JET PROPULSION LABORATORY

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Vol I

FINAL REPORT

VIDEO DATA MODULATION STUDY

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RE-ORDER No. 64-712
Vol I

TABLE OF CONTENTS

		Page
SECTION 1	INTRODUCTION	1-1
SECTION 2	INVESTIGATION	2-1
	2.1 Organization	2-1
	2.2 Net and Gross Compression	2-1
	2.3 Fidelity	2-3
	2.4 Performance Verification	2-10
	2.4.1 Introduction	2-10
	2.4.2 Compression Verification	2-10
	2.4.3 Fidelity Verification	2-21
	2.5 Recommended Video Modulation System	2-23
SECTION 3	CONCLUSIONS AND RECOMMENDATIONS	3-1
	3.1 Conclusions	3-1
	3.2 Recommendations	3-1
APPENDIX A	COMPRESSION TECHNIQUES	A-1
	A.1 Introduction	A-1
	A.2 Previous Element Coding	A-3
	A.3 Line-to-Line Coding	A-8
	A.3.1 Introduction	A-8
	A.3.2 PEPL Coding	A-8
	A.4 Area Coding	A-17
	A.4.1 Area Scanning	A-17
	A.4.2 Previous Element Area Scan Coding (PEAS)	A-18
	A.5 Bit Interlace Run Length Compression	A-20
	A.5.1 Concept	A-20
	A.5.2 Experimental Results	A-21
	A.6 Roberts Coding	A-28
	A.7 Delta Modulation	A-33
	A.7.1 Introduction	A-33
	A.7.2 Delta System	A-35
	A.7.3 Delta Results	A-36
	A.8 Split Band	A-38
	A.8.1 General	A-38
	A.8.2 Fidelity	A-39
	A.8.3 Compression	A-41
	A.9 Linear Approximation	A-43
	A.9.1 Introduction	A-43
	A.9.2 Linear-Approximation	A-45
	A.9.3 Computer Simulation	A-47
	A.9.4 Computer Results	A-50
APPENDIX B	SYNCHRONIZATION STUDY	B-1
	B.1 Introduction	B-1
	B.2 Synchronization of Fixed Word Length Data	B-1
	B.2.1 PCM Subsystem Configuration	B-1

RE-ORDER No. 64-712
Vol I

TABLE OF CONTENTS (CONT'D.)

			Page
		B.2.2 Subsystem Design Equations - Search Mode	B-2
APPENDIX B (Cont'd.)		B.2.3 Subsystem Design Equations - Check Mode	B-5
		B.2.4 Subsystem Design Equations - Lock Mode	B-6
		B.2.5 Subsystem Performance Requirements	B-6
		B.2.6 Calculation of Subsystem Parameters	B-7
		B.2.7 Summary of Subsystem Performance	B-9
	B.3	Synchronization of Variable Word Length (PEC-Huffman Data) Systems	B-10
		B.3.1 PEC-Huffman Subsystem Configuration	B-10
		B.3.2 Subsystem Design Equations - Search Mode	B-10
		B.3.3 Subsystem Design Equations - Check Mode	B-15
		B.3.4 Subsystem Design Equations - Lock Mode	B-15
		B.3.5 Subsystem Performance Requirements	B-15
		B.3.6 Calculation of Subsystem Parameters	B-15
		B.3.7 Summary of Subsystem Performance	B-19
APPENDIX C		FIDELITY	C-1
	.1	Analytical Fidelity	C-1
		C.1.1 Concept	C-1
		C.1.2 Relation of Pictorial Factors to TV Parameters	C-1
		C.1.3 Criteria	C-4
		C.1.4 Implementation	C-12
	C.2	Measured Fidelity	C-15
		C.2.1 General	C-15
		C.2.2 Purposes of the Measured Fidelity Study	C-16
		C.2.3 Classification of Measurements	C-16
		C.2.4 Schedule of Fidelity Measurements	C-20
APPENDIX D		RECOMMENDED SYSTEM	D-1
	D.1	General	D-1
	D.2	System Design	D-1
		D.2.1 Linear Approximation Encoder	D-1
		D.2.2 Buffer Storage	D-2
		D.2.3 Linear Approximation Decoder	D-2
	D.3	Circuit Design	D-3

RE-ORDER No. 64-712 Vol I

TABLE OF CONTENTS (CONT'D.)

			Page
APPENDIX E		FIDELITY MEASUREMENT PROGRAM.	.E-1
	E. 1	Background Information	.E-1
	E. 1. 1	EDITS System Summary.	.E-1
	E. 1. 2	Video Signal-to-Noise Calibration . .	.E-3
	E. 1. 3	Basic TV Camera Resolution	.E-5
	E. 2	Objective Measurements	.E-8
	E. 2. 1	Introduction	.E-8
	E. 2. 2	Objective Test Procedures and Results.	.E-8
	E. 3	Quasi-Objective Measurements.	.E-13
	E. 3. 1	Introduction	.E-13
	E. 3. 2	Discussion of Quasi-Objective Test Procedures and Results	.E-14
	E. 4	Subjective Measurements	.E-16
	E. 4. 1	Introduction	.E-16
	E. 4. 2	Discussion of Subjective Test Procedures and Results	.E-17

LIST OF ILLUSTRATIONS

SECTION 2	Figure 2.3-1	Analytic Fidelity Concept
APPENDIX A	Figure A.3-1	PEPL Modulator
	Figure A.3-2	PL as a Function of PE for the Conditions $\bar{C}_{PEC} = \bar{C}_{PEPL}$ and $\bar{C}_{PEPL} = 2\bar{C}_{PEC}$
	Figure A.3-3	Compression for Various Scenes Processed on EDITS at 256 Resolution
	Figure A.3-4	PE and PL Probabilities for Several Scenes Processed on EDITS at 256 Resolution
	Figure A.4-1	Subscan Sweep Generation
	Figure A.4-2	Subscan Sweep Pattern
	Figure A.4-3	2^2 Subscan Pattern
	Figure A.4-4	Horizontal/Vertical Pattern
	Figure A.5.2-1	Schematized Array of Bit Plane Pictures for Visual Analysis
	Figure A.5.2-2	Division of Bit Plane Pictures Into Organized and Disorganized Categories
	Figure A.5.2-3	Raw Data - Number of Occurrences per Frame of Element Pairs 00, 01, 10, and 11
	Figure A.5.2-4	Transition Probabilities of Bit Plane Information Sources (J Level Follows I Level) I, J One or Zero
	Figure A.5.2-5	Average Compression as a Function of Truncation for Transition Probabilities of 0.99
	Figure A.5.2-6	Average Compression as a Function of Truncation for Transition Probabilities Model, Assuming Transition Probabilities of 0.99
	Figure A.5.2-7	Compression for Bit Planes of 256 Lunar Picture for Various Coding Formulae
	Figure A.5.2-8	BIRL Compression as a Function of Run-Length Truncation for Model and Actual Statistics
	Figure A.6-1	PCM Television System with a Pseudo-random Noise Addition and Subtraction
	Figure A.7.1-1	Generalized Delta Modulation System
	Figure A.7.1-2	Block Diagram of Single-Bit Delta Modulator Encoder and Decoder

RE-ORDER No. 64-712

Vol I

LIST OF ILLUSTRATIONS (Cont'd.)

APPENDIX A (Cont'd.)

Figure A.7.2-1	Delta Modulation Encoder	
Figure A.7.2-2	Delta Modulation Thresholds	
Figure A.7.2-3	Exponential Integrator Equivalent Circuit	
Figure A.7.2-4	Exponential Integrator Waveforms	
Figure A.8.1-1	Basic Split Band Encoder with Roberts' Compression	
Figure A.8.1-2	Nomograph for Average Compression of Split-Band Encoding	
Figure A.8.2-1	Basic Elastic Encoding System	
Figure A.8.3-1	Split-Band Gross Compression Vs Average Run Length	
Figure A.9-1	Average Run-Length Compression as Function of Average Run-Length Assuming Flat Encoding of Runs for Six-Bit Quantization	
Figure A.9-2	Piecewise Linear Approximation Encoding	
Figure A.9-3	Four-Bit PCM Compared to ΔX-Slope Linear Approximation	
Figure A.9-4	Comparison of ΔX-Slope and ΔX-Amplitude Approximations	
Figure A.9-5	ΔX-Slope Approximation with Error Band	
Figure A.9-6	Linear Approximation Flow Diagram for General Fortran I Program	
Figure A.9-7	Composite Linear Approximation PE and PEQ Program Flow Diagram in Fortran II	
Figure A.9-8	Linear Approximation Compression Program in Fortran I	
Figure A.9-9	Flow Diagram of PE Simulation Program	
Figure A.9-10	Flow Diagram of PEQ Program on a Line Basis in Fortran II	
Figure A.9-11	Comparison of Optimum Coding and Flat Coding Compression for Linear Approximation Coding as a Function of Error Band and Allowable Errors	

RE-ORDER No. 64-712
Vol I

LIST OF ILLUSTRATIONS (Cont'd.)

APPENDIX A (Cont'd.)

Figure A.9-12 Comparison of Compression for 31 and 63 Slopes for Linear Approximation Coding Technique Using Flat Coding

Figure A.9-13 Comparison of Compression for 31 and 63 Slopes for Linear Approximation Coding Compression Technique Using Optimum Coding

Figure A.9-14 RMS Error for Linear Approximation and Equivalent PCM Quantization for Line 1 of Girl Picture as a Function of Error

Figure A.9-15 Average First Order Error Probability Distribution for 31 Constant Slopes for Four Lines Each with 45 Parameter Combinations with an Entropy of 4.84 Bits per Element

Figure A.9-16 Average First Order Probability Distribution for 63 Constant Slopes for Four Lines Each with 45 Parameter Combinations with an Entropy of 5.70 Bits per Element

Figure A.9-17 Average Computer Simulated Compression for Linear Approximation as a Function of PE and the Number of Tolerable Errors Outside the Error Band Parametric in the Error Bandwidth for 15 Lines from Five Different Sources Obtained from EDITS of 256 Resolution

Figure A.9-18 Least Squares Fit Flow Diagram in Fortran II

Figure A.9-19 Computer Simulated Linear Approximation Coding Compression as a Function of Sensor S/N for Three from the Canadian Arctic, Also as a Function of the Error Band and Tolerable Errors

RE-ORDER No. 64-712
Vol I

LIST OF ILLUSTRATIONS (Cont'd.)

APPENDIX C	Figure C.1-4-1	Analytical Fidelity Taker (Objective Fidelity Device)
APPENDIX D	Figure D.1-1	Block Diagram of Linear Approximation Encoder
	Figure D.1-2	Timing Diagram for Linear Approximation
	Figure D.1-3	Linear Approximation Timing Generator
	Figure D.1-4	Linear Approximation Decoder
APPENDIX E	Figure E.1-1	EDITS Block Diagram
	Figure E.1-2	DC Input Level with Gaussian Noise
	Figure E.1-3	PE Versus DC Position Between Quantizing Levels
	Figure E.1-4	PE Versus SPP/NRMS for a Six-Bit A/D Encoder
	Figure E.1-5	Experimental PE Versus S/N Setup
	Figure E.1-6	Comparison of Conventional Vidicon Scan Format with EDITS Scan Format
	Figure E.1-7	Resolution Improvement Effect Produced by a Small Scanning Aperture
	Figure E.2-1	Sinusoidal Test Object Spatial Frequencies
	Figure E.2-2	TV Camera and System Gamma

2

RE-ORDER No. 64-712
Vol I

LIST OF TABLES

SECTION 2	Table 1	Relation of Pictorial Factors to Fidelity Criteria
	Table 2	Fidelity Factors and Effects
	Table 3	Classification of Fidelity Measurements
	Table 4	Compression and Fidelity Performance of Linear Approximation
APPENDIX A	Table A. 2-1	PE as a Function of Sensor Signal-to-Noise for Six-Bit Quantization and 256 Resolution
	Table A. 2-2	Gross Average PEC Compression as a Function of Sensor Signal-to-Noise for Six-Bit Quantization and 256 Resolution
	Table A. 2-3	PE as a Function of Sensor Signal-to-Noise for Four-Bit Quantization and 256 Resolution
	Table A. 2-4	Gross Average Compression for Four-Bit PEC Relative to Six Bits of 256 Resolution as a Function of Sensor Signal-to-Noise
	Table A. 2-5	Net Average Compression for Four-Bit PEC Relative to Six Bits at 256 Resolution as a Function of Sensor Signal-to-Noise
	Table A. 3-1	PE, PL, PE', PL', and (1-PE-PL) Probabilities from EDITS at 256 Resolution
	Table A. 3-2	Compressions Achievable for Pictures Processed on EDITS at 256 Resolution
	Table A. 3-3	Amount of Redundancy in a Picture that is Either Like the Adjacent Element Horizontally or Vertically
	Table A. 5. 2-1	Bit Plane Transition Probabilities for Maximum Sensor Signal-to-Noise Ratio for the Lunar Picture
	Table A. 5. 2-2	Compression Ratios and Average Run Lengths for Bit Planes of Lunar Picture
	Table A. 5. 2-3	Average Compression of 4-Bit Planes for Lunar Picture
	Table A. 5. 2-4	Compression of Various Bit Planes as a Function of Truncation for Lunar Picture Quantized to 6 Bits at 512 Resolution for Both Flat Encoding and Upper Bound Using Entropy
	Table A. 6-1	PE as a Function of Quantization, Roberts Quantization and Number of Shift-Register Stages for 256 Resolution Using MIT Girl Picture on EDITS

RE-ORDER No. 64-712
Vol I

LIST OF TABLES

APPENDIX A (Cont'd.)	Table A. 9-1	MIT Piecewise-Linear-Approximation Computer Results
	Table A. 9-2	Identification of Computer Processed Lines Using Linear Approximation Coding Techniques
	Table A. 9-3	Computer Results for Linear Approximation Coding Simulation
	Table A. 9-4	Linear Approximation Simulation Computer Results for the Canadian Arctic Scene Processed on EDITS as a Function of Sensor Signal-to-Noise Ratio
	Table A. 9-5	Linear Approximation Simulation Computer Results for the Canadian Arctic Scene Processed on EDITS as a Function of Sensor S/N, with th Line Previous-Element Qualified (Peq'd) before Processing
APPENDIX B	Table B-1	Performance Probabilities of Line Sync Subsystem for Uncompressed Television Data
	Table B-2	Search Mode Performance of Line Sync Subsystem for Uncompressed Television Data
	Table B-3	Performance Probabilities of Line Sync Subsystem for Uncompressed Television Data
	Table B-4	Search Mode Performance of Line Sync Subsystem for Uncompressed Television Data
	Table B-5	Performance Probabilities for PEC - Huffman Television System
	Table B-6	Search Mode Performance of Line Sync for PEC-Huffman Television System
	Table B-7	Check Mode, 40 Bit Sync Pattern
APPENDIX C	Table C. 1. 2-1	Relation of Pictorial Factors to Fidelity Criteria
	Table C. 1. 4-1	System Construction Priorities
	Table C. 1. 4-2	Measurement Modes for Objective Fidelity Device
	Table C. 2-1	Classification of Fidelity Measurements
	Table C. 2-2	Photographic Test Parameters
	Table C. 2-3	Fidelity Measurement Schedule
APPENDIX D	Table D. 2-1	Linear Approximation Parts Count Estimate
	Table D. 2-2	Volume Allowed for Parts Estimate
	Table D. 2-3	Power Allowed for Parts Estimate

10

RE-ORDER No. 64-712
Vol I

SECTION 1

INTRODUCTION

It is the purpose of this study to recommend a video data modulation technique that yields a minimum data compression ratio of 2:1. Since there are losses associated with the process of compression, the gross compression must be greater than the desired net compression of 2:1. Factors contributing to the degradation of gross compression include the need for additional synchronization bits and the need for increased transmitted energy per bit to provide a lower bit error probability on the compressed system transmission link.

Taking these degradation factors into consideration a figure can be derived for the necessary gross compression, and the problem becomes one of devising a compression system to furnish it. One way in which this may be done is to reduce the fidelity of the processed picture. In order to keep reduced fidelity to a minimum, and to keep an exact account of the fidelity reduction to be expected in a processed picture, it was necessary to make the study of fidelity criteria an integral part of this video data modulation study. Also it was suspected that trade offs might exist between fidelity and reliability, size, weight, and power consumption of the compression system terminal equipment. Therefore, brief considerations of system trade offs were included in this study.

Once a system was discovered which met the gross compression requirements with apparently acceptable picture fidelity, then two additional parts of the study were carried out. The first was to verify experimentally the gross compression figures and the fidelity expectations through the use of real pictures and computer techniques. The second was to design and make block diagram level recommendations of the selected video modulation technique.

The previous paragraphs summarize the guiding line of thinking which has served to organize the requested study areas of the video data modulation study. However, there has developed much detailed data, particularly in the use of several compression schemes to meet the gross compression or fidelity requirements. The bulk

RE-ORDER No. 64-712
Vol I

of the detailed experimental investigation and results are set forth in the appendices. However, the essence of their findings is summarized in this, the main body of the report.

RE-ORDER No. 64-712
Vol I

SECTION 2

INVESTIGATION

2.1 Organization

The investigations that are a part of this study are grouped into four areas:

- o Net and gross compression
- o Fidelity
- o Performance verification
- o Video modulation recommendation.

2.2 Net and Gross Compression

An uncompressed television picture contains a fair amount of redundancy. This redundancy enables the human viewer to extract intelligence from a television picture at signal-to-noise ratios far less than those normally associated with "pleasant" viewing. When a television picture is compressed, that is, when the redundancy is partially or considerably removed from the signal, each signal bit must carry a greater proportion of the picture description making it reasonable to assume that the compressed bit stream is more "fragile" with respect to channel noise. This is indeed the case and has been experimentally verified. The relative fragility of the compressed signal bit stream implies a requirement of more signal energy per bit.

In addition to requiring more signal energy per bit, the compressed signal will require a more reliable synchronization word since the decoding of the compressed bit stream depends, generally, on a firm knowledge of the starting point. This means that though the compressed system requires fewer information bits per frame, it will usually require more synchronization bits per frame. For instance, in a 512 x 512 element PCM picture, 6 x 512 x 512 information bits and approximately 512 x 30 sync bits will be required. In a compressed system of 512 lines, about 512 x 50 sync bits are required. The difference, 10240 bits, is a negligible portion ($\approx$ 0.65%) of the total number of uncompressed bits. Even assuming a very large (10:1) compression ratio, the additional number of sync bits is still only 6.5% of the number of compressed information bits. To the degree of accuracy assumed in a preliminary design, the effect of the increased number of sync bits on compression is negligible.

The most serious reduction of gross (or design) compression, therefore, comes from the need to send each compressed information bit at a lower bit error probability than that satisfactory to send an uncompressed information bit. A bit error probability of 1×10^{-3} is satisfactory for many PCM video applications. A bit error probability of 1×10^{-5} is satisfactory for most compression systems wherein a bit error may destroy, at most, the remainder of the line in which it occurs. For instance, a 512 x 512 x 6 bit video frame will have about 1600 picture elements in error at 1×10^{-3}. At a gross compression of 4:1, a compressed 512 x 512 element video frame will have about 1000 picture elements in error in the decoded picture for a 1×10^{-5} bit error rate. We note, of course, that the errors in the PCM picture will be uniformly distributed "salt-and-pepper" type errors whereas the errors in the compressed picture will be contained within four lines (expected value) which are of average corrupted length, 256. The relative subjective merits of these two dissimilar noise phenomena will not be debated here. Suffice it to say, the absolute number of errors is so low as to render the pictures usable for general viewing, and probably usable for most serious scientific work as well.

What penalty is paid for the relative luxury of a 1×10^{-5} bit error probability compared with a 1×10^{-3} bit error probability? This depends on the choice of the communication link modulation system. A popular "standard" choice is coherent phase-shift keying (PSK). For this system approximately 2-1/2 db more signal energy per bit is required for 1×10^{-5} than is required for 1×10^{-3} bit error probability. In this instance the 4:1 gross compression postulated must be divided by 1.8 (2.5 db) to yield a net compression on the basis of signal energy per video frame of about 2.2:1. To a working first approximation, net and gross compression are related by equation (1) below.

$$C_n = \frac{C_g}{1.8}, \qquad (1)$$

where C_n and C_g are the net and gross compressions, respectively.

Within the context of this study, compression is referenced to six bit PCM. That is, a system with a compression (gross) of 2:1 is one in which the average number of bits per picture element is three. A compression of 6:1 implies one bit per element--a goal long-sought in the video compression field. Note that though "compression" is referenced to six-bit PCM, picture quality of the particular system need not necessarily be equal to six-bit picture quality in order that the system be labelled as an "X-to-one" compression system. However, for judging the operational value of a given system, its picture quality may be referenced to either six-bit picture quality or to some other, task oriented, fidelity criteria. Some of the latter are discussed in paragraph 2.3.

2.3 FIDELITY

In an analysis of the factors governing photographic image quality(1) there are at least three current views. The first holds that the factors controlling image quality are

- o Angular field
- o Definition
- o Distortion
- o Character of emulsion
- o Altitude
- o Ground speed
- o Vibration
- o Character of illumination

The second view holds that the amount of information in a photo depends on four image properties:

- o Graininess
- o Sharpness
- o Resolving power
- o Tone reproduction

Finally, the third view reduces the problem to three primary characteristics governing the photo image quality:

- o The tone or color contrast between the object and its background

(1) Robert N. Colwel, Editor, Manual of Photographic Interpretation, published by American Society of Photogrammetry, Washington, D. C., 1960.

RE-ORDER No. 64-712
Vol I

- o Image sharpness characteristics
- o Stereoscopic parallax characteristics

The first approach is system oriented in that it includes vehicle characteristics. While these are of ultimate importance, it is only the sensor oriented characteristics, typified by the second outlook that interest us in this study. In the third view, these are condensed essentially to two sensor-oriented characteristics (stereoscopic parallax being a trajectory and timing problem).

The problem of good television images is certainly analogous to the problem of good photographic images. The television system must render tones accurately and it must sharply render the boundaries between these tones. In television terms, this means that gray scale rendition and good step function response are essential.

There exist, however, three additional types of images and their electrical equivalents which are of interest. In observing aerial photographs of predominantly geological areas, one often encounters long, thin, winding lines at virtually the limit of resolution. These lines, when scanned, will give rise to impulses within the electro-optical system. If these lines are to be recorded, the impulse response of the system must give rise to a pulse which will exceed the next higher threshold level for a sufficient length of time to allow at least one digital sample to occur. The detection of thin, low contrast lines (roads, railroad beds, fault lines, lunar rays, etc.) implies a requirement for good impulse response in the overall system in addition to good step function response.

In other parts of photographs of predominantly geological interest, individual dots appear which are presumably trees or other isolated vegatation. These optical point images will give rise to small spread functions which will also result in impulses into the digital portion of the system when scanned. The digital system resolution (in both dimensions) must be set so that these spread functions will have a reasonable chance of being scanned near their central peak and the analog portion of the system will again be required to have a good impulse response.

In viewing geological pictures, adjacent sections of closely packed trees often occur which appear to have approximately equal <u>average</u> tone. However, there is a distinct demarcation between the sections of forest on the basis of texture. For instance, large old-growth timber closely packed differs in texture from

RE-ORDER No. 64-712
Vol I

similar species timber of new growth and medium density, yet the average tone for the areas is quite close. A similar effect can occur over farm crop lands. Even though individual trees and crops are below the system resolution limit, it appears that an electro-optical system of good midrange sinewave response is required for adequate texture distinction.

Table 1 summarizes the relation of pictorial factors and fidelity criteria.

The listing of fidelity factors is suitable for the evaluation of low-noise photographs. However, a somewhat more inclusive listing is presented in Table 2. This listing includes the effects of unwanted perturbations due to modulation, synchronization, and bit errors. Furthermore, it is seen that the six fidelity factors may be grouped into three simplifying categories: sharpness, tonal rendition, and purity.

Table 2 presents a qualitative estimate of the effect of resolution, filtering, contouring, gray scale, modulation and synchronization, and bit error upon the detection of each of the pictorial features; points, lines, edges, shades and textures. The effect of each fidelity factor is estimated for two conditions: (1) detection of all true occurrences and (2) detection of false (pseudo) occurrences of the pictorial feature and is rated as being of primary importance, of secondary importance, or of no (or relatively minor) importance.

The assessment of fidelity so far has been largely qualitative. A quantitative, or analytical, approach to fidelity is also needed. Such an approach, detailed in Appendix C. 1, consists of the simple concept illustrated in Figure 2. 3-1.

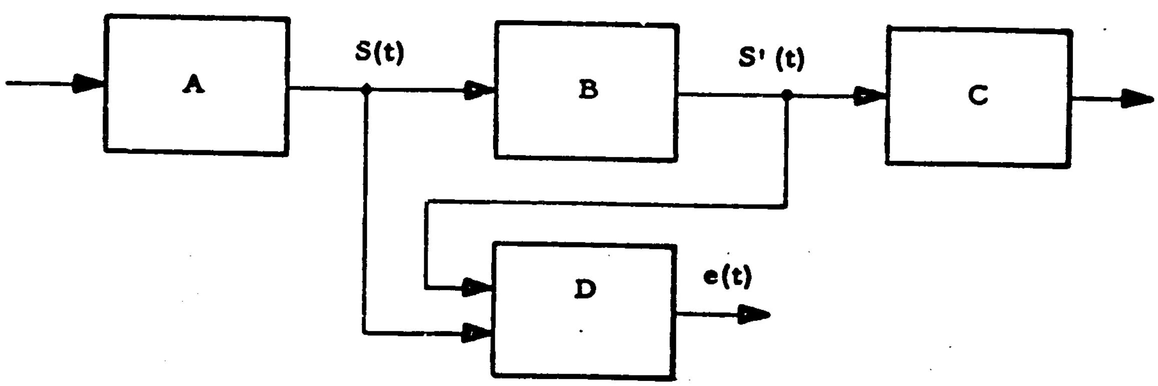


FIGURE 2. 3-1

ANALYTIC FIDELITY CONCEPT

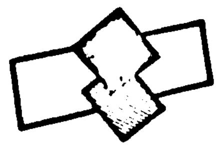
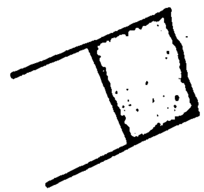
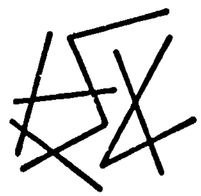
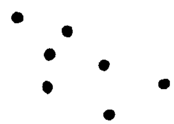
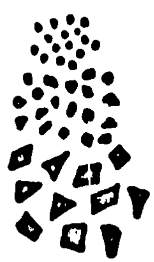
DETECTION OF	IMPLIES	PICTORIAL FACTOR WHICH REQUIRES	SYSTEM FIDELITY CRITERION
OBJECTS		TONE CONTRAST	LINEARITY, QUANTIZATION LEVELS, NUMBER AND SPACING
EDGES OF OBJECTS AND OTHER TONAL BOUNDARIES		SHARPNESS	STEP FUNCTION RESPONSE
LONG, THIN LINES		SHARPNESS	IMPULSE RESPONSE AND DIGITAL RESOLUTION (1 DIMENSION)
ISOLATED POINTS		SHARPNESS	IMPULSE RESPONSE AND DIGITAL RESOLUTION (2 DIMENSIONS)
COLLECTION OF POINTS AND/OR BLOBS AT OR NEAR TERMINAL RESOLUTION		TEXTURE CONTRAST	MID-RANGE SINE WAVE RESPONSE

TABLE 1

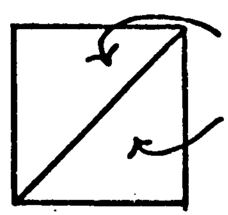
RELATION OF PICTORIAL FACTORS TO FIDELITY CRITERIA

TABLE 2

FIDELITY FACTORS AND EFFECTS

Fidelity Factors	Resolution	(Analog Low Pass) Filtering	(Coarse Quantization) Contouring	(Analog Linearity) Gray Scale	Modulation and Synchronization	Bit Errors
Pictorial Features	Sharpness		Tonal Rendition		Purity	
Points	P / N	S / N	P / S	P / N	P / N	N / P
Lines	P / N	S / N	P / N	P / N	P / N	N / N
Edges	P / N	P / P	P / P	P / N	P / N	N / N
Shades	N / N	P / P	P / P	N / N	P / S	N / N
Textures	S / N	S / S	P / N	S / S	P / P	S / P

Key:



Effect on detecting all true occurrences

Effect on producing false occurrences

P - Primary
S - Secondary
N - None (or relatively minor)

RE-ORDER No. 64-72
Vol I

The analytic measure of fidelity of a system is either

$$e(t) = S'(t) - S(t) \tag{1}$$

or some function of e(t), averaged over a suitable period of time. If the system is to be used for photometric work, then an analytical fidelity criterion of peak error $\widehat{e(t)}$ for one picture time is a very useful measure. With this measure either empirically determined or estimated from the system design, the engineer can assure the space science experimenter that all measurements will be within X% of true, or will so be with very high confidence. For instance, considering the analog-to-digital and digital-to-analog portions only, a PCM system will have a peak error equal to or less than one-half quantizing interval. However, it must be realized that some very useful systems, for instance Delta modulation, will at times exhibit instantaneous errors close to 100% of full scale. For such systems the analytical peak error criterion is unduly severe and another criterion should, in all fairness, be chosen. Such an analytical criterion may be the familiar one of rms error.

In the final analysis, however, most experimenters judge system performance subjectively. The final test of a system is picture quality, whether or not it is used photometrically.

The comprehensive fidelity measurement program defined in paragraph C.2 was designed to answer these questions. In this program the relative fidelity of the compression system incorporated into EDITS (Experimental Digital Television System) is determined by the comparison of pictures rendered by each system relative to the same test material rendered by a six-bit PCM system operating under the same condition.

A total of eight video modulation system variations were included in the test program. These include: six-bit PCM as the primary reference technique, 4, 3, and 2 bit PCM as secondary references together with three-four Roberts, two-three Roberts, 2-bit Delta, and 4-bit PEC/Huffman compression systems. Table 2.3-3 presents the organization of the test program which is shown to be divided into three measurement catagories. They are: (1) Objective, (2) Quasi-Objective, and (3) Subjective. Section 2.4 summarizes the more significant fidelity

TABLE 3

CLASSIFICATION OF FIDELITY MEASUREMENTS

Category	List of Measurements	Data Characteristics		Test Conditions
		Input	Output	
1. Objective	1. Sine wave response 2. Square wave response 3. Gamma characteristics	Calibrated	Objectively measured	40 db S/N (approx.)
2. Quasi-Objective	1. Overall image fidelity 2. Sine wave lim. resolution 3. Square wave lim. resolution 4. Brightness contouring	Calibrated	Subjectively measured	Includes 20, 30, and 40 db S/N
3. Subjective	Overall image fidelity	Partially calibrated	Subjectively measured	Combinations of 20, 30, and 40 db S/N and $10^{-\infty}$, 10^{-4}, 10^{-3}, and 10^{-2} P_e.

RE-ORDER No. 64-712
Vol I

RE-ORDER No. 64-712
Vol I

measurement results derived from this program.

2.4 PERFORMANCE VERIFICATION

2.4.1 Introduction

Performance verification is in two areas, compression and fidelity. In several of the compression systems discussed in paragraph 2.4.2 the compression achieved did not meet expectations. For these systems, fidelity considerations are academic. For those systems which do meet, or appear to meet, the JPL compression requirements, some discussion of fidelity is included as a part of the system commentary. However, in addition to this, there are several points of significant interest with regard to fidelity verification utilizing EDITS. These points are covered in paragraph 2.4.3.

2.4.2 Compression Verification

The compression study investigated both objective and subjective compression techniques that had the potential of achieving a minimum net compression ratio of at least 2:1. The compression analysis was conducted in the following manner: for each compression technique a mathematical model was formulated and the compression was calculated from the entropy of the model. Preliminary values of compression were obtained by taking limiting values for the statistics necessary to evaluate the model. If these statistics indicated that gross average compressions greater than 2:1 were possible, then the system was simulated by taking actual statistics on EDITS; alternately, the compression technique was simulated on the ASI-210 digital computer. If the statistical model then indicated that the system satisfied the minimum gross compression of greater than 2:1, the net compression was calculated by considering sensor noise, channel noise, and synchronization.

Eight separate and distinct compression techniques were analyzed and these were previous element coding (PEC); previous-element/previous-line coding (PEPL); Roberts coding; Delta modulation; split-band coding; bit-intensity run-length coding (BIRL); area coding; and linear approximation. Each of these techniques have definite upper boundaries on compression due to (1) the theoretical restrictions imposed by the basic concept of the compression technique, (2) the theoretical limitations imposed by the noise inherent in the sensor, (3) limitations imposed

RE-ORDER No. 64-712
Vol I

by the amount of redundancy inherent in the actual scene, and (4) the loss of compression as a result of an increase in power due to implementation and an increase in information content of each transmitted bit.

Previous Element Coding

Previous element coding (PEC) utilizes adjacent element correlation in a frame along with efficient coding to achieve compression. Assuming that the video is quantized and converted to fixed-length digital words, the system concept can be described as follows: Successive elements are compared and if they are alike, a single bit PE word is transmitted; however, if they are not alike, the quantized value is encoded and digitally transmitted with a suitable prefix bit.

The PE symbols are assigned one bit because they are more probable. The quantized amplitudes of picture elements that differ from the preceding element give rise to symbols longer than their PCM equivalent. However, because these 1-PE symbols are of relatively lesser occurrence than the PE symbols, the overall result is a reduction in average number of bits required to transmit a picture element.

The PEC system has the advantage of being a completely information preserving coding system since all of the coding and decoding steps are reversible processes. The only advantage taken is of the statistical redundancies inherent in the signal, and no further degradation is experienced in order to achieve compression. The digital video from a PEC system should be identical with the digital video from an equivalent PCM system.

The compression available, of course, depends very strongly on the statistics of the message source. For this reason a representative number of message sources (pictures) must be measured. Table A.2-3, A.2-4,and A.2-5 in Appendix A summarize the PE probability statistics, gross compression and net compression findings for the PEC system. Average net compressions between 1.43 and 1.98 were observed for good sensor signal-to-noise ratios. Since these figures are all less than the 2:1 required by the study, PEC is not a recommended system.

Previous-Element/Previous-Line (PEPL) Coding

In PEC there are a number of elements not like their previous neighbor (1-PE

RE-ORDER No. 64-712
Vol I

elements). It was thought possible that a number of these 1-PE elements would be like their counterpart on the immediately previous line. If enough were of this nature they could be given a special two-bit symbol called a PL symbol. While the PL symbol is longer than the one-bit PE symbol, it still is shorter than PCM encoding or PCM encoding with the appropriate prefix bits. It was believed that the average symbol length could be lowered below that exhibited by the PEC system. The PEPL system is more complex than the PEC system because it requires storage of a complete previous line of video. Any increase in compression observed must be substantial to justify the added terminal complexity. The system was implemented and PL statistics were taken. Unfortunately, the increase in compression observed was less than 10% over PEC. PEPL is not a recommended system.

Area Coding

With the failure of PEPL to provide a significant gain in compression through the use of line-to-line correlations, another approach was tried. In general, a picture element is most apt to be like its immediate neighbor, and the probability of similarity decreases with distance. If the scanning beam moves four picture elements away from a given picture element, the correlation on a per-element basis will drop, and there will exist some mean correlation. If the scanning beam was made to traverse a different path for four elements so that it remained, on the average, closer to the original element, then the mean correlation could be expected to increase. If the mean correlation increases, the increase can be expected to reflect in a higher PE probability.

In paragraph A.4, an expression is derived which implies that for PE = 0.5, the optimum area to scan is about four picture elements (2 x 2). This would take the form of a Z-scan, proceeding from upper left to upper right to lower left to lower right, thence to upper left of the next subscan area.

The elemental area scan technique was tried on EDITS and the results appeared to be discouraging. The PE probability for the PEA (Previous Element Area) system did not increase significantly with a two-element-square subscan.

31

RE-ORDER No. 64-712
Vol I

Examination of the video revealed that an alternation of video pulse heights occurred. It was determined that this was a result of the particular scanning technique imposed. Several possible solutions were suggested, but none adequately solved the problem. This unfortunate turn coupled with the results of the PEPL analysis indicates that unless sensor noise and other peculiarities are eliminated, a PEA coding technique does not offer a sufficient value of gross compression to satisfy the minimum JPL requirements. For six-bit quantization, an average of only 50% of the elements are redundant; therefore, utilizing the statistical redundancy directly could result in an upper boundary on compression of only 2:1. When synchronization, the effects of channel noise, and implementation problems are considered, it is evident that a six-bit PEA system will not satisfy the JPL compression requirement.

Roberts Coding

The Roberts coding technique of bandwidth compression is not an information preserving coding technique. Compression is achieved by reducing the quantization and breaking up the annoying contours by the addition and subtraction of identical pseudorandom noise at the transmitter and receiver, respectively. The technique indirectly utilizes the statistical redundancy and it displays the data in such a manner as to take advantage of the physiological properties of the human viewer.

Implementation of the technique is achieved by the addition of a nonlinear amplifier and a pseudorandom noise generator to the transmitting and receiving portions of a conventional PCM television system. A nonlinear amplifier is used to match the response of the eye to the displayed image and optimize the overall system transfer function. Since the response of the eye to incident light is a square-root function, the dynamic range could be compressed to concentrate intensity errors to those of high intensity where the eye would readily notice these errors. This portion of the technique could physically be implemented by passing the analog video signal from the sensor through an amplifier with a square-root transfer function which compands the video signal. Pseudorandom noise is then added to the companded signal before it is converted to a digital signal.

The subjective results of the comparison of compression techniques are given in paragraph E.2. Summarizing the Roberts results we find that on a single frame basis at 40 db sensor signal-to-noise ratio, the three-four Roberts is not subjectively equivalent to six-bit PCM as claimed by Roberts. However, it does appear from the photographs in paragraph E. 4 that, subjectively, the Roberts technique is preferred to Delta compression.

It appears from these photographs that there exists a relationship such that six-bit PCM and three-four Roberts are almost subjectively equivalent at X db and Y db signal-to-noise ratio for the two systems, respectively. Therefore, the six-bit PCM and three-four Roberts Polaroid pictures at 30 and 20 db signal-to-noise were enlarged to 8 x 10 to determine this relationship. Both the Canadian Arctic and the JPL Lunar scenes were used for the comparison since they represent extremes in high and low contrast scenes. From the Canadian Arctic and JPL Lunar photographs, A. 6P-9, A. 6P-10, A. 6P-13, and A. 6P-14, at 30 db sensor signal-to-noise ratio, six-bit PCM not only represents subjectively a more pleasing picture, but it is superior in line and small detail recognizability. It is apparent that the lower contrast photos are almost subjectively equivalent somewhere between 20 and 30 db sensor signal-to-noise ratio; however, for the high contrast scene, even at 20 db sensor signal-to-noise, the six-bit PCM is preferred to the three-four Roberts both subjectively and in objective detail recognizability.

It is seen from the Roberts photographs that net and gross compressions are substantially equivalent at about 2:1 for the Roberts system. However, from Figures A. 6 P-9 through A. 6 P-16, it is quite apparent that the three-four Roberts is not subjectively equivalent to the six-bit PCM. There does appear to be some relationship between the signal-to-noise ratio required for six-bit PCM and three-four Roberts. It then remains to be seen whether or not the space scientist can utilize the information from the Roberts display. Apparently, somewhere between 20 and 30 db sensor signal-to-noise ratio, the six-bit PCM and three-four Roberts could become subjectively equivalent.

Bit Interlace Run Length (BIRL)

A picture may be represented by one plane of six-bit binary numbers. Alternatively one can think of the six-bit binary numbers as standing on end upon their base plane and then being sliced horizontally by six parallel planes. After this is

RE-ORDER No. 64-712
V&I

accomplished, conceptually, there exists six planes each containing only one bit at each picture element position. Possibly, there is an advantage in having six one-bit planes per picture instead of one six-bit plane.

In run-length coding, the end of a run occurs when the value of the function changes from the value maintained during the run. For a six-bit PCM system the sixth (or low order) bit is largely of random value and rapidly fluctuating, even at high signal-to-noise ratios. This indicates that long runs of six-bit PCM words are quite improbable, even for a constant level input. However, the fifth bit and higher order bits are less susceptible to random changes. Even so, their lack of rapid change is not available for exploitation as straight run-length encoding. To exploit the suspected redundancy of the higher order bits, the bits (or bit planes) must somehow be separated and encoded separately. This concept is the basis for bit-interlace-run-length (BIRL) coding.

In BIRL, each separate bit plane is run-length encoded. At the receiver the bit planes are reconstituted from the reduced redundancy signal. The decoded bit planes would be interlaced together in the display with the proper analog weighting for each bit plane to reconstruct the original half-tone PCM picture from which the original bit planes were obtained. Unfortunately, the overall result of these analyses was a compression of only 1.3:1. Therefore, BIRL is not a recommended system.

Delta Modulation

A Delta modulation system includes a decision element and an approximation element. As an option it may include a pre-processing unit and a post-processing unit. The input waveform is compared to the approximated waveform in the decision element. Depending on the similarity between the input waveform and the approximating waveform the decision element alters the output of the approximation element. The signal by which the decision element changes the operation of the approximation unit is the signal which is transmitted. At the receiver there is another approximation element identical to the one at the transmitter. The receiver approximation element, under control of the transmitter decision element, produces a facsimile of the input waveform at the receiver output.

The "classical" form of Delta modulation uses both a simple decision element and

RE-ORDER No. 64-712
Vol I

an equally simple approximation element. The decision element is a difference amplifier with a threshold at zero. If the difference between the input waveform and the approximation waveform is positive, a negative pulse (in principle) is sent to the approximation element. If the difference is negative, a positive pulse is sent. The approximation element is a simple integrator of near infinite time constant. Thus, the stream of difference signal pulses is integrated into a (hopefully) reasonable facsimile of the input waveform.

This simple Delta system has numerous drawbacks and the Delta utilized in this study is a more complex Delta system in which the decision element having three thresholds emits a two-bit signal,and the approximation element is the equivalent of an exponential integrator with time constant equal to five picture elements.

The sample rate for the Delta system was chosen to be identical with the sample rate of the PCM system and other compression systems so that a valid pictorial comparison could be made. Since the decision element utilizes three thresholds, any of four corrective signals (two bits) are transmitted at each video sample time. This two bits of transmission per picture element results in a gross compression of 3:1.

As can be seen from photographs in Figure E. 4 P-16, the Delta image is "softer" than the PCM images. However, the Delta image at a bit error probability of 10^{-3} is probably acceptable relative to the Delta image with $10^{-\infty}$ bit error probability. So one concludes that (neglecting differences in synchronization bit requirements) Delta net compression approximately equals gross compression.

As can be seen from the pictures, the Delta rendering yields a "soft" picture. This softness is further brought out by examining the Delta step function response in Figure E. 2 P-2. While the Delta system yields pleasing pictures which may be usable for command and control functions, the three-to-one reduction in step function response rise time may eliminate the Delta system from serious consideration where sharpest pictures and best photometric accuracy are a system requisite. Therefore, Delta modulation is not a recommended system in the context of the JPL study, even though it may well have other uses in space video communications.

RE-ORDER No. 64-712
Vol I

Split-Band

The split-band technique consists of treating the video as though it were composed of two distinct signals: a low frequency component and a high frequency component. Each component is compressed by an appropriate technique. At the receiver, each part is reconstituted and combined with the other to reform the original video. The low frequency part is formed by passing the video through a low-pass filter with cutoff frequency about 1/10 to 1/16 the maximum video frequency. The high frequency component is formed by passing the video through a complementary high pass filter.

The most efficient compression technique for the low frequency component occurs simply because the sampling frequency for the lows is from 1/10 to 1/16 the sampling frequency needed for the whole video signal. Additional compression is obtained on the low frequencies by Roberts processing (see Appendix A, paragraph 6). The compression of the highs is accomplished by detecting spikes in the highs which correspond to edge transitions in the original video. The spikes are encoded in amplitude and in relative position to one another. The exact technique employed on the lows has relatively small influence (in the order of ±20%) on the overall compression

There appears to be a very good possibility that a split-band system will meet the 2:1 net compression required, or even the 3:1 net compression desired. To do so would require average run-lengths between encoded highs of 5.4 to 8.8. From evidence in the literature, these values appear feasible although the threshold setting on the highs channel necessary to achieve them is not specified. A high threshold setting (necessary to achieve long run lengths and high compressions) would mean lowered fidelity and loss of low amplitude, high frequency components.

Another factor bearing on fidelity is the fact that the video is split into two paths, each of which may be in error by as much as one-half quantizing level. For the case where the lows channel and the highs channel have equal quantization, this peak error is double that of a PCM system of equivalent quantization. In fact, the split-band system is specifically designed to take advantage of certain characteristics of human vision to achieve compression through a controlled reduction in fidelity. Whether or not such a system would meet the accuracy requirements

RE-ORDER No. 64-712
Vol I

of a photometric system is problematical. In summary, the split-band system is not the recommended system of this study because of the uncertainty in the trade-off between statistics affecting compression and resulting fidelity. If the system were simpler in concept or implementation, it would warrant further investigation into these matters on the basis of its apparent potential for compression.

Linear Approximation

The linear approximation coding technique represents the video signal by a series of straight line segments. If the number of bits required to represent these segments in slope and length is less than the number of bits required to represent the video signal on a point-by-point basis, then compression results. The linear approximation system faces two decisions: what slope to choose and when to give it up. In the EMR design these decisions are interrelated. All slopes are simultaneously generated. That slope which is sufficiently like the input video for the longest run is the slope which is chosen. A slope is abandoned when the absolute error between it and the video signal exceeds a predetermined threshold a given number of times in succession.

Within the linear approximation system, a direct trade-off between compression and fidelity exists. If the slope-video error threshold is reduced the fidelity of approximation is increased. However, the slopes will, on the average, terminate sooner and compression is reduced. Consequently, much of the work reported in Appendix A, paragraph 9, consists of trade studies between error thresholds (and other parameters) and compression. These trade studies were accomplished by computer (ASI-210 plus FORTRAN I and FORTRAN II) using actual video lines from EDITS. Since the lines must be manually typed into the computer, complete frames of video were not evaluated, but every effort was made to choose a wide variety of single video lines.

The computer program can be summarized as follows: the program to simulate the linear approximation coding technique has the following options: (1) the Δx runlengths were truncated to obtain compression as a function of truncation of the Δx runlengths; (2) the number of slopes was variable (31 and 63); (3) the error

PO ORDER No. 64-712
Vol I

band within which the function must lie was variable (±1, ±2, and ±4 levels out of 64 possible amplitude levels); (4) the number of consecutive errors that could be tolerated was variable (0, 1, and 2 errors); (5) the RMS error and the equivalent PCM error could be obtained; (6) the Δx runlength probabilities could be obtained; and (7) the compression could be obtained for either flat coding or optimum coding.

Like most other compression systems, the performance of linear approximation is divided into two parts--compression and fidelity. With respect to fidelity, we are not in as good an experimental position with linear approximation as we are with other systems such as Roberts and Delta which are incorporated into EDITS. Therefore, our measures of fidelity for linear approximation derive from three sources: (1) error band threshold setting, (2) rms error calculated for the computer runs on each line, and (3) full frame pictures produced at MIT some years ago using a similar technique with rather wide error bands. With respect to compression, we are considerably more certain of our conclusions since compression figures are derived from fifteen video lines (3,840 elements) taken from six photos of four generically different subjects.

The linear approximation technique is of that class wherein a single bit error has the possibility of corrupting the rest of the video data in that particular line. Therefore, the gross compression, for which the system parameters must be chosen, is about 1.8 times the desired net compression. To achieve the JPL study requirement minimum net compression of 2:1, the system must be designed to achieve a 3.6:1 gross compression. For the JPL desideratum of a net compression of 3:1, a gross compression of 5.4:1 is required. For pictures of average complexity (as measured by their PE probability) the choice of an error band threshold of ±2/64 of full scale and the choice of two consecutive threshold violations result in an average gross compression of 3.6:1. This is for a system in which 31 slopes are possible and slope run lengths to 32 elements may occur. It is also for "flat" coding; that is, where every slope is represented by a word with the same number of bits. In this case the number of bits is ten, equally distributed between slope specification and runlength specification.

As an illustration of the trade-off flexibility of linear approximation, the gross

RE-ORDER No. 64-712
Vol I

compression under the previous conditions may be raised to 4.3:1 by allowing three consecutive threshold violations. Or, it may be raised to 5.4:1 by retaining the two consecutive threshold violation rule and yet relaxing the error-band threshold from ±2/64 of full scale to ±4/64 of full scale. Finally, if one allows both three consecutive threshold violations before slope abandonment and the error-band threshold to be ±4/64 of full scale, an average gross compression of 6.5:1 occurs.

With these attractive compressions in mind, an estimate of system error is in order. The rms error was calculated on the basis of the difference in video amplitude given by the linear approximation system and that given by six-bit PCM. This rms error must be appropriately combined stated in terms of rms error with the rms error for PCM, which is $\frac{1}{2\sqrt{3}}$ of a quantizing interval. The error results for each compression are given in Table 4. The rms video error for six-bit PCM is about 0.3; for five-bit PCM, 0.6, for four-bit PCM, 1.2, for three-bit PCM, 2.3. In percent of full scale these errors become 0.47%, 0.94%, 1.88%, and 3.60%, respectively.

TABLE 4

COMPRESSION AND FIDELITY PERFORMANCE OF LINEAR APPROXIMATION

Gross Compression	Net Compression	RMS Error (relative to Six-Bit PCM)	Total Video RMS Error	Percent of Full-Scale RMS Error
3.6:1	2.0:1	1.6*	1.63*	2.55%
4.5:1	2.5:1	1.9	1.93	3.02%
5.4:1	3.0:1	2.2	2.22	3.47%
6.5:1	3.6:1	2.5	2.52	3.94%

*Units are in 64ths of full scale.

Six-bit PCM rms error relative to original video = 0.289 ≐ 0.3

As can be seen, on rms error a three-bit PCM system is not bad as analog instrumentation goes. It is predominantly the contouring in three-bit PCM which is objectionable. However, the very nature of the linear approximation video reconstitution at the receiver allows a continuum of gray shades. Therefore, we expect linear approximation output to be more pleasing than three bit PCM though it may possess an approximately equal rms error. In the absence of implementing linear approximation on EDITS (beyond the scope of this study) and actually

32

RE-ORDER No. 64-712
Vol I

processing whole frames of video, we still find reasonable assurance, on an rms basis, that satisfactory pictures will be produced at a net compression of 2:1 (3.6:1 gross). This assurance is further increased by observing the MIT linear approximation pictures with an error band tolerance as high as ±5/64 of full scale.

In summary, linear approximation has five attributes which are verified by simulation with an actual video input.

- o Gross compression >3.6:1
- o Net compression >2:1
- o Probably acceptable fidelity
- o RMS error of 2.5 to 3% of full scale
- o Flexibility of trade off between compression and fidelity

Accordingly, the linear approximation system is the recommended video data modulation system resulting from this study.

2.4.3 Fidelity Verification

Once the compression objective has been met by a system, then the verification of that system's fidelity becomes of paramount importance. Insofar as fidelity verification on EDITS is concerned there are several important test conditions common to all fidelity measurements. EDITS vertical and horizontal scans are adjusted to provide 256 horizontal element samples per scan line and 256 scan lines per frame. These parameters establish the nominal resolution limit for all input video data for this study. Due primarily to highlight and lowlight saturation of the Polaroid film which was used for data recording, the maximum contrast ratios of output photographic images are limited to approximately 35:1. The midrange gamma characteristic of the combined vidicon camera/display/film subsystem is approximately 2:0. This results in an increase in contrast of the output images compared with the input test images, but limits the useful contrast range of the input test photographs to approximately 20:1.

In addition to the common EDITS factors relating to fidelity, there exist several key fidelity measurement results. The reference six-bit PCM system consistently produces the best overall subjective picture fidelity under any given set of

RE-ORDER No. 64-712
Vol I

conditions. The degree of its superiority is, as would be expected, a function of the input photographic data characteristics. The degree of superiority of six-bit PCM also is very much a function of the TV sensor video signal-to-noise ratio. As the sensor signal-to-noise ratio is lowered, the fidelities of all modulation techniques converge downward toward a condition of equality. This results from the fact that the high fidelity images suffer a greater percentage of degradation than the lower fidelity image for an equal increase in sensor noise. This effect is particularly evident when three-four Roberts performance is compared with six-bit PCM at the three signal-to-noise conditions.

One modulation technique, viz. four-bit PEC/Huffman, is considerably more susceptible to degradation under simulated high level digital channel noise than any other technique tested. This is due to propagated data word errors resulting from the coding/decoding process.

The objective and quasi-objective fidelity measurement series graphically demonstrate the need for good step function response in the reproduction of black-white transitions encountered on standard television test charts. The two-bit Delta system fidelity under such conditions is far below PCM performance. However, on natural terrain photographs, the respective fidelity levels appear to be nearly equivalent. This illustrates the importance of accurately defining the required frequency response characteristics of the input video data to satisfy a particular mission.

Despite observed image fidelity differences in a given measurement series, there is still the question of the comparative usefulness of the image data. This determination is best left to the individual reader who can evaluate the pictorial results in relation to his particular set of requirements.

RE-ORDER No. 64-712
Vol I

2.5 RECOMMENDED VIDEO MODULATION SYSTEM

There are four major compression systems which satisfy the JPL minimum requirement for 2:1 compression. They are; three-four Roberts, Delta, Split-Band, and Linear Approximation. In this study EMR has taken the minimum JPL requirement to mean 2:1 net compression; i.e., gross compression that equals a minimum of 2 when it has been divided by a factor that accounts for additional signal energy per bit and additional synchronizing bits.

The Roberts system offers 2:1 compression, but upon judging the pictures relative to six-bit PCM, the usefulness of the resulting subjective fidelity is in doubt. Delta modulation provides 3:1 gross compression and 2.5:1 net compression, but here again adequacy of the fidelity as evidenced by the pictorial results is in question. Split-Band clearly offers gross and probably net compressions greater than 3:1, and yet the published literature is not sufficiently detailed to make an accurate trade off comparison between fidelity and compression. Based on the above comments, it is felt that these three systems do not qualify unreservedly as recommended video modulation systems.

Linear approximation also offers gross and net compressions of greater than 3:1. The trade off between fidelity and compression has been studied extensively during this program. Fidelity, in this sense, has been studied by examining line traces and computing rms error in the line traces, rather than processing full frame pictures. The linear approximation system is quite flexible, since, as an approximation system, it offers a relatively smooth varying trade off between the degree of approximation and the resulting compression achieved.

Linear approximation, as the name implies, approximates the given video function by a piece-wise, continuous function consisting of a series of straight-line segments. In this system, two questions immediately arise: how to choose the correct slope and when to give it up.

Several methods of choosing an initial slope for approximation exist . The most common method is to take the first derivative at the requisite point. If the first derivative is not convenient, one may take the first difference; however,

RE-ORDER No. 64-712
Vol I

neither method is used in the recommended video data modulation system because both the derivative and first difference are highly susceptible to video noise. The approach taken in the linear approximation system is to generate a complete family of slopes starting from the requisite point. In the system developed by EMR, a family of 31 slopes ranging roughly between $\pm 90^{\circ}$ are used. There is a system synthesis problem in determining the optimum spacing of the slopes. A good first approximation choice is to use equal angles between the slopes.

The question of when to give up the chosen slope is also an important one, and it has a major bearing on the trade off between fidelity and compression. Effectively, an error band is placed around the video to be approximated. All slopes in the family are simultaneously compared with the input video to the error band tolerance chosen. As the video progresses, more and more slopes exceed the error band tolerance. The last slope to exceed the error band tolerance is chosen as the valid approximating slope. The slope is terminated at the first sample point which lies beyond the error band tolerance.

It is quite possible that noise may occasionally corrupt the video to make an occasional single point on the approximating slope exceed the error band. The potential for combatting this situation is inherent in the linear approximation system because the rule for abandoning a chosen slope may be changed to allow two or even three consecutive points on the chosen slope to exceed the error band tolerance before the slope is abandoned. This procedure lengthens the average slope run length, increasing compression at the expense of introducing the possibility of additional error. Another modification is possible. The new family of slopes which will determine the next linear signal may be started at either the last consecutive point in error or at the first of the consecutive points in error. While the latter probably offers more accuracy, its electronic implementation is much more complex and we chose the former case of starting the next slope family at the last of the consecutive points in error.

As each new family of slopes is generated, the previous slope (five bits) and the X component length of the slope run (five bits) are transmitted. While the recommended system has 31 slopes, computer studies have been made of systems

RE-ORDER No. 64-712
Vol I

using 63 slopes. While the recommended system truncates the maximum run-length at 32 resolution elements, the computer studies have considered compression as a function of truncation between slope run lengths of 4 and 64. Paragraph A.9 discusses these studies in detail and Appendix D furnishes block diagrams and details of system construction with comments on circuit selection.

In summary, the recommended video system is a linear approximation system with 31 slopes and with run-lengths to 32 resolution elements. It possesses an error band adjustable to ±1, ±2, and ±4/64ths of full-scale video. It should have a selectable run error tolerance of one, two, or three consecutive errors.

This system provides a wide range of trade offs between fidelity and compression and offers gross compressions as low as 1:1 and as high as 9:1, the trade offs being dependent on parameter settings and pictorial material. Net compressions on the order of 3:1 with 3.5% of full-scale rms error are indicated from the experimental study. Net compressions of 2:1 likewise appear possible with rms error reduced to 2.5% of full-scale video. This is a relatively low-error instrumentation system, and all video levels are produced so that contouring should be minimal if not nonexistent at the output.

While the resulting power, weight, and volume of the system vary widely with system requirements, the following is representative of what can be done if high-speed Motorola "MECL" integrated circuits in TO-5 packages are used. Because of the circuit speed, the power estimate is 35.56 watts. The nonstructural volume (actual parts plus correction factor) is estimated at 53.5 cubic inches. The nonstructural weight is estimated at 1.80 pounds. This system will offer a video data processing rate of approximately 625,000 picture elements per second. If the system is designed with slower speed logic, (approximately 62,500 picture elements per second), the estimated power drops to 19.56 watts. Because of linear approximation's compression capabilities, flexible trade-off capability, and its possibility for integrated circuit implementation, we recommend highly the construction of a laboratory model as the next step in this program.

RE-ORDER No. 64-712
Vol I

SECTION 3

CONCLUSIONS AND RECOMMENDATIONS

3.1 CONCLUSIONS

Eleven major conclusions resulted from this study.

1. Video data modulation systems with a net compression of 2:1 are feasible.
2. Video data modulation systems with a net compression of 3:1 are possible.
3. A net compression of 2:1 can be achieved by the relatively simple three-four Roberts system at the expense of some loss in fidelity (image noise) relative to 6-bit PCM.
4. A net compression of 2.5:1 can be achieved by use of the 2-bit Delta system, but at the expense of some loss in fidelity (image softness) relative to 6-bit PCM.
5. A net compression of 3:1 can be obtained by the somewhat more complex linear approximation system.
6. A linear approximation system rms error of approximately 3.5% of full scale video is expected at net compressions in the vicinity of 3:1 (gross compression, 5.4:1).
7. The linear approximation system is a very flexible system offering a wide range of trade offs between compression and fidelity through adjustment of system parameters.
8. Data compression performance is relatively inconclusive without concomitant fidelity performance.
9. No single adequate measure of fidelity performance is believed to exist.
10. The relative fidelity penalty paid for compression becomes progressively less as sensor signal-to-noise ratio decreases.
11. The fidelity performance of a video data modulation system ultimately must be judged by the final user (space science experimenter).

3.2 RECOMMENDATIONS

Based on the above conclusions, the recommendations below are made.

- Investigate in greater detail the possibility of the three-four Roberts system satisfying the fidelity requirements of planetary imaging experiments.
- Investigate in greater detail the possibility of the 2-bit Delta system and variations thereof satisfying the fidelity requirements of planetary imaging experiments.
- Construct a laboratory-model linear-approximation system, which has (1) variable system parameters, (2) provision for injecting controlled amounts

RE-ORDER No. 64-712
Vol I

of sensor noise, and (3) provision for simulating a variable bit error probability communication link between the linear approximator and the picture reconstructor.

- Investigate the ability of the linear approximation video modulation system to satisfy the fidelity requirements of planetary imaging while achieving the highest possible data compression.

RE-ORDER No. 64-712
Vol I

APPENDIX A

COMPRESSION TECHNIQUES

A.1 INTRODUCTION

The compression study task was to undertake the investigation of both objective and subjective compression techniques that have the potential of achieving a minimum net compression ratio of at least 2:1. The compression analysis was conducted in the following manner. A mathematical model was formulated for each compression technique and the compression was calculated from the entropy of the model. Preliminary values of compression were obtained by taking limiting values for the statistics necessary to evaluate the model. If these statistics indicate that gross average compressions greater than 2:1 could exist, then the system was simulated by taking actual statistics on EDITS; alternately, the compression technique was simulated on the ASI-210 digital computer. If the statistical model then indicated that the system satisfied the minimum gross compression of greater than 2:1, the net compression was calculated by considering sensor noise, channel noise, and synchronization.

The net compression can be given for three separate and distinct considerations. The net data compression is

$$\bar{C}_{net_d} = \frac{6m + k_1}{\frac{6m}{\bar{C}_g} + k_2}; \quad (1)$$

where m is the television resolution for a square aspect ratio, k_1 is the number of bits for line synchronization for a six bit PCM system, k_2 is the number of bits for line synchronization for the compressed system, and $\bar{C}_g$ is the average gross compression for the compression technique. Although the net data compression is useful for the purposes of this study, the net signal power compression is a better measure for the study and is

RE-ORDER No. 64-712
Vol I

$$\bar{C}_{net_s} = \frac{6m + k_1}{\left(\frac{6m}{\bar{C}_g} + k_2\right)} \Delta S/N \quad = \bar{C}_n ; \tag{2}$$

where Δ S/N is the ratio of the six bit PCM signal power to the compressed signal power to maintain an equivalent bit error probability. The final formula to be calculated is the net prime power compression which is

$$\bar{C}_{net_p} = \frac{\left(6m + k_1\right) \quad f(\Delta P)}{\left(\frac{6m}{\bar{C}_g} + k_2\right) \Delta S/N} ; \tag{3}$$

where f(ΔP) is some functional relationship between the power required to implement a six bit PCM source and the power required to implement the compression technique. This, however, requires a knowledge of the rf parameters, beyond the scope of this study; therefore, for the purposes of the study, the net signal power compression formula will be referred to as the net compression, $\bar{C}_m$.

Eight separate and distinct compression techniques were analyzed. These were previous element coding (PEC), previous-element/previous-line coding (PEPL), Roberts coding, Delta modulation, split-band coding, bit-intensity run-length coding (BIRL), area coding, and linear approximation. Each of these techniques have definite upper bound on compression due to (1) the theoretical restrictions imposed by the basic concept of the compression technique, (2) the theoretical limitations imposed by the noise inherent in the sensor, (3) limitations imposed by the amount of redundancy inherent in the actual scene, and (4) the loss of compression as a result of an increase in power due to implementation and an increase in information content of each transmitted bit.

The theoretical limitations imposed by the basic concepts of PEC, PEPL, Roberts, and Delta modulation, force these systems to have an upper bound on compression of 6:1. BIRL, linear approximation, and split-band coding have theoretical upper bound on compression that are set by the maximum allowed run-length. Area coding does not have a definite number that can be assigned to the theoretical upper

RE-ORDER No. 64-712
Vol I

bound on compression. These upper bounds were calculated under ideal, but not practical considerations, including (1) an infinite signal-to-noise ratio in the sensor, (2) 100% redundancy in the picture, (3) noiseless channel, (4) no synchronization, and (5) the encoder requires the same power as the noncoded PCM source.

In the real world, the upper bound on compression decreases quite drastically. For realistic sensor signal-to-noise ratios and average picture redundancy, the 6-bit gross compressions become 1.5:1 for PEC, 1.5:1 for PEPL, 2:1 for Roberts, 3:1 for Delta, 4:1 for split-band, 1.5:1 for BIRL, 1.5:1 for area coding, and 4:1 for linear approximation. When consideration is given to the increase in signal required to reduce the effects of the noisy channel and the synchronization problems, the only techniques that appear to satisy a minimum net compression of 2:1 are Roberts, Delta, split-band and linear approximation. To determine the recommended compression techniques it becomes necessary to consider additional factors. These are, the subjective evaluation of the displayed image and the size, weight, power, and reliability considerations of the encoder. If the subjective evaluation is relative to 6-bit quantization, the only coding technique that may meet the JPL study goals is the linear approximation coding technique. In the following sections of this appendix the eight coding techniques are discussed in further detail to support the above discussion.

A.2 PREVIOUS ELEMENT CODING

Previous element coding (PEC) utilizes adjacent element correlation in a frame along with efficient coding to achieve compression. Assuming that the video is quantized and converted to fixed-length digital words, the system concept can be described as follows. Successive elements are compared and, if they are alike, a single bit called PE is transmitted; however, if they are not alike, the quantized value is encoded and digitally transmitted.

The conventional PCM source has an entropy of N bits/element where $N = \log_2 n$, n being the number of brightness bits. Compression, it is believed, will reduce the PCM source to a new source with a coded entropy, H_c, where $H_c \leq N$. In general, the entropy is given by

RE-ORDER No. 64-712
Vol I

$$H = -\sum_{i=1}^{n} p(i) \log_2 p(i) \tag{4}$$

for p(i), which is the probability of obtaining brightness level, i. However, for a decodeable set of words (such as a Huffman[1] code set) the entropy is the coded entropy, H_c, where

$$H_c = \sum_{i=1}^{n} p(i)\, L(i) \tag{5}$$

for p(i), the probability of obtaining event i, and L(i), the number of bits to encode level i. Hence, the gross compression for PEC is given by

$$\bar{C}_g = \frac{N}{H_c} = \frac{N}{\sum_{i=1}^{n} p(i)\, L(i)} . \tag{6}$$

PEC is an information preserving coding technique inasmuch as there is a one-to-one relationship (under noiseless channel considerations) between the quantized value after the sensor at the transmitter and the quantized value at the receiver before the display device. For purposes of the video data modulation study it was specified that compression was to be re. .ive to a quantized six-bit video signal. Hence, it would be possible to apply the PEC concept to the six-bit source and possibly achieve compression. For the six-bit quantized data, and assuming that the (1-PE) levels have a flat distribution, the gross average compression is given by

$$\bar{C}_{PEC} = \frac{6}{7 - 6PE} . \tag{7}$$

The flat coding assumption on the (1-PE) levels was used for simplicity since statistics are dependent on the picture-content. In actuality, the optimum code would vary from picture to picture, thus requiring an adaptive system. If the encoder in the spacecraft is to be nonadaptive, a reasonable compromise between compression,

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[1] Huffman, T. A., A Method of Construction of Minimum Redundancy Codes, Communications Theory, Jackson, Editor, Butterworth Scientific Publications, London, 1963.

