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DIGITAL DATA TRANSMISSION IN A PULSE AMPLITUDE MODULATION (PAM) COMMUNICATION SYSTEM

(AIRTASK A5355352 054D 5W47410030, Work Unit A535210000002)

By

D. A. KING
Instrumentation Development Division

14 August 1975

PACIFIC MISSILE TEST CENTER

Point Mugu, California

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Technical Director

This report describes work accomplished under AIRTASK A5355352 054D 5W47410030, Missile Flight Evaluation Systems, Work Unit A535210000002.

Mr. D. R. Knight, Head, Systems Development Branch and Mr. K. L. Berns, Head, Instrumentation Development Division, have reviewed this report for publication.

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This report presents the results of an experimental evaluation of the effectiveness of digital data transmission in a pulse-amplitude modulation (PAM) communication system. Some sources of PAM system errors are discussed to clarify the nature of the problem. A PAM communication system was simulated and equipment was developed to obtain experimental results indicative of PAM's effectiveness in transmission of digital data. The results consist of word, channel, and bit error probability curves for different numbers of digital bits per PAM channel. The digital bits are

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converted to analog voltage levels which modulate selected PAM channels. After decommutation, the digital bits are recovered from the analog voltage levels.

The data indicates that the number of bits per channel should be 4 or less with 4 being marginal depending upon the digital data quality requirements. The error probability curves can be used to determine whether or not a specific PAM communication system can telemeter digital data within given constraints. In addition, the results provide guidelines for the implementation of digital data in a PAM system.

ACKNOWLEDGMENT

The author wishes to thank F. R. Hartzler, technical leader, and D. R. Knight, supervisor, for their counseling during the project and to acknowledge the efforts of E. L. Law who provided invaluable assistance throughout the project.

FOREWORD

The work reported herein, involves an evaluation of digital data transmission via a PAM communication system. It was performed under AIRTASK A5355352 054D 5W47410030, Missile Flight Evaluation Systems, Work Unit A535210000002, to provide support to the Telemetry Group of the Range Commanders Council. The purpose of this report is twofold: (1) Part of a project concerning evaluation of telemetry systems with hybrid data requirements (combinations of analog, digital, voltage-controlled oscillator, and video doppler data signals), and (2) To satisfy one of the requirements for the Master of Science degree in Engineering obtained through the Technical Professional program at Pacific Missile Test Center, Point Mugu.

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SUMMARY

This report presents the results of an experimental evaluation of the effectiveness of digital data transmission in a pulse-amplitude modulation (PAM) communication system. Some sources of PAM system errors are discussed to clarify the nature of the problem. A PAM communication system was simulated and equipment was developed to obtain experimental results indicative of PAM's effectiveness in transmission of digital data. The results consist of word, channel, and bit error probability curves for different numbers of digital bits per PAM channel. The digital bits are converted to analog voltage levels which amplitude-modulate selected PAM channels. After decommutation the digital bits are recovered from the analog voltage levels.

The data indicates that the number of bits per channel should be 4 or less with 4 being marginal depending upon the digital data quality requirements. The error probability curves can be used to determine whether or not a specific PAM communication system can telemeter digital data within given constraints. In addition, the results provide guidelines for the implementation of digital data in a PAM system.

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INTRODUCTION

The effectiveness of any telemetry system is primarily determined by its informational quantity and quality characteristics within certain constraints. Pulse-amplitude modulation (PAM) systems have proven to be effective and are currently in wide use. However, application of existing telemetry systems such as PAM, to new telemetry requirements of newly developed missile systems requires that telemetry systems be flexible and adaptive. Furthermore, adapting an existing telemetry system to new requirements ideally should not reduce its effectiveness. If it does, then the system is of limited utility and the new application may require the development or acquisition of a costly new telemetry system. The application to PAM of a new telemetry data requirement and the resulting effectiveness of PAM in handling that data is the motivation behind this report.

In the past, information transmitted by PAM systems have mainly been continuous (analog) in nature, but the development of Navy missiles containing minicomputers for control and data processing has resulted in PAM telemetry being required to accommodate various amounts of digital data. Since many Navy telemetry systems involve PAM, and if these systems are to be used in support of missiles with digital data requirements, then it is necessary to know the digital data transmission characteristics of a PAM system. This report is aimed at determining these characteristics and their ramifications on the effectiveness of PAM with digital data. It presents the results of an experimental evaluation of digital data transmission in a PAM communication system. Some sources of PAM system errors are discussed to clarify the nature of the problem. A PAM communication system was simulated and interface, signal source, and error detection equipment were developed to obtain experimental results indicative of PAM's effectiveness in transmission of digital data. The results consist of word error, channel error, and bit error probability curves for different numbers of digital bits per PAM channel. The digital bits are converted to analog voltage levels which amplitude-modulate selected PAM channels. After decommutation, the digital bits are recovered from the analog voltage levels.

This report does not support PAM as a replacement of pulse-code modulation but is merely an evaluation of a method to telemeter limited amounts of digital data within existing hardware constraints.

PAM SYSTEM ERRORS

All PAM telemetry system components introduce errors in the telemetry data. Basic PAM system components consist of a commutator, a premodulation filter, a radio frequency (RF) transmitter, a receiver, a video filter, and a decommutator. An understanding of the major system errors introduced by these components is useful in characterizing the PAM data quality at the decommutator output. The major system errors are caused by receiver noise characteristics, RF power level, component nonlinearities, channel crosstalk, incidental frequency modulation (FM), and circuit noise.

For this report, noise is defined as random perturbations of voltage or frequency. Circuit noise results from induced random voltage fluctuations on a signal as it passes through components of the circuit. For example, resistor noise is caused by thermal agitation; semiconductor noise, commonly called shot noise, is caused by current flow across semiconductor junctions. The result of circuit noise is to limit the resolution of a signal from the system components.

Incidental FM is defined as random frequency fluctuations in the RF carrier of the transmitter. The carrier is frequency-modulated by random noise internal to the transmitter. The random frequency modulation of the carrier results in random voltage fluctuations in the video signal upon demodulation in the receiver.

Channel crosstalk is the interference of one channel with another and is usually caused by premodulation, video, or receiver filtering and commutator overmodulation of a data channel. Either can induce voltages in the data channels and cause signal distortion.

System component nonlinearities also result in signal distortion. They occur when the characteristic curve of a device is nonlinear.

The RF signal power level at the receiver input determines the signal-to-noise ratio (SNR) in the receiver and decommutator. At high RF signal levels, the SNR is relatively fixed by the noise floor of the system, as the signal power level drops so does the SNR. If the signal power level drops too low, the signal becomes buried in transmission and circuit noise causing difficulty in video signal recovery. Transmission noise is interference picked up in the transmission medium and the circuit noise, as defined previously, is from the transmitter and receiver.

Noise characteristics in a receiver are caused by signal filtering, RF signal power level, and the demodulation process.* The signal is filtered in the receiver before the demodulator to eliminate unwanted signals and to limit noise power. The demodulator output is proportional to the frequency of its input signal. As the signal power drops, the automatic gain control in the receiver amplifies the signal to maintain required receiver voltage levels. The amplification also amplifies the noise. When the noise magnitude at the demodulator input is less than that of the signal, random voltage fluctuations proportional to the noise frequency are induced on the video signal at the demodulator output. They are of a nature that, for the most part, allows them to be averaged to zero by the low-pass filter that follows the demodulator. When random noise fluctuations are larger in magnitude than the signal at the demodulator input, the demodulator is "captured" by the noise. The demodulator output voltage becomes heavily dependent upon the noise frequency at its input. The capture results in two types of noise at the demodulator output. One type is essentially averaged to zero in the low-pass filter but the other, called subtractive noise, is of a type that cannot be averaged to zero. Subtractive noise is characterized by voltage

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* E. R. Hill. A Nonstatistical Treatment of FM Demodulator Noise. Missile Systems Department, Naval Ordnance Laboratory, Corona (NOLC Report 724) 1967, p. 19-25 UNCLASSIFIED.

spikes which are usually directed toward the band center carrier modulation voltage. Filtering before demodulation concentrates the noise power about the band center frequency. Thus when the modulation level is deviated from center frequency, noise at band center usually captures the demodulator. Subsequent low-pass filter averaging suppresses the video signal. The point at which noise is comparable to the signal is called the FM threshold. As the SNR drops below threshold, a rapid degradation of the video signal occurs because of the subtractive noise.

The manner in which these factors affect the quality of the decommutator output signals can be illustrated by an examination of the decommutator operation. The decommutator used in this study was comprised of (1) An integrator which integrates each PAM channel sample over a fixed period within the channel to a final value equal to the sample's amplitude; (2) Sample/hold circuits that reconstruct the commutated signals by sampling the appropriate amplitude from the integrator and holding it until the next integrated sample of that signal arrives; (3) Timing circuits to control the integration and sample/hold functions and to maintain decommutator synchronization; and (4) Calibration circuits to establish internal calibration levels and to compare these levels with the data calibration correction signals.

The amplitudes of the samples of the PAM sequence after demodulation are biased by nonlinearities of preceding system components; in addition, superimposed on the amplitudes are random demodulator noise, circuit noise, and crosstalk. The low-pass filter tends to average out any mean-value-zero noise, but any noise that is not mean-value-zero cannot be averaged out and causes further biasing of the samples' amplitudes. Noise below the low-pass filter cutoff is passed along with the PAM sequence to the decommutator's calibrator and integrator.

The integration and sample/hold process in the decommutator acts to average the noise and crosstalk remaining on the biased amplitudes of the sample pulses of the video signal. Because of the nature of noise and crosstalk, and because of the decommutator's own nonlinearities, the averaging process generally results in further biasing of the amplitudes. Therefore, the decommutator output voltage level is usually not a completely accurate representation of the sampled signal. The degree to which the decommutator output amplitude differs in value from the actual sample amplitude depends upon the state of the system. State is defined as the RF signal power level and the degree of nonlinearities, crosstalk, circuit noise, incidental FM, and demodulator noise present at any given time in the system.

The biases at the decommutator output due to nonlinearities and crosstalk depend upon the modulation level of the channel and the modulation level of the preceding channel, respectively. Neglecting time and temperature variations, the net bias, at any given modulation level, is a static error. Since both demodulator noise (of which incidental FM is a part) and circuit noise are random, the biases they cause on the decommutator output amplitude are random. These random biases are known as decommutator noise. Consequently, the decommutator output voltage is randomly distributed about the true amplitude of the sample and is characterized by a probability density function (PDF). The mean of the PDF distribution is shifted by the biases of nonlinearities and crosstalk such that it may not be the amplitude of the actual sample. The distribution in general is dependent upon the channel modulation level, crosstalk and the system noise floor.

The PDF distribution is also dependent upon the power level of the RF signal because of the random demodulator noise. At high RF signal levels, the PDF distribution at the decommutator output is that of the basic system noise floor and is primarily determined by nonlinearities, crosstalk, circuit noise, and incidental FM. For a given system, these factors are usually specified and maintained within certain tolerances and thus determine the best accuracy obtainable at the decommutator output. As the RF level decreases, the demodulator noise increases and the decommutator output distribution increases. Accordingly, the overall accuracy at the decommutator output decreases with the RF level. Because of the rapid degradation of the video signal by subtractive noise, FM threshold should have a significant effect upon the decommutator output distribution and PAM data accuracy.

DIGITAL DATA QUALITY AND QUANTITY IN A PAM SYSTEM

Digital data can be handled by a PAM communication system if the digital data is first converted to analog voltage levels by means of a digital-to-analog (D/A) conversion. Once converted, the data can be sampled by the system commutator and transmitted with other data to a receiving station. The received analog voltage levels are separated from the other data in the decommutator and reconverted to digital data by an analog-to-digital (A/D) conversion.

When N bits of digital data are converted to an analog voltage level and this level is sampled by the commutator, transmitted, received, and decommutated, the analog voltage level out of the decommutator will, in general, not be exactly the same as that sampled by the commutator. The difference between the sampled value and the recovered value will be a randomly distributed variable whose distribution depends upon the RF power, the channel modulation level, crosstalk, and the system noise floor. Since recovery of the digital data requires an A/D conversion, an accurate recovery of the N bits requires that the difference be smaller than 1/2 the voltage representative of the least significant bit (LSB) of the correct code's voltage level; otherwise, the conversion will be in error.

If the difference distribution curve's end points lie within $\pm 1/2$ LSB of the correct code's voltage level, then there will essentially be no digital recovery errors. If the distribution curve's end points exceed $\pm 1/2$ LSB, there will exist some probability that the decommutator output voltage will cause an A/D conversion error. The probability is proportional to the area under those parts of the distribution curve exceeding $\pm 1/2$ LSB. Since the distribution of the decommutator output error is variable with RF power, the probability of a digital recovery error will also be variable. Digital errors can be reduced by increasing the size of the LSB. This increase is accomplished by decreasing N, the number of bits per channel.

