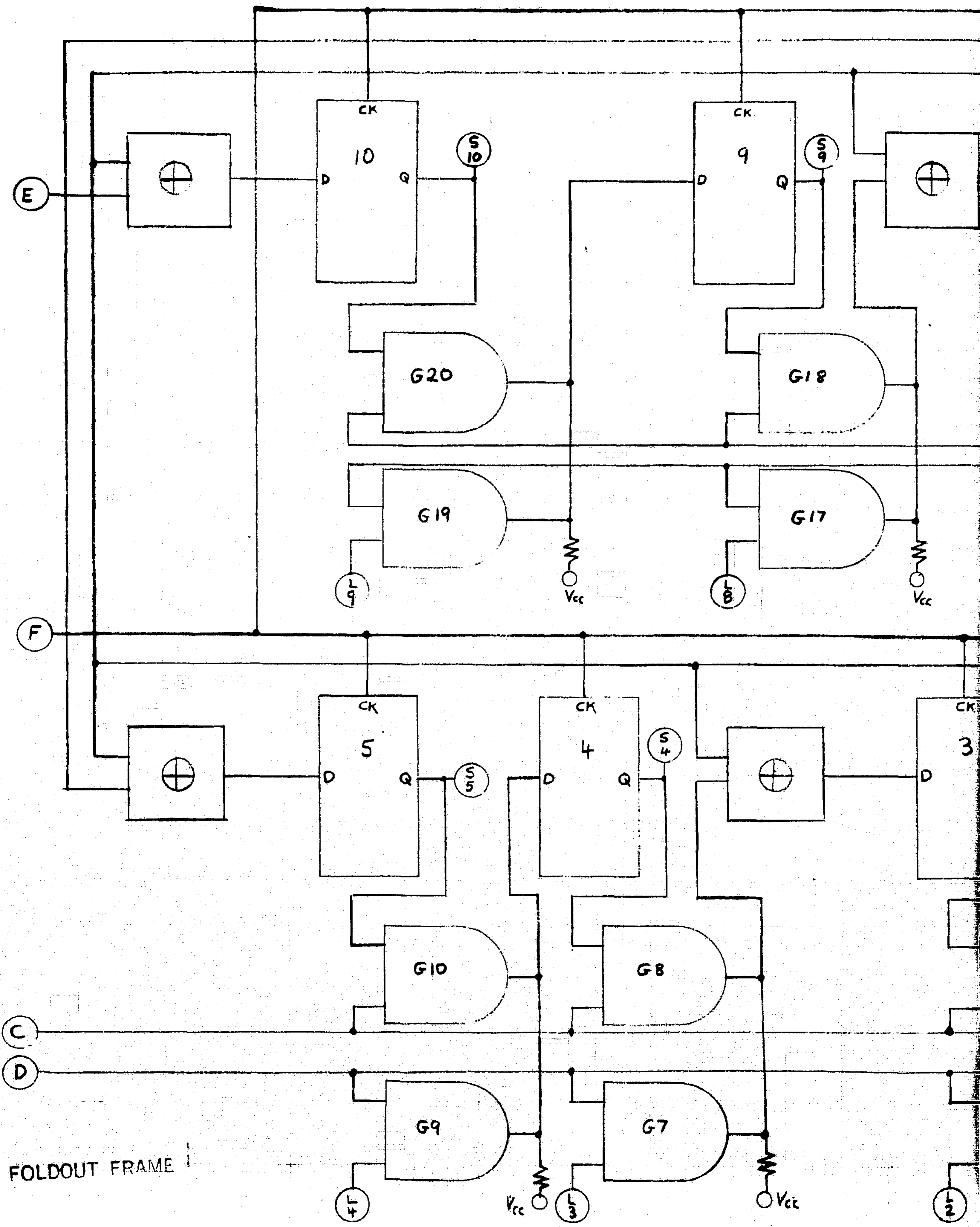

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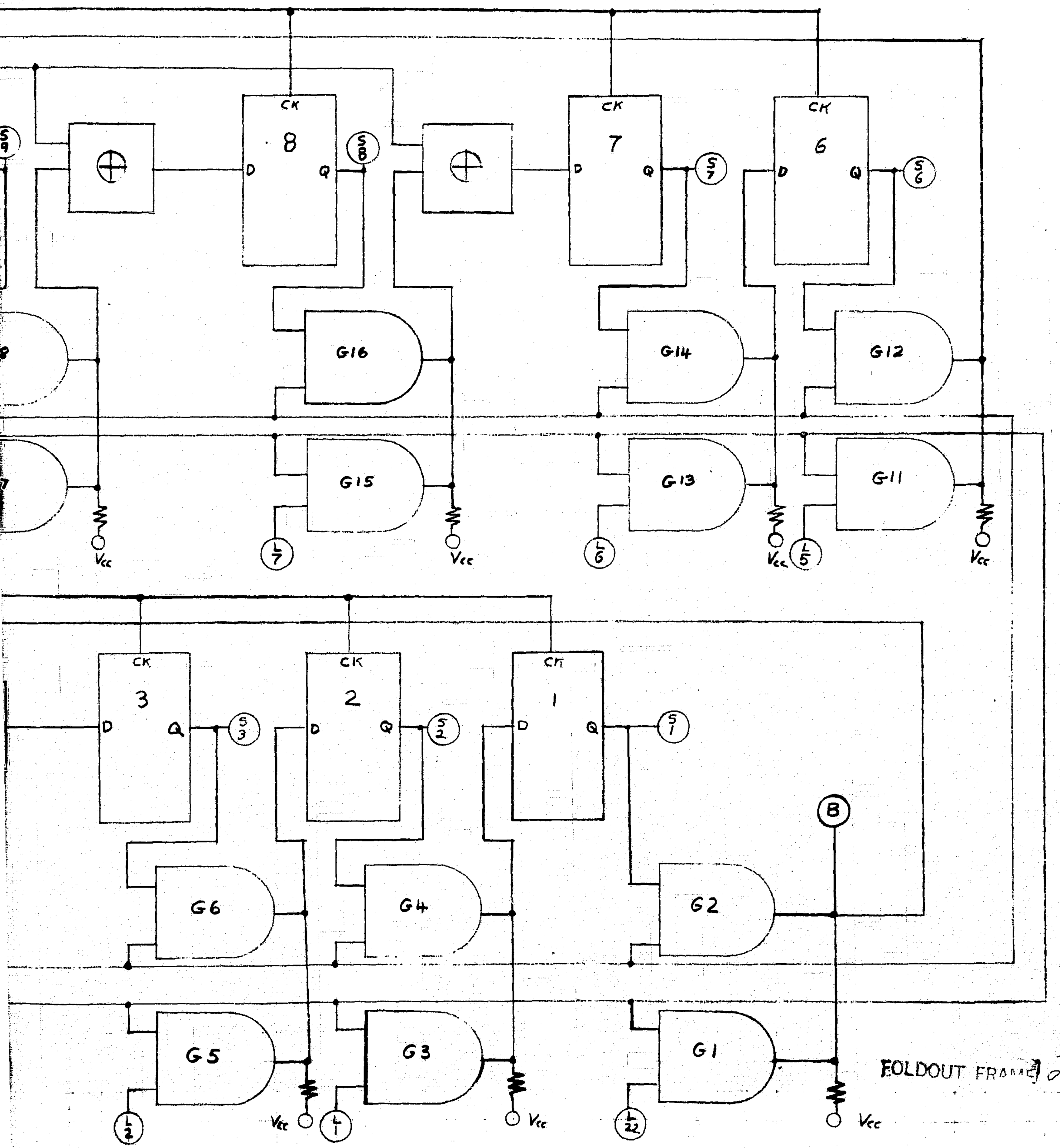
Fig. 4-12B Code Feedback Logic and Generating Register



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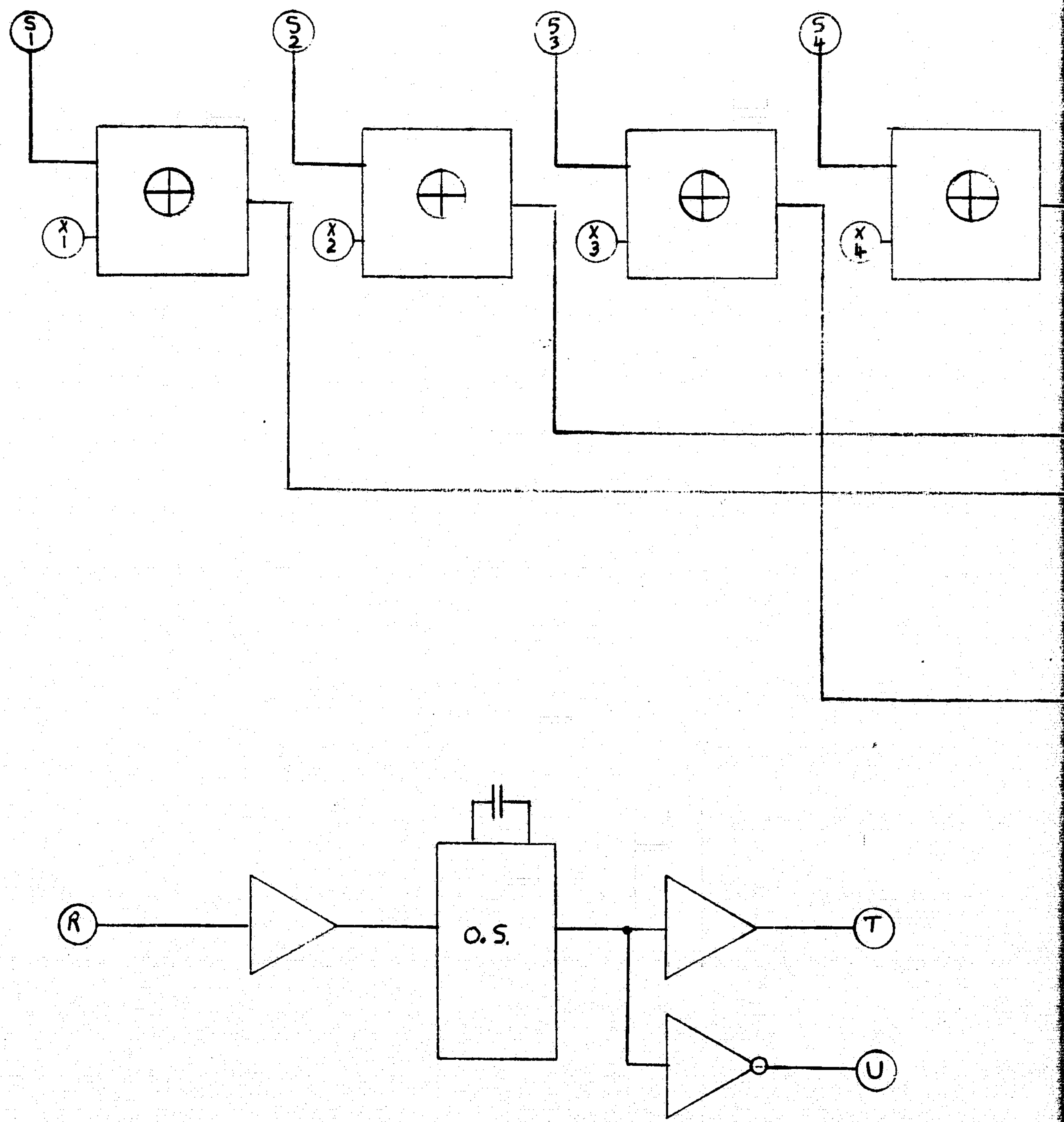

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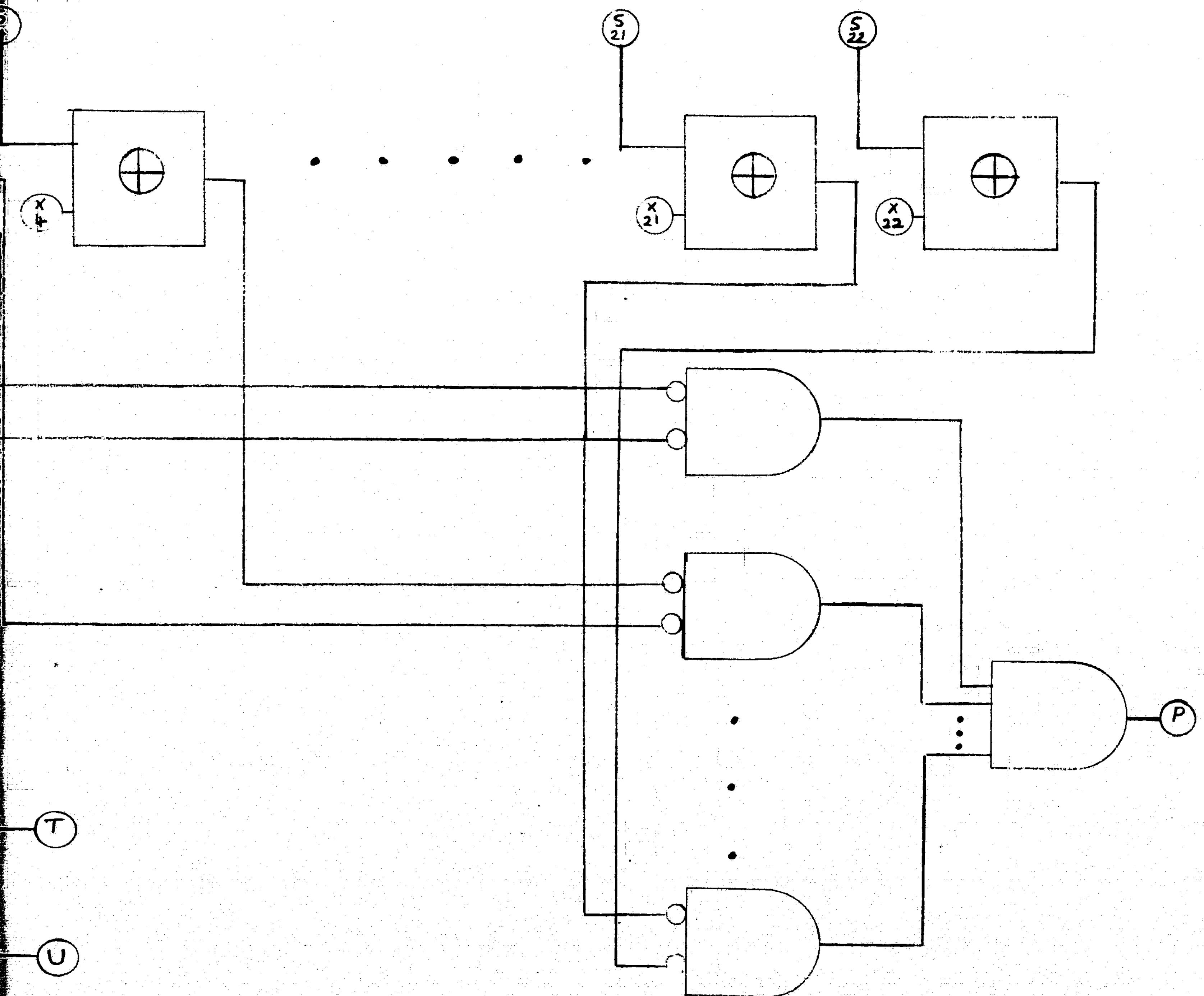
Fig. 4-13 Code Word Correlator and Initial Loader



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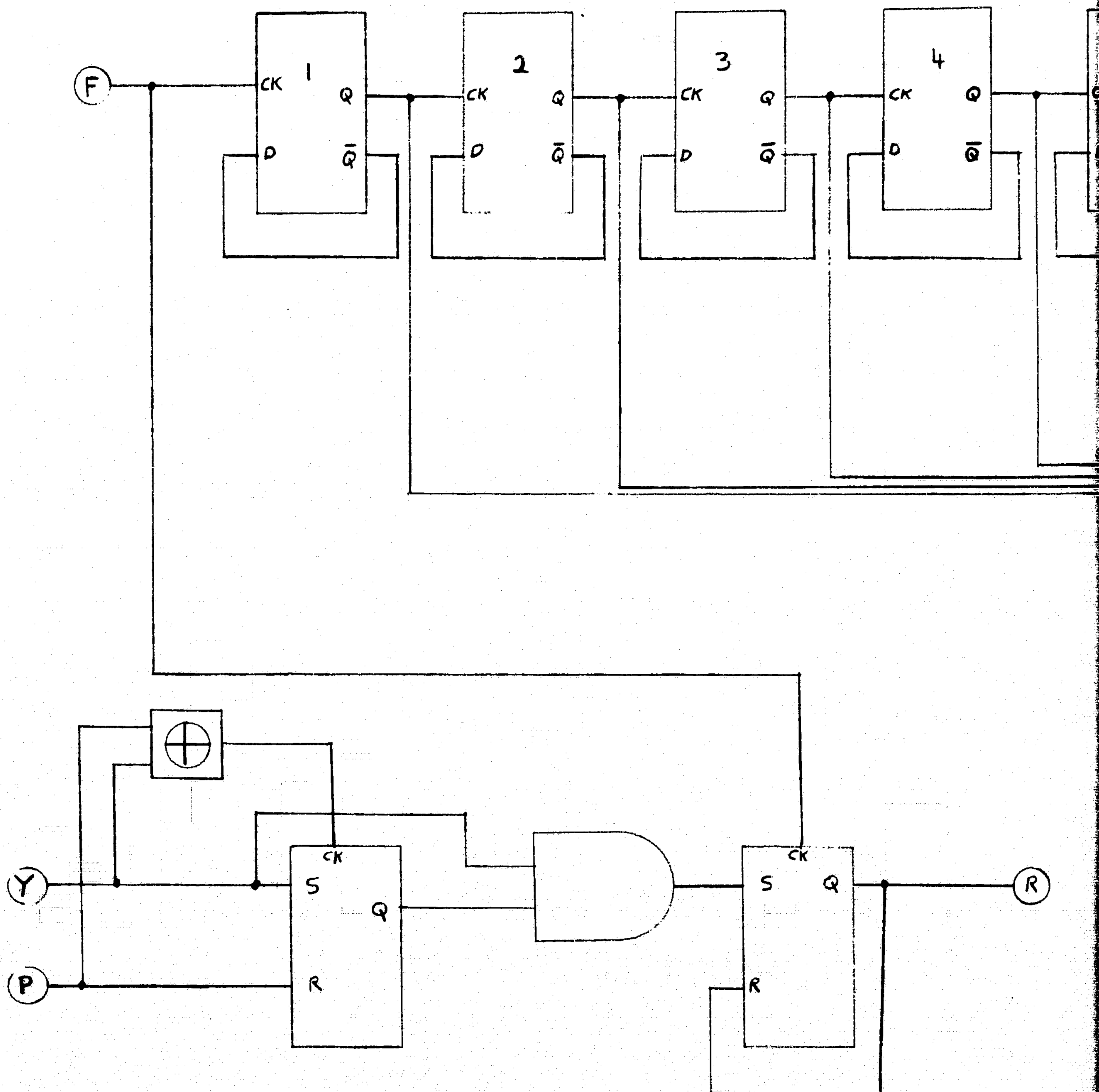

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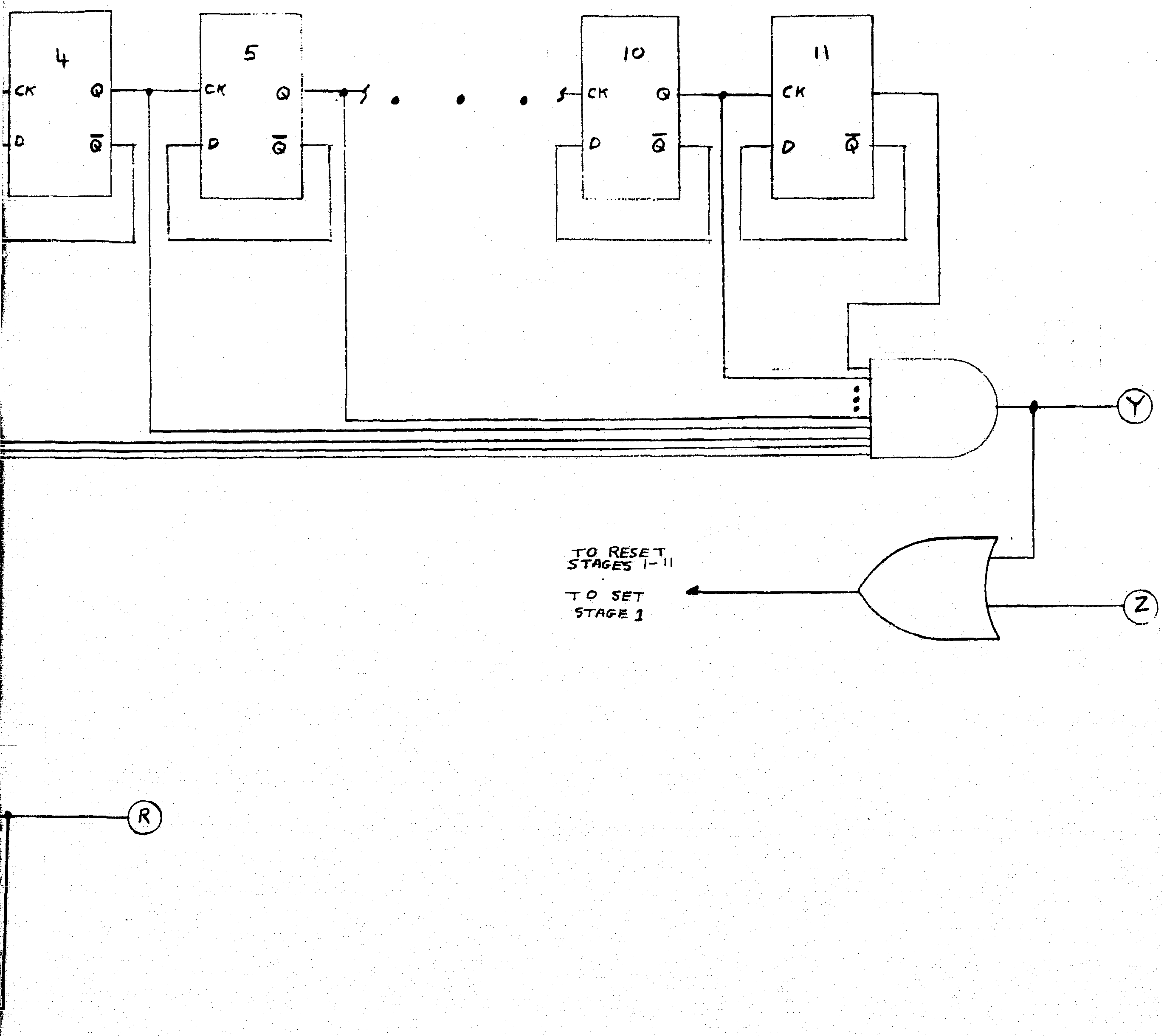
Fig. 4-14 Network for ÷2047



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5. CONVOLUTIONAL ENCODER DESIGN AND DECODER ALGORITHM

(a) Error Control Coding

A previous study of possible coding schemes for digital data in a satellite relay communications link has concluded that convolutional encoding in conjunction with soft-decision Viterbi decoding gives favorable performance gain with minimum increased hardware complexity.

Figure 5-1 is a result of a computer simulation of a rate 1/3, constraint length 7 convolutional coding scheme with a 3-bit soft decision Viterbi decoder. As can be seen from the figure, at a bit error rate of 10^{-5}, a 6 db coding gain results with coded ideal coherent PSK as compared with uncoded ideal coherent PSK. Figure 5-2 is a diagram of a rate 1/3, constraint length 7 convolutional encoder.

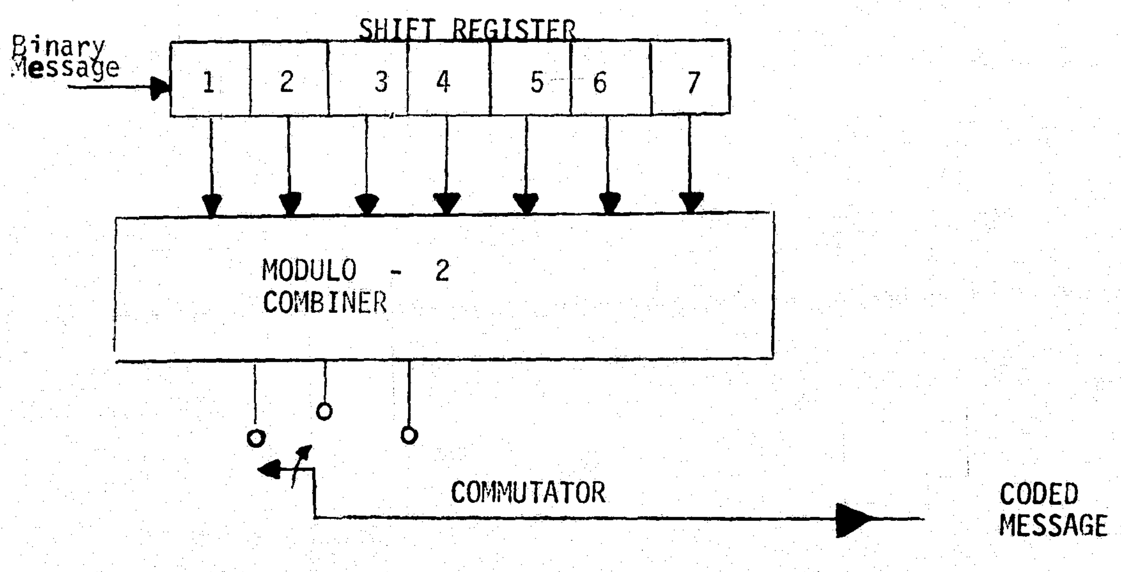


Figure 5-2 K=7, V=3 Convolutional Encoder

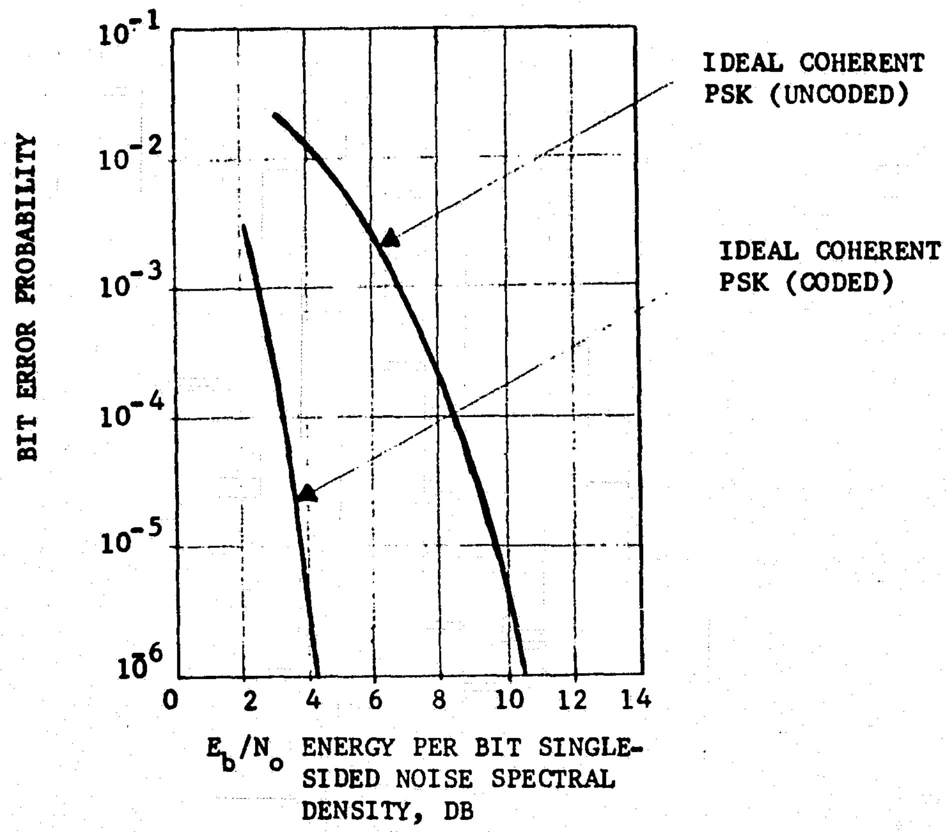


Figure 5-1 - Simulation results for K = 7, V = 3 convolutional encoding/soft-decision Viterbi decoding

The modulo-2 combiner forms a modulo-2 combination of selected register stages to form each of the three commutator nodes. This can be expressed as

$$G_i = (g_{1i}, g_{2i}, \cdots g_{7i}), \tag{5-1}$$

and

$$C_i = \sum_{j=1}^{7} g_{ji} X_j \qquad \text{MOD-2} \tag{5-2}$$

where X_j is the contents of the jth shift register stage and g_{ji} is 0 or 1 depending upon whether the jth stage contributes, modulo-2, to the ith commutator pole.

The operation of the encoder is as follows: The binary message may be much larger than the constraint length. The first bit of the message is switched into the shift register, whose other stages are logical zero, and a complete cycle of the commutator is made. The next bit of the sequence is switched into the register, the initial bit shifted to register stage-two and another synchronous cycle of the commutator is made. Using the synchronous shift and cycle procedure the message sequence is encoded. At the end of the binary message seven zeros are attached, and when they are shifted into the register and accompanying code generated by the commutator, the shift register is in the all zero state once more. For an L - bit message,

$$L_c = 3(L+6) \tag{5-3}$$

bits from the coded message.

Decoding may be accomplished by sequential or Viterbi algorithms.

The sequential decoding method may be described as a tree searching procedure, the exact details depending upon which particular algorithm is being used. The decoding procedure is best described by example, K = 7 is large for the purpose of an example, so a K = 4, V = 3 example is given.

The tree structure for a K = 4, V = 3 truncated code is shown in figure 5-4. The encoder for the code is shown in figure 5-3.

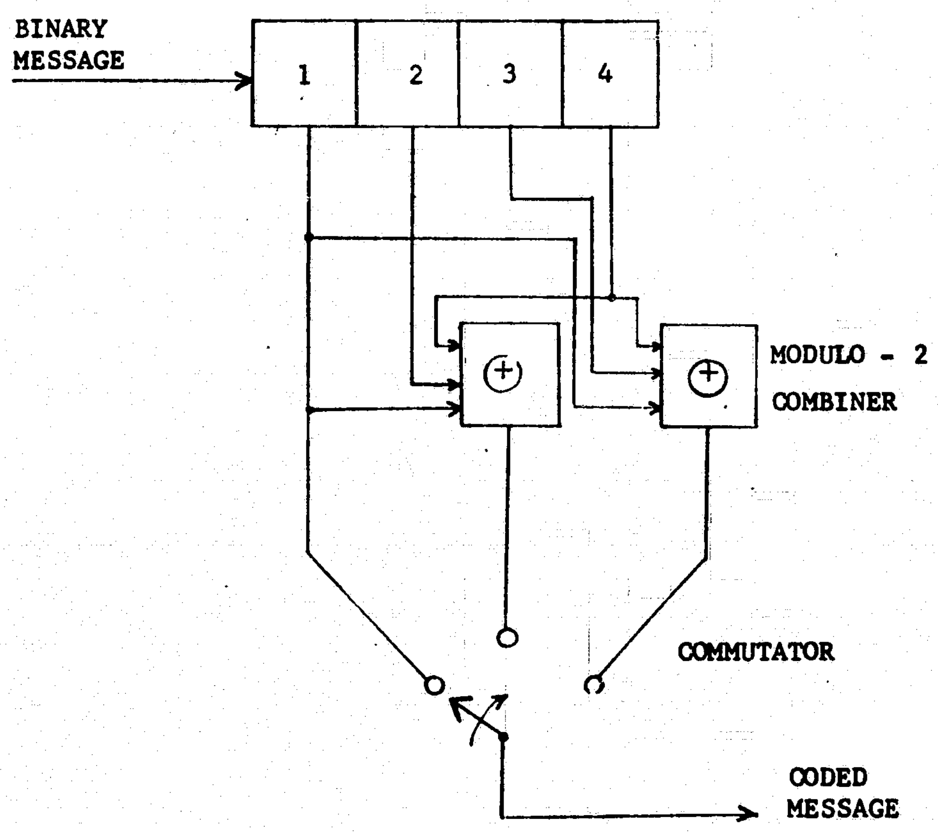


Figure 5-3 Encoder for tree structure of figure 5-4, K=4, V=3.

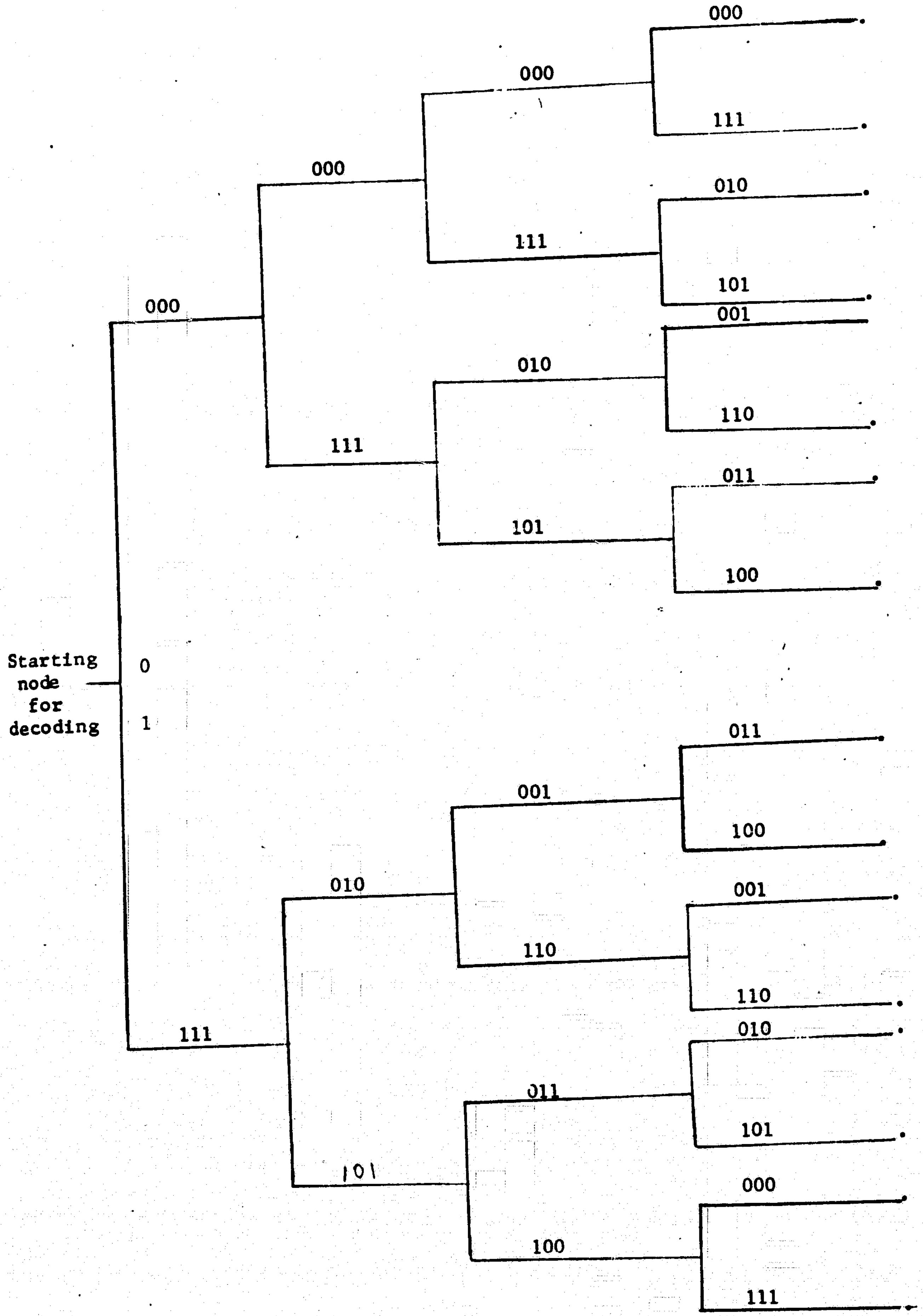


Figure 5-4 Tree Structure for K=4 V=3 truncated code

As an example assume the message

$$X=(1011) \tag{5-4}$$

is to be transmitted. The encoder of figure 5-3 provides the coded message,

$$Y=(111,010,110,110). \tag{5-5}$$

Assuming the channel introduces the noise

$$N=(100,101,000,010), \tag{5-6}$$

the received code is

$$R=(011,111,110,100). \tag{5-7}$$

The sequential decoder will form the quantity

$$d_i = w\left[R_i \oplus Y_i\right], \tag{5-8}$$

where i represents the ith three bit sequence, w represents the weight function, and d_i is called the Hamming distance. The decoder makes each decision at each mode of the code tree based on minimizing the Hamming distance. However the decisions are tentative, and if the decoder finds in successive steps that it has probably made a wrong bit decision it is able to backtrack and try another branch of the code tree.

In the example, the decoded message would begin

$$C = 11.. \tag{5-9}$$

the initial decision for the second bit being made in error. Proceeding down the error branch however significantly large values of d_i are encountered. Backtracking and trying the

$$C=10.. \tag{5-10}$$

branch gives significantly smaller values of d_i on successive steps.

The decoder algorithm is based on monitoring the statistical properties of the sum of d_i as the decoder proceeds into the code tree. If the sum of the d_i terms approaches a buildup rate of $\frac{1}{2}V$ then the decoder declares an error and backtracks to a new branch.

Expected buildup of the d_i sum for the correct branch is PV where P is the channel transision probability for the binary symmetric channel. The branch decision criteron is buildup somewhere between V/2 and PV. The decoder keeps track of the branches it has explored and avoids needless retracing of any branch.

The design selected for the HEAO-C transponder error control coding is the rate 1/3, constraint length 7 convolutional encoder (K=7, V=3). The equations for the three commutator nodes are:

$$G_1 = (1,0,0,0,0,0,0) \tag{5-11}$$

$$G_2 = (1,0,1,1,0,1,1) \tag{5-12}$$

$$G_3 = (1,1,0,0,1,1,1) \tag{5-13}$$

The encoder consists of the following components:

(1) Input storage buffer

(2) Word counter

(3) Timing and control circuit

(4) Encoding register

(5) Half-adder circuits

(6) Three node commutator

Figure 5-5 is the overall block diagram of the encoder. The data to be encoded is assumed to be organized in data frames of 64 data words, 32 bits being the length of each word. The 64th word is a frame-synchronization word. The 2048 frame bits of data are encoded into 7488 bits for the coded data frame. The data word is organized by the encoder into the form shown in figure 5-6.

25-BIT DATA	7-BIT ZERO STRING

(32-BIT)

FIGURE 5-6 DATA WORD

The operation of the encoding is as follows:

(1) Input data is directed into a serial-in, serial-out buffer register.

(2) The word counter allows 32 bits of data to be encoded, then indicates end-of-data-word to the timing and control circuit.

(3) The timing and control circuit inserts a 7-bit all zero string at the end of the 25 bits of data into the generating register.

(4) The half-adders form three nodes as shown by equations 5-11, 5-12, and 5-13.

(5) Under the control of the timing and control circuit the three node commutator samples the three nodes and places the resulting digital sequence on the output line.