RE-ORDER No. 64-71.2
Vol I

picture content, error effects, and implementation can be achieved by the flat coding assumption. The compression under this assumption is dependent only upon PE, which was measured on EDITS. Table A. 2-1 gives PE for various scenes processed on EDITS at various signal-to-noise ratios for six-bit quantization. The actual photographs that were obtained are given in Section A. 2P. The gross compression for flat coding is tabulated in Table A. 2-2. Although these are the gross compression figures, the final compression ratio must consider noise effects, synchronization, and implementation, the results of Table A. 2-2 show that the PEC compression technique does not satisfy a 2:1 gross compression. Therefore, since the net compression is less than gross, the PEC compression technique for six-bit data was eliminated as a possible compression system for the JPL study.

TABLE A. 2-1

PE AS A FUNCTION OF SENSOR SIGNAL-TO-NOISE
FOR SIX-BIT QUANTIZATION AND 256 RESOLUTION

Photo	40 db	30 db	20 db
Canadian Arctic	0.246	0.126	0.047
Canadian Arctic a	0.186	0.111	0.046
Canadian Arctic b	0.226	0.121	0.045
Canadian Arctic c	0.202	0.116	0.046
JPL Lunar	0.231	0.128	0.047
JPL Lunar a	0.279	0.137	0.047
JPL Lunar b	0.296	0.137	0.047
JPL Lunar c	0.264	0.137	0.046

Although PEC was eliminated for the recommended system for six-bit encoding, the PEC coding technique could be applied to four-bit-quantized data. Although subjectively the four-bit and six-bit pictures are not equivalent, it is still possible to compress the four-bit data. Since the study requirement was to compare the systems to six-bit data, the compression for the PEC technique applied to four-bit data is given by

$$\bar{C}_{PEC_4} = \frac{6}{5 - 4PE}, \qquad (8)$$

40

RE-ORDER No. 64-712
Vol I

TABLE A. 2-2

GROSS AVERAGE PEC COMPRESSION AS A FUNCTION OF SENSOR SIGNAL-TO-NOISE FOR SIX-BIT QUANTIZATION AND 256 RESOLUTION

Photo	40 db	30 db	20 db
Canadian Arctic	1.22	0.96*	0.89*
Canadian Arctic a	1.02	0.95	0.89
Canadian Arctic b	1.06	0.96	0.89
Canadian Arctic c	1.04	0.95	0.89
JPL Lunar	1.07	0.96	0.89
JPL Lunar a	1.12	0.97	0.89
JPL Lunar b	1.15	0.97	0.89
JPL Lunar c	1.11	0.97	0.89

Table A. 2-4 gives gross average four-bit PEC compression as a function of the sensor signal-to-noise ratio for the various scenes calculated from equation (5) and Table A. 2-3. It is apparent that gross average compressions larger than two-to-one are possible for this approach. It now becomes necessary to calculate the net compression which considers synchronization and the effects of channel noise. The net compression for this case is

$$\bar{C}_{PEC_{4N}} = \frac{(6m + s)}{\Delta S/N\left(\frac{6}{\bar{C}} m + q\right)} . \qquad (9)$$

TABLE A. 2-3

PE AS A FUNCTION OF SENSOR SIGNAL-TO-NOISE FOR FOUR-BIT QUANTIZATION AND 256 RESOLUTION

Photo	40 db	30 db	20 db
Canadian Arctic	0.616	0.433	0.184
Canadian Arctic a	0.540	0.400	0.175
Canadian Arctic b	0.608	0.433	0.180
Canadian Arctic c	0.549	0.403	0.179
JPL Lunar	0.613	0.445	0.186
JPL Lunar a	0.696	0.478	0.179
JPL Lunar b	0.701	0.457	0.180
JPL Lunar c	0.665	0.457	0.180

*Gross average compression less than 1 results in a bandwidth expansion equal to $1/\bar{C}$

RE-ORDER No. 64-712
Vol I

TABLE A.2-4

GROSS AVERAGE COMPRESSION FOR FOUR-BIT PEC RELATIVE TO SIX BITS OF 256 RESOLUTION AS A FUNCTION OF SENSOR SIGNAL-TO-NOISE

Photo	40 db	30 db	20 db
Canadian Arctic	2.36	1.84	1.42
Canadian Arctic a	2.04	1.76	1.40
Canadian Arctic b	2.34	1.84	1.40
Canadian Arctic c	2.14	1.77	1.40
JPL Lunar	2.36	1.86	1.41
JPL Lunar a	2.70	1.94	1.40
JPL Lunar b	2.72	1.89	1.40
JPL Lunar c	2.56	1.89	1.40

TABLE A.2-5

NET AVERAGE COMPRESSION FOR FOUR-BIT PEC RELATIVE TO SIX BITS AT 256 RESOLUTION AS A FUNCTION OF SENSOR SIGNAL-TO-NOISE

Photo	40 db	30 db	20 db
Canadian Arctic	1.66	1.30	1.00
Canadian Arctic a	1.43	1.24	0.99
Canadian Arctic b	1.65	1.30	0.99
Canadian Arctic c	1.50	1.25	0.99
JPL Lunar	1.66	1.31	1.00
JPL Lunar a	1.90	1.37	0.99
JPL Lunar b	1.98	1.33	0.99
JPL Lunar c	1.80	1.33	0.99

In equation (9), m is the number of elements along a line, s the number of bits for line sync for six-bit PCM data, $\bar{C}_{PEC_4}$ is the gross PEC four-bit compression, q is the number of bits for line sync for the four-bit PEC system, and Δ S/N is the power increase for the compressed data to have the equivalent bit error probability.

From experimental studies, the compressed system will require a bit error probability of 10^{-4} compared to the uncompressed bit error probability of 10^{-3}. Therefore, assuming coherent-phase-shift keying modulation, the Δ S/N is 1.42.

46

RE-ORDER No. 64-712
Vol I

From the synchronization study, s is 30 bits and q is 50 bits. Therefore, for the four-bit PEC system, the net compression is

$$\bar{C}_{PEC_{4N}} = \frac{2161}{1.42\ (2590-2048)} = \frac{1}{(1.199-0.947)} \text{ for 512 resolution.}$$

This result can now be applied to Table A. 2-4 and the net compression for the processed scene is given by Table A. 2-5.

It appears that the four-bit PEC system will not satisfy the JPL net compression requirement . At the higher signal-to-noise ratios, both scenes and the sub-scenes may satisfy the JPL requirements if the PEQ concept is applied. If the space scientist agrees that four-bit quantization is sufficient, and if the additional compression is picked up with PEQ, then the four-bit PEC may satisfy the minimum 2:1 net compression

A. 3 LINE-TO-LINE CODING

A. 3. 1 Introduction

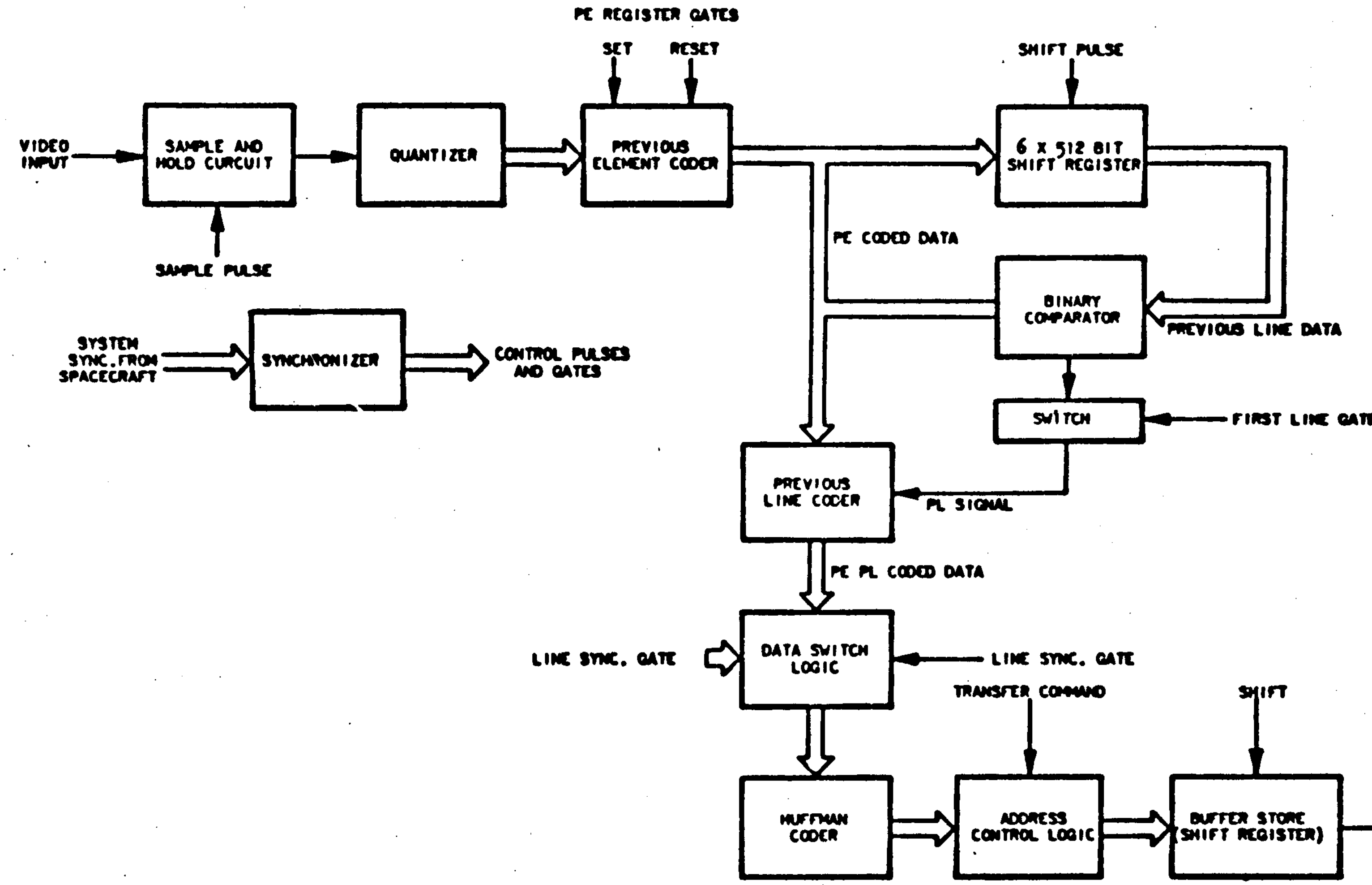
The PEC compression system efficiently utilizes the correlation between successive elements along a scanning line. This technique was extended in the vertical direction to take advantage of the correlation between adjacent elements between successive lines. There are two approaches open to utilize both the horizontal and vertical correlation.

A. 3. 2 PEPL Coding

PEPL coding is one technique that compares adjacent elements in successive lines and encodes adjacent elements of the same intensity with a PL signal if they are not PE elements. A block diagram of this system is given in Figure A. 3-1. The alternate approach is a PL' PE' coding technique that encodes all elements that are alike between successive lines with a PL' signal.

The crossover points where the compression for PEC equals the compression for PEPL will be determined: that is, given the value of PE, PL is determined so that $\bar{C}_{PE} = \bar{C}_{PEPL}$. For this condition

$$\frac{N}{N+1-N\,PE} = \frac{N}{N+2-(N+1)\,PE-NPL}; \qquad (10)$$



RE-ORDER No. 64-712
Vol I

FIGURE A. 3-1

PEPL MODULATOR

RE-ORDER No. 64-712
Vol I

where N is the number of intensity levels expressed in bits, and assuming a flat distribution for the (1-PE) and (1-PE-PL) levels. Solving for PL as a function of PE,

$$PL = \frac{1-PE}{N} . \qquad (11)$$

Figure A. 3-2 plots the above relationship for N=6 and four bits.

Although this set of curves sets the lower limit on PL, consider the value of PL required so that

$$\bar{C}_{PEPL} = 2\bar{C}_{PE} \qquad (12)$$

Substituting in (1),

$$\frac{N}{N+1-N\,PE} = \frac{N}{2\left[N+2-(N+1)\,PE-N\,PL\right]} \qquad (13)$$

Solving for PL

$$PL = \frac{N+3-(N+2)\,PE}{2N} . \qquad (14)$$

This relationship is also plotted on Figure A. 3-2. From the curve in order to increase the compression by a factor of two over PEC, PE for six bits must be less than 0. 5. If

$$\bar{C}_{PEPL} = 2\,\bar{C}_{PE} ,$$

with PE = 0. 5, PL must equal 0. 5 to satisfy this condition, and this is of course impossible. In fact, a minimum of 90% of a scene must be either PE or PL to double the compression.

Several pictures were processed on EDITS and the appropriate statistics were measured to determine the efficiency of a coding technique that takes advantage of this horizontal and vertical correlation and the results are compared to the PE and PL' coding compression values. The statistics measured were: (1) PE, the probability that an element along a scanning line has the same intensity as its neighboring element; (2) PL, the probability that an element has the same intensity as the adjacent element in the neighboring line, but not a PE element; (3) PE', the probability that an element has the same intensity as its neighboring element but not the same intensity as the neighboring element in the adjacent line; and (4) PL', the

47

RE-ORDER No. 64-712
Vol I

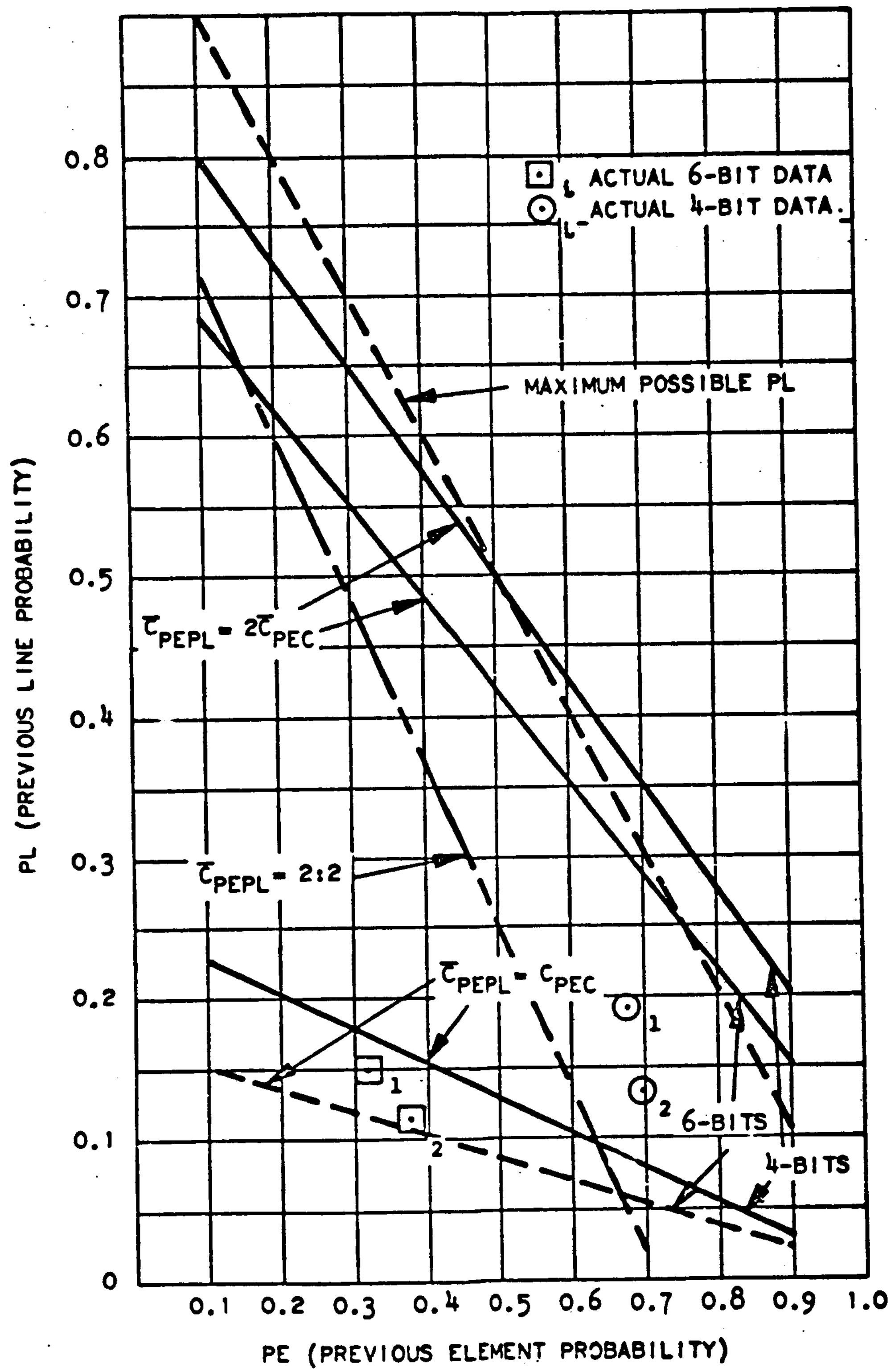


FIGURE A.3-2

PL AS A FUNCTION OF PE FOR THE CONDITIONS
$\bar{C}_{PEC} = \bar{C}_{PEPL}$ AND $\bar{C}_{PEPL} = 2\,\bar{C}_{PEC}$

07314

50

probability an element has the same intensity as the adjacent element in the adjacent line. From these probabilities, it is possible to obtain the gross average compression. These compressions are (1) for the PEC system the compression is

$$\bar{C}_{PE} = \frac{N}{N + 1 - N\,PE}, \tag{15}$$

(2) for the PL' system the compression is

$$\bar{C}_{PL'} = \frac{N}{N + 1 - N\,PL'}, \tag{16}$$

(3) for the PEPL system the compression is

$$\bar{C}_{PEPL} = \frac{N}{N + 2 - (N + 1)\,PE - N\,PL}, \tag{17}$$

and (4) for the PE'PL' system the compression is

$$\bar{C}_{PE'PL'} = \frac{N}{N + 2 - (N + 1)\,PL' - N\,PE'}. \tag{18}$$

The line-to-line compressions are valid under the assumptions that (1) the most probable event (PE or PL') is encoded with a 1; (2) the next most probable event (PE or PL) is encoded with a 01, and (3) the remaining n events have a flat distribution and are encoded with N + 2 bits. To determine the efficiency of these coding techniques relative to the optimum Huffman coding technique, the 1-PE-PL statistics were taken. The optimum compression for the PEPL system is

$$\bar{C}_{Opt.} = \frac{N}{\sum_{i=1}^{n} p(i) \, \mathrm{Log}_2 \, p(i)}. \tag{19}$$

Several pictures were processed on EDITS and the results are summarized in Tables A. 3-1, A. 3-2, and A. 3-3, and graphically in Figure A. 3-3.

These results seem to indicate that several concepts that have been stated in the past are not valid; that is, given a compression of X:1 in the horizontal direction and a compression of Y:1 in the vertical direction, then combing the two will result in a compression of XY:1. This is not valid since

$$\bar{C}_{PE} \cdot \bar{C}_{PL'} \neq \bar{C}_{PEPL} \text{ or } \bar{C}_{PE'PL}.$$

RE-ORDER No. 64-712
Vol I

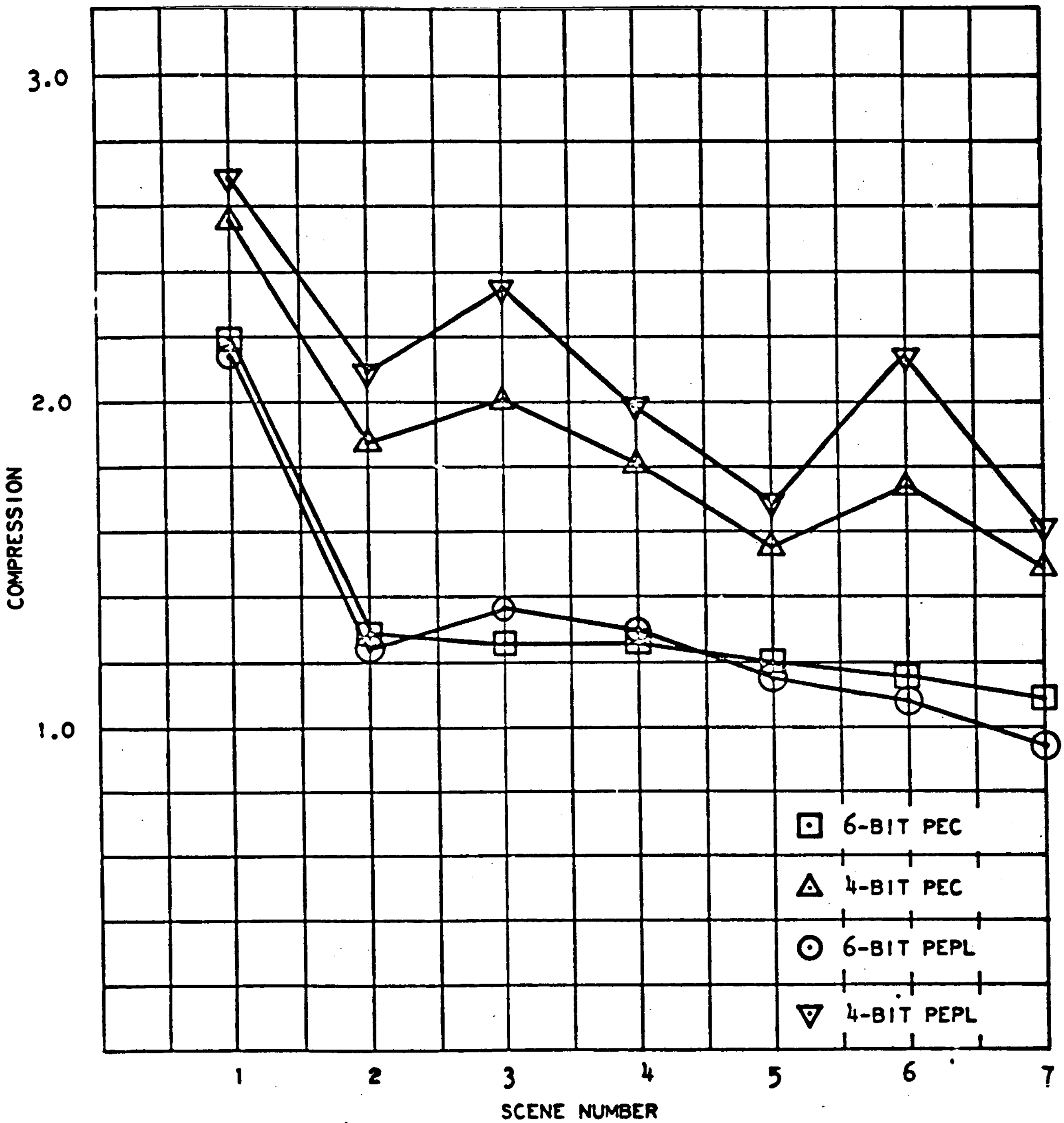


FIGURE A. 3-3

COMPRESSION FOR VARIOUS SCENES
PROCESSED ON EDITS AT 256 RESOLUTION

07314

In general, the compression is less; that is,

$$\overline{\bar{C}_{PE} \cdot \bar{C}_{PL}} = 1.85 \text{ and } \bar{C}_{PEPL} = 1.35$$

for the seven pictures processed.

The figures obtained by combining PE and PL' indicate that the amount of redundancy that could be eliminated in a picture quantized to six bits varies from a minimum of 34% to a maximum of 75% with the average redundancy for the seven pictures being 49%. This indicates that at six bits on the average, one would expect a 2:1 compression. From the published literature, the amount of redundancy is from 90% to 95%; therefore, approximately 40% of the redundancy is eliminated by sensor noise.

It appears that there is no clear analytical or experimental relationship between PE and PL; that is, it is not possible to estimate the value of PEPL compression from only the value of PE. This can be seem from Figure A. 3-4, which plots PE and PL for the several scenes processed on EDITS.

In the area of noise effects, the problems enlarge considerably relative to the error effects in PEC. For example, the near optimum code is now composed of three word lengths.

RE-ORDER No. 64-712
Vol I

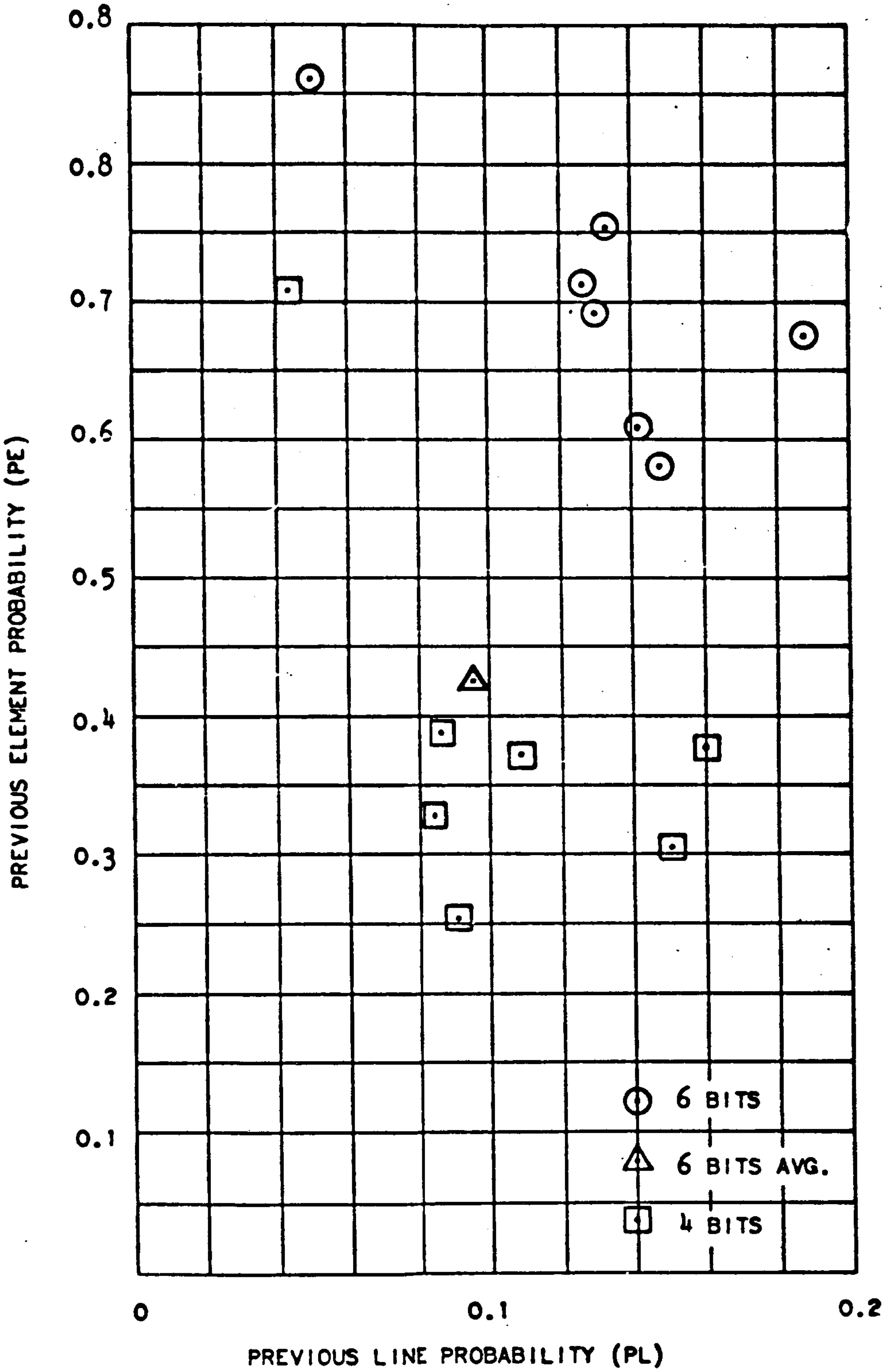


FIGURE A. 3-4

PE AND PL PROBABILITIES FOR SEVERAL SCENES
PROCESSED ON EDITS AT 256 RESOLUTION

07314

RE-ORDER No. 64-712
Vol I

TABLE A. 3-1

PE, PL, PE', PL', AND (1-PE-PL) PROBABILITIES
FROM EDITS AT 256 RESOLUTION

PICTURE	QUANTIZATION	PE	PL	PL'	PE'
Moon	6 bits	0.305178	0.150453	0.247194	0.134889
Moon	4 bits	0.674443	0.187684	0.635228	0.185487
Girl	6 bits	0.373080	0.109254	0.323794	0.174562
Girl	4 bits	0.694280	0.129701	0.629887	0.187227
Cooper	6 bits	0.251467	0.091553	0.340884	0.127717
Cooper	4 bits	0.579838	0.148011	0.582280	0.202028
Celescope	6 bits	0.707708	0.043259	0.750738	0.051575
Celescope	4 bits	0.859839	0.050660	0.864112	0.048218
Apollo	6 bits	0.387576	0.086823	0.437930	0.130311
Apollo	4 bits	0.713811	0.126344	0.668340	0.167507
Star	6 bits	0.375216	0.160066	0.344088	0.186922
Star	4 bits	0.753332	0.133058	0.724035	0.161744
JPL Moon	6 bits	0.329592	0.085450	0.286867	0.126191
JPL Moon	4 bits	0.607520	0.141280	0.508427	0.160981
JPL Moon	(1-PE-PL)(1)	0.003920	(1-PE-PL)(9)		0.01832
JPL Moon	(1-PE-PL)(2)	0.01520	(1-PE-PL)(10)		0.02696
JPL Moon	(1-PE-PL)(3)	0.010880	(1-PE-PL)(11)		0.02464
JPL Moon	(1-PE-PL)(4)	0.01360	(1-PE-PL)(12)		0.0224
JPL Moon	(1-PE-PL)(5)	0.1560	(1-PE-PL)(13)		0.02696
JPL Moon	(1-PE-PL)(6)	0.017584	(1-PE-PL)(14)		0.02112
JPL Moon	(1-PE-PL)(7)	0.01688	(1-PE-PL)(15)		0.00026
JPL Moon	(1-PE-PL)(8)	0.01856	(1-PE-PL)(16)		0.00069

55

RE-ORDER No. 64-712
Vol I

TABLE A. 3-2

COMPRESSIONS ACHIEVABLE FOR PICTURES
PROCESSED ON EDITS AT 256 RESOLUTION

PICTURE	SCENE NO.	QUANTIZATION (BITS)	$\bar{C}_{PE}$	$\bar{C}_{PEPL}$	$\bar{C}_{PL}$	$\bar{C}_{PL',PE'}$
Moon	6	6	1.16	1.21	1.09	1.10
		4	1.74	2.13	1.63	1.92
Girl	4	6	1.26	1.27	1.19	1.28
		4	1.80	1.99	1.61	1.90
Cooper	7	6	1.09	1.05	1.21	1.24
		4	1.49	1.59	1.50	1.75
Celescope	1	6	2.18	2.15	2.40	2.46
		4	2.56	2.67	2.59	2.69
Apollo	2	6	1.28	1.26	1.37	1.44
		4	1.87	2.08	1.72	1.99
Star	3	6	1.26	1.36	1.22	1.34
		4	2.01	2.35	1.90	2.31
JPL Moon	5	6	1.19	1.16	1.14	1.15
		4	1.56	1.67	1.35	1.42
Average		6	1.35	1.35	1.37	1.43
		4	1.86	2.07	1.76	2.00
$C_{Opt.}$	(JPL Moon)	4	1.75:1			

56

RE-ORDER No. 64-712
Vol I

TABLE A. 3-3

AMOUNT OF REDUNDANCY IN A PICTURE THAT IS EITHER LIKE THE ADJACENT ELEMENT HORIZONTALLY OR VERTICALLY

PICTURE	QUANTIZATION	PE+PL	PL'+PE'
Moon	6 bits	0.455631	0.382083
Moon	4 bits	0.862127	0.920715
Girl	6 bits	0.482334	0.498356
Girl	4 bits	0.823981	0.817114
Cooper	6 bits	0.343020	0.468601
Cooper	4 bits	0.727849	0.784308
Celescope	6 bits	0.750967	0.802313
Celescope	4 bits	0.910499	0.912330
Apollo	6 bits	0.474399	0.568241
Apollo	4 bits	0.840155	0.830847
Star	6 bits	0.535282	0.531010
Star	4 bits	0.886390	0.886379
JPL Moon	6 bits	0.415042	0.413058
JPL Moon	4 bits	0.748800	0.669408
Average	6 bits	0.493811	0.523380
Average	4 bits	0.828543	0.817300

RE-ORDER No. 64-712
Vol I

Some preliminary and basic error results can be obtained. Consider first the probability of obtaining a zero or one. This is given by

$$P(1) = PE + PL/2 + (1\text{-}PE\text{-}PL)\ (N)/2\ (N + 2), \text{ and} \tag{20}$$

$$P(0) = PL/2 + (1\text{-}PE\text{-}PL)\ (N + 4)/2\ (N + 2). \tag{21}$$

These equations are valid for PEPL coding and assuming that the (1-PE-PL) levels have a flat distribution. For six-bit quantization, equations (11) and (12) reduce to

$$P(1) = 3/8 + 5/8\ PE + 1/8\ PL, \text{ and} \tag{22}$$

$$P(0) = 5/8 - 5/8\ PE - 1/8\ PL. \tag{23}$$

For seven pictures processed on EDITS, the average value of $\overline{PE} = 0.426$ and $\overline{PL} = 0.096$. Therefore, $P(1) = 0.653$ and $P(0) = 0.347$ from which it can be stated that $P(1) \doteq 2P(0)$. For these pictures some of the probabilities of errors can be estimated.

The probability of a level change is

$$P(1\text{-}PE\text{-}PL) \longrightarrow \text{level change} = 0.75\ (0.478) = 0.359,$$

and there is no shortening of the line.

$$P_{2PE \longrightarrow PL} = \sum_{i=2}^{\infty} \frac{(i - 1)\ (PE)^i}{i} = PE/(1\text{-}PE) - \log_e\left(\frac{1}{1\text{-}PE}\right) = 0.188.$$

When this occurs, the line will be shortened by one element.

The probability that a PL code will be decoded as 2 PE's is

$$P_{PL \longrightarrow 2PE} = 0.096/2 - 0.048.$$

The line will be lengthened by one element. If P_P is defined as the predicatable error, $P_P = 0.595$ or approximately 40% of the errors require additional detailed examination. If the PEPL system is in state S, the synchronization state, the probability of remaining in S is the probability that an error occurs in the last N bits of a (1-PE-PL) word or $P_1 = N(1\text{-}PE\text{-}PL)/(N + 2)$. This type of error is quite similar to the PCM error; however, it may propagate for several

58

elements. But on the average the error will only propagate as long as the average run-length. These results are rather discouraging and it was decided to investigate an alternate and more common approach to line-to-line coding.

PE-ΔL-RL Coding

If there exists redundancy between successive lines, how can one effectively utilize this information? One possible manner would be to have a (PE - ΔL - RL) previous element delta-line, run-length coded system. The system will operate as follows. Every kth line will be PE encoded as used as a reference point. Relative to the reference line, the difference between successive lines will be calculated. If this difference, ΔL, the binary sum of the total differences of the line, is less than T, an arbitrary assigned value to be determined, a code is transmitted indicating that the two successive lines are equal. However, if ΔL is more than T, then the difference is PE encoded and transmitted. The compression when two successive lines satisfy the threshold requirement is

$$\frac{Nm}{N+4} : 1.$$

For six-bit quantization and 256 resolution, the compression for that line will be approximately 150:1. In general the compression will be given by

$$\bar{C} = \frac{Nm^2}{(N+4)\,L + P + (N+2)\,(m^2 - mL - P)} \; ; \tag{24}$$

where N = number of quantizing bits,
m = horizontal resolution,
L = number of identical lines, and
P = difference elements that are PE .

An estimate on the compression can be made by assuming that the redundancy is sensor limited to 50%. With 256 resolution, six-bit quantization, 50% of the remaining elements are PE, and 50% of the lines are redundant, the compression is only 2.5:1.

Conclusions

Based on the statistics obtained from EDITS, the effects of error and the increase

RE-ORDER No. 64-712
Vol I

in size, weight, and power, it is evident that the PEPL coding technique will not satisfy the requirements imposed by JPL of a minimum compression of 2:1. The average PEPL compression for the seven pictures was 1.35:1. A technique is posed that will attempt to take advantage of the line-to-line redundancy; however, statistics are not available, and intuitively it appears that the technique will not satisfy the compression requirements.

A.4 AREA CODING

A.4.1 Area Scanning

Several approaches on utilizing area correlation have been investigated by Schaphorst and Remm[1], Haagin[2], Altemus and Schaphorst[3], Posner[4], and Wichmann[5]. It is not the purpose of this section to discuss the generalized area coding approach, but to consider one specific coding technique that utilizes both the horizontal and vertical correlation and to propose an additional technique for area encoding.

It has been previously established that the statistical correlation is about the same in the vertical and horizontal directions; that is, the previous element probability along a line is about the same as the previous line probability obtained by scanning the picture vertically. Therefore, it may be feasible to use this correlation in determining a near optimum size for the area to be scanned.

Consider an average run of L_h elements in the horizontal direction. Using the PE concept, this average run requires $(N + 1) + (L_h - 1)$ bits under the previously

References

(1) Schaphorst, R. A., and Remm, R. L., An Image Coding System Which Reduces Redundancy in Two Dimensions, 1962 National Conference Proceedings on Aerospace Electronics.

(2) Haagen, W. C., Video Bandwidth Compression, Ohio State University Research Foundation Report 1222-7, 1 February 1962.

(3) Altemus, W. C., and Shaphorst, Data Conversion Equipment, AD 275 514, 18 May 1962.

(4) Posner, E. C., Pseudo-Random Area Coding for Television Using Modular Geometric Shapes, JPL Space Summary No. 37-19, Vol. 10, 28 February 1963.

(5) Wichmann, T., Optimum Encoding of Pictorial Data, Ninth National Communications Symposium, October 1963.

RE-ORDER No. 64-712
Vol I

stated flat coding assumption. The average number of bits per run $\bar{L}$ is

$$\bar{L} = \frac{N + L_h}{L_h} . \qquad (25)$$

With the modified Huffman coding assignment, the average number of bits per element is

$$\bar{L} = N + 1 - N\,PE_h . \qquad (26)$$

where PE_h is the previous element probability in the horizontal scanning direction. If equations (1) and (2) are equated, the average run length is

$$L_h = \frac{1}{1 - PE_h} . \qquad (27)$$

Similarly, in the vertical direction, the average run length can be found and is given by

$$L_v = \frac{1}{1 - PE_v} . \qquad (28)$$

Since experimentally it can be shown that $PE_h \doteq PE_v$, the average area size is

$$\bar{A} = \frac{1}{(1-PE)^2} . \qquad (29)$$

Given PE, it would be possible to utilize the horizontal and vertical correlation in a picture by scanning the picture in an area of size determined by the average area size. For example, eleven pictures were processed on EDITS which were quantized to six bits and the average PE was 0.5; therefore, the average area $\bar{A} = 4$ elements. Hence, if the picture were broken into a number of four element areas and efficiently encoded, compression could result.

4.4.2 Previous Element Area Scan Coding (PEAS)

Although the above area coding technique is considered the conventional approach to area coding, it may be possible to increase the efficiency of a PEC system by scanning the picture in a raster which sequentially covers the picture in the average size increments.

By scanning the picture in the elemental area approach, the expected results

RE-ORDER No. 64-712
Vol I

are: (1) PE should necessarily increase over the conventional line scanning PE; (2) the scanning technique utilizes line-to-line correlation, thus providing a simplified version of the PEPL system; and (3) the scanning technique will provide a mathematical model for the area coding technique.

Several hypothetical matrices of elements where $PE_h \doteq PE_v$ were composed and the PEA technique was applied to them. The manual simulation indicates that the elemental area scanning technique increased PE in the PEA system such that the net advantage in compression was greater than that which could be achieved by PEPL coding the matrices. The PEA technique was simulated on EDITS.

Implementation of the elemental area scan was achieved by scanning small subrasters instead of horizontal lines. To accomplish the subraster scan, the digital sweep circuits of EDITS were rewired as shown in Figure A. 4-1. The subscan counters can be any portion of the total; however, implementation is easy if they are powers of two. Addition of the two counter outputs is accomplished in the D/A ladder network. The subscan configuration for one-bit subscan counters is shown in Figure A. 4-2. The subscan size is $2^1 = 2$ bits square. If the subscan counters are each two stages, the size becomes $2^2 = 4$ bits square. The subscan raster as shown in Figure A. 4-3 scans vertically and then steps horizontally. If the vertical and horizontal subscan counters are reversed so that the horizontal is stepped before the vertical, the raster will take the form shown in Figure A. 4-4.

The elemental area scan technique was tried on EDITS and the results appeared to be discouraging. The PE for the PEA system did not increase significantly with a two-element square subscan. Examination of the video revealed that an alternation of video pulse heights occurred. It was determined that this was a result of the particular scanning technique imposed. Several possible solutions were tried, but none adequately solved the problem. This unfortunate turn coupled with the results of the PEPL analysis indicates that unless sensor noise is eliminated, a PEA coding technique does not offer a sufficient value of gross compression to satisfy the minimum JPL requirements. For six-bit quantization, an average of only 50% of the elements are redundant; therefore, directly utilizing statistical redundancy could result in an upper bound on compression of 2:1. When synchronization, effects of channel noise, and implementation problems are considered, it is evident that a six bit PEA system will not satisfy the JPL compression requirement.

2

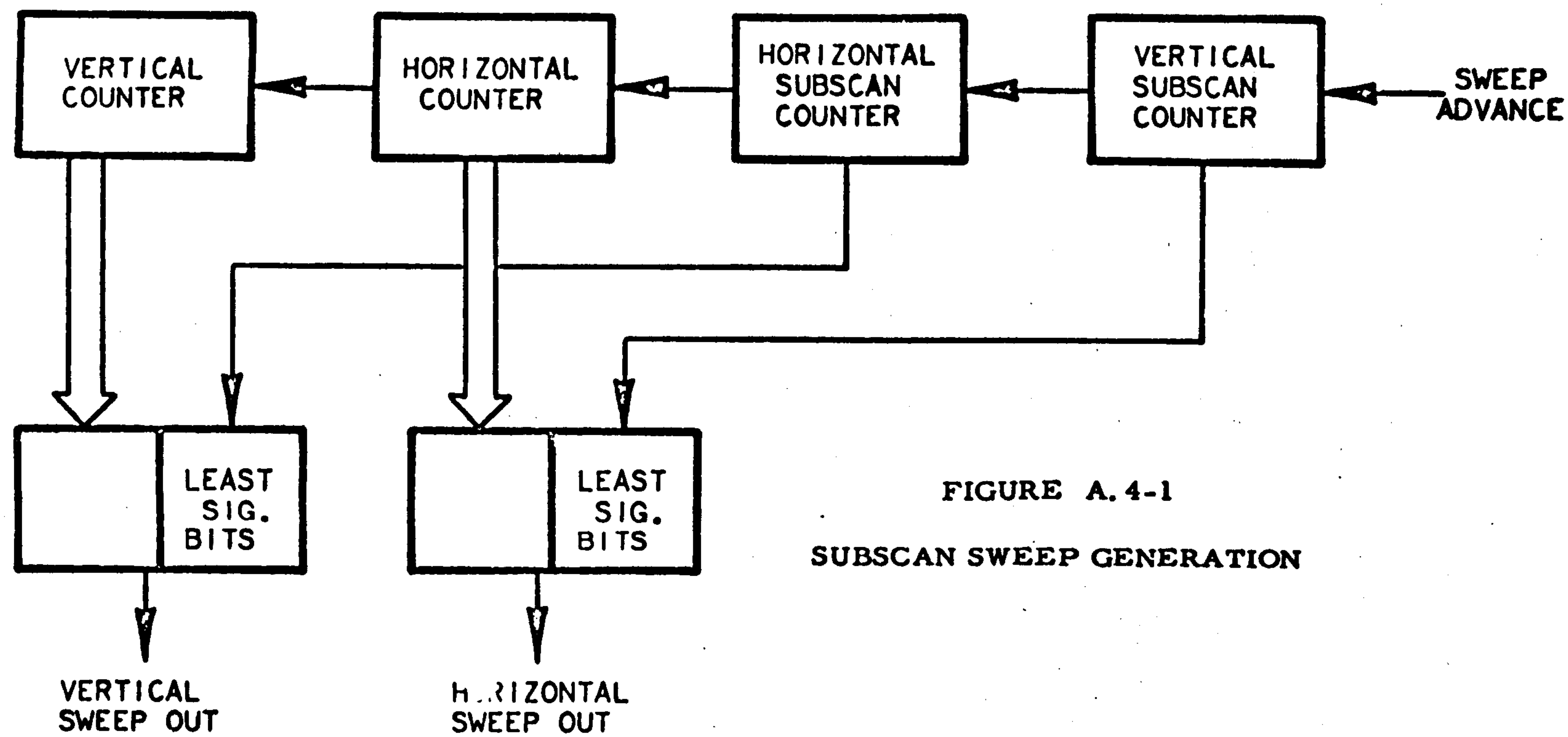


FIGURE A. 4-1

SUBSCAN SWEEP GENERATION

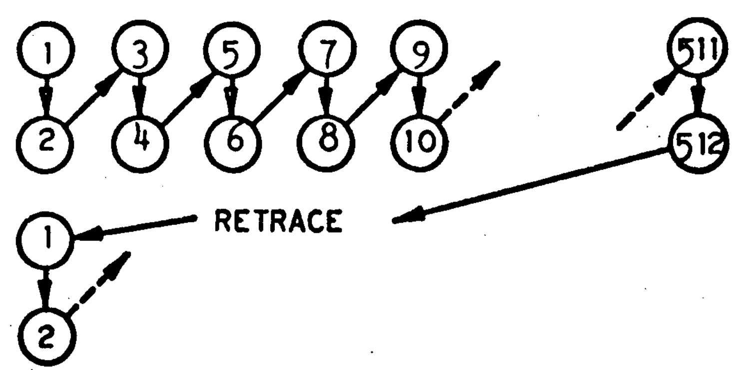


FIGURE A. 4-2

SUBSCAN SWEEP PATTERN

07314

RE-ORDER No. 64-712
Vol I

62

RE-ORDER No. 64-712
Vol I

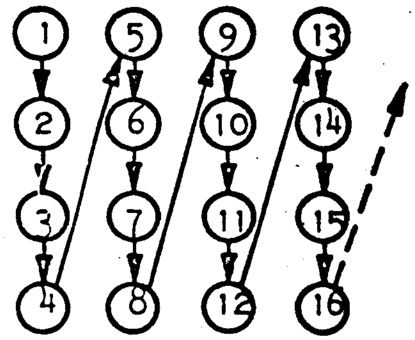


FIGURE A. 4-3

2^2 SUBSCAN PATTERN

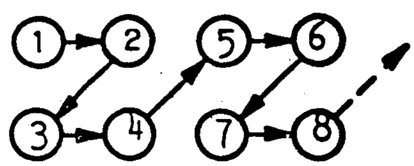


FIGURE A. 4-4

HORIZONTAL/VERTICAL PATTERN

07314

RE-ORDER No. 64-712
Vol I

A.5 BIT INTERLACE RUN LENGTH COMPRESSION

A.5.1 Concept

A picture may be represented by one plane of six-bit binary numbers. Alternatively, one can think of the six-bit binary numbers as standing on end, upon their base plane and then being sliced horizontally by six parallel planes. After this is accomplished, conceptually, there exists six planes each containing only one bit at each picture element position. There is, possibly, an advantage in having six one-bit planes per picture instead of one six-bit plane.

In run-length coding, the end of a run occurs when the value of the function changes from the value maintained during the run. For a six-bit PCM system, the sixth (low order) bit is largely of random value and rapidly fluctuating, especially at high signal-to-noise ratios. This indicates that long runs of six-bit PCM words are quite improbable, even for a constant level input. However, the fifth bit and higher order bits are less susceptible to random changes. Even so, their lack of rapid change is not available for exploitation as straight run-length encoding. To exploit the suspected redundancy of the higher order bits, the bits (or bit planes) must somehow be separated and encoded separately. This concept is the basis for bit-interlace-run-length (BIRL) coding.

In BIRL, each separate bit plane is run-length encoded. At the receiver the bit planes are reconstituted from the reduced redundancy signal. The decoded bit planes would be interlaced together in the display with the proper analog weighting for each bit plane to reconstruct the original half-tone PCM picture from which the original bit planes were obtained.

EDITS (see Appendix E.1) is sufficiently flexible to secure the signal from each bit decision in the analog-to-digital converter. This bit-decision signal is displayed to form a one-bit picture which may be photographed. Six such pictures constitute, in principle, the information in the original six-bit PCM picture. Indeed, a composite half-tone picture has been reconstructed by the expedient of multiple exposure of the recording film giving the low-order bit a one-frame time exposure, the second order bit a two-frame exposure, the third order bit a four-frame exposure, etc.

A. 5. 2 Experimental Results

The basic plan for evaluating BIRL potential is to secure six individual bit-plane pictures of a scene and analyze each bit plane for its run-length compression potential. The analysis can be carried out in three ways. (1) The bit-plane second-order statistics can be converted to 1-1 and 0-0 transition probabilities which in turn are inserted into a run length compression mathematical model equation to predict average compression. (2) Actual run-length statistics can be taken on each bit plane and compression can actually be measured. (3) The bit-plane pictures are laid out left-to-right, high-to-low order bits and a visual estimation of bit plane organization (and hence, redundancy) is made. All three methods were used to analyze BIRL potentiality. The subjective picture analysis is interesting and will be dealt with at some length, as will the analytical methods.

A lunar scene photograph (not the one furnished by JPL) was analyzed on a bit plane basis. Since it is well-known that effective picture statistics are a function of both the true picture content and the noise statistics, the picture was analyzed at six different signal-to-noise ratios:

- Maximum signal-to-noise ratio (>40 db)
- 35 db
- 30 db
- 25 db
- 20 db
- 15.4 db

Second order statistics were taken on all six-bit planes for a total of 36 signal-to-noise-ratio-bit-plane combinations. The raw data second order event counts were reduced to two of the four possible transition probabilities, specifically, the probability that a one is followed by a one, and the probability that a zero is followed by a zero. These transition probabilities can be used in connection with various run-length formulae and computer algorithms for predicting the compression inherent in each bit plane. A FORTRAN I computer program was written for the truncated transition probabilistic compression model. The program was run on the ASI-210 computer.

RE-ORDER No. 64-712
Vol I

The bit planes 1 through 6 were chosen for display with bit plane 1 indicating the high-order bit and bit plane 6 the low-order bit in the six-bit word describing the intensity of each picture element.

When the pictures are laid out as shown in Figure A. 5. 2-1, the following subjective appearances were noticed: proceeding from bit plane 1 to 6, a decreasing amount of "organization" in the pictures was observed at all signal-to-noise ratios. Even at the relatively high signal-to-noise ratios of $>$40 db and 35 db, bit planes 5 and 6 both present a homogeneous, salt-and-pepper appearance with occurrences and nonoccurrences (blacks and whites) being approximately equal. At the lower signal-to-noise ratios, 30 db, for example, bit plane 4 is on the threshold of showing no organization. At 25 db, bit plane 4 shows no organization and bit plane 3 is on the threshold of disorganization. At 20 db, bit plane 3 shows virtually complete disorganization; bit plane 2 retains fairly strong organization overlaid with a considerable amount of salt-and-pepper noise. At 15.4 db, even bit plane 2 is on the verge of disorganization. Reproductions of the actual pictures are in the photographic section of this report, part A. 5P.

The initial suggestion of these results is that there is a serious question as to whether it is worthwhile to use the channel capacity to send those bits which show considerable disorganization. A line can be drawn through the matrix of 36 pictures which separate bit planes with organization from bit planes with no organization. This is done in Figure A. 5. 2-2.

While the above visual interpretations of the experimental results are interesting, quantitative verification is always desirable. Second-order statistics were taken on all 36 bit planes. The raw data is reported in Figure A. 5. 2-3. From these data, transition probabilities were calculated and given in Figure A. 5. 2-4. Referring to Figure A. 5. 2-4, the following can be noted: for the two high-signal-to-noise-ratio cases, bit planes 5 and 6 both show transition probabilities ranging between 0.49 and 0.51. This indicates that, considering each bit plane as a second-order information source, very little information is produced regarding the occurrence of a zero or one, given that the occurrence in the preceding picture element is known.

7

RE-ORDER No. 64-712
Vol I

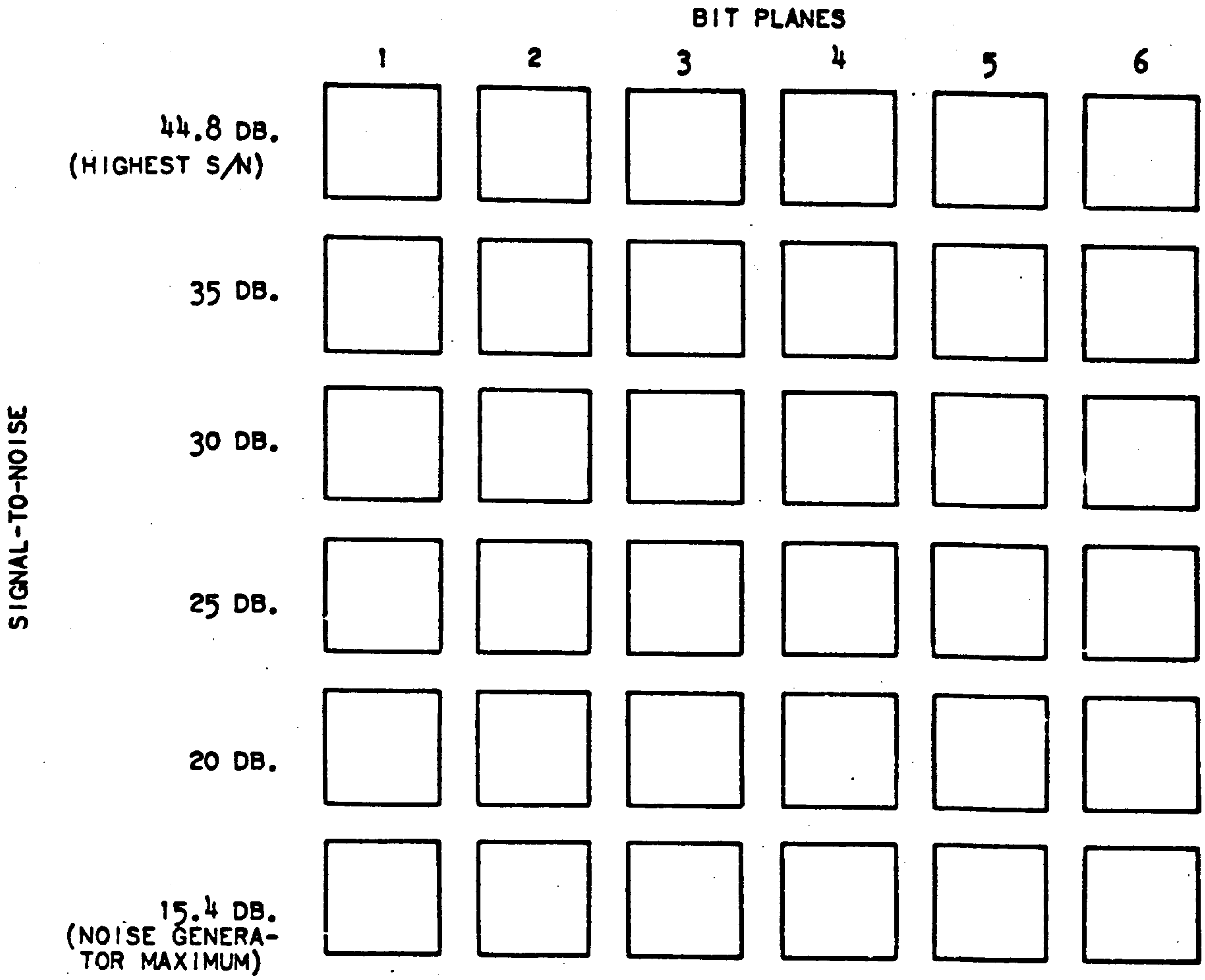


FIGURE A.5.2-1

SCHEMATIZED ARRAY OF BIT PLANE PICTURES FOR VISUAL ANALYSIS

07314

RE-ORDER No. 64-712
Vol I

SIGNAL-TO-NOISE

SIGNAL-TO-NOISE	BIT PLANES 1	2	3	4	5	6
44.8 DB. (HIGHEST S/N)	O	O	O	O	D	D
35 DB.	O	O	O	O	D	D
30 DB.	O	O	O	O	D	D
25 DB.	O	O	O	D	D	D
20 DB.	O	O	D	D	D	D
15.4 DB. (NOISE GENERATOR MAXIMUM)	O	O	D	D	D	D

KEY:

O = ORGANIZED

D = DISORGANIZED

FIGURE A.5.2-2
DIVISION OF BIT PLANE PICTURES
INTO ORGANIZED AND DISORGANIZED CATEGORIES

07314

-9

BIT PLANES

DB.		1		2		3		4		5		6	
				0	1								
44.8 (HIGHEST S/N)	0	63330	585	25600	5190	24400	9430	19360	14190	16500	16390	16360	16460
	1	575	1350	4810	29300	9470	22500	14100	17700	16350	16460	16300	16470
35		63640	536	35570	5470	24540	9215	17820	14575	16650	16330	16306	16300
		518	830	5250	18670	9250	22880	14550	18900	16280	16360	16400	16540
30		63700	540	33550	6620	21480	11584	16600	15840	17100	16380	16220	16400
		530	765	6380	19033	11606	20800	16025	16950	16350	15980	16470	16790
25		63630	629	32870	7760	19280	13409	17430	16340	16970	16250	16380	16250
		608	707	7600	17460	13703	18820	16177	15640	16380	15750	16460	16520
20		63350	870	33460	10420	19450	15390	19260	16022	18338	16113	17608	16317
		862	426	10230	10305	15230	15315	15930	14360	16220	14614	16097	15153
15.4 (NOISE GENERATOR MAXIMUM)		59770	2643	25960	14016	19730	15907	19120	16134	18600	16122	17680	16294
		2627	534	14030	11424	15800	14230	16036	14058	16174	14640	16290	14960

FIGURE A.5.2-3
RAW DATA - NUMBER OF OCCURRENCES PER FRAME
OF ELEMENT PAIRS 00, 01, 10, AND 11

07314

BIT PLANES

DB.		1		0 (2)	1 (2)	3		4		5		6	
44.8 (HIGHEST S/N)	0	0.99084	0.00915	0.83143	0.16856	0.72125	0.27874	0.57704	0.42295	0.50167	0.49832	0.49847	0.50152
	1	0.29870	0.70129	0.14101	0.85898	0.29621	0.70378	0.44339	0.55660	0.49832	0.50167	0.49740	0.50259
35		0.99164	0.00835	0.86671	0.13328	0.72700	0.27299	0.55008	0.44991	0.50485	0.49514	0.50009	0.49990
		0.38427	0.61572	0.21948	0.78051	0.28789	0.71210	0.43497	0.56502	0.49877	0.50122	0.49787	0.50212
30		0.99159	0.00840	0.83520	0.16479	0.64964	0.35035	0.51171	0.48828	0.51075	0.48924	0.49724	0.50275
		0.40926	0.59073	0.25105	0.74894	0.35814	0.64185	0.48597	0.51402	0.50572	0.49427	0.49518	0.50481
25		0.99021	0.00978	0.80900	0.19099	0.58980	0.41019	0.51613	0.48386	0.51083	0.48916	0.50199	0.49800
		0.46235	0.53764	0.30327	0.69672	0.42133	0.57866	0.50843	0.49156	0.50980	0.49019	0.49909	0.50090
20		0.98645	0.01354	0.76253	0.23746	0.55826	0.44173	0.54588	0.45411	0.53229	0.46770	0.51902	0.48097
		0.66925	0.33074	0.49817	0.50182	0.49860	0.50139	0.52591	0.47408	0.52604	0.47395	0.51510	0.48489
15.4 (NOISE GENERATOR MAXIMUM)		0.95765	0.04234	0.64938	0.35061	0.55363	0.44636	0.54234	0.45765	0.53568	0.46431	0.52039	0.47960
		0.83106	0.16893	0.55119	0.44880	0.52614	0.47385	0.53286	0.46713	0.52489	0.47510	0.52128	0.47872

FIGURE A. 5. 2-4
TRANSITION PROBABILITIES
OF BIT PLANE INFORMATION SOURCES
(J LEVEL FOLLOWS I LEVEL) I, J ONE OR ZERO

07314

RE-ORDER No. 64-712
Vol I

RE-ORDER No. 64-712
Vol I

Proceeding to the 30 db signal-to-noise ratio, and again referring to Figure A. 5. 2-4, it appears that bit plane 4 has poor organization from a second-order transitional-probability standpoint. The transitional probabilities at this bit plane vary between 0. 48 and 0. 52. However, referring to the array of the actual photographs, it can been seen that a small amount of organization inherent in bit plane 4 at 30 db signal-to-noise ratio. At 25 db signal-to-noise ratio, bit plane 4 also exhibits transitional probabilities between 0. 48 and 0. 52. However, in this instance a visual check of bit plane 4 fails to reveal any humanly detectable evidence of organization.

At 20 db and at 15. 4 db, bit plane 3--which visually shows no organization--still has transition probabilities between 0. 44 and 0. 56. On the surface, this would indicate statistically, that a reasonable amount of recognisable organization can be expected. However, at 20 db bit planes 5 and 6 show transition probabilities between 0. 46 and 0. 54. Since bit planes 5 and 6 are disorganized at 44. 8 db, these bit planes should have no organization. The answer to the seeming paradox of slight statistical organization at low signal-to-noise ratios in the high-order bit planes seems to lie in the consideration of edge effects. At a 20 db signal-to-noise ratio, the two-sigma limits of the noise extend over 20% of the dynamic coding range. At 15. 4 db, approximately three-eighths of the dynamic coding range is covered by the two-sigma limits of the noise. This means that the picture-plus-noise signal does indeed spend a considerable amount of time in the zero condition and in the fully saturated condition. On an intuitive basis, it is quite conceivable that normal noise spikes would drive the picture signal so that two or more minor bits (6, 5, 4, etc.) would remain unchanged for two or three picture elements, thus exhibiting some element-to-element transition probability correlation.

At this point we turn to a method of estimating compression for BIRL which is based on a transition probability model for run length coding of a one-bit source. The model is written in closed form for truncation at a run length of 256. The model was then applied to actual statistics obtained from EDITS and the compression for the BIRL technique was calculated and compared to the compression

74

RE-ORDER No. 64-712
Vol I

estimated from actual run-length probabilities of the bit planes. From these results several conclusions regarding BIRL are presented.