If quality constraints require N bits per channel and $N < M$ (where M is the number of bits per word) then the word must be divided into segments and each segment converted and sampled in order to transmit the word. The word must be reconstructed when the segments are recovered. The N bits per channel each has different error probabilities because each is not equally weighted in the D/A and A/D conversions. Consequently, consideration must be given to the arrangement and packing of the M bits in the N locations of the channels so as to maximize final data quality.

The previously mentioned best accuracy obtainable at the decommutator output limits the quantity of digital data possible per PAM channel for a given recovery quality. For example, if $N = 5$, then the voltage representative of the LSB for 10 volts (V) full scale is

$$\frac{1}{2^5}(10\text{V}) = 312 \text{ millivolts (mV)}$$

Accuracy of PAM systems used in the field is at best approximately $\pm$ 2 percent of full scale. Two percent of 10 volts is equal to 200 mV which is greater than 1/2 the LSB for $N = 5$. This indicates that digital errors will sometimes result for system operation under best conditions with $N = 5$. Therefore, the practical range of N is limited to 4 bits per channel or less. Obviously, the quality and quantity of recovered digital data on PAM are inversely proportional to one another. For a fixed RF signal power level, the quantity of digital data per PAM channel decreases as the quality requirement becomes more stringent; or, for a given quality requirement, the quantity of digital data per channel decreases as the RF power decreases.

DESCRIPTION OF EXPERIMENT

The measure of effectiveness used in evaluating digital data transmission in a PAM communications system was quantity and quality of recovered digital data. Accordingly, the experiment was designed to produce data relevant to this measure. The required experimental data is listed below:

1. Bit error probabilities for each of the N bits per channel.
2. Channel error probability.
3. Word error probability.
4. RF power level for 1, 2, and 3 above.
5. IF SNR for 1, 2, and 3 above.
6. Decommutator output SNR for 1, 2, and 3 above.
7. System parameters consisting of

    a. Sample rate
    b. Premodulation filter bandwidth
    c. Transmitter deviation
    d. IF filter and demodulator bandwidth
    e. Low-pass filter bandwidth

A bit error for any of the recovered N bits per channel is defined to be the occurrence of a digital "1" ("0") level when the correct digital level is a "0" ("1"). A channel error is defined to be the occurrence of one or more bit errors within the N bits of the channel. A word error is defined to be the occurrence of one or more bit errors within the M bits of the word or one or more channel errors within the K channels of the word. Items 4, 5, and 6 in the data list are related to each other and provide reference points for items 1, 2, and 3 in the list. Item 7 indicates that there are several operating modes in which the PAM system can operate. A mode is determined by selecting parameters a through c of item 7. Items 1 through 6 are dependent upon the mode. The experiment examined the digital data quality (error probabilities) as a function of signal power level at the receiver input for a given quantity of bits, N, per channel for three modes of the system.

The experiment required the simulation of a PAM communication system and the development of a digital signal generator; a D/A converter-interface; an A/D interface-converter; and a bit, channel, and word error detector. The digital signal generator simulated a digital information source. The schematic of the generator is shown in the appendix (figure A-1). This digital source was a pseudo-random bit generator capable of supplying up to six bits in parallel to the D/A converter-interface and the error detector. The generator was not free-running but supplied new bits to the D/A interface-converter and error detector upon command before the next sampling of the D/A occurred.

The D/A converter-interface acted as a buffer between the digital signal generator and the PAM commutator. The function of the D/A converter interface was to convert digital information to an analog voltage level and to condition the D/A output for interfacing with the commutator before the sampling occurred. The schematic of the D/A converter-interface is shown in the appendix (figure A-1). The circuit selected by switch the number of bits per channel, N, to be converted. The pseudo-random digital input to the D/A resulted in a pseudo-random sequence of voltage levels at the D/A output. For N = 4, there were 16 random levels; for N = 3, there were 8 random levels; for N = 2, there were 4 random levels; and for N = 1, there were 2 random levels.

The A/D interface converter was the digital recovery unit. The interface section coupled the decommutator's integrator output to the A/D converter. It consisted of a sample/hold (S/H) device and timing circuits. Decommutator timing was used to control the S/H device so that the sampled voltage levels of the integrator were those that represent digital data. The voltage level of the integrator was held by the S/H and an A/D conversion was performed to recover the digital data. The A/D interface-converter schematic is shown in the appendix (figure A-2).

The error detector compared the A/D interface-converter output with the original signal from the pseudo-random generator. The detector operated in three modes: bit error, channel error, and word error detection. Bit and channel comparisons occurred for every sample of digital data; a word comparison was made by comparing the K consecutive channels comprising the word and noting if any one was in error. Once the error comparison for a single channel was made, a command pulse was sent to the pseudo-random bit generator to clock a new code to the D/A converter interface. If an error was detected, it was summed with others on one channel of a ratio counter On the other channel of the ratio counter, pulses representative of the number of bits, channels, or words compared were counted. As the system operated, many comparisons occurred. The ratio counter counted to a preset number of bits, channels, or words to be compared and automatically divided the total number of errors counted by the number of comparisons to give the bit error, channel error, or word error probability. A schematic of the error detector is given in the appendix (figure A-3). A brief description of the interconnections between circuits is also given in the appendix.

The PAM simulation was made up of standard telemetry components. The transmitting end consisted of a commutator with selectable sample rate and premodulation filter, and an L-band transmitter operating at 1483 MHz. The transmission medium was simulated with RF coaxial cable and 1- and 10-decibel (dB) step attenuators. The receiving end consisted of an L-band receiver with selectable IF filter and demodulator bandwidths, and a decommutator with selectable sample rate, video filter bandwidth, and manual or automatic synchronization. The IF filter precedes the receiver demodulator and limits the noise power to the demodulator. A diagram of the PAM simulation is shown in figure 1.

Several system parameters were held constant during the experiment: The decommutator was always operated in the automatic synchronization mode, the transmitter frequency was held constant at 1483 megahertz, the receiver automatic gain control time constant was set to 1 millisecond, and M, the number of bits per word, was fixed at 12.

Three operating modes of the PAM simulation were selected for experimentation. They are typical of many PAM systems. These modes are as follows:

Mode I. Sample Rate: 25 kHz, non-return to zero (NRZ), 64 channels
Premodulation Filter: 50 kHz
Transmitter Deviation: ±125 kHz
IF Filter: 1 MHz
Demodulator Bandwidth: 1 MHz
Video Filter: 50 kHz

Mode II. Sample Rate: 25 kHz, NRZ, 64 channels
Premodulation Filter: 50 kHz
Transmitter Deviation: ±125 kHz
IF Filter: 500 kHz
Demodulator Bandwidth: 500 kHz
Video Filter: 50 kHz

Mode III. Sample Rate: 100 kHz, NRZ, 64 channels
Premodulation Filter: 200 kHz
Transmitter Deviation: ±250 kHz
IF Filter: 1 MHz

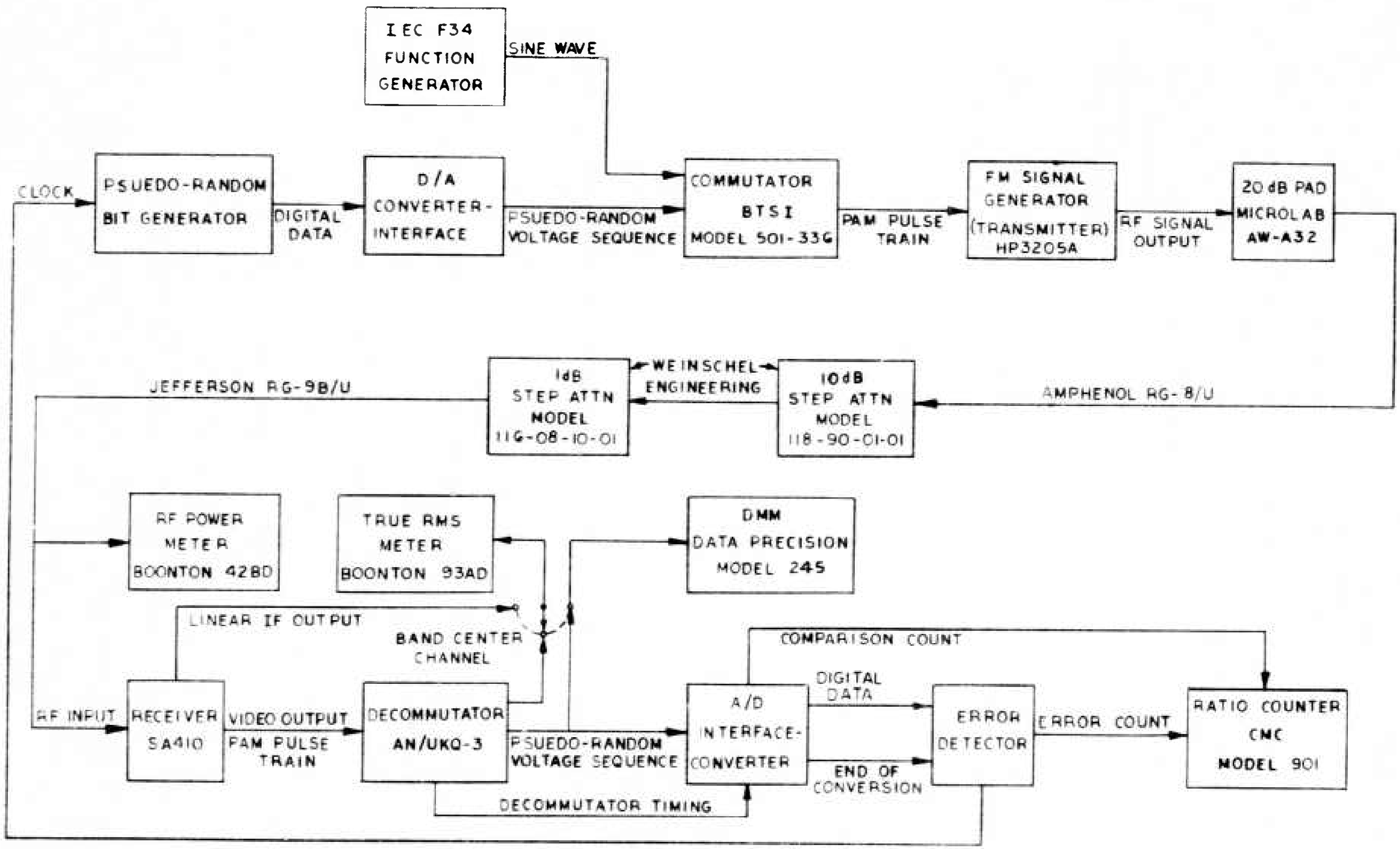

Figure 1. Block Diagram of PAM Simulation and Test Equipment.

Demodulator Bandwidth: 1 MHz
Video Filter: 200 kHz

Mode I was selected for most of the experimentation. The other two modes were examined to determine the effects of a change in IF filter bandwidth (a change in noise power to the demodulator) and sampling rate.

A block diagram of the experimental setup is shown in figure 1. The equipment used is listed in table 1 and listed in table 2 are the PAM frame channel assignments. Channels 60 through 64 comprise the decommutator frame synchronization code. Every fourth channel from 3 to 59 was reserved for digital information. Every fourth channel from 2 to 58 was modulated with a low frequency (1K at 25K sample rate and 2K at 100K sample rate) sine wave to introduce typical crosstalk distortion in the digital information channels. Every eighth and ninth channel from 8 to 56 was tied to full-scale high and low modulation levels respectively to ensure enough channel voltage transitions to maintain decommutator timing. All other channels were grounded. Figure 2 illustrates the PAM frame without the sine wave modulation. The lower trace clearly shows the eight modulation levels for N = 3 on a digital information channel.

