

FINAL REPORT

MANNED SPACECRAFT ADVANCED TELEVISION STUDY

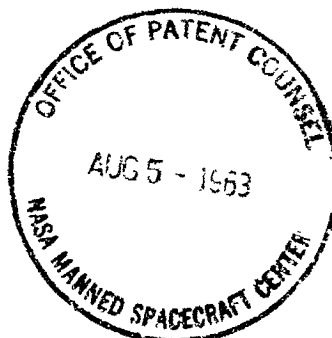
FACILITY FORM 802	N65-24844	(ACCESSION NUMBER)	(THRU)
	399	(PAGES)	/
	CR 65012	(NASA CR OF TMA OR AD NUMBER)	07
		(CATEGORY)	

GPO PRICE \$ _____

CTS PRICE(S) \$ _____

Hard copy (HC) 7.00

Microfiche (MF) 2.00



FOR:

NASA Manned Spacecraft Center,

Houston, Texas,

Contract NAS 2-991

31 July 1963

SARASOTA DIVISION

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Page	Location	Change From:	To
2-2	Table 2.0-1, first column, sixth item	(MSPS)	(MBPS)
2-2	Table 2.0-1, first column, eighth item	De^h	P_e^b
2-2	Table 2.0-1, second column, last item	1	0
2-3	Paragraph 2.1.1.1, line 9	original	original
2-3	Paragraph 2.1.1.1, line 13	with	either
2-3	Equation	$P_{y,i} (y+1), j \quad x=k$	$P_{y,i} [(y+1), j] \quad x=k$
2-4	Line 8	$P_{x,y,f,i} (x,y), (f+1), j$	$P_{x,y,f,i} [(x,y), (f+1), j]$
2-4	Line 16	Robert's	Roberts'
2-6	Table 2.1.2-1, third column, 11th item	Girl	MIT girl
2-6	Paragraph 2.1.3, line 3	... parameters $P_i, i = k \dots$	... parameters $P_i, i \neq k \dots$
2-7	Line 20	... is closer.	... is closest.

Page	Location	Change From:	To
2-9	Line 16	...in Section 6.2.	...in Section 6.2.5.
2-9	Table 2.1.3-2, first column, 2nd set of items	6 4 3	6 Q 4 3
2-10	Line 14	...are not including...	...have not included...
2-11	Line 6	...through 2.6.	...through 2.5.
2-21	Paragraph 2.2.1, line 1	...source can use...	...source use...
2-22	Equation 12	$C_f...$	$\bar{C}_f...$
2-23	Line 5	...minimum PE $\neq 0.833...$	...minimum PE = 0.833...
2-23	Line 12	...levels $N/2...$	...levels $n/2...$
2-23	Line 13	... $N/2 + 1$ are...	... $\frac{n}{2} + 1$ are...
2-24	Table 2.2.1-1, second column, 2nd item	0.0005	0.00015
2-24	Lines 3 and 4	The appropriate Huffman word to the PE and (1-PE) signals.	Delete
2-26	Bottom of page	...	Add: The joint probabilities can now be defined in terms of the first order and transition probabilities, that is, $P(i, j) = P(i)P_j(j)$.
2-29	Equation 33	$C = \sum_{y=1}^{\infty} \dots$	$C = \sum_{y=1}^{\infty} \dots$

Page	Location	Change From:	To
2-29	Equation 35	$\bar{C} = \sum_{y=1}^{\infty} \dots$	$\bar{C} = \sum_{y=1}^{\infty} \dots$
2-31	Equation 43	$\dots (1 - P_{bb}^l) \dots (1 - P_{ww}^l) \dots$	$\dots (1 - P_{bb}^l) \dots (1 - P_{ww}^l) \dots$
2-32	Equations 45, 46, and 47	$\dots (2 - P_{bb}^l - P_{ww}^l) \dots$	$\dots (2 - P_{bb}^l - P_{ww}^l) \dots$
2-36	Equation 60	$\dots a_{ll}^P \dots$ $l \neq j$	$\dots a_{ll}^P \dots$ $l \neq j$
2-37	Equation 61	$\dots a_{ll}^P \dots$ $l \neq j$	$\dots a_{ll}^P \dots$ $l \neq j$
2-38	Line 10	$\dots y-1 \text{ PE} \dots$	$\dots (y-1) \text{ PE} \dots$
2-38	Lines 11 and 13	$\dots y-1 \dots$	$\dots (y-1) \dots$
2-38	Line 26	$\dots \sum_{j=1}^n P(j)^l \dots$	$\dots \sum_{j=1}^n P(j)^l \dots$
2-39	Line 1	$\dots \text{PE}^l \dots$	$\dots \text{PE}^l \dots$
2-39	Line 2	$\dots \text{where } P(j)^l$	$\dots \text{where } P(j)^l$

Page	Location	Change From:	To
2-39	Equation 70	$\left\{ \sum_{j=1}^n \frac{P(j)^1 + \left[1 - \sum_{j=1}^n P(j) \right]^1 \overline{PERL}}{H \left[P(R_{PE}^y) + \sum_{j=1}^n H \left[P(j) \right] \right]} \right\}$	$\left\{ \sum_{j=1}^n \frac{P(j)' + \left[1 - \sum_{j=1}^n P(j)' \right] \overline{PERL}}{H \left[P(R_{PE}^y) + \sum_{j=1}^n H \left[P(j) \right] \right]} \right\}$
2-39	Equation 71	 $P(j)^1 \dots P(j)^1 \dots$	 $P(j)' \dots P(j)' \dots$
2-39	Line 15	 the $P(j)^1$ and $1 - P(j)^1 \dots$	 the $P(j)'$ and $1 - P(j)' \dots$
2-40	Equations 74 and 75	$P(j)$ (all cases)	$P(j)'$ (all cases)
2-40	Equation 76	$P \left[\sum_{j=1}^n P(j) \right]$	$P \left[\sum_{j=1}^n P(j)' \right]$
2-40	Equation 77	$\bar{C}_{PERL} = \frac{\log_2 n}{H(\overline{PERL})} \frac{\overline{PERL}}{(1-PE) \overline{PERL} + PE}$	$\bar{C}_{PERL} = \frac{\left[\log_2 n \right] \overline{PERL}}{H(\overline{PERL}) \left[(1-PE) \overline{PERL} + PE \right]}$
...	Figure 2.2.3-2, third box from left, top line	$(z-)^2$	$(z-\alpha)^2$
2-44	Line 18	 would defect...	 would defeat...

Page	Located	Change From:	To
2-62	Paragraph 2.3, line 23	$H = \sum_{i=1}^h \dots$	$H = \sum_{i=1}^n \dots$
3-2	Equation 2	$P_r = \frac{P_t^1 G_1 G_2}{P \ell}$	$P_r = \frac{P_t^1 G_1 G_2}{P \ell}$
3-2	Equation 3	$P_t^1 = \frac{P_t}{L} \text{ watts,}$	$P_t^1 = \frac{P_t}{L} \text{ watts,}$
3-2	Equation 4	$P_r = \frac{P_t G_1 G_2}{L P \ell}$	$P_r = \frac{P_t G_1 G_2}{L P \ell}$
3-10	Paragraph 3.4, eighth item	P_i	$P \ell$
3-10	Paragraph 3.4, 13th item	P_b^e	P_b^b P_e
4-2	Paragraph 4.2, third line	The basic recorder...	The basic recorder...
4-5	Paragraph 4.4, line 16	...80,000 (w/1 day)...	...80,000 seconds (w/1 day)...
5-5	Paragraph 5.3, line 3	...from the A to D converter...	...from the Roberts' decoder...

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SECTION 1

DESIGN TRADE-OFF COMPARISON

1.0 EXECUTIVE SUMMARY

During the course of the theoretical and experimental portions of this study program (which are described elsewhere in this report) various digital television compression techniques were evaluated. As determinations were made of the most suitable techniques to be combined into the final compressed digital TV system, design studies were initiated to arrive at estimates of reliability, size, weight, and input power for a prototype equipment version of the spacecraft portion of this system. To provide the necessary input data for these estimates, digital logic and analog functional block diagrams were conceived, preliminary circuit designs were established, and component types and quantities were estimated.

To provide standards of comparison for the compressed digital TV equipment, similar design studies were conducted on noncoded analog TV, straight PCM TV (four-bit encoding) and standard analog TV (analog TV camera with six-bit encoding of the camera video signal and with PCM video output).

Two basic hardware implementation techniques were considered. These were: (1) the exclusive use of miniature discrete parts (all solid state except for the vidicon camera tube), and (2) the use of semiconductor integrated circuits combined with discrete parts where necessary, thus termed hybrid SCIC circuits. In total, eight separate equipment configurations were studied.

Based on the parts (and SCIC circuit) counts obtained on each configuration, and on the latest available part and circuit failure rate data, reliability analyses were performed to obtain the estimated mean-time-between-failure (MTBF) for each spacecraft equipment design. MTBF's based on mid-1963 and projected 1964 failure-rate estimates were obtained.

Table 1-1 summarizes the results of these studies. The comparison of the compressed digital TV estimates with those of the standard analog TV equipment are of particular interest and importance, since the video data compression ratios achievable, typically three to one, are measured relative to standard analog TV (see Section 2 for a detailed discussion of data compression). Note that the compressed digital TV system achieves this information-bandwidth reduction, with the attendant savings in video transmitter power and weight, and still permits reductions in spacecraft TV equipment size, weight, and input power relative to the standard analog TV design.

Table 1-1 also contains columns of data on a "Straight PCM TV" and on a "Non-coded Analog TV." Although these systems appear to offer terminal equipment

power and reliability advantages relative to "Compressed Digital TV" and to "Standard Analog TV," they would both require more transmitter power and weight than the "Compressed Digital TV." The "Straight FCM TV" utilizes a parametric value of four bits per picture element instead of the approximately two bits per picture element of the "Compressed Digital TV." Therefore, it will require approximately twice the bandwidth and twice the transmitter power. The "Noncoded Analog TV" will require some analog form of modulation (otherwise it becomes the "Standard Analog TV" system.) This analog form of modulation will probably be some variant of the FM technique such as frequency-compressive-feedback FM or straight FM utilizing a phase-locked loop as a threshold-improving receiver. The efficiency of these analog modulation systems in power and bandwidth requirements relative to optimum PCM techniques for transmission of a given quality video signal has not been experimentally determined as yet. Theoretically, *orthogonal systems such as Digi-lock, coherent and possibly noncoherent phase modulation require 5 to 10 db less signal-to-noise ratio (and hence less transmitter power) for a given transmission bandwidth-to-bit rate ratio than do the FM, FM phase-locked, and Enloe FM feedback systems. Therefore, it can be assumed, until proven otherwise, that the apparent advantage of the "Noncoded Analog TV" shown in Table 1-1 is largely, if not completely, nullified by the increased transmitter requirements.

Table 1-2 presents estimated savings in spacecraft prime power and weight** achieved by using the compressed digital TV system in place of the standard analog TV system under the conditions specified in Appendix II of the RFQ.

1.1.0 Compressed Digital Television System Design

1.1.1 Spacecraft Equipment

1.1.1.1 Summary of Equipment Operation

An over-all diagram of the compressed digital spacecraft equipment is shown in Figure 1.1-1. Summaries of the principal functions of each system unit are presented below.

*Keenan, R. K.; Mannex, H. R.: "Spacecraft Communications Systems," Space/Aeronautics, Part 2, Aerospace Electronics, November, 1962.

** The above estimates of spacecraft weight savings are conservative since the reduced prime power requirements were not considered as contributing to reductions in the weight of the prime power source. Presently there are insufficient data available on specific mission profiles, video equipment operating duty cycles, and spacecraft prime power source characteristics, to permit meaningful estimates to be made in this area.

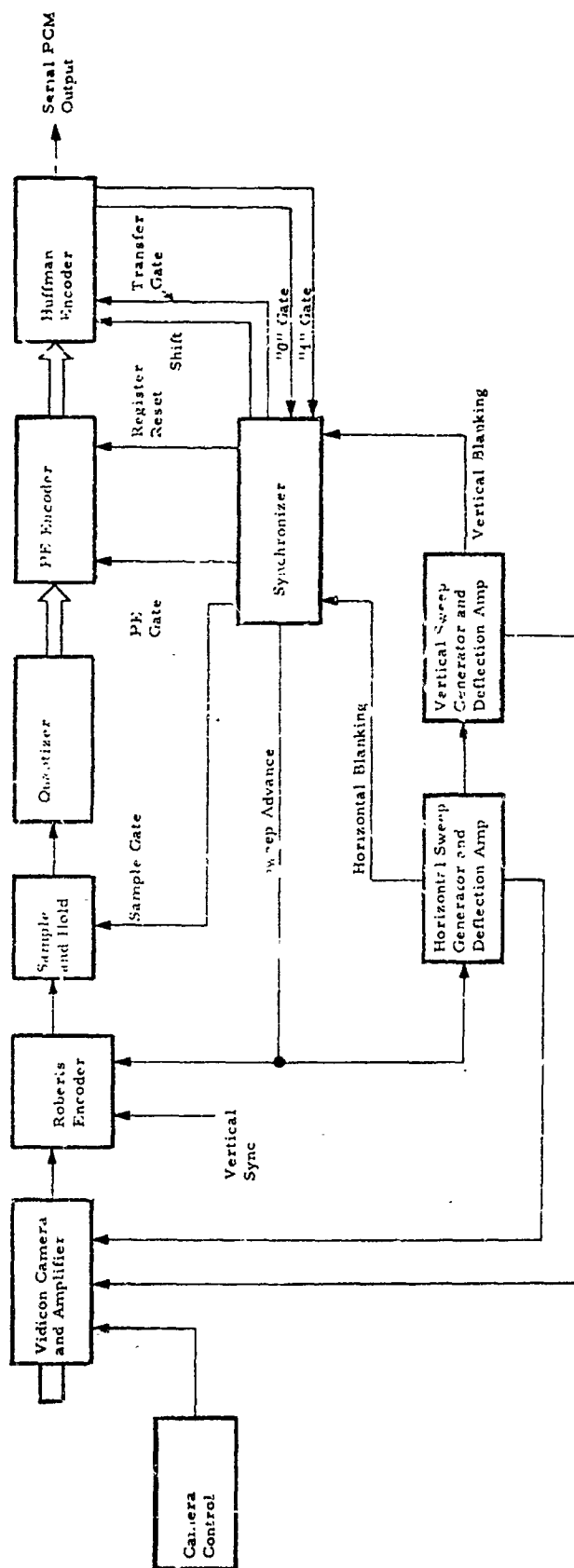


FIGURE 4.4-1 COMPRESSED DIGITAL TV SPACECRAFT EQUIPMENT

TABLE 1-1 SUMMARY OF PROTOTYPE SPACECRAFT EQUIPMENT DESIGN STUDY ESTIMATES

PARAMETER	COMPRESSED DIGITAL TV		STRAIGHT PCM TV		NONCODED ANALOG TV		STANDARD ANALOG TV	
	Hybrid SCIC	Discrete Parts	Hybrid SCIC	Discrete Parts	Hybrid SCIC	Discrete Parts	Hybrid SCIC	Discrete Parts
Power (Watts)	34.0	26.5	36.0	45.8	18.4	17.0	50.0	67.6
Volume (Cu. In.)	81.4	107.5	77.8	126.4	51.7	63.1	134.5	190.6
Nonstructural Weight (Lb.)	2.92	3.93	2.81	4.34	1.93	2.29	4.83	6.73
Reliability (Mid-1963) MTBF (Hours)	11,800	45,500	16,200	45,500	37,800	104,200	12,300	31,000
Reliability (Mid-1964) MTBF (Hours)	54,100	89,000	71,100	86,500	153,400	191,100	41,900	60,500

NOTE: These figures are for a particular set of design criteria, such as 10^6 picture elements per second, and should not be quoted out of context for systems of either higher or lower speed capability.

TABLE 1-2 ESTIMATED SAVINGS IN SPACECRAFT PRIME POWER AND WEIGHT

Condition (in RFQ)	Range (statute miles)	Prime power savings (watts)	Spacecraft weight saving (lb)
1	250,000	130	8
2	53,000,000	2,040	28

1. TV Camera

- a. Generates the basic analog video signals.
- b. Shapes video frequency response characteristic and provides necessary signal amplification to permit coupling of the signal to the Roberts encoder.
- c. Re-establishes the DC components of video signal by a clamping technique.
- d. Incorporates automatic gain control (AGC) to maintain a uniform peak video output level with variations in scene brightness.

2. Camera Sweep Generators and Deflection Amplifiers

- a. Generates variable-length horizontal and modified-vertical stair-step scanning waveforms as a function of sweep-advance pulses from the system synchronizer.
- b. Furnishes line and frame sync (blanking) pulses to the synchronizer to achieve proper line and frame synchronization of the compressed video data.

3. System Synchronizer

- a. Serves as the timing reference for all time-dependent system functions.
- b. Controls the TV camera beam dwell time on each target element as a function of the "fullness" of the bit storage register in the Huffman encoder.

4. Roberts Encoder

- a. Generates a pseudo-random sequence of variable amplitudes occurring at the element scanning rate of the camera.
- b. Performs an analog addition of these levels to the basic video data (to "randomize" the transitions between video quantization levels).

5. Sample and Hold

Discretely samples the camera analog video signal level associated with each scanned element, and holds each sampled level constant during the video quantizing process.

6. Quantizer

Quantizes the video (plus pseudo-random noise signal) to the desired digital bit level.

7. PE Encoder

Converts the PCM signal from the quantizer unit into previous element (PE) and not-previous-element (1-PE) signals. These signals have the proper format for optimum encoding.

8. Huffman Encoder

- a. Optimally encodes the PE and (1-PE) signals to produce a serial Huffman pulse train of minimum constant bit rate for transmission to the ground station.
- b. Generates sweep advance sync signals as new data are required in the Huffman shift register.

1.1.1.2 System Design Concepts and Operating Parameters

One unique feature of the system is that an asynchronous digital camera scan is employed to facilitate Huffman encoding of the video data without employing a buffer store prior to transmission of the Huffman bit train. To accomplish this, the Huffman encoder generates control signals which are fed back to the synchronizer; these control signals allow the synchronizer to generate "sweep advance" signals only when new data are required by the Huffman coder.

The system is designed for a resolution of 256 x 256 elements and 3-bit quantization following Roberts encoding. Although the use of an actual Huffman code would depend on picture statistics, a "flat distribution" code is assumed in this system design for the sake of best performance in the face of unknown picture statistics. The assumed code is as follows:

Previous element	-1
Intensity level 0	0111
Intensity level 1	0110
Intensity level 2	0101
Intensity level 3	0100
Intensity level 4	0011
Intensity level 5	0010
Intensity level 6	0001
Intensity level 7	00001
Horizontal sync	000001
Vertical sync	000000

The constant output bit rate of the Huffman encoder is set by the synchronizer at two megabits per second. With a total of $256 \times 256 \approx 65,500$ elements per TV frame, and an average of two data bits required per element (based on the three-to-one compression ratio) the average TV frame rate approximates 15 frames per second.

1.1.1.3 TV Camera Operation

Figure 1.1-2 illustrates a two-feed-back-loop automatic gain control (AGC) camera unit employing a slow-scan type all-electrostatic vidicon tube. The two loops allow for a very wide signal dynamic range in conjunction with gain stability. A signal from the vidicon target is amplified in the low-noise video preamplifier. If the amplified signal is greater than the preset threshold of the high-level AGC threshold circuit, then the amplifiers and detector circuits in the high-level AGC circuits act to reduce the Loop 1 gain. From the video preamplifier, the signal is fed to the gain-controlled video amplifier where signal clamping, frequency shaping, and additional signal amplification are performed. The Loop 2 gain is controlled by the amplifiers and detector of the low-level AGC circuits. Thus, the video output signal is a function of both Loop 1 and Loop 2 gains. For small target signals, the output of the video preamplifier will be too small to pass through the high-level AGC threshold circuit. Thus, only Loop 2 will control the video output amplitude.

The camera control circuits contain circuits and adjustments for the several vidicon operating potentials including electrical focus, beam, and target.

1.1.1.4 Camera Sweep Generators and Deflection Amplifiers

Figure 1.1-3 is a block diagram of the camera sweep generators and deflection amplifiers. Sweep advance pulses from the system synchronizer are fed to the horizontal binary counter chain, which consists of eight binary counter circuits. The eight binary counter outputs are fed to the eight horizontal analog gates and to the horizontal sync generator. Also, the reference voltage is applied to the horizontal gate circuits. Each horizontal analog gate is associated with one input of the sweep ladder, and an analog gate either grounds or provides the reference voltage to its associated ladder input. Thus, as the binary counter chain counts down, the output of the ladder increases by equal increments in synchronization with the sweep advance pulses. Horizontal deflection amplifiers are required to convert the single-ended sweep ladder output to the push-pull high voltages required by the vidicon deflection plates. Essentially, the horizontal sync generator is an eight-input "OR" circuit with an inverted output which is used to sweep advance the vertical sweep and also to provide blanking to the system synchronizer.

Operation of the vertical sweep generation circuits is quite similar to those employed in horizontal sweep generation. One circuit addition is the scan

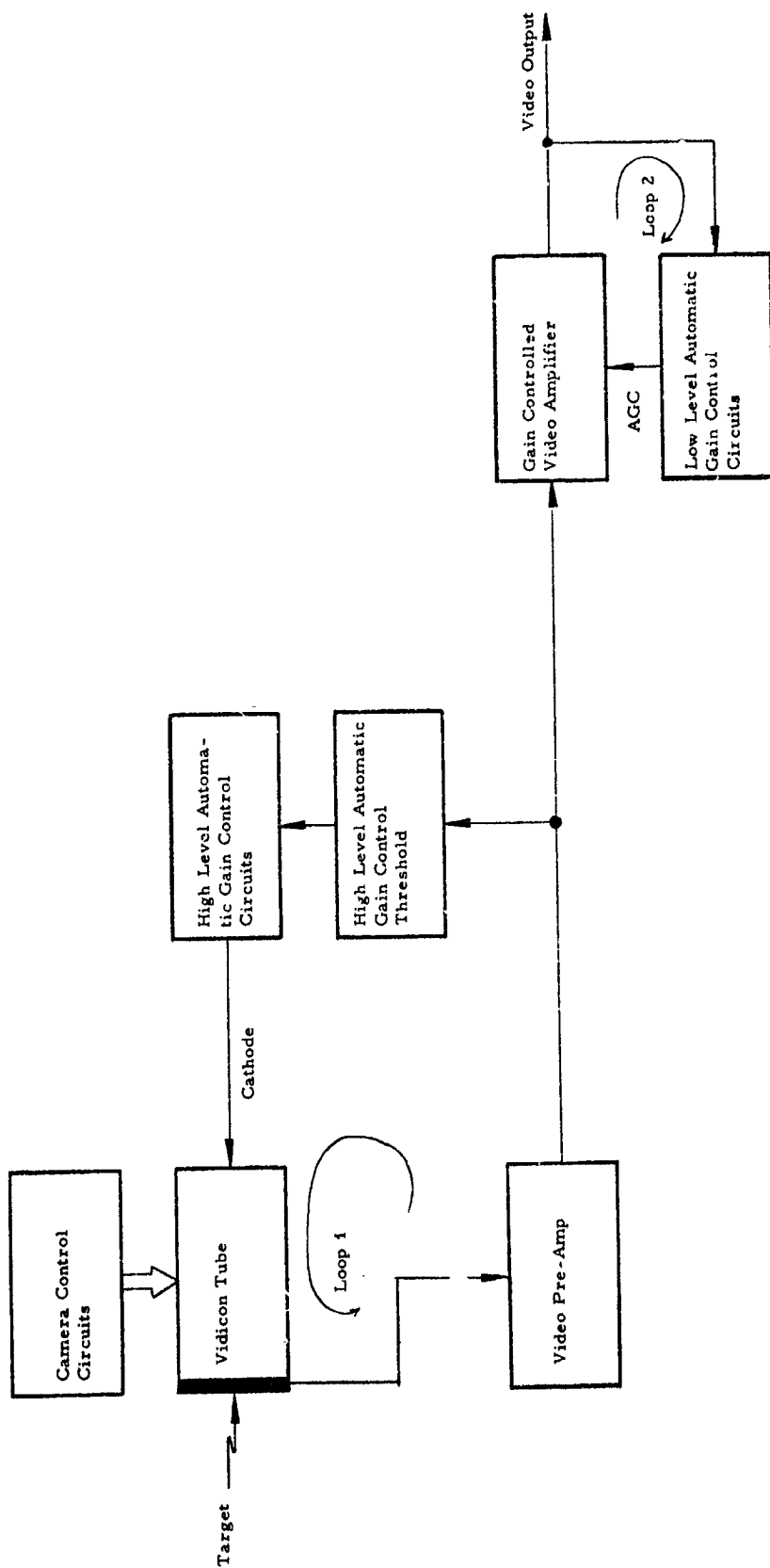


FIGURE 1.1-2 T. V. CAMERA BLOCK DIAGRAM

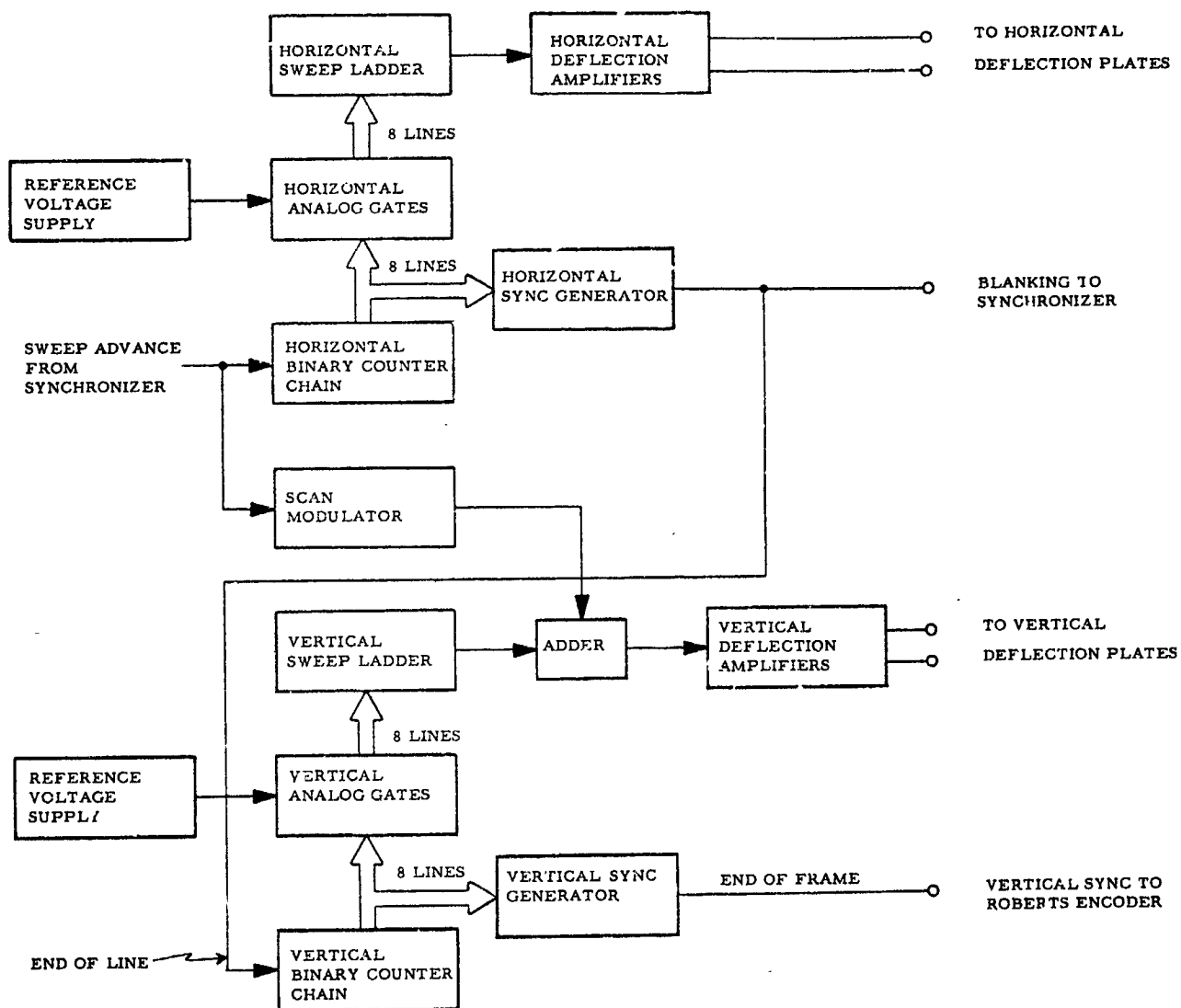


FIGURE 1.1-3 CAMERA SWEEP GENERATOR AND DEFLECTION AMPLIFIERS

modulator which generates pulses in synchronization with the sweep advance pulses from the system synchronizer. They are combined with the vertical sweep waveform in the adder circuit shown in front of the vertical deflection amplifiers. In conjunction with the sample and hold circuit, they permit uniform camera signal generation at varying scanning rates. A separate reference voltage supply is provided for the vertical analog gates so that interaction of the two sweeps will be kept to a minimum.

1.1.1.5 Synchronizer Operation

Figure 1.1-4 is a functional block diagram of the synchronizer unit. The 16-megacycle clock frequency is divided by eight in the three-stage binary counter circuit. The five following "AND" gates produce the pulse trains identified as B_0 , B_1 , B_5 , B_6 and B_7 ; the time relationships of these pulse trains are as shown in Figure 1.1-5. B_0 is applied as a shift pulse directly to the Huffman code shift register, thus establishing the PCM bit rate of two megabits per second. B_1 is utilized as the sweep advance pulse to the horizontal staircase generator; logic circuits are employed to inhibit this output except when new data are required by the Huffman coder. This condition is defined by the presence of a "0" gate from the coder or by a "1" gate from the coder during the sync code readout interval.

Horizontal and vertical sync code gates are created by setting flip-flop No. 1 with the "end of line" gate from the horizontal staircase generator and combining the flip-flop output with horizontal and vertical blanking gates from the sweep generators.

B_5 , B_6 and B_7 are combined to produce the "ADD" gate. This gate serves as the Huffman register transfer pulse after passing through logic gates which inhibit it except when the register is empty. B_0 is utilized as the sample gate and logic is included to place this output in proper sequence with the sweep advance pulse. B_0 is also used, in conjunction with the proper logic, as the PE gate, which is used to transfer video data to the PE encoder storage register.

1.1.1.6 Roberts Encoder Operation

The Roberts encoder is shown in Figure 1.1-6. This is an N-stage shift register which is reset by the vertical synchronizing pulse. The input to the register is determined by a modulo two adder connected to the appropriate stages. The resulting pseudo-random binary sequence is then decoded in the digital-to-analog ladder to generate a pseudo-random four-bit amplitude sequence that is added to the video signal before encoding.

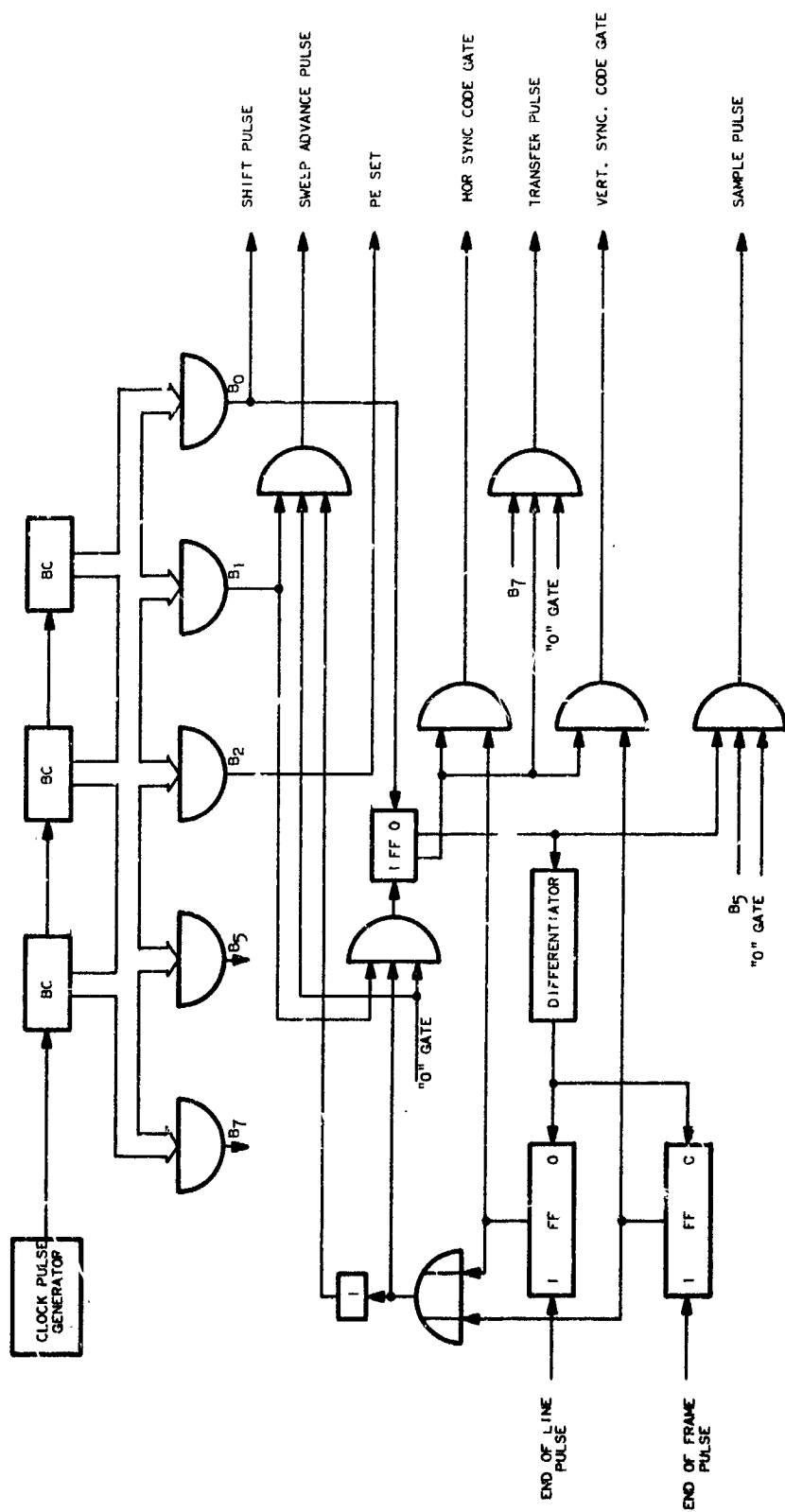


**FIGURE 1.1-4
SYNCHRONIZER**

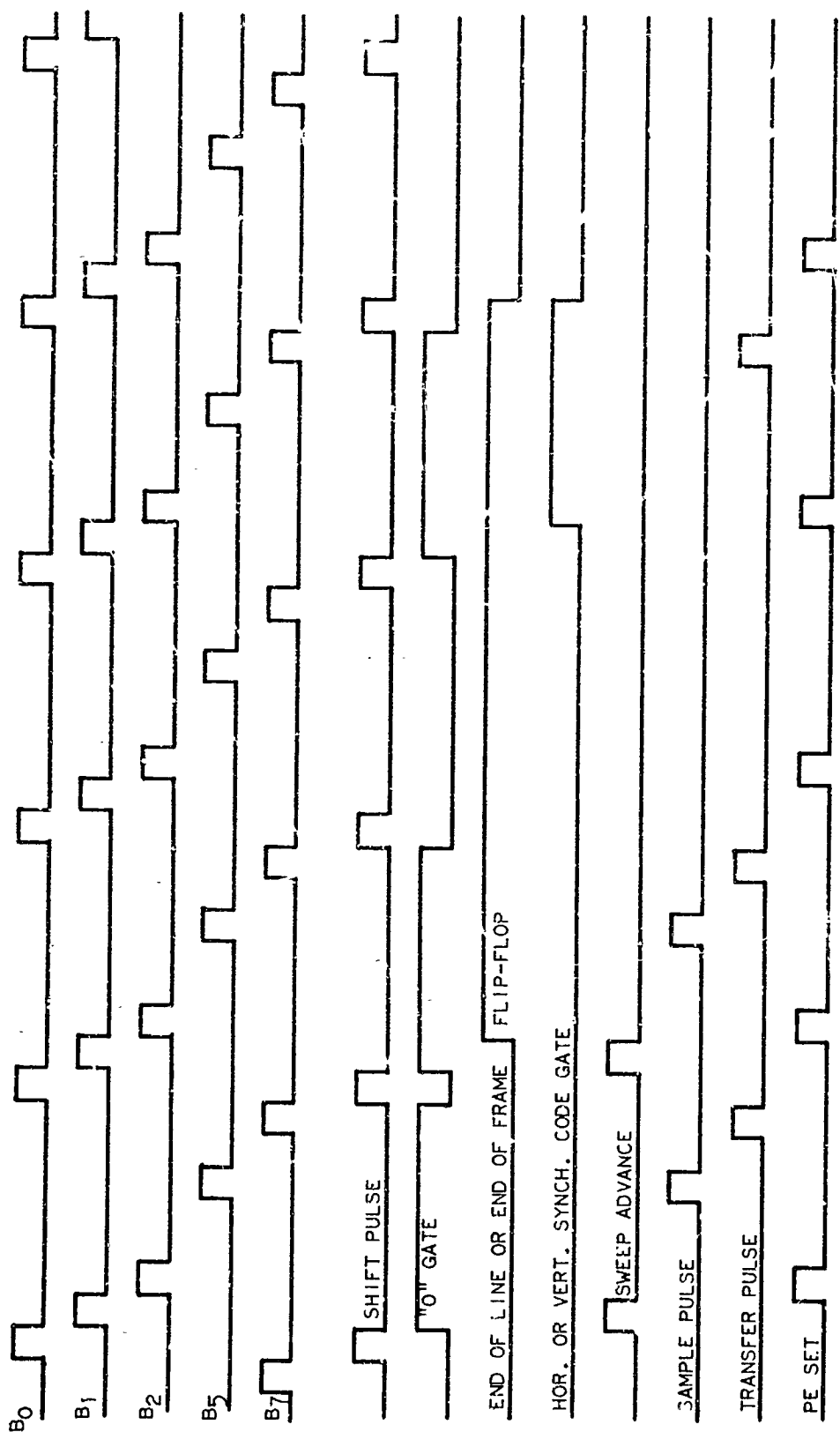


FIGURE 1.1-5
COMPRESSED DIGITAL TV WAVEFORMS

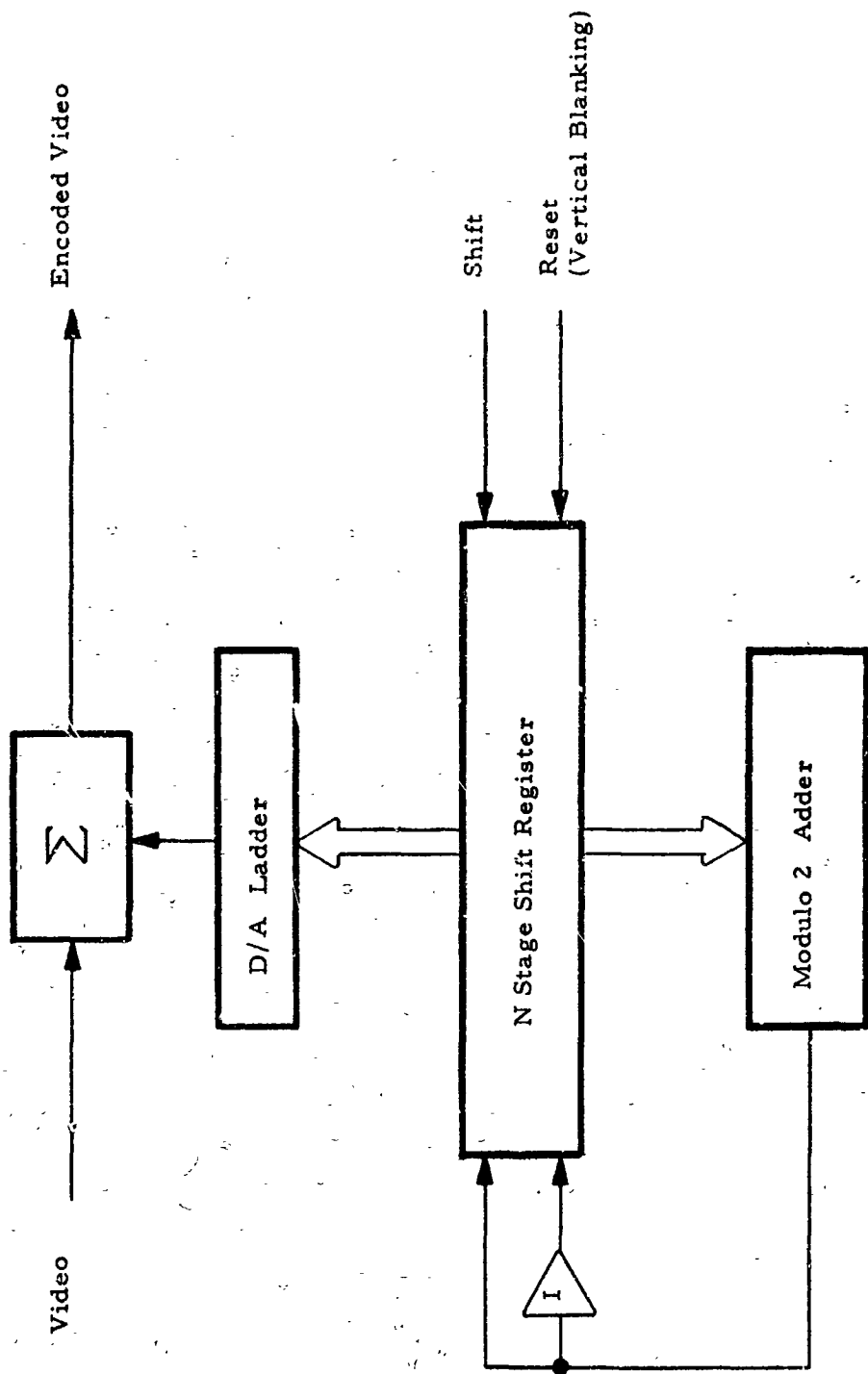


FIGURE 1.1-6 ROBERTS ENCODER

1.1.1.7 Sample and Hold

Figure 1.1-7 is a diagram of the Sample and Hold Circuitry. This circuit is necessary to prevent the video level from changing during the encoding time. During the application of the sample pulse, the sample gate acts as closed switch. This charges the holding capacitor to the video voltage present during the sampling time. At the end of the sample pulse, the sample gate opens. The holding capacitor voltage remains essentially unchanged, assuming a high input impedance in the amplifier stage, until the next sample pulse charges or discharges the holding capacitor to the new video level corresponding to the next scanned element.

1.1.1.8 Quantizer Operation

This unit (shown in Figure 1.1-8) utilizes seven parallel decision circuits to produce one "hot" wire out of eight; the hot wire indicates the intensity level of the signal from the sample and hold unit. The eight output wires are routed to the PE encoder.

1.1.1.9 PE Encoder Operation

The PE encoder is shown in Figure 1.1-9. A PE gate transfers the incoming hot line into the appropriate flip-flop in the eight stage register. When the next element is generated, it is compared with the stored element in the comparison gates. If they are equal, the output PE line is hot. If not, the appropriate level line is enabled by the (1-PE) gate. The register reset gate clears the store. The PE gate then reads in the present element and the cycle repeats.

1.1.1.10 Huffman Encoder Operation

This unit is shown in Figure 1.1-10. The "OR" gates at the input of the unit route the incoming hot line to the appropriate shift register stages. These interconnections are determined by the specific Huffman code used. The transfer gate shifts this parallel Huffman word into the register. Each Huffman word has been modified by the addition of a "1" to the end.

The Huffman word is shifted out serially by the shift pulses. Two "AND" gates are used to recognize the two states of interest of the shift register; namely, when there is one bit remaining and when there are no bits remaining. When all Huffman bits have been shifted out the additional "1" remains in the last register and all other registers contain "0". This condition is recognized by the "0" gate. The "1" gate recognizes when the shift register contains only one Huffman bit. This condition is identified by a "1" in the second stage and "0" in all higher stages.

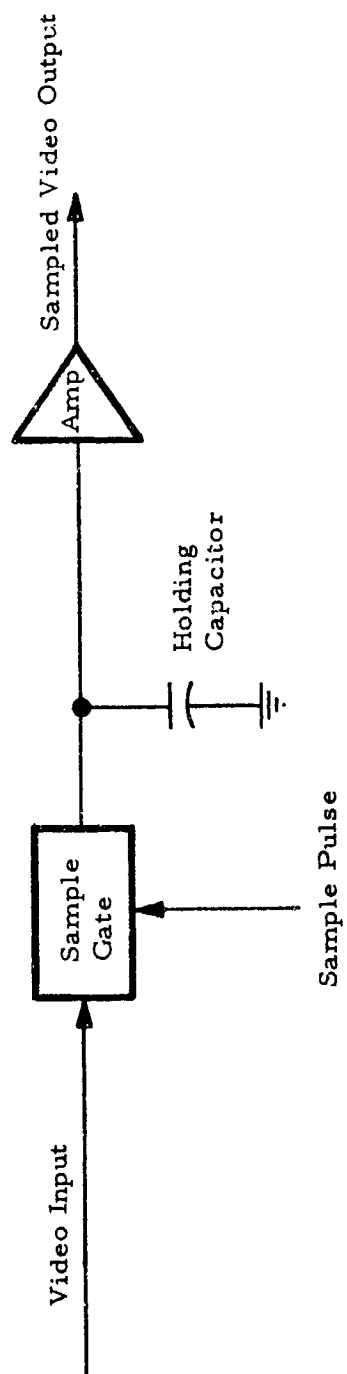


FIGURE 1.1-7 SAMPLE AND HOLD

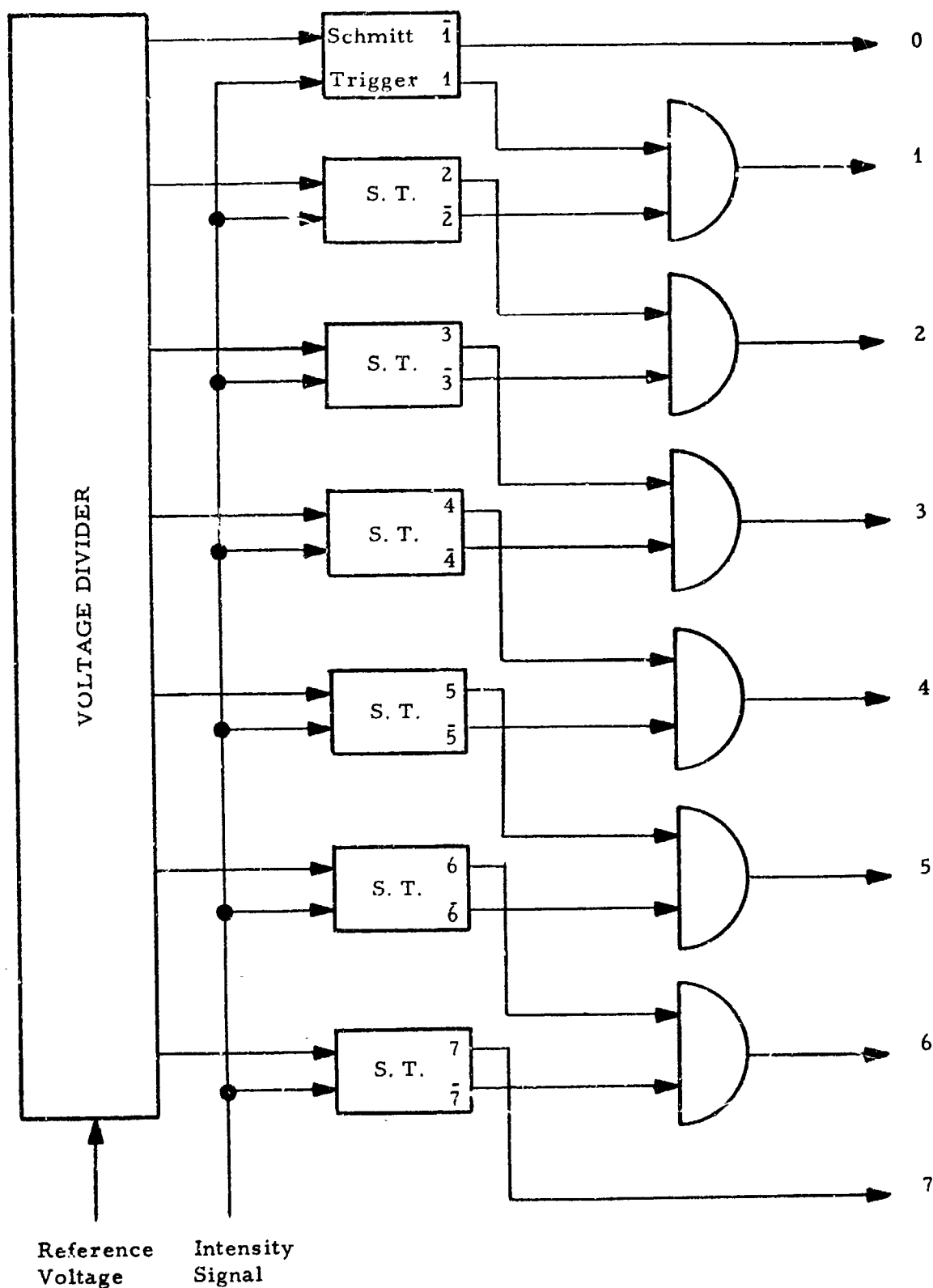


FIGURE 1.1-8 QUANTIZER

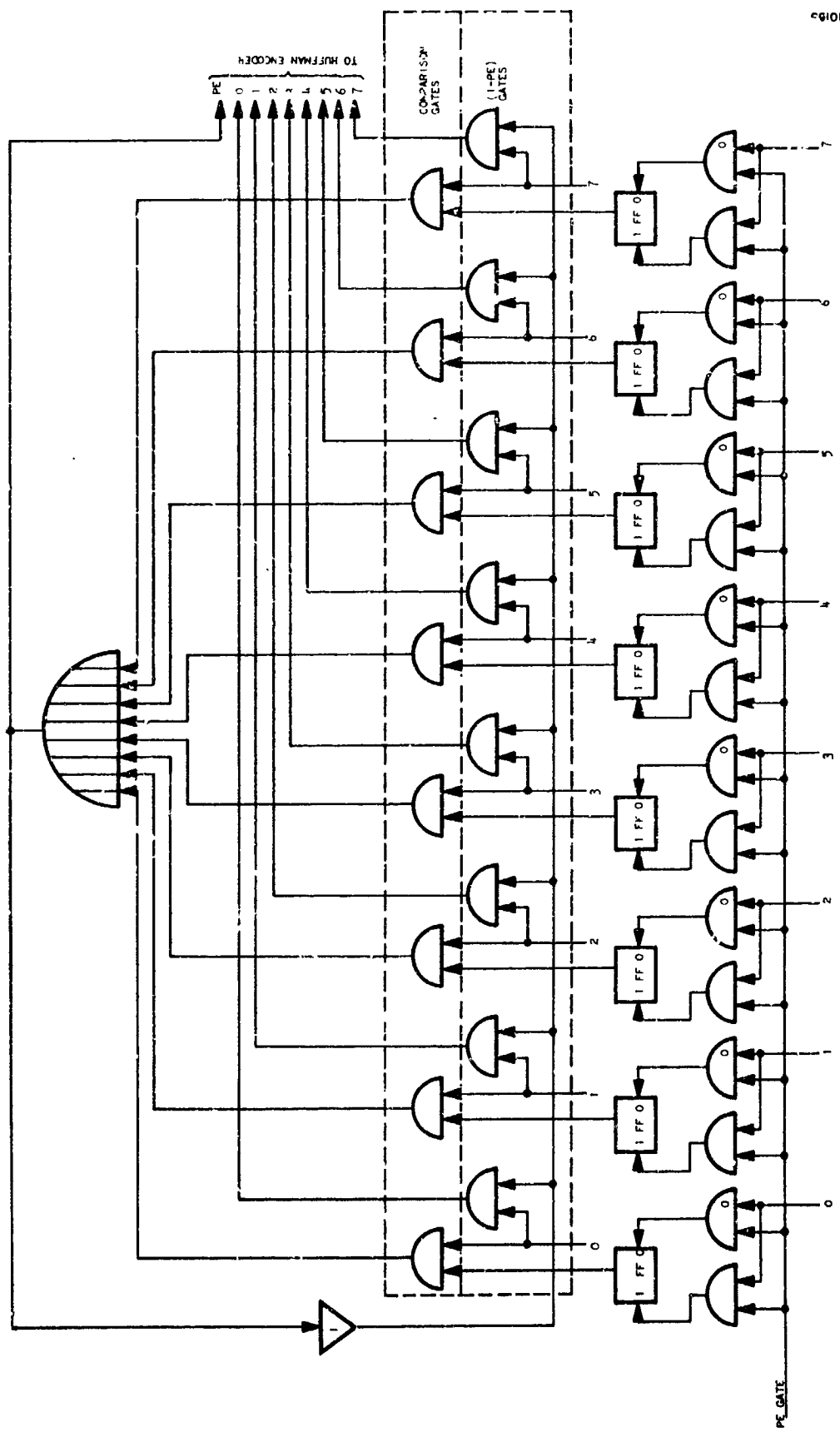


FIGURE 1.1-9 PE ENCODER

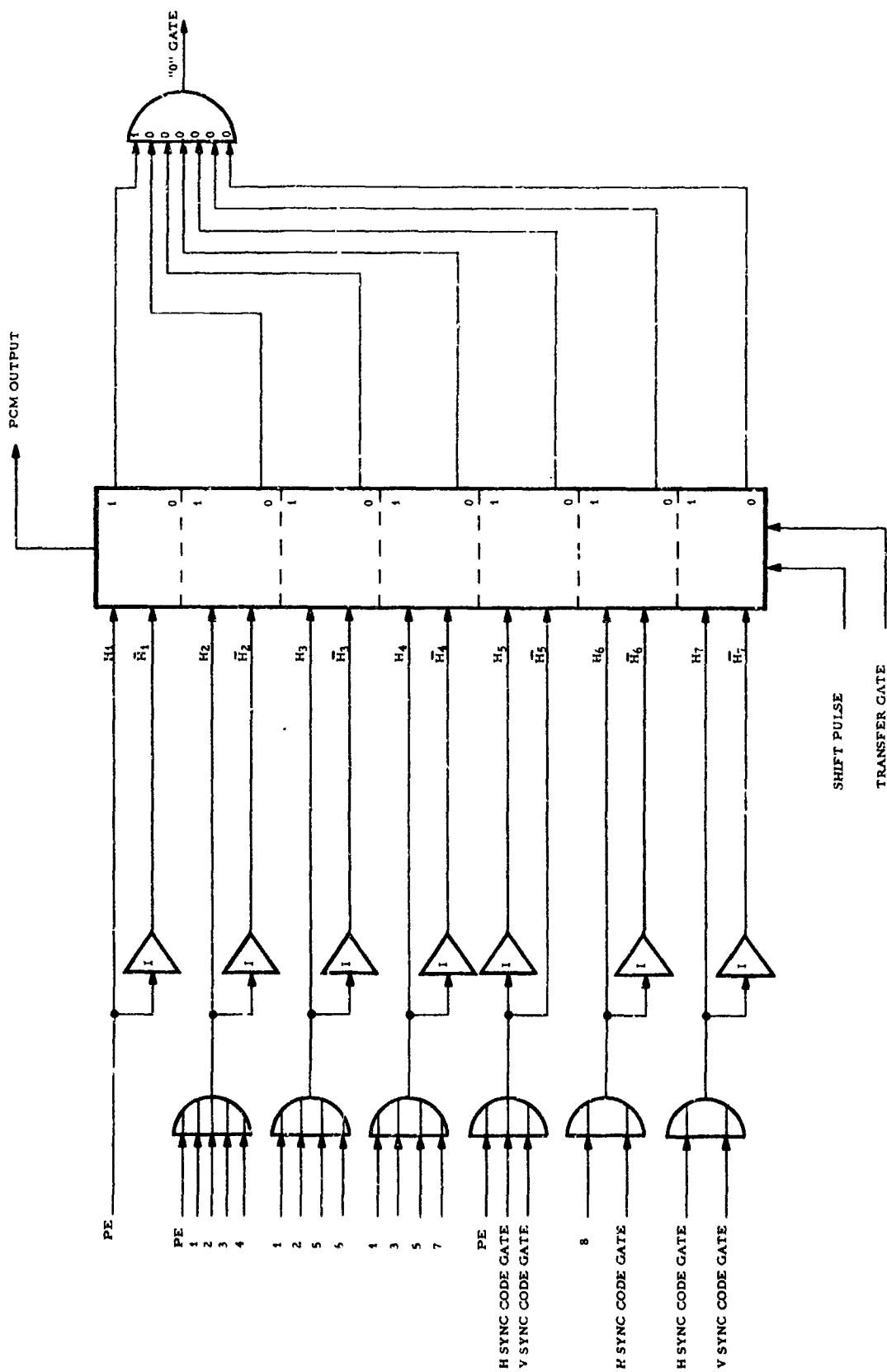


FIGURE 1.1-10 HUFFMAN ENCODER

1.1.2 Ground Equipment

1.1.2.1 Summary of Equipment Operation

Figure 1.1-11 is an overall diagram of the compressed digital ground equipment. The principal function of each system unit is summarized below:

1. Huffman Decoder

- a. Converts the received serial Huffman pulse train back to the PE and (1-PE) signal format.
- b. Generates necessary system timing pulses (sweep advance, display pulse, (1-PE) shift pulse, and vertical and horizontal synchronization pulses).

2. Video Digital-to-Analog Converter

Converts the Huffman output into the original analog video plus pseudo-random noise signal. This process includes PE decoding.

3. Roberts Decoder

- a. Generates a pseudo-random sequence of variable amplitude, which is synchronized with and identical to that produced in the Roberts encoder.
- b. Performs analog subtraction of this sequence from the combined video plus pseudo-random noise (effectively increasing the number of quantizing levels, thus reducing contouring effects in the displayed image).

4. Display Deflection

- a. Uses element, line, and frame synchronizing pulses from the Huffman Decoder to generate variable-length horizontal and vertical staircase waveforms synchronized with the camera scan waveforms.

5. Display

- a. Displays the final low-frame-rate analog video signal.
- b. Incorporates special erasure techniques to minimize brightness flicker effects normally associated with low-frame-rate TV systems. (Assumes multimode Tonotron display.)

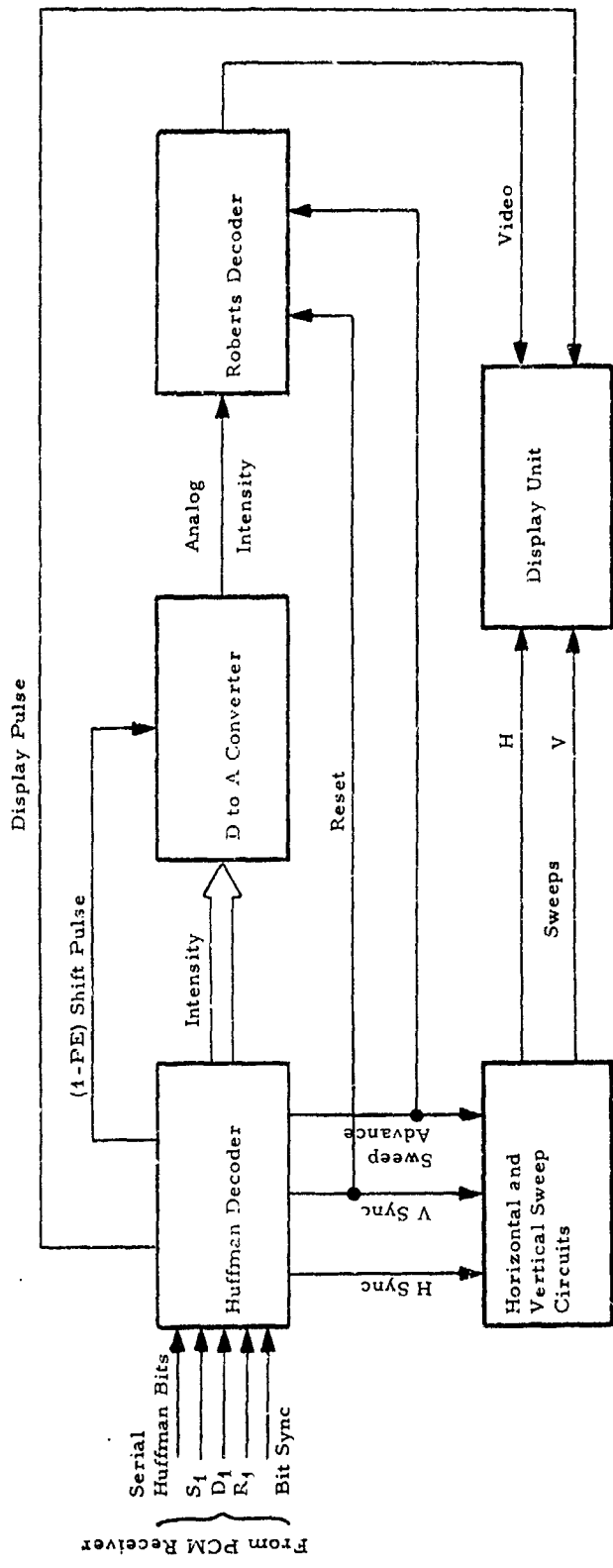


FIGURE 1.1-11 COMPRESSED DIGITAL TV GROUND EQUIPMENT

- c. Operates in an asynchronous scan mode to eliminate the need for a buffer store.

1.1.2.2 Design Concept

The entire ground station operates asynchronously. The system receives only the serial Huffman bits, at a constant rate, and four synchronous timing waveforms (which are assumed to be generated in the ground video telemetry receiving equipment). These waveforms are shown in Figure 1.1-12.

1.1.2.3 Huffman Decoder Operation

Figure 1.1-13 is a diagram of the Huffman decoder. The input to this unit consists of the four waveforms shown in Figure 1.1-12 plus the serial Huffman bit stream. The first Huffman bit applied to the flip-flops can set only FF_1 or FF_2 . All others are not enabled because the flip-flops preceding them are set to "one's." If the first bit is "one," FF_1 is set and this enables the PE hot line. If the first bit is a "zero," FF_2 is set, which enables FF_3 and FF_4 . If the second bit is a "one," FF_3 is set, enabling FF_5 and FF_6 . This sequential decoding continues until one of the hot lines is enabled.

The coincidence of an enabled hot line and pulse R resets all flip-flops to "zero" to ready the decoder for the next word.

The Huffman decoder also generates a sweep-advance pulse used to advance the sweep counters and to shift the Roberts register, a vertical synchronization pulse used to reset the vertical counter and the Roberts register, and a horizontal synchronization pulse used to reset the horizontal counter.

A (1-PE) shift pulse is generated to read new data into the digital-to-analog converter. A display pulse is also generated to display the video for the proper length of time on the CRT face.

1.1.2.4 Digital-to-Analog Converter Operation

This section is shown in Figure 1.1-14. The incoming hot line is converted to a binary code in a matrix and read into a three-stage register by the (1-PE) shift pulse. This level will remain stored until the occurrence of another level which is not PE, thus decoding the PE signal.

The stored binary level is converted to an analog signal by a conventional resistive ladder.

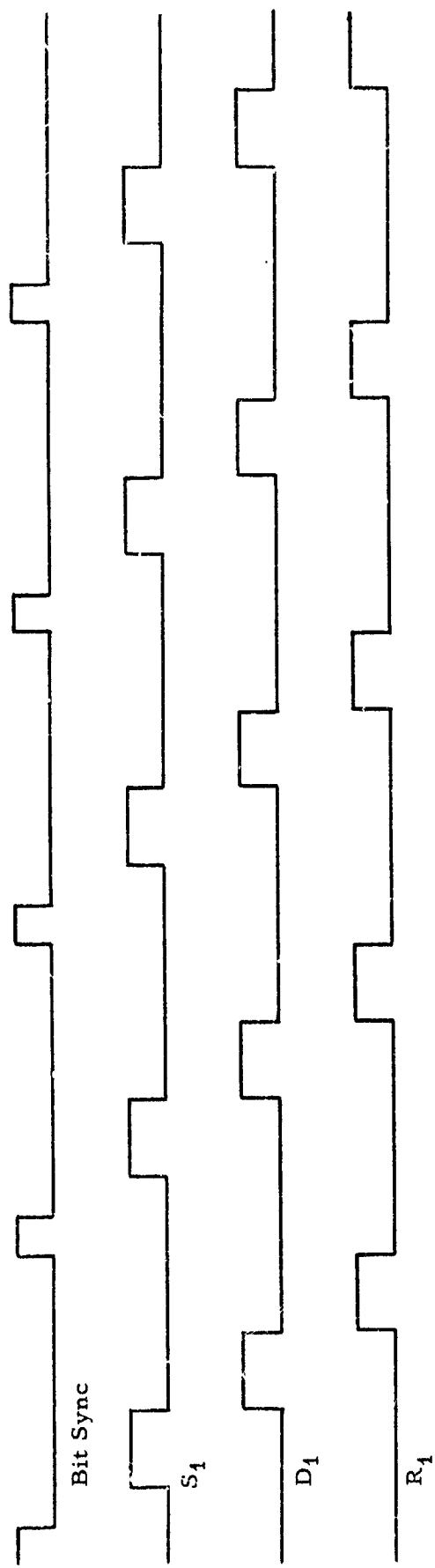


FIGURE 1.1-12 COMPRESSED DIGITAL TV GROUND EQUIPMENT
INPUT WAVEFORMS

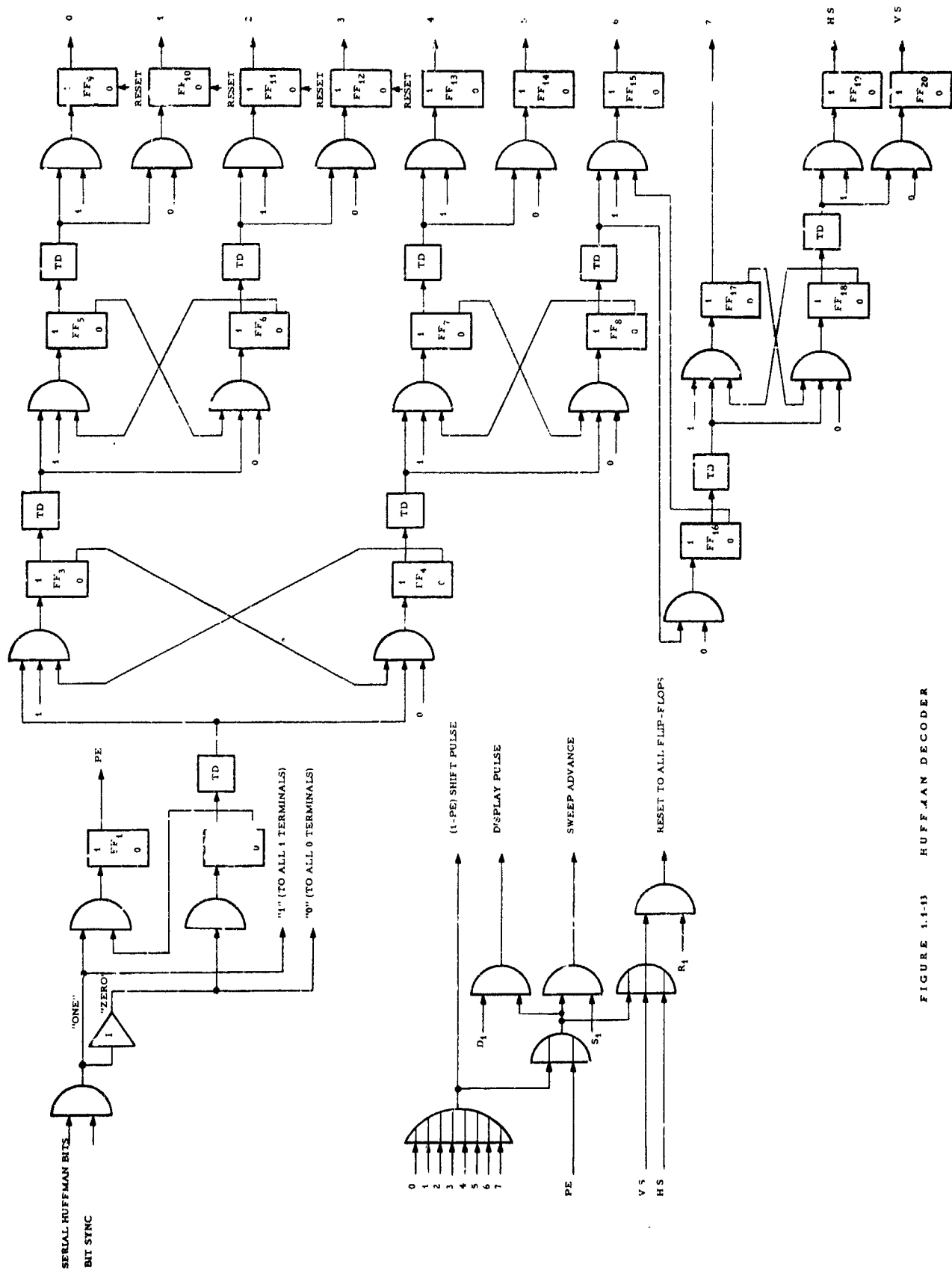


FIGURE 1.1-13 HUFFMAN DECODER

Analog Intensity to Robert's Decoder

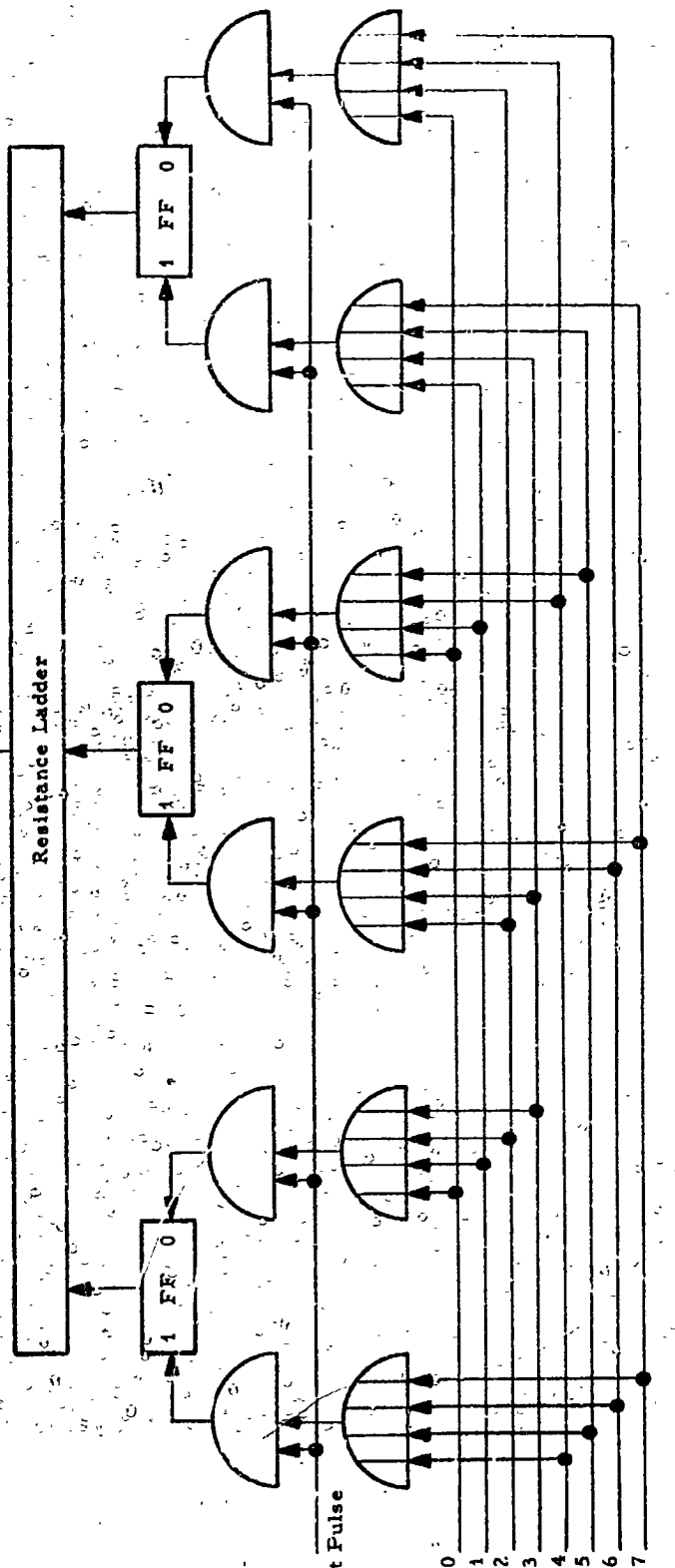
Resistance Ladder

(1-PE) Shift Pulse

Intensity Levels from
Huffman Decoder

0
1
2
3
4
5
6
7

FIGURE 1.1-14 D TO A CONVERTER



1.1.2.5 Roberts Decoder Operation

The Roberts Decoder is shown in Figure 1.1-15. A pseudo-random sequence generator identical to the one in the spacecraft equipment generates an identical sequence. The start of both generator sequences is synchronized with the vertical blanking pulse. These binary data are converted to an analog waveform and subtracted from the video signal.

1.1.2.6 Display and Sweep Circuit Operation

The display sweep is generated by counting the sweep-advance pulses in a binary counter and converting this count to an analog-sweep voltage. It is necessary to reset the sweep counters with the vertical and horizontal synchronization pulses to ensure synchronization of the display raster with the spacecraft camera scan.

The display incorporates circuitry which transforms the analog video signal into constant-width, amplitude-modulated pulses, and then displays these pulses at the appropriate element scan times.

1.2 STANDARD ANALOG TV SYSTEM

The "standard analog TV" considered in this study is by Manned Spacecraft Center definition* an analog TV Camera whose output feeds a 6 bit analog to digital converter. Thus, the complexity of the defined system is somewhat more than one would expect for an ordinary analog TV system. Figure 1.2-1 is a block diagram of both the spacecraft and ground equipments.

1.2.1 Spacecraft Equipment

The camera tube, camera control, video amplifier, automatic gain control, and the sample-and-hold blocks are essentially identical to those corresponding blocks in the compressed digital TV. The quantizer is a high-speed, parallel analog-to-digital converter with a six-line output. These six parallel lines are fed to the PCM output unit. The synchronizer provides a system clock, sample pulse for the sample and hold, sweep advance pulses for the sweep generator circuits, and gating and synchronizing signals for the PCM output unit. These synchronizing signals and the quantized video signals are serialized in the PCM output unit and fed to the PCM link. Horizontal and vertical sweep generators and sweep amplifiers are conventional analog sawtooth scan circuits.

1.2.2 Ground Equipment

Serial information from the PCM link is decoded into synchronization and video digital signals in the PCM decoder. The synchronization signal is sent to the horizontal and vertical sweep generator circuits. The digital-to-analog con-

*Paragraphs II-B and II-C of RFP (See Appendix A) Telecon: Knight/Engert and Stanwood, 6-12-63; Follow-up: Bi-Weekly Technical Report No. 12, 6-19-63.