Probabilistic Model

Assuming that a run length t units long of level s is defined as

$$R_s^t = \bar{s}\ \underbrace{s\, s, \ldots s}_{t}\ \bar{s}\ , \tag{30}$$

where $\bar{s}$ is any other level but s. Using the transition probabilities of $P_0(0) = x$ and $P_1(1) = y$, the run length probabilities are given by equations (31) and (32) assuming that runs to infinity could exist.

$$P(R_x{}^n) = \frac{(1-x)\, x^n}{2x} \tag{31}$$

$$P(R_y{}^n) = \frac{(1-y)\, y^n}{2y} \tag{32}$$

If the maximum run length is limited to some maximum finite value, k, the run-length probabilities are given by equations (33) and (34):

$$P(R_x{}^k) = \frac{(1-x)\, x^n}{x(2-x^k - y^k)}\ , \tag{33}$$

$$P(R_y{}^k) = \frac{(1-y)\, y^n}{y(2-x^k - y^k)}\ . \tag{34}$$

The maximum, average, run-length compression is

$$\bar{C}_{H\ max.} = \frac{\sum_{n=1}^{\infty} \left[n\ P(R_x{}^n) + n\ P(R_y{}^n)\right]}{\sum_{n=1}^{\infty} \left[P(R_x{}^n) \log_2\ P(R_x{}^n) + P(R_y{}^n) \log_2\ P(R_y{}^n)\right]}\ . \tag{35}$$

and the lower bound on average run-length compression by flat encoding is given by equation (36).

$$\bar{C}_{FC} = \frac{\sum_{n=1}^{k} \left[n\ P(R_x{}^n) + n\ P(R_y{}^n)\right]}{1 + \log_2 k} \tag{36}$$

Using the following summation formulae (equations (38)-(40), the various compression ratios can be defined.

$$\sum_{n=1}^{n} x^n = \frac{x(1-x^n)}{(1-x)} \tag{37}$$

$$\sum_{n=1}^{n} n\,x^n = \frac{x}{(1-x)^2}\left[1-(n+1)\,x^n + n\,x^{n+1}\right] \tag{38}$$

$$\sum_{n=1}^{\infty} x^n = \frac{x}{(1-x)} \tag{39}$$

$$\sum_{n=1}^{\infty} n\,x^n = \frac{x(2-x)}{(1-x)^2} \tag{40}$$

Assuming that runs to infinity exist, from equations (31), (32), (35), and the summation formulae, the compression is

$$\bar{C}_{H\infty} = \frac{\dfrac{2-x}{1-x} + \dfrac{2-y}{1-y}}{2 - \left\{\log(1-x) + \dfrac{\log_2 x}{(1-x)} + \log_2(1-y) + \dfrac{\log_2 y}{(1-y)}\right\}} \,. \tag{41}$$

Assuming that runs to 256 exist, the compression is

$$\bar{C}_{H256} = \left[\frac{\dfrac{(1-257x^{256}+256x^{257})}{(1-x)}}{-\left\{(1-x^{256})\log_2\left[\dfrac{(1-x)}{x(2-x^{256}-y^{256})}\right] + \dfrac{(1-257x^{256}+256x^{257})}{(1-x)}\log_2 x\right.}\right.$$
$$\left. + \left[\frac{\dfrac{(1-257y^{256}+256y^{257})}{(1-y)}}{\left.(1-y^{256})\log_2\left[\dfrac{(1-y)}{y(2-x^{256}-y^{256})}\right] + \dfrac{(1-257y^{256}+256y^{257})}{(1-y)}\log_2 y\right\}}\right]\right] \tag{42}$$

Assuming x = y = 0.9, a plot of compression as a function of truncation for entropy calculation is given in Figures A. 5.2-5 and A. 5.2-6. If it is assumed that runs to

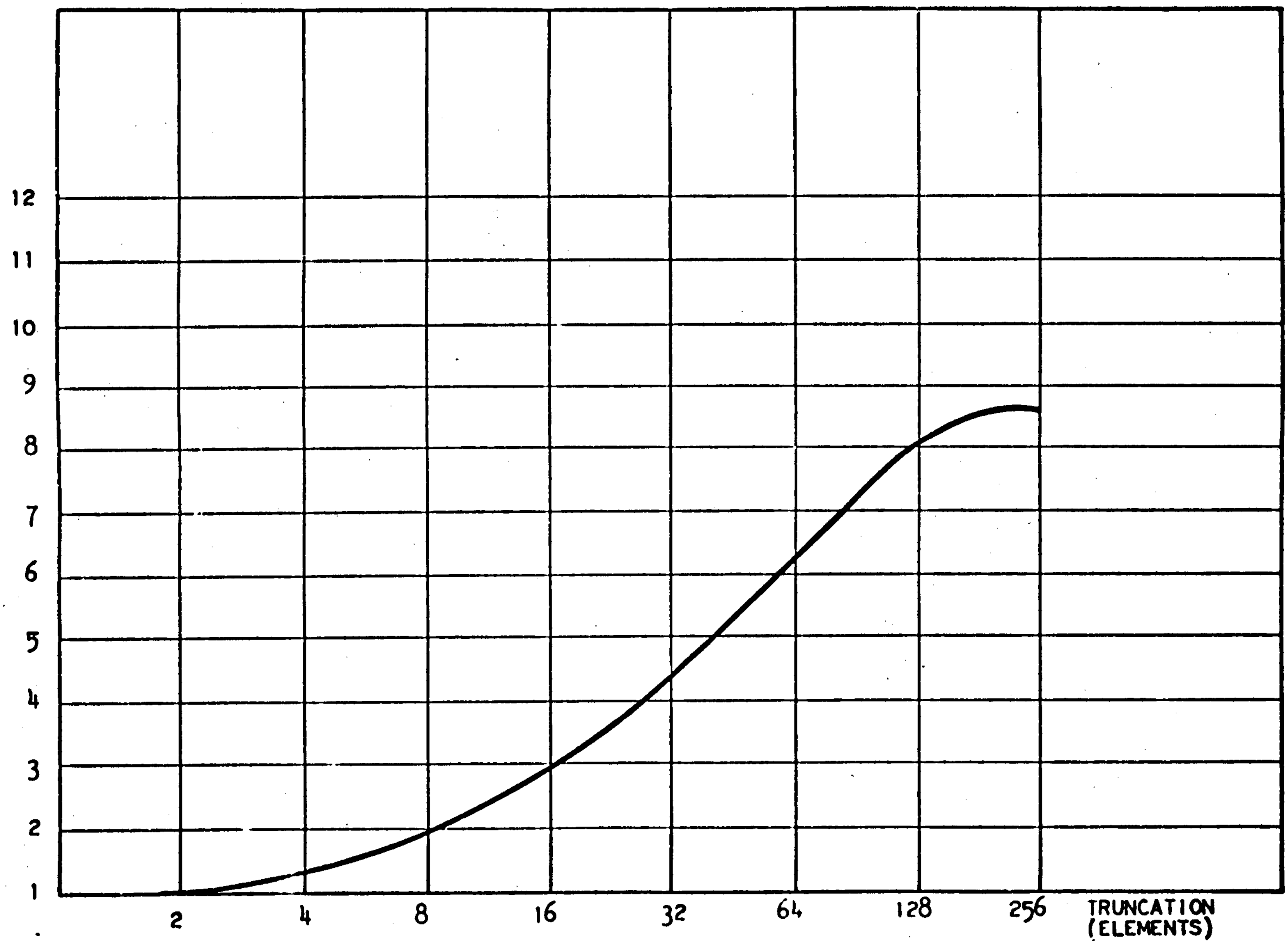


FIGURE A. 5. 2-5
AVERAGE COMPRESSION AS A FUNCTION
OF TRUNCATION FOR TRANSITION PROBABILITIES OF 0.99

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RE-ORDER No. 64-712
Vol I

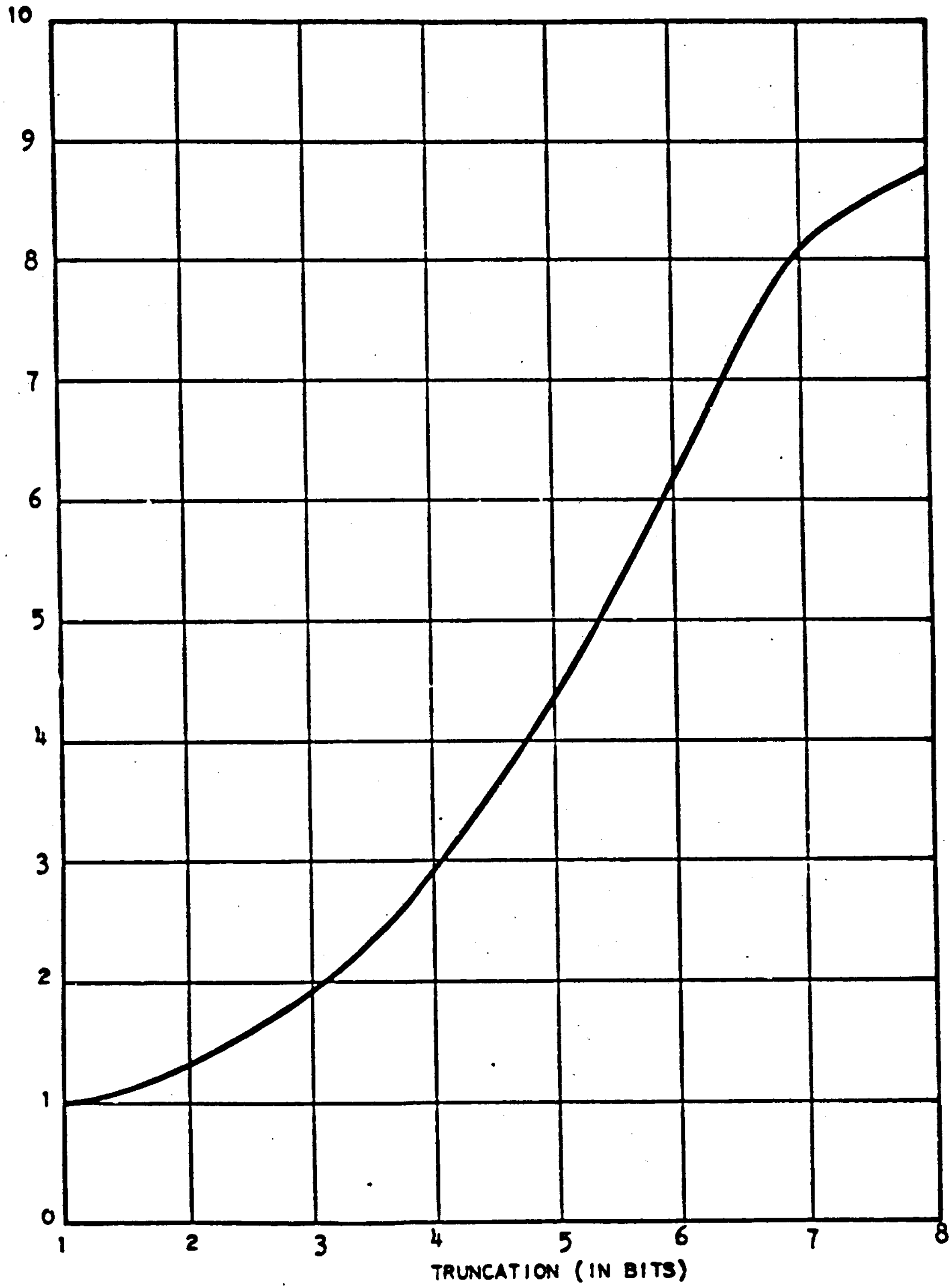

FIGURE A.5.2-6

AVERAGE COMPRESSION AS A FUNCTION
OF TRUNCATION FOR TRANSITION PROBABILITIES MODEL,
ASSUMING TRANSITION PROBABILITIES OF 0.99

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RE-ORDER No. 64-712
Vol I

256 exist, and the runs are encoded with a fixed length code,

$$\bar{C}_{FC256} = \frac{(1-257x^{256} + 256x^{257})}{9(1-x)(2-x^{256}-y^{256})} + \frac{(1-257y^{256} + 256y^{257})}{9(1-y)(2-x^{256}-y^{256})} .$$

BIRL Coding

Taking the four most significant bit planes of a 256 lunar picture, the transition probabilities are given by Table A. 5. 2-1

For these transistion probabilities, the various compression ratios for each bit plane can be tabulated along with the average run length. For the Lunar picture the results are given in Table A. 5. 2-2 and graphically presented in Figure A. 5. 2-7.

TABLE A. 5. 2-1

BIT PLANE TRANSITION PROBABILITIES FOR MAXIMUM SENSOR SIGNAL-TO-NOISE RATIO FOR THE LUNAR PICTURE

BIT PLANE	X	Y
1	0. 9908	0. 7013
2	0. 8314	0. 8590
3	0. 7213	0. 7038
4	0. 5770	0. 5566

TABLE A. 5. 2-2

COMPRESSION RATIOS AND AVERAGE RUN LENGTHS FOR BIT PLANES OF LUNAR PICTURE

BIT PLANE	$\bar{R}_L$	$\bar{R}_{L256}$	$\bar{C}_H$	$\bar{C}_{H256}$	$\bar{C}_{FC256}$
1	114. 05	40. 79	8. 35	6. 58	4. 53
2	15. 02	6. 51	1. 43	1. 30	0. 72
3	8. 96	3. 48	1. 0	0. 87	0. 39
4	6. 62	2. 31	0. 81	0. 70	0. 26

The net compression for BIRL is

$$\bar{C}_{BIRL} = \frac{N}{\sum_{i=1}^{N} \frac{1}{C_i}} . \qquad (43)$$

RE-ORDER No. 64-712
Vol I

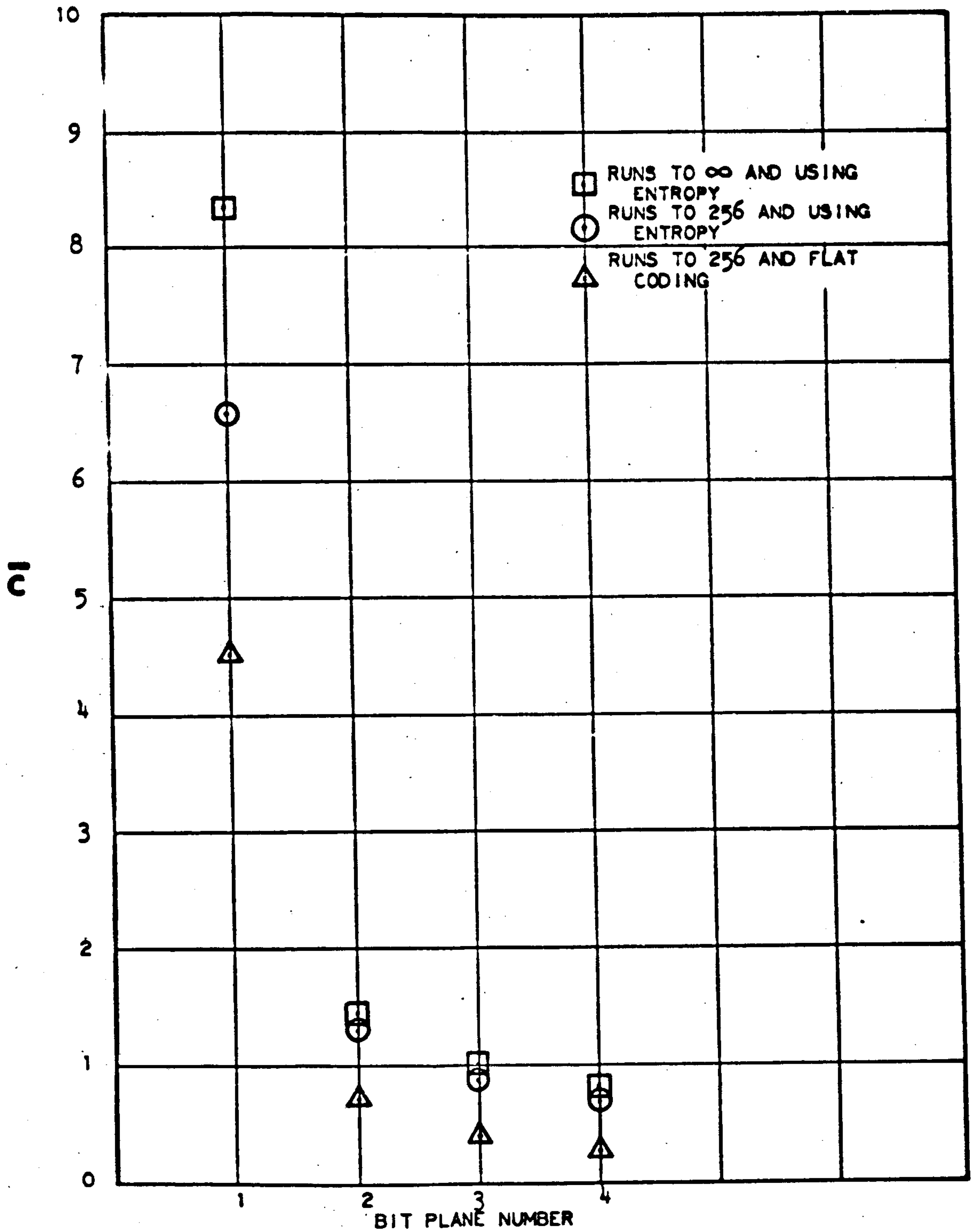

FIGURE A. 5. 2-7
COMPRESSION FOR BIT PLANES OF 256 LUNAR PICTURE
FOR VARIOUS CODING FORMULAE

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Vol I

where C_i is the average compression for the individual bit planes and N is the number of bit planes.

The results for truncations at ∞ and 256 with entropy and flat coding are given below:

TABLE A. 5. 2-3

AVERAGE COMPRESSION OF 4-BIT PLANES FOR LUNAR PICTURE

TRUNCATION	∞	H256	FC256
BIRL COMPRESSION	1. 31	1. 14	0. 5

Actual truncated run-length probabilities were obtained on EDITS at 512 resolution and the results are tabulated in Table A. 5. 2-4 and presented in Figure A. 5. 2-8.

TABLE A. 5. 2-4

COMPRESSION OF VARIOUS BIT PLANES AS A FUNCTION OF TRUNCATION FOR LUNAR PICTURE QUANTIZED TO 6 BITS AT 512 RESOLUTION FOR BOTH FLAT ENCODING AND UPPER BOUND USING ENTROPY

Bit Plane	No. 1		No. 2		No. 3		No. 4		No. 5		No. 6	
Truncation	$\bar{C}_{FC}$	$\bar{C}_H$	$\bar{C}_{FC}$	$\bar{C}_H$	$\bar{C}_{FC}$	$\bar{C}_H$	$\bar{C}_{FC}$	$\bar{C}_H$	$\bar{C}_{FC}$	$\bar{C}_H$	$\bar{C}_{FC}$	$\bar{C}_H$
4	1. 32						0. 94		0. 55			
8	1. 84		1. 32		1. 46		0. 60					
16	2. 71	11. 93	1. 67	2. 84	1. 23							
32	4. 0		1. 95									
64	5. 49		1. 99									
128	6. 64											

Conclusions

The results of this investigation are summarized in Figure A. 5. 2-8. The actual run length statistics were taken at 512 resolution. Therefore, whatever compression can be obtained at 512, the compression at 256 will be less, since as the resolution decreases the redundancy decreases and hence, the average run length decreases.

RE-ORDER No. 64-712
Vol I

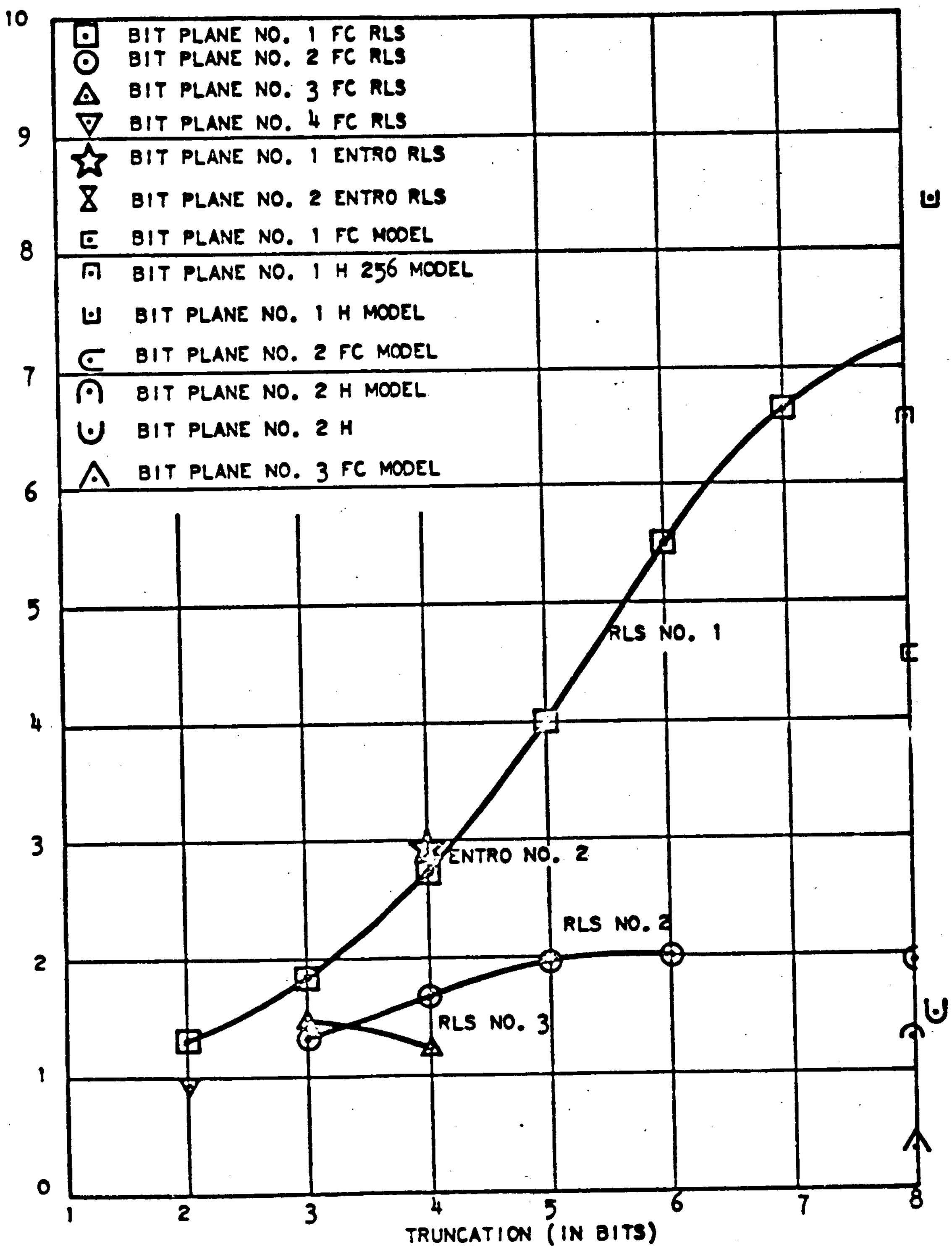

FIGURE A.5.2-8

BIRL COMPRESSION AS A FUNCTION OF RUN-LENGTH TRUNCATION
FOR MODEL AND ACTUAL STATISTICS

07314

RE-ORDER No. 64-712 Vol I

The gross average compression for BIRL is given by equation (15) and, using the actual run-length statistics, the compression is 1.3:1. Therefore, at 256 resolution the gross average compression will be even less. When considerations of implementation, data link noise, buffer store, and synchronization are considered, the net compression will be extremely small and may actually result in an increase in bandwidth. Based on the actual run-length probabilities and the results obtained from the transition-probability model, the BIRL technique is not a suitable digital television compression technique for the JPL study.

A.6 ROBERTS CODING

The Roberts [1, 2] coding technique of bandwidth compression is not an information preserving coding technique. Compression is achieved by reducing the quantization and breaking the annoying contours up by the addition and subtraction of identical pseudorandom noise at the transmitter and receiver. The technique indirectly utilizes statistical redundancy and displays the data in such a manner as to take advantage of the physiological properties of the human viewer.

Implementation of the technique is achieved by the addition of a nonlinear amplifier and a pseudorandom noise generator to the transmitting and receiving portions of a conventional PCM television system. A nonlinear amplifier is used to match the response of the eye to the displayed image and optimize the overall system transfer function. Since the response of the eye to brightness stimulation is a square-root function, the dynamic range could be compressed to concentrate errors of intensity to regions of high intensity where the eye would not appreciably notice these errors. This portion of the technique could physically be implemented by passing the analog video signal from the sensor through an amplifier with a square-root transfer function which compands the video signal.

Pseudorandom noise is then added to the companded signal before it is converted to a digital signal. It is necessary that the pseudorandom noise has well defined properties, according to Roberts. Roberts stated that if the video is quantized to n levels and that if the number of elements in the picture is m^2, then the maximum amplitude of the pseudorandom noise must be limited to one quantizing level; that is, $1/n$, if the peak-to-peak excursion of the video signal is normalized to

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(1) Roberts, L. G., PCM Television Bandwidth Reduction Using Pseudo-Random Noise, S. M. Thesis, Mass. Inst. Tech., Feb., 1961.
(2) Roberts, L. G., Picture Coding Using Pseudo-Random Noise, IRE Transactions On Information Theory; Feb., 1962.

21

RE-ORDER No. 64-712
Vol I

one volt. The maximum amplitude, 1/n, is then divided into p distinct levels where, as a function of time, the p levels each occur with an equal probability, have equal run-length probabilities, and have a correlation function that is two-valued over the period of occurrence; that is, the p levels occur in a pseudorandom manner.

This pseudorandom noise can be generated by a k-stage shift register with linear-modulo-two feedback. Roberts stated that if the pseudorandom noise was quantized to P bits, where $P = \log_2 p$, then the number of stages in the shift register must satisfy the relationship that $2^k - 1$ be evenly divisable by P and equal to m^2; that is, the period of the pseudorandom sequence must be equal to the number of elements in a frame. For this condition, the P bit sequence generated by the pseudorandom shift register sequence generator will not repeat itself in one frame.

Several experiments were conducted to determine if this condition imposed by Roberts was necessary. Experimentally, the approach was as follows: EDITS was used with a pseudorandom shift register generator with a variable number of stages. The television resolution was 256 x 256 elements/frame. If one bit of Roberts noise is to be added, the number of stages required is 18, if the sequence is not to be repeated in one frame. However, if three bits of Roberts noise is to be added, the sequence will repeat after 87, 381 elements if an 18 stage device is used. Several experiments were performed to determine whether or not fewer shift register stages could be used.

The 18 stage shift-register pseudorandom number generator in EDITS was shortened to operate as a four stage, three stage, and two stage shift register. For three and two bit quantization and various Roberts quantization PE was measured and Polaroid prints were made from the displayed image at the 256 resolution. The results of PE as a function of the Roberts quantization and the number of shift register stages is given in Table A. 6-1.

From Table A. 6-1, decreasing the number of stages reduces PE slightly, but not a significant amount to vary the compression appreciably. There is a substantial decrease in PE when Roberts is added, but the change for various Roberts quantization is not significant although PE does increase Roberts quantization.

RE-ORDER No. 64-712
Vol I

TABLE A.6-1

PE AS A FUNCTION OF QUANTIZATION, ROBERTS QUANTIZATION AND NUMBER OF SHIFT-REGISTER STAGES FOR 256 RESOLUTION USING MIT GIRL PICTURE ON EDITS

Quantization	18 Stages	4 Stages	3 Stages	2 Stages
3-0	0.836	0.855	0.860	--
3-1	0.738	0.730	0.722	0.713
3-2	0.757	0.746	0.740	0.728
3-3	0.761	0.751	0.745	--
3-4	0.765	0.752	--	--
2-0	--	0.913	--	--
2-1	0.814	0.814	0.804	0.781
2-2	0.827	--	--	0.801
2-3	0.831	--	0.827	--
2-4	0.833	0.836	--	--

It can be concluded that as far as compression for a PEC with Roberts is concerned, the compression is nearly independent of Roberts quantization and the number of shift-register stages; however, there is a slight decrease in the PEC compression when Roberts is added to the data stream. Therefore it is apparent the true evaluation of Roberts be a subjective evaluation of the displayed image. Figures A6P-1 through A. 6P-8 are the photographs obtained from EDITS for the various conditions of quantization, Roberts quantization, and the number of shift-register stages. It is apparent that there is a relationship between the Roberts quantization and the PCM quantization for various shift-register lengths. It would be possible to decrease the number of shift-register stages without appreciably affecting the subjective picture quality; however, if the number of stages is decreased beyond a reasonable number, annoying contours are present in the displayed image as seen from the photographs of Figure A. 6P-1 through A. 6P-8.

Assuming that a three-four Roberts system with a four stage shift register gives a suitable picture, the compression can be calculated. The gross compression is 2:1 since, instead of six bits per element, three bits per element are being transmitted. The Roberts technique of pseudorandom noise addition and subtraction randomizes the average video level to many levels instead of only one which reduces the effect of contouring present in coarse quantization. Subjectively, a three-bit picture with Roberts technique becomes acceptable relative to a six-bit picture.

A block diagram of the Roberts system is given in Figure A. 6-1 and the technique has been built into EDITS. This allowed basic Roberts experimentation and experimentation of combining Roberts with other coding techniques.

The subjective results of the comparison of compression techniques is given in Appendix E. 2. Summarizing these results subjectively on a single frame basis at 40 db signal-to-noise ratio, the three-four Roberts is not subjectively equivalent to six-bit PCM as claimed by Roberts. However, it appears from the photographs in Appendix E. 4 that, subjectively, the Roberts is preferred to the Delta compression technique. Therefore, this starts a foundation to eliminate the Delta system from being the recommended system.

It appears from these photographs that there exists a relationship such that six-bit PCM and three-four Roberts are almost subjectively equivalent at X db and Y db signal-to-noise ratio for the two systems, respectively. Therefore, the six-bit PCM and three-four Roberts Polaroid pictures at 30 and 20 db signal-to-noise ratio were enlarged to determine this relationship. Both the Canadian Arctic and the JPL Lunar scenes were used for the comparison since they represent extremes in high and low contrast scenes. From photographs A. 6P-9, A. 6P-10, A. 6P-13, and A. 6P-14 the Canadian Arctic and JPL Lunar photographs at 30 db sensor signal-to-noise ratio at six-bit PCM not only represents a more pleasing picture, but it is also superior in line and small detail recognizability. It is apparent that the lower contrast photos are almost subjectively equivalent somewhere between 20 and 30 db sensor signal-to-noise ratio. However, for the high contrast scene, even at 20 db sensor signal-to-noise ratio, the six-bit PCM is preferred to the three-four Roberts both subjectively and in detail recognizability

Although the gross compression is 2:1, it is necessary to determine the net compression. Utilizing the same constraints imposed previously, the net compression is

84

RE-ORDER No. 64-712
Vol I

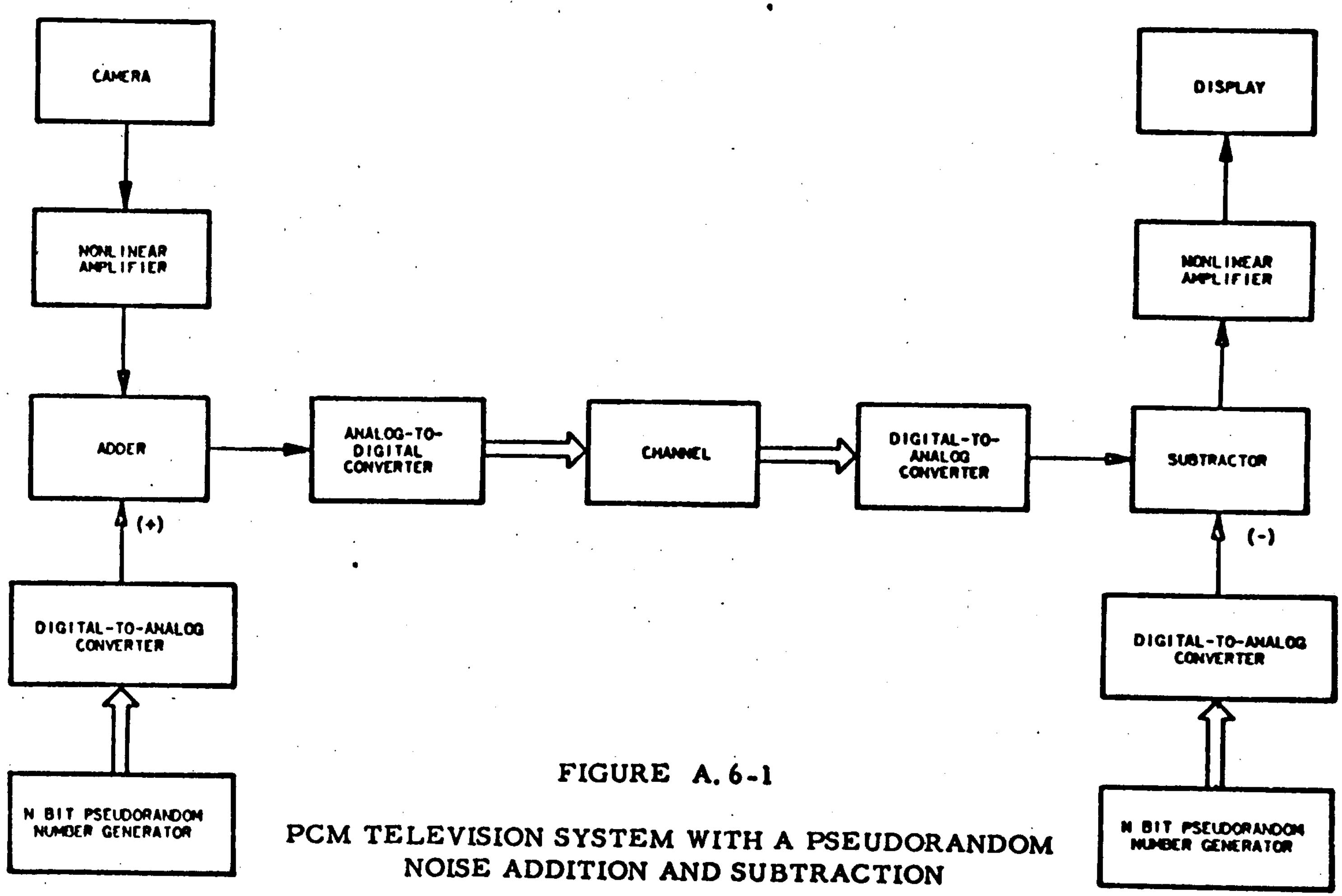

FIGURE A.6-1

PCM TELEVISION SYSTEM WITH A PSEUDORANDOM NOISE ADDITION AND SUBTRACTION

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Vol I

$$\bar{C}_{net} = \frac{6m + k_1}{\Delta S/N\ (3m + k_3)}\ ; \tag{44}$$

where k_1 is the number of bits for synchronization of a six bit PCM source, k_3 is the synchronization bits for the Roberts source, and $\Delta S/N$ the change in S/N for the compressed source. Based on experiments performed at EMR and the results of other organizations, $\Delta S/N$ is 0.85. This assumes coherent phase-shift keying since with the pseudorandom noise more errors can be tolerated over conventional PCM. From the synchronization study, k_1 equals 30 bits, and k_3 equals 30 bits; therefore, the net compression is

$$\bar{C}_{net} = \frac{1566}{0.85\ (798)} = 2.31. \tag{45}$$

This compression is independent of the source statistics, whereas, for the majority of the compression techniques the compression is picture dependent.

It appears that the Roberts encoding technique satisfies the desired JPL compression requirement; however, from the photos, Figures A.6P-9 through A.6P-16, it is quite apparent that, subjectively, the three-four Roberts is not equivalent to the six bit PCM. There does appear to be some relationship between the signal-to-noise required for six bit PCM and three-four Roberts. It appears that somewhere between 20 and 30 db sensor signal-to-noise ratio, the six bit PCM and three-four Roberts are subjectively equivalent. It then remains to be seen whether or not the space scientist can utilize the information from the Roberts display.

RE-ORDER No. 64-712
Vol I

A.7 DELTA MODULATION

A.7.1 Introduction

Like split-band, Delta has been covered in some detail in a previous NASA sponsored study.(1) Numerous references are cited in this study which lead to even more detailed consideration of many of the particular Delta modulation variants. Delta modulation has so many variants because, even though the conceptual diagram (Figure A.7.1-1) is exceedingly simple, each of the constituent blocks is capable of several widely varying implementations, and most of these can be used in combination.

In concept, a Delta modulation system includes a decision element and an approximation element. As an option it may include a pre-processing unit and a post-processing unit. The input waveform is compared to the approximated waveform in the decision element. Depending on the similarity between the input waveform and the approximating waveform, the decision element alters the output of the approximation element. The signal by which the decision element changes the operation of the approximation unit is the signal which is transmitted. At the receiver there is another approximation element, identical to that at the transmitter. The receiver's approximation element, under control of the transmitter's decision element, produces a facsimile of the input waveform at the receiver output.

The classical form of Delta modulation uses both a simple decision element and an equally simple approximation element. The decision element is a difference amplifier with a threshold at zero. If the difference between the input waveform and the approximation waveform is positive, a negative pulse (in principle) is sent to the approximation element. If the difference is negative, a positive pulse is sent. The approximation element is a simple integrator of near-infinite time constant. Thus, the stream of difference signal pulses are integrated into a reasonable facsimile of the input waveform. Figure A.7.1-2 illustrates this system.

There are several disadvantages to this simple system. (1) It is subject to over load, i.e., a fast-rising input waveform may yield a positive difference in the

(1) Final Report, Manned Spacecraft Advanced Television Study, NASA Manned Spacecraft Center, Houston, Texas. Contract NAS 9-991, 31 July 1963.

RE-ORDER No. 64-712
Vol I

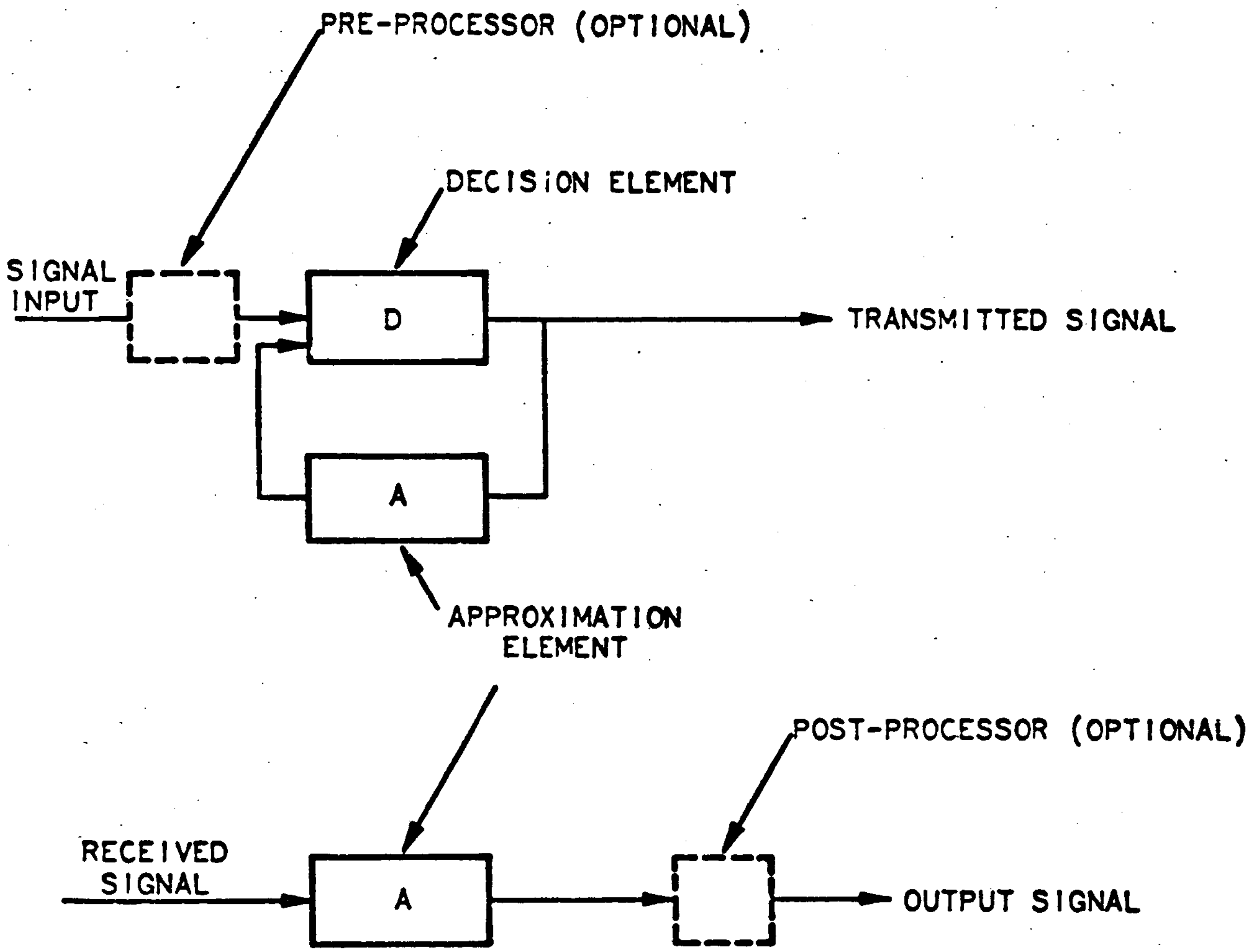

FIGURE A.7.1-1

GENERALIZED DELTA MODULATION SYSTEM

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88

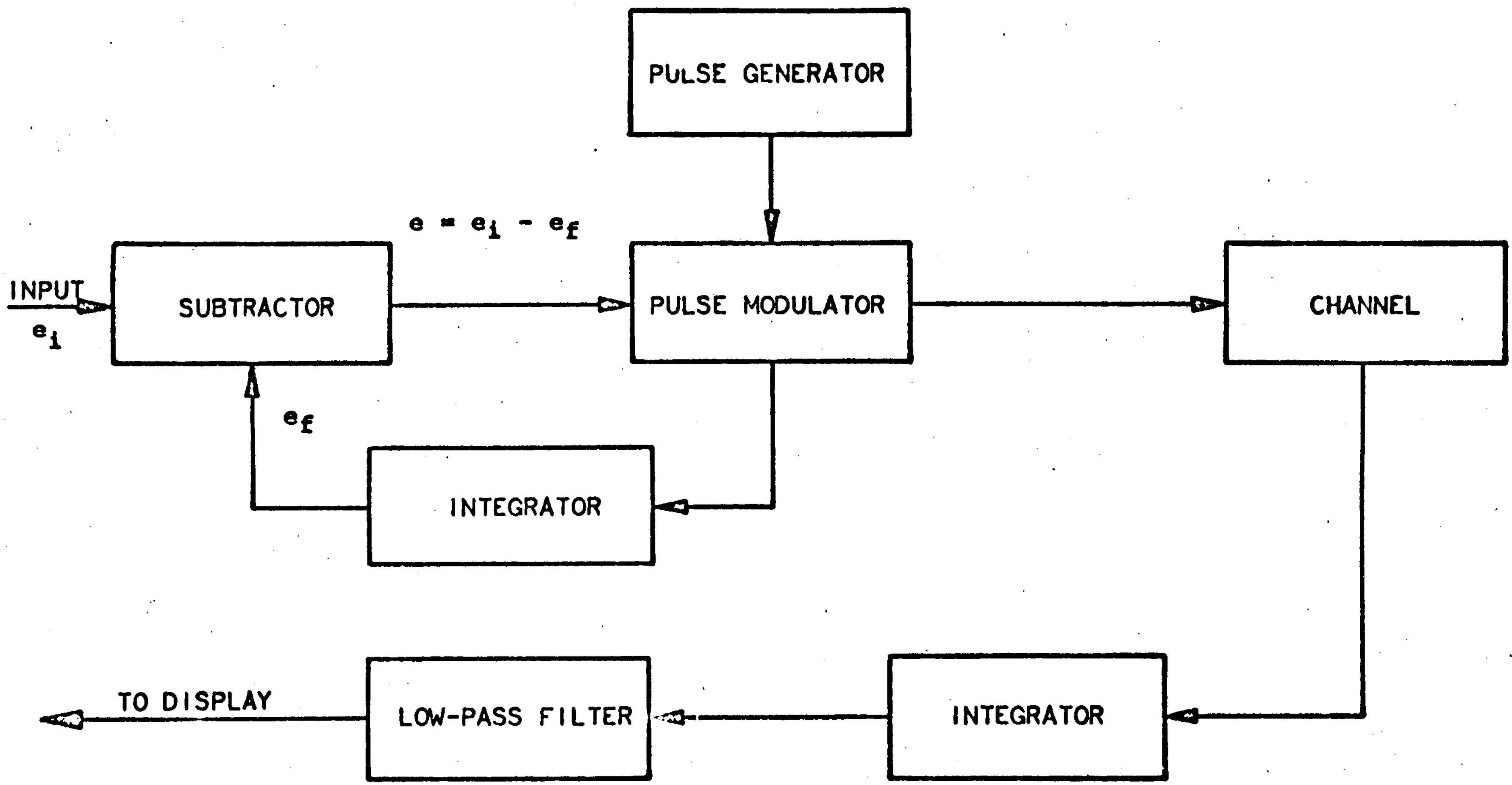

FIGURE A.7.1-2

BLOCK DIAGRAM OF SINGLE-BIT
DELTA MODULATOR ENCODER AND DECODER

07314

RE-ORDER No. 64-712
Vol I

RE-ORDER No. 64-712
Vol I

decision element for many sample times; (2) it has poor step function response; (3) it has a $\frac{1}{\omega}$ frequency response characteristic; (4) it yields a low signal-to-noise ratio (on speech) unless the pulse rate equals or exceeds that of a high quality PCM system.(1) Virtually the sole advantage of this system is the simplicity of its terminal equipment.

The vast proliferation of Delta schemes is fueled by the hope of making improving modifications to either the decision element, the approximation element, or both, yet without unduly sacrificing the Delta virtue of terminal simplicity. Unfortunately most fixes (Δ^2, $\sum \Delta$ and other systems) do increase terminal complexity. The extent of the proliferation of schemes is indicated by the mere partial listing of functionally possible options in the decision element and in the approximation elements. The decision element may have 1, 3, 7, or more thresholds. It may produce 2^1, 2^2, 2^3 or more correction signals and these signals may have a virtual infinitude of relative weightings.

The approximation element, usually built around an integrator, can have several options: (1) the correction signals from the decision unit may be unequally weighted between positive and negative corrections, (2) the correction signals may have either a time- or event-independent meaning to the approximator, or their interpretation may depend on the immediate time history of several correction signals, (3) there may be one, two, or even three integrators (with one partially bypassed) in the approximation unit, (4) the integrators themselves may be of near infinite time constant, or relatively short (a few samples) time constant and the decay characteristics may be either exponential or trapezoidal. Accounting for all combinations of all options listed provides some 108 possible Delta systems. Each of these systems has several quantitative parameters which may be manipulated to achieve a reasonable balance between signal-to-noise ratio, channel noise immunity, step function response, and dc response.

During the course of this study, EMR has not, of course, investigated all these possible Delta configurations. Rather we have taken one promising Delta configuration from a previous work(2) and have optimized its parameters.

(1) Pieruschka, Erich, The Signal-to-Noise Ratio in Delta-Modulation Systems, Contribution Number 63, Research Laboratories, Ordnance Missile Laboratories, Redstone Arsenal, 10 March 1958.
(2) Cotten, R. V., et. al. Demonstrations of the Feasibility of Using Delta Modulation for Pictorial Transmission, ASD-TDR-62-1037, November, 1962. AF-33(657)-8478.

RE-ORDER No. 64-712
Vol I

A. 7. 2 Delta System

The Delta system chosen has three thresholds in the decision element. They are at relative values of 0 volts +0.82 volts and -0.82 volts, with 8.0 volts as video full scale. The approximation element is a single exponentially decaying integrator with a decay time constant of about 5.0 video sample times. This Delta system is locked into synchronism with EDITS, of which it is a part. Since the approximated waveform is utilized at only certain times its generation need not be accomplished by the usual integrator circuit so long as the waveform at the sample times would be that seen from a true integrator. This fact allows a certain freedom in implementation which, for the case of the experimental unit, has resulted in the configuration shown in Figure A. 7. 2-1.

The operation of this circuit is as follows: The video is sampled and held in EDITS, and this waveform is used as the input to the Delta system. The reconstructed video is amplified, inverted, and added to the incoming video. This results in the reconstructed video being subtracted from the incoming video. The difference signal is then applied to the threshold circuits. In practice, these are resistor-weighted tunnel-diode thresholds. One threshold is at zero volts difference, the other two are at ±10% of full scale as shown in Figure A. 7. 2-2.

The threshold outputs are sampled and stored in three flip-flops. The output of each of these three flip-flops is either zero or minus three volts. Because these three flip-flops are actually an m-out-of-n code, there are only four permutations of their outputs. These are 000, 001, 011, and 111. A Thevenin equivalent circuit of the flip-flop outputs and the integrating circuit is shown in Figure A. 7. 2-3. The time constant of this circuit is approximately five element periods (50 μs).

A textbook type of Delta modulation with exponential integration would take an abrupt step when a Delta function is applied and then decay towards zero volts as shown in the curve and Figure A. 7. 2-4.

The implemented version will suddenly have -3 volts applied and will charge towards this value with a time constant of five elements. Examination of these two curves shows that they coincide at the element times. The actual output

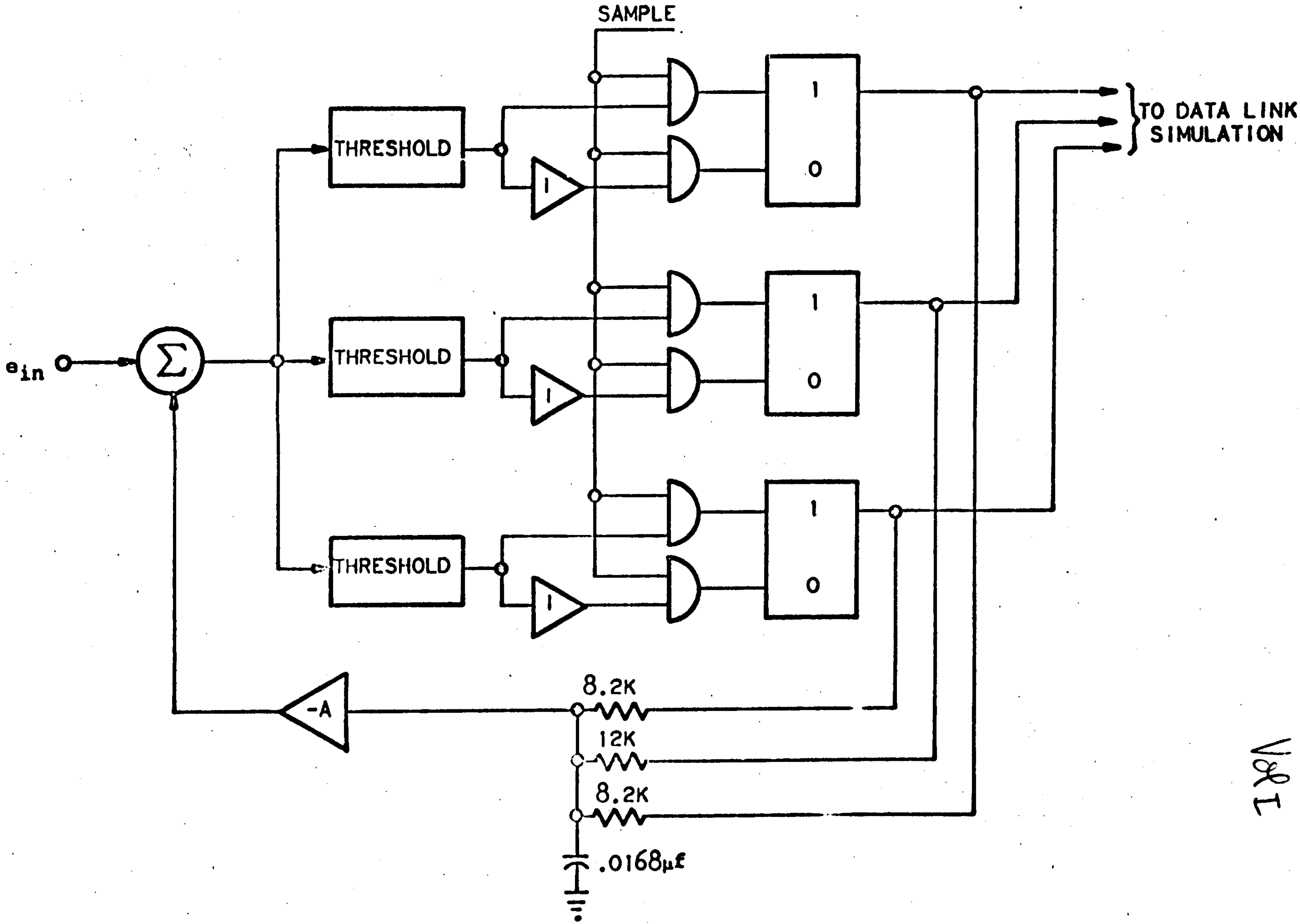

FIGURE A.7.2-1

DELTA MODULATION ENCODER

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RE-ORDER No. 64-712
Vol I

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Vol I

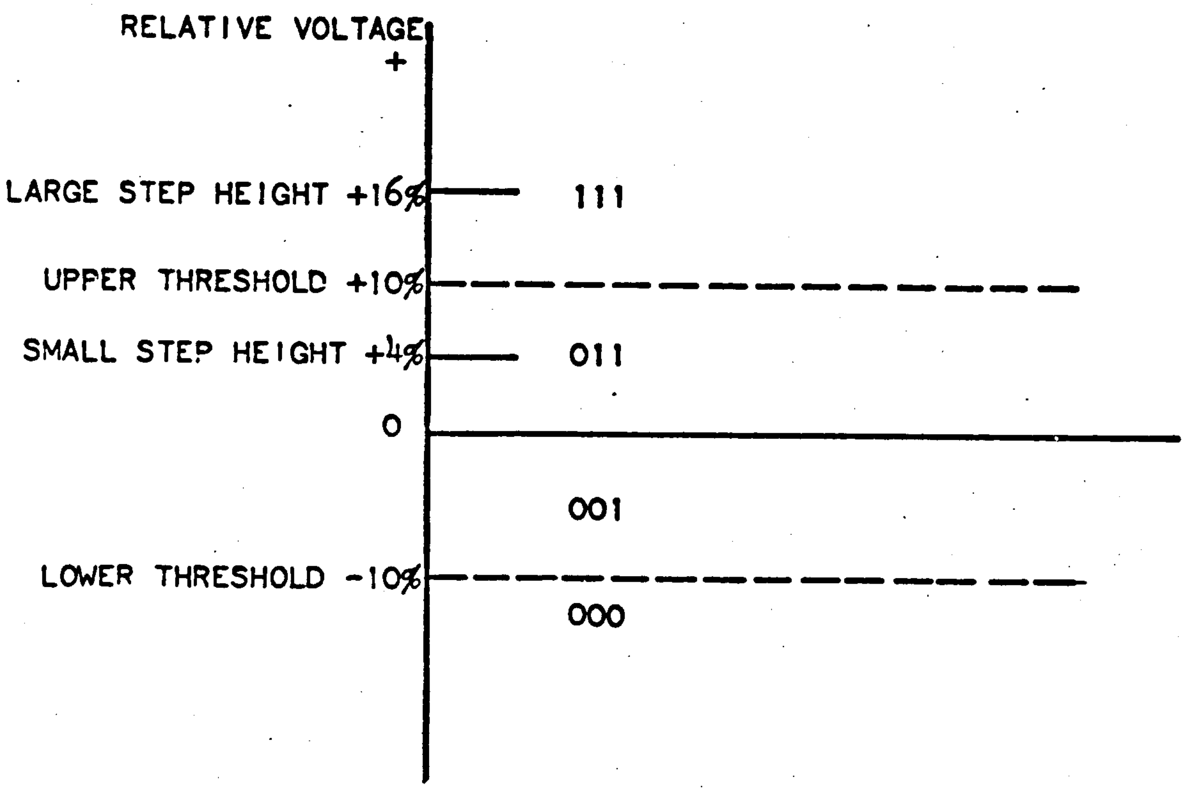

FIGURE A.7.2-2

DELTA MODULATION THRESHOLDS

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RE-ORDER No. 64-712
Vol I

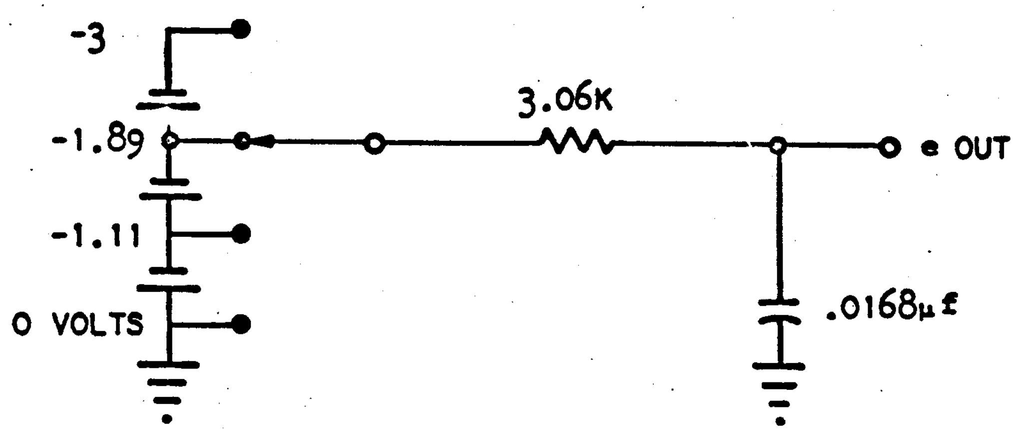

FIGURE A. 7. 2-3

EXPONENTIAL INTEGRATOR EQUIVALENT CIRCUIT

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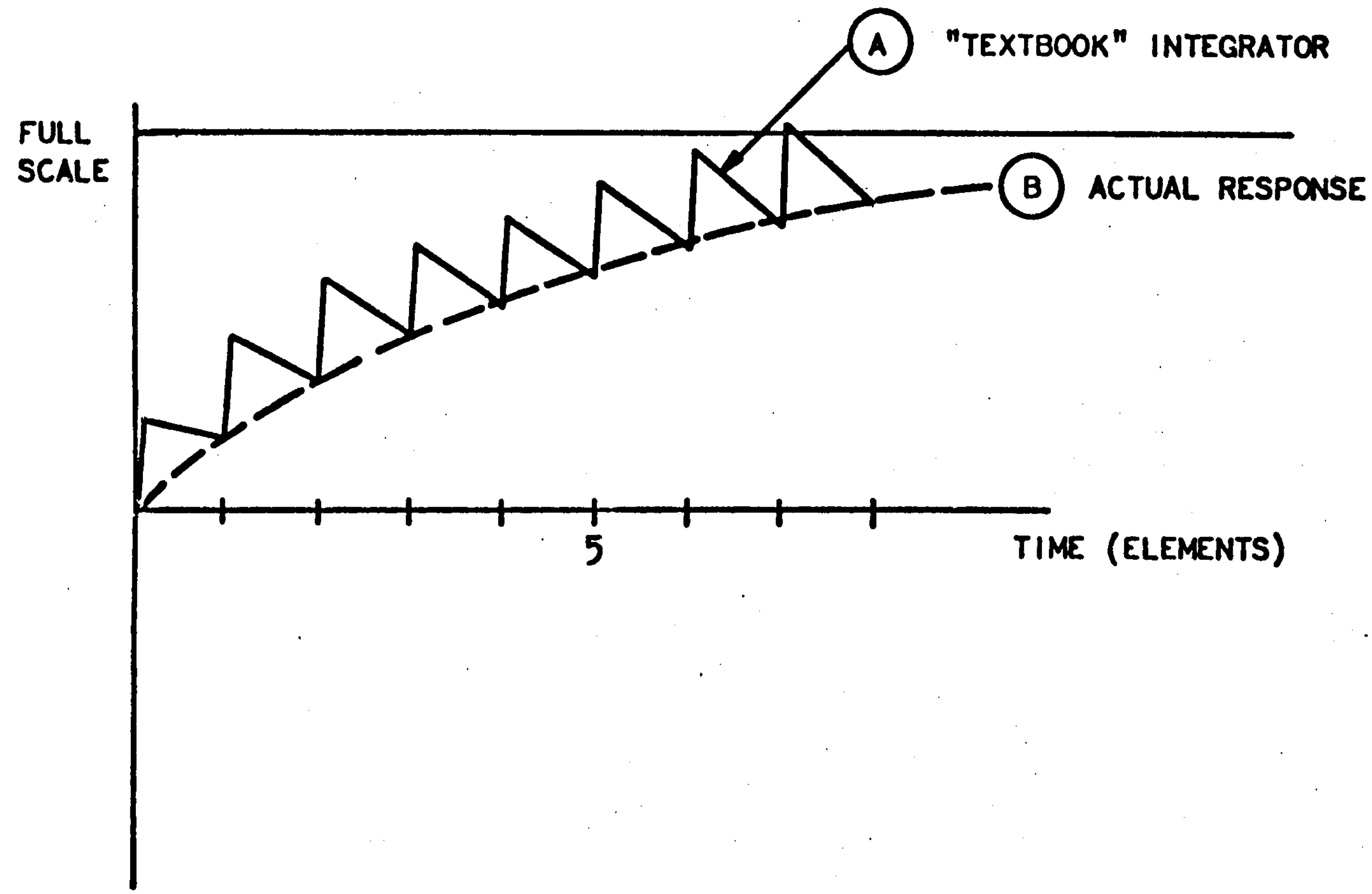

FIGURE A.7.2-4

EXPONENTIAL INTEGRATOR WAVEFORMS

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RE-ORDER No. 64-712
Vol I

RE-ORDER No. 64-712
Vol I

waveform is sampled at this equal time and displayed. Thus, the displayed voltage is identical to the textbook case.

This reconstructed video is then subtracted from the incoming video and the decision is again made. The continuation of this process results in the Delta approximation to the video.

A.7.3 Delta Results

The sample rate for the Delta system was chosen to be identical with the sample rate of the PCM and the other compression systems so that a valid pictorial comparison could be made. Since the decision element utilizes three thresholds, any of four corrective signals (two bits) are transmitted at each video sample time. This two bits of transmission per picture element results in a gross compression of 3:1.

As can be seen from photographs in Figure E.4P-16, the Delta image is "softer" than the PCM images. However, the Delta image at a bit error probability of 10^{-3} is probably acceptable relative to the Delta image with $10^{-\infty}$ bit error probability. So one concludes that (neglecting differences in synchronization bit requirements) Delta net compression approximately equals Delta gross compression.

Fidelity remains to be evaluated. The perceptible softness in the Delta images is more quantitatively exhibited in Figure E.2P-2. Comparing the Delta response in the figure to the response for PCM, it will be observed that the full-scale rise time for the Delta system requires 9 or 10 video sample times. In contrast, the EDITS PCM system has, at most, three video sample times per full-scale rise time. This indicates that the Delta system in effect is acting as a low-pass filter of about 1/3 the equivalent cut-off frequency of the PCM system. The overall softness of the lunar and arctic scenes provide subjective verification. From this observation, we see that the Delta system, as optimized and implemented on EDITS, requires $9 \times 2 = 18$ bits per rise time, while the PCM system requires $3 \times 6 = 18$ bits per rise time, also. In principle, one could low-pass the analog video into the PCM system by a factor of three, and with lowered sample rate, get the same pictorial result. Thus, on the basis of one fidelity criterion

RE-ORDER No. 64-712
Vol I

(equal rendition of step functions) the Delta system offers no compression advantage relative to six-bit PCM. This is not surprising since as early as 1957, Bowers[1] showed even a fairly complex Delta TV system as requiring over 40% more bits/second than a PCM TV system for equal (44 db) signal-to-noise ratio. Delta investigations have come full circle: from the initial concept of Delta, not as a "compression system" but as a cheap-terminal speech digitizing system, to the reluctant conclusion that Delta today is, once again, not a "compression system" but a (hopefully) cheap-terminal way of digitizing TV.

(1) Bowers, F. K., What Use is Delta Modulation to the Transmission Engineer?, Communication and Electronics No. 3, published by AIEE, May, 1957.

RE-ORDER No. 64-712
Vol I

A.8 SPLIT BAND

A.8.1 General

The history and general description of the split-band technique is already covered in the NASA sponsored literature.[1] Briefly, the split-band technique (Figure A.8.1-1) consists of treating the video as though it were composed of two distinct signals: a low frequency component and a high frequency component. Each component is compressed by an appropriate technique. At the receiver, each part is reconstituted and combined with the other to reform the original video. The low frequency part is formed by passing the video through a low-pass filter with cut-off frequency about 1/10 to 1/16 the maximum video frequency. The high frequency component is formed by passing the video through a complementary high pass filter.

Effectively, a compression technique for the low frequency component occurs simply because the sampling frequency for the lows is from 1/10 to 1/16 the sampling frequency needed for the whole video signal. Additional compression is obtained on the low frequencies by Roberts processing (see A.6). The compression of the highs is accomplished by detecting spikes in the highs which correspond to edge transitions in the original video. The spikes are encoded in amplitude and in relative position to one another. The exact technique employed on the lows has relatively small influence (in the order of ±20%) on the overall compression. For PCM or for Roberts encoding, the statistics of the lows signal has no influence on compression. However, the statistics of the highs signal has great bearing on the compression obtained. The statistics of interest are the distribution of run-lengths between the spikes and the average run-length between the spikes. In the previously cited NASA report the formula for split-band compression was derived as

$$\overline{C} = \frac{6}{ab + cd};$$

[1] Final Report, Manned Spacecraft Advanced Television Study, for NASA Manned Spacecraft Center, Houston, Texas, Contract NAS 9-991, 31 July 1963.

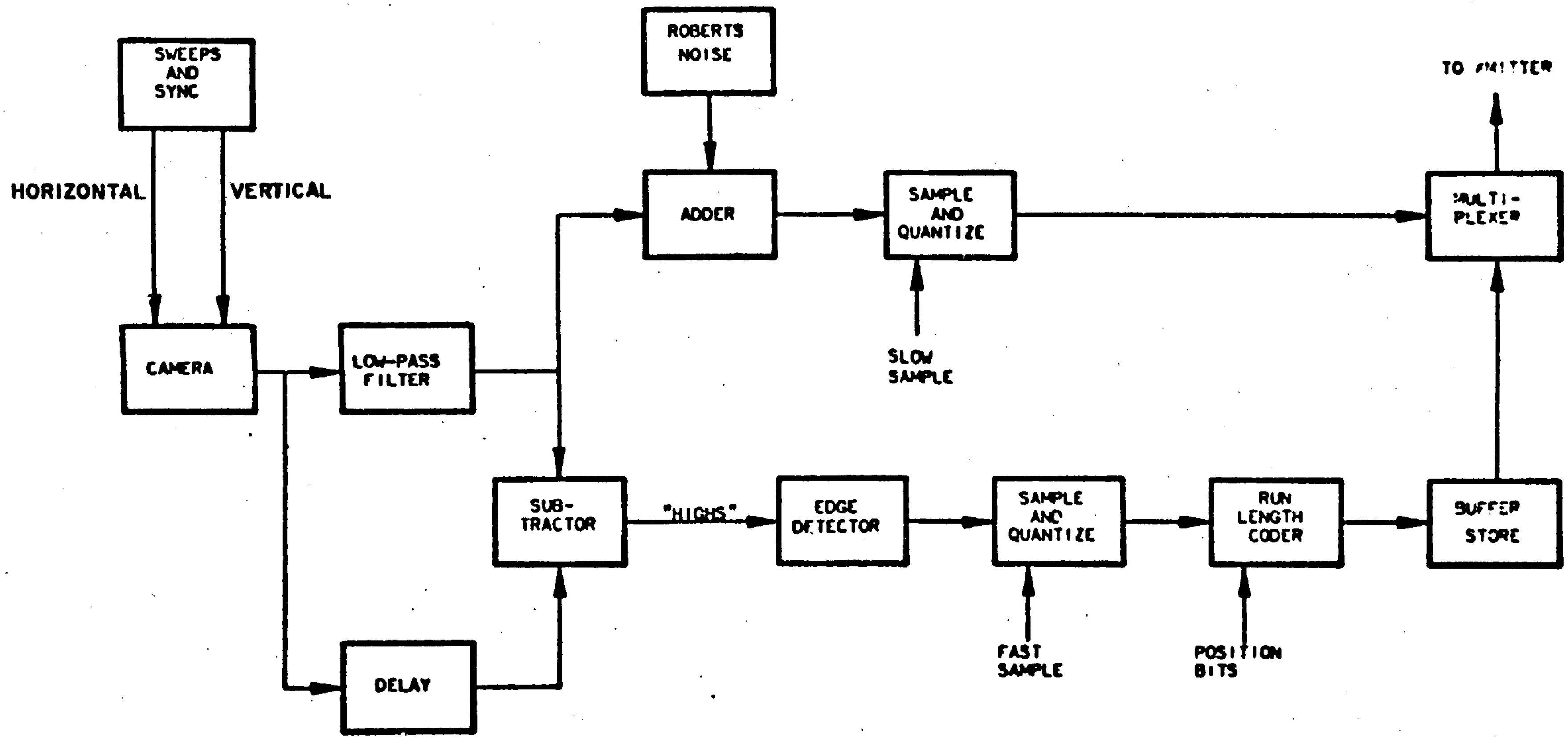

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FIGURE A.8.1-1
BASIC SPLIT BAND ENCODER WITH ROBERTS' COMPRESSION

RE-ORDER No. 64-712
Vol I

RE-ORDER No. 64-712
Vol I

where

- a = reciprocal average run-length (1/ average run length in elements)
- b = edge (spike amplitude), quantization in bits
- c = low pass frequency cut-off (fraction of total video bandwidth)
- d = lows, quantization in bits

Equation (1) gives rise to a nomograph, Figure A. 8. 1-2 reproduced from the cited literature.

A. 8. 2 Fidelity

Fidelity of signal reproduction is of prime concern in any reconnaissance system. In other TV applications such as entertainment, test observation, vehicular control, ordnance neutralization, etc., only a pleasing (nonfatiguing) picture of reasonable detail and contrast is required. Indeed, the literature on split-band and its variants indicate that it was explicitly developed to exploit the subjective characteristics of the human viewer. The eye is relatively insensitive to signal amplitude errors in regions of sharp transition, yet, to prevent contouring, it is demanding of rather fine quantization in broad grey areas. Typical system parameter choices reported in the very recent (fall 1963) literature[1] are three-bit (Roberts) quantization of the lows and three-bit (PCM) quantization of the highs. Although yielding acceptable subjective fidelity, these parameters may permit large instantaneous errors in the output video.

In split-band systems the video signal is split, separately processed to a given accuracy, and recombined; the maximum error between the output signal and the input signal is the sum of the maximum errors in each processing chain. The peak error in a three-bit quantization system (PCM or Roberts) is one-half a quantizing interval which is 1/16 of the peak-to-peak video signal. When the high and low channels are recombined, the peak error is doubled to 1/8

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[1] Gicca, F. A., Optimizing Space Television, International Telemetry Conference, London, 1963.