Table 1. Equipment List

PAM Simulation

1. Signal Simulator (commutator); Base Ten Systems, Inc; Model 501-336

2. FM Signal Generator (Transmitter); Hewlett-Packard 3205A

3. Coaxial Cable; Amphenol RG-8/u and Jefferson RG-9B/u

4. 20 dB Pad; Microlab AW-A32

5. One and ten dB step attenuators; Weinschel Engineering; Models 118-90-01-01 and 116-08-01-01

6. Receiver; Scientific Atlanta, Inc; Model 410

7. Signal Conditioner and Converter (decommutator); AN/UKQ-3

Miscellaneous

1. Frequency Counter; Computer Measurements Co; Model 901

2. RF Power Meter; Boonton Electronics; Model 42BD

3. True RMS Meter; Boonton Electronics; Model 93AD

4. Digital Multi-Meter; Data Precision; Model 245

5. Function Generator; Interstate Electronics Corporation; Model F34

6. Oscilloscope; Tektronix; Type 585A

7. Power Supply, ± 15V; Instant Instruments, Inc; Model D152

8. Power Supply, + 5V; Instant Instruments, Inc; Model 510

Table 2. Frame Channel Assignments

1. GND	26. EXT 2	51. EXT 1
2. EXT 2	27. EXT 1	52. GND
3. EXT 1	28. GND	53. GND
4. GND	29. GND	54. EXT 2
5. GND	30. EXT 2	55. EXT 1
6. EXT 2	31. EXT 1	56. +2.5V
7. EXT 1	32. +2.5V	57. −2.5V
8. +2.5V	33. −2.5V	58. EXT 2
9. −2.5V	34. EXT 2	59. EXT 1
10. EXT 2	35. EXT 1	60. −2.5V
11. EXT 1	36. GND	61. +2.5V
12. GND	37. GND	62. +2.5V
13. GND	38. EXT 2	63. +2.5V
14. EXT 2	39. EXT 1	64. GND
15. EXT 1	40. +2.5V	
16. +2.5V	41. −2.5V	
17. −2.5V	42. EXT 2	
18. EXT 2	43. EXT 1	
19. EXT 1	44. GND	
20. GND	45. GND	
21. GND	46. EXT 2	
22. EXT 2	47. EXT 1	
23. EXT 1	48. +2.5V	
24. +2.5V	49. −2.5V	
25. −2.5V	50. EXT 2	

EXT 1: Digital information

EXT 2: Sine wave

The data collection procedure is listed below.

1. Allow 2-hour equipment warm-up.
2. Set and record system parameters.
3. Set transmitter deviation.
4. Tune receiver.
5. Select N.
6. Align D/A to A/D.
7. Measure true root mean square (RMS) of pseudo-random sequence at decommutator output for high SNR.
8. Set total comparison count on ratio counter.
9. Set attenuators such that errors are just detectable.
10. Record bit error, channel error, and word error probabilities.
11. Record attenuator settings.
12. Record decommutator output RMS noise of a grounded channel.
13. Increase attenuation by 1 dB of the RF signal.
14. Repeat 9 through 13 until decommutator loses synchronization.

15. Repeat 5 through 14 for N = 1 through 4 for mode I only.
16. Calibrate IF SNR to decommutator output noise.
17. Calibrate RF power to attenuator settings.
18. Repeat 2 through 17 to modes II and III for N = 3 only.

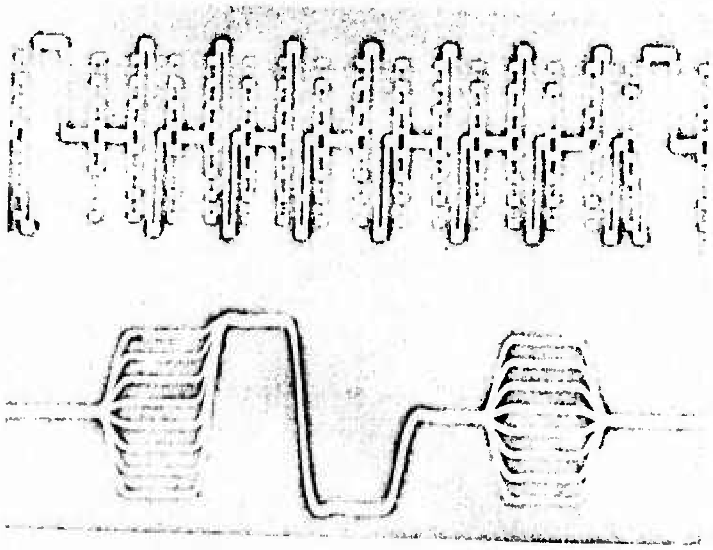

Figure 2. PAM Frame and Digital Data Channel Modulation.

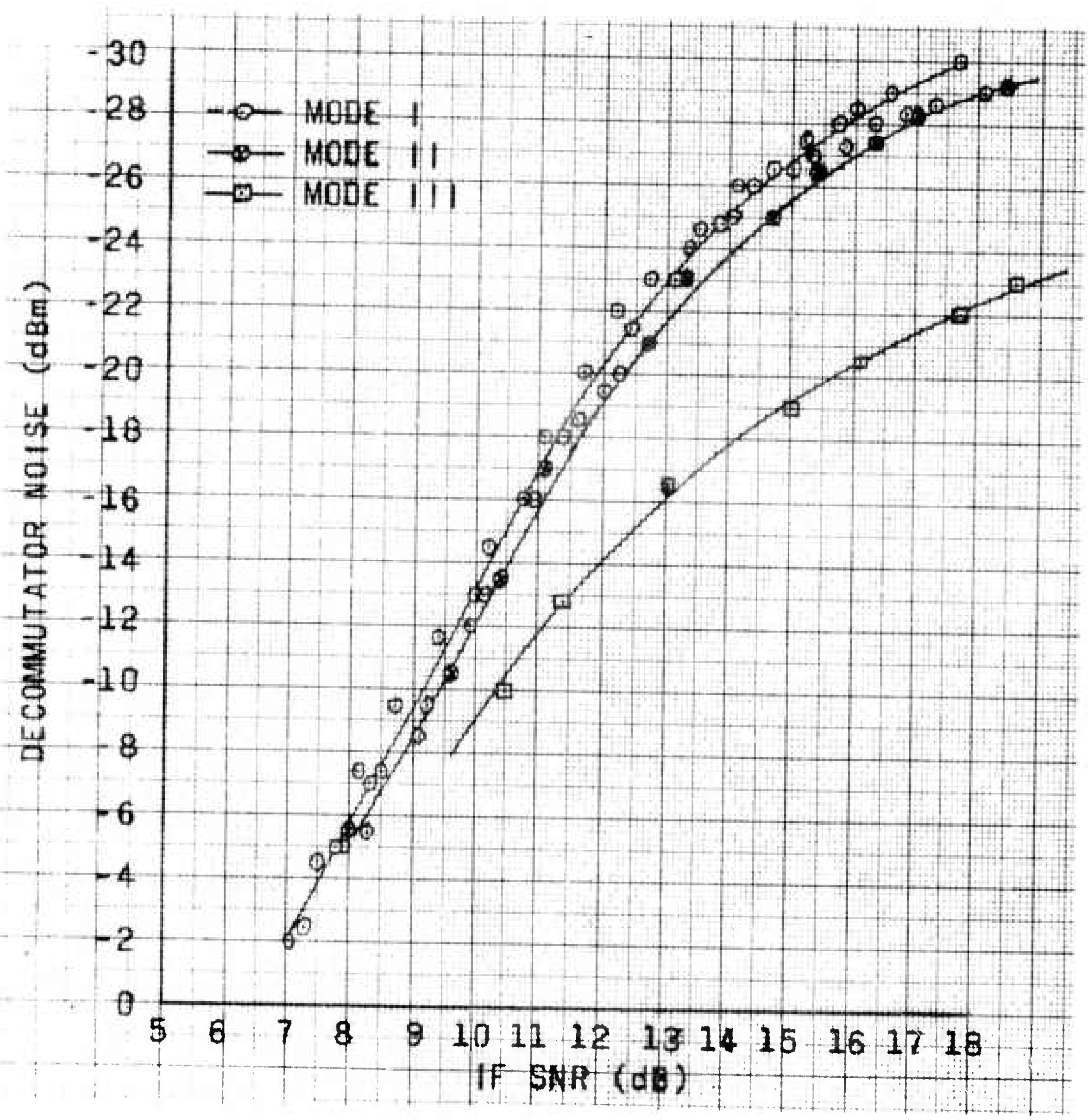

Figure 3. Relation Between Decommutator Noise and IF SNR.

For control of nonlinearities, the D/A output voltage levels were aligned through the system to approximately halfway between the A/D decision points. The nonlinearity at the decommutator output was held to approximately ±1/2 percent of full scale.

The IF SNR and RF power were not directy measurable; therefore, all data were taken with respect to the true root mean square (RMS) of the decommutator output noise on a mid-scale (bandcenter) channel. The IF SNR was then calibrated to the decommutator noise and the RF power was calibrated to the attenuator settings. The decommutator SNR in dB was calculated by subtracting the decommutator noise measurement in dBm from the RMS of the pseudo-random voltage sequence in dBm. Figure 3 shows the relation between decommutator RMS output noise and IF SNR. The RMS values of the pseudo-random voltage sequences for 10 volts full scale are: 10.8 dBm for N = 4, 3, and 2; and 9.8 dBm for N = 1. As RF power level drops below -95 dBm for modes I and II, and below -92 dBm for mode III, subtractive noise begins to seriously affect the decommutator noise measurements such that the measurement is no longer indicative of the actual noise. Decommuator noise measurements made below these levels serve only as reference points for the experimental data.

The calibration of IF SNR to decommutator noise was accomplished by measuring decommutator noise, IF noise, and IF signal plus noise for various attenuator settings. IF SNR was calculated from the following expression:

$$\text{IF SNR} = 20 \log \sqrt{[\overline{(S+N_o)^2} - \overline{N_o^2}] / \overline{N_o^2}}$$

$\overline{(S+N_o)^2}$ is the mean square value of the signal plus noise in the receiver IF. $\overline{N_o^2}$ is the mean square value of the receiver IF noise. $\overline{(S+N_o)^2}$ is equal to the expected value of $(S+N_o)^2$.

$$\begin{aligned}\overline{(S+N_o)^2} &= E[(S+N_o)^2] \\ &= E(S^2) + E(2SN_o) + E(N_o^2) \\ &= \overline{S^2} + \overline{N_o^2}\end{aligned}$$

because S and N_o are uncorrelated, independent, random, mean-value-zero signals, $E(2SN_o)$ is zero.*

High RF power was measured at the receiver input for several low attenuator settings for each of the system operating modes. The relation between attenuator settings and RF power was noted so that low RF power levels could be determined from the attenuator settings. Low RF power levels could not be measured with the power meter used in the experiment. The recording of decommutator noise and attenuator settings during the course of the experiment resulted in the decommutator noise versus RF power curves in figure 4. The increased baseband bandwidth of mode III relative to mode I resulted in higher decommutator noise for the same IF SNR and RF power level. The decreased IF filter bandwidth of mode II relative to mode I resulted in an approximate 2 dB improvement in RF power for the same decommutator noise level.

STATISTICAL SIGNIFICANCE OF ERROR PROBABILITY MEASUREMENTS

In an experiment dealing with measured error probabilities, it is important to know the statistical accuracy of the measurements. For example, how much confidence is there in measuring an error probability to within 1 percent of its true value? In this study, the confidence depends upon the number of comparisons, n, made in the error detector for the measurement. For example, if an attempt to measure a bit error probability on the order of 10^{-6} is accomplished by making only 10^4 bit comparisons, then the bit error probability (BEP) will be measured as 10^{-4} or more if errors occur, or zero if no errors occur. Obviously, neither measurement is correct. This section will show how many comparisons, n, are necessary to ensure an accurate measurement for a given statistical confidence level.

It will be assumed that the bit, channel, and word error probabilities are constant for any given SNR or RF power level. In reality, the probabilities vary with time, temperature, and receiver tuning drift. Any bit, channel, or word appearing at the system output has been subjected to a random process and will either be in error (failure) with probability p or not in error (success) with probability q = (1-p). Let each bit, channel, or word compared for error be called a statistical sample. Since theoretically there are an infinite number of statistical samples to which this process can be applied, then the distribution of successes and failures is binomial if

* P. G. Hoel. Introduction to Mathematical Statistics. Fourth ed.; New York: Wiley and Sons, Inc., 1971, pp. 119 and 150.