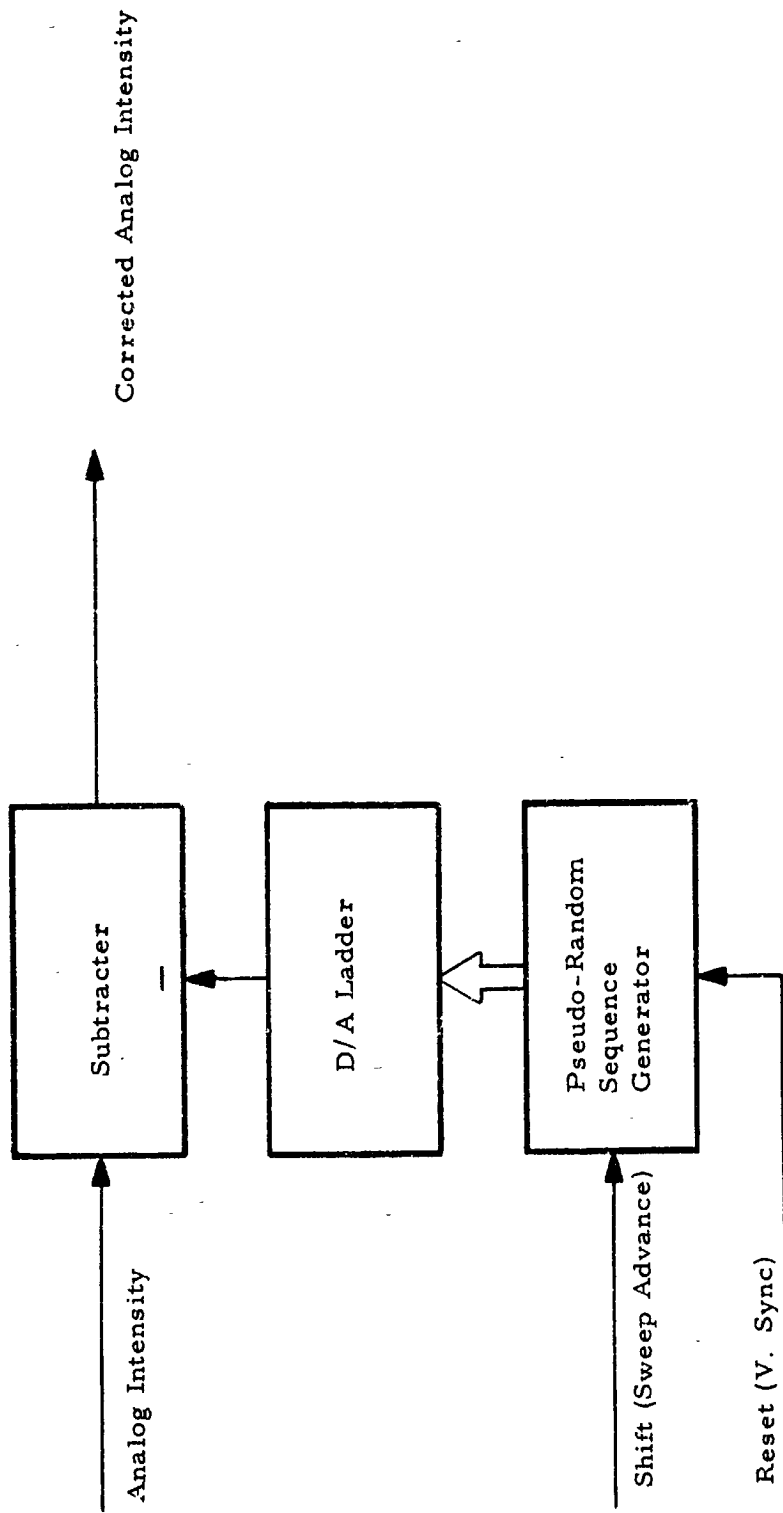


FIGURE 1.1-15 ROBERTS DECODER

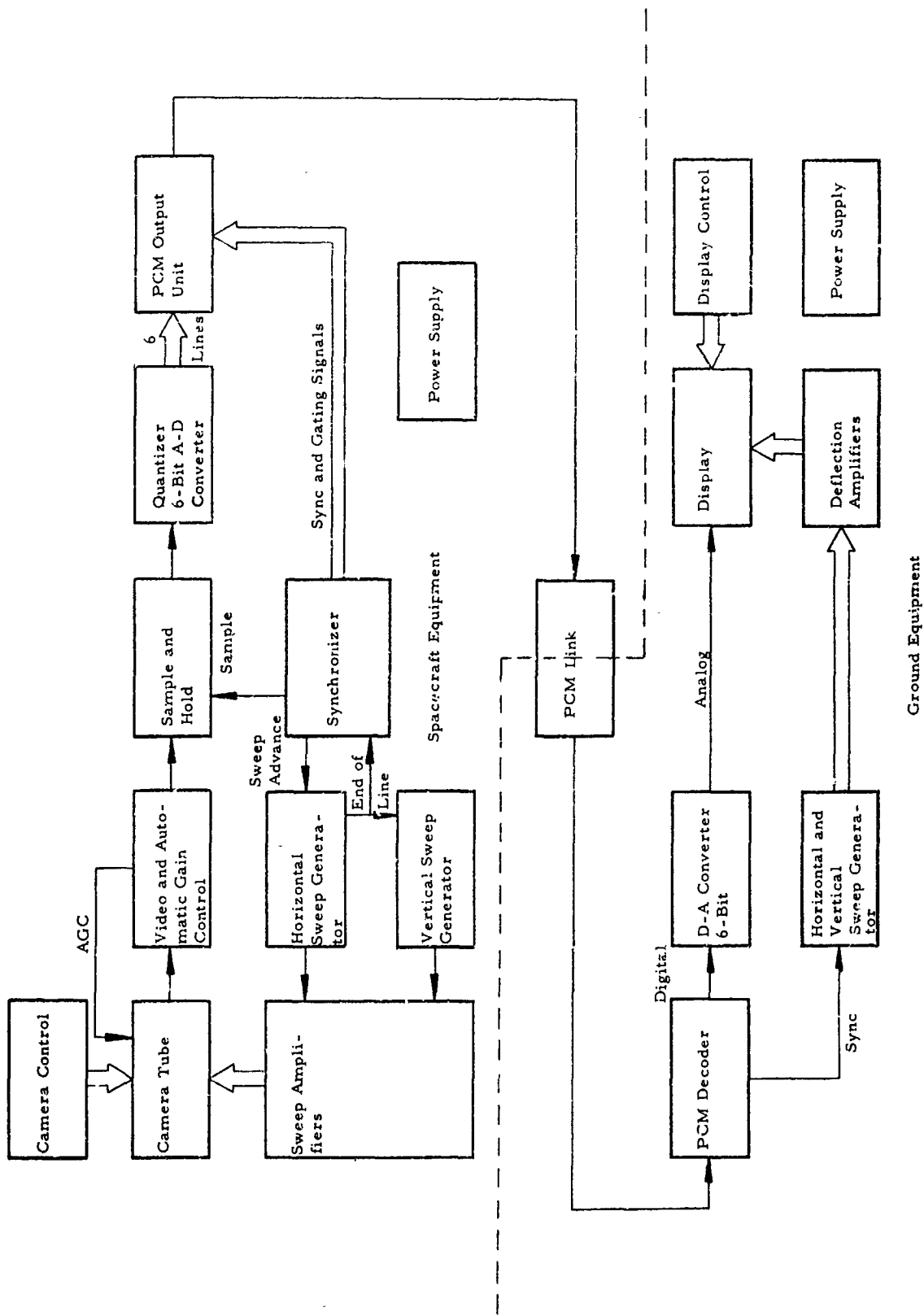


FIGURE 1.2-4 STANDARD ANALOG T. V. BLOCK DIAGRAM

verter performs the inverse operation of the spaceborne analog-to-digital converter. Deflection amplifiers drive the deflection plates or coils of the display tube and the display control supplies other voltages required to reproduce the images received from the camera system.

1.3 NONCODED ANALOG AND STRAIGHT PCM TV EQUIPMENT DESIGNS

As in the case of the two previously discussed systems, logic and block diagrams were prepared on the spacecraft portions of both the above designs to accomplish the required circuit design and reliability studies.

To permit direct equipment comparisons, the TV-camera scanning parameters selected were the same as those used for the compressed digital and standard analog systems; that is,

Horizontal resolution = 256 lines

Vertical resolution = 256 lines

Frame rate = 15 per second

Since the previously discussed standard analog system is essentially a composite of the noncoded analog and straight PCM systems, it would be duplicative to include design details in this section.

1.4 CIRCUIT DESIGN STUDIES

Currently, it is both possible and practicable 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 is 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 an 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. The logic designs generated in the MSC study are initial and are not intended to be optimum insofar as a particular circuit implementation of the logic designs is concerned.

A complete logic-design-circuit-design loop, required in the creation of functioning equipment, is beyond the scope of the MSC study. However, it is of interest to comprehend the limiting equipment parameters of several of the TV systems considered. Thus, the criteria of high reliability, minimum size, and reasonable cost were established as a basis for the circuit selection.

For highest reliability, without resorting to redundancy techniques, miniature discrete parts were considered for the circuits of the several TV systems estimated. Miniature parts in this instance mean small transistor cases, small glass diodes, standard-size metal film resistors, and small-size capacitors. For minimum-size, hybrid semiconductor integrated circuits were considered for the several TV systems estimated. It was necessary to use hybrid semiconductor integrated circuits instead of total semiconductor integrated circuits (SCIC), since, in some of the subsystems, either speed or high voltage requirements obviated SCIC circuits. There are a number of manufacturers of SCIC's. The estimates in this study were made primarily around Westinghouse SCIC's since these units are representative of available logic elements and also since Westinghouse has the widest available line of analog circuits. One penalty of using Westinghouse's circuits in place of, for example, Texas Instruments is that the latter company's circuits require less volume.

Hybrid thin film circuits are certainly applicable to building the considered TV systems. This approach would fall between the two estimated approaches in terms of equipment size. It may be logically and quite strongly argued that hybrid thin film circuits are more reliable than hybrid SCIC's. Actually, neither approach can be supported on the basis of extensive reliability testing, but there is more reliability history available on the SCIC's than on thin films. On this basis, the SCIC approach was selected for this study. Should one of the considered TV systems be implemented in the future, hybrid thin film should strongly be considered and not eliminated on the basis of the estimates appearing in this study.

Reasonable cost is considered only insofar as not using quad redundancy techniques in the logic circuits. The cost of quad redundancy is more than just a factor of four for several reasons: First, transistor specifications for quad circuits generally have to be more constraining than for single circuits; second, fan-out and speed of quad circuits are degenerated; third, on a per-function basis, quad circuits require more power than the single circuits; fourth, building and testing of quad circuits are time consuming and the testing is particularly difficult.

Table 1-3 is a tabulation of the parts estimate for both implementation approaches estimated for the four considered TV systems.

The volumes of various parts shown in Table 1-4 were used in the estimates of Table 1-3. The TO-5 transistor-case volume was used for SCIC BC's, F.F.'s and NAND's. It should be noted that the volumes given allow for projecting leads from the parts. Power estimates used are given in Table 1-5. Resistor power

TABLE 1.3 PARTS COUNT ESTIMATES

System		HYBRID-SCIC PARTS										DISCRETE PARTS									
		SCIC PARTS					DISCRETE PARTS					SCIC PARTS					DISCRETE PARTS				
		Power (watts)	Volume (cu. in.)	Amp.	B. C.	F. F.	NAND	Q	CR	R	C	Other	Power (watts)	Volume (cu. in.)	Q	CR	R	C	Other		
Compressed digital TV	Video and AGC	822	0.76	11	...	...	...	7	6	22	8	...	672	1.79	30	6	55	13	...		
	Camera control	1000	4.00	...	...	...	...	18	...	108	...	Potentiometers 9	1000	4.00	18	...	108	...	Potentiometers 9		
	Horizontal and vertical sweep generator	3720	3.40	...	16	...	...	40	2	64	...	...	3250	5.21	88	50	160	32	...		
	Horizontal and vertical deflection amp.	1000	2.40	...	...	...	...	14	...	48	...	Potentiometers 4	1000	2.40	14	...	48	...	Potentiometers 4		
	Synchronizer	942	1.94	...	2	3	18	9	4	27	2	Crystal 1	700	2.80	64	4	76	8	Crystal 1		
	Sample and hold	60	0.29	...	...	...	...	6	6	8	1	...	60	0.29	6	6	8	1	...		
	Quantizer	1320	1.92	...	...	...	6	21	1	60	14	...	1200	2.09	33	1	72	14	...		
	PE encoder	658	2.04	1	...	8	43	...	...	...	...	...	510	2.02	102	...	130	...	...		
	Koberta encoder	1390	0.98	...	18	...	4	...	...	...	...	...	900	4.20	62	40	120	36	...		
	Huffman encoder	1138	1.08	10	6	2	15	...	...	...	...	Half adder 3	615	3.32	75	15	100	...	...		
Power supply		5000	4.00	...	...	...	Note 1	...	...	...	...	4500	3.13	...	...	Note	...	...	...		
Totals		17.0 watts	22.81	...	...	...	...	...	...	...	...	13.4 watts	31.45	...	...	...	...	...	...		
Straight PCM TV	Video and AGC	822	0.76	11	...	...	...	7	6	22	8	...	672	1.79	30	6	55	13	...		
	Camera control	1000	4.00	...	...	...	...	18	...	108	...	Potentiometers 9	1000	4.00	18	...	108	...	Potentiometers 9		
	Horizontal and vertical sweep generator	3720	3.40	...	16	...	...	40	2	64	...	...	3250	5.21	88	50	160	32	...		
	Horizontal and vertical deflection amp.	1000	2.40	...	...	...	...	14	...	48	...	Potentiometers 4	1000	2.40	14	...	48	...	Potentiometers 4		
	Synchronizer	2798	2.36	...	32	1	11	6	4	20	6	...	6100	10.90	134	74	350	100	...		
	Sample and hold	60	0.29	...	...	...	...	6	6	8	1	...	60	0.29	6	6	8	1	...		
	Quantizer	2640	3.84	...	...	...	12	42	1	120	23	...	2640	4.18	66	1	144	28	...		
	PCM output unit	780	0.52	...	10	...	3	...	...	...	...	...	1600	5.08	59	60	300	60	...		
	Power supply	5200	4.00	...	...	...	Note 1	...	...	...	...	...	6500	4.00	...	...	Note	...	...		
	Totals		18.0 watts	21.57	...	...	...	...	...	...	...	...	22.9 watts	37.84	...	...	...	...	...	...	
Norced analog TV	Video and AGC	822	0.76	11	...	...	...	7	6	22	8	...	677	1.79	30	6	55	13	...		
	Camera control	1000	4.00	...	...	...	...	18	...	108	...	Potentiometers 9	1000	4.00	18	...	108	...	Potentiometers 9		
	Horizontal and vertical sweep generator	1082	1.16	...	...	...	...	18	2	48	4	...	1082	1.16	18	2	48	4	...		
	Horizontal and vertical deflection amp.	1000	2.40	...	...	...	...	14	...	48	...	Potentiometers 4	1000	2.40	14	...	48	...	Potentiometers 4		
	Synchronizer	1260	1.29	4	8	...	1	6	...	20	8	Crystal 1	1035	3.97	41	20	132	40	Crystal 1		
	Power supply	4000	3.33	...	...	...	Note 1	...	...	...	...	...	3700	3.33	...	...	Note	...	...		
	Totals		9.2 watts	12.94	...	...	...	...	...	...	...	...	8.5 watts	16.65	...	...	...	...	...	...	
	Standard analog TV	Video and AGC	822	0.76	11	...	...	...	7	6	22	8	...	672	1.79	30	6	55	13	...	
		Camera control	1000	4.00	...	...	...	...	18	...	108	...	Potentiometers 9	1000	4.00	18	...	108	...	Potentiometers 9	
		Horizontal and vertical sweep generator	1082	1.16	...	...	...	...	18	2	48	4	...	1082	1.16	18	2	48	4	...	
Horizontal and vertical deflection amp.		1000	2.40	...	...	...	...	14	...	48	...	Potentiometers 4	1000	2.40	14	...	48	...	Potentiometers 4		
Synchronizer		2903	2.50	...	3	...	14	6	4	20	6	...	6823	11.70	146	77	380	103	...		
Sample and hold		60	0.29	...	...	...	...	6	6	8	1	...	60	0.29	6	6	8	1	...		
Quantizer		10560	15.36	...	...	...	48	168	4	480	112	...	10560	16.72	264	4	576	112	...		
PCM output unit		1080	0.68	...	14	...	3	...	...	...	...	...	2388	6.51	51	68	248	76	...		
Power supply		7500	6.67	...	...	...	Note 1	...	...	...	...	...	9200	8.33	...	...	Note	...	...		
Totals		25.0 watts	40.48	...	...	...	...	...	...	...	...	33.8 watts	59.24	...	...	...	...	...	...		

NOTE: 1. Estimates based on known high-reliability power supply.

TABLE 1-4

VOLUME ALLOWED FOR PARTS ESTIMATE

Item	Volume (cubic inches)
SCIC amplifier	0.004
TO-18 transistor case	0.02
TO-5 transistor case	0.04
Diode	0.005
Resistor	0.014
Capacitor	0.03
Potentiometer	0.2
Crystal	0.2
Vidicon	13.0

TABLE 1-5

POWER ALLOWED FOR PARTS ESTIMATE

Item	Power (milliwatts)
SCIC Amplifier	36
SCIC Binary Counter	75
SCIC Flip-Flop	24
SCIC NAND	10
Amplifier Transistor	60
High-Speed Switching Transistor	60
Low-Speed Switching Transistor	10
Diode	1
Vidicon Filament	1000

is considered in the transistor power allowed except in the few cases where resistors are used without transistors. In these cases, the resistor power is considered separately.

Table 1-3 data relate only to actual parts, power, and volume. To arrive at a realistic prototype equipment estimate, correction factors were applied to the Table 1-3 data. These correction factors were arrived at on the basis of previous experience. For example, consider volume. A known high-reliability discrete-parts circuit module consists of 16 TO-size transistors, 16 glass diodes, and 32 resistors. Using allowed volumes in Table 1-3, the estimated parts volume is 0.85 cubic inch. However, the measured volume of the module is 2.5 cubic inches. Thus, a multiplication factor of three is required to get equipment volume. In addition, the vidicon volume was added. Also, power corrections and weight computations were made on all units. The factors used were:

1. Power--Prototype power equals two times parts power.
2. Volume--Prototype volume equals three times parts volume plus vidicon volume.
3. Weight--Prototype weight equals 0.5 ounce per cubic inch for electronic parts and vidicon plus one ounce per cubic inch for the power supply.

Table 1-6 is a compilation of summary data from Table 1-3 and prototype data derived from applying the foregoing factors.

1.5 INFLUENCE OF EQUIPMENT DESIGN ON SPACECRAFT PRIME POWER AND WEIGHT

The results of the preceding circuit design study show that the compressed digital TV equipment is smaller and lighter and requires less input power than a comparable standard analog TV design. Also, it has been pointed out that the three-to-one reduction in information bandwidth afforded by the compression techniques results in savings in video transmitter power and weight. To determine the total savings in spacecraft prime power and weight, it is first necessary to estimate the weights and efficiencies of the associated video transmitters operating at their required output power levels.

Table 1-7 presents estimates of weights, efficiencies and input-output power levels for four video transmitters operating at the 2.3 kmc frequency specified in Appendix II of the RFQ. For defined Condition 1, it is assumed here for comparison purposes that the standard analog TV system will require the specified transmitter average output power of 25 watts. Under equivalent conditions

TABLE 1-6 SUMMARY DATA-EQUIPMENT COMPARISONS

TYPE DATA	PARAMETER	COMPRESSED DIGITAL TV		STRAIGHT PCM TV		NONCODED ANALOG TV		STANDARD ANALOG TV	
		Hybrid SCIC	Discrete Parts	Hybrid SCIC	Discrete Parts	Hybrid SCIC	Discrete Parts	Hybrid SCIC	Discrete Parts
Parts data	Power (watts)	17.0	13.4	18.0	22.9	9.2	8.5	25.0	33.8
	Volume (cu. in)	22.8	31.5	21.6	37.8	12.9	16.7	40.5	59.2
Prototype data	Power (watts)	34.0	26.8	36.0	45.8	18.4	17.0	50.0	67.6
	Nonstructural volume (cu. in)	81.4	107.5	77.8	126.4	51.7	63.1	134.5	190.6
	Nonstructural weight (pounds)	2.92	3.93	2.81	4.34	1.93	2.29	4.83	6.73

TABLE 1-7 ESTIMATED VIDEO TRANSMITTER CHARACTERISTICS

Frequency (kmc)	Average output of power (watts)	Total input power (watts)	System efficiency (%)	Total weight (lb).
2.3	8	50	16	8
2.3	25	140	18	13
2.3	330	1,000	33	25
2.3	1,000	3,000	33	50

(equal receiver input signal-to-noise ratio and equal TV scanning parameters, etc.) the compressed digital TV system, achieving a three-to-one bandwidth compression, will require approximately eight watts of r.f. power. For Condition 2, assuming that the standard analog TV system requires the specified 1,000 watts transmitter output power, the corresponding figure for the compressed digital TV system is 330 watts.

Under the assumed conditions, the compressed digital TV system, compared to the standard analog TV system, achieves the following total savings in spacecraft prime power and spacecraft equipment weight.

Condition 1 (assuming discrete parts design).

- (a) Spacecraft prime power saving = TV equipment power saving + video-transmitter power saving $\cong 40 + 90 = 130$ watts.
- (b) Spacecraft weight saving = TV equipment weight saving + video-transmitter weight saving $\cong 3 + 5 = 8$ lb.*

Condition 2 (assuming discrete parts design)

- (a) Spacecraft prime power saving $\cong 40 + 2000 = 2,040$ watts.
- (b) Spacecraft weight saving $\cong 3 + 25 = 28$ lb.*

*As mentioned in the Executive Summary (Paragraph 1.0), these estimates of spacecraft-equipment weight saving do not include any reduction in the weight of the prime power source permitted by the reduced prime power requirement. In an actual system, some reduction should be forthcoming; however, this is very much a function of the prime power source characteristics, the length of the mission, the operating duty cycle of the video transmitting equipment, and the nature of the other spacecraft equipment loads.

If the video equipment operation is intermittent, it is possible that a time-sharing arrangement with other spacecraft equipment could be arrived at, such that the video equipment would not require a significant increase in prime source power capacity. Under these conditions, it would only be necessary to assure that sufficient additional energy were available from the prime source to accommodate the total energy requirements of the video equipment. In the case of solar-powered and nuclear-powered sources, their sizes are dictated primarily by total power requirements rather than total energy requirements.

It is certainly beyond the scope of this study program to attempt to define overall spacecraft equipment and operational requirements. Rather, the above discussion is included only to show that the determination of total spacecraft weight savings achievable by use of video data compression techniques is considerably influenced by other system considerations.

1.6 PACKAGING CONCEPTS AND CONSIDERATIONS

An optimum spacecraft electronic package can only be designed with detailed information of the number and type circuits, expected mechanical stresses, restricting component sizes, maximum dimensions, and many other inputs. It is not the purpose of this study to give the details of a mechanical design. However, it is of interest to consider some of the factors that would influence an optimum package design of the compressed digital TV and standard analog TV using both hybrid SCIC and discrete parts approaches.

The TV systems considered may be divided logically into three functional units: camera and video, digital, and power supply. It is reasonable to consider the camera and video unit as a separate entity connected via a cable to an electronic package containing the digital and power supply units. The vidicon tube and its shielding restrict the minimum dimensions of the camera and video unit to a width of two inches and a length of eight inches. Only the power supply imposes dimensional limits on the other electronic package, since it is approximately cubic in form factor. With allowance for the foregoing dimensional limits and the calculated prototype equipment volume plus 10% for structure, the dimensions of the packages shown in Figure 1.6-1 were derived. It is anticipated that the digital unit and power supply unit may be easily separated to facilitate replacement of a defective unit. The construction lends itself to shielding against RFI and radiation, urethane foam potting, minimum connectors and connections, welded sub-modules, tension-membered cover, and flexibility in using the camera at a separate location from the other electronics.

Any desired optics may be externally attached to the camera unit. The drawings of Figure 1.6-1 do not include optics.

The details of the modular construction would depend upon the equipment reliability and maintainability requirements. For instance, if standby redundancy is desired, replaceable plug-in modules would be preferred with the portion of the equipment that exhibits the highest probability of failure (i. e., the analog circuits) grouped in one replaceable module.

The use of either soldered or welded construction would be satisfactory. However, for optimum reliability, welding is preferred since the component parts are subjected to less thermal stress during production. This is not meant to imply that a welded connection is more reliable than a soldered one. Conversely, if the necessary precautions are taken to protect parts from thermal stress during soldering, a soldered assembly can equal or even surpass the reliability of welded construction since less stringent process controls are required for soldering.

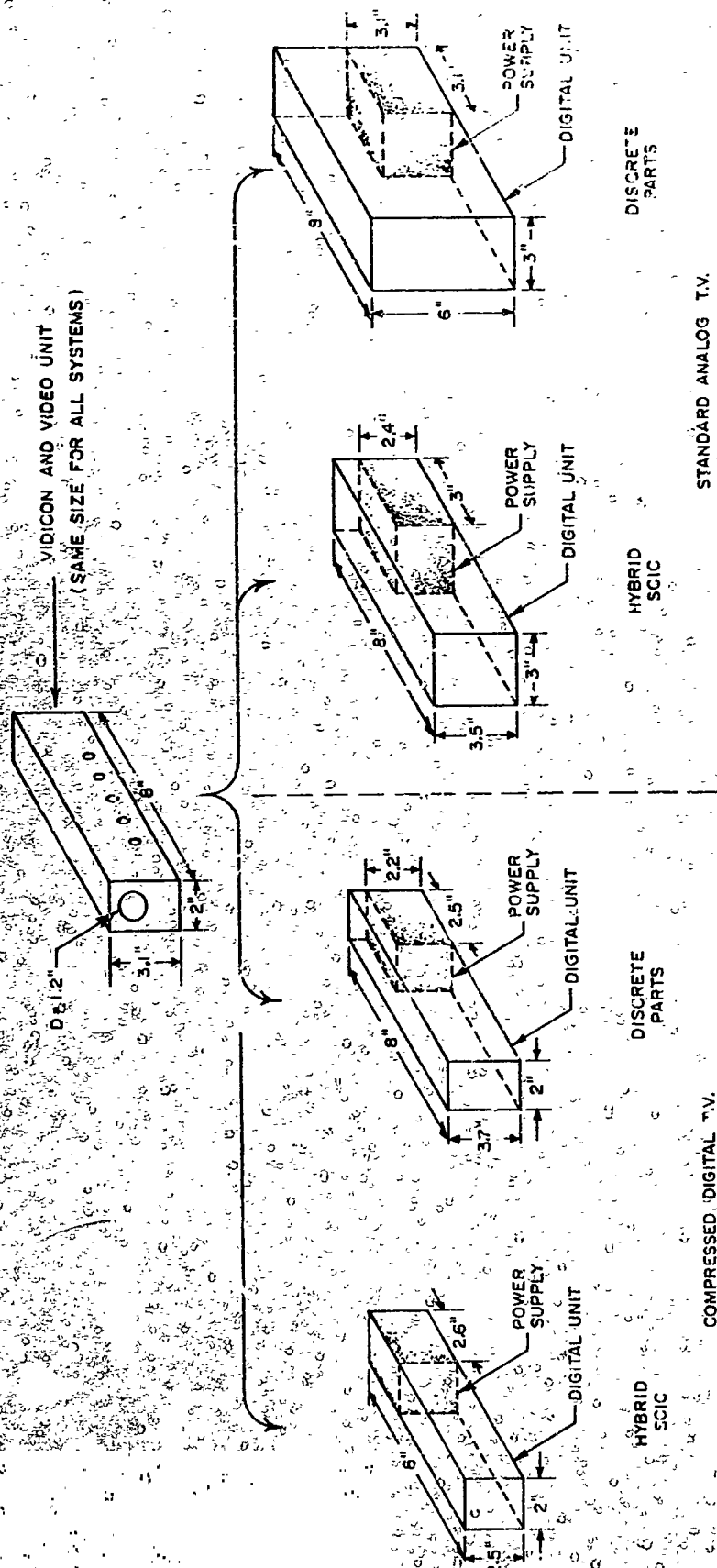


FIGURE 1.6-1 PROTOTYPE SPACECRAFT EQUIPMENT PACKAGING

1.7 EQUIPMENT RELIABILITY STUDIES

1.7.1 Introduction

The reliability analyses were made from part counts of each of the eight spacecraft equipment configurations previously discussed. No detailed stress analyses were performed.

All available sources of failure rates were investigated in an effort to arrive at realistic, up-to-date estimates. Among these sources investigated were Minute-man Failure Rate Goals, RADC Reliability Notebook, Martin Company Failure Rates, and vendor test data. Vendor test data were the only available sources for integrated circuits.

After evaluation of each individual failure rate source and the application of EMR's own experience with high-reliability parts, failure rates were derived for mid-1963 and mid-1964. These failure rates are listed in Table 1-8. The mean-time-between-failure (MTBF) of each system was calculated using these derived failure rates.

1.7.2 Analysis of Integrated Circuits

Most of the vendor data available on integrated circuits were published in late 1962 and included the results of testing during that year. The failure rates that were achieved at the end of 1962 varied considerably from vendor to vendor. After evaluation of each failure rate and the test data, several reasons can be given for these variances.

Because of the newness of integrated circuits, relatively little test time was accumulated and the samples were too small to obtain a failure rate comparable to discrete parts with a high degree of confidence.

Some published failure rates were much lower than the average and were backed up by considerably more test hours than the average. The analysis of these test data revealed that the failure rates were misleading. The total test time from which the failure rates were derived was actually the total unit test time multiplied by the number of equivalent discrete parts which would have been necessary to perform the same function as was performed by the integrated circuit.

For example, if an integrated circuit performed a logical function that required 20 discrete parts, the total number of unit test hours was multiplied by 20. This total number of hours is actually the total equivalent discrete part test time. However, the design of the integrated circuits dictated that only the entire circuit be monitored during test, and a failure of an equivalent discrete part is not necessarily reflected in performance. Therefore, the data used to derive the failure rate per unit comprised the total equivalent part test time and the total circuit

TABLE 1-8 ESTIMATED FAILURE RATES

PART	TYPE	MID-1963 FR $\times 10^{-8}$	MID-1964 FR $\times 10^{-8}$
SCiC	Logic	50.0	7.0
SCiC	Analog	100.0	14.0
Transistor	Logic	2.0	1.0
Transistor	General purpose	4.0	2.0
Diode	Logic	0.2	0.1
Diode	General purpose	0.9	0.5
Resistor	Precision	1.0	0.5
Resistor	Composition	0.5	0.3
Resistor	Variable	5.0	3.0
Capacitor	Solid tantalum	1.0	0.5
Capacitor	Glass or paper	0.6	0.3
Crystal	...	10.0	8.0
Transformer power	...	50.0	35.0
Transistor power	...	10.0	5.0

failures. The calculated failure rate is then a meaningless number and was not considered in the estimated failure rates used in this analysis.

The integrated circuit failure rates used in this analysis were derived by statistically projecting the actual failure rates achieved in 1962 to mid-1963 and mid-1964. A significant improvement in reliability from mid-1963 to mid-1964 is expected, and the failure rates used for mid-1964 are based on the assumption that integrated circuits will be available on a mass-production basis by 1964.

Because of the limited amount of test data available on integrated circuits, the failure rates should be considered a relative number to be used for comparing the four systems and should not be considered a precise numerical estimate of reliability.

1.7.3 Analysis of Discrete Parts

The failure rates derived for discrete parts were taken from several sources which are widely used in aerospace applications. In addition to these published failure rates, consideration was given to EMR's past experience with components in high reliability programs.

The failure rates derived for mid-1963 were statistically projected to mid-1964. These numerical estimates for mid-1964 were based on the assumption that the present reliability improvement trend will continue through mid-1964. An improvement of about 2:1 was predicted.

1.7.4 Reliability Analysis Discussion

The MTBF of each system was calculated by taking the reciprocal of the summation of the individual part and integrated circuit failure rates. It is stressed that the total equipment failure rates for the discrete part designs are based on the assumption that a failure of any one part causes total equipment failure. The failure rates for the integrated circuits are derived from failure rates of entire circuits. (A failure is defined as either catastrophic or an out-of-tolerance condition). In many cases a discrete part could also drift out of tolerance without causing a circuit failure. To make a more realistic comparison of discrete vs SCIC designs, a detailed parts application study would be required.

Because of the unavailability of statistically valid vidicon tube reliability information, it was not included in the reliability estimates. The same vidicon would be used in each system, so the omission of this part from the analysis does not affect the comparisons between systems. Based on the performance of vidicons employed in the TIROS satellite series, acceptable reliability should be achievable with a generally similar pick-up device in the MSC camera equipment.

Redundancy and other reliability improvement measures were not considered in the equipment for this study. If higher reliability should be desired in operational hardware, these reliability enhancement measures should be considered.

1.7.5 System Comparisons and Discussion of Results

Table 1-9 lists the calculated failure rates for each equipment design on a functional block level. Both mid-1963 and projected mid-1964 estimates are included. The reciprocal of the total failure rate for each equipment, expressed as MTBF, is presented as a histogram in Figure 1.7-1.

From these data, it is concluded that the exclusive use of discrete parts will result in higher equipment reliability. Using the mid-1963 failure rates the ratio of predicted MTBF of designs using discrete parts to designs using hybrid integrated circuitry is on the order of three to one. Using the projected failure rates for mid-1964, the reliability of hybrid integrated circuits begins approaching discrete circuit reliability. However, if integrated circuits do not reach a mass-production basis by 1964 and a substantial number of test hours are not accumulated, the expected improvement in reliability from mid-1963 to mid-1964 may not be achieved.

Comparing the spacecraft TV equipment on the basis of similar design approaches, it is concluded that there are no significant reliability differences between compressed digital TV, standard analog TV and straight PCM TV. As would be expected, noncoded analog TV, because of its greater simplicity, is the most reliable equipment.

MTBF
(1000 hrs.)



Discrete Part Circuits



Integrated Circuits

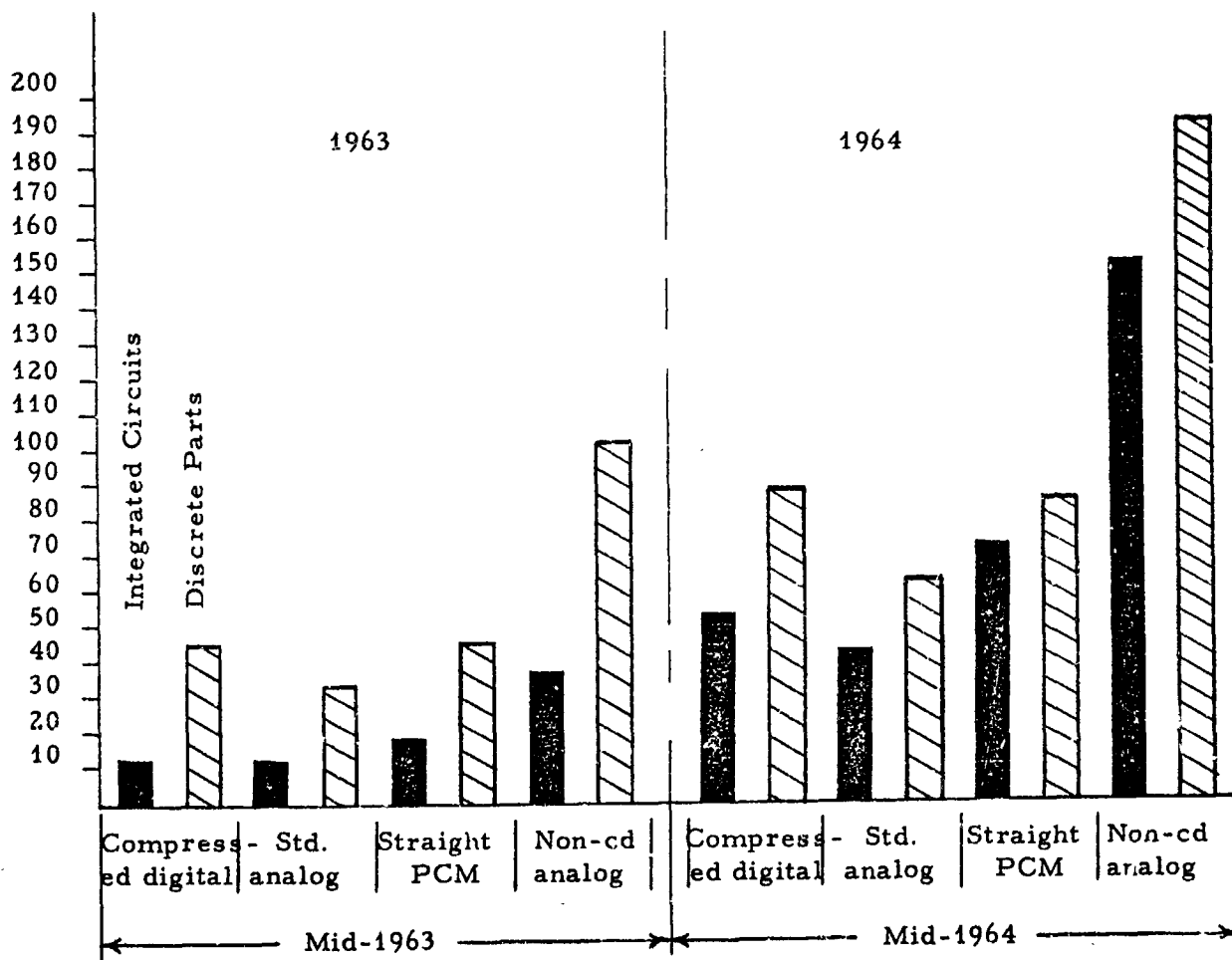


FIGURE 1.7-1 MTBF COMPARISON HISTOGRAM

TABLE 1-9 FAILURE RATES FOR SPACECRAFT TV EQUIPMENT

FUNCTION	COMPRESSED DIGITAL				STANDARD ANALOG				STRAIGHT PCM				NONCODED ANALOG			
	SCIC		Discrete		SCIC		Discrete		SCIC		Discrete		SCIC		Discrete	
	'63	'64	'63	'64	'63	'64	'63	'64	'63	'64	'63	'64	'63	'64	'63	'64
Video and AGC	1159	180	167	86	1149	180	167	86	1149	180	167	86	1149	180	167	86
Camera controls	198*	94*	198	94	198*	94*	198	94	198*	94*	198	94	198*	94*	198	94
Horizontal and vertical sweep generator	994	212	381	191	75*	42*	79	42	994	212	381	191	79*	42*	79	42
Horizontal and vertical deflection circuit	112*	59*	112	59	112*	59*	112	59	112*	59*	112	59	112*	59*	112	59
Synchronizer	1214	198	302	198	2442	358	545	307	2242	400	533	285	908	152	205	109
Sample and hold and add	34*	18*	34	18	34*	18*	34	18	34*	18*	34	18	...	...	...	...
Quantizer (analog-to-digital)	441	114	129	82	3530	1365	1365	740	881	228	341	175	...	...	...	...
PE encoder	715	371	269	141	...	...	...	...	...	...	...	...	...	...	...	...
Roberts encoder	1100	154	214	113	...	...	...	...	...	...	...	...	...	...	...	...
Huffman encoder	2300	322	203	107	...	...	...	...	...	...	...	...	...	...	...	...
PCM output	...	...	...	...	340	119	285	157	450	91	226	123	...	...	...	...
Power supply	200*	100*	200	100	250*	150*	250	150	200*	100*	200	100	200*	100*	200	100
Total FR x 10 ⁻⁸ failures per hour	8457	1872	2209	1149	8134	2385	3035	1653	6260	1382	2192	1131	2646	627	961	490

*NOTE: NO SCIC USED

SECTION 2 CODING TECHNIQUES

2.0 EXECUTIVE SUMMARY

In this section the field of picture compression is broken down into four constituent parts (1) general considerations of philosophy of coding, definitions of pictures to be coded, grade of pictorial rendition, and special bibliography; (2) types of redundancy elimination such as those involving element-to-element line-to-line, frame-to-frame and other correlations; (3) optimum encoding of the reduced redundancy source, including a consideration of the effects of error, and (4) a brief consideration of the storage requirements generated by the foregoing coding systems.

The redundancy reduction and optimum encoding effort continued throughout the course of this study; however, since the major output of this study is a system recommendation, that coding system (Roberts with PEC) which was decided upon during the midterm of this study has received the majority of the investigation effort. Nevertheless, the relative position of seven different video coding systems can now be summarized. This summary takes the form of Table 2.0-1.

The recommended system (Roberts with PEC) is seen from Table 2.0-1 to yield an expected power saving of about 3.7 db relative to straight PCM and an expected bandwidth saving of about 3:1, also relative to straight PCM. For these benefits one must pay the price of median (five out of eight) relative complexity and a picture quality judged acceptable. Systems exist with greater complexity and greater savings, but detailed knowledge of them is less complete. Relatively simpler systems exist that do not offer quite the expected savings. The expected trade-off between relative complexity and compression savings does indeed occur. Compression ratios of two to six can be forecast with reasonable certainty.

Two systems exist which are relatively farther out in development and yet which offer the potential of large ($> 6:1$) compressions. These are split-band and area encoding. One is tempted to add run-length encoding to this list since its potential compression is greater than 6:1, but actual video statistics on run-lengths have been disappointing and the split-band system itself makes use of a form of run-length encoding of the high band.

In summary, currently foreseeable systems can be expected to yield experimental picture compressions of 3:1 by the end of 1963. Future systems probably can exhibit compressions of about 6:1 by the end of 1964, if study and development proceed at the current pace.

2.1 GENERAL CONSIDERATIONS

Table 2.0-1

COMPARISON OF VIDEO CODING SYSTEMS

Parameter	Basic	PEC(4)	RLC	PCM and Roberts	PEC and Roberts	SE	Δ	PEPL
Resolution	256	256	256	256	256	256	256	256
Frame rate (fps)	15	15	15	15	15	15	15	15
Upper bound on compression	1:1	6:1	> 6:1	2:1	6:1	> 6:1	6:1	6:1
Actual compression	1:1	3:1	2:1*	2:1	3.2:1	6:1*	3:1	4:1
Bits per element	6	2	3	3	1.9	1	2	1.5
Compressed digital bit rate (MSPS)	5.9	1.97	2.95	2.95	1.84	0.98	1.97	1.48
Bandwidth (MC)	5.9	1.97	2.95	2.95	1.84	0.98	1.97	1.48
D_e^b	10^{-3}	10^{-4}	10^{-4}	10^{-3}	10^{-4}	10^{-5}	10^{-4}	10^{-5}
S/N (db)	8.0	9.4	9.4	8.0	9.4	10.4	9.4	10.4
Picture quality	G	G	G	A	A	A	A	G
Complexity (relative rank)	3	4	7	2	5	8	1	6
Bandwidth relative to basic	1	0.334	0.500	0.500	0.312	0.166	0.334	0.251
Power relative to basic (db)	1	-3.2	-1.6	-3.0	-3.7	-5.4	-3.2	-3.6

* Estimated

G = Good

A = Acceptable

2.1.1 General Remarks on Coding

Bandwidth reduction of a digital video source by coding is extremely desirable and is also feasible under certain conditions. The coding techniques should take advantage of the signal redundancy or cleverly degrade the information in such a manner that the observer does not notice the degradation. The position taken in the coding analysis is a unified viewpoint which regards the television system as an information source in the Shannon sense. The redundancy elimination techniques which are considered result in the original information source being changed into a new source, the entropy of which hopefully becomes less than that of the original source. To measure the effectiveness of this approach, statistics are required on both the original and the modified sources. The statistics will vary depending upon both the coding technique and the basic pictorial content as modified by the television sensor. The redundancy reducing techniques to be discussed will be classified as either information-preserving or information-destroying techniques.

Information-preserving techniques are completely reversible processes as far as the encoder and decoder are concerned and the only error introduced in the system, assuming a noiseless channel, is the quantizing error in the digital converter. The efficiency of an information-preserving technique that utilizes redundancy elimination depends upon the statistics of the digital source. Statistics are necessary to determine the correlation between adjacent elements, successive adjacent elements along a line, adjacent elements between successive lines, adjacent elements in an area, and adjacent elements between successive frames. The statistics that are required to calculate these correlations are the first-order and second-order probabilities $p(i, j)$. Although it is possible to obtain the remaining statistics from these probabilities, certain other probabilities can be defined and measured on EDITS that will be extremely useful for analyzing some coding techniques. For instance, the previous element probability is defined as the probability that a picture element is at the same intensity level as the immediately preceding one. This probability, designated PE (previous element probability) is necessary to determine the efficiency of previous-element coding (PEC) techniques.

Extending the previous-element probability concept to many successive picture elements requires obtaining the transition-probability distribution $p_i(j)$, the probability that a given picture element is at level j given that the immediately preceding picture element is at level i . This probability enables the efficiency of a coding technique that uses the correlation between many successive elements, such as run-length coding, to be determined.

To evaluate the efficiency of coding techniques that utilize line-to-line correlation, it is necessary to determine the statistical relationship between adjacent picture elements on successive lines; that is,

$$P_{y,i}(y+1), j \quad x = k$$

elements the probability that an element in the k th position of line $y + 1$ is at level j , given that the element on line y is at level i . This probability will be used in a coding technique that utilizes previous-line correlation such as the previous line-

previous element (PEPL) coding technique. Although the correlation between successive elements along a line or between lines is high, the correlation between elements in an area is even higher. Hence, it would be necessary to determine the area correlation to evaluate an area coding technique.

The final statistic necessary to evaluate the efficiency of the information-preserving techniques is the correlation between successive frames. Given the probability that an element at point x, y is at level j ; that is,

$$P_{x, y, f, 1}(x, y), (f + 1), j$$

A system that would utilize the majority of the statistics mentioned would be an elastic system that on a short-term basis, matches the coding technique to the source statistics. This would then be an optimum information-preserving coding technique.

Although the statistics are not as well defined for the information-destroying techniques, the addition of certain information-preserving coding schemes would enable the over-all efficiency of the techniques to be improved. For example, Roberg's technique of nonlinear amplification with the addition and subtraction of pseudo-random noise uses the physiological aspects of the viewer and the picture redundancy to obtain an efficient digital system. However, it is possible to apply any of the information-preserving coding techniques to Robert's video source in expectation of improving the efficiency.

Another information-destroying technique is delta modulation which uses the correlation between successive elements indirectly. Since the correlation between successive elements is high, only the difference between adjacent elements is to be transmitted in a delta system. The efficiency of this system could also be improved by statistically determining the correlation between the delta approximation of the original and the original itself. Possibly this correlation would lead to an increase in efficiency; that is, by applying an information-preserving technique to a delta-modulation source the over-all system efficiency could be improved. Other information destroying techniques could utilize information-preserving coding techniques. One of these is the split-band coding system in which the original video signal is separated into its low and high spectral components. Both components may be regarded as information sources possessing both statistics and redundancy to be removed by an information-preserving coding technique.

Statistics will enable an evaluation of each coding technique; however, to compare the many different coding techniques it is necessary to compare all techniques on a common basis. For example, the average number of bits per element (coded entropy) is one common parameter that could be used since compression is related to the coded entropy; however, the compression also must be relative to some basic video system with a basic number of bits per element. With the statistics specified it is possible to apply optimum coding to all the previously mentioned coding techniques. An optimum coding technique with the prefix

property that is quite efficient is the Huffman code. The Huffman code depends directly upon the statistics of the source. Should the source statistics vary, the Huffman code should change to remain optimum. Therefore, it is necessary to determine a Huffman code that is optimum for a specified set of source statistics and near optimum if the source statistics should change. Such a code can be determined and the resulting code fortunately simplifies the error analysis considerably. The combination of parameter reduction, redundancy eliminations and efficient (Huffman) coding results in an extremely efficient digital television system.

2.1.2 Definition of Scenes

For a typical manned spacecraft mission it is possible to classify the events or scenes that may be viewed with a television sensor into four categories. These are:

1. Interior scenes in the spacecraft
2. Exterior scenes of the spacecraft
3. Discovery or exploration scenes
4. Standard scenes or test patterns

For each category some typical or expected scenes that may be encountered are given below. The first category may include scenes such as (1) a full-face view of a control panel, (2) an oblique view of a control panel with an operator, (3) a wide-angle view of the cabin, (4) a close-up of a crew member, (5) a view of systems within the spacecraft, (6) a view of experiments performed, and (7) microscopic views.

In the second category of scenes some possible views are (1) an exterior view of the spacecraft, (2) monitoring of a reactor on a boom, (3) antenna and sensing equipment orientation, (4) damage assessment of the spacecraft, (5) monitoring of repairs to exterior of spacecraft, and (6) rendezvous operations.

The third category may contain scenes such as (1) planetary approach, (2) orbital reconnaissance, (3) re-entry, (4) landing, (5) countryside surveillance, and (6) crew member exploring the planet.

The scenes for the fourth category are the standard test patterns for adjustment and alignment. However, this category will also contain "standard" scenes used by previous investigators in determining system performance. Included in this latter classification will be typical scenes such as the "MIT Girl" and "crowd" pictures and the Bell Laboratory standard scenes.

In view of the vast number of scenes that could be processed, several scenes from each category were chosen for the analysis. Included in these were interior scenes of an existing spacecraft, mockups of proposed spacecraft, exterior view of a spacecraft, planetary scenes, and several standard scenes. These were

processed on EDITS in the form of either positive prints, positive transparencies, or live scenes. The nomenclature and classification of the scenes are summarized in Table 2.1.2-1.

Table 2.1.2-1

SCENES FOR STATISTICAL AND SUBJECTIVE ANALYSIS

Scene	Category	Description	Type
1	1	Apollo 1	Positive transparency
2	1	Apollo 2	Positive transparency
3	1	Cooper 1	Positive transparency
4	1	Cooper 2	Positive transparency
5	1	Cooper 3	Positive print
6	1	Laboratory scene	Live
7	2	Telescope	Positive print
8	3	Emission nebula	Positive print
9	3	Moon	Half-tone reproduction
10	4	Test chart	Positive print
11	4	Girl	Positive print

These scenes are shown in the photographic section of 6.2 and the results of the subjective and statistical analysis are given in Section 6.2.

2.1.3 Grade of Service

In general a "grade" is a simplified designation representing a more detailed specification. For instance, a system or object may be specified by n parameters $p_1, p_2, \dots, p_i, p_n$. If all the parameters $p_i, i = 1, \dots, k$, are held fixed and parameter p_k is allowed to take on values $v_{k-\min} \leq v_k \leq v_{k-\max}$, then this range of parametric values may be broken up into G regions and each system whose k^{th} parameter falls within a particular region G_j is said to be a "grade j " system. Another grading system is conceptually possible where a value trade-off among the parameters p_n is justifiable. Let such system value V_s be a function $V_s(p_1, p_2, \dots, p_n)$ of the n system parameters and let it also be restricted in range $V_{s-\min} \leq V_s \leq V_{s-\max}$. This range may be divided into G regions also and a system k whose system value V_{sk} falls within region G_j is also said to be a "grade j " system. Finally, it is emphasized that a "grade" is a simplification and that if the original parameters are available then there is little use in specifying a system grade. Nevertheless, the concept of a simplified system grade is useful when it is necessary to designate in an approximate manner many systems which are to be considered simultaneously.

With respect to digital television, a system is defined largely by specifying scan resolution R , quantization Q and frame rate F . There exists no simple system-value function V_s which relates trade-offs between R , Q and F . Therefore, the approach has been taken here to "grade" a system on the basis of individual grades for each of the parameters R , Q and F . The basis for starting the grading system is approximately commercial standards: $R = 512$ lines per picture, $Q = 6$ bits, and $F = 30$ frames per second. These three parametric values are given the grade A and the system is graded AAA, which is possibly more familiar in concept than "512-6-30." Another rule, rather arbitrarily chosen, but borne out approximately in experiments on EDITS, is that a factor of two degradation is required to be a "really noticeable" degradation. Thus, grade B is 256-line resolution, or 4 bits quantization or 15 frames per second. Normally if 6 bits is taken for grade A quantization, then 5 bits should constitute grade B. However, there appears to be a just barely noticeable difference in five- and six-bit quantization so that actually four-bit quantization is about one-half the noticeably maximum quantization. Grade C is 128-line resolution or 3 bits quantization or 7.5 frames per second. Note that these grades AAA through CCC could apply to systems of arbitrary R , Q , and F if the definition were made that a system parameter grade is determined by that grade point parameter value to which it is closer. For instance, a system with $R = 400$, $Q = 5$, $F = 10$ would be a grade AAC system. The parameters R , Q and F have not been carried below 128, 3, and 7.5, respectively, as these are considered to be approximately the lowest usable values for any system which claims to be a moving-target, half-tone television system. For certain special pictorial cases (e. g., sun spots) involving high-brightness, high-contrast detail, two-bit quantization may suffice. However, for general pictorial material this quantization is so coarse as to cause loss of resolution because the optics/camera system modulation transfer function falls below a small fraction of the quantization interval.

Table 2.1.3-1 lists the 27 conceptual digital television systems from grade AAA through CCC. Although there exists no useful system value function V_s there certainly exists a meaningful system cost function C_s given by

$$C_s = R^2 Q F, \quad (1)$$

where R , Q and F are as defined previously. The units of C_s are bits per second and, given an appropriate signal-to-noise ratio, represent the communication channel capacity necessary to transmit the requisite picture grade. This is tabulated in Column 5. This system cost function C_s is for an uncompressed system. Another system cost function C_{sc} , listed in Column 7 and given by Equation 2, is defined for a compressed digital TV system.

$$C_{sc} = \frac{R^2 Q F}{\bar{C}} \quad (2)$$

TABLE 2.1.3-1

ANALYSIS OF GRADE SYSTEMS

System	R	Q	F	C _a	PE	C _{ac}	C _o	Remarks
AAA	512	6	30	47,185,920	0.4595	33,365,164	1,414	Commercial grade
AAB			15	23,592,960	0.4595	16,682,582	2,828	
AAC			7.5	11,796,480	0.4595	8,341,291	5,656	Commercial grade but slow-or no moving targets
ABA		4	30	31,457,280	0.8452	12,733,906	3,705	
ABF			15	15,728,640	0.8452	6,366,953	7,411	High-resolution spacecraft system
ABC			7.5	7,864,320	0.8452	3,183,476	14,820	
ACA		3	30	23,592,960	0.9230	9,680,191	4,874	Minimum rendition of fast-moving targets over a wide field of view
ACB			15	11,796,480	0.9230	4,840,095	9,748	
ACC			7.5	5,898,240	0.9230	2,420,047	19,490	Good resolution of high-contrast, slow-moving targets
BAA	256	6	30	11,796,480	0.4150	8,866,234	5,321	
BAB			15	5,898,240	0.4150	4,433,117	10,640	
BAC			7.5	2,949,120	0.4150	2,216,558	21,280	Good rendition of slow-moving targets
BBA		4	30	7,864,320	0.7743	3,741,057	12,610	
BBB			15	3,932,160	0.7743	1,870,528	25,220	Manned space craft system
BBC			7.5	1,966,080	0.7743	935,264	50,450	Moderate definition of high-contrast, fast-moving targets
BCA		3	30	5,898,240	0.8735	2,712,010	17,390	
BCB			15	2,949,120	0.8735	1,355,005	34,790	
BCC			7.5	1,474,560	0.8735	678,092	69,590	Minimum rendition of slow moving targets in a moderate field of view
CAA		6	30	2,949,120	0.3466	2,418,278	19,510	
CAB			15	1,474,560	0.3466	1,209,139	39,020	
CAC			7.5	737,280	0.3466	604,569	78,040	Good rendition of slow-moving targets in a restricted field of view
CBA		4	30	1,966,080	0.6591	1,161,756	40,610	
CBB			15	983,040	0.6591	580,878	81,230	Low-resolution spacecraft system
CBC			7.5	491,520	0.6591	290,439	162,400	
CCA		3	30	1,474,560	0.7836	810,565	58,210	Lowest-bit-rate, fast-moving target system
CCB			15	737,280	0.7836	405,282	116,400	
CCC			7.5	368,640	0.7836	202,641	232,800	Minimum half-tone, moving target system

R, Q and F are as before and $\bar{C}$ is the "average" compression of the system. Within certain assumptions $\bar{C}$ can be defined by

$$\bar{C} = \frac{Q}{PE + (Q + 1) (1 - PE)} \quad (3)$$

where Q is quantization and PE is the probability that a picture element is like its previous neighboring picture element to within one quantization level. It can be readily appreciated that for a given picture-optics-camera combination PE will be a function of both Q and R with the expected relation being that a decrease in PE should result from either a decrease in R or an increase in Q or both in combination. Algebraic simplification of Equation 3 results in

$$\bar{C} = \frac{Q}{1 + (1 - PE) Q} \quad (4)$$

Combining Equations 4 and 2 gives

$$C_{sc} = R^2 F [1 + (1 - PE) Q] \quad (5)$$

Considering PE functionally, Equation 5 becomes

$$C_{sc} = R^2 F \{1 + [1 - PE(R, Q)] Q\} \quad (6)$$

with the function PE (R, Q) being experimentally measured on EDITS (see Table 2.1.3-2 for an "average picture, that labeled "Cooper 2" in Section 6.2.

Table 2.1.3-2
PE MEASUREMENTS ON THE PICTURE "COOPER 2" FOR VARIOUS
R AND Q VALUES

R^2	16,384	65,536	262,144
R	128	256	512
6	5,680	27,200	120,480
4	10,800	50,750	221,570
3	12,840	57,250	241,980

Number of elements per frame like their previous neighboring element.

The function $PE(R, Q)$ is tabulated in Column 6 of Table 2.1.3-1 and the resulting C_{sc} is tabulated in Column 7.

Another compression function C_o may be defined. C_o is the "over-all compression" of the given system plus its PE encoding when its C_{sc} is compared to the C_s of the AAA system. In this definition "compression" is considered to come about by information degradation in combination with true "scientific," or bits-per-element type compression. C_o is given by

$$C_o = \frac{47,185,920}{C_{sc}} \quad (7)$$

This function is tabulated in Column 8 of Table 2.1.3-1. Values of C_o range as high as 232.8; and it is just possible that for some restricted uses a CCC television system may be adequate at a saving of 232.8 to 1 over a straight AAA system. It is to be emphasized again the C_o does not represent true compression in the scientific sense. Any compression claims in excess of about six to one should be very closely scrutinized to ensure that such claims are not including picture-parameter degradation as part of the compression claims.

Column 9 of Table 2.1.3-1 lists a few comments which speculate on the areas of possible utility for some of the graded TV systems. It is seen that the system recommended to MSC is the "median" or BBB system. The resolution and frame rate of this system are not far from those of the Apollo 278-line, 10-frame-per-second system.

It is interesting to place the 27 graded TV systems of Table 2.1.3-1 in decreasing order C_s and C_{sc} . This is done in Figure 2.1.3-1. From this figure one may choose an available channel bit rate and then locate all those digital television systems, compressed or uncompressed, that will operate within the available channel bit rate. Notice that many of the uncompressed systems have bit rates identical with bit rates of one or two other systems. Even in the compressed case several systems of diverse parameters have relatively the same bit rate requirement. From Figure 2.1.3-1 the system designer may obtain very quickly a feeling for the R, Q, F trade-offs possible within a given bit rate. By reference to the right-hand column, he may also assess very quickly the utility of a simple (PE) compression system.

2.1.4 Coding Bibliography

A rather comprehensive survey of the field of digital video data compression techniques has been conducted as part of the EMR company-sponsored digital video program. These results are presented here because of their current interest. The techniques were classified as either a digital compression technique or a digital compression technique

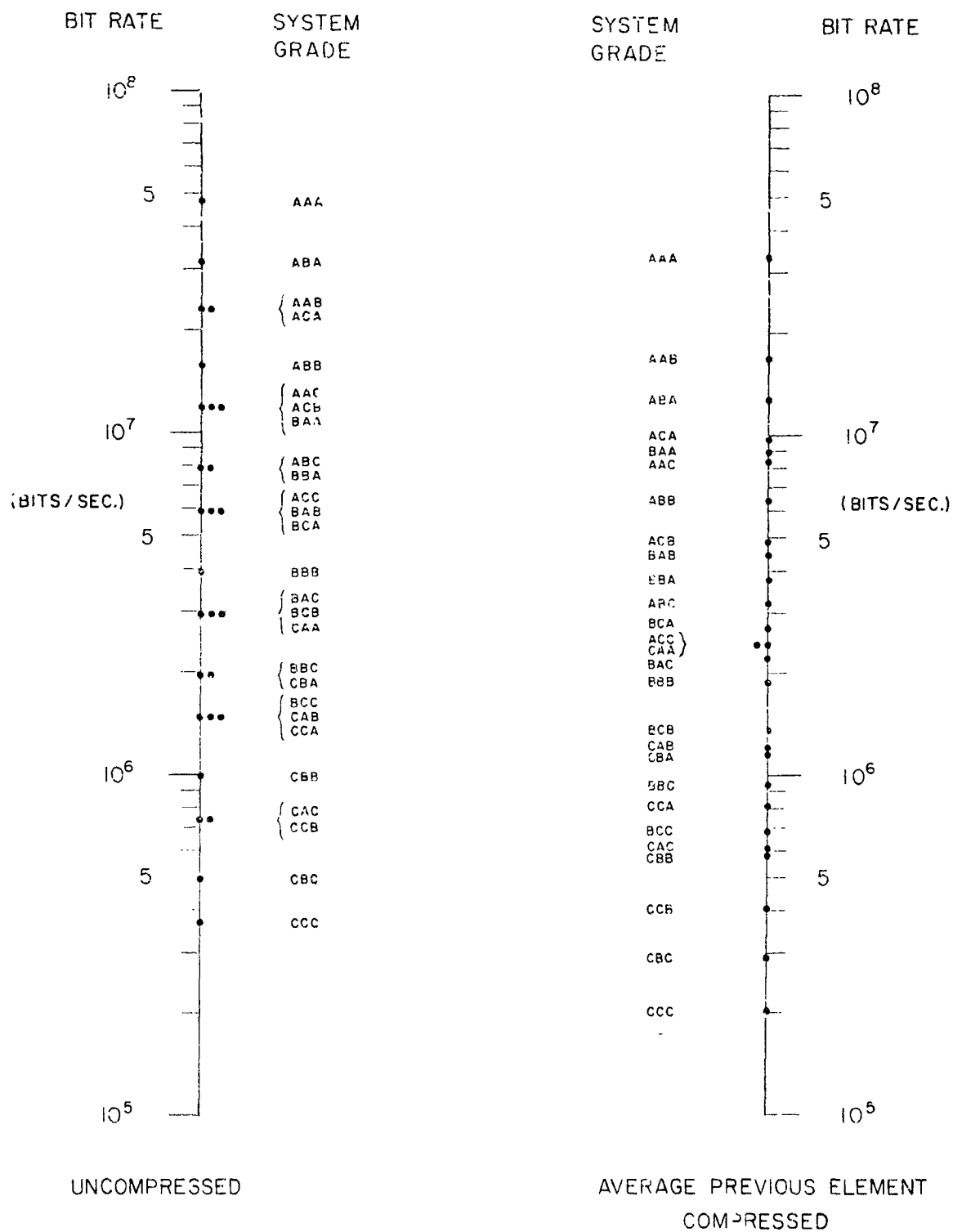


FIGURE 2.1.3-1
ANALYSIS OF GRADED SYSTEMS

with analog techniques. Table 2.1.4-1 lists 54 digital video-compression techniques that have been either theoretically proposed, experimentally determined, computer simulated, or actually implemented. The table with its associated bibliography is intended to allow a quick look at the field of digital video data compression.

Some of the more promising techniques will be discussed in detail in Sections 2.2 through 2.6. The table does not attempt to normalize the compression ratios claimed to some common parameter, but merely lists the compression ratio claimed. In general, the compression ratios listed are the gross amount; that is, the effects of noise and synchronization were not considered. Certain compression techniques may require a rather sizable increase in signal-to-noise ratio, thus reducing the gross compression considerably. Synchronization also has the possibility of decreasing the gross compression such that the net compression is considerably less than the gross compression.

2.1.5 Bibliography for 2.1

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TABLE 2.1.4-1
DIGITAL VIDEO COMPRESSION AND RELATED SYSTEMS

Systems	Investigator	Year	Country	Type	Method	Compression	Remarks
Digital with Analog Techniques	Gouriet	1956	England	Brightness- position coding	Theoretical	8.3:1	Variable-velocity scanning
	Gouriet	1957	England	Constant slope	Computational	6:1	Variable-velocity scanning
	Graham	1960	U. S.	Predictive interpolation	Experimental	3:1	...
	Schaphorst and Remm	1961	U. S.	Area encoding	System	1.63:1	For high-contrast picture; 5.5:1 for low picture detail; variable velocity scan
	Graham, R. E.	1958	U. S.	Differential quantizing	Computer simulation	2:1	...
	Youngblood	1958	U. S.	Piecewise linear approximation	Computer simulation	1.8:1	For high-picture detail while 4.6:1 for low picture detail
	Kharkevich	1958	U. S. S. R.	Co-ordinate interceptions	Theoretical	6:1	For two-level video and simple line drawings
	Deriugin	1958	U. S. S. R.	Statistical encoding	Theoretical	2:1	...
	Michel	1958	U. S.	Run-length coding	Theoretical	10:1	For two-level video with optimum encoding
	Novik	1958	U. S. S. R.	Optimum coding	Theoretical	...	Optimum variable-length coding

TABLE 2.1.4-1

DIGITAL VIDEO COMPRESSION AND RELATED SYSTEMS (Cont'd)

Systems	Investigator	Year	Country	Type	Method	Compression	Remarks
Digital with Analog Techniques (continued)	Novik	1959	U. S. S. R.	Optimum coding	Theoretical experimental	...	Decoding of variable-length codes
	Coombs	1959	U. S.	Delta modulation	Experimental System	2:1	...
	Schreiber	1959	U. S.	Split-band	Experimental system	4:1	...
	Corradetti	1959	U. S.	Planar prediction	Computer simulation	1.8:1	For picture of high detail while 7.6:1 for low-detail picture
Digital with Analog Techniques (continued)	Capon	1959	U. S.	Run-length coding	Theoretical	2:1	For two-level video
	Jules	1959	U. S.	Edge-detection	Computer simulation	2.3:1	2.9 for lower picture content
	Newell and Geddes	1961	England	Quantized video	System	2:1	Subjective results not as good as low-pass filtering for compression
	Wholey	1961	U. S.	Predictive coding	Computer simulation	3:1	For two-level video, 7.8:1 with pattern recognition

TABLE 2.1.4-1

DIGITAL VIDEO COMPRESSION AND RELATED SYSTEMS (Cont'd.)

Systems	Investigator	Year	Country	Type	Method	Compression	Remarks
Digital with Analog Techniques (continued)	Roberts	1961	U. S.	Pseudorandom noise	Computer simulation	2:1	If some degradation acceptable, 3:1 possible
	EMR	1961	U. S.	Redundancy removal	Theoretical	2.3:1	Previous-element, line, and frame-redundancy removal
	Gabor and Hill	1961	England	Contour interpolation	Experimental system	2:1	4:1 predicted, but only measured 2:1 with variable- velocity scanning
	Maurice and Monteath	1961	England	Cable film	System	1:1	660:1 with picture degradation
	Gluss	1961	U. S.	Optimum encoding	Theoretical	...	Uses curve fitting for compression
	Dobren	1962	U. S.	Run-length coding	Experimental system	2:1	For high detail picture; 10:1 for low detail black and white picture
	Cherry	1962	England	Elastic encoding	Experimental system	2.85:1	Variable-velocity scan
	Pratt	1962	U. S.	Stop-scan edge detection	Experimental system	6:1?	Variable-velocity scan

TABLE 2.1.4-1

DIGITAL VIDEO COMPRESSION AND RELATED SYSTEMS (Cont'd)

System	Investigator	Year	Country	Type	Method	Compression	Remarks
Digital	Laemmel	1951	U. S.	Differential co-ordinate coding	Theoretical	5:1	Two-level video and line drawings
	DeJager	1952	Netherlands	Delta modulation	Theoretical	...	...
	Schouten, et al	1952	Netherlands	Delta modulation	Theoretical	...	...
	Harrison	1952	U. S.	Linear prediction	Experimental system	2.3:1	Translated by coding to bandwidth saving
	Deutsch	1956	U. S.	Run-length coding	Theoretical	2:1	For two-level video with optimum encoding
	Hinckfuss, I. C.	1956	England	Delta modulation	Experimental	2:1	...
	Kretzmer		U. S.	Split-band	System	...	Pictures degraded
	Tsukkerman	1957	U. S. S. R.	Co-ordinate transmission	Theoretical	...	Only general results
	Powers and Staras	1957	U. S.	Statistical encoding	Experimental	2:1	...
	Cunningham	1957	U. S.	Two-dimensional prediction	Computer simulation	2.6:1	...

TABLE 2.1.4-1

DIGITAL VIDEO COMPRESSION AND RELATED SYSTEMS (Cont'd.)

Systems	Investigator	Year	Country	Type	Method	Compression	Remarks
Digital (continued)	Schreiber and Knapp	1958	U. S.	Split-band	Experimental system	4:1	...
	Clinger and Ziemer	1961	U. S.	Predictive coding	Experimental	7:1	Huffman coding
	Altman and Schaphorst	1962	U. S.	Area coding	Experimental	1.63	For high detail picture; 8:1 low detail picture
	Swerling	1962	U. S.	Contour statistics	Theoretical	...	No results reported from theoretical presentation
	Van Blerkom, et al	1962	U. S.	Split-band	Computer simulation	...	General results reported
	Eisenberger, et al	1962	U. S.	Line following	Experimental	10:1	Black-and-white pictures
	Stiffler	1962	U. S.	First differences	Theoretical	...	Mathematical analysis
	Blasbalg and Van Blerkom	1962	U. S.	Message compression	Theoretical	...	General discussion of compression; no results reported
	Easterling	1962	U. S.	Quantiles	Theoretical	...	...
	Stiffler	1962	U. S.	Optimum coding	Theoretical	2:1	...

TABLE 2.1.4-1

DIGITAL VIDEO COMPRESSION AND RELATED SYSTEMS (Cont'd.)

Systems	Investigator	Year	Country	Type	Method	Compression	Remarks
Digital (continued)	Rostrom	1962	U. S.	Floating aperture	Theoretical	...	...
	Eisenberger	1962	U. S.	Quantized floating aperture	Theoretical	...	...
	Cotton	1962	U. S.	Delta modulation	Computer simulation	...	...
	Gicca	1962	U. S.	Elastic encoding system		6:1	Picture degradation in system
	Jet Propulsion Lab	1963	U. S.	Polynomial interpolation	Experimental	8:1	Requires storage
	Posner	1963	U. S.	Area encoding	Theoretical	...	...
	Knight	1963	U. S.	Previous-element experiment and coding	Computation	2:1	...
	Lockwood and Huggins	1963	U. S.	Δ^2 modulation	Experimental	...	...

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2.2 ELEMENT-TO-ELEMENT

2.2.1 PE Coding

Relatively few systems proposed for the compression of a television source can use efficient coding of the digital data. Previous-element coding uses efficient encoding to take advantage of the correlation between adjacent elements along a line. The system will transmit the absolute magnitude of an element only if it is not exactly equal to the magnitude of the immediately preceding element. If the magnitude of the element is exactly equal to the previous element, a unique symbol, designated PE, is transmitted. If the source that transmits the PE and remaining signals is efficiently encoded with an optimum code such as the Huffman (1) code, a compression of the data can be achieved. To encode the source with an optimum code for maximum efficiency, the probability distribution of the source must be known since the Huffman code depends upon the probability distribution of the source.

The Huffman code is an optimum, variable-length code that has the prefix property. The prefix property enables the set of code words to be decoded since no longer word can be an extension of a shorter word. It is optimum because no other coding technique with the prefix property can have an entropy less than the entropy of the Huffman set of code words. Since the correlation between successive elements is high, the PE probability will also in general be high; therefore, it is assigned the shortest word in the Huffman set. To illustrate the coding technique, assume that a signal is quantized to n levels and converted to an N -bit PCM signal.

The previous-element encoder with the Huffman encoder operates on the PCM source to reduce the entropy by converting the PCM source into a PE signal and a PCM signal for those elements which are not PE. The two signals are then encoded with the optimum code to reduce the entropy.

The Huffman encoder assigns code words according to the probabilities of the source. The amount of reduction depends upon the PE and $(1-PE)$ $P(j)$ probabilities. It is not possible to obtain a general expression for the probabilities of the various levels since the probabilities depend upon the picture being scanned, the resolution, and number of quantizing levels. However, if the resolution and quantization are

fixed, it is possible to determine an efficient code for a majority of the pictures that may be encountered. This code can be determined by assuming that the (1-PE) levels have a flat distribution. With this assumption it is only necessary to measure PE to obtain the efficiency of the system where the efficiency is defined as the maximum coded entropy divided by the compressed coded entropy. The maximum coded entropy is the entropy of the PCM source. The coded entropy is given by

$$H_c = \sum_{i=1}^n P(i) L(i) \quad (8)$$

where $P(i)$ is the probability of obtaining level i and $L(i)$ is the number of bits used to digitally code level i . The maximum coded entropy of the PCM source is N ; that is, $H_{\max} = N$. Therefore, the efficiency or compression of a system is

$$\bar{C} = N/H_c \quad (9)$$

The coding technique must minimize H_c for maximum compression.

A lower bound on compression can be obtained by assuming that the (1-PE) levels have a flat distribution. Under this assumption the coded entropy is merely a function of the PE probability if N is specified. In general, the (1-PE) levels will each be assigned $N + 1$ bits by the Huffman coding technique while PE will have a single bit code word. The statistical entropy or the minimum number of bits per element mathematically achievable is

$$H_s = - \sum_{i=1}^n P(i) \log_2 P(i) \quad (10)$$

This entropy is always less than or equal to the coded entropy; however, the Huffman coded entropy is the minimum entropy for a decodable set of code words. For the flat (1-PE) probability distribution the coded entropy

$$H_{cf} = N + 1 - N PE \quad (11)$$

The gross average compression is

$$C_f = \frac{H_{\max}}{H_{cf}} = N / (N + 1 - N PE) \quad (12)$$

Figure 2.2.1-1 plots the average compression as a function of the number of quantizing levels in bits parametric in PE while Figure 2.2.1-2 plots the average compression as a function of PE parametric in the number of quantizing levels in bits. To achieve a compression of 2:1 for a 3-bit PCM source requires a minimum PE $\neq$ 0.833, while to achieve a compression of 2:1 for a 4-bit source requires a minimum PE = 0.75. Therefore, for the two cases considered, 83.3 and 75 percent of the picture must be redundant.

Although this gives a lower-bound on compression, higher values of compression can be achieved if the (1-PE) probabilities do not have a flat distribution. Deriugen⁽²⁾ experimentally determined that the probability distribution of a quantized television source followed a Gaussian distribution. Assume that the brightness-amplitude-probability distribution is Gaussian where levels $N/2$ and $N/2 + 1$ are most probable. The probability of level n can be determined by using the normal probability integral,

$$P(n) = \frac{1}{2\sqrt{2\pi}} \int_{\frac{-4.5}{n}}^{\frac{4.5}{n}} e^{-\frac{t^2}{2}} dt \quad (13)$$

If the normal distribution is divided into the appropriate number of intervals, the probabilities for a Gaussian distribution can be obtained. For $n = 8$ the coded entropy for a Gaussian distribution is given by

$$H_g = 3.207 - 2.207 \text{ PE.} \quad (14)$$

Figure 2.2.1-3 plots the average compression as a function of PE for a Gaussian distribution for the (1-PE) levels and $N = 3$ and $N = 6$.

To achieve a 2:1 compression for 8 levels with a Gaussian distribution requires a minimum PE of 0.773. The Gaussian distribution requires PE = 0.773 and for this value of PE and flat encoding, the compression is only 1.79:1 or a 10.5% decrease in compression by assuming a flat distribution when the actual distribution is Gaussian. For the Gaussian distribution assume the probabilities and PE that are given in Table 2.2.1-1.