/60

$$\bar{C} = \frac{6}{AB + CD}$$

$\bar{C}$ = AVERAGE SPLIT-BAND COMPRESSION
A = AVERAGE EDGE RUN-LENGTH
B = EDGE QUANTIZATION BITS
C = LOW-PASS CUT-OFF POINT
D = "LOW" QUANTIZATION IN BITS

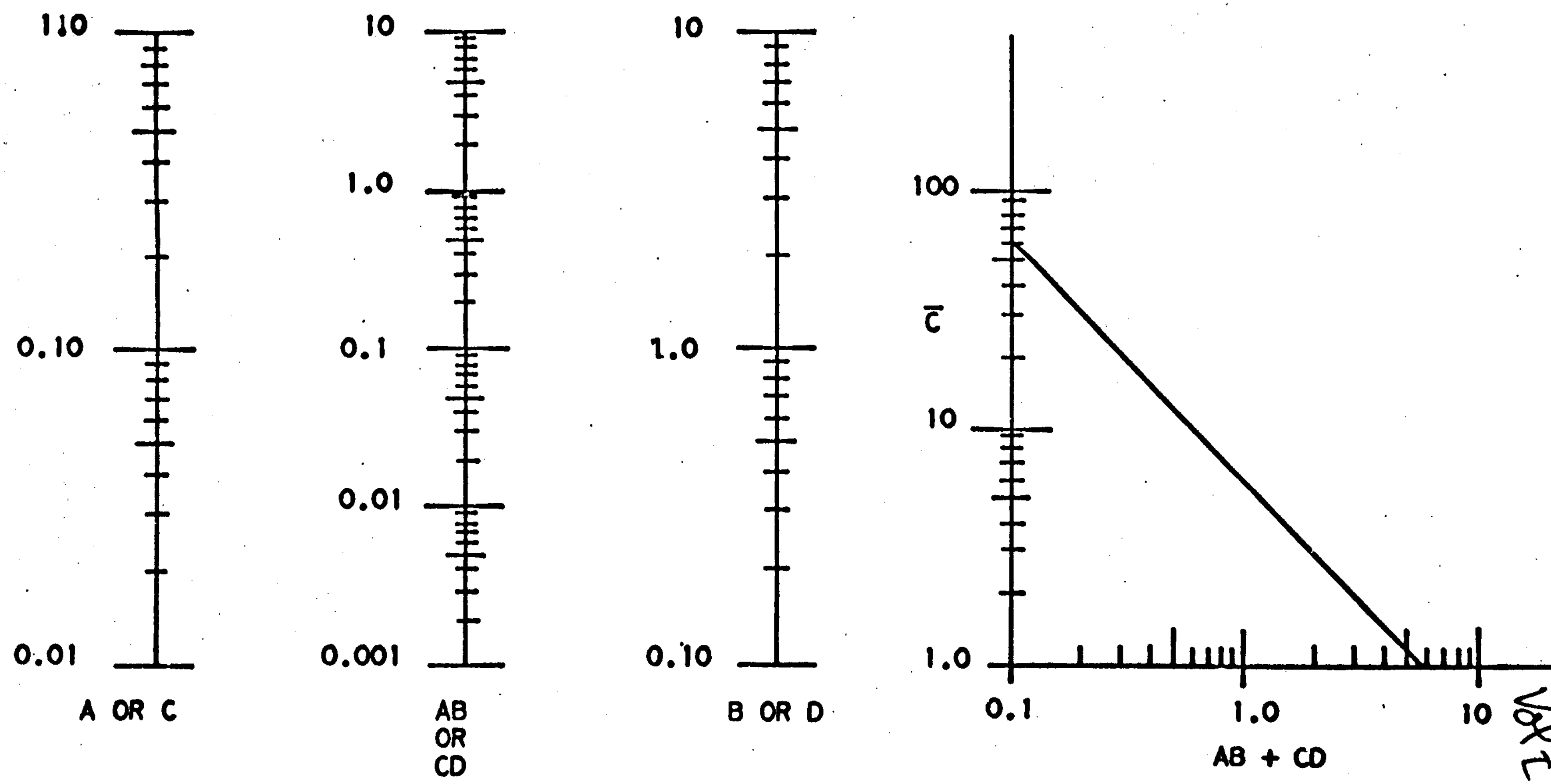

FIGURE A.8.1-2

NOMOGRAPH FOR AVERAGE COMPRESSION
OF SPLIT-BAND ENCODING

RE-ORDER No. 64-712 Vol I

(12.5%) of the peak-to-peak video signal. Possibly, even with this large maximum error the picture may still be subjectively pleasing for some subjects. Gicca provided no split-band pictures with his article. However, we do know that the lows processer is a three-bit Roberts system so the subjective fidelity can be expected to approximate the three-four Roberts system illustrated in the photographic volume of this report.

For photometric work the possibility of a 12.5% maximum error may eliminate a split-band system from consideration in the mind of the space sciences experimenter. On the other hand, perhaps his subjective judgment of split-band pictures of known objects may lead to a declaration that split-band fidelity is adequate. This can be determined only by implementing a representative split-band system and producing pictures -- a task which is beyond the scope of this particular study, but which deserves further consideration.

Other factors which weigh against using the split-band system also have a bearing on fidelity. A split-band system is complex. Either the video must be processed continuously, in parallel, and read into a large buffer store or it must be processed sequentially, a line-at-a-time, first for the lows, then for the highs. The latter system, called an elastic coding system, is reported by Gicca and Figure A.8.2-1 is adapted from his article.

Processing a given line of video twice involves reading it out of a sensor twice. While this technique can be accomplished using an image dissector camera tube or a flying spot scanner (on photocopy), this technique could have dire consequences when applied to ordinary sensors (such as vidicons) which normally erase the image as they read it out. If beam currents were low and target voltages adjusted accordingly, it may be possible to read "half" the signal each time -- once to process the lows and once to process the highs. However, such an operation would surely reduce dynamic range and sensor signal-to-noise ratio. This would ultimately show as a reduction in fidelity and pictorial quality. On the other hand, the sequential read-out technique could imply the necessity of using a storage sensor which could be read out many times and then erased during the interframe periods.

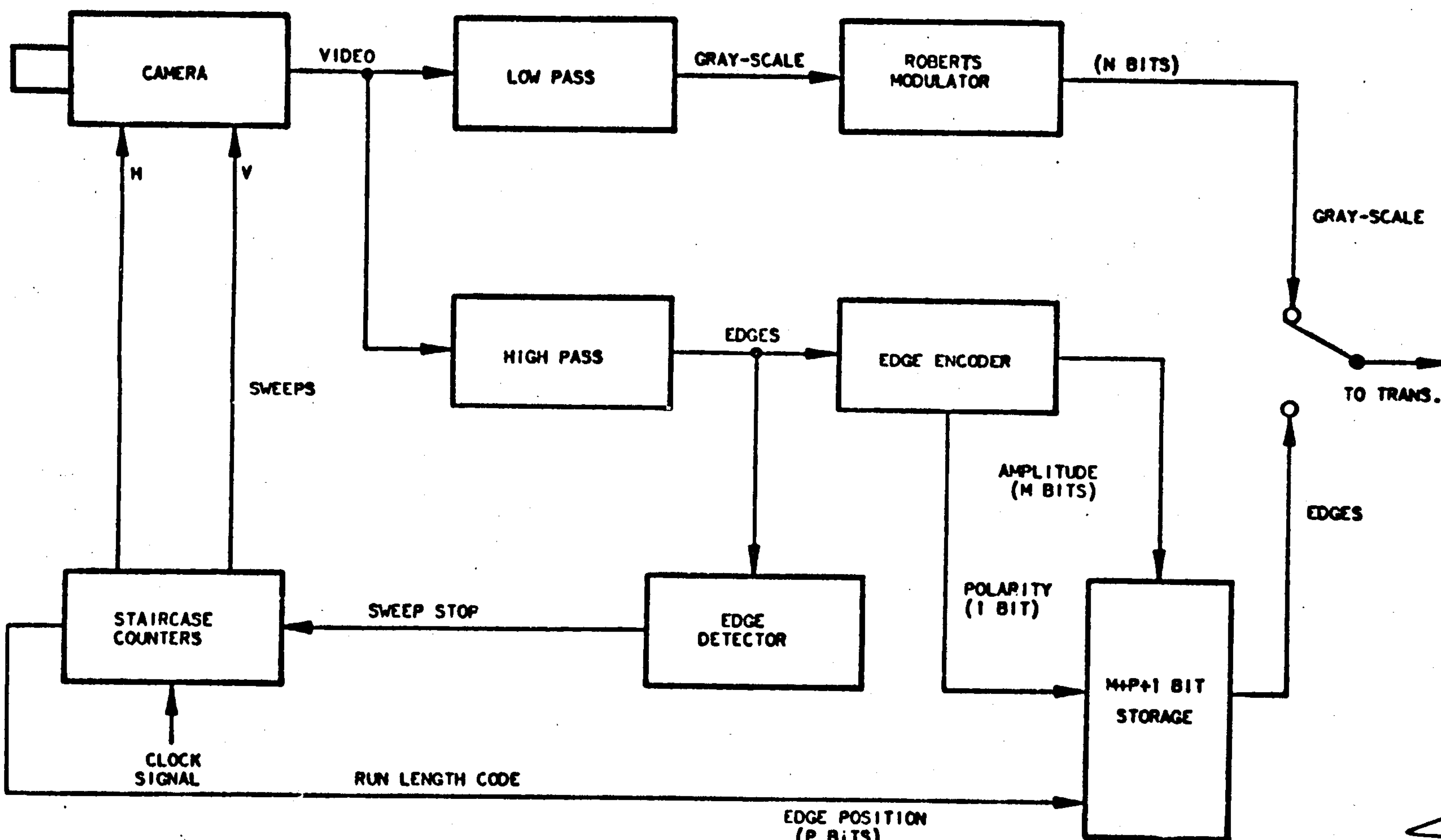

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FIGURE A.8.2-1

BASIC ELASTIC ENCODING SYSTEM

RE-ORDER No. 64-712
Vol I

RE-ORDER No. 64-712
Vol I

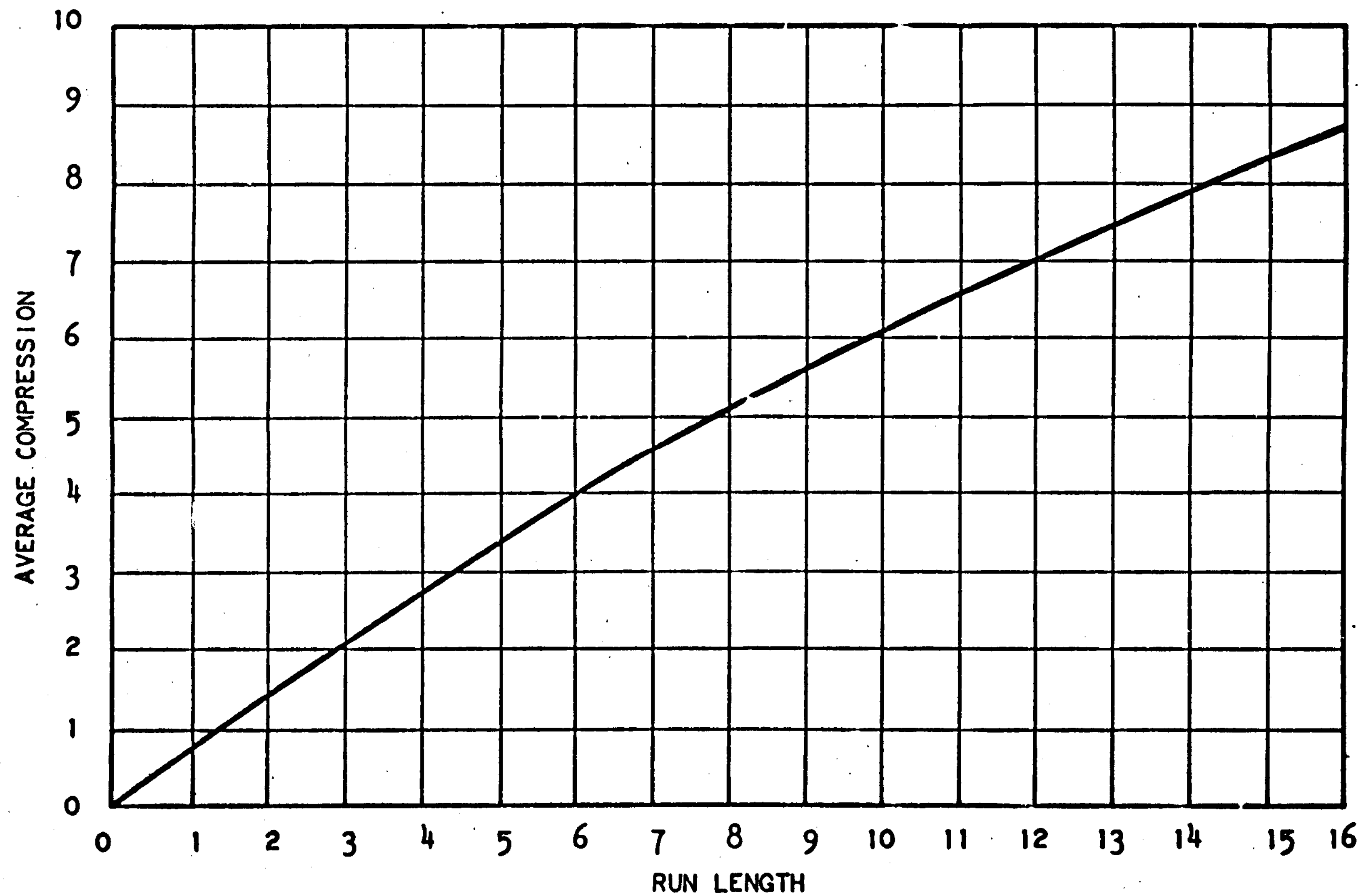
Finally, the elastic technique involves stopping the scan of the sensor when spikes in the highs video are detected and encoded. Unless properly done, start-stop scanning introduces very troublesome transients into the video from those sensors (such as vidicons) whose signal output amplitude is directly related to scan speed. In any event, the control of the scan beam by the compression system has been specifically excluded by the ground rules of this study which specify that continuous video is to be fed to the compression system. Nevertheless, it is felt that the split-band concept of compression should be discussed within the context of this study because, for all the apparent fidelity problems, it undoubtedly possesses the compression potential to meet the JPL requirement of two, minimum, and three, desired.

A.8.3 Compression

Applying the parameters of paragraph A.8.2 to equation (1) of paragraph A.8.1 yields a gross compression for split-band of 6.08. This assumes that the average run length is, as reported, about 10 picture elements and that the low-pass frequency cut-off is 1/16 the total video bandwidth. The highs are quantized to three bits, but an additional bit is required to indicate spike polarity, and four additional bits to locate the given video spike from the previous spike for a maximum distance of 16 picture elements. The compression figure is extremely sensitive to variations in average run length. Figure A.8.3-1 plots compression versus run length for the other parameters held as before.

The average run length can be made longer by setting the highs threshold at a larger value. This ignoring of lower amplitude edges provides the basis of trade-off between fidelity (rendition of small-edge detail) and compression. There is no indication in Gicca's article of the highs threshold setting which is necessary to get an average run length of 10, nor is there any indication of the sensor signal-to-noise ratio. It has been found that a noisy sensor can effectively break up long runs of video, and split-band does make use of a form of run-length coding.

RE-ORDER No. 64-712
Vol I



FICURE A.8.3-1

SPLIT-BAND GROSS COMPRESSION VS AVERAGE RUN LENGTH

07314

RE-ORDER No. 64-712
Vol I

A run-length coding system is quite susceptible to communication channel errors. For the parameters listed, about 80% of the bits will pertain to the highs signal and the remainder to the lows. Of the highs bits, 50% pertain to relative edge amplitude (including sign) and 50% to edge location. From these percentages it is evident that 40% of the errors in the bit stream will displace not only the edge location of the immediately affected edge, but will displace all subsequent edges (since edge location is relatively addressed to all previous edges) until the end of the TV line when resynchronization will occur.

In order to disrupt no more than a couple of lines per frame, it is necessary to raise the signal-energy-per-bit in the channel to lower the bit error probability below the 1×10^{-3} which is deemed acceptable for PCM. To a first (conservative) approximation, gross compression should be divided by 1.8 (see Section 2.2). Applying this factor to the gross compression figure of 6.08 yields a net compression of 3.38--still a very desirable figure relative to the requirements of this study. A run length between 5.4 and 8.8 will give net compressions between 2.0 and 3.0.

However, before the above figures can be accepted at face value, and hence result in split-band being considered as a recommended video data modulation system, at least the information listed below, in addition to that given in the recent literature, is needed.

(1) Sensor type.
(2) Sensor signal-to-noise ratio.
(3) Highs threshold setting (or other analytical fidelity measure).
(4) Experimental (pulse counting) verification of postulated run lengths.
(5) Pictures of
   (a) Original test material.
   (b) Original test material rendered by the given sensor and six-bit PCM.
   (c) Original test material rendered by the given sensor and split-band.
   (d) Split-band pictures corrupted by variable bit error probability channel noise.

RE-ORDER No. 64-712
Vol I

A.9 LINEAR APPROXIMATION

A.9.1 Introduction

In theory the concept of run-length coding offers the potential of achieving large (greater than 4:1) compressions as seen from Equation (46) which gives the average compression for run-length coding. The average compression is

$$\bar{C}_{RL} = \frac{N\overline{RL}}{N + R}, \quad \text{where} \tag{46}$$

N = number of bits to encode intensity levels,

$\overline{RL}$ = average constant intensity run-length, and

R = number of bits to encode the run-length distance.

Figure A.9-1 plots theoretical run-length compression for six-bit quantization as a function of average run-length for flat encoding of the run-lengths. It is evident from the curve that if long average run-lengths exist, gross average compressions greater than 4:1 could be achieved. However, for six-bit quantization it can be shown that with practical sources the maximum theoretical run-length is four elements; therefore, the maximum possible run-length compression from equation (1) is 2.67:1 if it is assumed that the run-lengths are truncated at eight elements. When the actual redundancy in a picture is coupled with the increase in signal power required for the compressed data, the net compression for run-length coding decreases considerably (at least by a factor of two).

Run-length coding can be thought of as the coding of zero-slope signals. If, in addition, all constant slopes are run-length encoded along with the zero slopes, there again exists the possibility of obtaining large compressions. It is this idea that leads naturally to the linear-approximation class of compression systems. Although the promising concept of run-length coding of constant slopes is not a new idea, there does not yet exist (to EMR's knowledge) an implemented system for video data compression.

The linear approximation coding technique can be thought of as a Delta-line system coupled with run-length coding. In a conventional Delta system, constant amplitudes force the Delta steps to alternate in polarity, and the average run-length of these intervals is one. This makes it difficult to achieve compression

RE-ORDER No. 64-712
Vol I

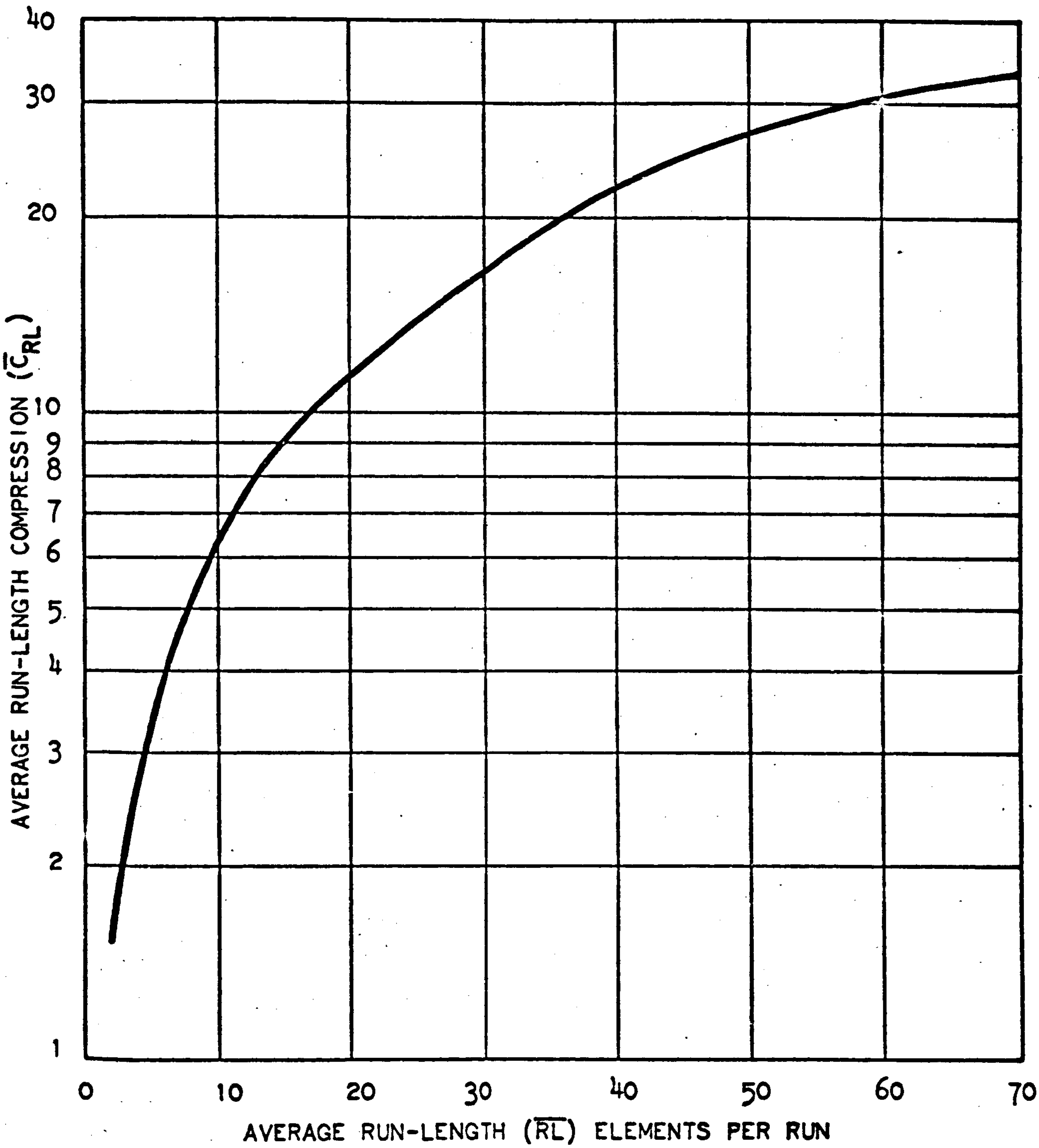


FIGURE A. 9-1

AVERAGE RUN-LENGTH COMPRESSION AS FUNCTION OF AVERAGE RUN-LENGTH ASSUMING FLAT ENCODING OF RUNS FOR SIX-BIT QUANTIZATION

07314

RE-ORDER No. 64-712
Vol I

by run-length coding a Delta system since the alternating polarity occurs with a high probability as compared to the constant slope probabilities. There are several approaches to a linear-approximation system. Before considering these, a brief background of the basic conept will be given.

Youngblood [1], [2] conceived the approach which he called piecewise-linear-approximation coding. Figure A. 9-2 illustrates the approach, which operates as described below.

The origin and sample point 2 are connected by a straight line. The straight-line value at sample point 1 is compared to the actual sample point 1. If the error between the two points does not exceed a specified error criterion, the process continues and sample point 3 is connected to the origin by a straight line; sample points 1 and 2 are then compared to the straight line to determine whether they satisfy the error criterion. However, if the error exceeds the error criterion, a straight line is connected to sample point 1, which then becomes the up-dated origin and the process continues.

This particular technique was computer simulated at MIT and the results are tabulated in Table A. 9-1. Subjectively, the pictures were quite acceptable; however, with the ±5 error criteria there were some noticeable artifacts present. Nevertheless, these results do indicate that the constant-slope run-length coding technique has the potential of achieving a large gross compression.
Medlin [3] applied a modification of this technique to telemetry data and achieved very large compressions. However, a computer was used to process the data and the technique proposed was essentially that of sending constant first differences with a floating error band. This technique is not as efficient as the peicewise-linear-approximation technique, but it is easier to implement.

(1) Youngblood, W. A., Picture Processing, Quarterly Progress Report, M.I.T., Research Laboratory of Electronics, January, 1958.

(2) Youngblood, W. A., Estimation of the Channel Capacity Required for Picture Transmission, DSc Thesis, M.I.T., June, 1958.

(3) Medlin, J.E., The Comparative Effectiveness of Several Telemetry Data Compression Techniques, International Telemetering Conference, London, 1963 Vol. I.

RE-ORDER No. 64-712
Vol I

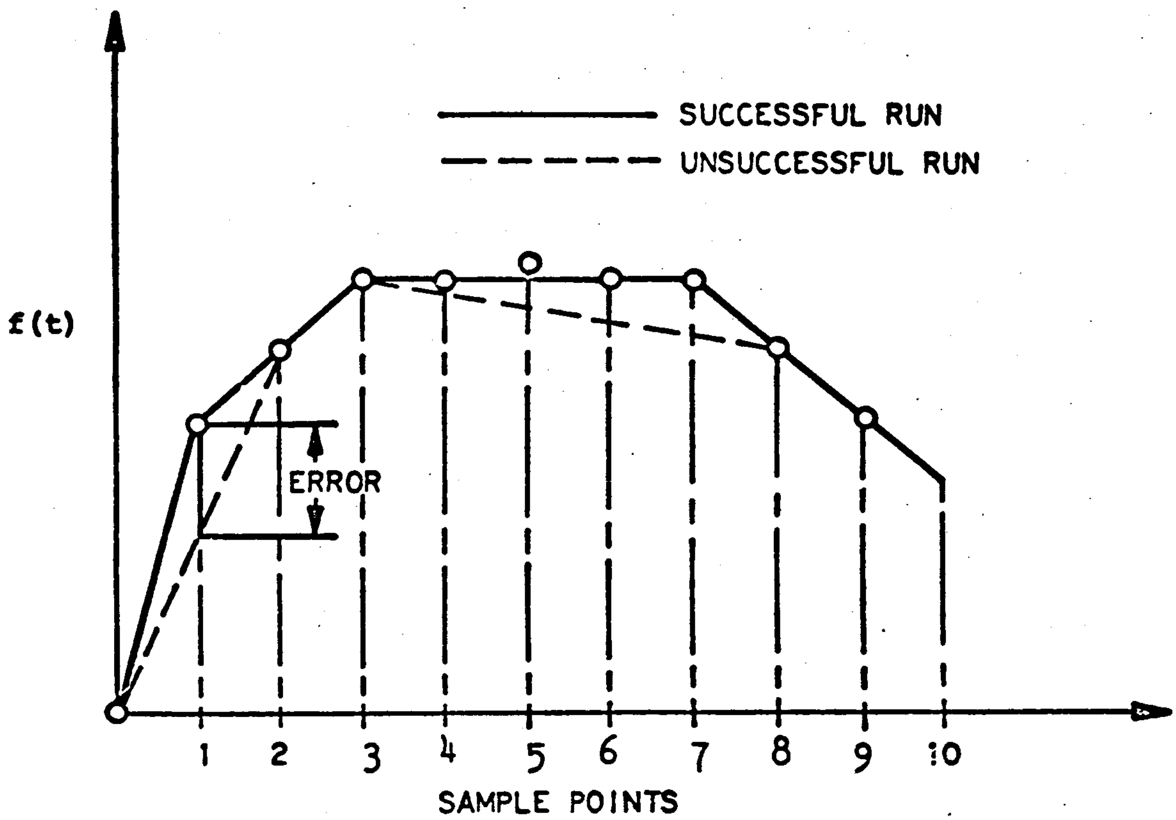


FIGURE A. 9-2

PIECEWISE LINEAR APPROXIMATION ENCODING

07314

RE-ORDER No. 64-712
Vol I

TABLE A.9-1

MIT PIECEWISE-LINEAR-APPROXIMATION COMPUTER RESULTS

Picture	RL (Elements)	Entropy (Bits/Element)	Error Criterion	Gross Average Compression
MIT girl	7.46	1.29	±3 out of 64 levels	4.55
MIT girl	10.12	0.97	±5 out of 64 levels	6.19
MIT crowd	2.48	3.34	±3 out of 64 levels	1.80
MIT crowd	3.21	2.79	±5 out of 64 levels	2.15

McClure(1) processed a large group of aerial photographs on a computer and obtained estimates on the compression possible by constant-slope run-length coding. The estimates were relative to six-bit quantization and indicated that compressions of two to three are possible. However, if some degradation can be tolerated (as in Youngblood's error-band technique), compressions will be much larger. He concluded that if one is concerned with contrasts rather than absolute intensities, errors in slopes can be tolerated if they are not too frequent.

A.9.2 Linear-Approximation

Late in the JPL Video Data Modulation Study, EMR independently arrived at a linear-approximation coding technique similar to Youngblood's, but with several distinct differences as described below. This linear-approximation coding technique will eventually become an integral part of EDITS.

Several approaches are possible for a linear-approximation coding technique. Basically, the idea is that given a function, $f(t)$, the function can be represented by a series of straight lines. It will be assumed that at $t = 0$, the function is zero; with this assumption, all line segments will be relative to the origin and the receiver can then reconstruct the approximated waveform.

Consider now one of EMR's approaches to linear approximation in which the function is specified by a number of slopes and Δx's. Assume that the function can be approximated by 32 slopes from $\pm 90^\circ$ in increments of 6° with the necessary number of Δx's. The function and an initial slope are sampled at the Nyquist rate

(1) McClure, G.W. and Dute?, J.C., Probability Characteristics of Sensor Output Data, Institute of Science and Technology, The University of Michigan, September, 1963.

RE-ORDER No. 64-712
Vol I

and if the difference exceeds a specified threshold, T, the distance from the origin, Δx, is transmitted along with the slope number. Consider Figure A.9-3 where the approach has been graphically plotted and compared to four-bit PCM encoding.

Several interesting comparisons can be made from this curve. The value of PE is 0.5 for four-bit quantization, and from this the average compression for PEC is

$$\bar{C}_{PEC} = \frac{4}{5-0.5(4)} = 1.33:1 \; . \tag{46}$$

For the run-length coding, the average compression is given by

$$\bar{C}_{RL} = \frac{4\overline{RL}}{(4+K)} = \frac{4(2)}{8} = 1:1 \; .$$

K is the number of bits required to specify the run-length. Given an average run-length of two elements, there is no compression.

The Δx-slope compression technique requires P bits to specify the slope and R bits for the Δx. The compression is

$$\bar{C} = \frac{N\,E}{(P+R)S} \; . \tag{47}$$

E is the number of elements in a line, S is the number of segments required to specify the function, and N is the number of PCM bits for intensity that is equivalent to the linear-approximation fidelity. From Figure A.9-3, E=80, S=9, P=5, and N=4; therefore, for the linear-approximation technique, the average compression is

$$\bar{C} = \frac{4 \times 80}{10 \times 9} = 3.56:1.$$

In order to appreciate the advantages of the linear-approximation technique, the average run length for run-length coding and linear-approximation coding will determined. For run-length coding of a four-bit quantization of f(t) the average run length is

$$\overline{RL} = \sum_{y=1}^{K} \sum_{i=1}^{16} y\, P(R_i^{\,y}) = 2 \text{ elements.} \tag{48}$$

112

RE-ORDER No. 64-712
Vol I

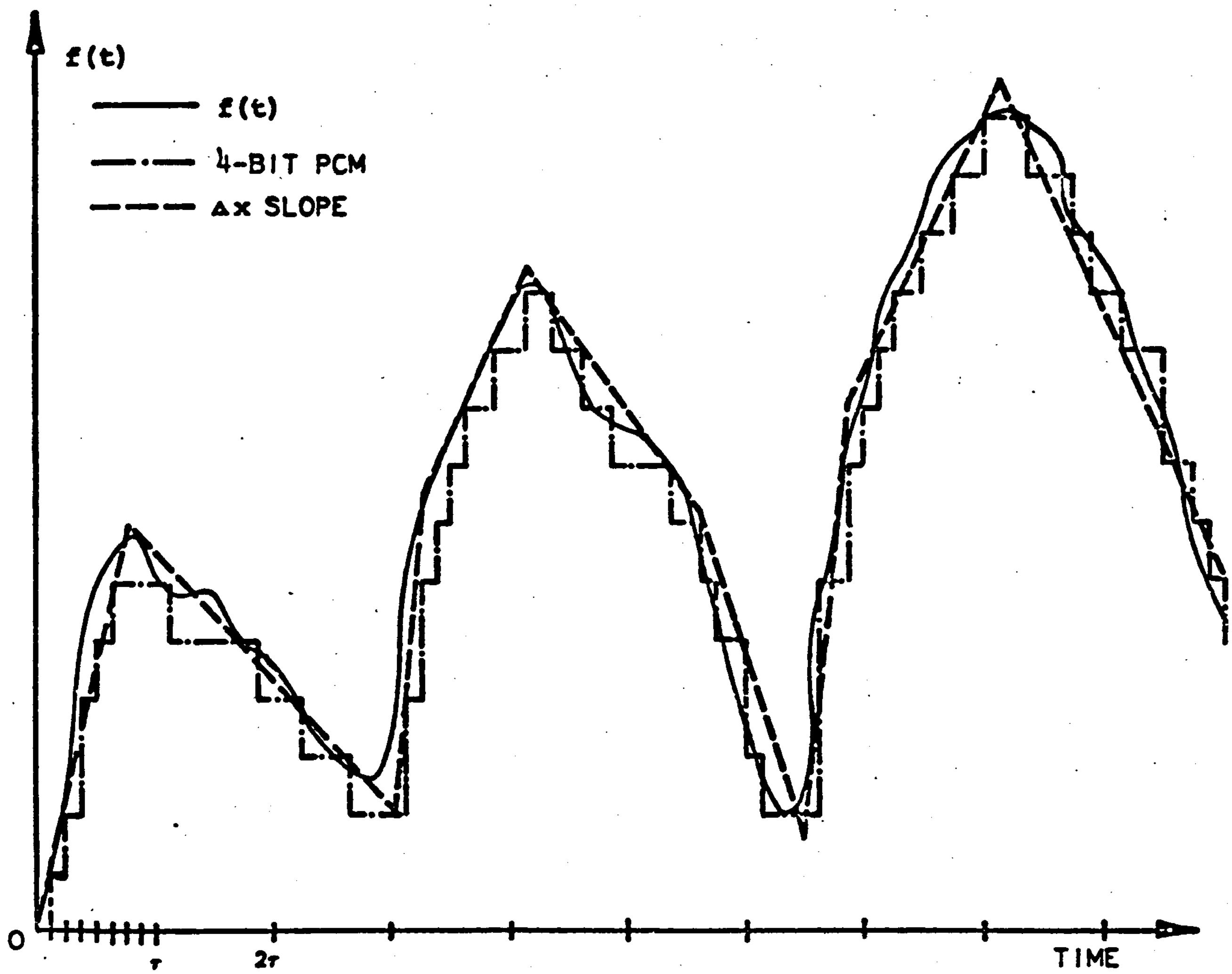


FIGURE A. 9-3

FOUR-BIT PCM COMPARED TO Δ X-SLOPE LINEAR APPROXIMATION

07314

RE-ORDER No. 64-712
Vol I

K is the maximum run-length. However, the average run-length for the linear approximation is

$$\overline{L} = E/S = 8.89 \text{ elements.} \tag{49}$$

This is over four times that of conventional run-length coding. This increases the compression approximately by a factor of four. Another feature of the coding technique is that it produces a continuum of gray scales similar to analog video rather than producing PCM discrete steps which can create annoying contours.

An alternate approach to linear approximation is to transmit the amplitude and Δ x where the slope of the approximating function can be determined from these two quantities and must always be within one quantizing level of the function f(t). Figure A.9-4 graphically represents both the Δx-slope and Δx-amplitude techniques with an error band plus and minus one quantizing level of f(t) wherein the approximation must lie. For the Δx-amplitude technique the compression is 3.45:1, and for the Δx-slope approach the compression is 3.2:1.

Basically, the Δx amplitude approach is analogous to taking f(t) and displacing it plus or minus one quantizing level, forming a new composite function within which the approximating function must remain. This is illustrated in Figure A.9-5.

It would be very desirable to obtain an analytical expression that relates the zero slope average run-length to the composite average constant-slope run-length. If this was determined, it may be possible to measure PE and obtain the linear approximation compression since it has been shown in area coding that PE can be related to the average zero-slope run length. Whatever compression one can achieve with run length coding, the linear-approximation technique gives a larger value of compression.

A.9.3 Computer Simulation

A program was written for the ASI-210 digital computer to simulate and evaluate the linear slope approximation on a line-by-line basis.

The computer sequentially processes a single line trace and utilizes a mathematical model in its memory to approximate the trace by a series of linear slopes.

By appropriate choice of sense-switch settings, the following may be read out:

114

RE-ORDER No. 64-712
Vol I

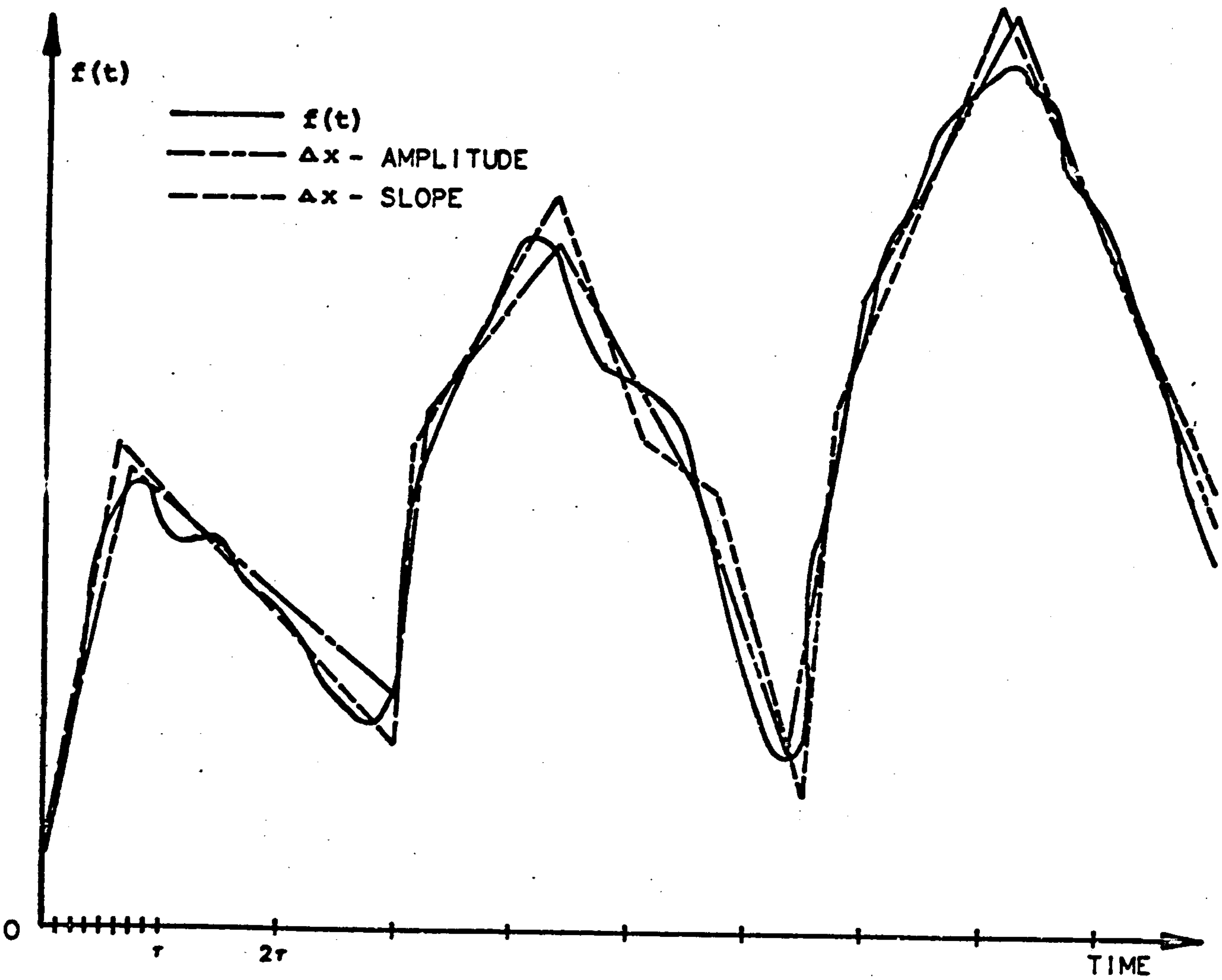


FIGURE A. 9-4

COMPARISON OF ΔX-SLOPE AND ΔX-AMPLITUDE APPROXIMATIONS

07314

115

RE-ORDER No. 64-712
Vol I

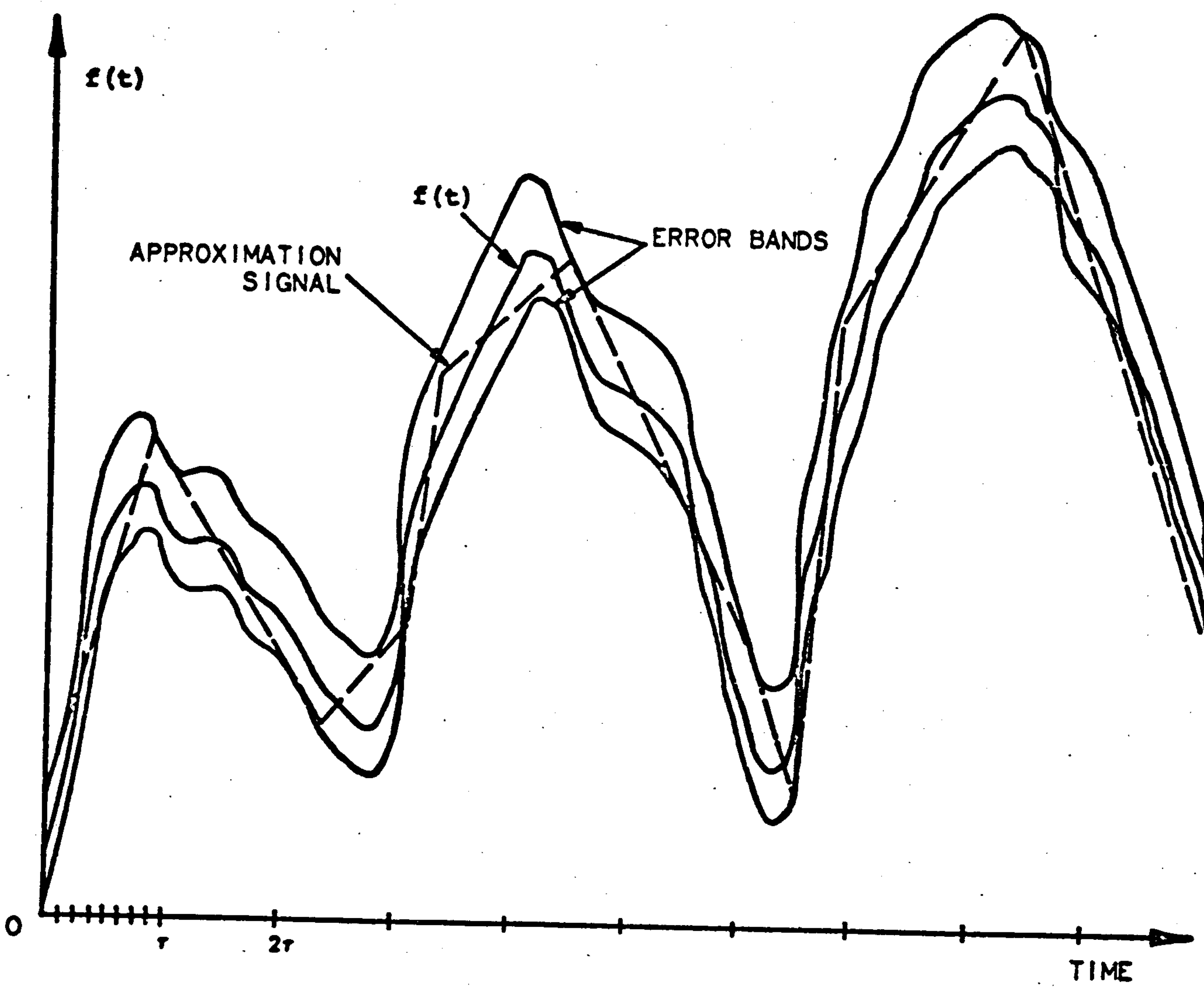


FIGURE A.9-5

ΔX-SLOPE APPROXIMATION WITH ERROR BAND

07314

RE-ORDER No. 64-712
Vol I

a slope-by-slope description of how the system is processing the TV line; the lower-bound on compression with flat coding (i.e., the ratio of total bits required to transmit the line by PCM as opposed to the linear-approximation system); and data tapes linking auxiliary programs. Through the use of these auxiliary programs the options are: (1) the probability of a particular slope may be determined; (2) the probability of a particular run length (length of slope, Δx in elements); (3) the average slope run length; (4) the average mathematical entropy of the source; (5) the upper-bound on compression (obtainable through use of the mathematical entropy); and (6) statistics on the errors due to the approximation technique (based on signal-in-minus-signal-out on an element-by-element basis). Another auxiliary program will average linking data tapes for secondary data on the basis of multiple lines or multiple parameter variations.

Various parameter combinations can be obtained by specifying the appropriate initial conditions. It is necessary to specify how closely the slope approximations are to conform to the original signal. The error band is the number of quantizing levels, plus or minus, that the approximated signal may deviate from the actual one. In general, a run is allowed to continue as long as the absolute difference between the actual and the approximated signal is less than or equal to the error band. When this condition is exceeded, the run is terminated at the element previous to the violation, the slope being used is discarded, and a new one is initiated. Sometimes it is not desirable to terminate a run at the discovery of a single "error". A single "error" may be the result of signal noise and it might be advantageous to disregard it. If it is desired that a slope not be abandoned until more than one consecutive error is recorded, a parameter, maximum consecutive errors, may be so adjusted. For example, if maximum consecutive errors is set equal to zero, no errors will be tolerated; that is, the first error will disqualify the slope--if it is equal to one, a run will continue until two errors in successive elements disqualify the slope, and so forth. Besides the error band and maximum consecutive errors being variable, the truncation is also variable. The truncation is the longest run length allowed. When a run reaches the maximum length, it is terminated regardless of error band violations. In addition to the trade-off between the number of runs needed per line, the number of bits needed to specify a run's

117

RE-ORDER No. 64-712
Vol I

length the hardware limitations in a real system (which is being simulated) will force the imposition of a truncation.

Four additional parameters are relatively constant over a number of program runs. They are: (1) the resolution (the number of elements in the line); (2) the number of slopes which specifies the bits per slope code (the smallest integer greater than or equal to the logarithm to the base 2 of the number of slopes); (3) the magnitude of the angle of the slopes; and (4) the number of bits to which the data are quantized (the number of bits required to specify the brightness level if ordinary PCM were used).

A flow diagram of the computer program is given in Figure A. 9-6.

The computer program can be summarized as follows: The program to simulate the linear approximation coding technique has the following options: (1) the Δx run lengths were truncated to obtain compression as a function of truncation of the Δx run lengths; (2) the number of slopes were variable(31 and 63); (3) the error band within which the function must be was variable (±1, ±2, and ±4 levels out of 64 possible amplitude levels); (4) the number of consecutive errors that could be tabulated was variable (0, 1, and 2 errors); (5) the RMS error and the equivalent PCM error could be obtained; (6) the Δx-run-length probabilities could be obtained; and (7) the compression could be obtained for either flat coding or optimum coding.

Although Figure A. 9-6 gives the flow diagram of the generalized linear approximation compression program in Fortran I, part of the program was rewritten in Fortran II for the ASI-210 to allow formatting of the output data.

The linear approximation program in Fortran II has the following options:

(1) for linear approximation, the variables are

- (a) number of slopes (31 or 63),
- (b) Δx truncation (64 to 4 elements),
- (c) error band (±1, ±2, ±4 levels out of 64 levels), and
- (d) tolerable errors outside error band (0, 1, 2 errors).

(2) For PEC coding they are

- (a) value of PE, and

118

RE-ORDER No. 64-712
Vol I

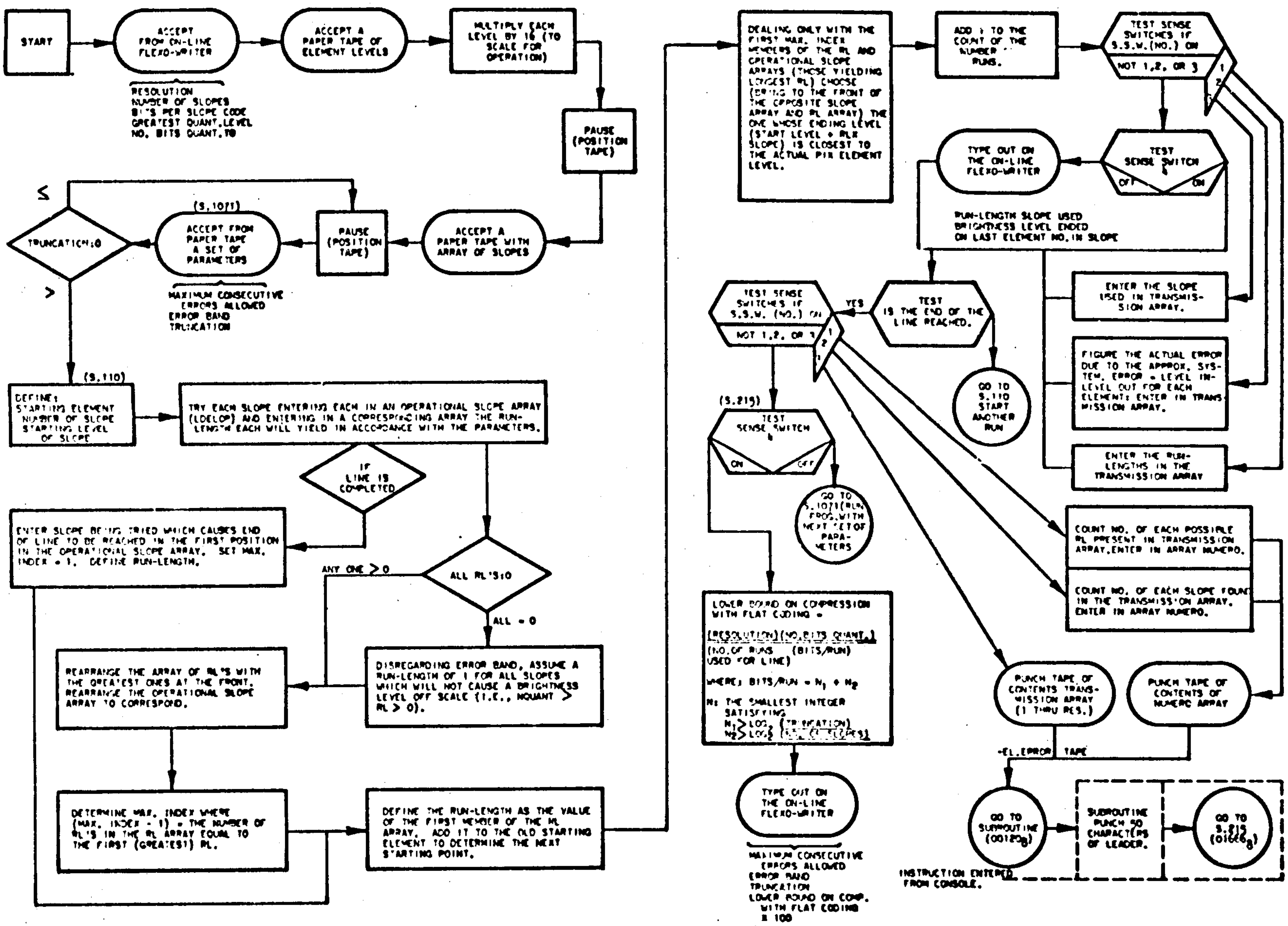


FIGURE A. 9-6

LINEAR APPROXIMATION FLOW DIAGRAM
FOR GENERAL FORTRAN I PROGRAM

RE-ORDER No. 64-712
Vol I

(b) the flat coding value of PEC compression

(3) For PEQ (previous element qualifier) the options are

(a) linear approximation coding compressions results after the line has been PEQ'd and

(b) the PEC results after the line has been PEQ'd.

Figure A.9-7 represents the flow diagram for the generalized overall linear approximation program including the three options. Figure A.9-8 represents only the linear approximation program of the generalized program while Figures A.9.9 and A.9-10 represent the associated PE and PEQ programs respectively.

A.9.4 Computer Results

Although it is necessary to subjectively evaluate a compression technique to determine whether or not it will meet a desired compression goal, the computer simulation will determine if the compression technique will meet the JPL compression goal. For the linear approximation technique, fifteen lines were processed on the ASI-210 computer. The photo-indentification and the number of lines from each photo processed are given in Table A.9-2.

TABLE A.9-2

IDENTIFICATION OF COMPUTER PROCESSED LINES USING LINEAR APPROXIMATION CODING TECHNIQUES

Photo	Number of Lines
Lunar Photo 66-2	1
Lunar Photo 66-1	3
Canadian Arctic A-15420-43	3
Tiros Cloud (Frame 24)	3
Tiros Cloud (Frame 18)	1
MIT Girl	4

Table A.9-3 tabulates some of the results of the various parameter combinations. Several remarks regarding these computer results can be made.

As an example of how computer results were used to determine the final system design of the linear approximation coding technique, consider the computer results for lines 1, 2, 3, and 4. For this particular case it is clearer if the results are graphically summarized.

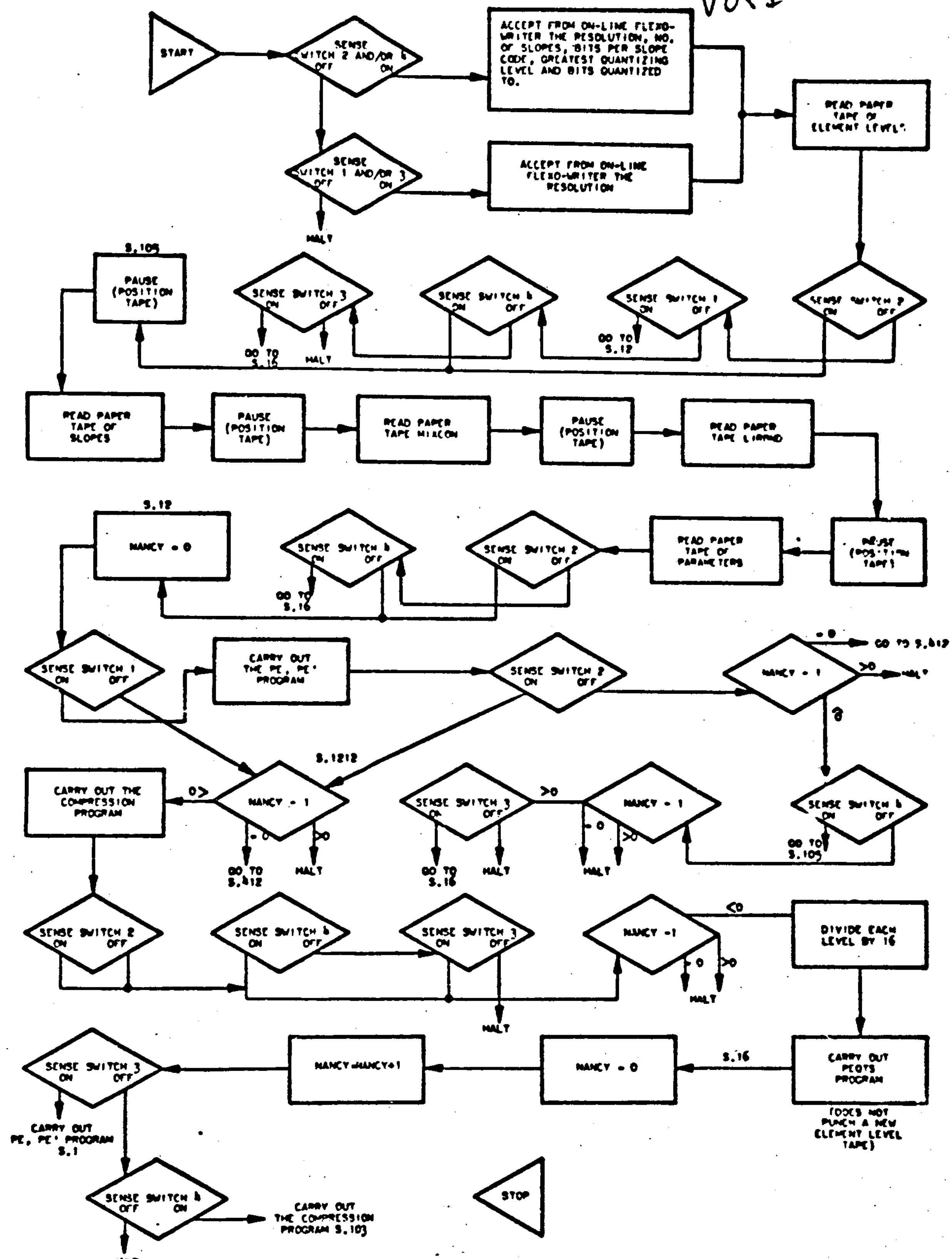


FIGURE A. 9-7

COMPOSITE LINEAR APPROXIMATION PE AND PEQ PROGRAM FLOW DIAGRAM IN FORTRAN II

RE-ORDER No. 64-712
Vol I

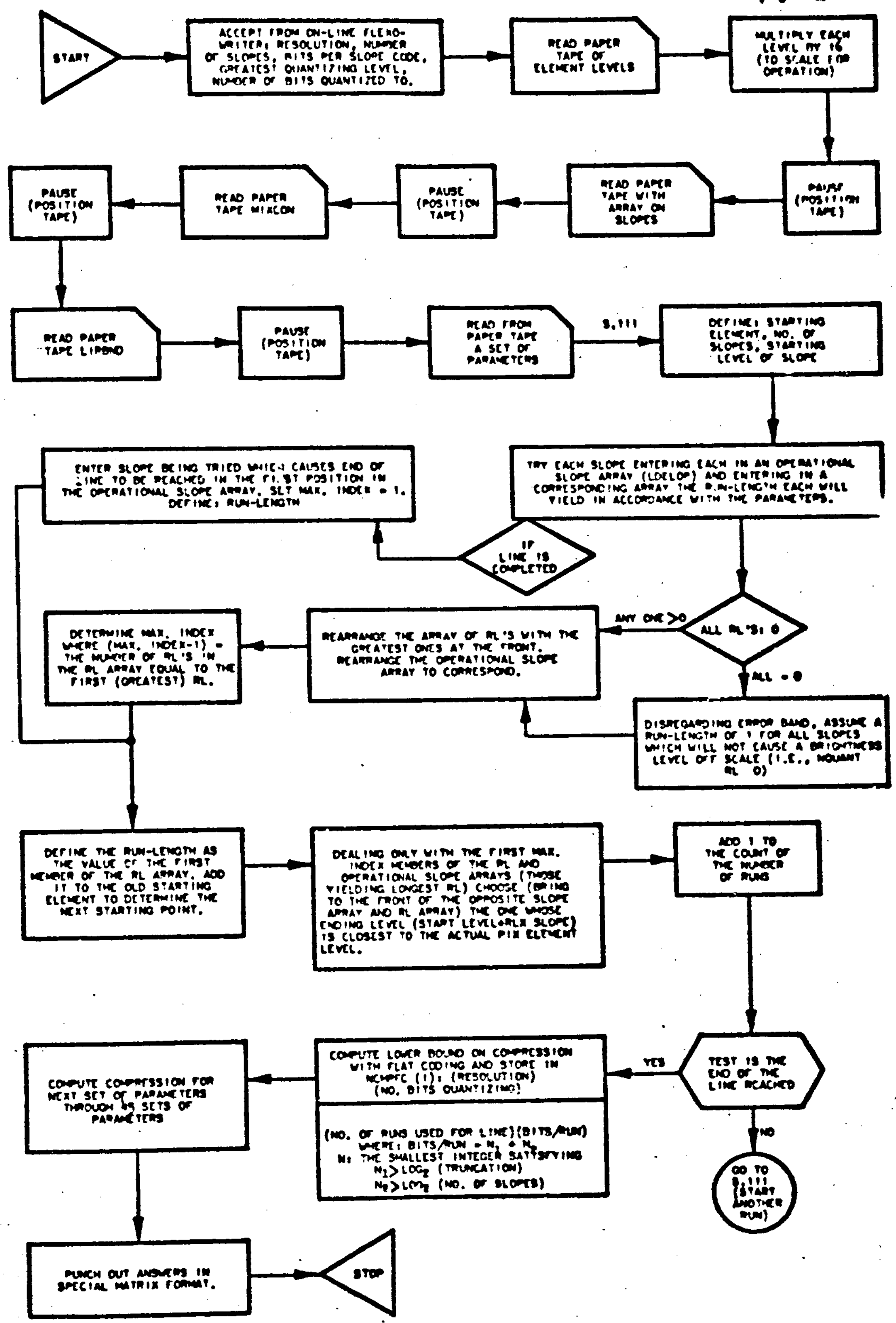


FIGURE A.9-8

LINEAR APPROXIMATION COMPRESSION PROGRAM IN FORTRAN I

RE-ORDER No. 64-712
Vol I

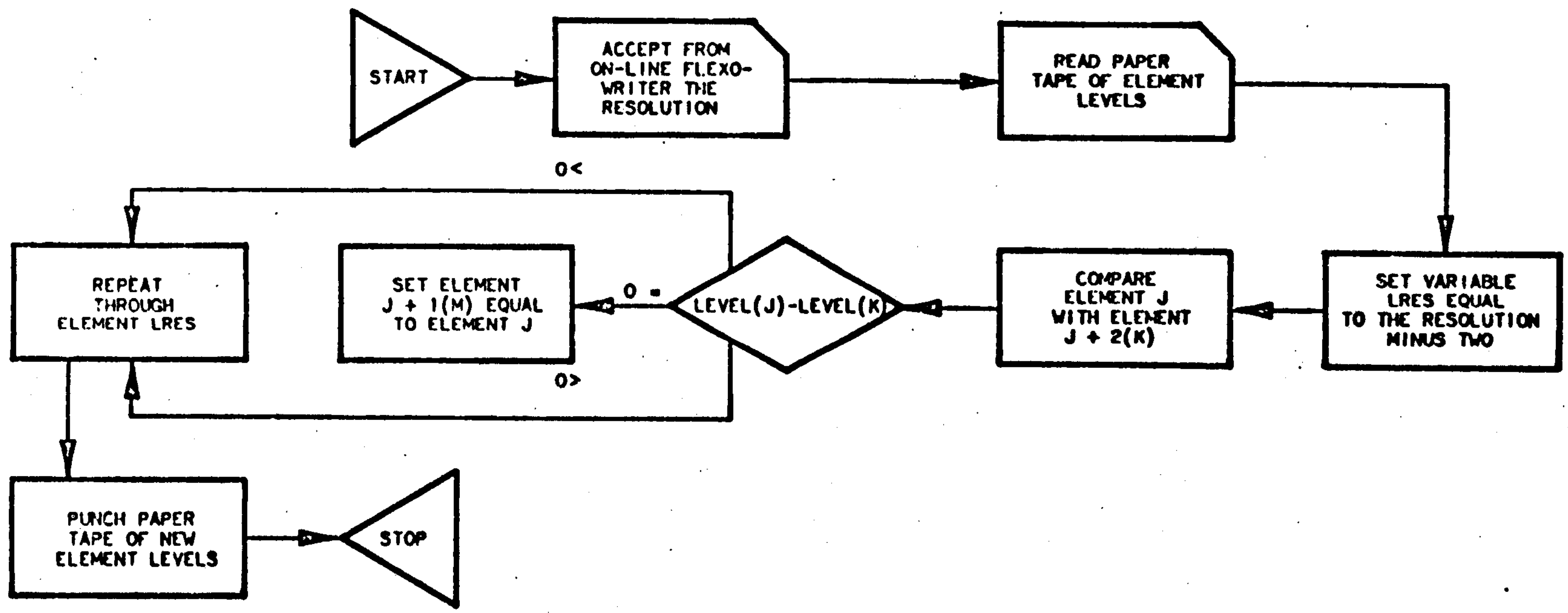


FIGURE A.9-9

FLOW DIAGRAM OF PE SIMULATION PROGRAM

07314

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Vol I

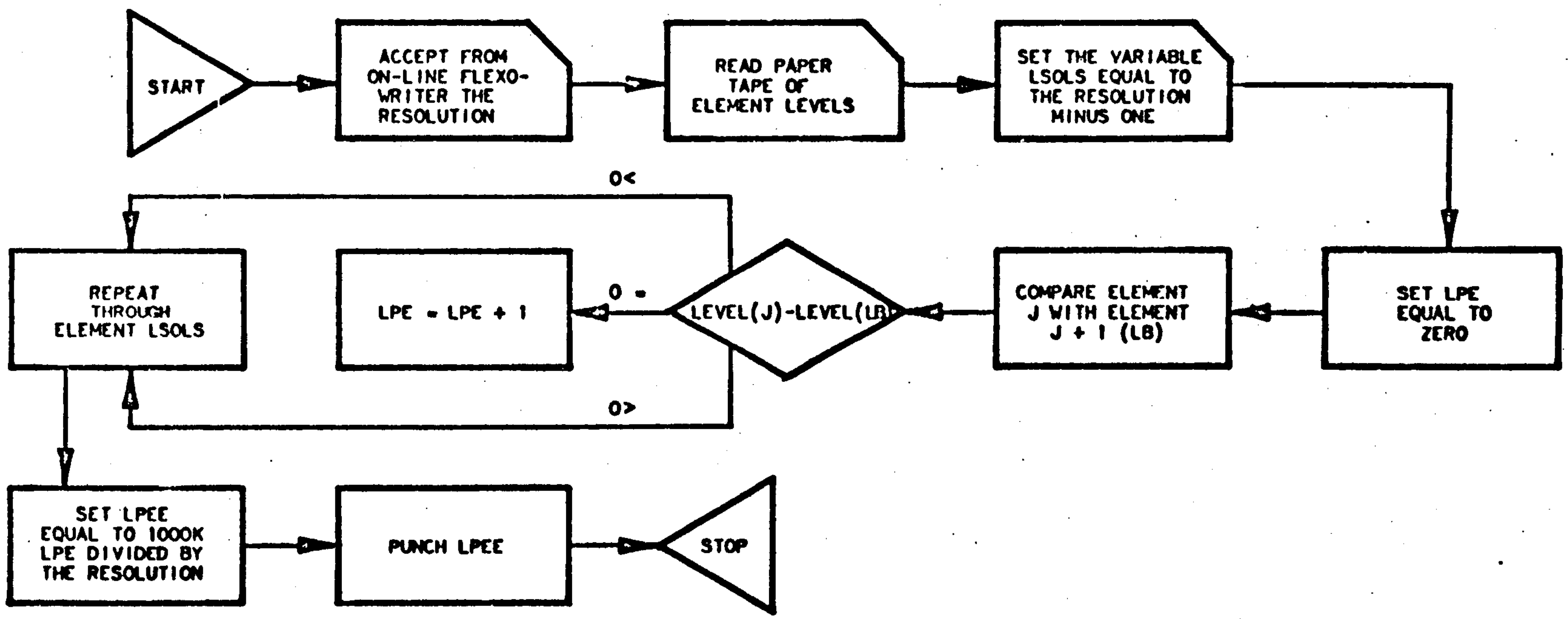

FIGURE A.9-10

FLOW DIAGRAM OF PEQ PROGRAM ON A LINE BASIS IN FORTRAN II

0734

124

RE-ORDER No. 64-712
Vol I

Figure A. 9-11 shows a plot of compression as a function of PE parametric in the number of consecutive errors outside the error band, the size of the error band, and the type of encoding employed for the slope approximation. Compression increases with optimum encoding of Δx run lengths. However, the increase in compression does not warrant the increase in equipment complexity or the channel noise problems encountered with a variable word length code.