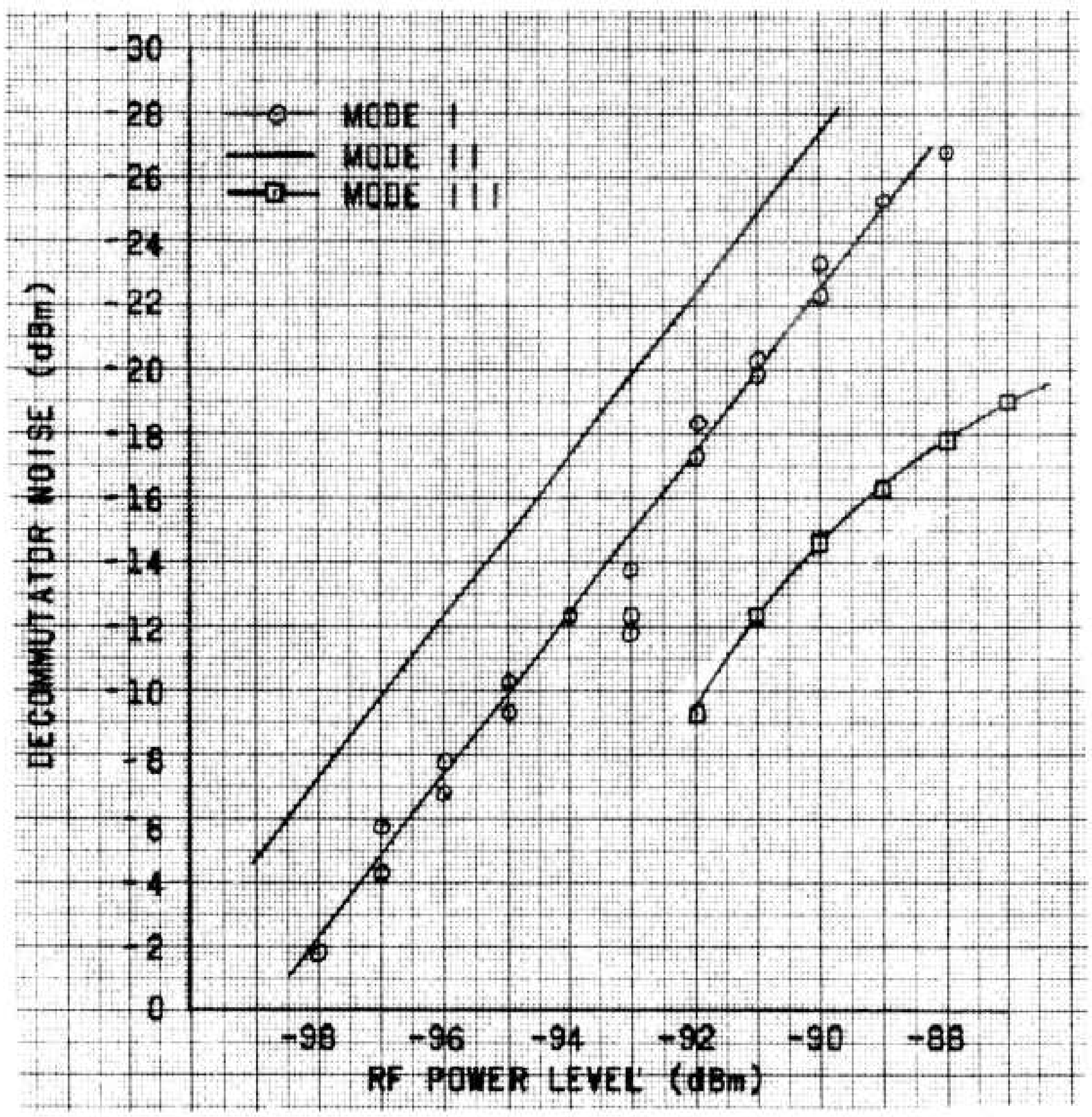

Figure 4. Relation Between Decommutator Noise and RF Power Level.

the statistical samples are independent. Each bit sampled will be independent of all other sampled bits and each word sampled will be independent of all other words if the PAM channels are independent. Channel independence is affected by noise, crosstalk, and decommutator calibration and synchronization. When the decommutator loses synchronization, the channel errors are not independent because the PAM data appears on the wrong channels. Channel crosstalk introduces a slight dependence on the channel preceding the channel of interest. Integrated calibration channel noise in the decommutator biases the calibration channel's voltage levels. Calibrator amplification of the biased calibration channels back to calibrated levels also amplifies the data channels and results in digital channel error dependenee upon noise in the calibration channels and upon calibrator operation.

To determine if PAM channels are independent of each other, the data in table 3 was collected at approximately 8 dB IF SNR for $N = 3$ in mode I operation. Applying the chi-squared goodness of fit test to the data resulted in an acceptable fit of the data to a hypothesized binomial distribution of channel errors. If the channel errors are approximately binomially distributed, then the channels are relatively independent. Channel independence seems to hold at low SNR levels as long as the decommutator remains in synchronization.

The normal approximation to the binomial distribution can be used if $np > 5$ for $p \leq 1/2$ or if $n(1-p) > 5$ for $p > 1/2$.* Let Y be a random variable representing the number of bit errors, channel errors, or word errors in a statistical sample of size n. The random variable Y/n may be treated as normally distributed with mean p and varianee $p(1-p)/n$. Y/n represents the measured bit, ehannel, or word error probability and p is the actual error probability. If 95 percent confidence (the 2σ point on a normal curve) is desired in measuring the error probabilities within E percent of their respective means p; then the required number of comparisons can be found as follows:

$$2\sigma = \text{E percent of p}$$

$$2\sqrt{p(1-p)/n} = \text{E percent of p}$$

$$n = 4p\,(1-p)/(\text{E pereent of p})^2$$

The number of comparisons (statistical samples), n, is practically limited by the time required to make the n comparisons. If the time is too long, p will not be constant but will vary with system drift. However, if the time is too short, not enough comparisons can be made to ensure an accurate measurement. Table 4 was compiled to indicate the number of comparisons required and the accuracy of measured error probabilities, p, with 95 percent confidence.

The table indicates that time seriously limits the accuracy of the measurement. The smaller the error probability, the larger the number of comparisons needed and the longer the time required to make the measurement within a specified accuracy. In general, more comparisons were made for each measurement than indicated in the table and the time required was generally held to 5 minutes or less except for low error probability measurements. The key point to be noted is that the accuracy of the measurements decrease as the error probability decreases.

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* See Hoel's Introduction to Mathematical Statistics.

Table 3. Channel Independence Data

Y*	Frequency of Y	P[Y=y]**	Expected Frequency*** of Y
0	10	0.113	7
1	20	0.275	18
2	14	0.302	20
3	17	0.196	13
4	4	0.084	6
5	0	0.0245	2
6	1	0	0
7	0	0	0
8	0	0	0
9	0	0	0
10	0	0	0
	Total 66		Total 66

*Y random variable of the number of channel errors out of 10 channels; y is a value of Y.

$$**P(Y = y) = \frac{10!}{y!\,(10-y)!}\ \hat{p}\,(1-\hat{p})^{10-y}$$

where $\hat{p} = \dfrac{[10(0) + 20(1) + 14(2) \dots 0(10)]}{66(10)}$

$= 0.196$ (the average error probability from the table)

***Expected frequency $= 66\ P(Y = y)$

Chi squared (χ^2) test for hypothesis that channel error data is binomially distributed:

Degrees of freedom $= 6 - 1 - 1 = 4$

$$\chi^2 = \frac{(10-7)^2}{8} + \frac{(20-18)^2}{18} + \frac{(20-14)^2}{20} + \frac{(17-13)^2}{13} + \frac{(6-4)^2}{6} + \frac{(2-0)^2}{2}$$

$$= 7.04$$

At the 5 percent significance level, the chi-squared tables with 4 degrees of freedom give $\chi_o^2 = 9.5$; since $\chi^2 < \chi_o^2$, then the binomial distribution fits the distribution of the channel error data.

Table 4. Accuracy of Error Probability Measurements

P	E %	n	Time Required for n Comparisons, Mode I, N = 4 (12 digital channels per frame): Bit and Channel Measurements	Word Measurements
		(Thousands of Samples)	(Minutes)	(Minutes)
0.9	1	4.5	0.016	0.067
0.5	1	40	0.150	0.567
0.1	5	14.4	0.050	0.200
0.05	5	30.4	0.117	0.433
0.01	5	158.4	0.567	2.260
5×10^{-3}	5	318	1.133	4.520
10^{-3}	10	400	1.417	5.700
5×10^{-4}	15	355	1.267	5.050
10^{-4}	30	444	1.583	6.320
5×10^{-4}	40	500	1.78	7.100
10^{-5}	100	400	1.417	5.700
5×10^{-6}	120	555	1.98	7.900
10^{-6} *	200	680	2.42	9.700

* At 90 percent confidence level.

EXPERIMENTAL RESULTS

The experimental error probabilities are presented in figures 5 through 10. The graphs illustrate how word error probability (WEP), channel error probability (CEP), and bit error probability (BEP) vary as the IF SNR is changed. They are an indication of the degradation of digital data quality with degradation in signal to noise. The error probabilities can be related back to decommutator SNR and RF power level through figures 3 and 4.

The error probability curves of this report were obtained from a controlled experiment. The static biases due to nonlinearities were held to approximately $\pm 1/2$ percent or less of full scale of the decommutator output voltage. The control was applied by aligning the D/A to the A/D through the system. This control need not restrict the application of the experimental results to practical (uncontrolled) PAM systems. The static biases of nonlinearities in practical systems are typically around ± 1 percent of full-scale decommutator output voltage. This increase in static biases from the controlled $\pm 1/2$ percent merely shifts the decommutator output noise distribution. Errors due to one tail of the distribution curve exceeding a digital bit's decision level are reduced and errors due to the other tail are increased by the shift. Whether or not there is a net increase or decrease in errors depends on how the distribution is skewed and the direction of the shift. From observations during the course of the experiment, an increase in the static biases to around ± 1 percent resulted in no more than a 0.2 dB IF SNR shift of the error probability curves to the right. If the difference in error probabilities is considered significant for a system with ± 1 percent nonlinearities, then a 0.2 dB shift in the curves may be applied as a correction factor.

Observations indicated that subtractive noise was becoming significant around -17 dBm (11 dB IF SNR) decommutator noise for modes I and II. However, no significant change in the error probability curves appeared. Some of the errors that should have been caused by subtractive noise were suppressed by decommutator calibrator operation. Comparison of subtractive-noise-suppressed calibration channels with internal calibration levels resulted in amplification of the PAM signal. The amplification offset the suppression and resulted in fewer digital errors than expected. Subtractive noise never really became a factor in mode III operation. Most of the data quality degradation was caused by the wider baseband filter before subtractive noise appeared. The decommutator began dropping out of synchronization around 8 dB IF SNR in modes I and II. Experimental data below 8 dB are not reliable.

Figure 5 shows the improvement in the word error probability as N is decreased and the improvement in word error probability obtained for a specified N by placing bits of the word in only the most significant bit (MSB) locations of the channel as opposed to placement in all the channel's bit locations. To distinguish between WEP curves, the notation WEP_{XN} was used. X is the number of the most significant channel bit locations used out of the N bits of the channel. This notation was also used to distinguish between CEP curves. The bits of the word were always placed in the X most significant bit positions of the channel. The other N-X bit locations of the channel were filled with pseudo-randomly selected bits. The pairs of curves for each value of N in figure 5 represent maximum and minimum WEP possible for a given N.