The Huffman coding tree is given in Table 2.2.1-2.

To determine whether or not a coding technique will be useful, it will be necessary to determine the range of PE experimentally as a function of the picture material.

A simplified block diagram of the previous element encoder is given in Figure 2.2.1-4. Since the operation of the camera, video processor, and analog-to-digital converter are described elsewhere, only the encoder and decoder will be considered. The encoder compares successive words and transmits a PE signal to the Huffman encoder when two successive words are identical.

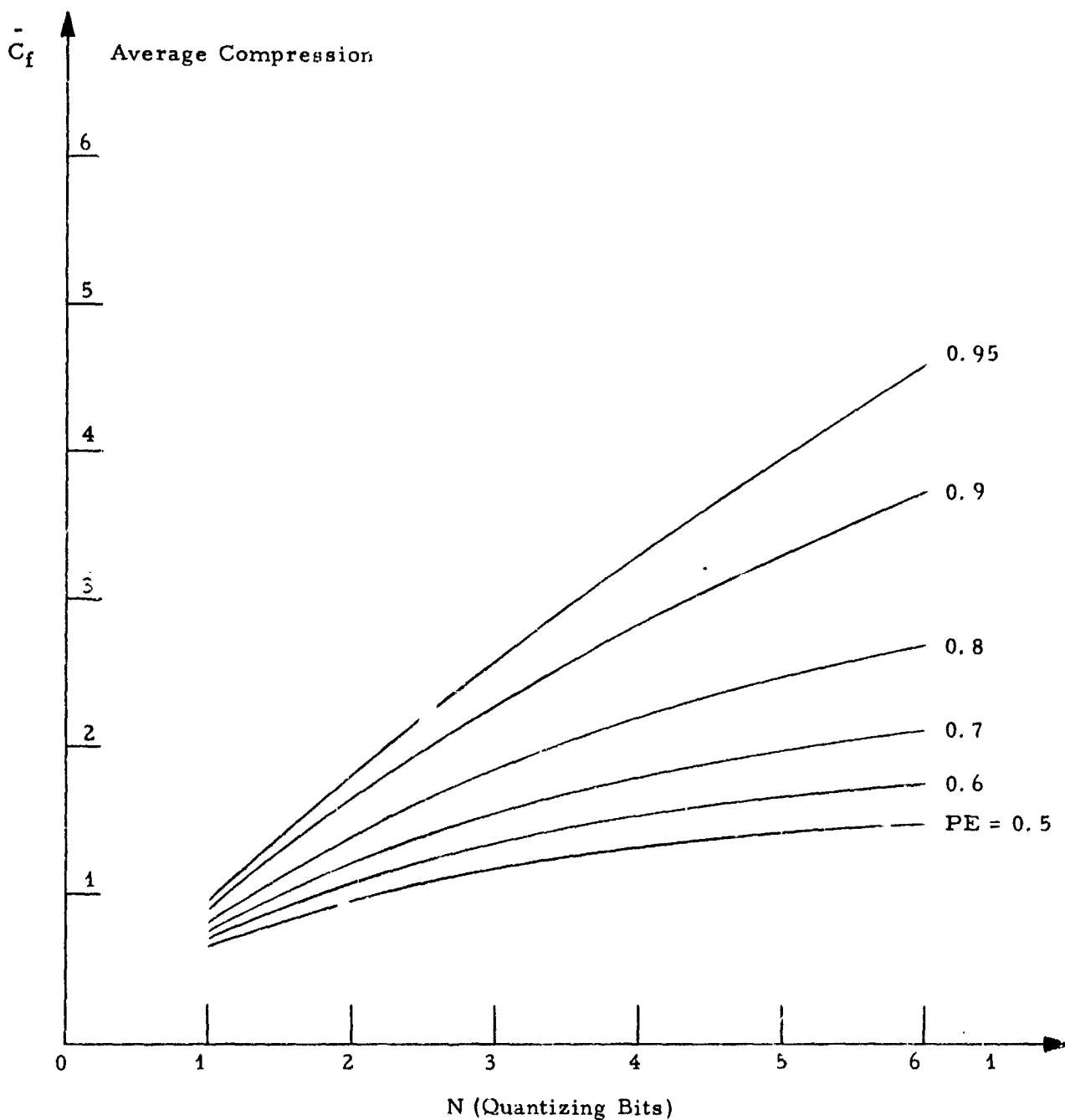


FIGURE 2.2.1-1 AVERAGE COMPRESSION AS A FUNCTION OF QUANTIZING BITS FOR VARIABLE PE ASSUMING (1-PE) LEVELS HAVE FLAT DISTRIBUTION

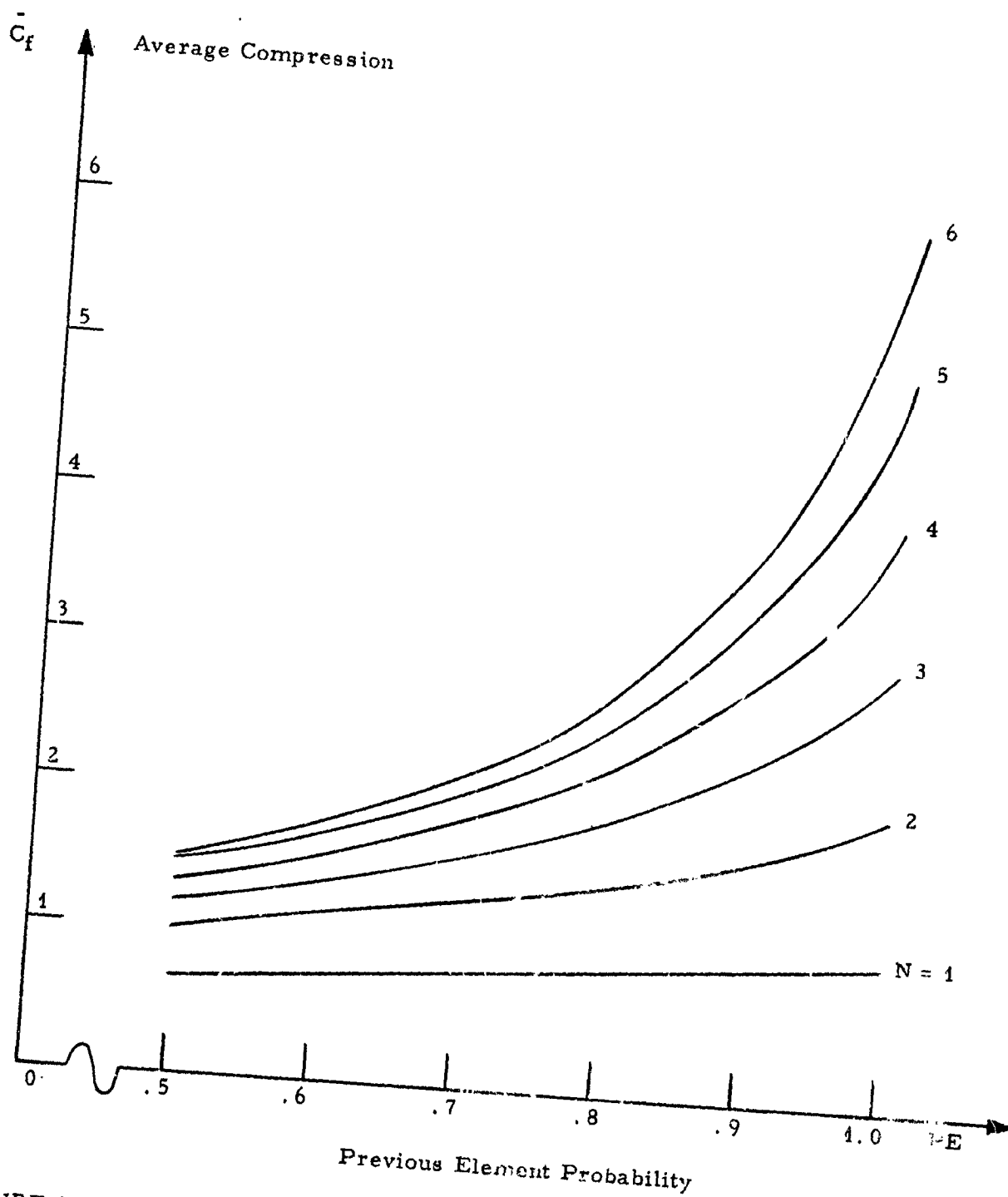


FIGURE 2.2.1-2 AVERAGE COMPRESSION AS A FUNCTION OF PE FOR VARIABLE QUANTIZATION ASSUMING (1-PE) LEVELS HAVE FLAT DISTRIBUTION

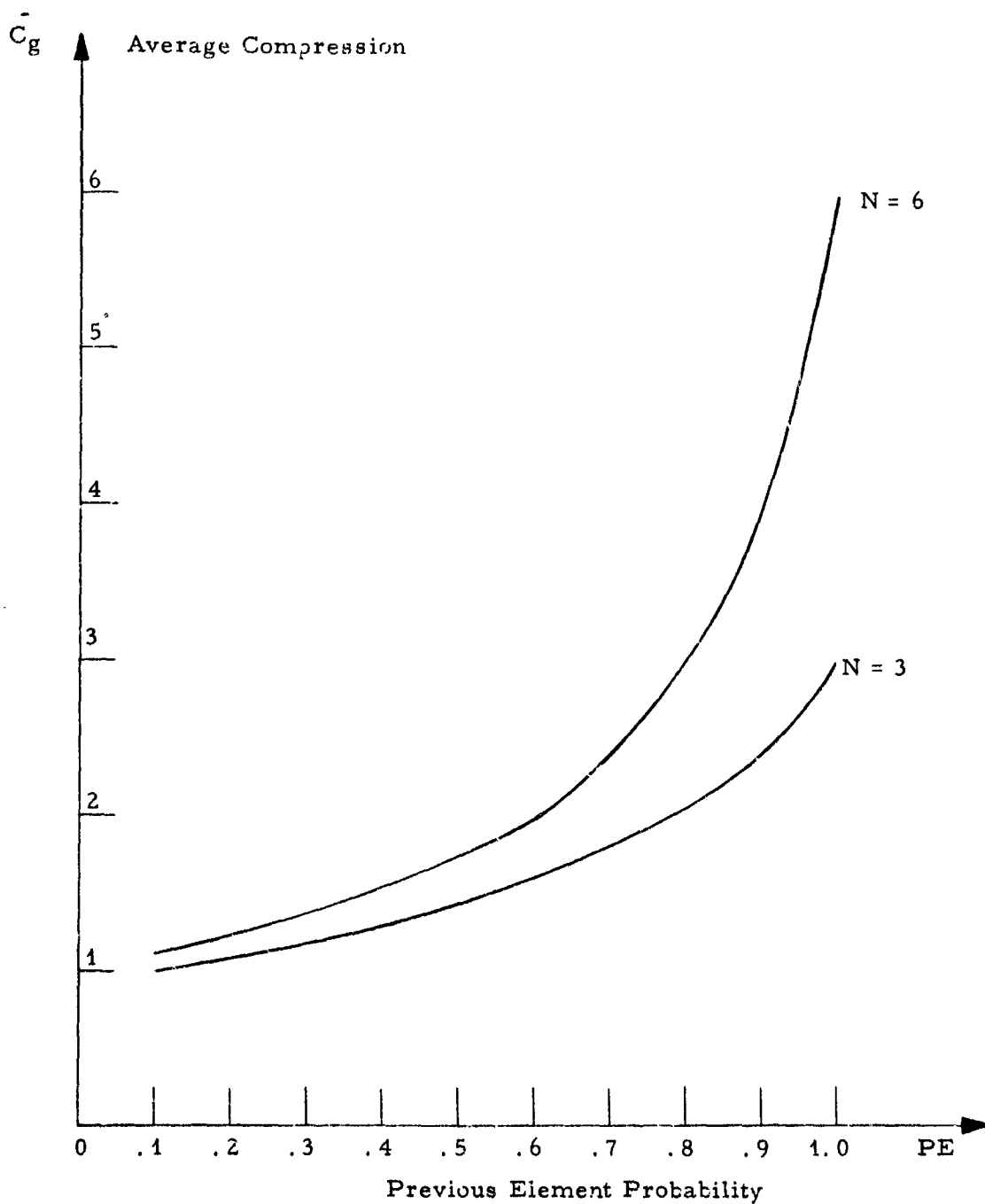


FIGURE 2.2.1-3 AVERAGE COMPRESSION AS A FUNCTION OF PE FOR VARIABLE QUANTIZATION ASSUMING GAUSSIAN DISTRIBUTION FOR (1-PE) LEVELS

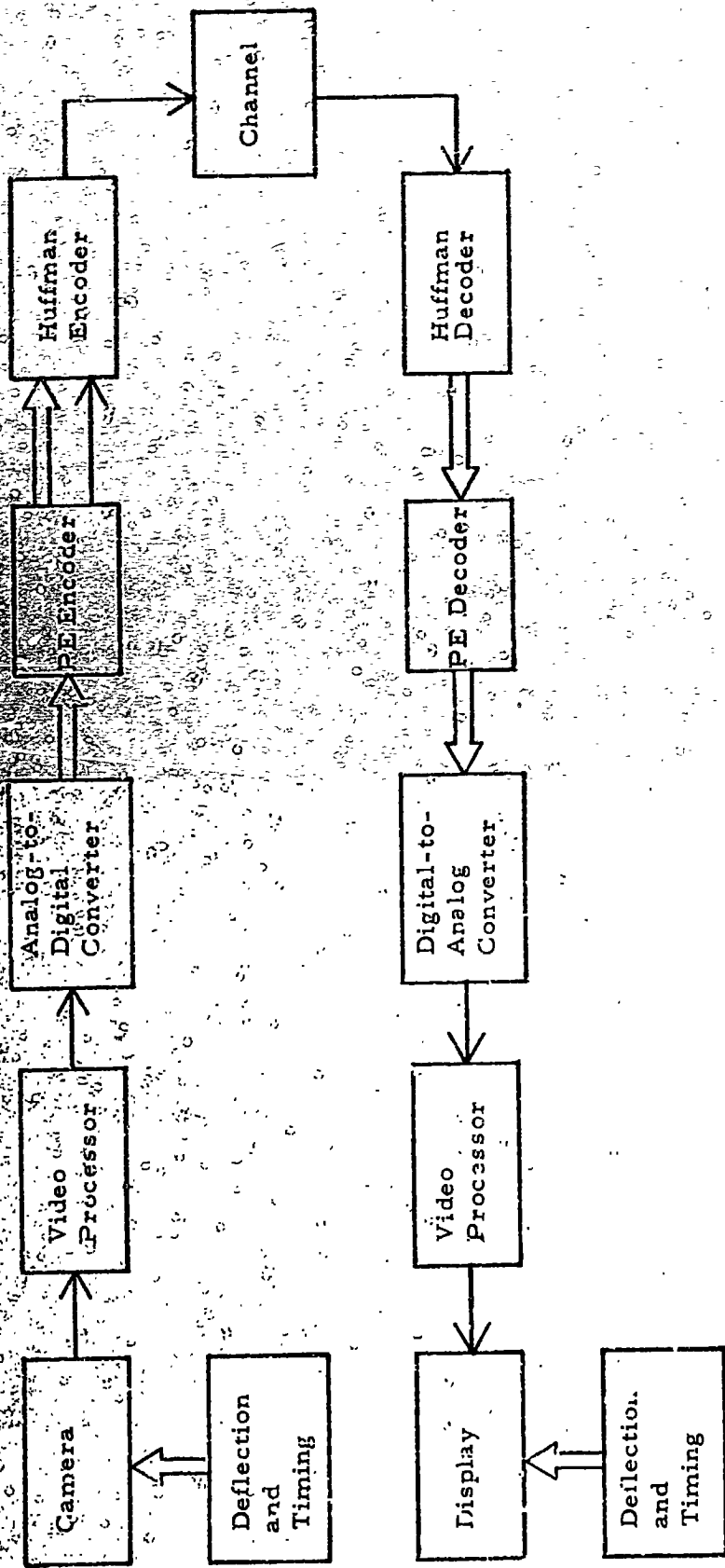


FIGURE 2.2.1-4 PREVIOUS-ELEMENT-CODING TELEVISION SYSTEM

TABLE 2.2.1-1
PROBABILITIES FOR A GAUSSIAN DISTRIBUTION

Level	Probability	Level	Probability
PE	0.848	5	0.05244
1	0.0005	6	0.02037
2	0.00304	7	0.00304
3	0.02037	8	0.00015
4	0.05244		

TABLE 2.2.1-2
HUFFMAN CODING WORDS FOR GAUSSIAN DISTRIBUTION

Probability	Huffman coding tree	Code words
PE	0.848	1
(1-PE)P(4)	0.05244	01
(1-PE)P(5)	0.05244	001
(1-PE)P(3)	0.02037	0001
(1-PE)P(6)	0.02037	00001
(1-PE)P(2)	0.00304	000001
(1-PE)P(7)	0.00304	0000001
(1-PE)P(1)	0.00015	00000001
(1-PE)P(8)	0.00015	00000000

If the second word is not like the first word in store, then it is transmitted to the Huffman encoder as an absolute magnitude. The Huffman encoder merely assigns the appropriate Huffman word to the PE and (1-PE) signals. The appropriate Huffman word to the PE and (1-PE) signals. The Huffman encoder has been investigated by Novik ⁽³⁾, Gilbert ⁽⁴⁾, and others. The encoder's function is to receive the original data at some rate r , and transmit the data to the channel at a rate r_2 where $r_2 < r$, for r_1 a constant. Since the coding technique is an information-preserving technique, the decoding operation is merely the inverse of the encoding.

The preliminary results from EDITs indicate the potential of the PE coding system. For the 11 standard pictures described in Section 2.1.2 (Table 2.1.2-1) and assuming that a 4-bit PCM signal is to be compressed, the average entropy

of the compressed source with flat encoding is 2.012 bits per element with the maximum and minimum entropies 2.562 and 1.496 bits per element, respectively. If it is further assumed that subjectively the picture quality of a 4-bit picture is acceptable, then the average compression relative to a 6-bit PCM source with flat previous element coding is 2.982 for the 11 pictures with the maximum and minimum compressions of 4.011 and 2.343, respectively. A general analysis of the Huffman coding technique and some preliminary effects of errors on the Huffman code is given in Section 2.6. Present indications are that for reasonable error rates and real-time transmission the effects of errors in the previous-element coding system will not appreciably degrade the displayed picture. The effects of errors are quite similar to errors in a delta-modulation system where a single error will cause streaks versus the single error in a PCM system causing a "salt-and-pepper" effect in the displayed image. However, compressions in the order of 2:1 are reasonable for the coding technique and the system can be implemented.

2.2.2 Run-Length Coding

The upper-bound on compression for previous-element coding is $\log_2 n$ where n is the number of quantizing levels. However, for certain probability distributions the statistical entropy is less than one indicating that compressions greater than $\log_2 n$ exist. To achieve these higher compressions the coding technique should take advantage of the correlation between many adjacent elements. Run-length coding is a two-dimensional coding technique that encodes both the amplitude of an element and the number of successive times the element at the particular amplitude repeats itself; therefore, it can be considered as an extension of previous element coding.

Several investigators have postulated that television is a higher than a first-order process; hence, it will be assumed that for run-length coding, the source is a second-order process. Using this assumption with the appropriate second-order probabilities, a lower-bound on compression can be obtained. For the coding analysis a run-length will be defined as the number of successive elements at a specified amplitude immediately preceded and followed by an element not of the same amplitude as the successive elements. To obtain the compression for run-length coding it is necessary to know the statistics of the source.

Although it would be possible to measure the run-length probabilities directly and obtain the compression, it is also possible to obtain the compression by measuring the transition probabilities and apply these to a model of the source. For the model the measurements necessary are $n(n+1)$ while for the direct run-length probabilities the measurements necessary are nR where n is the number of quantizing levels and R is the maximum run-length. For a typical system where $n = 8$, R is in the order of 10^3 . Therefore, by using the model, there are 7.936×10^3 fewer measurements required; however, using the model does require a rather lengthy evaluation of the compression expression, but this can be eliminated by solving the expression in general with a computer and obtaining the compression as a function of the transition probabilities.

Although it appears that run-length coding will result in rather large compressions, this in general is not valid for multibit coding due to the large number of code words required to specify all the possible runs of each amplitude level. However, the technique is extremely useful for a two-level source where long runs could and do exist. For example, consider a two-level source where runs to 512 could exist. Here, 1024 code words are required to specify runs to 512 of each level; therefore, each run-length will be described by a 10-bit code word. Consider now a run of 256. This run would require a 10-bit code word and a 256:1 compression would result for that portion of the picture. Unfortunately, a run of one also requires a 10-bit word and results in a 10:1 increase in bandwidth. For the system to effect compression, it is necessary to have an average run-length greater than 10.

However, since there exists the possibility of obtaining "large" compressions, consider a derivation of the compression for a model for run-length coding. Although a general derivation for an n-level video source will be presented later for illustrative purposes, assume that the picture is composed of black and white elements.

From elementary probability theory, the following probabilities can be defined:

$P(w)$ = first-order probability of obtaining a white element

$P(b)$ = first-order probability of obtaining a black element

$P_b(w)$ = transition probability of obtaining a white element given the immediately preceding element is black

$P_b(b)$ = transition probability of obtaining a black element given the immediately preceding element is black

$P_w(b)$ = transition probability of obtaining a black element given that the immediately preceding element is white

$P_w(w)$ = transition probability of obtaining a white element given that the immediately preceding element is white

$P(b, w)$ = joint probability of obtaining a black-and-white element in succession

$P(b, b)$ = joint probability of obtaining a black-and-black element in succession

$P(w, b)$ = joint probability of obtaining a white and-black element in succession

(15)

A run-length of level black y elements long will be defined by the joint occurrence of a white element followed by y black elements followed by a white element. Let this sequence be defined by the variable R_b^y . Similarly, a run-length of level white y elements long will be defined by the joint occurrence of a black element followed by a y white element followed by a black element. Let this sequence be defined by the variable R_w^y .

The run-length probability of a run-length y elements long of level black is then given by:

$$P(R_b^y) = P(w, \underbrace{b, \dots, b}_y, w) . \quad (16)$$

This can now be rewritten in terms of the first order and transition probabilities using Equation 15.

$$P(R_b^y) = k \left[P(w) P_w(b) \underbrace{P_b(b) \dots P_b(b)}_{y-1} P_b(w) \right] , \quad (17)$$

where k is a constant to be evaluated.

Equation 17 can now be written in as:

$$P(R_b^y) = k \left[P(w) P_w(b) P_b(b)^{y-1} P_b(w) \right] . \quad (18)$$

However,

$$P_b(b) + P_b(w) = 1 . \quad (19)$$

Therefore 17 becomes

$$P(R_b^y) = k \left\{ P(w) P_w(b) P_b(b)^{y-1} \left[1 - P_b(b) \right] \right\} . \quad (20)$$

The expression for the run-length probability of a white run can also be derived which is

$$P(R_w^y) = k \left\{ P(b) P_b(w) P_w(w)^{y-1} \left[1 - P_w(w) \right] \right\} . \quad (21)$$

In order to achieve the maximum possible run-length, assume that y varies from one to infinity.

Therefore

$$\sum_{y=1}^{\infty} \left[P(R_b^y) + P(R_w^y) \right] = 1 \quad (22)$$

Substituting 20 and 21 into 22 the constant k can be evaluated.

$$k \sum_{y=1}^{\infty} \left\{ P(w) P_w(b) P_b(b)^{y-1} [1 - P_b(b)] + P(b) P_b(w) P_w(w)^{y-1} [1 - P_w(w)] \right\} = 1.$$

(23)

However, for the two-level video

$$P(b, w) = P(w, b) ; \quad (24)$$

that is,

$$P(w) P_w(b) = P(b) P_b(w) . \quad (25)$$

Applying this to 23 it becomes

$$k P(w) P_w(b) \sum_{y=1}^{\infty} \left\{ P_b(b)^{y-1} [1 - P_b(b)] + P_w(w)^{y-1} [1 - P_w(w)] \right\} = 1. \quad (26)$$

It can be shown that

$$\sum_{y=1}^{\infty} a^y = \frac{a}{(1-a)} ; \quad a < 1 . \quad (27)$$

Therefore,

$$\begin{aligned} & k P(w) P_w(b) \sum_{y=1}^{\infty} \left\{ P_b(b)^{y-1} [1 - P_b(b)] + P_w(w)^{y-1} [1 - P_w(w)] \right\} = \\ & k P(w) P_w(b) \left\{ \frac{[1 - P_b(b)]}{P_b(b)} + \frac{P_b(b)}{1 - P_b(b)} + \frac{1 - P_w(w)}{P_w(w)} + \frac{P_w(w)}{1 - P_w(w)} \right\} = \\ & k P(w) [P_w(b)] (2) = 1 \end{aligned} \quad (28)$$

Solving for k,

$$k = \frac{1}{2 P(w) P_w(b)} . \quad (29)$$

Applying this constant to Equations 20 and 21,

$$P(R_b^y) = 1/2 P_b(b)^{y-1} (1 - P_b(b)) \quad (30)$$

$$P(R_w^y) = 1/2 P_w(w)^{y-1} [1 - P_w(w)] \quad (31)$$

The average compression obtained for run-length coding will be defined as

$$\bar{C} = \sum_{y=1}^{\infty} \frac{y [P(R_b^y) + P(R_w^y)]}{H [P(R_b^y)] + [H P(R_w^y)]} \quad (32)$$

where $P(R_b^y)$ and $P(R_w^y)$ are given by Equations 30 and 31, respectively.

$$C = \sum_{y=1}^{\infty} \frac{1/2 \{ y P_b(b)^{y-1} [1 - P_b(b)] + y P_w(w)^{y-1} [1 - P_w(w)] \}}{-1/2 \{ P_b(b)^{y-1} [1 - P_b(b)] \log_2 \{ 1/2 P_b(b)^{y-1} [1 - P_b(b)] \} + P_w(w)^{y-1} [1 - P_w(w)] \log_2 \{ 1/2 P_w(w)^{y-1} [1 - P_w(w)] \} } \quad (33)$$

$y = 1$ (34)

However, $\log AB = \log A + \log B$ and 33 can be rewritten as

$$\bar{C} = \sum_{y=1}^{\infty} \frac{\left\{ y \left[\frac{1 - P_b(b)}{2 P_b(b)} \right]^{y-1} (1 - P_b(b)) + y P_w(w)^{y-1} \left[\frac{1 - P_w(w)}{2 P_w(w)} \right] \right\}}{v P_b(b)^y \log_2 P_b(b) + P_b(b) \log_2 \left[\frac{1 - P_b(b)}{2 P_b(b)} \right] + \left\{ y P_w(w)^y \log_2 P_w(w) + P_w(w) \log_2 \left[\frac{1 - P_w(w)}{2 P_w(w)} \right] \right\}} \quad (35)$$

$y = 1$

Equation 27 and 36 below can be applied to 35.

$$\sum_{y=1}^{\infty} ya^y = \frac{a(2-a)}{(1-a)^2}, \quad a < 1. \tag{36}$$

Therefore

$$\bar{C} = \frac{\left[\frac{2-P_b(b)}{1-P_b(b)} \right] \left[\frac{2-P_w(w)}{1-P_w(w)} \right]}{\left\{ \frac{\left[\frac{2-P_b(b)}{1-P_b(b)} \right]}{1-P_b(b)} \log_2 P_b(b) + \log_2 \frac{\left[\frac{1-P_b(b)}{2P_b(b)} \right]}{1-P_w(w)} + \frac{\left[\frac{2-P_w(w)}{1-P_w(w)} \right]}{1-P_w(w)} \log_2 P_w(w) + \log_2 \frac{\left[\frac{1-P_w(w)}{2P_w(w)} \right]}{1-P_w(w)} \right\}} \tag{37}$$

Therefore, the average compression as given by 37 is defined by the transition probabilities $P_b(b)$ and $P_w(w)$.

Since this expression is a function of only the transition probabilities $P_b(b)$ and $P_w(w)$, it is possible to obtain a family of curves to obtain the expected compression as a function of $P_w(w)$ for constant $P_b(b)$. These curves as shown in Figures 2.2.2-1 and 2.2.2-2 show the possible compressions available using the model. The compression is at least upper-bound since the statistical entropy implies an upper-bound and it is the least lower-bound since runs to infinity were assumed.

The assumption of runs to infinity is rather impractical since the maximum run length is determined by the resolution of the system; however, it is used as a mathematical tool. For the practical case, assume that the maximum run length is limited to some value, ℓ . Consider now how this effects the compression expression previously defined. From Equation 20 and 21 the probability of obtaining a black-and-white run-length y is

$$P(R_b^y) = k \left[P(w) P_{wb} (1-P_{bb}) P_{bb}^{y-1} \right] ; \quad (38)$$

$$P(R_w^y) = k \left[P(b) P_{bw} (1-P_{ww}) P_{ww}^{y-1} \right] . \quad (39)$$

If it is assumed that the maximum run length is ℓ , the constant k can be evaluated from

$$\sum_{y=1}^{\ell} \left[P(R_b^y) + P(R_w^y) \right] = 1. \quad (40)$$

Expanding 40 out, it takes the form of

$$a + ax + ax^2 + \dots (n \text{ terms}), \quad (41)$$

which is equal to

$$\frac{(1-x^n)}{(1-x)} . \quad (42)$$

Therefore,

$$\begin{aligned} 1 &= \sum_{y=1}^{\ell} \left[P(R_b^y) + P(R_w^y) \right] \\ &= k \left[P(w) P_{wb} (1-P_{bb}^{\ell}) + P(b) P_{bw} (1-P_{ww}^{\ell}) \right]. \end{aligned} \quad (43)$$

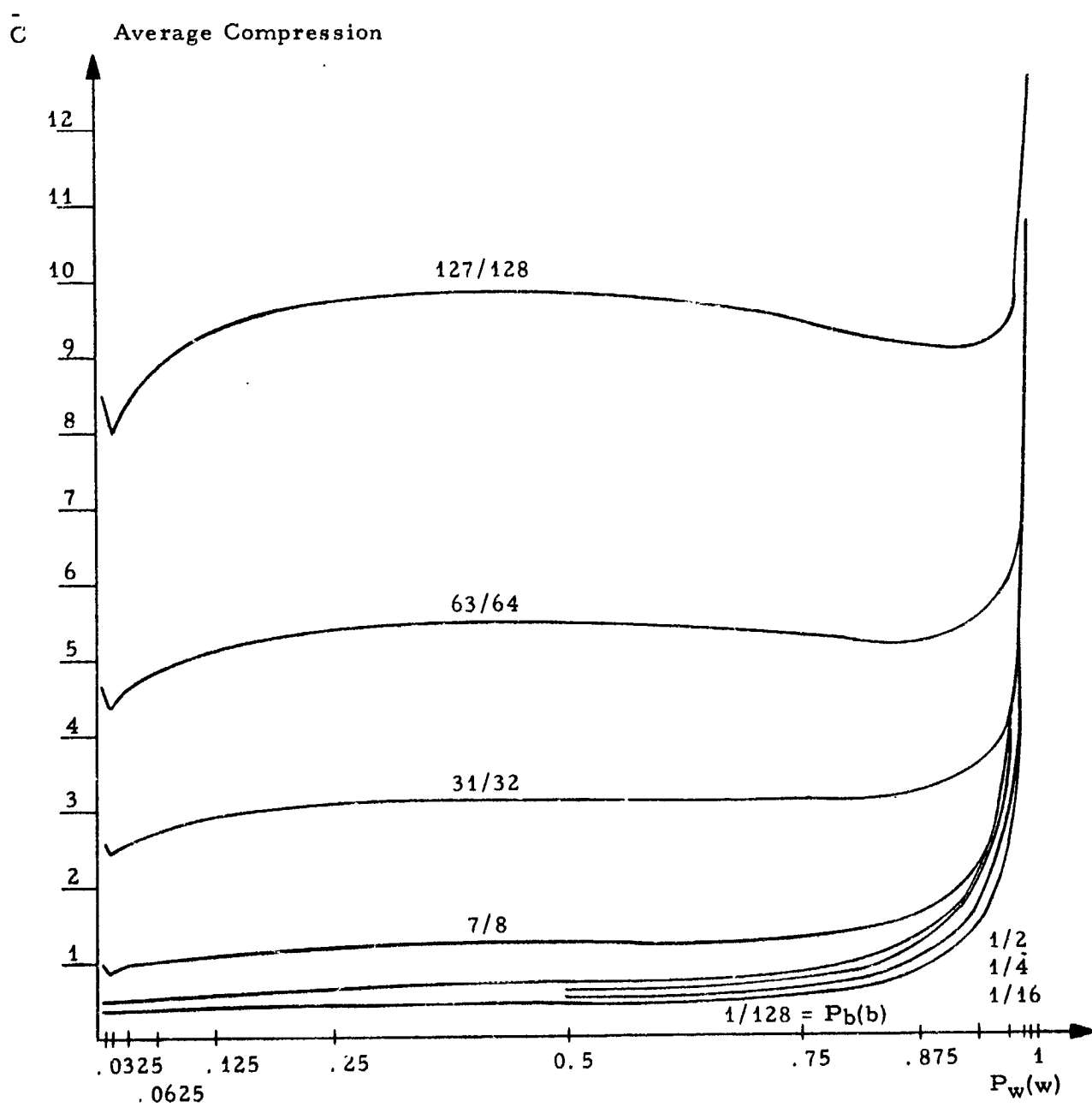


FIGURE 2. 2. 2-1 PLOT OF AVERAGE COMPRESSION AS A FUNCTION OF THE TRANSITION PROBABILITIES $P_w(w)$ FOR CONSTANT $P_b(b)$ FOR A TWO-LEVEL PICTURE

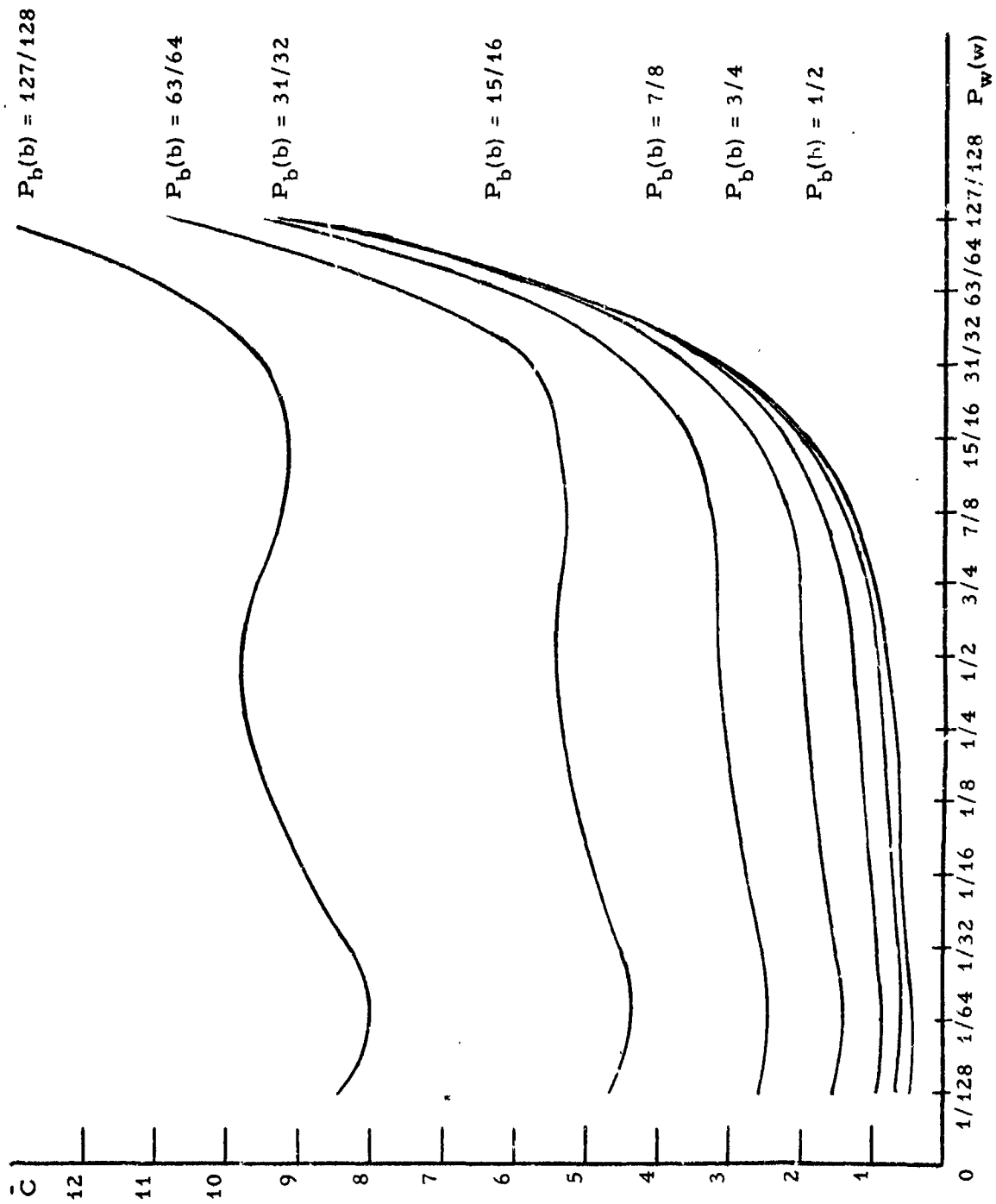


FIGURE 2.2.2-2 PLOT OF AVERAGE COMPRESSION AS A FUNCTION OF THE TRANSITION

However,

$$P(w) P_{wb} = P(b) P_{bw} . \quad (44)$$

Therefore, solving for k

$$k = \frac{1}{P(w) P_{wb} [2 - P_{bb} \ell - P_{ww} \ell]} . \quad (45)$$

Substituting this into 38 and 39 ,

$$P(R_b^y) = \left[\frac{(1 - P_{bb})}{(2 - P_{bb} \ell - P_{ww} \ell) P_{bb}} \right] P_{bb}^y \quad (46)$$

and

$$P(R_w^y) = \left[\frac{(1 - P_{ww})}{(2 - P_{bb} \ell - P_{ww} \ell) P_{ww}} \right] P_{ww}^y \quad (47)$$

Since the bracketed terms of Equations 46 and 47 are constants, the probabilities become

$$P(R_b^y) = k_1 P_{bb}^y , \quad (48)$$

and

$$P(R_w^y) = k_2 P_{ww}^y . \quad (49)$$

It is now possible to obtain the run-length probabilities for any value of ℓ , P_{bb} , P_{ww} , and y .

Assume that the transition probabilities are specified, therefore enabling calculation of the run-length probabilities. In the past the average run-length compression has been obtained by assuming or measuring an average run length. However, if a maximum run length, ℓ , is assumed, the probability set normalized to one (by truncating the run length to same value less than ℓ), higher values of compression can be predicted.

Assume that the normalization process has been performed for a maximum run length of L . Consider the effects on compression if the equipment is designed such that the maximum run length allowable is $L/2$. Therefore, for a run greater than $L/2$, say k units long where $L/2 < k < L$, this will be counted as a run of $L/2$ and a run of $k - L/2$; It is now possible to take the original probability set for runs to L and convert it to a new probability set, normalized to one, truncated at $L/2$. It is possible then to truncate the runs at any value of k for $1 \leq k \leq L$. The compression available is a function of k and there exists a k such that a maximum compression exists. Using this technique the data from EDITS were operated on and applied to Equation 23 to obtain the lower bound for compression as a function of the truncation length.

It was assumed that runs to 256 could exist and the data from EDITS were used to compute the compression for truncating the run-lengths at values less than 256. The probabilities for the chosen picture are given in Table 2.2.2-1.

TABLE 2.2.2-1 FIRST-ORDER AND TRANSITIONAL PROBABILITIES OF LINEARITY RETMA CHART TAKEN ON EDITS

$P(b) = 0.304$	$P_b(b) = 0.869$	$P_w(w) = 0.943$
$P(w) = 0.696$	$P_b(w) = 0.131$	$P_w(b) = 0.057$

The lower bound on compression for run-length coding was obtained by flat-encoding the run lengths and is tabulated in Table 2.2.2-2.

TABLE 2.2.2-2 LOWER-BOUND ON COMPRESSION FOR VARYING TRUNCATION LENGTHS

Truncation Length	256	128	64	32	16	8	4	2
Compression	1.34	1.50	1.67	1.78	1.76	1.53	1.23	0.99

Figure 2.2.2-3 graphically presents the results of terminating the run lengths at various values of k . The model and the actual statistics compare quite favorably; for example, terminating at 16, the model predicted a compression of 2.09 for run-length coding of the linearity-RETMA chart while the actual value obtained from the statistics was 2.10. Truncating the maximum runs at 8, the model predicted 1.93 while the actual value was 1.96. The probabilities obtained for these calculations are tabulated in Table 2.2.2-3.

TABLE 2.2.2-3 RUN-LENGTH PROBABILITIES
FOR MODEL TRUNCATED AT EIGHT

Run Length	Level	
	Black	White
1	0.04844	0.03692
2	0.04124	0.03464
3	0.03582	0.03272
4	0.03113	0.03084
5	0.02680	0.02906
6	0.02351	0.02740
7	0.02043	0.02582
8	0.13449	0.42068

Therefore, for the values of k chosen the model proved to be quite successful and illustrates the point that the maximum possible compression for run-length coding is a function of the terminating length. From the flat encoding it appears that for the linearity RETMA Chart, the maximum compression would be obtained if the run-length truncation is between 16 and 32. However, this value would vary as the picture statistics change. It would be possible to use the model by writing a computer program and plot compression as a function of terminating length for various values of transition probabilities. Such a plot would enable an elastic encoder to choose the optimum truncation length for maximum compression. As a further check on the model the encoded entropy and the mathematical entropy were compared. For the linearity RETMA Chart and truncating at 8 and 16, the mathematical and encoded entropy were ± 0.001 of each other. One other check was performed on the model to check its validity. The first-order probabilities were measured and were 0.304 and 0.696 from the raw data. The model reduced these probabilities to 0.311 and 0.689, respectively.

Although the model is extremely useful for a two-level source, the majority of the applications for space transmission require an n -level source. Therefore, the model will be extended and an expression for compression will be given using the first-order and transition probabilities.

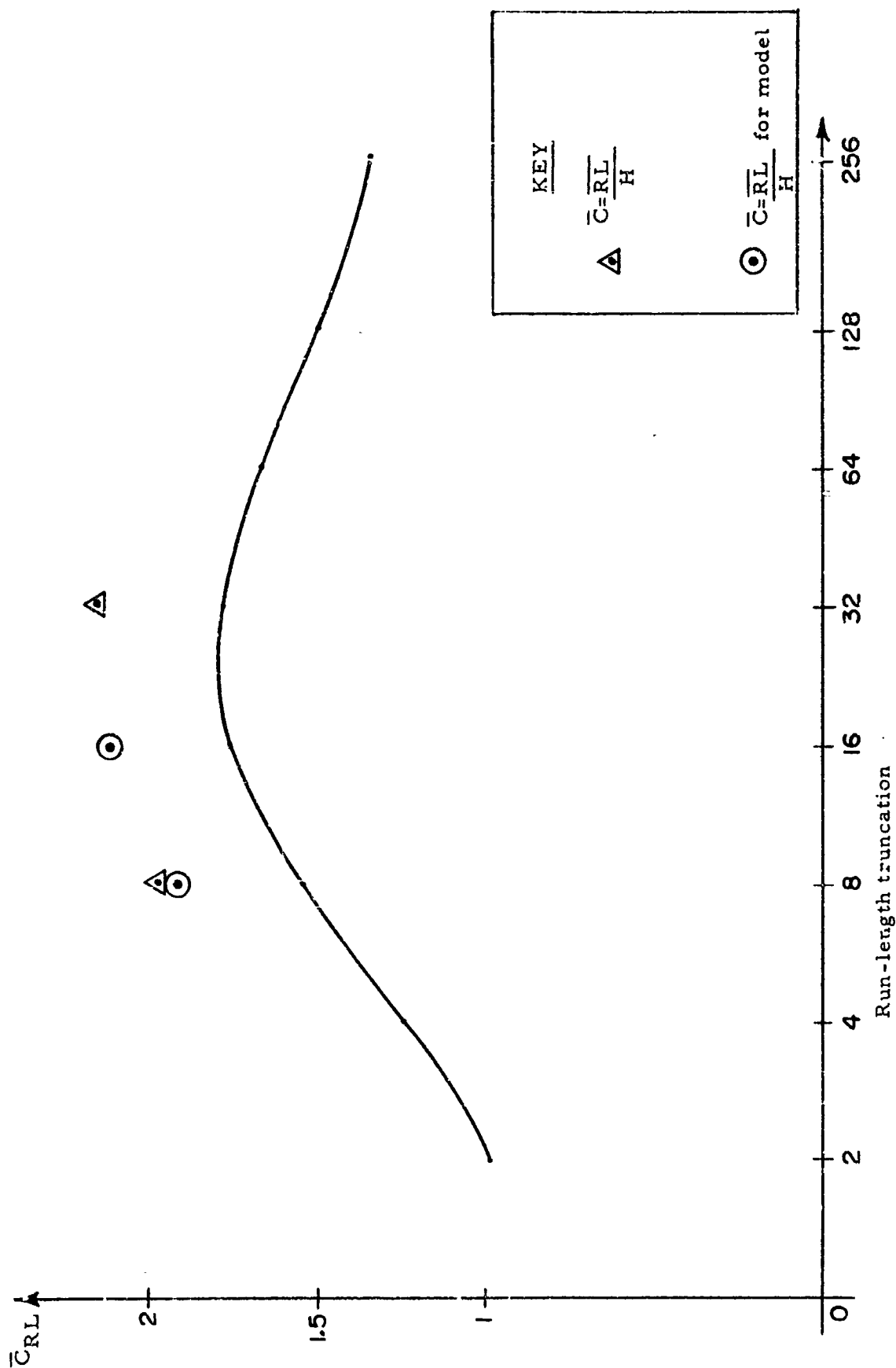


FIGURE 2.2-2-3 AVERAGE COMPRESSION AS A FUNCTION OF RUN-LENGTH TRUNCATION FOR THE VARIOUS DATA SETS AND THE RECORDING OF RUN LENGTHS

Consider n events $E_1, E_2, \dots, E_n$ with individual probabilities $a_1, a_2, \dots, a_n$ and transition probabilities P_{ij} . The joint probability of any two events is given by $P(E_i, E_j)$ where

$$P(E_i, E_j) = a_i P_{ij} \tag{50}$$

This concept can be extended to the joint occurrence of as many events as required, that is

$$P(E_i, E_j, \dots, E_s, E_t) = a_i P_{ij} \dots P_{st} \tag{51}$$

Define a run length y units long of event E_j as the joint occurrence of y successive E_j 's immediately preceded and followed by any other event other than E_j .

Consider a run-length of one and let $R_1^{y_1}$ be a random variable that describes a run length y_1 units long of E_j . Therefore,

$$\begin{aligned} P(R_j^1) &= P(E_1, E_j, E_1 \text{ or } E_2 \text{ or } \dots E_{j-1} \text{ or } E_{j+1} \text{ or } \dots \text{ or } E_n) + \\ &P(E_2, E_j, E_1 \text{ or } E_2 \text{ or } \dots \text{ or } E_{j-1} \text{ or } E_{j+1} \text{ or } \dots \text{ or } E_n) + \dots + \\ &P(E_n, E_j, E_1 \text{ or } E_2 \text{ or } \dots \text{ or } E_{j-1} \text{ or } E_{j+1} \text{ or } \dots \text{ or } E_n). \end{aligned} \tag{52}$$

This can be rewritten as

$$\begin{aligned} P(R_j^1) &= m \left[a_1 P_{1j} \cdot \sum_{k=1}^n P_{jk} \Big|_{k \neq j} + a_2 P_{2j} \cdot \sum_{k=1}^n P_{jk} \Big|_{k \neq j} + \dots + \right. \\ &\left. a_n P_{nj} \cdot \sum_{k=1}^n P_{jk} \Big|_{k \neq j} \right], \end{aligned} \tag{53}$$

where m is a constant to be evaluated. However,

$$\sum_{k=1}^n P_{jk} \Big|_{k \neq j} = (1 - P_{jj}). \tag{54}$$

Therefore,

$$P(R_j^1) = m \sum_{\ell=1}^n a_{\ell} P_{\ell j} (1-P_{jj})^{\ell-1}. \quad (55)$$

Now consider the case of obtaining a run length of $E_j y_1$ units long.

$$P(R_j^{y_1}) = m \sum_{\ell=1}^n a_{\ell} P_{\ell j} \left| \begin{array}{c} \ell \neq j \end{array} \right| \cdot P_{jj}^{y_1-1} (1-P_{jj}). \quad (56)$$

The constant m can be evaluated from

$$\sum_{y_1=1}^{y_1 \max} \sum_{j=1}^n P(R_j^{y_1}) = 1. \quad (57)$$

Assume that $y_1 \max$ is infinite in order to obtain a least-upper-bound, therefore,

$$\sum_{y_1=1}^{\infty} \sum_{j=1}^n P(R_j^{y_1}) = \sum_{y_1=1}^{\infty} \sum_{j=1}^n m \sum_{\ell=1}^n a_{\ell} P_{\ell j} \left| \begin{array}{c} \ell \neq j \end{array} \right| \cdot P_{jj}^{y_1-1} (1-P_{jj}). \quad (58)$$

For $(P_{jj})^2 < 1$,

$$\sum_{y_1=1}^{\infty} P_{jj}^{y_1} = P_{jj} (1-P_{jj})^{-1}, \quad (59)$$

and

$$\sum_{y_1=1}^{\infty} \sum_{j=1}^n P(R_j^{y_1}) = 1 = m \sum_{j=1}^n \sum_{\ell=1}^n a_{\ell} P_{\ell j} \left| \begin{array}{c} \ell \neq j \end{array} \right|. \quad (60)$$

Solving for m,

$$m = \frac{1}{\sum_{j=1}^n \sum_{\ell=1}^n a_{\ell} P_{\ell j} \ell \neq j} \quad (61)$$

Therefore,

$$P(R_j^{y_1}) = \frac{\sum_{\ell=1}^n a_{\ell} P_{\ell j} \cdot P_{jj}^{y_1-1} (1-P_{jj})}{\sum_{j=1}^n a_j (1-P_{jj})} \quad (62)$$

To obtain the average compression it is necessary to obtain the average run length ($\bar{RL}$) where it is defined as

$$\bar{RL} = \sum_{y_1=1}^{\infty} \sum_{j=1}^n y_1 P(R_j^{y_1}) \quad (63)$$

The compression for obtaining the least-upper-bound for run-length coding is

$$\bar{C}_R = \frac{N \bar{RL}}{H[P(R_j^{y_1})]} \quad (64)$$

which reduces to

$$\bar{C}_R = N \sum_{y_1=1}^{\infty} \sum_{j=1}^n \frac{y_1 P(R_j^{y_1})}{H[P(R_j^{y_1})]} \quad (65)$$

However,

$$H[P(R_j^{y_1})] = - \sum_{y_1=1}^{\infty} \sum_{j=1}^n P(R_j^{y_1}) \log_2 P(R_j^{y_1}) \quad (66)$$

Therefore, Equation 65 becomes

$$\bar{C}_R = \frac{N \sum_{y_1=1}^{\infty} \sum_{j=1}^n y_1 P(R_j^{y_1})}{\sum_{y_1=1}^{\infty} \sum_{j=1}^n P(R_j^{y_1}) \cdot \log_2 P(R_j^{y_1})} \quad (67)$$

Therefore, given the necessary transition probabilities, it is possible to obtain the least-upper-bound on the compression for run-length coding of an n-level video source.

The expression for run-length coding is quite involved; however, by considering run-length coding of the PE and (1-PE) source, the expression is simplified somewhat. Previous-element-run-length coding (PERL) is related directly to run-length coding. Consider a run of level j , y elements long. In this run there are $y - 1$ PE elements or a run length y elements long is a PE run length of $y - 1$ elements long. However, the analysis of the PERL truncated system is different from the truncation process of conventional run-length coding.

Consider a run length of any j , y elements long. A PE run-length $y - 1$ elements long would be the summation of all the possible values of j . However, this would require sending first a level, then the PE run length. This system does have the advantage that since the previous-element encoder has been implemented, the PERL system could be implemented with very few modifications to the existing system.

To define the compression available with PERL coding, it is necessary to make the distinction between PE run-lengths and the remaining elements. Let the least-upper-bound for compression be given by $\bar{C}_{PERL}$ where

$$\bar{C}_{PERL} = \frac{k_1 (\log_2 n)}{H(\text{message set})} + \frac{k_2 (\log_2 n) \bar{RL}}{H(\text{message set})} \quad (68)$$

and k_1 is the percentage of the elements only requiring one code word. The value of k_1 will then be the number of times an element is not a PE signal. From previous-element coding this is the equivalent of the (1-PE) statistic and is computed by

$$k_1 = (1-PE) = \sum_{j=1}^n P(j)^1$$

Therefore, k_2 is merely equal and equivalent to the PE value, or

$$k_2 = PE^1 = 1 - \sum_{j=1}^n P(j)^1,$$

where $P(j)^1$ is the probability of obtaining level j after PEC has been performed. Let the average previous-element run length be $\overline{PERL}$. Given

$$P(R_{PE}^y),$$

the probability of obtaining a PE run of y elements, the average PE run-length is

$$\overline{PERL} = \sum_{y=1}^m y P(R_{PE}^y). \quad (69)$$

The upper-bound on average compression is given by $\bar{C}_{PERL}$ and is equal to

$$\bar{C}_{PERL} = \log_2 n \frac{\left\{ \sum_{j=1}^n P(j)^1 + \left[1 - \sum_{j=1}^n P(j) \right]^1 \overline{PERL} \right\}}{\sum_{y=1}^m H \left[P(R_{PE}^y) \right] + \sum_{j=1}^n H \left[P(j) \right]}. \quad (70)$$

The lower-bound for $\overline{PERL}$ can be obtained by flat encoding. The total number of code words necessary is the $s(m+n)$ since each run requires both a magnitude and PE run length, but not all require two code words and s is a constant for this. Therefore, the number of bits required is $\log_2 s(m+n) = \log_2 s + \log_2(m+n)$.

The average compression for flat encoding is

$$\bar{C}_{PERL} = \frac{\log_2 \left\{ \sum_{j=1}^n P(j)^1 + \left[1 - \sum_{j=1}^n P(j) \right]^1 \overline{PERL} \right\}}{\log_2 [s + \log_2 (m+n)]}. \quad (71)$$

It would be extremely useful to relate the $P(j)^1$ and $1-P(j)^1$ to the well-known PE and (1-PE) statistics. Consider now Equations 68 and 71. The number

$$\sum_{j=1}^n P(j)$$

is equal to k_1 , or

$$k_1 = \left[\sum_{j=1}^n P(j) \right] \quad (72)$$

If $k_1 + k_2 = k$, then the probability of obtaining the original 1-PE statistics

$$P \left\{ \left[\sum_{j=1}^n P(j) \right] = \frac{k_1}{k} = 1-PE \right. \quad (73)$$

The new value of the probability from the original $P(j)$ can be found. Now

$$P \left[\sum_{j=1}^n P(j) \right] = \frac{k_1}{k-k_1} = \frac{k_1 \overline{PERL}}{k_1 \overline{PERL} + k-k_1} \quad (74)$$

Dividing numerator and denominator of 74 by k_1 , it becomes

$$P \left[\sum_{j=1}^n P(j) \right] = \frac{\frac{k_1}{k} \overline{PERL}}{\frac{k_1}{k} \overline{PERL} + 1 - \frac{k_1}{k}} \quad (75)$$

However, $k_1/k = (1-PE) = \sum_{j=1}^n P(j)$ and 75 can be rewritten as

$$P \left[\sum_{j=1}^n P(j) \right] = \frac{(1-PE) \overline{PERL}}{(1-PE) \overline{PERL} + PE} \quad (76)$$

It is now possible to rewrite Equation 70 in terms of the average PE run length and the PE signal. Therefore,

$$\bar{C}_{\overline{PERL}} = \frac{\log_2 n}{H(\overline{PERL})} \frac{\overline{PERL}}{(1-PE) \overline{PERL} + PE} \quad (77)$$

The upper-bound on PERL coding can be found by taking the limit as PE approaches one; 77 now becomes

$$\lim_{PE \rightarrow 1} \bar{C}_{PERL} = \overline{PERL} \frac{\log_2 n}{H(PERL)} \quad (78)$$

PE $\rightarrow$ 1

Run-length coding is a highly efficient system for a small class of pictures. Compressions of 2:1 to 12:1 can be shown for a black and white picture, but at the expense of a highly complex system. Compression of an eight-level source is in the order of 2:1 to 4:1 for a small class of pictures.

The compression that can be realized by PERL coding is less than the amount by conventional run-length coding, but the PERL system is much easier to implement, thus less complex. The transition probability model for run-length coding is a valid and extremely accurate model for predicting compression. In general, the model verified that there exists an optimum truncation length for compression since the compression is a function of the truncation run-length. The compression for truncated run-length coding in general is higher than conventional run-length coding. This is valid not only for a two-level source, but for a multilevel source also.

In general, if complexity is traded for compression, run-length coding is probably not as efficient as PEC. This is valid for both conventional, previous-element, and truncated run-length coding due to the buffer store requirements on run-length coding.

2.2.3 Roberts Encoding

Previous attempts at bandwidth compression have attempted to either utilize the statistical properties of the video signal or to eliminate redundancy. Robert's (5,6) technique of bandwidth compression does not utilize either of these properties directly, but takes advantage of the viewer's eye and presents the video data in such a manner as to best take advantage of the redundancy and statistics.

The system is a conventional PCM system with a digital pseudo-random noise generator and associated circuitry at the transmitter and receiver. The digital pseudo-random noise is produced by a pseudo-random number generator and is added to the video signal prior to quantization. At the receiver, the same noise (synchronized with the transmitter) is subtracted after the digital video signal has been converted to an analog voltage. Effectively the pseudo-random noise addition and subtraction randomizes the average video and quantizes it to several intermediate levels instead of only one level. As a result, the normal quantization noise of a three-bit, analog-to-digital converter is effectively reduced and the subjective results indicate that the three bits can represent the tonal scale. Therefore, the continuum of pseudo-random noise, limited in maximum amplitude to one basic PCM quantization level, added to the signal before quantization, reduces the tonal error but at the expense of increasing the total mean-square error. Therefore, an improved picture is obtained by trading noise for annoying contours. Subjectively, the noisy picture is more pleasing to the eye than the coarsely quantized signal with contours since the eye tends to average out the noise.

Although the addition and subtraction of the pseudo-random noise effectively reduces the contouring effect, a further refinement is added to take advantage of the visual process. Since the eye is more tolerant of errors in bright areas than errors in dark areas, the video signal is companded prior to the addition of the pseudo-random noise. An improvement will result if the majority of the errors are concentrated in the bright areas since the eye responds more nearly to percentage of errors than to the absolute error. Thus by employing a nonlinear amplifier with a square root transfer function, an improvement will result since the eye responds to the square root of the intensity; therefore, if x is the video output, the output of the nonlinear device will be $\sqrt{x}$. The companded signal is then added to the pseudo-random noise.

The pseudo-random noise has rather precisely defined properties. Assume that the video is quantized to n levels and that the total number of elements in the picture is m^2 . (See 2.2.3.1 for discussion of pseudo-random numbers.)

The maximum amplitude of the pseudo-random noise is limited to one quantizing level. The maximum amplitude is then divided into $p = 2^P$ levels where, as a function of time, the p levels are generated in a pseudo-random manner. If the pseudo-random noise is generated by a shift-register with linear modulo-two feedback, the number of stages required is S where S must satisfy the relationship that if $(2^S - 1)$ is evenly divisible by P , then the quotient must equal m^2 . However, if $(2^S - 1)$ is not evenly divisible by P , S must be chosen appropriately in conjunction with P . This will ensure that the P bit sequence generated by the

pseudo-random number generator will not repeat itself in one frame, be random in time, and have a flat distribution. The companded video with pseudo-random noise is then quantized to N bits where $N = \log_2 n$.

The signal to be quantized will be $y = f(\sqrt{x} + \alpha)$. If a perfect channel is assumed, the signal y can be converted back to an analog voltage at the receiver. From this analog voltage the same noise α , synchronized with the transmitter, is subtracted. Finally this voltage must be squared to reconstruct a close approximation to the original waveform. Figure 2.2.3-1 is a diagram of a typical pseudo-random noise transmitter.

The receiver has the same pseudo-random noise generator as the transmitter and the two are synchronized. The signal from the transmitter, y , quantized to 3 bits is passed through a digital-to-analog converter where the pseudo-random noise with a maximum amplitude of one basic quantization level and variable is subtracted. This noise subtracted from z effectively increases the 8 levels going to the display to 64 levels since each quantization level may be broken up into 8 additional levels. This signal is then squared to return the voltage to a good approximation of the original video. Figure 2.2.3-2 shows a block diagram of the receiver.

A block diagram of the complete Roberts system is given in Figure 2.2.3-3. The physical operation of the system imposes no problems with regard to implementation. The Roberts technique has been built into EDITS as an optional feature and allows experimentation of combining Roberts technique with other coding techniques.

Before discussing combining Roberts and previous-element coding, a short graphical analysis of Roberts technique will be given to illustrate how the technique performs. Figure 2.2.3-4 shows a ramp function to which eight levels of pseudo-random noise is added. The ramp function plus pseudo-random noise is quantized to eight levels and is shown in Figure 2.2.3-5. This waveform is then transmitted over the channel and should be compared to Figure 2.2.3-6 which is the ramp function encoded to eight levels by conventional PCM. From the PCM with Roberts waveform quantized to eight levels the same pseudo-random noise added at the transmitter is now subtracted as shown by Figure 2.2.3-7. Figure 2.2.3-8 plots the displayed waveform for the PCM and PCM with Roberts. From this figure it can be seen how the technique has reduced the effects of contouring. For example, consider the time interval between 5 and 10. For the conventional PCM the value of the signal is one, whereas with Roberts the value varies from 0.7, 0.2, 1.1, 1.4, and 1.6. Therefore, the contour has been broken up into five levels instead of only one. In theory, it is possible to display a maximum of 64 levels with three-bit PCM and three-bit Roberts, whereas with conventional PCM only eight levels can be displayed. However, this is at the expense of having a larger total mean-square error which can be seen from Figure 2.2.3-9. Probably one of the most interesting aspects of the system is