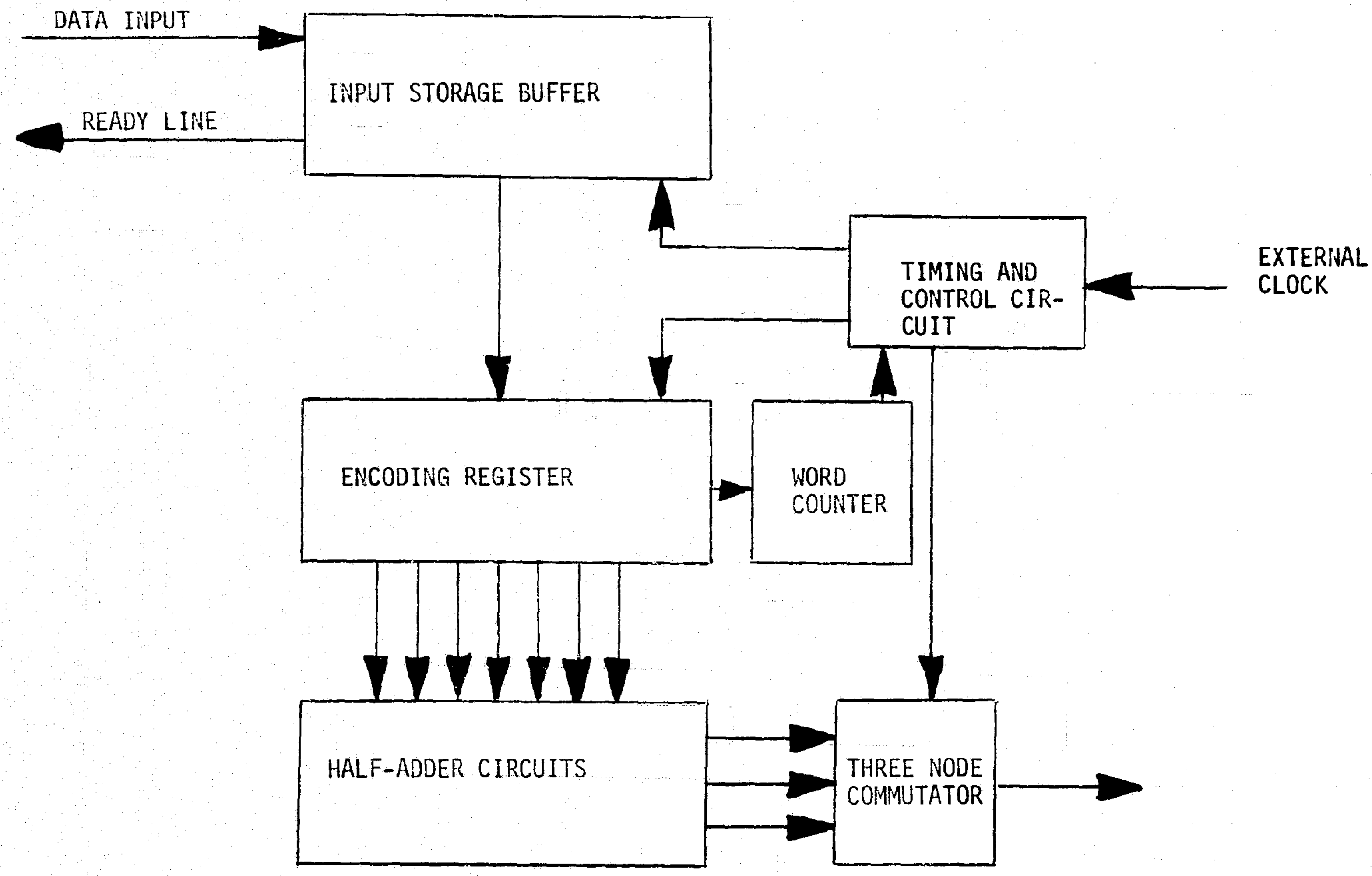
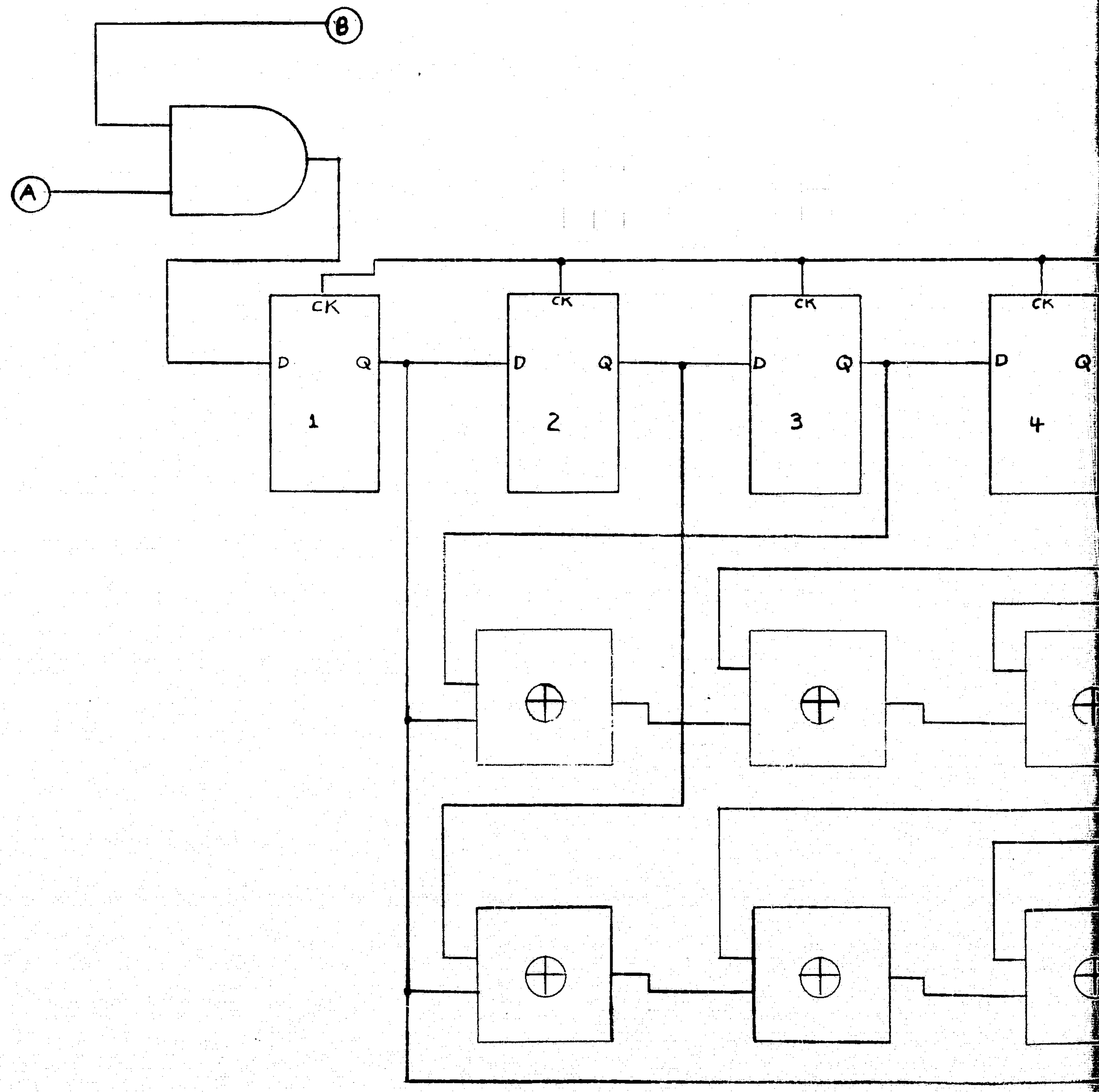


FIGURE 5-5 ENCODER BLOCK DIAGRAM

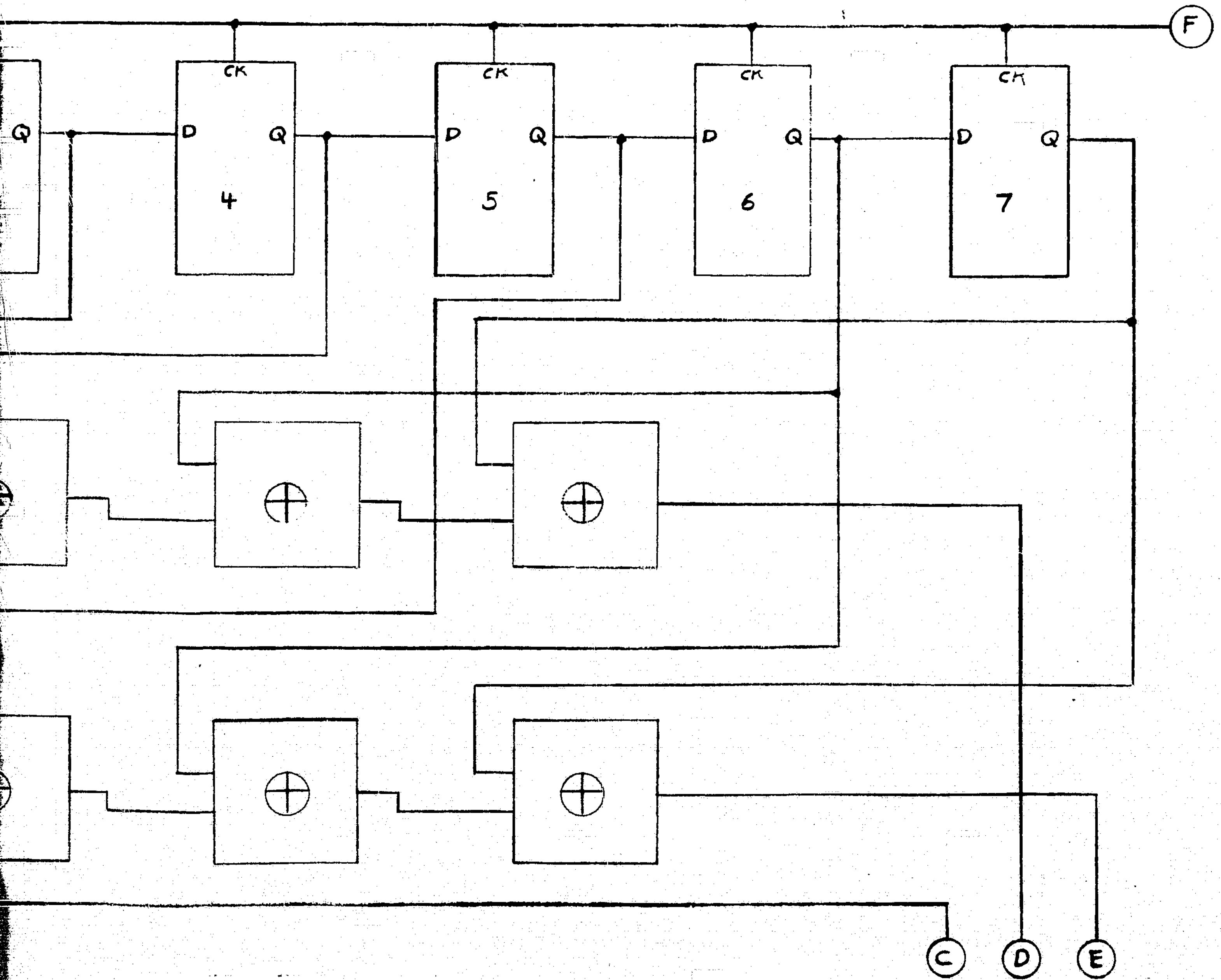
BY ____________ DATE ________ SUBJECT ________________________________ SHEET NO. ________ OF ______ Fig.

CHKD. BY ______ DATE ________ JOB NO. ____________



FOLDOUT FRAME /

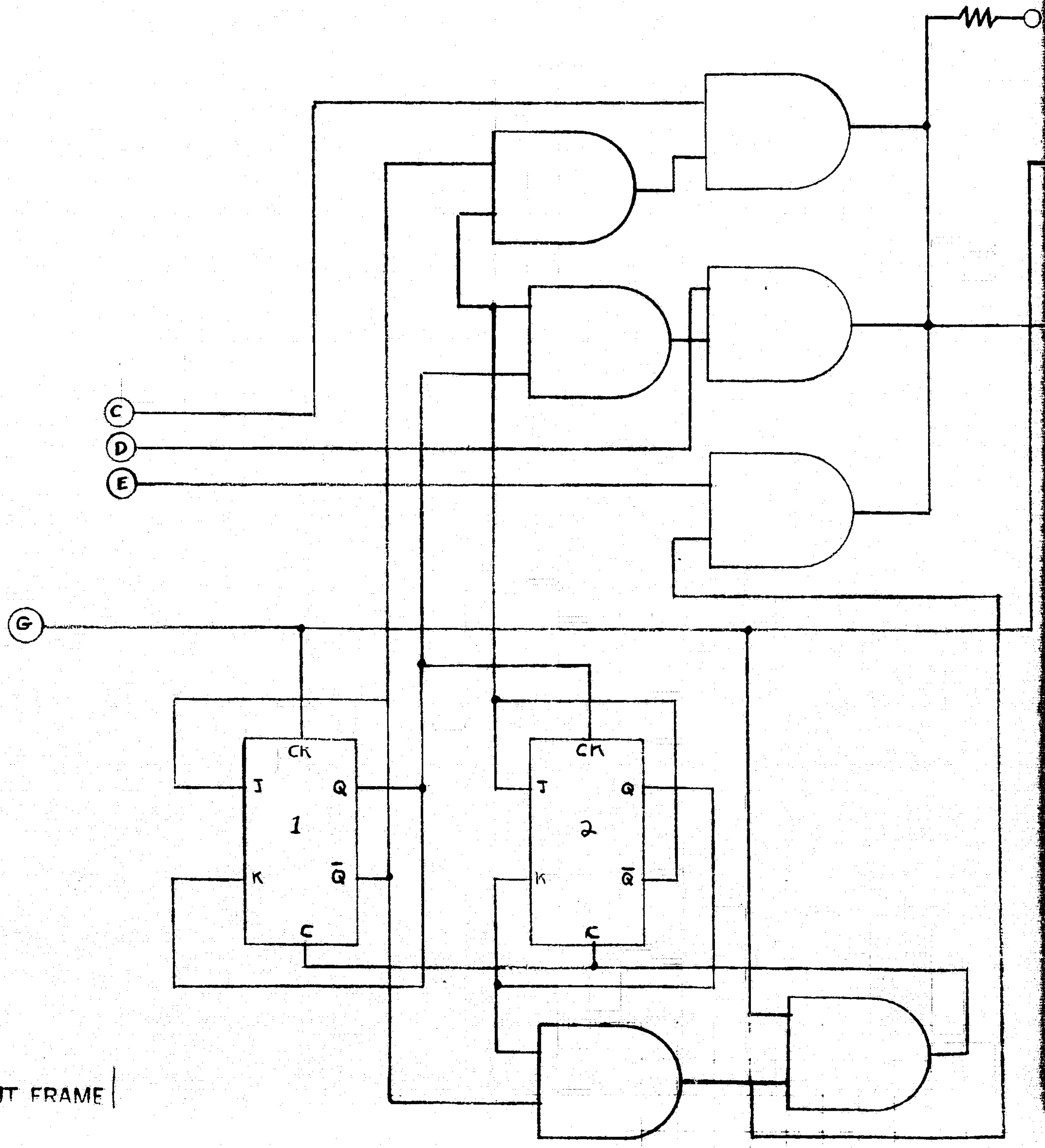
Fig. 5-6 Encoding Register and Half-Adder Circuits



FOLDOUT FRAME 2

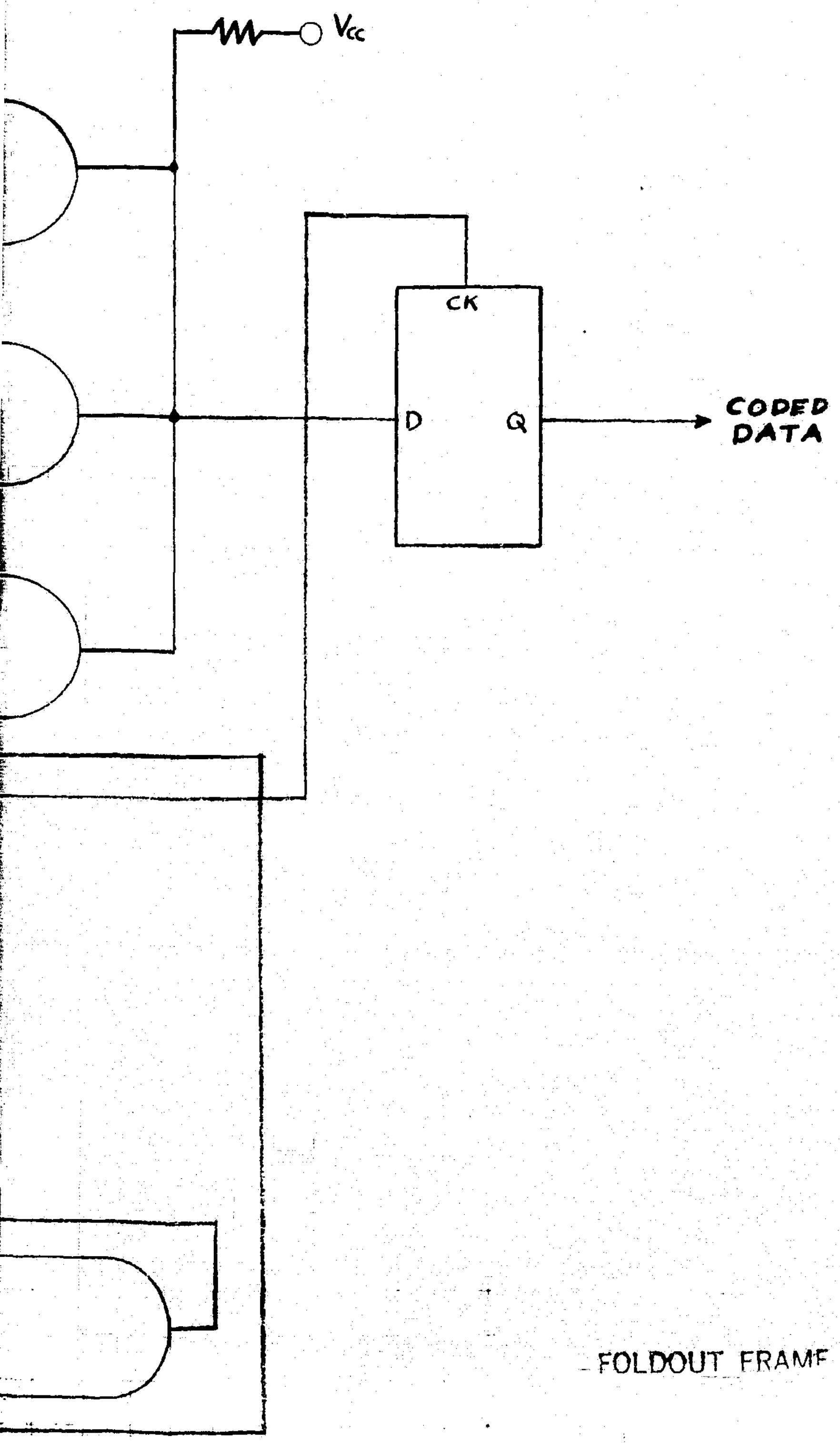
BY ________ DATE ________ SUBJECT THREE NODE COMMUTATOR SHEET NO. ________ OF ______

CHKD. BY ________ DATE ________ JOB NO. ________



FOLDOUT FRAME

Fig. 5-7 Three Node Commutator



FOLDOUT FRAME 2

6. GENERATION OF HIGH-SPEED MAXIMUM LENGTH DIGITAL SEQUENCES

This section is the result of a study of sampled maximum length digital sequences. The purpose of the study was to establish the mathematical basis for the design of a high speed digital PSEUDORANDOM SEQUENCE GENERATOR FOR USE IN A SPREAD SPECTRUM TRANSPONDER SYSTEM. The proposed procedure for generating the high speed ML sequence involves sampling several slower speed ML generations. Figure 6-1 illustrates the sequence generator.

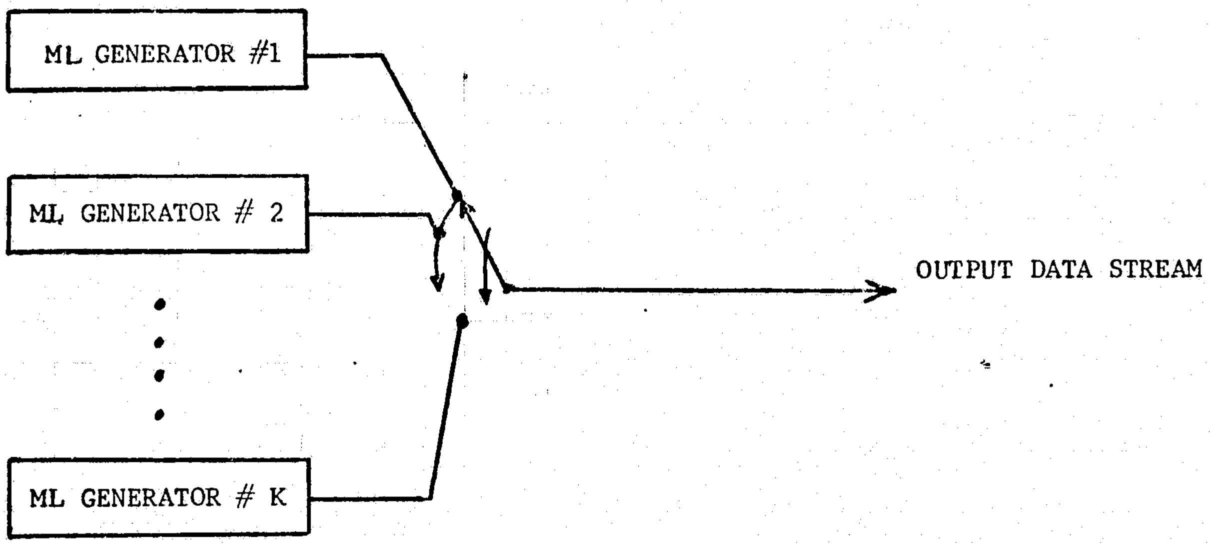


Figure 6-1 HIGH SPEED SEQUENCE GENERATOR

If there are K ML generators forming the sequence generator where K=2, an integer, then the commutation rate should be K times the clock rate of ML generators. Each generator is sampled once during a clock interval, and the output data stream would consist of K digits during the clock-interval. The advantage of this configuration is that a higher speed digital bit stream can be generated with ML sequence generators operating with a clock-frequency that is only a fraction of the data-rate.

Specifically, if the data-rate is F_B bits/sec, the required clock-rate is F_B/K. For example, if it is desired to operate with a data rate of 40×10^6 Bits/sec, and assuming K=4 (four ML generators) then the generator clock rates would be 10 MHZ. This allows the use of less-expensive, more-reliable digital components from lower speed logic families. The only component required to operate at the 40 MHZ rate is the commutating switch.

One important technical consideration involved in the high speed sequence generator is the phasing of the maximum length sequence generators, shown in figure 1, to provide the desired output data stream.

<u>Sequence Generator Phasing</u>

The phasing problem can be stated: "what initial phasing of the K ML-generators shown in figure 6-1 is required to provide the ML-sequence in the output data stream when the sampling procedure is used." The rule with K ML-generators for phasing the ith generator relative to the first generator is

1. Advance by (i-1) $[(L+1)/K]$ bits

   or

2. Delay by (i-1) $[L-(L+1)/K]$ Bits

The rationale for the above choice of phase relation is as follows:

1. Sampling a ML sequence provides a shifted version of the same sequence if the sampling rate is an integer power of two.
2. Consider the synthesized sequence as being reconstructed from K, K-sampled versons of itself.
3. Consecutive digits in a component sequence must be separated by K digits in the composite ML sequence.
4. Consecutive digits in the composite ML sequence must be separated by (L+1)/K bits in the component sequence.
5. Arranging K, K-sampled sequences, each advanced by (L+1)/K bits relative to its adjacent sequence, and sampling from each as shown in figure 6-1, must yield the same ML sequence.

6. A phase advance of (L+1)/K bits in a ML sequence is equivalent to a delay of [L-(L+1)/K] bits.

1. abcdefg $\longleftrightarrow$ acegbdf
   same sequence

2. $\left.\begin{array}{l}\text{acegbdf}\\ \text{bdfaceg}\end{array}\right\}$ abcdefgabcdefg

3. ADJ
   $\left.\begin{array}{l}\text{acegbdf}\\ \text{bdfaceg}\end{array}\right\}$ K
   abcdefgabcdefg

4. $\frac{L+1}{K}$
   $\left.\begin{array}{l}\text{acegbdf}\\ \text{bdfaceg}\end{array}\right\}$ ADJ
   abcdefgabcdefg

As an example consider the ML sequence abcdefg. Sampling every other bit yields acegbdf which must be a shifted version of the same sequence. Advancing this sequencing by (L+1)/K = 8/2 = 4 bits yields bdfaceg. Synthesizing by sampling, in-turn, from the two sequence yields abcdefabcdef, which is the original ML sequence repeated twice.

As a practical illustration consider the ML sequence generator with the characteristic polynomial

$$G(z) = 1 + z + z^3 \tag{1}$$

The sequence generated by this ML generator is S(z), where

$$\frac{1}{G(z)} = \frac{S(z)}{1+z^L} \tag{2}$$

and where

$$L = 2^N - 1 \tag{3}$$

for an N-stage shift-register generator. For the generator in question

$$S(z) = 1 + z + z^2 + z^4 \tag{4}$$

which represents the sequence 1110100. Forming $S^*(z)$ by advancing the phase by

$$(L+1)/K = 4 \text{ for } K = 2 \tag{5}$$

and sampling in turn from S(z) and $S^*(z)$ yields

SEQUENCE: 1110100
SEQUENCE
Advanced
by 4 BITS: 1001110 } 11101001110100

as expected. Performing the similar analysis for K = 4 yields

1110100
1010011
1001110
0111010 } 1110100111010011101001110100.

The output for the sampling generator is

$$S^2(z) + z\,(S(z)z^{(L-1)/2})^2 = SS(z) \qquad \text{MOD } 2L \qquad (6)$$

For K=2
or

$$SS(z) = S^2(z)(1+z^L) \qquad \text{MOD } 2L \qquad (7)$$

The sequence can also be expressed as

$$SS(z) = (1+z^L)^3/(G^2(z)) \qquad \text{MOD } 2L \qquad (8)$$

In general, for K component generators

$$SS(z) = \sum_{i=1}^{K} z^{i-1}\, S^K(z) z^{(i-1)(KL-L-1)} \qquad (9)$$

or

$$SS(z) = \sum_{i=1}^{K} \frac{z^{(i-1)}L(1-z^L)^K \; z^{(i-1)(KL-L-1)}}{G^K(z)} \qquad \text{MOD } KL \qquad (10)$$

The synthesized sequence can be expressed as a shifted version of the original ML sequences,

$$SS(z) = z^x(1+z^L)\, S(z) \qquad \text{MOD } 2L \qquad (11)$$

for the case K=2, or in general,

$$SS(z) = z^x S(z) \sum_{i=1}^{K} z^{(i-1)L} \qquad \text{MOD } KL \qquad (12)$$

or

$$SS(z) = \frac{z^x(1+z^L)}{G(z)} \sum_{i=1}^{K} z^{(i-1)^L} \qquad \text{MOD KL,} \qquad (13)$$

Equating (9) and (12) yields

$$S^K(z)\sum_{i=1}^{K} z^{(i-1)(KL-L)} = z^x S(z)\sum_{i=1}^{K} z^{(i-1)L} \qquad \text{MOD KL,} \qquad (14)$$

or

$$S^K(z)\sum_{i=1}^{K} z^{(i-1)(KL-L)} + z^x\ S(z)\sum_{i=1}^{K} z^{(i-1)L} = 0 \qquad \text{MOD KL.} \qquad (15)$$

For the case K=2 this reduces to

$$S^2(z)(1+z^L) + z^x S(z)(1+z^L) = 0 \qquad \text{MOD 2L.} \qquad (16)$$

Simplification of (15) yields

$$S^{K-1}(z) + z^x = 0 \qquad \text{MOD KL,} \qquad (17\text{-A})$$

which must be satisfied by the sequence.

Equation (17-A) holds because

$$z^{KL} = 1 \qquad \text{MOD KL,} \qquad (17\text{-B})$$

and

$$\sum_{i=1}^{K} z^{-(i-1)L} = \sum_{i=1}^{K} z^{(i-1)L} \qquad \text{MOD KL.} \qquad (17\text{-C})$$

If equation (17) describes the sequence generated by $S(z)$ then

$$G(z) \mid (S^{K-1}(z) + z^x) \qquad \text{MOD KL.}$$

Now

$$\frac{S^{K-1}(z) + z^x}{G(z)} = \frac{(1+z^L)^{K-1} + z^x G^{K-1}(z)}{G^K(z)} \qquad \text{MOD KL,} \qquad (18)$$

or

$$G^K(z) \mid ((1+z^L)^{K-1} + z^x G^{K-1}(z)) \qquad \text{MOD KL.} \qquad (19)$$

For the example:

$$G(z) = 1 + z + z^3 \qquad (20)$$

and

$$K = 2 \qquad (21)$$

yields

$$X = 0 \tag{22}$$

and

$$\frac{(1+z^L)^{K-1} + z^x G^{K-1}(z)}{G^K(z)} = \frac{z + z^3 + z^7}{1+z^2+z^6} = z \qquad \text{MOD } 14. \tag{23}$$

A similar example for

$$G(z) = 1 + z^2 + z^3$$

where

$$S(z) = 1 + z^2 + z^3 + z^4$$

represents the sequence 1011100.

A sampling arrangement for K=2, requires a delay = (L-1)/2=3, shown below

$$\left.\begin{matrix} 1011100 \\ 1001011 \end{matrix}\right\} 11001011100101$$

observe for this case X=4,

and

$$\frac{(1+z^L)^{K-1} + z^x G^{K-1}(z)}{G^K(z)} = \frac{1 + z^4 + z^6}{1 + z^4 + z^6} = 1 \qquad \text{MOD } 14. \tag{24}$$

An algorithm to calculate X is as follows:

1. Starting with the all zero (N-1) - tube generate the sequence, S(z), with the characteristics equation G(z).
2. Generate $S^K(z)$ From S(z) Or $G^K(z)$.
3. Find X such that $(1+z^L)^{K-1} + z^x G^{K-1}(z)$ forms a recursive relation that holds over the all zero 2(N-1) - tuple of $S^K(z)$.

Sample calculations are shown below For K=2.

Sample 1:

$$G(z) = 1 + z^2 + z^3 \tag{25}$$

0 0 1 0 1 1 1 0 0 1 0 1 1 1

00 00 10 00 10 10 10 00 00 10 00 10 10 10

$$1 + z^7 + z^4 (1+z^2+z^3) = 1 + z^4+z^6 = G^2(z) \tag{26}$$

Sample 2

$$G(z) = 1 + z + z^3 \tag{27}$$

0 0 1 1 1 0 1 0 0 1 1 1 0 1

00 00 10 1010001000001010100010

$$1 + z^7 + z^0\,(1+z+z^3) = z(1+z^2+z^6) = zG^2(z) \tag{28}$$

A similar problem involves the phase of the sequence resulting from sampling a sequence at a rate

$$r = 2^g \tag{29}$$

with g an interger. For example,

$$G(z) = 1 + z^2+z^3 \tag{30}$$

yields the sequence

1011100.

Sampling this sequence with r=2 yields

1110010

which is a phase shift corresponding to z^5.

As another example

$$G(z) = 1 + z + z^3 \tag{31}$$

yields the sequence

1110100.

Sampling the sequence with r=2 yields

1110100

which is a phase shift corresponding to Z^o.

A procedure for determining the phase shift can be found if an expression of the form S(z) can be found for the sequence formed as a result of sampling, and

$$f(z) = z^x S(z) \qquad \text{MOD L.} \tag{32}$$

If an ML sequence is sampled at a rate of

$$r=2 \tag{33}$$

then an adjacent bit will be sampled $(\frac{L+1}{2})$ bits after the sampled bit in the sequence formed from sampling. Extending this analysis, a sampled sequence can be expressed as

$$f(z) = \sum_{i=o}^{\frac{L-1}{2}} z^i (z^{2i}\ S(z))^{(L+1)/2} \qquad \text{MOD } \frac{L+1}{2}L \tag{34}$$

or alternately

$$f(z) = \sum_{i=o}^{\frac{L-1}{2}} z^i (z^{L-2i}\ S(z))^{(L+1)/2} \qquad \text{MOD} (\frac{L+1}{2})L \tag{35}$$

For example, the sequence with characteristic equation

$$G(z) = 1+z^2+z^3 \tag{36}$$

and

$$S(z) = 1 + z^2 + z^3 + z^4,\ 1011100 \tag{37}$$

with

$$s^4(z) = 1 + z^8 + z^{12} + z^{16} \qquad \text{MOD } 28 \tag{38}$$

$$z^1(z^5)^4 s^4(z) = z + z^5+z^9+z^{21} \qquad \text{MOD } 28 \tag{39}$$

$$z^2(z^3)^4 s^4(z) = z^2+z^{14}+z^{22}+z^{28} \qquad \text{MOD } 28 \tag{40}$$

and
$$z^3(z)^4 s^4(z) = z^7+z^{15}+z^{19}+z^{23} \qquad \text{MOD } 28 \tag{41}$$

yields from equation (30)

$$f(z) = 1+z+z^2+z^5+z^7+z^8+z^9+z^{12}+z^{14}+z^{15}+z^{16}+z^{19}+z^{21} + z^{22}+z^{23}+z^{26} \qquad \text{MOD } 28 \tag{42}$$

or

1110010111001011100101110010.

Reference Sequences:			
1011100 = $z^7 S(z)$	1110010 = $z^5 S(z)$	1001011 = $z^3 S(z)$	0101110 = $zS(z)$
Sum Sequence:			
1000000001000100010000000000	= $[z^7 S(z)]^4$	MOD 28	
0100010001000000000000100 0000	= $[z^5 S(z)]^4$	MOD 28	
0010000000000001000000001000l0	= $[z^3 S(z)]^4$	MOD 28	
00000001000000001000100010000	= $[zS(z)]^4$	MOD 28	
1 110010111001011100101110010	= $f(z)$	MOD 28	

Table 1 Formation of synthesized sequence for the case $G(z) = 1 + z^2 + z^3$

The table above also illustrates the result of equation (30).

Now from equation (37).

$$\sum_{i=0}^{\frac{L-1}{2}} z^i \left[z^{L-2i}\, S(z)\right]^{\frac{L+1}{2}} = z^x S(z) \sum_{i=0}^{\frac{L-1}{2}} z^{iL} \qquad \text{MOD } \frac{L+1}{2}L \tag{43}$$

or the sequence must satisfy

$$[S(z)]^{\frac{L-1}{2}} \sum_{i=0}^{\frac{L-1}{2}} z^i [z^{L-2i}]^{\frac{L+1}{2}} + z^x \sum_{i=0}^{\frac{L-1}{2}} z^{iL} = 0 \tag{44}$$

and

$$G(z) \,\Big|\, \left[(S(z))^{\frac{L-1}{2}} \sum_{i=0}^{\frac{L-1}{2}} z^i \, [z^{L-2i}]^{\frac{L+1}{2}} + z^x \sum_{i=0}^{\frac{L-1}{2}} z^{iL} \right] \tag{45}$$

Also,

$$\frac{[S(z)]^{\frac{L-1}{2}} \sum_{i=o}^{\frac{L-1}{2}} z^i[z^{L-2i}]^{\frac{L+1}{2}} + z^x \sum_{i=o}^{\frac{L-1}{2}} z^{iL}}{G(z)} =$$

$$[1+z^L]^{\frac{L-1}{2}} \sum_{i=o}^{\frac{L-1}{2}} z^i[z^{L-2i}]^{\frac{L+1}{2}} + z^x G^{\frac{L-1}{2}}(z) \sum_{i=o}^{\frac{L-1}{2}} z^{iL} /[G(z)]^{(L+1)/2} \quad (46)$$

For example if

$$G(z) = 1+z+z^2 \quad (47)$$

$$S(z) = 1+z \ : \ 110 \quad (48)$$

$$L = 3 \quad (49)$$

and

$$X = 2 \ : \ 101 \quad (50)$$

for

$$r = 2 \quad (51)$$

Evaluating the terms in (41) yields:

$$(1+z^L)^{\frac{L-1}{2}} = 1 + z^3 \quad (52)$$

$$\sum_{i=o}^{\frac{L-1}{2}} z^i[z^{L-2i}]^{\frac{L+1}{2}} = \sum_{i=o}^{1} z^i[z^{3-2i}]^2 = 1 + z^3 \quad \text{MOD } 2L \quad (53)$$

$$[G(z)]^{\frac{L-1}{2}} = G(z) = 1 + z + z^3 \quad (54)$$

$$[G(z)]^{\frac{L+1}{2}} = [G(z)]^2 = 1 + z^2 + z^4 \quad (55)$$

Equation (46) becomes

$$\frac{(1+z^3)(1+z^3) + z^2(1+z+z^7)(1+z^3)}{1+z^2+z^4} = \frac{(1+z^3)(1+z^2+z^4)}{1+z^2+z^4} = 1+z^3 \tag{56}$$

Notice that MOD $\left(\frac{L+1}{2}\right)$ L, or MOD 6, arithmetic was not used in (56).

As a second example, if

$$G(z) = 1 + z^2 + z^3 \tag{57}$$

$$S(z) = 1 + z^2 + z^3 + z^4 : 1011100 \tag{58}$$

$$L = 7 \tag{59}$$

and

$$X = 5 \tag{60}$$

Corresponding to the sequence 1110010 formed by sampling.

Evaluating the terms in (41)

$$(1+z^L)^{\frac{L-1}{2}} = (1+z^7)^3 = 1 + z^7 + z^{14} + z^{21} \tag{61}$$

$$\sum_{i=o}^{\frac{L-1}{2}} z^i \left(z^{L-2i}\right)^{\frac{L+1}{2}} = \sum_{i=o}^{3} z^i (z^{7-2i})^4$$

$$= 1+z^7+z^{14}+z^{21} \tag{62}$$

$$(G(z))^{\frac{L-1}{2}} = (1+z^2+z^3)^3 = 1+z^2+z^3+z^4+z^7+z^8+z^9 \tag{63}$$

$$\sum_{i=o}^{\frac{L-1}{2}} z^{iL} = \sum_{i=o}^{3} z^{i7} = 1+z^7+z^{14}+z^{21} \tag{64}$$

and

$$[G(z)]^{\frac{L+1}{2}} = (1+z^2+z^3)^4 = 1+z^8+z^{12} \qquad (65)$$

Equation (41) becomes

$$\frac{(1+z^7+z^{14}+z^{21})(1+z^7+z^{14}+z^{21})+z^5(1+z^2+z^3+z^4+z^7+z^8+z^9)(1+z^7+z^{14}+z^{21})}{1+z^8+z^{12}} =$$

$$\frac{(1+z^7+z^{14}+z^{21})(1+z^5+z^8+z^9+z^{12}+z^{13}+z^{21})}{1+z^8+z^{12}} = (1+z^7+z^{14}+z^{21}) \cdot (1+z^5+z^9) \qquad (66)$$

Notice for these two examples, (46) reduces to finding X such that

$$\frac{(1+z^L)^{\frac{L-1}{2}} + z^x(G(z))^{\frac{L-1}{2}}}{[G(z)]^{\frac{L+1}{2}}} \qquad (67)$$

is a rational fraction.

As another example, consider

$$G(z) = 1+z+z^3 \qquad (68)$$

$$S(z) = 1+z+z^2+z^4\text{: } 1110100 \qquad (69)$$

$$L = 7 \qquad (70)$$

and

$$X = 0 \qquad (71)$$

Corresponding to the sequence 110100 formed by sampling.

Evaluating the terms in (46)

$$[G(z)]^{\frac{L-1}{2}} = (1+z+z^3)^3 = 1+z+z^2+z^5+z^6+z^7+z^9 \qquad (72)$$

and

$$(G(z))^{\frac{L+1}{2}} = 1+z^4+z^{12} \qquad (73)$$

Equation (46) becomes

$$\frac{(1+z^7+z^{14}+z^{21})(1+z^7+z^{14}+z^{21}+z^x)(1+z+z^2+z^5+z^6+z^7+z^9)}{(1+z^4+z^{12})} \qquad (74)$$

Now if X=0 this becomes

$$\frac{(1+z^7+z^{14}+z^{21})(z)(1+z+z^4+z^5+z^8+z^{13}+z^{20})}{1+z^4+z^{12}} =$$

$$(1+z^7+z^{14}+z^{21})(z)(1+z+z^8) \tag{75}$$

The algorithm for finding the shift X for a sampling rate S=2 (corresponding to sampling every other bit) is to find an integer X such that

$$\frac{(1+z^L)^{\frac{L-1}{2}} + z^x\,(G(z))^{\frac{L-1}{2}}}{(G(z))^{\frac{L+1}{2}}}$$

is rational.

Now for the general case

$$S=2^{\vartheta} \tag{76}$$

ϑ an integer,

the generalization of (43) becomes

$$\sum_{i=o}^{\frac{L+1}{S}-1} z^i[z^{L-Si}(S(z)]^{\frac{L+1}{3}} = z^x S(z) \sum_{i=o}^{\frac{L+1}{S}} z^{iL} \qquad \text{MOD } \frac{L+1}{S}L \tag{77}$$

For example with

$$G(z) = 1+z^2+z^3 \tag{78}$$

and

$$S(z) = 1+z^2+z^3+z^4 \;:\; 1011100 \tag{79}$$

Sampling at the rate

$$S=4 \tag{80}$$

yields

$$1100101:\; 1+z+z^4+z^6$$

or

$$X=4. \tag{81}$$

Equation (77) becomes

$$\sum_{i=o}^{1} z^i[z^{7-4i}][S(z)]^2 = z^x S(z)\sum_{i=o}^{1} z^{7i} \tag{82}$$

Evaluating the term in (82),

$$S^2(z) = 1 + z + z^4 + z^6 \tag{83}$$

$$\sum_{i=o}^{1} z^i\ [z^{7-4i}]^2 = z^{14}+z^7 = 1 + z^7 \tag{84}$$

$$\sum_{i=o}^{1} z^{7i} = 1 + z^7 \tag{85}$$

Substituting into (82)'

$$(1+z+z^4+z^6)(1+z^7) = z^4(1+z^2+z^3+z^4)(1+z^7) \quad \text{MOD}(7)2 \tag{86}$$

or

$$(1+z+z^4+z^6) = z^4(1+z^2+z^3+z^4) \quad \text{MOD}(7) \tag{87}$$

Equation (77) reduces in general to

$$[S(z)]^{\frac{L+1}{S}} = z^x S(z) \quad \text{MOD L} \tag{88}$$

in general

or

$$G(z) \,|\, ((S(z))^{\frac{L+1}{S}} + z^x S(z))$$

and

$$G(z) \,|\, ((S(z))^{\frac{L+1}{S}-1} + z^x)$$

Now

$$\frac{[S(z)]^{\frac{L+1}{S}-1} + z^x}{G(z)} = \frac{(1+z^L)^{\frac{L-S+1}{S}} + z^x(G(z))^{\frac{L-S+1}{S}}}{[G(z)]^{\frac{L+1}{S}}} \tag{89}$$

and the fraction in the latter portion of (89) must be rational.