Figures A. 9-12 and A. 9-13 show plots of compression as a function of PE parametric in the number of errors allowed outside the error-band, the size of the error-band, and the number of slopes (31 or 63). On the basis of compression, it is evident from the figures referenced above that the slight increase in compression going to 63 slopes is not of sufficient magnitude to warrant the increase in complexity. This conclusion is further justified by comparing the RMS error for 63 and 31 slopes as shown in Figure A. 9-14 which plots RMS error for 63 and 31 slopes as a function of error-band with a comparison to the equivalent PCM.

Optimum encoding is not desired, as can be seen from Figures A. 9-15 and A. 9-16 which plot the first-order probability distributions for 31 and 63 slopes,respectively. the statistical entropy for the two cases is 4.84 and 5.70 bits/element, respectively,as compared to five and six bits/element for flat encoding.

Based on these results the preliminary design of the recommended linear approximation coding system is as follows: (1) 31 slopes (2) flat encoding of the run-lengths and slope numbers, (3) error bands of ±1, ±2, and ±4 out of 64 levels, and (4) the number of errors outside the error band of 0, 1, and 2 errors. The 15 lines were then processed on EDITS and the ASI-210 Computer and the results are summarized graphically in Figure A. 9-17. Included on this plot are the result of a least square fit to the data program where the computer flow diagram is given in Figure A. 9-18.

The qualification of this coding technique will be achieved when a subjective evaluation can be performed on images reproduced by an operating linear approximation system, and when the effects of noise in the data link are observed. If these tests can be successfully carried out, it appears that the theoretical compression results

RE-ORDER No. 64-712 Vol I

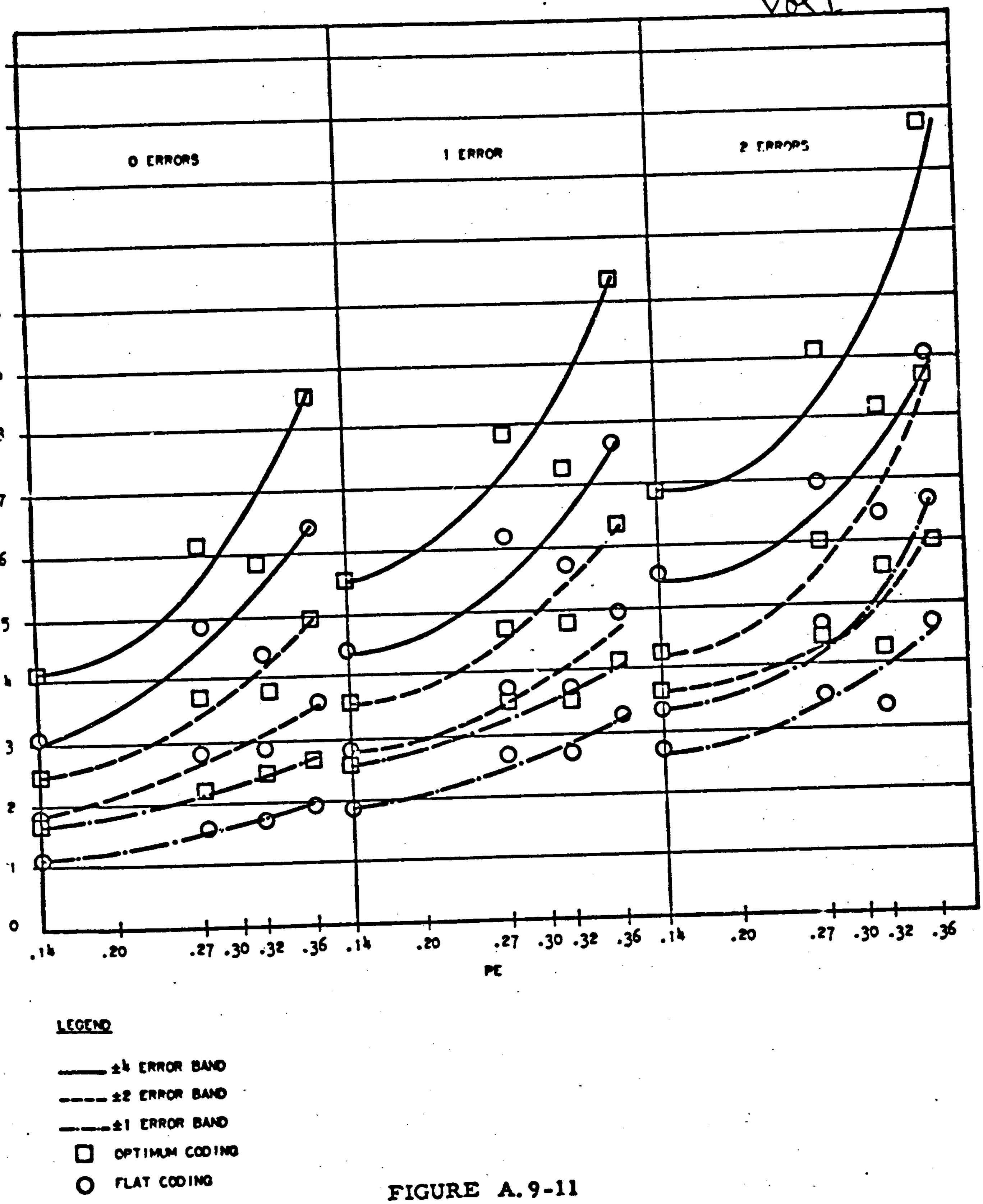

FIGURE A.9-11

COMPARISON OF OPTIMUM CODING AND FLAT CODING COMPRESSION FOR LINEAR APPROXIMATION CODING AS A FUNCTION OF ERROR BAND AND ALLOWABLE ERRORS

RE-ORDER No. 64-712
Vol I

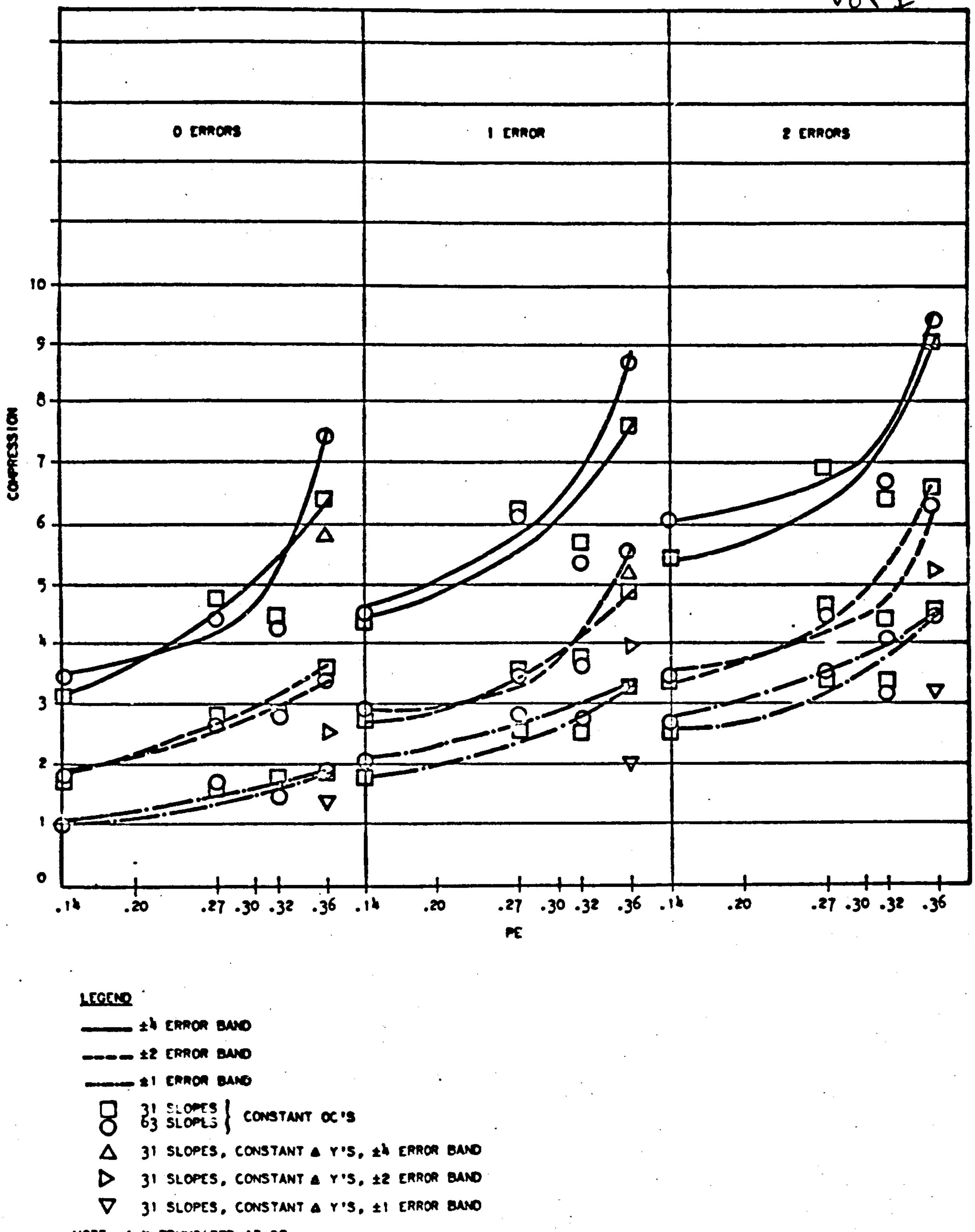

FIGURE A. 9-12

COMPARISON OF COMPRESSION FOR 31 AND 63 SLOPES FOR LINEAR APPROXIMATION CODING TECHNIQUE USING FLAT CODING

RE-ORDER No. 64-712 Vol I

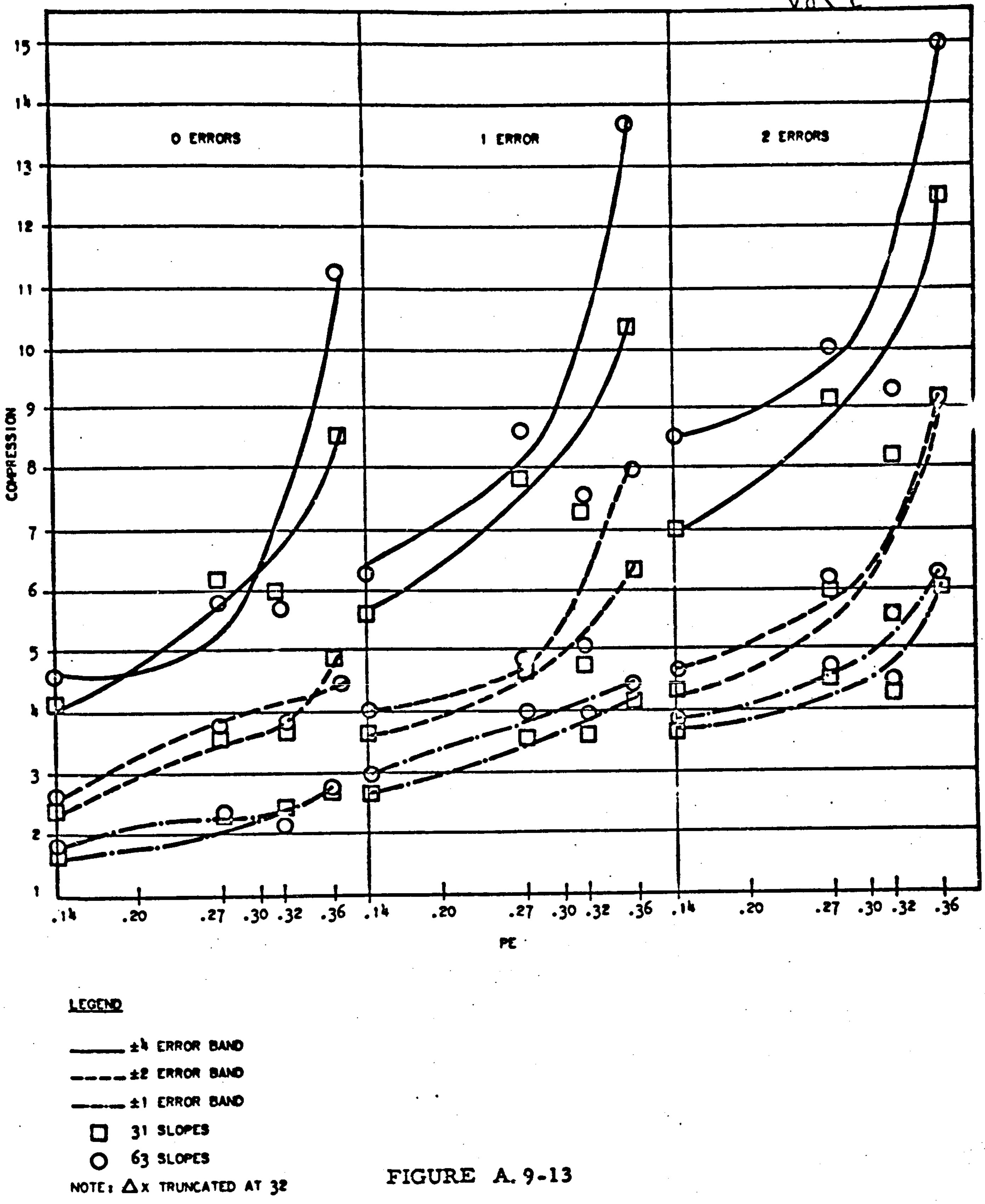

FIGURE A. 9-13

COMPARISON OF COMPRESSION FOR 31 AND 63 SLOPES
FOR LINEAR APPROXIMATION CODING COMPRESSION
TECHNIQUE USING OPTIMUM CODING

RE-ORDER No. 64-712
Vol I

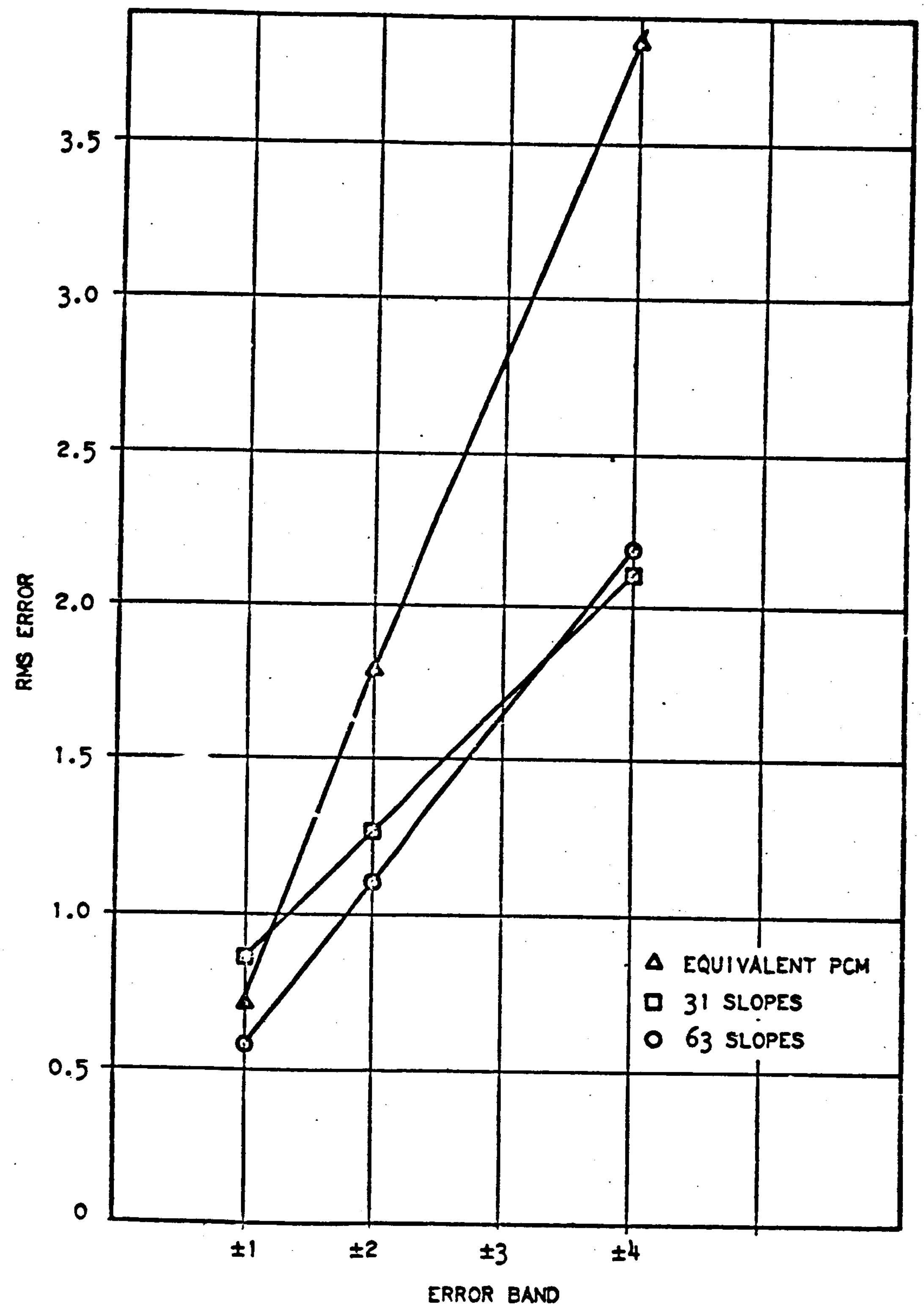

FIGURE A. 9-14

RMS ERROR FOR LINEAR APPROXIMATION
AND EQUIVALENT PCM QUANTIZATION FOR LINE 1
OF GIRL PICTURE AS A FUNCTION OF ERROR

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RE-ORDER No. 64-712
Vol I

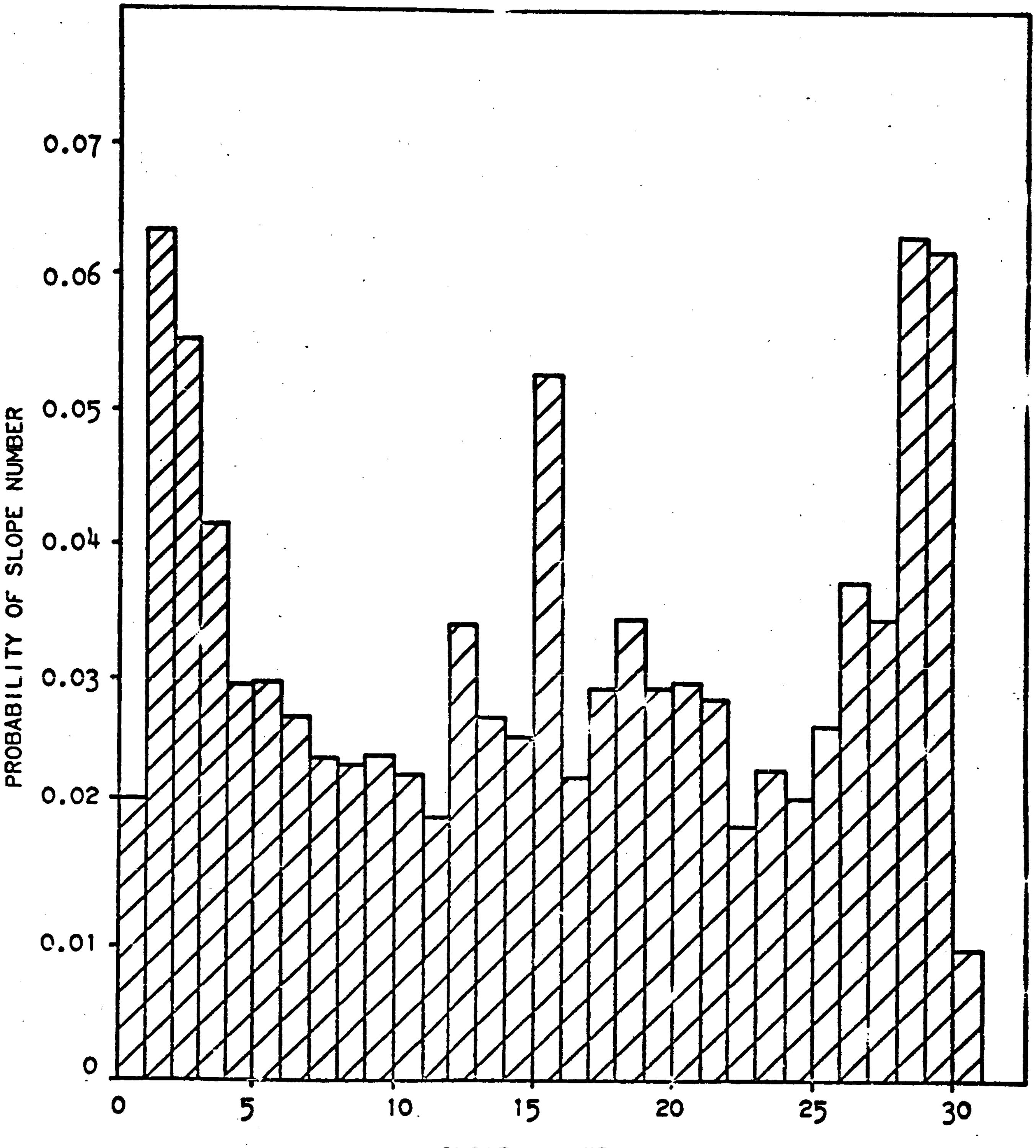

FIGURE A. 9-15

AVERAGE FIRST ORDER ERROR PROBABILITY DISTRIBUTION FOR 31 CONSTANT SLOPES FOR FOUR LINES EACH WITH 45 PARAMETER COMBINATIONS WITH AN ENTROPY OF 4.84 BITS PER ELEMENT

07314

RE-ORDER No. 64-712
Vol I

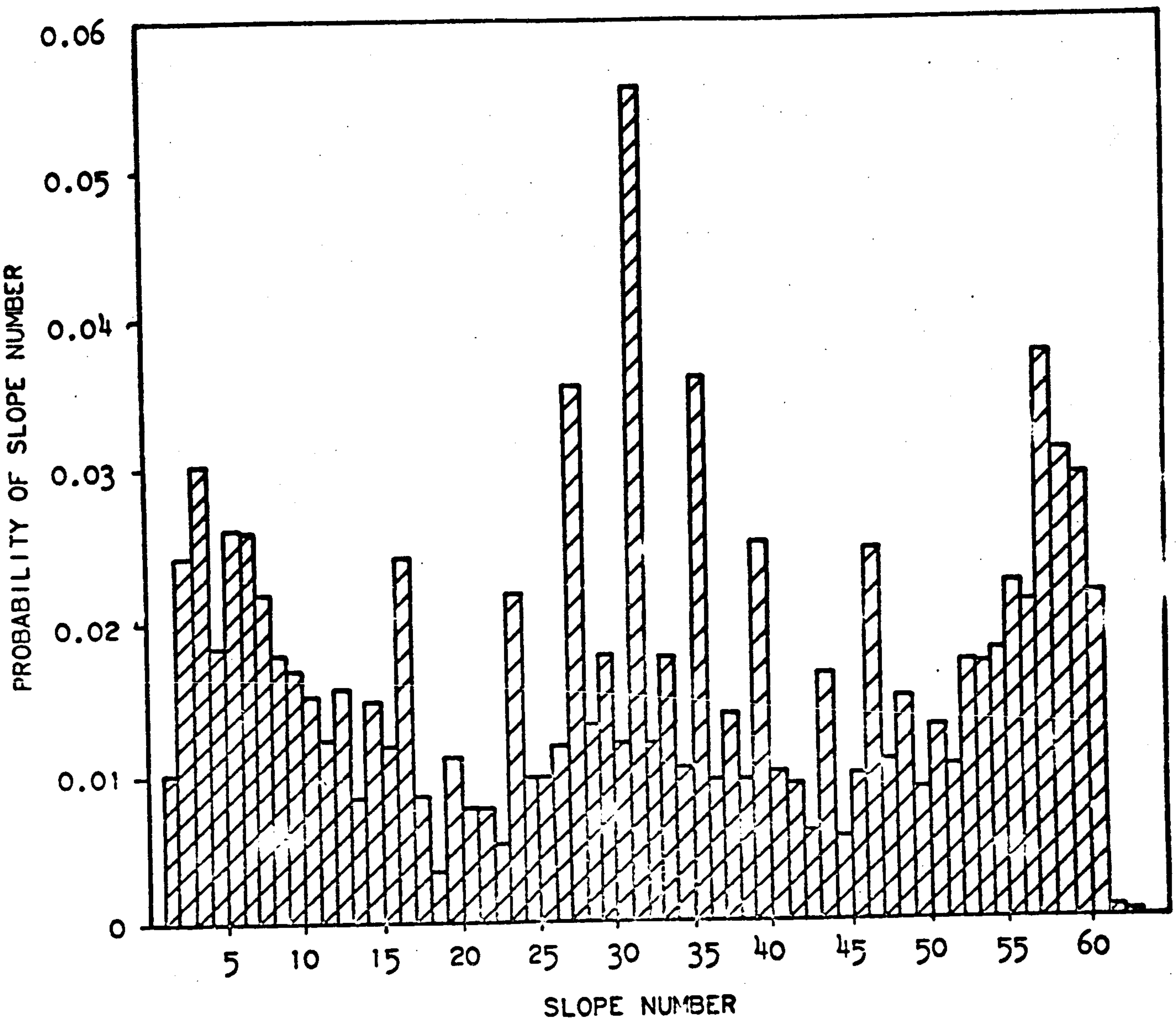

FIGURE A. 9-16

AVERAGE FIRST ORDER PROBABILITY DISTRIBUTION FOR 63 CONSTANT SLOPES FOR FOUR LINES EACH WITH 45 PARAMETER COMBINATIONS WITH AN ENTROPY OF 5.70 BITS PER ELEMENT

07314

RE-ORDER No. 64-712
Vol I

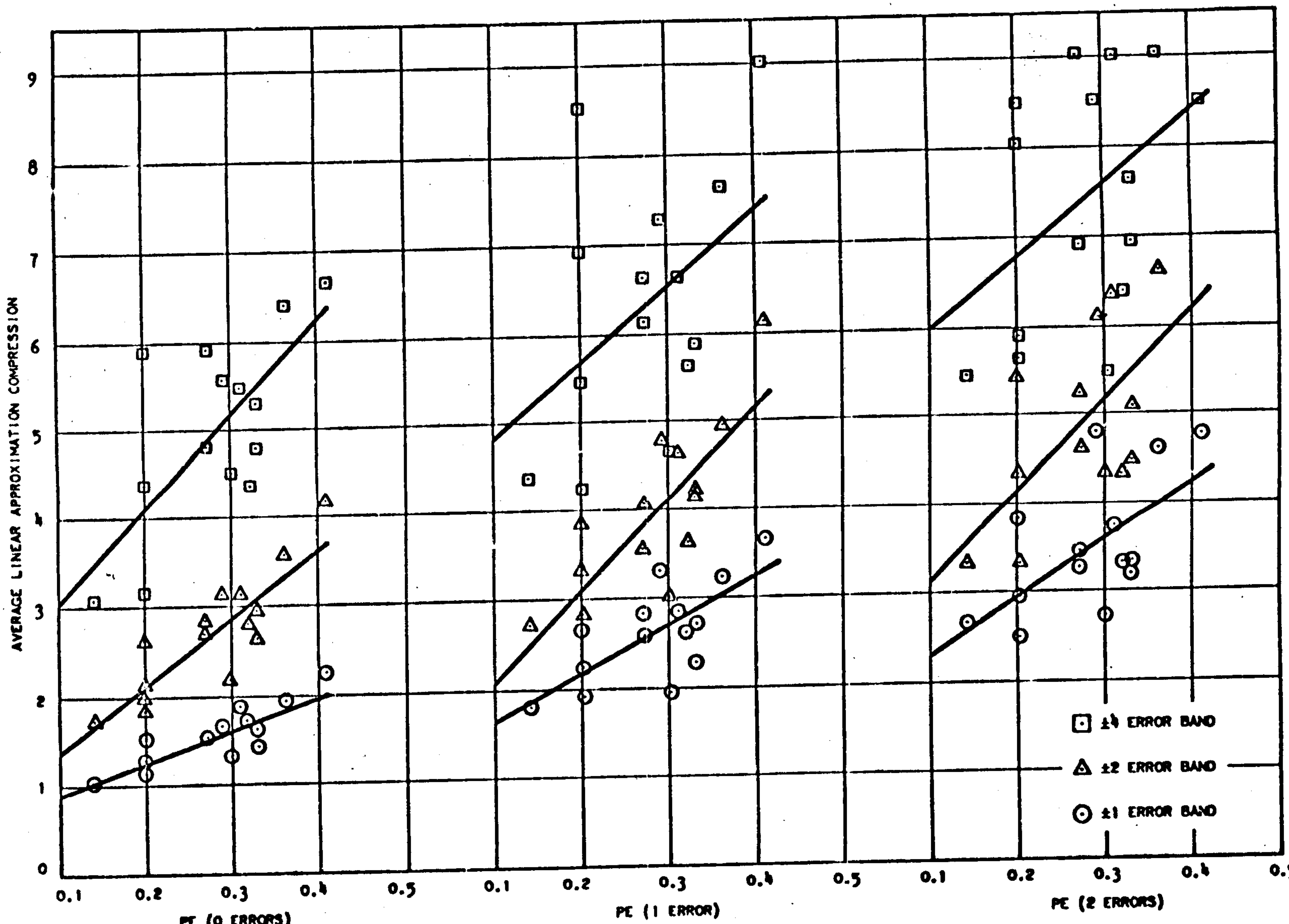

FIGURE A.9-17

07314

AVERAGE COMPUTER SIMULATED COMPRESSION FOR LINEAR APPROXIMATION AS A FUNCTION OF PE AND THE NUMBER OF TOLERABLE ERRORS OUTSIDE THE ERROR BAND PARAMETRIC IN THE ERROR BANDWIDTH FOR 15 LINES FROM FIVE DIFFERENT SOURCES OBTAINED FROM EDITS OF 256 RESOLUTION

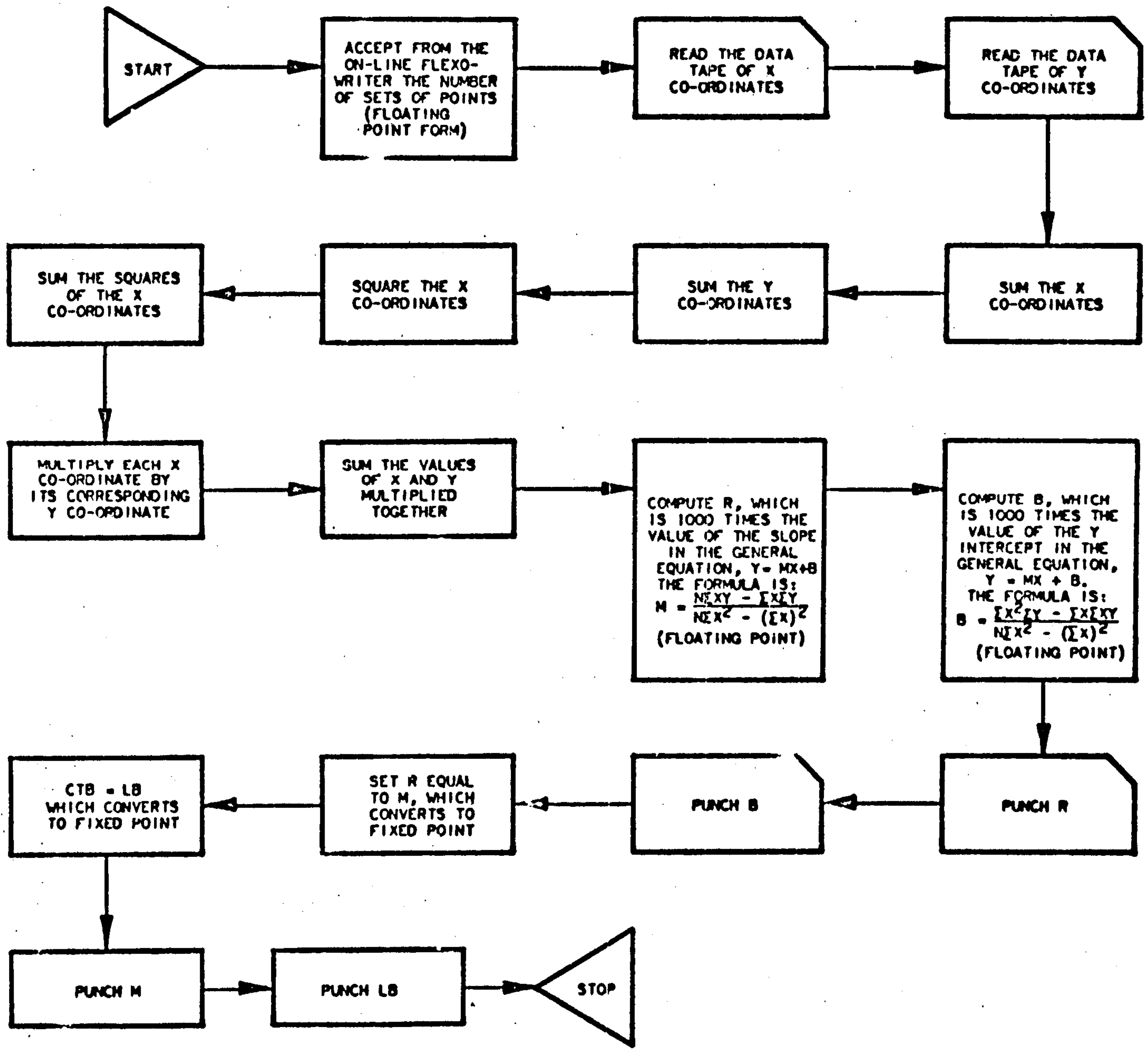

FIGURE A. 9-18

LEAST SQUARES FIT FLOW DIAGRAM IN FORTRAN II

RE-ORDER No. 64-712
Vol I.

achieved from the several experimental video scan lines as of sufficient magnitude to consider that the technique will satisfy the JPL requirements.

The results of this simulation can be summarized and are: (1) it is apparent that doubling the number of slopes increases the compression, but not of sufficient magnitude to warrant the increase in complexity of the implemented system; (2) it is questionable whether using optimum encoding of the slope-run lengths is desirable in view of the small increase in compression realized; (3) it is also questionable whether increasing the number of errors, thus degrading the picture, will warrant the small increase in compression going from 1 to 2 errors; (4) large (>2:1) compressions can be obtained using linear approximation; (5) for an average line, it appears that the coding technique could satisfy the JPL requirements on net compression.

Although the results of the fifteen lines appear to be quite encouraging, it is necessary to attempt to compare the linear approximation coding technique on a basis similar to the other coding techniques. Therefore, three lines from the Canadian Arctic scene at 256 resolution were processed at 40, 30, and 20 db signal-to-noise ratio. Table A.9-4 presents the results of the linear approximation coding computer program to these lines. Graphically these results are summarized in Figure A.9-19 for Δx truncated at 32 elements.

As anticipated, the compression decreased as the total number of errors decreased; however, with a large number of errors the compression varied considerably from line to line as a function of sensor signal-to-noise ratio. It is apparent that at realistic sensor signal-to-noise ratios, and an error band of ± 4 levels out of 64 levels, gross average compressions about 3.8:1 appear to be reasonable. This value of compression is of sufficient magnitude to satisfy the net compression requirements.

Although the error band concept is a form of noise stripping, the PEQ concept (another form of noise removal) was applied to the video lines and the linear approximation compression results for lines PEQ'd before processed on the computer are given in Table A.9-5. These results indicate that there exists a relationship between the tolerable errors, error band, and PEQ concept; however,

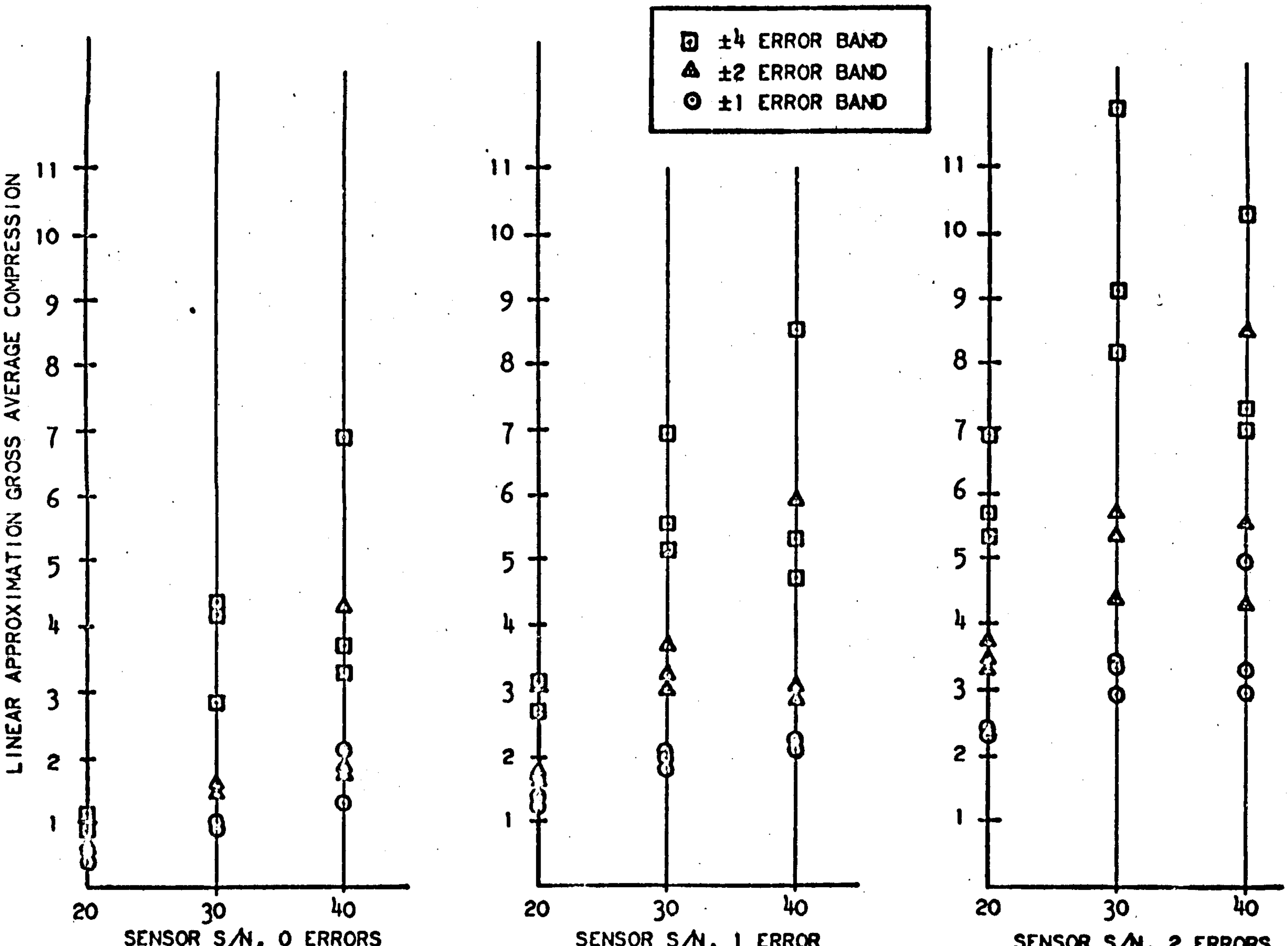

FIGURE A.9-19

COMPUTER SIMULATED LINEAR APPROXIMATION CODING COMPRESSION AS A FUNCTION OF SENSOR S/N FOR THREE FROM THE CANADIAN ARCTIC, ALSO AS A FUNCTION OF THE ERROR BAND AND TOLERABLE ERRORS

07314

RE-ORDER No. 64-712
Vol I

135

RE-ORDER No. 64-712
Vol I

Error Band / T	0 Errors			1 Error			2 Errors			
	±1	±2	±4	±1	±2	±4	±1	±2	±4	
64	1.68	2.44	4.09	2.55	3.59	5.54	3.63	4.23	6.69	Line 1
32	1.68	2.44	4.09	2.55	3.59	5.54	3.63	4.23	6.89	MIT Girl
16	1.68	2.44	4.19	2.55	3.59	5.27	3.50	4.17	6.68	PE = .14
8	1.68	2.46	4.08	2.52	3.40	4.91	3.54	4.49	5.66	31 Slopes
4	1.65	2.32	3.28	2.42	2.91	4.09	3.08	3.39	4.12	Optimum Coding
64	2.21	3.69	6.37	3.51	4.54	8.24	4.69	6.34	9.45	Line 2
32	2.21	3.69	6.15	3.51	4.67	7.84	4.48	6.00	9.11	MIT Girl
16	2.21	3.60	5.16	3.51	4.58	7.67	4.36	5.96	8.61	PE = .27
8	2.22	3.46	4.83	3.33	4.27	6.04	3.98	5.56	6.99	31 Slopes
4	2.14	3.01	3.72	2.98	3.51	4.05	3.63	4.08	5.00	Optimum Coding
64	2.42	3.76	6.46	3.54	4.92	7.32	4.29	5.56	9.34	Line 3
32	2.42	3.76	5.84	3.54	4.72	7.27	4.29	5.56	8.15	MIT Girl
16	3.21	3.82	5.46	3.56	4.37	6.51	4.36	5.11	9.36	PE = .32
8	2.42	3.55	4.95	3.27	4.53	5.17	3.83	5.21	6.77	31 Slopes
4	2.23	2.98	3.64	2.90	3.40	3.85	3.42	4.00	4.52	Optimum Coding
64	2.63	4.90	10.10	4.14	6.60	10.72	6.28	8.99	12.75	Line 4
32	2.63	4.93	8.57	4.14	6.36	10.33	6.03	9.21	12.52	MIT Girl
16	2.63	4.79	8.44	4.17	5.67	10.03	6.02	8.52	10.03	PE = .36
8	2.71	4.66	6.27	4.57	5.21	6.67	5.59	6.16	6.73	31 Slopes
4	2.47	3.78	4.70	3.29	3.82	4.75	4.32	4.37	4.86	Optimum Coding

TABLE A.9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION

Sheet 1 of 13

RE-ORDER No. 64-712
Vol I

	0 Errors			1 Error			2 Errors			
Error Band / T	±1	±2	±4	±1	±2	±4	±1	±2	±4	Line 9
64	1.01	1.68	2.84	1.74	2.58	3.87	2.28	3.03	4.98	Canadian A 15420-43
32	1.12	1.85	3.13	1.92	2.84	4.26	2.51	3.33	5.68	PE = .20
16	1.24	2.05	3.71	2.13	2.84	4.74	2.84	3.71	5.68	31 Slopes
8	1.40	2.37	3.49	2.43	3.09	4.08	3.09	3.76	4.92	Flat Coding
4	1.59	2.23	2.81	2.35	2.77	3.18	2.67	2.88	3.32	
64	1.42	2.58	5.37	2.58	3.87	6.07	2.97	4.81	7.75	Line 10
32	1.56	2.84	5.90	2.84	4.04	6.67	3.26	5.29	9.03	Canadian A 15420-41
16	1.74	2.89	6.09	2.99	4.26	7.42	3.63	5.33	7.42	PE = .27
8	1.90	3.20	5.05	3.09	4.00	5.05	3.69	4.46	5.48	31 Slopes
4	1.95	2.64	3.22	2.67	3.13	3.27	3.00	3.18	3.27	Flat Coding
64	1.51	2.84	4.81	3.10	4.36	6.07	4.10	5.58	8.72	Line 11
32	1.66	3.13	5.48	3.33	4.80	7.31	4.80	6.14	8.53	Canadian A 15420-43
16	1.85	3.34	5.88	3.41	4.87	6.56	5.01	6.32	7.42	PE = .29
8	2.04	3.76	4.68	3.36	4.26	4.92	4.26	5.05	5.48	31 Slopes
4	2.03	2.92	3.27	2.88	3.13	3.27	3.18	3.27	3.32	Flat Coding
64	1.21	1.99	4.23	1.79	2.73	4.36	2.49	3.77	4.81	Line 12
32	1.33	2.19	4.51	1.96	3.01	4.65	2.74	4.15	5.48	Tiros - Fr. 24
16	1.48	2.40	4.26	2.18	3.34	5.60	2.94	4.37	5.50	PE = .30
8	1.62	2.49	4.00	2.40	3.25	4.57	2.90	4.26	4.80	31 Slopes
4	1.75	2.38	3.09	2.28	2.81	3.22	2.77	3.13	3.27	Flat Coding

TABLE A.9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

Sheet 2 of 13

RE-ORDER No. 64-712 Vol I

	0 Errors			1 Error			2 Errors			
T \ Error Band	±1	±2	±4	±1	±2	±4	±1	±2	±4	Line 5
64	1.04	1.81	3.17	1.81	2.58	4.98	2.68	3.49	5.17	Lunar 66-2
32	1.15	1.99	3.19	1.99	2.84	5.48	2.95	3.84	5.90	PE = .20
16	1.28	2.21	3.96	2.21	3.16	5.68	3.28	4.16	6.09	31 Slopes
8	1.44	2.31	3.91	2.46	3.49	4.92	3.55	4.46	5.33	Flat Coding
4	1.61	2.38	3.00	2.49	2.96	3.22	2.92	3.22	3.27	
64	1.38	2.40	5.37	2.44	3.49	7.75	3.49	4.98	7.75	Line 6
32	1.52	2.64	5.90	2.69	3.84	8.53	3.84	5.48	8.53	Lunar 66-1
16	1.68	2.94	5.88	2.99	4.26	8.12	4.26	5.88	8.53	PE = .20
8	1.97	3.20	5.33	3.36	4.26	5.64	4.20	4.92	5.48	31 Slopes
4	1.99	2.88	3.37	2.84	3.18	3.37	3.18	3.32	3.37	Flat Coding
64	1.18	1.96	3.98	2.05	3.03	6.07	2.63	3.98	7.34	Line 7
32	1.30	2.16	4.38	2.25	3.33	6.98	2.89	4.38	8.08	Lunar 66-1
16	1.44	2.40	4.49	2.50	3.63	5.50	3.22	5.68	7.75	PE = .20
8	1.62	2.63	4.36	2.63	3.69	5.18	3.31	4.68	5.64	31 Slopes
4	1.75	2.64	3.18	2.67	3.09	3.37	3.13	3.32	3.42	Flat Coding
64	1.72	2.90	6.34	2.63	4.50	6.34	3.40	6.07	8.21	Line 8
32	1.89	3.13	5.48	2.89	4.65	6.67	3.74	6.40	9.03	Lunar 66-1
16	2.10	3.41	5.50	3.28	4.87	7.11	4.06	5.88	8.53	PE = .31
8	2.31	3.55	4.92	3.49	4.46	5.18	4.08	4.92	5.64	31 Slopes
4	2.21	2.81	3.18	3.00	3.18	3.27	3.18	3.32	3.37	Flat Coding

TABLE A.9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

Sheet 3 of 13

RE-ORDER No. 64-712
Vol I

Error Band / Δx	0 Errors			1 Error			2 Errors			
	±1	±2	±4	±1	±2	±4	±1	±2	±4	Line 1
64	0.93	1.56	2.79	1.66	2.49	3.98	2.44	3.03	4.65	MIT Girl
32	1.03	1.72	3.07	1.82	2.74	4.38	2.69	3.33	5.48	PE = 0.14
16	1.14	1.91	3.63	2.03	2.99	4.49	2.89	3.63	5.50	31 Slopes
8	1.28	2.15	2.62	2.25	3.09	4.26	3.14	4.00	4.68	Flat Coding
4	1.42	2.15	2.81	2.26	2.64	3.13	2.77	2.92	3.18	
64	1.42	2.49	4.36	2.36	3.17	5.81	3.24	4.36	6.34	Line 2
32	1.56	2.74	4.80	2.60	3.57	6.14	3.49	4.65	6.98	MIT Girl
16	1.74	2.94	4.49	2.89	3.87	5.88	3.89	5.01	6.56	PE = .27
8	1.97	3.04	4.17	3.04	3.84	4.57	3.69	4.57	5.18	31 Slopes
4	1.99	2.61	3.00	2.74	2.92	3.13	3.00	3.18	3.37	Flat Coding
64	1.55	2.58	4.36	2.40	3.40	4.98	3.03	3.98	6.34	Line 3
32	1.70	2.84	4.38	2.64	3.65	5.68	3.33	4.38	6.40	MIT Girl
16	1.89	3.22	4.61	2.99	3.87	5.17	3.71	4.49	6.82	PE = .32
8	2.13	3.14	4.08	2.95	3.91	4.26	3.49	4.36	5.05	31 Slopes
4	2.08	2.61	3.00	2.61	2.92	3.09	2.92	3.18	3.32	Flat Coding
64	1.76	3.32	6.64	2.97	4.65	7.34	4.36	6.07	8.72	Line 4
32	1.94	3.57	6.40	3.26	4.95	7.68	4.65	6.67	9.03	MIT Girl
16	2.16	3.87	6.09	3.63	4.74	7.11	5.01	6.32	7.11	PE = .36
8	2.43	3.76	4.68	3.62	4.08	4.80	4.36	4.57	4.92	31 Slopes
4	2.28	2.88	3.13	2.77	3.00	3.22	3.09	3.13	3.27	Flat Coding

TABLE A. 9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

RE-ORDER No. 64-712
Vol I

Error Band / T	0 Errors ±1	0 Errors ±2	0 Errors ±4	1 Error ±1	1 Error ±2	1 Error ±4	2 Errors ±1	2 Errors ±2	2 Errors ±4	
64	1.45	2.68	4.81	2.49	3.58	6.07	2.90	3.98	6.64	Line 13
32	1.60	2.95	5.29	2.74	4.15	5.90	3.20	4.51	6.98	Tiros - Fr. 24
16	1.77	3.04	5.17	3.34	4.49	5.88	3.79	4.6[illegible]	6.82	PE = .33
8	2.00	3.00	4.68	3.09	3.84	5.05	3.69	4.08	5.33	31 Slopes
4	2.03	2.70	3.22	2.74	3.00	3.32	3.00	3.18	3.32	Flat Coding
64	1.31	2.40	4.23	2.08	3.87	5.17	3.03	4.65	6.98	Line 14
32	1.44	2.64	4.80	2.29	4.26	5.90	3.33	5.12	7.68	Tiros - Fr. 18
16	1.55	2.89	5.01	2.50	4.26	5.68	3.63	4.87	6.82	PE = .33
8	1.76	2.82	4.26	2.82	4.08	4.92	3.69	4.46	5.33	31 Slopes
4	1.89	2.67	3.22	2.64	3.13	3.22	2.88	3.22	3.32	Flat Coding
64	1.96	3.77	6.07	3.32	5.58	7.75	4.50	6.34	8.21	Line 15
32	2.13	4.15	6.67	3.65	6.14	9.03	4.80	6.40	8.53	Tiros - Fr. 24
16	2.40	4.37	6.09	4.06	5.68	7.75	5.17	6.09	8.12	PE = .41
8	2.31	3.91	4.92	4.00	4.80	5.64	4.36	4.80	5.33	31 Slopes
4	2.30	3.00	3.22	3.00	3.27	3.32	3.18	3.27	3.37	Flat Coding
64	1.31	2.36	5.58	1.86	3.49	4.98	3.03	4.98	8.72	Line 4
32	1.44	2.60	5.90	2.04	3.84	5.29	3.26	5.29	9.03	MIT Girl
16	1.61	2.79	5.68	2.24	3.67	6.32	3.48	5.17	8.12	PE = .36
8	1.74	2.90	4.80	2.40	3.76	5.05	3.31	4.57	5.48	31 Slopes (Constant 1st diff.)
4	1.84	2.81	3.22	2.38	3.18	3.37	2.81	3.18	3.42	Flat Coding

TABLE A.9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

RE-ORDER No. 64-712
Vol I

Error Band	0 Errors ±1	0 Errors ±2	0 Errors ±4	1 Error ±1	1 Error ±2	1 Error ±4	2 Errors ±1	2 Errors ±2	2 Errors ±4	Line 4
64	2.28	1.22	0.59	1.45	0.91	0.56	0.95	0.67	0.47	MIT Girl
32	2.28	1.22	0.70	1.45	0.94	0.58	0.99	0.65	0.48	PE = .36
16	2.28	1.25	0.71	1.44	1.06	0.60	1.00	0.70	0.60	31 Slopes
8	2.21	1.29	0.96	1.31	1.15	0.89	1.07	0.97	0.89	Entropy
4	2.43	1.59	1.27	1.83	1.57	1.26	1.38	1.37	1.24	
64	2.13	1.32	0.59	1.38	0.66	0.48	0.96	0.61	0.40	Line 4
32	2.18	1.36	0.53	1.38	0.76	0.44	0.96	0.66	0.40	MIT Girl
16	2.22	1.32	0.61	1.39	0.86	0.59	0.94	0.72	0.44	PE = .36
8	2.37	1.27	0.86	1.35	0.98	0.80	1.09	0.96	0.68	63 Slopes
4	2.38	1.59	1.19	1.61	1.20	1.04	1.31	1.12	0.95	Entropy
64	3.23	6.07	12.14	5.43	8.50	13.42	7.97	11.09	15.94	Line 4
32	3.23	5.93	10.63	5.43	8.23	12.75	7.72	11.09	15.00	MIT Girl
16	3.23	5.80	9.11	5.43	7.08	10.63	7.50	9.44	10.63	PE = .36
8	3.23	5.00	6.22	4.81	5.43	6.38	5.80	6.07	6.54	31 Slopes
4	2.66	3.36	3.64	3.23	3.49	3.75	3.59	3.64	3.81	Average Run Length
64	3.54	6.38	12.75	5.93	11.59	15.94	8.23	12.75	18.21	Line 4
32	3.59	6.22	13.42	5.93	10.63	15.94	8.23	11.59	17.00	MIT Girl
16	3.54	6.22	10.63	5.93	8.50	11.09	7.73	9.44	12.75	PE = .36
8	3.31	5.31	6.89	5.31	6.22	6.89	5.93	6.22	7.29	63 Slopes
4	2.83	3.40	3.81	3.45	3.70	3.92	3.70	3.81	3.98	Average Run Length

TABLE A. 9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

Sheet 6 of 13

RE-ORDER No. 64-712
Vol I

Error Band Δx	0 Errors			1 Error			2 Errors			
	±1	±2	±4	±1	±2	±4	±1	±2	±4	Line 1
64	.93	1.64	3.12	1.88	2.72	4.12	2.56	3.20	5.56	MIT Girl
32	1.01	1.79	3.40	2.05	2.97	4.50	2.79	3.49	6.07	PE = .14
16	1.12	1.94	3.57	2.25	3.13	4.65	2.95	3.74	5.48	63 Slopes
8	1.24	1.98	3.63	2.47	3.10	4.37	3.16	4.06	4.49	Flat Coding
4	1.31	1.93	2.74	2.25	2.59	2.82	2.56	2.74	2.90	
64	1.54	2.46	4.26	2.66	3.28	6.09	3.20	4.41	6.40	Line 2
32	1.68	2.68	4.36	2.90	3.49	6.07	3.58	4.50	6.98	MIT Girl
16	1.80	2.95	4.95	3.13	3.74	6.14	3.74	4.51	6.40	PE = .27
8	2.00	2.84	3.79	2.94	3.48	4.49	3.41	3.96	6.74	63 Slopes
4	1.90	2.40	2.94	2.52	2.66	2.86	2.70	2.86	3.00	Flat Coding
64	1.40	2.56	4.00	2.50	3.28	5.33	2.97	4.26	6.40	Line 3
32	1.53	2.79	4.23	2.79	3.58	5.37	3.24	4.10	6.64	MIT Girl
16	1.68	2.95	4.38	3.01	3.84	5.46	3.41	4.65	6.40	PE = .32
8	1.91	2.75	3.63	2.89	3.63	4.16	3.34	4.26	4.49	63 Slopes
4	1.95	2.31	2.70	2.40	2.66	2.82	2.63	2.82	2.95	Flat Coding
64	1.77	3.20	6.40	2.97	5.81	8.00	4.12	6.40	9.14	Line 4
32	1.96	3.40	7.34	3.24	5.81	8.72	4.50	6.34	9.30	MIT Girl
16	2.13	3.74	6.40	3.57	5.12	6.67	4.65	5.68	7.68	PE = .36
8	2.21	3.55	4.61	3.55	4.16	4.61	3.96	4.16	4.87	63 Slopes
4	2.13	2.56	2.86	2.59	2.78	2.95	2.78	2.86	3.00	Flat Coding

TABLE A.9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

Error Band / T	0 Errors ±1	0 Errors ±2	0 Errors ±4	1 Error ±1	1 Error ±2	1 Error ±4	2 Errors ±1	2 Errors ±2	2 Errors ±4	
										Line 1
64	.859	1.274	2.099	1.577	1.866	2.823	2.235	2.629	3.020	MIT Girl
32	.859	1.274	2.099	1.577	1.866	2.823	2.235	2.629	3.140	PE = .14
16	.859	1.274	2.209	1.577	1.883	2.730	2.187	2.661	2.943	31 Slopes
8	.859	1.236	2.019	1.555	1.875	2.494	2.336	2.509	2.902	RMS Error
4	.852	1.153	1.703	1.531	1.746	2.182	2.320	2.392	2.527	
64	.807	1.223	2.229	1.191	1.586	2.354	1.683	1.946	2.741	Line 2
32	.807	1.223	2.114	1.191	1.599	2.190	1.626	1.928	2.506	MIT Girl
16	.807	1.217	2.161	1.191	1.555	2.092	1.699	1.974	2.450	PE = .27
8	.812	1.191	1.794	1.177	1.465	2.002	1.661	1.857	2.239	31 Slopes
4	.818	.983	1.461	1.134	1.369	1.667	1.656	1.702	1.843	RMS Error
64	.688	1.124	2.050	.989	1.553	2.383	1.605	2.012	2.538	Line 3
32	.688	1.124	2.075	.989	1.505	2.312	1.605	2.012	2.879	MIT Girl
16	.688	1.091	1.961	1.000	1.379	2.097	1.519	1.888	2.641	PE = .32
8	.684	1.001	1.713	1.064	1.352	1.895	1.530	1.935	2.198	31 Slopes
4	.667	.889	1.367	.878	1.302	1.479	1.527	1.722	1.990	RMS Error
64	.735	1.229	2.166	.999	1.368	2.243	1.288	1.630	2.294	Line 4
32	.735	1.139	2.085	.999	1.327	2.089	1.288	1.614	2.261	MIT Girl
16	.735	1.026	1.808	.979	1.446	1.771	1.352	1.448	1.902	PE = .36
8	.720	1.004	1.452	.898	1.092	1.482	1.291	1.377	1.625	31 Slopes
4	.663	.917	1.146	.826	1.003	1.273	1.292	1.307	1.600	RMS Error

TABLE A. 9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

RE-ORDER No. 64-712
Vol I

	0 Errors			1 Error			2 Errors			
Error Band / T	±1	±2	±4	±1	±2	±4	±1	±2	±4	Line 1
64	1.59	2.56	4.48	2.86	4.03	6.17	3.79	4.58	8.53	MIT Girl
32	1.59	2.56	4.48	2.86	3.99	6.17	3.79	4.58	8.44	PE = .14
16	1.59	2.53	4.31	2.86	3.79	5.97	3.61	4.54	6.94	63 Slopes
8	1.59	2.38	4.39	2.90	3.66	5.73	3.66	5.12	6.04	Optimum Coding
4	1.53	2.20	3.74	2.67	3.26	4.01	3.08	3.65	4.45	
64	2.33	3.62	6.35	3.95	4.87	9.05	4.62	6.70	9.93	Line 2
32	2.33	3.62	5.77	3.95	4.69	8.63	4.70	6.12	9.98	MIT Girl
16	2.29	3.62	6.77	3.85	4.65	8.31	4.49	5.52	9.11	PE = .27
8	2.29	3.24	4.99	3.38	4.22	6.37	3.92	5.04	7.22	63 Slopes
4	2.08	2.96	4.02	3.11	3.62	4.42	3.65	4.30	5.47	Optimum Coding
64	2.11	3.73	5.90	3.83	4.90	8.00	4.43	6.04	10.44	Line 3
32	2.11	3.73	5.65	3.90	4.95	7.49	4.43	5.53	9.23	MIT Girl
16	2.11	3.71	5.69	3.67	4.82	7.82	4.10	5.73	9.17	PE = .32
8	2.18	3.34	4.69	3.39	4.55	5.87	3.92	5.61	6.09	63 Slopes
4	2.25	2.74	3.76	2.94	3.65	4.23	3.58	4.12	5.22	Optimum Coding
64	2.81	4.53	10.16	4.36	9.16	12.54	6.24	9.85	14.91	Line 4
32	2.75	4.41	11.26	4.36	7.91	13.66	6.24	9.03	14.93	MIT Girl
16	2.71	4.55	9.87	4.31	6.97	10.11	6.38	8.33	13.53	PE = .36
8	2.53	4.72	7.01	4.43	6.11	7.53	5.52	6.28	8.78	63 Slopes
4	2.52	3.79	5.04	3.74	4.99	5.78	4.60	5.36	6.30	Optimum Coding

TABLE A. 9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

RE-ORDER No. 64-712
Vol I

Error Band Δx	0 Errors ±1	0 Errors ±2	0 Errors ±4	1 Error ±1	1 Error ±2	1 Error ±4	2 Errors ±1	2 Errors ±2	2 Errors ±4	Line 1
64	.93	1.64	3.12	1.88	2.72	4.12	2.56	3.20	5.56	MIT Girl
32	1.01	1.79	3.40	2.05	2.97	4.50	2.79	3.49	6.07	PE = .14
16	1.12	1.94	3.57	2.25	3.13	4.65	2.95	3.74	5.48	63 Slopes
8	1.24	1.98	3.63	2.47	3.10	4.37	3.16	4.06	4.49	Flat Coding
4	1.31	1.93	2.74	2.25	2.59	2.82	2.56	2.74	2.90	
64	1.54	2.46	4.26	2.66	3.28	6.09	3.20	4.41	6.40	Line 2
32	1.68	2.68	4.36	2.90	3.49	6.07	3.58	4.50	6.98	MIT Girl
16	1.80	2.95	4.95	3.13	3.74	6.14	3.74	4.51	6.40	PE = .27
8	2.00	2.84	3.79	2.94	3.48	4.49	3.41	3.96	6.74	63 Slopes
4	1.90	2.40	2.94	2.52	2.66	2.86	2.70	2.86	3.00	Flat Coding
64	1.40	2.56	4.00	2.50	3.28	5.33	2.97	4.26	6.40	Line 3
32	1.53	2.79	4.23	2.79	3.58	5.37	3.24	4.10	6.64	MIT Girl
16	1.68	2.95	4.38	3.01	3.84	5.48	3.41	4.65	6.40	PE = .32
8	1.91	2.75	3.63	2.89	3.63	4.16	3.34	4.26	4.49	63 Slopes
4	1.95	2.31	2.70	2.40	2.66	2.82	2.63	2.82	2.95	Flat Coding
64	1.77	3.20	6.40	2.97	5.81	8.00	4.12	6.40	9.14	Line 4
32	1.96	3.40	7.34	3.24	5.81	8.72	4.50	6.34	9.30	MIT Girl
16	2.13	3.74	6.40	3.57	5.12	6.67	4.65	5.68	7.68	PE = .36
8	2.21	3.55	4.61	3.55	4.16	4.61	3.96	4.16	4.87	63 Slopes
4	2.13	2.56	2.86	2.59	2.78	2.95	2.78	2.86	3.00	Flat Coding

TABLE A.9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

Sheet 10 of 13

RE-ORDER No. 64-712
Vol I

	0 Errors			1 Error			2 Errors			
Error Band / T	±1	±2	±4	±1	±2	±4	±1	±2	±4	Line 4
64	.617	1.182	2.138	.846	1.533	2.624	1.165	1.790	2.392	MIT Girl
32	.617	1.182	2.138	.846	1.533	2.624	1.165	1.790	2.392	PE = .36
16	.617	1.182	2.138	.846	1.559	2.542	1.165	1.790	2.335	31 Slopes
8	.617	1.176	2.302	.846	1.518	2.376	1.165	1.790	2.875	RMS Error (Const. 1st Diff.
4	.623	1.206	1.588	.856	1.384	1.612	1.170	1.790	1.868	

TABLE A. 9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

Sheet 12 of 13

RE-ORDER No. 64-712
Vol I

	0 Errors			1 Error			2 Errors			
Error Band / T	±1	±2	±4	±1	±2	±4	±1	±2	±4	Line 1
64	.572	1.133	2.175	1.615	1.768	2.901	2.137	2.589	3.352	MIT Girl
32	.572	1.133	2.175	1.615	1.773	2.901	2.137	2.589	3.330	PE = .14
16	.572	1.133	2.027	1.615	1.753	2.794	2.117	2.755	3.166	63 Slopes
8	.572	1.085	2.021	1.561	1.706	2.500	2.240	2.438	3.133	RMS Error
4	.556	1.009	1.848	1.510	1.658	2.141	2.108	2.455	2.532	
64	.650	1.199	2.254	1.085	1.543	2.590	1.665	2.044	2.813	Line 2
32	.650	1.199	2.243	1.085	1.540	2.438	1.836	2.132	2.781	MIT Girl
16	.654	1.183	1.865	1.095	1.453	2.359	1.836	2.017	2.667	PE = .27
8	.662	1.083	1.715	1.075	1.371	1.969	1.760	1.873	2.327	63 Slopes
4	.602	.856	1.270	1.050	1.171	1.633	1.729	1.745	1.800	RMS Error
64	.656	1.107	2.269	1.162	1.653	2.684	1.681	2.331	2.689	Line 3
32	.656	1.107	2.192	1.131	1.585	2.429	1.681	2.064	2.918	MIT Girl
16	.656	1.020	1.877	1.164	1.562	2.229	1.685	2.036	2.590	PE = .32
8	.635	1.082	1.729	1.095	1.511	2.010	1.657	1.945	2.380	63 Slopes
4	.617	.870	1.264	1.101	1.298	1.538	1.556	1.660	1.862	RMS Error
64	.732	1.207	2.129	.958	1.339	1.835	1.343	1.620	2.516	Line 4
32	.718	1.184	1.702	.958	1.233	1.879	1.343	1.638	2.256	MIT Girl
16	.718	1.138	1.601	.958	1.233	1.812	1.312	1.549	1.937	PE = .36
8	.694	.953	1.446	.934	1.197	1.552	1.216	1.385	1.712	63 Slopes
4	.683	.836	1.071	.914	1.074	1.231	1.148	1.367	1.618	RMS Error

TABLE A. 9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

Sheet 11 of 13

RE-ORDER No. 64-712
Vol I

COMPARABLE PCM RMS ERROR

Error Band \ Line No.	Line 1	Line 2	Line 3	Line 4
±1	.699	.696	.723	.656
±2	1.800	1.932	2.025	1.712
±4	3.821	4.288	4.207	4.088