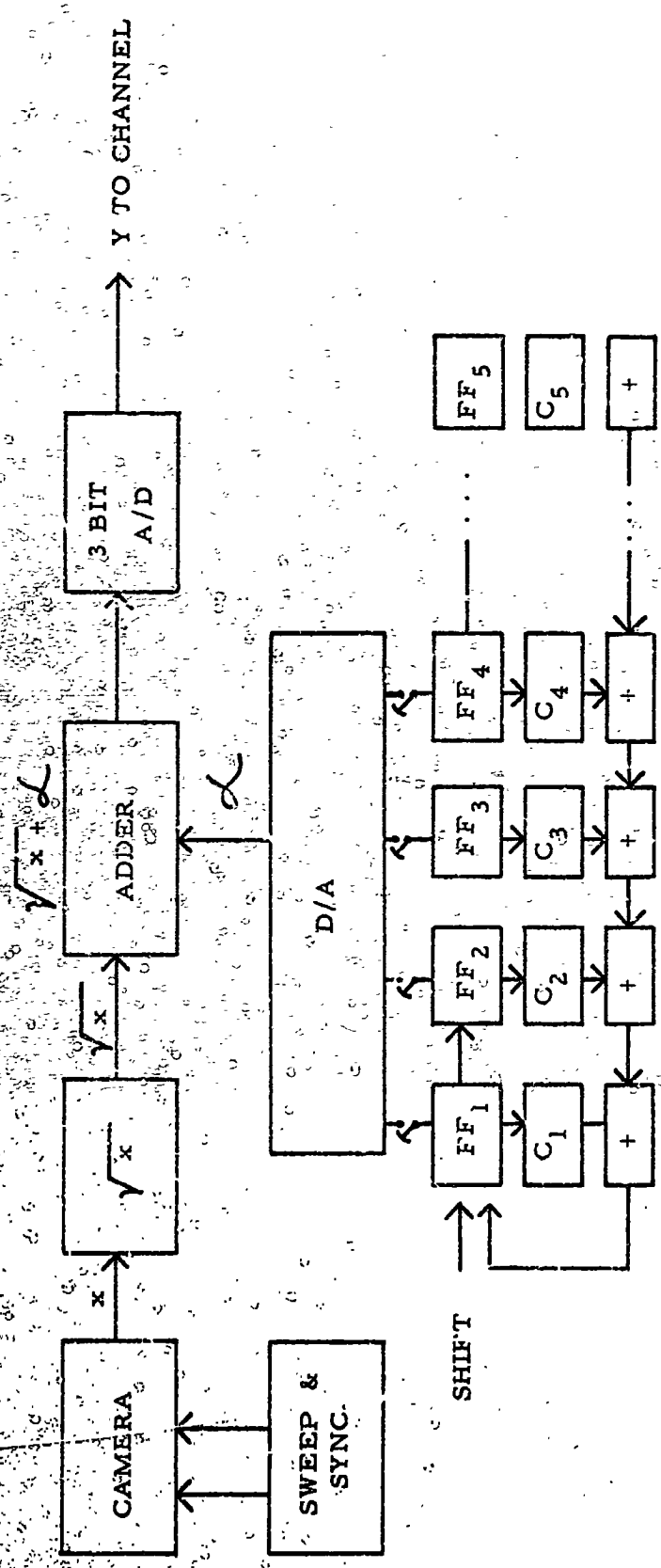


Figure 2.2.3-1
PSEUDO-RANDOM NOISE TRANSMITTER

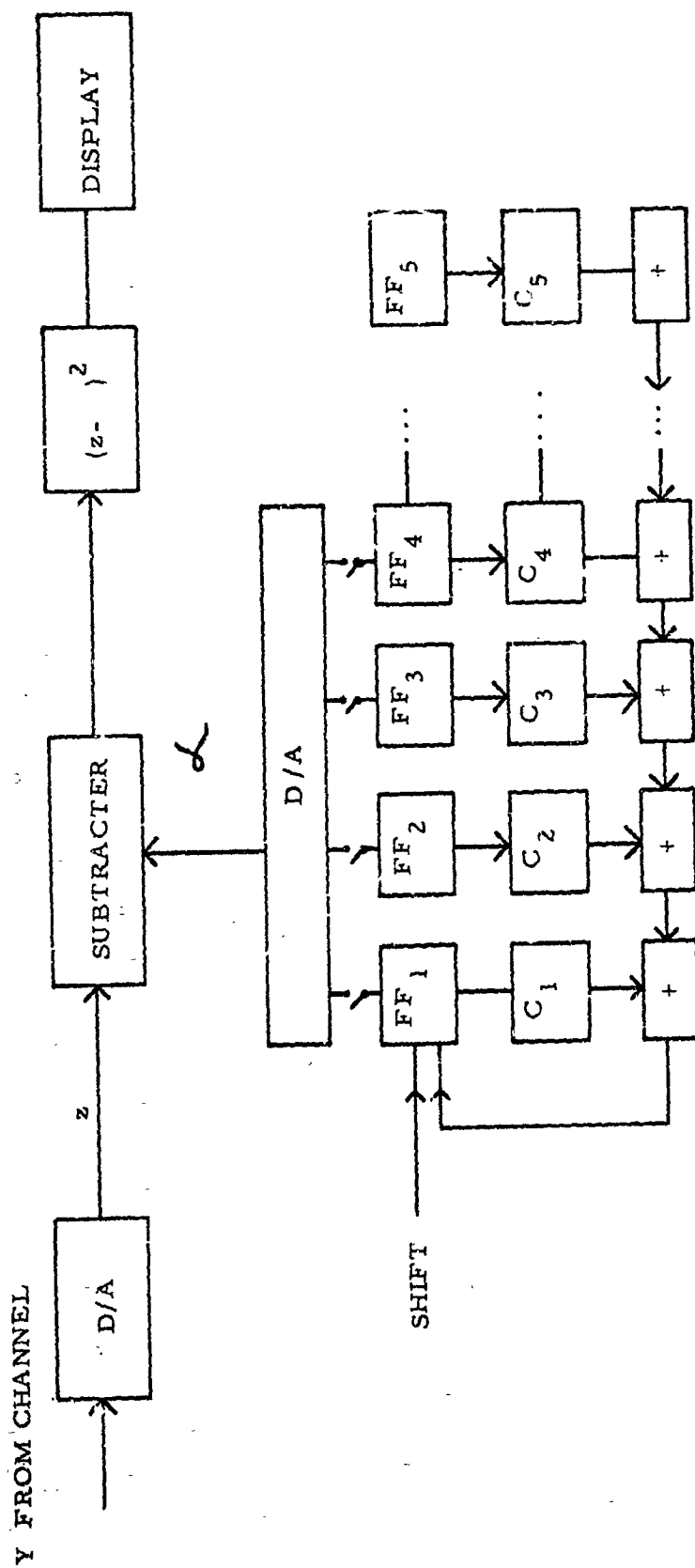


Figure 2.2.3-2
PSEUDO-RANDOM NOISE RECEIVER

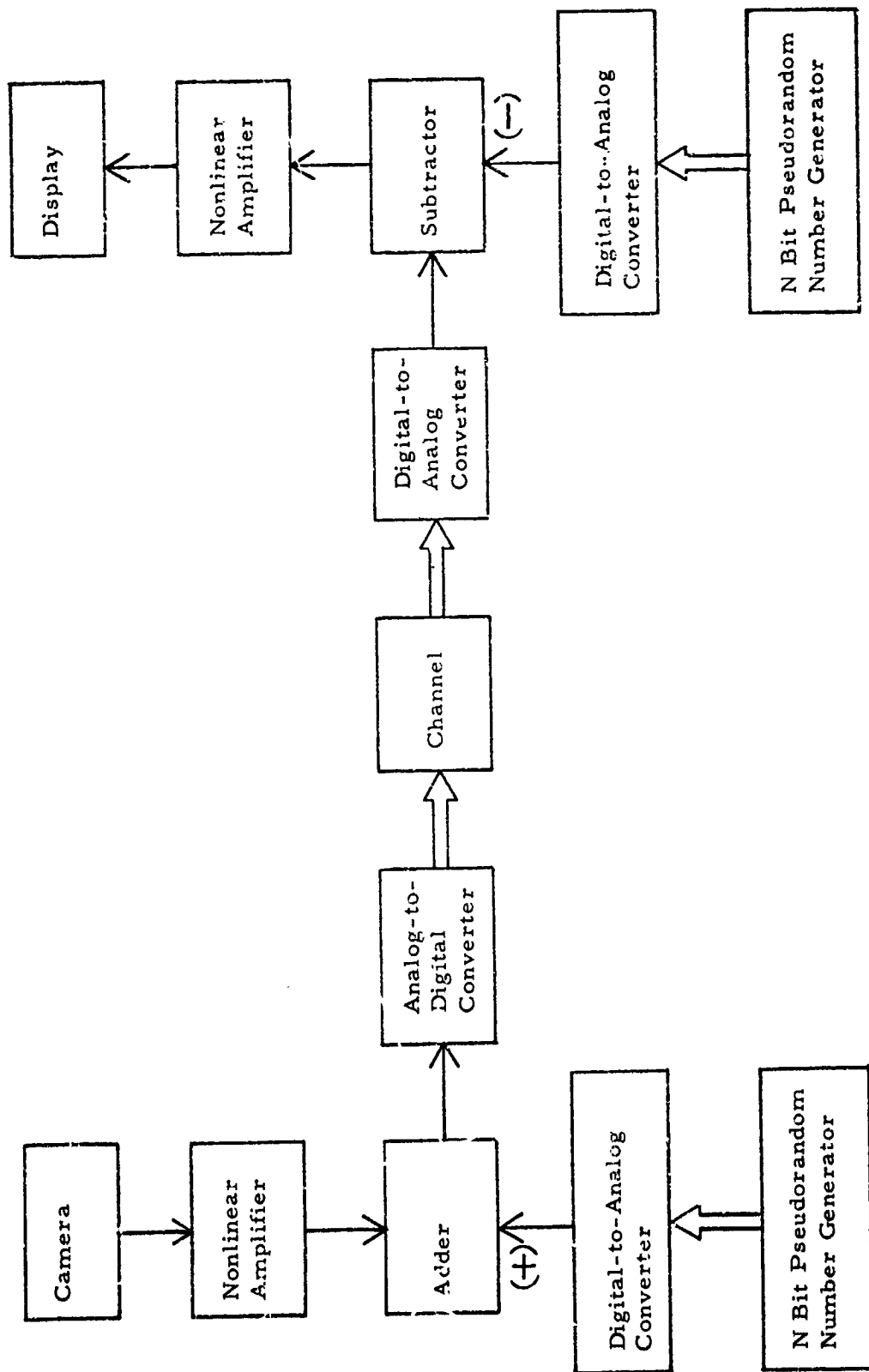


FIGURE 2.2.3-3 PCM TELEVISION SYSTEM WITH PSEUDORANDOM NOISE ADDITION AND SUBTRACTION

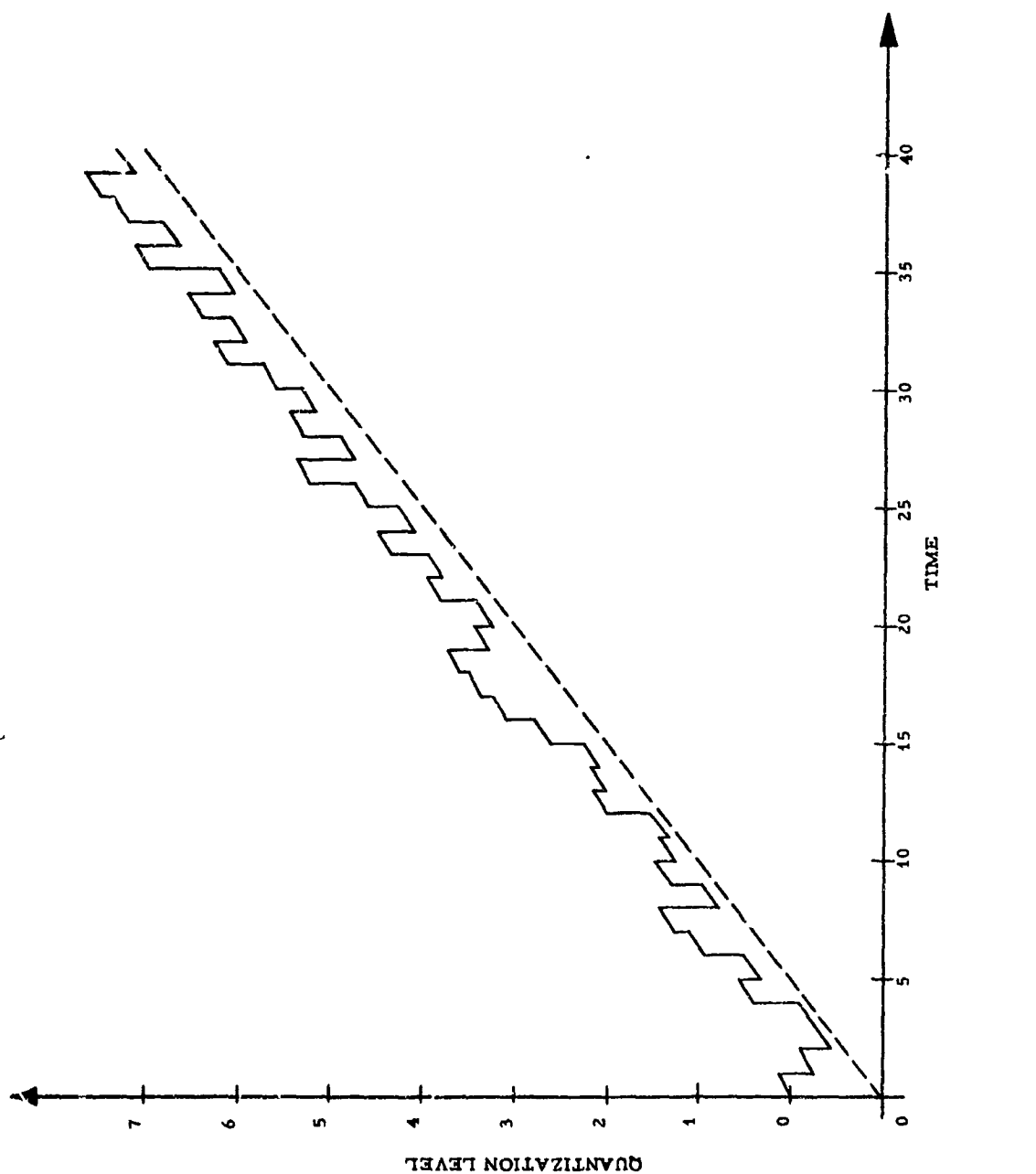


FIGURE 2. 2. 3-4 COMPARISON OF ORIGINAL WAVEFORM AND WAVEFORM AFTER PSEUDORANDOM NOISE HAS BEEN ADDED

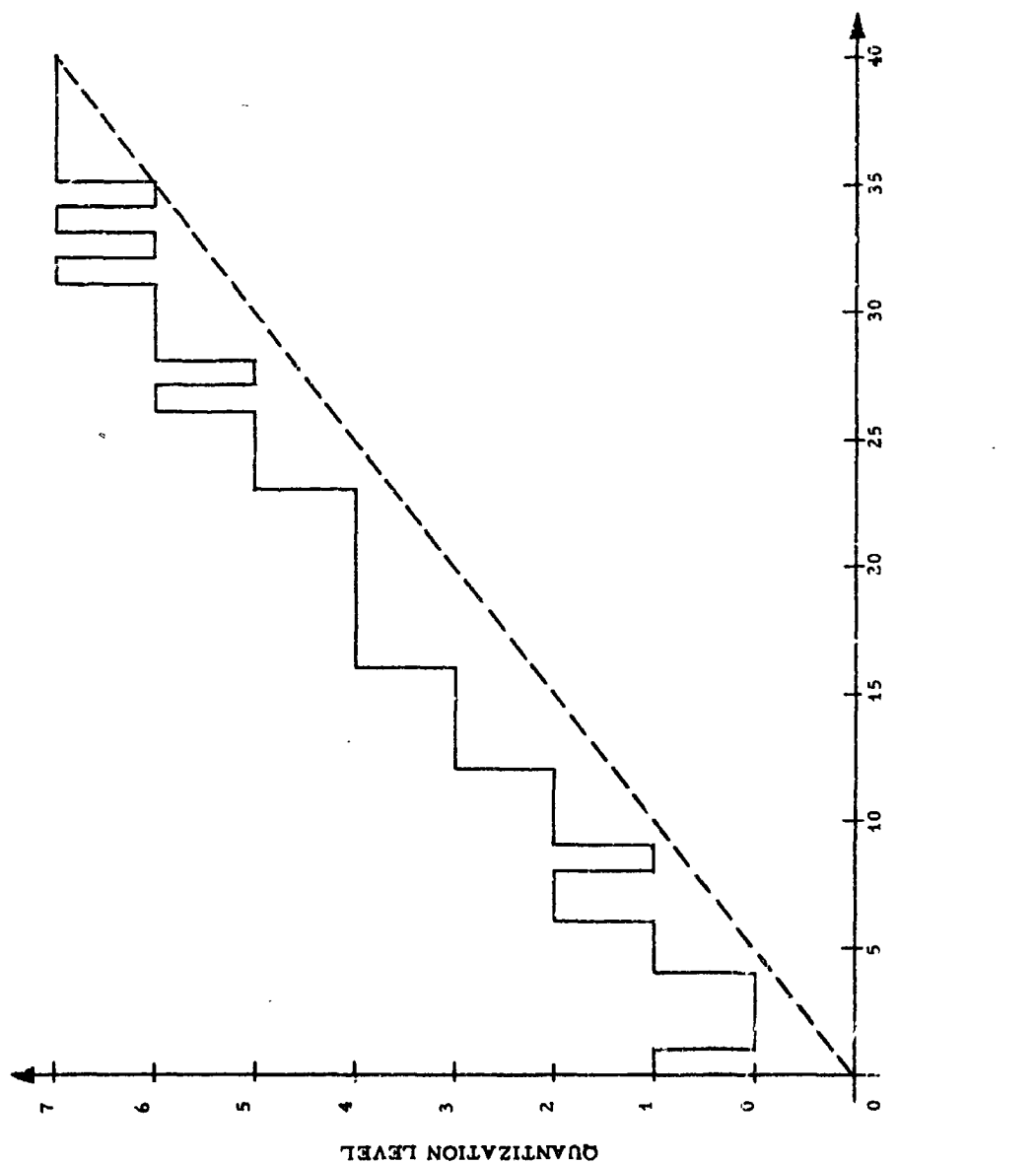


FIGURE 2.2.3-5 COMPARISON OF ANALOG TRANSMITTED WAVE AND PCM TRANSMITTED WAVE WITH PSEUDORANDOM NOISE ADDED

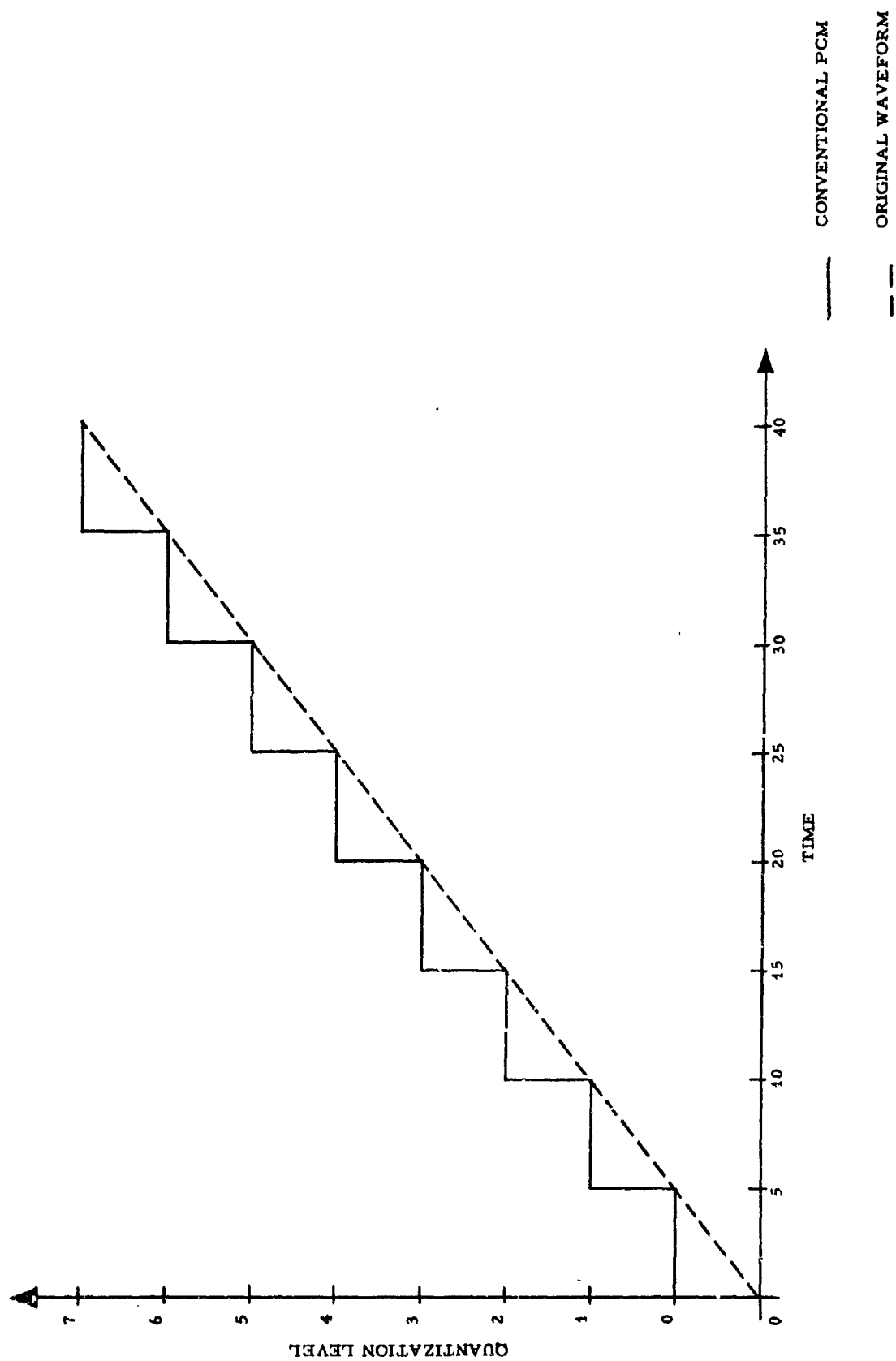


FIGURE 2.2.3-6 COMPARISON OF ANALOG OUTPUT AND CONVENTIONAL PCM OUTPUT

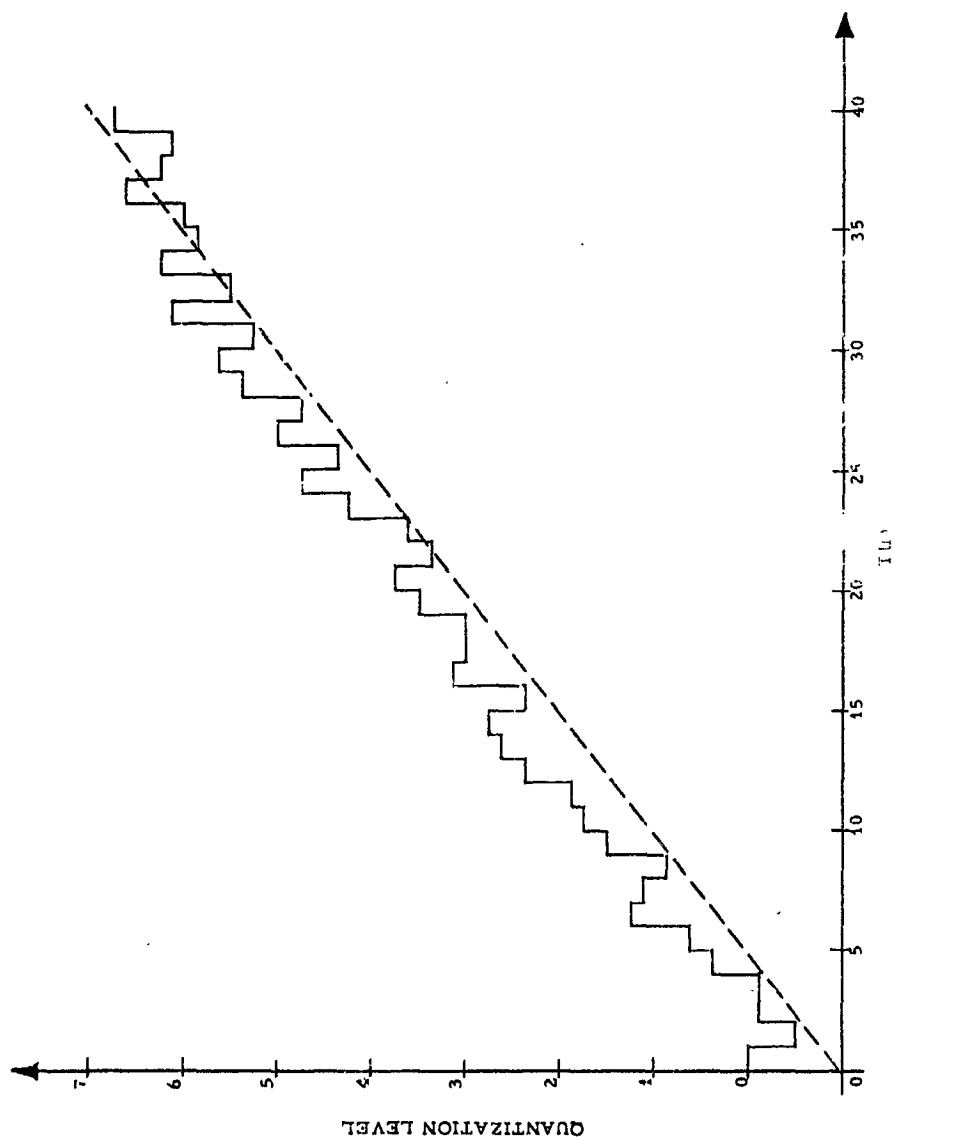


FIGURE 2.2.3-7 COMPARISON OF ANALOG OUTPUT AND CONVENTIONAL PCM WITH PSEUDORANDOM NOISE ADDED PRIOR TO QUANTIZATION AT TRANSMITTER AND THE SAME PSEUDORANDOM NOISE SUBTRACTED AFTER QUANTIZATION AT THE RECEIVER

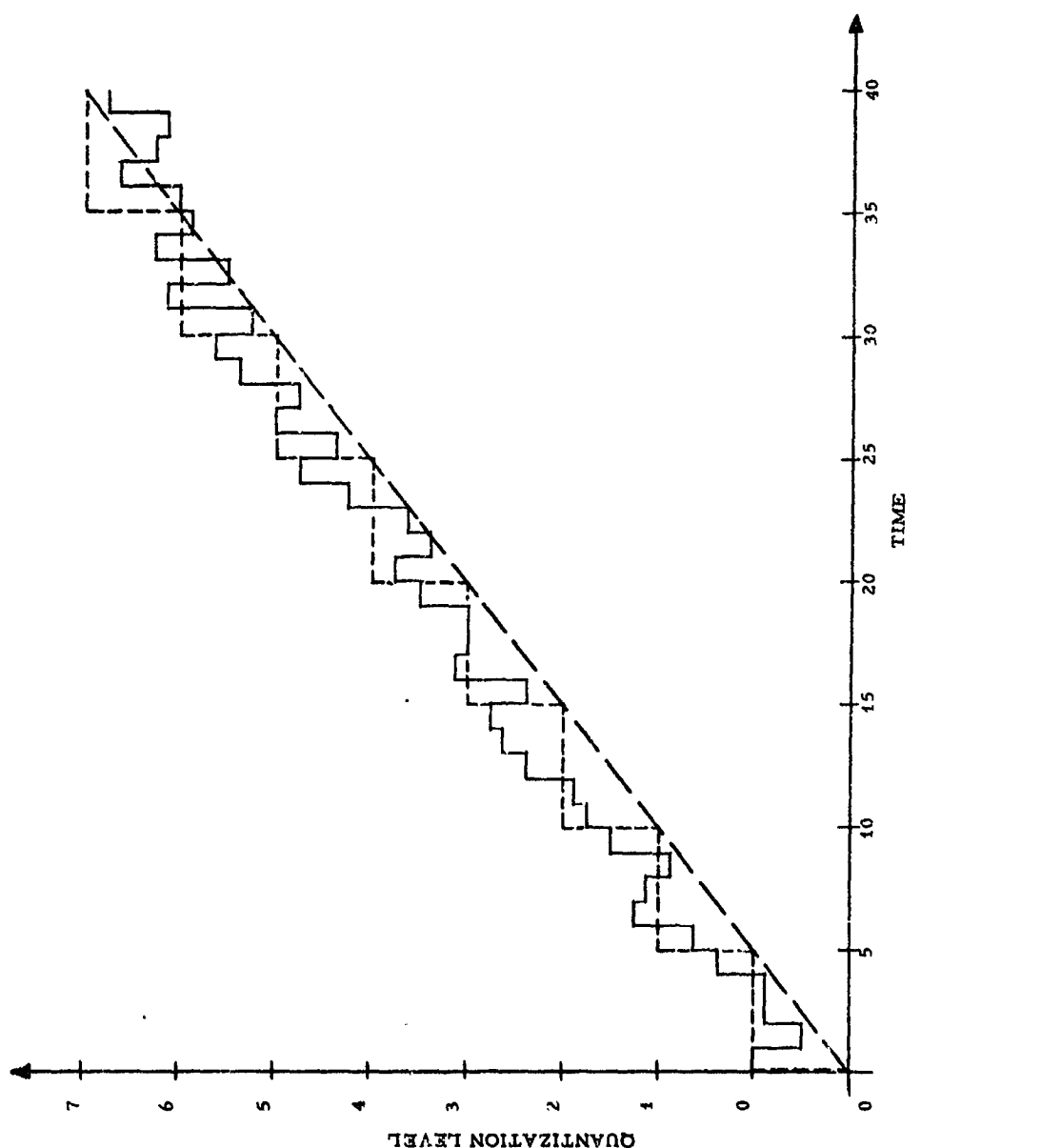


FIGURE 2.2.3-8 COMPARISON OF OUTPUT WAVEFORMS FOR CONVENTIONAL PCM AND CONVENTIONAL PCM WITH PSEUDORANDOM NOISE ADDED AND SUBTRACTED

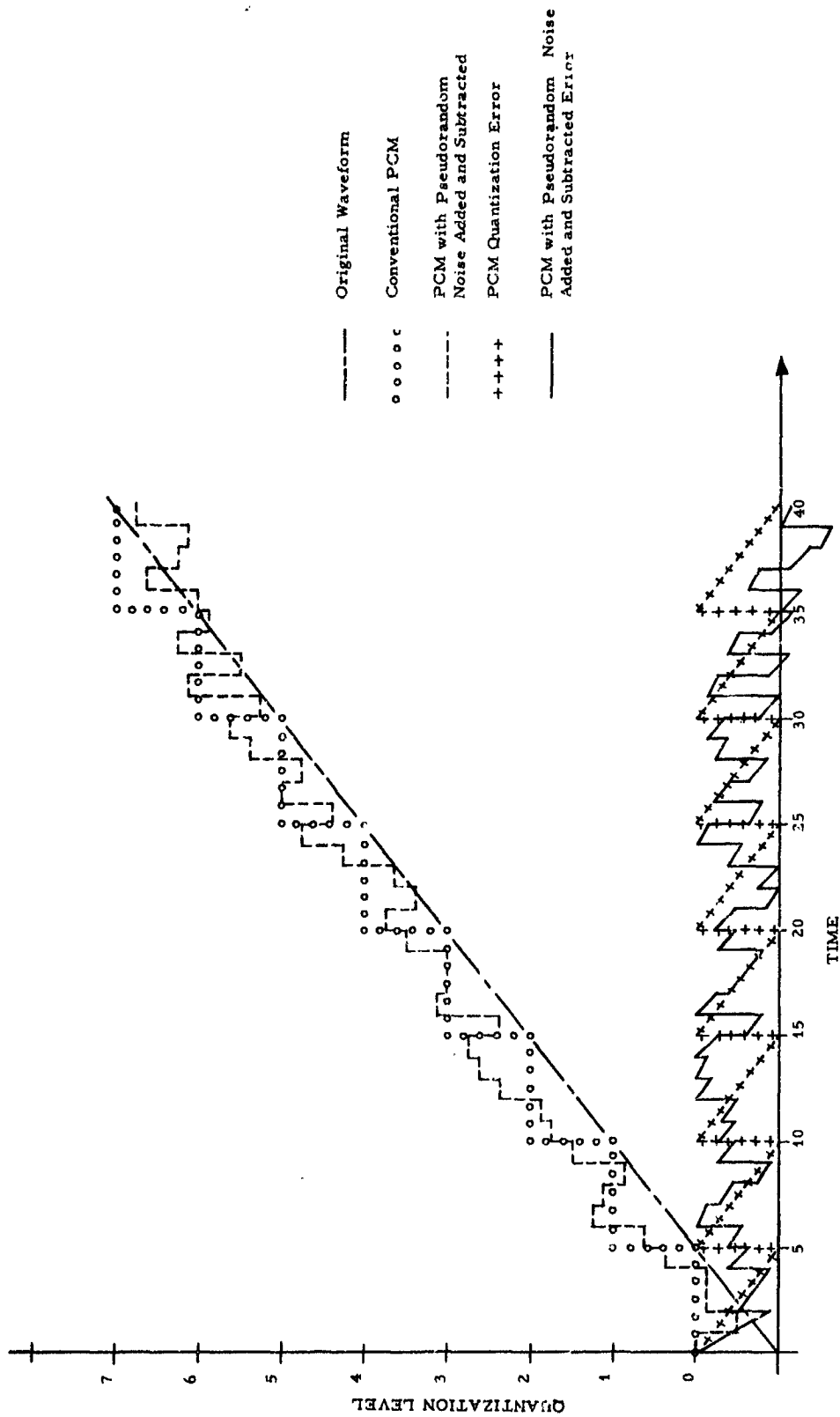


FIGURE 2.2.3-9 PLOT OF QUANTIZATION ERROR AS A FUNCTION OF TIME FOR PCM AND PCM WITH PSEUDORANDOM MODULATION

that the system requires the same signal-to-noise ratios and synchronization as conventional PCM.

The compression for Roberts technique is approximately 2:1 if the subjective results of using three bits of quantization with four bits of pseudo-random noise is judged subjectively acceptable.

Some preliminary results on subjectively judging the effects of Roberts and decreased quantization can be obtained from the comparison of EMR-developed EDITS I of two subjective reproduction compression systems shown in Section 6.2.5. This figure presents pictures obtained from EDITS of conventional PCM and PCM with Roberts for varying resolutions. It is apparent that the normal annoying contours present for three-bit quantization are not present with the Roberts technique; however, the six-bit PCM is subjectively preferred to the PCM with Roberts. There are, however, applications where the Roberts technique with the 2:1 compression would be highly desired.

The Roberts technique can be combined with other coding techniques to increase the compression by considering the output of the Roberts encoder as an information source. Although it appears that applying previous-element coding to the noisy Roberts source would defeat the purpose of previous-element, experimental results have proven otherwise. For six-bit quantization the previous-element signal is approximately 0.5; however, for three-bit quantization the PE is approximately 0.8. When Roberts is added to three-bit quantization, there does exist a slight decrease (about 15 percent) in PE, but there is already the 2:1 compression of the Roberts technique to begin with. The only change in the PE signal is due to the pseudo-random noise addition, but since the maximum change in quantization level that can be effected by the addition of the pseudo-random noise is limited to a maximum of one quantizing level, and the probability of obtaining any intermediate level is the same as the probability of obtaining a maximum change, the signal statistics do not change drastically for the relatively coarse quantization to 3 bits.

First order, (1-PE), PE, and run-length statistics were taken on EDITS at 256 resolution of the "girl picture". These statistics were used to calculate the various values of compression available. Three values of compression will be used to predict the compression for the various coding techniques. The upper bound is found by using the statistical entropy of the message set, the lower bound by flat encoding, and the actual compression by using the encoded entropy from Huffman encoding the message set.

The encoding techniques to be used are truncated run-length and PEC with and without Roberts technique. For PEC and assuming a 2:1 compression for Roberts technique, the results are tabulated in Tables 2.2.3-1 and 2.2.3-2.

TABLE 2.2.3-1

PEC COMPRESSION FOR GIRL PICTURE

Coding	Compression	
	Upper Bound	Actual
Without Roberts	3.26	2.18
With Roberts	3.37	3.18

TABLE 2.2.3-2

ENTROPY OF GIRL PICTURE

Coding	Entropy	
	Minimum	Actual
Without Roberts	0.92	1.38
With Roberts	1.78	1.59

Although basic PEC gives a 2:1 compression without Roberts, this decreases to 1.59 when used with Roberts. However, a 2:1 gain is obtained with Roberts; therefore, a net compression of 3.18 results.

Truncated run-length coding was discussed earlier and Figure 2.2.3-10 gives a plot of average run-length as a function of the truncation length. The black-and-white model predicted compression as a function of truncation length and this was verified for the 8-level picture also as shown by Figure 2.2.3-11. The probabilities were plotted in Figure 2.2.3-12. If the probabilities are considered it is possible to relate these to the average run-length. The average run-length can then be used to predict the compression. Notice how compression varies with truncation run-length. Table 2.2.3-3 tabulates the entropy of the girl picture as a function of the truncated run-length. For analysis the run-lengths were allowed to vary from 2 to 64.

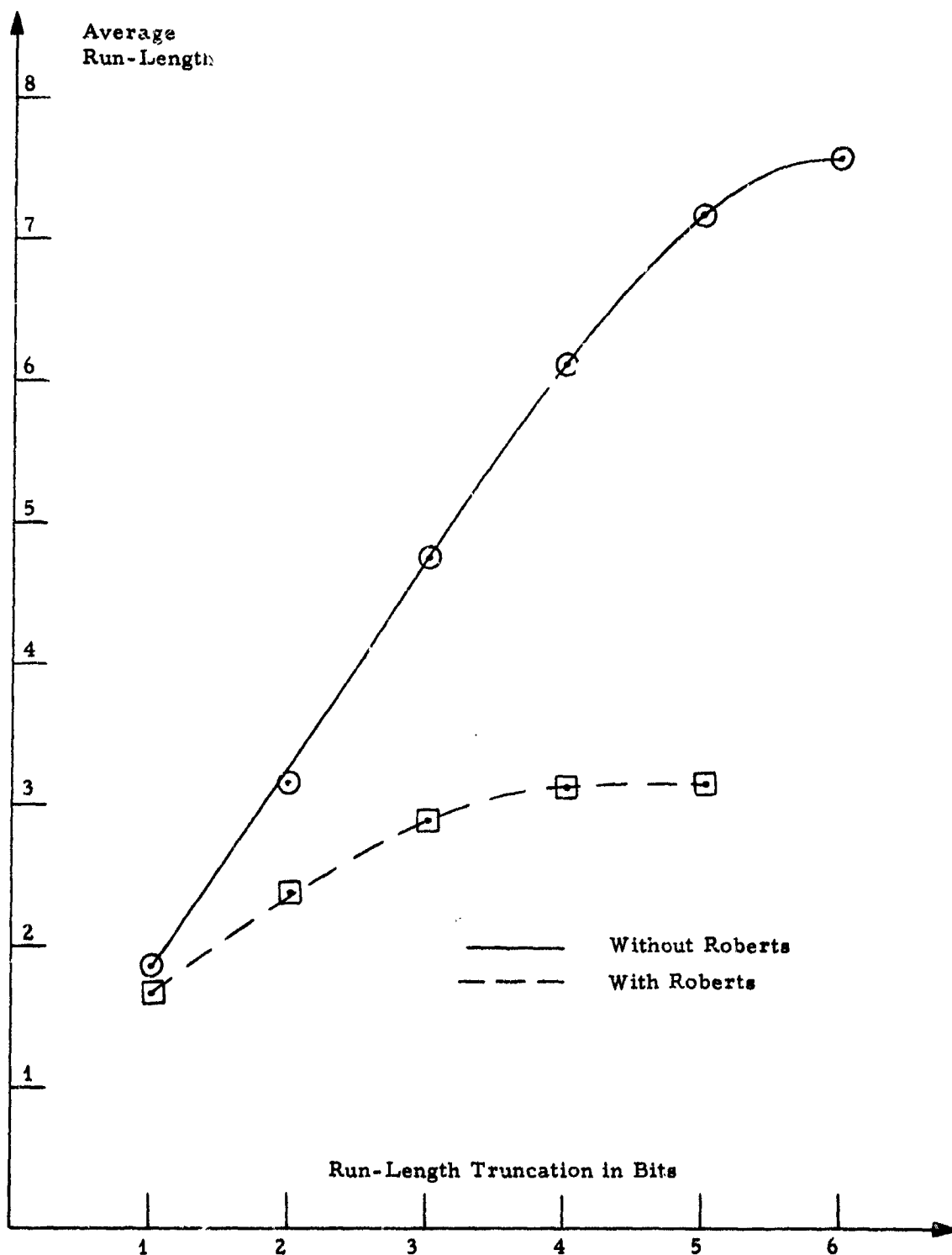


FIGURE 2.2.3-10 AVERAGE RUN-LENGTH AS A FUNCTION OF TRUNCATION LENGTH WITH AND WITHOUT PSEUDO-RANDOM NOISE MODULATION

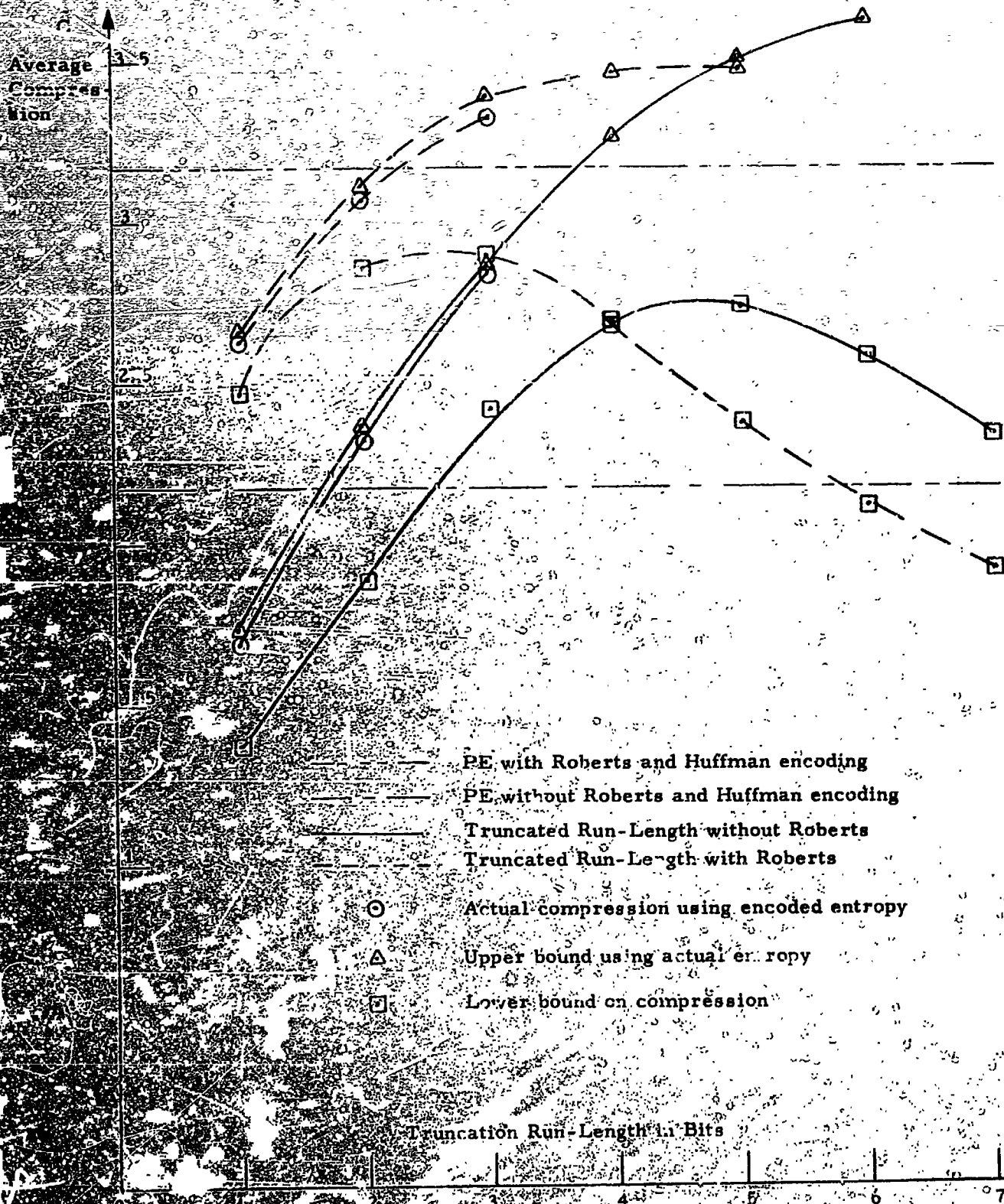


FIGURE 2-2-11 NET COMPRESSION FOR RUN-LENGTH WITH AND WITHOUT ROBERTS AND FOR PEC WITH AND WITHOUT ROBERTS ASSUMING A 2:1 COMPRESSION FOR ROBERTS

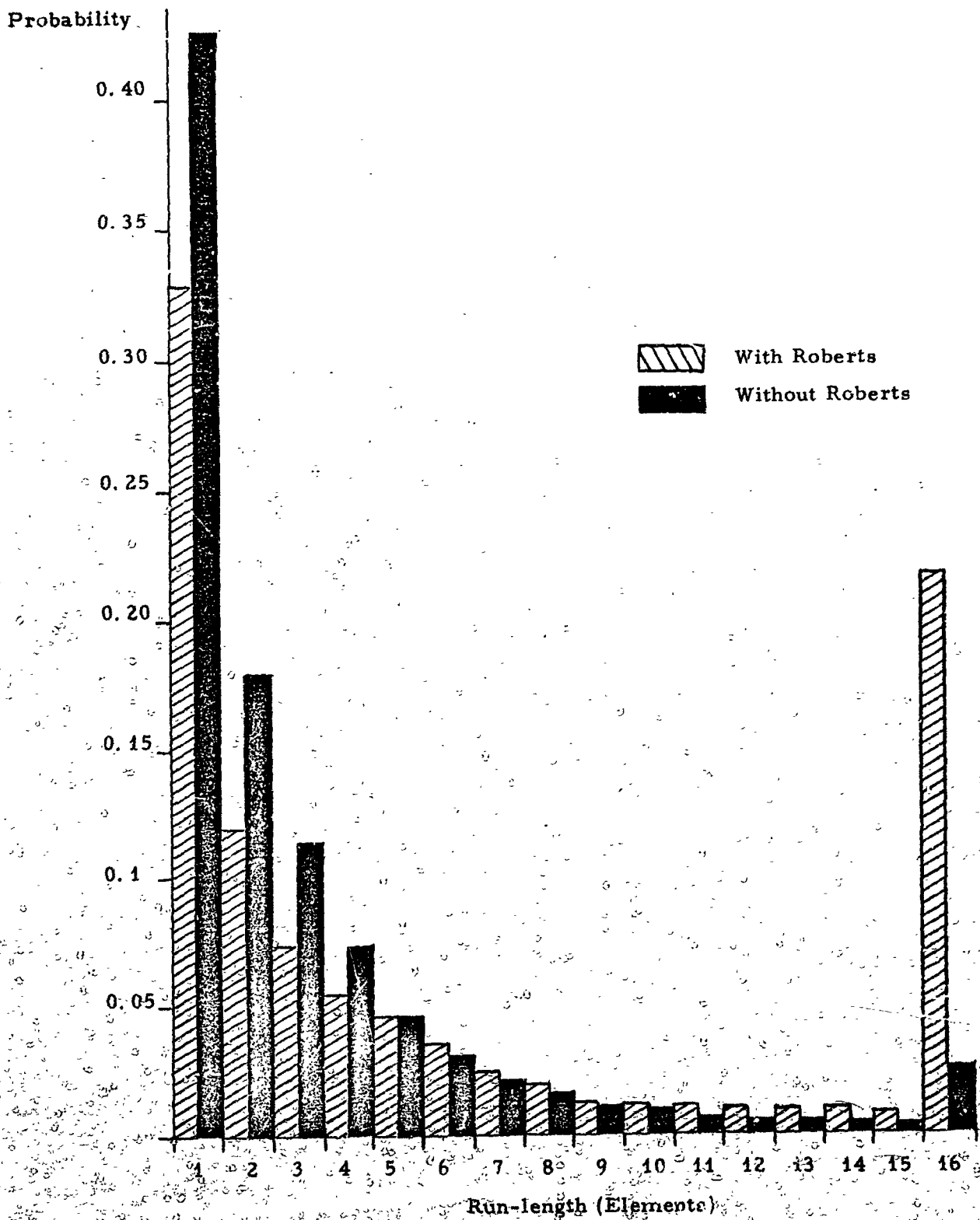


FIGURE 2.2.3-12

PROBABILITY OF RUN-LENGTH FOR GIRL PICTURE WITH
AND WITHOUT ROBERTS PSEUDO-RANDOM NOISE MODU-
LATION

TABLE 2.2.3-3
ENTROPY (BITS PER ELEMENT)

	Maximum	Minimum	Actual	PE Actual	Maximum	Minimum	Actual	PE Actual
H_2	2.16	1.73	1.77	1.38	2.41	2.24	2.26	1.59
H_4	1.59	1.28	1.29	1.38	2.10	1.94	1.95	1.59
H_8	1.26	1.04	1.05	1.38	2.07	1.76	1.81	1.59
H_{16}	1.14	0.92	...	1.38	2.24	1.74	...	1.59
H_{32}	1.11	0.85	...	1.38	2.52	1.72	...	1.59
H_{64}	1.18	0.83	...	1.38	...	...	...	1.59

Figure 2.2.3-11 summarizes the results of truncating the run-lengths with and without Roberts and compares the results to PE coding. From this figure the following statements can be made. PEC compression with Roberts technique is slightly less than truncated run-length with Roberts (3.2 to 3.4), but it is far easier to implement PE than run-length coding.

Truncated run-length coding compression without Roberts is greater than truncated run-length with, but subjectively the picture is unacceptable due to contouring.

The best over-all system at present considering complexity compression and picture quality is PEC with Roberts.

Compressions of 3:1 are possible for PEC with Roberts.

An additional amount of compression may be possible by combining PEC and Roberts with the split-band technique.

2.2.3.1. Pseudo-Random Number Generation

Pseudo-random number generation has progressed rapidly to a state-of-the-art status over the past few years, mainly through the mathematical efforts of Colomb⁽⁷⁾ and other investigators,^(8 - 15) The practical application of pseudo-random number generation has followed the theoretical development rather closely in such fields as error-correcting codes,^(16 - 21) modulation,⁽²²⁾

bandwidth compression,⁽²³⁾ secure communications,⁽²⁴⁾ and scanning. The generation of pseudo-random sequences can be accomplished through the use of binary shift-registers with feedback. The number of sequences that can be generated and their length depends upon, among other factors, the type of feedback employed which can be either linear or nonlinear.

Before going into the actual properties of pseudo-random sequences, it will be advantageous to discuss the mechanization elements necessary. One possible method of generation is through the use of re-entrant shift-registers with modulo-two addition for feedback. A shift-register is composed of a series of bistable memory elements with the ability to transfer the contents of each element into the next element when interrogated by a shift pulse. Normally, the shift pulse returns each element to the zero state with an appropriate delay for the transferring signal. The outputs of the appropriate shift-registers are connected with the appropriate logic stages such as modulo-two adders, which can be described by the truth table of Figure 2.2.3.1-1.

The outputs of the modulo-two adders are then fed back to the first stage of the shift-register. When the shift-pulse arrives, the time output of any one of the shift-registers forms a digital sequence with properties depending upon the initial condition of the registers, the type of feedback used, and the number of modulo-two adders.

The shift-register generator can be best described by Figure 2.2.3.1-2. The F_j 's are flip-flops connected in series to form an m stage shift-register. There are m switches $C_1, \dots, C_m$, which can either be opened or closed depending upon the desired sequence. Associated with each switch are the modulo-two adders with the properties of Figure 2.2.3.1-1. The final portion of the generator is the shift-clock pulses. The sequence of pulses generated by this circuit will be a periodic sequence of period $2^m - 1$ where m is the number of stages. The time structure of the sequence will be determined by (1) the original sequence set into the shift-register, (2) the type of logic used to combine the output of the last and intermediate stages to the first stage, and (3) the number and location of the intermediate stages used.

Two types of sequences can be obtained; these are: (1) a maximal sequence where m is the number of stages, referred to as m sequences, and (2) non-maximal sequences where a sequence is defined as nonmaximal if the period of the sequence is less than $2^m - 1$. For the maximal or m sequences the total number of sequences can be determined from $[\phi(2^m - 1)]/m$, where $\phi(2^m - 1)$ is the Euler function equal to the number of integers (one included) smaller than $(2^m - 1)$ and coprime (no common factor) with $(2^m - 1)$. An m sequence contains one run of m consecutive ones, one run of $(m-1)$ zeros, a run of ones of length $(m-1)$, and one of zeros of length $(m-2)$. In general, there are 2^i runs of ones and 2^i runs of zeros of length $(m-i-2)$ for $0 \leq i \leq m-3$; however, a run of m zeros does not occur. If the digits in each m consecutive positions of the shift-register are considered as an m -place binary number, a sequence of $(2^m - 1)$ m -place binary numbers are generated. Each of these binary numbers occurs

	0	1
0	0	1
1	1	0

FIGURE 2.2.3.1-1
MODULO-TWO ADDER TABLE

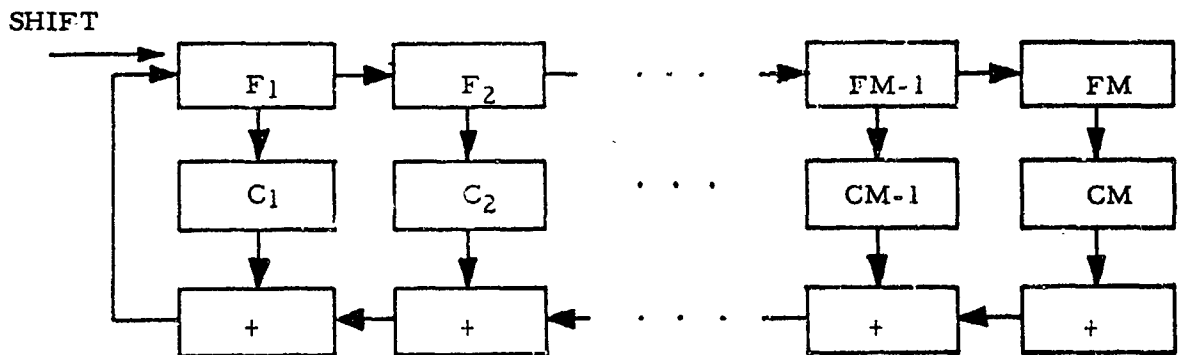


FIGURE 2.2.3.1-2

PSEUDORANDOM SHIFT REGISTER GENERATOR

once and only once with the exception of all zeros, and the sequence repeats itself after the generation of the complete sequence. Such a sequence will be called an M sequence. Although in general it is possible to generate an M sequence of any length, sometimes it is necessary to generate a sequence of period $2^m - 1$, but length S where $S < M$. It is possible to use the m-stage generator to generate the S sequence. Consider now an m-stage generator where only s of the m stages are used to generate the S sequence as shown in Figure 2.2.3.1-3.

Assume that the period of the S sequence must be greater than $2^s - 1$. This can be accomplished provided that $(2^m - 1)/s$ is not evenly divisible, and that m and s are chosen appropriately. If, for example, $(2^m - 1)/s$ is evenly divisible, then the length of the S sequence will be $(2^m - 1)/s$.

Therefore, it is possible to generate M sequences of length $2^m - 1$ or S sequences of length greater than or equal to $(2^m - 1)/s$, and each of these sequences have definite special applications.

Although these sequences generated look random in nature, no finite sequence can ever be truly random. However, the sequences generated by the shift-register generator technique will have certain properties which can be associated with randomness. For example, if the output of any stage is taken as a time sequence composed of zeros and ones, the total number of zeros is nearly equal to the total number of ones with the maximum difference being one. Stated in terms of conventional probability terminology, the first-order probability distribution must be flat. This of course implies that the total number of runs of zeros must equal the total number of runs of ones. Finally, the autocorrelation function of such a sequence is two-valued, which gives another mathematical tool to analyze generated sequences to determine whether or not they can be classified as pseudo-random in nature.

Although the earlier investigators spent much time and effort determining the proper connections for generating m sequences, such an analysis for linear feedback is not necessary for M sequences less than 20. Fortunately, it is possible to relate the feedback stages to the roots of irreducible polynomials for specified m lengths. Marsh (9) has tabulated these characteristics for not only m sequences, but for nonmaximal sequences also. Therefore, given the sequence length, it is only necessary to refer to the table to obtain the proper linear feedback connections. It is also possible to generate nonmaximal sequences of K bit numbers and have these K bit numbers presented in time in a pseudo-random manner. Therefore, it can be concluded that pseudo-random number generation has achieved such a state-of-the-art status that generation is relatively simple to implement and prediction of the statistical properties of the sequence can be accomplished without too much difficulty.

2.2.4 Split-Band Coding

Split-band coding is an efficient bandwidth-reduction-coding technique that uses the statistical redundancies in video signals and certain characteristics of human

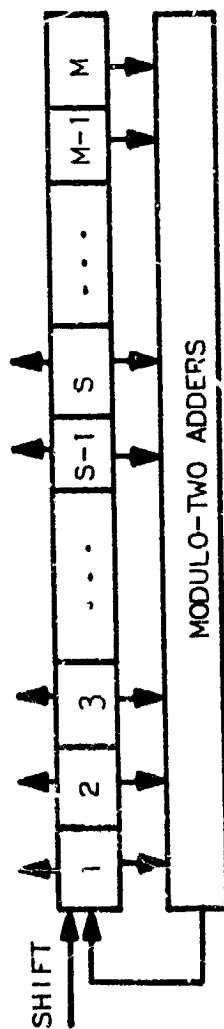


FIGURE 2.2.3.1-3
S-BIT NUMBER GENERATOR

vision to reduce the transmission bandwidth. The system attempts to match the video system to the information channel by reducing the information rate to some lower average value. A television picture consists basically of two kinds of picture detail: (1) sharp tonal transitions or picture edges, and (2) gradual changes in tonal scale between the edges. The eye mainly perceives the edges of objects and has little success in judging the absolute brightness of any large area; therefore, since there are only a finite number of edges in an average picture and they are contained in the high-frequency components of the video, it is necessary to sample them at a high sampling rate. The highs can be quantized coarsely, stored, and transmitted at a slower, uniform rate.

However, the low-frequency components require more levels of quantization to prevent contouring. This signal is sampled at a much lower rate because it is composed of the low-frequency components. Therefore, it is possible to reduce the bandwidth required by separating the video signal into its low- and high-frequency components and requiring the average information transmission rate to be much less than the conventional system. Kretzmer ⁽²⁵⁾ in 1956 proposed such a system where the low-frequency components or the "lows" were obtained by low-pass filtering the original video signal. The high-frequency components or the "highs" were obtained by subtracting the "lows" from the original signal with a delay equal to the delay of the low-pass filter. However, this system exhibited spurious edge effects.

The spurious edge effects were eliminated in the system proposed by Schreiber ⁽²⁶⁾ in 1958. Schreiber's system used separate channels for the "lows" and "highs." Since the "lows" were transmitted in the analog mode, there was no contouring as there is with coarsely quantized PCM. However, the eye is extremely critical of edge position and will not tolerate edge displacement; therefore, the digital edge information requires a high signal-to-noise ratio and the edges require both a brightness and position code word. This system requires an elaborate buffer store to transmit the edge information at a uniform rate since the edges appear at a nonuniform rate. However, the average compression for such a system is in the order of 4:1. This system did exhibit a picture whose subjective appearance was inferior to the original picture.

As a possible improvement to the split-band system, Schreiber ⁽²⁷⁾ in 1959 proposed the "synthetic highs" system. The system is a basic split-band system, but the "highs" are synthesized from a waveform of pulses carrying the edge and amplitude information only. The leading and trailing edges of the "highs" are artificially generated and the original waveform can be reconstructed by adding the "lows" and "synthetic highs." However, the system still requires two channels for operation along with the buffer store problem. A system was built and average bandwidth reductions of 4:1 were obtained where the overall picture quality was acceptable. To locate the "highs," a form of run-length coding called "run-end positioning" was employed to obtain the average compression.

Newell and Geddes ⁽²⁸⁾ in 1961 conducted tests of the last two systems described and the results were startling. The test assumed that the basic operation for the

"quantized-highs" and for the "synthetic-highs" systems was the same as just described. A 2:1 bandwidth reduction was reported but at the expense of reduced accuracy in reproduction; thus, the bandwidth reduction was not due to reduction of redundancy entirely. The reduced-redundancy pictures of both systems were compared to a conventional system after reducing the bandwidth by means of only a low-pass filter and simple "crispener" circuit. The split-band systems were subjectively inferior to the low-pass system. Both the brightness and position code words required a higher S/N ratio than the conventional system. Neither system considered the subjective effects of noise, which will be a limiting factor. However, it was concluded that the split-band system could possibly result in a 4:1 average bandwidth compression.

In September of 1962, Pratt (29) presented a system utilizing the split-band-technique, but the over-all system is a hybrid of many compression techniques. Variable-velocity scanning, split-band, delta modulation, run-length coding, and the synthetic highs were all combined in one system. Although a 7:1 bandwidth reduction was claimed, this figure was not verified. However, if the system could be implemented, an average 7:1 bandwidth reduction seems entirely feasible.

The final split-band system was reported by Gicca(30) in 1962. This system as with Pratt's system was a combination of several compression techniques. The system combined the compression techniques of variable-velocity scanning, split-band, Roberts pseudo-random noise modulation, and multiplexing of the "highs" and "lows" for single-channel operation. A block diagram of the system is given by Figure 2.2.4-1. In view of this accomplishment, split-band, PEC, Roberts, and run-length coding could result in a feasible compression system. However, before considering these systems, consider the general results of split-band coding.

Split-band coding requires separation of the original video into the "lows" and "highs;" however, there is no general agreement as to what percentage of the original will be called the "lows." Instead of assuming a definite percentage, a general expression for split-band coding will be derived. The following definitions are necessary for the derivation.

Let

- x = the number of horizontal elements per line,
- y = the number of lines per frame,
- f = the number of frames per second,
- e = average number of edges per run-length,
- r = the number of bits necessary for quantization of "highs",

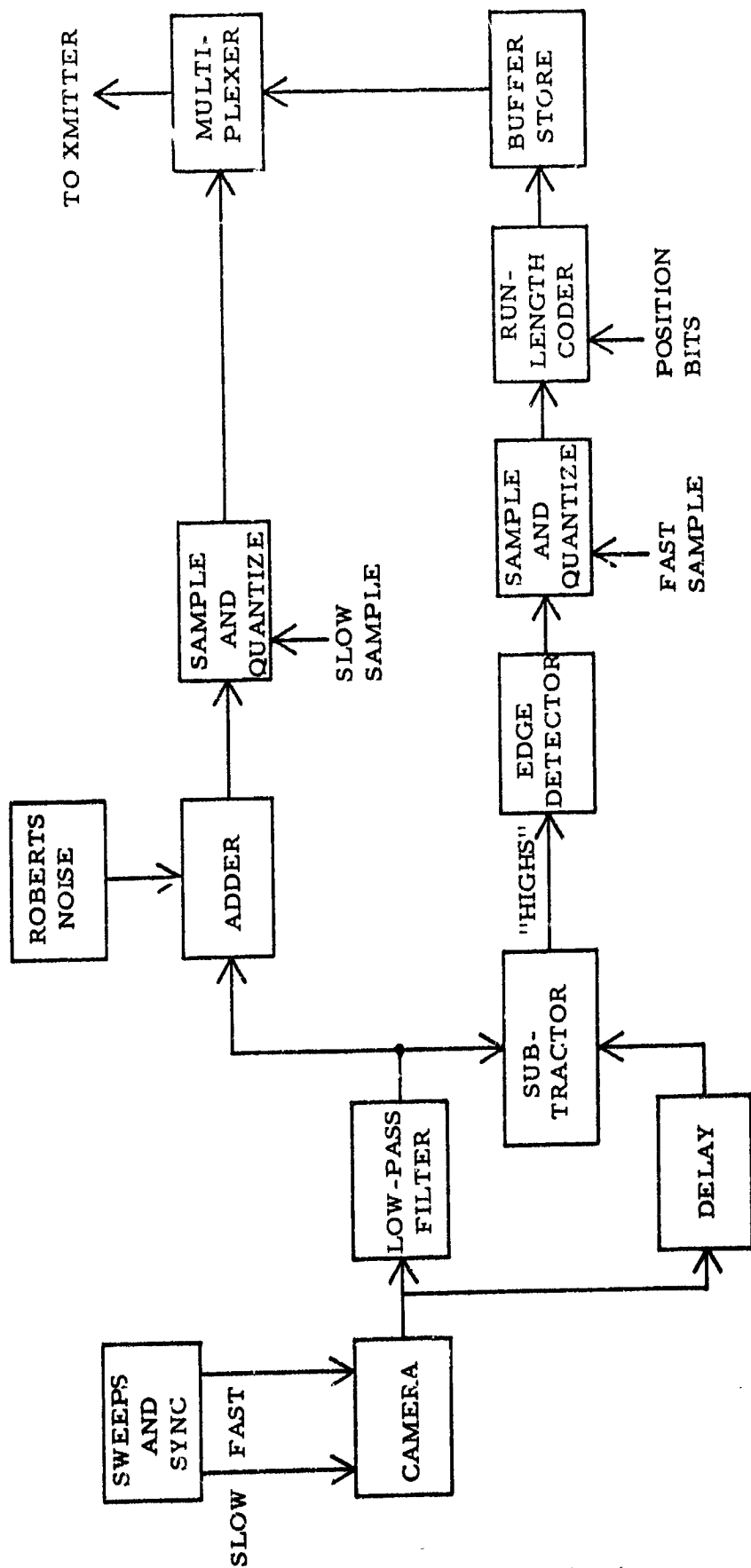


FIG. 2.2.4-i BASIC SPLIT-BAND ENCODER WITH ROBERT'S

ℓ = percentage of the video low passed, and

p = the number of bits necessary for quantization of "lows."

From the above definitions it is possible to calculate the "lows" and "highs" digital bit rates. Consider the "lows" digital bit rate.

Let

xy = total elements per frame,

ℓxy = total elements per second,

ℓf_{xy} = low-pass scan rate,

$(\ell xy)/2$ = low-pass filter cut-off frequency, and

$\ell p f_{xy}$ = "lows" digital bit rate.

Therefore, since the total number of elements per second is a fixed quantity, the "lows" digital bit rate depends directly on ℓp , the low-pass cut-off percentage and the "lows" quantization. No specific statement can be made for the low-pass percentage since this point is a function of the picture detail; however, the maximum quantization for the lows is six bits. To calculate the savings in bandwidth for the split-band technique, it is necessary to calculate the "highs" digital bit rate.

The "highs" digital bit rate can be calculated after determining the average edge run-length. Assume the element rate is fixed and is equal to f_{xy} . Therefore,

f_{xy} = element rate,

e = average edge run-length,

$e f_{xy}$ = average number of edges per second, and

$e f_{xy}$ = "highs" digital bit rate.

The quantization of the "highs" will depend upon the average edge run-length and the amplitude quantization of the edges. Therefore, with the "lows," element rate fixed, the digital bit rate will depend upon the average edge run-length and quantization. A general expression for the average compression can be found by obtaining the digital bit rate for the conventional PCM system. The PCM digital bit rate is $6 f_{xy}$. If it is assumed that the split-band system incorporates the necessary buffer stores, the average digital bit rate of the split-band system will be the sum of the "lows" and "highs" digital bit rates. Therefore, the average compression $\bar{C}$ can be found.

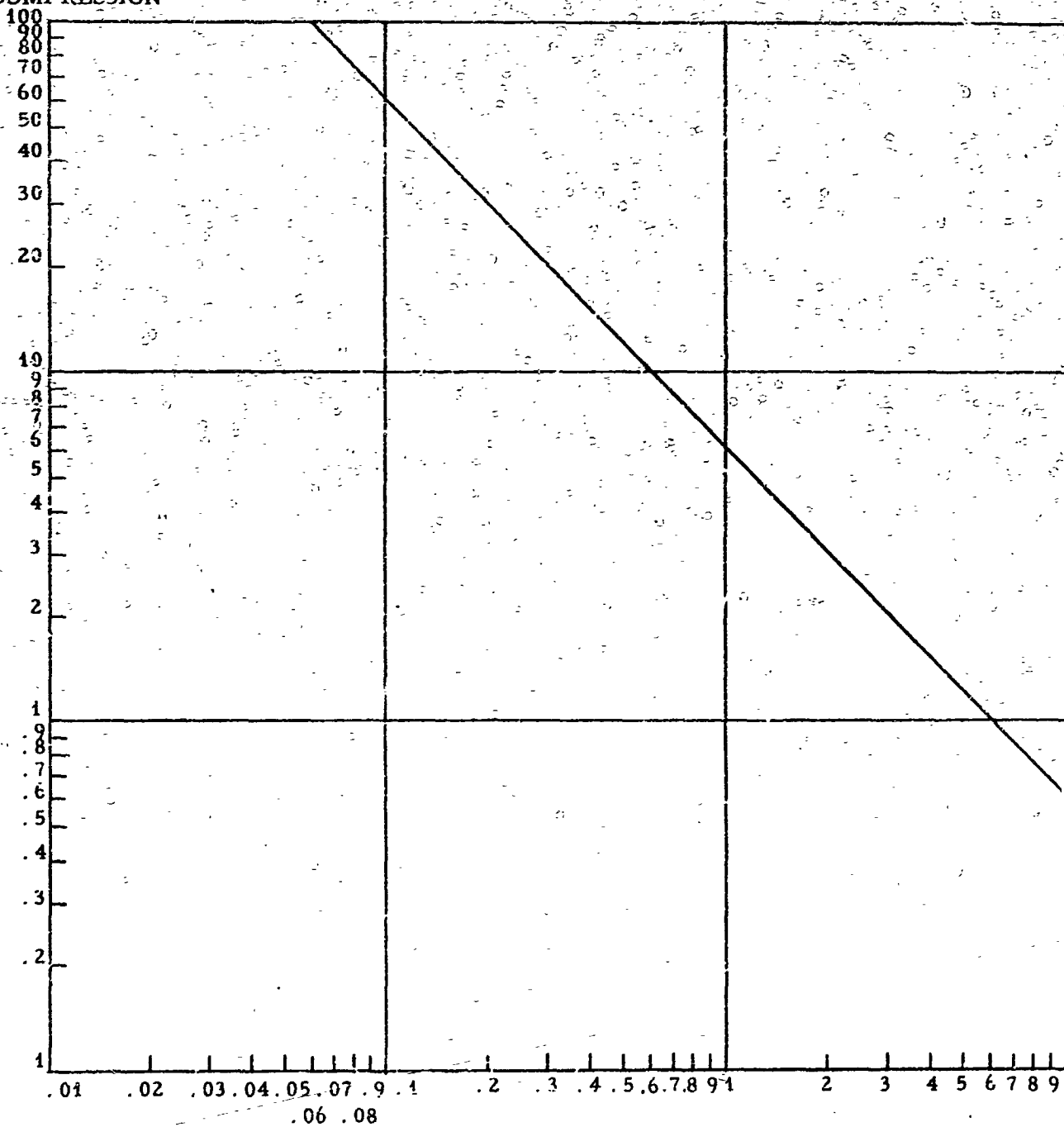
$$\begin{aligned}
\bar{C} &= \frac{\text{Basic PCM Digital Bit Rate}}{\text{Split-Band System Bit Rate}}, \\
&= \frac{6f_{xy}}{e r_{xy} + l_p f_{xy}}, \text{ which can be reduced to} \\
&= \frac{6}{e r + l_p}.
\end{aligned}$$

This expression shows that the average number of bits per element for the split-band system is $e r + l_p$. Therefore, it is possible to plot the average compression as a function of the average number of bits per split-band element, which is presented in Figure 2.2.4-2. Figure 2.2.4-3 gives a plot of the average compression as function of the average edge-run-length. It is also possible to form a nomograph for obtaining split-band compression. The nomograph was constructed and is given by Figure 2.2.4-4.

While the average compression for split-band coding is relatively simple to determine the split-band concept is quite involved to implement. The compression expression was derived from the basic split-band concepts without regard to the type of encoding and depends merely on quantities that can be statistically measured. However, before considering a possible hybrid system, assume that the channel is limited to one megabit per second. It is possible under this condition to obtain the maximum frame rate for various resolutions and bits per element. This is plotted in Figure 2.2.4-5.

Since the average compression is a function of the average number of split-band elements per second, decreasing this average could result in an increase in compression. Therefore, several experiments were conducted on EDITS to investigate the possibility of combining split-band coding with Roberts and PEC. Among the points of interest is a determination of the effects of the low-pass filter cut-off frequency on the PE signal. Previous investigators have determined that the low-pass cut-off frequency should be between 1/16 to 1/10 the bandwidth of the video signal. In this region the PE value varied from 0.910 to 0.880 as compared with a PE of 0.843 without low passing the video signal and without the Roberts technique. However, when the Roberts technique is added, the PE varied from 0.910 to 0.895 as compared with 0.875 without the low-pass filter. Figure 2.2.4-6 plots PE as a function of the low pass filter cut-off. Although this indicates that there is not a substantial increase in PE in the "lows," the increase is sufficient to warrant further investigation of split-band with Roberts and PEC. Consider the expression for split-band compression. Combining Roberts and PEC with the "lows" can effectively reduce the average number of bits per "lows" element by 1/2. However, this decrease does not imply that there is a 2:1 additional increase in compression. In fact, assuming the average edge run-length is 1 edge per 10 elements, using PEC on the "lows"

COMPRESSION



Average Bits Per Element

FIGURE 2.2.4-2 PLOT OF AVERAGE COMPRESSION AS A FUNCTION OF AVERAGE BITS PER ELEMENT FOR SPLIT-BAND CODING

COMPRESSION AS A FUNCTION OF AVERAGE EDGE RUN-LENGTH FOR SPLIT-BAND CODING WITH ROBERTS WITH AND WITHOUT PEC ON THE "LOWS"

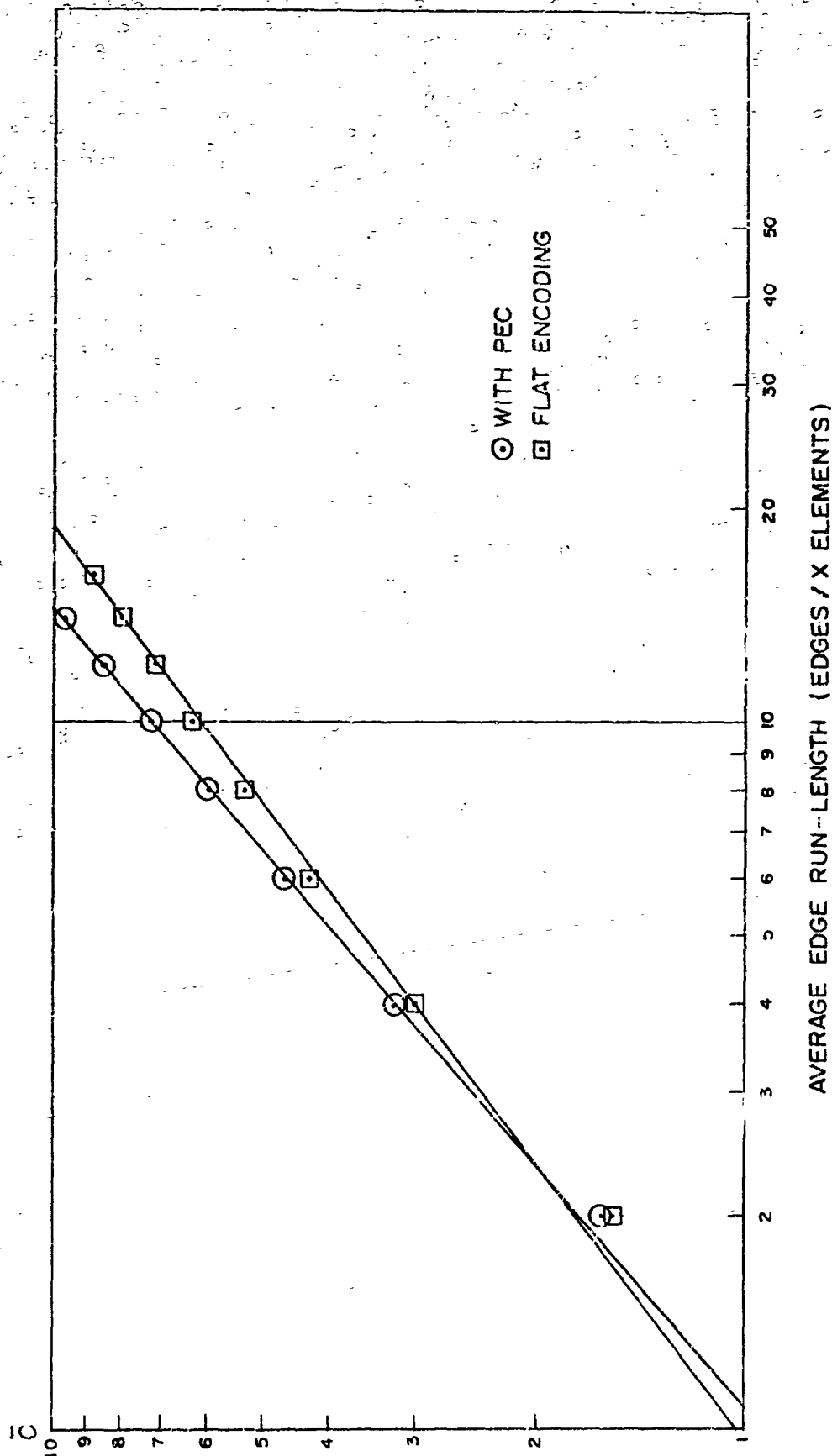


FIGURE 2. 2. 4-3 AVERAGE COMPRESSION AS A FUNCTION OF AVERAGE EDGE RUN-LENGTH FOR SPLIT-BAND CODING WITH ROBERTS WITH AND WITHOUT PEC ON THE "LOWS"

$\bar{C}$ = AVERAGE SPLIT-BAND COMPRESSION
 a = AVERAGE EDGE RUN LENGTH
 b = EDGE QUANTIZATION BITS
 c = LOW-PASS CUT-OFF POINT
 d = "LOW" QUANTIZATION IN BITS

$$\bar{C} = \frac{6}{ab + cd}$$

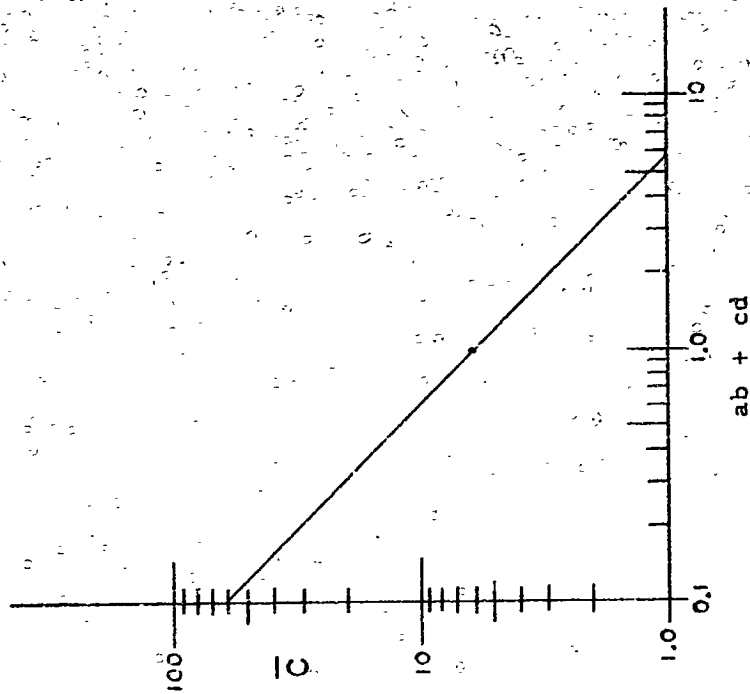
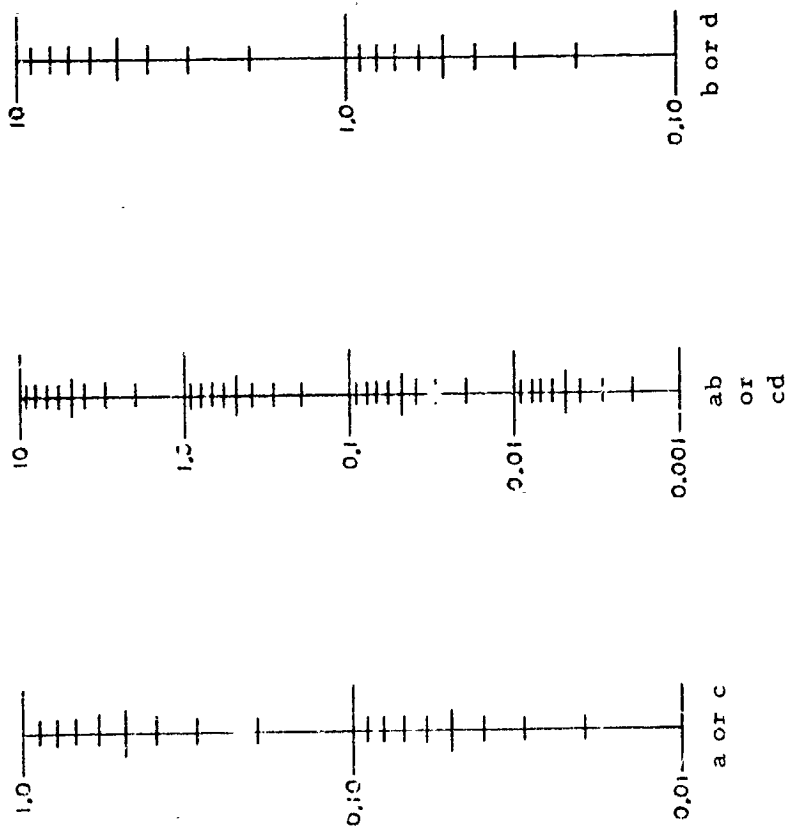


Figure 2.2.4-4 NCMOGRAPH FOR AVERAGE COMPRESSION OF SPLIT-BAND CODING

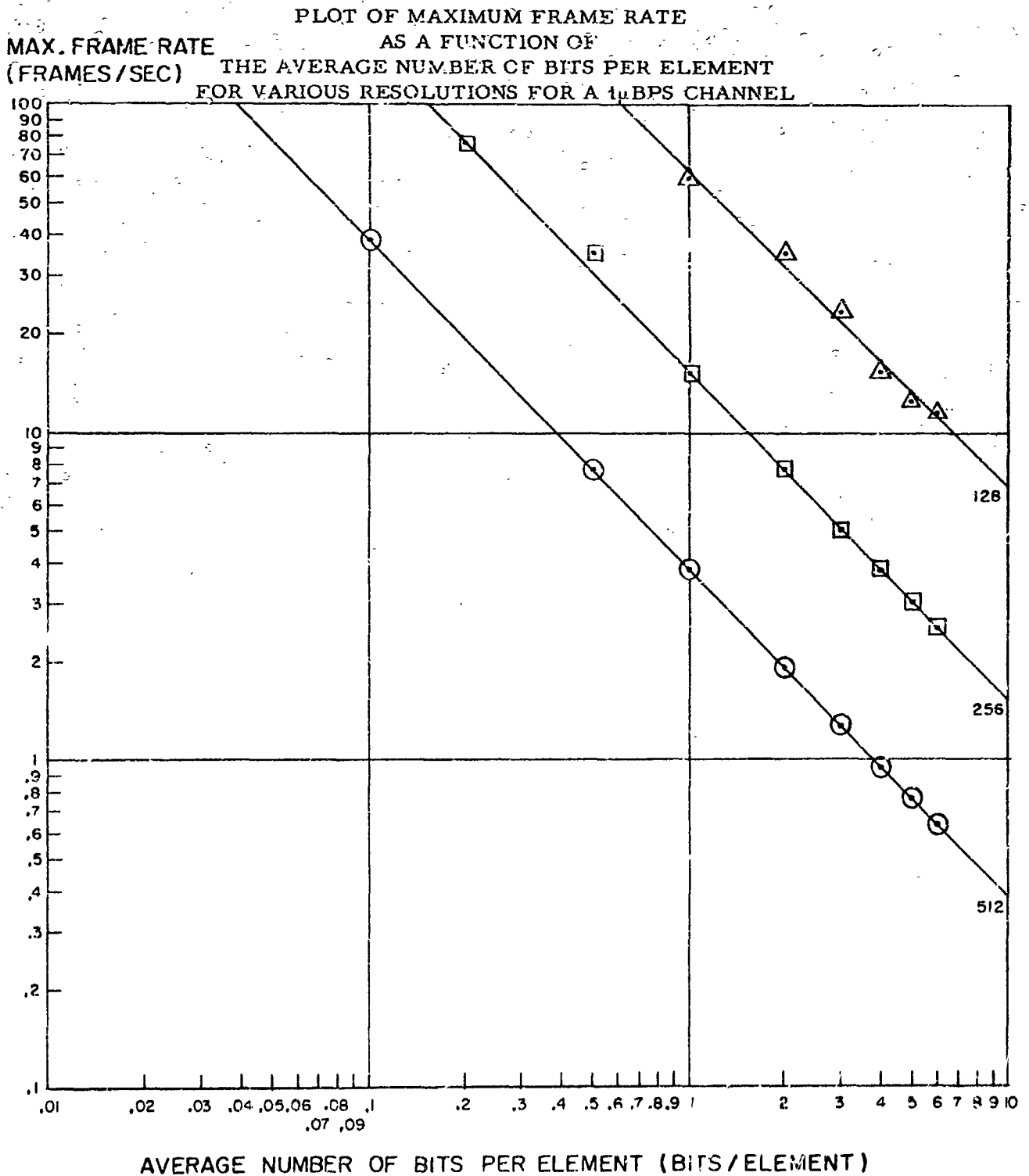


FIGURE 2. 2. 4-5 PLOT OF MAXIMUM FRAME RATE AS A FUNCTION OF THE AVERAGE NUMBER OF BITS PER ELEMENT FOR VARIOUS RESOLUTIONS FOR A 1 μ BPS CHANNEL

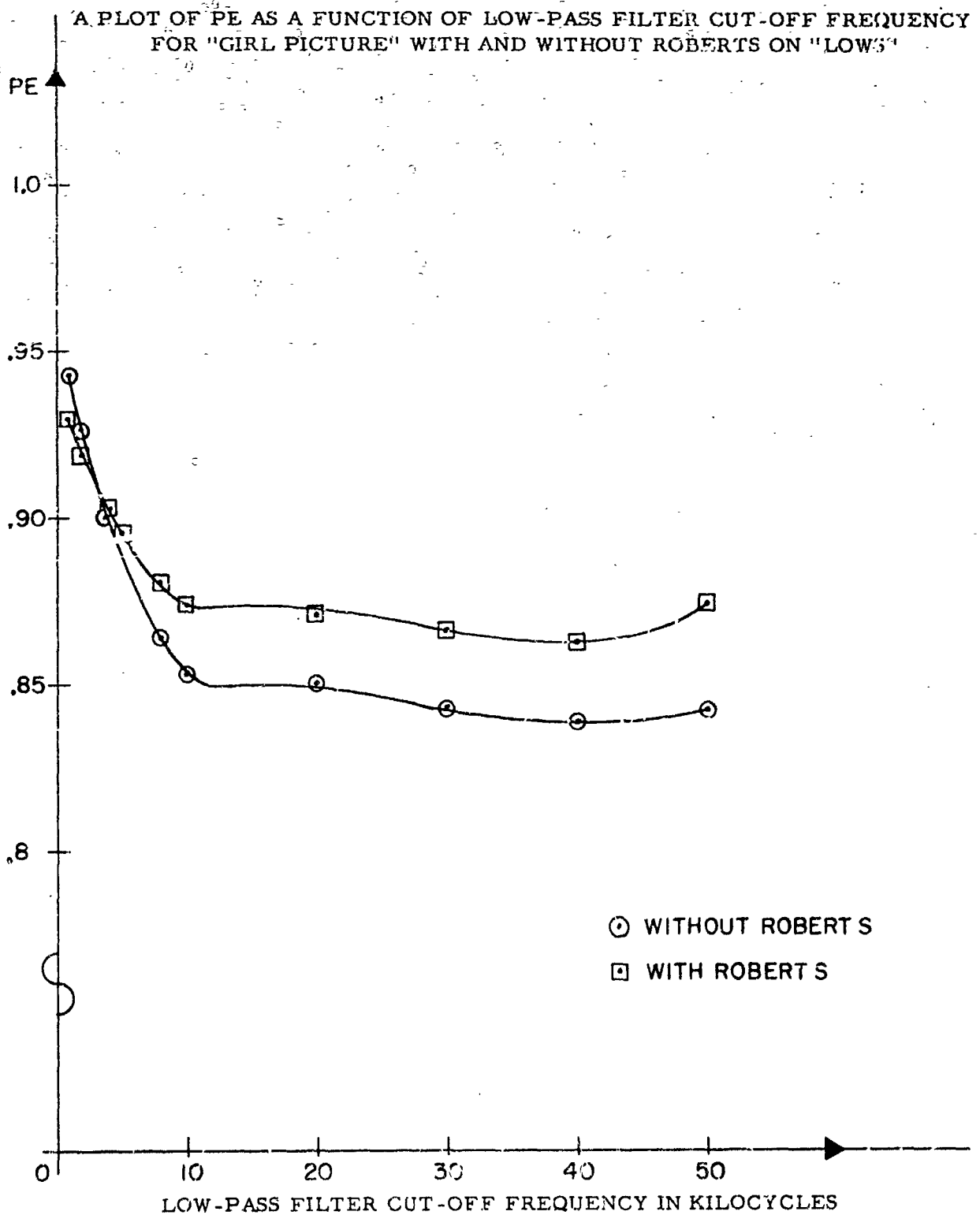


FIGURE 2.2.4-6

A PLOT OF PE AS A FUNCTION OF LOW-PASS FILTER CUT-OFF
FREQUENCY FOR "GIRL PICTURE" WITH AND WITHOUT ROBERT
ON "LOWS"

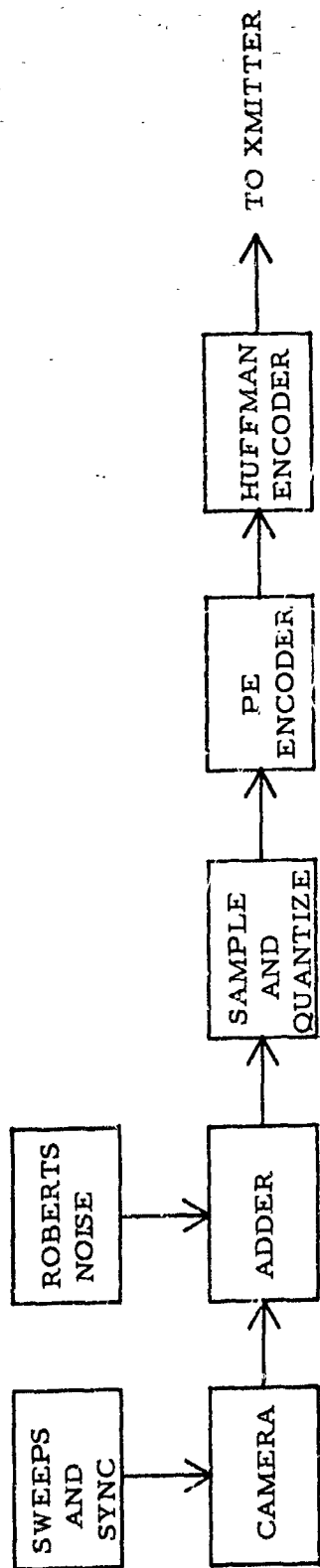


FIGURE 2.2.4-7 BASIC PEC ENCODER WITH ROBERTS

results in only a 1.15:1 increase in compression over conventional flat encoding of the "lows." Even if the average run-length is assumed to be 1 edge per 32 elements, PEC with Roberts would only result in a 1.4:1 increase in compression. Therefore, the added complexity of Roberts and PEC must be weighed quite carefully before considering the addition of these techniques to split-band coding.