For example,

$$G(z) = 1 + z^2 + z^3 \tag{90}$$

$S=4$

$$\frac{(1+z^7)^1 + z^4(G(z))}{(G(z))^2} = \frac{1+z^4+z^6}{1+z^4+z^6} = 1 \tag{91}$$

The table below is intended to summarize the use of equation (89) for several example sequences.

$G(z)$	L	S	X	Ratio From (89)
$1 + z + z^2$	3	2	2	1
$1 + z^2 + z^3$	7	2	5	$(1+z^5+z^9)$
$1 + z^2 + z^3$	7	4	4	1
$1 + z + z^3$	7	2	0	$z(1+z+z^8)$
$1 + z + z^3$	7	4	0	z
$1 + z + z^4$	15	2	8	—
$1 + z + z^4$	15	4	12	$1+z^4+z^8+z^{13}+z^{14}+z^{17}+z^{29}$
$1 + z + z^4$	15	8	14	$1+z^2+z^4+z^6+z^{10}$
$1 + z^3 + z^4$	15	2	9	—
$1 + z^3 + z^4$	15	4	6	$1+z^6+z^9+z^{10}+z^{14}+z^{17}+z^{21}+z^{25}+z^{29}$
$1 + z^3+z^4$	15	8	12	$1+z^6+z^8$

Breadboard Circuit

A breadboard circuit was assembled to demonstrate the feasibility of using the sampling technique for generating high speed sequences. A (4,3,0) ML code was selected for the demonstration, and the circuit designed to generate and sample this sequence is shown in figure 6-2.

The phasing function for the given code is

$$P = \frac{(i-1)\ (L+1)}{K} = \frac{(i-1)\ (16)}{4} = -(i-1)(4) \qquad (92)$$

The set of sampled sequences are

$$S$$
$$Sz^{-4}$$
$$Sz^{-8}$$
$$Sz^{-12}$$

where S is the sequence at stage 1 of the generating register shown in figure6-2. The technique for generating the sequence shown in figure 6-2 utilizes a technique for minimizing the number of stages required by the generating register. This method is useful when L is relatively small. The single register method requires

$$N = \frac{(K-1)\ (L+1)}{K} \qquad (93)$$

stages while the separate register method requires

$$N = Kn$$

stages, where n is the order of the code.

For the configuration of figure 2

$$N = 12,$$

while for separate generating registers

$$N = 16.$$

For a case where K = 16,

$$N = 15$$

for the single register generator, and

$$N = 64$$

for the separate register generator.

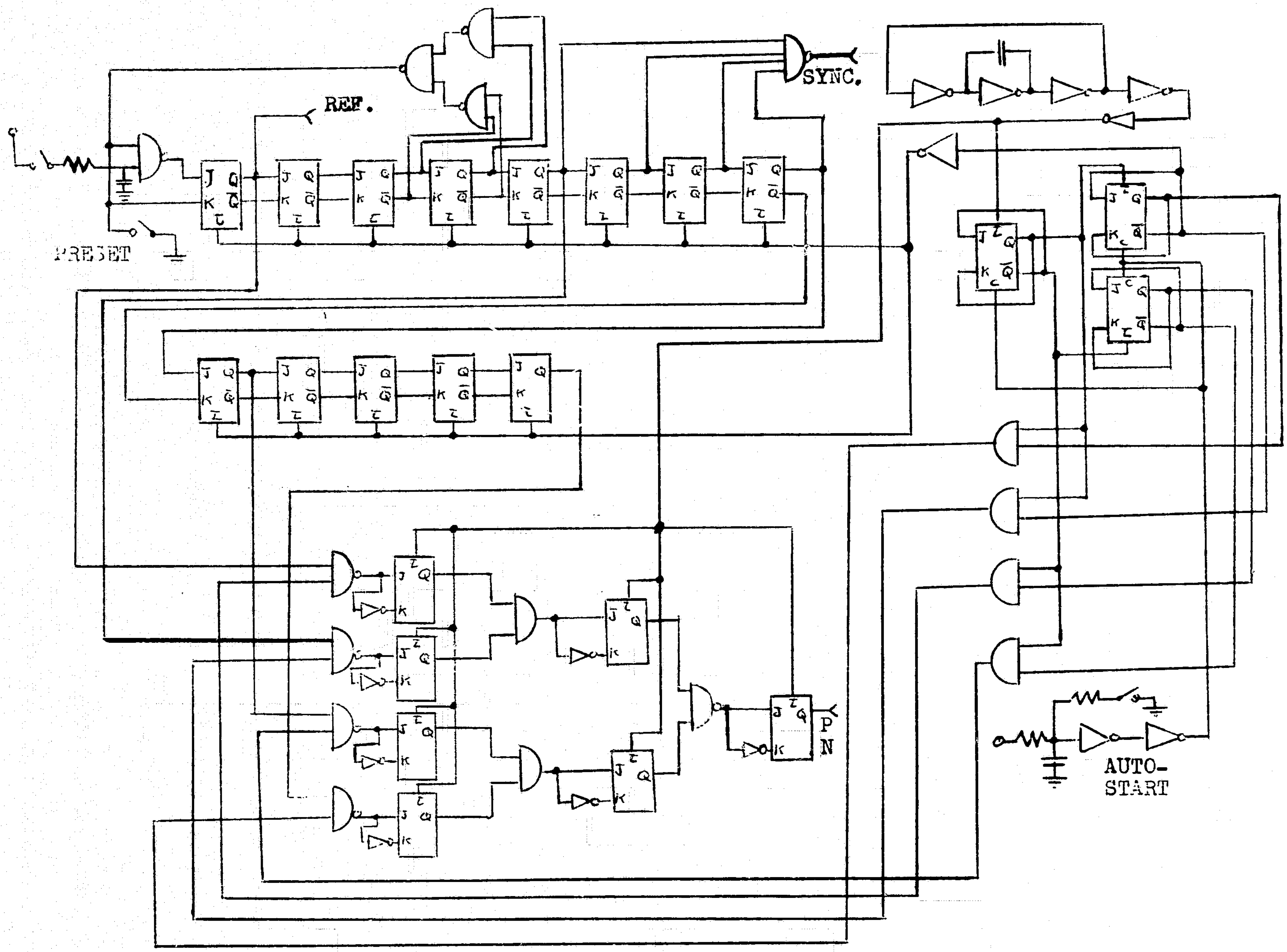


Figure 6-2 Sequence Generator Circuit

Figures 6-3, 4, and 5 are photographs of an oscilloscope display of sequences generated by the circuit shown in figure 6-2. In each case the upper trace is the low-speed sequence generated by the feedback shift register, and the lower trace is the high-speed sequence generated by the sampling technique. The rate of the low speed sequence corresponds to a 2.5 M BITS/SEC clock, and the rate of the high speed sequence corresponds to a 10 M BITS/SEC clock.

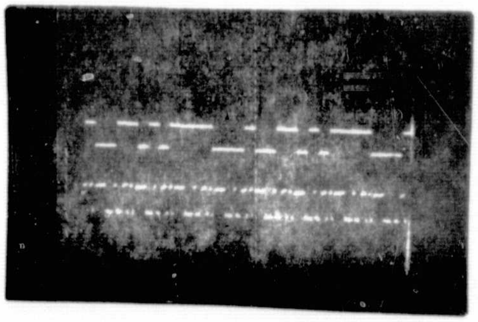


Figure 6-3

Sequence Oscilloscope Trace
Verticle Scale: 5 V/DIV
Horizontal Scale: 1 μSEC/DIV

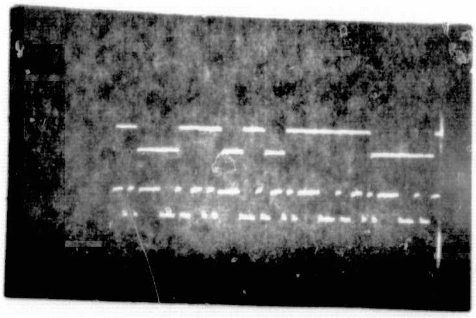


Figure 6-4

Sequence Oscilloscope Trace
Verticle Scale 5V/DIV
Horizontal Scale .5 μSEC/DIV

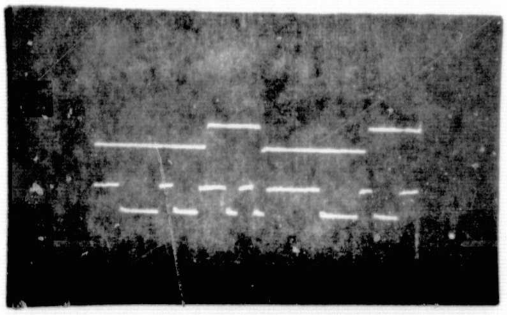


Figure 6-5

Sequence Oscilloscope Trace
Verticle Scale 5V/DIV
Horizontal Scale .2 μSEC/DIV

6-B 250 M BIT/SEC SEQUENCE GENERATOR USING EMITTER-COUPLED-LOGIC

The next step in the development of high speed pseudo-random coders for space application was the construction of a 250 M BIT/SEC sequence generator. This unit was designed to utilize the sequence multiplexing technique as was the previous demonstration breadboard. The coder was designed with emitter-coupled-logic to obtain the sequence speed of 250 M BIT/SEC. The coder was configured with two (4,3,0) ML code generators designed with the MECL 10k logic family. These generators, operating at 125 MHz are multiplexed to 250 MHz. The x2 multiplexer includes MECL 10k components and one MECL III component.

The coder, designed for evaluation purposes, uses MECL 10k components with their controlled edge speeds so that wire-wraping could be used in the breadboard. The faster edge speeds of the MECL III logic prohibits wire-wrap, and controlled impedance lines must be used.

Figure 6-6 is the electrical schematic of the 250 M BIT/SEC coder. The fundamental sequence generator includes the first Mc 10141 four - bit universal shift register and the Mc 10107 2-input exclusive-or/exclusive-nor. The MC 10131 type-D master-slave flip-flop and the second MC 10141 provide five clock periods of sequence delay for the multiplex operation. The MC-10104 Quad 2-input AND gate chip provides a logical AND of each stage of the sequence generator and gives an indication of the all-1 condition of that register. This provides a synchronization pulse once each repetition of the code sequence.

The MC 10216 triple line receiver is configured to operate as a clock generator. The second MC 10104 and the MC 1690 UHF prescaler type D

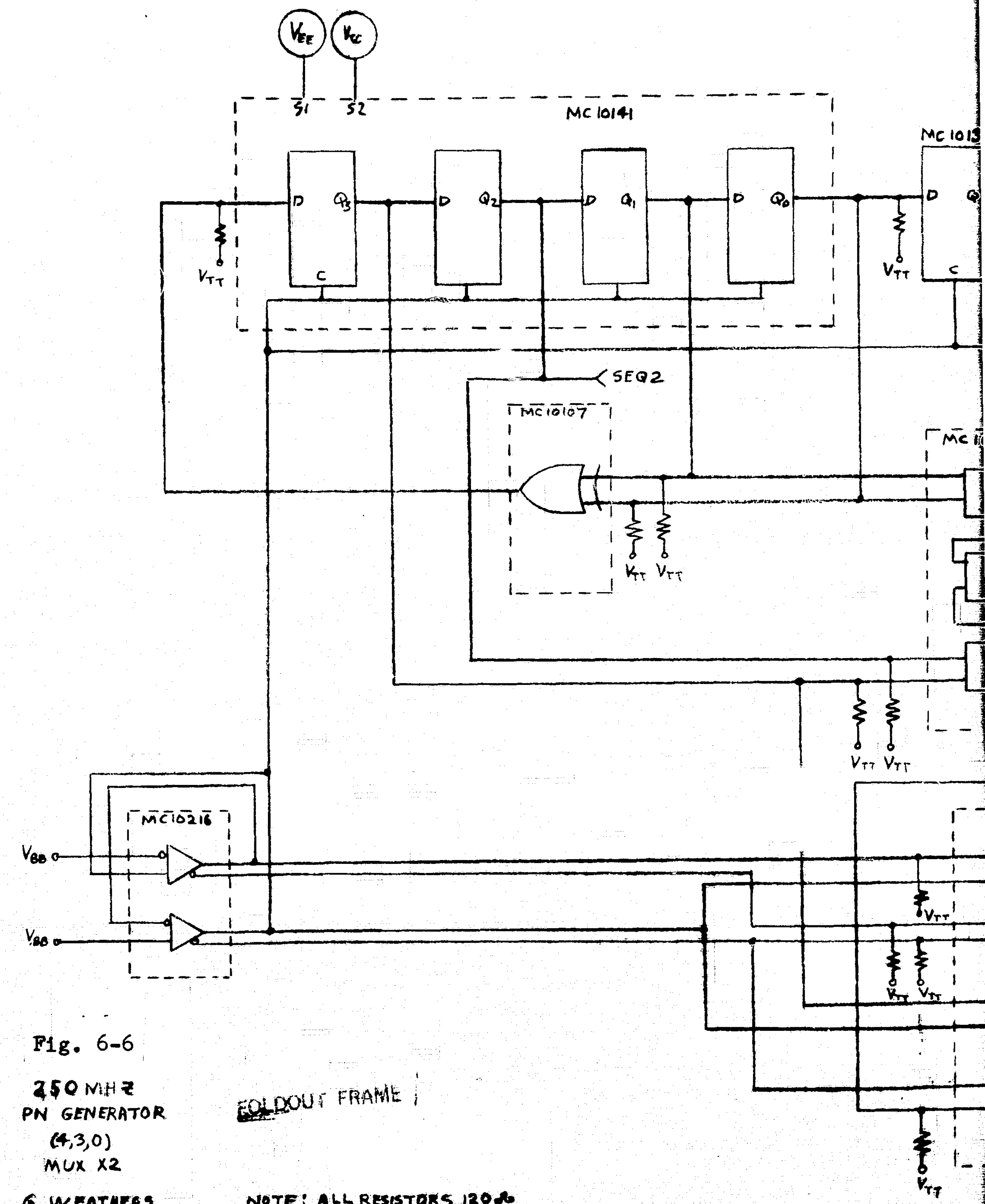


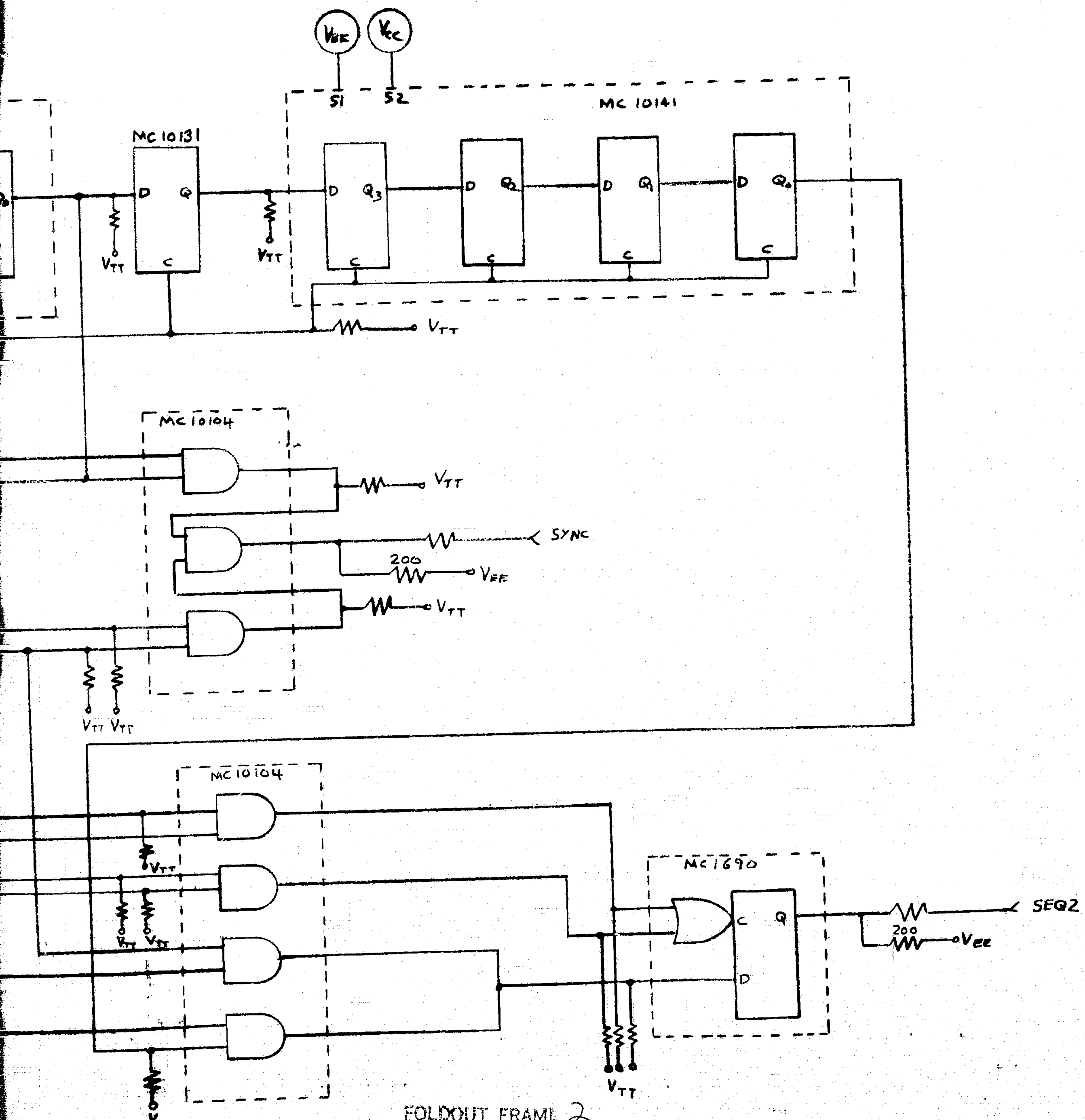
Fig. 6-6

250 MHz
PN GENERATOR
(4,3,0)
MUX X2

G. WEATHERS
2/14/75

FOLDOUT FRAME

NOTE: ALL RESISTORS 120Ω
UNLESS OTHERWISE
INDICATED



FOLDOUT FRAME 2

flip-flop perform the multiplexing operation operation of the two 125 MBIT/SEC sequences to provide the 250 M BIT/SEC sequence output.

The phasing function for the x2-Mux operation for the (4,3,0) code is

$$P = -\frac{(i-1)\ (L + 1)}{K} = -\frac{(i-1)(16)}{2} = -(i-1)\ (8)\ .$$

The set of sampled sequences are:

$$S$$
$$S\ z^{-8}$$

S is the sequence out of stage 1 of the first MC 10141 and Sz^{-8} is the sequence out of the fourth stage of the second MC 10141.

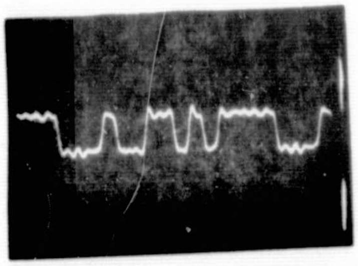
The clock (MC 10216) provides four clock phases for the multiplex operation. These are labled CK, $\overline{CK}$, CKQ, and $\overline{CKQ}$. The third and fourth AND gates of the MC 10104 form CK•S and $\overline{CK} \cdot Sz^{-8}$. The outputs of these two AND gates are combined using the "wired-or" capability of the MECL 10 K family to provide

$$CK \cdot S + \overline{CK} \cdot Sz^{-8}$$

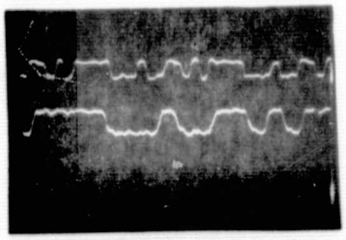
At the input to the MC 1690.

The first two AND gates of the MC 10104 in the multiplexer and the OR gate internal to the MC 1690 form an exclusive-or operation between CK and CKQ. This effectively gives a frequency doubling operation. If CK is at a 125 MHz rate, CK $\oplus$ CKQ is at a 250 MHz rate. The 250MHz synthetic clock exists only internal to the MC1690 chip. It does not exist on the wire-wrap board.

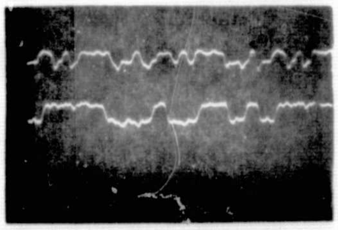
Figure 6-7 shows the waveforms of the sequence generated by the circuit for two basic clock rates. In this figure the lower trace is the basic sequence and the upper trace is the composit high rate sequence resulting from the multiplex operation. At the time of this writing the unit has been operated at a composit 200 M BIT/SEC rate. It is expected that the full 250 M BIT/SEC rate will soon be achieved.



110 MHz (4,3,0) Code
20 nsec/div. , .5 V/div.



175 MHz and 87.5 MHz (4,3,0) Code
20 nsec/div. , 1V/div.



200 MHz and 100 MHz (4,3,0) Code
20 nsec/div. , 1V/div.

Fig. 6-7

Code Generator Waveforms

Important MECL 10K and MECL III family characteristics are shown in the Table.

	MECL 10K	MFCL III
Gate propagation Delay	2 ns	1 ns
Gate edge speed	3.5 ns	1 ns
Flip-flop toggle speed	125 MHz	500 MHz
Wired-wrap capibility	Yes	No

Table 6-2 MECL 10K and MECL III Family Characteristics

From the table it is seen that a 500 MHz coder could be designed using a x4 multiplexing procedure, with four 125 MHz sequences multiplexed to 500 MHz. The four basic sequence generators could be constructed with MECL 10K components with the multiplexer being constructed with MECL III components.

7 Statistical Evaluation of Candidate Code Sequences

7A Amplitude Moments

The statistical properties of candidate code sequences for use in a **spread spectrum transponder** can be based upon the calculation of the amplitude moments of the filtered sequences. The expected value of the first five central moments of a random sequence are:

MOMENT (i)	S_c^i (i_{th} CENTRAL MOMENT)
1	0
2	M
3	0
4	M+12M(M-1)
5	0

In the above table M is the impulse response of the filter measured in bit periods of the sequence.

The central moments of a filtered pseudorandom digital sequence can be calculated from the sequence characteristic polynomial. It is assumed that the characteristic polynomial of the code sequence in question factors into primitive irreducible polynomials of order such that their code lengths are relatively prime. Making the following definitions:

B_γ: The number of trinomials of power less than or equal to M-1 that have the γ_{th} characteristic polynomial as a factor.

E_γ: The number of quadrinomials of power less than or equal to M-1 that have the γ_{th} characteristic polynomial as a factor.

F_γ: The number of peritanomials of power less than or equal to M-1 that have the γ_{th} characteristic polynomial as a factor.

The approximation for the first five central moments are:

APPROXIMATIONS FOR THE FIRST FIVE CENTRAL MOMENTS

Moment (i)	Approximation for $S_c^{\,i}$
1	0
2	M
3	$3!\sum_{\gamma=1}^{k}\left[B_\gamma L_\gamma (-1)^{k-1}\right] \Big/ \prod_{\gamma=1}^{k} L_\gamma$
4	$M + 12M(M-1) + 4!\sum_{\gamma=1}^{k}\left[E_\gamma L_\gamma (-1)^{k-1}\right] \Big/ \prod_{\gamma=1}^{k} L_\gamma$
5	$10M3!\sum_{\gamma=1}^{k}\left[B_\gamma L_\gamma (-1)^{k-1}\right] \Big/ \prod_{\gamma=1}^{k} L_\gamma$
	$+\ 5!\sum_{\gamma=1}^{k}\left[F_\gamma L_\gamma (-1)^{k-1}\right] \Big/ \prod_{\gamma=1}^{k} L_\gamma$

The approximations hold for the case $M < L_\gamma$ for all γ, $M \ll L$ where $L = \prod_{\gamma=1}^{k} L_\gamma$, and when no common trinomials, quadrinomials, and pentanomials of order M-1 or less contain the sequence characteristic polynomials as factors for each maximum-length sequence comprising the sum sequence.

A test involving the formation of a weighted sum of the difference in the first N moments for the sum sequence and a random sequence may be used to evaluate sequences from sum generators. This can be expressed as

$$T(M) = \sum_{i=1}^{N} \omega_i \left(S^i{}_c - S_{cr}^{\,i}\right) , \tag{1}$$

where $S_{cr}^{\,i}$ is the ith-central moment for weights of M-tuples from a random sequence, ω_i is a weighting factor, and $S_c{}^i$ is the ith-central moment for weights of M-tuples from a pseudorandom sequence.

Using the results from the table, equation (2) becomes

$$T(M) = \frac{1}{\prod_{\gamma=1}^{k} L_\gamma} \left[\sum_{\gamma=1}^{k} \left[\omega_3 B_\gamma 3! + \omega_4 4! E_\gamma + \omega_5 (10M3! B_\gamma + 5! F_\gamma) \right] L_\gamma (-1)^{k-1} \right] . \tag{2}$$

For a particular selection of the weighting functions, the smaller the value of T(M), the better the sequence approximates a random sequence.

The weighting values can be selected to place emphasis on a particular aspect of the distribution of M-tuple weights. For example, the term

$$\omega_3 \sum_{\gamma=1}^{k} \left[B_\gamma \; L_\gamma \; (-1)^{k-1} \right] / \prod_{\gamma=1}^{k} L_\gamma$$

indicates the relative symmetry or skewing of the distribution. The term

$$\omega_5 \sum_{\gamma=1}^{k} \left[(10M3! \; B_\gamma + 5! \; F_\gamma) \; L_\gamma \; (-1)^{k-1} \right] / \prod_{\gamma=1}^{k} L_\gamma$$

indicates skewing of the distribution with more emphasis on the shape of the distribution of M-tuple weights beyond the variance of the distribution. The term

$$\omega_4 \sum_{\gamma=1}^{k} \left[4! \; E_\gamma \; (-1)^{k-1} \right] / \prod_{\gamma=1}^{k} L_\gamma$$

indicates the kurtosis of the distribution of M-tuple weights. Assuming $\omega_4 > 0$, positive values of this term indicate a leptokurtic distribution, and negative values of the term indicate a platykurtic distribution. If k is odd the distribution is leptokurtic, and if k is even the distribution is platykurtic.

A computer algorithm for evaluating the sequence test parameter, T(M), has been developed. The algorithm calculates B_γ, the number of trinomials of order M-1 or less that contains the γth-sequence characteristic polynomial or a factor; E_γ, the number of quadrinomials of order M-1 or less that contains the γth-sequence characteristic polynomial or a factor; and F_γ, the number of pentanomials of order M-1 or less that contains the γth-sequence characteristic polynomial or a factor.

Lindholm developed an efficient algorithm for calculating B_γ, and the algorithm developed for E_γ and F_γ are essentially extensions of Lindholm's method [16].

If a sequence is generated by an n-stage register, and the sequence is a maximum-length type, then any 2n-1 digits of the sequence can define the particular stages that contribute to the feedback. This is equivalent to solving n-1 simultaneous equations, since for a maximum-length sequence the last stage is always fed back. If the sequence characteristic polynomial is a factor of a trinomial of the form

$$g(x) = 1 + x^{d-c} + x^{d} , \tag{3}$$

then the sequence satisfies the recursive relation

$$x_i = x_{i-c} x_{i-d} \tag{4}$$

when the sequence is from the set {-1, +1}.

One particular content vector in a maximum-length sequence is $x_1 = 1, x_2 = 1, \cdots x_{n-1} = 1, x_n = -1$. Using this content vector as a starting point, the next M-1 content vectors are calculated using the sequence recursive relation. Then M + n digits of the sequence are known. These digits can be represented as

$$x_o, x_1, \cdots x_{M+n-1} =$$

$$-1, 1, 1, 1, \cdots -1, x_{n+1}, x_{n+2}, \cdots x_{M+n-1} . \tag{5}$$

Because the tuple

$$(x_1, x_2, \cdots x_{n-1}) = (1, 1, \cdots 1) , \tag{6}$$

and if the sequence characteristic polynomial is a factor of

$$1 + x^{d-c} + x^d , \tag{7}$$

then the tuple

$$(x_{d+1}, x_{d+2}, \cdots x_{d+n-1}) \; (x_{c+1}, x_{c+2}, \cdots x_{c+n-1})$$

$$= (1, 1, 1, \cdots 1, 1) . \tag{8}$$

If X_d is a vector representing the tuple $(x_{d+1}, x_{d+2}, \cdots x_{c+n-1})$, and similarly for X_c, the relation

$$\overset{*}{X}_d \overset{*}{X}_c = I \tag{9}$$

can be expressed where I is the identity matrix of order n-1, and $\overset{*}{X}_d$ and $\overset{*}{X}_c$ are (n-1) by (n-1) matrices with the elements of X_d and X_c respectively on the main diagonal, with all other elements equal to zero. Extending this procedure to quadrinomials and pentanomials that contain the sequence characteristic polynomial as factors the required vector relations are

$$\overset{*}{X}_d \overset{*}{X}_c \overset{*}{X}_e = I \tag{10}$$

and

$$\overset{*}{X}_d \overset{*}{X}_c \overset{*}{X}_e \overset{*}{X}_f = I. \tag{11}$$

By finding tuples for which these equations hold using the first M + n digits after the state vector (1, 1, 1, ··· -1), all trinomials, quadrinomials, and pentanomials that contain the sequence characteristic polynomial as a factor are yielded.

The computer program POLTE 1 was written to solve for the vector relations in equation (9), (10), and (11). The results from this program can be used to evaluate T(M) from equation (2). The procedure is as follows:

(a) Select M, the size of the M-tuple

(b) Select k sequences to form the sum sequence

(c) Using the computerized algorithm, calculate B_γ, E_γ, and F_γ

(d) Select the set of weightings, ω_i, depending on the characteristics of the distribution of M-tuple weights that are critical

(e) Evaluate $T(\omega)$ from equation (1).

This procedure can be used to evaluate candidate designs for pseudorandom sequence generators of the type under study.

An indication of the reduction in the amount of calculations required to evaluate the statistics of a filtered hybrid-sum sequence as compared to a filtered maximum-length sequence can be determined as follows: The limit ratio of the number of pseudorandom sequences statistically evaluated compared to the amount of calculations required is

$$R = \frac{\prod_{i=1}^{k} \frac{2^{n_i}}{n}}{\sum_{i=1}^{K} \frac{2^{n_i}}{n}}, \tag{12}$$

where the upper limit of equation (12) is used. For k=1 the ratio is unity, but as k increases the ratio tends to increase as previously illustrated. This means the hybrid-sum sequence generator configurations can potentially provide many pseudorandom digital sequences with a minimum number of calculations required.

As an example of the potential increase in computational efficiency using the hybrid-sum approach, assume the computer algorithm was efficient enough so that each moment could be calculated in 1 second. It would then require over 12 days of computer time to completely analyze the statistics of all possible maximum-length sequences from a 23-stage register. If, however, the sequence group is established from the hybrid sum of sequences from 11- and 12-stage registers, then the analysis of sequences, which are approximately 99.9 percent as long as the maximum-length sequences from the 23-stage register, can be accomplished at a rate 120 times faster than the analysis in the maximum-length case. The more maximum-length sequences that form the hybrid-sum sequence, the greater the efficiency in forming the sequences in this manner.

As an illustration of the theory presented, a comparison is made

8-2

of the statistics of a filtered maximum-length sequence from the 11-stage generator, and a filtered hybrid-sum sequence from a 5- and 6-stage generator. The filter impulse-response length is assumed to be 20 digital clock periods. The 11-stage maximum-length sequence generator is described by the polynomial (11, 9, 0), and the hybrid-sum generator by the pair of polynomials (5, 2, 0) and (6, 1, 0).

Evaluation of equation (2) with

ω_3 equal to 1,

ω_4 equal to 2, and

ω_5 equal to 0

gives an indication of the skewing of the amplitude distribution of filtered pseudorandom sequences. Using the results of POLTE 1 given in Appendix B, this parameter is evaluated as follows:

FILTERED MAXIMUM-LENGTH SEQUENCE (11, 9, 0)

$$T(M=20,\ \omega_3 = 1,\ \omega_4 = 0,\ \omega_5 = 0) = 54, \tag{13}$$

indicating dominate positive skewing.

FILTERED HYBRID-SUM SEQUENCE (5, 2, 0) + (6, 1, 0)

$$T(M=20,\ \omega_3 = 1,\ \omega_4 = 0,\ \omega_5 = 0) = -8, \tag{14}$$

indicating slight negative skewing.

A computer program was written to evaluate the distribution of weights of the filtered sequence for both the maximum-length sequence and the hybrid-sum sequence directly. Figure 7-1 is the result for the maximum-length sequence. The weight distribution skews to the positive side and is a poor approximation to the normal distribution. Figure 7-2 is the result for the hybrid-sum sequence, and shows very little skewing tendency.

This remaining portion of this section contains example results of the computer program POLTE 1. The program was run for three irreducible polynomials of order 11, 6, and 5. The results of POLTE 1 can be used to evaluate the statistics of filtered, pseudorandom digital sequences using equation (2).

The procedure for evaluating equation (2) using the results from POLTE 1 is as follows:

For a given irreducible polynomial that generates a maximum-length sequence the polynomial representation is printed in binary and octal form. For example,

POLYNOMIAL	110000100000
OCTAL	0103

represents the polynomial

$$x^6 + x^5 + 1.$$

Following the polynomial octal form, the representations of the trinomials, quadrinomials, and pentanomials that contain the characteristic polynomial as a factor are printed. For example,

(7, 5, 1, 0)

represents the polynomial

$$x^7 + x^5 + x + 1.$$

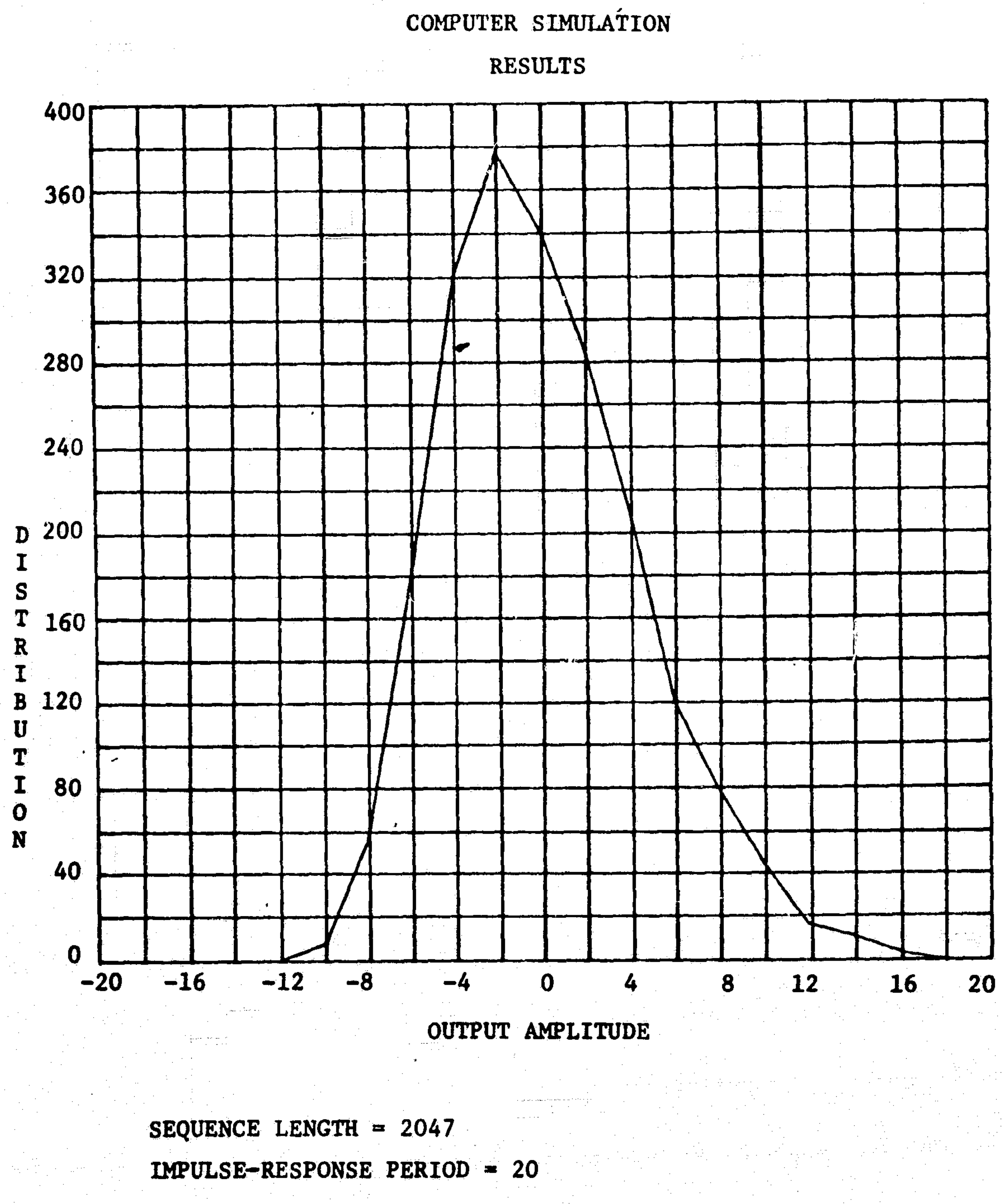
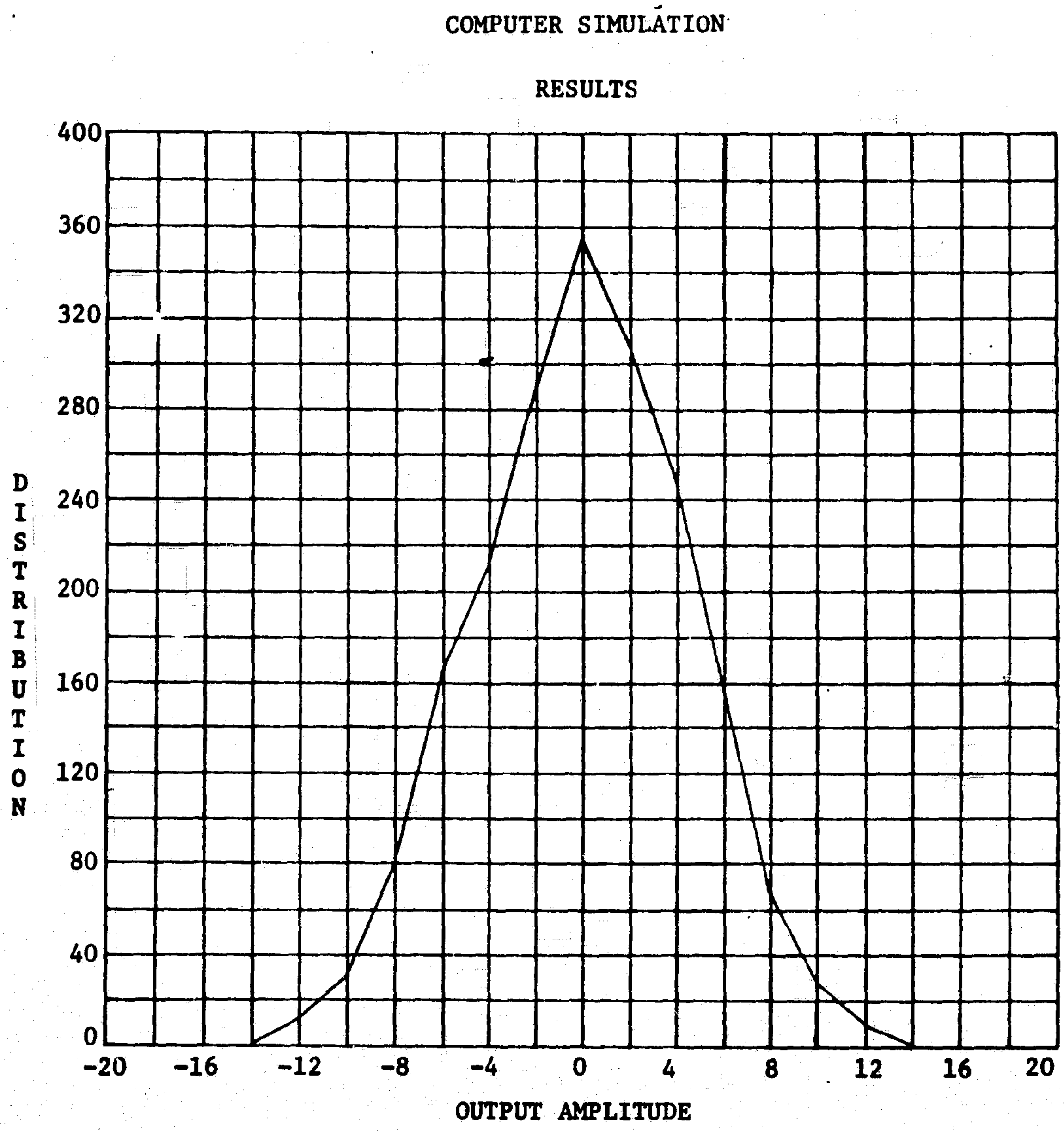


Figure 7-1 --Maximum-length sequence (11, 9, 0)



SEQUENCE LENGTH = 1953

INPULSE-RESPONSE PERIOD = 20

Figure 7-2--Hybrid-sum sequence (5,2,0)+(6,1,0)

This quadrinomial contains the characteristic polynomial,

$$x^7 + x^5 + x + 1 = (x^6 + x^5 + 1)\ (x + 1)\ (\text{mod-}2). \tag{15}$$

For a filter impulse-response period M, the parameters B_γ, E_γ, and F_γ are respectively, the number of trinomials, quadrinomials, and pentanomials of order M-1 or less of the form

$$x^p\ (x^d + x^c + 1)$$

$$x^p\ (x^d + x^c + x^b + 1)$$

and

$$x^p\ (x^d + x^c + x^b + x^a + 1)$$

that contain the γth-characteristic polynomial as a factor. In the above polynomials, p can range from 0 to M-1-d. Therefore, for each basic polynomial of the forms

$$x^d + x^c + 1,$$

$$x^d + x^c + x^b + 1,$$

or

$$x^d + x^c + x^b + x^a + 1$$

that contains the sequence characteristic polynomial as a factor, there are M-d product polynomials that also contain the sequence characteristic polynomial as a factor.