All of the WEP_{1N} curves were calculated from their respective MSB error probability curves of figure 8. Measurement of WEP_{1N} was not feasible due to the length of measurement time. The calculation was based upon the MSB errors being binomially distributed for a fixed IF SNR. This requires that the MSB bit error probability be constant for a given SNR and that the MSB's of all the channels be independent of each other. The previous section established independence and the MSB bit error probability was held fairly constant by measuring it over a

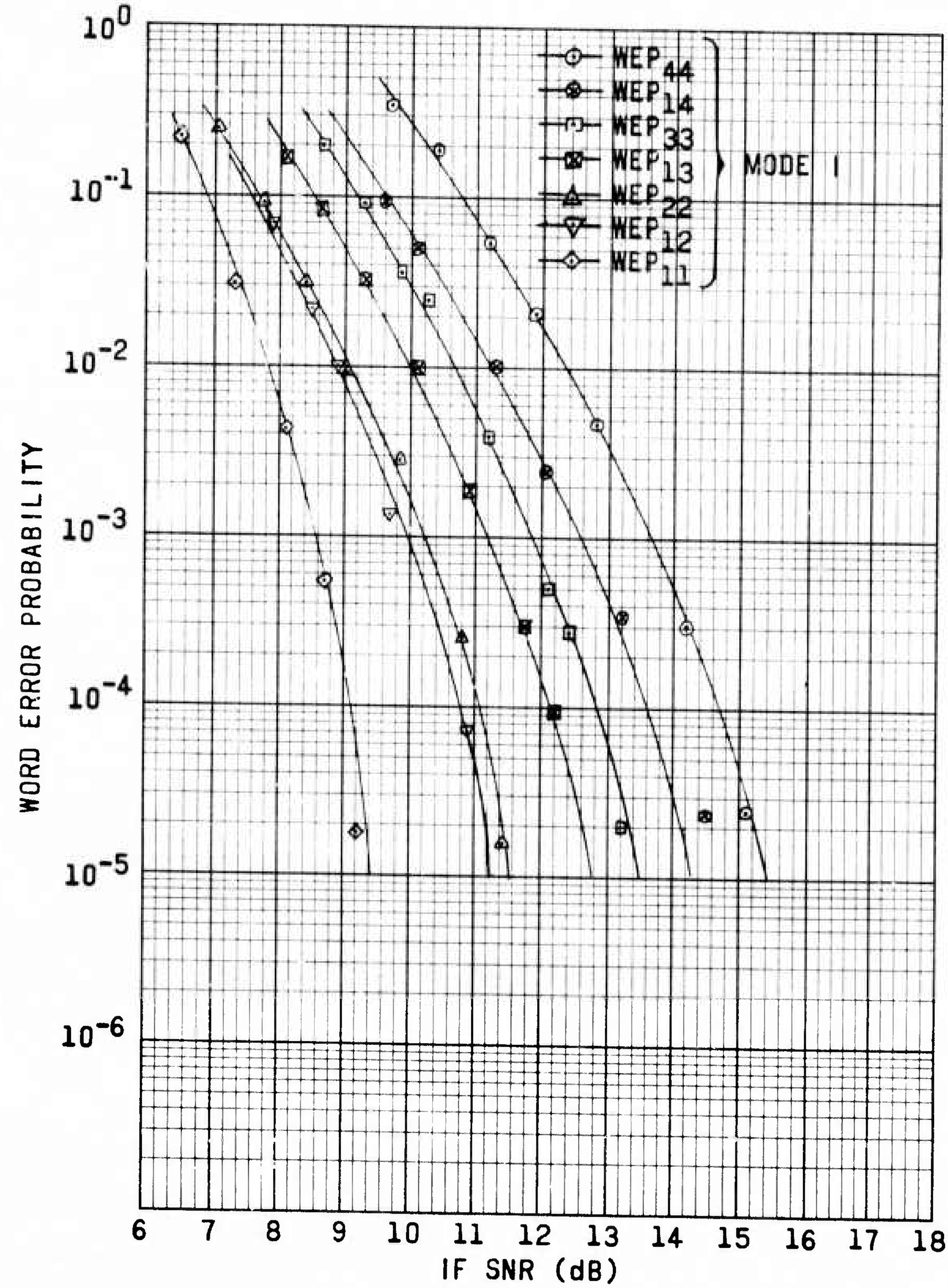

Figure 5. Variation of WEP With IF SNR for N Bits/Channel.

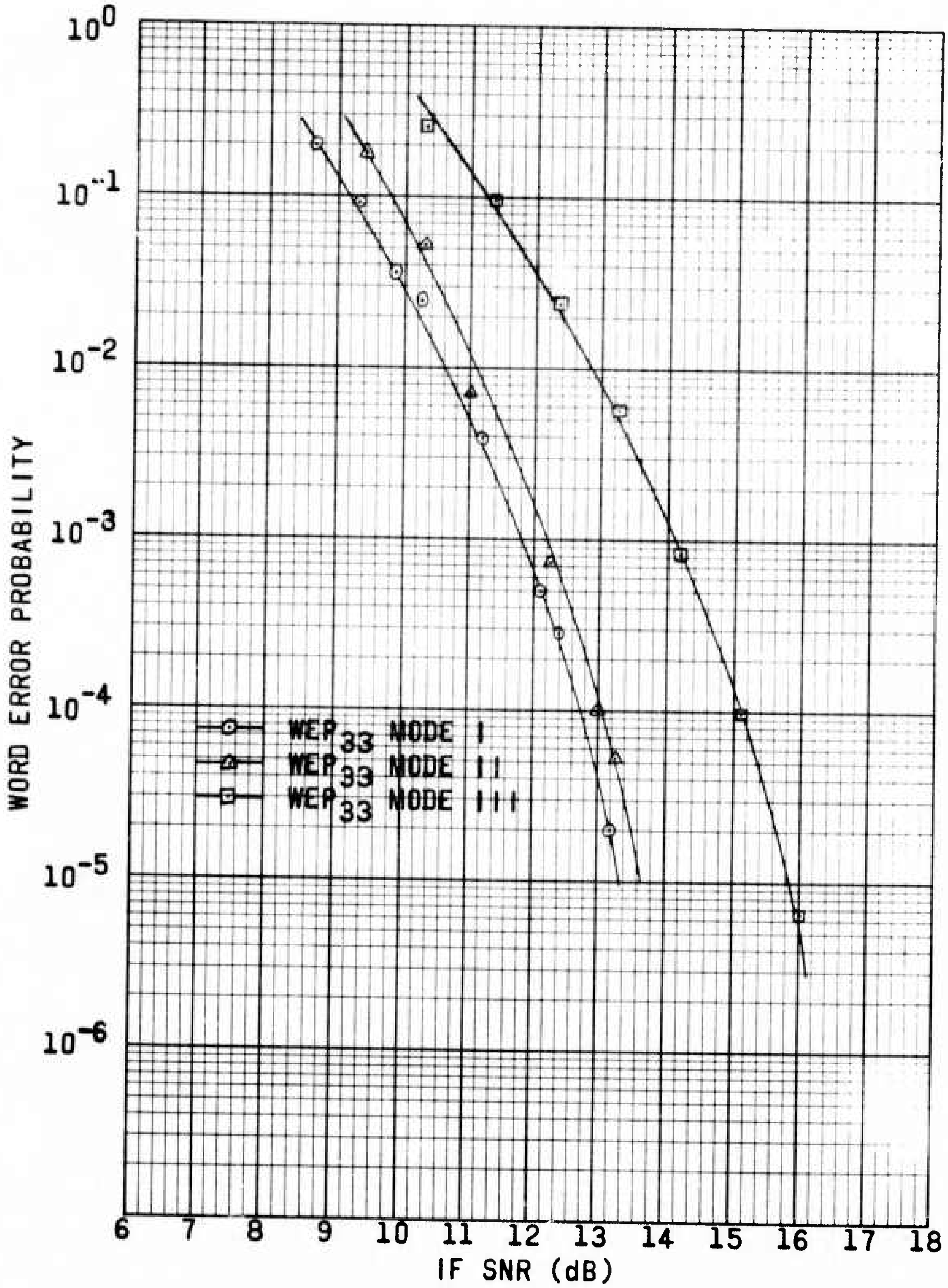

Figure 6. Variation in WEP With IF SNR for a Change in IF Bandwidth or Sample Rate.

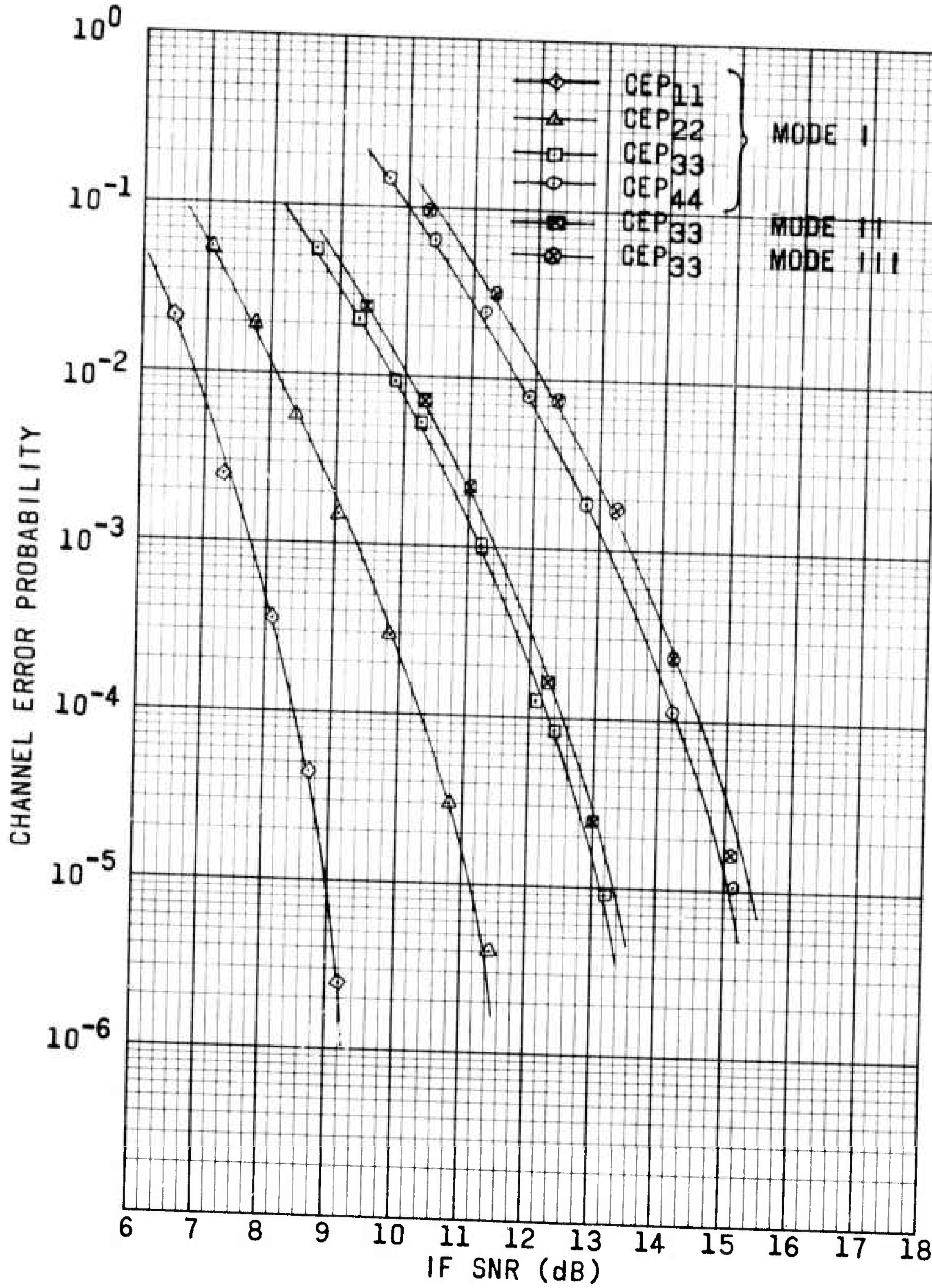

Figure 7. Variation in CEP With IF SNR.

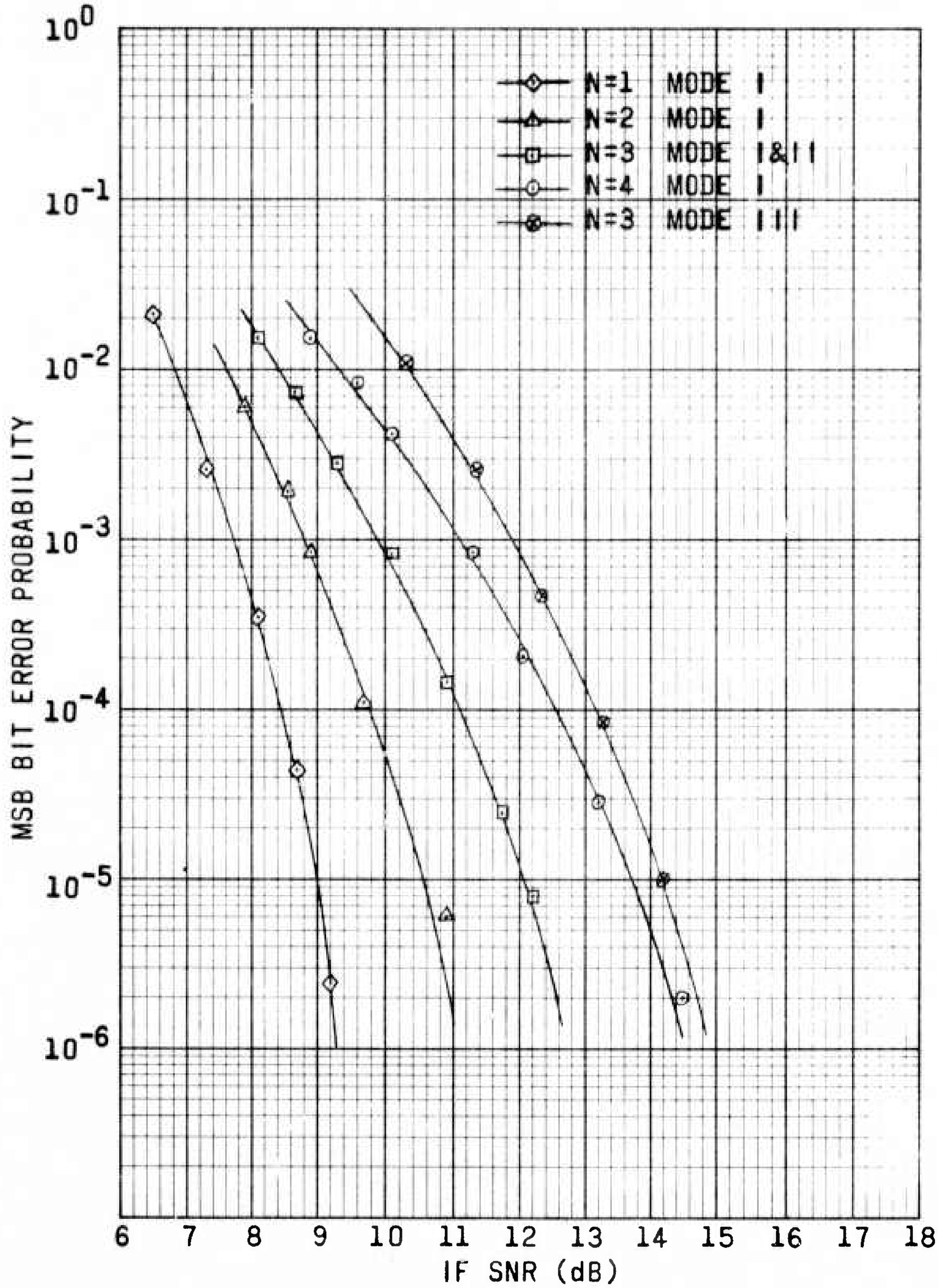

Figure 8. Variation in MSB BEP With IF SNR.