It can be concluded the compression for split-band coding for average pictures is in the order of 6:1. Combining split-band coding with Roberts could increase the compression approximately 20% to 30%. Combining split-band coding with Roberts and PEC does not appear to be too feasible when comparing the increase in compression with the increased complexity. For example, a 2:1 increase in compression could result by combining split-band with PEC and Roberts, but at the expense of a 4:1 increase in complexity as indicated by comparing the addition of boxes in Figure 2.2.4-7 to those in Figure 2.2.4-1.

2.2.5 Delta Modulation

Delta modulation is a digital technique similar to pulse code modulation; however, the delta system takes advantage of the information previously sent to reduce the number of bits per element. Delta modulation had its conception in France in 1946⁽³¹⁾ and since that time many investigators⁽³²⁻⁴⁸⁾ have applied the technique and modifications thereof to the transmission of video and speech signals. The basic idea of delta modulation can be obtained by considering the single-bit system. The single-bit system block diagram is given in Figure 2.2.5-1 and its operation is quite similar to a servo loop.

The input analog voltage to be transmitted digitally, e_i , is compared to an approximation of the input voltage fed back, e_f , through an integrator. The difference voltage between these two signals, e , is applied to a pulse modulator which generates a pulse whose polarity depends upon the polarity of the difference signal. This pulse is sent to the transmitter and to an integrator to form an approximation of the input waveform. During the next sampling interval, the input and the approximation of the input are compared and the new difference signal is sent to the pulse modulator. Therefore, the time output of the pulse modulator is a sequence of positive and negative pulses whose polarity depends upon the difference signal. If it is assumed that the receiver conditions the transmitted pulses, the single-bit encoder consists of merely an integrator and low-pass filter.

The single-bit delta system can be modified to optimize the performance by the appropriate trade-off of system parameters. The integrator could be a simple linear integrator with a long time constant in comparison to the sampling period; however, it has many undesirable characteristics. For example, a noise pulse in the data link can cause an error to propagate over many elements until the integrator is reset. Also, the total number of levels possible is limited by the integrator characteristics. Both these problems can be reduced by employing either double integration or an exponential integrator.

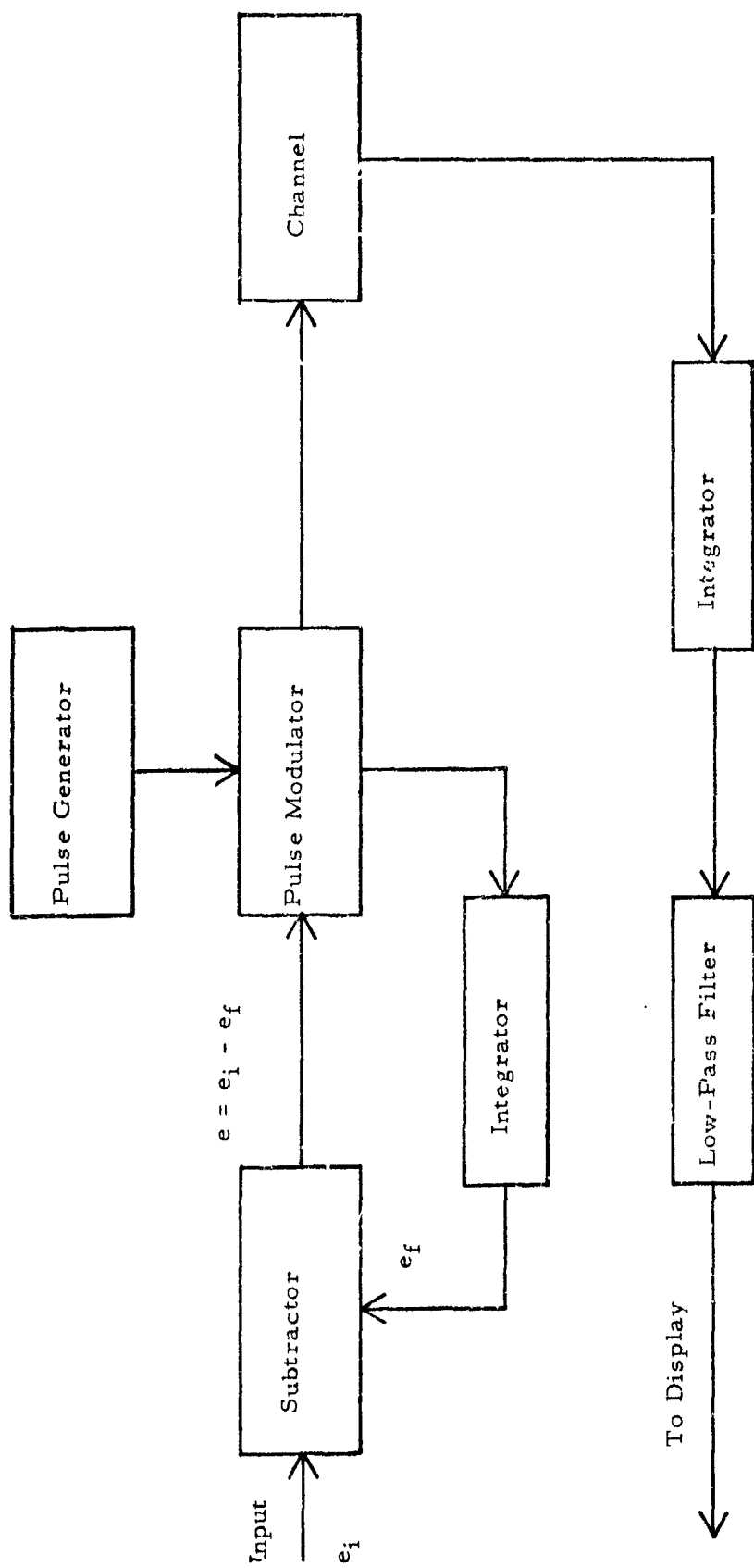


FIGURE 2.2.5-1 BASIC BLOCK DIAGRAM OF SINGLE-BIT DELTA MODULATOR ENCODER AND DECODER

One of the major limitations of the delta system is the high sampling rate required to obtain a subjectively acceptable picture. Unless the sampling rate is high enough compared to the highest frequency component in the video, the advantages gained by going to a nonlinear integrator are lost. From this it is apparent that a delta system is slope-limited rather than amplitude-limited. The maximum slope or rise time that can be reproduced is dependent upon the delta step height and the sampling rate. This maximum value can be increased for a fixed sampling rate by a multi-bit delta system. However, as the number of delta quantizing bits is increased, the system complexity increases, thus defeating somewhat the main advantage of the delta system, simplicity. An alternate method would be to use Δ^2 modulation. Here, the steps are weighted in increasing amplitude provided the difference signal maintains the same polarity. This system has a better step function response.

In the multibit system the difference signal is quantized into several discrete positive and negative levels. Figure 2.2.5-2 gives a block diagram of an N-bit delta modulation system. If the delta step height is h and the sampling period is k for a single-bit system, the maximum slope that can be reproduced is h/k . However, if the difference signal is quantized to n levels where $N = \log_2 n$, the maximum slope will be greater than h/k . The subjective performance of an N-bit system increases, but at the expense of the increase in complexity and an increase in the output bit rate by N for the same sampling frequency as a single-bit system. A compromise must be reached between the step height size, sampling rate, fidelity, and bandwidth for the correct integrator.

A linear integrator definitely places rather severe restrictions on the system performance. The number of reproducing levels is limited and constant because of the relatively long time constant of the integrator. With a finite number of possible levels, a single-bit system will display a certain amount of contouring. To decrease the contouring a smaller step height could be used but this limits the maximum slope that could be reproduced. It is necessary to compromise between fidelity and contouring. The noise performance of a linear integrator in the system is rather poor since a transmission error will smear along an entire line instead of effecting only one element as the PCM system. These particular features of the linear integrator can be reduced by employing an exponential integrator.

If the response of the integrator is related to the sampling period, the compromise between the step height and sampling period can be reduced. Contouring is reduced because of the nonlinear characteristics of the integrator, that is, the integrator rarely decays to the same intensity value for two successive steps. Therefore, the number of intensity levels for the exponential integrator is almost infinite and a continuum of intensity levels is produced. The noise performance of a delta system with an exponential integrator is increased considerably since the integrator has a finite response; thus, transmission errors could propagate to only a few succeeding elements. However, if the sampling rate is too low in comparison to the video, an annoying oscillation occurs. It is then necessary to compromise between the sampling period and

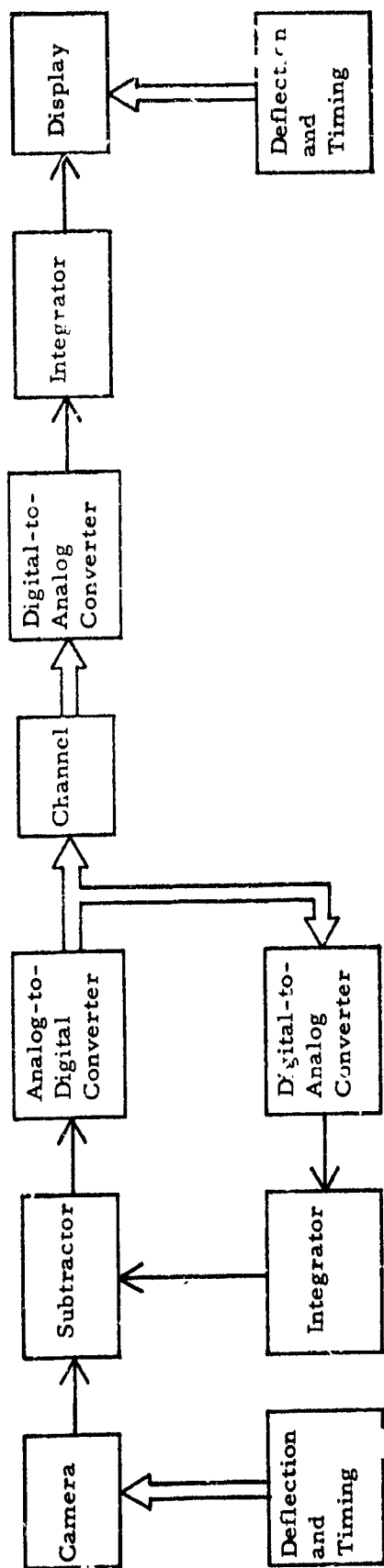


FIGURE 3.2.5-2 N-BIT, DELTA-MODULATION COMPRESSION SYSTEM

the time constant of the integrator. Fortunately, an optimum relationship can be achieved between these two quantities. The advantages of the exponential integrator for the delta system are: reduced contouring since the number of possible reproducing levels is nearly continuous with variable steps, good noise immunity since the time constant of the integrator is finite, and advantageous use of the statistical properties of the video since the correlation between successive elements is high.

The final delta system will represent a compromise between (1) image fidelity, (2) data link noise sensitivity, (3) transmission bit rate, (4) physical size, (5) weight, (6) power, and (7) complexity. In general, a PCM system uses the bandwidth more efficiently than a delta system, but requires a sizable increase in complexity to achieve this. The major limitations in the subjective quality of the delta system are the magnitude of the minimum change in signal that can be detected, and the time in which a change in amplitude can occur. Although a delta system would appear to require a much higher signal-to-noise ratio than a PCM system, it has been shown that the delta system requires only one db more signal-to-noise ratio than a PCM system.

To illustrate some of the previously mentioned characteristics of a delta system, a graphical analysis of a typical waveform is performed. The fidelity of the displayed waveform can be improved by increasing the sampling rate. Figure 2.2.5-3 plots the displayed waveform for a single-bit system sampled at f_s and $2f_s$ and assuming a linear integrator for graphical convenience. If, however, the sampling rate is fixed, the fidelity of the delta system can be improved by employing a multibit delta encoder; therefore, Figure 2.2.5-4 shows the output of a two-bit delta system. The quantization was accomplished under the assumption that the large step should be $\log_2 n$ where n is the number of levels required. For the graph the small step was arbitrarily chosen to be $E_{\max}/28$ which requires that the large step be $5E_{\max}/28$. To obtain a better reproduction of the waveform, the large step is taken if the difference signal is greater than or equal to $4E_{\max}/28$. Since there is an inherent delay in the delta encoder, the displayed output waveform was advanced in time one sampling interval and plotted in Figure 2.2.5-5 to illustrate the fidelity of the two-bit delta system. Since delta modulation will be compared to PCM, the waveform was encoded to three bits and is plotted in Figure 2.2.5-6. A three-bit PCM system will display contouring; therefore, three bits of pseudo-random noise is added to the waveform before quantization and the waveform plus noise is quantized as shown in Figure 2.2.5-7. This is one-half of Roberts technique. Figure 2.2.5-8 shows the displayed waveform after the pseudo-random noise is subtracted. Therefore, for the given waveform, it is possible to compare the two-bit delta and three-bit Roberts. The displayed waveforms for both systems are given in Figure 2.2.5-9. From this figure the fidelities of the two systems appear to be about the same. However, the results obtained from EDITS do not agree with this assumption, since the Roberts pictures are subjectively better than the delta pictures at the same resolution. The average compression for the delta system is $\bar{C} = 3:1$.

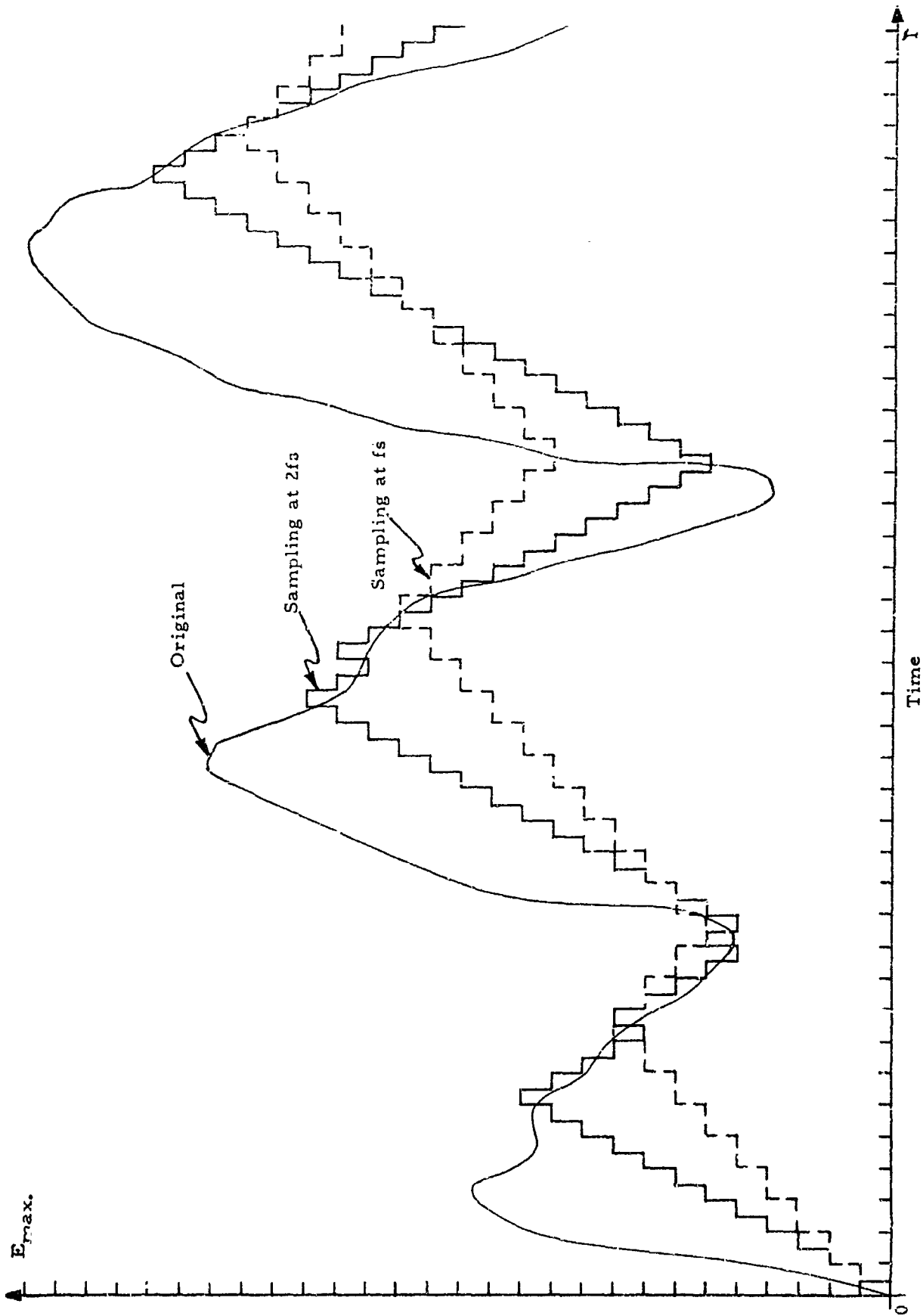


FIGURE 2.2.5-3 A PLOT OF A SINGLE-BIT DELTA-MODULATION OUTPUT AS A FUNCTION OF SAMPLING FREQUENCY

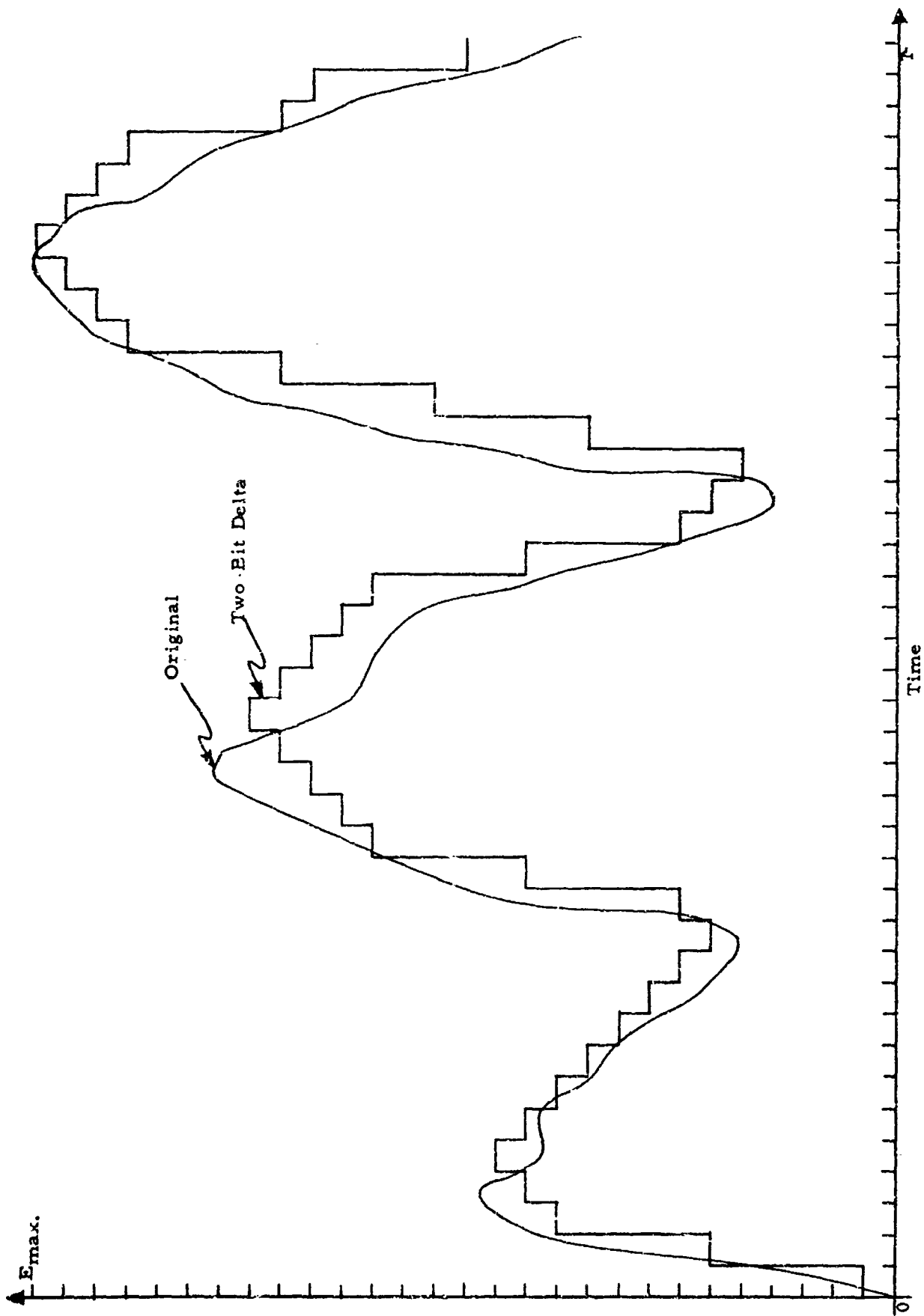


FIGURE 2.2.5.4 A PLOT OF OUTPUT OF A TWO-BIT DELTA-MODULATION SYSTEM WHERE THE LARGE STEP

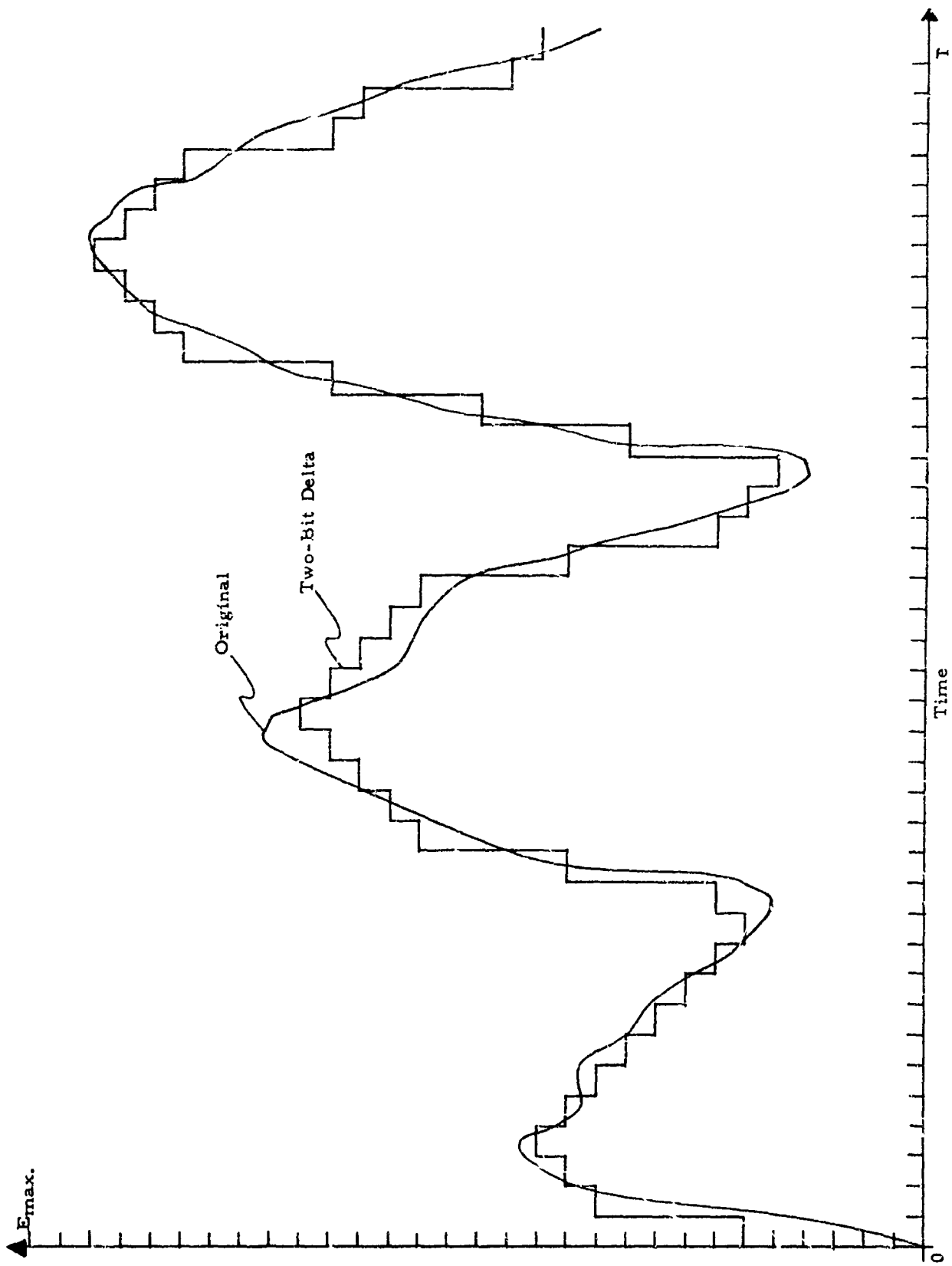


FIGURE 2.2.5-5 A PLOT OF TWO-BIT DELTA SYSTEM OUTPUT NOT DELAYED, FOR LARGE STEP EQUAL TO FIVE TIMES SMALL STEP SIZE, WITH THRESHOLD SET AT FOUR TIMES SMALL STEP SIZE

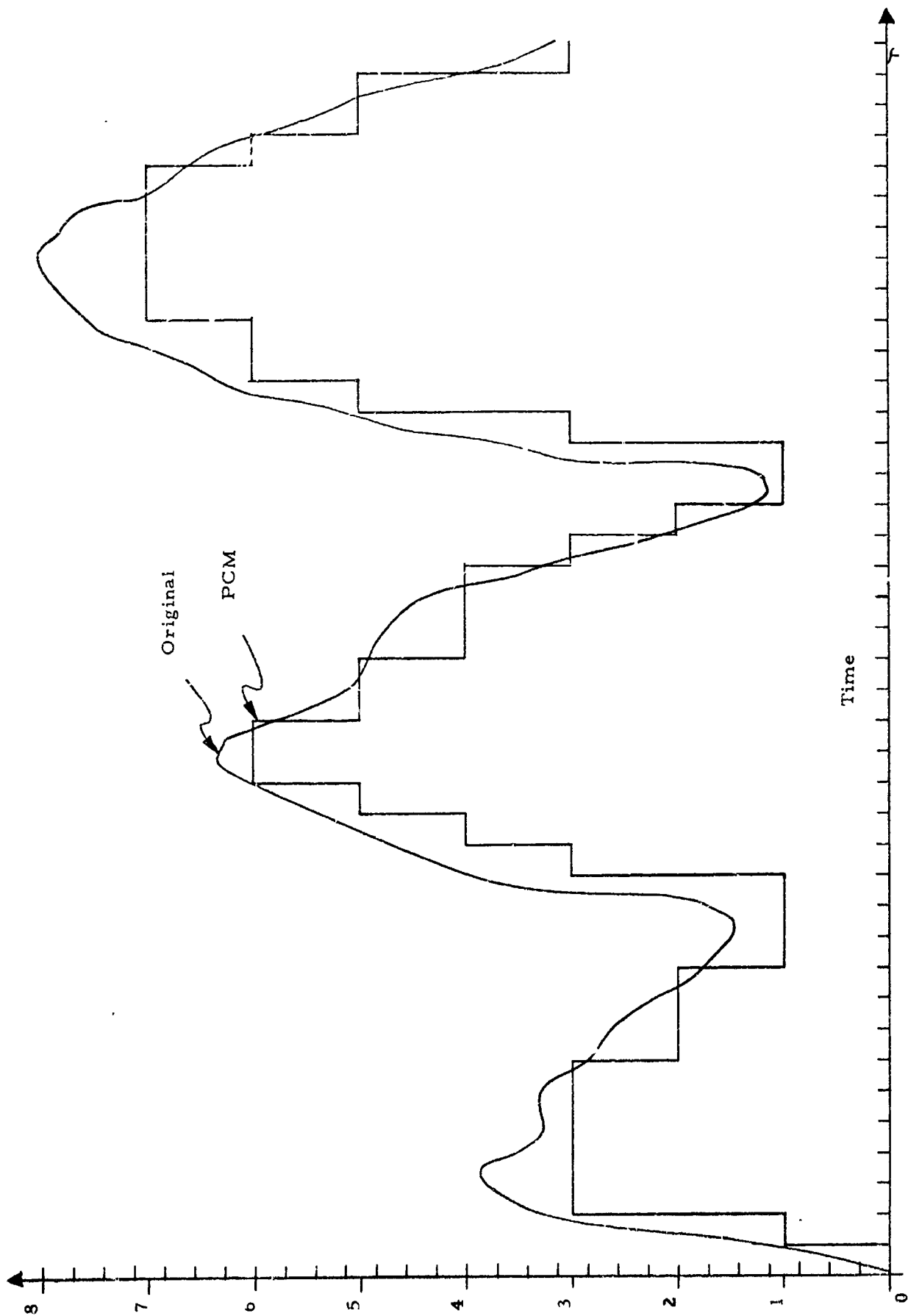


FIGURE 2.2.5-6 OUTPUT OF A THREE-BIT PCM SYSTEM AS A FUNCTION OF TIME FOR SAME SAMPLING RATE AS DELTA SYSTEM

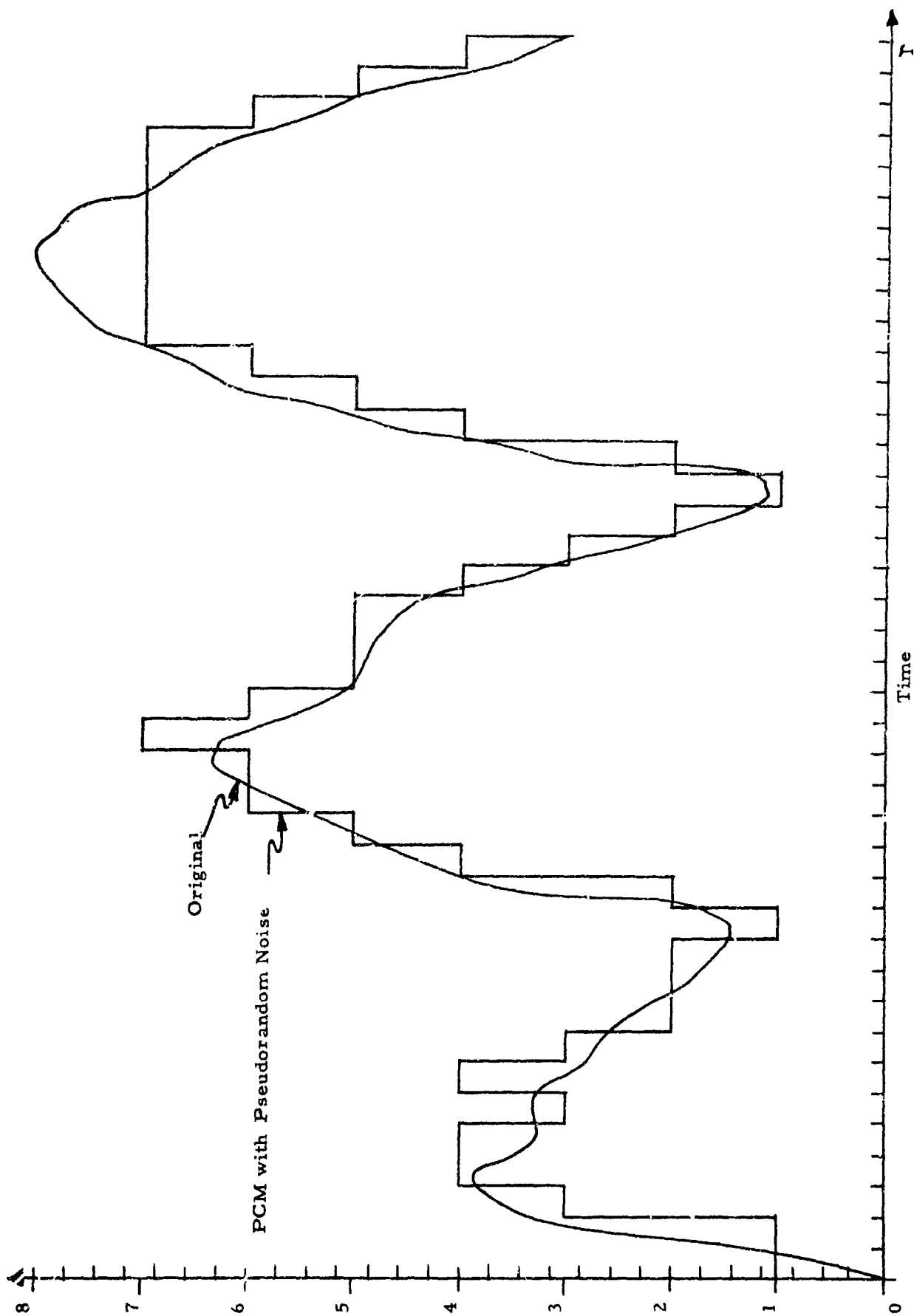


FIGURE 2.2.5-7 TRANSMITTED WAVEFORM OF THREE-BIT PCM SYSTEM WITH THREE BITS OF PSEUDO-RANDOM NOISE ADDED BEFORE QUANTIZATION AT THE SAME SAMPLING RATE AS THE DELTA SYSTEM

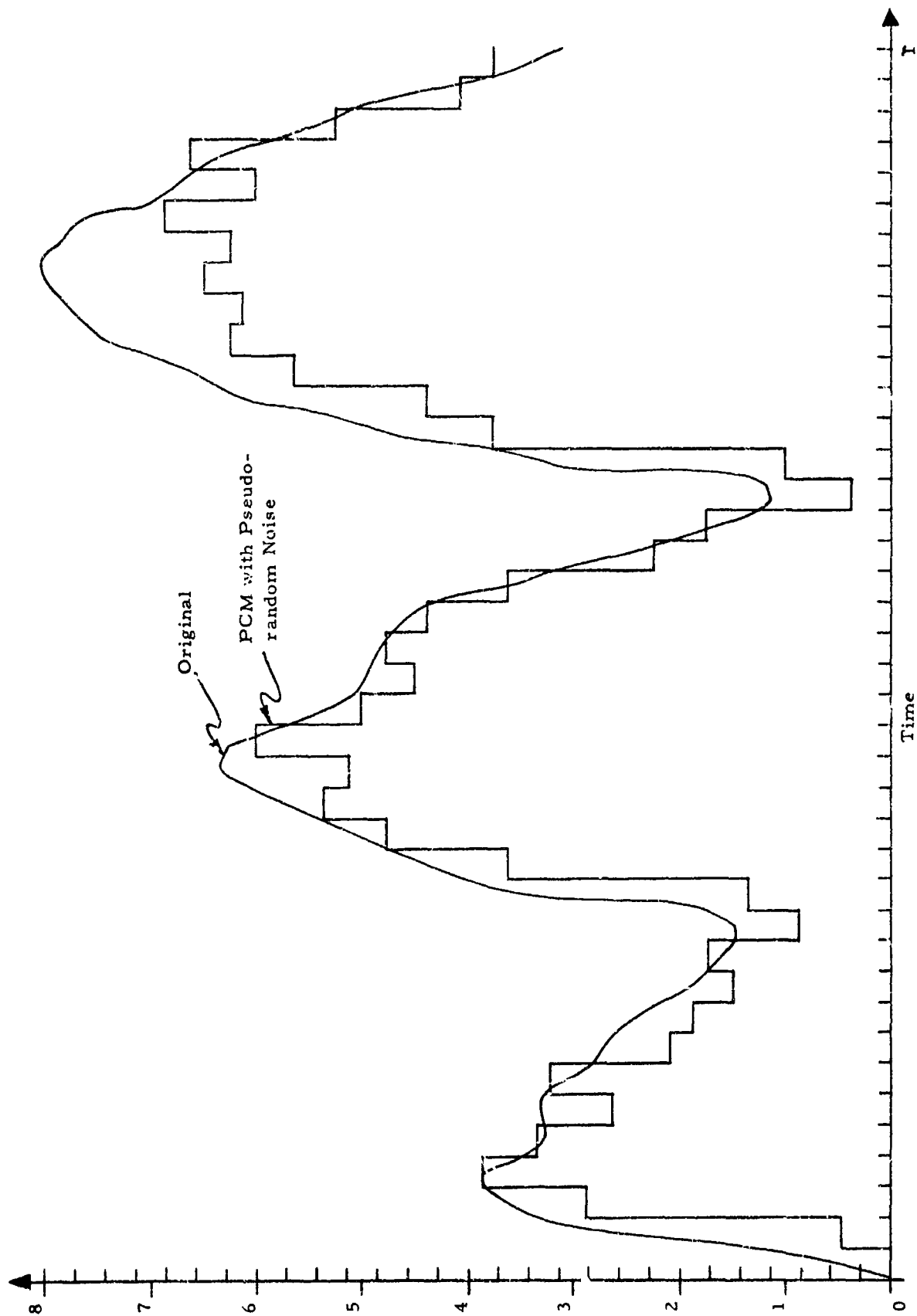


FIGURE 2.2.5-8 DISPLAYED WAVEFORM FOR A THREE-BIT PCM SYSTEM WITH THREE BITS OF PSEUDO-RANDOM NOISE ADDED AND SUBTRACTED AT THE SAME SAMPLING RATE OF THE DELTA SYSTEM

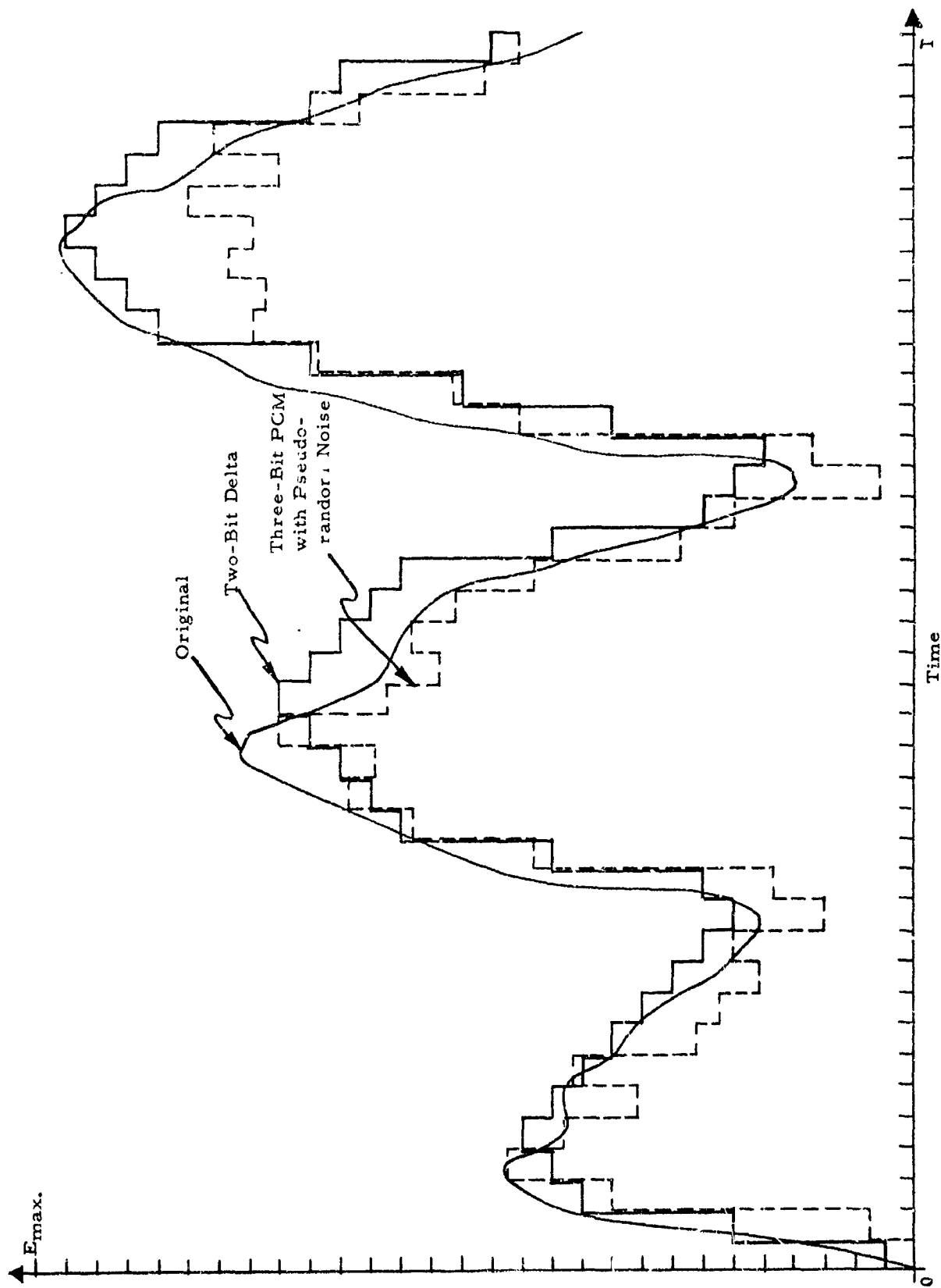


FIGURE 2.2.5-9 OUTPUT OF A TWO-BIT DELTA SYSTEM AND A THREE-BIT PCM SYSTEM WITH PSEUDO-RANDOM NOISE ADDED AND SUBTRACTED

The discussion of the delta system up to this point represents the basic concepts of the system. In keeping with EMR's philosophy in coding, the delta encoder is considered as an information source to determine whether further compressions could be achieved. Normally the statistical properties of the source can be defined from the first-order probabilities; however, the statistics for the delta source must be determined from the difference signal and not the original video. Without too much difficulty it can be shown that the probability distribution of the difference signal is given by Figure 2.2.5-10.

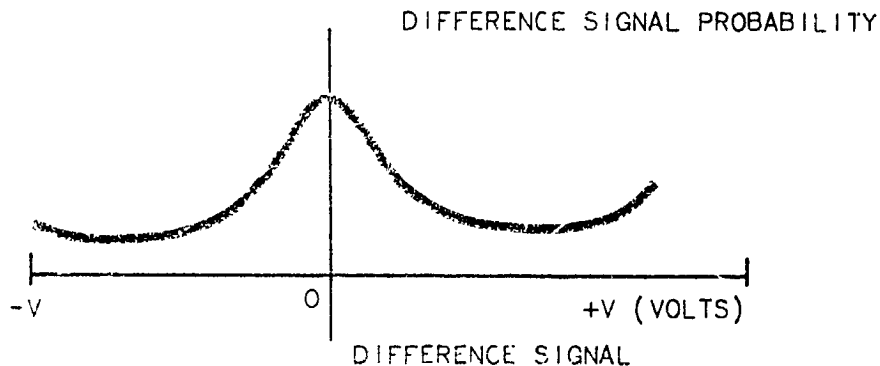


FIGURE 2.2.5-10
DIFFERENCE SIGNAL PROBABILITY DISTRIBUTION

The type of distribution is well-suited for previous-element coding to decrease the entropy of the source. For example, assume that difference signal is quantized to four levels and previous-element coded. Assume that the large step occurs with a probability of x . A good approximation to the probability of the small step would be kx where $k \geq 1$. If the probability of PE is y , then from probability theory $y + 2x(k + 1) = 1$. Assuming the $y \geq 2x(k + 1)$, it is possible to assign a Huffman code for the five events and obtain the encoded entropy or the average number of bits per element. Table 2.2.5-1 gives the Huffman code for $k = 2$.

Unfortunately, the entropy of the delta source depends upon the Huffman code, which in turn depends upon the probability of the event; therefore, a general expression cannot be given. Figure 2.2.5-11 was constructed to show the relationship between encoded entropy, delta probability, and PE probability. This imposes some rather severe restrictions on the PE. To achieve a gross compression with previous-element coding and delta of 3:1, the PE must be greater than 0.5 for $k = 2$. In fact, to increase the compression from 3:1 to 4:1 with PEC, it is necessary that the PE be greater than 0.75. For $k > 2$, the requirements on the PE are lessened somewhat, but a new set of Huffman code words must be defined.

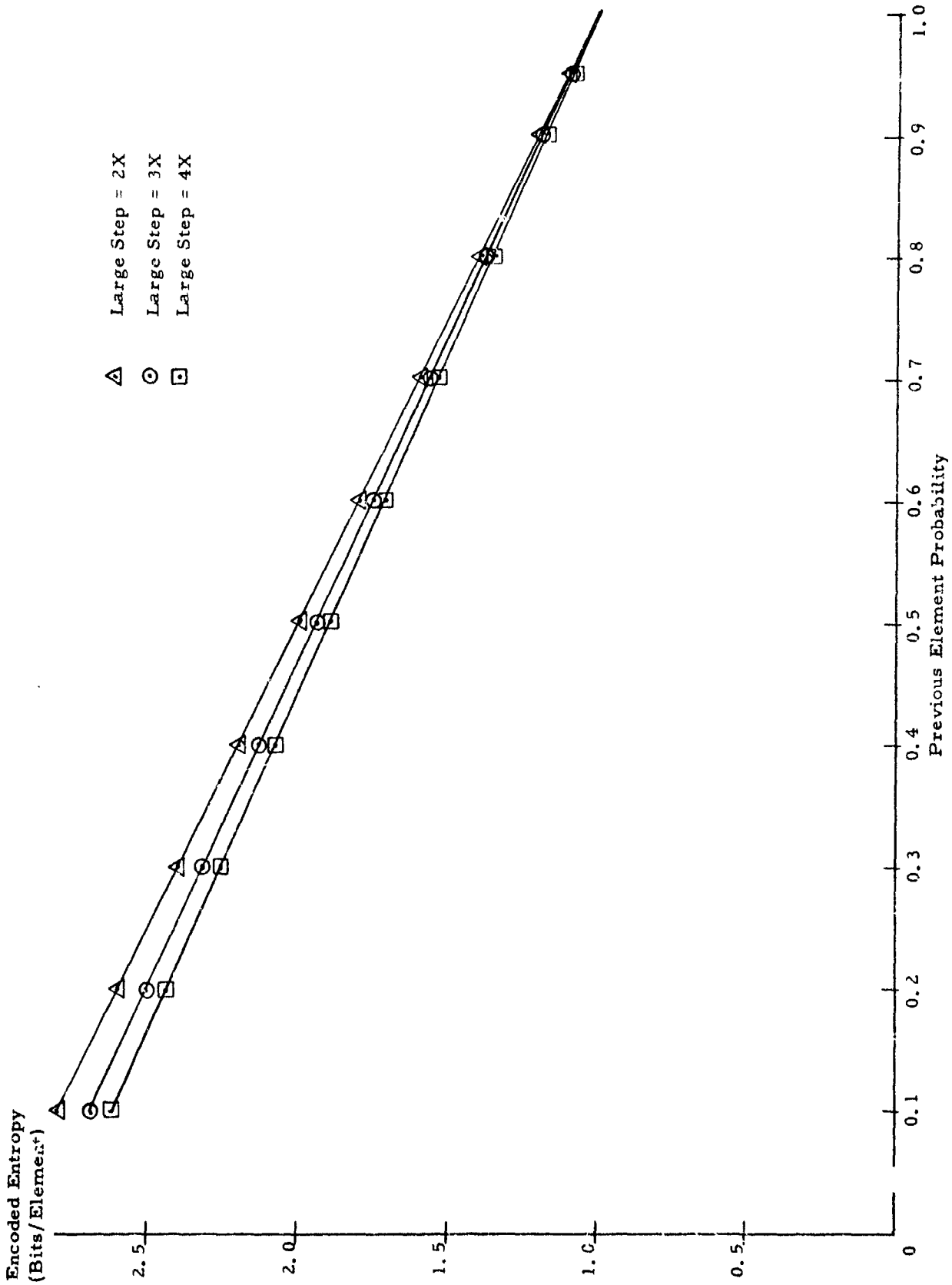


FIGURE 2.2.5-11 A PLOT OF ENCODED ENTROPY AS A FUNCTION OF PREVIOUS ELEMENT PROBABILITY FOR A TWO-BIT DELTA-MODULATOR SOURCE

TABLE 2. 2. 5-1

HUFFMAN CODE FOR PREVIOUS ELEMENT
CODING OF A TWO-BIT DELTA SOURCE

EVENT	PROBABILITY	HUFFMAN CODE
PE	'	1
+1	2x	011
-1	2x	010
+5	x	001
-5	x	000

If the sampling rate is high enough, a two-bit delta system is subjectively equivalent to a six-bit PCM system. At lower sampling rates PCM is superior to delta. It would be possible to increase the compression of a delta system by employing previous-element coding if the statistics of the difference signal can satisfy the previously mentioned requirements. However, this increases the complexity of the over-all system by a considerable factor; thus eliminating simplicity -- a major advantage of a delta system.

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2.3 LINE-TO-LINE CODING

Previous-element coding takes advantage of the redundancy that exists between adjacent elements along a line. Previous-element previous-line (PE PL) coding extends previous element coding in the vertical direction to take advantage of the correlation between adjacent elements in successive lines. Thus, PE PL coding is a two-dimensional coding technique that utilizes both element-to-element and line-to-line redundancy.

The basic system operation will be briefly described. A picture will be PEC coded into the PE and (1-PE) signals. Adjacent lines will be compared and if the (1-PE) signals on the two adjacent lines are alike, a previous line (PL) signal will be transmitted. Although it would have been possible to encode the PE signals on adjacent lines with the PL signal, experiments indicate this would degrade the over-all system performance. Obviously encoding the PE signals with the PL criteria would decrease the PE value, thus reducing the effectiveness of the Huffman code. However, it is possible to estimate the compression that could be achieved by assuming statistics for the video.

As a lower bound on compression, assume that the (1-PE-PL) statistics have a flat distribution. Further, assume that the PL probability is greater than the sum of the probabilities of the signals that are neither PE or PL. For a flat distribution of the (1-PE-PL) statistics the Huffman code words can be determined. Each of these elements will then have $N + 2$ bits, the PL signal will have two bits, and the PE signal will have one bit. The encoded entropy can be determined by

$$H = \sum_{i=1}^h L(i) P(i) ,$$

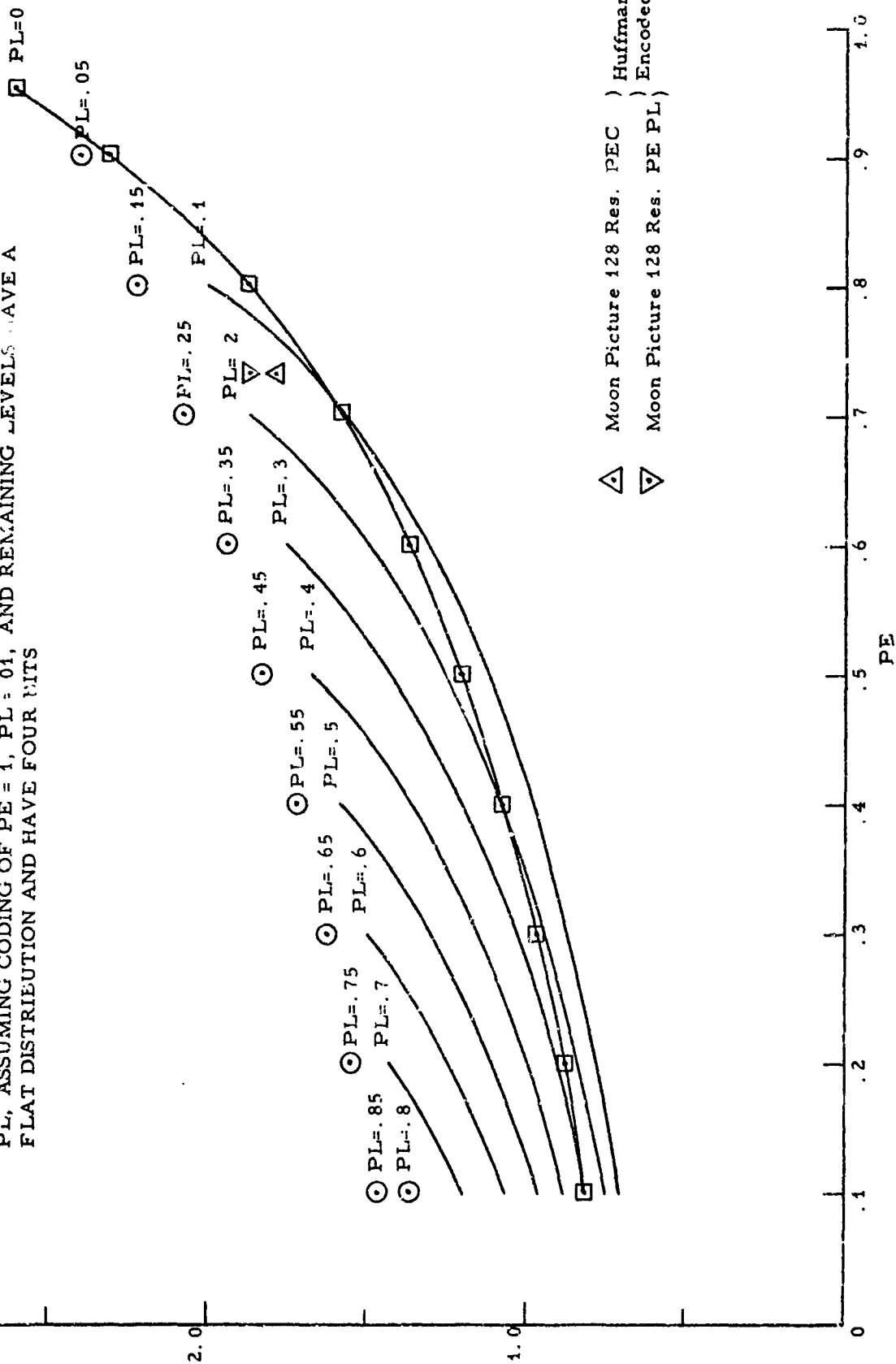
where L is the number of bits for event i and $P(i)$ the probability of event i . For the flat distribution the encoded entropy is $H_F = PE + 2PL + (1-PE-PL)(N + 2)$ which reduces to $H_F = N + 2 - (N + 1) PE - NPL$.

The compression can then be determined from $\bar{C}_F = N/(N + 2 - (N + 1) PE - NPL)$.

Two cases will be given to illustrate the effectiveness of PE PL coding of a six-bit and three-bit PCM source. For certain applications the three-bit PCM with Roberts will be subjectively acceptable. Therefore, assuming a three-bit PCM source is to be compressed, the compression for a flat distribution is $\bar{C}_{F3} = 3/(5-4PE-3PL)$. Since the compression is a function of PE and PL and at present the statistical relationship between PE and PL has not been determined, a plot of compression as a function of PE parametric in PL was plotted in Figure 2.3-1. Given the values of PE and PL, the compression can be

$\hat{C}_{F3}$

FIGURE 2.3-1 PLOT OF COMPRESSION FOR PE PL CODING AS A FUNCTION OF PE FOR VARYING PL, ASSUMING CODING OF PE = 1, PL = 01, AND REMAINING LEVELS HAVE A FLAT DISTRIBUTION AND HAVE FOUR BITS



△ Moon Picture 128 Res. PEC) Huffman
▽ Moon Picture 128 Res. PE PL, Encoded)

determined. As an illustration, assume that it is desired to have a minimum compression of 2:1. What must be the statistics in order to achieve this?

As an approximation to an upper and lower bound, it has been experimentally determined that for three-bit quantization, PE generally varies from 0.7 to 0.9. Therefore, for PE = 0.7, PL must be at least 0.233 to achieve the desired compression; however, for PE = 0.9, PL must be at least 0.033. That is, the average number of bits per element must be 1.5 for a PE PL system. Results must be obtained from EDITS to determine whether or not the system will be feasible.

During preliminary investigations, it was noticed that the PE signal was approximately the same regardless of whether or not the picture was scanned vertically or horizontally. Therefore, if the correlation is reduced in one direction, then the elements that are alike in the other direction should be related to PE. It seems reasonable that the approximation of $PL = PE(1-PE)$ could be valid. Under this assumption it is only necessary to specify PE to obtain an estimate of the compression. The compression is given by $\bar{C}_{FR} = 3/(4-5PE + 2PE^2)$. This expression is plotted in Figure 2.3-2. To achieve a 2:1 compression it is necessary that PE be greater than or equal to 0.69. The value of PE does not appear to be too unrealistic for three-bit quantization.

Although the flat distribution sets a lower bound on compression, it is possible to increase the compression by making maximum use of the Huffman code. Although the Gaussian distribution is not an optimum distribution, it does provide an improvement over assuming a flat distribution. Several investigators have stated that the video statistics are Gaussian in nature. Therefore, the Gaussian distribution was divided into eight equal parts and the associated probabilities were obtained. These probabilities were coded with the Huffman code and the encoded entropy for a Gaussian distribution with the PE and PL signals were obtained. The compression is given by $\bar{C}_{G3} = 3/(4.207 - 3.207 PE - 2.207 PL)$. For a 2:1 compression with PE = 0.7, PL must be at least 0.209. However, with PE = 0.9, the compression for a Gaussian distribution is greater than 2:1 independent of PL.

For various other applications it may be necessary to compress a six-bit source. For a flat distribution of the (1-PE) statistics the compression is $C_{F6} = 6/(8-7PE-6PL)$. Figure 2.3-3 was constructed plotting compression as a function of PE parametric in PL. To achieve a compression of 2:1 for the six-bit source it is necessary to determine the relationship between PE and PL. Fortunately, for six-bits statistics the average value of PE was obtained. Nine pictures were processed on EDITS and the average value of PE was 0.519. The pictures and PE are tabulated below in Table 2.3-1.

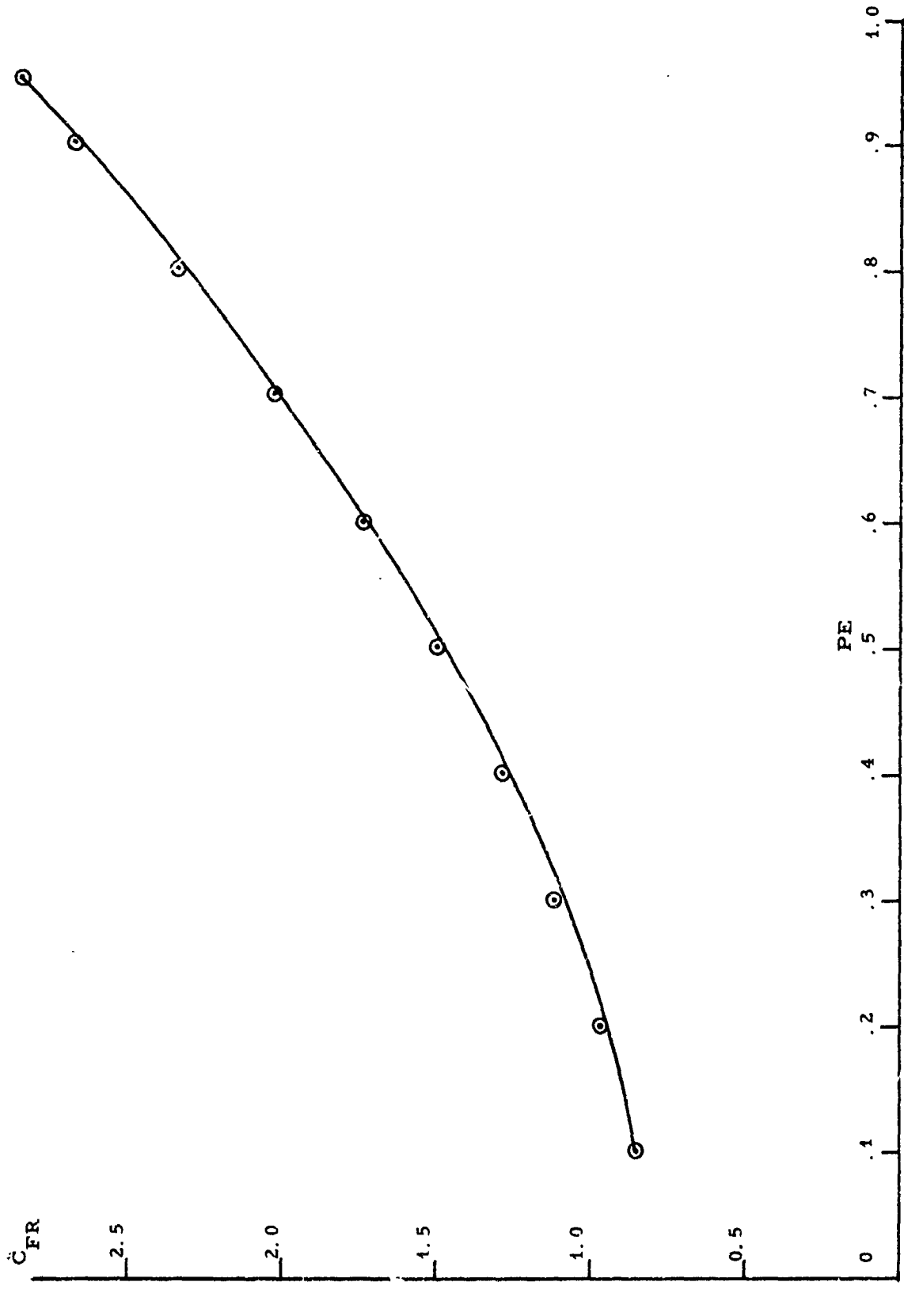


FIGURE 2.3-2 PLOT OF COMPRESSION FOR PE PL CODING AS A FUNCTION OF PE UNDER THE ASSUMPTION THAT PL = PE(1-PE) AND WITH HUFFMAN CODE SUCH THAT PE = 1, PL = 01, AND REMAINING = FOUR

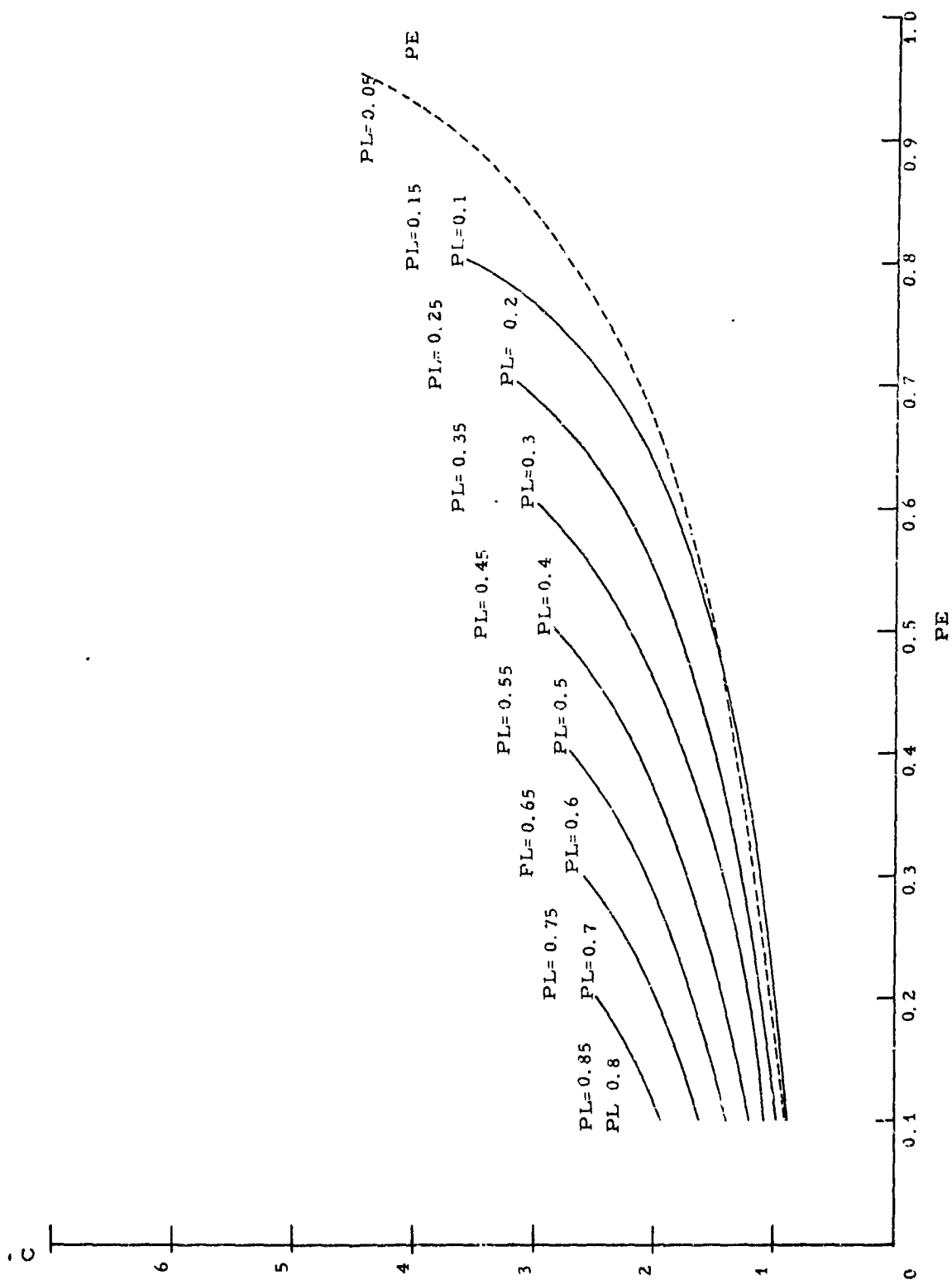


FIGURE 2.3-3 PLOT OF COMPRESSION FOR PEPL CODING AS A FUNCTION OF PE FOR VARYING PL , ASSUMING $PE = 1$, $PL = 0.1$, AND REMAINING LEVELS HAVE FLAT DISTRIBUTION AND HAVE 8 BITS, LOWER BOUND

TABLE 2.3-1 PE AS A FUNCTION OF PICTURE MATERIAL

Picture	PE
MIT girl	0.509
Moon	0.479
Inside view, Mercury capsule No. 1	0.416
Inside view, Mercury capsule No. 2	0.526
Interior view, Apollo No. 1	0.517
Interior view, Apollo No. 2	0.544
OAC satellite	0.605
RETMA test chart	0.466
Starfield	0.609

Figure 2.3-4 was constructed assuming that $PE = 0.519$ and PL allowed to vary to a maximum of 0.431 . The restriction of 0.431 was imposed such that the minimum number of $(1-PE-PL)$ elements are five percent of the maximum number of elements. For the flat distribution and a PE of 0.5 , PL must equal 0.228 to obtain a $2:1$ compression, while PL must increase to 0.395 to obtain a $3:1$ compression.

Assuming that the $(1-PE-PL)$ statistics are Gaussian, the requirements on PL will lessen. For the Gaussian distribution the entropy can be calculated. With the restriction of $PE + PL \leq 0.95$, the compression is given by $\bar{C}_{G6} = 6 / (5.909 - 4.909 PE - 3.909 PL)$. Assuming the $PE = 0.519$, to achieve a $2:1$ compression, PL must be at least 0.118 , while to achieve a $3:1$ compression PL must be at least 0.374 . The value of PL for a $2:1$ compression for a Gaussian distribution is less than the value required for a flat distribution. This indicates that compressions larger than those that can be predicted from the assumed statistics could exist; however, the flat distribution does establish a lower bound on the compression that can be achieved with $PE PL$ coding.

A block diagram of the system given in Figure 2.3-5. Since the system is an extension of PEC in the vertical direction, the system operation will not be discussed.

$PE PL$ coding was designed to fill the void left by PEC for lower values of PE . At high quantization the value of PE decreases to approximately 0.5 , which is not extremely well suited for PEC ; however, for this value of PE , the $PE PL$ system has the possibility of increasing the compression by a factor of two over PEC . Experimental results will verify or disprove this later; however, for $PE = 0.519$ and PEC , the expected compression is $1.8:1$, but assuming $PE + PL = 0.95$, the $PE PL$ compression would be $3.6:1$, or there is a potential increase in compression of $2:1$ by using previous line correlation.

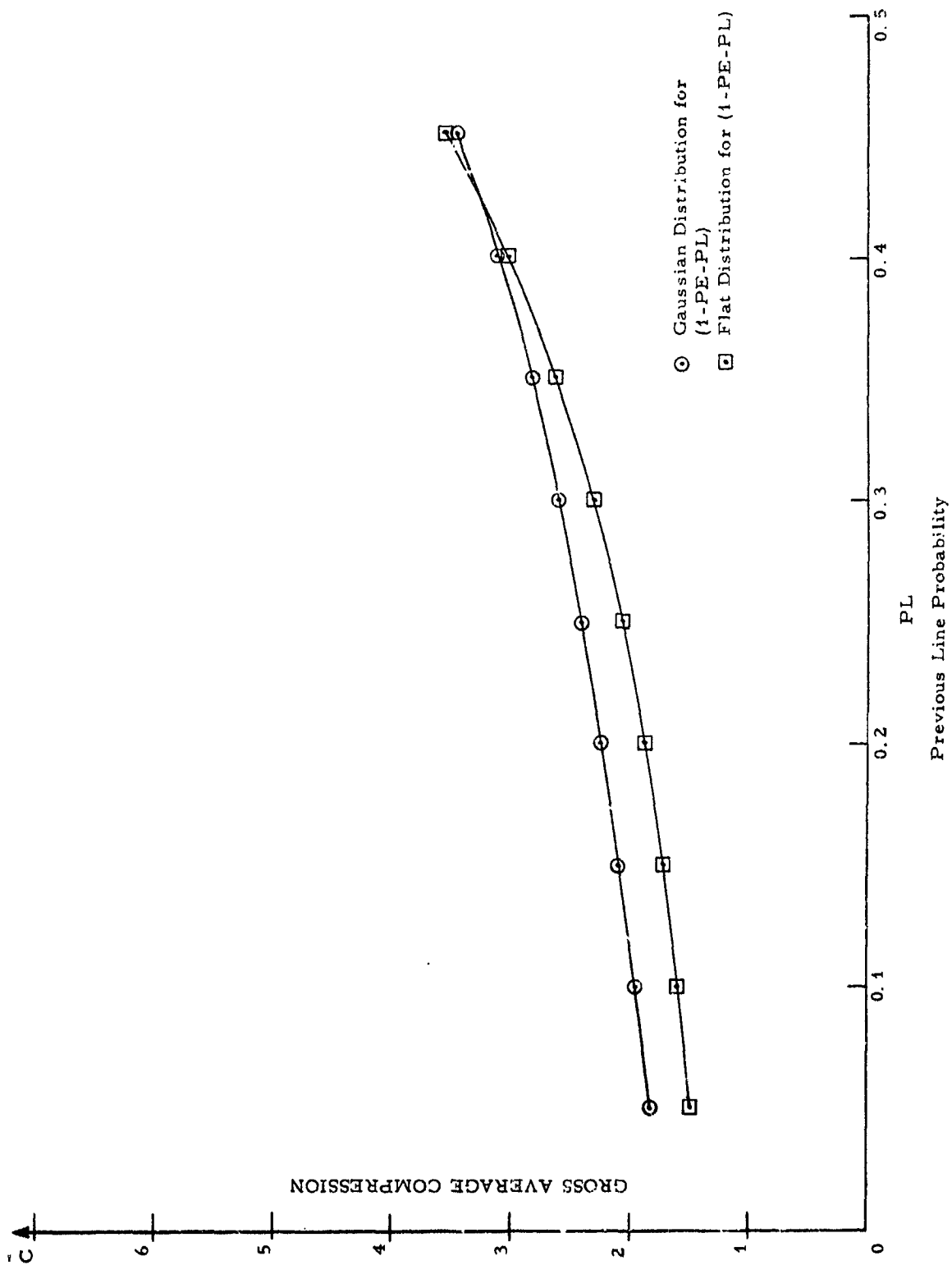


FIGURE 2.3-4 A PLOT OF GROSS AVERAGE COMPRESSION AS A FUNCTION OF PL PROBABILITY FOR $PE = 0.519$ FOR A FLAT DISTRIBUTION AND GAUSSIAN DISTRIBUTION OF (1-PE-PL) ELEMENTS

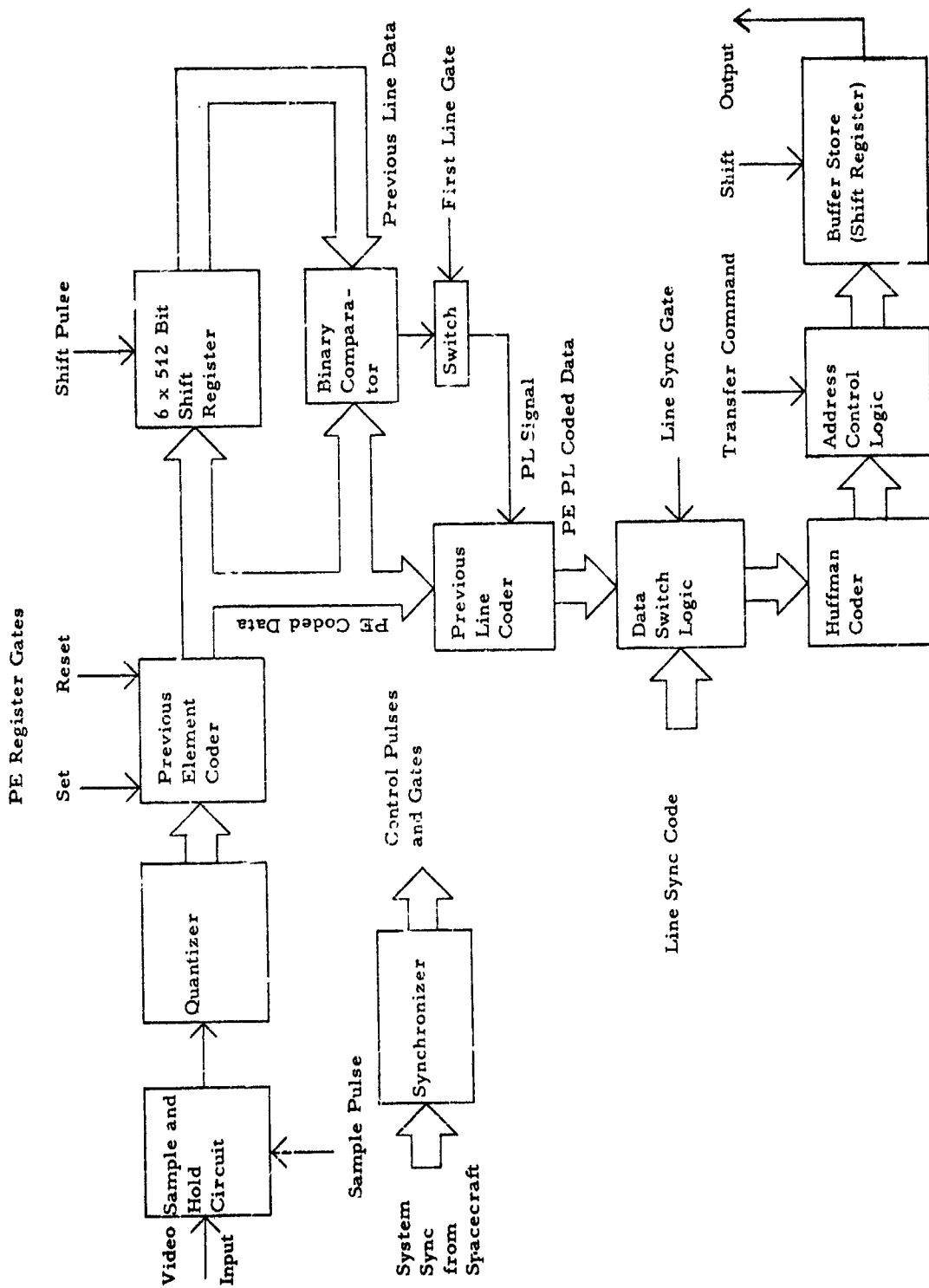


FIGURE 2.3-5 PE PL MODULATOR

For three-bit quantization two pictures were processed on EDITS to determine PL. These pictures were the Moon Picture and the MIT Girl Picture. As an aid to statistically determining PL, each level was displayed as a black and white picture if successive elements were not PE. It is then possible to manually count the number of PL's present in each level. Once these are determined, it is possible to obtain the (1-PE-PL) statistics and PE statistics and Huffman encode these events. Figure 2.3-6 shows the displayed results for levels 0 through 7 for the Moon Picture and Figure 2.3-7 for the Girl Picture. The statistical results are tabulated in Tables 2.3-2 and 2.3-3.