The algorithm in the computer program POLTE 1 detects the number of basic polynomials that contains the sequence characteristic equation as a factor. The parameters B_γ, E_γ, and F_γ are calculated from the results of POLTE 1 by forming the sums

$$B_\gamma = \sum_{i=1}^{D_1} (M-d_i) \tag{16}$$

$$E_\gamma = \sum_{i=1}^{D_2} (M-d_i)-f \sum_{i=1}^{M-1} (M-i) \tag{17}$$

and

$$F_\gamma = \sum_{i=1}^{D_3} (M-d_i) , \tag{18}$$

where D_1, D_2, and D_3 are the number of trinomials, quadrinomials, and pentanomials of order M-1 or less that are detected for the γth-component sequence by the program POLTE 1. If the characteristic equation of the γth sequence is a trinomial, f = 1; f = 0 otherwise.

The program POLTE 1 was used to evaluate the parameter, B_γ, for maximum-length sequences (11, 9, 0), (6, 1, 0), and (5, 2, 0), where M was equal to 20. The results are given in Table below .

POLTE 1 FOR THREE MAXIMUM-LENGTH SEQUENCES

SEQUENCE	B_γ
(11, 9, 0)	9
(6, 1, 0)	22
(5, 2, 0)	38

These values were previously used to evaluate the statistics of the maximum-length sequence (11, 9, 0) and the hybrid-sum sequence (6, 1, 0) + (5, 2, 0).

POLTE 1 was used to calculate B_γ the number of trinomials of power less than or equal to M-1 that have the γth characteristic polynomial as a factor, and E_γ, the number of quadrinomials of power less than or equal to M-1 that have the γth characteristic polynomial as a factor, for several codes. These included (5, 3, 0), (5, 4, 3, 1, 0) (5, 4, 3, 2, 0), (6, 5, 3, 2, 0) and (11, 9, 0). Figures 7-3 and 7-4 contain the results of these calculations.

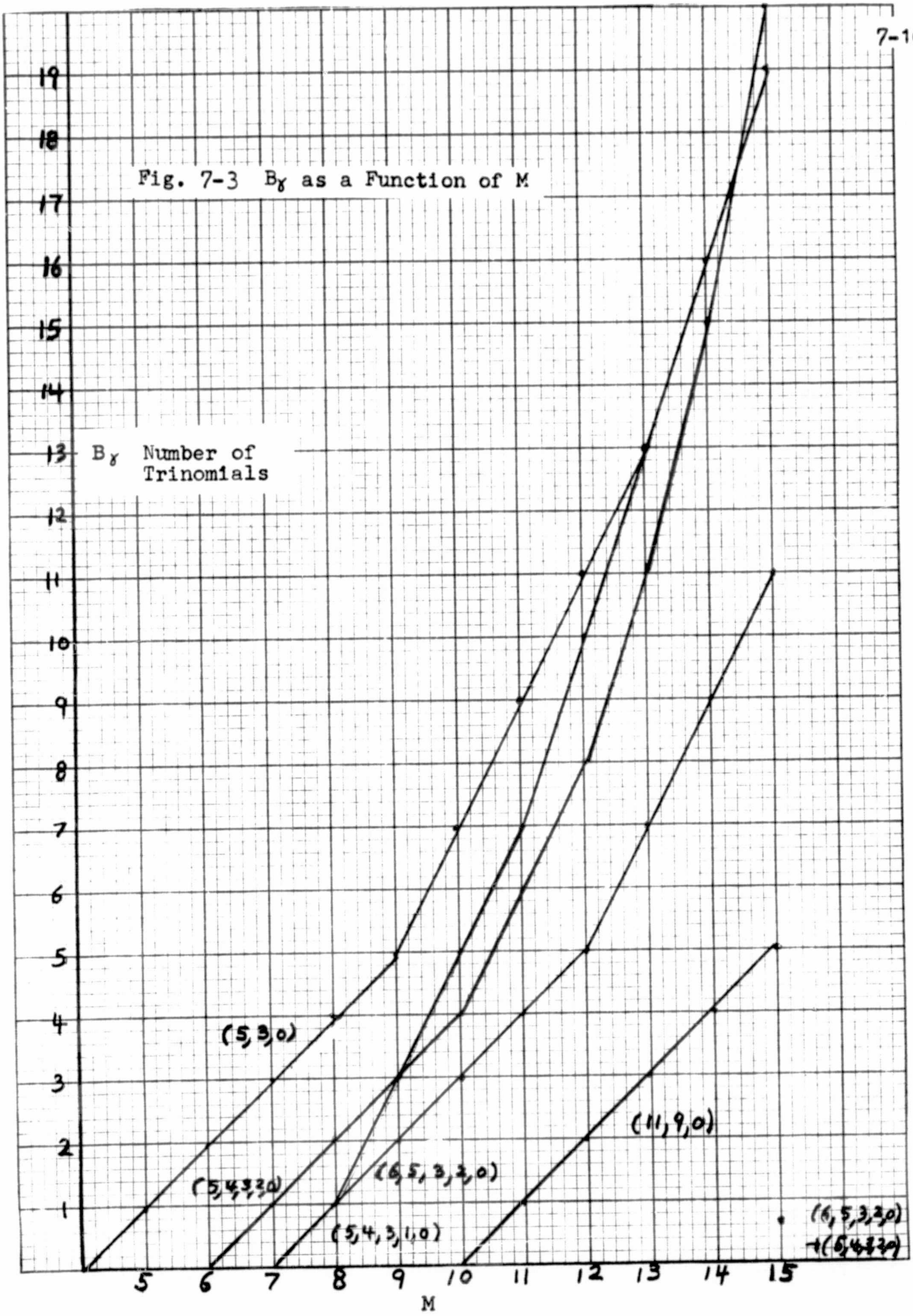


Fig. 7-3 B_γ as a Function of M

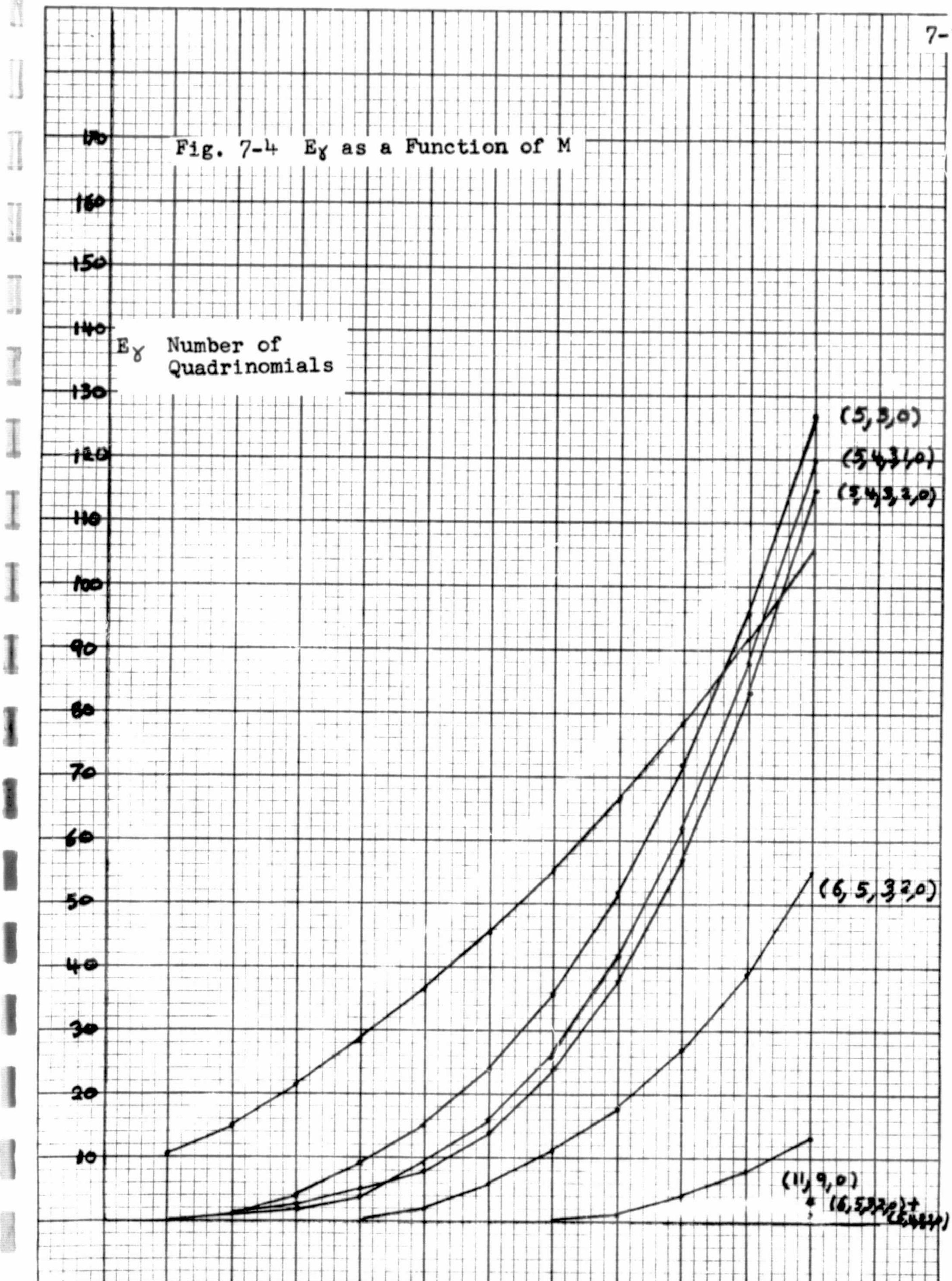
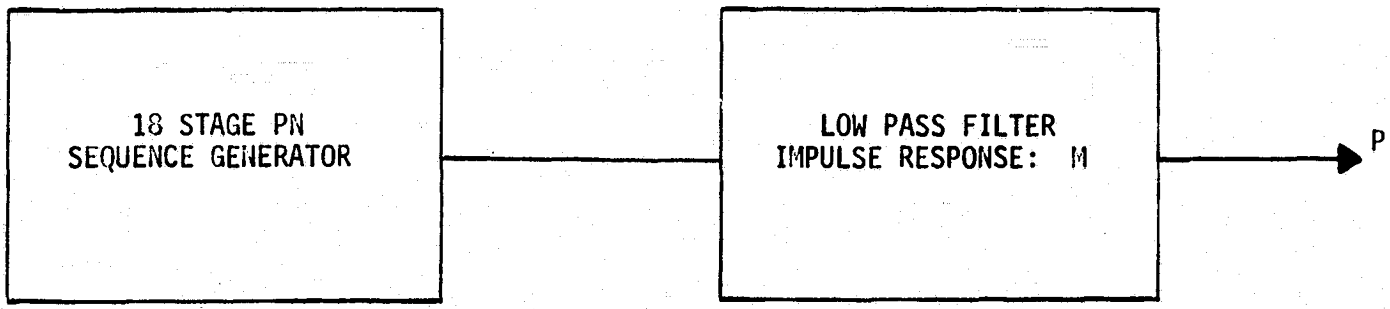


Fig. 7-4 E_γ as a Function of M

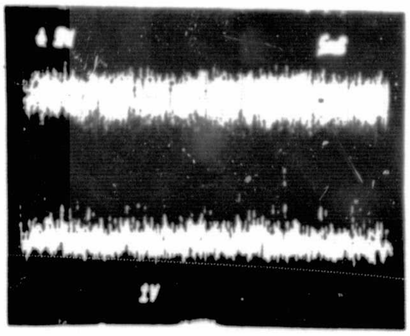
The non-random characteristics of a sequence that has significant skewing tendency was illustrated by experimentation. The arrangement is shown in figure 7-4B, and includes a eighteen stage sequence generator and a low-pass filter with adjustable cut-off frequency.



CODE LENGTH: 262143

FIGURE 7-4B CODE EVALUATION EXPERIMENTAL ARRANGEMENT

The code generated is described by the trinomial characteristic equation (18,11,0). Since the characteristic polynomial is a trinomial, it will be a factor of many trinomials of order $M > 18$. The experiment was run with impulse response length M equal to 20 and 500. Figure 7-4C shows the results of the experiment.



M = 20

M = 500

L = 262143

FIGURE 7-4C Results of Filtering the (18,11,0) code with Low-Pass Filters with Inpulse Response Lenth M = 20 and M = 500

For the case M = 500, the pulses in the positive direction only indicate positive shewing of the amplitude density function, and would be indicated by the program POLTE 1 by a large value of B_{γ}. For the case M = 20, the amplitude density function is approximately normally distributed.

The following pages give the flow diagram of the program POLTE 1

```
                              /.....................\
                              !        BEGIN        !
                              \...................../
                                         !
                                         !
               ..........................................
               ! INTEGER STAGE,STATE,COUNT,SHIFT,PART,A,B,
               ! C,D,TUPLES,X,TT,DW,CW,BW,AW
               ! DIMENSION NN(9),N(9),MM(12),STATE(12),X(
               ! 5(5),TT(12)
               ..........................................
                                         !
                                         !---C PROGRAM FOR EVALUATION OF
                                         !   CFILTERED HYBRID SUM SEQUENCES
                                         !   C READ THE FILTER IMPULSE
                                         !   CRESPONSE PERIOD AND NUMBER OF
                                         !   CMAXIMUM
                                         !   C LENGTH SEQUENCES FORMING HYBRID
                                         !   CSUM SEQUENCE
                                         !
               ..........................................
               !                   II=0
               ..........................................
                                         !
                 C  137                  !
               /......................................../
              /            READ(5,83)M,NT              /
             /......................................../
                                         !
                                         !---C 83 FORMAT(2I5)
                                         !
               ..........................................
               !                  MNM=M
               ..........................................
                                         !
A------------------>!           DO 21 JJ=1,NT
A              ..........................................
A                                        !---C NN= OCTAL ML SEQUENCE POLYNOMIA
A                                        !   C CODE
A                                        !
A              ..........................................
A              !                  NCTRI=0
A              !                  NCQUD=0
A              !                  NCPEN=0
A              ..........................................
A                                        !---C READ CHARACTERISTIC POLYNOMIAL
A                                        !   CEQUATIONS OF COMPONENT MAXIMUM
A                                        !   C LENGTH SEQUENCES IN OCTAL FORM
A                                        !
A              /......................................../
A             /             READ(5,10) NN              /
A            /......................................../
A                                        !
A                                        !---C 10 FORMAT(9I1)
A                                        !
A B---------------->!            DO 113 I=1,9
A B            ..........................................
A B                                      !
A B            ..........................................
A B            !                   J=5-I
A B            ..........................................
A B              C  113                  !
A B------------------!           N(I)=NN(J)
A              ..........................................
                                         !
```

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OF POOR QUALITY

```
I---C TRANSFORM OCTAL POLYNOMIAL CODE]
I    [TO BINARY CODE FORM             ]

DO 13 I=1,4

M=3*I

IF(N(I))75,11,12        .LT.  .EQ.
.GT.

C   12)
NNN=N(I)

GO TO (1,2,3,4,5,6,7),NNN

C   11)
MM(M-2)=0
MM(M-1)=0
MM(M)=0

GO TO 13

C   1)
MM(M-2)=1
MM(M-1)=0
MM(M)=0

GO TO 13

C   2)
MM(M-2)=0
MM(M-1)=1
MM(M)=0

GO TO 13

C   3)
MM(M-2)=1
MM(M-1)=1
MM(M)=0
```

```
GO TO 13

C     4)
MM(M-2)=0
MM(M-1)=0
MM(M)=1

GO TO 13

C     5)
MM(M-2)=1
MM(M-1)=0
MM(M)=1

GO TO 13

C     6)
MM(M-2)=0
MM(M-1)=1
MM(M)=1

GO TO 13

C     7)
MM(M-2)=1
MM(M-1)=1
MM(M)=1

C     13)
CONTINUE
    ---[ DETERMINE THE NUMBER OF STAGES ]
       [ FROM THE BINARY CODE           ]

I=13

C     14)
I=I-1

IF(MM(I))75,14,15
.LT.  .EQ.  .GT.
```

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OF POOR QUALITY

```
     [   15]              !
  ..........................................
  !             NOSTG=I-1                  !
  !            L=2**NOSTG-1                !
  !........................................!
                          !
                          !---[ L IS THE SEQUENCE LENGTH          ]
                          !   [ NOSTG IS THE NUMBER OF STAGES     ]
                          !
                          !
   /........................................\
  <             IF(I-II)34,33,34             >----.EQ.-----------O
   \......................................../                     X
     [   34]                                                     X
    /........................................./                  X
   /           WRITE(6,61) NOSTG,I           /                   X
  /........................................./                    X
                          !                                      X
                          !---[ 61 FORMAT(/////,10X,1H 'NUMBER ] X
                          !   [OF STAGES =',I3, 13X,'MAXIMUM   ] X
                          !   [SEQUENCE LENGTH =',I6,///)      ] X
                          !                                      X
                          !                                      X
  ..........................................                     X
  !                 II=I                   !                     X
  !........................................!                     X
                          !                                      X
                          O<-------------------------------------O
     [   33]              !
  ..........................................
  !               CONTINUE                 !
  !........................................!
                          !
                          !---[ WRITE OUT POLYNOMIAL CODE IN   ]
                          !   [BINARY AND OCTAL FORM           ]
                          !
                          !
    /........................................./
   /            WRITE(6,50) MM,NN            /
  /........................................./
                          !
                          !---[ 50 FORMAT(1H 'POLYNOMIAL ',2X, ]
                          !   [I2I1,/,1H 'OCTAL',2X,9I1)       ]
                          !   [ SET INITIAL CONTENT VECTOR     ]
                          !
                          !
  ..........................................
  !            STATE(NOSTG)=-1             !
  !             NOST3=NOSTG-1              !
  !........................................!
                          !
                          !
  ..........................................
B--------------->!        DO 16 STAGE=1,NOST3                !
B !........................................!
B                         !
B    [   16]              !
B ..........................................
 ----------------!        STATE(STAGE)=1                     !
  !........................................!
                          !
                          !
  ..........................................
  !                COUNT=0                 !
  !             MAX=NNN+NOST3              !
  !........................................!
                          !
                          !---[ GENERATE FIRST M+NOSTG BITS    ]
                          !
     [  100]              !
  ..........................................
  !                INPUT=1                 !
  !........................................!
                          !
```

```
A                                        I                                            Z
A                   ...........................................                        Z
A B---------------->I           DO 200 K=1,MAX              I                        Z
A B                 I.........................................I                        Z
A B                                      I                                            Z
A B                                      I                                            Z
A B                 ...........................................                        Z
A B C-------------->I           DO 17 STAGE=1,NOSTG         I                        Z
A B C               I.........................................I                        Z
A B C                                    I                                            Z
A B C                                    I                                            Z
A B C               /.........................................\      .LT.     /------V
A B C              <       IF(MM(STAGE+1))17,17,18             >----.EQ.------O---O   Z
A B C               \........................................./                   X   Z
A B C                                .GT. I                                      X   Z
A B C                                    I                                       X   Z
A B C                                    I---C GENERATE 'INPUT' AS THE NEXT    ]  X   Z
A B C                                    I   CCOMPONENT FOR THE STATE (CONTENT]  X   Z
A B C                                    I   C VECTOR                          ]  X   Z
A B C                                    I                                       X   Z
A B C                 C    18            I                                       X   Z
A B C               ...........................................                  X   Z
A B C               I      INPUT=INPUT+STATE(STAGE)         I                  X   Z
A B C               I.........................................I                  X   Z
A B C                                    I                                       X   Z
A B C                                    O<--------------------------------------O   Z
A B C                 C    17            I                                            Z
A B C               ...........................................                        Z
A B  ---------------I              CONTINUE                 I                        Z
A B                 I.........................................I                        Z
A B                                      I                                            Z
A B                                      I---C THE X(K) ARRAY CONTAINS THE     ]      Z
A B                                      I   CFIRST N*NOSTG BITS               ]      Z
A B                                      I                                            Z
A B                                      I                                            Z
A B                 ...........................................                        Z
A B                 I         X(K)=STATE(NOSTG)             I                        Z
A B                 I.........................................I                        Z
A B                                      I                                            Z
A B                                      I---C SET THE VARIABLE X(K) TO STATE(]      Z
A B                                      I   C1) FOR EACH SHIFT                ]      Z
A B                                      I                                            Z
A B                                      I                                            Z
A B                 ...........................................                        Z
A B                 I            NSH=NOSTG-1                I                        Z
A B                 I.........................................I                        Z
A B                                      I                                            Z
A B                                      I                                            Z
A B                 ...........................................                        Z
A B C-------------->I          DO 19 SHIFT=1,NSH            I                        Z
A B C               I.........................................I                        Z
A B C                                    I                                            Z
A B C                                    I                                            Z
A B C               ...........................................                        Z
A B C               I          PART=NOSTG-SHIFT             I                        Z
A B C               I.........................................I                        Z
A B C                                    I                                            Z
A B C                                    I---C ARRAY 'STATE' CONTAINS STATE (  ]      Z
A B C                                    I   CCONTENT ) VECTOR                 ]      Z
A B C                                    I   C SHIFT BITS IN SIMULATED         ]      Z
A B C                                    I   CREGISTER                         ]      Z
A B C                                    I                                            Z
A B C                 C    19            I                                            Z
A B C               ...........................................                        Z
A B  ---------------I     STATE(PART+1)=STATE(PART)         I                        Z
A B                 I.........................................I                        Z
A B                                      I                                            Z
A B                                      I                                            Z
A B                 ...........................................                        Z
A B                 I           STATE(1)=INPUT              I                        Z
A B                 I             INPUT=1                   I                        Z
A B                 I.........................................I                        Z
A B                                      I                                            Z
A B                   C    200           I                                            Z
A B                 ...........................................                        Z
A  -----------------I              CONTINUE                 I                        Z
A                   I.........................................I                        Z
A                                        I                                            Z
```

```
C TEST FOR POLYNOMIALS
C THAT CONTAIN FACTORS OF THE
CCHARACTERISTIC EQUATION

N21=NOSTR+2
M21=NNM+1

DO 21 D=N21,M21

D2=D-2
ND22=D-1

DO 22 C=3,ND22

C TEST FOR TRINOMIALS

C2=C-2

C GENERATE TUPLES
C TEST TUPLE LENGTH IS NOSTG - 1

TUPLES=1
N31=NOSTG-1

DO 31 I=1,N31

TT(I)=X(D-I+1)+X(C-I+1)

IF(TT(I))533,75,31          .EQ.
                            .GT.
C 533
TUPLES=0

C TUPLE ADDITION TEST HAS FAILED

C 31
CONTINUE

IF(TUPLES.EQ.0) GO TO 211     TRUE
FALSE
```

```
NCTRI=NCTRI+1

C WRITE DETACHED EXPONENT FORM OF
  TRINOMIAL

WRITE(6,220)DW,CW

C 220 FORMAT(5X,I9,5X,I9)

C 2113
CONTINUE
NC23=C-1

IF(C.EQ.3) GO TO 22          TRUE
FALSE

C TEST FOR QUADRINOMIALS

DO 23 B=3,NC23

BB=B-2

C GENERATE TUPLES

TUPLES=1
NN32=NOSTG-1

DO 32 I=1,NN32

TT(I)=X(D-1+1)*X(C-1+I)*X(B-1+I)

IF(TT(I))1534,75,32          .EQ.   .GT.

C 5347
TUPLES=0

C TUPLE ADDITION TEST HAS FAILED

C 393
CONTINUE
```

```
        IF(TUPLES.EQ.0) GO TO 212          TRUE
                 FALSE

        NCQUD=NCQUD+1

        ---C WRITE DETACHED EXPONENT FORM OF
             EQUADRINOMIAL

        WRITE(6,221)DW,CW,AW

        ---C 221 FORMAT(5X,I4,5X,I4,5X,I4)

  C  212
        CONTINUE
        NB24=A-1

        IF(A.EQ.3) GO TO 23                TRUE
                 FALSE

        ---C TEST FOR PENTANOMIALS

  E--->  DO 24 A=3,NB24

        AW=A-2

        ---C GENERATE TUPLES

        TUPLES=1

  F--->  DO 42 I=1,N31

        TT(I)=X(D-I+1)*X(C-I+1)*X(B-I+1)*X(A-I+1)

        IF(TT(I))44,75,42                  .EQ.
                                           .GT.
  C   44
        TUPLES=0

        ---C TUPLE ADDITION TEST HAS FAILED
```

```
[   42]
CONTINUE

IF(TUPLES.EQ.0) GO TO 213        TRUE
FALSE

NCPEN=NCPEN+1

[ WRITE DETACHED EXPONENT FORM OF]
[PENTANOMIAL                     ]

WRITE(6,222)DW,CW,BW,AW

[ 222 FORMAT(5X,I4,5X,I4,5X,I4,5X]
[I4)                             ]

[  213]
CONTINUE

[   24]
CONTINUE

[   23]
CONTINUE

[   22]
CONTINUE

[ WRITE NUMBER OF TRINOMIALS,     ]
[QUADRINOMIALS, AND PENTANOMIALS  ]
[OF ORDER M                       ]
[ -1 OR LESS THAT CONTAINS THE    ]
[CHARACTERISTIC EQUATION AS A     ]
[FACTOR                           ]

WRITE(6,313) DW,NCTRI,NCQUD,NCPEN

[   21]
CONTINUE
```

```
                              I                                                 I
                              I---[ 313 FORMAT(5X,//,13X,'M = ',I5,]            I
                              I      [,5X,'TRINOMIALS ',I10,/,5X, 1'    ]       I
                              I      [QUADRINOMIALS ',I6,/,5X, 2'       ]       I
                              I      [PENTANOMIALS ',I6,////)           ]       I
                              I                                                 I
                              I                                                 I
                    /.....................\                                     I
                    I        GO TO 193      I-----------------------------0     I
                    \...................../                               X     I
                                                                          X     I
                                                                          X     I
                              O<------------------------------------------------O
     [   75]                  I                                           X
      /........................................./                         X
     /                  WRITE(6,173)          /                           X
    /......................................../                            X
                              I                                           X
                              I---[ 173 FORMAT(5X,'ERROR CONDITION ]      X
                              I      [)                            ]      X
                              I                                           X
                              I                                           X
                    /.....................\                               X
                    I        GO TO 193      I                             X
                    \...................../                               X
                                                                          X
                              O<------------------------------------------O
     [   193]                 I
    ...........................................
    I I              CALL EXIT              I I
    I.I.....................................I.I
                              I
                              I
                    /.....................\
                    I          END          I
                    \...................../
```

7B Statistical Method for the Analysis of the Phase Distribution of Harmonic Components of Potential Spread-Spectrum Radar Codes

The computer program POLTE 1 can be used to evaluate the statistical properties of pseudorandom codes for potential transponder application. POLTE 1 calculates the third, forth, and fifth central moments of a filtered sequence where the filter impulse response period is M. The algorithm requires that an array of size M be generated (M bits from the sequence) and a search algorithm operater on this array. The number of iterations required are

THIRD MOMENT: $$\sum_{K=N}^{m} (K-1)$$

FOURTH MOMENT: $$1/2 \sum_{K=N}^{m} (K-1)\ (K-2)$$

FIFTH MOMENT: $$1/4 \sum_{K=N}^{m} (k-1)\ (K-2)\ (K-3)$$

For large values of M this requires a potential large number of iterations. In the above equations N is the order of the code. In the case of candidate codes where M is large and L is relatively small a more efficient evaluation technique is to perform a statistical analysis of the phase distribution of harmonic components of potential spread spectrum codes. The phase of the α th harmonic of a PR code is

$$\Theta\ (\alpha)\ \mathrm{TAN}^{-1} \left\{ \frac{\sum_{g=1}^{m} \sin \frac{(2\Pi\alpha g)}{L} (Ag-A-g)}{A_0 + \sum_{g=1}^{m} \cos \frac{(2\alpha\pi g)}{L} (Ag-A-g)} \right\} \qquad (1)$$

where L is the sequence length, and

$$m = \frac{L-1}{2} \tag{2}$$

Ag is the gth member of the A- array, the sequence itself.

The phase distribution statistics for a filter passing H harmonics of the code can be evaluated by evaluating.

$$(\overline{\theta(\alpha)})_H = \frac{1}{H} \sum_{\alpha=1}^{H} \theta(\alpha) \tag{3}$$

where

$$H = \frac{L}{M} \tag{4}$$

This is a first order evaluation based on the first moment of the phase distribution. Futher evaluation can be made by calculating

$$\overline{(\theta^2(\alpha))_H} \quad \text{and} \quad \overline{(\theta^3(\alpha))_H}$$

In general

$$\overline{(\theta^n(\alpha)\ )_H} = \frac{1}{H} \sum_{\alpha=1}^{H} \left[\mathrm{TAN}^{-1} \frac{\sum_{g=1}^{m} \sin\left(\frac{(2\pi\alpha g)}{L}\right) Pg}{A_0 + \sum_{g=1}^{m} \cos\left(\frac{(2\pi\alpha g)}{L}\right) S_g} \ n \right]^n \tag{5}$$

where

$$P_g = A_g - A{-}g \tag{6}$$

$$S_g = Ag + A{-}g \tag{7}$$

where Ag is from $\{-1, 1\}$

To show the utility of the method a few examples will be given.

Figure 7-1 shows the sequential output of a filtered sequence for M = 50, H = 40. As can be seen the output skews to the positive side, and the distribution function of figure 7-2 shows the skewing effect. Figure 7-3 and figure 7-4 are similiar illustrations for M = 70, H = 30. In each case the skewing is caused by a pulse embeded in the output that otherwise appears random. POLTE 1 detects this problem for the (11, 9, 0) code because it finds many trinomials of order less than or equal to M-1 that contain the sequence characteristic equation as a factor. However, for M = 70 the iterations required for POLTE 1 to evaluate the statistics are large.

Figure 7-5 shows the results of equation (1) for $\alpha \leq 300$. The phase components appear to be distributed in a random manner except for values of $\alpha \leq 20$. Figure 7-6 shows the distribution of phase for M = 80, H = 25. As can be seen there are no phase components between + 1 and = $+\pi$ radians

The sum of a pulse and random noise would tend to form a phase distribution as shown in figure 7-6 since the harmonic components of a pulse are co-phased.

Figure 7-7 and 7-8 give the output distribution for M = 20, H = 100, and M = 40, H = 50 for the hybird-sum code (5, 2, 0) + (6, 1, 0). Figure 7-9 and 7-10 give the sequential output and distribution for M = 80, H = 25. The sequence appears random and does not suffer the problems of the (11, 9, 0) code. Figure 7-11 is the phase distribution of the first 200 harmonics of (5, 2, 0) + (6, 1, 0). Figures 7-12 and 7-13 show the phase distribution for M = 80, H = 25 and M = 40, H = 50.

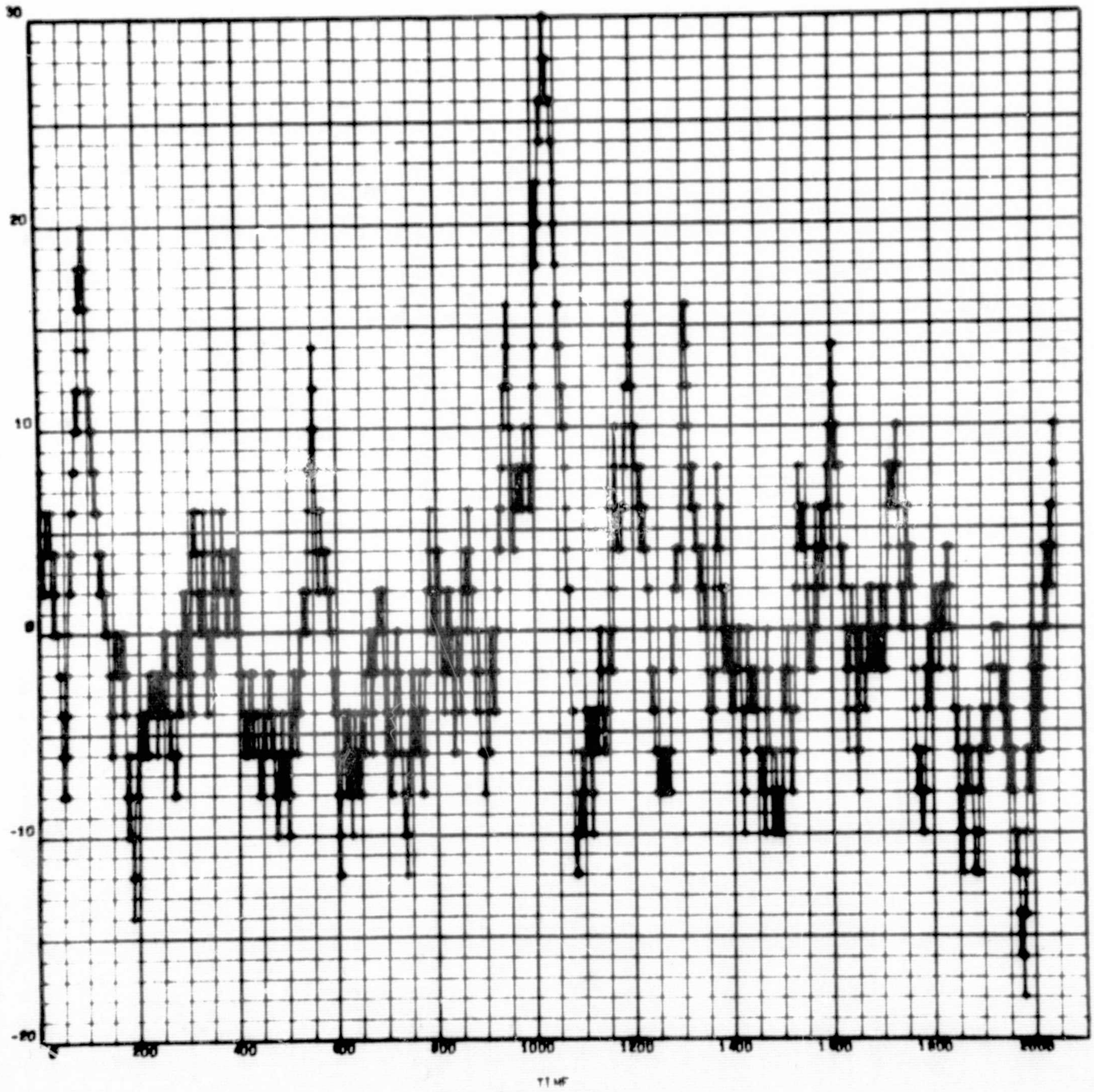
For comparison, figure 7-14 and 7-15 give the output distribution for M = 20, H = 100, and M = 40, H = 50 for the hybrid-sum code (5, 4, 2, 1, 0) + (6, 1, 0). Figure 7-16B and 7-17B give the sequential output and distribution for M = 80, H = 25. The sequence shows a pulse form in figure 7-16B and skewing in figure 7-17B. Figure 7-18B is the phase distribution of the first 200 harmonics of (5, 4, 2, 1, 0) + (6, 1, 0). Figures 7-19B and 7-20B show the phase density for M = 80, H = 25 and M = 40, H = 50. In figure 7-19B there are no phase components distributed between 0 and -1 radiaus. This distribution and the affect on the output are similar to the (11, 9, 0) code case.

A procedure for evaluation of the pseudorandomness of codes based upon their phase moments can be provided by the calculation of these phase moments and comparing to expected phase moments. The phase probability density for a random set of spectral lines is uniform between $-\pi$ and π. The phase moments for this random set of spectral lines are:

MOMENT	EXPECTED VALUE
1	0
2	$\pi^2/3 = 3.28987$
3	0
4	$\pi^4/5 = 19.5234$
5	0

The comparison of calculated phase moments for a particular sequence and value of H (harmonics passed by the filter) provides a measure of the pseudo-random quality of the sequence.

A computer program was prepared to perform the calculation of the first four phase moments. This program is called PHASE 1 and calculates the moments as a function of the number of spectral lines within the filter bandwidth. As an example of the use of PHASE 1, phase moments were calculated for the filtered sequence described by the polynomial (8, 7, 6, 1, 0). Figure 7-21 shows the plot of the first four phase moments as a function of H.