COMPUTER TRUNCATION LINE 1

Errors \ Error Band	31 Slopes			63 Slopes		
	±1	±2	±4	±1	±2	±4
0	.859	1.274	2.077	.572	1.133	1.615
1	1.577	1.866	2.823	1.615	1.768	2.901
2	2.235	2.629	3.070	2.137	2.589	3.352

TABLE A. 9-3

COMPUTER RESULTS FOR LINEAR APPROXIMATION CODING SIMULATION (CONT'D)

Sheet 13 of 13

RE-ORDER No. 64-712
Vol I

TABLE A. 9-4

LINEAR APPROXIMATION SIMULATION COMPUTER RESULTS FOR THE CANADIAN ARCTIC SCENE PROCESSED ON EDITS AS A FUNCTION OF SENSOR SIGNAL-TO-NOISE RATIO

	0 Errors			1 Error			2 Errors			
Error Band Δx	±1	±2	±4	±1	±2	±4	±1	±2	±4	
64	1.15	1.70	2.97	1.93	2.68	4.81	2.68	4.98	6.64	Line 1
32	1.26	1.87	3.26	2.13	2.95	5.29	2.95	5.48	7.31	Canadian Arctic
16	1.41	2.08	3.71	2.37	3.10	5.50	3.22	5.01	7.75	PE = .246
8	1.57	2.31	3.76	2.52	3.36	4.80	3.36	4.36	5.64	40 db
4	1.67	2.41	2.96	2.46	2.92	3.32	3.09	3.32	3.37	Flat coding
64	1.15	1.76	3.32	2.02	2.79	4.23	2.97	3.87	7.75	Line 2
32	1.26	1.94	3.65	2.22	3.07	4.65	3.26	4.26	6.98	Canadian Arctic
16	1.39	2.13	4.49	2.40	3.34	5.01	3.55	4.49	7.11	PE = .269
8	1.56	2.34	4.00	2.66	3.55	4.80	3.69	4.08	5.48	40 db
4	1.68	2.26	3.09	2.46	2.88	3.27	3.00	3.18	3.42	Flat coding
64	1.91	3.77	6.34	2.79	5.37	8.21	4.50	7.75	9.97	Line 3
32	2.10	4.26	6.98	3.07	5.90	8.53	4.95	8.53	10.24	Canadian Arctic
16	2.18	3.96	6.82	3.48	5.88	8.12	4.87	7.75	8.53	PE = .394
8	2.34	4.17	5.05	3.62	4.92	5.64	4.68	5.48	5.81	40 db
4	2.26	3.00	3.37	2.96	3.27	3.37	3.27	3.37	3.42	Flat coding

RE-ORDER No. 64-712
Vol I

TABLE A. 9-4

LINEAR APPROXIMATION SIMULATION COMPUTER RESULTS FOR THE CANADIAN ARCTIC SCENE PROCESSED ON EDITS AS A FUNCTION OF SENSOR SIGNAL-TO-NOISE RATIO (CONT'D)

	0 Errors			1 Error			2 Errors			
Error Band Δx	±1	±2	±4	±1	±2	±4	±1	±2	±4	
64	.83	1.45	4.23	1.86	3.32	4.98	3.24	5.17	10.74	Line 1
32	.91	1.60	4.26	2.04	3.65	5.48	3.57	5.68	11.81	Canadian Arctic
16	1.01	1.77	4.37	2.27	3.87	6.82	3.55	5.68	8.98	PE = .226
8	1.14	2.00	4.57	2.63	4.08	5.64	3.76	5.18	5.81	30 db
4	1.29	2.15	3.13	2.55	3.09	3.37	3.04	3.32	3.42	Flat coding
64	.90	1.45	2.63	1.74	2.73	4.50	2.58	3.98	6.64	Line 2
32	.99	1.60	2.89	1.92	3.01	5.12	2.84	4.38	8.08	Canadian Arctic
16	1.10	1.75	3.16	2.13	3.34	5.17	3.16	4.74	6.82	PE = .203
8	1.24	1.90	3.25	2.37	3.55	4.80	3.36	4.57	5.33	30 db
4	1.37	1.99	2.74	2.49	3.00	3.32	2.96	3.27	3.32	Flat coding
64	.84	1.38	3.67	1.81	3.03	6.34	2.97	4.81	8.72	Line 3
32	.92	1.52	4.15	1.99	3.33	6.98	3.26	5.29	9.03	Canadian Arctic
16	1.02	1.68	4.26	2.21	3.41	6.82	4.06	5.88	8.12	PE = .183
8	1.15	i.88	4.08	2.56	3.69	5.05	3.91	4.80	5.33	30 db
4	1.27	2.03	3.13	2.52	2.96	3.32	3.04	3.27	3.32	Flat coding

RE-ORDER No. 64-712
Vol I

TABLE A. 9-4

LINEAR APPROXIMATION SIMULATION COMPUTER RESULTS FOR THE CANADIAN ARCTIC SCENE PROCESSED ON EDITS AS A FUNCTION OF SENSOR SIGNAL-TO-NOISE RATIO (CONT'D)

	0 Errors			1 Error			2 Errors			
Error Band Δx	±1	±2	±4	±1	±2	±4	±1	±2	±4	
64	.59	.69	.87	1.12	1.50	2.49	2.18	3.40	5.81	Line 1
32	.65	.76	.96	1.23	1.65	2.74	2.40	3.74	6.98	Canadian Arctic
16	.73	.84	1.07	1.37	1.83	3.04	2.66	4.06	6.09	PE = .117
8	.82	.95	1.20	1.54	2.04	3.09	2.74	3.69	4.92	20 db
4	.94	1.08	1.36	1.70	2.17	2.74	2.64	3.00	3.27	Flat coding
64	.60	.70	.92	1.07	1.41	2.79	2.05	2.97	4.81	Line 2
32	.66	.77	1.01	1.18	1.55	3.07	2.25	3.26	5.29	Canadian Arctic
16	.73	.86	1.13	1.31	1.72	3.41	2.50	3.63	5.68	PE = .144
8	.82	.96	1.27	1.47	1.93	3.09	2.90	3.55	4.80	20 db
4	.94	1.09	1.41	1.64	2.03	2.81	2.52	2.84	3.22	Flat coding
64	.60	.68	.94	1.12	1.43	2.49	2.02	3.10	5.17	Line 3
32	.67	.75	1.03	1.23	1.58	2.74	2.22	3.41	5.68	Canadian Arctic
16	.74	.84	1.15	1.37	1.75	3.04	2.47	3.79	5.88	PE = .128
8	.83	.94	1.28	1.53	1.93	3.14	2.86	3.76	5.18	20 db
4	.95	1.08	1.40	1.75	2.01	2.74	2.67	2.92	3.27	Flat coding

Sheet 3 of 3

RE-ORDER No. 64-712.
Vol I

TABLE A. 9-5

LINEAR APPROXIMATION SIMULATION COMPUTER RESULTS FOR THE CANADIAN ARCTIC SCENE PROCESSED ON EDITS AS A FUNCTION OF SENSOR S/N, WITH THE LINE PREVIOUS-ELEMENT QUALIFIED (PEQ'D) BEFORE PROCESSING

	0 Errors			1 Error			2 Errors			
Error Band Δx	±1	±2	±4	±1	±2	±4	±1	±2	±4	
64	1.19	1.76	3.24	1.83	2.68	4.81	2.58	3.67	7.34	Line 1'
32	1.31	1.94	3.57	2.02	2.95	5.29	2.84	4.04	7.31	Canadian Arctic
16	1.45	2.16	4.06	2.24	3.22	5.50	3.16	4.61	7.11	PE' = .324
8	1.60	2.31	4.00	2.52	3.36	4.92	3.42	4.57	5.64	40 db
4	1.72	2.41	3.04	2.52	2.92	3.32	3.00	3.32	3.37	Flat coding
64	1.19	1.83	3.32	1.99	2.73	4.50	2.97	3.87	7.75	Line 2'
32	1.31	2.02	3.65	2.19	3.01	4.51	3.26	4.26	6.98	Canadian Actic
16	1.44	2.18	4.06	2.40	3.34	5.17	3.55	4.61	7.11	PE' = .355
8	1.60	2.49	4.00	2.59	3.62	4.80	3.62	4.08	5.48	40 db
4	1.72	2.28	3.04	2.46	2.88	3.37	2.96	3.18	3.42	Flat coding
64	1.99	3.87	6.64	2.90	4.98	8.21	4.10	7.34	9.97	Line 3'
32	2.19	4.38	6.98	3.20	5.48	8.08	4.51	8.08	10.24	Canadian Arctic
16	2.40	3.96	6.82	3.71	5.88	7.42	4.87	7.75	8.53	PE' = .542
8	2.52	4.08	5.05	3.76	5.18	5.64	4.68	5.64	5.81	40 db
4	2.46	3.00	3.37	3.00	3.27	3.37	3.27	3.32	3.42	Flat coding

RE-ORDER No. 64-712
Vol I

TABLE A. 9-5

LINEAR APPROXIMATION SIMULATION COMPUTER RESULTS FOR THE CANADIAN ARCTIC SCENE PROCESSED ON EDITS AS A FUNCTION OF SENSOR S/N, WITH THE LINE PREVIOUS-ELEMENT QUALIFIED (PEQ'D) BEFORE PROCESSING (CONT'D)

	0 Errors			1 Error			2 Errors			
Error Band Δx	±1	±2	±4	±1	±2	±4	±1	±2	±4	
64	1.01	1.66	3.98	1.81	2.68	6.64	2.79	3.98	9.97	Line 1'
32	1.11	1.82	4.38	1.99	2.95	6.98	3.07	4.38	10.97	Canadian Arctic
16	1.23	2.03	5.17	2.21	3.28	5.50	3.41	4.87	8.98	PE' = .398
8	1.38	2.25	4.68	2.40	3.69	5.33	3.69	4.80	5.81	30 db
4	1.50	2.21	3.13	2.61	3.13	3.42	3.22	3.37	3.42	Flat coding
64	1.01	1.60	2.97	1.76	2.68	4.65	2.53	3.98	5.58	Line 2'
32	1.12	1.76	3.26	1.94	2.95	5.12	2.79	4.38	6.40	Canadian Arctic
16	1.24	1.96	3.63	2.16	3.22	5.01	3.10	4.87	6.56	PE' = .328
8	1.41	2.02	3.55	2.43	3.55	4.68	3.31	4.36	5.33	30 db
4	1.54	2.05	2.81	2.46	3.00	3.32	2.96	3.27	3.32	Flat coding
64	.92	1.62	3.67	1.83	3.10	7.34	2.73	4.81	8.21	Line 3'
32	1.01	1.78	4.38	2.02	3.41	6.98	3.01	5.29	8.53	Canadian Arctic
16	1.13	1.98	4.06	2.24	3.71	6.82	3.41	5.50	8.12	PE' = .339
8	1.30	2.13	4.26	2.46	3.69	5.05	3.62	4.80	5.33	30 db
4	1.42	2.17	3.09	2.43	3.04	3.32	2.96	3.27	3.32	Flat coding

Sheet 2 of 3

RE-ORDER No. 64-712
Vol I

TABLE A. 9-5

LINEAR APPROXIMATION SIMULATION COMPUTER RESULTS FOR THE CANADIAN ARCTIC SCENE PROCESSED ON EDITS AS A FUNCTION OF SENSOR S/N, WITH THE LINE PREVIOUS-ELEMENT QUALIFIED (PEQ'D) BEFORE PROCESSING (CONT'D)

	0 Errors			1 Error			2 Errors			
Error Band Δx	±1	±2	±4	±1	±2	±4	±1	±2	±4	
64	.62	.71	.95	1.13	1.55	2.58	2.25	2.97	6.34	Line 1'
32	.68	.79	1.05	1.24	1.70	2.84	2.47	3.26	6.98	Canadian Arctic
16	.76	.87	1.16	1.38	1.89	3.16	2.75	3.63	6.32	PE' = .187
8	.86	.98	1.31	1.56	2.13	3.25	2.90	3.84	5.18	20 db
4	.98	1.13	1.43	1.75	2.21	2.84	2.64	3.04	3.32	Flat coding
64	.63	.74	.94	1.11	1.34	2.63	1.91	2.97	5.17	Line 2'
32	.69	.81	1.04	1.22	1.47	2.89	2.10	3.26	5.68	Canadian Arctic
16	.77	.90	1.16	1.36	1.64	3.22	2.33	3.63	5.50	PE' = .222
8	.86	1.02	1.30	1.53	1.95	3.04	2.63	3.49	4.68	20 db
4	.99	1.16	1.47	1.68	2.03	2.77	2.49	2.77	3.27	Flat coding
64	.62	.73	.97	1.16	1.45	2.73	2.02	3.40	4.81	Line 3'
32	.69	.80	1.07	1.28	1.60	3.01	2.22	3.74	5.29	Canadian Arctic
16	.76	.89	1.19	1.42	1.77	3.34	2.47	3.79	5.88	PE' = .191
8	.86	1.01	1.33	1.57	1.95	3.25	2.86	3.91	5.18	20 db
4	.98	1.14	1.48	1.79	2.07	2.81	2.64	2.96	3.27	Flat coding

RE-ORDER No. 64-712
Vol I

APPENDIX B

SYNCHRONIZATION STUDY

B.1 INTRODUCTION

The synchronization study was divided into two separate areas, one for fixed word length and fixed bits per frame systems, and the other for variable word length and variable bits per frame systems (PEC-Huffman). The ground rules under which the investigations were made are stated below.

- o The PCM signal conditioning function is not included in the study. Bit sync at the receiver output is assumed to be perfect for all conditions which yield a bit error probability of 10^{-1}, or less.
- o $P_1 = P_0 = 0.5$ for the data-bit stream from an uncompressed source, where P_1 is the probability that a given bit is a "one" and P_0 is the probability that a given bit is a "zero".
- o $P_1 = P_0 = 0.5$ for the data-bit stream from a compressed source if all code words have the same length. This assumption applies to delta modulation and the linear approximation technique, for example.
- o The video data source is a 512 line television camera having 512 samples per line. An anticipated mode of operation is aperiodic transmission of single television frames. This places severe requirements on sync acquisition speed.

B.2 SYNCHRONIZATION OF FIXED WORD LENGTH DATA

B.2.1 PCM Subsystem Configuration

The obvious method of synchronizing fixed word length digital data is to employ the techniques which have been developed for use in modern PCM telemetry systems. The basic technique is to insert, usually at the beginning of each commutation sequence (data frame), an n bit binary word which identifies the location of the first bit of the data frame. Synchronization at the receiver, is achieved by employing correlation techniques to recognize the frame sync word; the usual implementation is a shift register recognizer. Each group of n data bits is compared

156

to the sync pattern and a sync signal is generated for each comparison which results in a correlation above a prescribed threshold.

The synchronization process usually involves three distinct modes of operation which can be described as follows:

Search Mode In this mode the correlator is set to tolerate relatively few errors. Operation in this mode terminates with generation of the first frame-sync signal. B...ause of the high correlation threshold, there is a high probability that this initial sync signal is a true one. The probabilities of true sync and false sync are functions of sync pattern length only.

Check Mode The purpose of this mode is to increase the confidence that true sync has been detected. The procedure is to count bit-sync pulses and establish an aperture at the correlator exactly one frame length after the first frame sync signal. The correlation threshold is lowered in this mode to achieve a high probability of recognizing the true sync pattern. If the result of the check is negative, the system reverts to the search mode. After one or more positive checks, the exact number depending on the system requirements, the Lock mode is initiated.

Lock Mode This mode utilizes the aperture at the correlator, as in the Check mode. but the criteria for returning to the Search or Check mode vary from system to system, depending on requirements and design philosophy. In general, selection of the Lock mode parameters is aimed at minimizing the percentage of data lost due to being out of sync. The system usually includes a means of measuring the bit error rate; knowledge of the error rate is then used to optimize the criteria for returning to the Search or Check mode.

B. 2. 2 Subsystem Design Equations - Search Mode

The sync subsystem design is based entirely on various probabilities which must be optimized in accordance with the system requirements. A number of papers exist which give the derivations of equations for calculating the required probabilities. Most of the symbols and equations in the following analysis are taken from Magnin.[1]

[1] Magnin, J. P., Digital Synchronization of PCM Telemeters, Electro-Mechanical Research, Incorporated, Sarasota, Florida.

RE-ORDER No. 64-712
Vol I

The probability that the correlator recognizes the transmitted sync pattern, when the received pattern is in the shift-register recognizer, is a function of the bit error rate, the pattern length and the number of errors allowed in the correlation. Elementary combinatorial analysis leads to the following equation for the probability that the transmitted code arrives in the correlator with e or fewer errors:

$$P_c(e) = \sum_{i=0}^{e} (q)^i (1-q)^{n-i} C_i^n ; \qquad (1)$$

where, q = independent bit error rate,

n = number of bits in sync pattern,

e = number of errors allowed in the correlator, and

C_i^n = the binomial coefficient symbol.

The probability that the correlator recognizes a random group of n data bits as the sync pattern is

$$P_d(e) = (1/2)^n \sum_{i=0}^{e} C_i^n . \qquad (2)$$

If the analysis is made for the worst case, when the correlator begins its search just after a sync code pattern, the probability P_{fm} that a false sync be detected in one frame can easily be calculated.

$$P_{fm} = 1 - \left[1 - P_d(e)\right]^{m-1} ; \qquad (3)$$

where m equals the number of bits in the frame, including the sync pattern.

This equation assumes a sync pattern having autocorrelation properties equal to, or better than, a random group of n data bits for all degrees of partial overlap in the correlator. In other words,

$$P_x(e) \leq P_d(e), \quad o < x < n;$$

where, x equals degree of overlap in the correlator.

Willard[1] shows that patterns meeting this criterion exist for all $n \geq 2$ with

(1) Williard, Merwin W., Technical Report on Optimum Code Patterns for PCM Synchronization, Dynatronics, Inc., Orlando, Florida, 3 October 1960, rev. 15 July 1961.

RE-ORDER No. 64-712
Vol I

$q \geq 0.1$. Thus, equation (3) is valid at least over the range of parameters to be considered in this study.

The probability P_t that the true sync pattern is recognized in the first search frame is,

$$P_t = (1 - P_{fm}) P_c (e). \tag{4}$$

The probability P_n that no sync signal is generated in the first search frame is

$$P_n = (1 - P_{fm}) (1 - P_c). \tag{5}$$

The accumulated probabilities of false sync, P_{fk}, true sync, P_{tk}, and no sync, P_{nk}, after K frames in the search mode are:

$$P_{fk} = \frac{P_{fm} (1 - P_n^K)}{1 - P_n}, \tag{6}$$

$$P_{tk} = \frac{P_t (1 - P_n^K)}{1 - P_n}, \text{ and} \tag{7}$$

$$P_{nk} = P_n^K. \tag{8}$$

As K increases, P_{nk} approaches zero, and

$$\lim_{K \to \infty} \left(P_{fk} \right) = \frac{P_{fm}}{1-P_n}, \tag{9}$$

$$\lim_{K \to \infty} \left(P_{tk} \right) = \frac{P_t}{1-P_n}. \tag{10}$$

A useful design parameter is the ratio, R, of true sync to false sync outputs from the correlator:

$$R = \frac{P_{tk}}{P_{fk}} = \frac{P_c (1-P_{fm})}{P_{fm}} = \frac{P_t}{P_{fm}}. \tag{11}$$

Finally, the number of frames, K, necessary to make P_{tk} reach any desired value less than the maximum can be found with the aid of the identity:

157

RE-ORDER No. 64-712
Vol I

$$P_{nk} + P_{fk} + P_{tk} = 1. \tag{12}$$

Substituting the appropriate values leads to

$$K = \frac{\log\left[1 - P_{tk}\left(1 + \frac{1}{R}\right)\right]}{\log P_n} . \tag{13}$$

One more parameter of interest: In the Search mode, is the average number of frames, K_{av}, to obtain the first sync pulse, either true or false.

$$K_{av} = \frac{1}{1-P_n} . \tag{14}$$

B. 2. 3 Subsystem Design Equations - Check Mode

In the Check mode, the a priori knowledge of the number of bits in the frame is used to determine whether the pattern detected in Search was actually the transmitted sync word or a suitable arrangement of random data bits. The object of Check mode analysis is to determine the optimum number of Checks and the number of errors to be tolerated in the correlator to meet the system requirements at a given bit error rate. A trade-off must be made between confidence in acquiring true sync and speed of acquisition.

The correlator makes its first check exactly one frame (m bits) after the first sync detection in the Search mode. The probability that the Check mode begins on a true sync pattern is P_{tk}, while the probability that the Check mode is begun on a false sync pattern is P_{fk}.

The probability, P_{fsc}, that the correlator discovers a false pattern in Search, checks it j times in the Check mode, and does not discard it, is

$$P_{fsc} = P_{fk} \left[P_d (e_c)\right]^j ; \tag{15}$$

where, e_c = number of errors allowed in the correlator during Check, and

j = number of frames checked.

The probability, P_{tsc}, that the correlator discovers the true sync pattern in Search, checks it j times in the Check mode, and does not discard it, is

RE-ORDER No. 64-712
Vol I

$$P_{tsc} = P_{tk} \left[P_c(e_c)\right]^j \quad . \tag{16}$$

These are only two of the possible events which can occur during Check. For instance, the true code may fail to meet check number (j-1), be rediscovered, and again fail to meet the required number of checks. It can be shown, however, that for normal Search and Check mode parameters, the probability of events other than P_{fsc} and P_{tsc} is small.

B. 2. 4 Subsystem Design Equations - Lock Mode

It was possible to give a general analysis of the search and check modes because the requirements and implementation of these modes are well defined. This is not true in the case of the Lock mode, because there is no generally accepted implementation concept. Once the implementation is established, however, the analysis of the Lock mode is simpler than the Search and Check mode analyses. The necessary equations for Lock mode analysis are included in those developed for the other two modes.

B. 2. 5 Subsystem Performance Requirements

In implementing the sync subsystem, each line of television data will be treated as a PCM data frame; that is, each line will be preceded by an n bit code pattern for the purpose of establishing horizontal synchronization.

Since transmission of single television frames is assumed, it is essential to have a high probability of acquiring sync in a very short time to avoid missing substantial portions of the televised scenes. Maintaining a high probability of staying in sync once it is acquired is of equal importance. Although the transmission link will undoubtedly be designed to produce low bit error rates, it is desirable to design the sync subsystem to perform under abnormal conditions, when the bit error rate may be as high as 0. 1.

The above qualitative requirements must be translated into a set of quantitative performance specifications which can be used to select the subsystem parameters. Thus, the specifications below can be expected to yield optimum sync performance in the system, as defined by the assumptions in the introduction.

161

RE-ORDER No. 64-712
Vol I

Search Mode Performance. In this mode we specify a probability of at least 0.999 that true sync be acquired within five frames or less. This means that the search parameters must be such that when 0.999 is substituted for P_{tk}, in equation 13, the calculated value of K must be five or less.

Check Mode Performance. To keep the total acquisition time small we specify that only one check be made. In addition, we specify that P_{tsc} be at least 0.99 and that P_{fsc} be no greater than 10^{-6}.

Lock Mode Performance. For our special case of single frame transmission it is convenient to define an additional performance factor. Let $P_{t\ell}$ be the probability that the correlator recognizes every horizontal sync code for the entire frame in the Lock mode. In other words, $P_{t\ell}$ is the probability of staying in line sync, provided that bit sync is maintained in the receiver. In order to assure high quality picture reproduction, we specify that the Lock mode parameters must yield a $P_{t\ell}$ of at least 0.999.

At the same time, we must require a high probability that out-of-sync conditions be rapidly detected. Let us define $P_{f\ell y}$ as the probability that an out-of-sync condition is not detected within y lines. Finally, we specify that $P_{f\ell y}$ must be less than 10^{-3} for y equals 2.

B.2.6 Calculation of Subsystem Parameters

The subsystem parameters to meet the above performance specifications will be computed assuming four bit intensity quantization and a 10^{-1} bit error probability. Tables B-1 and B-2 contain all the performance criteria necessary to design the subsystem. The calculations were done on EMR's ASI-210 computer. Although in this case the computer was programmed to print out the results for a wide range of variation in the parameters, the program is easily modified such that the computer selects the parameters required to meet the specifications.

Search Mode Parameters Reference to Table B-2 shows that to obtain a P_{tk} of 0.999 requires at least 36 frames with a 30 bit pattern and at least 7.3 frames with a 40 bit pattern. With a 50 bit pattern, and allowing seven errors in the correlator there is a 0.999 probability of acquiring true sync in 3.4 frames; there is only a 2.5×10^{-4} probability that a false sync will occur in the same

162

number of frames. Thus, the Search mode parameters meeting our specifications are established:

$$n = 50,$$
$$e = 7.$$

Check Mode Parameters Using equations 15 and 16 and substituting values of P_d and P_c from Table B-1, we arrive at the following tabulation of P_{tsc} and P_{tfc}, for a 50 bit sync pattern.

e_c	P_{tsc}	P_{fsc}
10	0.989	2.98×10^{-9}
11	0.996	1.13×10^{-8}
12	0.998	3.83×10^{-8}
13	0.998+	1.17×10^{-7}

The check mode specifications are satisfied with $e_c = 11$, 12, or 13; $e_c = 11$ is preferred due to the fact that P_{fsc} is changing much faster than P_{tsc} in the wrong direction. This establishes the Check mode parameters:

$$j = 1,$$
$$e_c = 11.$$

Lock Mode Parameters For this system, the probability P_t is calculated by the equation:

$$P_{t\ell} = P_c^{\,512}$$

if the criterion for return to Search mode is a single correlation failure. Even for 13 errors allowed in the correlator, which yields $P_c = 0.999$ according to Table B-1, $P_{t\ell}$ is approximately 0.6.

If we change the criterion for return to the Search mode to two consecutive correlation failures, then the expression for $P_{t\ell}$ becomes

$$P_{t\ell} = \left[1 - (1-P_c)^2\right]^{512}.$$

RE-ORDER No. 64-712
Vol I

N	ε	$P_d(e)$	$P_c(e)$	P_{fm}	P_t	P_n	R	$\lim_{K\to\infty} P_{fl}$	$\lim_{K\to\infty} P_{th}$	$K_{avg} = \frac{1}{1-P_n}$
30		9.31^{-10}	.042	1.93^{-6}	.042	.958	2.19^{+4}	4.96^{-5}	.999954371	23.8
40	0	9.09^{-13}	.015	1.90^{-9}	.015	.985	7.79^{+6}	1.28^{-7}	.999999871	66.6
50		8.88^{-15}	.005	1.06^{-11}	.005	.995	2.67^{+9}	3.61^{-10}	→ 1	200.0
30		2.09^{-8}	.184	6.00^{-5}	.184	.816	3.06^{+3}	3.26^{-4}	.999673649	5.4
40	1	3.73^{-11}	.080	7.78^{-8}	.081	.920	1.03^{+6}	9.67^{-7}	.999999033	12.5
50		4.53^{-14}	.034	9.50^{-11}	.034	.966	3.55^{+8}	2.01^{-9}	.999999997	29.4
30		4.34^{-7}	.411	9.01^{-4}	.411	.588	4.56^{+2}	2.19^{-3}	.997811403	2.4
40	2	7.47^{-10}	.223	1.56^{-6}	.223	.777	1.63^{+5}	6.99^{-6}	.999993006	4.5
50		1.13^{-12}	.112	2.38^{-9}	.112	.888	4.70^{+7}	2.13^{-8}	.999999979	9.0
30		4.22^{-6}	.647	8.75^{-3}	.642	.349	7.33^{+1}	1.35^{-2}	.986541810	1.5
40	3	9.73^{-9}	.423	2.03^{-5}	.423	.576	2.08^{+4}	4.80^{-5}	.999951990	2.4
50		1.83^{-11}	.250	3.39^{-8}	.250	.750	6.44^{+6}	1.55^{-7}	.999999844	4.0
30		..97^{-5}	.825	5.99^{-2}	.775	.165	1.29^{+1}	7.17^{-2}	.928267836	1.2
40	4	9.29^{-8}	.629	1.94^{-4}	.629	.371	3.25^{+3}	3.08^{-4}	.999691967	1.6
50		2.23^{-10}	.431	4.68^{-7}	.431	.569	9.22^{+5}	1.08^{-6}	.999998915	2.3
30		1.62^{-4}	.927	2.06^{-1}	.661	.052	2.31	3.02^{-1}	.697804358	1.1
40	5	6.91^{-7}	.794	1.44^{-3}	.793	.206	5.49^{+2}	1.82^{-3}	.998182985	1.3
50		2.10^{-9}	.616	4.41^{-6}	.616	.384	1.40^{+5}	7.16^{-6}	.999992836	1.6
30		7.15^{-4}	.974	7.74^{-1}	.220	.006	.285	7.78^{-1}	.222	1.0
40	6	4.10^{-6}	.900	8.72^{-3}	.892	.099	1.02^{+2}	9.68^{-3}	.990316278	1.1
50		1.62^{-8}	.770	3.40^{-5}	.770	.230	2.26^{+4}	4.42^{-5}	.999955844	1.3
30		2.61^{-3}	.992	9.96^{-1}	.434	3.41^{-5}	4.36^{-3}	9.95^{-1}	4.34454437^{-3}	-
40	7	2.11^{-5}	.958	4.32^{-2}	.917	.040	2.12^{+1}	4.50^{-2}	.955039429	1.0
50		1.05^{-7}	.878	2.20^{-4}	.878	.122	3.99^{+3}	2.51^{-4}	.999749344	1.1
30		8.06^{-3}	.998	.999+	.497	1.01^{-10}	4.97^{-8}	.999+	4.98^{-8}	-
40	8	9.11^{-5}	.985	1.73^{-1}	.814	.013	4.70	1.75^{-1}	.824630947	1.0
50		5.82^{-6}	.942	1.22^{-3}	.941	.058	7.71^{+2}	1.29^{-3}	.998705175	1.1
30		-	-	-	-	-	-	-	-	-
40	9	3.40^{-4}	.995	5.08^{-1}	.490	.003	.964	$.509^{-1}$	.491	1.0
50		2.81^{-6}	.975	5.89^{-3}	.970	.024	1.65^{+2}	6.03^{-3}	.993966436	1.0
30		-	-	-	-	-	-	-	-	-
40	10	1.11^{-3}	.999	9.02^{-1}	.098	1.45^{-4}	1.09^{-1}	9.08^{-1}	.0982067608	1.0
50		1.19^{-5}	.990	2.47^{-2}	.966	.009	3.91^{+1}	2.49^{-2}	.975064179	1.0
30		-	-	-	-	-	-	-	-	-
40	11	3.21^{-3}	.999+	9.99^{-1}	.001	4.61^{-7}	1.21^{-3}	$.999^{-1}$	.001209708	1.0
50		4.51^{-5}	.997	9.02^{-2}	.907	.003	1.00^{+1}	9.05^{-2}	.909478185	1.0
30		-	-	-	-	-	-	-	-	-
40	12	8.29^{-3}	.999+	.999+	2.84^{-8}	2.51^{-12}	2.84^{-8}	.999+	2.84^{-8}	1.0
50		1.53^{-4}	.999	2.74^{-1}	.724	.001	2.64	2.74^{-1}	.725424177	1.0
50	13	4.68^{-4}	.999+	6.25^{-1}	.375	1.07^{-4}	.560	6.25^{-1}	.375	1.0
50	14	1.30^{-3}	.999+	9.35^{-1}	.065	4.81^{-6}	6.98^{-2}	9.35^{-1}	.0009760158	1.0
50	15	3.30^{-3}	.999+	9.99^{-1}	.001	1.70^{-8}	9.77^{-4}	9.99^{-1}	9.64^{-8}	1.0

INDIVIDUAL BIT ERROR RATE = 0.1
INTENSITY QUANTIZATION = 4 bits
DATA BITS PER LINE = 2048

(Note: 9.31^{-10} means 9.31×10^{-10}, etc.)

TABLE B-1

PERFORMANCE PROBABILITIES OF LINE SYNC SUBSYSTEM FOR UNCOMPRESSED TELEVISION DATA

RE-ORDER No. 64-712
Vol I

For $P_c = 0.999$, $P_{t\ell} = 0.9995$ which meets our specification. For y = 2, $P_{f\ell y} = 2\,P_d$. If 12 errors are allowed in the correlator, $P_{f\ell y} = 3.06 \times 10^{-4}$ and is well within our specifications. The lock mode parameters are now established: e = 12, return to Search after two consecutive correlation failures.

3.7.7 Summary of Subsystem Performance

A horizontal synchronization subsystem has been designed for use with a single frame digital television system having 512 x 512 line resolution and four bit intensity quantization.

With a 0.1 bit error rate, the odds are 1000 to 1 that the true synchronization pattern is found within 3.4 lines. Consulting Table B-1, we find that, on the average, only 1.1 lines are required to find true sync. The odds are 4000 to 1 against the possibility of finding a false sync pattern during the first 3.4 lines.

By adding one line of Check mode operation, we have a total acquisition time of 4.4 lines with a 0.996 probability of true sync as opposed to a 1.13×10^{-8} probability of false sync. These figures represent a virtual certainty that the Lock mode is established with true synchronization.

The Lock mode strategy adopted yields a probability of 0.9995 that synchronization will not be lost during the entire frame because of failure to recognize the true sync pattern. At the same time, the odds are 3270 to 1 that an out-of-sync condition is detected within two lines.

The 50-bit sync code recommended is relatively long compared to many conventional PCM systems, but considering the fact that each line contains 2048 data bits, this amounts to an increase of only 2.4% in the total number of bits transmitted. This is a small price to pay for the synchronization performance achieved.

Tables B-3 and B-4 are design tables identical to Tables B-1 and B-2, but calculated for a bit error rate of 10^{-2}; these tables are included to show how sync performance varies as a function of bit error rate.

RE-ORDER No. 64-712
Vol I

P_{th}	C	n = 30 K	n = 30 P_{fl}	n = 30 R	n = 30 K_{avg}	n = 40 K	n = 40 P_{fl}	n = 40 R	n = 40 K_{avg}	n = 50 K	n = 50 P_{fl}	n = 50 R	n = 50 K_{avg}
0.9	0	53.2	4.10^{-5}	2.19^{+4}	23.8	154.6	1.16^{-7}	7.79^{+6}	66.6	445	3.25^{-10}	2.67^{+9}	200.0
	1	11.4	2.94^{-4}	3.06^{+3}	5.4	27.4	8.70^{-7}	1.03^{+6}	12.5	67.0	2.53^{-9}	3.55^{+8}	29.4
	2	4.4	1.97^{-3}	4.56^{+2}	2.4	9.1	6.29^{-6}	1.43^{+5}	4.5	19.4	1.91^{-8}	4.70^{+7}	9.0
	3	2.3	1.23^{-2}	7.33^{+1}	1.5	4.2	4.32^{-5}	2.08^{+4}	2.4	8.0	1.40^{-7}	6.44^{+6}	4.0
	4	1.9	6.92^{-2}	1.29^{+1}	1.2	2.3	2.77^{-4}	3.25^{+3}	1.6	4.1	9.76^{-7}	9.22^{+5}	2.3
	5					1.5	1.64^{-3}	5.49^{+2}	1.3	2.4	6.45^{-6}	1.40^{+5}	1.6
	6					1.0	8.00^{-3}	1.02^{+2}	1.1	1.6	3.97^{-5}	2.26^{+4}	1.3
	7					0.9	4.23^{-2}	2.12^{+1}	1.0	1.1	2.26^{-4}	3.99^{+3}	1.1
	8									0.8	1.17^{-3}	7.71^{+2}	1.0
	9									0.6	5.46^{-3}	1.65^{+2}	1.0
	10									0.5	2.30^{-2}	3.91^{+1}	1.0
	11												
0.99	0	106.4	4.52^{-5}	2.19^{+4}	23.8	399.3	1.27^{-7}	7.79^{+6}	66.6	891	3.58^{-10}	2.67^{+9}	200.0
	1	22.8	3.23^{-4}	3.06^{+3}	5.4	54.9	9.57^{-7}	1.03^{+6}	12.5	134.0	2.78^{-9}	3.55^{+8}	29.4
	2	9.1	2.17^{-3}	4.56^{+2}	2.4	18.3	6.92^{-6}	1.43^{+5}	4.5	38.9	2.11^{-8}	4.70^{+7}	9.0
	3					8.4	4.75^{-5}	2.08^{+4}	2.4	16.0	1.54^{-7}	6.44^{+6}	4.0
	4					4.7	3.05^{-4}	3.25^{+3}	1.6	8.2	1.07^{-6}	9.22^{+5}	2.3
	5					3.0	1.80^{-3}	5.49^{+2}	1.3	4.8	7.09^{-6}	1.40^{+5}	1.6
	6									3.1	4.37^{-5}	2.26^{+4}	1.3
	7									2.2	2.40^{-4}	3.99^{+3}	1.1
	8									1.7	1.28^{-3}	7.71^{+2}	1.0
	9									1.4	6.01^{-3}	1.65^{+2}	1.0
	10												
	11												
0.999	0	160.5	4.56^{-5}	2.19^{+4}	23.8	463.9	1.28^{-7}	7.79^{+6}	66.6	1337	3.61^{-10}	2.67^{+9}	200.0
	1	36.0	3.26^{-4}	3.06^{+3}	5.4	82.3	9.66^{-7}	1.03^{+6}	12.5	201.0	2.81^{-9}	3.55^{+8}	29.4
	2					27.4	6.99^{-6}	1.43^{+5}	4.5	58.3	2.12^{-8}	4.70^{+7}	9.0
	3					12.6	4.80^{-5}	2.08^{+4}	2.4	24.0	1.55^{-6}	6.44^{+6}	4.0
	4					7.3	3.08^{-4}	3.25^{+3}	1.6	12.2	1.08^{-6}	9.22^{+5}	2.3
	5									7.2	7.16^{-6}	1.40^{+5}	1.6
	6									4.7	4.41^{-5}	2.26^{+4}	1.3
	7									3.4	2.50^{-4}	3.99^{+3}	1.1
	8												
	9												
	10												
	11												
0.9999	0	226.7	4.56^{-5}	2.19^{+4}	23.8	618.6	1.28^{-7}	7.79^{+6}	66.6	1781	3.61^{-10}	2.67^{+9}	200.0
	1					109.9	9.67^{-7}	1.03^{+6}	12.5	268.0	2.81^{-9}	3.55^{+8}	29.4
	2					36.8	6.99^{-6}	1.43^{+5}	4.5	77.7	2.13^{-8}	4.70^{+7}	9.0
	3					17.9	4.80^{-5}	2.08^{+4}	2.4	32.0	1.55^{-6}	6.44^{+6}	4.0
	4									16.3	1.08^{-6}	9.22^{+5}	2.3
	5									9.7	7.16^{-6}	1.40^{+5}	1.6
	6									6.6	4.42^{-5}	2.26^{+4}	1.3
	7												
	8												
	9												
	10												
	11												

INDIVIDUAL BIT ERROR RATE = 0.1
INTENSITY QUANTIZATION = 4 bits
DATA BITS PER LINE = 2048

TABLE B-2

SEARCH MODE PERFORMANCE OF LINE SYNC SUBSYSTEM FOR UNCOMPRESSED TELEVISION DATA

RE-ORDER No. 64-712
Vol I

N	ℓ	P_d	P_e	P_{fm}	P_t	P_n	R	$\lim_{K\to\infty} P_{fk}$	$\lim_{K\to\infty} P_{tk}$	$K_{avg} = \frac{1}{1-P_n}$
15	0	3.05^{-5}	.860	6.10^{-2}	.860	.131	1.32^{+1}	7.02^{-2}	.929784037	1.1
20		9.53^{-7}	.818	1.97^{-3}	.816	.182	4.44^{+2}	2.41^{-3}	.997590948	1.2
25		2.98^{-8}	.778	6.18^{-4}	.778	.222	1.26^{+4}	7.94^{-5}	.999920613	1.3
30		9.31^{-10}	.749	1.93^{-6}	.740	.260	3.82^{+5}	2.62^{-6}	.999997385	1.4
35		2.91^{-11}	.703	6.06^{-8}	.703	.297	1.16^{+7}	8.61^{-8}	.999999914	1.4
15	1	4.88^{-4}	.990	6.35^{-1}	.362	.004	5.70^{-1}	6.37^{-1}	.363042509	1.0
20		2.00^{-5}	.983	4.06^{-2}	.943	.016	2.33^{+1}	4.12^{-2}	.958782252	1.0
25		7.75^{-7}	.974	1.61^{-3}	.972	.026	6.06^{+2}	1.65^{-3}	.998352109	1.0
30		2.89^{-8}	.964	6.08^{-5}	.964	.036	1.61^{+4}	6.22^{-5}	.999937786	1.0
35		1.05^{-9}	.952	2.18^{-6}	.952	.048	4.36^{+5}	2.29^{-6}	.999997709	1.0
15	2	3.69^{-3}	.999+	9.99^{-1}	4.86^{-4}	2.02^{-7}	4.87^{-4}	.9995	.00486	1.0
20		2.01^{-4}	.999+	3.40^{-1}	.659	.001	1.93	3.40^{-1}	.659470995	1.0
25		9.72^{-6}	.998	2.01^{-2}	.978	.002	4.86^{+1}	2.01^{-2}	.979830814	1.0
30		4.34^{-7}	.997	9.01^{-4}	.996	.003	1.10^{+3}	9.04^{-4}	.999095597	1.0
35		1.84^{-8}	.994	3.82^{-5}	.995	.005	2.60^{+4}	3.84^{-5}	.999961567	1.0
15	3	-	-	-	-	-	-	-	-	-
20		1.29^{-3}	.999+	9.30^{-1}	.070	2.97^{-6}	7.48^{-2}	.9304	.06961	1.0
25		7.83^{-5}	.999+	1.50^{-1}	.850	9.09^{-5}	5.68	1.50^{-1}	.850289109	1.0
30		4.22^{-6}	.999+	8.75^{-3}	.991	2.20^{-4}	1.13^{+2}	8.76^{-3}	.991243168	1.0
35		2.09^{-7}	.999+	4.35^{-4}	.999	4.09^{-4}	2.30^{+3}	4.35^{-4}	.999564998	1.0
15	4	-	-	-	-	-	-	-	-	-
20		5.91^{-3}	.999+	.999+	4.78^{-6}	6.50^{-12}	4.78^{-6}	.999+	4.78^{-6}	1.0
25		4.55^{-4}	.999+	6.11^{-1}	.389	1.75^{-6}	6.37^{-1}	6.11^{-1}	.389256523	1.0
30		2.97^{-5}	.999+	5.99^{-2}	.940	1.09^{-5}	1.57^{+1}	5.99^{-2}	.940101784	1.0
35		1.73^{-6}	.999+	3.61^{-3}	.996	2.32^{-5}	2.76^{+2}	3.61^{-3}	.996392374	1.0
15	5	-	-	-	-	-	-	-	-	-
20		-	-	-	-	-	-	-	-	-
25		2.03^{-3}	.999+	9.85^{-1}	.015	2.91^{-9}	1.48^{-2}	9.85^{-1}	.0146	1.0
30		1.62^{-4}	.999+	2.86^{-1}	.713	3.33^{-7}	2.49	2.86^{-1}	.713587193	1.0
35		1.12^{-5}	.999+	2.30^{-2}	.977	1.22^{-6}	4.25^{+1}	2.30^{-2}	.976990953	1.0
15	6	-	-	-	-	-	-	-	-	-
20		-	-	-	-	-	-	-	-	-
25		7.32^{-3}	.999+	.999+	2.46^{-7}	3.44^{-16}	2.46^{-7}	.999+	2.46^{-7}	1.0
30		7.15^{-4}	.999+	.774	.226	3.16^{-10}	.292	.774	.226156374	1.0
35		5.84^{-5}	.999+	1.15^{-1}	.885	3.18^{-8}	7.73	1.15^{-1}	.885470815	1.0
30	7	2.61^{-3}	.999+	9.96^{-1}	4.38^{-3}	6.12^{-12}	4.40^{-3}	9.96^{-1}	.00437847665	1.0
35		2.54^{-4}	.999+	4.11^{-1}	.059	8.23^{-10}	1.43	4.11^{-1}	.589095772	1.0
30	8	8.06^{-3}	.999+	.999+	4.98^{-8}	4.64^{-17}	4.98^{-8}	.999+	4.98^{-8}	1.0
35		9.39^{-4}	.999+	8.59^{-1}	.014	1.98^{-10}	1.65^{-1}	8.59^{-1}	.141401618	1.0
35	9	2.99^{-3}	.999+	9.98^{-1}	.002	2.72^{-12}	1.95^{-3}	9.98^{-1}	.00194404507	1.0
35	10	8.34^{-3}	.999+	.999+	2.70^{-8}	3.77^{-17}	2.70^{-8}	.999+	2.70^{-8}	1.0

INDIVIDUAL BIT ERROR RATE = 10^{-2}
INTENSITY QUANTIZATION = 4 bits
DATA BITS PER LINE = 2048

(Note: 3.05^{-5} means 3.05×10^{-5}, etc.)

TABLE B-3

PERFORMANCE PROBABILITIES OF LINE SYNC
SUBSYSTEM FOR UNCOMPRESSED TELEVISION DATA

P_{tk}	ε	n = 15				n = 20				n = 25				n = 30				n = 35			
		K	P_{fk}	R	K_{avg}	K	P_{fk}	R	K_{avg}	K	P_{fk}	R	K_{avg}	K	P_{fk}	R	K_{avg}	K	P_{fk}	R	K_{avg}
0.9000	0	1.7	6.76^{-2}	1.32^{+1}	1.1	1.4	2.17^{-3}	4.14^{+2}	1.2	1.5	7.15^{-5}	1.26^{+4}	1.3	1.7	2.35^{-6}	3.82^{+5}	1.4	1.9	7.75^{-8}	1.16^{+7}	1.4
	1					0.7	3.86^{-2}	2.33^{+1}	1.0	0.6	1.49^{-3}	6.06^{+2}	1.0	0.7	5.60^{-5}	1.61^{+4}	1.0	0.8	2.06^{-6}	4.36^{+5}	1.0
	2									0.4	1.85^{-2}	4.86^{+1}	1.0	0.4	8.15^{-4}	1.10^{+3}	1.0	0.4	3.46^{-5}	2.60^{+4}	1.0
	3													0.3	7.95^{-3}	1.13^{+2}	1.0	0.3	3.92^{-4}	2.30^{+3}	1.0
	4													0.3	5.71^{-2}	1.57^{+1}	1.0	0.2	3.26^{-3}	2.76^{+2}	1.0
	5																	0.2	2.12^{-2}	4.25^{+1}	1.0
0.9900	0					2.9	2.39^{-3}	4.14^{+2}	1.2	3.1	7.86^{-5}	1.26^{+4}	1.3	3.4	2.59^{-6}	3.82^{+5}	1.4	3.8	8.53^{-8}	1.16^{+7}	1.4
	1									1.3	1.63^{-3}	6.06^{+2}	1.0	1.4	6.16^{-5}	1.61^{+4}	1.0	1.5	2.27^{-6}	4.36^{+5}	1.0
	2													0.8	8.96^{-4}	1.10^{+3}	1.0	0.9	3.81^{-5}	2.60^{+4}	1.0
	3													0.8	8.75^{-3}	1.13^{+2}	1.0	0.6	4.31^{-4}	2.30^{+3}	1.0
	4																	0.5	3.58^{-3}	2.76^{+2}	1.0
	5																				
0.9990	0									4.7	7.93^{-5}	1.26^{+4}	1.3	5.1	2.61^{-6}	3.82^{+5}	1.4	5.7	8.61^{-8}	1.16^{+7}	1.4
	1													2.1	6.22^{-5}	1.61^{+4}	1.0	2.3	2.29^{-6}	4.36^{+5}	1.0
	2													1.6	9.04^{-4}	1.10^{+3}	1.0	1.3	3.84^{-5}	2.60^{+4}	1.0
	3																	1.0	4.35^{-4}	2.30^{+3}	1.0
	4																				
	5																				
0.9999	0									7.2	7.94^{-4}	1.26^{+4}	1.3	4.9	2.61^{-6}	3.82^{+5}	1.4	7.6	8.61^{-8}	1.16^{+7}	1.4
	1													3.1	6.22^{-5}	1.61^{+4}	1.0	3.0	2.29^{-6}	4.36^{+5}	1.0
	2																	4.0	3.84^{-5}	2.60^{+4}	1.0
	3																				
	4																				
	5																				

INDIVIDUAL BIT ERROR RATE = 10^{-2}
INTENSITY QUANTIZATION = 4 Bits
DATA BITS PER LINE = 2048

TABLE B-4

SEARCH MODE PERFORMANCE OF LINE SYNC SUBSYSTEM FOR UNCOMPRESSED TELEVISION DATA

RE-ORDER No. 64-712
Vol I

B. 3 SYNCHRONIZATION OF VARIABLE WORD LENGTH (PEC-HUFFMAN DATA) SYSTEMS

B. 3. 1 PEC-Huffman Subsystem Configuration

The PEC-Huffman coded television system is a variable word length and variable bits per frame system where the synchronization requires adapting the results of the PCM fixed word length synchronization analysis. The considerations for the PEC-Huffman analysis are: (1) the number of words per data frame are constant; (2) the total number of data bits per frame is variable; (3) the probabilities of obtaining a zero and one are no longer equal, but depend upon the PE statistic; (4) a property of the Huffman coding technique is that it is a self-synchronizing code, and (5) errors in the transmitted data bits may propagate over several words.

The basic correlation technique to recognize the frame sync will be employed, but the aperture location will be determined by counting the number of decoded words rather than bits as in the PCM case. The general remarks of paragraph B. 2. 1 will be applicable regarding the Search, Check, and Lock modes of operation.

B. 3. 2 Subsystem Design Equations - Search Mode

The sync subsystem design is based on the various probabilities which will be optimized in accordance with a set of specified system operation requirements. The development of the PEC-Huffman subsystem design will parallel the PCM design in paragraph B. 2. 2, but under the PEC-Huffman constraints. The probability that the correlator recognizes the transmitted sync pattern is given by equation (17) for PCM.

$$P_c(e) = \sum_{i=1}^{e} (p)^i (1-p)^{n-i} C_i^n \qquad (17)$$

p = independent bit error probability

n = number of bits in the sync pattern

e = number of errors allowed in the correlator

C_i^n = binomial coefficient symbol

RE-ORDER No. 64-712
Vol I

Since this equation is independent of the coding technique, the equation is also valid for the variable word length data systems. For the remainder of this section, the primes will indicate the appropriate PEC-Huffman sync probabilities. Therefore, for PEC-Huffman, the probability of correctly recognizing the sync words is

$$P_c'(e) = \sum_{i=0}^{e} (p)^i (1-p)^{n-i} C_i^n . \tag{18}$$

The probability that the correlator recognizes a random group of n data bits as the sync pattern for PCM is given by

$$P_d(e) = (1/2)^n \sum_{i=0}^{e} C_i^n . \tag{19}$$

This equation must be modified since, in general, for the PEC-Huffman technique, $P_0 \neq P_1$. Furthermore, it is necessary not only to calculate P_0 and P_1 for the transmitted data, but also P_0^* and P_1^*, the probability of obtaining a zero and one in the receiver after the data is corrupted with channel noise. For the purposes of this analysis it will be assumed that the transmitted data is composed of words of only two lengths. If the data is quantized to N bits where $N = \log_2 n$ for n, the number of quantized data levels, then PE will be transmitted with a one and the remaining events (1-PE) levels will be transmitted with (N + 1) bits, that is, a zero followed by a N-bit PCM word.

With this condition it is possible to find the probability of obtaining a zero and one in the transmitted (noiseless case) data.

The probability of obtaining a one for the noiseless case is the total number of ones in a data frame divided by the total number of data bits in the data frame.

The number of ones is given by

$$O = PE + (1-PE)\,(N+1)\,(N)/(2)\,(N+1), \tag{20}$$

and the number of data bits in the frame is given by

$$D = PE + (1-PE)\,(N+1). \tag{21}$$

RE-ORDER No. 64-712
Vol I

The probability of obtaining a one is then given by

$$P_1 = O/D = \frac{PE + (1-PE)\ N/2}{PE + (1-PE)\ (N+1)} \ . \tag{22}$$

The probability of obtaining a zero for the noisless case can also be calculated and is given by

$$P_0 = \frac{(1-PE)\ (1 + N/2)}{PE + (1-PE)\ (N+1)} \ . \tag{23}$$

If it is assumed that $P_1 > P_0$, a is the number of ones in the sync word, b is the number of zeros also in the sync word, and n=a + b, the total number of bits in the sync word, the probability of false sync given in equation (18) can be modified for the noisless PEC-Huffman case and is given by

$$P'_{dn}(e) = \sum_{e=0}^{k} \sum_{r=0}^{e} P_1^{(a - e + 2r)} P_0^{(b + e - 2r)} C^a_{e-r} \cdot C^b_r \ . \tag{24}$$

In this case, k is the maximum number of errors allowed in the sync word, and the subscript n indicates that it is for the noiseless case.

The total number of ones in the received data is dependent upon the channel bit error probability, p. Assuming a binary symmetrical channel, the received number of ones P_1^* and P_0^* are given by

$$P_1^* = P_1 + p(1-2P_1), \tag{25}$$

and

$$P_0^* = 1-P_1 + p(1-2(1-P_1)). \tag{26}$$

If these two equations are substituted the probability of false sync is given by

$$P'_d(e) = \sum_{e=0}^{k} \sum_{r=0}^{e} P_1^{*(a - e + 2r)} P_0^{*(b + e - 2r)} C^a_{e-r} \ C^b_r \tag{27}$$

Although for the current analysis it is not necessary to calculate PE^* of the received data, the parameter is useful in estimating the data change efforts of channel noise. From equation (22), PE can be obtained as a function of N and P_1

171

and is given by

$$PE = \frac{\left[P_1 - N/(2)\,(N+1)\right]\ (N+1)}{P_1 N - N(N+1)\ /2(N+1) + 1}\ . \tag{28}$$

To determine PE^* it is necessary to substitute P_1^*, P_2^*, and $\left[N/(2)(N+1)\right]^*$ which is the ratio of the ones in the non-PE bits to the total number of non-PE bits and it is given by

$$\left[\frac{N}{2(N+1)}\right]^* = \frac{N}{2(N+1)} + p\left(1 - \frac{N}{(N+1)}\right). \tag{29}$$

Hence,

$$PE^* = \frac{\left(P_1^* - \left[N/2(N+1)\right]^*\right)\ (N+1)}{P_1^* N - \left[N/2\,(N+1)\right]^*\ (N+1) + 1}\ . \tag{30}$$

which becomes

$$PE^* = \frac{\left[P_1 + p(1-2P_1) - (N/2\,(N+1) + p(1 - N/(N+1)))\right]\ (N+1)}{\left[P_1 + p(1 - 2P_1)\right] N - \left[N/2(N+1) + p(1 - N/(N+1))\right] (N+1) + 1}\ . \tag{31}$$

With P_d' and P_c' calculated the remaining probabilities merely are those in paragraph B.2 with P_d' and P_d' substituted. Hence, for completeness this will be given. The probability that a false sync is detected in one frame is given by

$$P'_{fm} = 1 - \left[1 - P'_d\,(e)\right]^{m-1}\ . \tag{32}$$

The probability, P'_t, that the true sync pattern is recognized in the first Search frame is

$$P'_t = (1 - P'_{fm})\ P'_c(e)\ . \tag{33}$$

The probability, P'_n, that no sync signal is generated in the first Search frame is

$$P'_n = (1 - P'_{fm})\ (1 - P'_c(e))\ . \tag{34}$$

The accumulated probabilities of false sync, P'_{fk}, true sync, P'_{tk}, and no sync, P'_{nk}, after K' frames in the Search mode are

RE-ORDER No. 64-712
Vol I

$$P'_{fk} = \frac{P'_{fm}\left(1 - P'^{K'}_{n}\right)}{(1 - P'_{n})} , \tag{35}$$

$$P'_{tk} = \frac{P'_{t}\left[1 - P'^{K'}_{n}\right]}{(1 - P'_{n})} , \text{ and} \tag{36}$$

$$P'_{nk} = P'^{K'}_{n} . \tag{37}$$

As K' increases, P'_{nk} approaches zero, and

$$\lim_{K \to \infty} P'_{fk} = \frac{P'_{fm}}{(1 - P'_{n})} . \tag{38}$$

$$\lim_{K \to \infty} P'_{tk} = \frac{P'_{t}}{(1 - P'_{n})} . \tag{39}$$

A useful design parameter is the ratio R' of true sync to false sync outputs from the correlator:

$$R' = \frac{P'_{tk}}{P'_{fk}} = \frac{P'_{t}}{P'_{fm}} . \tag{40}$$

Finally, the number of frames, K', necessary to make P'_{tk} reach any desired value less than maximum can be found from the identity

$$P'_{nk} + P'_{fk} + P'_{tk} = 1. \tag{41}$$

Substituting in the appropriate values leads to

$$K' = \frac{\log\left[1 - P'_{tk}(1 + 1/R')\right]}{\log(P'_{n})} . \tag{42}$$

Another parameter of interest in the Search mode is the average number of frames, K'_{ave}, to obtain the first sync pulse, either true or false

$$K'_{ave} = 1/(1 - P'_{n}) . \tag{43}$$

RE-ORDER No. 64-712
Vol I

B. 3. 3 Subsystem Design Equations - Check Mode

This is analgous to the Check mode in conventional PCM; therefore the equations of paragraph B. 2. 3 are altered according to the PEC-Huffman concept and become:

$$P'_{fsc} = P'_{fk} \left[P'_{d}(e_c) \right]^{j}, \text{ and} \tag{44}$$

$$P'_{tsc} = P'_{tk} \left[P'_{c}(e_c) \right]^{j}. \tag{45}$$

B. 3. 4 Subsystem Design Equations - Lock Mode

This analysis is the same as that given in paragraph B. 2. 4.

B. 3. 5 Subsystem Performance Requirements

The subsystem performance requirements are identical to PCM requirements.

B. 3. 6 Calculation of Subsystem Parameters

The subsystem parameters to meet the specifications will be computed assuming a four-bit PEC-Huffman system with PE = 0.65 and a 10^{-2} bit error probability. Tables B-5 and B-6 contain the necessary performance criteria to design the subsystem and were obtained from EMR's ASI-210 digital computer.

Search Mode Parameters

Reference to Table B-6 shows that to obtain a P'_{tk} of 0.999 and allowing zero errors requires at least 5.13 frames with a 30-bit pattern, at least 6.25 frames with a 40-bit pattern, and with a 50-bit pattern, at least 7.43 frames. Therefore, it appears that if it is desirable to minimize the sync word length, then we can operate with a 40-bit pattern and allowing four errors in the correlator there is a probability of 0.999 of obtaining true sync in one frame, there is only a 9.7×10^{-5} probability that false sync will occur in one frame. Therefore, for the Search mode it will be specified that n = 40 bits and e = 4 satisfies the Search mode parameters.

Check Mode Parameters

Using equations (44), (45) and substituting $P'_{d}(e)$ and P'_{c} from Table B-7, we arrive at the tabulation that, for a 40-bit sync word, the P'_{tsc} and P'_{fsc} satisfies the requirements for y = 1.

175

RE-ORDER No. 64-712

Vol I

N		a	b	P'_d	P'_c	P'_{fm}	P'_t	P'_n	R'	P_{hm}	P_{hm}
20	0	10	10	*.821E-6	.818E+0	.102E-2	.817E+0	.182E+0	.798E+3	.125E-2	.998747766
30	0	15	15	.744E-9	.740E+0	.936E-6	.740E+0	.260E+0	.791E+6	.126E-5	.999998735
40	0	20	20	.674E-12	.669E+0	.854E-9	.669E+0	.331E+0	.783E+9	.128E-8	.999999999
50	0	25	25	.610E-15	.605E+0	.780E-12	.605E+0	.395E+0	.775E+12	.129E-11	.100000000
20	+1	10	10	.177E-4	.983E+0	.219E-1	.962E+0	.165E-1	.439E+2	.223E-1	.977743775
30	+1	15	15	.237E-7	.964E+0	.299E-4	.964E+0	.361E-1	.323E+5	.310E-4	.999969029
40	+1	20	20	.284E-10	.939E+0	.361E-7	.939E+0	.607E-1	.260E+8	.384E-7	.999999962
50	+1	25	25	.321E-13	.911E+0	.410E-10	.911E+0	.894E-1	.222E+11	.450E-10	.100000000
20	+2	10	10	.183 E-3	.999E+0	.204E+0	.795E+0	.799E-3	.390E+1	.204E+0	.795873400
30	+2	15	15	.366 E-6	.997E+0	.461E-3	.996E+0	.332E-2	.216E+4	.462E-3	.999537520
40	+2	20	20	.585E-9	.993E+0	.742E-6	.993E+0	.750E-2	.134E+7	.748E-6	.999999252
50	+2	25	25	.825E-12	.986E+0	.105E-8	.986E+0	.138E-1	.936E+9	.107E-8	.999999999
30	+3	15	15	.365E-5	.999E+0	.459E-2	.995E+0	.222E-3	.217E+3	.459E-2	.995409465
40	+3	20	20	.783E-8	.999E+0	.993E-5	.999E+0	.686E-3	.101E+6	.994E-5	.999990062
50	+3	25	25	.139E-10	.998E+0	.177E-7	.998E+0	.160E-2	.564E+8	.177E-7	.999999982
30	+4	15	15	.263E-4	** 4/9E+0	.326E-1	.967E+0	.112E-4	.297E+2	.326E-1	.967414948
40	+4	20	20	.766E-7	4/9E+0	.971E-4	4/9E+0	.491E-4	.103E+5	.971E-4	.999902886
50	+4	25	25	.171E-9	4/9E+0	.219E-6	4/9E+0	.146E-3	.457E+7	.219E-6	.999999781
30	+5	15	15	.147E-3	7/9E+0	.169E+0	.831E+0	.389E-6	.493E+1	.169E+0	.831274841
40	+5	20	20	.584E-6	6/9E+0	.740E-3	.999E+0	.285E-5	.135E+4	.740E-3	.999260092
50	+5	25	25	.165E-8	6/9E+0	.211E-5	5/9E+0	.109E-4	.473E+6	.211E-5	.999997885
40	+6	20	20	.361E-5	7/9E+0	.457E-2	.995E+0	.120E-6	.218E+3	.457E-2	.995425925
50	+6	25	25	.130E-7	6/9E+0	.167E-4	4/9E+0	.660E-6	.600E+5	.167E-4	.999983324
40	+7	20	20	.186E-4	9/9E+0	.233E-1	.977E+0	.136E-8	.419E+2	.233E-1	.976687801
50	+7	25	25	.863E-7	8/9E+0	.110E-3	4/9E+0	.121E-7	.907E+4	.110E-3	.999889699
40	+8	20	20	.817E-4	9/9E+0	.984E-1	.902E+0	.126E-8	.917E+1	.984E-1	.901630507
50	+8	25	25	.489E-6	[illegible]/9E+0	.624E-3	.999E+0	.140E-8	.160E+4	.624E-3	.999375569
40	+9	20	20	.310E-3	9/9E+0	.325E+0	.675E+0	.943E-9	.208E+1	.325E+0	.675034903
50	+9	25	25	.240E-5	9/9E+0	.307E-2	.997E+0	.139E-8	.324E+3	.307E-2	.996927290
50	+10	25	25	.104E-4	9/9E+0	.132E-1	.987E+0	.138E-8	.747E+2	.132E-1	.986784893
50	+11	25	25	.400E-4	9/9E+0	.499E-1	.950E+0	.133E-8	.190E+2	.499E-1	.950113102
50	+12	25	25	.138E-3	9/9E+0	.162E+0	.838E+0	.117E-8	.519E+1	.162E+0	.838349074

$P_e^b = 10^{-2}$

PE = 0.65

Resolution = 512 elements/line

* .821E-6 = .821 X 10^{-6}

** 4/9 = .9999+

TABLE B-5

PERFORMANCE PROBABILITIES
FOR PEC - HUFFMAN TELEVISION SYSTEM

RE-ORDER No. 64-712
Vol I

N = 10

P^j_{sh}		K^j	P^j_{fs}	P^j_{ms}	K^j_{ave}	a	b
0.9000	0	.136E+1	.113E-2	.909E-1	.122E+1	10	10
	+1	.617E+0	.205E-1	.800E-1	.102E+1	10	10
	+2						
	+3						
	+4						
0.9900	0	.274E+1	.126E-2	.876E-2	.122E+1	10	10
	+1						
	+2						
	+3						
0.9990	0						
	+1						
	+2						
0.9999	0						
	+1						

* 8/9 = .99999+

$P_e^b = 10^{-2}$

PE = 0.66

N = 15

P^j_{sh}		K^j	P^j_{fs}	P^j_{ms}	K^j_{ave}	a	b
0.9000	0	.171E+1	.114E-5	* 8/9E-1	.135E+1	15	15
	+1	.694E+0	.279E-4	3/9E-1	.104E+1	15	15
	+2	.404E+0	.416E-3	.796E-1	.100E+1	15	15
	+3	.279E+0	.415E-2	.959E-1	.100E+1	15	15
	+4	.234E+0	.303E-1	.707E-1	.100E+1	15	15
0.9900	0	.342E+1	.125E-5	4/9E-2	.135E+1	15	15
	+1	.139E+1	.307E-4	.997E-2	.104E+1	15	15
	+2	.815E+0	.458E-3	.954E-2	.100E+1	15	15
	+3	.620E+0	.457E-2	.544E-2	.100E+1	15	15
0.9990	0	.513E+1	.126E-5	.999E-3	.135E+1	15	15
	+1	.209E+1	.399E-4	.969E-3	.104E+1	15	15
	+2	.132E+1	.462E-3	.538E-3	.100E+1	15	15
0.9999	0	.685E+1	.126E-5	.987E-4	.135E+1	15	15
	+1	.289E+1	.310E-4	.690E-4	.104E+1	15	15

Resolution = 512 elements/ line

.136E+1 = .136 X 10^1

N=20

P^j_{sh}		K^j	P^j_{fs}	P^j_{ms}	K^j_{ave}	a	b
0.9000	0	.208E+1	.115E-8	*8/9E-1	.149E+1	20	20
	+1	.822E+0	.346E-7	6/9E-1	.106E+1	20	20
	+2	.471E+0	.673E-6	5/9E-1	.101E+1	20	20
	+3	.316E+0	.894E-5	4/9E-1	.100E+1	20	20
	+4	.232E+0	.874E-4	.999E-1	.100E+1	20	20
	+5	.181E+0	.666E-3	.993E-1	.100E+1	20	20
	+6	.147E+0	.414E-2	.959E-1	.100E+1	20	20
	+7	.125E+0	.215E-1	.790E-1	.100E+1	20	20
	+8	.308E+0	.972E-1	.115E-1	.100E+1	20	20
	+9						
	+10						
	+11						
0.9900	0	.417E+1	.126E-8	7/9E-2	.149E+1	20	20
	+1	.164E+1	.300E-7	5/9E-2	.106E+1	20	20
	+2	.941E+0	.740E-6	4/9E-2	.101E+1	20	20
	+3	.632E+0	.984E-5	.999E-2	.100E+1	20	20
	+4	.465E+0	.962E-4	.990E-2	.100E+1	20	20
	+5	.367E+0	.733E-3	.927E-2	.100E+1	20	20
	+6	.327E+0	.455E-2	.547E-2	.100E+1	20	20
	+7						
	+8						
	+9						
0.9990	0	.625E+1	.128E-8	6/9E-3	.149E+1	20	20
	+1	.247E+1	.304E-7	4/9E-3	.106E+1	20	20
	+2	.141E+1	.747E-6	.999E-3	.101E+1	20	20
	+3	.950E+0	.993E-5	.990E-3	.100E+1	20	20
	+4	.707E+0	.970E-4	.903E-3	.100E+1	20	20
	+5	.646E+0	.740E-3	.261E-3	.100E+1	20	20
	+6						
	+7						
	+8						
0.9999	0	.833E+1	.128E-8	5/9E-4	.149E+1	20	20
	+1	.329E+1	.304E-7	3/9E-4	.106E+1	20	20
	+2	.188E+1	.704E-6	.993E-4	.101E+1	20	20
	+3	.128E+1	.994E-5	.901E-4	.100E+1	20	20
	+4	.129E+1	.971E-4	.290E-5	.100E+1	20	20
	+5						
	+6						

*8/9 = .99999999+

$P_e^b = 10^{-2}$

PE = 0.66

N=25

P^j_{sh}		K^j	P^j_{fs}	P^j_{ms}	K^j_{ave}	a	b
0.9000	0	.248E+1	.116E-11	.100E+0	.165E+1	25	25
	+1	.954E+0	.405E-10	.100E+0	.110E+1	25	25
	+2	.538E+0	.962E-9	8/9E-1	.101E+1	25	25
	+3	.350E+0	.160E-7	6/9E-1	.100E+1	25	25
	+4	.261E+0	.197E-6	5/9E-1	.100E+1	25	25
	+5	.201E+0	.190E-5	4/9E-1	.100E+1	25	25
	+6	.162E+0	.158E-4	3/9E-1	.100E+1	25	25
	+7	.126E+0	.993E-4	.999E-1	.100E+1	25	25
	+8	.113E+0	.562E-3	.994E-1	.100E+1	25	25
	+9	.114E+0	.277E-2	.972E-1	.100E+1	25	25
	+10	.119E+0	.121E-1	.881E-1	.100E+1	25	25
	+11	.144E+0	.471E-1	.551E-1	.100E+1	25	25
0.9900	0	.496E+1	.128E-11	.100E-1	.165E+1	25	25
	+1	.191E+1	.445E-10	.100E-1	.110E+1	25	25
	+2	.108E+1	.106E-8	7/9E-2	.101E+1	25	25
	+3	.715E+0	.176E-7	5/9E-2	.100E+1	25	25
	+4	.521E+0	.217E-6	4/9E-2	.100E+1	25	25
	+5	.403E+0	.209E-5	3/9E-2	.100E+1	25	25
	+6	.324E+0	.165E-4	.998E-2	.100E+1	25	25
	+7	.253E+0	.109E-3	.989E-2	.100E+1	25	25
	+8	.229E+0	.619E-3	.938E-2	.100E+1	25	25
	+9	.244E+0	.305E-2	.696E-2	.100E+1	25	25
0.9990	0	.744E+1	.129E-11	.100E-2	.165E+1	25	25
	+1	.286E+1	.449E-10	.100E-2	.110E+1	25	25
	+2	.161E+1	.107E-8	6/9E-3	.101E+1	25	25
	+3	.107E+1	.177E-7	4/9E-3	.100E+1	25	25
	+4	.782E+0	.219E-6	.999E-3	.100E+1	25	25
	+5	.605E+0	.211E-5	.998E-3	.100E+1	25	25
	+6	.487E+0	.167E-4	.983E-3	.100E+1	25	25
	+7	.385E+0	.110E-3	.890E-3	.100E+1	25	25
	+8	.387E+0	.624E-3	.376E-3	.100E+1	25	25
0.9999	0	.992E+1	.129E-11	.100E-3	.165E+1	25	25
	+1	.382E+1	.450E-10	.100E-3	.110E+1	25	25
	+2	.215E+1	.107E-8	5/9E-4	.101E+1	25	25
	+3	.143E+1	.177E-7	3/9E-4	.101E+1	25	25
	+4	.104E+1	.219E-6	.998E-4	.100E+1	25	25
	+5	.808E+0	.211E-5	.979E-4	.100E+1	25	25
	+6	.660E+0	.167E-4	.833E-4	.100E+1	25	25

Resolution = 512 elements/ line

.200E+1 = .200 X 10^1

TABLE B-6

SEARCH MODE PERFORMANCE OF LINE SYNC FOR PEC-HUFFMAN TELEVISION SYSTEM

176

RE-ORDER No. 64-712
Vol I

TABLE B-7

CHECK MODE, 40 BIT SYNC PATTERN

e_c	P'_{tsc}	P'_{fsc}
1	0.938	1.10×10^{-18}
2	0.992	4.39×10^{-16}
3	0.997	7.76×10^{-14}
4	0.999	7.43×10^{-12}
5	0.999+	4.32×10^{-10}

The check mode specifications are satisfied for $e_c = 2, 3, 4, 5$ and probably for 6 and 7 errors if the if the data is extrapolated; $e_c = 3$ is chosen since P'_{fsc} is approaching the limit of acceptability for the Check mode. Therefore, the Check mode parameters are:

$$j = 1$$
$$e_c = 3.$$

Lock Mode Parameters

For this system the probability P'_t is calculated from the equation

$$P'_{t\ell} = P'_c{}^{(512)} \tag{46}$$

if the criterion for return to Search mode is a single correlation failure. Even for three errors allowed in the correlator which yields $P'_c(e) = 0.999$, according to Table B-5, $P'_{t\ell}$ is approximately 0.6, whereas increasing the allowable errors merely increase $P'_{t\ell}$ since $P_c(e)$ steadily increases. For the case where the criterion to return to Search mode is two consecutive correlation failures the expression for $P'_{t\ell}$ becomes

$$P'_{t\ell} = \left[1 - (1 - P'_c)^2\right]^{512}. \tag{47}$$

For $P'_c = 0.999$, $P'_{t\ell} = 0.9995$, whereas for $y = 2$, $P'_{f\ell y} = 2 P'_d$. If nine errors are allowed in the correlator, $P'_{f\ell y} = 6.2 \times 10^{-4}$ which is well within the desired specifications. Therefore, the Lock mode parameters are e = 9 errors and return to Search after two consecutive failures.