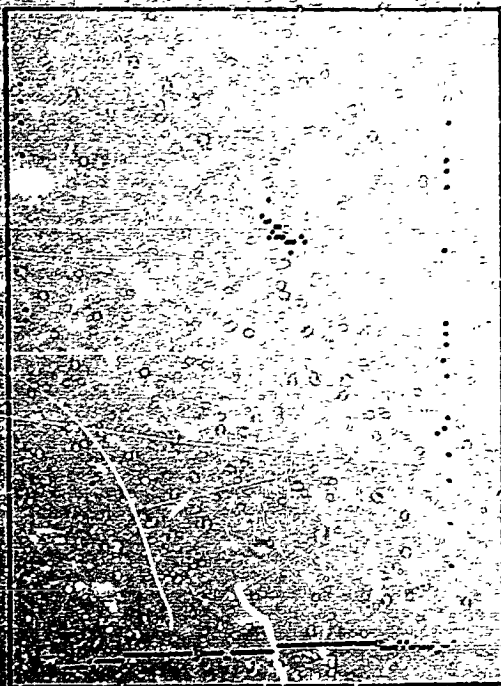
TABLE 2.3-2 PE AND PE PL STATISTICS FOR GIRL PICTURE

Event		Probability	Event		Probability
	PE	0.7820		PE	0.7820
(1-PE)	0	0.0211	(1-PE-PL)	0	0.0988
(1-PE)	1	0.0351	(1-PE-PL)	1	0.0079
(1-PE)	2	0.0333	(1-PE-PL)	1	0.0203
(1-PE)	3	0.0318	(1-PE-PL)	2	0.0165
(1-PE)	4	0.0361	(1-PE-PL)	2	0.0191
(1-PE)	5	0.0342	(1-PE-PL)	3	0.0199
(1-PE)	6	0.0165	(1-PE-PL)	4	0.0222
(1-PE)	7	0.0043	(1-PE-PL)	5	0.0106
			(1-PE-PL)	6	0.0030
			(1-PE-PL)	7	

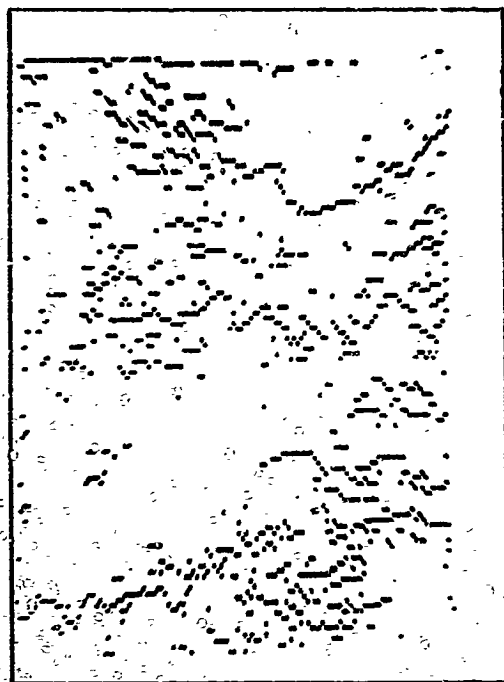
TABLE 2.3-3 PE AND PE PL STATISTICS FOR MOON PICTURE

Event		Probability	Event		Probability
	PE	0.7320		PE	0.7320
(1-PE)	0	0.0100		PL	0.1290
(1-PE)	1	0.0746	(1-PE-PL)	0	0.0037
(1-PE)	2	0.0960	(1-PE-PL)	1	0.0377
(1-PE)	3	0.0459	(1-PE-PL)	1	0.0467
(1-PE)	4	0.0222	(1-PE-PL)	2	0.0266
(1-PE)	5	0.0116	(1-PE-PL)	3	0.0130
(1-PE)	6	0.0064	(1-PE-PL)	4	0.0077
(1-PE)	7	0.0031	(1-PE-PL)	5	0.0037
			(1-PE-PL)	6	0.0014
			(1-PE-PL)	7	

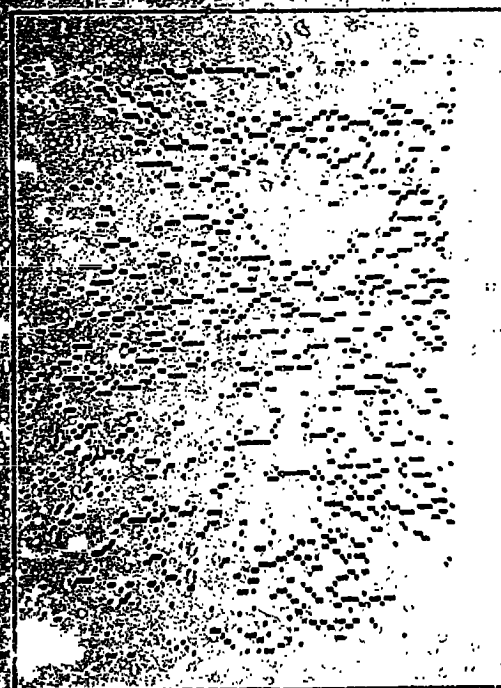
If these statistics are now encoded with the optimum Huffman code, the compressions for a three-bit picture using PEC and PE PL coding can be tabulated. These results are given in Table 2.3-4.



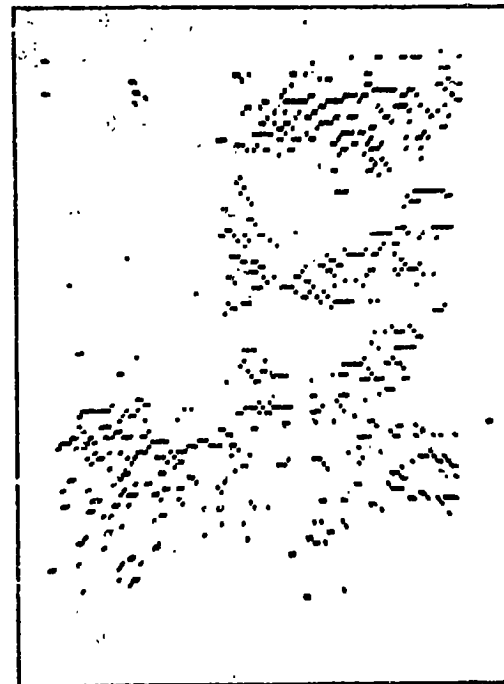
LEVEL 0 (1-PE)



LEVEL 1 (1-PE)

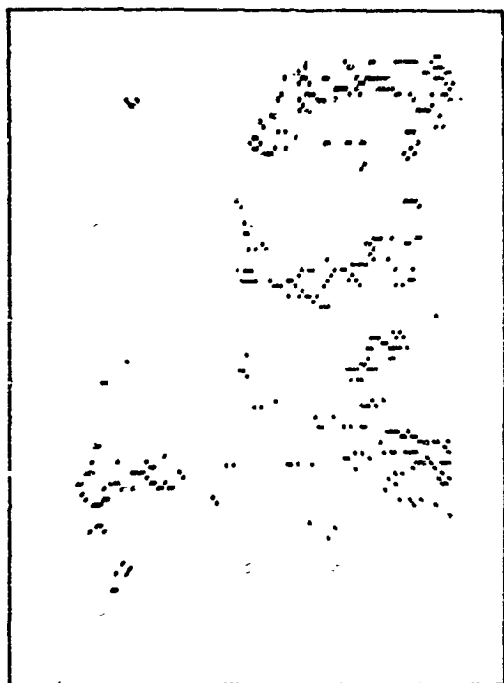


LEVEL 2 (1-PE)

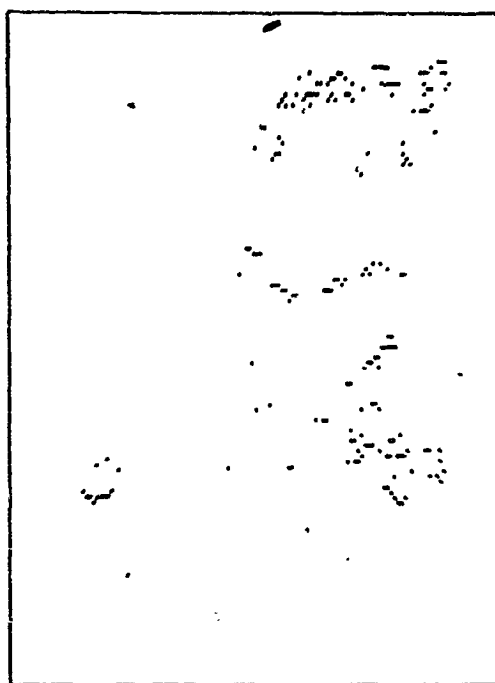


LEVEL 3 (1-PE)

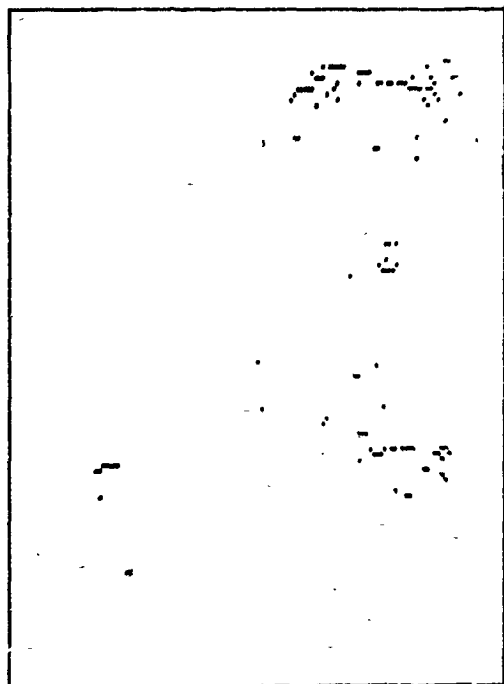
FIGURE 2-3-6. MOON PICTURE. PHOTOGRAPHS FROM EDITS OF (1-PE) SIGNALS FOR EACH OF EIGHT LEVELS QUANTIZED TO 8 BITS AT 128-LINE RESOLUTION



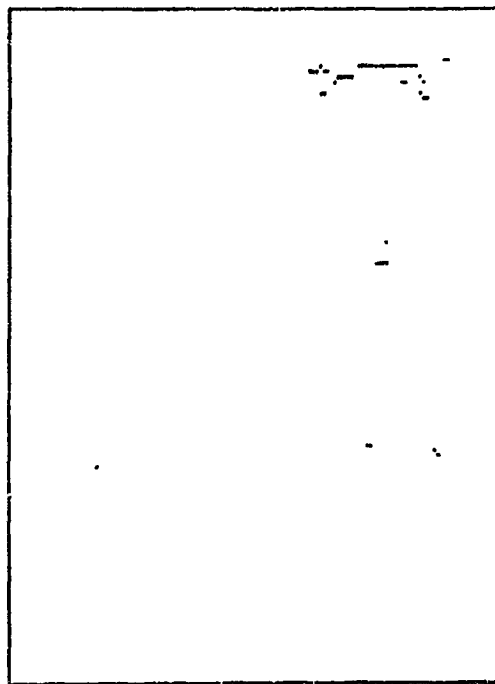
LEVEL 4 (1-PE)



LEVEL 5 (1-PE)

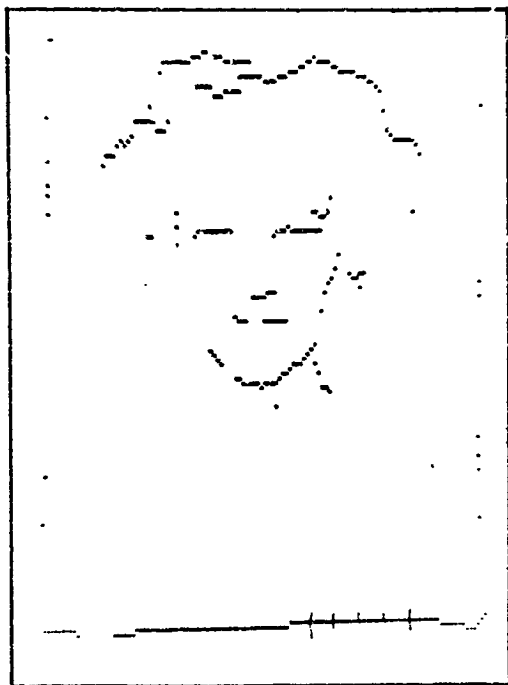


LEVEL 6 (1-PE)

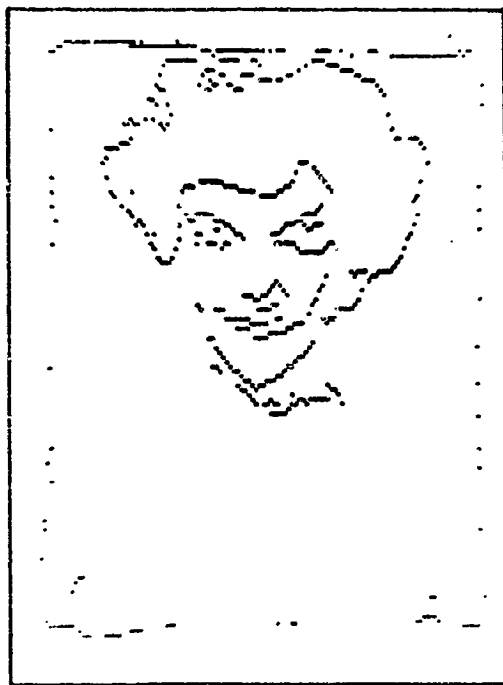


LEVEL 7 (1-PE)

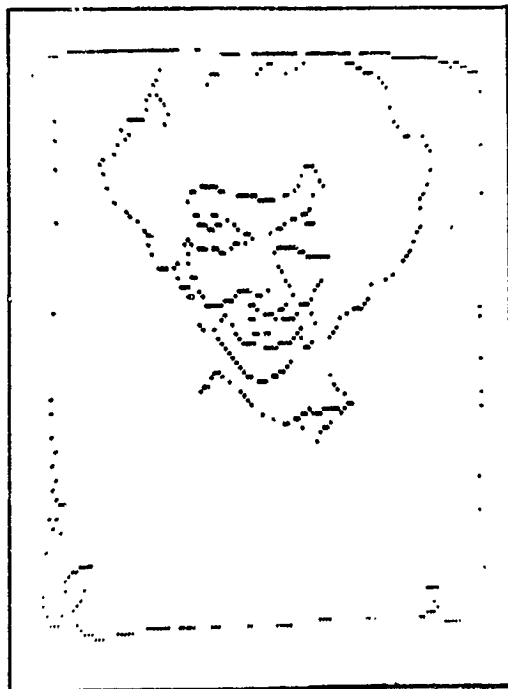
FIGURE 2.3-6 (Continued)



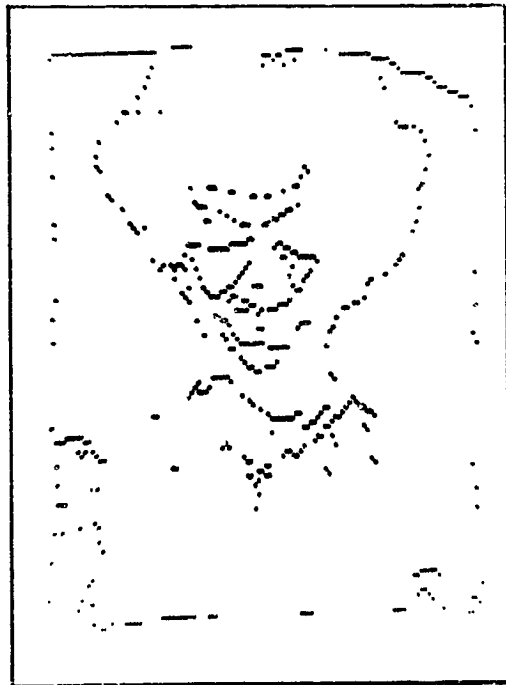
LEVEL 0 (1-PE)



LEVEL 1 (1-PE)

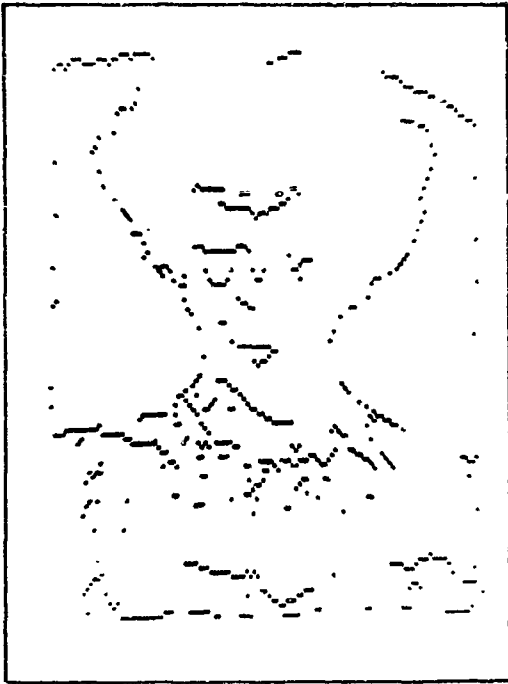


LEVEL 2 (1-PE)



LEVEL 3 (1-PE)

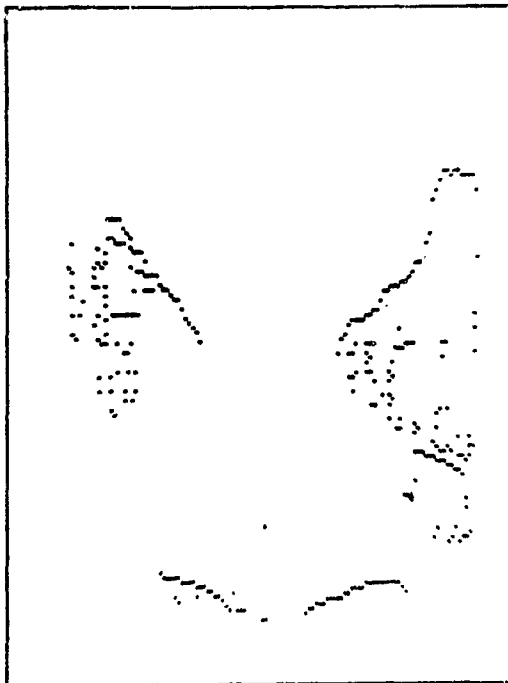
FIGURE 2.3-7. GIRL PICTURE. PHOTOGRAPHS FROM EDITS OF (1-PE) SIGNALS FOR EACH OF EIGHT LEVELS QUANTIZED TO 3 BITS AT 128-LINE RESOLUTION.



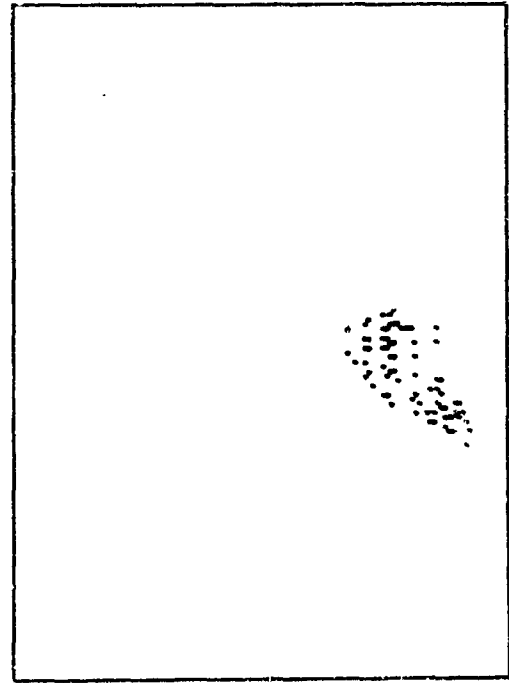
LEVEL 4 (1-PE)



LEVEL 5 (1-PE)



LEVEL 6 (1-PE)



LEVEL 7 (1-PE)

FIGURE 2.3-7 (Continued)

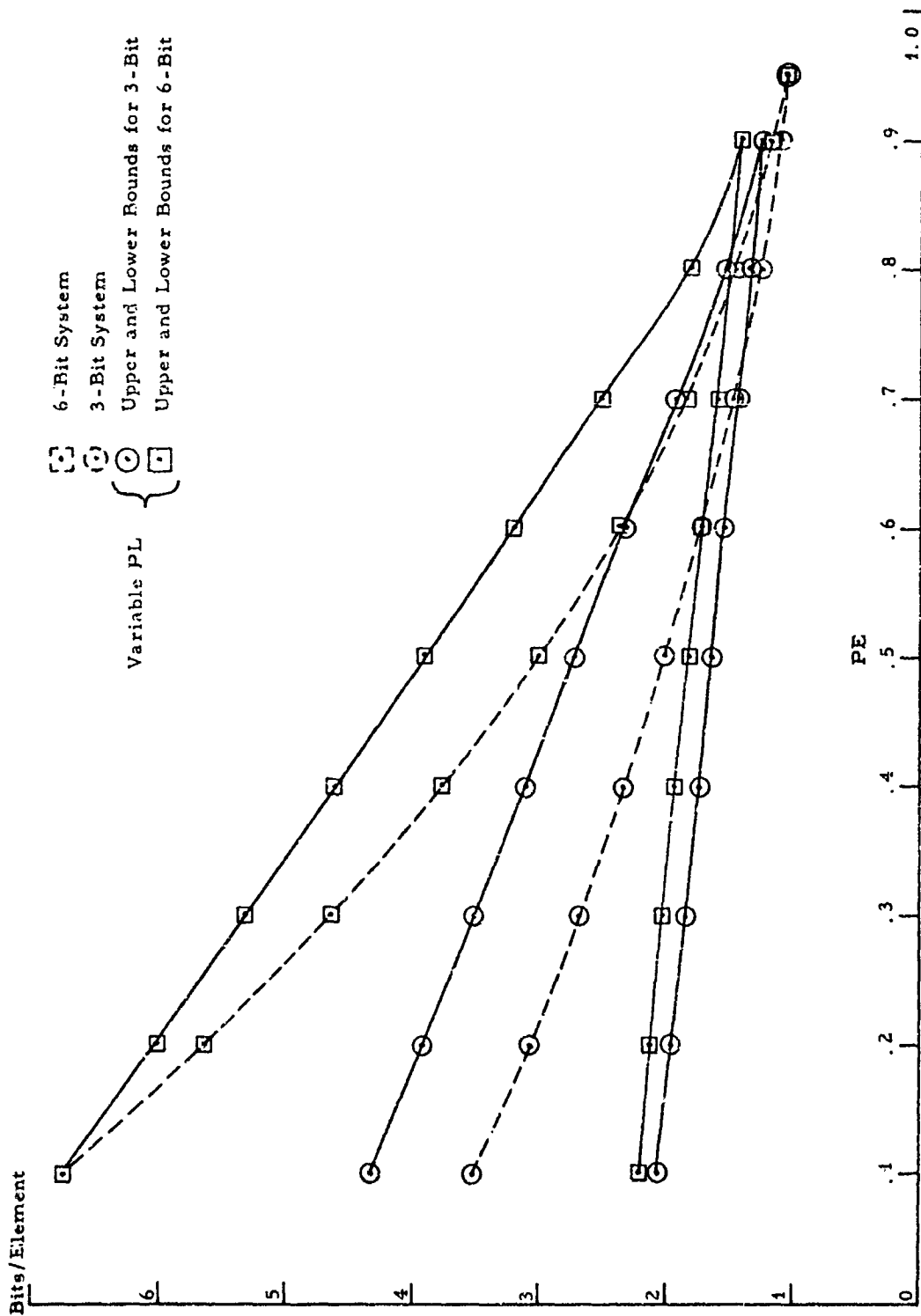


FIGURE 2.3-8 PLOT OF AVERAGE BITS PER ELEMENT AS A FUNCTION OF PE FOR PE PL CODING, ASSUMING THAT $PL = PE(1-PE)$ FOR 6-BIT AND 3-BIT COMPRESSION SYSTEMS

TABLE 2.3-4 COMPRESSION OF TWO THREE-BIT PICTURES
FOR PEC AND PE PL CODING.

Picture	Coding	
	PEC	PE PL
Moon	1.81	1.36
Girl	1.86	1.92

Although these results for three-bit coding do not appear too promising, the PE PL system could possibly be extremely useful. For example, Figure 2.3-8 plots the upper and lower bounds on bits per element for PE PL coding of a three and six-bit source.

It is apparent that PE coding is a lower bound to PE PL and implementation is straightforward extension of PEC. The PE PL system uses both element-to-element and line-to-line correlation and statistics can be obtained from EDITS. PE PL designed to operate in regions where $PE = 0.5$ for 6 bit quantization where PE PL has the potential to increase the compression by a factor of two over PEC.

2.4 FRAME-TO-FRAME CODING

The correlation between successive frames is large because of the repetition in time necessary to achieve continuity of vision. The bandwidth required for real-time transmission is directly proportional to the frame rate, therefore, if the correlation between successive frames is by necessity high; then a considerable savings in bandwidth could be achieved by transmitting only the difference between successive frames. A signal is transmitted when there is a change between frames. The only information transmitted is the new information represented by changes in the picture; that is, the amount by which each frame differs from the previous one. If only this difference is transmitted, the bandwidth should be reduced considerably.

The previously discussed coding systems utilized only the short-term redundancy such as element-to-element or line-to-line, whereas the frame-to-frame coding uses the long-term redundancy. Therefore, the long-term redundancy elimination techniques by necessity require extremely large buffer stores since an entire frame must be stored. However, the size of the buffer store could possibly be reduced by reducing or eliminating the short-term redundancy. For example, consider PE coding where a 2:1 bandwidth reduction is achieved. The size of the buffer store could be reduced by 50% if PE coding is used prior to frame-to-frame coding. It appears that if frame-to-frame coding is to be practical, it must be used in cascade with other coding techniques.

There are several possible methods of frame-to-frame coding. The most logical is an extension of the PEPL coding techniques, that is, code the source with the PEPL technique and then previous-frame (PF) encode the (1-PE-PL) elements. However, this does not appear to be too practical or profitable. For example, assume that a source is encoded to three bits. The value of PE is approximately 0.8 and PL must be less than 0.2. Therefore, this implies that at most only 20% of the total elements have the possibility of being PF coded. In addition, by PF coding another code word is added to the code set and one more digit is added to the (1-PE-PL-PF) words, thus, the coded entropy could be increased if PF is not high enough. The coded entropy is given by

$$H'_C = N + 3 - (N + 2) PE - (N + 1) PL - N PF,$$

which must be compared to the coded entropy of the PEPL system, which is

$$H_C = N + 2 - (N + 1)PE - N PL$$

For $N = 3$, $H'_C = 6 - 5PE - 4PL - 3PF$, and $H_C = 5 - 4PE - 3PL$, and if it is assumed that $PE = 0.8$, then $H'_C = 2 - 4PL - 3PF$ and $H_C = 1.8 - 3PL$. Now $PL + PF < 0.2$ for H'_C while $PL < 0.2$ for H_C . To achieve the minimum H'_C and H_C for maximum compression, let $PL = 0.2$; therefore, $H_C = 1.2 = H'_C$ or the PF system does not improve the maximum efficiency. In fact, if the

additional size, weight, and power are considered, the PF coding would actually decrease the net compression. Therefore, consider an alternate approach to PF coding.

For this system assume that frames k and $k + 1$ are PE PL coded and then the two frames are compared and PF coded. This system achieves the compression of the PE PL on the two frames, but on the $k + 1$ frame additional compression is achieved. Therefore, it is necessary to determine the correlation between frames k and $k + 1$. This correlation in general is high, but once again the maximum achievable compression is $\log_2 n$ since the PF must have one bit to describe it. However, it is necessary to perform a trade-off in assigning the Huffman code words to the PE, PL, and PF probabilities. Therefore, it becomes quite obvious that frame-to-frame coding using PF correlation in the same manner as PE coding does not provide a sufficient increase in the amount of compression.

It would be advantageous to determine an entirely new approach to frame-to-frame coding. The most obvious choice is to transmit only the difference between successive frames, that is, both the coordinates and intensity of each difference signal is transmitted. However, this imposes rather severe restrictions on the amount of redundancy between successive frames. For example, consider a picture composed of m^2 elements where each element is encoded to n levels. The total number of bits required to describe this picture is Nm^2 where $N = \log_2 n$. Further, assume that the PEC system achieves a 2:1 compression of the number of bits for PEC is $\frac{Nm^2}{2}$. The number of bits to specify the location of each element is $\log_2 m^2 = 2M$ and intensity is N or each element in the PF system requires $2M + N$ bits. Now, determine a value of p , the correlation between successive frames such that

$$(1 - p)(2M + N)m^2 \leq \frac{Nm^2}{2}.$$

That is, $(1-p)(2M + N) \leq N/2$. The minimum p is when the equality sign is valid; that is $p = \frac{N}{2(2M + N)}$. For $N = 3$ and $M = 8$, $p = 1 - 3/2(19) = 0.921$, which

implies that 92.1% of the elements in the successive frames must be of the same magnitude to achieve a 2:1 compression with previous-frame coding. However, this is the same compression as with PEC. One desires that the PF system have a compression greater than the 2:1 of PEC; therefore it is necessary to determine what percentage of the elements in frame $k + 1$ must be the same as the elements in frame k . To achieve a 4:1 compression $p = 0.961$ or 96.1 percent of the information between successive frames must be redundant. This may or may not appear to be a realistic estimate on frame-to-frame redundancy. A system of this type is extremely sensitive to errors and it is conceivable that if for example every T frames a new picture is transmitted, that an error in frame $T-k$ could totally destroy the remaining frames. Therefore, an alternate possibility exists where, instead of every T frames sending the complete frame, x percent of each frame is transmitted in

real-time, thus at least x percent of each picture would not allow an error to propagate through the remaining frames.

In general, it appears that frame-to-frame coding is somewhat unrealistic in view of the large buffer store and error effects. Statistics are not now available to determine whether or not further investigation of frame-to-frame coding should be undertaken.

2.5 OTHER

2.5.1 Area Coding

The correlation between n elements in an area or block of elements is much higher than the correlation between n elements along a line or between n lines. Therefore, if this correlation is properly used, the compression for encoding areas should be higher than the compression for line-to-line or element-to-element coding. The results presented by Schaphorst^{1, 2} on area encoding appear to be encouraging; however, closer examination of the experimental results indicate that some of the results could be in error. The statistics are based on a picture containing 3.2×10^6 elements; however, only 10^4 elements are actually sampled and measured.

A picture was reduced to approximately 100 randomly located samples and each sample was comprised of 100 elements. This means that only one out of every 320 elements was actually measured. This in general will not give a very true indication of the actual statistics. Since the compression is based on these statistics, it is debatable whether or not the compression ratios obtained are valid. However, ignoring this, the system operation will be described.

The area-coding technique had as a design goal a 10:1 compression for pictures of "average" redundancy where "average" redundancy was not defined. The coding technique utilizes both the horizontal and vertical redundancy in a single-frame and achieves compression by combining variable-velocity scanning with pattern-recognition techniques. The overall system employs three modes of operation where the mode operation is determined by the intensity distribution in a 5×5 area of elements. Block diagrams of a simplified area-character encoder and decoder are given in Figures 2.5.1-1 and 2.5.1-2 respectively. In Mode 1 the 25-element area is scanned and if all 25 elements have the same intensity level, the area is rescanned at a rate 25 times the normal scanning rate and transmitted in analog form in one Nyquist interval. In areas where the conditions necessary for the operation of Mode 1 are satisfied, a 25:1 bandwidth compression is achieved. If an intensity change occurs during Mode 1, the system goes into Mode 2 operation. In Mode 2 operation the area is rescanned at the normal scanning rate and 27 different area characters must be recognized. If any one of the area characters is recognized, the information is transmitted in six Nyquist intervals as amplitude-modulated pulses. This results in a 4:1

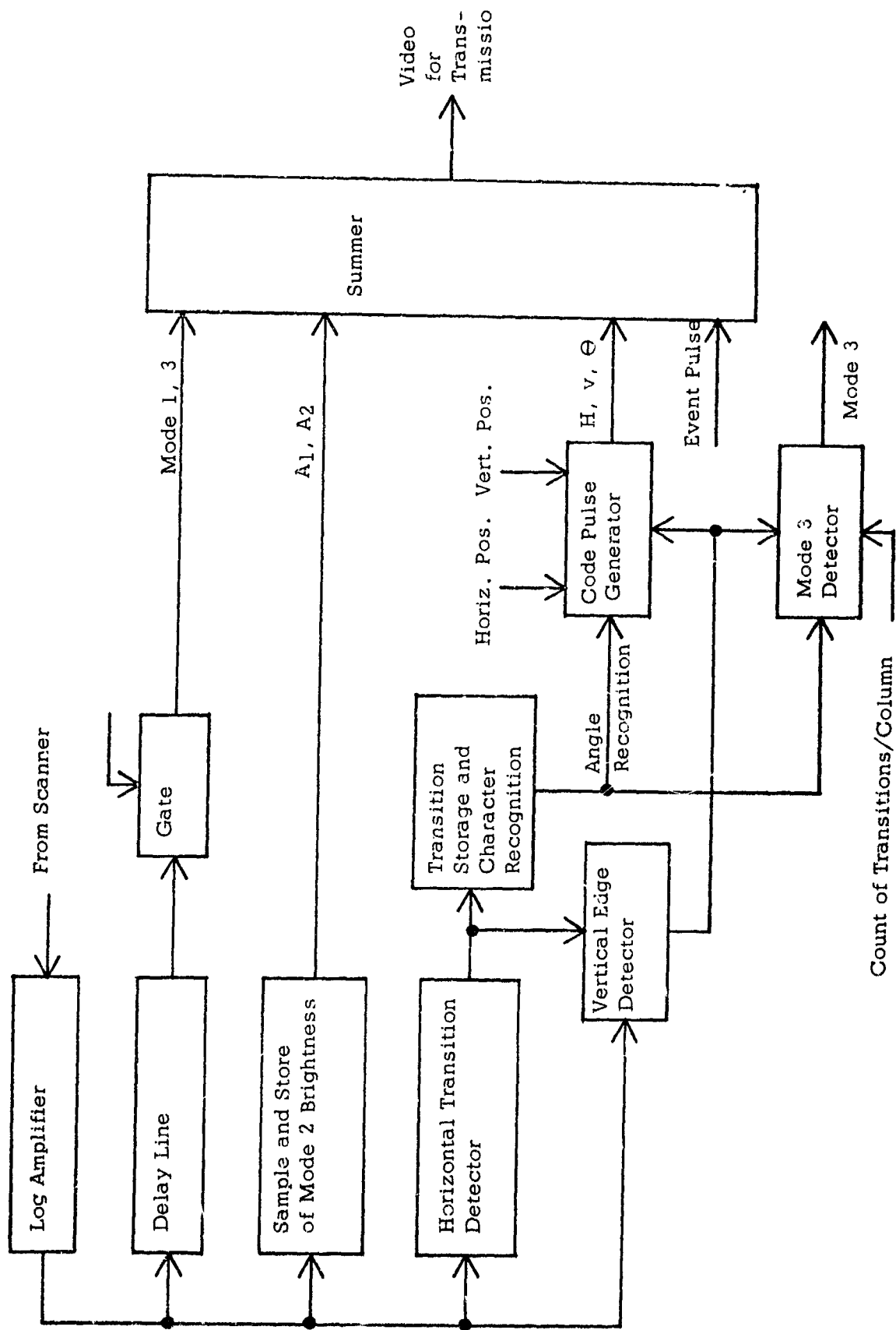


FIGURE 2.5.1-1 AREA - CHARACTER ENCODER

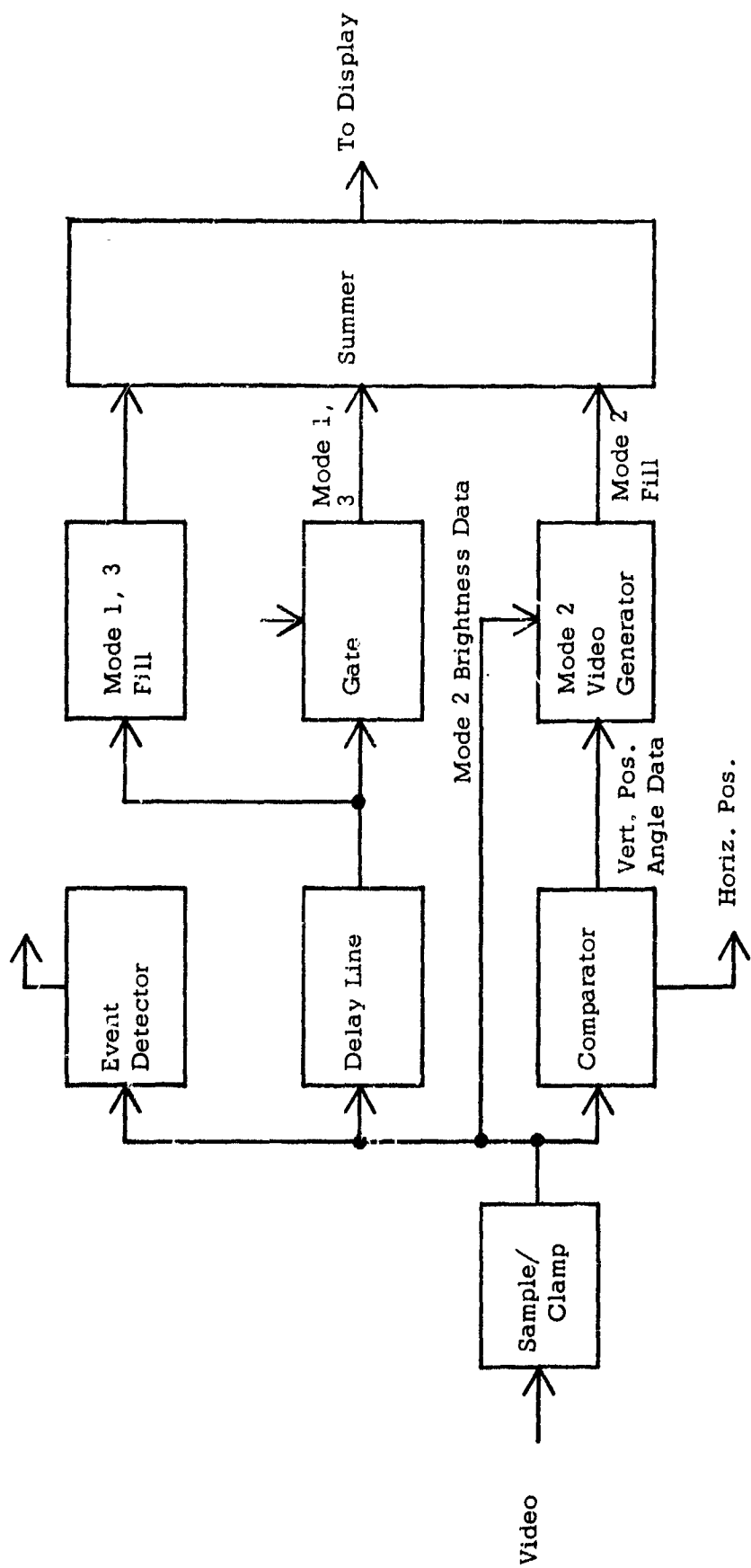


FIGURE 2. 5. 1-2 AREA - CHARACTER DECODER

bandwidth compression for Mode 2 operation. However, if none of the 27 area characters are present in the scanned area, the system enters Mode 3 operation. In Mode 3 each of the 25 elements is transmitted in one Nyquist interval, but since the mode operation must be specified, there is an increase in bandwidth by a factor of 1.12:1 for Mode 3 operation.

The compression was determined by manually simulating the modes of operation for several pictures. The simulation consisted of measuring the intensity of each element in the randomly selected areas. This made it possible to tabulate the percentage of time each mode will be in operation. From these percentages it is possible to estimate the compression for each particular picture. However, stressing once again that only one out of every 320 elements was measured, it can be inferred that this approach may yield results that are only an approximation to the actual compression.

The manual simulation was compared to experimental results by implementing Modes 1 and 3. Unfortunately, experimental results were not obtained for Mode 2 operation. The experimental results and manual simulation were in close agreement; however, the reconstructed picture exhibited a certain amount of distortion due to the interpretation of slightly curved transitions as straight-line transitions and thinning or broadening of transitions. Experimental results indicate that at least one percent of the areas in Mode 2 will be distorted; however, since the video was quantized to only 15 levels, this percentage of distortion may vary with quantization.

Although the area-character coding technique is a combination of analog and digital transmission, it is possible to analyze the coding technique by assuming a true digital operation. Fortunately, the percent of operation of the three modes for various pictures was given. Assume the area-character coding parameters of a 5 x 5 elemental area, 16 brightness levels, and 27 area-characters for Mode 2. For a typical picture the percent occurrence of the three modes are: Mode 1--40%, Mode 2--40%, and Mode 3--20%. To describe Mode 1, each area-character requires four bits for intensity and two bits to indicate the operating mode, or 6 bits per area for Mode 1. In Mode 2 operation, the area-characters require five bits, four bits for intensity, and two bits for the mode of operation, or to describe each area in Mode 2 operation requires eleven bits per area. In Mode 3 operation, each element in the area requires four bits for intensity and two bits for mode operation or each element requires six bits per element; therefore, each area for Mode 3 operation requires 150 bits. Using these quantities, the number of bits per mode can be tabulated and are given in Table 2.5.1-1.

TABLE 2.5.1-1 BITS PER MODE FOR DIGITAL AREA CODING

Mode	Percentage of Picture	Bits Per Mode
1	40	9.6×10^4
2	40	17.6×10^4
3	20	120.0×10^4

For this particular picture the total number of bits per frame is 1.472×10^6 . The uncoded picture requires 4×10^6 bits per frame; therefore, the compression is 2.72:1. The simulated digital-analog area-character coding reported a 3.33:1 compression for the picture.

Although the experimental results for area coding suggest that compressions from 1.54 to 12.5 exist, it remains to be seen whether or not these compressions actually exist. Area coding should provide an improvement in compression over line-to-line or element-to-element, but implementation of the area-coding system is quite a problem and synchronization quite difficult.

The general conclusion is that area coding should be investigated further to determine whether some of the practical problems of implementation, such as buffer store, can be eliminated as well as the effects of errors on the system which may prove to be the limiting factor of the system. The implemented system is well suited for single-frame operation and the size and complexity of the system as it now exists probably excludes the possibility of obtaining a space-borne model.

2.5.2 "Elastic" Systems

Since the statistics of the source change from picture to picture, frame to frame, line to line, and element to element, the optimum compression technique would be an elastic coding system where the coding technique would match the source statistics. With the exception of the delta modulation and the Roberts coding techniques the amount of compression achievable is totally dependent upon the statistics of the source. If the source statistics were known prior to transmission, an optimum coding technique could be applied and maximum compression could be obtained; however, for space applications the source statistics are not known in advance. Therefore, it is essential that the coding technique be elastic if maximum efficiency is desired.

Elastic coding is a computer technique that operates on the source and determines the optimum or near optimum code to be used during a specified time interval. It is necessary that the coding technique has the capability of recognizing certain

statistical characteristics of the source and of determining the optimum or near-optimum code to be used. The statistical characteristics of the source need not be determined precisely, but an estimate is sufficient since the coding efficiency remains relatively insensitive to small errors. However, should violent changes in the source statistics take place, then the coding technique must adapt itself to the change.

The amount of compression that can be achieved by elastic coding is a function of the fidelity criteria of the user. The fidelity criteria will determine the amount of redundancy that can be removed. Once the amount of redundancy that can be removed has been determined, it is necessary to ensure that during periods of little or no redundancy, the coding technique is capable of reproducing the data to the desired fidelity criteria. This is to say that an appropriate trade off must be made between the amount of redundancy to be removed and the buffer store size. The larger the redundancy to be removed, the larger the buffer store required along with its many associated problems. However, it is possible to reduce the fidelity of the system during periods of little redundancy, thus reducing the buffer store requirements.

Although the elastic system would be optimum for a finite time interval, the coding must be such that during periods of little or no redundancy, it must be capable of reproducing the data to some specified fidelity criterion. Not only must it have this capability, it is also necessary during these intervals that the coding technique achieves compression. If the statistics of the source should change drastically, it is possible that the coding will lead to a bit rate that exceeds the maximum entropy of the source. This would increase the bandwidth thus defeating the elastic coding purpose. Not only would the bandwidth be increased but it is possible that during these intervals the buffer store would be taxed to a point exceeding the maximum overflow, thus degrading the information to a point where the original information of the source could not be recovered. Therefore, it is necessary to trade theoretical entropies for practical considerations to achieve compression. Thus, elastic coding is a compression technique that is a compromise between the most efficient coding (when the source statistics are completely known) and maximum entropy coding (straight transmission when the source statistics are unknown).

Consider a set of source statistics to illustrate the value of the elastic system. For the Cooper No. 1 standard picture at 256 resolution the (1-PE) and PE probabilities are given in Table 2.5.2-1.

Assume that for the probabilities of Table 2.5.2-1 that three coding techniques are to be used. The first coding technique will encode the data with the optimum Huffman code. The second coding technique will assume that the (1-PE) levels have a flat distribution and encode accordingly. The third coding technique will assume that the (1-PE) levels have a Gaussian distribution and the Huffman code will be assigned. It is possible to calculate the coded entropy for each coding technique from the probabilities. Table 2.5.2-2 tabulates the number of bits $[L(i)]$ for each level for each of the coding techniques.

TABLE 2.5.2-1

COOPER NO. 1 PE AND
(1-PE) PROBABILITIES

LEVEL	PROBABILITY
0	0.0303
1	0.0448
2	0.0306
3	0.0254
4	0.0225
5	0.0179
6	0.0171
7	0.0135
8	0.0123

LEVEL	PROBABILITY
9	0.0099
10	0.0089
11	0.0070
12	0.0053
13	0.0027
14	0.0012
15	0.0005
PE	0.7501

TABLE 2.5.2-2

BITS PER LEVEL FOR
CODING TECHNIQUES

Level	L(i)		
	Huffman	Flat	Gaussian
0	4	5	13
1	4	5	11
2	4	5	9
3	4	5	8
4	4	5	5
5	5	5	5
6	5	5	4
7	5	5	3
8	5	5	3
9	6	5	4
10	6	5	5
11	6	5	6
12	7	5	7
13	8	5	10
14	9	5	12
15	9	5	13
PE	1	1	1

From Table 2.5.2-2 it is possible to calculate the entropy and compression for each of the coding techniques. The compression is tabulated in Table 2.5.2-3.

TABLE 2.5.2-3
COMPRESSION FOR CODING VARIOUS DISTRIBUTIONS

Distribution	Coding		
	Huffman	Flat	Gaussian
Actual	2.1083	2.0004	1.4957
Flat	1.8420	2.0004	1.5424
Gaussian	1.9382	2.0004	2.3036

In row one of Table 2.5.2-3 the statistics of Table 2.5.2-1 are encoded with the three coding techniques. This illustrates the possible potential of an elastic system. For example, if the elastic system were using Gaussian coding with a Gaussian distribution and the statistics changed to the actual statistics of Cooper No. 1, the compression would drop to 1.4957 from 2.3036 or the compression would decrease 35%, whereas if the coding could change to the Huffman code there would be little decrease in compression due to changing of the source statistics.

There are several possible methods of implementing an elastic system. The simplest method would be to have a computer with a stored catalog of probability distributions, compare the source statistics with the stored distributions, and encode the source with an optimum or near optimum code; that is, the source statistics would be continually monitored and when a specified threshold is exceeded, the most efficient coding technique is employed. An alternate approach to eliminate the large buffer store would be to transmit a sample of the statistics of the source to the ground station, have the statistics analyzed, and transmit back to the vehicle a command to actuate the most efficient coding technique. It may also be possible to determine the statistics on the spacecraft and have the operator on the vehicle determine the optimum code for transmission.

The possibilities of an elastic system are unlimited, but require a sizable trade-off between fidelity, buffer store requirements, and compression. However, with the advent of microminiature circuits and more efficient buffer stores, elastic encoding may prove to be quite feasible.

2.5.3 References for 2.5

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2.6 OPTIMUM ENCODING

2.6.1 Efficient Codes

In order to achieve maximum compression the compression technique must use efficient coding of the digital data. Among the coding techniques with the prefix property, that is, in which the set of code words is decodable, the Huffman code is an optimum code while the Shannon-Fano code is optimum or near optimum in many cases. Therefore, consider how the coding effects the efficiency of the overall system.

Assume that an information source consisting of an ensemble of μ of M messages $\mu_1, \mu_2, \dots, \mu_k, \dots, \mu_M$ with corresponding probabilities $P(\mu_k)$ is given where

$$\sum_{k=1}^M P(\mu_k) = 1.$$

It is necessary to find a coding technique that will assign to each message a sequence of symbols belonging to a prescribed alphabet where the average number of symbols for all the messages is minimum.

Mathematically a lower bound to this average number of symbols can be determined. The entropy⁽¹⁾ of the message ensemble is the lower bound of the average (over the message ensemble) number of binary digits sufficient to represent the messages of the ensemble. The entropy is given by

$$H(\mu) = - \sum_{k=1}^M P(\mu_k) \log_2 P(\mu_k).$$

With the exception of the case when $P(\mu_k) = 1/M$, that is, the probability of each message is the same, no coding technique can achieve the lower bound. For the "flat" distribution of message probabilities the fixed-length coding technique achieves the lower bound. However, when this is not true, and in general this is the case, it is possible to determine a code such that the average number of binary digits per message will not exceed the entropy of the message set by more than one binary digit.

Although it is possible to find the mathematical lower bound, a more practical value can be obtained by finding the encoded entropy of the message ensemble. The encoded entropy is given by

$$H(E) = \sum_{k=1}^M P(\mu_k) L(k),$$

where $L(k)$ is the number of binary digits assigned to message μ_k of the ensemble. The basic coding theorem states that it is possible to encode an ensemble of messages by means of sequences of symbols belonging to the prescribed alphabet in such a way that average number of symbols (digits) per message [encoded entropy $H(E)$] satisfies the inequality $H(\mu) \leq H(E)$. That is, the encoded entropy cannot be smaller than the entropy. The problem now is to find the coding technique that gives the minimum $H(E)$. The Shannon-Fano⁽²⁾ code is an encoding procedure that approaches the minimum $H(E)$, but it is not necessarily the optimum set.

A systematic coding procedure was discovered by D. A. Huffman⁽³⁾ and this procedure always yields an optimum set of code words; that is, no other coding technique exists that yields a smaller encoded entropy. However, this coding technique requires a priori knowledge of the statistics to achieve the minimum encoded entropy. The coding technique will assign the shortest code words to the most probable messages and the longest words to the least probable messages. It is a coding technique that matches the encoding to the source statistics. It is a minimum redundancy encoding technique that among all variable-length binary encoding techniques having the prefix property (that is, it is decodable since no longer word is an extension of a shorter word) none of these other encoding techniques can have a lower encoded entropy.

An example of the coding technique will be given to illustrate how the coding is performed and a comparison with another coding technique will be made later to verify some of the previously mentioned properties. The procedure will be given in general and an example will follow. Therefore, consider an ensemble of M messages $\mu_1, \mu_2, \dots, \mu_M$ occurring with probabilities $P(\mu_1), P(\mu_2), \dots, P(\mu_M)$ and an alphabet of D symbols. The following procedure leads to an optimum set of code words and is taken directly from Fano⁽⁴⁾.

Step 1. The M messages are arranged in order of decreasing probability.

Step 2. Let m_0 be the integer satisfying the requirements that

$$(1) \quad 2 \leq m_0 \leq D,$$

$$(2) \quad \frac{M - m_0}{D - 1} = \text{positive integer.}$$

Group together the m_0 least probable messages and compute the total probability of such a message subset.

Step 3. Construct an auxiliary ensemble of messages from the original ensemble by regarding the subset of m_0 messages formed in Step 2 as a single message with probability equal to the probability of the whole subset. Rearrange the messages of this auxiliary ensemble in order of decreasing probability.

Step 4. Form a subset of the m_0 least-probable messages of the auxiliary ensemble and compute its total probability.

Step 5. Construct a second auxiliary ensemble from the first auxiliary ensemble by regarding the subset of the m_0 messages formed in Step 4 as a single message with probability equal to the whole subset. Rearrange the messages of this second auxiliary ensemble in order of decreasing probability.

Step 6. Form successive auxiliary ensembles by repeating steps 4 and 5 until a single message of unit probability is left in the ensemble.

Step 7. The preceding steps, when formally carried out by drawing lines joining the messages of successive subsets, yield a tree of which specified messages are terminal nodes. Code words can then be constructed by assigning different symbols from the prescribed alphabet to the branches stemming from each intermediate node.

As an example of how the coding technique is applied, assume that the number of messages to be encoded (M) is eight, the number of symbols of the alphabet (D) is two, and the message probabilities $[P(\mu_k)]$ are given in Table 2.6.1-1.

TABLE 2.6.1-1 ASSUMED MESSAGE PROBABILITIES

Message u_k	1	2	3	4	5	6	7	8
Probability $P(\mu_k)$	0.70	0.15	0.06	0.04	0.03	0.01	0.005	0.005

The encoding steps are given in Figure 2.6.1-1.

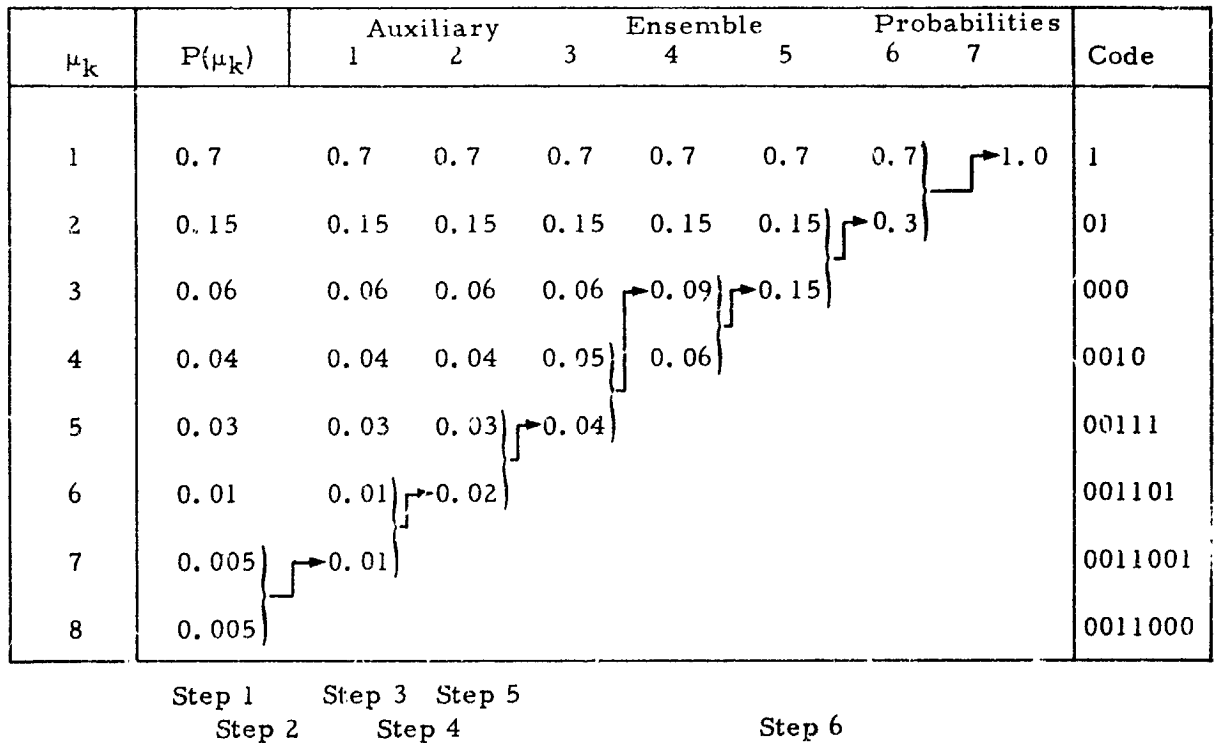


FIGURE 2.6.1-1 HUFFMAN CODING PROCEDURE (STEP-BY-STEP ANALYSIS)

Although Figure 2.6.1-1 is useful to illustrate the step-by-step coding procedure, the Huffman coding tree appropriately drawn is given in Figure 2.6.1-2.

Step 1. The M messages are arranged in order of decreasing probabilities.

Step 2. The ensemble of probabilities is divided into two auxiliary ensembles of near-equal probabilities.

Step 3. Each auxiliary ensemble of probabilities is assigned a symbol from the alphabet.

Step 4. The auxiliary ensembles formed by Steps 2 and 3 are divided into successive auxiliary ensembles until no further divisions are possible.

Step 5. The code words have then been constructed with the coding procedure.

As an example of the coding procedure, consider the same probabilities used for the Huffman code given by Figure 1. The step-by-step coding procedure is given by Figure 2.6.1-3.

μ_k	$P(\mu_k)$	Auxiliary		Ensemble	Probabilities		
		1	2	3	4	5	6
1	0.7 }	A					
2	0.15 }	0.15		A			
3	0.06 }	0.06		0.04 }	0.04 }	A	
4	0.04 }	0.04				0.03 }	B
5	0.03 }	0.03		0.06 }	A		
6	0.01 }	0.01			0.01 }	0.01 } A	
7	0.005 }	0.005		0.005 }		0.005 }	0.005 } A
8	0.005 }	0.005			0.005 }		

FIGURE 2.6.1-3 SHANNON-FANO CODING PROCEDURE (STEP-BY-STEP ANALYSIS)

Although Figure 2.6.1-3 is useful to illustrate the step-by-step coding procedure, the Shannon-Fano coding tree is given in Figure 2.6.1-4.

$P(\mu_k)$	Code Words	$L(k)$	$P(\mu_k)L(k)$
0.7	1	1	0.70
0.15	01	2	0.30
0.06	0001	4	0.24
0.04	0010	4	0.16
0.03	0011	4	0.12
0.01	00001	5	0.05
0.005	000001	6	0.03
0.005	000000	6	0.03

FIGURE 2.6.1-4 SHANNON-FANO CODING

From Figure 2.6.1-4 it is possible to calculate the encoded entropy $H(SF)$ for the Shannon-Fano coding which is given by

$$H(SF) = \sum_{k=1}^8 P(\mu_k) L(\mu_k) = 1.630 \frac{\text{bits}}{\text{message}}.$$

However, notice that in auxiliary ensemble there are many possible ways to divide the probabilities near equal. This choice will then affect the encoded entropy. For example, an alternate choice, but a valid choice, for the auxiliary ensembles will result in an encoded entropy:

$$H(SF)_2 = 1.65 \frac{\text{bits}}{\text{message}}.$$

Therefore, the Shannon-Fano coding technique is dependent upon the ensemble choice whereas the Huffman coding technique is a systematic coding technique yielding an optimum set of code words. It is now possible to state the following relationship:

$$H(\mu) \leq H(E) \leq H(SF);$$

That is, the encoding entropy for the Shannon-Fano code is always greater than or equal to the encoded entropy of the Huffman code.

The Huffman coding technique is the only optimum code that yields an entropy less than any other coding technique. The Shannon-Fano coding technique in general

is not as efficient as the Huffman coding technique. The Huffman code has a little-recognized property in that it is self-synchronizing;⁽⁵⁾ that is, if the decoder should get out of synchronization, the Huffman decoder will regain word synchronization automatically after a few elements, whereas for PCM it is necessary to transmit word synchronizing and if the message becomes out of synchronization, the PCM message is completely lost without word synchronization. This is an advantage to be considered fully in comparing the Huffman code with other codes. However, the effects of errors on the Huffman code must be studied to fully evaluate the coding technique.

2.6.2 Noise Performance

A general analysis of errors on the Huffman code would be a formidable task since the code words are assigned according to the probabilities of the various events. Therefore, a specialized Huffman code was developed that is optimal for a specified criterion. Consider the ensemble of messages $\mu_0, \mu_1, \dots, \mu_n$ with probabilities $P(\mu_0), \dots, P(\mu_n)$ which are to be encoded with the optimum Huffman code. Assume the statistics $P(\mu_i)$ are as follows:

$$P(\mu_0) = PE$$

$$P(\mu_k) = \frac{(1-PE)}{n} \text{ for } k = 1 \dots n.$$

$$P(\mu_0) + \sum_{k=1}^n P(\mu_k) = 1,$$

(78)

and for this set of statistics the Huffman coding technique will assign to the element μ_k of the ensemble a binary alphabet of $L(k)$ bits where $L(0) = 1$ and $L(k) = N+1$ for $k \neq 0$.

Assume now that there is a data stream from an N -bit digital TV source. This can be described by a homogeneous Markov chain of first order; i.e., it can be described by a series of 2^N states S_i , the source being in the state S_i if the previous element in the data was the level i , and a series of probabilities $P_i(j)$ of transition from a state S_i to a state S_j . This transition accompanies the occurrence in the data of the level j . Since the Markov chain is ergodic,* the

The problem of ergodicity is discussed by Shannon in "The Mathematical Theory of Communication," University of Illinois Press, 1959, beginning on page 15. This property of the Markov process above follows from its irreducibility and the fact that $PE > 0$ [hence, $P_i(i) > 0$ for some i and the existence of loops of length 1 in Shannon's graphical representation].

first order probabilities $P(i)$ (which is the limiting distribution of the random variable i) is the only stationary state and may be found from solution of the equations, (14)

$$P(j) = \sum_i P(i) P_i(j). \quad (79)$$

These data will then be encoded by elements of the foregoing ensemble ($\mu_0 \dots \mu_n$) by encoding the level i as μ_0 or μ_i according as i was or was not preceded by itself in the data stream. The probability of occurrence of μ_0 will then be related to the Markov process by

$$PE = \sum_i P(i) P_i(i). \quad (80)$$

Classification of Errors

Consider the effects of a random error occurring in the sequence of encoded information. Since this sequence is composed of words of variable length, the effects of a particular error will depend strongly on the exact location of the error. The main possibilities are as follows:

- o The error can occur in a μ_0 . This will cause this element and the next N bits to be read together as a spurious level. In addition it will cause a loss of "word synchronization"; i.e., because of the variable word length the receiver and transmitter will no longer come to the end of a "word" at the same point in the data stream. The Huffman code has a self-synchronizing property, but it will take time for re-synchronization to occur.
- o The error may occur in the first bit of a member of the μ_k ensemble ($k \neq 0$). In this case the first bit of μ_k would be read as μ_0 and word synchronization would again be lost.
- o The error may occur in any of the N succeeding bits of μ_k . In this case there will be no loss of word synchronization, but merely an error in reading μ_k .

Probabilities of Occurrence

The average number of bits per word in the encoded data is given by

$$W = (PE) 1 + (1-PE) (N+1) = 1 + N(1-PE) \quad (81)$$

The average number of bits per element that refer to μ_0 is 1 (PE). Therefore, the probability that a random error in the encoded data will occur in a μ_0 is

$$P_1 = \frac{PE}{1 + N(1-PE)} \quad (82)$$

which is the probability that a random error will be of the first type.

If the error does not occur in a μ_0 , then the probability is $1/N + 1$ that it will occur in the first bit of one of the μ_k . The probability that a random error be of this type is therefore

$$\begin{aligned} P_2 &= \left[1 - \frac{PE}{1 + N(1-PE)} \right] \left(\frac{1}{N+1} \right) \\ &= \frac{1-PE}{1 + N(1-PE)} \end{aligned} \quad (83)$$

The last type of error occurs when the random error falls in any of the last N bits of a μ_k word, and is therefore exactly N times as likely as one of the second type. The probability of occurrence of this type of error is

$$P_3 = \frac{N(1-PE)}{1 + N(1-PE)} \quad (84)$$

Figure 2.6.2-1 is a plot of P_3 vs PE .

Effects of Errors

The most serious problems in analyzing the effects of errors on a Huffman code stem from the loss of word synchronization encountered with errors of the first two types. As previously remarked, Huffman codes are self-synchronizing in the sense that this loss of word synchronization will eventually be restored by the structure of the code itself, but the time required is dependent upon the structure of succeeding words and is too complex to analyze completely at this juncture. In addition to the propagation of the error for an undetermined distance there is also the problem that synchronization may occur with significant shortening of the message length. In the case of a TV picture, this would require some type of outside synchronization at the end of each line and would mean that a portion of a line might easily become displaced by a number of places. The effects of this on picture quality have not yet been determined.

If the random error is of the first type, the first effect will be to introduce into the data a spurious level. If the error is followed by a succession of N or more PE elements, the code will regain word synchronization at that point, and the message will be shortened by N elements. The probability that an error of type 1 will follow this pattern is PE^N .

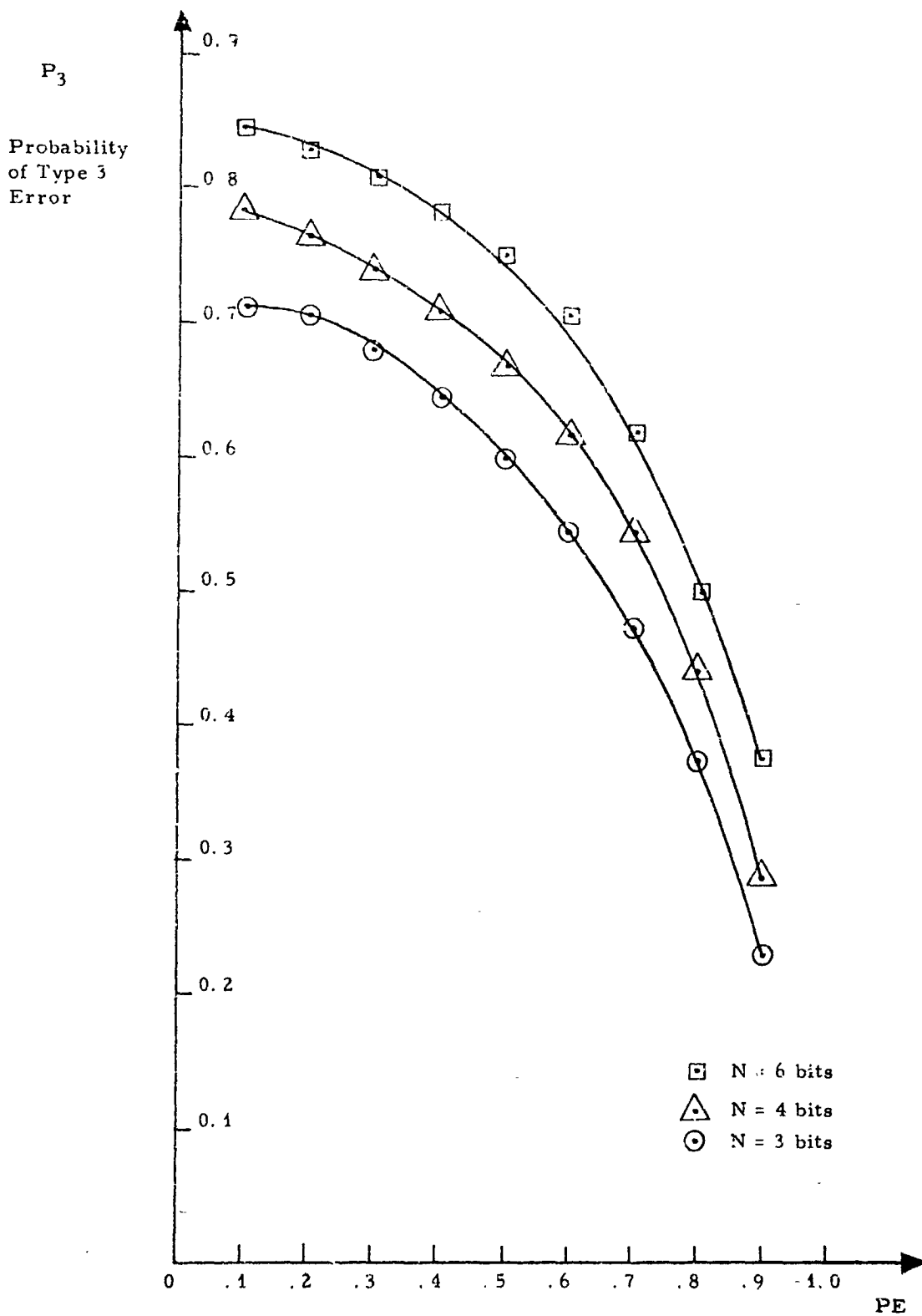


FIGURE 2.6.2-1 PROBABILITY OF ERROR OF TYPE 3 PER RANDOM ERROR AS A FUNCTION OF PE PARAMETRIC IN N

If the random error is of the second type, an element μ_k will fail to be recognized as such and will be read as a message beginning with the element μ_0 . If it was the element μ_n of the ensemble which contains N ones, then the message will be lengthened by N elements, the N + 1 bits of the μ_n word will all be read as μ_0 , and the code will then regain word synchronization at the end of the word. The probability that an error of type 2 will follow this pattern is $(1-PE)/N$.

In neither of these cases does regaining word synchronization mean regaining accuracy in the data. The code is resynchronized, but it is synchronized on an incorrect level, and this error will remain until corrected by the next μ_k word. The change in message length will also remain until corrected by external synchronization at the end of the line.

If the error is of the third type, disturbance of the data is minimal and the loss of synchronization that was characteristic of the first two types of error does not occur. However, the error will not in general be confined to a single word but will be propagated by the μ_0 words in the encoded message. In this case one will be able to determine the average picture degradation in terms of the average length of this sequence of erroneous data.

The average length L_i at a sequence of elements $\mu_i \mu_0 \mu_0 \dots$, referring to the i th level, is

$$L_i = 1 + P_i(i) + P_i(i)^2 + \dots = \frac{1}{1 - P_i(i)} \quad (85)$$

Therefore, if an error occurs in one of the last N bits of μ_i it will cause an average of L_i errors in the encoded data. Since a run of M elements $\mu_i \mu_0 \dots \mu_0$ requires $N + 1 + (M-1) \cdot 1 = N + M$ bits in the encoded one can also conclude from Equation 8 that an average of $\frac{N + L_i}{L_i}$ bits of encoded data are required, each element referring to the level i . Thus, the probability that a random error refer to the level i is $\frac{P(i) W_i}{W}$. If the error does refer to this level, the probability is $N/N + L_i$ that it will be an error of type 3. Therefore, the probability that an error of type 3 will refer to the i th level (and hence introduce L_i errors into the data) is the probability that a random error will be an error of type 3 referring to the i th level divided by the probability that the random error will be a level error, or

$$\text{Prob.} = P(i) \frac{W_i}{W} \cdot \frac{N + L_i}{L_i} = \frac{N(1-PE)}{1 - N(1-PE)} \quad (86)$$

which simplifies to

$$\text{Prob.} = \frac{P(i) [1 - P_i(i)]}{1 - PE} \quad (87)$$

The number of errors introduced by this error is

$$L_i = \frac{1}{1 - P_i(i)} ,$$

and hence the average number of errors (R) introduced by an error of type 3 is

$$\begin{aligned} R &= \sum_i \frac{P(i) [1 - P_i(i)]}{1 - PE} \cdot \frac{1}{1 - P_i(i)} = \sum_i \frac{P(i)}{1 - PE} \\ &= \frac{1}{1 - PE} , \end{aligned} \quad (88)$$

which is also the average run length.

A plot of R as a function of PE is given in Figure 2.6.2-2.

Finally one can ask for the expected number of errors of this type per random error in the encoded data. The probability that a random error will be an error of type 3 in μ_i is given in (86) by

$$\frac{P(i) W_i}{W} \cdot \frac{N + L_i}{L_i} \cdot \frac{P(i) N (1 + P_i(i))}{W} \quad (89)$$

This introduces an expected L_i errors, and therefore the expected number of errors of this type (E_3) introduced by a random error is

$$E_3 = \sum_i \frac{P(i) N [1 - P_i(i)]}{W} \cdot \frac{1}{1 - P_i(i)} = \frac{N}{1 + N(1 - PE)} . \quad (90)$$

A plot of E_3 vs PE is given in Figure 2.6.2-3.