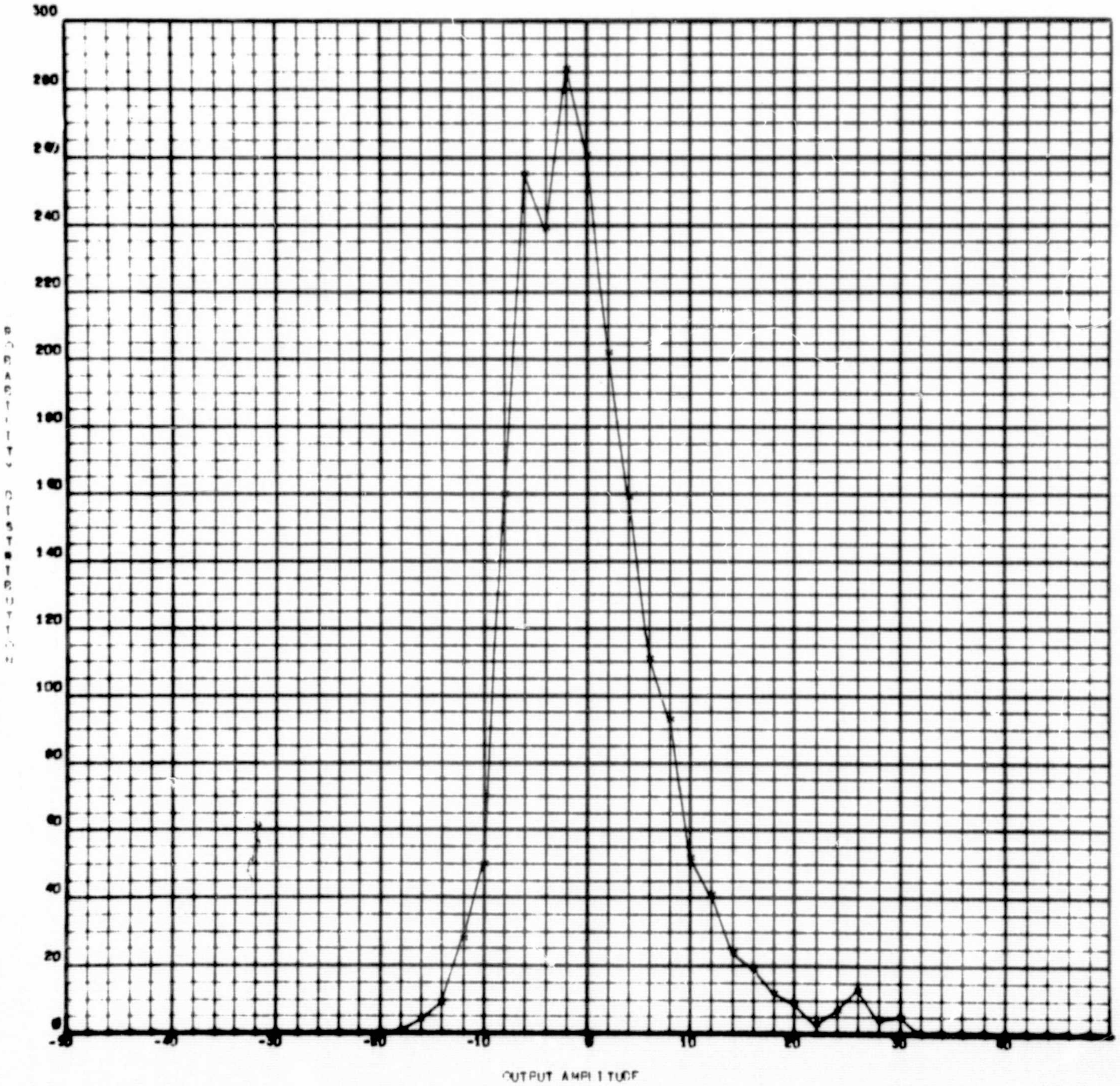
PLOT NUMBER 15

SEQUENCE LENGTH = 2047

IMPULSE RESPONSE PERIOD = 50

(11,9,0)
M=50
H = 40

FIG. 7-1B

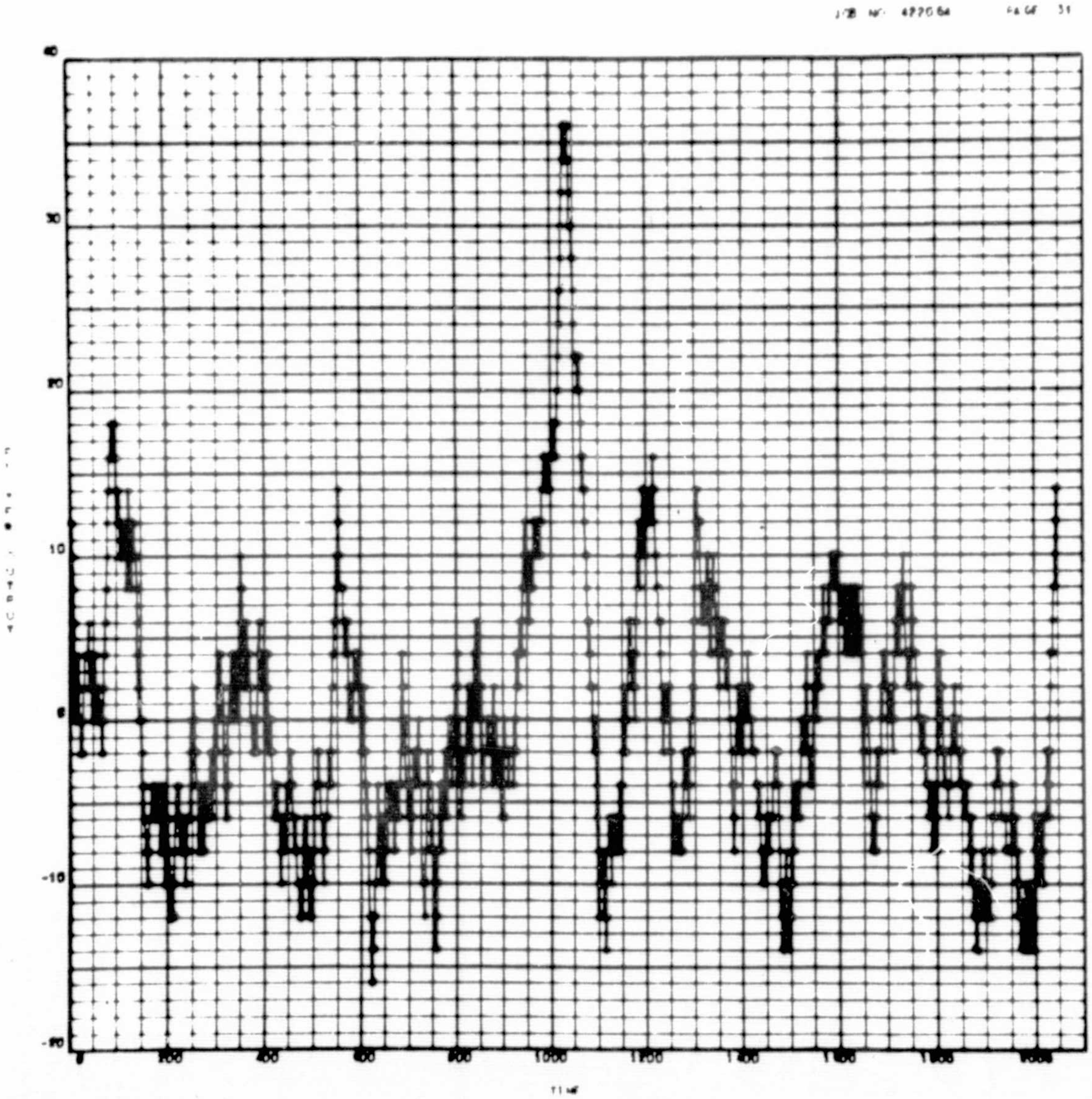


PLOT NUMBER 15

SEQUENCE LENGTH = 2047

IMPULSE RESPONSE PERIOD = 50

FIG. 7-2B



PLOT NUMBER 16

SEQUENCE LENGTH = 2047

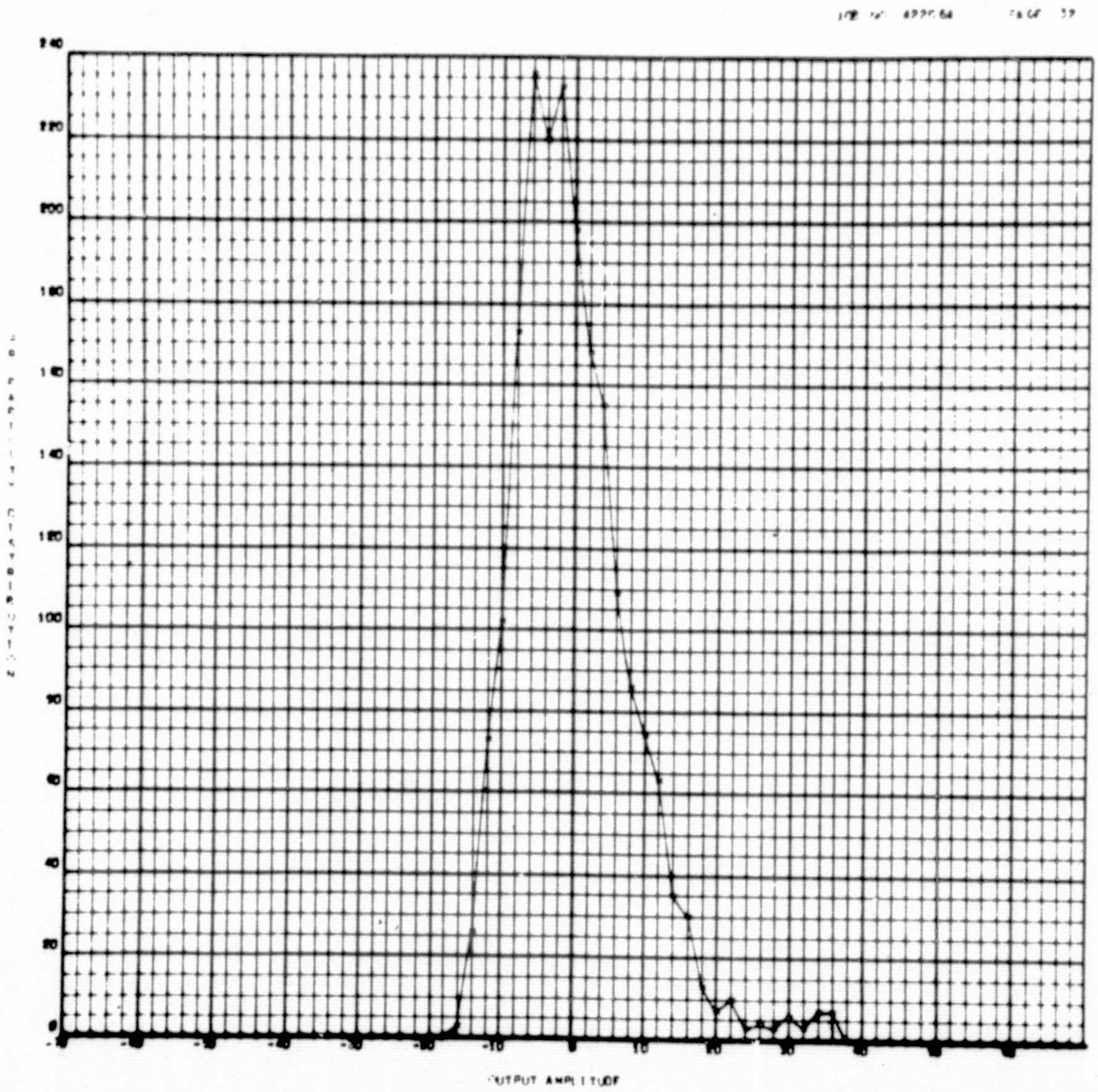
IMPULSE RESPONSE PERIOD = 70

(11, 9, 0)

M = 70

H ≈ 30

FIG. 7-3B



PLOT NUMBER 16

SEQUENCE LENGTH = 2047

IMPULSE RESPONSE PERIOD = 70

FIG 7-4B

(11,9,0)
M = 70

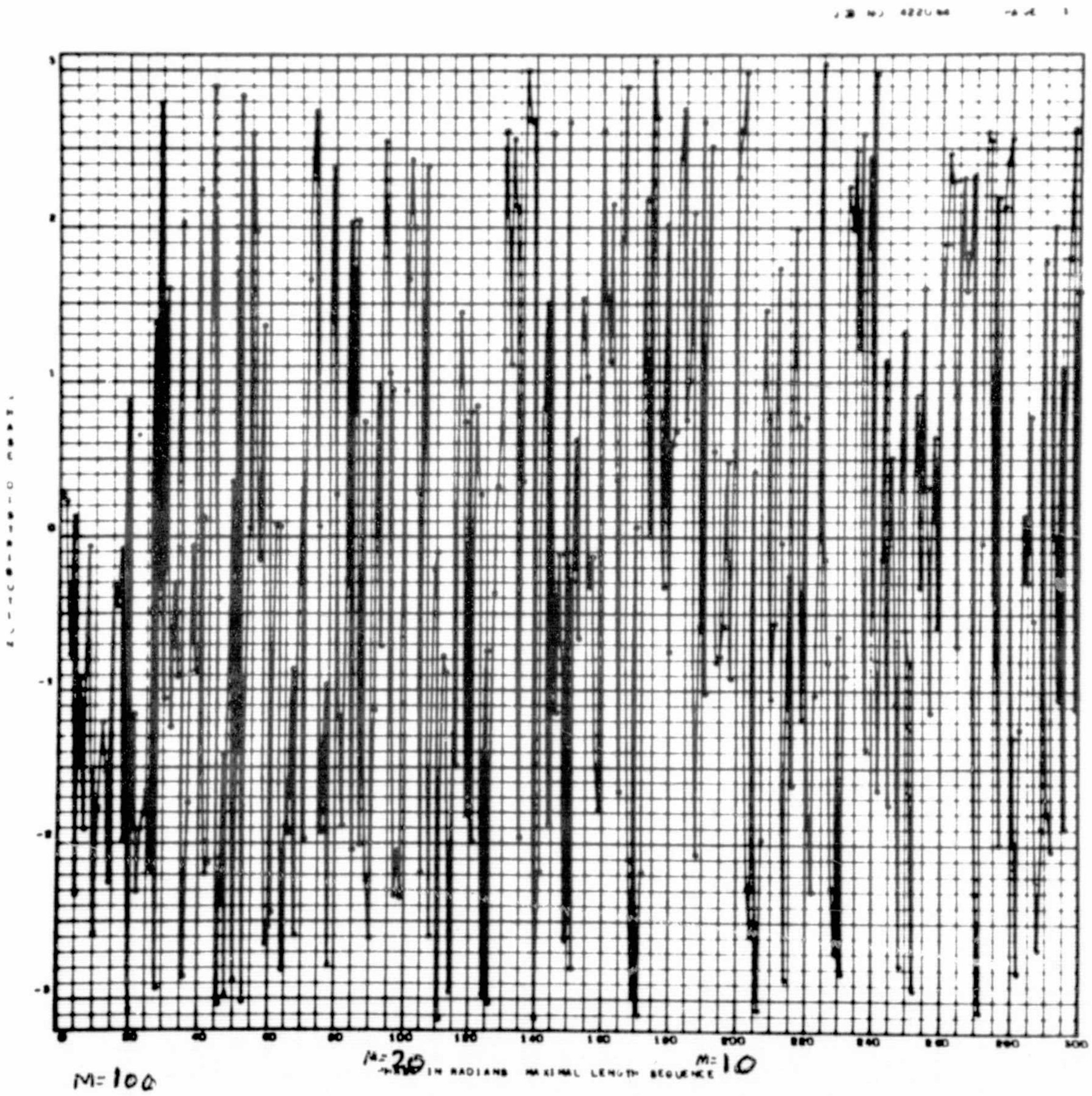
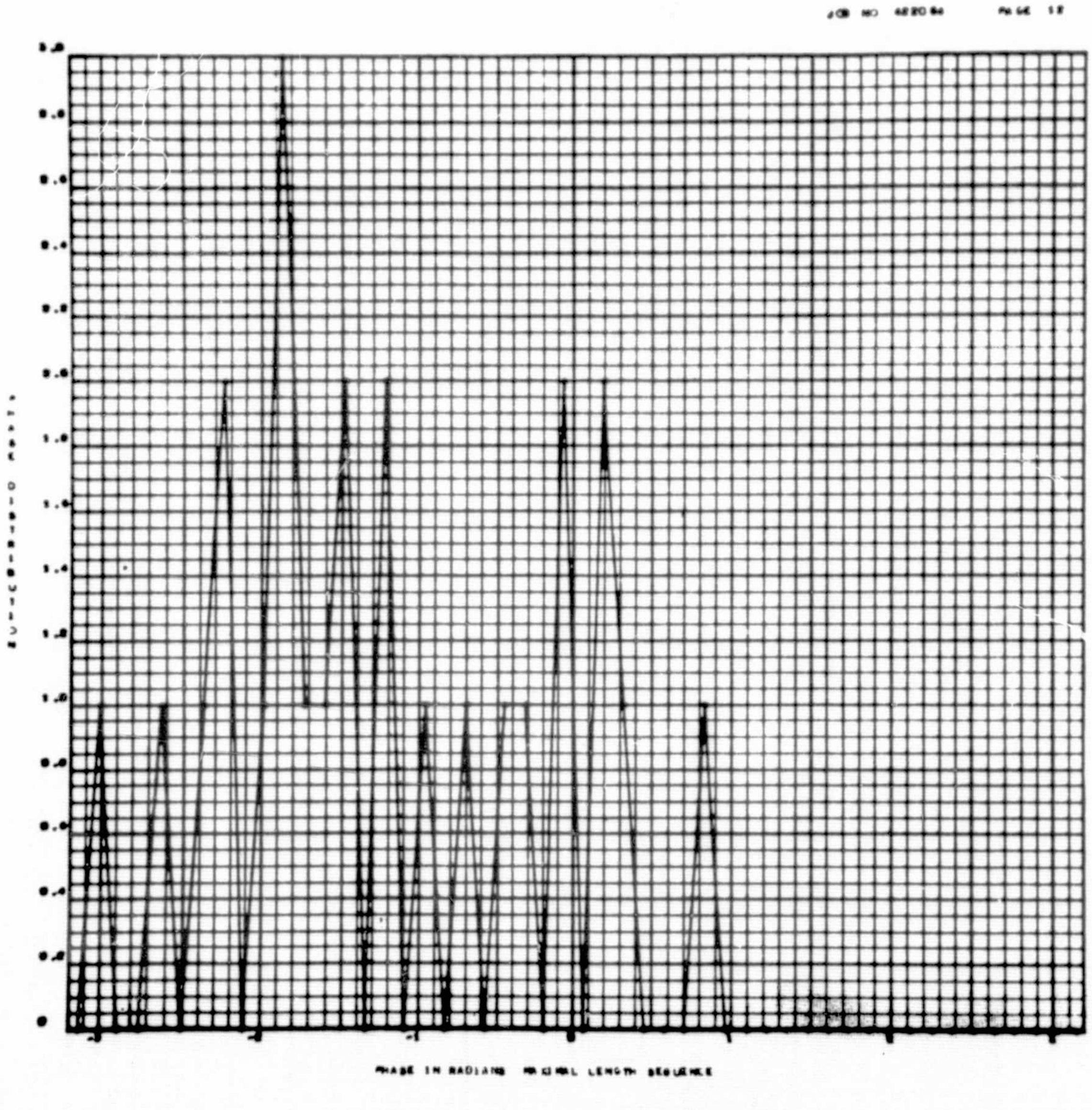


FIG 7-5B

(11,9,0)
FIRST 300 HARMONICS

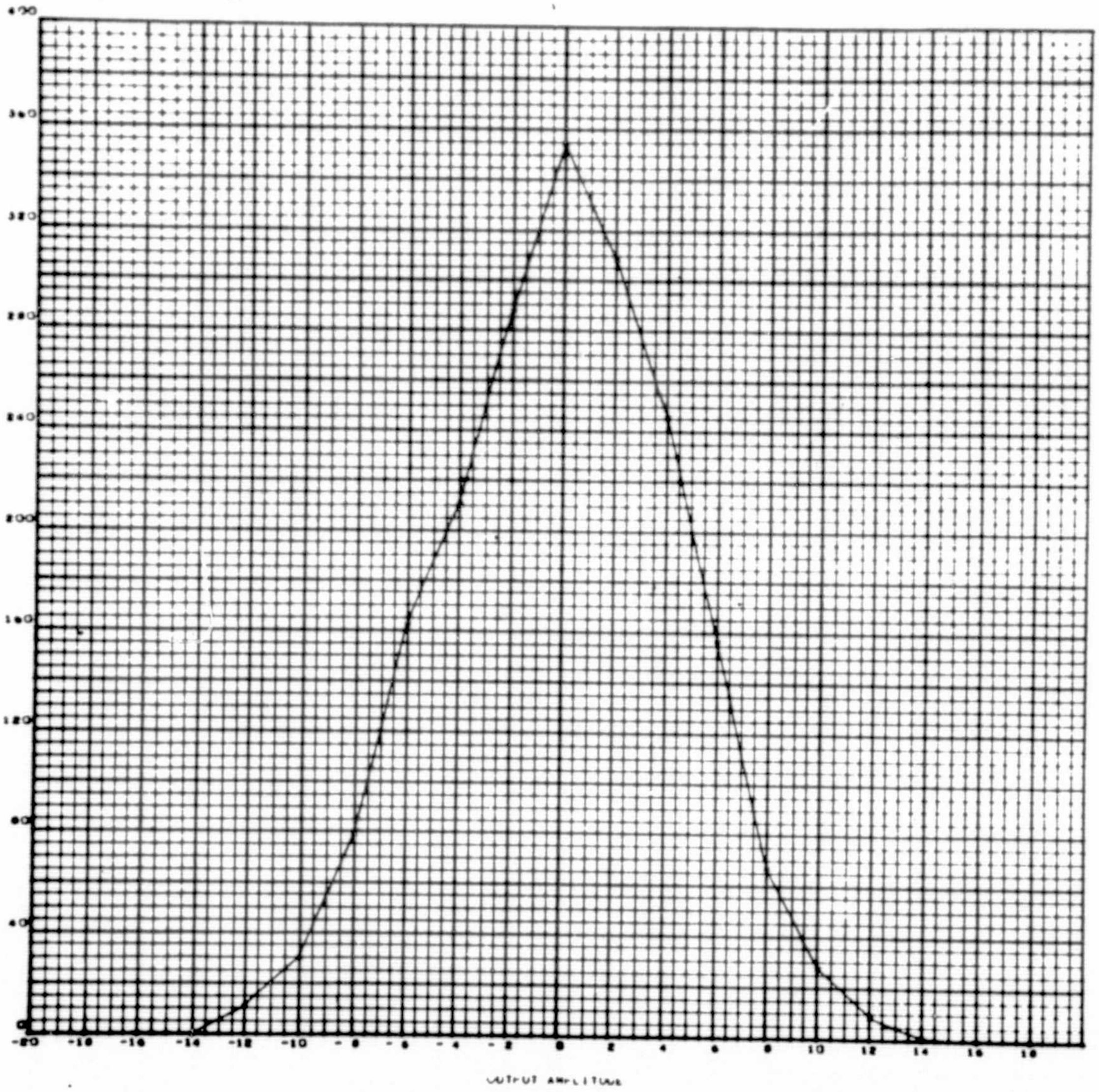


PLOT NUMBER 11

SEQUENCE LENGTH = 11
IMPULSE RESPONSE PERIOD = 25

(11,9,0)
M = 80
H = 25

FIG 7-6B



2σ

PLOT NUMBER 9

SEQUENCE LENGTH = 1953

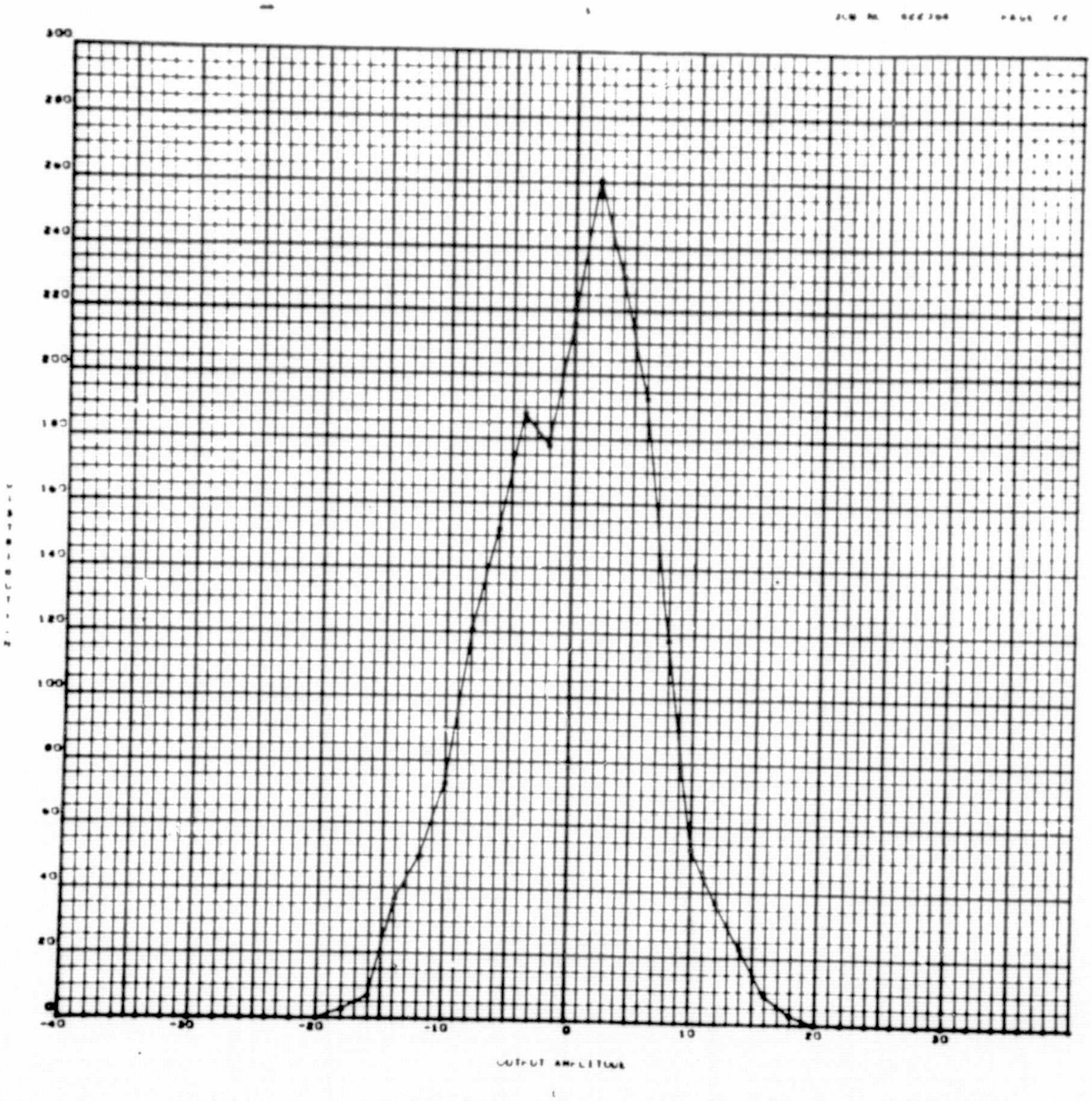
IMPULSE RESPONSE PERIOD = 20

(5,2,0) ⊕ (6,1,0)

M=20

FIG 7-7B

H ≈ 100



PLOT NUMBER 11

SEQUENCE LENGTH = 1953

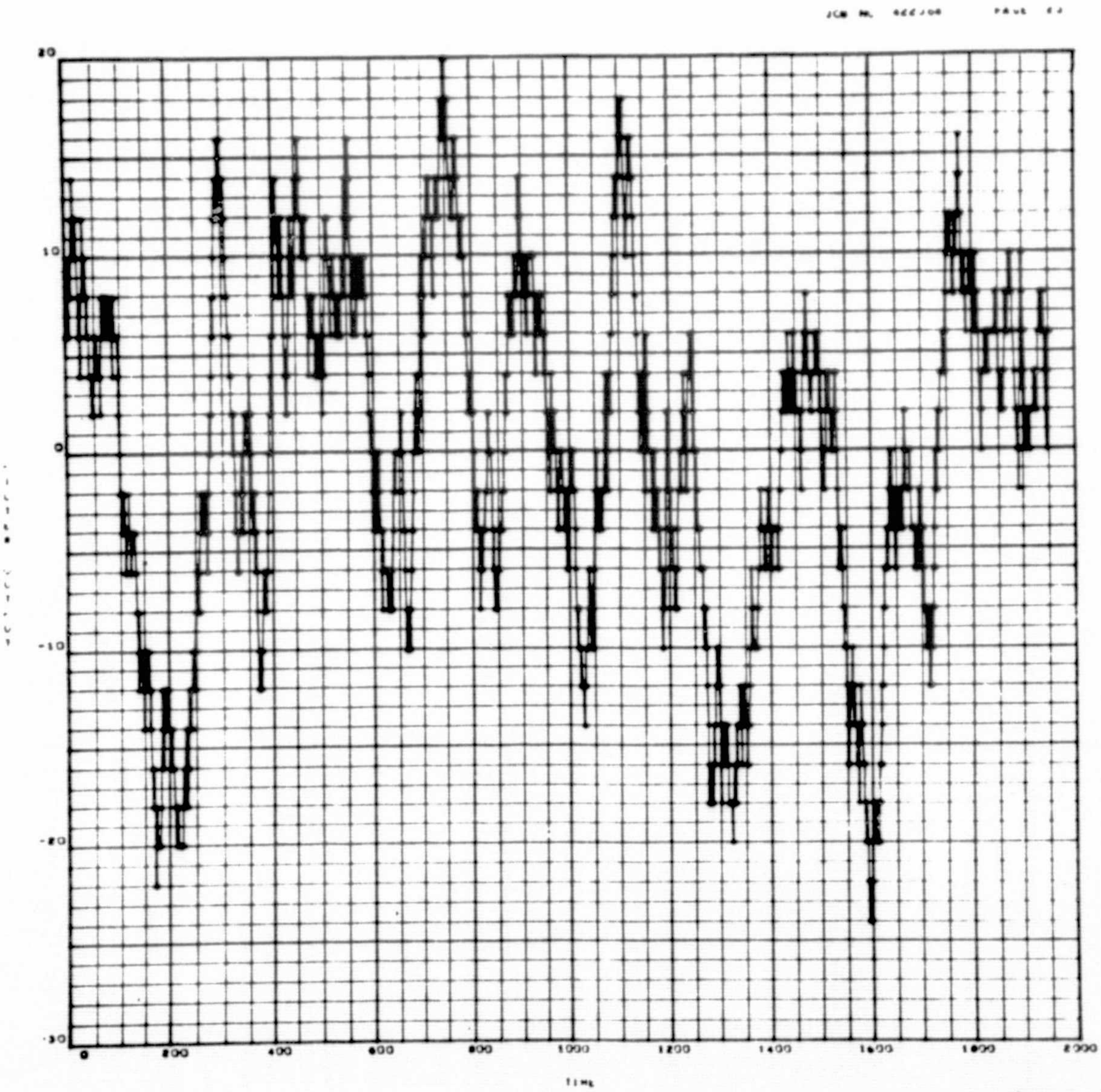
IMPULSE RESPONSE PERIOD = 40

$(5,2,0) \oplus (6,1,0)$

$M = 40$

$H \approx 50$

FIG 7-8B



$(5,2,0) \oplus (6,1,0)$

$M = 80$

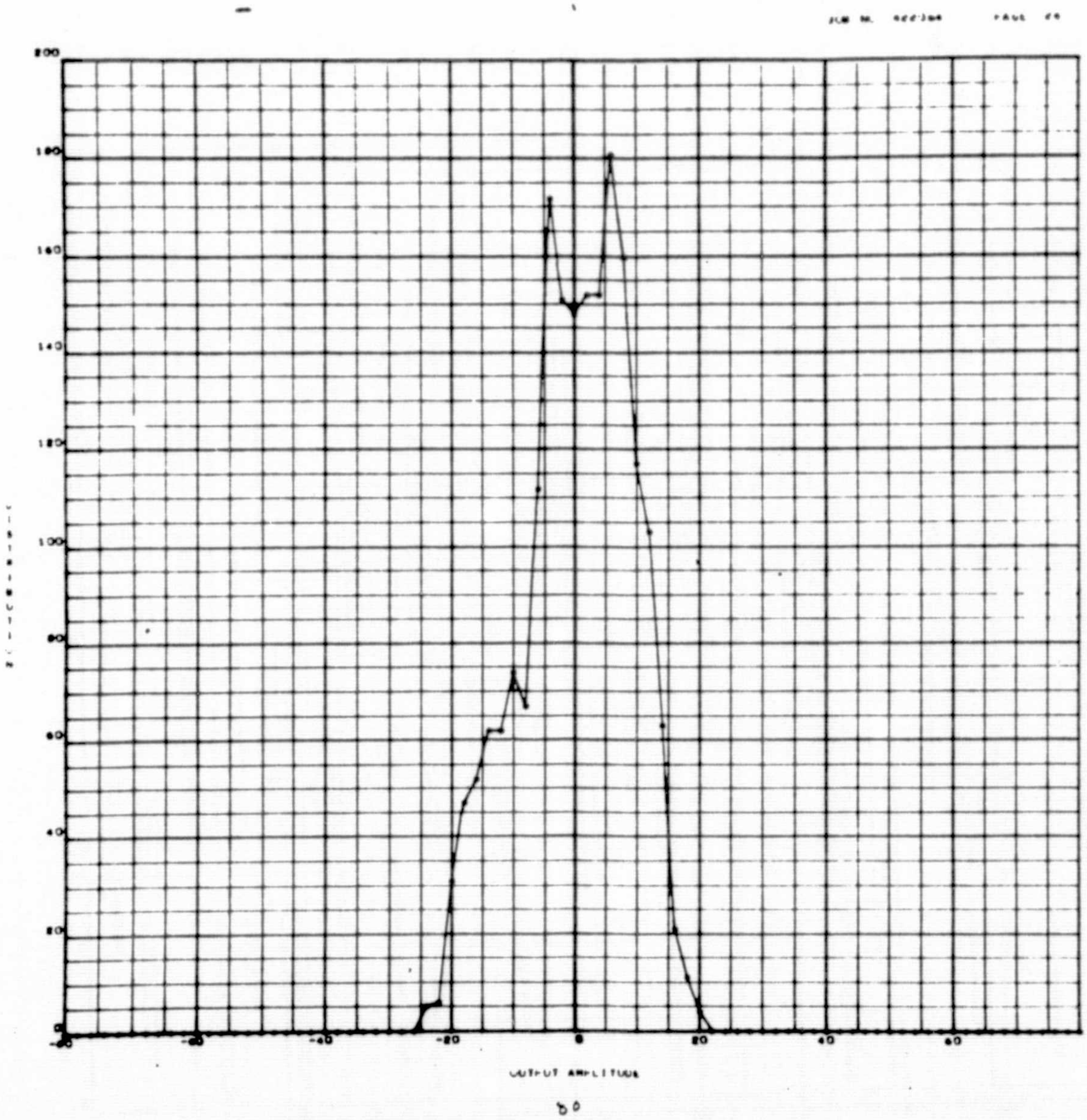
$H \approx 25$

PLOT NUMBER 12

SEQUENCE LENGTH = 1953

IMPULSE RESPONSE PERIOD = 80

FIG 7-9B



PLOT NUMBER 12

SEQUENCE LENGTH = 1953

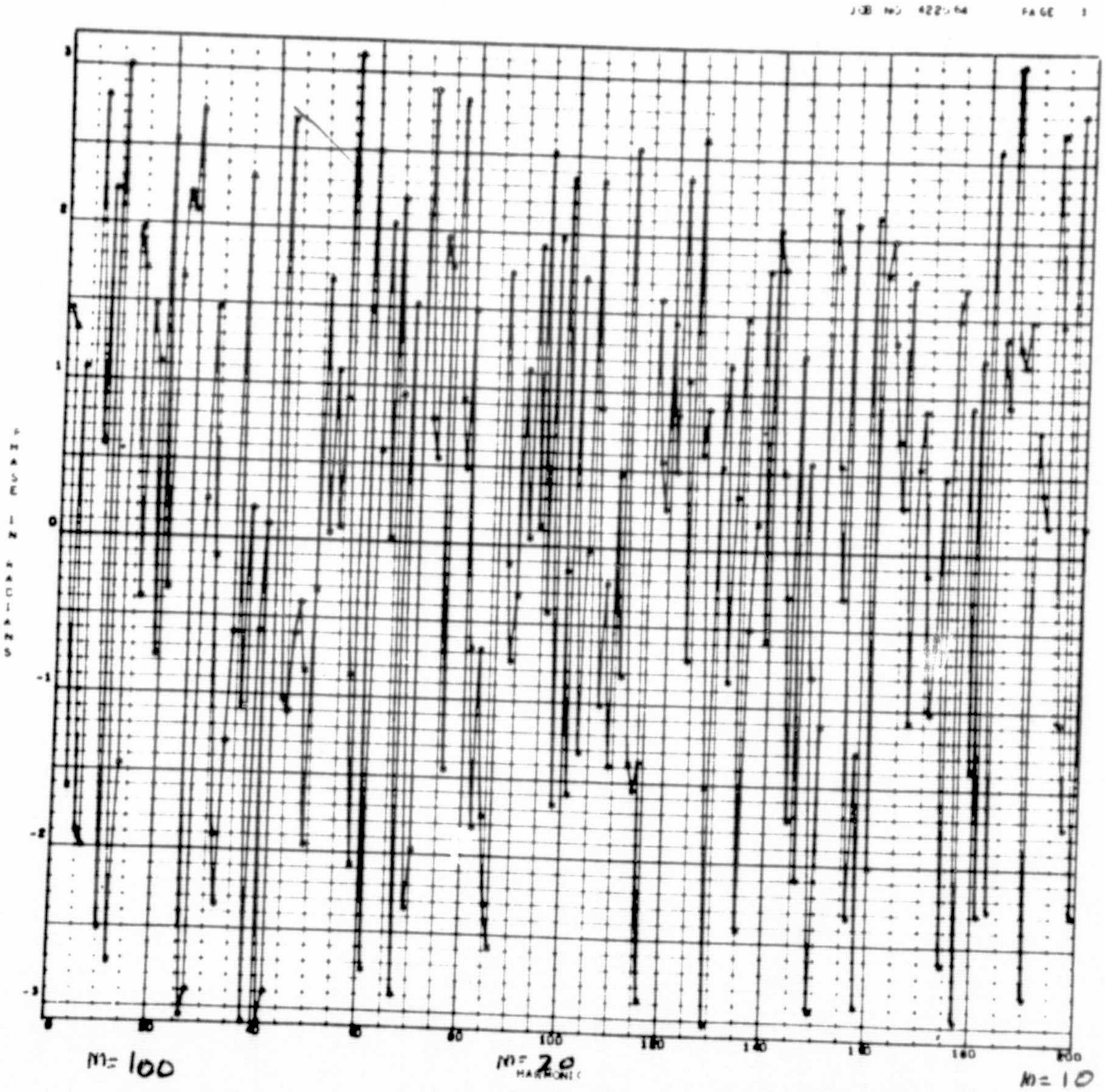
IMPULSE RESPONSE PERIOD = 80

(5,2,0)⊕(6,1,0)

M = 80

H = 25

FIG 7-10B



$(5,2,0) \oplus (6,1,0)$

FIRST 200 HARMONICS

FIG 7-118

PLOT NUMBER 3

N1 = 6

N2 = 5

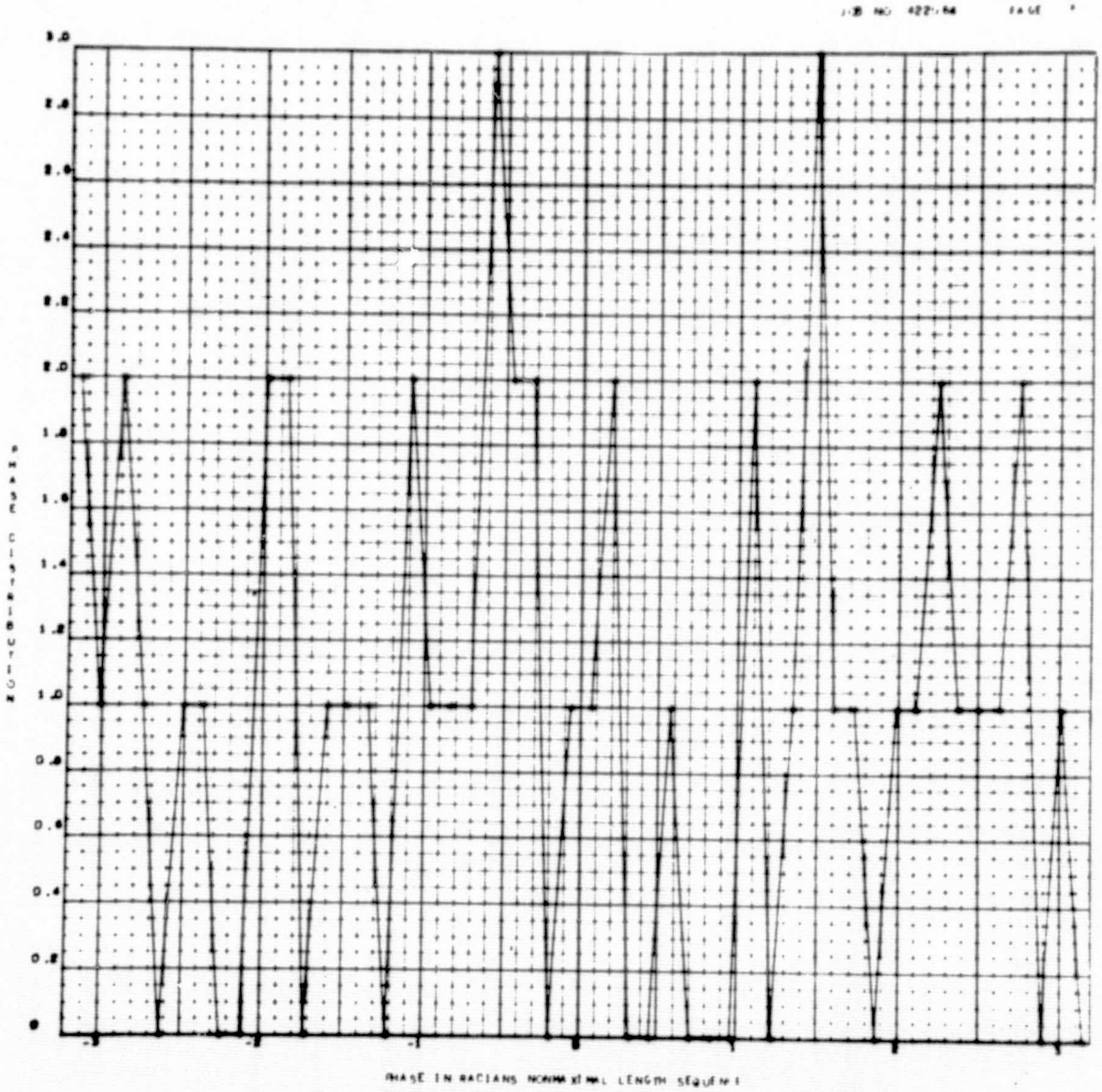
24 HARMONICS INCLUDED

(5,2,0)⊕(6,1,0)