177

RE-ORDER No. 64-712
Vol I

B. 3.7 Summary of Subsystem Performance

The horizontal synchronization subsystem for the PEC-Huffman single frame television system has been theoretically designed from the design tables for a 512 x 512 line resolution with an average PE = 0.65 for four-bit intensity quantization.

With a bit error probability of 10^{-2}, the odds are 1000:1 that the true synchronization pattern is found within one line. The odds are 10, 300:1 against the possibility of finuing a false sync pattern in the first line. By adding one line of Check mode operation, we have a total acquisiton time of two lines with a 0.998 probability of true sync as opposed to a 7.76×10^{-14} probability of false sync. The Lock mode results are identical to those of the PCM analysis and ensures that the Lock mode is established with true synchronization. The 40 bit sync word code recommended for the 10^{-2} bit error probability for the PEC-Huffman system with a PE equaling 0.65 amounts to an increase in an average of 3.3% in the total bits transmitted. This is an extremely small price to pay for the synchronization performance achieved.

RE-ORDER No. 64-712
Vol I

APPENDIX C

FIDELITY

C.1 ANALYTICAL FIDELITY

C.1.1 Concept

An analytical fidelity measure is some defined function of the difference between the output and input signals. However, in order to decide which function to use and the input signal set, the ultimate use of the system must be specified. In general, within this study, we are dealing with a reconnaissance system whose use is to provide pictorial information. The pictorial information, being of an exploratory nature, is largely unknown. This requires that features which are common to all pictures be identified. Then these pictorial features are translated to system input signals. Also, criteria for the goodness of response to these signals must be identified and means for their measurement chosen. In the course of this study it was not possible to go into the actual implementation of analytical fidelity criteria. However, for illustrative purposes a block diagram design is given.

C.1.2 Relation of Pictorial Factors to TV Parameters

In an analysis of the factors governing photographic image quality[1] there are at least three current views. The first view holds that the factors controlling image quality are angular field, definition, distortion, character of emulsion, altitude, ground speed, vibration, and character of illumination.

The second view holds that the amount of information in a photograph depends on four image properties: graininess, sharpness, resolving power, and tone reproduction.

Finally, the third view reduces the problem to three primary characteristics governing the photo image quality: the tone or color contrast between the object and its background; image sharpness characteristics, and stereoscopic parallax characteristics.

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[1] Colwel, Robert N., Editor, Manual of Photographic Interpretation, American Society of Photogrammetry, Washington, D. C., 1960.

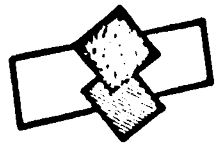
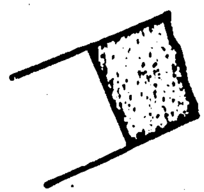
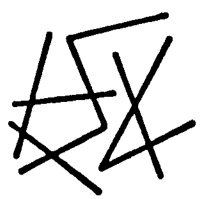
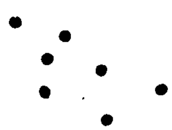
DETECTION OF	IMPLIES	PICTORIAL FACTOR	WHICH REQUIRES	SYSTEM FIDELITY CRITERION
OBJECTS		TONE CONTRAST		LINEARITY, QUANTIZATION LEVELS, NUMBER AND SPACING
EDGES OF OBJECTS AND OTHER TONAL BOUNDARIES		SHARPNESS		STEP FUNCTION RESPONSE
LONG, THIN LINES		SHARPNESS		IMPULSE RESPONSE AND DIGITAL RESOLUTION (1 DIMENSION)
ISOLATED POINTS		SHARPNESS		IMPULSE RESPONSE AND DIGITAL RESOLUTION (2 DIMENSIONS)
COLLECTION OF POINTS AND/OR BLOBS AT OR NEAR TERMINAL RESOLUTION		TEXTURE CONTRAST		MID-RANGE SINE WAVE RESPONSE

TABLE C.1.2-1

RELATION OF PICTORIAL FACTORS TO FIDELITY CRITERIA

07314

RE-ORDER No. 64-712
Vol I

The first view is "system oriented" in that it includes vehicle characteristics. While these are of ultimate importance, it is only the sensor oriented characteristics, typified by the second view, that interest us in this study. In the third view these are condensed essentially to two sensor-oriented characteristics (stereoscopic parallax being a trajectory and timing problem).

The problem of good television images is certainly analogous to the problem of good photographic images. The television system must render tones accurately and it must sharply render the boundaries between these tones. In television terms this means that gray scale rendition and good step function response are essential.

There exist, however, three additional types of images and their electrical equivalents. In observing aerial photographs of predominantly geological areas, long, thin, winding lines at virtually the limit of resolution are often encountered. These lines, when scanned, will give rise to impulses within the electro-optical system. If these lines are to be recorded, the impulse response of the system must give rise to a pulse which will exceed the next higher threshold level for a sufficient length of time to allow at least one digital sample to occur. The detection of thin, low-contrast lines (roads, railroad beds, fault lines, lunar rays, etc.) implies a requirement for good impulse response in the overall system in addition to good step function response.

In other parts of photographs of predominantly geological interest, individual dots appear which are presumably trees or other isolated vegetation. These optical point images will give rise to small spread functions which, when scanned, will also result in impulses into the digital portion of the system. The digital system resolution (in both dimensions) must be set such that these spread functions will have a reasonable chance of being scanned near their central peak and the analog portion of the system will again be required to have a good impulse response. Table C.1.2-1 summarizes the relation of pictorial factors and fidelity criteria.

RE-ORDER No. 64-712
Vol I

In viewing geological pictues there often occur adjacent sections of closely packed trees which appear to have approximately equal average tone. However, there is a distinct demarcation between the sections of forest on the basis of texture. For instance, large, old-growth timber closely packed differs in texture from even similar species of new-growth timber of medium density. Yet the average tone for both areas is quite close. A similar effect can occur over farm crop lands. Even though individual trees and crops are below the system resolution limit, it appears that an electro-optical system of good mid-range sine-wave response is required for adequate texture distinction.

Suppose we have two texture target images whose spectral response is H(f) and G(f), with H(0) = G(0), (i.e., equal average tone - dc component). If the response of our eye is essentially unity over the range 0 - f_o where $H(f_o) = G(f_o) = 0$, then the eye's perception of textural difference can, in some sense, be represented by

$$\Delta t(f) = \left| \, |H(f)|^2 - |G(f)|^2 \right| \tag{1}$$

Now, let each texture image, H and G, be rendered by an idealized, linear electro-optical system of response R(f). Then,

$$H'(f) = R(f)\,H(f), \text{ and} \tag{2}$$

$$G'(f) = R(f)\,G(f). \tag{3}$$

Now, the texture difference of the two rendered images is

$$\Delta' t(f) \left| \, |R(f)\,H(f)|^2 - |R(f)\,G(f)|^2 \right| . \tag{4}$$

Under the condition that

$$|R(f)|^2 \, |H(f)|^2 = |R(f)\,H(f)|^2 , \tag{5}$$

182

RE-ORDER No. 64-712

Vol I

equation (4) reduces to

$$\Delta' t(f) = \left|R(f)\right|^2 \left| \left|H(f)\right|^2 - \left|G(f)\right|^2 \right| . \tag{6}$$

If we have two systems $R_1(f)$ and $R_2(f)$ with equal limiting resolutions ($R_1(0) = R_2(0) = 1$ and $R_1(f_0) = R_2(f_0) = 0$) then we can define a measure of average differential texture as

$$\bar{\Delta}'_i t = \frac{1}{f_0} \int_0^{f_0} \left| R_i(f) \right|^2 \left| \left| H(f)\right|^2 - \left|G(f)\right|^2 \right| df \quad i = 1, 2, \tag{7}$$

for each system, R_1 and R_2.

If, for each midrange value of f,

$$\left| R_2(f) \right|^2 > \left| R_1(f) \right|^2 \quad f \neq 0 \neq f_0, \text{ then} \tag{8}$$

$$\bar{\Delta}'_2 t > \bar{\Delta}'_1 t,$$

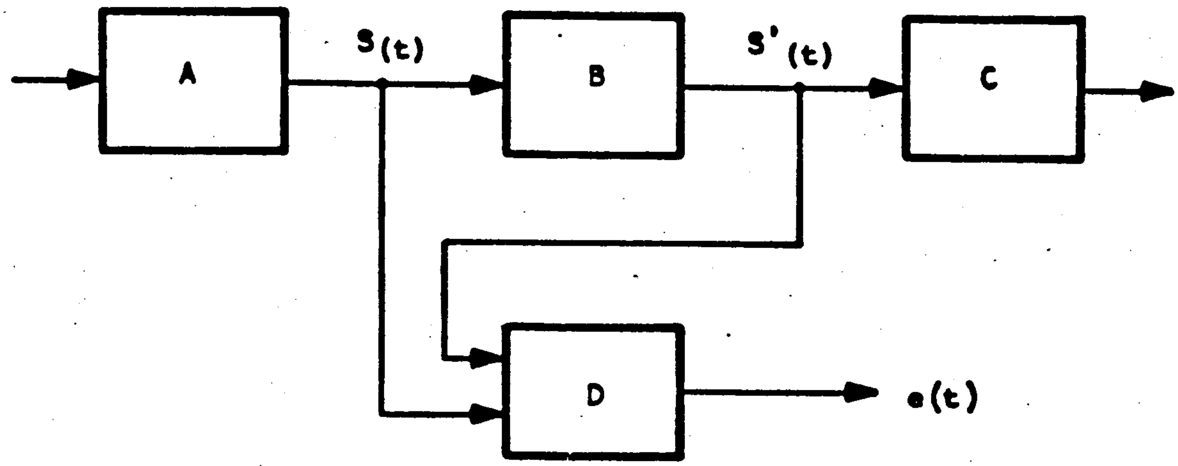
showing that the measure of average differential texture is higher with that system which has everywhere better midrange response.

C.1.3 Criteria

In the previous section it has been shown that, without specifying the specific pictorial content, we can at least list four system fidelity criteria: linearity, step function response, impulse response and midrange sine-wave response. The next step is to put these and other possibly useful criteria into analytical language.

RE-ORDER No. 64-712
Vol I

In the diagram below,



a signal S(t) enters block B and a signal S'(t) leaves block B. S'(t) is a function of S(t) and the parameters $b_1, b_2 \ldots b_i$ of block B. Thus,

$$S'(t) = B(S, b_1, b_2 \ldots b_i). \tag{9}$$

Block D generates an error signal, e(t), which in general is given by

$$e(t) = D(S, S'), \tag{10}$$

$$e(t) = D\left[S, B(S, b_1, b_2 \ldots b_i)\right]. \tag{11}$$

In principle, e(t) may be any function of S and S', but in practice, an error implies some deviation between what is desired (input signal) and what is obtained (output signal). The simplest function is simply the difference between S and S'. Since it occurs so frequently in error measures, it will be given the special symbol, Δ.

$$e(t) = \Delta(t) = S'(t) - S(t). \tag{12}$$

184

RE-ORDER No. 64-712
Vol I

Notice that error functions are functions of <u>both</u> the input signal S(t) and the parameters $b_1, b_2 \ldots b_i$ of box B. In principle, the usefulness of an error measure, E, is such that, over a suitable ensemble of S, the ensemble average, $\overline{E}$ of E, can be made usefully small by adjusting the B parameters, $b_1, b_2 \ldots b_i$.

Many estimates of the fidelity of B can be formulated as functions of Δ. Before proceeding to this step, a restriction which is pertinent to this study will be placed on S(t). While S(t) may not be periodic, it is possible to divide S(t) into meaningful segments called frames. That is, any given S(t) under discussion is understood to be that S(t) lying between two distinct limits.

$$S_n(t) = S(t) \text{ for} \tag{13}$$

$$n\tau \leq t < (n+1)\tau \quad n = 0, 1, 2 \ldots \infty ,$$

where τ is a frame time (order of 20 sec.).

Now measures of error can be defined that, while being a single number, reflect the effect of many instantaneous errors occurring throughout one frame time, τ.

Some possible error functions, e(t), are given below along with some error measures, E.

$$e(t) = \Delta(t) \tag{14}$$

$$e(t) = \left| \Delta(t) \right| \tag{15}$$

$$E_n = \widetilde{\Delta(t)} = \frac{1}{\tau} \int_{n\tau}^{(n+1)\tau} \Delta(t)\, dt \tag{16}$$

RE-ORDER No. 64-712
Vol I

$$E_n = \sqrt{\left[\widetilde{\Delta(t)}\right]^2} = \sqrt{\frac{1}{\tau}\int_{n\tau}^{(n+1)\tau}\left[\Delta(t)\right]^2 dt} \tag{17}$$

$$E_n = \sqrt{\widetilde{\left[\widetilde{\left[\Delta(t)\right]} - \left[\Delta(t)\right]\right]}^2} \tag{18}$$

In equations (14)-(18) $\sim\!\sim$ indicates a time average. The error terms given by equations (14)-(18) may be labelled, respectively,

(a) Instantaneous error
(b) Instantaneous absolute error
(c) Mean error (per TV frame)
(d) Root mean square (rms) error (per TV frame)
(e) Standard deviation of error (per TV frame)

In addition to these functions, we have three others of a "nonlinear" but potentially useful nature.

$$E_n = \hat{\Delta}(t)\Big|_{t = n\tau}^{t = (n+1)\tau} \tag{19}$$

Peak error within TV frame, n.

$$E_n = \left|\widehat{\Delta(t)}\right|\Bigg|_{t = n\tau}^{(n+1)\tau} \tag{20}$$

RE-ORDER No. 64-712
Vol I

Peak absolute error within TV frame, n.

$$E_n = P(\Delta_i \leq \Delta < \Delta_i + d\Delta) \Bigg|_{t = n\tau}^{t = (n+1)\tau} \tag{21}$$

Probability distribution of error.

Note that mean error per frame, rms error per frame, peak error per frame. and peak absolute error per frame can all be derived from (21). The probability density of error can be measured experimentally by EDITS.

The above error functions are, for perfectly general inputs, S(t). However, in many cases systems are required to perform quite well for very special types of input signals, $S_i(t)$, $i = 1, 2, \ldots \infty$. Suppose a typical TV picture line could be regarded as a superposition of special signals, $S_i(t)$, such that

$$S(t) = \sum_{i=1}^{n} S_i(t). \tag{22}$$

One set of these special signals might include a dc level,

$$S_1(t) = c, \tag{23}$$

a sine wave function,

$$S_2(t) = a \operatorname{Cos} (\omega t + \phi), \tag{24}$$

a random function,

$$S_3(t) = r(t), \text{ such that} \tag{25}$$

$$R(\omega) = \frac{1}{2\pi} \int_{-\infty}^{\infty} r(t)\, e^{-j\omega t}\, dt, \text{ and}$$

177

RE-ORDER No. 64-712
Vol I

$$\int_{-\infty}^{\infty} \left| R(\omega) \right|^2 d\omega = \int_{-\infty}^{\infty} \left| r(t) \right|^2 dt < \infty,$$

an impulse function,

$$S_4(t) = b\,\delta\,(t - t_o). \tag{26}$$

a step function,

$$S_5(t) = d \int_{-\infty}^{t} \delta\,(t - t_o)\, dt = d\, U(t - t_o). \tag{27}$$

a ramp function,

$$S_6\,(t) = g \int_{-\infty}^{t} U(t - t_o)\; dt = g \int_{-\infty}^{t} \int_{-\infty}^{t} \delta\,(\tau - t_o)\, d\tau\; dt \tag{28}$$

and a pulse function,

$$S_7(t) = h \left[U(t - t_o) - U(t - t_o - \epsilon\,) \right] \tag{29}$$

where ϵ is the pulse width.

For each of these special functions it is possible to devise special measures of fidelity which may not correspond to the more conventional measures (equations (14)-(21)) of error. For the step function, one measure of "fidelity" of the system is rise-time, i.e., the time it takes the output signal $S'(t)$ to go from 10% to 90% of the final value. This measure is hard to define mathematically, but it is easy to measure given a clean, amplitude-time display.

RE-ORDER No. 64-712
Vol I

Another fidelity measure for the step function response is percent overshoot, also difficult to specify mathematically but easy to measure in practice. In TV this measure would amount to a measure of "edge enhancement."

For the sine wave input, there is, of course, the familiar amplitude and phase response criteria. Analogously, for the random input, there exists the power spectral response. If box B is a linear system with impulse response $B(\tau)$ then its spectral density response is given by

$$\left| B(\omega) \right|^2 = \frac{1}{4\pi^2} \left| \int_{-\infty}^{\infty} B(\tau)\, e^{-j\omega\tau}\, d\tau \right|^2 , \tag{30}$$

and its output spectral density,

$\left| S'(\omega) \right|^2$, is given by

$$\left| S'(\omega) \right|^2 = \left| R(\omega) \right|^2 \left| B(\omega) \right|^2 \tag{31}$$

for an input random function $r(t)$ with input spectral density $\left| R(\omega) \right|^2$.

One conceivable measure of fidelity in this case could be the relative power efficiency, η , given by

$$\eta = \frac{\int_{-\infty}^{\infty} \left| S'(\omega) \right|^2 d\omega}{\int_{-\infty}^{\infty} \left| R(\omega) \right|^2 d\omega} , \tag{32}$$

with $0 \leq \eta \leq 1$.

172

RE-ORDER No. 64-712
Vol I

For the pulse input, where the pulse width ϵ is small compared to the predominating time constant in $B(\tau)$, the B system will give some small output. The maximum value of this pulse output relative to the value h, for example, would be an index of the system's fidelity in reproducing the narrow pulse's peak amplitude. Also, the maximum value of this pulse output relative to one-half the peak output signal, $(\widehat{S'(t)}/2)$, would be a measure of the system's ability to deliver "detectable", though by no means faithful, representations of the presence of isolated, subresolution pulses in the input signal S(t). A similar criterion, the maximum value of S'(t) relative to some standard, can be applied with regard to the system's response to an impulse function.

In the servomechanism field, the ramp function is sometimes used as a special input function and the steady-state delay in the output following the input is taken as a measure of system quality. The visual effect of this phenomena would be a slight shift in the direction of scanning of those areas whose luminosity rose linearly with distance. This effect may not be relatively serious since all video features will experience some phase lag.

The dc level input at first glance may appear to be a trivial special test signal. Yet, there exist video processing systems which render a dc level by "hunting" around the input dc level. Also, even a straight PCM system which employs an analog-to-digital converter will have regions of uncertainty about each A-to-D decision level due to noise in the decision amplifier. This will cause a dc signal near one of the decision levels to be rendered, on a probabilistic basis, as a signal of either the correct quantizing level or as a signal of the adjacent quantizing level. In this case the fidelity criteria could be the percentage of time a dc level is rendered correctly by the system. For the "hunting" systems maximum error excursion (equation (20)) could be used.

In summary, there are at least sixteen fidelity criteria that can be analytically expressed. Eight of these can be expressed easily in conventional mathematical terms, and eight others can be evaluated numerically, although their most efficient specification is of an empirical or "verbal" nature. There are at least

RE-ORDER No. 64-712
Vol I

seven special input test signals which can be specified, and associated with these special test signals are the majority of the latter eight fidelity criteria.

C.1.4 Implementation

Another approach to analytical fidelity is to consider the compression system as producing an output waveform which approximates the input waveform. The various measures by which the degree of approximation may be quantitatively expressed are then built into auxiliary equipment which monitors the real time performance of the "approximating" system. The function and block diagram of one example of an analytic fidelity measuring system is given in this section. It is devised within the context of EDITS, although its applicability is by no means so limited.

In numerical analysis[(1)] where it is required that one function approximate another, the customary representation is by a rational function, in particular by a polynomial P_n. Likewise, the customary measures of approximation are

$$\max_{x \in I} \left| f(x) - P_n(x) \right| = \left\| f - P_n \right\|_{\infty}, \text{ and} \tag{33}$$

$$\left\{ \int_I \left| f(x) - P_n(x) \right|^2 dx \right\}^{1/2} = \left\| f - P_n \right\|_2 \tag{34}$$

where $f(x)$ is the function to be approximated over the interval, I. Equation (33) is referred to as the maximum, or the "maximum norm," and is ordinarily preferable in order to assert that nowhere within the interval I does $\left| f(x) - P_n(x) \right|$ exceed a certain ϵ. The form of equation (34) comes from the definition of the norm of a matrix as being the square root of the sum of the

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(1) Forsythe, G. E., and Rosenbloom, P. C., Numerical Analysis and Partial Differential Equations, John Wiley and Sons, Inc., 1958, pp. 18-21

191

RE-ORDER No. 64-712
Vol I

squares of the absolute values of the elements, where the values $f(x_i) - P_n(x_i)$ are considered to constitute the elements of the matrix.

The above discussion applies directly to the JPL Video Data Modulation Study wherein the approximating polynomial function $P_n(x)$ is replaced by some modulation compression system function $C_n\left[f(x)\right]$. For each input signal $f(x)$ the modulation system delivers an output $f_o(x)$ given by

$$f_o(x) = C_n\left[f(x)\right], \tag{35}$$

and it is desired to say

$$\max_{x \in I} \left| f(x) - f_o(x) \right| = \epsilon$$

for some interval, I, which may be taken as one image frame.

In order to give a relative assessment with respect to maximum norm among the compression systems studied for JPL, it would be useful though not absolutely necessary to add to EDITS an Objective Fidelity Taker.[1] As near as practicable, this facility should have the inputs, outputs and functional design as given in Figure C. 1. 4-1. The functions of the boxes labelled A, B, C, etc. are given in Table C. 1. 4-1. A facility of this nature will enable us to give experimental values for equations (33) and (34) for PCM and all the compression systems actually fully implemented on EDITS. It will enable us to do this for a wide variety of picture inputs. In addition, it will enable us to form the first order statistics of the modulation system error from which the range, the mean, and higher moments can be obtained numerically, if desired. In view of the many functions included in Figure C. 1. 4-1, these functions have the priority as given in Table C. 1. 4-1 where $\Delta = f_o(x) - f_i(x)$.

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(1) Time and scope of study did not permit this.

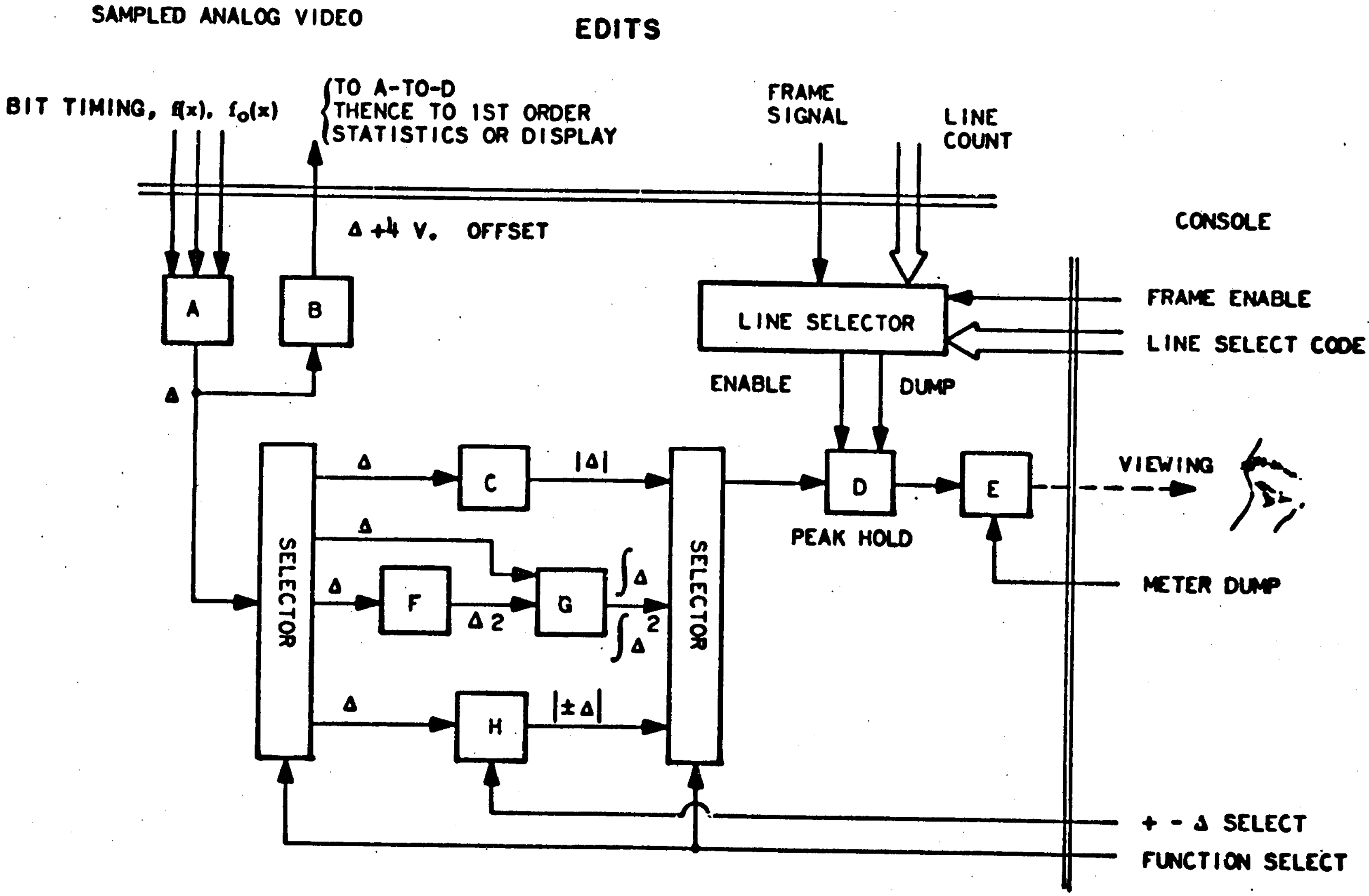

FIGURE C.1.4-1

ANALYTICAL FIDELITY TAKER (OBJECTIVE FIDELITY DEVICE)

07314

RE-ORDER No. 64-712
Vol I

RE-ORDER No. 64-712
Vol I

TABLE C.1.4-1

SYSTEM CONSTRUCTION PRIORITIES

Priority	Mathematical Function	Function
A	Δ	Difference signal
B	$\Delta + 4\,v$	Offset and 1st order statistics driver
C	$\lvert\Delta\rvert$	Absolute value of difference
D	$\underset{x \subset I}{\text{Max}} \lvert\Delta\rvert$	Peak hold (and associated frame selector D_1; line selector D_2)
E	$\underset{x \in I}{k\ \text{mx}} \lvert\Delta\rvert$	Meter output
F	Δ^2	Squaring
G	$\int \Delta^2$	Integrator
H	$\Delta > 0,\ \Delta < 0$	Positive and negative differences

This system would have the overall measurement modes given in Table C.1.4-2. Most of these error measurement modes are now being included in the ASI-210 digital computer code for simulating the linear approximation compression system.

TABLE C.1.4-2

MEASUREMENT MODES FOR OBJECTIVE FIDELITY DEVICE

System Modes	Measurement
AB	Δ display, 1st order statistics on Δ over a frame.
ACDE	Peak error $\lvert\Delta_{max}\rvert$ over either a single, selectable line or over a whole frame.
AFGDE	Mean square error $\int \Delta^2$ over either a line or frame
AGDE	Mean error $\int \Delta$ over a line or frame.
AHDE	Peak positive or peak negative error $\lvert +\Delta_{max}\rvert$ or $\lvert -\Delta_{max}\rvert$ over a line or frame.

194

RE-ORDER No. 64-712
Vol I

C.2. MEASURED FIDELITY

C.2.1 General

At the outset of this study it was recogn'zed that it would be very desirable to establish those fidelity criteria against which the performance of each video data modulation technique could be measured on an absolute basis. However, it was decided during the early stage of this program[1] that, due to the many unknowns presently encountered in defining planetary reconnaissance mission requirements, determination of such absolute criteria was not within the scope of this study.

As an alternative, we have elected to use PCM television performance as the the standard against which other video modulation techniques may be compared. Six-bit PCM serves as the primary reference and four, three and two-bit PCM serve as secondary references in judging certain equipment performance characteristics.

In the general case, therefore, fidelity assessment is a two-step process wherein the performance of the reference PCM technique is initially evaluated relative to the original input test data characteristics. Then the fidelity of the succeeding video modulation technique is compared with the performance obtained using the reference PCM method.

Four modulation methods in addition to the four straight PCM techniques are included in the fidelity measurement program. These are:

- 2-Bit PCM with 3-Bit Roberts
- 3-Bit PCM with 4-Bit Roberts
- 2-Bit Delta
- 4-Bit Previous Element Coding (PEC) with Huffman efficient encoding.

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[1] Based to a large extent on the discussions at the joint JPL-EMR technical meeting held at JPL on October 25, 1963.

RE-ORDER No. 64-712
Vol I

C.2.2 Purposes of the Measured Fidelity Study

The principal purposes of this study are defined as follows:

(1) To determine the types and extent of the measurements which are required to compare the performance of the several video data modulation techniques.

(2) To establish the test controls and procedures required to carry out the defined measurement program.

(Note: These measurements were performed on the EDITS equipment, which incorporates each of the above eight modulation modes.)

C.2.3 Classification of Measurements

The selected measurements are divided into three general categories; specifically, objective, quasi-objective, and subjective. Table C.2-1 summarizes these categories. Further definition is given below, together with a listing of specific measurements performed in each area. The discussion of actual measurement procedures employed will be found in Appendix E of this report.

<u>Objective Measurements</u>

<u>Definition.</u> The measurements included in this category make use of calibrated input data in the form of known test object spatial frequencies, measured (or known) photometric characteristics, and calibrated input video signal-to-noise (S/N) conditions. The output from the video data modulation system under test is objectively measured by using test instruments (CRT and photometric densitometer) in which first-order errors are established by the basic instrument design, and human operator errors produce only second-order effects. Since such measurements are effectively divorced from operator interpretive errors, they are categorized as objective types.

<u>List of Measurements.</u> Objective measurements include the following:

- Sine wave response (using 30% modulation Kodak Sine Wave Test Object shown in Figure C.2 P-1)
- Square wave response (using Westinghouse Resolution Chart ET-1332A shown in Figure C.2 P-2).

TABLE C. 2-1

CLASSIFICATION OF FIDELITY MEASUREMENTS

Category	List of Measurements	Data Characteristics		Test Conditions
		Input	Output	
1. Objective	1. Sine wave response 2. Square wave response 3. Gamma characteristic	Calibrated	Objectively measured	40 db S/N (Approx.)
2. Quasi-Objective	1. Overall image fidelity 2. Sine wave lim. resolution 3. Square wave lim. resolution 4. Brightness contouring	Calibrated	Subjectively measured	Includes 20, 30, and 40 db S/N
3. Subjective	Overall image fidelity 1. Canadian Arctic aerial photo 2. JPL Lunar photo	Partially calibrated	Subjectively measured	Combinations of 20, 30, and 40 db S/N and $10^{-\infty}$, 10^{-4}, 10^{-3}, and 10^{-2} P_e.

RE-ORDER No. 64-712
Vol I

RE-ORDER No. 64-712
Vol I

- Gamma characteristic (using Foto-Video Grey Scale Chart F-63C #668 shown in Figure C.2 P-3).

(Note: These tests are performed at approximately 40 db video S/N $\left(\frac{\text{Peak-to-peak video}}{\text{rms noise}}\right)$.)

<u>Quasi-Objective Measurements</u>

<u>Definition.</u> As in the case of the objective measurements, similar calibrated input data also are employed in this test series. This includes several calibrated levels of input video S/N. However, output measurements, which are performed on photographic reproductions of the TV display images, are primarily of a qualitative nature. This is true because the human eye is used here as the measuring device; hence, the element of subjectivity is present in the output measurements.

<u>List of Measurements.</u> Quasi-objective measurements include the following types:

- Overall image fidelity (using RETMA test chart input shown Figure C.2 P-4).
- Limiting resolution (sine wave input using Kodak Sine Wave Test Object).
- Limiting resolution (square wave input using Westinghouse Resolution Chart).
- Brightness contouring (hot-spot incandescent light source used as the input).

(Note: These tests include three video S/N conditions; 20, 30 and 40 db.)

<u>Subjective Measurements</u>

<u>Definition.</u> The basic input data used in this test series is derived from two photographs. The first is a photograph of a telescopic image of the lunar surface (one of a group of such photographs furnished by JPL, see Figure C.2 P-5) and the second is a photograph of an aerial camera image of Canadian Arctic terrain (see Figure C.2 P-6). These particular photographs were chosen

RE-ORDER No. 64-712
Vol I

because they represent wide extremes in terms of terrain types, scale factors and photometric contrasts. Figures C.2 P-7 and C.2 P-8 include grid overlays which permit easy and accurate coordinate location of terrain areas of interest.

In comparison with the input data employed in the previously discussed test areas, these photographic input data are only partially calibrated. To achieve total calibration it would be necessary to completely scan the complex terrain image patterns with a microdensitometer to obtain precision spatial and photometric image characteristics. For purposes of this measurement program, it was deemed sufficient to take relative reflection density readings of larger areas with a simple Kodak densitometer and then compute corresponding relative reflectances from which contrast ratios of selected areas could be determined. Relative reflectance values are presented in Figures C.2 P-9 and C.2 P-10. Knowledge of basic photo scale factors has permitted calculation of simulated (terrain-related) scale factors and TV camera terrain resolutions used in the laboratory measurement program. Table C.2-2 presents a summary of this data. Also, as in the above tests, several calibrated levels of video S/N are included. In addition, in order to assess the effects of bit errors in the digital video data, several calibrated bit error probability (P_e) conditions are employed in this subjective test series.

TABLE C.2-2

PHOTOGRAPHIC TEST PARAMETERS

Photo Type	(1) Basic Photo Scale Factor	(2) Simulated Photo Scale Factor	(3) Simulated TV Ground Resolution	(4) Max. Photo Contrast
Lunar	1:10,000,000 (Max)	1:160,000,000 (Max)	7 Km/TV line (Max)	11 to 1
Canadian Arctic A-15420-43	1:60,000	1:960,000	0.043 KM/ TV line	50 to 1

RE-ORDER No. 64-712
Vol I

Notes: (1) Defined as the ratio of photograph unit dimension to corresponding terrain dimension.

(2) 7" x 7" photo area focused on 0.44" x 0.44" vidicon photocathode area provides a scale factor reduction of 16 to 1.

(3) Assumes 256 line TV resolution.

(4) Represents highest photographic contrast where the average contrast of adjacent areas is lower.

$$\text{Contrast} = \frac{\text{Highlight Reflectance}}{\text{Lowlight Reflectance}}$$

The fidelity of the video output data may be subjectively assessed by viewing photographic reproductions of the TV display images[1] and comparing this image quality with photographic reproductions of reference PCM video output data.

List of Measurements. Subjective measurements include the following:

- Overall image fidelity -- Canadian Arctic photo input
- Overall image fidelity -- lunar photo input.

(Note: These measurements were performed with combinations of video S/N of 40 db, 30 db and 20 db, and bit error probabilities (P_e) of $10^{-\infty}$, 10^{-4}, 10^{-3} and 10^{-2}.)

C.2.4 Schedule of Fidelity Measurements

Table C.2-3 presents a schedule of measurements recommended for the purpose of comparing the fidelities of the various video modulation techniques. This table lists objective, quasi-objective, and subjective measurements together with the parametric combinations of video S/N and bit error probabilities (P_e) applicable to each modulation technique under test. The actual fidelity measurement program discussed in Appendix E is set up on the basis of this schedule.

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[1] Direct assessment of the display image quality is not feasible due to the short decay time of the display phosphor relative to the EDITS frame scan time.

RE-ORDER No. 64-712
Vol I

Measurement Type[4]	EDITS Analog and Digital Noise Condition[2]	6 Bit PCM	4 Bit PCM	3 Bit PCM	2 Bit PCM	3/4 Roberts	2/3 Roberts	2 Bit Delta	4 Bit PEC/ Huffman
		Systems Under Measurement							
Objective Measurements									
1. Sine Wave Response	40 db[3]	X[1]						X	
2. Square Wave Response	40 db[3]	X						X	
3. Gamma a. Camera	40 db[3]	X							
b. System		X							
Quasi-Objective Measurements									
1. Overall Image Fidelity (RETMA Chart)	S/N = 40 db	X	X	X	X	X	X	X	X
	S/N = 20 db	X	X	X	X	X	X	X	X
2. Limiting Resolution (Sine Wave Chart)	S/N = 40 db[3]	X						X	X
3. Limiting Resolution (Square Wave Chart)	S/N = 40 db[3]	X	X	X	X	X	X	X	X
4. Brightness Contouring (Hot Spot Light Source)	S/N = 40 db	X	X	X	X	X	X	X	X
	S/N = 30 db	X	X	X	X	X	X	X	X
	S/N = 20 db	X	X	X	X				
Subjective Measurements									
1. Image Fidelity Canadian Arctic	S/N = 40 db	X	X	X	X	X	X	X	X
Lunar	$P_e = 10^{-\infty}$	X	X	X	X	X	X	X	X
2. Image Fidelity Canadian Arctic	S/N = 30 db	X	X	X	X	X	X	X	X
Lunar	$P_e = 10^{-\infty}$	X	X	X	X	X	X	X	X
3. Image Fidelity Canadian Arctic	S/N = 20 db	X	X	X	X	X	X	X	X
Lunar	$P_e = 10^{-\infty}$	X	X	X	X	X	X	X	X
4. Image Fidelity Canadian Arctic	S/N = 40 db	X	X	X	X	X	X	X	X
Lunar	$P_e = 10^{-4}$	X	X	X	X	X	X	X	X
5. Image Fidelity Canadian Arctic	S/N = 40 db	X	X	X	X	X	X	X	X
Lunar	$P_e = 10^{-3}$	X	X	X	X	X	X	X	X
6. Image Fidelity Canadian Arctic	S/N = 40 db	X	X	X	X	X	X	X	X
Lunar	$P_e = 10^{-2}$	X	X	X	X	X	X	X	X

Notes:
(1) X denotes scheduled measurement.
(2) When P_e condition is not specified, it may be assumed to be a zero bit error condition ($P_e = 10^{-\infty}$).
(3) 40 db is the nominal value for these tests, although 42 db is measured in certain cases.
(4) All measurements are taken with scan parameters adjusted for 256 elements per horizontal and vertical line.

TABLE C.2-3

FIDELITY MEASUREMENT SCHEDULE

RE-ORDER No. 64-712
Vol I

APPENDIX D

RECOMMENDED SYSTEM

D.1 GENERAL

This section describes one possible method of implementing the linear approximation bandwidth compression technique as described in paragraph A.9.

D.2 SYSTEM DESIGN

D.2.1 Linear Approximation Encoder

A logic diagram of this system is shown in Figure D.1-1. The incoming video signal is applied to a series of three sample-and-hold circuits. Sample and Hold 1 samples the video at each element and holds it until the Reset 2 pulse occurs as shown in the timing diagram, Figure D.1-2. During this time the output of 1 is sampled by 2. The output of 2 is the element prior in time to the element which is currently being examined by the encoder. Sample and Hold 3 in turn samples the output of 2, but only when enabled by the transmit gate. This series results in the last good element being held in Sample and Hold 3.

The last good element, held in Sample and Hold 3, is applied to an operational adder along with the inverted video and a slope from the slope generator. If this slope plus the last good element is a close approximation to the video, the output summation will be zero volts. This output is tested with a threshold device to determine if it is within the preset error band. The threshold device is sampled during period "Sample 1." If it is out of tolerance, a flip-flop is set to the zero state. At the same time the output of this flip-flop, with built-in delay, is sampled and used to set a second flip-flop to the zero state if the threshold was out of tolerance at the previous sample time. These circuits are repeated 31 times in parallel, once for each slope which must be tested.

The output of each of the 31 flip-flop A's is OR'ed so that if any one slope is still in tolerance the output is an "in-tolerance" signal. However, if all slopes are out of tolerance an appropriate signal is generated. This signal activates a chain of events which results in the transmission of the code word representing the best slope and its length.

202

RE-ORDER No. 64-712

Vol I

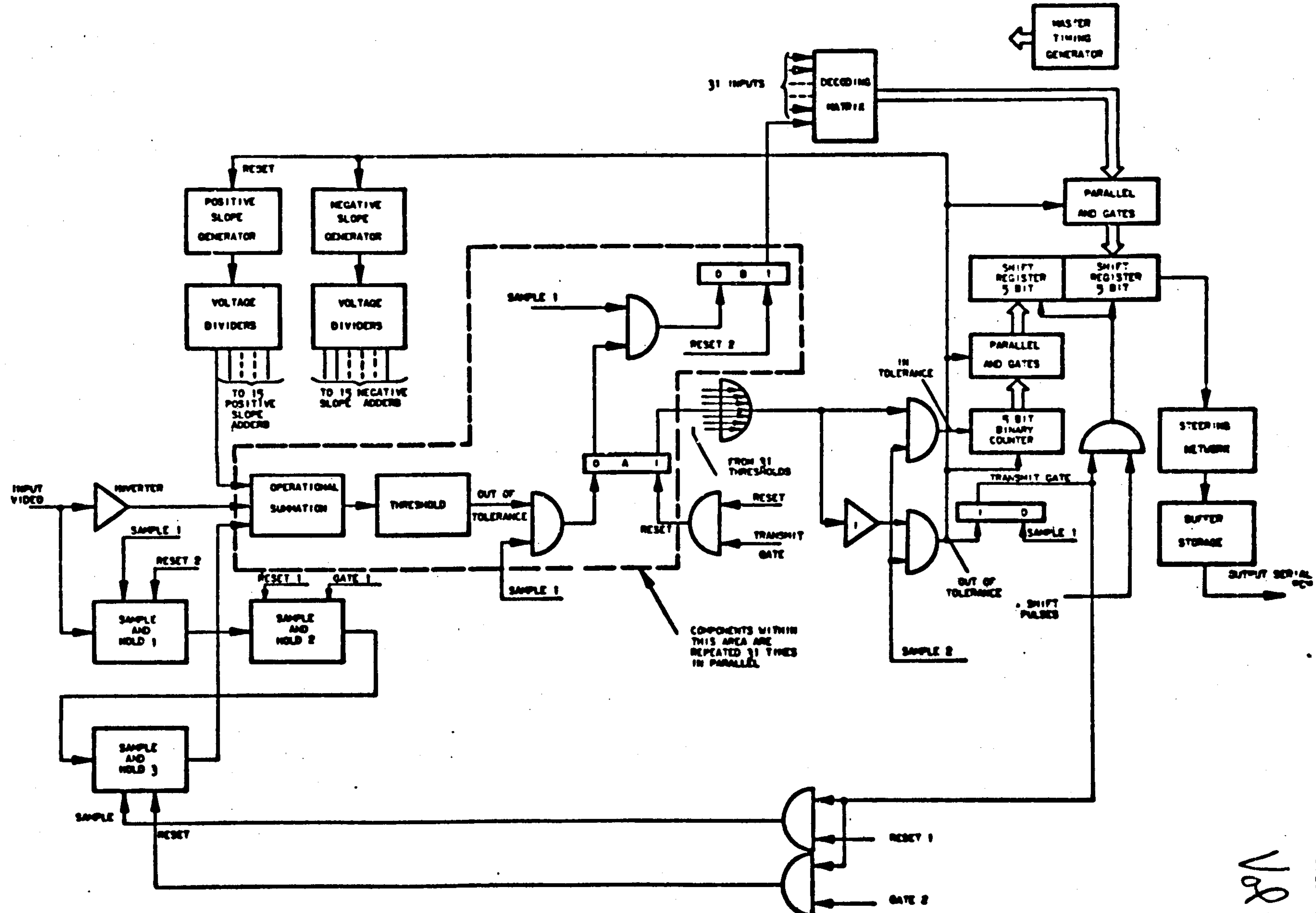

FIGURE D. 1-1

BLOCK DIAGRAM OF LINEAR APPROXIMATION ENCODER

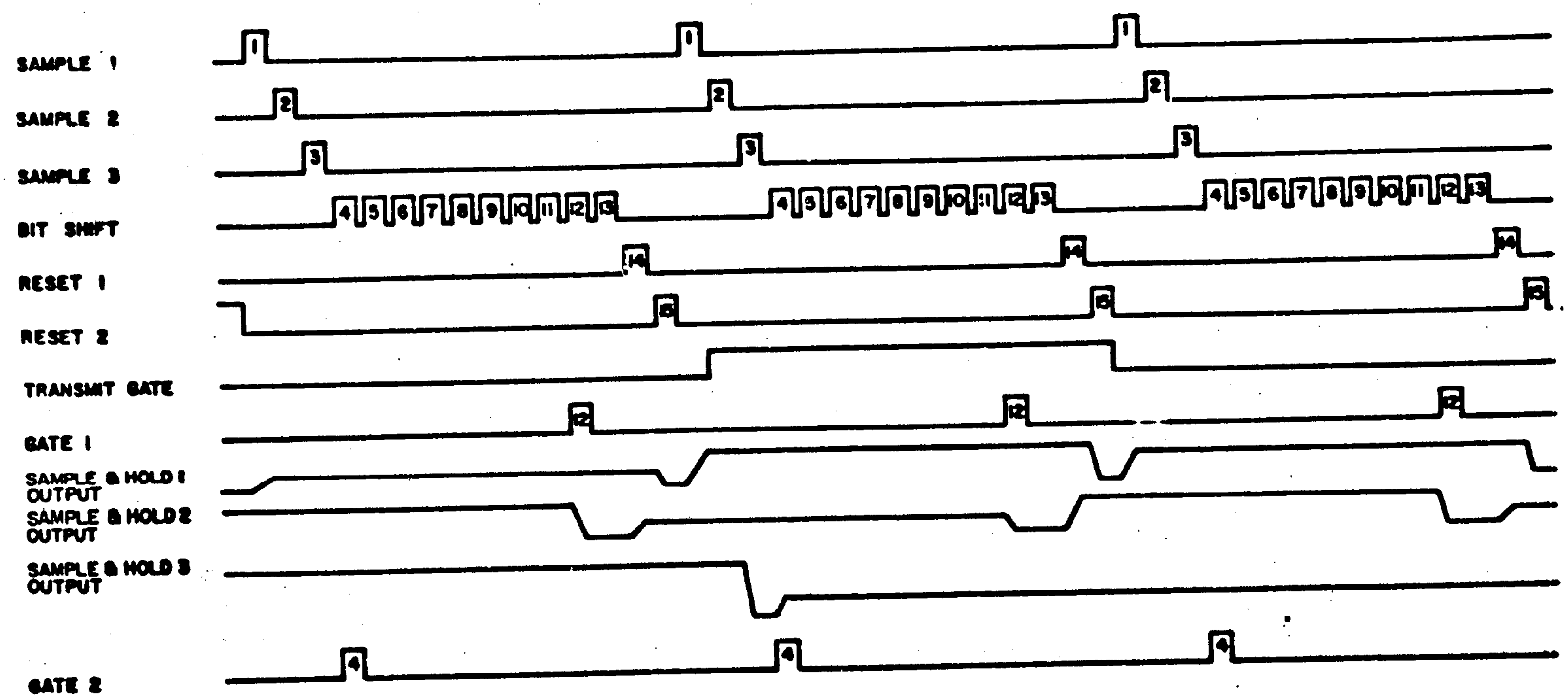

FIGURE D. 1-2

TIMING DIAGRAM FOR LINEAR APPROXIMATION

RE-ORDER No. 64-712
Vol I

RE-ORDER No. 64-712
Vol I

A five-stage binary counter counts the number of "in-tolerance" pulses. The "out-of-tolerance" pulse transfers this count into a shift register and resets the binary counter. The pulse also reads a code representing the proper slope from the decoding matrix into the shift register. These two five-bit code words comprise the ten-bit word which is transmitted.

Timing is controlled by the timing generator shown in Figure D. 1-3. A crystal oscillator is counted down by 16; AND gates are used to separate the 16 various element pulses. The timing generator also provides the horizontal and vertical blanking waveforms.

D. 2. 2 Buffer Storage

The buffer store is not shown in detail as its design is completely dependent on system boundary conditions which at this time are unspecified by operational requirements. A possible system would use a tape recorder as the main store, with a small register storing data in blocks which would then be recorded asynchronously.

D. 2. 3 Linear Approximation Decoder

The incoming serial pulse train is passed through a bit synchronizer and a sync separator, shown in Figure D. 1-4, which extracts the synchronization. The video data is directed into a shift register buffer storage. The PCM is then loaded ten bits at a time into the decoding register. The five bits on the left determine the slope to be generated. The first bit represents positive or negative slope. The next four are decoded in the digital to analog converter to provide amplitude information to the current sources.

One of the two current sources is gated on by the first bit; the magnitude of the current is determined by the decoded slope amplitude. This current is fed into the holding capacitor resulting in a voltage ramp which reconstitutes the original video.

The next five bits which represent the length of the slope are compared with the number of elements which have been displayed. When the two are equal, an End-of-Segment Pulse is generated. This pulse resets the counter and loads another ten bit word into the decoding register.

RE-ORDER No. 64-712
Vol I

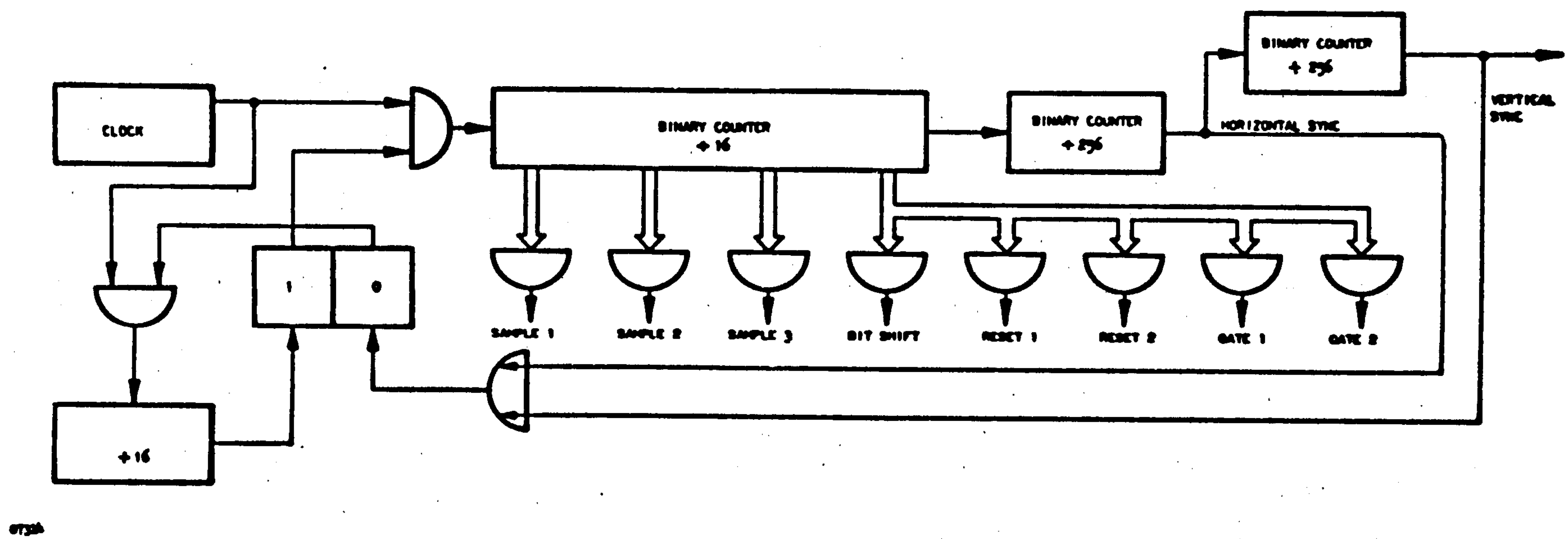

FIGURE D. 1-3

LINEAR APPROXIMATION TIMING GENERATOR

RE-ORDER No. 64-712
Vol I

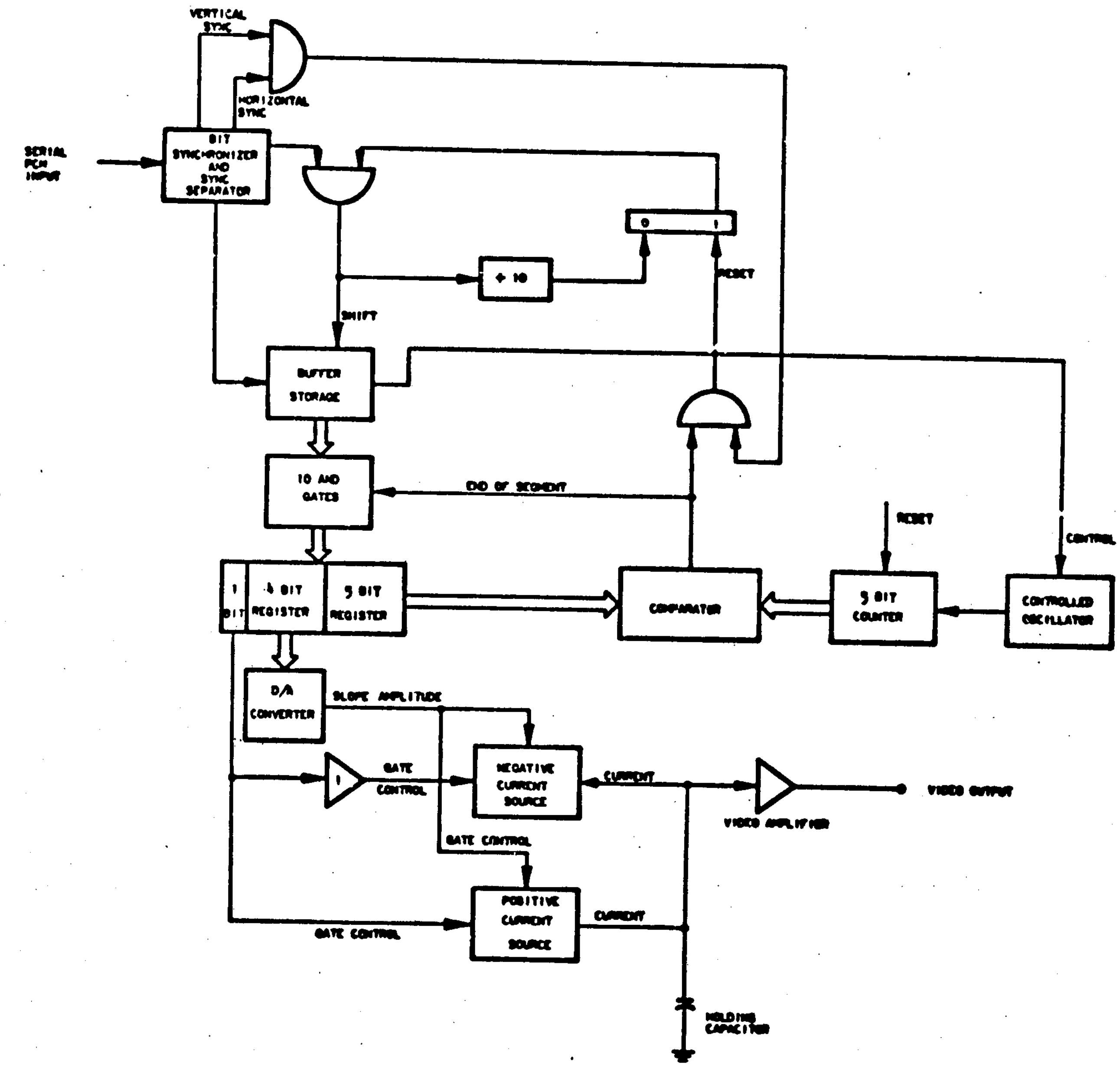

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FIGURE D. 1-4

LINEAR APPROXIMATION DECODER

RE-ORDER No. 64-712
Vol I

Timing of the decoder is controlled by an oscillator. The oscillator is in turn controlled over a small range by signals from the buffer store to prevent the storage from overflowing or underflowing.

D. 3 CIRCUIT DESIGN

Currently, it is both possible and practical to implement a system design in a number of ways using solid state active devices in conjunction with either diffused or more conventional passive devices. The limitations of the various electronic devices are well understood in all areas with the possible exception of particle radiation, particularly protons. For low-power circuits, a wide variety of form factors for the various electronic devices are available. Also, most circuits may be implemented in several different ways. Thus, the implementation of any system design is quite readily made to favor any prime requirement, i.e., size, weight, reliability, etc.

Initial logic design (digital system design) usually is made without consideration of circuit implementation. Thus, a literal circuit implementation of any initial logic design stands little chance of functioning properly. Initial logic designs are intended to verify that a system concept is practical and to give some idea of the relative magnitude of the system. This logic design is initial and is not intended to be optimum insofar as a particular circuit implementation of the logic design is concerned.

For minimum-size, hybrid semiconductor integrated circuits were used in the following estimates. It was necessary to use hybrid semiconductor integrated circuits (SCIC) instead of total semiconductor integrated circuits, since some of the circuits are not available in integrated form. However, if cost was of minor importance, the complete encoder system could be constructed with integrated circuits.

Table D. 2-1 is a tabulation of the parts estimate for the hybrid-SCIC configuration of the encoder.

208

RE-ORDER No. 64-712

Vol I

Circuit	Integrated Circuit Parts							Discrete Parts						
	Operational Amplifier	Thresholds	NAND Gates	Flip-Flop	Diode Array 6 diodes	Binary Counter (mw)	Power Dissipation (mw)	Transistors	Diodes	Resistors	Capacitors	Other	Power Dissipation (mw)	Volume (in 3)
Video Inverter	1	-	-	-	-	-	70	-	-	2	-	-	-	.068
Sample and Hold (3 units)	-	-	-	-	-	-	-	21	3	33	18		1263	1.851
Slope Generators (2 units)	-	-	-	-	-	-	-	6	4	42	6	-	364	1.028
Comparison Ckts. (31 units)	31	31	62 (2 in-put)	62	-	-	7360	-	-	31	-	-	-	6.634
Decoding Matrix	-	-	10 (3 in-put)	-	30	-	530	-	-	-	-	-	-	1.60
Shift Register	-	-	-	-	-	10	500	-	-	-	-	-		.4
Binary Counter	-	-	-	-	-	5	250	-	-	-	-	-		.2
Misc. Gates	-	-	17 (2 in-put) 2 (3 input)	-	6	-	531	-	-	-	-	-		.64
Timing Generator	-	-	19 (2 in-put)	1	-	24	1710	-	-	-	-	Crystal		1.40
Power Supply (known high reliability supply) TOTAL POWER							10951 5200 17.78 watts						1627	4.00 17.827

TABLE D. 2-1

LINEAR APPROXIMATION PARTS COUNT ESTIMATE

RE-ORDER No. 64-712
Vol I

The volumes of various parts shown in Table D. 2-2 were used for the estimates of Table D. 2-1. The TO-5 transistor-case volume was used for all SCIC circuits. The volumes given allow for projecting leads and parts. Power estimates used are given in Table D. 2-3. Resistor power is considered in the transistor power allowed.

These estimated values for power and size relate only to actual component parts and power. Experience has shown a need for correction factors to arrive at the values for the finished component. The following correction factors were applied to the calculated data:

(1) Power -- Prototype power equals two times parts power.

(2) Volume -- Prototype volume equals three times parts volume.

(3) Weight -- Prototype weight equals 0.5 ounce per cubic inch for electronic parts plus one ounce per cubic inch for the power supply.

Using these correction factors the prototype encoder would have the following parameters:

Power -- 35.56 watts

Nonstructural volume -- 53.5 cubic inches

Nonstructural weight -- 1.80 lbs.

The above estimates were based on the use of Motorola, "MECL" integrated circuits. At the present time these are the highest speed integrated circuits; however, they consume a large amount of power. If system speeds permitted, it is possible to trade speed for lower power consumption. Also, the TO-5 package was chosen over the flat-pac as being more compatible with the discrete components. There are many trade-offs which could possibly be made and the combination chosen here is not necessarily the optimum for any one particular set of system design parameters. However, it does give an idea of the relative size and complexity of the recommended system.

The estimated system would be operational up to logic rates on the order of 10 megabits/second which would allow element rates of approximately 625 kc. Motorola specifies a speed/power trade-off of 20 times power for 10 times in

RE-ORDER No. 64-712
Vol I

speed. If special low power "MECL" components were used, the total parts power would drop to 9.78 watts. This gives a final power of 19.56 watts and would result in an element rate of approximately 62.5 kc. If each picture element were compressed to two bits (average), then this low power implementation would provide average bit rates as high as 130,000 bits/second. According to a recent[1] projected bit rate performance of a deep space telemetry system, bit rates of 130 kc or higher will not be available from Venus until after 1967, nor from Mars until after 1969.

Although the MECL components offer a 20:1 trade in power for a 10:1 trade in speed, an overall system power saving greater than 2:1 could not be realized in this analysis because the power consumption of the operational amplifiers and schmitt triggers was not lowered. If cost were of secondary consideration, special, low power, narrower bandwidth operational amplifiers and schmitt triggers could be designed. The speed of these redesigned, lower power components would then be more compatible with expected spacecraft data rates.

[1] Thatcher, John W., Mgr., Deep Space Networks, JPL, "Deep Space Communication," Space/Aeronautics, July 1964.