A future program to continue this analysis of the effects of noise on Huffman code would be to investigate the effects of loss of word synchronization and alteration of line length on picture quality. In the first case the problem would center on finding the mean resynchronization time and the average change in line length, while the second problem would be concerned with evaluation of the loss of data involved in alteration of line length. The final step would be to investigate the effects of errors on synchronization of the overall system. Such a program would be more fruitful if accompanied by an experimental program which included transmitting the Huffman data stream over a controlled-noise, binary symmetric channel.

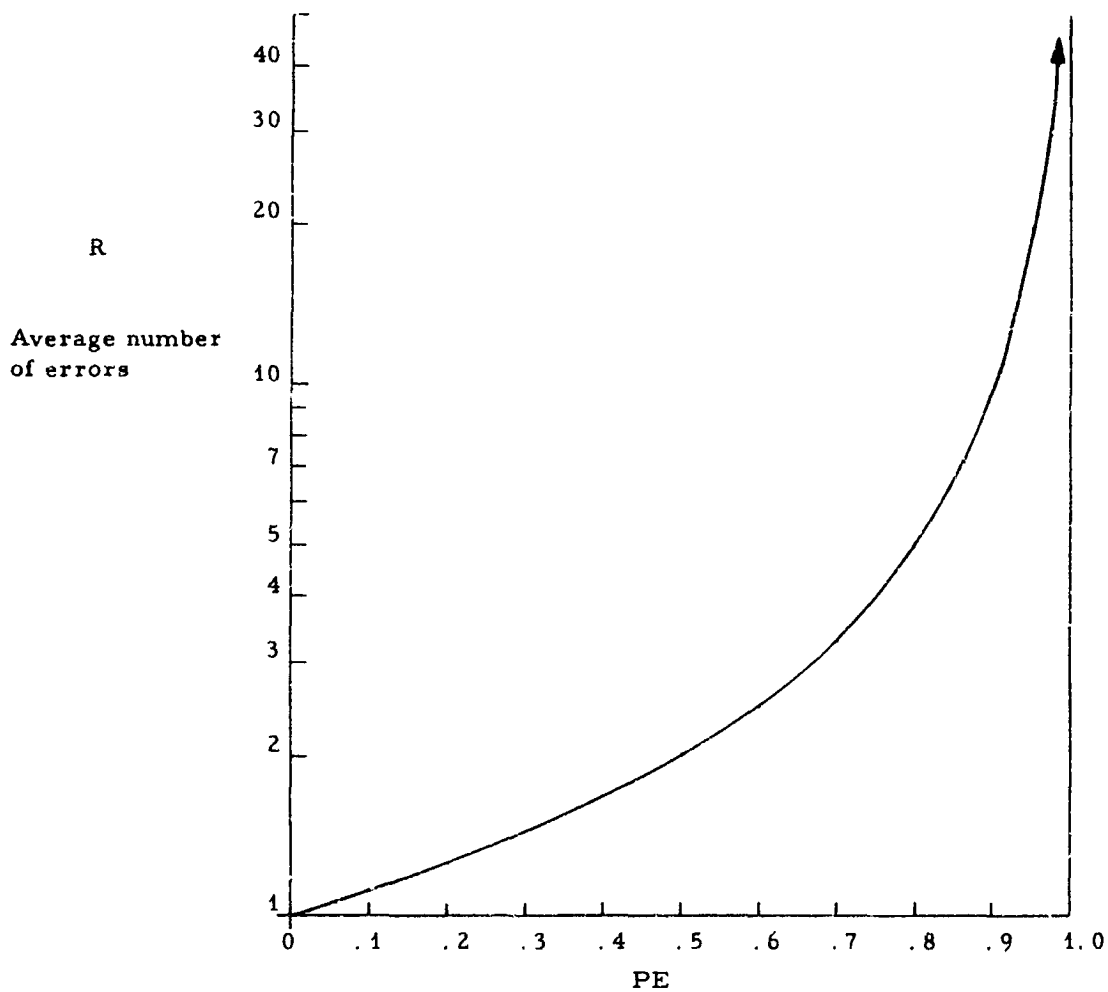


FIGURE 2.6.2-2 AVERAGE NUMBER OF ERRORS PER ERROR OF TYPE 3
AS A FUNCTION OF PE

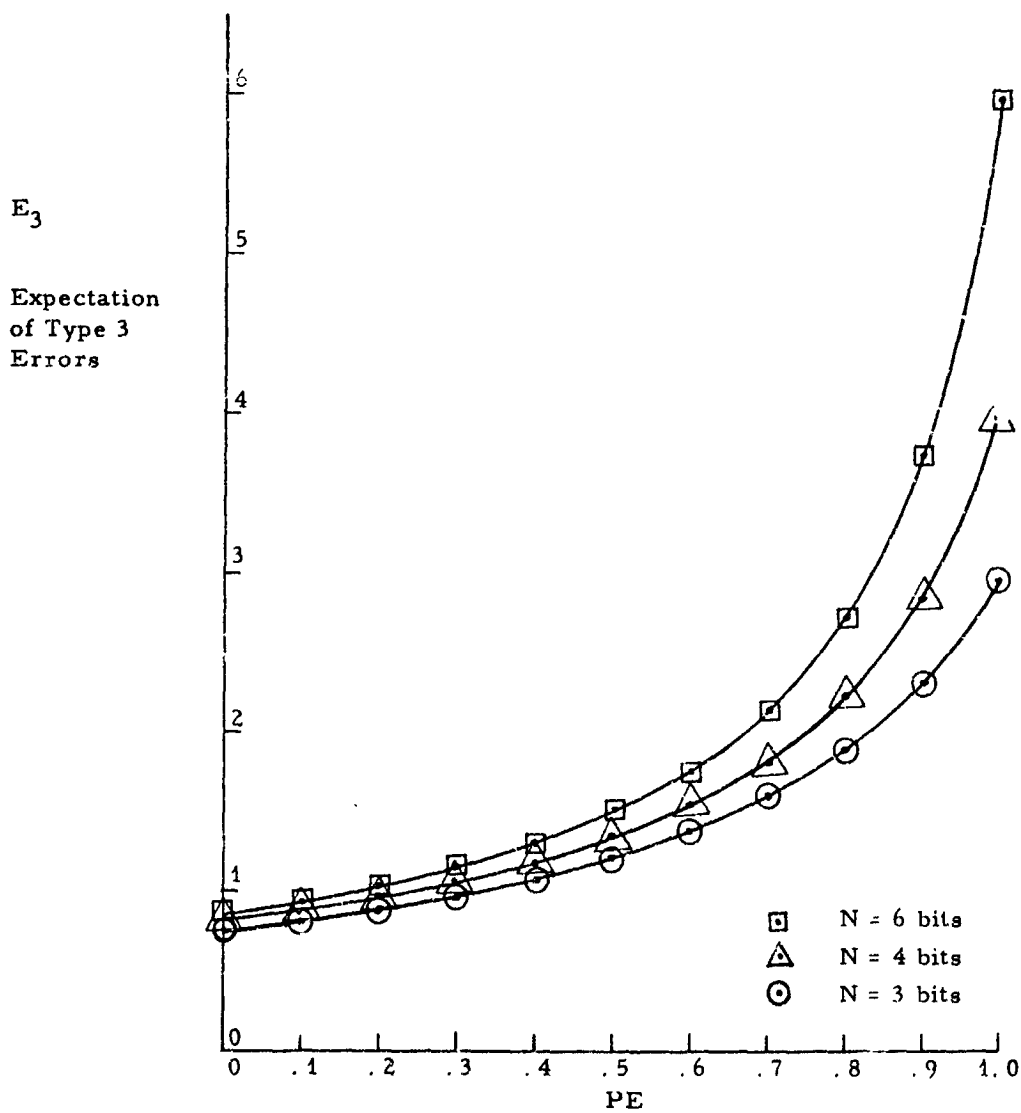


FIGURE 2.6.2-3 NUMBER OF ERRORS INTRODUCED BY ERRORS OF TYPE 3 PER RANDOM ERROR AS A FUNCTION OF PE PARAMETRIC IN N

2.6.3 References for 2.6

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2.7 NOTE ON TEMPORARY STORAGE REQUIREMENTS

The buffer storage requirement in the previous-element coding system is easily met using flip-flops since a delay of only 1μ sec is required for 3 or 4 bits in order to compare the current picture element with its previous neighbor. The other area of buffer storage, that of matching a constant word frequency-variable word length source to a constant bit rate channel is avoided in the system discussed in Section 1 by using the encoder to control the rate of the source (vidicon). In fact, this is a very significant point in the design of the system.

The line-to-line compression systems (PE-PL, for instance) impose a much more severe storage requirement since the three or four bits must be delayed (on a continuous basis) for one line time. One line time is, in the MSC case, 256μ sec. If the system is operated synchronously, then this delay-bit number product is easily met (See Figure 2.7-1) with a wire-sonic NRZ delay line (Tempo Instrument, Inc., Plainview, New York, 4.0 mc. with delay to 800μ sec). However, if the system operates asynchronously, as in Section 1, then the previous line delay store must operate asynchronously also. In this case three or four (depending on the number of bits quantization) 256 stage, 1 mc., shift registers are required for previous-line storage. This is a formidable shift register requirement, yet it is not an impossible one considering the current state of the semiconductor integrated circuitry art.

Compression systems involving frame-to-frame correlations impose the most severe requirements on storage equipment. The PE-PL-PF compression system compares a given element in the current picture frame with the equivalent element in the preceding picture frame. This is done on a continuous basis and necessitates the storage on a continuous, and probably asynchronous, basis of $256 \times 256 \times 3$ or 4 bits. This 200- to 260-kilobit, 1 -mc. storage requirement indicates that the use of magnetic core storage is mandatory. This order-of-magnitude of storage requirement is similar to that indicated in Section 5.2.2 where a core store is considered as the heart of a scan-conversion system.

General spacecraft storage of program material is the next higher step in storage requirements. One hour of MSC video requires the storage of $3600 \times 10^6 \times 4 = 1.44 \times 10^{10}$ bits at 10^6 4-bit picture elements per second. This magnitude of storage requires either photographic or magnetic-tape storage. Some aspects of the problem are treated further in Section 4.

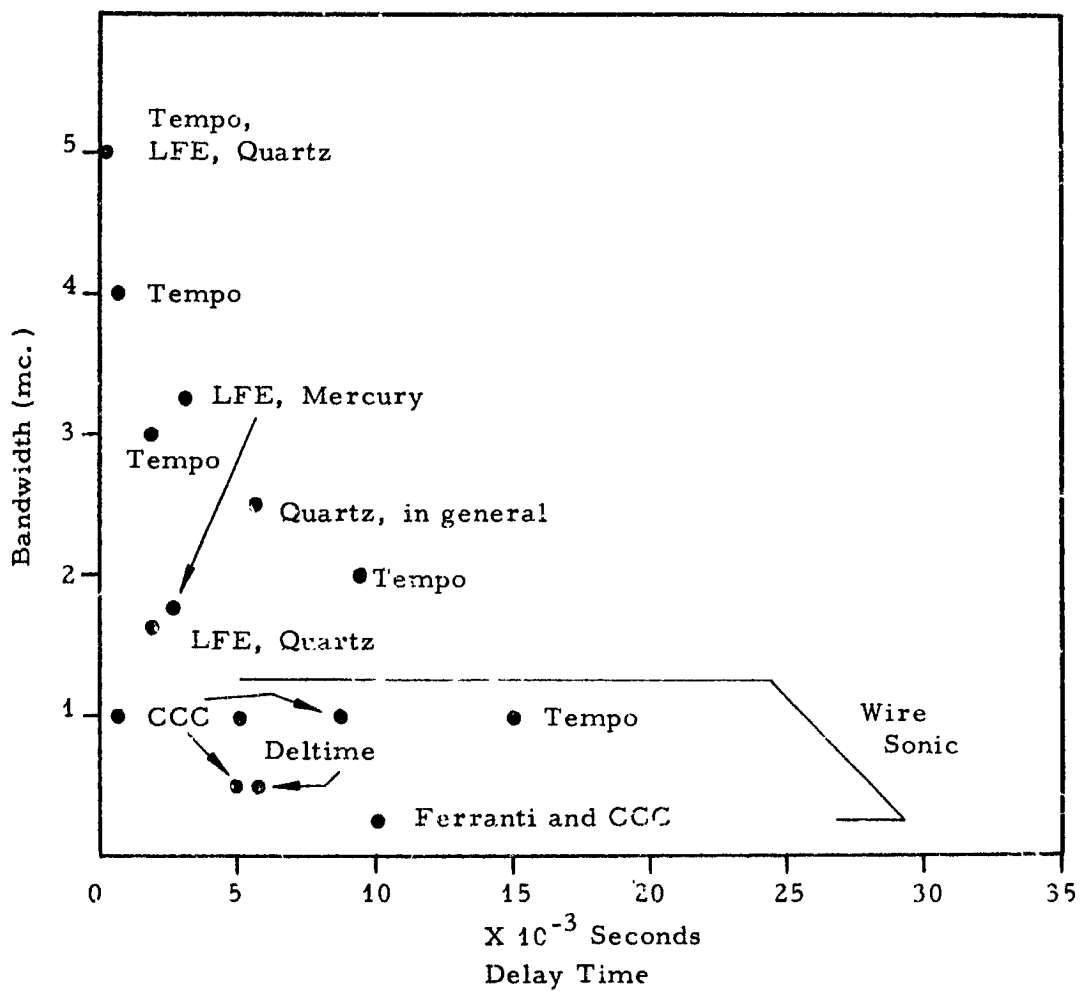


FIGURE 2.7-1 DELAY-BANDWIDTH AVAILABILITIES

SECTION 3

RF LINKS

3.0 EXECUTIVE SUMMARY

The bandwidth required for the transmission of digital television data from a spacecraft was calculated for the specified parameters of Conditions I and II; the upper bound on the bandwidth is 23.4 mc and 6.46 mc for Conditions I and II, respectively. If it is assumed that the transmitter power is fixed at 25 and 1000 watts for Conditions I and II, the maximum frame rate that can be achieved at 525-line resolution for a compressed and uncompressed system can be calculated. For a 525-line resolution and six bits per element the maximum frame rate for Conditions I and II is 14.15 fps and 3.91 fps, respectively. However, should the coding result in a 3:1 compression, then at 525-line resolution and two bits per element, the maximum frame rate is 42.45 fps and 11.72 fps for Conditions I and II, respectively.

Although compression would result in a subjectively acceptable picture, additional compression could be achieved by decreasing the resolution, and the displayed picture would be subjectively acceptable. Experiments indicate that 256-line resolution is sufficient for most of the applications. Assuming for Condition I that the minimum frame rate is 15 fps, a compression of 3:1 would result in a savings of prime power (assuming 33-1/3% efficiency) of 12.60 watts, whereas for the same compression and frame rate for Condition II, but assuming 20% efficiency, the compressed system would result in a saving of 30.39 watts of prime power. The power required for these systems would be 2. and 303.9 watts, respectively, for Condition I and II.

3.1 RF ANALYSIS

The general derivation of the bandwidth formula for digital television transmission from a spacecraft is presented below. From this equation it is possible to relate the frame rate, resolution, coded bits per element, bit rate, bandwidth, transmitter power, and losses for a specified set of parameters. The nomenclature used in the derivation is given in Section 3.4.

The received power from the spacecraft is given by the familiar "radar range equation" stated in Equation 1.

$$P_r = \frac{P_t G_1 G_2 \lambda^2}{16 \pi^2 R^2}, \text{ watts;} \quad (1)$$

However,

$$\frac{16\pi^2 R^2}{\lambda^2}$$

is designated as the path loss and Equation 1 can be rewritten as

$$P_R = \frac{P_t G_1 G_2}{PL}, \text{ watts.} \quad (2)$$

The net transmitter power is the total available power divided by the losses; that is,

$$P_t^1 = \frac{P_t}{L} \text{ watts,} \quad (3)$$

and Equation 2 can be rewritten as

$$P_R = \frac{P_t G_1 G_2}{LPL}, \text{ watts.} \quad (4)$$

Although it would be possible to continue the derivation in watts, it is convenient to use decibels. Equation 4 can be rewritten in terms of db's using capital subscripts indicating decibels; therefore, in decibels, Equation 4 becomes

$$P_R = P_T + G_1 + G_2 - P_L - L, \text{ db.} \quad (5)$$

The received power can be related to the signal-to-noise ratio and noise power by

$$P_R = \frac{S}{N} + N, \text{ db,} \quad (6)$$

which can be substituted into Equation 5, and by re-arranging terms it becomes

$$N = P_T + G_1 + G_2 - (P_L + \frac{S}{N} + L), \text{ db.} \quad (7)$$

However, the noise power N is related to the bandwidth B by

$$N = k + T + B, \text{ db}, \quad (8)$$

which can be substituted into Equation 7, which becomes

$$B = P_T + G_1 + G_2 - (P_L + \frac{S}{N} + k + T + L), \text{ db}. \quad (9)$$

If the parameters of Equation 9 are specified, it is possible to obtain the bandwidth required in decibels and then convert to cycles per second. Consider for example the parameters of Table 3-1 as conditions for a space mission.

TABLE 3-1. PARAMETERS FOR TWO SPACE MISSIONS

Parameter	Condition I	Condition II
Range (R)	250,000 stat. mi.	53,000,000 stat. mi.
Spacecraft transmitter power (P_T)	25 watts	1 kw
Spacecraft transmission line loss (L_1)	1.5 db	1.5 db
Frequency (f)	2.3 kmc	2.3 kmc
Spacecraft antenna (G_1)	5-ft. parabola	10-ft. parabola
Spacecraft antenna pointing loss (L_2)	1.0 db	1.0 db
Ground station antenna (G_2)	85-ft. parabola	225-ft. parabola
Ground station receiver bandwidth (B)	To be determined	To be determined
Ground station excess noise temperature (T)	330°K	30°K

The total losses for the parameters of Table 3-1 are a lower bound on losses. To determine the upper bound on space communication system losses a short survey was conducted to tabulate some of the losses in the RF link. Among the losses to be considered along the transmission path are: gaseous absorption, Faraday rotation, ionosphere reflection, rain attenuation, galactic noise, transmission line, system noise, antenna pointing, system deterioration, system design margin, theoretical design, fading, and miscellaneous spacecraft power losses. These losses are tabulated in Table 3-2 along with the reference.

Although Faraday rotation has been considered, it will be assumed that this loss can be neglected for future space communications systems. The theoretical design loss is used in conjunction with the signal-to-noise ratio. The theoretical curves of bit error probability versus signal-to-noise ratio for various systems necessitate the theoretical design loss. JPL has assumed that it is possible to design within 1.5 db of the theoretical expected value. Therefore, for upper-bound calculations the theoretical design loss should be considered. The system design margin varies from 6 to 10 db; however, JPL has determined from ground evaluation of Mariner that 8 db of margin is sufficient. Antenna pointing and transmission-line losses are generally specified and are not discussed.

The fading loss, like the system design margin, is quite ambiguous at the present time. In Explorer VI, 10 to 20 db of fading was experienced at VHF, which was

not explained. Several authors have conducted theoretical studies of fading, such as Rayleigh, Gaussian, and exponential for the various types of modulation. The area of fading requires additional information before the fading loss can be determined. For example, to have confidence limits of 50 to 99 per cent for Rayleigh fading requires a system design margin of 18 db, while the margin for Gaussian and exponential fading is only 1.5 db. For the lower-bound calculations the losses to be considered are 1.0 db for antenna pointing, 1.5 db for transmission-line loss or a total loss excluding free-space transmission of 2.5 db.

TABLE 3-2. LOSSES ASSUMED BY VARIOUS AUTHORS

Loss (db) ^a	Reference													
	1	2	3	4	5	6	7	8	9	10	11	12	13	14
La	-	-	-	-	-	-	-	-	-	0	1.0	-	-	1
Lc	-	-	-	-	-	3.0	-	-	-	0	-	-	3.0	3.0
Le	-	-	1.5	1.5	1.5	-	-	-	-	-	-	-	-	-
Lf	-	-	-	-	-	-	1.5 ^b	-	-	-	-	-	? ^c	-
Lm	2.0	1.3	-	2.0	6.0	1.0	-	-	-	-	-	3.0	2.0	-
Ls	8.0	-	-	8.0	6.8	-	-	-	-	5.0	-	6.0	8.0	7.8
Lt	-	-	-	-	-	-	-	2.0 ^d	18.5 ^e	1.0	1.5	-	-	1

a. La = antenna pointing loss

Lc = polarization loss --since the path loss is referenced between two isotropic antennas, the antenna gain must also be referred to isotropic. However, it is also necessary to know whether the receiving antenna is of the same polarization as the transmitted wave. If not, the effective receiving antenna gain will be approximately 3 db lower.

Le = theoretical design loss--in general when considering mechanizing a system, the system will perform to within 1.5 of theoretical performance.

Lf = fading loss

Lm = miscellaneous spacecraft power loss

Ls = system design margin--in general the system design margin should include deterioration of equipment and unusual propagation conditions from either solar flares or man-made interference.

Lt = transmission line loss

b. For $P_e = 10^{-5}$, exponential fading auto-correlation and second order diversity.

c. At VHF Explorer IV experienced 10 to 20-db fading loss.

d. At 2000 Mc

e. For Rayleigh fading of desired signal and normal fading of the undesired signal and for 50 to 99 per cent satisfactory operation.

For upper-bound calculation the losses will be 1.0 db for antenna pointing, 1.5 db for theoretical design, 1.5 db for transmission line loss, 8.0 db for system design margin, and 0.5 db for fading--or a total loss excluding free space transmission of 12.5 db.

It is now possible to calculate the upper and lower bounds on the bandwidth required for conditions I and II.

3.2 REAL-TIME TRANSMISSION

For real-time transmission the parameters for calculating the upper and lower bound of the bandwidth necessary to evaluate Equation 9 can be obtained from Table 3-1 and are tabulated in Table 3-2.

TABLE 3-2. PARAMETERS FOR REAL-TIME TRANSMISSION

Parameter (db)	P_T	G_1	G_2	P_L	K	T	L	S/N
Upper-Bound	14	28.4	53.0	211.8	-228.6	25.2	2.5	10.3
Lower-Bound	14	28.4	53.0	211.8	-228.6	25.2	12.5	10.8

Substituting Table 3-2 into Equation 9 the upper bound on the bandwidth for real-time transmission is

$$B_{IU} = 73.7 \text{ db}$$

which expressed in megacycles becomes

$$B_{IU} = 23.4 \text{ mc.}$$

The lower bound on bandwidth for real-time transmission is.

$$B_{IL} = 63.7 \text{ db}$$

which expressed in megacycles becomes

$$B_{IL} = 2.34 \text{ mc.}$$

Although this gives an upper and lower bound on bandwidth, Equation 9 was written as a function of a constant minus losses and is given by

$$B_I = 76.2 - L, \text{ db.} \quad (10)$$

Therefore, given the losses, the bandwidth as a function of losses for Condition I was plotted in Figure 3.2-1. Although Figure 3.2-1 is sufficient for losses less than 10 db, the readability of the graph is decreased for losses greater than 10 db. For large losses, Equation 10 was plotted on semi-log graph paper and is given in Figure 3.2-2. However, for both figures the upper and lower-bound bandwidth points are appropriately located.

Once the losses are determined, it is possible to obtain a graph of the transmitter power required as a function of the digital bit rate. For differential coherent-phase-shift-keying modulation, the bandwidth is related to the bit rate by the relationship that the bandwidth required is one cycle per bit or the bandwidth is equal to the bit rate. With this relationship, Equation 9, and Table 3-2, the transmitter power as a function of the bit rate is

$$P_t = \text{Bit Rate} - 59.7 \text{ db}, \quad (11)$$

which can be rewritten in terms of the bit rate in bits per second. Therefore, Equation 11 becomes

$$P_t = 1.07 (\text{Bit Rate}) \times 10^{-6}, \text{ watts}. \quad (12)$$

Figure 3.2-3 plots Equation 12 for the upper- and lower-bound conditions for Condition I. Although this curve is extremely useful in relating the bit rate to the required transmitter power, additional information could be achieved by relating the frame rate and resolution to the transmitter power required.

The digital bit rate is determined by the resolution, frame rate, and quantization. If it is assumed that the horizontal and vertical resolution are equal and equal to R , f is the frame rate in frames per second, and b is the coded bits per element, the bit rate is

$$D = R^2 f b. \quad (13)$$

For illustration, assume that the coding used is previous-element coding that encodes each element on the average with two bits. Therefore, using this and substituting Equation 13 into Equation 12, the transmitter power required is

$$P_t = 2.14 R^2 f \times 10^{-6}, \text{ watts}. \quad (14)$$

Equation 14, which is the transmitter power as a function of the resolution and frame rate, is given by Figure 3.2-4.

Although Figure 3.2-4 gives sufficient information concerning frame rate and transmitter power at various resolutions, additional information can be achieved by plotting transmitter power as a function of frame rate for various encoding techniques. Therefore, Equation 13 is rewritten in terms of variable-encoding bits at a fixed resolution $R = 256$ and is given by

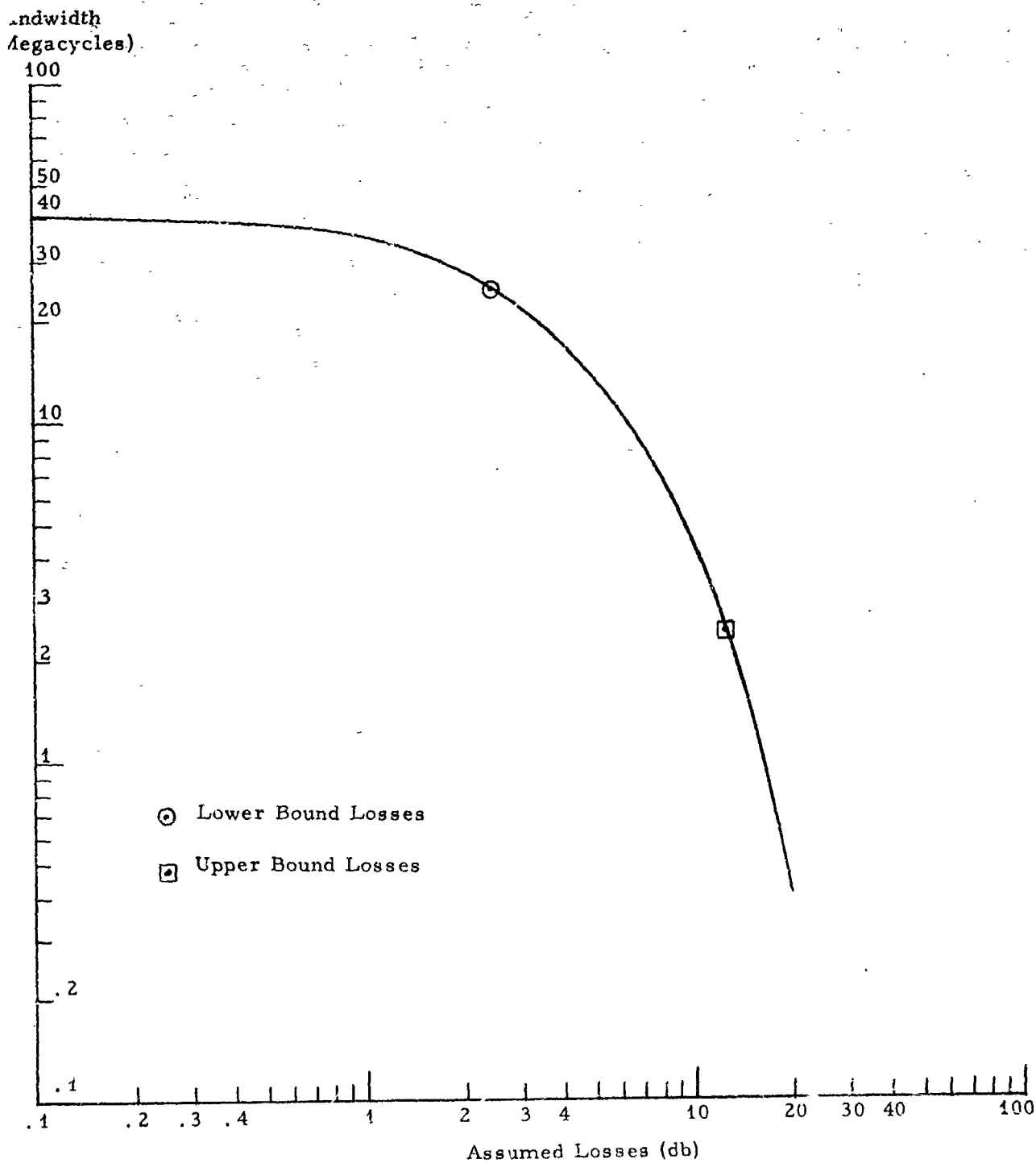


FIGURE 3.2-1 A PLOT OF BANDWIDTH AS A FUNCTION OF LOSSES FOR
CONDITION I

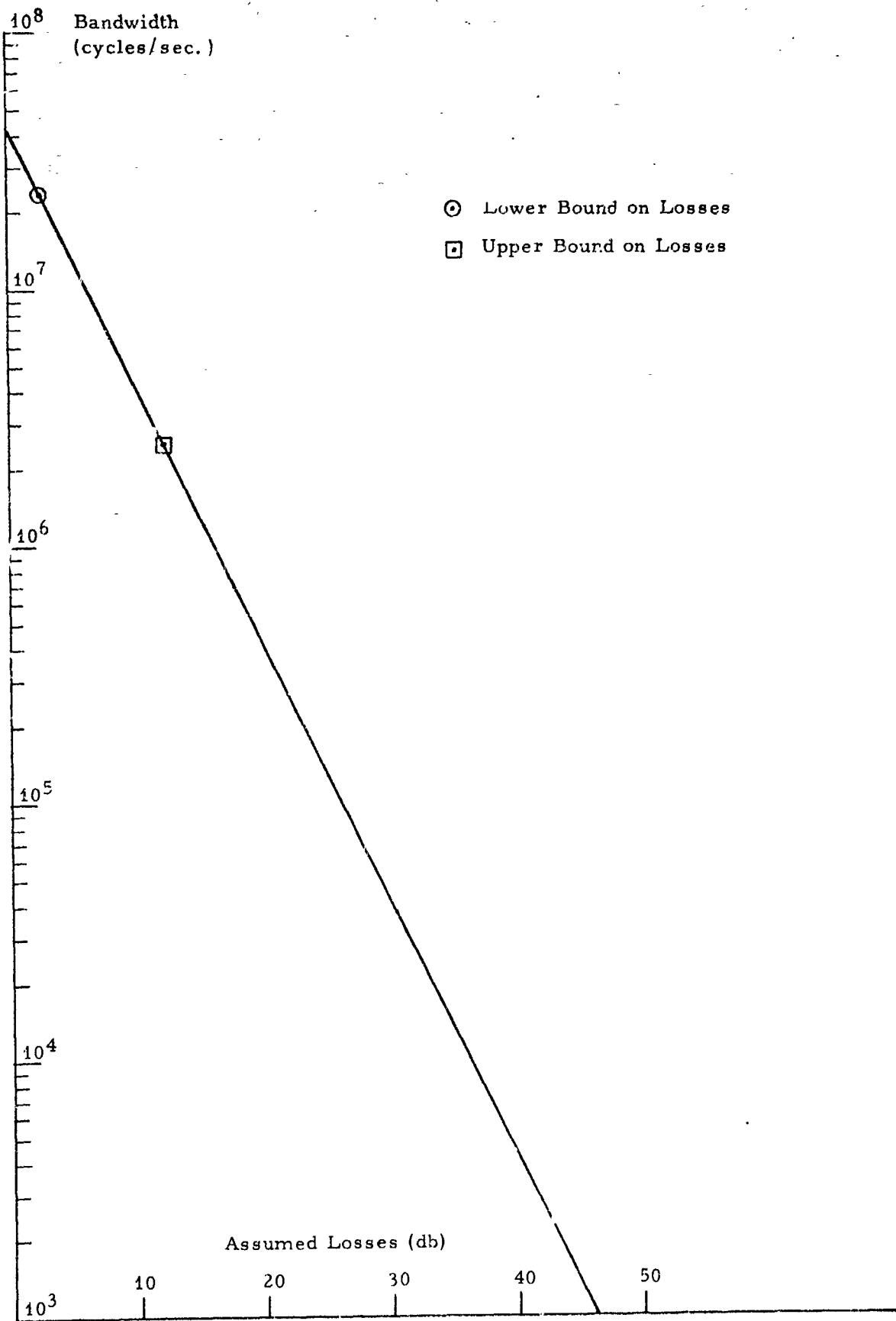


FIGURE 3.2-2 A PLOT OF BANDWIDTH AS A FUNCTION OF LOSSES FOR
CONDITION I

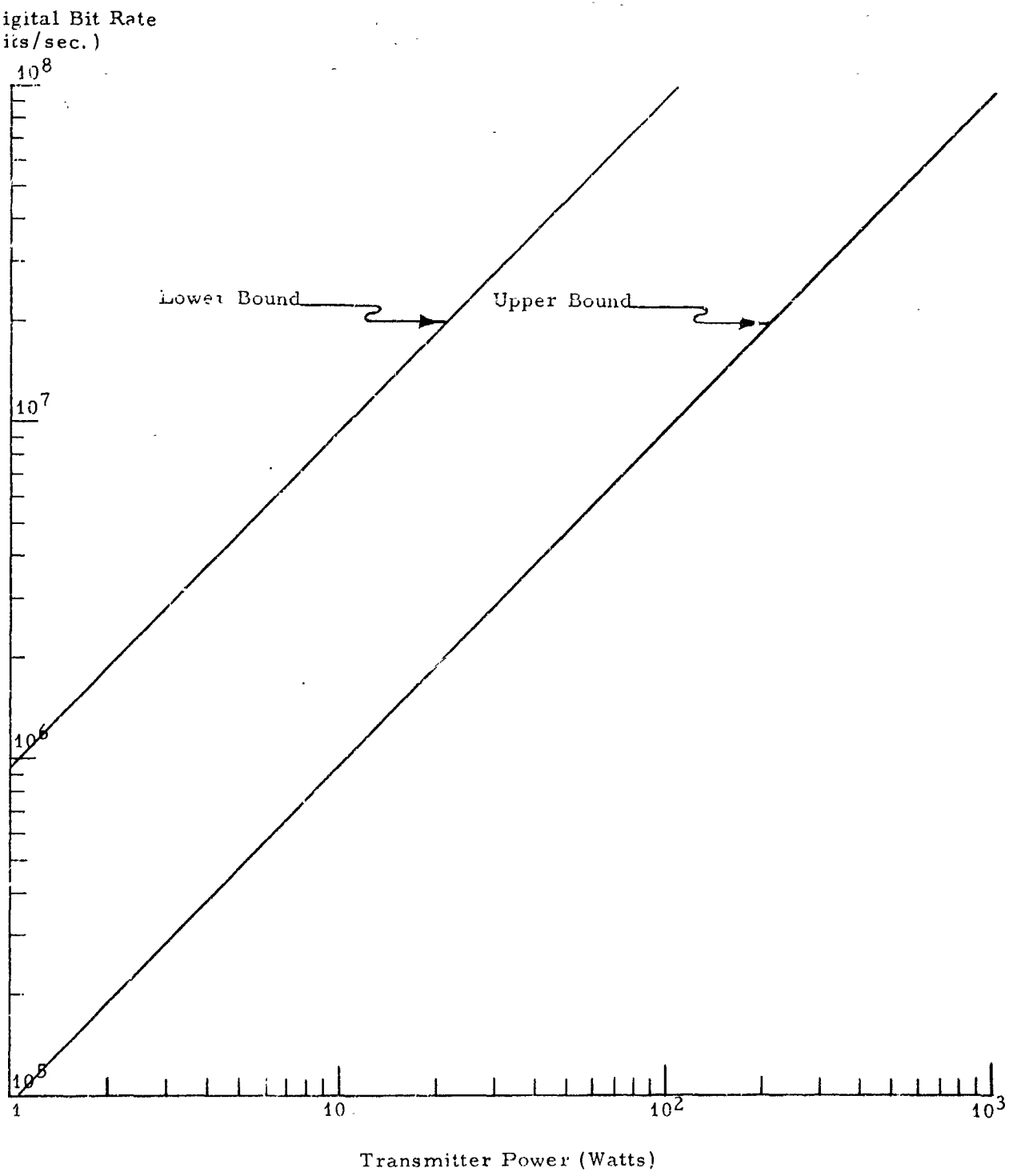


FIGURE 3.2-3 A PLOT OF TRANSMITTER POWER AS A FUNCTION OF DIGITAL BIT RATE FOR CONDITION I FOR UPPER AND LOWER BOUNDS ON LOSSES

Frame Rate
(frames/sec.)

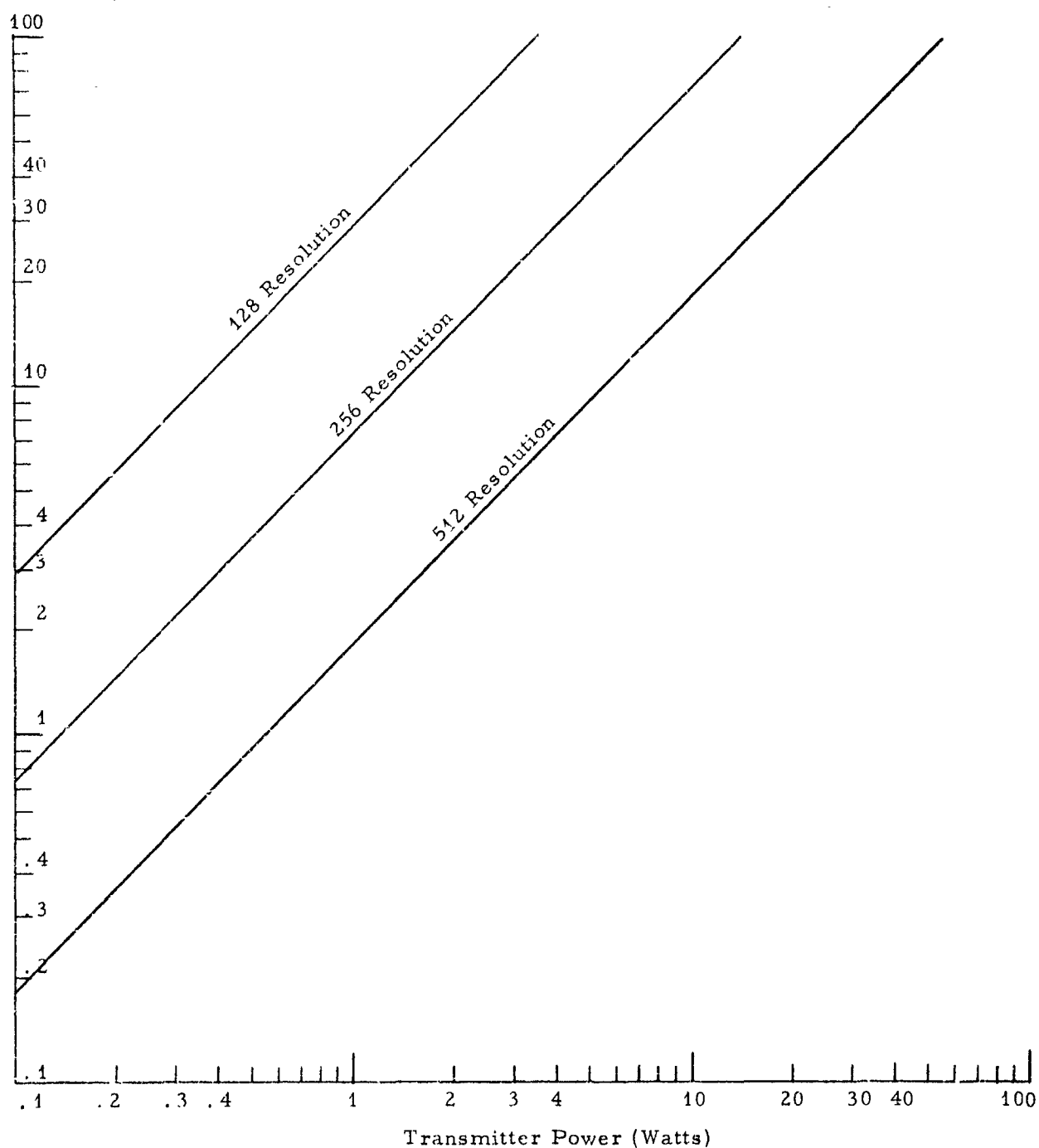


FIGURE 3.2-4

A PLOT OF TRANSMITTER POWER AS A FUNCTION OF FRAME
RATE FOR PREVIOUS ELEMENT CODING FOR CONDITION I
AT VARIOUS RESOLUTIONS

$$P_t = 7.01 f b \times 10^{-2}, \text{ watts.} \quad (15)$$

Therefore, Equation 15 is plotted in Figure 3.2-5 for previous-element coding, PCM with Roberts, four-bit PCM, and six-bit PCM coding. This curve is extremely useful to illustrate the savings in power due to coding. For example, assuming 30 fps, 6-bit PCM requires 12.6 watts while previous element coding requires only 0.2 watts. Assuming a transmitter efficiency of 33-1/3%, the savings in prime power for PEC is 25.2 watts.

However, assume that the transmitter power is fixed at 25 watts. It is possible to obtain the relationship between frame rate, resolution, and coding bits per element.

Assume again that the horizontal and vertical resolutions are equal and equal to R. From Equation 9 the relationship between the three parameters is given by

$$23.4 \times 10^6 = R^2 f b. \quad (16)$$

Therefore, Equation 16 is plotted in Figure 3.2-6, which plots the frame rate as a function of the resolution parametric in f. Therefore, assuming 600-line resolution, the maximum frame rate varies from 10.83 fps at 6 bits per element to 32.5 fps at 2 bits per element. Here, it is assumed that the transmitter power is fixed at 25 watts; however, it is necessary to minimize the power, which was pointed out in Figure 3.2-5.

3.3 STORAGE TRANSMISSION

For storage transmission the parameters for calculating the upper and lower bound of the bandwidth necessary to evaluate Equation 9 can be obtained from Table 3-1 and are tabulated in Table 3-3.

TABLE 3-3. PARAMETERS FOR STORAGE TRANSMISSION

Parameter (db)	P_T	G_1	G_2	P_L	k	T	L	S/N
Upper-Bound	30	34.4	61.5	258.3	-228.6	14.8	2.5	10.8
Lower-Bound	30	34.4	61.5	258.3	-228.6	14.8	12.5	10.8

Substituting Table 3-3 into Equation 9, the upper bound on the bandwidth for storage transmission is

$$B_{III} = 68.1 \text{ db,}$$

Frame Rate
(frames/sec.)

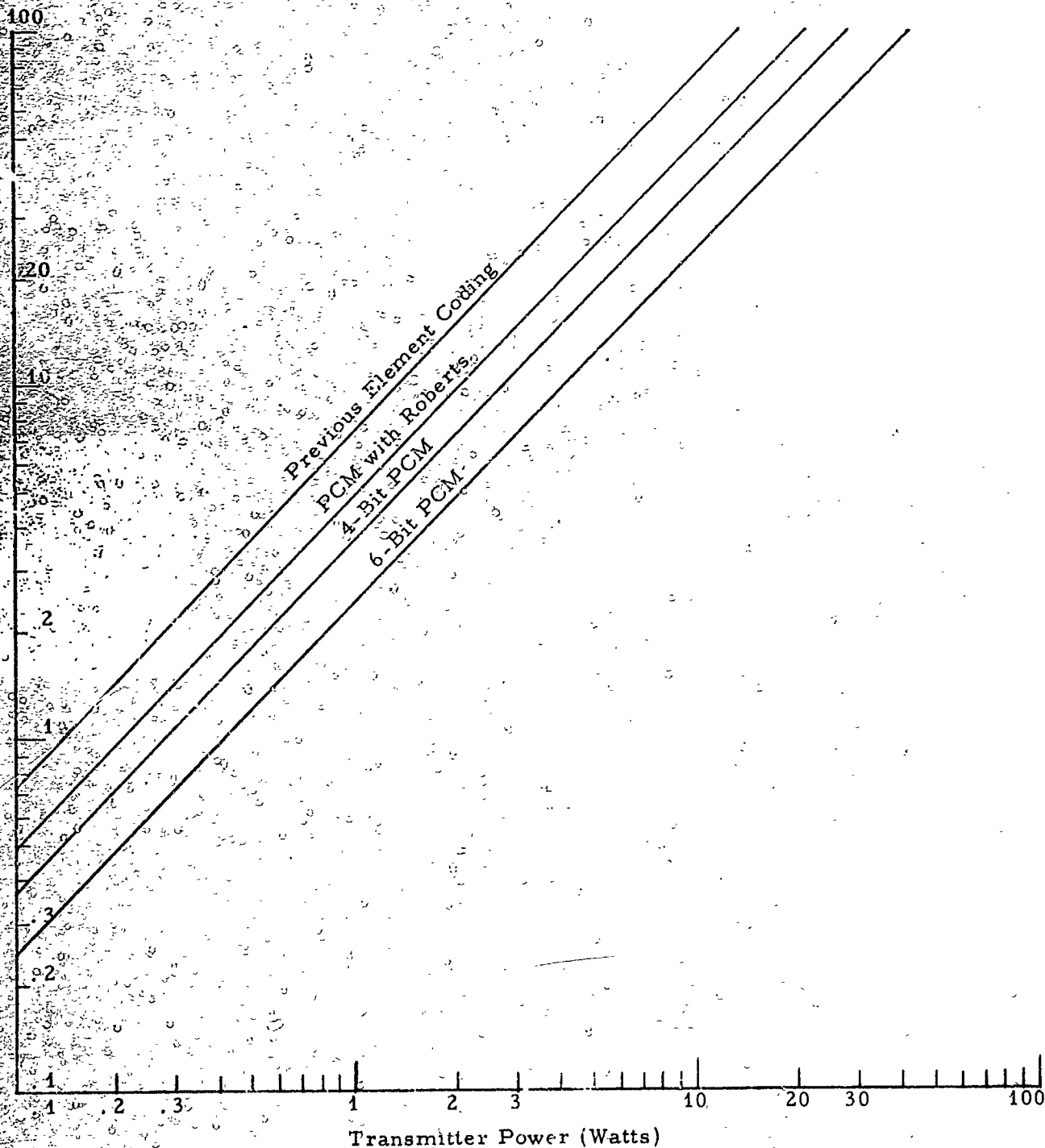


FIGURE 3.2-5 A PLOT OF TRANSMITTER POWER AS A FUNCTION OF FRAME RATE FOR VARIOUS CODING TECHNIQUES FOR CONDITION I AT 256 RESOLUTION

Frame Rate
(frames/sec)

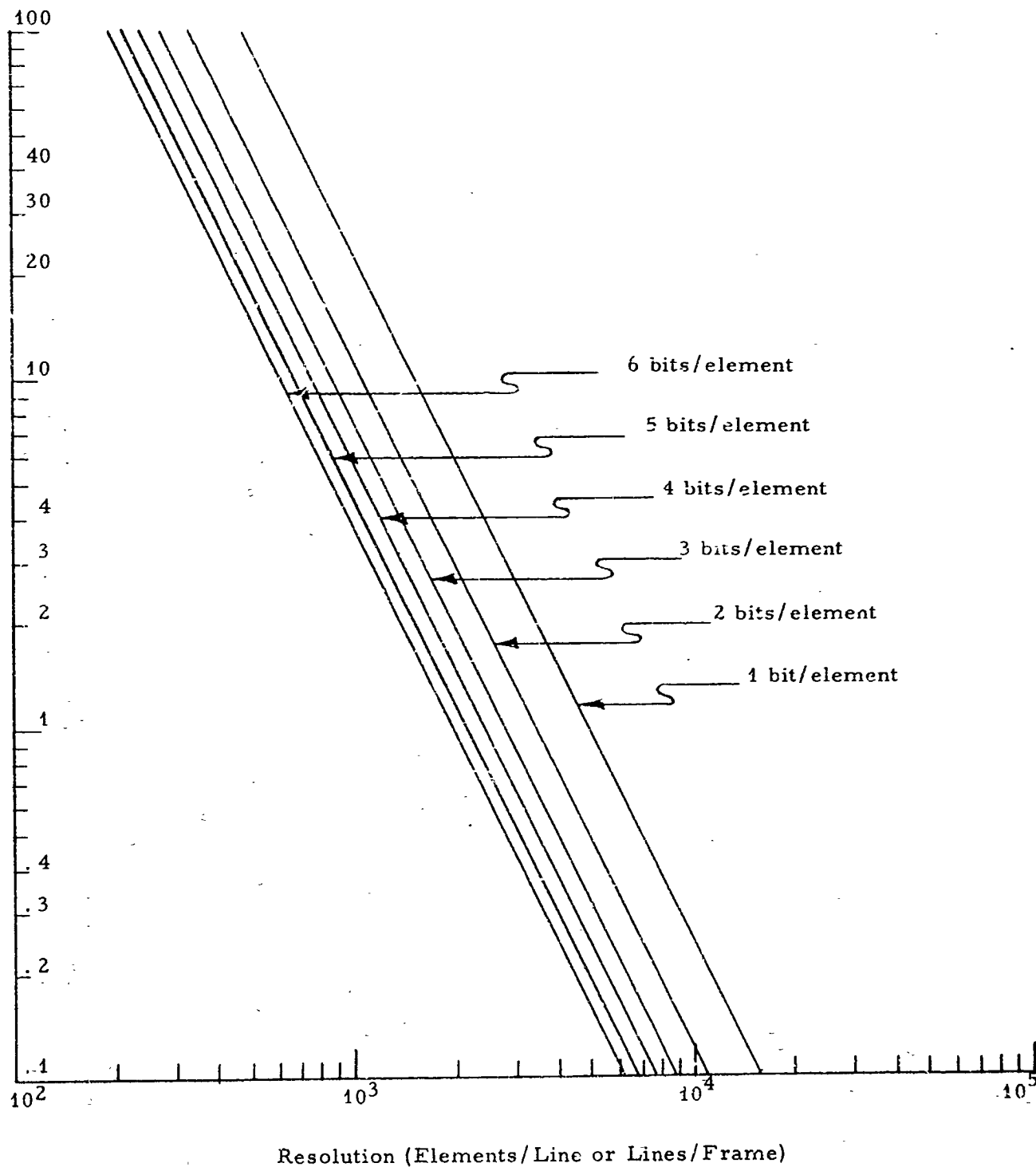


FIGURE 3.2-6

A PLOT OF FRAME RATE AS A FUNCTION OF RESOLUTION (ASSUMING HORIZONTAL RESOLUTION EQUALS VERTICAL RESOLUTION) FOR CONSTANT BITS/ELEMENT FOR CONDITION I OF MSC AT 25 WATTS TRANSMITTER POWER FIXED

which expressed in megacycles becomes

$$B_{IIIU} = 6.46 \text{ mc.}$$

The lower bound on bandwidth for storage transmission is

$$B_{IIIL} = 58.1 \text{ db,}$$

which expressed in megacycles becomes

$$B_{IIIL} = 0.646 \text{ mc.}$$

Although this gives an upper and lower bound on bandwidth, Equation 9 was written as a function of a constant minus losses and is given by

$$B_{II} = 60.6 - L, \text{ db.} \quad (17)$$

Therefore, given the losses, the bandwidth as a function of losses for Condition II was plotted in Figure 3.3-1. Although Figure 3.3-1 is sufficient for losses less than 10 db, the readability of the graph is decreased for losses greater than 10 db. For large losses, Equation 17 was plotted on semi-log graph paper and is given in Figure 3.3-2. However, for both figures the upper and lower-bound bandwidth points are appropriately located.

Once the losses are determined, it is possible to obtain a graph of the transmitter power required as a function of the digital bit rate. For differential coherent-phase-shift keying modulation, the bandwidth is related to the bit rate by the relationship that the bandwidth required is one cycle per bit or the bandwidth is equal to the bit rate. With this relationship Equation 9, and Table 3-3, the transmitter power as a function of the bit rate is

$$P_t = \text{Bit Rate} - 38.1 \text{ db} \quad (18)$$

which can be rewritten in terms of the bit rate in bits per second. Therefore, Equation 18 becomes

$$P_t = 154.8 (\text{Bit Rate}) \times 10^{-6}, \text{ watts.} \quad (19)$$

Figure 3.3-3 plots Equation 19 for the upper and lower-bound conditions for Condition II. Although this curve is extremely useful in relating the bit rate to the required transmitter power, additional information could be achieved by relating the frame rate and resolution to the transmitter power required.

The digital bit rate is determined by the resolution, frame rate, and quantization. If it is assumed that the horizontal and vertical resolution are equal and equal to R , f is the frame rate in frames per second, and b is the coded bits per element, the bit rate is (as was shown before)

Bandwidth
(Megacycles)

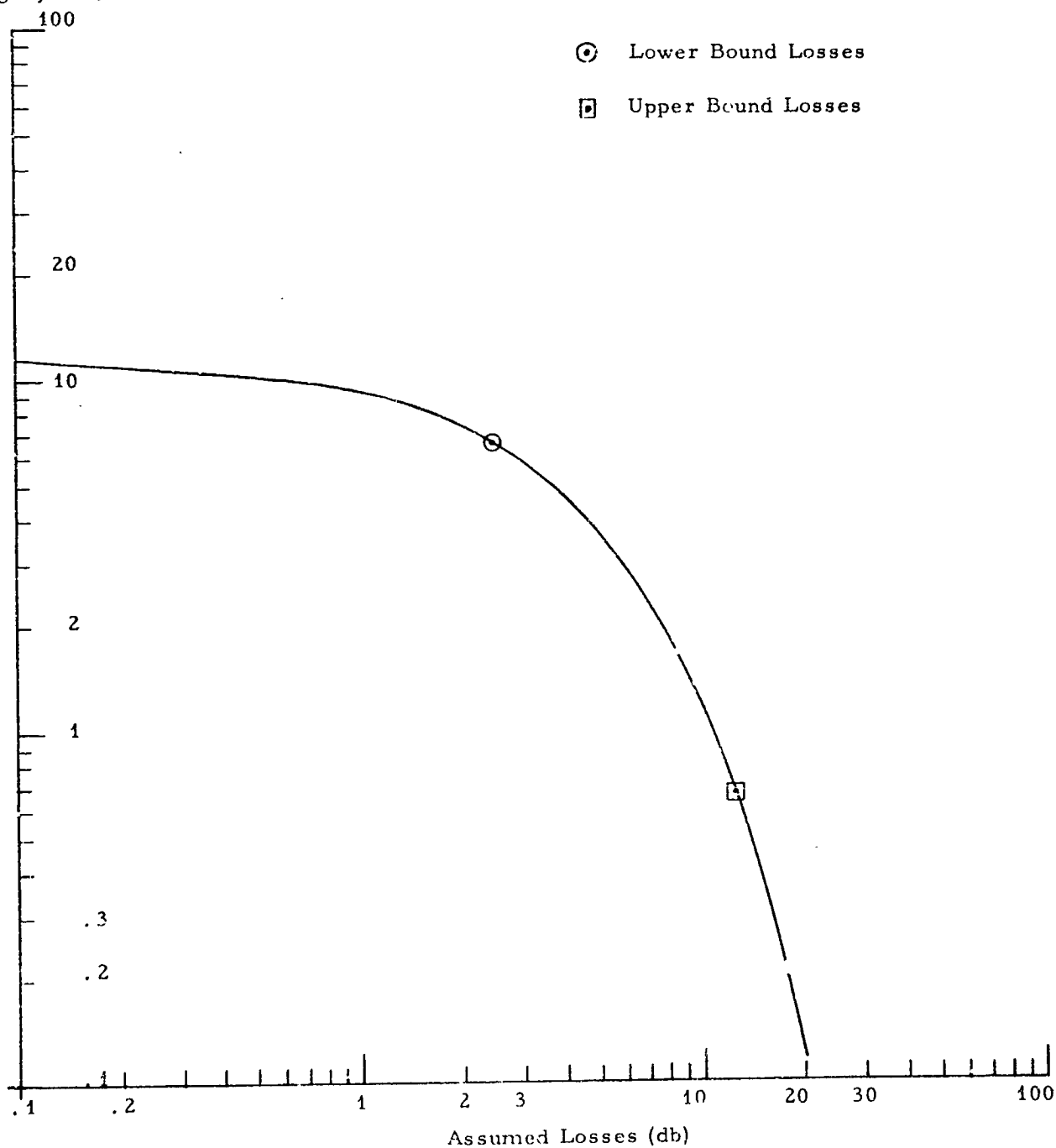


FIGURE 3.3-1 A PLOT OF BANDWIDTH AS A FUNCTION OF LOSSES
FOR CONDITION II

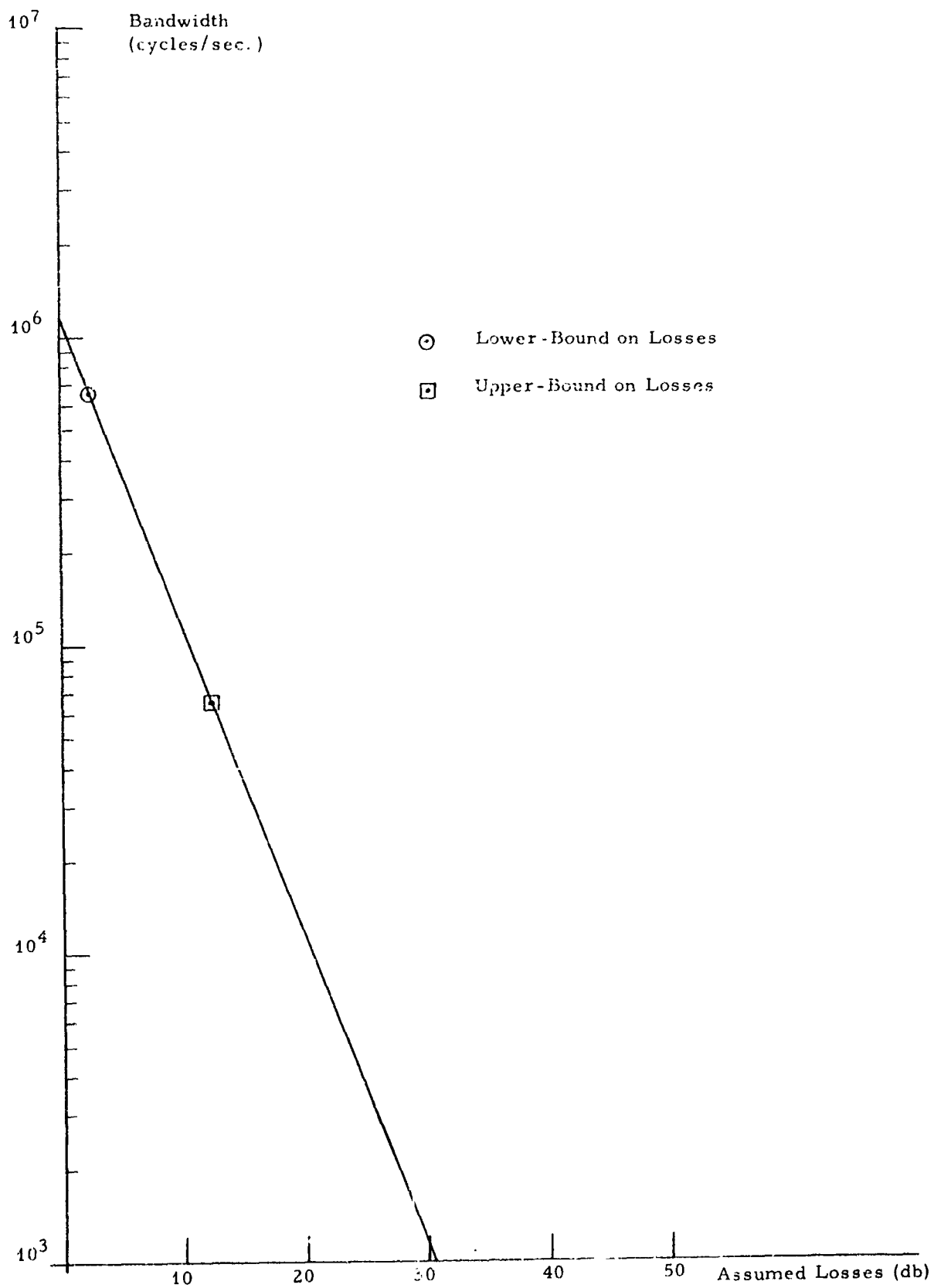


FIGURE 3.3-2 PLOT OF BANDWIDTH AS FUNCTION OF LOSSES, CONDITION II

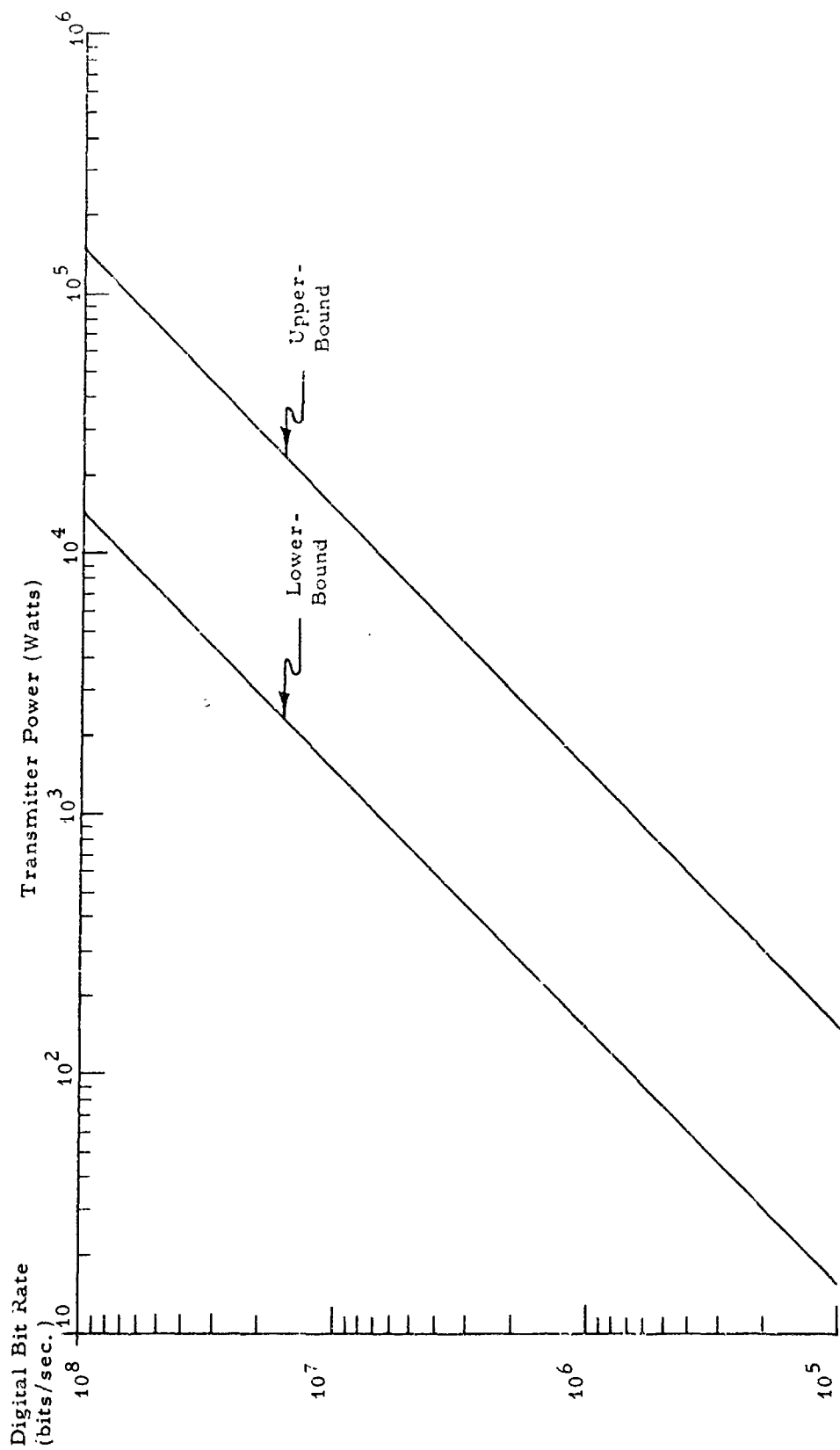


FIGURE 3.3-3 A PLOT OF TRANSMITTER POWER AS A FUNCTION OF DIGITAL BIT RATE FOR CONDITION II FOR UPPER AND LOWER BOUNDS ON LOSSES

$$D = R^2 f b. \quad (13)$$

For illustration, assume that the coding used is previous-element coding that encodes each element on the average with two bits. Therefore, using this and substituting Equation 13 into Equation 19, the transmitter power required is

$$P_t = 3.096 R^2 f \times 10^{-4}, \text{ watts.} \quad (20)$$

Equation 20 which is the transmitter power as a function of the resolution and frame rate is given by Figure 3.3-4.

Although Figure 3.3-4 gives sufficient information concerning frame rate and transmitter power at various resolutions, additional information can be achieved by plotting transmitter power as a function of frame rate for various encoding techniques. Therefore, Equation 13 is rewritten in terms of variable-encoding bits at a fixed resolution $R = 256$ and is given by

$$P_t = 10.13 f b, \text{ watts.} \quad (21)$$

Therefore, Equation 15 is plotted in Figure 3.3-5 for previous-element coding, PCM with Roberts, 4-bit-PCM, and 6-bit PCM coding. This curve is extremely useful to illustrate the savings in power due to coding. For example, assuming 30 fps, 6-bit PCM requires 1824 watts while previous-element coding requires only 608 watts. Assuming a transmitter efficiency of 20%, the savings in prime power for PEC is 6080 watts.

However, assume that the transmitter power is fixed at 1000 watts. It is possible to obtain the relationship between frame rate, resolution, and coding bits per element.

Assume again that the horizontal and vertical resolutions are equal and equal to R . From Equation 9 the relationship between the three parameters is given by

$$6.46 \times 10^6 = R^2 f b. \quad (22)$$

Therefore, Equation 22 is plotted in Figure 3.3-6, which plots the frame rate as a function of the resolution parametric in f . Therefore, assuming 600-line resolution, the maximum frame rate varies from 2.99 fps at 6 bits per element to 8.97 fps at 2 bits per element. Here, it is assumed that the transmitter power is fixed at 1000 watts; however, it is necessary to minimize the power, which was pointed out in Figure 3.3-5.

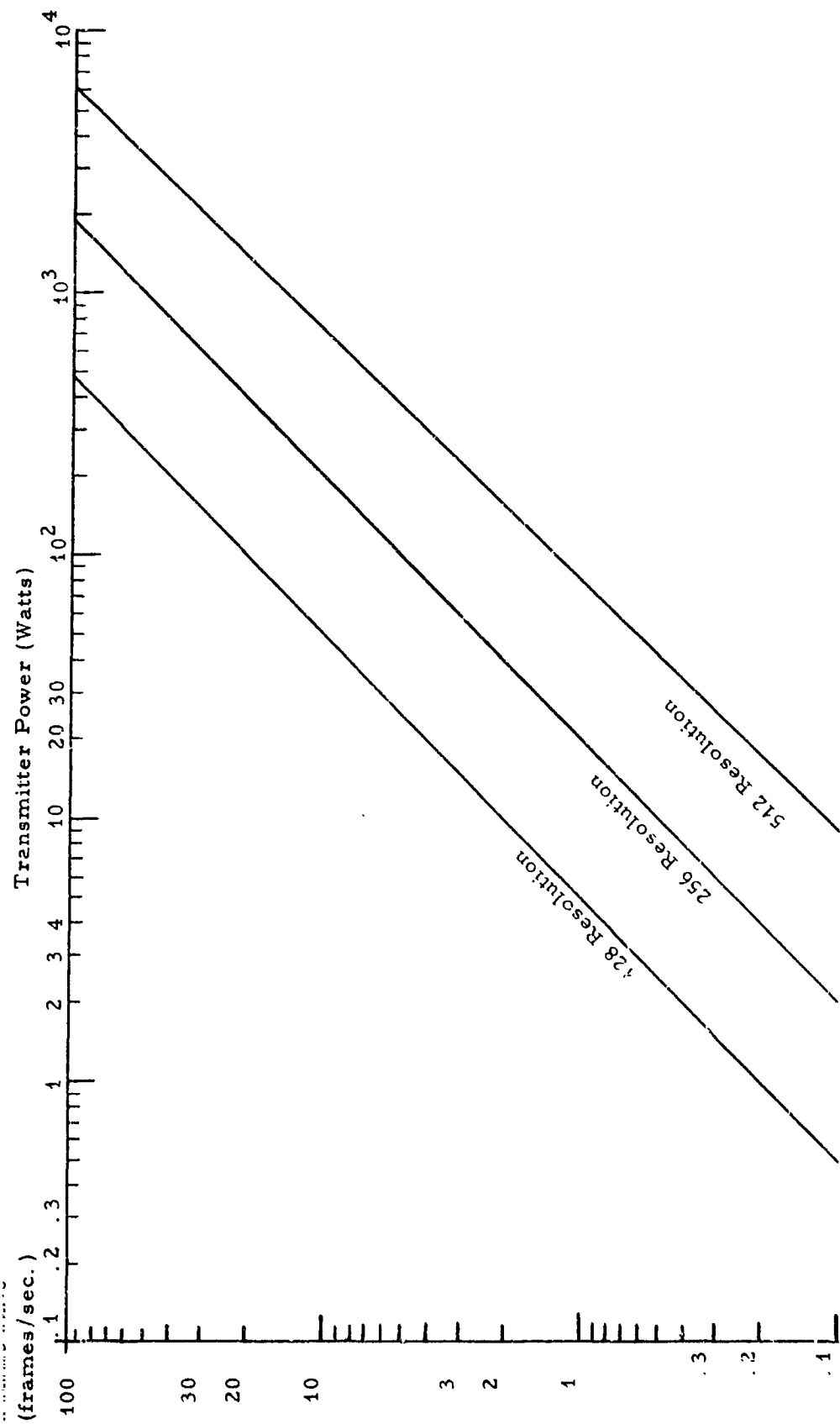


FIGURE 3.3-4 PLOT OF TRANSMITTER POWER AS A FUNCTION OF FRAME RATE FOR PREVIOUS ELEMENT CODING FOR CONDITION II AT VARIOUS RESOLUTIONS

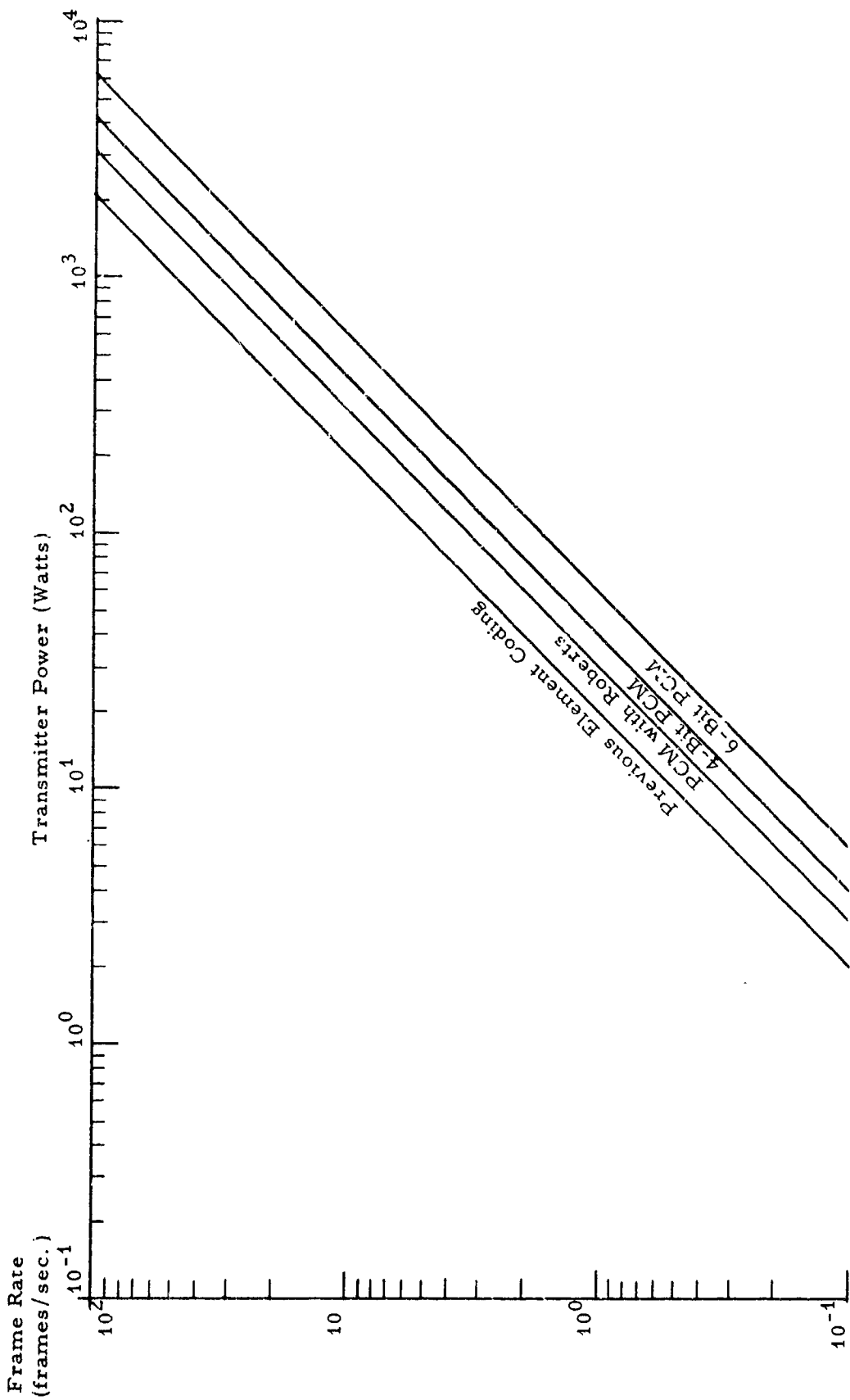


FIGURE 3.3-5

A PLOT OF TRANSMITTER POWER AS A FUNCTION OF FRAME RATE FOR VARIOUS CODING TECHNIQUES FOR CONDITION II AT 256 RESOLUTION

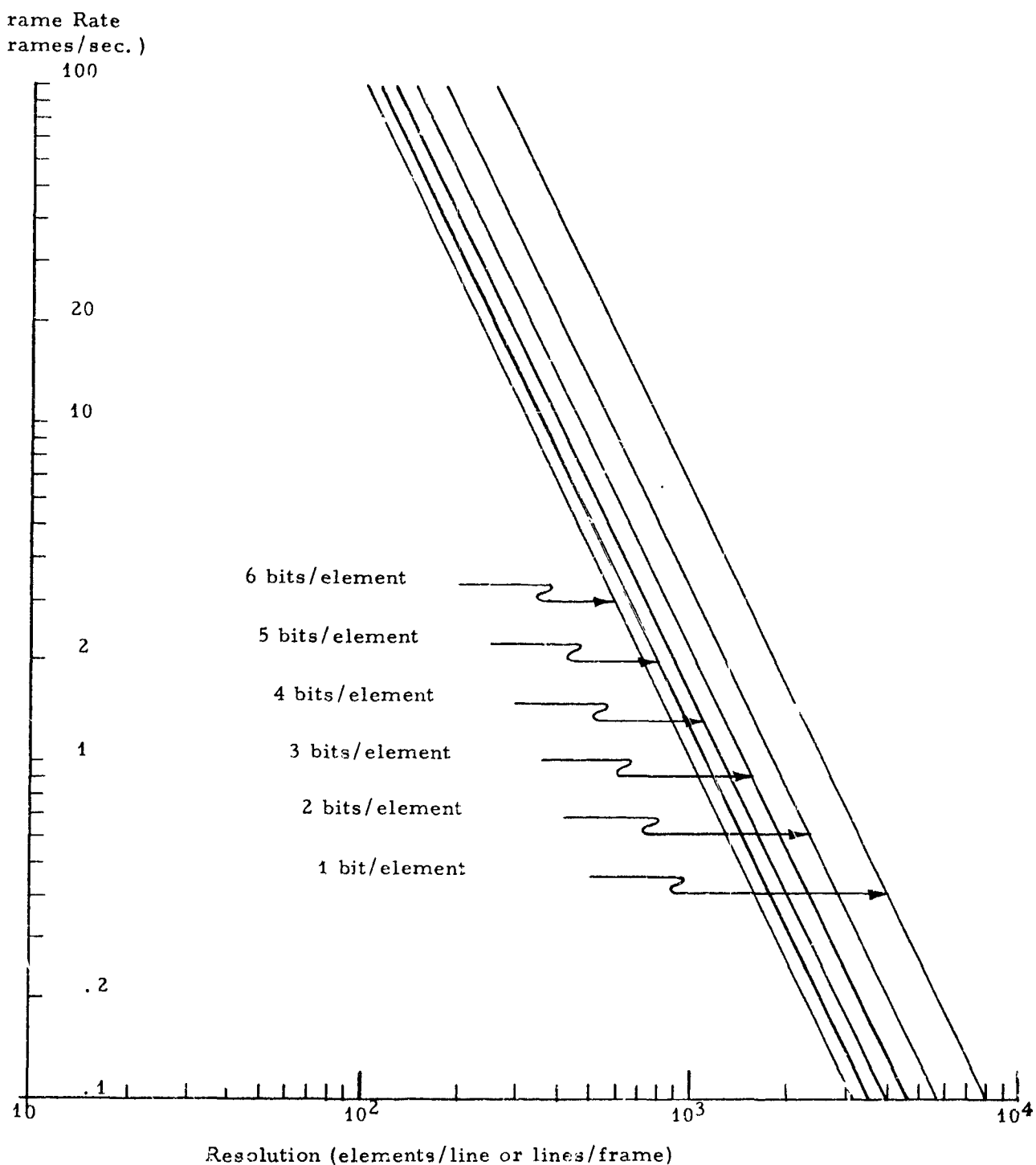


FIGURE 3.3-6 A PLOT OF FRAME RATE AS A FUNCTION OF RESOLUTION (ASSUMING HORIZONTAL RESOLUTION EQUALS VERTICAL RESOLUTION) FOR CONSTANT BITS/ELEMENT FOR CONDITION II OF MSC AT 1000 WATTS TRANSMITTER POWER FIXED.