M = 80

H = 25

FIG 7-12B



PHASE IN RADIANS NONMAXIMAL LENGTH SEQUENCE

PLOT NUMBER 5

N1 = 6

N2 = 5

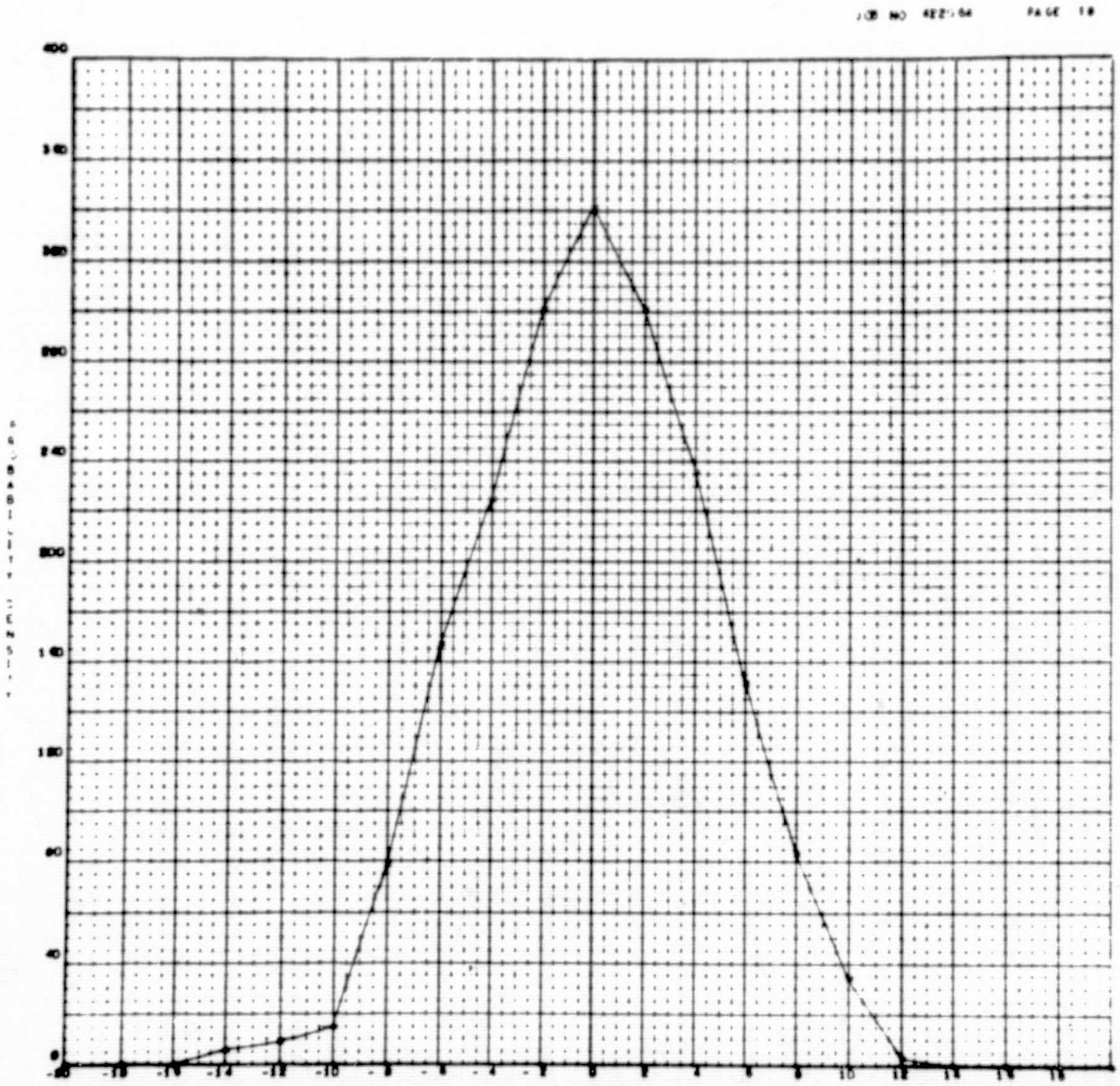
50 HARMONICS INCLUDED

$(5,2,0) \oplus (6,1,0)$

$M \approx 40$

$H = 50$

FIG 7-13B



PLOT NUMBER 9

SEQUENCE LENGTH = 1753

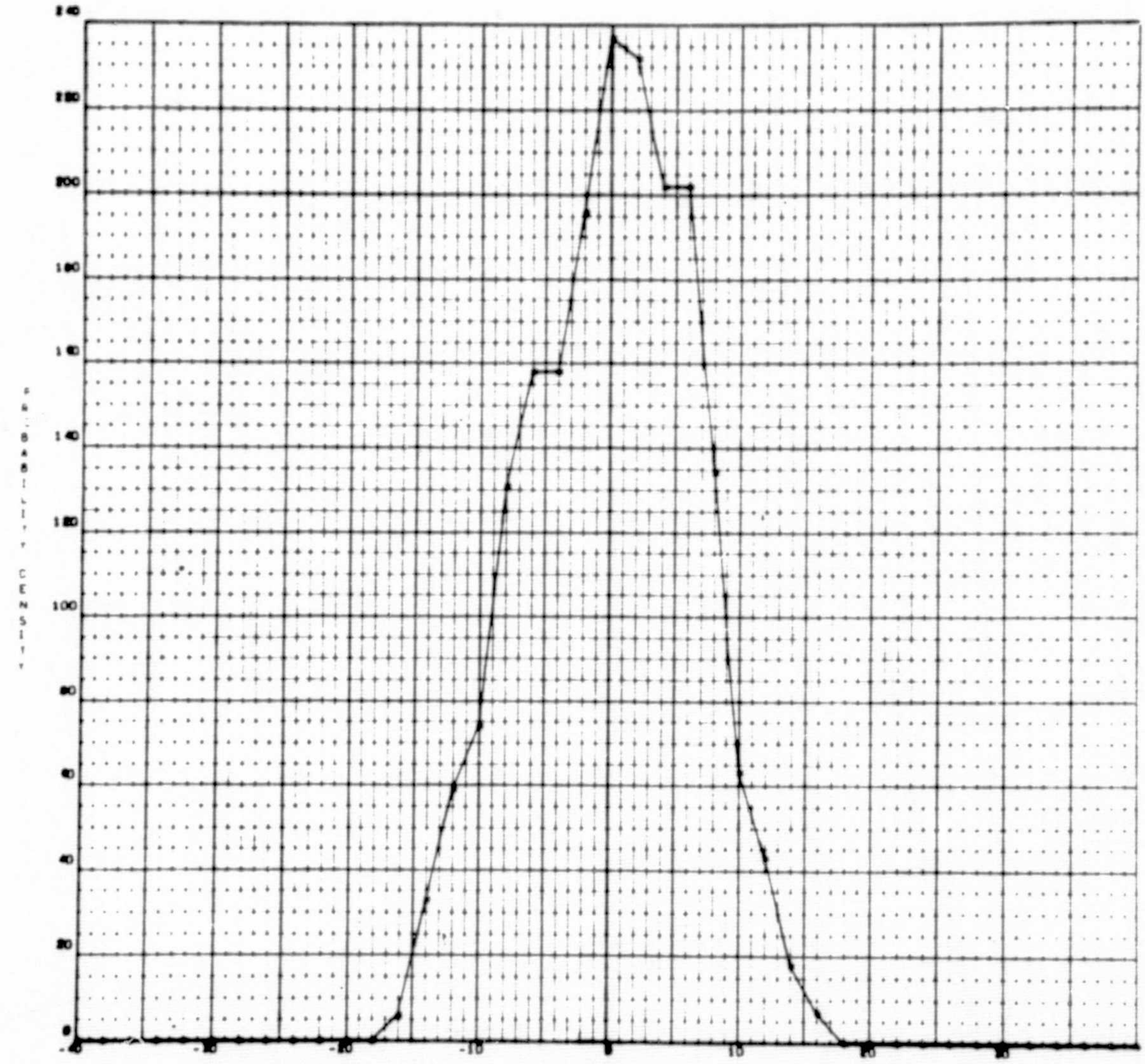
IMPULSE RESPONSE PERIOD = 2.0

(5, 4, 2, 1, 0) ⊕ (6, 1, 0)

M = 20

H ≈ 100

FIG 7-14B



PLOT NUMBER 11

SEQUENCE LENGTH = 1953

IMPULSE RESPONSE PERIOD = 43

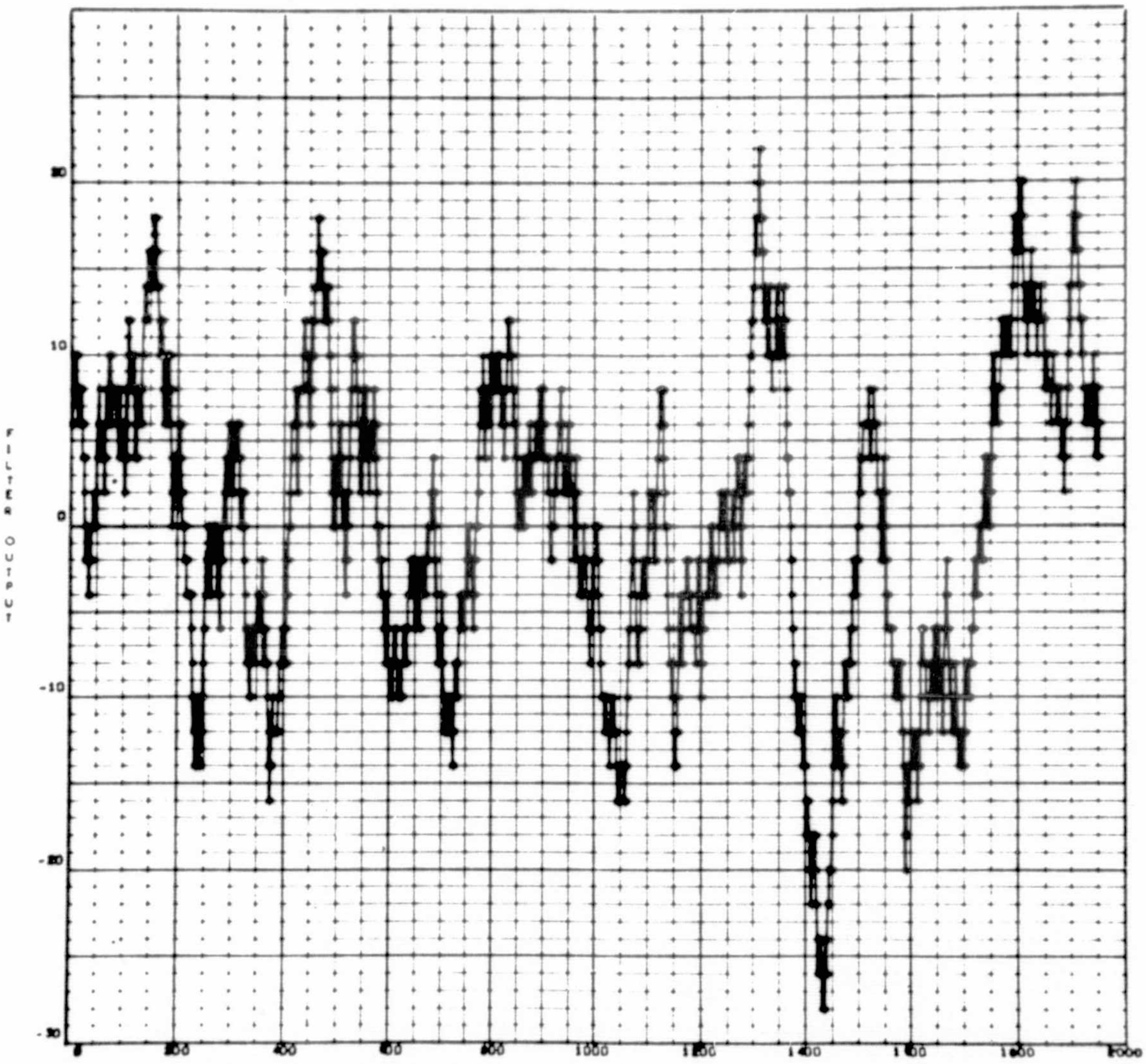
(5,4,2,1,0) ⊕ (6,1,0)

M = 40

H = 50

FIG 7-15B

JOB NO 422584 PAGE 23



PLOT NUMBER 12

SEQUENCE LENGTH = 1953

IMPULSE RESPONSE PERIOD = 80

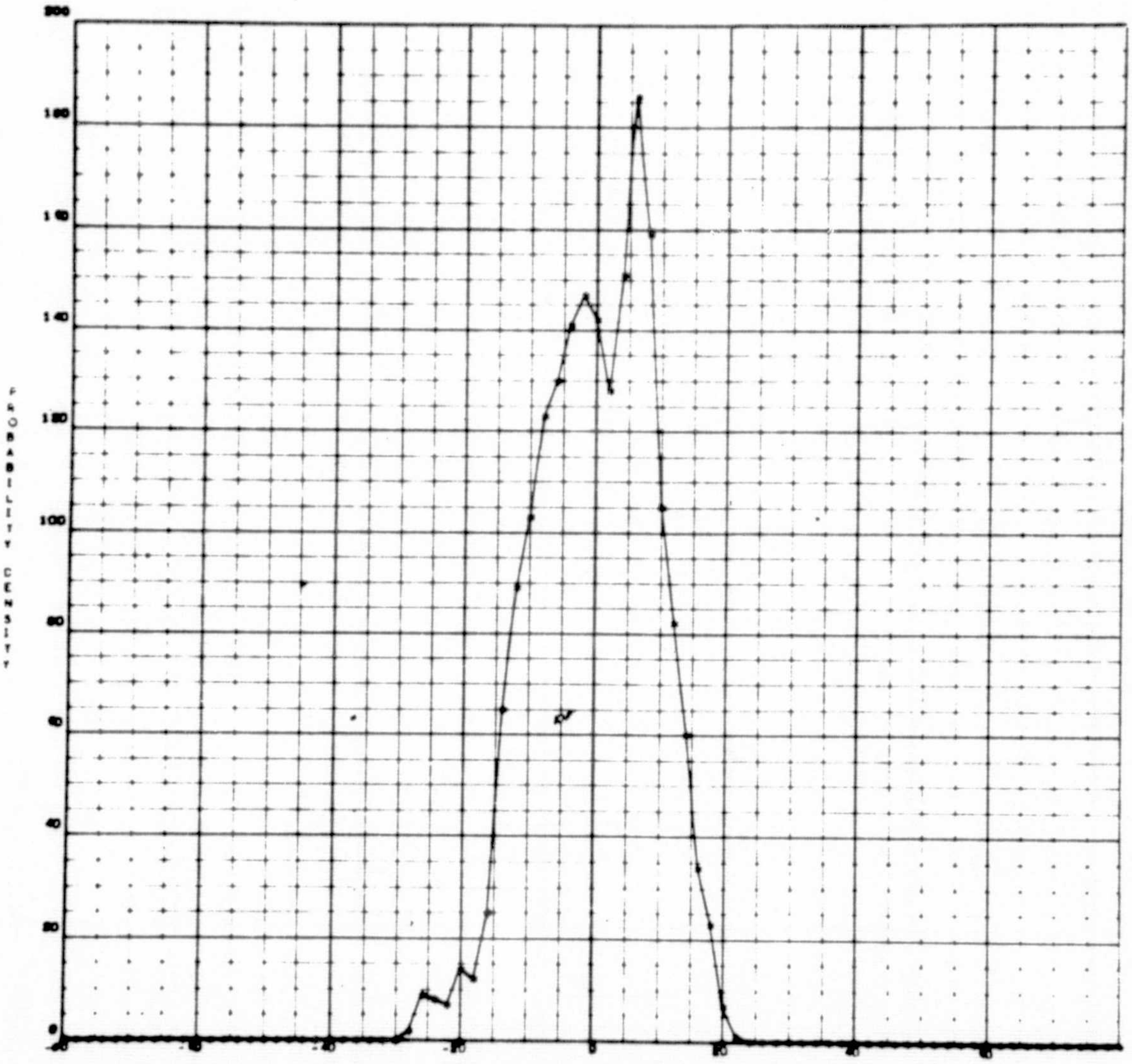
$(5,4,3,1,0) \oplus (6,1,0)$

$M = 80$

$H \approx 25$

FIG 7-16B

JOB NO 422064 PAGE 24



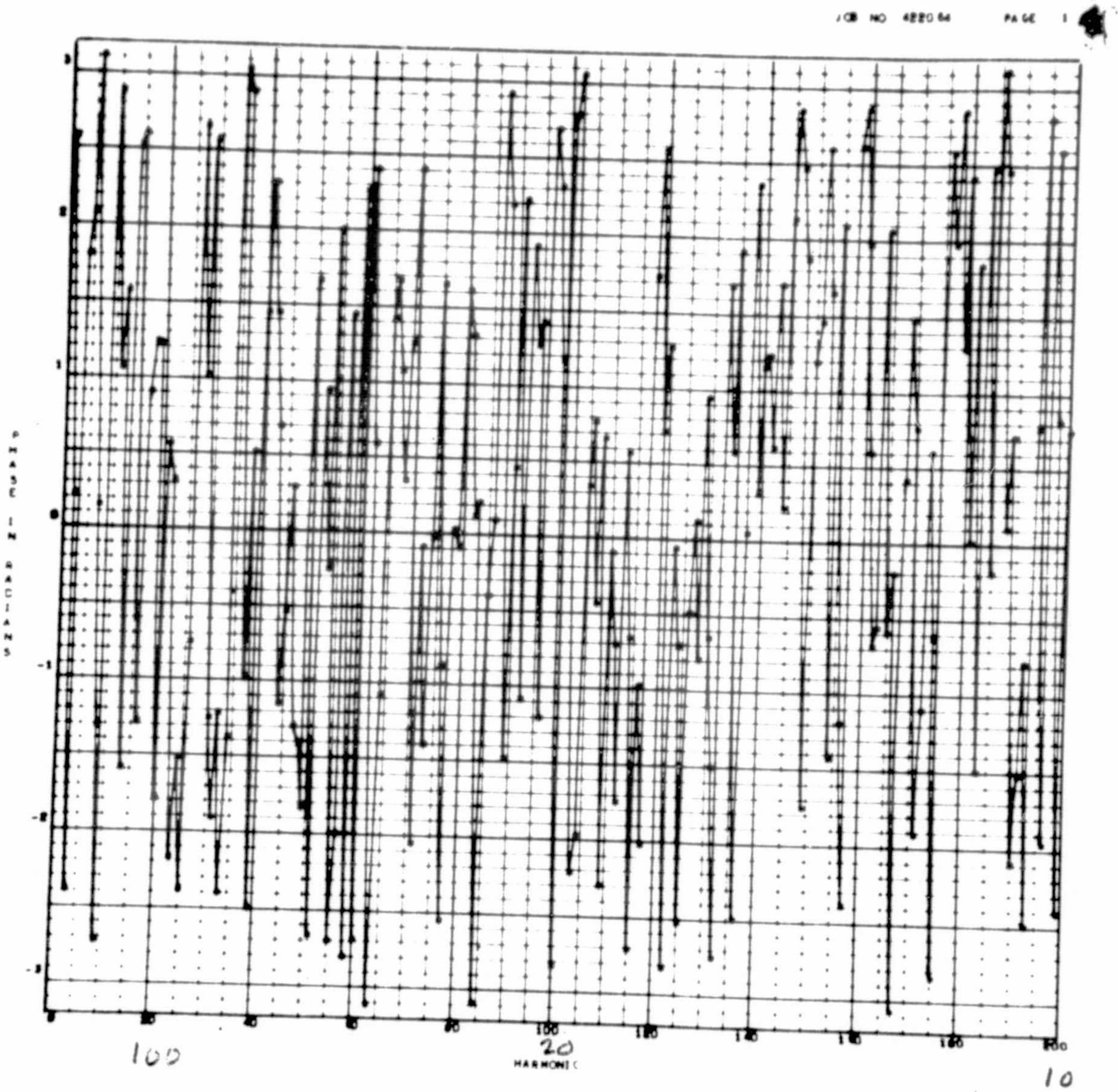
PLOT NUMBER 12

SEQUENCE LENGTH = 1953

IMPULSE RESPONSE PERIOD = 81

(5,4,2,1,0)⊕(6,1,0)
M = 80
H ≈ 25

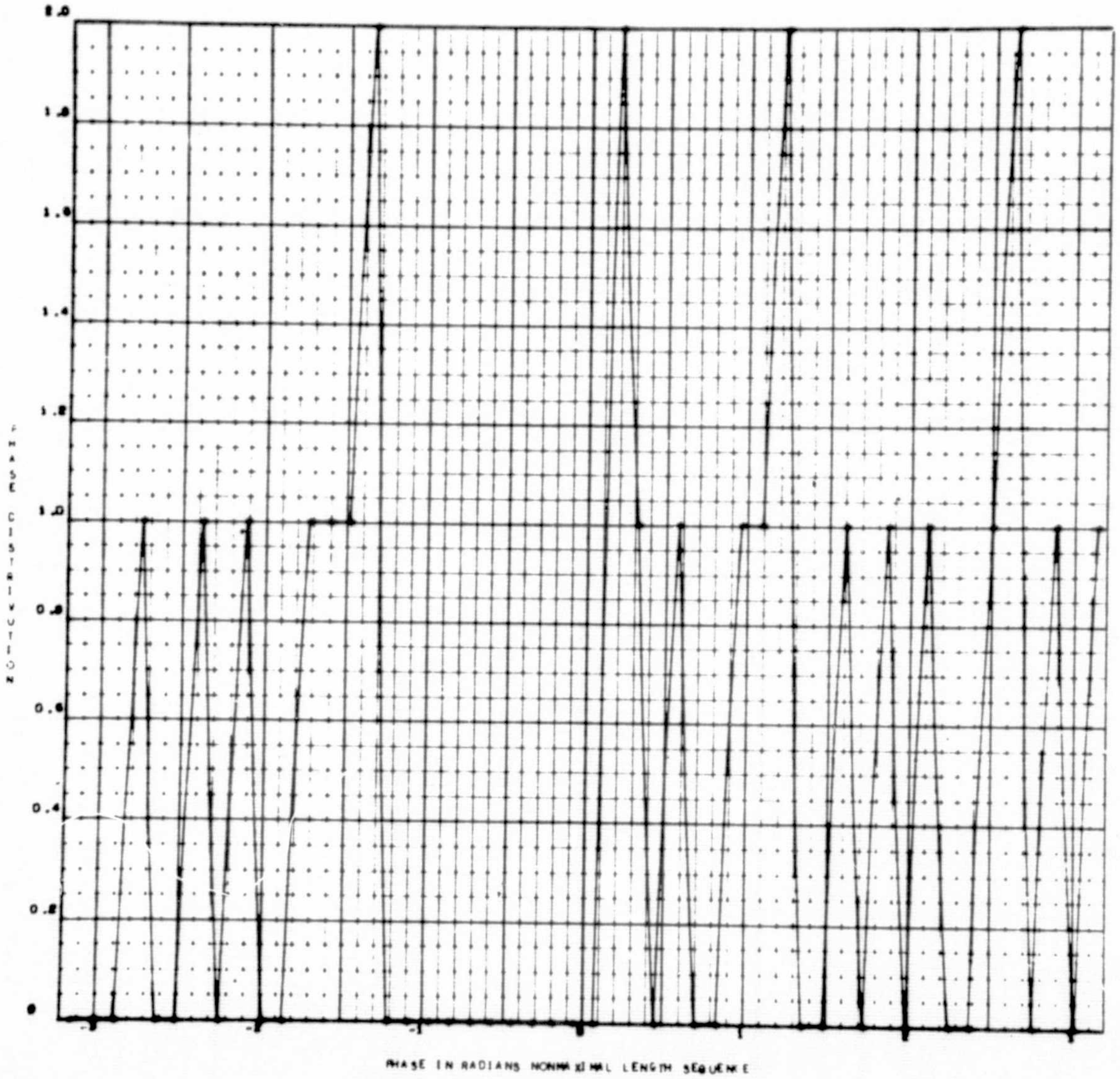
FIG 7-17B



(5,4,2,1,0)⊕(6,1,0)

FIRST 200 HARMONICS

FIG 7-18B



PLOT NUMBER 3
N1 = 5
N2 = 6
24 HARMONICS INCLUDED

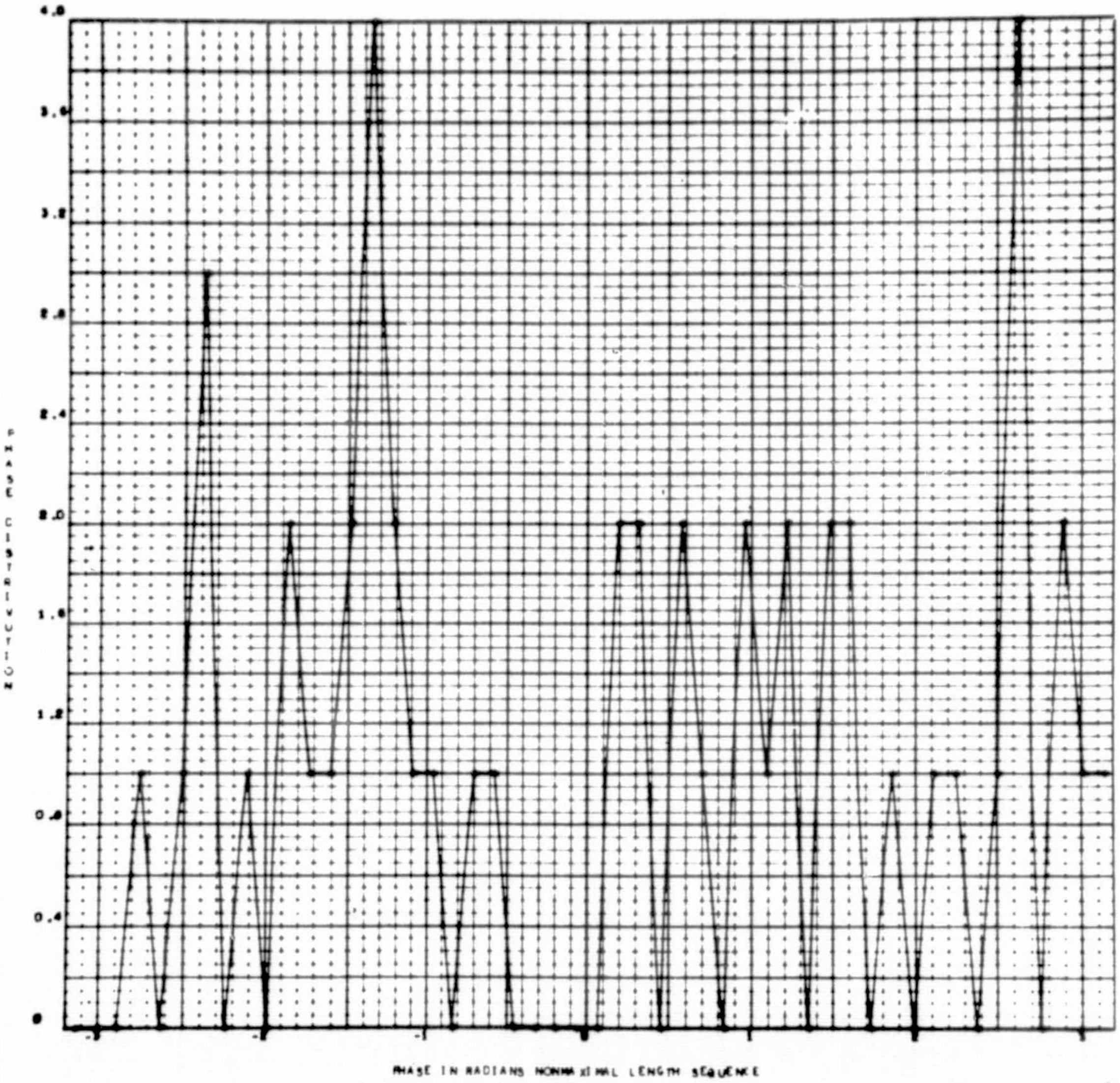
$(5,4,2,1,0) \oplus (6,1,0)$

$M \approx 80$

FIG 7-19B

$H \approx 25$

JOB NO 422064 PAGE 6



PHASE IN RADIANS NONMAXIMAL LENGTH SEQUENCE

PLOT NUMBER 5

N1 = 5

N2 = 6

50 HARMONICS INCLUDED

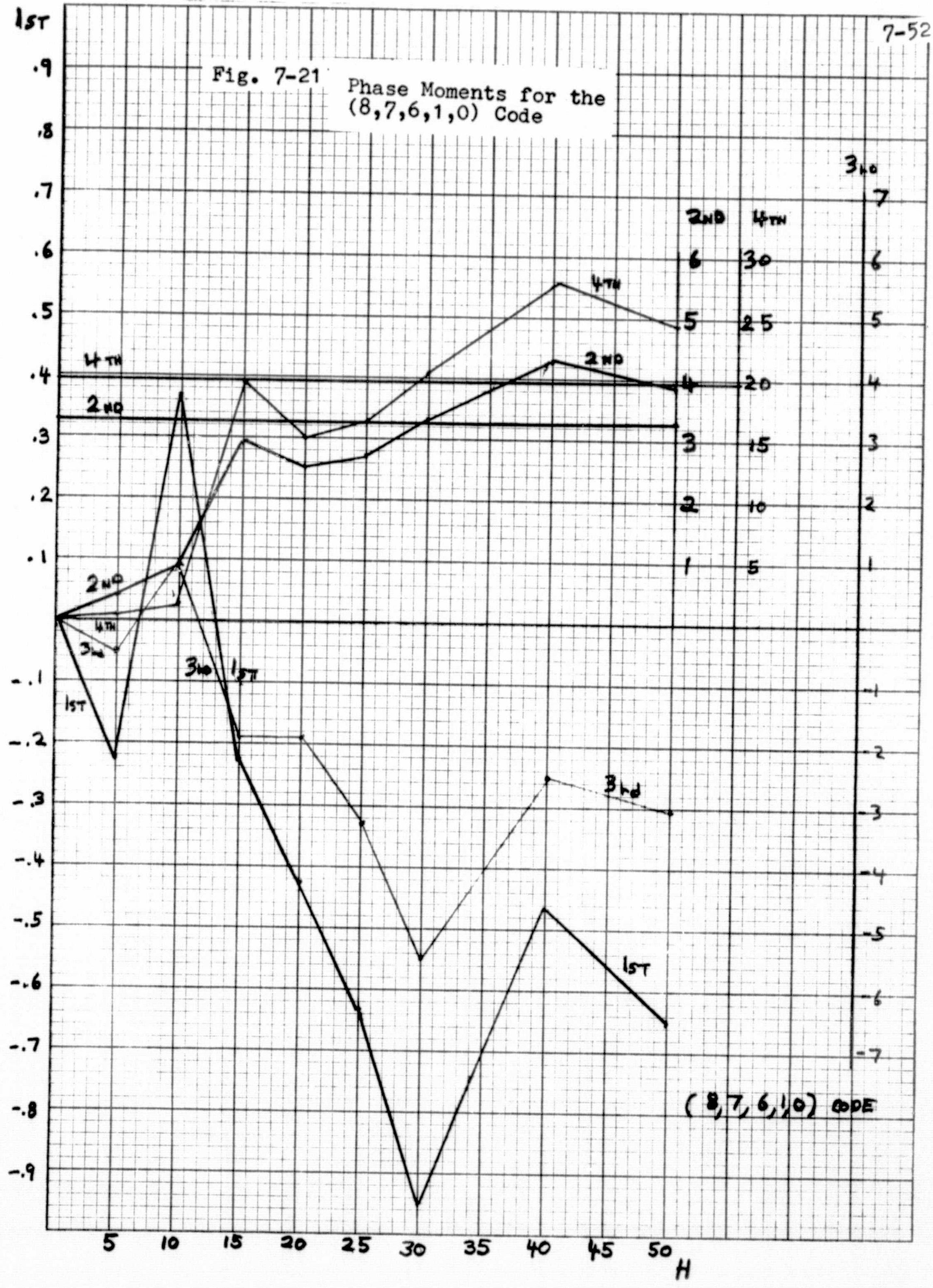
(5,4,2,1,0) ⊕ (6,1,0)

M = 40

H = 50

FIG 7-20 B

Fig. 7-21 Phase Moments for the (8,7,6,1,0) Code



8. CODE AND CARRIER PHASE LOCK LOOPS

This section gives the results of the design of the spectrum spreading delay lock code loop and carrier phase lock loops for use in the HEAO-C transponder. The code lock loop tracks the TDRSS-HEAO-C forward link spread spectrum modulation, and modulates the return link signal to provide range and range-rate tracking. The code lock loop also provides a coherent reference signal to perform the correlation function in the command receiver.

A simplified block diagram of a candidate HEAO-C PN transponder is shown in figure 8-1. The PN generator code output is shown as a single signal, but actually early/late gate signals are included in the design to implement a delay-lock-loop system.

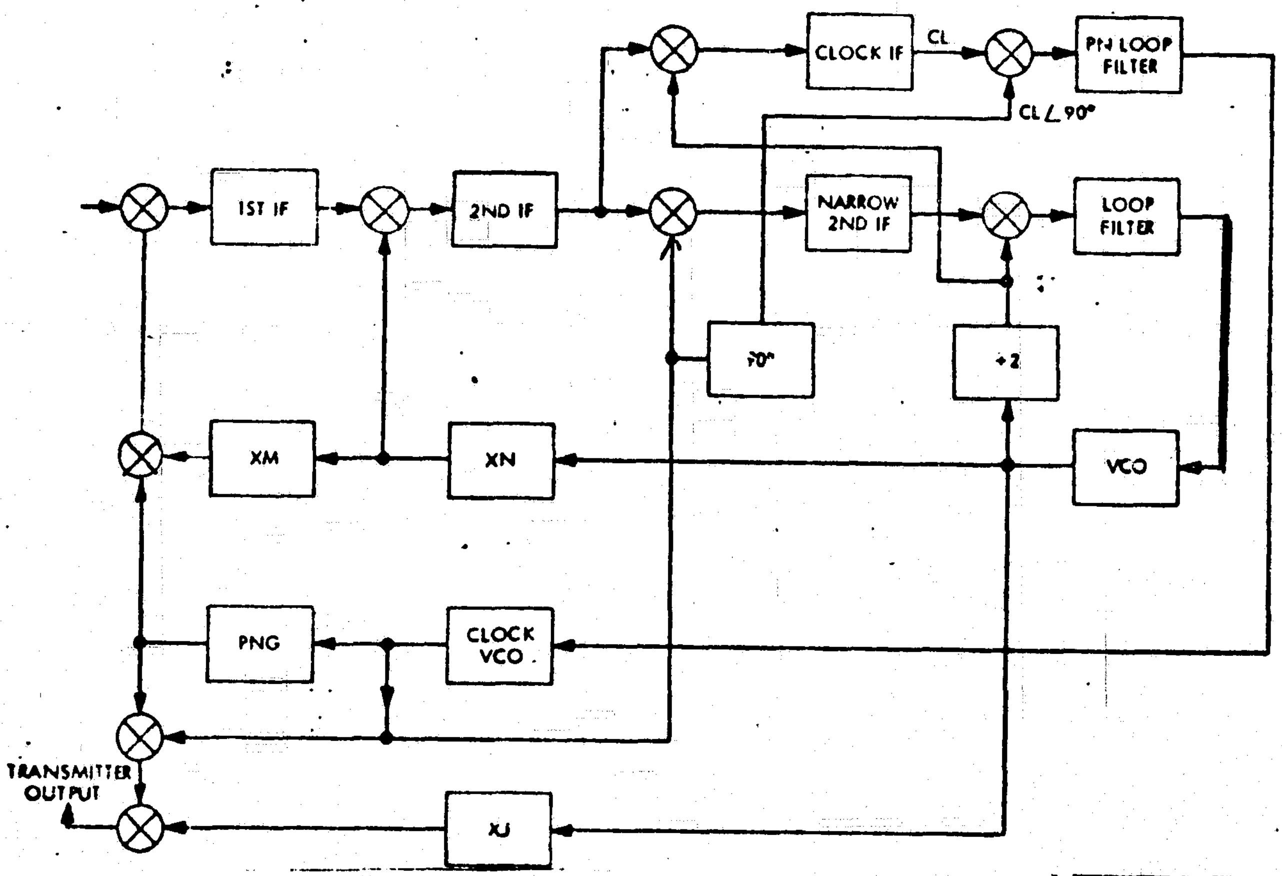


FIGURE 8-1 Complete PN Transponder

Figure 8-2 is a further simplified block diagram of the PN transponder including a symbolic representation of the signals. The definition of the symbols are as follows:

CA: Carrier
CL: Clock
PN: Code

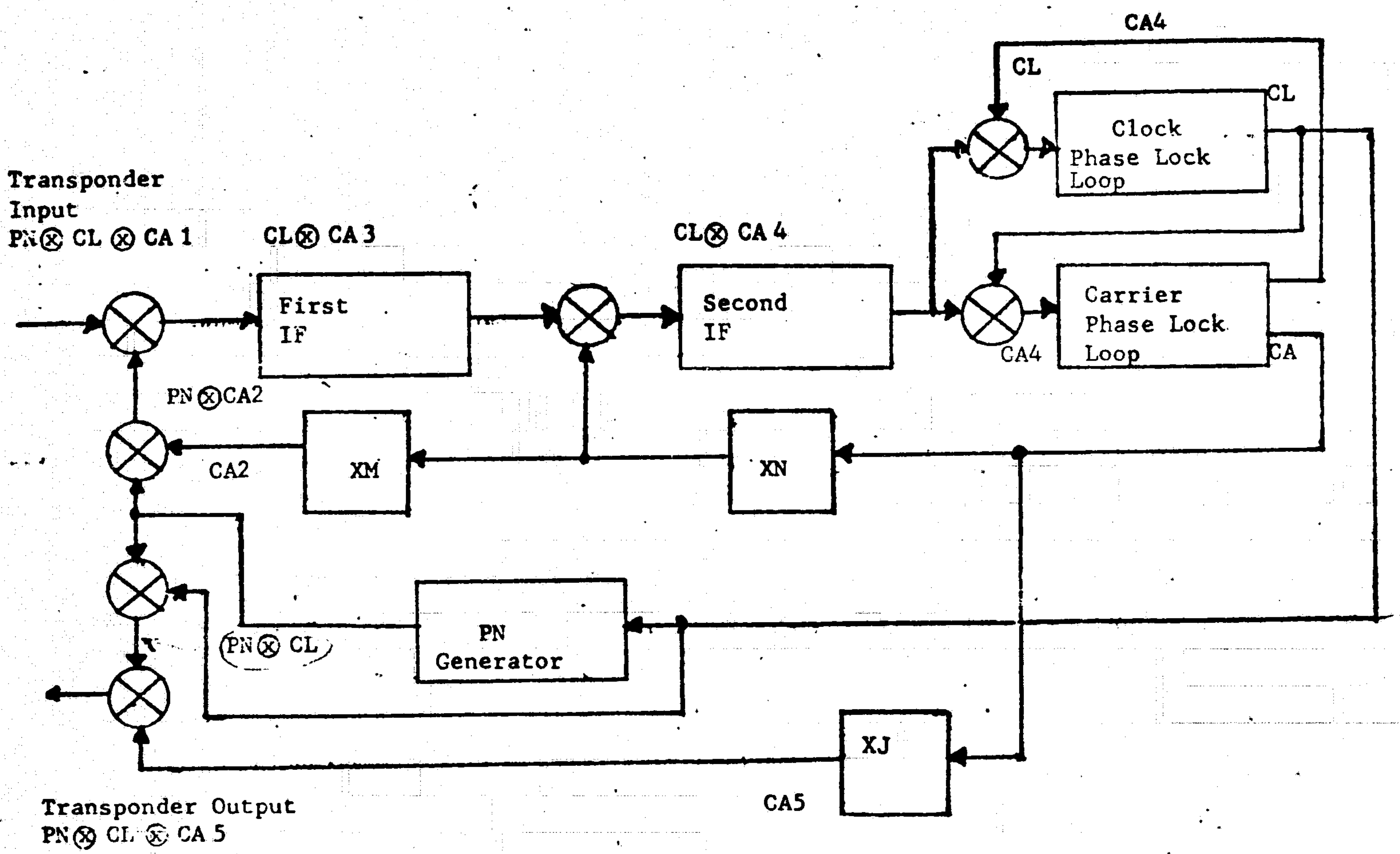


Figure 8-2 Simplified Transponder Block Diagram

Figure 8-3 and 8-4 are alternate designs of the IF - code delay lock loops. Early - late PN code signals, delayed by half a clock period provide the local code reference to paralled I and Q processors. ϕ_1, and ϕ_2 are clock phases separated by 90^o, and provide the half-clock delay for the late gate.