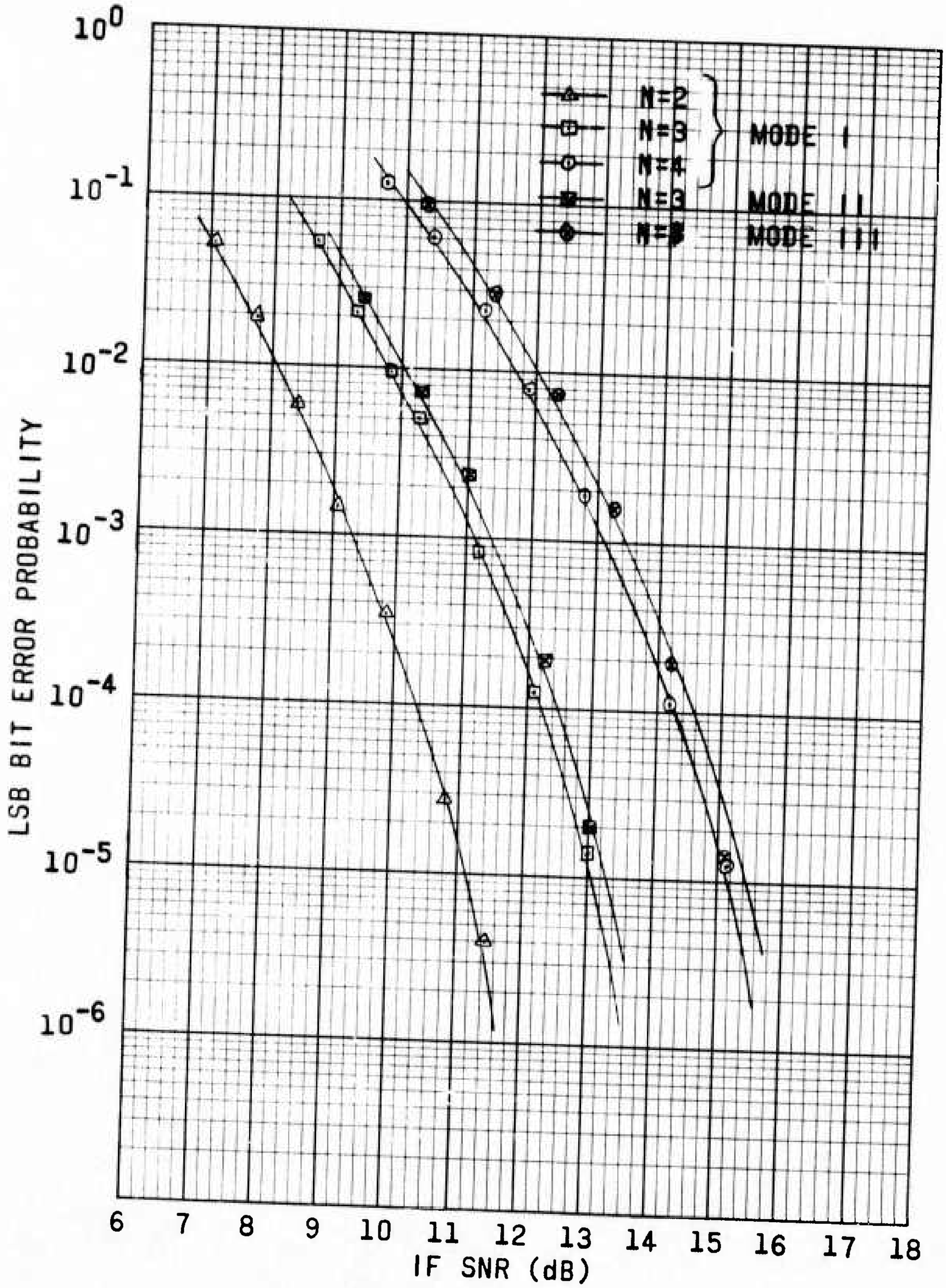

Figure 9. Variation in LSB BEP With IF SNR.

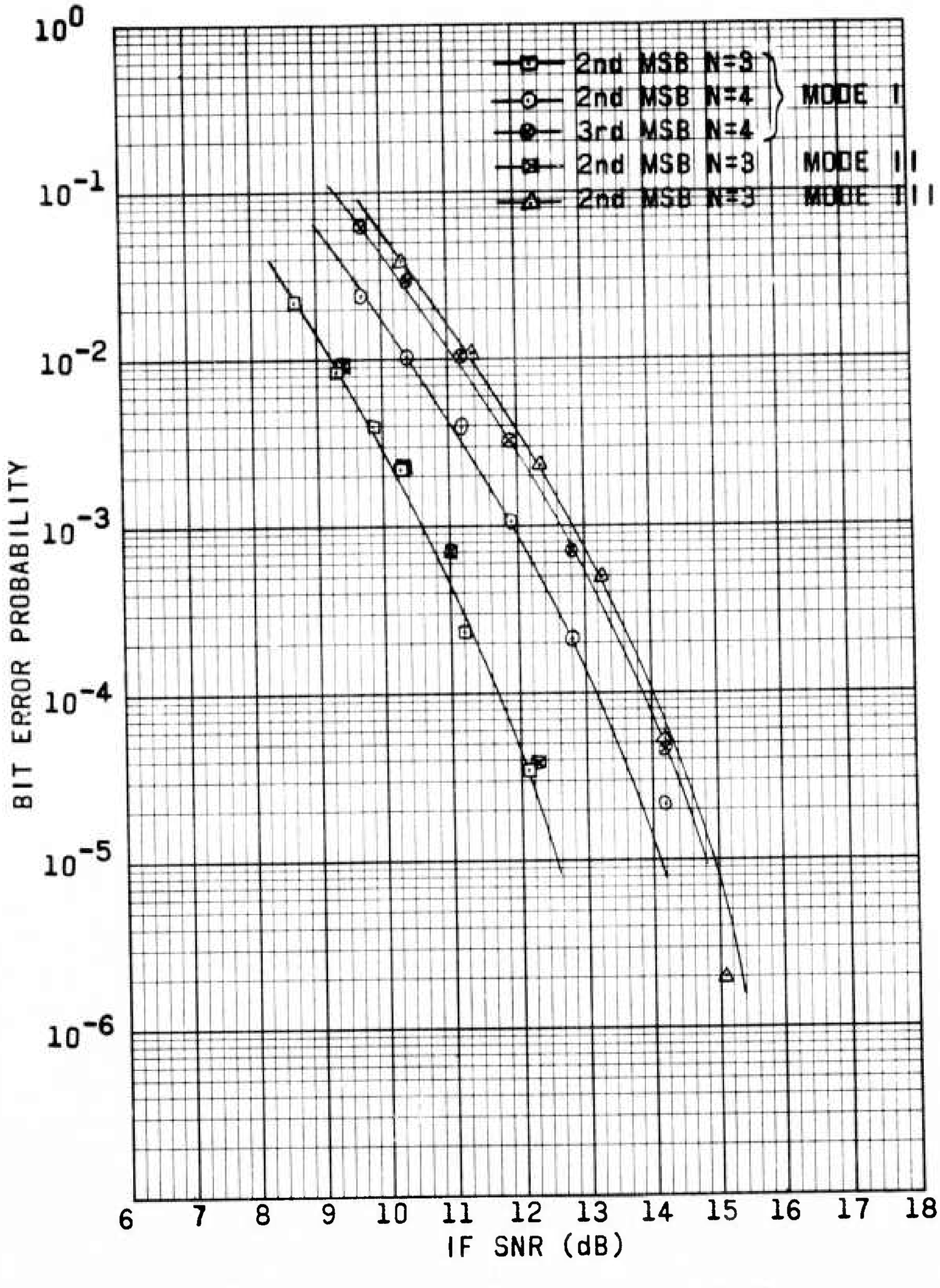

Figure 10. Variation of BEP of Middle Bits for N = 3 and 4 With IF SNR.

relatively short period. The probability of a word error for $X = 1$ was calculated from the formula shown below for a 12-bit word:

$$WEP_{1N} = 1-(1-p_{msb})^{12}$$

where p_{msb} is the MSB error probability. This formula and subsequent ones are derived on the basis of the errors being binomially distributed for a given SNR or RF power level.* The probability of a word error is one minus the probability of no word error. The probability of no word error, in this case, is the probability of no MSB errors in the 12 channels carrying the word bits. Because the MSB locations of different channels are independent and their errors binomially distributed, the probability of no MSB errors is the quantity one minus the probability of a MSB error to the 12th power. Similar formulas for WEP can be derived in the same manner.

For $N = 4$, there is a 1 dB improvement in WEP_{14} with respect to WEP_{44}. For $N = 3$, the improvement of WEP_{13} with respect to WEP_{33} has decreased to about 1/2 dB and; for $N = 2$, the improvement has decreased to roughly 0.2 dB. The improvement between WEP_{1N} and WEP_{NN} results from the analog weighting of the bits in the D/A and A/D conversion process. For example, at 12 dB IF SNR, for $N = 4$, no matter what four word bits are placed on a channel, a word error will occur any time the decommutator noise exceeds $\pm 1/2$ the LSB voltage of the channel. On the other hand, for word bits placed only in the MSB locations of the channels, a word error will occur only when the decommutator noise exceeds the MSB decision point of the A/D.

The CEP curves are shown in figure 7 for $X = N$. For the conditions of mode I, there is a 2-dB improvement between curves of CEP_{NN}. This improvement relates directly to the increasing size of the LSB voltage as N decreases. Comparison of the respective LSB bit error curves in figure 9 with the CEP curves in figure 7 indicate that CEP is very nearly the same as the LSB BEP. This is to be expected. The only time a channel error occurs in the absence of an LSB error is when the decommutator noise jumps two (or a multiple of two) LSB decision levels. Therefore, the CEP and LSB BEP curves should be identical at their lower ends and only slightly different at their upper ends. This result, as shown later, is important in determining WEP_{XN} when $1 < X < N$.

The MSB bit error curves are shown in figure 8. They were used to calculate the WEP_{1N} curves of figure 5. The MSB decision point never changes, but as N decreases, the closest modulation level (pseudo-random voltage level) on either side of it gets further away. Consequently, the 1.5 dB improvement for decreasing N is because a lower SNR is required to cause decommutation noise to span the increased distance.

Figure 10 presents the BEP's of the second MSB of $N = 3$ and the second and third MSB's of $N = 4$. Because the A/D decision levels of these bits are further apart than their respective LSB decision levels but not as far apart as their respective MSB decision levels, these curves are equally spaced between their respective LSB and MSB BEP curves.

The effect of increasing the sample rate and decreasing the IF filter and the demodulator bandwidth on WEP is shown in figure 6 for $N = X = 3$. The increased sample rate resulted in a 2 dB degradation in WEP. A degradation is to be expected because of the wider baseband bandwidth. The decreased IF bandwidth resulted in a slight degradation of WEP for a given IF SNR because of phase nonlinearities and signal power loss in the narrower IF bandwidth. However, the reduced bandwidth actually results in a 2 dB RF power improvement as noted in figure 4. For a given RF power level, the narrower bandwidth decreases the noise power and increases the SNR.