211

RE-ORDER No. 64-712
Vol I

TABLE D. 2-2

VOLUME ALLOWED FOR PARTS ESTIMATE

Item	Volume (cubic inches)
TO-5 transistor case	0.04
Diode	0.005
Resistor	0.014
Capacitor	0.03
Crystal	0.2

TABLE D. 2-3

POWER ALLOWED FOR PARTS ESTIMATE

Item	Power (milliwatts)
SC1C operational amplifier	70
SC1C threshold circuits	50
SC1C 2-input NAND 3-input NAND	25 35
SC1C flip-flop	35
SC1C diode array	6
SC1C binary counter	50 avg.
Amplifier transistors	60
Diodes	1

212

RE-ORDER No. 64-712
Vol I

APPENDIX E

FIDELITY MEASUREMENT PROGRAM

E. 1 BACKGROUND INFORMATION

This section describes the procedures employed in carrying out the three-phase fidelity measurement program defined in Appendix C. 2, Measured Fidelity, and measurement results in the form of photographic recordings of EDITS output data are discussed in this section. With reference to the several video modulation techniques under evaluation, this discussion will serve to amplify the fidelity appraisal comments included in the compression sections of Appendix A.

In order to provide a firm basis for interpreting and evaluating the results obtained from the fidelity measurement program, the following background information is presented. This includes a description of the salient features of the Experimental Digital Television System (EDITS) developed by EMR, a review of the video signal-to-noise calibration procedure employed therein, and a discussion of basic EDITS television camera resolution considerations applicable to the measurement program.

E. 1. 1 EDITS System Summary

Basic Function

The system is designed primarily as a laboratory tool to investigate statistical and pictorial characteristics associated with various digital television processing techniques. Its operating video data modulation systems include:

- 1, 2, 3, 4, 5, and 6 Bit PCM
- Roberts (any combination of 7 bits total, except the 0-7 combination)
- PEC/Huffman
- 2-Bit Delta

Physical Characteristics

The electronic equipment comprising EDITS, with the exception of the TV camera proper and the analog noise generator, are rack mounted. The equipment,

RE-ORDER No. 64-712
Vol I

including the auxiliary digital noise generator, rms voltmeter and standard low and high voltage power supplies, is contained in five, 30-inch racks, as shown in Figure E. 1 P-1. The basic electronic system element is the plug-in module, a considerable number of which are standard DEC (Digital Equipment Corp., Maynard, Mass.) logic units. Figure E. 1 P-2 shows a view of the TV camera unit, mounted in the light shield housing viewing the test fixture which mounts and illuminates the input test photographic data.

Input/Output Equipment Characteristics

Sensor	Westinghouse one-inch all-electrostatic vidicon, Type WX-4384 (slow-scan type), rated limiting resolution: 500 TV lines in center, and 350 TV lines in corners.
TV Display	Tektronix Type RM503, 5" all electrostatic CRT with P4 phosphor. Rated limiting resolution - approximately 350 to 400 TV lines.
Display Recorder	Polaroid camera - employ speed/type 47 film. Film speed - ASA 3000.

System Operating Characteristics

Scan elements/line	128, 256, or 512 (horizontal and vertical), selectable
Frame periods	0.16, 0.66 or 2.6 seconds, respectively
Frame rates	6, 1.5 or 0.4/seconds, respectively
Element period	10 microseconds
Scan type and format	Horizontal and vertical stairstep scan with square format
Digital encoding	1 to 6 bits. Element sampling rate - 100 kc

Signal Statistical Measurement Capability

- First Order -- number of times a scan element of a given level occurs within a selected period. This period is selectable by means of gate switching, from a total frame or line period to a fractional frame or line.
- Second Order -- number of times a given element level follows another given element within a selected period.
- Run Length (RL) -- number of times a given element level successively repeats within a selected period.

RE-ORDER No. 64-712
Vol I

- Previous Element (PE) -- number of times within a selected period that adjacent elements along a horizontal line have equal levels.
- Not Previous Element (1-PE) -- number of times within a selected period that adjacent horizontal elements do not have equal levels.
- Previous Line (PL) -- number of times within a selected period that matching levels occur on vertically adjacent elements (which are not PE elements).

Noise Simulation Capability

- Channel (digital) Noise -- P_e from $10^{-\infty}$ to 5×10^{-1}. Noise source is a General Radio Random Noise Generator Type 1390-BR plus adjustable threshold and sampling circuitry.
- Sensor (analog) Noise -- S/N from more than 40 db (limited by vidicon/preamplifier combination) to 15 db (lower if desired).

Test-Chart/Test-Photo Illumination

- Front -- incident illumination is approximately 150 foot candles.
- Back -- screen brightness is approximately 250 foot lamberts.

Typical Vidicon Exposure

Using normal 25 or 50 mm lens at an aperture of f/8 (effective), and operating at 256-element scan condition, typical photocathode highlight exposure is approximately 0.3 foot-candle/sec.

Block Diagram

Figure E.1-1 is a simplified block diagram of the EDITS system. This diagram depicts the various functions listed in the above section.

E.1.2 Video Signal-to-Noise Calibration

Introduction

If a nonvarying voltage is applied to an ideal analog-to-digital encoder, the resultant binary code will identify one particular level for each sample. Unfortunately, there is always some noise in the applied voltage and in the threshold levels. This paragraph derives the probability that an encoded level will be the same as the previous level. This is defined as the PE (previous element) probability and results in a curve of the maximum possible PE for a given system S/N ratio.

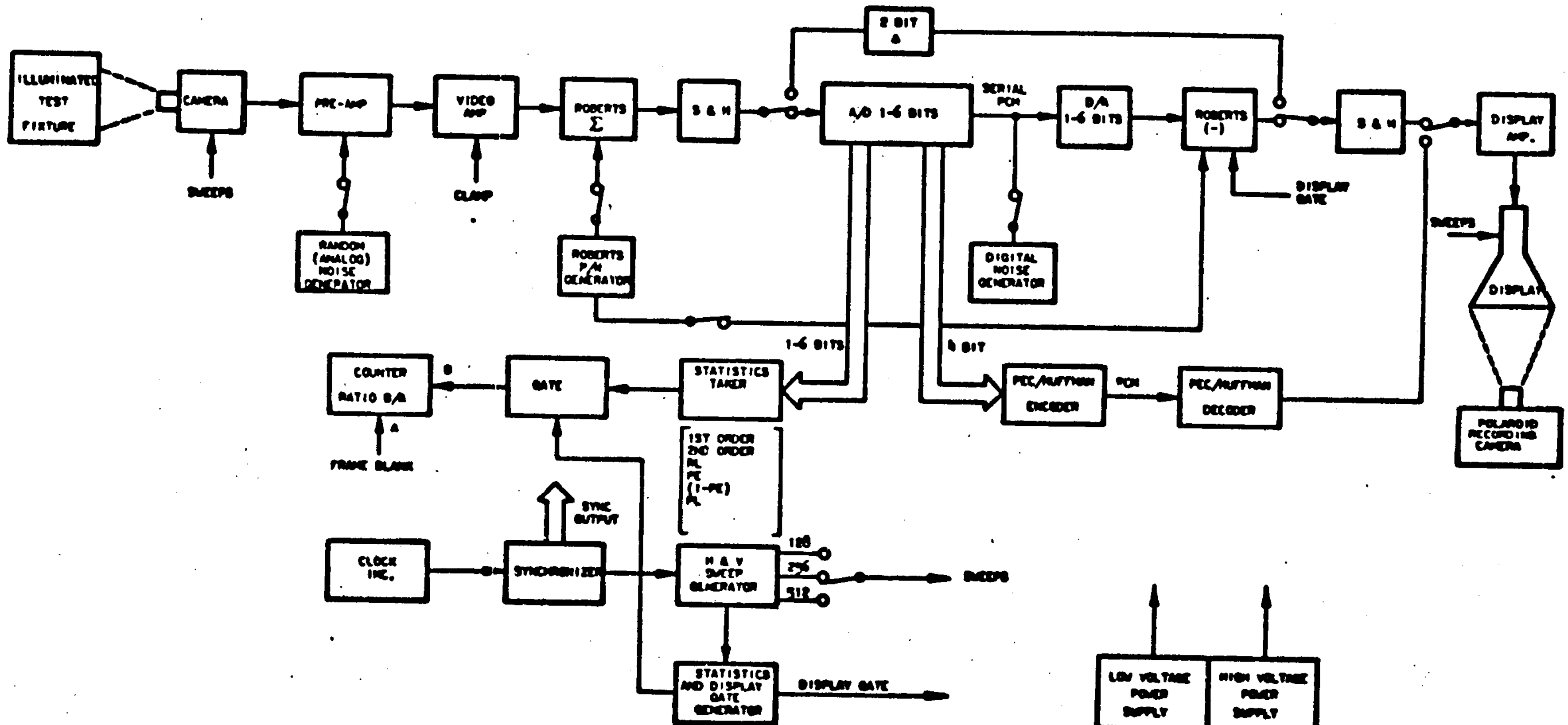

FIGURE E. 1-1

EDITS BLOCK DIAGRAM

RE-ORDER No. 64-712
Vol I

RE-ORDER No. 64-712
Vol I

Analysis

Figure E. 1-2 shows an input signal level which falls a distance X above level 19. The curve is the Gaussian probability density function for the noise which is riding on the input signal level. The noise has an RMS value of δ, and Δ is the distance between quantizing levels.

The probability that the signal is correctly identified as level 19 is equal to the area under the Gaussian curve labeled (2). The probability that the signal is identified as level 20 is equal to the area labeled (1). Similarly, P(21) = area (5), P(17) = area (4), and P(18) = area (3). This must be continued until the probabilities become insignificant.

The previous element probability is equal to the probability that the signal is encoded to the same level for the present sample as it was for the last sample. For the figure as shown this becomes

$$PE = \ldots + P(17)^2 + P(18)^2 + P(19)^2 + P(20)^2 + P(21)^2 + \ldots$$

This statement assumes the two samples are statistically independent. For a given ratio of δ / Δ, PE will vary as $\frac{X}{\Delta}$ varies from 0 to 1. Figure E. 1-3 is a family of curves of PE versus the ratio $\frac{X}{\Delta}$ for various signal-to-noise ratios. For a six-bit encoder with 64 levels, the signal (peak-to-peak) to noise (rms) ratio is $64 \, \Delta/\delta$. For the curve of $\delta = \Delta$, Spp/Nrms = $64 \Delta/\Delta = 64 = 36.1$ db. The curves in Figure E. 1-3 are separated by 6-db steps.

If a slowly varying waveform is applied to the A/D encoder, the average PE will be equal to the average value of the curve in Figure E. 1-3 for the actual S/N ratio of the system. Therefore, it is assumed that a given S/N ratio will result in the average PE as X/Δ varies from 0 to 1.

Figure E. 1-4 is a curve of PE averaged over X for each S/N ratio. The S/N ratio scale is applicable to a six-bit encoder only. If the encoder is reduced to five bits, the curve shifts downward by 6 db (i. e., 36. 1 db on the scale becomes 30. 1 db).

RE-ORDER No. 64-712

Vol I

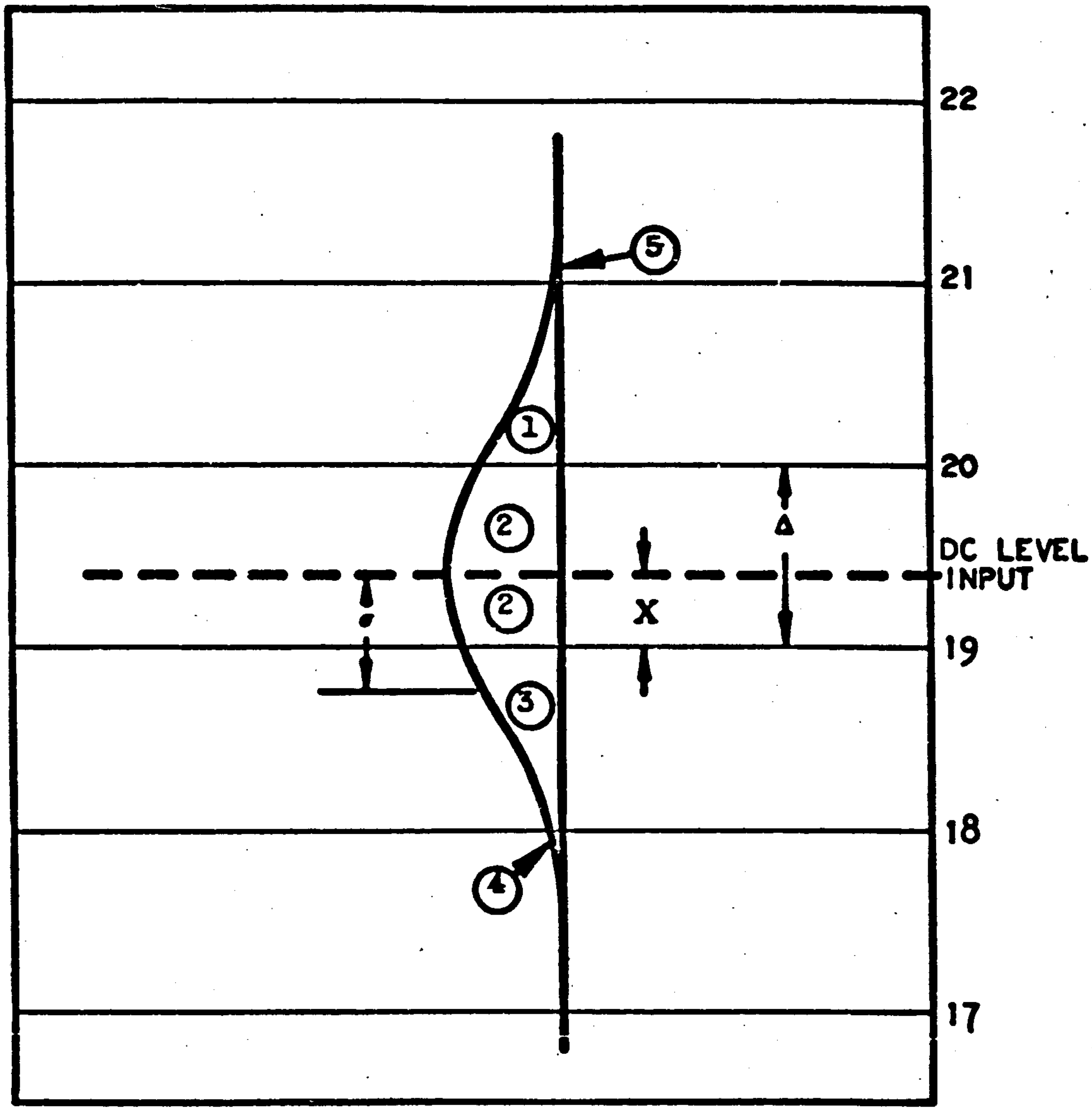

FIGURE E.1-2

DC INPUT LEVEL WITH GAUSSIAN NOISE

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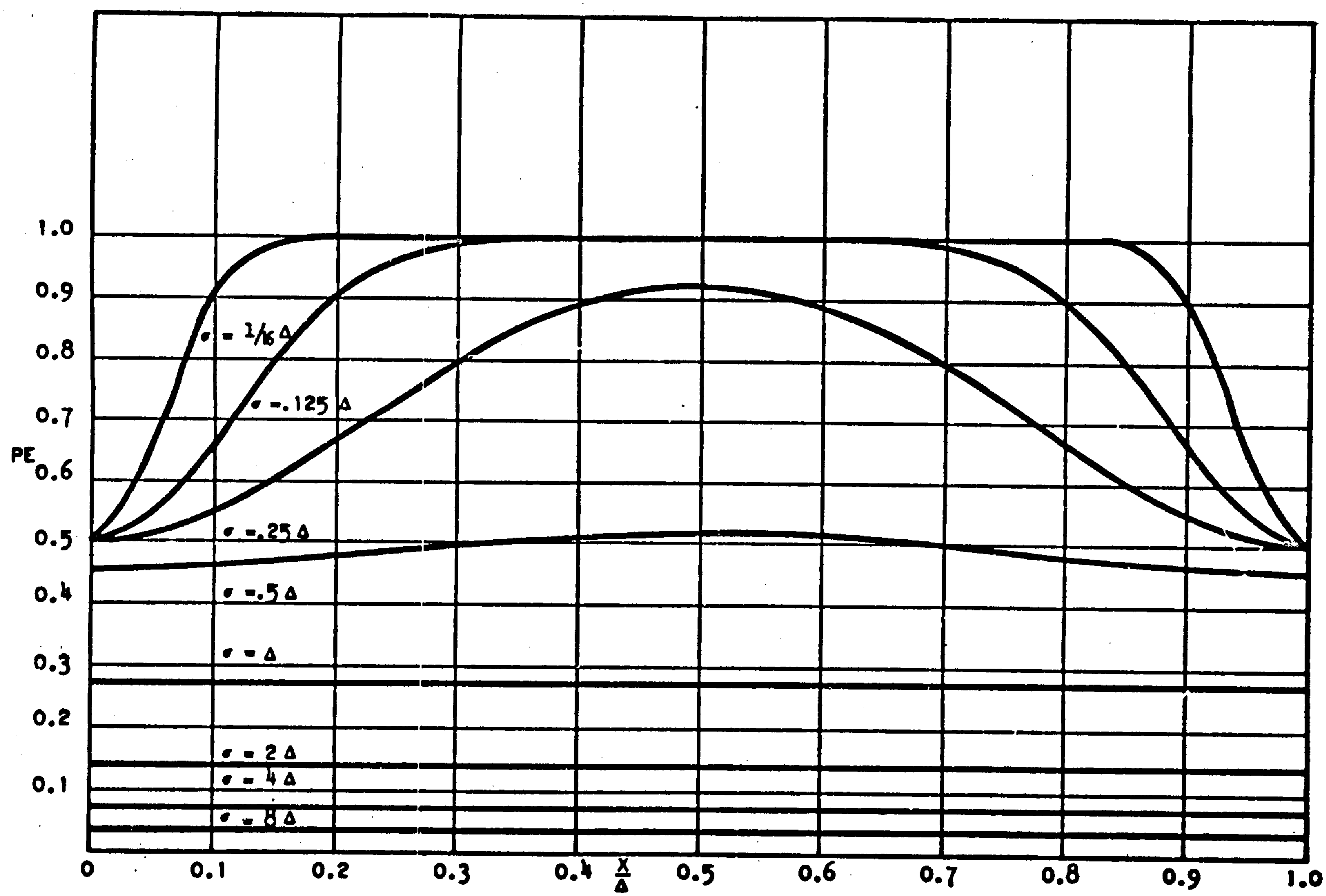

FIGURE E. 1-3

PE VERSUS DC POSITION BETWEEN QUANTIZING LEVELS

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RE-ORDER No. 64-712
Vol I

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RE-ORDER No. 64-712
Vol I

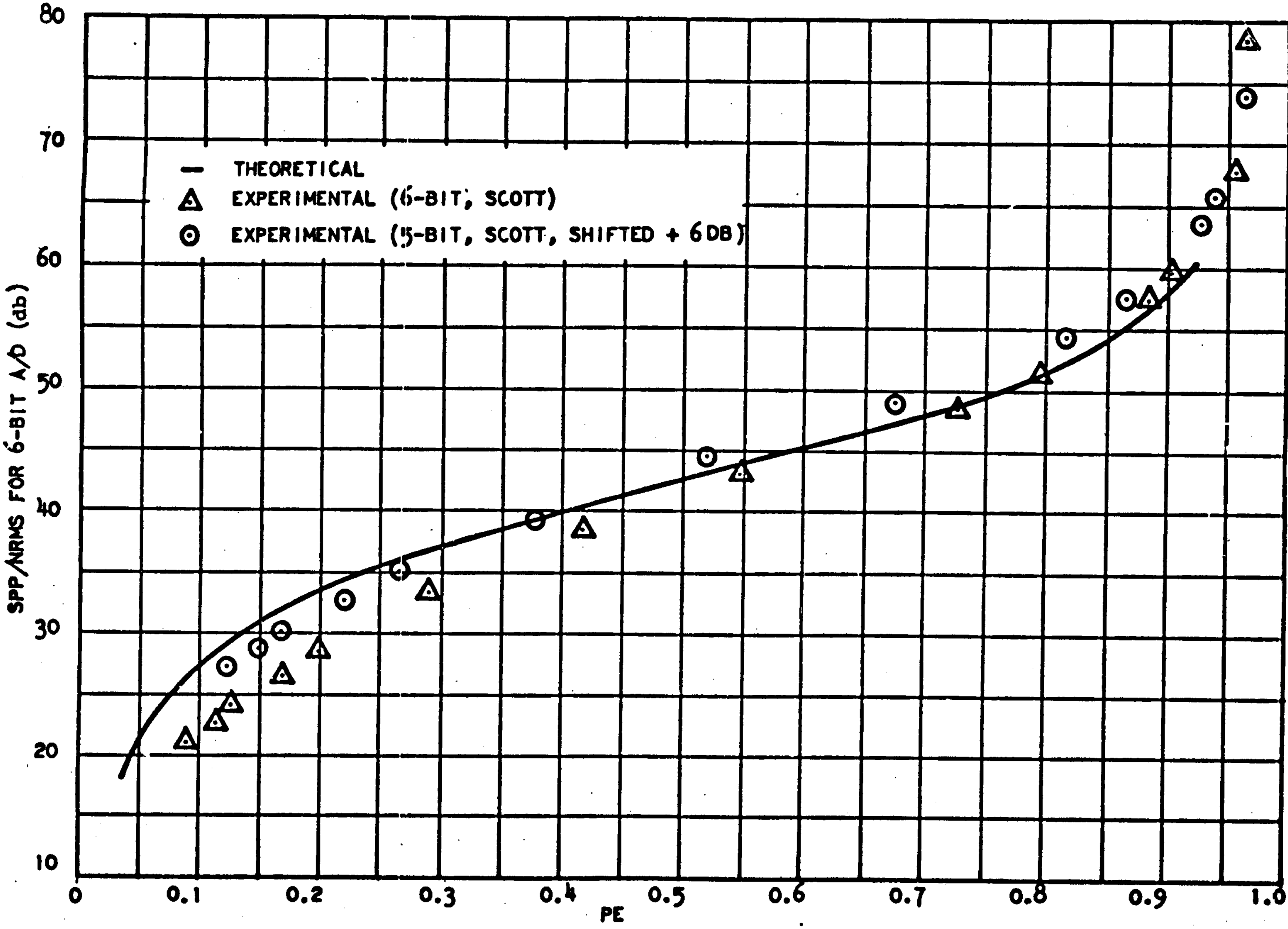

FIGURE E.1-4

PE VERSUS SPP/NRMS FOR A SIX-BIT A/D ENCODER

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RE-ORDER No. 64-712
Vol I

Experimental Results

Experimental verification of this curve was attempted using the equipment setup shown in Figure E. 1-5. Results are plotted in Figure E. 1-4 directly for the six-bit curve. The five-bit curve is plotted, but it is shifted upward by six db for comparison with the theoretical curve.

Conclusion

The results show reasonable agreement, although for the lower values of S/N, PE is higher than predicted. This could be caused by the noise being non-Gaussian or by the assumption of independent samples being invalid.

Whereas the absolute accuracy of the S/N measurement method may possibly be subject to a maximum error of several db, the technique has proved to be very useful in establishing repeatable S/N conditions for all relative fidelity measurements employing the EDITS equipment.

E. 1. 3 Basic TV Camera Resolution

The camera scanning method, format and size employed in EDITS depart from the conventional analog vidicon TV camera system. In the standard analog system, the camera beam sweeps along a given slanted line (the slant created by the action of the vertical deflection waveform)--ideally, at a constant velocity. It rapidly retraces and repeats the scan process for the next and successive lines. In the EDITS equipment, the horizontal scanning action is a discontinuous one in which the camera beam dwells on successive elements along a line and moves from element to element in a relatively short period of time. We classify this scan operation as digital indexing, since the horizontal deflection waveform is in the form of a staircase which is derived from a resistance-ladder network, which in turn is fed by binary-counter/analog-gate circuitry controlled by the synchronizer unit. This discontinuous sweep action also occurs in the vertical direction on a line-to-line basis.

All of the fidelity measurements discussed in this report were taken at a scan condition of 256 elements per horizontal line and 256 lines per frame. Allowing for a small overscan beyond the vidicon photocathode mask, the number of active elements along a horizontal or vertical scan line is approximately 240.

RE-ORDER No. 64-712
Vol I

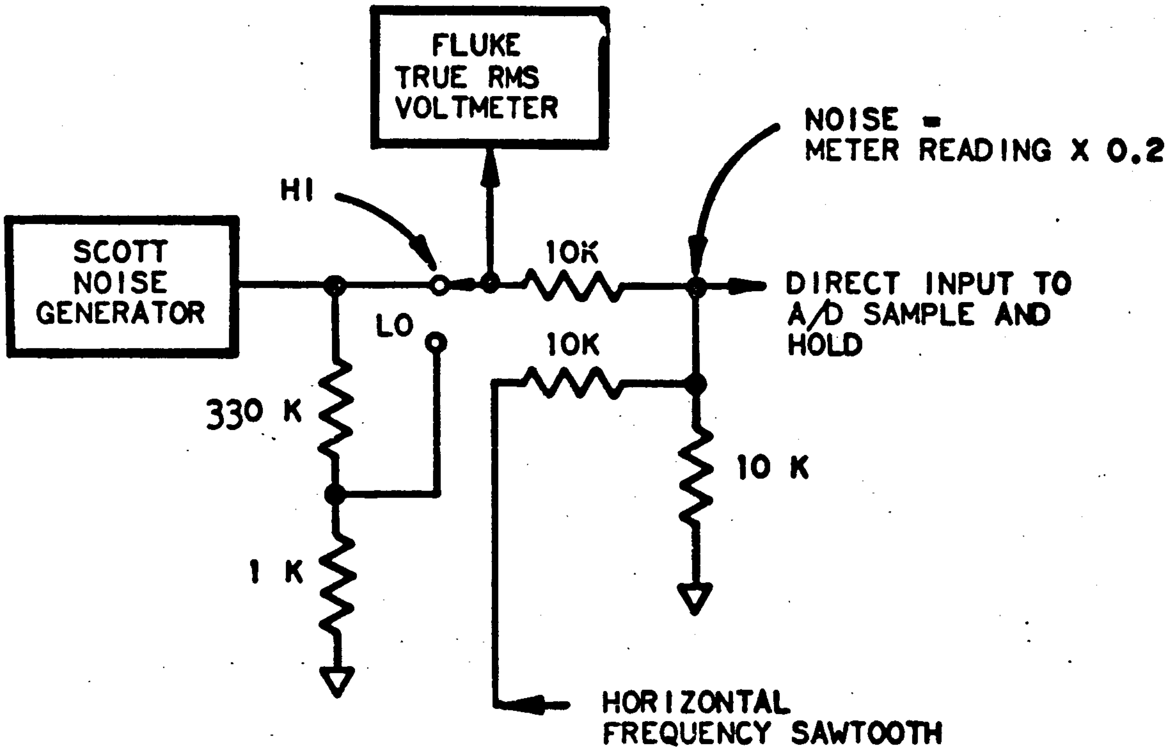

FIGURE E.1-5

EXPERIMENTAL PE VERSUS S/N SETUP

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RE-ORDER No. 64-712
Vol I

The format used in these measurements is square, as opposed to the 3 by 4 aspect ratio employed in conventional systems. (EDITS readily possesses the flexibility to operate with the 3 by 4 scan ratio; however, it was determined that a square format was more desirable here based on the nature of the majority of input test material employed.) A square mask in front of the vidicon faceplate defines this active raster area in the EDITS camera. A 0.44-inch square area is employed, which provides a 0.625 inch diagonal (equal in length to the diagonal of the conventional scan of 0.375 inch x 0.50 inch). This relationship is shown in Figure E.1-6(a).

In the measurement of camera-system resolution, the spatial frequency of the test pattern imaged on the photocathode surface may serve as a reference against which camera response is judged. Based on 240 active elements per line, the number of element samples per inch equals approximately 550 $(\frac{240}{0.44})$. This provides 275 element pairs (line pairs) per inch, which is equivalent to eleven line pairs per millimeter. Thus, based on the sampling theorem requiring a minimum of two samples per cycle of reproduced frequency, the camera system should be capable of reproducing image spatial frequencies up to eleven cycles per millimeter on the photocathode. A comparison of measured versus theoretical sine wave response of the EDITS camera is given in paragraph E.2.

A second approach to resolution measurement is that of directly relating the camera response to the standard "TV resolution lines" method employed in conventional practice. In order to compare the EDITS camera performance directly with that of a standard vidicon camera, it is necessary to produce equivalent test chart image spatial frequencies on the EDITS vidicon photocathode. Figure E.1-6(b) depicts the condition described. In this example, it is assumed that spacing of the black and white resolution bars is such that a total of 200 bars occupies the full 0.375-inch conventional raster height. Since the EDITS vertical scan is 0.44 inches high, this would be equivalent to 235 black and white bars in the vertical direction. Correspondingly, in the horizontal direction, where the conventional system requires 267 (4/3 x 200) similarly spaced bars, the EDITS camera requires 235. Therefore, when viewing the standard resolution charts used in subsequently described measurements only

RE-ORDER No. 64-712
Vol I

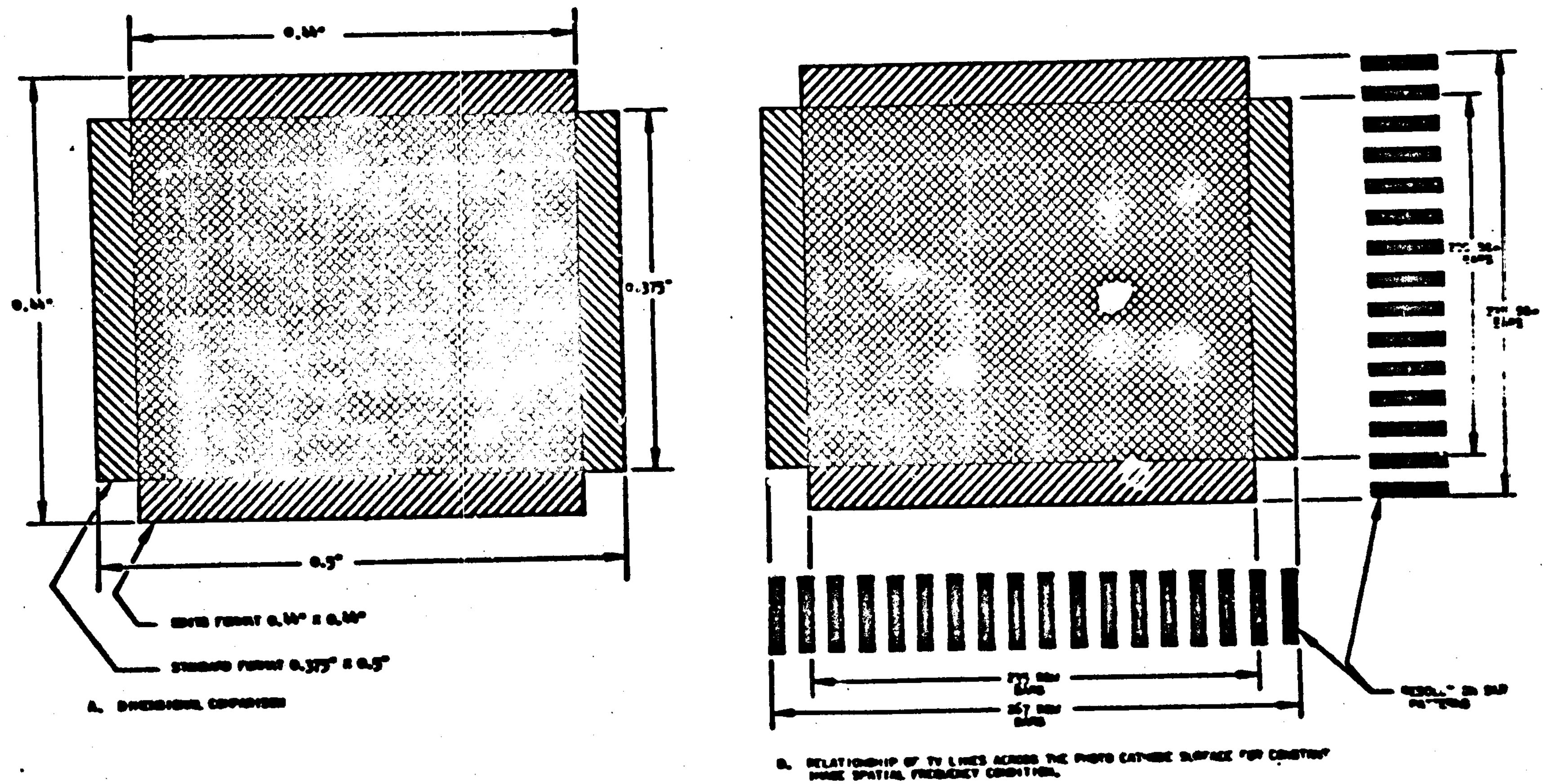

FIGURE E. 1-6

COMPARISON OF CONVENTIONAL VIDICON SCAN FORMAT WITH EDITS SCAN FORMAT

RE-ORDER No. 64-712
Vol I

$88\% \left(\frac{0.44 \text{ inches}}{0.5 \text{ inches}}\right)$ of the normal test chart width is imaged on the active vidicon photocathode area, while the image height is correspondingly greater than the chart height. In this manner, the resulting camera resolution measurements in TV lines may be directly related to conventional vidicon resolution figures.

In the example given, the EDITS vidicon must generate 235 elements along horizontal and vertical lines to provide one element sample per resolution bar. At first thought, it might be assumed that this 200 TV lines represents the maximum resolution potential of EDITS operating in the 256 element scan condition (240 active elements). In practice, it has been found that this does not represent the limiting case as is evidenced by some of the limiting square wave resolution measurement photos presented in paragraph E.3. It appears that a more reasonable maximum resolution figure is 250 conventional TV lines, which is equivalent to almost 300 black and white bars across the 0.44 inch photocathode. The natural question asked is how can there be more resolution bars along a line than scan elements. Figure E.1-7 offers one possible explanation for this. Since the EDITS TV camera and TV display (at lower brightness levels) both are capable of producing considerably better resolution than is presently set by the indexed scan, a condition similar to that depicted could occur. On the 200 line scan, beam positions (1) through (5) would produce full video modulation. On the 250 line scan with the scanning aperture size shown, beam positions (7), (8) and (9) still produce full video modulation, while positions (6) and (10) produce some intermediate video level. Thus, although the spacing of elements is larger than the resolution bar spacing by approximately 25%, the higher resolution bars are still reproduced.

It is recognized that the above represents a somewhat special case, since other positions of the scan beam relative to the resolution bars can produce a degradation in response, even below the 200 TV line level. This of course is where the well-known television Kell factor enters into the picture. However, it is apparent that as the scanning aperture size is reduced, the resolution degradation (normally accounted for in the Kell factor) becomes less; therefore, the EDITS equivalent factor should be increased from the figure of approximately 0.7 used in commercial TV practice up nearer to the unity value.

RE-ORDER No. 64-712
Vol I

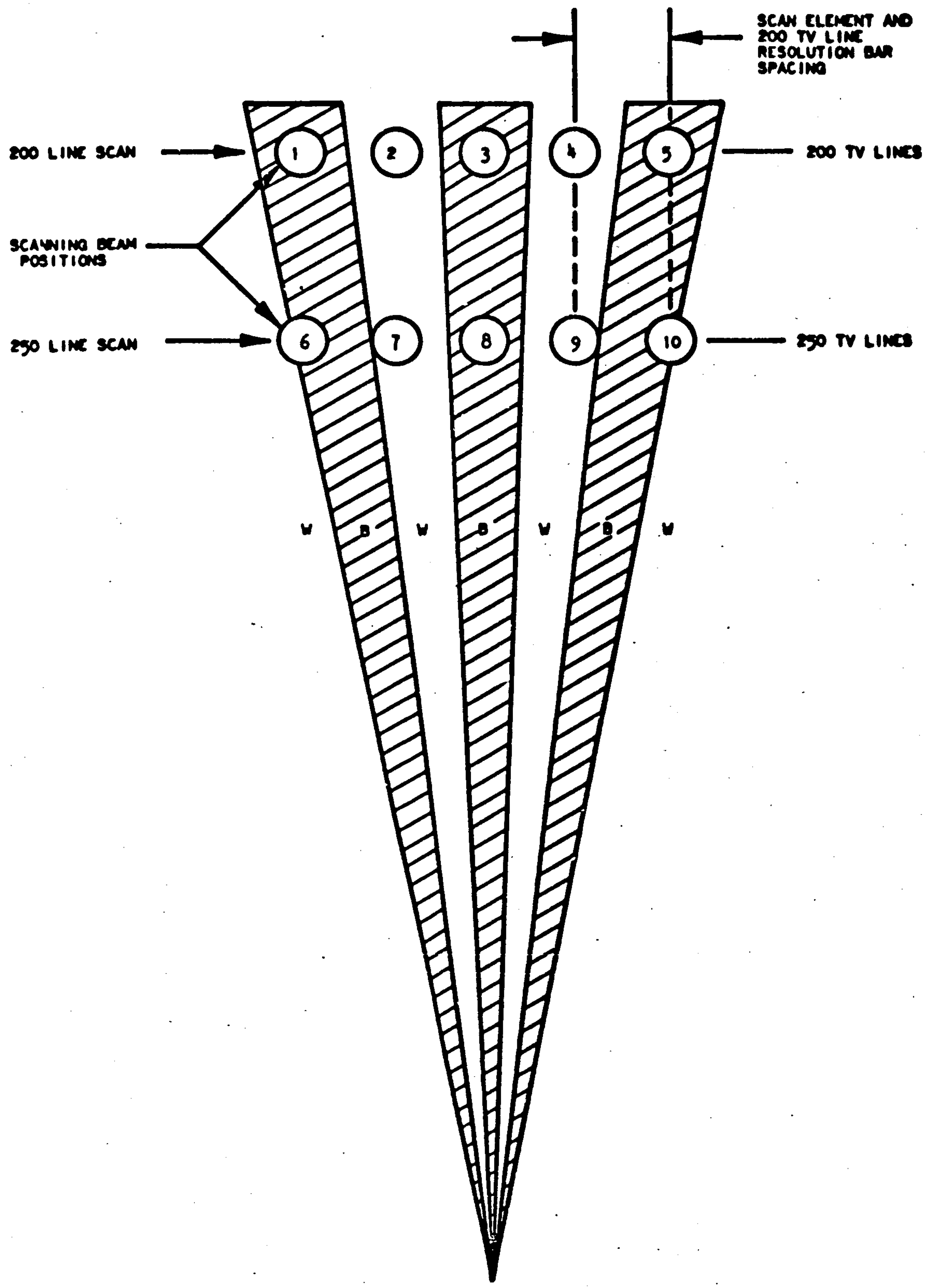

FIGURE E.1-7

RESOLUTION IMPROVEMENT EFFECT
PRODUCED BY A SMALL SCANNING APERTURE

RE-ORDER No. 64-712
Vol I

E.2 OBJECTIVE MEASUREMENTS

E.2.1 Introduction

Paragraph C.2 contains a summary of the measurement types and conditions employed in this test series. The three types of tests are (1) sine wave response, (2) square wave response, and (3) gamma characteristic. The principal objective of tests (1) and (2) is that of providing absolute measures of signal amplitude response versus input image spatial frequency for the two fundamentally different video modulation methods under evaluation. These are straight PCM (6 bits in this case), and Delta modulation (2-bit Delta is used). The output of test (3) is of considerable value in the subjective measurement program in determining the dynamic operating range expected of the system relative to the total contract range of the input photographic test data.

E.2.2 Objective Test Procedures and Results

Sine Wave Response

Test Procedure. The vidicon camera is positioned to view a selected 1-1/4 inch wide section of the back-illuminated 30% modulation Kodak Sine Wave Test Object transparency. This is accomplished by use of a 50 mm focal length lens plus the additional required lens-to-photocathode extension to provide the proper image and object distances. Figure E.2-1 illustrates the layout of the basic chart sinusoidal spatial frequencies and denotes the area selected for this particular test (the surrounding chart area is masked off).

Since the 1-1/4 inch wide image just fills the photocathode width of 0.44 inches, this represents a size reduction of 2.84 times. Therefore, relative to the photocathode, the six selected groups of spatial frequencies are multiplied by the factor of 2.84, resulting in a sinusoidal pattern series of the following approximate values:

RE-ORDER No 64-712
Vol I

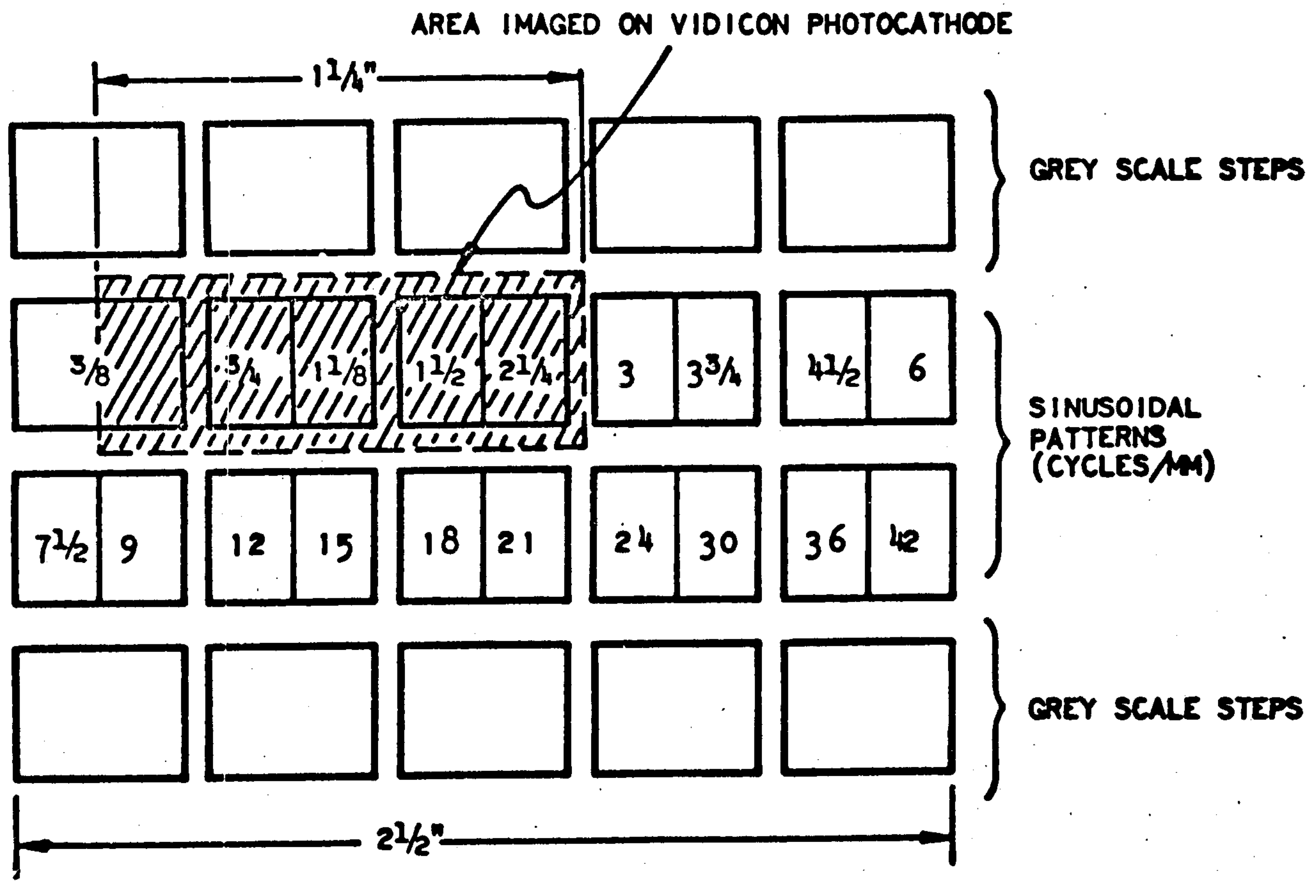

FIGURE E.2-1

SINUSOIDAL TEST OBJECT SPATIAL FREQUENCIES

07314

RE-ORDER No. 64-712
Vol I

Basic Frequency		Image Reduction Ratio		Spatial Frequency on Photocathode
3/8 cy/mm	x	2.84	≅	1 cy/mm
3/4 cy/mm	x	2.84	≅	2 cy/mm
1-1/8 cy/mm	x	2.84	≅	3 cy/mm
1-1/2 cy/mm	x	2.84	≅	4 cy/mm
2-1/4 cy/mm	x	2.84	≅	6 cy/mm

Measurements are performed at a video S/N $\geq$ 40 db to minimize the masking effects of noise on the basic system response characteristics. The most convenient point of measurement is the output of the display amplifier. EDITS gating circuits are adjusted to select a single scan line passing through the sine wave patterns. The resulting waveform is displayed on and recorded from the CRT.

Test Results. Figure E.2 P-1 presents the results of the objective sine wave response measurements for both six-bit PCM and two-bit Delta modulation methods in the form of CRT amplitude traces. Also, for ease and directness of comparison, the corresponding quasi-objective sine-wave outputs from the EDITS TV display are included in this figure; however, the latter results will be discussed in paragraph E.3.

From the figure, six-bit PCM response is seen to hold up well out to 4 cycles/mm (approximately 80% of the 1 cycle/mm response) and still shows about 40% response at 6 cycles/mm. Delta performance is considerably poorer. First, note the reduction in 1 cycle/mm response (to approximately 60% of the comparable PCM amplitude). The higher delta frequency components also are degraded more rapidly than with PCM resulting in a very low percent response at 6 cycles/mm.

It is concluded that, even with six-bit PCM, the theoretical response capability of 11 cycles/mm is at best barely achievable, and is not achievable with Delta. Limiting factors common to the two systems are as follows: first, the basic contrast, $\frac{\text{highlight brightness}}{\text{lowlight brightness}}$, of the input sinusoidal patterns is low (less than 2-to-1 for a 30% modulation chart). With the combination of low input

RE-ORDER No. 64-712
Vol I

contrast and falling response, the noise in the video signal assumes significance in determining limiting detectable response. Also, since EDITS performs discrete horizontal and vertical element sampling processes with the vidicon scanning beam, it is possible to encounter an out-of-phase condition where the sampling element locations on the photocathode occur between positions of maximum and minimum image brightness. In this respect, the discrete beam scanning/sampling process is directly equivalent to the more standard PCM television process of time-sampling the normal analog video signal from a camera which employs normal sawtooth scanning. This effect is related to the Kell factor which is applied to vertical resolution performance of conventional TV cameras. However, for reasons discussed in paragraph E.1.3, the conventional Kell factor of approximately 70% should not apply to this particular EDITS digitally-indexed scan mode. In addition, Delta inherently suffers more from a limitation of high frequency response than does PCM, which accounts for its more rapid response fall-off with increasing image spatial frequency.

Square Wave Response

Test Procedure. The vidicon camera with the extended 50 mm lens is adjusted to scan an image corresponding to a total width of two inches on the Westinghouse resolution chart. The area selected includes one set of the square wave black and white resolution bars, which covers the range from 100 to 1000 TV lines. Based on the previously discussed relationship between the EDITS camera raster format and size and those of a conventional TV camera, and based on the special optical magnification ratio established for this test, there is a conversion factor of 1-to-4 which must be applied to the chart spatial frequencies to relate them to the EDITS vidicon camera photocathode. Thus, the camera is caused to scan a bar-pattern image group ranging from 25 TV lines to 250 lines resolution. Measurements were performed at a video S/N $\geq$ 40 db.

Test Results. Figure E.2 P-2 shows the resulting CRT single line amplitude traces for six-bit PCM and two-bit Delta modulation systems. The two photographs on the left are taken under the same test input conditions as the two traces on the right. The left traces are presented on an expanded CRT time base to

RE-ORDER No. 64-712
Vol I

permit easier interpretation of the response corresponding to the first four input bar pattern groups, which extend over a range of 25, 50, 75, and 100 TV lines. The time base on right photographs equal one full camera horizontal scan line period. In each case, only a single line through the bar pattern group is recorded.

As expected, the absolute and relative square wave responses from the two modulation systems are in substantial agreement with the previous sine wave results. The data shows that PCM has approximately 75% response at 100 TV lines (relative to 25 TV line amplitude), 50% at 150 TV lines, and 20% (in this particular case) at 200 TV lines. The results beyond 200 lines are not easily interpreted on these traces.

Delta starts out at 25 TV lines with approximately 75% of the amplitude level of PCM due to the time constant of the integrator circuit. Its relative response at 100 lines is down to about 25%, and at 200 lines it is down to only a few percent.

It is concluded from these measurements that PCM rendition of points, lines and edges will be superior to Delta performance. However, the degree of superiority can best be determined by studying the fidelity of the test photographs obtained in the subjective test series, which is covered in paragraph E.4

Gamma Characteristics

Test Procedure. The gamma measurement of primary concern here is the overall characteristic including the camera, TV display film, and recording system. In addition, the gamma of the vidicon camera is of interest, and this measurement also is included.

Common test input data is used in both these measurements. The basic input data is in the form of a Foto-Video Grey Scale test object consisting of ten steps of increasing transmission densities covering a relative density range from 0 to 1.7 in average step increments of 0.2 density. Since the step-to-step change is not uniform (varying from 0.14 to 0.27) it is necessary to measure the actual density values corresponding to each step. This is readily accomplished

RE-ORDER No. 64-712
Vol I

using a simple Kodak transmission/reflectance densitometer. Then the system output response is related to the corresponding calibrated step densities (converted to relative percent of transmission for graphical plotting purposes).

In the case of the camera gamma measurement, the camera amplitude response is recorded on the CRT connected to the EDITS display amplifier output (it was determined that the amplitude linearity of the display amplifier was sufficiently good and did nc affect the measurement accuracy). By means of video gating circuits incorporated in EDITS, the total frame of video is gated out, with the exception of a small line sector falling in the central area of the camera scan. Then the gray scale chart is positioned to cause the photocathode image of the first gray scale step to correspond with the active portion of the camera scan line. The signal output amplitude is then recorded on the CRT. The gray scale image is then moved one step interval, the CRT horizontal centering is moved one centimeter division, and a second film exposure is made. This process is repeated until all ten steps (ten, multiple film exposures) are completed. The resultant step waveform pattern represents the relative amplitude response of the vidicon camera to the corresponding input step brightness values. This particular procedure is followed to avoid any errors resulting from uneven sensitivity across the vidicon photocathode surface or uneven shading in the back-illumination light fixture.

A reasonable approximation of the overall gamma of the complete system is obtained by photographing the display output produced as the camera scans the same gray scale chart. Typical values of display drive voltage and display brightness are employed in this test. The output photographic reflectance density values are then measured using the Kodak densitometer, and relative percent reflectance values are then calculated for purposes of graphical presentation.

Test Results. Figure E.2-2 contains plots of the relative camera and overall system responses versus the corresponding relative gray scale step percent brightness values. The camera gamma is seen to be uniform and is calculated to have a value of 0.7, approximately.

RE-ORDER No. 64-712
Vol I

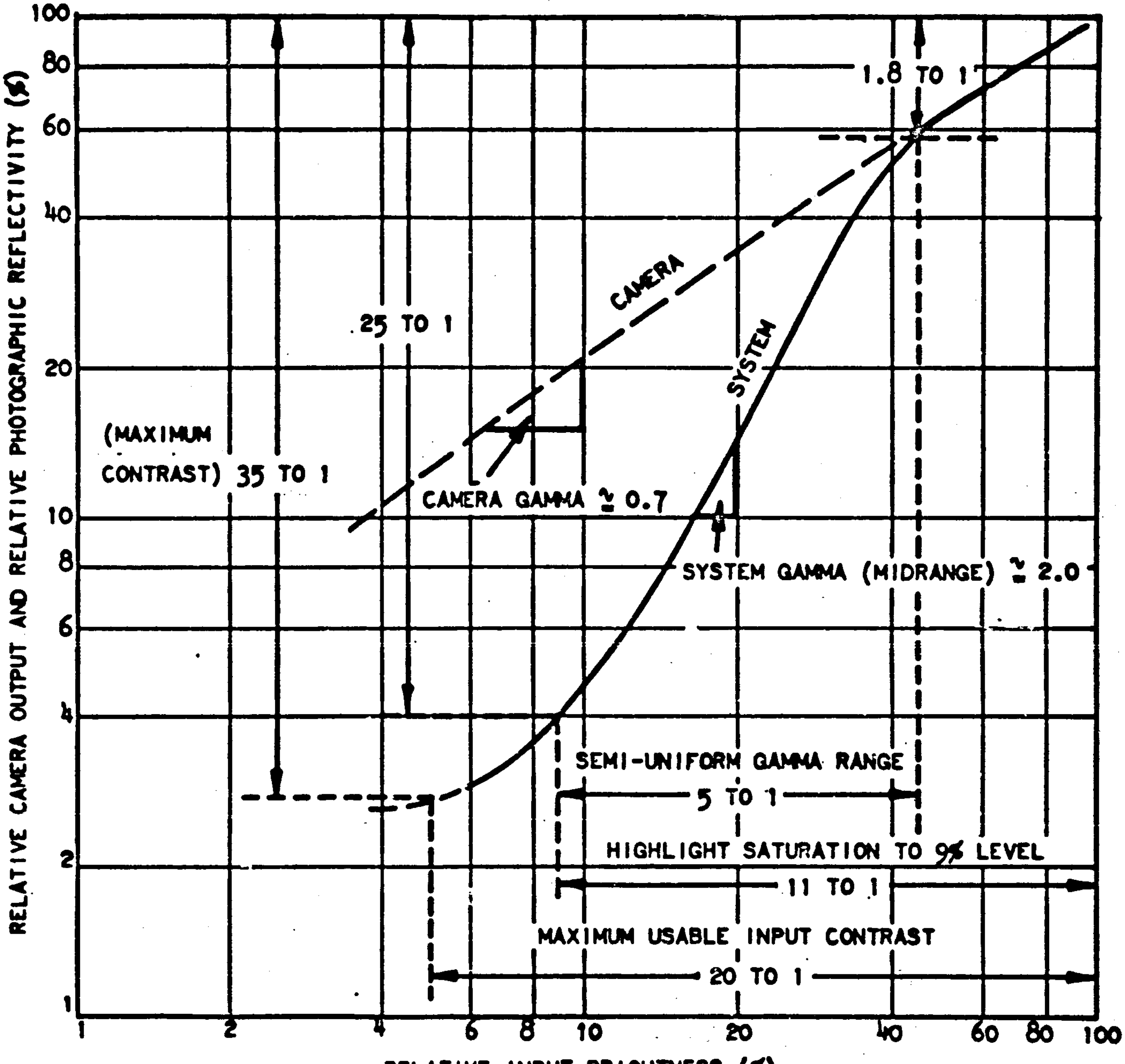

FIGURE E.2-2

TV CAMERA AND SYSTEM GAMMA

07314

RE-ORDER No. 64-712
Vol I

The calculated midrange gamma for the complete system is approximately 2.0; however, highlight and lowlight saturation effects, primarily in the Polaroid film, limit the overall output contrast to about 35-to-1. This corresponds to an input contrast range of 20-to-1. However, if system operation were limited to a fairly straight portion of the gamma characteristic corresponding to a contrast range of about 14-to-1, $\left(\frac{25 \text{ to } 1}{1.8 \text{ to } 1}\right)$ the usable input contrast would be restricted to a ratio of 5-to-1. In the more normal case, the system is adjusted to reproduce scene highlights at the maximum photographic exposure permissible without significant loss of highlight detail. Under this condition, the maximum usable input contrast equals approximately 11-to-1 and this produces an output photographic contrast of 25-to-1.

The above figures represent various dynamic operating ranges obtainable with EDITS under the photometric conditions specified.

E.3 QUASI-OBJECTIVE MEASUREMENTS

E.3.1 Introduction

Paragraph C.2 presents a summary of the types of tests and test conditions employed in this measurement series. In summary, the tests include: (1) overall image fidelity (RETMA test chart), (2) limiting sine wave resolution, (3) limiting square wave resolution, and (4) brightness contouring. All eight video data modulation techniques listed in paragraph C.2 are included in this test phase.

The principal purpose of the quasi-objective measurements is to provide a means of bridging the gap which would otherwise exist between the purely objective (waveform) test results previously discussed and the primarily subjective test evaluations covered in paragraph E.4. By employing known or calibrated input test conditions, various system parameters can be tested up to and beyond the limits of acceptability as determined by subjective assessment of the output photographic data.

Because of the large number of photographic recordings contained in this section, no attempt is made to comment specifically on every photograph. Rather, discussion in general will be limited to items on unusual interest and importance.

RE-ORDER No. 64-712
Vol I

E. 3. 2 Discussion of Quasi-Objective Test Procedures and Results

Overall Image Fidelity (RETMA Chart Input)

Test Procedure. The EDITS camera system is set up using the 25 mm lens to view the proper RETMA test chart dimensions in order to provide for direct reading of TV line resolution (as discussed in paragraph E. 1. 3). Electrical and optical focus are optimized, a standard video gain setting is selected, and the camera lens aperture is adjusted to achieve a full video amplitude level into the A/D encoder. A standard video gain setting is maintained for an entire series of measurements in order that the external analog noise generator output can be calibrated for the desired video S/N conditions required in the tests.

TV display grid drive and display brightness settings are carefully adjusted and controlled to achieve, insofar as is possible, similar operating conditions for each measurement condition which follows. The film camera aperture setting remains fixed for all measurements, as does the exposure time, since in each case this equals one complete frame time.

Test Results. Referring to Figures E. 3. P-1 through E. 3. P-4, as would be expected the reference 6-bit PCM technique produces the highest fidelity test chart image at 40 db signal-to-noise. A limiting resolution of approximately 250 TV lines is discernible in the central chart area (see paragraph E. 1. 3 for a discussion of this resolution enhancement effect). Seven gray-scale steps are in evidence, and the overall photographic image is clean and reasonably sharp. The most pronounced departures from this level of fidelity occur with 2 bit PCM (as would be expected on this type of image), on two-three Roberts which shows a definitely objectionable residual pseudorandom noise effect, and on 2-bit Delta. The Delta image possesses an overall "soft" appearance resulting from the limited step function response of the system. Horizontal resolution of 200 TV lines (maximum) is barely discernible in the center of the chart. Vertical resolution, determined by scan-line spacing, appears comparable to

22

RE-ORDER No. 64-712
Vol I

PCM response. Added to the Delta picture softness is a background graininess or coarseness caused by the fluctuations of the Delta signal above and below the actual levels of low frequency input video signal components. The 4-bit PEC/Huffman system also exhibits a raster graininess. This effect has been studied and is known to result from crosstalk effects in the present TV display scan system employed with the PEC system. Similar effects were originally observed in the EDITS camera and were removed by circuit design refinements. Therefore, the observed graininess is not inherent in the technique, rather, it is a present circuit limitation which can be overcome by redesign.

The 20 db signal-to-noise condition is observed to be something of a performance equalizer since a considerable amount of the fine-line and point detail present in the higher fidelity, low noise PCM photos is now obliterated by the high noise content of the signal. Still, however, the Delta image exhibits a softness not present with the other techniques. Three-four Roberts, 4-bit PCM and 6-bit PCM appear to have approximately equal image fidelities.

<u>Limiting Sine Wave Resolution</u>

<u>Test Results.</u> The test procedure is described in paragraph E.2.2. Referring to Figure E.2.P-1, 6-bit PCM and 2-bit Delta sine wave resolution patterns are presented adjacent to the corresponding CRT amplitude traces previously discussed. The subjective evaluation of the resulting sinusoidal images substantiates the objective measurement discussion. The 6-cycle/mm pattern is reproduced with reasonable clarity by 6-bit PCM; the Delta reproduction is approaching the limiting resolution condition.

<u>Limiting Square Wave Resolution</u>

<u>Test Procedure.</u> The procedure followed here is essentially the same employed with the RETMA chart fidelity measurement discussed above.

<u>Test Results.</u> Referring to Figures E.3.P-1 through E.3.P-4, the square wave resolution on all modulation systems except Delta is clearly 200 TV

RE-ORDER No. 64-712
Vol I

lines, and actually appears to extend nearly to 300 TV lines (limiting) on 6 bit and 4 bit PCM and three-four Roberts. Delta produces a clear 100-line response and an absolute limiting response at 200 lines.

Brightness Contouring

Test Procedure. In order to bring out the effects of brightness contouring resulting from signal quantizing, a roughly symmetrical input brightness gradient is produced by a 75-watt incandescent lamp placed several inches behind the glass diffusing plate in the EDITS back-illuminated test fixture. This creates the hot-spot area seen on the 40-db 6-bit PCM photo on Figure E. 3. P-5. The corresponding camera video signal level is adjusted to cover the full encoding range of the A/D converter.

Test Results. Figures E. 3. P-5 through E. 3. P-8 show the contouring effects of the various modulation techniques at 40, 30, and in certain cases, 20 db signal-to-noise. It is significant that the three-four Roberts modulation incorporating pseudorandom noise addition and subtraction does a very effective job of eliminating contouring, even a 40 db. All other techniques exhibit varying degrees of countouring, with the exception of 6-bit PCM. The 40-db Delta photo exhibits an unusual effect consisisting of two concentric rings corresponding to dead bands in the modulation process. This is not thought to be inherent in the system, but rather is a small operational defect in the present Delta implementation.

The 30 db and 20 db signal-to-noise conditions serve to illustrate the effectiveness of sensor noise in eliminating contouring on the more coarsely quantized signals.

E. 4 SUBJECTIVE MEASUREMENTS

E. 4. 1 Introduction

Paragraph C. 2 presents the summary of various test conditions employed for this test phase in comparing the performance of the eight modulation techniques under review. In all, 96 photographs are included in this section and are divided equally between the two selected natural terrain input test photographs; specifically the Canadian Arctic and Lunar scenes.

RE-ORDER No. 64-712
Vol I

The main objective of this last major measurement phase is to produce a series of carefully controlled photographs suitable for detailed subjective study, which represent typical image-fidelity characteristics expected of the eight modulation methods operating on an interplanetary reconnaissance mission. "Typical" image fidelity is the difficult part of this objective to satisfy since it implies detailed knowledge of the planetary terrain--information which of course is unavailable at present--and many other specific mission operational characteristics. EMR has attempted to satisfy the objective by selecting input test photographs representing wide extremes in terms of terrain types, scale factors, and photometric contrasts. We have endeavored to simulate other critical operational conditions by specifying a wide range of test parameters; i.e., video signal-to-noise of 40, 30, and 20 db and bit error probabilities of $10^{-\infty}$, 10^{-4}, 10^{-3}, and 10^{-2}.

Due to the very large number of photographs contained within this section, no attempt will be made to discuss each one. Rather, comments will be limited to items of special significance, and will serve to supplement discussions of the individual modulation techniques presented in Appendix A.

E. 4. 2 Discussion of Subjective Test Procedures and Results

Overall Image Fidelity - Canadian Arctic Photo

Test Procedure. The general equipment set-up procedure closely follows that outlined in paragraph E. 3. 2 for the RETMA chart fidelity measurement. The only significant differences are in optical magnification ratios employed plus the additional care being taken to obtain the maximum possible dynamic range from the system.

To produce the desired simulated photographic scale factors, the camera is positioned to scan a seven inch square area of the photograph. Critical adjustments of vidicon beam current and lens aperture are required to reproduce detail appearing in the highlight and lowlight areas. Even at best, this is a compromise since the maximum usable input contrast range for the system is 20-to-1, while maximum contrast ratios as high as 50-to-1 are present in the input photograph.

Test Results. Referring to Figures E. 4. P-1 through E. 4. P-4, the reference

RE-ORDER No. 64-712
Vol I

a 6 bit PCM photo (40 db, $10^{-\infty}$),again represents the fidelity standard for the group. It is important to note that the limited input dynamic range of the system very nearly results in a total loss of lowlight detail in coordinate locations B-9, B-10, and C-8 through C-10 (Figure C.2 P-7). Highlight detail is much better preserved due to the camera set-up procedure used, although some highlight saturation is present.

Still referring to the 6-bit PCM photograph, midrange tonal elements are reproduced with reasonable fidelity; that is, lines, edges, tonal shades, and textural features. Of course, due to the midrange system gamma of approximately 2.0, there is a noticable increase in contrast in these areas.

In terms of relative image fidelity, at 40 db major degradation appears only with the 3-bit and 2-bit PCM and the two-three Roberts methods. Delta, although lacking the subjective sharpness of the PCM picture, produces a surprising amount of image detail, which emphasizes the fact that standard test pattern image fidelity cannot be used as the sole measure of system usefulness.

Another interesting point here concerns the three-four Roberts system. Except for a moderate increase in fine-grain noise, the three-four Roberts fidelity at 40 db is not far below 6-bit PCM. Comparison of the 6-bit PCM at 30 db versus the three-four Roberts at 40 db shows nearly equal image fidelities.

All systems appear to hold up pretty well at the 30 db condition, and at 20 db the differences between 6-bit PCM, 4-bit PCM, and three-four Roberts are slight, with 2-bit Delta quality not far below other three.

Figures E.4.P-5 through E.4.P-8 present the effects of simulated channel noise on the output bit streams. The most significant results of this test series are: (1) all PCM techniques hold up quite well at increasing P_e. Even at $P_e = 10^{-2}$, little picture information is lost as a result of the "salt and pepper" error effects; (2) Delta holds up surprisingly well

39

RE-ORDER No. 64-712
Vol I

with only minor image information lost at $P_e = 10^{-2}$; (3) four-bit PEC/-Huffman exhibits increasingly severe data word errors[1] until at $P_e = 10^{-2}$, very severe image degradation is evidenced.

Overall Image Fidelity - Lunar Scene

Test Procedure. Since the photometric contrast ratio of the Lunar scene is only 11-to-1, camera set-up entails less of a compromise than with the Canadian Arctic set-up discussed in above. Other comments included above also pertain to this test procedure.

Test Results. Figures E. 4. P-9 through E. 4. P-12 depict the modulation system fidelities resulting from a variation of analog signal-to-noise. First, however, a brief comparison will be made between the reference 6-bit PCM (40 db, $10^{-\infty}$) photo and the input Lunar scene. The much smaller contrast range of the Lunar scene has permitted a camera/display system set-up in which the lowlight and midtone range area detail (for example, see coordinate locations B-2, B-3, C-2, C-3 of Figure C. 2P-8) is reproduced with reasonable fidelity. This includes lines, edges, textural features and, in particular, isolated points. Detail is lost in the highlight areas (for example coordinate location G-6 on the same figure) however, due to the saturation effect previously discussed.

From a relative fidelity comparison standpoint, the 40 db condition shows 4-bit PCM, three-four Roberts, and 2-bit Delta of almost equivalent quality below 6-bit PCM, with Delta suffering slightly more on reproduction of isolated point images. Decreasing video signal-to-noise to 30 db brings these four techniques into closer equivalence, as does the 20-db condition.

Figures E. 4. P-13 through E. 4. P-16 show the effects of increasing bit errors. The comments made on the corresponding series of Canadian Arctic photographs apply here also. The only exception is that some of the PCM "salt and pepper" errors conceivably could be mistaken for isolated point images.

(1) Wired line and frame scan synchronization is employed in this and all other performance tests reported herein.

240