Figure 8-5 is the final difference amplifier, loop filter, and clock VCO. The form of the loop filter implies a high gain second order tracking loop. Fundamental loop characteristics such as capture range, loop bandwidth, capture time, and transient response are controlled primarily by the loop filter. The loop phase transfer function is

$$\frac{\theta_o\ (s)}{\theta_i\ (s)} = \frac{K\phi\ F\ (s)\ K_v}{S + K\phi\ K_f\ K_v/N} \qquad (1)$$

where

$K\phi$ = Phase Detecter Gain

F (s) = Filter Transfer Function

Kv = VCO Gain

N = Integer Division

The filter shown in figure 8-5 has transfer function of the form

$$F(s) = \frac{1+R_2CS}{R_1CS} \qquad (2)$$

where R_1 is the filter input resistor, R_2 is the feedback resistor, and C is the feedback capacitor. With F (s) as shown in (1) , (2) becomes

$$\frac{\theta_o\ (s)}{\theta_i\ (s)} = \frac{N\ (1+T_2S)}{\frac{S^2NT_1}{K\phi\ K_v} + T_2S+1} \qquad (3)$$

where

$$T_1 = R_1 C \qquad (4)$$

and

$$T_2 = R_2 C \qquad (5)$$

The loop natural frequency and damping factor, two particularly important parameters when considering loop dynamic characteristics, are:

$$\omega_n = \sqrt{\frac{K_\phi \, Kv}{NT_1}} \qquad (6)$$

and

$$\xi = \sqrt{\frac{K\phi \, Kv}{NT_1}} \left[\frac{T_2}{2}\right] \qquad (7)$$

Loop acquisition time is an important consideration for space spread spectrum transponders. For the case of a second order high gain loop with $\xi = .707$, the pull in time is given by the approximation

$$T_p \approx \frac{4.2\,(\Delta f)^2}{B_L{}^3} \quad \text{SEC} \qquad (8)$$

where B_L is the loop bandwidth and Δf is the offset. Since for the doppler offset frequencier in question for the TDRS-HEAO-C application Δf can be large compared to B_L, T_p would be excessive without some acquisition aiding such as a sweep search. With the sweep search, the loop ring up time multiplied by the number of doppler cells to be searched gives

$$T_S = \frac{\Delta f}{B_L{}^2} \qquad (9)$$

as time required for a sweep doppler search. For the eleventh order gold codes there are 2047 range cells to be searched. The total acquisition could be as large as

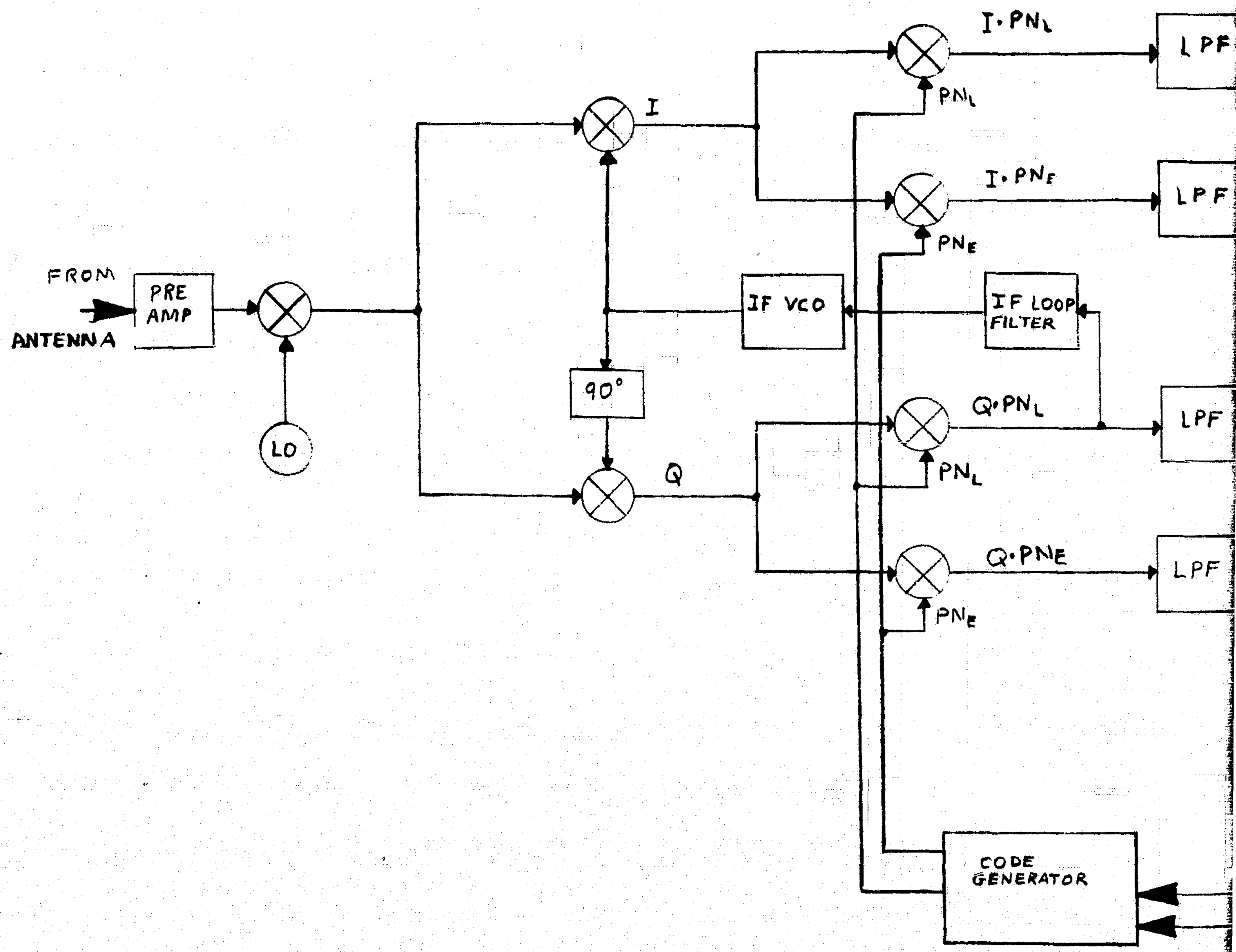
$$T_T = 2047 \quad \frac{\Delta f}{(B_L)^2} \, , \tag{10}$$

but the average acquisition time would be

$$T_T = \frac{2047}{2} \quad \frac{\Delta f}{(B_L)^2} \tag{11}$$

Figure 8-6 is the design for the mixer drivers for the alternate code delay lock loop designs.

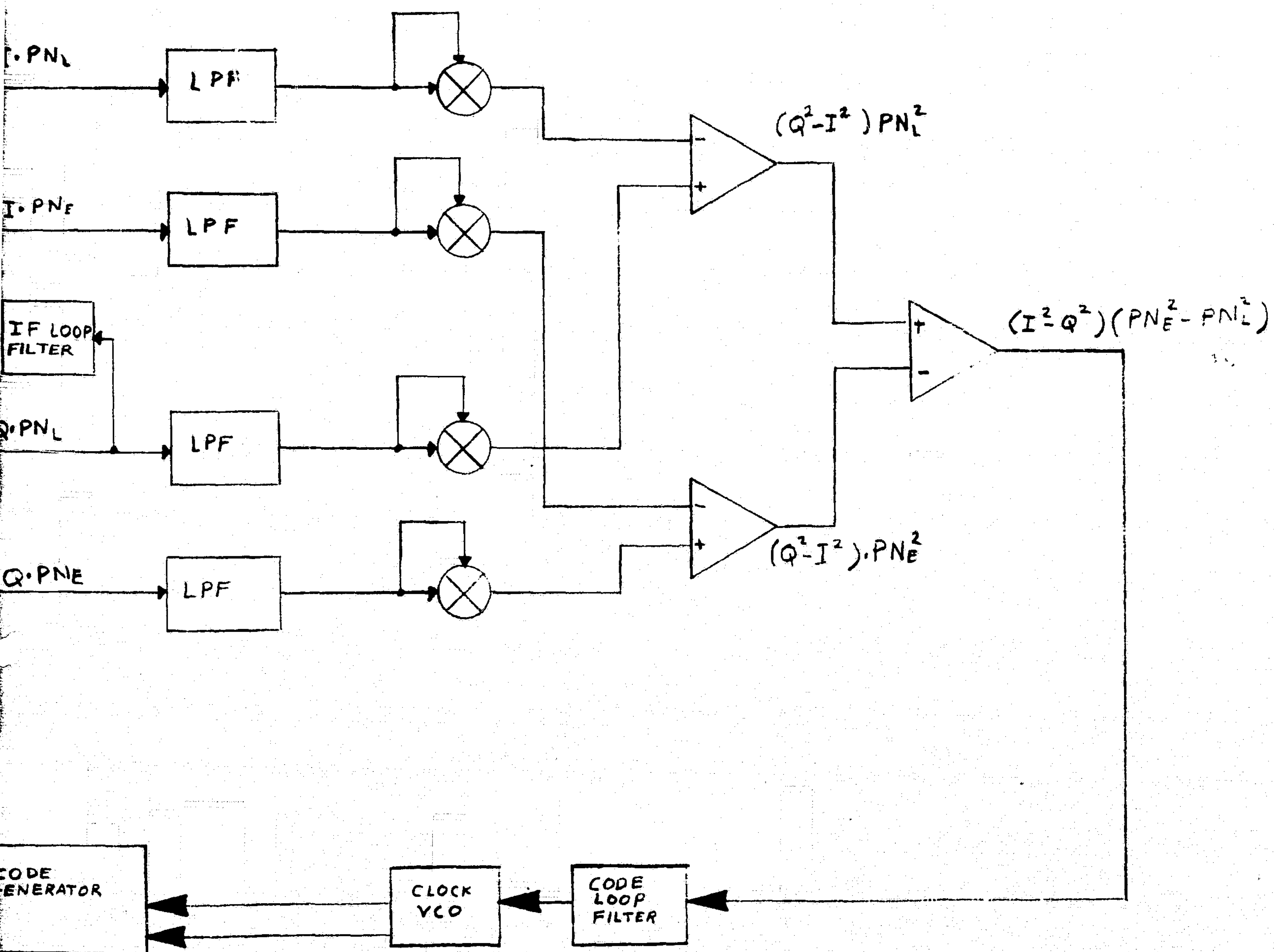
BY______________DATE________ SUBJECT____ SHEET NO.__________OF:______
CHKD. BY________DATE________ JOB NO. ____________________



FOLDOUT FRAME

-------OF:-------

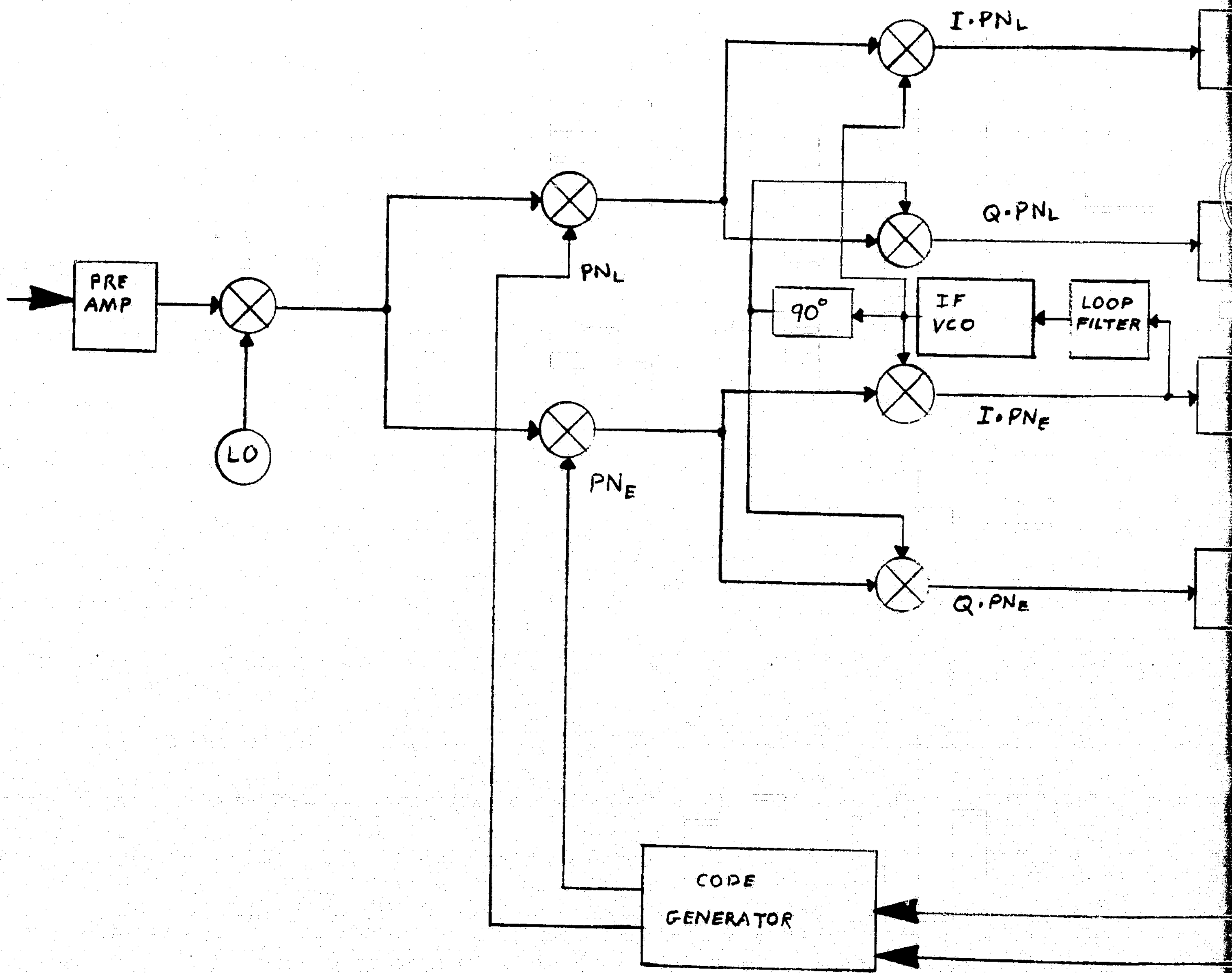
Fig. 8-3 Code Delay Lock Loop Configuration #1



BY .... DATE .... SUBJECT .... SHEET NO. .... OF ......

CHKD. BY .... DATE .... JOB NO. ....

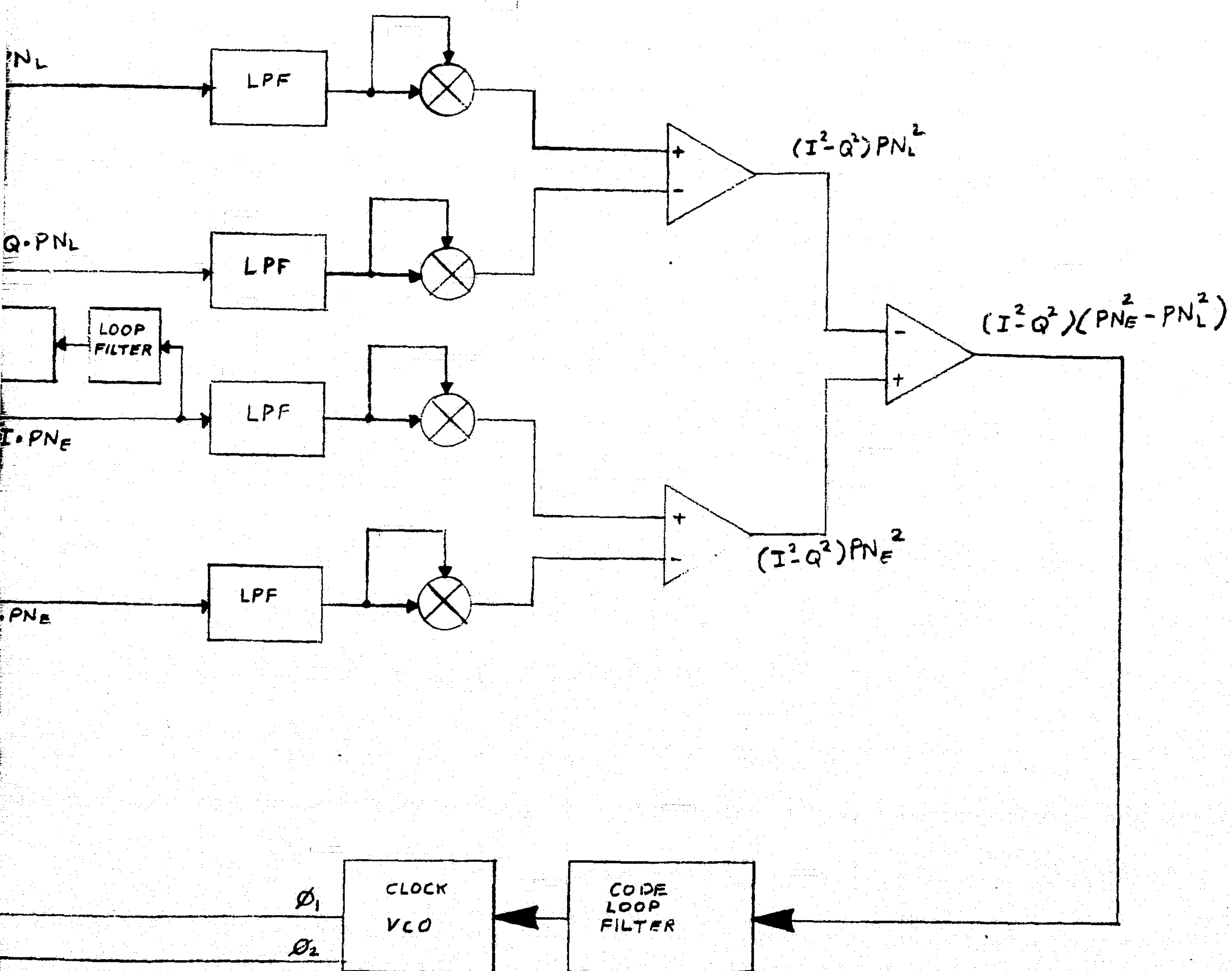
Fig



FOLDOUT FRAME

----OF------

Fig. 8-4 Code Delay Lock Loop Configuration #2



FOLDOUT FRAME 2

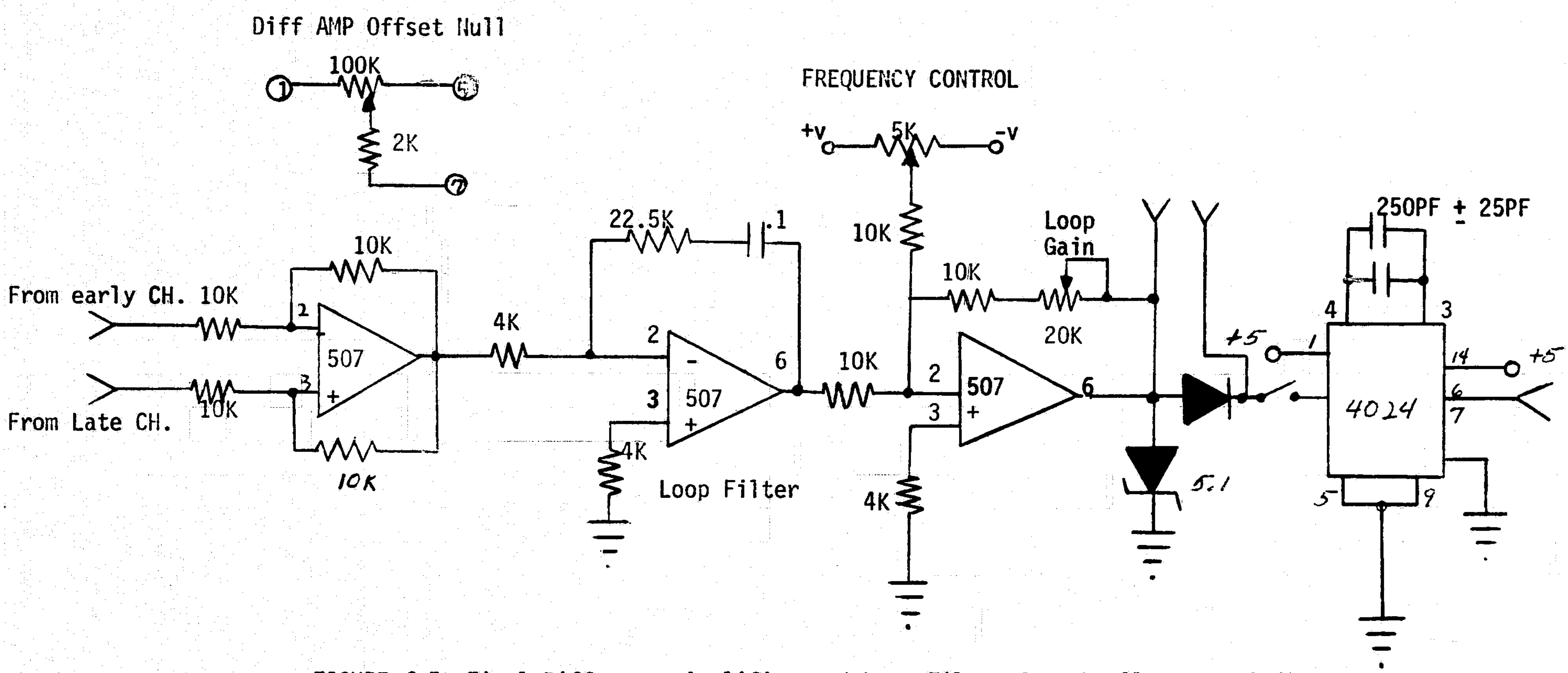


FIGURE 8-5 Final Difference Amplifier and Loop Filter for the Alternate Code Delay Lock Loop Designs

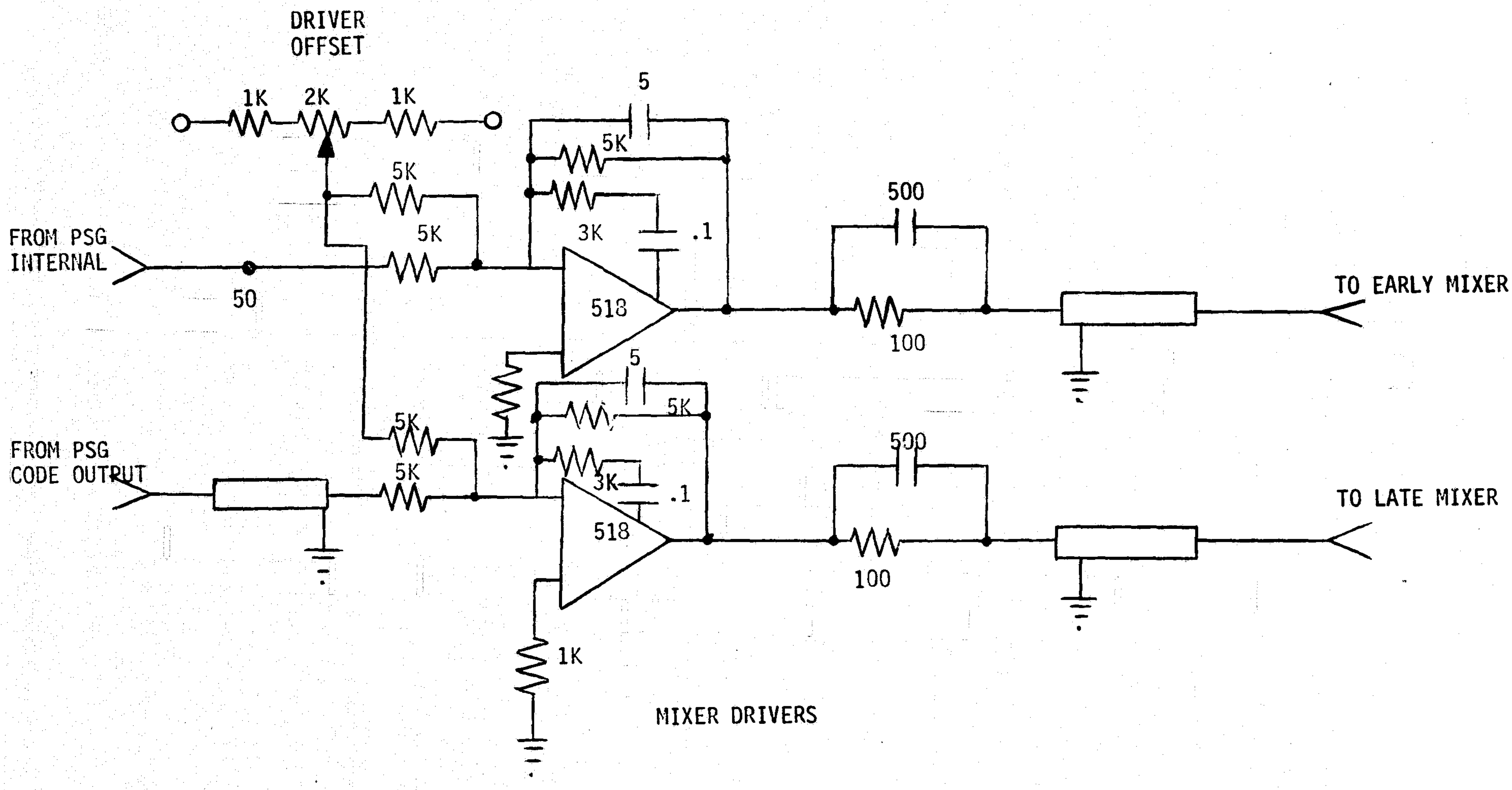


FIGURE 8-6 MIXER DRIVERS FOR THE CODE DELAY LACK LOOP

9. TOTAL SPREAD SPECTRUM TRANSPONDER SYSTEM

The block diagram of the total transponder system is shown in figure 9-1. The antenna system that would meet the HEAO-C-TDRESS requirements has been described in a previous report on this study (FINAL REPORT NASA GRANT NCR-01-001-021) and that description will be given here.

The proposed phased array antenna for the HEAO-C-TDRS return link will be a command pointing type with pointing commands formulated and communicated from the ground. An example of a phased array airborne steerable antenna system that would be applicable for this application was developed by Texas Instruments Incorporated for NASA under contract NAS8-24847. The antenna is an 128-element spiral array and achieved the performance parameters listed below.

Subsystem Performance	Value
Antenna (2232 MHz)	
Boresight gain (dB)	23.9
60-degree scan gain (dB)	20.3
Boresight axial ratio (dB)	0.3
60-degree scan axial ratio (dB)	2.0
Weight (pounds)	6.48
Boresight sidelobe level (dB)	19.5
60-degree scan sidelobe level (dB)	9.0
Module	
Noise figure (dB)	6.0
Receive gain (dB)	24
Diplexer isolation (dB)	35
Peak phase shifter phase error (degrees)	10

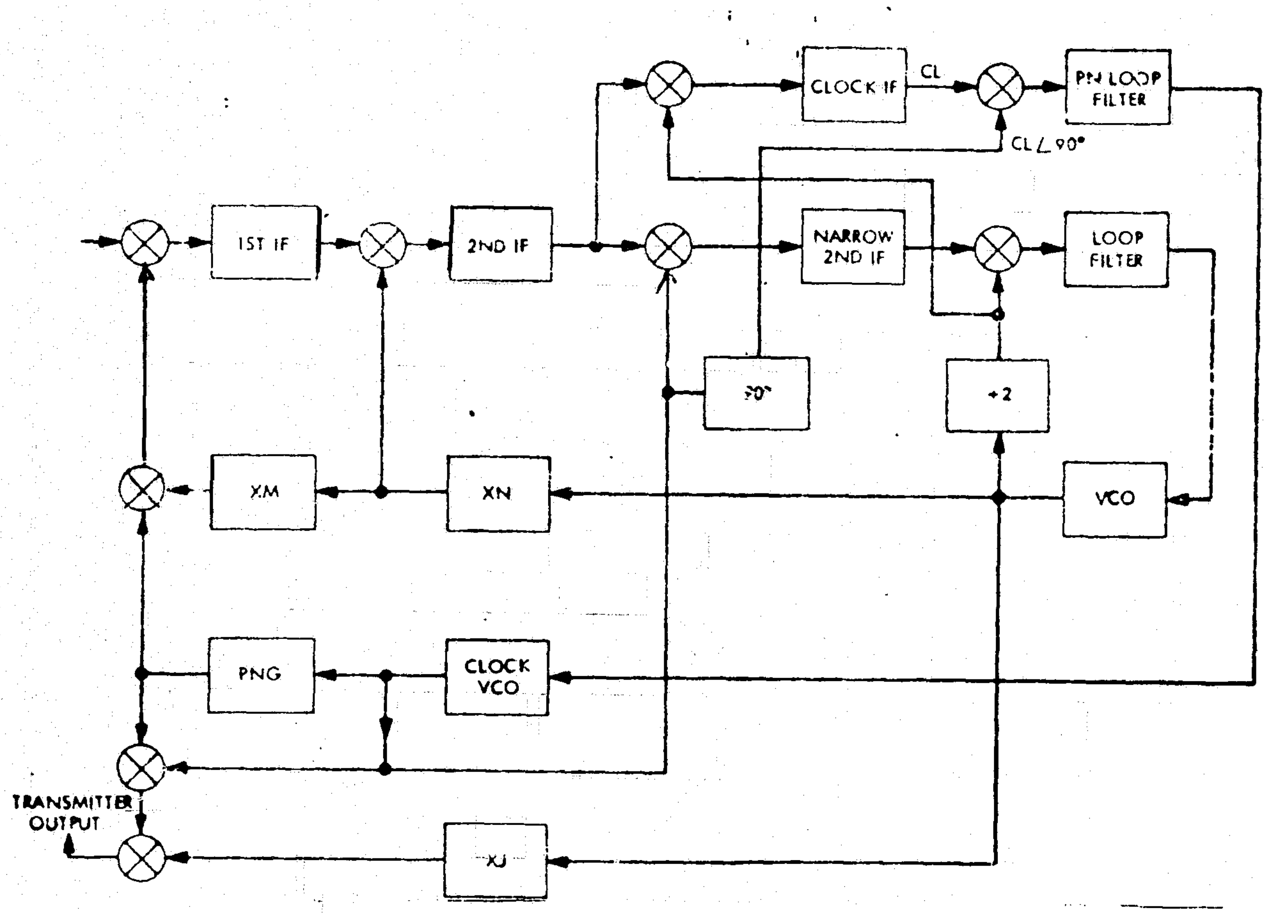


FIGURE 9-1 FIG 9-1 Complete PN Transponder

Module	Value
RMS phase shifter phase error (degrees)	5.2
Phase shifter amplitude error (dB)	0.5
Phase linearity (degrees)	5
Power output (dBW)	0.5
Transmit gain (dB)	19.0
Transmit efficiency (percent)	25
Weight (pounds)	0.275
Transmit Manifold (128-Element)	
Peak phase error (degrees)	±5.25
Peak amplitude variation (dB)	±0.6
Peak output VSWR (Ratio:1)	1.65
Loss (dB)	3.2
Weight (pounds)	6.48

These performance parameters were used in the TDRS-HEAO-C power budget calculations.

The block diagram of the HEAO-C communications system with error control coding and phased array antenna implementations is shown in figure 9-2.

The TDRS-HEAO-C forward link is established on the low gain (near isotropic) antenna. Pointing control commands are coded and the return-forward links are established with the phased array.

The current HEAO-C, NON-TDRS communication system block diagram is shown in figure 9-3.

Typical user characterstics for link budget calculations were:

SINGLE-ACCESS AND MULTIPLE-ACCESS, S-BAND

Ts	(^{o}K)	824
KTs (dbw/HZ		199.4
Ts (db)		29.4

This implies a pre-amp noise figure of 4.6 db.

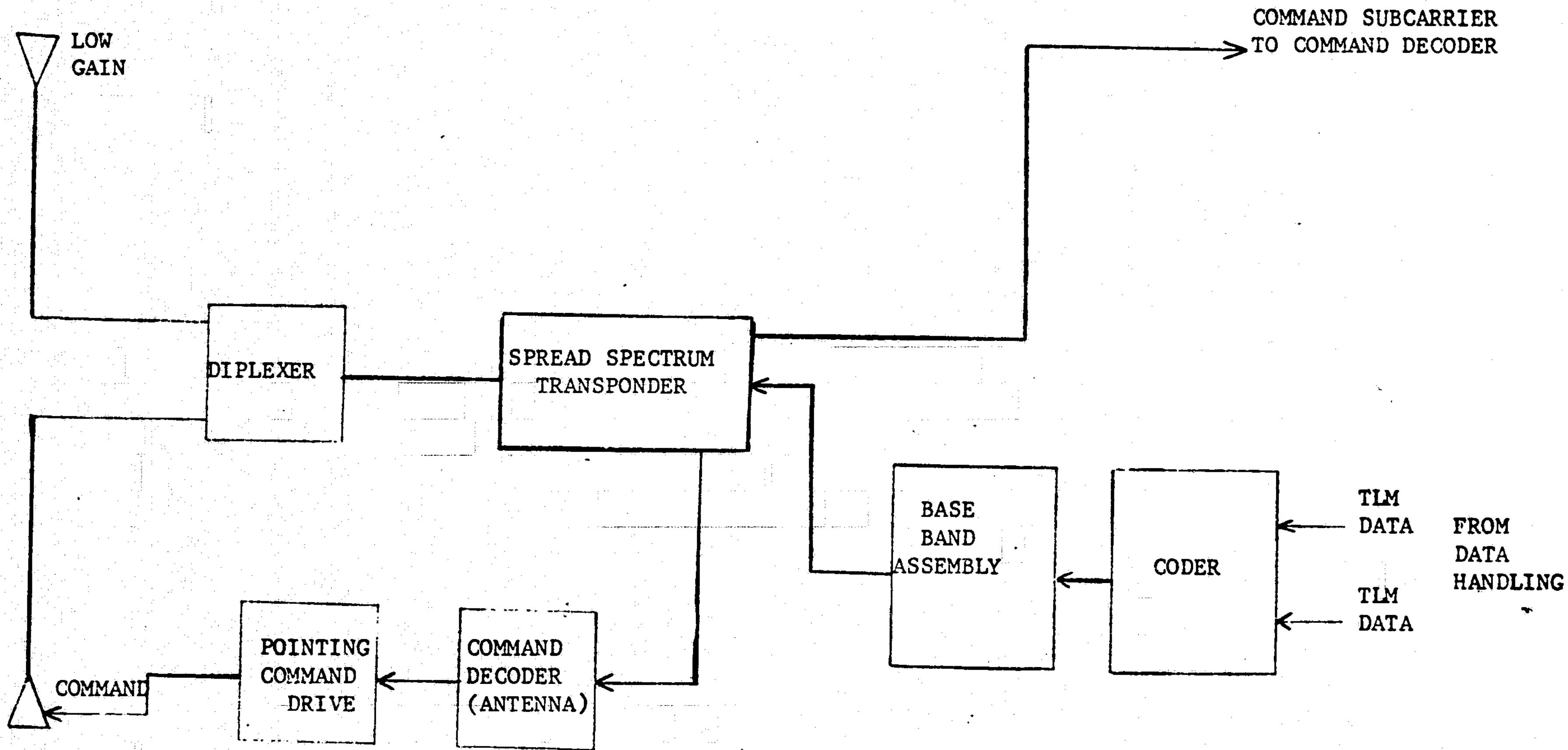


FIGURE 9-2 HEAO-C Communication System Block Diagram

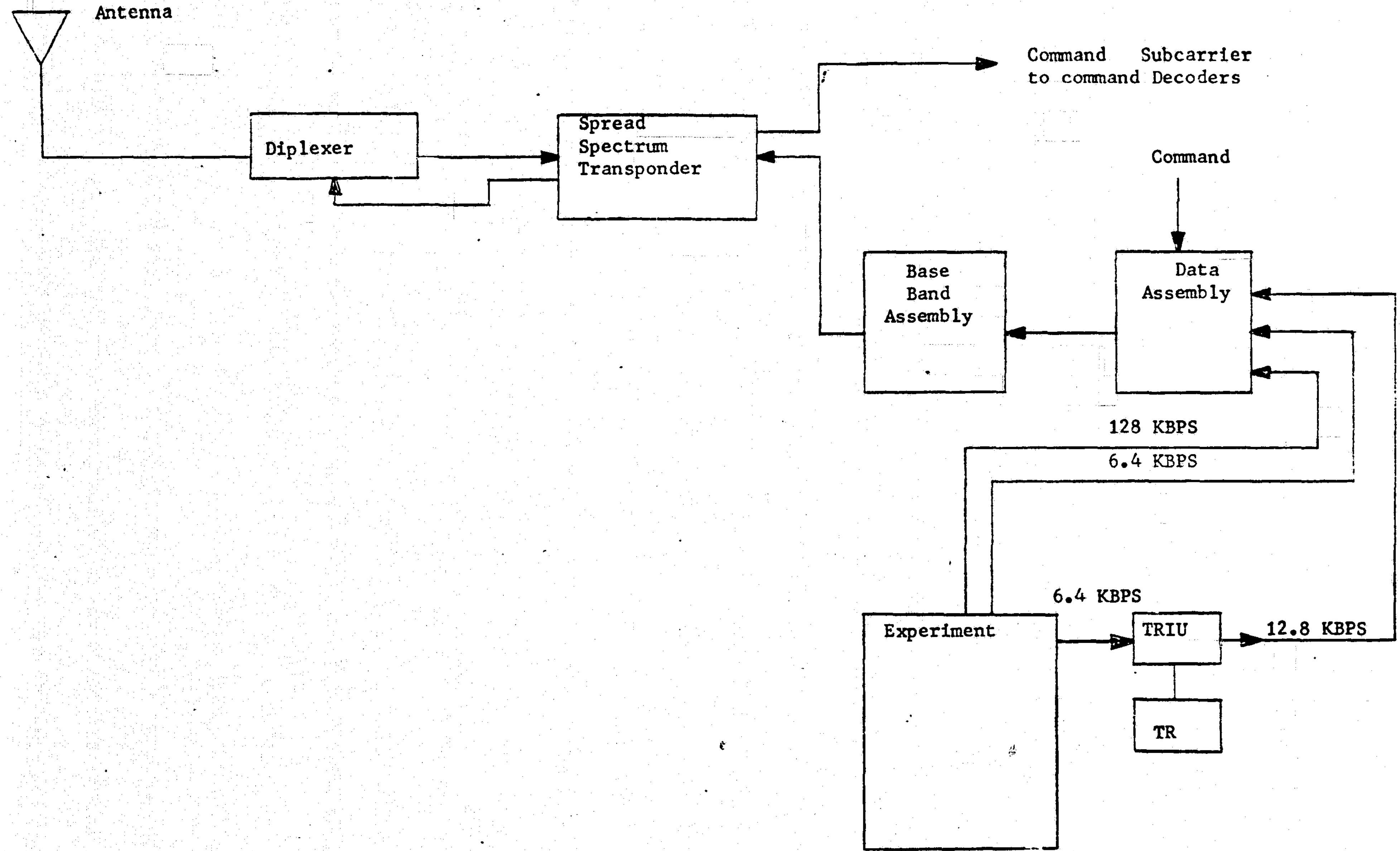


FIGURE 9-3 HEAO-C,NON - TDRS, COMMUNICATION SYSTEM BLOCK DIAGRAM

This noise figure can be achieved with a low noise solid-state pre-amplifier. Typical of such an amplifier is the WJ-5004-323 manufactured by the Watkins-Johnson Company. The specifications for this unit are:

MODEL ----- WJ-5004-325
FREQUENCY RANGE ----- 2-4 GHZ
NOISE FIGURE ----- 4.5db MAX, 3.8db TYP
SMALL SIGNAL GAIN ----- 35db MIM
POWER OUTPUT ----- +7dbm **MIM**
PRIMARY POWER ----- +15 VOLTS DC (19 REG) 110 MA
SIZE ----- 2.5X1.3X3.5 INCHES
WEIGHT ----- 8 OZ
VSWR (5006) -----IN 2db (5002) OUT 2db (5002)
TVP. INTERCEPT ----- +17 dbm
POINT FOR IM
PRODUCTS (dbm)
TEMP ----- 54^{o} **c**- $+71^{o}$c
ENVIRONMENT ----- MIL-E-5400, CLASS 2
MIL-E-16400, CLASS 2

The amplifier outline drawing is shown in figure 9-4.

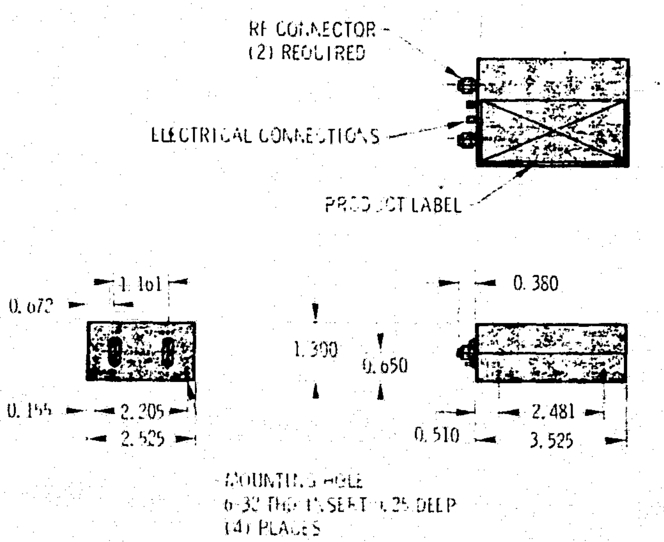


Figure 9-4 Pre Amplifier Package

10. REFERENCE LITERATURE SEARCH

The following pages are the results of a search of NASA computer listings in the area of HEAO and TDRSS Telecommunications.