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* See Hoel's Introduction to Mathematical Statistics.

PAM WITH DIGITAL DATA

The quantity of digital data that can be transmitted per PAM channel is limited by the system's noise floor and static biases. The SNR at the decommutator output for the noise floor level was observed in the laboratory to be approximately 46 dB with respect to sine wave modulation. Assuming an approximately gaussian distribution of decommutator noise at this RF level, the standard deviation of the noise will be 36 mV for 10 V full scale. If nonlinearities at the decommutator output for some modulation levels are around ± 1 percent, then for 10 V full scale, a 100 mV offset of the noise floor distribution from the correct amplitude of the sample occurs. For N = 5, the LSB decision levels are 156 mV $\left(\frac{1}{2}\text{ LSB} = \frac{1}{2} \times \frac{1}{2^5} \times 10\text{V}\right)$ from the correct amplitude of the digital data at the decommutator output. Consequently, one LSB decision level falls within the 2σ point of the shifted noise distribution and significant errors will result. For N = 4, the LSB decision levels are 312 mV $\left(\frac{1}{2}\text{ LSB} = \frac{1}{2} \times \frac{1}{2^4} \times 10\text{V}\right)$ from the correct amplitude of a sample and lie outside of the 5σ point of the noise distribution. Therefore, relatively few errors will occur for 4 bits per channel at high RF levels. This re-establishes N = 4 as the previously determined maximum transmission quantity of digital bits per channel. However, it is not uncommon for higher static biases and noise floor to exist in the field such that it is possible for N = 4 to be of marginal use. Other constraints will limit N even further. As an example of the determination of N, consider the following system. Assume that a 25 kHz, 64-channel, PAM system with 1 MHz IF is required to telemeter a 12-bit word every PAM frame. Additional digital data include 20 discrete information bits, each of which must be telemetered every 2 PAM frames. Furthermore, only 12 PAM channels can be allocated for the digital data and the maximum word error probability must be no more than 10^{-4} and the maximum discrete error probability must be no more than 10^{-2} at 30 dB decommutator SNR. The 30 dB SNR is with respect to full-scale sine wave channel modulation. Can all the data be transmitted? What should N be? How should the word bits and discrete bits be arranged so that channel allocation and error constraints are met?

First, the 30-dB SNR must be converted to an equivalent decommutator SNR for the pseudo-random voltage sequences used in the experiment. For $N \geq 2$, the RMS of each sequence is approximately that of a triangle. The RMS of a triangle is 11 dBm for 10-volt, full-scale decommutator output. SNR with respect to full scale (10 volts) triangle modulation is 1.7 dB less than SNR measured with full-scale sine wave modulation. Therefore, the decommutator SNR requirement with respect to the sequences for $N \geq 2$ is 28.3 dB (11.1 dB IF SNR from figure 3). For N = 1, the pseudo-random voltage sequence has degenerated to two levels and has the RMS characteristics of a square wave. The decommutator SNR with respect to a full-scale square wave is 3 dB better than a full-scale sine wave. However, for N = 1, the voltage range of the two modulation levels was only 1/2 of full scale, therefore, the decommutator SNR for N = 1 is 3 dB less than the SNR for full-scale sine wave modulation. The equivalent decommutator SNR for N = 1 is 27 dB (11.1 dB IF SNR).

Figure 5 indicates that the only way to meet the word error constraint at 11.1 dB IF SNR is with N = 2. N = 1 will not meet the channel constraint. For the discrete data, figure 9 shows N = 3 or 2 will meet the discrete error constraint.

There are three possible methods to handle the digital data using these results. Two of the methods involve N = 2 and the other involves a combination of N = 2 and 3. The first method for

$N = 2$ is to break the word up into 6 two-bit segments leaving 6 channels for the discrete bits. Since the 20 discrete bits are required only once every two frames, then the subcommutation of 5 channels at 2 bits/channel results in 20 bits every 2 frames and one channel is saved for other applications. WEP_{22} for this method is, from figure 5, 8×10^{-5}. The discrete error probability depends on the location of the bit in the channel; for the MSB location, the BEP is, from figure 8, 10^{-6}; for the LSB location, the BEP is, from figure 9, 9×10^{-6}. The difference in BEP of the MSB and LSB for $N = 2$ suggests that the six most significant bits of the word should occupy the MSB positions of the 6-word channels. This assumes that the MSB's of the word are of more value than the LSB's. Such is not the case if the word represents a code in which all bits are of equal importance.

If the bits of the word are all of equal importance, then the second method for $N = 2$ may be preferred over the other methods. This method places all 12 bits of the word in the MSB locations of the 12 channels and results in equal error probability for all word bits. The discrete bits fill the 12 LSB channel positions, 8 of whose positions must be subcommutated. WEP_{12} for this method is, from figure 5, 3×10^{-5}. The discrete error probability is, from figure 9, 9×10^{-6}. This method uses all 12 channels, decreases the word error probability, and increases the discrete error probability.

If it is desired to allocate as few channels as possible to the digital data, then method one above is a possibility. However, a third method using 2 and 3 bits per channel results in a savings of an additional channel. The word is transmitted, as in method one, on 6 two-bit channels. The discretes are transmitted on 4 three-bit channels. The 8 bit locations comprising the most significant bits and the second most significant bits of the four discrete channels are subcommutated. The number of channels used is 10. WEP_{22} is once again 8×10^{-5}. The discrete error probabilities depend on the bit location in the channel. For the MSB location, the BEP is, from figure 8, 10^{-4}; for the second MSB location, the BEP is, from figure 10, 4×10^{-4}; for the LSB location, the BEP is, from figure 9, 10^{-3}.

This example illustrates some of the key considerations involved in transmitting digital data. For a given PAM telemetry system, the quantity of digital data is known. The quality requirements of this data must be determined before it is possible to select N. In some cases N will be completely specified by the quantity of data and quality constraints. In cases where it is not, as in the above example, N may be chosen with one of two objectives in mind: i.e. maximization of data quality of either WEP or BEP of discretes, or both; or minimization of digital data channels. Maximization of data quality will generally result in utilization of all available digital channels, whereas minimization of digital channels reduces the data quality.

The weighting of the bits in the D/A conversion results in some important considerations. The difference in BEP of the bit locations of a channel is due to this weighting. The more significant the bit of the channel, the lower the error probability for a given SNR. Consequently, the most important bits of the digital data should be located in the MSB positions of the data channels. For a word representing a value, these positions are filled with the MSB's of the word. If the word is a code where all bits are of equal importance, then the bits must be placed in the same bit location of each channel or they will have different weights and BEP's. Since WEP constraints will usually be the most difficult to meet, the most significant bits of a channel should be reserved for word bits. The remaining bit locations are filled with discretes. The most important discretes are placed in the most significant position of the remaining channel bits.

If needed, WEP curves not shown, such as WEP_{23}, can be generated from the other experimental curves. As noted in the last section, channel error probability and LSB bit error probability are, for all practical purposes, the same. This suggests that CEP_{23} is the same as the BEP of the second most significant bit for $N = 3$ of figure 10. In general then, WEP_{XN} for

$1 < X < N$ can be calculated from:

$$WEP_{XN} = 1-(1-CEP_{XN})^{M/X}$$

Where N is bits per channel; X is the number of bit locations used for word bits out of the N per channel; M is the bits per word = 12; and CEP_{XN} is the probability of one or more of the X most significant bits in a channel being in error. In these cases, CEP_{XN} is approximated by the BEP of the X*th* bit and M/X must be an integer.

Because of the nature of the weighting in the D/A and A/D conversions, WEP_{23} should fall about halfway between WEP_{13} and WEP_{33}. WEP_{24} and WEP_{34} should fall from WEP_{14} about and 1/3 and 2/3 respectively, the distance between WEP_{14} and WEP_{44}.

The experimental WEP curves were generated for M = 12 bits per word. For longer or shorter words, the WEP can be determined from:

$$WEP_{XN} = 1-(1-CEP_{XN})^{M/X}$$

Where $1 \leq X \leq 4$, $1 \leq N \leq 4$, and M/X is an integer. For X = N, CEP_{NN} is given in figure 7; for $X < N$, CEP_{XN} is given by the appropriate BEP curve in figures 8 and 10. The CEP and BEP curves do not change with M, they vary with N, filter bandwidths, and sample rate. If M/X is not an integer, then the above formula can be modified to:

$$WEP_{XN} = 1-(1-CEP_{ZN})(1-CEP_{XN})^{T_I(M/X)}$$

Where Z is the remaining bits of M/X and $T_I(M/X)$ is the truncation of M/X to its integral number part.

Error probability curves for N = X for modes II and III may be approximated by utilizing the relatively fixed difference between the N = X = 3 curves of modes I and II and modes I and III of figures 6 through 10. The difference between modes I and II and modes I and III results in a fixed change in noise power of the video and IF signals respectively. Consequently, the error probability curves of mode II or III for any N = X, should fall relatively the same distance and manner from their respective mode I curves as those do for N = X = 3 in figures 6 through 10.

CONCLUSIONS

1. The quality of digital data is determined by the state of the system (i.e., the amount of crosstalk, RF signal power level, incidental FM, circuit noise, and system nonlinearities) and the placement of bits in the N bit locations of a channel.

2. The quantity of digital data per PAM channel that can be transmitted on any given PAM telemetry system is determined by the worst expectation of RF signal power at the receiver, the digital quality requirements, and the quantity of analog data in the system.

3. Practical accuracy limitations of PAM systems restrict the number of digital bits that can be accurately transmitted in a single PAM channel to 4 bits for modes I and II and to 3 bits for mode III.

4. The implementation of digital data in PAM telemetry systems requires:

   a. The determination of required digital and analog data quantities.

b. The determination of the digital data quality constraints with minimum expected RF signal power at the receiver.

c. The determination of the number of channels to be allocated to the digital data based on the primary (analog and/or digital) data requirements.

d. The selection of N to meet channel allocation and digital quantity and quality requirements.

e. The determination of the channel bit packing of words and discretes to meet channel allocation and data quality requirements.

f. The design of commutation, conversion, and decommutation equipment for the packing, encoding, recovery, and reconstruction of digital words and discrete bits.

5. The feasibility of digital data transmission through a PAM communication system must be resolved on a system by system basis by determining in 4 above that all the digital data can be transmitted within quality constraints and within channel limitations.

APPENDIX A

CIRCUIT INTERCONNECTIONS AND OPERATIONS

The clock input to the pseudo-random bit generator in figure A-1 controls the pseudo-random bits input to the D/A converter. The full-scale (F.S.) ZERO switch (SW1) of figure A-1 sets all bits to the zero level when grounded. SW2 in the F.S. ONES position sets all bits to the one level when clocked in the F.S. ONES position. These two positions are used for calibration. Normal operation is with SW2 in the shift right position and SW1 in the $\pm 5V$ position. N is selected by switches S1A, S1B, S2A, and S2B. The pseudo-random voltage sequence from the D/A is conditioned by an operational amplifier before being input to the commutator. The pseudo-random bits (PN1, PN2, PN3, PN4, PN5, PN6) applied to the D/A are also applied to the error detector.

The integrator output of the decommutator is applied directly to the sample/hold (S/H) of figure A-2. The hold command for the S/H is derived from the Boolean "anding" of the channel gating, patch panel, strobe signal and the sample trigger signal, both of which are from the decommutator. A channel gating, patch panel, strobe pulse is programmed in the decommutator to occur only when a digital data channel is being decommutated. A sample trigger pulse occurs at the end of the integration period of all channels. The S/H samples the final value of the integrator. After acquisition of a hold, a start conversion pulse is applied to the A/D. An end of conversion (EOC) pulse from the A/D signals the error detector to compare the A/D output with the pseudo-random generator output.

The channel and bit sample pulses indicate the number of channel and bit error comparisons made. The pulses go to switch 10 of figure A-3. They are derived from the programmed channel gating, patch panel, strobe signal of the decommutator. The end of word and word sample pulses are derived from the programmed end of word, patch panel, strobe pulses of the decommutator. A pulse occurs only when a complete word (consisting of K channels) has been received. These pulses are also applied to switch 10 of figure A-3.

Error detector operation is determined by switches 1 through 10 of figure A-3. N is set by switches 1 through 5; with all five of them in the "a" position, N is equal to 6. With SW5 grounded, N is equal to 5; with SW4 and SW5 grounded, N is equal to 4, etc. The bit error comparisons were made by comparing each bit of the A/D with the respective bit of the pseudo-random generator in six parallel "exclusive or" gates (N7486). Channel errors were detected by paralleling the bit error comparison outputs of the "exclusive or" gates into an "or" gate such that an error pulse occured when one or more bits in the channel were in error. Switches 6, 7, and 8 selected between seven lines of a multiplexer (N74153) which consisted of a channel error line and six individual bit error lines. Word errors are detected by selecting the channel error line of the multiplexer with switches 6, 7, and 8 and placing SW9 and SW10 in their word error and word sample pulse positions. In this mode, channel errors of K digital channels, selected by the decommutator, are applied to a flip-flop (N7474). If one of the K channels is in error, the word is in error and a channel error pulse sets the flip-flop. The flip-flop is reset after K channels by the end of word pulse. The flip-flop output constitutes the word error count signal.

The clock signal to the pseudo-random generator is derived and delayed from the EOC pulse in a one-shot multivibrator (SN74123). The delay permits the error comparisons to be made before clocking the pseudo-random generator to a new digital code.

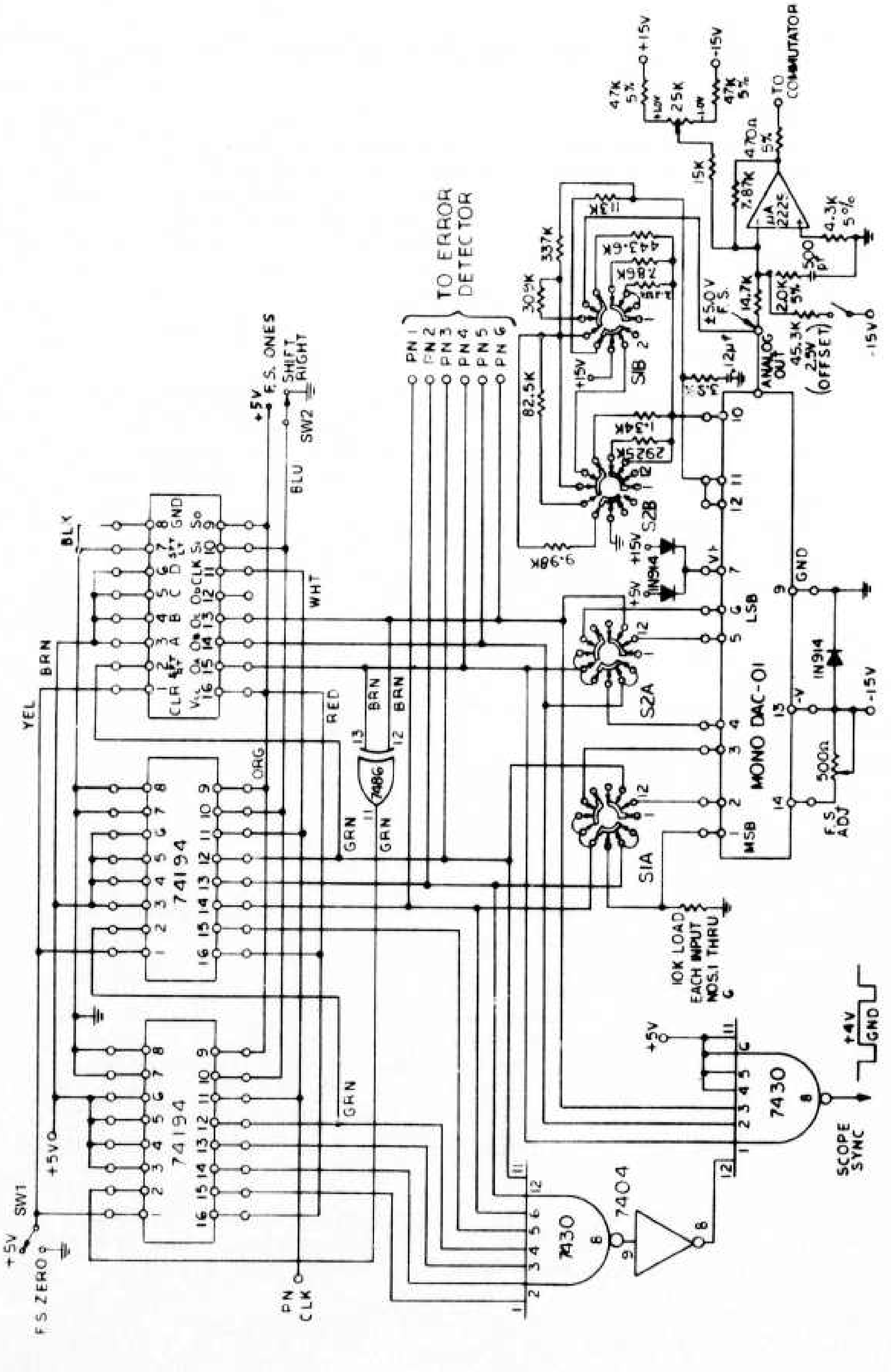

Figure A-1. Pseudo-Random Bit Generator and Digital-Analog Converter-Interface.

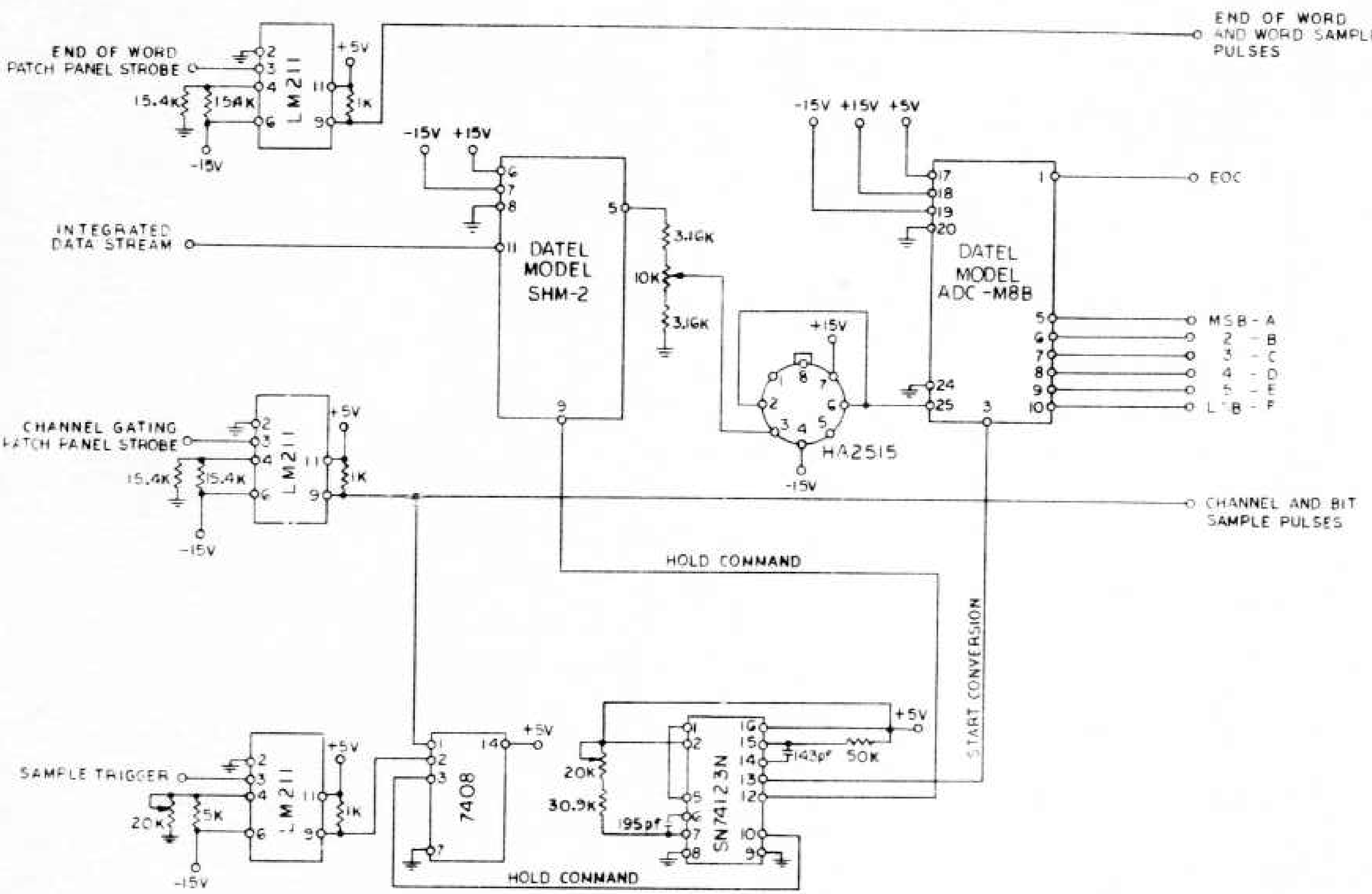

Figure A-2. Analog-to-Digital Interface Converter.

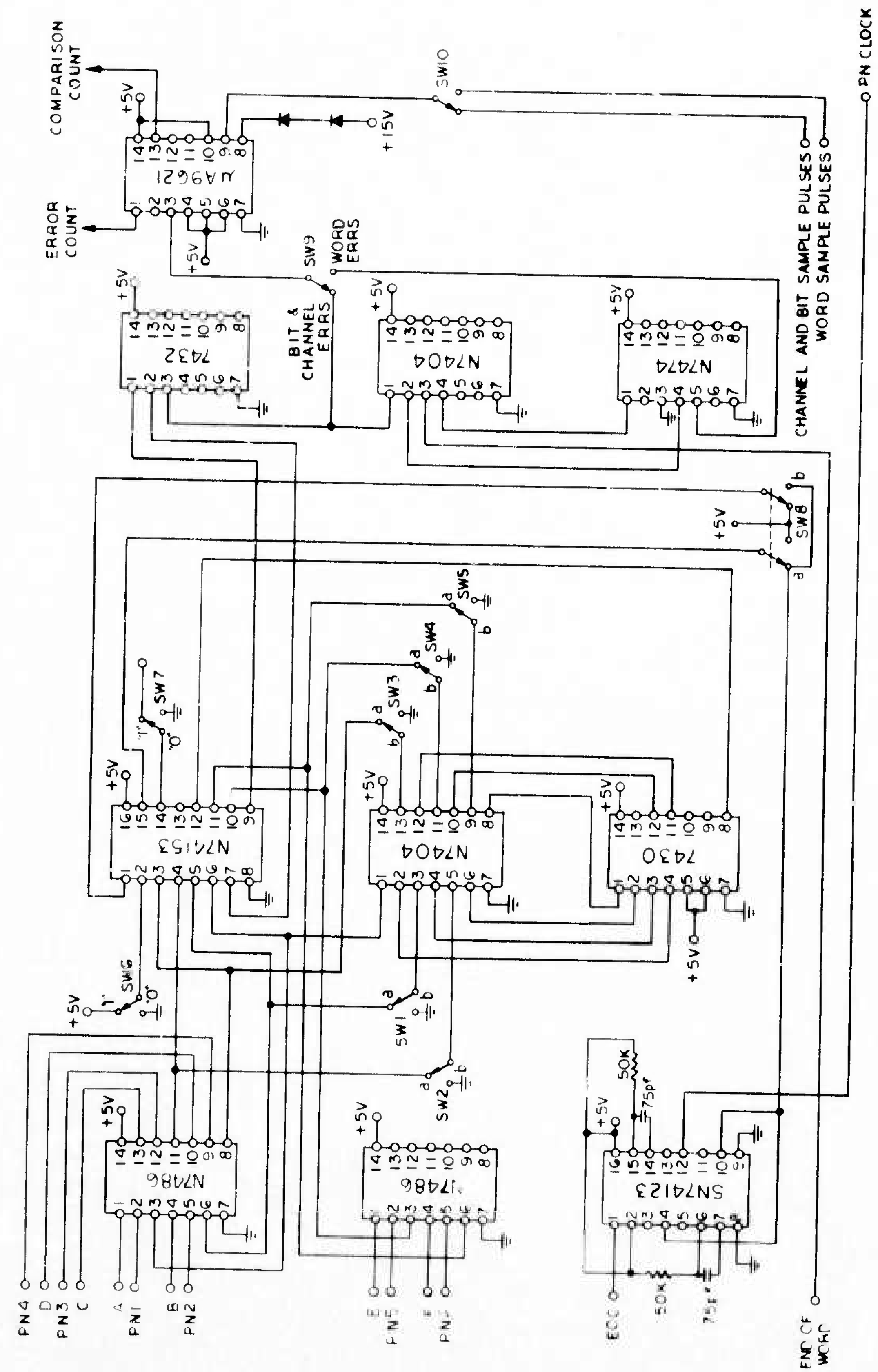

Figure A-3. Error Detector.

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