PRINT 06/2/1-12 TERMINAL=48

72X10394*# ISSUE 3 PAGE 72 CATEGORY 11 NASA-CR-127512
JPL-760-40 69/09/30 531 PAGES UNCLASSIFIED DOCUMENT GOVT.+
CONTR.
TRACKING AND DATA RELAY SATELLITE NETWORK (TDRSN)
(TECHNICAL AND COST DATA ON TRACKING AND DATA RELAY SATELLITE
NETWORK AND FEASIBILITY OF TELECOMMUNICATIONS SYSTEM) FINAL STUDY
REPORT
A/DIETL, M. G. A/COMP.
JET PROPULSION LAB., CALIF. INST. OF TECH., PASADENA.
SPONSORED BY NASA
/*COST ANALYSIS/*DATA ACQUISITION/*TDR
SATELLITES/*TELECOMMUNICATION/ PROJECT PLANNING/ RANGE AND RANGE RATE
TRACKING/ VERY HIGH FREQUENCIES

71X10685*# ISSUE 3 PAGE 149 CATEGORY 31
HMA-2065-5TRW-V-2-SECT-6/7/8/9 NASA-CR-119809
DRL-182-VOL-2-SECT-6/7/8/9 NAS8-26273 71/04/23 68 PAGES
UNCLASSIFIED DOCUMENT NASA + CONTR.
HEAO HIGH ENERGY ASTRONOMY OBSERVATORY. VOLUME 2 - TECHNICAL
DESCRIPTION/DESIGN DEFINITION AND ENGINEERING. SECTION 6 - RELIABILITY
ASSESSMENT. SECTION 7 - HIGH RISK AND LONG LEAD ITEMS. SECTION 8 -
COST ANALYSIS AND TRADEOFFS DATA. SECTION 9 - SUPPORTING RESEARCH AND
TECHNOLOGY
(SUBSYSTEM RELIABILITY, HIGH RISK/LONG LEAD ITEMS, TRADEOFFS, COST
ANALYSIS, AND SUPPORTING RESEARCH AND DEVELOPMENT - VOL. 2 - SECTS.
6-9)
TRW SYSTEMS GROUP, REDONDO BEACH, CALIF.
/*COMPONENT RELIABILITY/*COST ANALYSIS/*HEAO/*RESEARCH AND
DEVELOPMENT/*SPACECRAFT COMPONENTS/*TRADEOFFS/ ATTITUDE (INCLINATION)/
PROPULSION/ RELIABILITY ENGINEERING/ SYSTEMS ENGINEERING/
TELECOMMUNICATION

71X10682*# ISSUE 3 PAGE 148 CATEGORY 31 NASA-CR-119808
HMA-2065-4TRW-VOL-2-SECT-5 DRL-182-VOL-2-SECT-5 NAS8-26273 71/04/23
645 PAGES UNCLASSIFIED DOCUMENT NASA + CONTR.
HEAO HIGH ENERGY ASTRONOMY OBSERVATORY. VOLUME 2 - TECHNICAL
DESCRIPTION/DESIGN DEFINITION AND ENGINEERING. SECTION 5 - SUBSYSTEM
DEFINITION FINAL REPORT
(SPACECRAFT SUBSYSTEMS AND COMPONENTS FOR HEAO - VOL. 2, SECT. 5)
TRW SYSTEMS GROUP, REDONDO BEACH, CALIF.
/*HEAO/*SPACECRAFT COMPONENTS/ ATTITUDE CONTROL/ DATA SYSTEMS/
ELECTRIC POWER TRANSMISSION/ PROPULSION/ SPACECRAFT POWER SUPPLIES/
TELECOMMUNICATION/ TEMPERATURE CONTROL

71X10671*#   ISSUE 3   PAGE 147   CATEGORY 31   NASA-CR-119828
DRD-MA-082-U2-VOL-2-APP HMA-2055-4GAC-VOL-2-APP DRL-182-VOL-2-APP
NAS8-26272   71/04/00   419 PAGES   UNCLASSIFIED DOCUMENT   NASA + CONTR.

HIGH ENERGY ASTRONOMY OBSERVATORY. VOLUME 2, APPENDIX - DESIGN DEFINITION/ ENGINEERING ANALYSIS FINAL REPORT

(HEAO SPACECRAFT DESIGN AND SYSTEMS ENGINEERING STUDY - VOL. 2, APPENDIX)

GRUMMAN AEROSPACE CORP., BETHPAGE, N.Y.

/*HEAO/*SPACECRAFT DESIGN/*SYSTEMS ENGINEERING/ ATTITUDE CONTROL/ DATA PROCESSING/ GROUND SUPPORT EQUIPMENT/ TELECOMMUNICATION/ TELEMETRY/ WEIGHT (MASS)

73W70717   150-22-32

HIGH RELIABILITY CONTROL SYSTEMS FOR ANTENNAS

RAUMANN, N. A.   301-982-6579

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION. GODDARD SPACE FLIGHT CENTER, GREENBELT, MD.

CONCENTRATION OF DATA ACQUISITION RESPONSIBILITIES AND INCREASING DATA BANDWIDTHS RESULTING FROM REDUCTION IN THE NUMBER OF NETWORK STATIONS ARE PLACING GREATER LOADS ON THE NETWORK LINKS. THUS, THE COST OF LINK DOWN TIME IS INCREASED, REQUIRING A CORRESPONDING INCREASE IN LINK RELIABILITY. THE ANTENNA CONTROL SYSTEM IS ONE OF THE FEW COMPONENTS TO WHICH REDUNDANCY CANNOT BE ECONOMICALLY APPLIED. IN ADDITION, LINK DOWN TIME DUE TO ALIGNMENT REQUIREMENTS AND ROUTINE MAINTENANCE HAS TO BE MINIMIZED. AT THE SAME TIME A REDUCTION IN MAINTENANCE AND OPERATION (M AND O) MANPOWER IS HIGHLY DESIRABLE. ABOVE OBJECTIVES ARE MET BY THE TASKS IN THIS RTOP. THE COMPUTER CONTROLLED ANTENNA SYSTEM HAS DEMONSTRATED A POTENTIAL FOR MARKED REDUCTION IN (M AND O) MANPOWER AND THE FUNCTIONS OF SEVERAL EQUIPMENTS HAVE BEEN SUCCESSFULLY INTEGRATED. THIS SYSTEM IS OPERATING EXPERIMENTALLY AT THE NETWORK TEST AND TRAINING FACILITY (NTTF) AND PROTOTYPE DESIGN HAS BEGUN FOR FY 73 OPERATION. IT WILL SUPPORT THE STADAC SYSTEM AT NTTF TO BE INSTALLED IN THE SAME TIME FRAME. THE ACOUSTICAL ANALYSIS EQUIPMENT FOR DETECTING AND IDENTIFYING INCIPIENT FAILURES IN HYDRAULIC AND MECHANICAL SYSTEMS IS BEING OR HAS BEEN INSTALLED ON TEN NETWORK ANTENNAS. IN ADDITION TO DIRECT SUPPORT TO THE NETWORK, THESE INSTALLATIONS WILL PROVIDE FIELD DATA FOR FURTHER EVALUATION AND ANALYSIS TECHNIQUE DEVELOPMENT UNDER THIS RTOP. STUDY EFFORTS IN PROGRESS WILL DEFINE THE DESIGN CHARACTERISTICS FOR A HIGH ACCURACY CONTROL SYSTEM WHICH IS REQUIRED FOR FUTURE ANTENNAS OPERATING IN THE KU-BAND SUCH AS THE GROUND STATION IN SUPPORT OF THE TRACKING AND DATA RELAY SATELLITE (TDRS).

/ ANTENNAS/ DATA ACQUISITION/ GROUND STATIONS/ SERVOMECHANISMS/ TDR SATELLITES/ TELECOMMUNICATION

73W70709 150-22-20
TRACKING AND DATA RELAY SATELLITE TECHNOLOGY DEVELOPMENT
CLARK, G. Q. 301-982-6331
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION. GODDARD SPACE FLIGHT CENTER, GREENBELT, MD.
THE TWO OBJECTIVES ARE (1) TO PROVIDE FOR THE DEFINITION OF A TRACKING AND DATA RELAY SATELLITE SYSTEM TO BE USED FOR SUPPORT OF NASA MISSIONS, AND (2) TO PROVIDE FOR THE ORDERLY DEVELOPMENT OF THE TECHNOLOGY REQUIRED FOR IMPLEMENTING A FIRST-GENERATION TDRSS BY 1977. VARIOUS STUDIES WILL BE PERFORMED TO ESTABLISH THE CRITERIA FOR A TDRSS, WHILE OTHER STUDIES WILL LOOK FOR SOLUTIONS TO PROBLEMS INHERENT IN THE SYSTEM. IN ADDITION, TECHNOLOGY WILL BE DEVELOPED AS REQUIRED FOR A FIRST-GENERATION TDRSS.
/ DATA TRANSMISSION/ SATELLITE NETWORKS/ SATELLITE TRACKING/ TDR SATELLITES/ TELECOMMUNICATION

73W70539 164-21-55
TRACKING AND DATA RELAY SATELLITE TECHNOLOGY DEVELOPMENT
CLARK, G. Q. 301-982-6331
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION. GODDARD SPACE FLIGHT CENTER, GREENBELT, MD.
THE TWO OBJECTIVES ARE 1) TO PROVIDE FOR THE DEFINITION OF A TRACKING AND DATA RELAY SATELLITE SYSTEM TO BE USED FOR SUPPORT OF NASA MISSIONS, AND 2) TO PROVIDE FOR THE ORDERLY DEVELOPMENT OF THE TECHNOLOGY REQUIRED FOR IMPLEMENTING A FIRST GENERATION TDRSS BY 1977. VARIOUS STUDIES WILL BE PERFORMED TO ESTABLISH THE CRITERIA FOR A TDRSS WHILE OTHER STUDIES WILL LOOK FOR SOLUTIONS TO PROBLEMS INHERENT IN THE SYSTEM. IN ADDITION, TECHNOLOGY WILL BE DEVELOPED AS REQUIRED FOR A FIRST-GENERATION TDRSS.
/ DATA TRANSMISSION/ SATELLITE NETWORKS/ TDR SATELLITES/ TELECOMMUNICATION/ TRACKING NETWORKS

73N22821*# ISSUE 13 PAGE 1582 CATEGORY 31 NASA-CR-130217
SD-73-SA-0018-2-VOL-2 NAS5-21705 73/04/00 251 PAGES UNCLASSIFIED DOCUMENT
TRACKING AND DATA RELAY SATELLITE SYSTEM CONFIGURATION AND TRADEOFF STUDY. VOLUME 2 PART 2 TELECOMMUNICATIONS DESIGN
(DESIGN AND DEVELOPMENT OF TELECOMMUNICATIONS EQUIPMENT FOR USE WITH TRACKING AND DATA RELAY SATELLITE SYSTEM - VOL. 2) FINAL REPORT
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(CONFIGURATION DATA AND DESIGN INFORMATION FOR SPACE SHUTTLE LAUNCH OF TRACKING AND DATA RELAY SATELLITE SYSTEM - VOL. 4)
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73N22799*#   ISSUE 13   PAGE 1579   CATEGORY 31   NASA-CR-130219
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TRACKING AND DATA RELAY SATELLITE SYSTEM CONFIGURATION AND TRADEOFF STUDY. VOLUME 1 SUMMARY. PART 2 FINAL REPORT, 22 AUGUST 1972 - 1 APRIL 1973

(DEVELOPMENT OF TRACKING AND DATA RELAY SATELLITE SYSTEM CONCEPT FOR SERVICE OF LOW, MEDIUM, AND HIGH DATA RATE USER SPACECRAFT - VOL. 1)

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(COMPUTER PROGRAM COMPUTES MAXIMUM POSSIBLE STRENGTH OF INTERFERENCE PATTERN SENT FROM HIGH ALTITUDE TRACKING AND DATA RELAY SATELLITE TO LOW ALTITUDE USER SATELLITE.)
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TRACKING AND DATA RELAY SATELLITE NETWORK (TDRSN)
(TECHNICAL AND COST DATA ON TRACKING AND DATA RELAY SATELLITE NETWORK AND FEASIBILITY OF TELECOMMUNICATIONS SYSTEM) FINAL STUDY REPORT
A/DIETL, M. G. A/COMP.
JET PROPULSION LAB., CALIF. INST. OF TECH., PASADENA.
SPONSORED BY NASA
/*COST ANALYSIS/*DATA ACQUISITION/*TDR SATELLITES/*TELECOMMUNICATION/ PROJECT PLANNING/ RANGE AND RANGE RATE TRACKING/ VERY HIGH FREQUENCIES

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THE EFFECTS OF MULTIPATH AND RFI ON THE TDRSS COMMAND AND TELEMETRY LINKS
(MULTIPATH AND RADIO FREQUENCY INTERFERENCE EFFECTS ON TDR SATELLITE COMMAND AND TELEMETRY LINKS)
FINAL REPORT
A/JENNY, J.; B/SHAFT, P.
ESL, INC., SUNNYVALE, CALIF.
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THE EFFECTS OF MULTIPATH AND RFI ON THE TRACKING AND DATA RELAY SATELLITE SYSTEM
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A/JENNY, J. A.; B/WEISS, S. J.
/*MULTIPATH TRANSMISSION/*RADIO FREQUENCY INTERFERENCE/*TDR SATELLITES/ DATA PROCESSING/ IBM 360 COMPUTER/ MATHEMATICAL MODELS/ RADIO WAVES/ WAVE SCATTERING

74W70727 310-30-35
NETWORK UTILIZATION AND SHUTTLE STUDIES 1979-1990
UVAAS, C. M. 301-982-2357
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION. GODDARD SPACE FLIGHT CENTER, GREENBELT, MD.
THE OBJECTIVES ARE TO PERFORM ADVANCED SYSTEM PLANNING TO FORMULATE AND DEVELOP COMPARATIVE MODELS OF NETWORK SUPPORT CAPABILITIES AND NETWORK RESOURCES THAT WILL BE REQUIRED TO PROVIDE GROUND SUPPORT OF SHUTTLE AND SHUTTLE LAUNCHED PAYLOADS IN THE 1979-1990 TIME FRAME. THE NETWORK RESOURCES WOULD INCLUDE A TRACKING AND DATA RELAY SATELLITE (TDRS) SYSTEM PLUS 8 TO 11 GROUND STATIONS FOR SUPPORTING SHUTTLE ORBITER, SORTIE LABS, SPACE TUGS, AND PAYLOADS INJECTED INTO SYNCHRONOUS ORBIT AND BEYOND OR ORBITS ABOVE 350 N.MI. WITH THE SPACE TUG, AS WELL AS PAYLOADS LAUNCHED VIA CONVENTIONAL DELTA BOOSTERS DURING THE INTERIM PHASE-OVER PERIOD TO SHUTTLE LAUNCHES. THE PLANNING MODEL WILL INDETIFY SYSTEM CAPABILITIES, OPERATIONAL PHILOSOPHY, AND NEW TECHNOLOGY ASSOCIATED WITH THE NEW GENERATION OF SPACECRAFT AND SHUTTLE LAUNCHED VEHICLES IN SUFFICIENT DETAIL TO DEFINE HARDWARE SYSTEM REQUIREMENTS FOR THE GROUND SUPPORT NETWORK. THE APPROACH WILL BE TO INVESTIGATE SUPPORT REQUIREMENTS OF FUTURE MANNED AND UNMANNED MISSIONS SUCH AS SHUTTLE, LARGE SPACE TELESCOPE, SPACE STATIONS/PLATFORMS, TDRS, EARTH OBSERVATORY SATELLITE, HIGH ENERGY ASTRONOMY OBSERVATORY, ORBITING SOLAR OBSERVATORY, EARTH RESOURCES TECHNOLOGY SATELLITE, SYNCHRONOUS EARTH OBSERVATIONAL SATELLITE, ETC. THESE ARE PRESENTLY BEING PROGRAMMED FOR THE 1979-1990 TIME FRAME AND DEFINE THE IMPACT OF THESE SUPPORT REQUIREMENTS ON NETWORK RECEIVING AND TRANSMITTING SYSTEMS, THE NETWORK CONTROL CENTERS, AND REMOTE SITE COMPUTER AND DATA HANDLING SYSTEMS.
/ DATA SYSTEMS/ GROUND SUPPORT EQUIPMENT/ SPACE SHUTTLES/ SYSTEMS ENGINEERING/ TDR SATELLITES/ TRACKING NETWORKS

74W70725 310-30-21
ADVANCED NETWORK PLANNING
UNDERWOOD, C. H. 301-982-2357
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION. GODDARD SPACE FLIGHT CENTER, GREENBELT, MD.

THIS TASK ADDRESSES THE TOTAL SCOPE OF PROBLEMS WHICH ARE RELATED TO THE TECHNICAL INTEGRATION OF THE STADAN AND THE MSFN INTO THE STDN AND THE DEVELOPMENT OF PLANS, PROGRAMS, AND TECHNIQUES REQUIRED TO UPDATE THE NETWORK. THIS TASK WILL EMPHASIZE THOSE AREAS WHICH MAXIMIZE THE EFFECTIVENESS OF THE SUPPORT PROVIDED AND INCREASE THE COST EFFECTIVENESS OF THE TOTAL NETWORK. ADVANCED AND STATE OF THE ART TECHNIQUES WILL BE IDENTIFIED AND THEIR POTENTIAL IMPACT UPON THE NETWORK WILL BE EVALUATED ALONG WITH THEIR MISSION SUPPORT CAPABILITIES. SPECIFIC OBJECTIVES OF THIS TASK WHICH WILL AFFECT ALL ELEMENTS OF THE NETWORK, INCLUDING REMOTE SITES AND DATA HANDLING SYSTEMS, ARE IDENTIFIED IN THE FOLLOWING BROAD AREAS (1) INTEGRATION OF MSFN AND STADAN NETWORKS, (2) TDRS IMPACT ON THE NETWORK, (3) ADVANCED NETWORK SYSTEM SUPPORT/COST TRADE-OFF DATA, (4) ADVANCED TELECOMMUNICATIONS SYSTEMS, AND (5) TRACKING COVERAGE MODELING.

/ DATA SYSTEMS/ MANNED SPACE FLIGHT NETWORK/ RESOURCES MANAGEMENT/ STADAN (SATELLITE TRACKING NETWORK)/ TDR SATELLITES

74W70720 310-20-32
HIGH RELIABILITY CONTROL SYSTEMS FOR ANTENNAS
RAUMANN, N. A. 301-982-6579
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION. GODDARD SPACE FLIGHT CENTER, GREENBELT, MD.
CONCENTRATION OF DATA ACQUISITION RESPONSIBILITIES AND INCREASING DATA BANDWIDTHS RESULTING FROM REDUCTION IN THE NUMBER OF NETWORK STATIONS ARE PLACING GREATER LOADS ON THE NETWORK LINKS. THUS, THE COST OF LINK DOWN TIME IS INCREASED, REQUIRING A CORRESPONDING INCREASE IN LINK RELIABILITY. THE ANTENNA CONTROL SYSTEM IS ONE OF THE FEW COMPONENTS TO WHICH REDUNDANCY CANNOT BE ECONOMICALLY APPLIED. IN ADDITION, LINK DOWN TIME DUE TO ALIGNMENT REQUIREMENTS AND ROUTINE MAINTENANCE HAS TO BE MINIMIZED. AT THE SAME TIME A REDUCTION IN MAINTENANCE AND OPERATION (M AND O) MANPOWER IS HIGHLY DESIRABLE. ABOVE OBJECTIVES ARE MET BY THE TASKS IN THIS RTOP. THE COMPUTER CONTROLLED ANTENNA SYSTEM HAS DEMONSTRATED A POTENTIAL FOR MARKED REDUCTION IN (M AND O) MANPOWER AND THE FUNCTIONS OF SEVERAL EQUIPMENTS HAVE BEEN SUCCESSFULLY INTEGRATED. THIS SYSTEM IS OPERATING EXPERIMENTALLY AT THE NETWORK TEST AND TRAINING FACILITY (NTTF) AND IT WILL SUPPORT THE STADAC SYSTEM. THE ACOUSTICAL ANALYSIS EQUIPMENT FOR DETECTING AND IDENTIFYING INCIPIENT FAILURES IN HYDRAULIC AND MECHANICAL SYSTEMS HAS BEEN INSTALLED ON TEN NETWORK ANTENNAS. IN ADDITION TO DIRECT SUPPORT TO THE NETWORK, THESE INSTALLATIONS WILL PROVIDE FIELD DATA FOR FURTHER EVALUATION AND ANALYSIS TECHNIQUE DEVELOPMENT UNDER THIS RTOP. STUDY EFFORTS IN PROGRESS WILL DEFINE THE DESIGN CHARACTERISTICS FOR A HIGH ACCURACY CONTROL SYSTEM WHICH IS REQUIRED FOR FUTURE ANTENNAS OPERATING IN THE KU-BAND SUCH AS THE GROUND STATION IN SUPPORT OF THE TRACKING AND DATA RELAY SATELLITE (TDRS).
/ ANTENNAS/ AUTOMATIC CONTROL/ SATELLITE TRACKING/ SERVOMECHANISMS/ SUPERHIGH FREQUENCIES/ TDR SATELLITES

74W70719 310-20-31

A GROUND ANTENNA FOR WIDEBAND DATA TRANSMISSION SYSTEMS

DOD, L. R. 301-982-5319

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION. GODDARD SPACE FLIGHT CENTER, GREENBELT, MD.

FUTURE ADVANCED SPACECRAFT SYSTEMS WILL TRANSMIT DATA TO THE GROUND AT RATES MUCH HIGHER THAN THAT OF CURRENT OPERATIONAL SYSTEMS. THE EARTH OBSERVATION SATELLITE (EOS) WILL TRANSMIT HIGH RESOLUTION COLOR TV EITHER DIRECTLY TO A GROUND STATION OR VIA A TRACKING AND DATA RELAY SATELLITE (TDRS). THE TDRS WILL TRANSMIT SIGNALS FROM EOS AND OTHER SATELLITES WHICH REQUIRED TOTAL TDRS BANDWIDTHS APPROACHING 1 GHZ. EXISTING NASA GROUND STATIONS ARE NOT EQUIPPED FOR SUCH DATA RATES. FUTURE WIDEBAND COMMUNICATION BY TDRS, EOS AND OTHER PROJECTS, REQUIRE USE OF FREQUENCIES AT WHICH THE NECESSARY BANDWIDTH CAN BE ALLOCATED. A WIDEBAND (APPROXIMATELY 1 GHZ) SYSTEM REQUIRES A HIGH PERFORMANCE GROUND ANTENNA SYSTEM. EMPHASIS ON OVERALL SYSTEM EFFICIENCY WILL BE ESSENTIAL TO AN ECONOMICALLY FEASIBLE GROUND STATION. IN PARTICULAR, TECHNIQUES AND COMPONENTS WILL BE DEVELOPED WHICH YIELD HIGH EFFICIENCY ANTENNA SYSTEMS, FEED SYSTEMS, AND LOW NOISE PREAMPLIFIERS. IN ADDITION, DICHROIC SUBREFLECTOR TECHNIQUES PERMITTING SIMULTANEOUS AND EFFICIENT OPERATION OF AN ANTENNA AT DIFFERENT FREQUENCIES WITHOUT DEGRADATION OF OVERALL PERFORMANCE OR FLEXIBILITY WILL BE REFINED. ANALYTICAL PROCEDURES AND DESIGN TOOLS WILL BE FURTHER DEVELOPED TO SUPPORT THE SPECIFIC REQUIREMENTS OF THESE ADVANCED ANTENNA SYSTEMS AND THE GENERAL ANTENNA DEVELOPMENT PROGRAM.

/ ANTENNAS/ DATA TRANSMISSION/ PARAMETRIC AMPLIFIERS/ TDR SATELLITES/ WIDEBAND COMMUNICATION

74W70716 310-20-20

TRACKING AND DATA RELAY SATELLITE TECHNOLOGY DEVELOPMENT

CLARK, G. Q. 301-982-6331

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION. GODDARD SPACE FLIGHT CENTER, GREENBELT, MD.

THE TWO OBJECTIVES ARE (1) TO PROVIDE FOR THE SIMULATION AND PRELIMINARY DESIGN OF A TRACKING AND DATA RELAY SATELLITE SYSTEM TO BE USED FOR SUPPORT OF NASA MISSIONS, AND (2) TO PROVIDE FOR THE ORDERLY DEVELOPMENT OF THE TECHNOLOGY REQUIRED FOR IMPLEMENTING A FIRST-GENERATION TDRSS BY 1977. VARIOUS STUDIES, SIMULATIONS, AND MODEL FABRICATIONS WILL BE PERFORMED TO ESTABLISH THE PARAMETERS FOR A TDRSS, WHILE OTHER STUDIES WILL IDENTIFY AND PROVIDE SOLUTIONS TO PROBLEMS INHERENT IN THE SYSTEM. IN ADDITION, TECHNOLOGY WILL BE DEVELOPED AS REQUIRED FOR A FIRST-GENERATION TDRSS.

/ SIMULATION/ SYSTEMS ENGINEERING/ TDR SATELLITES

73W70717 150-22-32
HIGH RELIABILITY CONTROL SYSTEMS FOR ANTENNAS
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CONCENTRATION OF DATA ACQUISITION RESPONSIBILITIES AND INCREASING DATA BANDWIDTHS RESULTING FROM REDUCTION IN THE NUMBER OF NETWORK STATIONS ARE PLACING GREATER LOADS ON THE NETWORK LINKS. THUS, THE COST OF LINK DOWN TIME IS INCREASED, REQUIRING A CORRESPONDING INCREASE IN LINK RELIABILITY. THE ANTENNA CONTROL SYSTEM IS ONE OF THE FEW COMPONENTS TO WHICH REDUNDANCY CANNOT BE ECONOMICALLY APPLIED. IN ADDITION, LINK DOWN TIME DUE TO ALIGNMENT REQUIREMENTS AND ROUTINE MAINTENANCE HAS TO BE MINIMIZED. AT THE SAME TIME A REDUCTION IN MAINTENANCE AND OPERATION (M AND O) MANPOWER IS HIGHLY DESIRABLE. ABOVE OBJECTIVES ARE MET BY THE TASKS IN THIS RTOP. THE COMPUTER CONTROLLED ANTENNA SYSTEM HAS DEMONSTRATED A POTENTIAL FOR MARKED REDUCTION IN (M AND O) MANPOWER AND THE FUNCTIONS OF SEVERAL EQUIPMENTS HAVE BEEN SUCCESSFULLY INTEGRATED. THIS SYSTEM IS OPERATING EXPERIMENTALLY AT THE NETWORK TEST AND TRAINING FACILITY (NTTF) AND PROTOTYPE DESIGN HAS BEGUN FOR FY 73 OPERATION. IT WILL SUPPORT THE STADAC SYSTEM AT NTTF TO BE INSTALLED IN THE SAME TIME FRAME. THE ACOUSTICAL ANALYSIS EQUIPMENT FOR DETECTING AND IDENTIFYING INCIPIENT FAILURES IN HYDRAULIC AND MECHANICAL SYSTEMS IS BEING OR HAS BEEN INSTALLED ON TEN NETWORK ANTENNAS. IN ADDITION TO DIRECT SUPPORT TO THE NETWORK, THESE INSTALLATIONS WILL PROVIDE FIELD DATA FOR FURTHER EVALUATION AND ANALYSIS TECHNIQUE DEVELOPMENT UNDER THIS RTOP. STUDY EFFORTS IN PROGRESS WILL DEFINE THE DESIGN CHARACTERISTICS FOR A HIGH ACCURACY CONTROL SYSTEM WHICH IS REQUIRED FOR FUTURE ANTENNAS OPERATING IN THE KU-BAND SUCH AS THE GROUND STATION IN SUPPORT OF THE TRACKING AND DATA RELAY SATELLITE (TDRS).

/ ANTENNAS/ DATA ACQUISITION/ GROUND STATIONS/ SERVOMECHANISMS/ TDR SATELLITES/ TELECOMMUNICATION

73W70709 150-22-20
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/ DATA TRANSMISSION/ SATELLITE NETWORKS/ SATELLITE TRACKING/ TDR SATELLITES/ TELECOMMUNICATION

73W70539 164-21-55
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THE TWO OBJECTIVES ARE 1) TO PROVIDE FOR THE DEFINITION OF A TRACKING AND DATA RELAY SATELLITE SYSTEM TO BE USED FOR SUPPORT OF NASA MISSIONS, AND 2) TO PROVIDE FOR THE ORDERLY DEVELOPMENT OF THE TECHNOLOGY REQUIRED FOR IMPLEMENTING A FIRST GENERATION TDRSS BY 1977. VARIOUS STUDIES WILL BE PERFORMED TO ESTABLISH THE CRITERIA FOR A TDRSS WHILE OTHER STUDIES WILL LOOK FOR SOLUTIONS TO PROBLEMS INHERENT IN THE SYSTEM. IN ADDITION, TECHNOLOGY WILL BE DEVELOPED AS REQUIRED FOR A FIRST-GENERATION TDRSS.
/ DATA TRANSMISSION/ SATELLITE NETWORKS/ TDR SATELLITES/ TELECOMMUNICATION/ TRACKING NETWORKS

72A27523* ISSUE 12 PAGE 1970 CATEGORY 30 72/04/00 6 PAGES UNCLASSIFIED DOCUMENT
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(SPACE TECHNOLOGY APPLICATION TO ATS F AND G PROGRAM, DISCUSSING HIGH POWER REQUIREMENTS, PARABOLIC ANTENNA DESIGN, TRACKING ACCURACY AND GROUND STATION SIMPLIFICATION)
A/GERWIN, H. A/(NASA, GODDARD SPACE FLIGHT CENTER, GREENBELT, MD.)
JOURNAL OF ENVIRONMENTAL SCIENCES, VOL. 15, MAR.-APR. 1972, P. 12-17.
/*APPLICATIONS TECHNOLOGY SATELLITES/*SATELLITE ANTENNAS/*SPACECRAFT DESIGN/*TECHNOLOGY UTILIZATION/ ANTENNA DESIGN/ DIRECTIONAL ANTENNAS/ GROUND STATIONS/ PARABOLIC ANTENNAS/ POWER GAIN/ SYNCHRONOUS SATELLITES/ TDR SATELLITES

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/*COST EFFECTIVENESS/*REAL TIME OPERATION/*TDR SATELLITES/ GROUND STATIONS

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   (ADAPTIVE PHASED ANTENNA ARRAY SIMULATION FOR OPTIMAL DESIGN OF
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   A/SPEARING, R. E.
   NATIONAL AERONAUTICS AND SPACE ADMINISTRATION. GODDARD SPACE FLIGHT
CENTER, GREENBELT, MD.
    IN ITS SIGNIFICANT ACCOMPLISHMENTS IN TECHNOL., 1972   P 103-108
(SEE N73-27816 18-34)
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SATELLITES/ COMPUTERIZED SIMULATION/ GROUND BASED CONTROL/ PHASED
ARRAYS/ RADIO FREQUENCY INTERFERENCE

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   (DESIGN, DEVELOPMENT, AND CHARACTERISTICS OF SPACECRAFT SYSTEMS AND
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EQUIPMENT/ SPACE COMMUNICATION/ SYSTEMS ENGINEERING

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/*COMPUTER PROGRAMS/*HEAO/*MANAGEMENT PLANNING/*PROJECT MANAGEMENT/*SPACECRAFT CONFIGURATIONS/ MANAGEMENT METHODS/ ORGANIZATIONS/ POLICIES

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/*HAZARDS/*HEAO/*SAFETY MANAGEMENT/ EDUCATION/ SAFETY FACTORS/ SYSTEMS ENGINEERING/ WARNING SYSTEMS

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/*DATA PROCESSING/*GROUND OPERATIONAL SUPPORT SYSTEM/*HEAO/ COMPUTER PROGRAMS/ GROUND SUPPORT EQUIPMENT/ OGO/ TRACKING NETWORKS/ TRACKING STATIONS

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/*HEAO/*QUALITY CONTROL/*RELIABILITY ENGINEERING/ COMPONENT RELIABILITY/ COST EFFECTIVENESS/ SPACECRAFT COMPONENTS

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(SUBSYSTEM RELIABILITY, HIGH RISK/LONG LEAD ITEMS, TRADEOFFS, COST ANALYSIS, AND SUPPORTING RESEARCH AND DEVELOPMENT - VOL. 2 - SECTS. 6-9)
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(MISSION AND OPERATIONS ANALYSIS, SYSTEM DESIGN AND ANALYSIS, LAUNCH VEHICLE, GROUND SUPPORT EQUIPMENT, AND OPERATIONS PLANNING - VOL 2, SECTS. 4 AND 5)
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/*HEAO/*NASA PROGRAMS/*PROJECT PLANNING/ COSTS/ LOGISTICS/ PROJECT MANAGEMENT/ QUALITY CONTROL/ RELIABILITY ENGINEERING/ SAFETY/ TEST FACILITIES

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/*HEAO/*SPACECRAFT DESIGN/*SYSTEMS ENGINEERING/ ATTITUDE CONTROL/ DATA PROCESSING/ GROUND SUPPORT EQUIPMENT/ TELECOMMUNICATION/ TELEMETRY/ WEIGHT (MASS)

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/*HEAO/*SPACECRAFT DESIGN/*STRUCTURAL DESIGN/*SYSTEMS ENGINEERING/ LAUNCH VEHICLES/ PAYLOADS/ STRUCTURAL MEMBERS/ WEIGHT (MASS)

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(TECHNICAL ANALYSIS AND DETAILED DESIGN DATA FOR HEAO - VOL. 2, PART 1)

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HIGH ENERGY ASTRONOMY OBSERVATORY. VOLUME 1 - EXECUTIVE SUMMARY FINAL REPORT

(SPACECRAFT DESIGN AND MISSION EXPERIMENTS FOR HIGH ENERGY ASTRONOMY OBSERVATORY PROGRAM - VOL. 1, EXECUTIVE SUMMARY)

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/*HEAO/*MISSION PLANNING/*SPACECRAFT DESIGN/ COST REDUCTION/ GAMMA RAYS/ TECHNOLOGY UTILIZATION/ TRADEOFFS/ X RAYS

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FEASIBILITY STUDY OF A HIGH ENERGY ASTRONOMY OBSERVATORY /HEAO/
SPACECRAFT. VOLUME 1 - SUMMARY REPORT

(CONCEPTUAL DESIGN OF HIGH ENERGY ASTRONOMY OBSERVATORY SPACECRAFT
AND SUBSYSTEMS - VOL. 1)

A/DUFFIE, J. M.; B/WATSON, R. C., JR.   (AAED. ABED.)

BROWN ENGINEERING CO., INC., HUNTSVILLE, ALA.   (SCIENCE AND
ENGINEERING GROUP.)

/*HEAO/*SPACECRAFT COMPONENTS/*SPACECRAFT DESIGN/ ASTRONOMY/
ATTITUDE CONTROL/ ELECTROMAGNETIC RADIATION/ GROUND BASED CONTROL/ LIFE
(DURABILITY)/ SPACECRAFT COMMUNICATION/ TEMPERATURE CONTROL/ WEIGHT
(MASS)

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EFFECTS OF THE MAGNETIC SPECTROMETER EXPERIMENT ON HEAO-B AND HEAO-D
SPACECRAFT SUMMARY REPORT

(EFFECTS OF MAGNETIC SPECTROMETER EXPERIMENT ON OPERATION OF HEAO-B
AND HEAO-D SPACECRAFT)

A/DUFFIE, J. M.; B/ROSNER, H. R.; C/SCARBOROUGH, J. M.

TELEDYNE BROWN ENGINEERING, HUNTSVILLE, ALA.   (RESEARCH AND
ENGINEERING DEPT.)

/*HEAO/*MAGNETIC SPECTROSCOPY/*SPACECRAFT PERFORMANCE/ FIELD COILS/
GEOMAGNETISM/ MAGNETIC MOMENTS/ MAGNETIC SHIELDING/
TORQUE

74W70651   188-46-64

ASTROPHYSICAL INVESTIGATIONS ON THE SPACE SHUTTLE

OPP, A. G.   202-755-3698

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION, WASHINGTON, D.C.

THE SPACE SHUTTLE REPRESENTS THE NEXT MAJOR DEVELOPMENT OF A FLIGHT
OPPORTUNITY IN HIGH ENERGY ASTROPHYSICS BEYOND HEAO. THE CONCEPTS AND
PARAMETERS FOR THE NEXT GENERATION OF SPACECRAFT INSTRUMENTATION HAVE
BEGUN TO EVOLVE FROM THE SPACE SHUTTLE WORKING GROUP. MOST OF THE
INSTRUMENTATION EXISTS IN CONCEPTUAL FORM ONLY. IN ORDER TO ASSURE THAT
THE INSTRUMENTS ARE DEVELOPED AND TESTED ON A TIME SCALE COMMENSURATE
WITH THE FLIGHT SCHEDULES OF THE SHUTTLE, IT IS NECESSARY TO BEGIN AT
THIS TIME THE SUPPORT OF SEVERAL INVESTIGATORS WHO ARE INTERESTED IN
CARRYING OUT SUCH INVESTIGATIONS ON THE SHUTTLE. THE FUNDS PROVIDED
UNDER THIS RTOP WILL SUPPORT THE DEVELOPMENT OF VERY HIGH ENERGY
CHARGED PARTICLE DETECTORS, LARGE GAMMA RAY DETECTORS AND THE STUDY OF
DISCIPLINE UNIQUE REQUIREMENTS, WHICH MIGHT BE PLACED ON A SHUTTLE
FACILITY.

/ ASTROPHYSICS/ GAMMA RAYS/ HEAO/ RADIATION COUNTERS/ RADIATION
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THE ORIGIN OF COSMIC RADIATION

(GALACTIC NUCLEI, PULSARS AND SUPERNOVAE AS SOURCES OF PRIMARY COSMIC RAYS FROM GROUND BASED AND SATELLITE OBSERVATIONS, RELATING CHEMICAL COMPOSITION TO ORIGIN)

A/AUDOUZE, J.; B/MENEGUZZI, M.

LA RECHERCHE, VOL. 4, JUNE 1973, P. 549-555. IN FRENCH.

/*CHEMICAL COMPOSITION/*GALACTIC NUCLEI/*PRIMARY COSMIC RAYS/*PULSARS/*SUPERNOVAE/ ABUNDANCE/ ENERGY SPECTRA/ HEAO/ HEAVY IONS/ HIGH ENERGY ELECTRONS/ PARTICLE ACCELERATION/ PROTON ENERGY/ SATELLITE OBSERVATION

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A POSITION-SENSITIVE X-RAY DETECTOR FOR THE HEAO-A SATELLITE.

A/HELD, D.; B/WEISSKOPF, M. C. B/(COLUMBIA UNIVERSITY, NEW YORK, N.Y.)

(IEEE, AEC, AND NASA, NUCLEAR SCIENCE SYMPOSIUM, 19TH, MIAMI, FLA., DEC. 6-8, 1972.) IEEE TRANSACTIONS ON NUCLEAR SCIENCE, VOL. NS-20, FEB. 1973, P. 140-144.

/*HEAO/*PROPORTIONAL COUNTERS/*RADIATION DETECTORS/*SATELLITE-BORNE INSTRUMENTS/*X RAYS/ ENERGY DISTRIBUTION/ POSITION INDICATORS/ SIGNAL PROCESSING/ TELEMETRY

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HIGH-ENERGY RADIATIONS FROM SPACE.

(HIGH ENERGY ASTRONOMY RESEARCH IN SPACE, DISCUSSING HEAO A AND B, UV ASTRONOMY, X RAY ASTRONOMY, GAMMA RAYS, COSMIC RAYS, HOT STARS, STELLAR ENERGY SOURCES AND ELEMENTARY PARTICLES)

A/STUHLINGER, E.; B/BAILEY, C. B/(NASA, MARSHALL SPACE FLIGHT CENTER, HUNTSVILLE, ALA.)

(NEW YORK ACADEMY OF SCIENCES, CONFERENCE ON PLANETOLOGY AND SPACE MISSION PLANNING, 3RD, NEW YORK, N.Y., OCT. 28-30, 1970.) NEW YORK ACADEMY OF SCIENCES, ANNALS, VOL. 187, JAN. 25, 1972, P. 234-250.

/*COSMIC RAYS/*GAMMA RAYS/*HEAO/*SPACEBORNE ASTRONOMY/*X RAY ASTRONOMY/ ELEMENTARY PARTICLES/ ENERGY SOURCES/ HOT STARS/ STELLAR SPECTRA/ ULTRAVIOLET SPECTRA

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/*CHEMICAL COMPOSITION/*GALACTIC NUCLEI/*PRIMARY COSMIC RAYS/*PULSARS/*SUPERNOVAE/ ABUNDANCE/ ENERGY SPECTRA/ HEAO/ HEAVY IONS/ HIGH ENERGY ELECTRONS/ PARTICLE ACCELERATION/ PROTON ENERGY/ SATELLITE OBSERVATION

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/*HEAO/*PROPORTIONAL COUNTERS/*RADIATION DETECTORS/*SATELLITE-BORNE INSTRUMENTS/*X RAYS/ ENERGY DISTRIBUTION/ POSITION INDICATORS/ SIGNAL PROCESSING/ TELEMETRY

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HIGH-ENERGY RADIATIONS FROM SPACE.

(HIGH ENERGY ASTRONOMY RESEARCH IN SPACE, DISCUSSING HEAO A AND B, UV ASTRONOMY, X RAY ASTRONOMY, GAMMA RAYS, COSMIC RAYS, HOT STARS, STELLAR ENERGY SOURCES AND ELEMENTARY PARTICLES)

A/STUHLINGER, E.; B/BAILEY, C. B/(NASA, MARSHALL SPACE FLIGHT CENTER, HUNTSVILLE, ALA.)

(NEW YORK ACADEMY OF SCIENCES, CONFERENCE ON PLANETOLOGY AND SPACE MISSION PLANNING, 3RD, NEW YORK, N.Y., OCT. 28-30, 1970.) NEW YORK ACADEMY OF SCIENCES, ANNALS, VOL. 187, JAN. 25, 1972, P. 234-250.

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D/(CALIFORNIA, UNIVERSITY, BERKELEY, CALIF.)
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A/GURSKY, H. A/(AMERICAN SCIENCE AND ENGINEERING, INC., CAMBRIDGE, MASS.)

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A/OSTRIKER, J. P.

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A/DAILEY, C. C.   A/(NASA, MARSHALL SPACE FLIGHT CENTER, HUNTSVILLE, ALA.)
INTERNATIONAL ASTRONAUTICAL FEDERATION, INTERNATIONAL ASTRONAUTICAL CONGRESS, 23RD, VIENNA, AUSTRIA, OCT. 8-15, 1972, PAPER. 32 P.
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A/FRIEDMAN, H. A/(U.S. NAVY, E. O. HULBURT CENTER FOR SPACE RESEARCH, WASHINGTON, D.C.)
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A/WEIDNER, D. K.; B/WEST, G. S. (AAED. ABED.)
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A/DUFFIE, J. M.   (AAEC.)

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