3.4 NOMENCLATURE

P_r	=	Received power, watts
P_t	=	Transmitter power, watts
P_t'	=	Net transmitter power, watts
G_1	=	Ground antenna directivity
G_s	=	Spacecraft antenna directivity
λ	=	Wavelength
R	=	Range
P_l	=	Path loss (or free space transmission loss)
S/N	=	Signal-to-noise ratio
N	=	Noise power
k	=	Boltzmann's constant
B	=	Bandwidth
P_b^e	=	Data bit error probability
L_p	=	System design pad
L_m	=	Miscellaneous spacecraft losses
L	=	Total losses for system
D	=	Digital bit rate
f	=	Frame rate
R	=	Horizontal and vertical resolution
b	=	Coded bits per element

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SECTION 4

STORAGE MEDIA COMPARISON

4.0 EXECUTIVE SUMMARY

A consideration of total recording system weight as a function of time leads to the conclusion that for manned interplanetary missions a videotape recorder system is to be chosen in preference to a rapid film processor as an on-board TV program storage medium. A survey of the current and estimated future state of the art in both the videotape recorder field and the rapid film processor field was conducted with the conclusion that although the initial weight of the rapid film processor is less than the initial weight of the videotape system, the growth in film weight to keep up with the videotape system with a reasonable storage medium soon overbalances the total system weight in favor of the videotape system. Characteristics of 18 systems are presented in tabular form and a numerical example is given to quantify the above conclusions.

4.1 INTRODUCTION

Section 4 considers the question of choosing between two types of storage media, rapidly developed film and video tape, for the case of Condition 2 of Appendix II of the RFP. This case is the interplanetary case. Before going into this question on the basis of the limited state-of-the-art predictions available some comments are in order.

The first factor bearing on this study is the extremely limited response (with one notable exception) on the part of the manufacturers contacted. Unless one is willing to enter into detailed negotiations relative to a specific system requirement (which is beyond the scope of this study) one can obtain almost no reliable estimate of the trends in size, weight, and power. Therefore, the data on which this discussion is based are limited to personal industry contacts, current product data sheet, and some reports.

Considering the problem from a purely theoretical standpoint the weight of any system can be represented by

$$W = W_0 + ct \quad (1)$$

where W is the total weight committed at time t , given an initial weight W_0 and recording consumption (film, liquids, tape, reels, etc.) of c weight per unit time. If the tape recorder can transmit its data to earth every T seconds and then start recording again, its maximum weight is limited and given by

$$W_{\max} = W_0 + cT \quad (2)$$

However, the film recorder must use (and the spacecraft must carry) enough film to cover the entire picture taking time, regardless of the opportunities to transmit a portion of the data to earth. The weight of film recorder is therefore given by

$$W'' = W_o'' + c'' t \quad (3)$$

and there is no maximum limiting weight. Functions 2 and 3 are graphically illustrated in Figure 4.1-1, and it is readily seen that no matter how much smaller W_o'' is than W_o' , at some time t_1 the photographic storage system will outweigh the videotape system. Before making an actual system choice, one must obtain an estimate of W_o' , W_o'' , c' , c'' , T and also the total length of time t_2 that the television sensor will be in operation during the mission time, t_m . In general,

$$t_2 \leq t_m \quad (4)$$

and if $t_1 < t_2$ choose the tape recorder; if $t_1 > t_2$ choose the film recorder.

The argument just given is simplified in the interest of presentation, but it could equally well apply to total volume or total energy consumed as well as to total system weight. There are also additional considerations such as reliability, maintenance, radiation resistance, storability of photomaterial and chemicals, possible contamination of spacecraft interior by photochemicals, and probably many others.

Another aspect of the problem is that on any long, manned exploration, cameras and photoprocessing equipment of one sort or another will be included for its own sake in the table of organization and equipment. Therefore, it seems desirable to have on board the spaceship a TV film pickup capability. However, to have 100% film storage in the basic TV loop is open to serious question on the basis of the large amount of film expected to be used. Actually, the study of the film versus videotape question is worthy of a six-months' study in itself, and such study should be related to a relatively well-defined interplanetary mission wherein t_m and t_2 have been reasonably well determined.

4.2 VIDEOTAPE RECORDING

Small videotape recorders on the market today are made by two companies: Ampex, Redwood City, California, and Machtronics, Mt. View, California. The basic recorder characteristics are given in Table 4.2-1. We appreciate the personal assistance of Mr. David C. Thompson of Machtronics, Inc. in obtaining a look into the future trends of videotape recording. His comments follow:

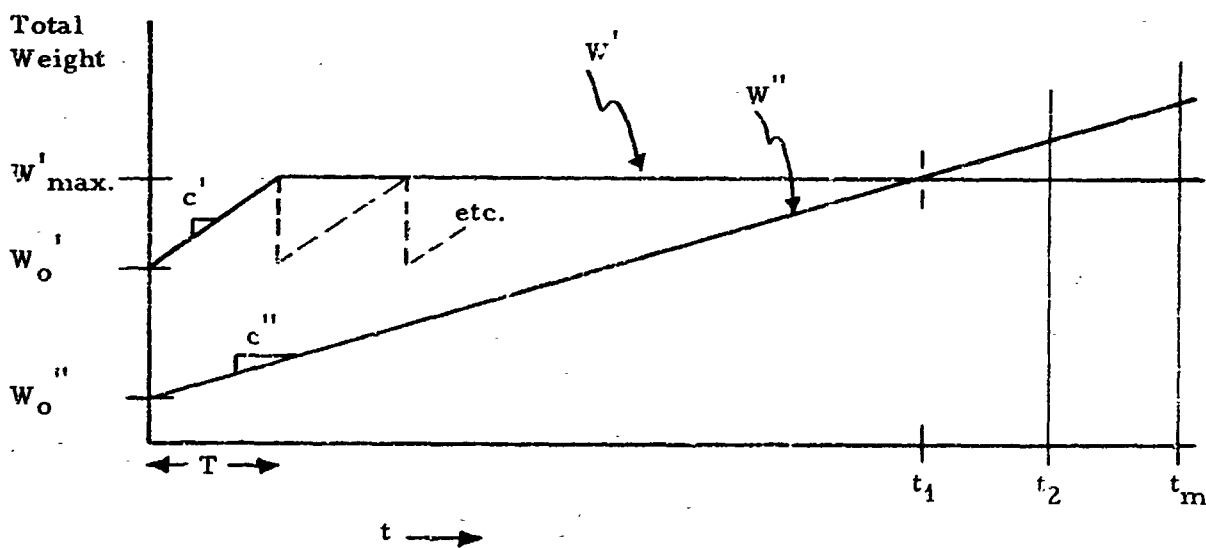


FIGURE 4.1-1 STORAGE WEIGHT FUNCTIONS

Volume of tape
for 1 hour -
.001" tape in
cubic inches

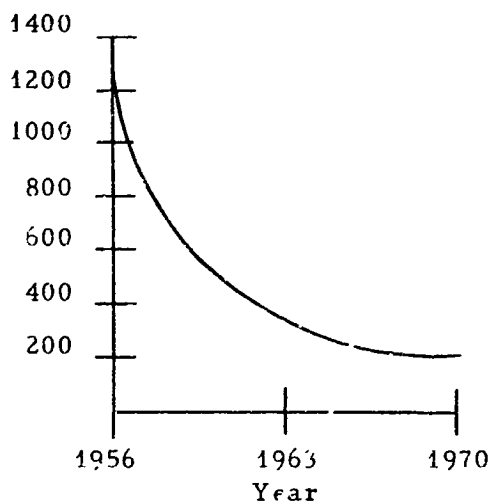


FIGURE 4.2-1 GRAPH OF TAPE VOLUME REQUIRED FOR ONE HOUR OF TV RECORDING

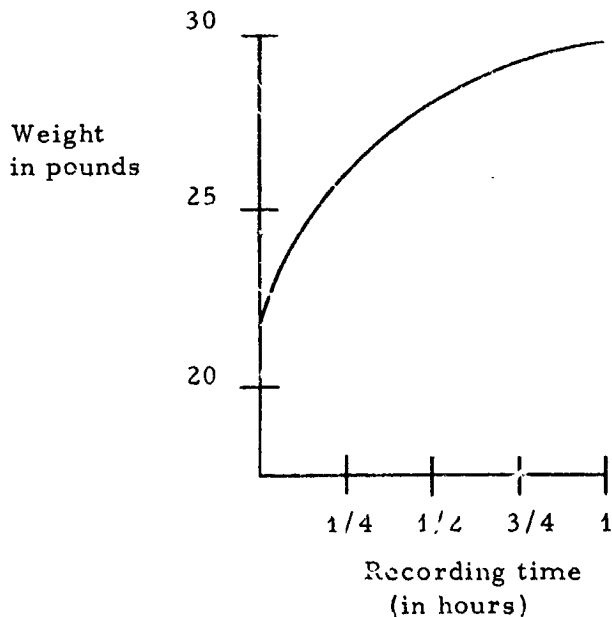
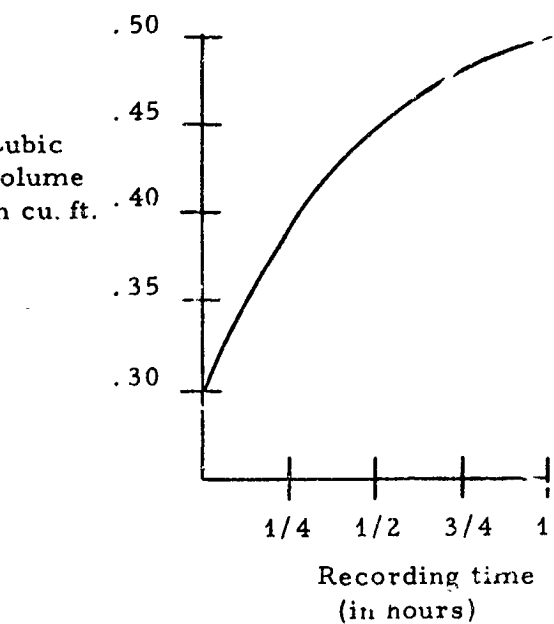


FIGURE 4.2-2 GRAPH OF ESTIMATED VOLUME AND WEIGHT OF A ONE-HOUR TV RECORDER

TABLE 4.2-1 SOME VIDEOTAPE RECORDER PARAMETERS

Manufacturer	Model	Size (cu ft)	Weight (lb)	Power (watts)	BW (mc)	Recording Time (hours)
Ampex	AR-300	3.5	175	?	4	1
Ampex	VR-1500	4.4	120	200	4	1
Mac Electronics	MVR-10	2	65	350	3.5	1.6
Machtronics	MVR-?	0.5	30	?	3-4	1

TABLE 4.2-2 STATISTICS ON TAPE PARAMETERS

Reel Diameter (inches)	Length of Tape (ft)	Tape Speed (IPS)	Tape Width (inches)	Volume/ Reel (cu. in.)	Thickness (inches)
14	4800	15	2	1230	0.001
10-1/2	2400	7-1/2	1	345	0.001
8	1200	3-3/4	1	200	0.001

"General Comments Regarding Advancement in the
State of Television Tape Recording Through 1970

1. Tape will be 1/2 mil (0.0005 inch) thick.
2. Tape speed will be 3-3/4 ips.
3. Helical scanning will be used.
4. The reel volume and size will be a minor over-all factor in recorder.
5. Writing speed will drop down to maybe 350 to 400 ips. (This will still have 3- to 4-mc. range.)
6. For strict military use, the size may go down to even 0.35 to 0.40 cu ft. Weight will be about the same (30 lb minimum). Here the tape would be built into unit instead of on standard removable reels."

Table 4.2-2, from Machtronics, gives some statistics on tape parameters. Figure 4.2-1, from Machtronics, estimates the tape volume required for one hour of TV recording as a function of year. Figure 4.2-2, from Machtronics, estimates the weight and volume of a one-hour TV recorder. Finally, Figure 4.2-3 shows the configuration of a possible videotape recorder as projected by Machtronics.

State-of-the-art figures may be taken as 30 lb., 0.5 cu. ft., and approximately 200 watts power for a one-hour, 3- to 4-mc., videotape recorder developed for space use in the later 1960's. The MSC-recommended system parameters of 256-line resolution, square format, approximately 15 frames per second and compression result in a digital bit rate on the order of 2 mc. By using a subcarrier this digital signal may be placed on the one-hour, 4-mc. tape recorder.

4.3 RAPID FILM PROCESSING

Data in the field of miniaturized, airborne, rapid-processing systems is even harder to obtain than equivalent tape-recorder data. Much information is published about the sensitometric qualities of the various rapid-processing techniques, and a fair amount of information may be found on the methods by which the photochemicals are applied to the film. However, for miniature systems, information on over-all system size, and in some cases weight, must be inferred from the illustrations of typical systems now in existence.

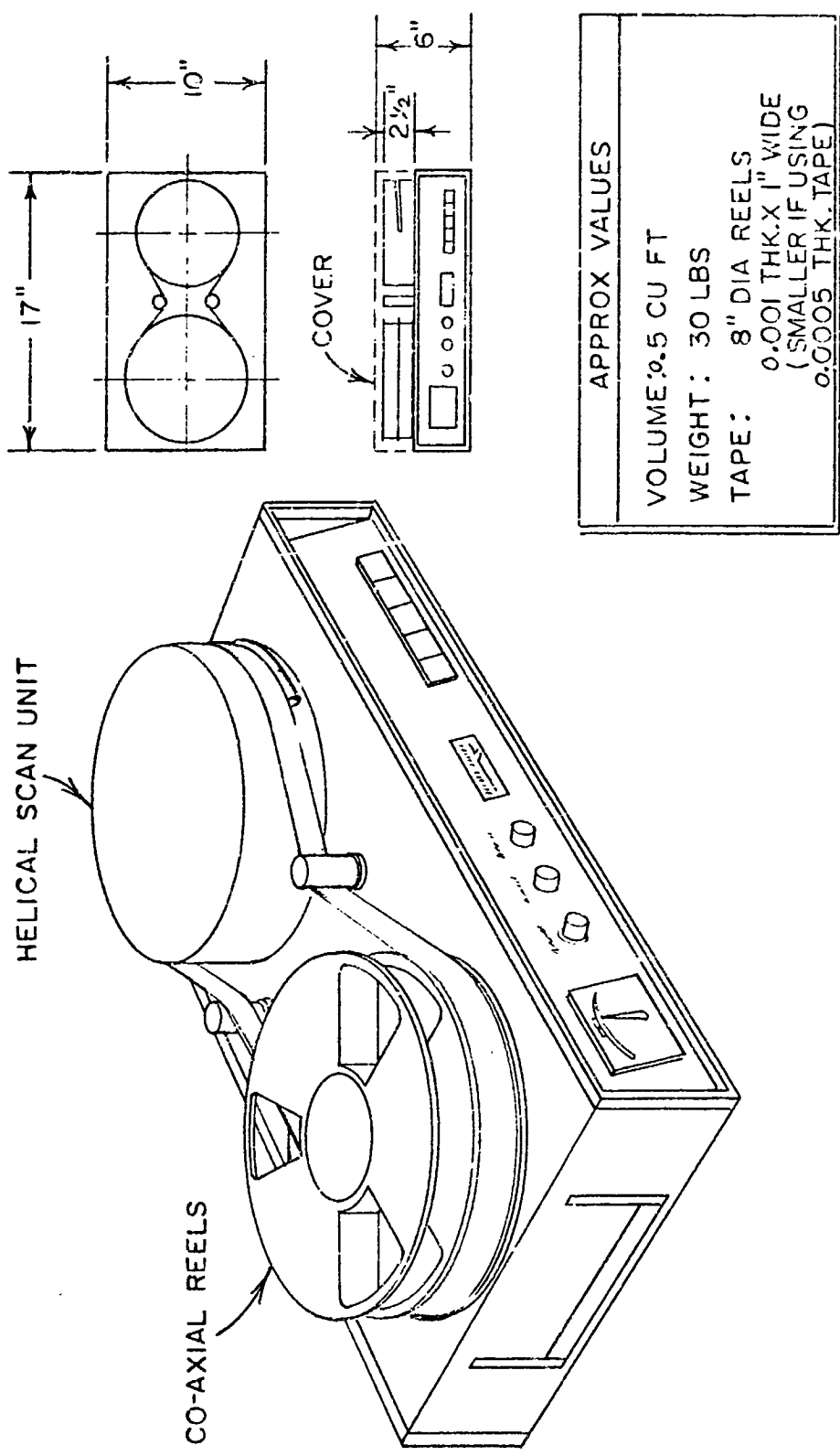


FIGURE 4.2-3 PROJECTED CONFIGURATION OF A FUTURE VIDEOTAPE RECORDER BY MACHTRONICS, INC.

Almost no information is commonly available on power consumption. Nevertheless, Table 4.3-1 summarizes the data available and inferred. Size, weight, and power data on the larger, commercial-type processors is easier to obtain from data sheets, but even though some of it is included in Table 4.3-1, it is of little use in inferring ultimate size, weight, and power miniaturizations.

From the table it can be concluded that for complete (tanks, film, reels, chemicals, etc.) airborne systems approximately one cu. ft. of volume and 25 lb. of weight should be allowed. However, for the processing station alone, exclusive of reels and tanks, on the order of 0.01 cu. ft. and 1 lb or less is needed. Considering that the data in Table 4.3-1 are for film sizes of 16-mm or larger, it appears reasonable to assume that an 8-mm processor could be made for a couple of pounds with reel drives and miniature pumps included. Reels, film, tanks and chemical weight are extra "expendables" to be averaged on a per-frame basis.

4.4 COMPARISON AND CHOICE

Consideration of parametric values from the preceding sections allow the placement of typical values on the quantities W_o' , W_o'' , and c'' defined in Section 4.1. Let

$$W_o' = 30 \text{ lb} \quad (5)$$

$$W_o'' = 2 \text{ lb} \quad (6)$$

$$c'' = 10 \text{ mg per frame} \quad (7)$$

The 10 mg per frame is derived from measuring a strip of 8-mm home movie film and finding its weight at about 5.5 mg. per frame. Photomechanisms, Inc. indicate that chemical consumption can be as low as 0.1 cc per square inch of film. Assuming a chemical specific gravity close to unity, this works out to about 3 mg per 8-mm frame. Finally, allow approximately 0.5 mg per frame for the fluid tanks (7:1 mass-ratio of tank plus fluid to tank) and 1 mg per frame for reels (6.5:1 mass-ratio of loaded reel to empty reel).

The film weight nomogram in Figure 4.4-1 yields a quick estimate of expendable weight c'' . Since the difference between W_o' and W_o'' is 28 lb (12.75 kg.) and since the MSC frame rate is 15 per second, 80,000 (~ 1 day) of picture taking will result in a rapid-processing system having an over-all weight equal to the videotape recorder. Any amount of total picture-taking time in excess of 24 hours or so will result in a weight balance in favor of the video tape. Since mission times of 400 days and such are being considered for the manned Mars missions, and since two hours a day TV for earth transmission is also a reasonable (probably low) assumption, it would appear that putting a rapid processor in the

TABLE 4.3-1 SOME RAPID FILM PROCESSOR PARAMETERS

MANUFACTURER	MODEL	SIZE	WEIGHT	POWER	REMARKS	REF.
Chicago Aerial	KA-30	1 cu. ft. *	25 lbs. *	?	1960 - Airborne on F-101	4.3-1
Chicago Aerial	70 mm	0.1 cu. ft. *	5 lbs. *	?	1962 - (Ground) Portable - 250' of 70 mm film	4.3-2
Fairchild	F-316A	4.5 cu. ft.	65 lbs.	1,335 W	1962 - (Ground) 400' of 16 mm film	4.3-3
Fairchild	70 mm	1 cu. ft. *	24 lbs.	<400 W	1961 - (Airborne) 1000' of 70 mm film	4.3-4
Fairchild	35 mm	<0.1 cu. ft. * (Less Reels and Dryer)	5 lbs. *	?	1956 - (Airborne) 1000' of 35 mm film	4.3-5
Kelvin and Hughes	RP 3	10 cu. ft.	250 lbs.	1,500 W	1962 - (Ground or Ship-board) Cont. Processing 16 mm film	4.3-6
Kelvin and Hughes	RP 3	22.5 cu. ft.	600 lbs.	?	1962 - (Ground or Ship-board) Cont. Processing 35 mm film	4.3-6
Optomechanisms	380	~1.7 cu. ft.	40 lbs. *	?	1961 - (Ground and Ship-board) Includes Camera and Projector	4.3-7
Photomechanisms	904	3.1 cu. ft.	<40 lbs.	?	1962 - (Ground) 16, 35 or 70 mm film	4.3-8
Photomechanisms	?	0.2 cu. ft. *	<10 lbs *	?	1961 - Report on Generic Systems	4.3-9

TABLE 4.3-1 SOME RAPID FILM PROCESSOR PARAMETERS (continued)

MANUFACTURER	MODEL	SIZE	WEIGHT	POWER	REMARKS	REF
Photomechanisms	?	<0.01 cu. ft.* (No Tanks)	1 lb.*	?	1962 - (Station Only) 35 mm "RAPIDATA"	4.3-10
Photomechanisms	?	~0.1 cu. ft.*	<10 lbs.	?	1962 - 35 mm Miniature Processor-Viewer	4.3-10
Photomechanisms	?	~0.03 cu. ft.*	<10 lbs.	?	1962 - 70 mm Processing Station Only	4.3-10
Photomechanisms	?	<1 cu. ft.	25 lbs.*	?	1962 - (Ground - Experimen- tal) Complete 35 mm Dark- room	4.3-10

* Estimated from illustrations.

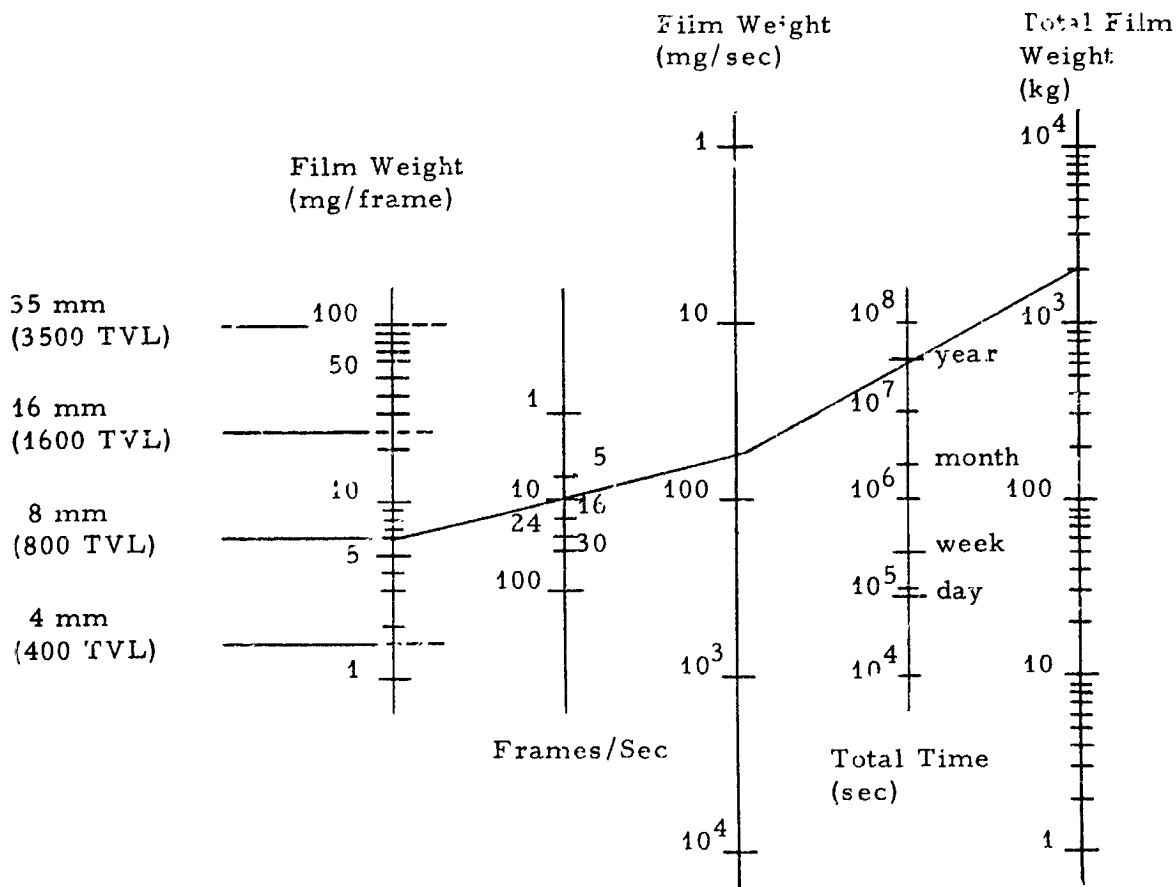


FIGURE 4. 4-1

FILM WEIGHT NOMOGRAM

spacecraft TV-to-earth loop for temporary program storage purposes could result in a total system weight some 30 times that of the videotape recorder.

It is concluded that a videotape unit is the preferred spacecraft TV program storage medium. It is also concluded, however, that a small, rapid processor should be included in the spacecraft equipment roster for its proper and very useful place of processing exploration movies of short duration but high value.

4.5 REFERENCES

1. Chicago Aerial Industries, Inc., Rapid Processor for the KA-30 Camera.
2. Chicago Aerial Industries, Inc., Product Information, Rapid Processing, Report I, Technical Discussion Characteristics Applications.
3. Fairchild Camera and Instrument Corporation, Industrial Products Division, Fairchild Automatic Film Processor Model F-316A.
4. Fairchild Camera and Instrument Corporation, Defense Products Division, Viscous Techniques for Airborne Processing.
5. Fairchild Camera and Instrument Corporation, Reconnaissance Systems Division, Airborne 35-mm Roll Film Processor Final Engineering Report No. 1 on Contract No. AF 33(616)-3695.
6. Kelvin and Hughes, Ltd., Kelvin Hughes Photographic Processing and Recording Equipment.
7. OPTOmechanisms, Inc., Standard Corporate Catalog.
8. Photomechanisms, Inc., Tentative Specifications for RAPIDATA Model 904 Tabletop Photographic Processor.
9. Photomechanisms, Inc., An Advanced Technique for Short Delay Processing of Photographic Emulsions.
10. Photomechanisms, Inc., RAPIDATA Photographic Processors.

SECTION 5

GROUND DATA SCAN CONVERSION AND RECORDING

5.0 EXECUTIVE SUMMARY

Operation of the ground station post-detection demodulation equipment portion of the compressed digital TV system has been described in Section 1.1.2. The present section describes those portions of the ground station equipment concerned with scan conversion and recording of the spacecraft demodulated video data.

The recommended method of scan conversion is the TV display-to-TV camera technique. This method has been successfully employed in commercial TV broadcast practice for some years in converting between U.S. and European standards. This equipment will transform the asynchronous 15 frame per second, 256-line spacecraft video data into synchronous video data operating with the standard broadcast scan format of 60 fields per second, 525 lines interlaced.

Standard high-quality kine-photo and film-processing equipment is then used to provide permanent film data records. Broadcast-type magnetic-video-tape recording equipment will provide storage for the video data to be transmitted to the commercial TV networks for rebroadcast.

5.1 SCAN CONVERSION SYSTEM

The block diagram of the scan-converter system is shown in Figure 5.1-1. The function of this equipment is to convert from the asynchronous 15 frame per second input video data format to the 60 field per 30 frame second commercial TV scan standards.

The video data from the Roberts decoder unit located in the compressed digital TV demodulation equipment (see Section 1.1.2) is fed to the automatic gain control (AGC) amplifier. This circuit automatically maintains a constant peak-white-to-peak-black video signal amplitude to provide the desired range of modulation of the TV display scanning beam. This circuit is followed by the aperture corrector which inserts a zero-phase-shift amplitude boost to the video signal to compensate for the finite aperture of the display beam. This results in improved display horizontal resolution.

The gamma corrector which follows is a video amplifier with an adjustable non-linear transfer characteristic which provides compensation for image-brightness compression in the black region (primarily caused by light-dispersion effects in

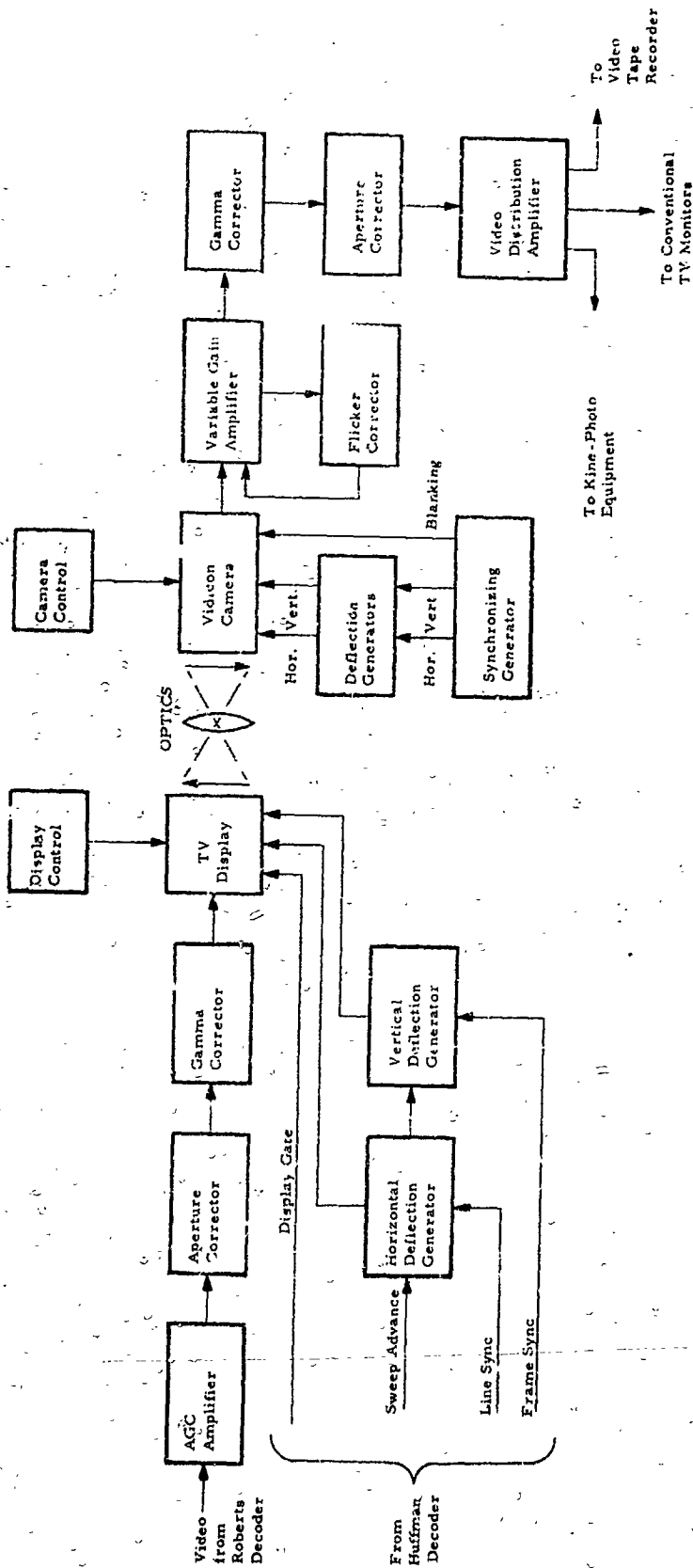


FIGURE 5.1-1 SCAN-CONVERSION SYSTEM

the converter optical system) and in the white region (primarily produced by display-tube saturation effects). The compensated video signal is fed to the TV display and amplitude-modulates the display beam during each display gating period (see Section 1.1.2 for a discussion of this technique which is related to the asynchronous operation of the compressed digital TV system).

Horizontal and vertical deflection voltages, which are locked to the appropriate synchronization output from the Huffman decoder, feed the display deflection plates. Electrostatic deflection is used to achieve the desired sweep response speed in the display.

Selection of the optimum display phosphor is intimately tied to the selection of the vidicon tube photoconductor type. The optimum signal storage and decay characteristics of these transducers will be determined on the basis of the best compromise between allowable signal carry-over from frame to frame (producing smear on moving images) and allowable brightness "flicker" or "shutter-bar" signal (caused by variations in display brightness as seen by the TV camera, due to differences between input and output frame rates).

The displayed image is optically coupled to the vidicon camera, which operates with the standard broadcast scan format. The camera video output feeds the variable gain amplifier, which, in conjunction with the flicker correction circuit* acts to reduce the above-mentioned flicker effect to an acceptable minimum. The flicker correction circuit consists of a peak white detector and low-pass filter combination which is "tuned" to the flicker frequency. The detected flicker voltage triggers a correction waveform generator, the output of which modulates the variable-gain amplifier in the proper phase to reduce the flicker component in the camera video signal.

This circuit is followed by gamma-correction and aperture-correction circuits which compensate for camera deficiencies in these particular areas. Finally, the video distribution amplifier provides multiple signal outputs which are fed to the data recording and display equipment.

5.2 ALTERNATE SCAN CONVERSION TECHNIQUES

Several other scan conversion methods have possible application here. The TV display-to-TV camera method discussed in the previous section is believed to be the best answer, since it makes use of relatively simple, established techniques and directly provides the desired video signal at broadcast standards. Other possible approaches which may, in certain special instances, offer advantages to the Manned Spacecraft TV System are discussed in the following sections.

*K. Belair Benson: "CBS Television Standards Conversion Techniques." Journal of the SMPTE. Vol. 70, 1961.

5.2.1 Recording Storage Tube Scan Converter

This conversion technique makes use of recording storage tube devices, such as produced by Raytheon (QK-685 single gun and QK-7702 dual gun types, for example).

These tubes operate in an electronic read-in/electronic read-out mode without the requirement for intermediate optical coupling as used in the previous system.

Presently, the principal use of equipment utilizing this type of tube is in radar-scan conversion applications. In these systems, the variable signal storage and erase characteristic of these tubes permit enhancement of moving targets, which, of course, is in addition to the principal tube function of converting from radar scan standards to the desired TV scan standards (to achieve greater operating ease and flexibility of the radar display complex).

For the present compressed digital TV system application, video conversion is required from 15 frames per second to the 60-field-per-second rate. The erase characteristic of the storage tubes which works to their advantage in the radar case becomes a disadvantage for this TV system. Due to the time required to erase the storage surface of previous data and "prime" it in preparation for a new frame of video data (this time is typically one to several tenths of a second), and with the requirement to continuously convert from a low to a higher frame rate, it becomes necessary to use three storage tubes operated in a time-sequencing arrangement. While the first tube is reading out data at the desired new rate, the second tube (which has just been erased and primed during the previous frame) is writing in the current data at the original rate, and the third tube (having just completed the "read" cycle) is being erased and primed. Thus, a fairly complex, though feasible system, results from having to meet the Manned Spacecraft TV System requirements.

One potential advantage which this technique offers is complete elimination of the "shutter bar" or "flicker" effect previously described. Should this prove to be a serious problem with the display-to-camera conversion approach, the storage tube type system could be adopted.

5.2.2 Magnetic Devices Scan Conversion

This section is included to indicate the possibility of using addressable, magnetic devices (cores, drums, perhaps discs) as scan-conversion devices. In this case scan conversion is taken in the context of the MSC ground equipment whose purpose is to convert from a low (~ 15 fps) to a high (~ 60 fps) frame rate. Basically, this operation entails sequentially reading out in a nondestructive manner the store at a high rate while simultaneously sequentially reading into the store in a "write-over" mode the new (updating) information at a low rate.

Let the read address $a_r(t)$ be given by

$$a_r(t) = \text{rep } T_1(k_1 t), \quad (1)$$

and the write address $a_w(t)$ be given by

$$a_w(t) = \text{rep } T_2(k_2 t), \quad (2)$$

where $\text{rep } T(\)$ is read "repeat function () with period T ", and where k_1 and k_2 are constants. Further, let

$$k_1 = A k_2, \text{ and} \quad (3)$$

$$T_2 = A T_1. \quad (4)$$

These relations are shown in Figure 5.2.2-1 for the case of $A = 4$ where A is the ratio of read-out to read-in speed. Actually, the functions $a_r(t)$ and $a_w(t)$ are point functions defined only at specific instants of time $t_i = t_0 + \Delta i$, $i = 0, 1, 2, \dots$. From Figure 5.2.2-1 it can be seen that during one read-in period T_2 the following relation

$$a_r(t) = a_w(t)$$

occurs four times, at points P_1, P_2, P_3 , and P_4 . At that time the information read out is the same as the information being read in, and read-out is switched to read the input information straight through to the display.

In order to ascertain the demands placed on the storage device by this scan conversion system, consider the following parameters:

$$T_1 = \Delta \ell_1 \quad (6)$$

and

$$T_2 = A \Delta \ell_1 \quad (7)$$

where

$$\Delta = 0.25 \text{ } \mu\text{sec},$$

$$\ell_1 = 256 \times 256 = 65,536,$$

$$A = 4.$$

Thus, $1/T_1$ is about 61 frames per second and $1/T_2$ is approximately the 15-frames-per-second average for the MSC system. These parameters place upon the store a nondestructive read-out time of $0.25 \mu\text{sec}$ and a write-over read-in cycle time of $4 \times 0.25 = 1 \mu\text{sec}$. The total storage capacity is 65,536 q bits where q, the quantization level, is recommended as 4 in the MSC system. In order to

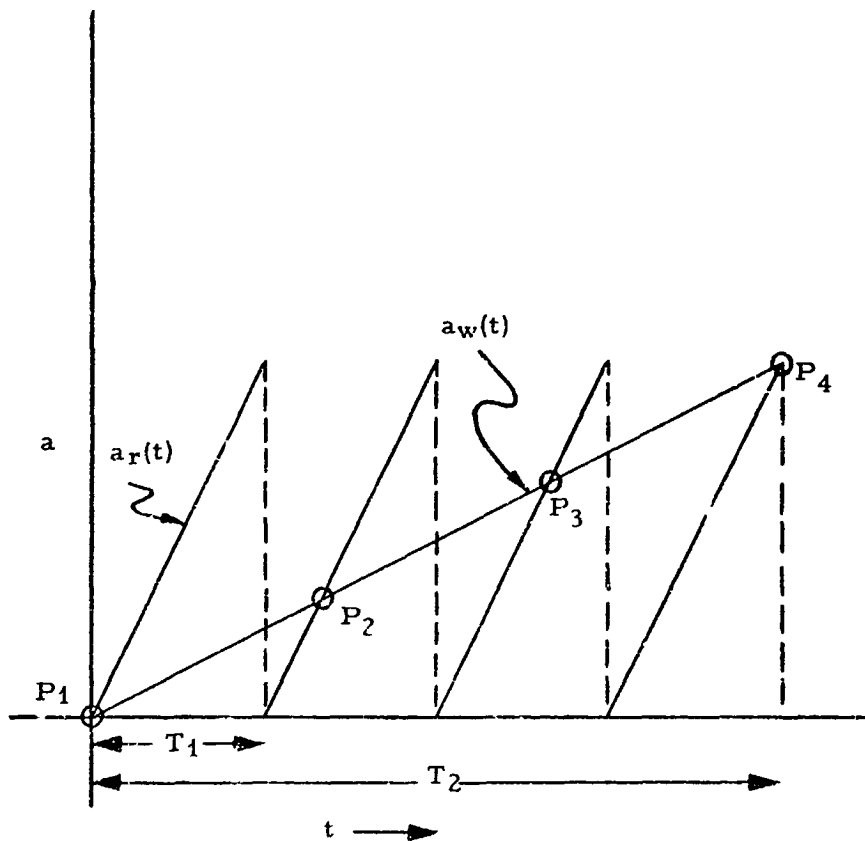


FIGURE 5. 2. 2-1 SCAN-CONVERSION FUNCTION

be economically competitive with the \$25K to \$100K storage-tube scan-conversion systems the complete storage unit must become available for about 10 cents per bit. Current prices (Burroughs) for high speed ($< 1 \mu\text{sec}$) ferrite storage run about 50 cents per bit for the storage alone, exclusive of address circuitry.

Perhaps drum storage is competitive and a drum rotation speed of 3600 rpm (60 per second) is certainly compatible with both existing drum technology and display frame rate characteristics. Drum read-out at current technology is in the one to five μsec range (example+ 180 kb rate for Bryant drums) so there exists a distinct problem of securing 0.25 μsec read-outs. This problem could be approached from the standpoint of staggered read heads side-by-side, displaced around the drum periphery by one bit each and read in overlapping sequence. Since one bit is on the order of $1 \text{ to } 3 \times 10^{-3}$ inches around a circumference of perhaps 30 inches, the head displacement and alignment problem is a formidable one.

In spite of the above-mentioned mechanical and economical difficulties, it is nevertheless instructive to return to the basic mathematical definition of a function to be performed and then to proceed from this point into alternate technologies to see if their current state-of-the-art is capable of supporting the required parametric values. It is with this spirit that this somewhat speculative discussion of scan conversion is offered.

5.3 DATA RECORDING

Due to the asynchronous nature of the spacecraft camera scan (i.e., variable element, line, and frame rates, which change as a function of picture statistics), the demodulated video signal from the A to D converter is not directly amenable to recording by conventional kine-photo or magnetic-video-tape techniques. This is because of the problems involved in asynchronous operation of film and tape transport mechanisms. Considerable simplification of the data recording processes would result from operation with synchronous video data employing commercial broadcast scan standards. This is the approach adopted in this study program. The scan converter equipment described in Section 5.1 furnishes the output video signals with the desired format.

To obtain permanent data recordings for detailed analysis and for archival storage, the scan-converter video output is fed to standard kinescope film recording and processing equipment. Also, the video output feeds Ampex magnetic-video-tape equipment which provides a recording that may be easily and rapidly edited for later transmission over the commercial television networks.

SECTION 6

EXPERIMENTS IN DIGITAL TELEVISION

6.0 EXECUTIVE SUMMARY

EDITS is a digital television research tool with variable parameters capable of taking statistics on an uncompressed and compressed video source. Not only does it have the capability of obtaining subjective results of parameter variation, but it also has the capability of determining the subjective effects of bit errors in the channel. The parameter variations are:

- (1) The resolution--512, 256, and 128.
- (2) The frame rate at 512 resolution--19.2 (maximum) to 0.2 (minimum) frames per second.
- (3) The quantization--1, 2, 3, 4, 5, and 6 bits.

As a result of the subjective experiments performed on EDITS, the standard digital television parameters for a manned space application are (1) 256 resolution, (2) 15 frames per second, and (3) four-bit quantization. This results in a 12:1 over-all reduction in bandwidth by parameter reduction alone. This parameter reduction combined with an efficient coding technique results in an extremely efficient system.

Several of the more promising coding techniques have been simulated on EDITS, and when the statistics warranted it, the system was actually implemented on EDITS. It is then possible to combine the efficient coding techniques with the parameter reductions. The conventional PCM system was compared to a two-bit delta system and to the PCM system with Roberts and PE coding. Although the complexity of the delta system is much less than the Roberts system, the picture quality is not as good. The three-bit PCM system with four bits of Roberts pseudo-random noise and PE coding appears to be the most efficient system with a compression of 3:1. Although this comparison was performed independently of noise, the effects of noise in the channel can be simulated on EDITS.

For the uncompressed PCM system a 10^{-3} bit error probability will allow adequate rendition; however, this bit error probability must be less for the compressed systems utilizing efficient coding. Both the delta system and the 3-4 Roberts-PEC system are expected to require about the same bit error probability for a subjectively acceptable picture. The effects of errors in both systems is quite similar in as much that a single bit error will cause a streak in the displayed scene instead of the normal salt and pepper effect observed in noisy PCM. These conclusions are based not only on the subjective experiments performed

on EDITS, but also on the statistics obtained from EDITS.

For the 11 standard scenes the statistics are given in Section 6.2.4. These statistics include (1) first-order, (2) second-order, (3) PE and (1-PE), and run-length statistics. For the standard scenes the average entropy of the 3-4 Roberts-PEC source is 1.9 bits per element, which represents a true 3:1 coding reduction in bandwidth relative to a 6-bit PCM system.

6.1 EDITS

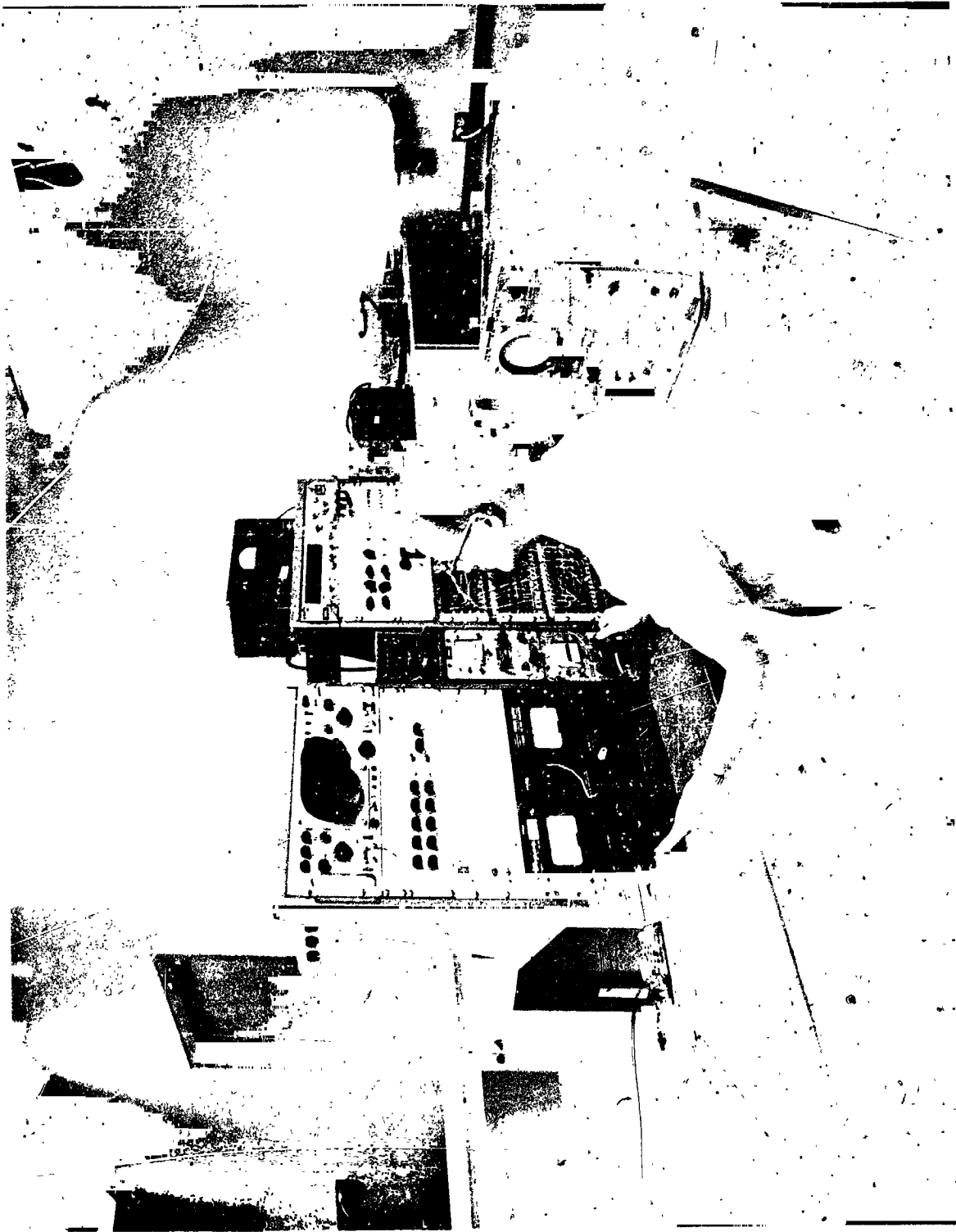
EDITS I (Experimental Digital Television System) is one answer to a long-standing need for a variable-parameter laboratory television system that is digital, highly flexible, and capable of taking information-theory statistics from an actual television video signal. Designed primarily as a tool for television compression research in earth reconnaissance and planetary exploration systems, EDITS I offers such a variety of parameter changes and combinations that its applications are practically unlimited. Potential uses range from a determination of the optimum design characteristics of high-data-rate systems to subjective studies of human reaction to resolution-limited surveillance systems.

The EDITS I is specifically designed to explore:

- o Effects of bit errors on PCM pictures
- o Redundancy elimination and efficient encoding
- o Picture (actual video) statistics
- o Display and conversion problems
- o Subjective picture parameters

The system is currently supporting on a daily basis a comparative study of television compression systems. In addition, it is being used to receive, reproduce, and statistically analyze pictures of weather, reconnaissance material, aircraft, line drawings, and other subjects sent to EMR from government agencies. In these applications EMR's system design has been characterized by flexibility, moderate cost, low engineering time requirements, and short time schedules. To meet the requirements of these programs, EDITS I was implemented with digital modules, thus enabling EMR to proceed virtually from logic diagrams directly to a working digital subsystem.

EDITS I, block diagrammed in Figure 6.1-1, is truly digital in the sense that the camera and display sweeps are staircase waveforms derived digitally from a master clock. The sweep speeds may be varied continuously over a 10-to-1 range, affording the television experimenter exceptional flexibility in studying



EDITS I

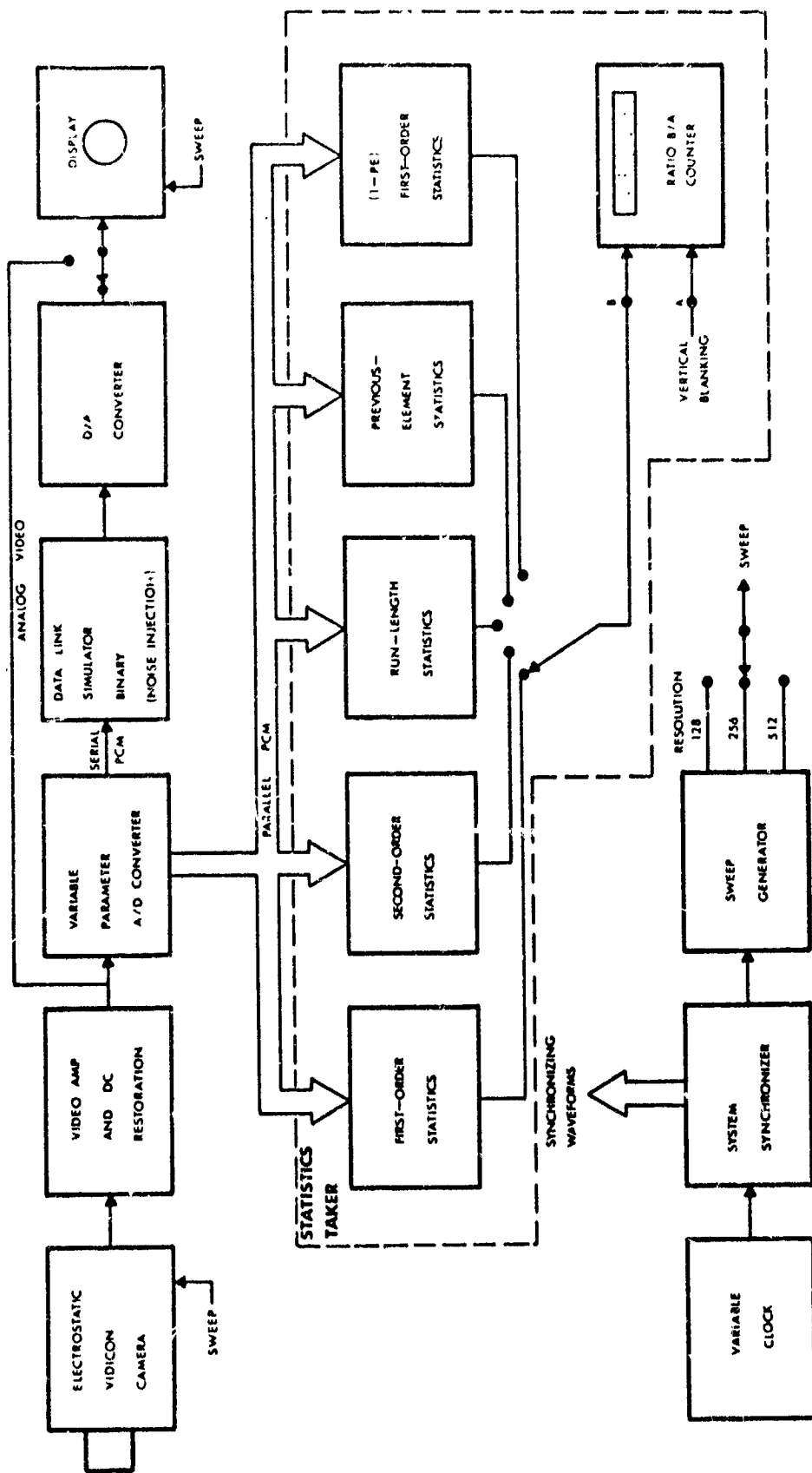


FIGURE 6.1-1 BLOCK DIAGRAM OF EDITS I

such effects as flacker, flicker, and display phosphor decay memory. The video obtained at each step of the sweep is sampled and the resulting level is held until it is encoded by a successive-approximation, analog-to-digital converter. This converter can encode the held video signal to 1, 2, 3, 4, 5 or 6 bits at the experimenter's command. The PCM-encoded video is sent to the display system via a controlled noise-corruption circuit which allows the experimenter to simulate the binary symmetric channel by inverting PCM bits at random on a "controlled-average" basis. For instance, one may wish to assess visually the effect of a 10^{-2} bit error rate on a given picture scene. With EDITS I the experimenter adjusts the level of a threshold circuit so that an average of one out of every 100 bits is inverted. Thus, one may experimentally assess the maximum tolerable bit error rate for digital television encoded to any of six quantization levels. Other parameters of the system are also variable. For example, the digitally derived sweeps may be set to either 128, 256, or 512-line resolutions; a 1024-line resolution is planned for use in assessing the signal characteristics of high-detail reconnaissance and planetary exploration photographs. The EDITS I system design will be expanded to incorporate redundancy-elimination circuits, efficient encoding circuits, sync-transmitting circuits, efficient decoding circuits, redundancy-restoration circuits, and sync-separation circuits to check out compression implementation designs.

EMR's camera system is built around the all-electrostatic Westinghouse WX-4384, electro-optically equivalent to the Westinghouse WL-7290 magnetic slow-scan vidicon. The tube's slow-scan capability, when coupled with the EDITS display storage capability, makes possible an exact assessment of how low the television frame rate may be for adequate rendition of given types of motion.

A future display system will employ a selective-erasure, direct-view storage tube that receives its analog input from a digital-to-analog converter. A conventional cathode-ray-tube monitor is available for current comparison studies. The converter receives its video from the noise-corruption circuit simulating the channel. The direct-viewing storage tube is one of the most advanced obtainable: it can be used to simulate a variable-decay phosphor; it may be operated so that it erases one written television frame and then erases it as the next one appears; it may be operated to erase the old information line-by-line just prior to writing in the new information; and it is capable* (according to manufacturer's specification) of storing and displaying half-tones to seven levels of gray.

6. EXPERIMENTAL VARIABLES AND EXPERIMENTAL RESULTS

6.2.1 Variable Parameters

If maximum compression is desired, then along with efficient encoding and redundancy elimination, the three basic parameters of frame rate, quantization, and resolution must be reduced. Although there is some published literature on the minimum frame rate for real-time television, the frame rate on EDITS was made variable to allow for a subjective evaluation of the displayed picture at various resolutions and quantization. The frame rate on EDITS for the stand-

and clock rate is 6.4, 1.6, and 0.4 frames per second for 128, 256, and 512 resolutions, respectively. However, the frame rate is variable up to 19.2 frames per second. From the subjective observations it was concluded that 15 frames per second is adequate for a space television system. It is felt that this is a conservative estimate relative to the Apollo system at 10 fps. Other applications for various other frame rates are tabulated in Table 2.1.3-1. The 15 frames per second is close to the 16-fps, home movie standards, and represents a 2:1 reduction in bandwidth relative to commercial TV standards. The bandwidth can be reduced further by considering the effects of reducing the resolution.

The present EDITS display is scan limited to about 350 to 400 lines. Nevertheless, from comparisons of the subjective results of the three resolutions, 128, 256, and 512, it is quite apparent that for specified applications reducing the resolution is possible without an appreciable loss in required information. It is necessary to specify the expected scene content and scientific information desired. Once this is specified it is possible to custom design the television system for the application. For example, for the majority of the anticipated manned spacecraft television applications it appears that the 256 resolution is sufficient. Therefore, a resolution reduction of 2:1 achieves a bandwidth reduction of 4:1 and the overall reduction is 8:1 by considering the reduced frame rate. The final basic parameter which can be reduced is the quantization.

The experiments performed with variable quantization far exceeded the reduced-resolution and frame-rate experiments. To illustrate these experiments the photographic matrix was constructed. A subjective analysis is necessary to determine the effects of the reduced-frame rate, resolution, and quantization. Coupled with this matrix are the experiments performed on reducing the quantization in conjunction with coding techniques. It has been established that a 6-bit PCM system is necessary to almost totally prevent the annoying contouring effect. However, there are coding techniques such as the Roberts coding technique that utilizes the properties of vision to avoid contouring and yet reduce the quantization to 3 bits; hence achieving a 2:1 reduction in bandwidth.

It has been further established that a 5-bit PCM system is roughly equivalent to home receiver television. During the course of the experiments, it was found that a 4-bit PCM is adequate for space television since, for most scenes observed, there is an insufficient amount of contouring to be objectionable. Therefore, this indicates that a 1.5:1 reduction in bandwidth is possible by merely reducing the quantization to four bits. Combining this reduction with reducing frame rate and resolution, a 12:1 overall reduction in bandwidth is possible by parameter reduction alone. This parameter reduction combined with an efficient coding technique results in an extremely efficient system.

2.1.2. Variable Noise

It is necessary to consider the effects of noise in the data link to fully analyze any coding technique. EDITS has the capability of simulating the digital noise;

that is, the data are passed through a binary symmetric channel and the noise-corrupted picture can be displayed. The effects of noise on the uncompressed data must be obtained along with the effects of noise on the compressed data to evaluate the compression system.

The uncompressed data were subjected to bit-error probabilities from 10^{-5} to 10^{-1} . For the uncompressed data a bit-error probability of 10^{-3} is sufficient for space applications since the digital errors cause only a salt-and-pepper effect in the displayed image. In a real-time system at 15 frames per second and assuming that the errors are random, the eye would tend to integrate out the 10^{-3} bit-error noise. The minimum bit-error probability for uncompressed data is a function of the frame rate and resolution. Experiments were performed on EDITS on varying the bit-error probability, resolution, frame rate, and quantization, and some of the subjective results are given in the photographic matrix in Section 6.2.5. The noise effects on the uncompressed data are quite readily obtainable; however, the ramifications of compression complicates the error effects for compression systems.

In general, one expects that the compressed data will be more vulnerable to errors since a single-bit error in the uncompressed system will effect only a single picture element, whereas a single bit error in the compressed data will effect many picture elements or lines. In many of the compressed data systems a single-bit error will cause streaks to appear in the displayed image instead of the usual digital salt-and-pepper effect. For example, the photographic matrix in Section 6.2.5 compares the digital noise performance of an uncompressed system with two different compressed systems.

The compressed system in general will require a low bit-error probability; however, there are exceptions. For example, the Roberts coding technique requires the same bit-error probability as the uncompressed system; however, other coding techniques such as redundancy elimination techniques (i.e., PE coding), do require a lower bit-error probability. Experimentally it has been determined that the delta system requires only 1 db signal-to-noise ratio more than the equivalent PCM system.

It is absolutely necessary to perform a complete error analysis accompanied by experimentation on each coding technique before the net or over-all compression can be stated. It is conceivable that a compression technique could achieve an apparent compression that could be nullified by a corresponding increase in required signal-to-noise ratio. This possibility emphasizes the necessity for studying the effects of noise.

6.2.3 Variable Encoding

EDITS currently has the capability of simulating 10 compression techniques. These are (1) previous-element coding, (2) run-length coding, (3) truncated run-length coding, (4) previous element run-length coding, (5) previous-element previous-line coding, (6) delta with previous-element coding, (7) delta with run-length

coding, (8) Roberts with previous-element coding, (9) Roberts with run-length coding, and (10) split-band coding. The simulation is not necessarily sufficient to obtain subjective results, but rather to obtain the statistics of each coding technique as an information source to determine whether or not the coding technique can achieve a sufficient amount of compression to warrant further investigation. The statistical results of EDITS compression technique simulation are given in Section 6.2.4. Along with the source simulation, EDITS also has built into it three digital coding techniques. The digital coding techniques are (1) pulse code modulation, (2) Roberts coding, and (3) delta modulation. The pulse-code modulation experiments are summarized in Section 6.2.1 where the three basic parameters were varied.

The delta system of EDITS is a two-bit delta with an exponential integrator. The delta system required many experimental trade-offs to determine the optimum system. These were performed to determine the appropriate trade-offs between the delta step-height sizes, integrator time constant, resolution, and noise effects. In general, the delta system performed as predicted especially with regard to the effects of noise. Although the two-bit tunnel diode analog-to-digital converter is somewhat novel, the overall complexity of the delta system is much less than the PE coding system. While the delta system is a promising candidate for a compression system, the subjective results to date are not quite as good as the Roberts technique results.

The Roberts coding technique is among the most promising coding techniques. A partial implementation on EDITS has been accomplished. This partial implementation appears to be one of the major reasons why the subjective results of the Roberts technique are not as good as the four-bit PCM, although L. G. Roberts* contends that when the technique is fully utilized, the three-bit PCM system with Roberts is subjectively equivalent to six-bit PCM. At the present time the squaring amplifier of the Roberts technique has not been constructed; however, it is anticipated that when this problem is eliminated and the over-all system transfer function is optimized with respect to the coding technique, the Roberts coding technique will perform as specified. The Roberts coding technique, combined with previous-element coding to obtain the most promising and efficient coding technique, is the coding technique recommended.

Complete statistics were obtained on the effects of varying the basic PCM quantization with the Roberts pseudo-random noise quantization. The end result of these experiments is that the three-bit PCM with four bits of Roberts pseudo-random noise is the optimum combination. This decision was based on not only subjective results, but also the compression achieved when the previous-element coding technique is combined with Roberts.

*Roberts, L. G.: "Picture Coding Using Pseudo-Random Noise." IRET on Information Theory, February 1962.

The photographic matrix in section 6.2.5 compares the subjective and noise results of the basic six-bit PCM system, the 2-bit delta system, and the 3-bit PCM-4 bit Roberts system with and without noise. The PCM-Roberts system is about as tolerant of channel noise as is the straight PCM system. The delta noise is subjectively different in appearance due to the propagating effects of the delta receiver integrator.

6.2.4 Picture Statistics

EDITS was designed not only to be an experimental tool for the subjective evaluation of compression techniques, but also to be a statistics taker for the many coding techniques. Although in the past the statistics of a television source have been based on the actual photographic or scene content, the statistics from EDITS are the statistics of the actual video. The statistics that have been taken on EDITS are: (1) First-order, (2) Second-order, (3) PE and (1-PE), and (4) Run-length statistics. The 11 standard scenes were processed on EDITS and the results tabulated. Table 6.2.4-1 gives the first-order probabilities for the standard scenes quantized to four bits.

Although this table enables the evaluation of efficient encoding of the data, the PE and (1-PE) statistics were taken to evaluate the efficiency of the PE coding technique. The PE and (1-PE) probabilities for the standard scenes for PE coding are given in Table 6.2.4-2. These statistics, when properly applied, will give the efficiency of the coding technique.

While the PE coding technique alone appears to be a very promising compression technique, the theoretical results of the Roberts plus PE technique appear to be more promising. Therefore, the first-order probabilities for a 3-bit PCM source with 4 bits of Roberts pseudo-random noise were taken. These statistics are tabulated in Table 6.2.4-3. However, to achieve the optimum efficiency, the coding technique should employ efficient encoding of the digital data. PE coding was applied to the Roberts source and the PE and (1-PE) probabilities for the standard pictures were tabulated in Table 6.2.4-4.

These probabilities were then converted to the common parameter of the entropy of the source, that is, the average number of bits per element. Two distinct and separate entropies were calculated and tabulated. These entropies are defined by

$$H_{1N} = - \sum_{j=1}^n P(i) \log_2 P(i), \quad (1)$$

$$H_{1PE} = - \sum_{j=1}^n P(i)' \log_2 P(i)' - PE \log_2 PE, \text{ and} \quad (2)$$

$$H_{CN} = \sum_{j=1}^n P(i) L(i). \quad (3)$$

TABLE 6.2.4-1 FIRST ORDER PROBABILITIES FOR STANDARD SCENES FOR 4-BIT CODING

SCENE EVENT	1	2	3	4	5	6	7	8	9	10	11
0	0.65361	0.64117	0.57074	0.57705	0.39471	0.29187	0.77811	0.73124	0.07811	0.06817	0.20377
1	0.08434	0.08269	0.11222	0.11373	0.15917	0.21699	0.01665	0.12032	0.07204	0.04688	0.06833
2	0.03766	0.03402	0.06651	0.05618	0.09520	0.08435	0.01619	0.03950	0.22533	0.04488	0.03727
3	0.02478	0.02350	0.05275	0.04351	0.06765	0.06861	0.01680	0.01944	0.20672	0.04979	0.04863
4	0.02013	0.01907	0.03731	0.03725	0.05892	0.06800	0.01756	0.01258	0.17758	0.08425	0.04848
5	0.01852	0.01495	0.02691	0.02824	0.04331	0.04172	0.01985	0.00793	0.10267	0.09559	0.05242
6	0.01952	0.01571	0.02446	0.02717	0.03352	0.03026	0.02123	0.00732	0.05919	0.11336	0.07742
7	0.01531	0.01480	0.01835	0.02275	0.02449	0.02460	0.01756	0.00564	0.02586	0.10417	0.10757
8	0.01454	0.01831	0.02293	0.02099	0.02525	0.02812	0.01283	0.00618	0.01526	0.07491	0.12120
9	0.01378	0.01770	0.01789	0.01710	0.02097	0.02269	0.00534	0.00549	0.00885	0.05208	0.06848
10	0.01760	0.02212	0.01529	0.01756	0.02097	0.02246	0.01183	0.00549	0.00702	0.04289	0.04250
11	0.02112	0.03021	0.01009	0.01450	0.01837	0.01757	0.01665	0.00564	0.00557	0.05913	0.05788
12	0.02878	0.03646	0.01032	0.01298	0.01607	0.01436	0.02016	0.00991	0.00549	0.05974	0.03409
13	0.01760	0.01877	0.00742	0.00687	0.01071	0.00764	0.01054	0.01128	0.00366	0.04442	0.02409
14	0.00918	0.00854	0.00527	0.00244	0.00704	0.00711	0.01107	0.01052	0.00320	0.04366	0.02242
15	0.00352	0.00198	0.00153	0.00168	0.00367	0.05364	0.00764	0.00152	0.00343	0.01608	0.00545

TABLE 6.2.4-2 PE AND (1-PE) PROBABILITIES FOR STANDARD SCENES FOR PE CODING

SCENE EVENT	1	2	3	4	5	6	7	8	9	10	11
0	0.02456	0.01973	0.03760	0.03027	0.03355	0.01778	0.00764	0.02257	0.00481	0.00441	0.01192
1	0.03540	0.03167	0.05595	0.04479	0.06389	0.03909	0.00993	0.02745	0.02169	0.00911	0.02139
2	0.02281	0.02126	0.04066	0.03057	0.05261	0.03419	0.01100	0.01411	0.05987	0.01382	0.01840
3	0.01755	0.01652	0.03210	0.02538	0.04193	0.02499	0.01314	0.00991	0.08186	0.01899	0.01834
4	0.01449	0.01316	0.02568	0.02247	0.03599	0.02345	0.01375	0.00763	0.07483	0.02962	0.01948
5	0.01221	0.01185	0.01865	0.01789	0.02790	0.02024	0.01512	0.00593	0.05269	0.03205	0.01948
6	0.01129	0.01178	0.01681	0.01712	0.02348	0.01686	0.01397	0.00526	0.03360	0.03737	0.02231
7	0.00977	0.01132	0.01337	0.01345	0.01830	0.01518	0.01069	0.00427	0.01863	0.03403	0.02361
8	0.01068	0.01254	0.01314	0.01231	0.01754	0.01487	0.00733	0.00458	0.01252	0.03068	0.02414
9	0.00992	0.01224	0.01070	0.00994	0.01449	0.01303	0.00458	0.00381	0.00779	0.02370	0.01627
10	0.01144	0.01300	0.00902	0.00887	0.01266	0.01150	0.00458	0.00397	0.00611	0.01747	0.00793
11	0.01106	0.01358	0.00688	0.00703	0.01067	0.00858	0.00530	0.00381	0.00489	0.01686	0.00733
12	0.01144	0.01247	0.00611	0.00535	0.00938	0.00690	0.00534	0.00442	0.00458	0.01641	0.00520
13	0.00748	0.00673	0.00390	0.00275	0.00595	0.00483	0.00351	0.00381	0.00290	0.01177	0.00504
14	0.00381	0.00214	0.00237	0.00122	0.00329	0.00399	0.00244	0.00214	0.00244	0.01435	0.00351
15	0.00107	0.00018	0.00069	0.00046	0.00091	0.00360	0.00122	0.00031	0.00145	0.00273	0.00076
PE	0.78502	0.78983	0.79637	0.75013	0.62746	0.74090	0.87046	0.87602	0.60935	0.68663	0.77475

TABLE 6.2.4-3 FIRST-ORDER PROBABILITIES FOR STANDARD SCENES FOR ROBERTS - PE CODING

SCENE EVENT	1	2	3	4	5	6	7	8	9	10	11
0	0.55868	0.53122	0.51477	0.50710	0.36166	0.29074	0.71037	0.59967	0.07710	0.06974	0.16880
1	0.21617	0.22898	0.23607	0.24097	0.29223	0.30760	0.09939	0.28883	0.31141	0.09675	0.14359
2	0.04491	0.04228	0.08727	0.07793	0.12254	0.12771	0.03338	0.03225	0.35978	0.13124	0.09471
3	0.03806	0.03144	0.05148	0.05521	0.07615	0.07611	0.03994	0.01553	0.16780	0.20418	0.12532
4	0.03045	0.03419	0.04264	0.04575	0.05234	0.05358	0.03110	0.01253	0.04686	0.18159	0.21509
5	0.03227	0.04030	0.03274	0.03431	0.04090	0.04384	0.01753	0.01039	0.01542	0.09583	0.12022
6	0.04719	0.06213	0.02071	0.02745	0.03495	0.03197	0.03430	0.01574	0.01134	0.11528	0.07241
7	0.03227	0.02946	0.01432	0.01128	0.01923	0.06842	0.03399	0.02506	0.01028	0.10438	0.05988

TABLE 6.2.4-4 PE AND (1-PE) PROBABILITIES FOR STANDARD SCENES FOR ROBERTS - PE CODING

SCENE EVENT	1	2	3	4	5	6	7	8	9	10	11
0	0.08185	0.08480	0.07222	0.07639	0.07265	0.06349	0.04421	0.10707	0.01341	0.01235	0.03648
1	0.09757	0.10090	0.09930	0.09598	0.10996	0.08928	0.05185	0.11671	0.08107	0.02910	0.05236
2	0.02657	0.02461	0.04313	0.03766	0.06163	0.04594	0.02233	0.01390	0.11733	0.04926	0.3282
3	0.02092	0.02039	0.02938	0.02786	0.04152	0.03357	0.02233	0.00871	0.07009	0.07540	0.04274
4	0.01878	0.01932	0.02264	0.02113	0.02936	0.02335	0.01644	0.00695	0.02735	0.06517	0.05388
5	0.01802	0.02024	0.01576	0.1515	0.02065	0.01984	0.00948	0.00580	0.00442	0.04547	0.04091
6	0.01863	0.02085	0.00994	0.00980	0.01575	0.01396	0.01040	0.00618	0.00808	0.00637	0.01984
7	0.00916	0.00874	0.00413	0.00344	0.00630	0.00702	0.00597	0.00382	0.00434	0.01849	0.00916
PE	0.70853	0.70015	0.70150	0.71259	0.64219	0.70353	0.81696	0.73086	0.66590	0.66838	0.71181

The minimum or statistical and the flat coded entropy are tabulated in Table 6.2.4-5. Although the entropy of each standard scene is given, the average entropy for the 4- and 3-bit sources is also tabulated. As expected the average entropy of the three-bit source with Roberts is less than the entropy of the conventional 4-bit source.

Although the entropy is a common parameter for coding comparisons, the final parameter desired is the compression. Therefore, the average, maximum, and minimum compressions for each technique are tabulated in Table 6.2.4-6. This table along with the size, weight, power, and noise considerations enabled the most efficient system to be specified. The statistics for several other promising coding techniques were taken.

Although run-length probabilities are somewhat time-consuming to obtain, the Moon scene was processed on EDITS and the run-length probabilities were obtained for run-lengths truncated at eight. These probabilities are given in Table 6.2.4-7. The compression for run-length coding is given by

$$\bar{C} = \frac{(\log_2 n) \left[\sum_{y=1}^k \sum_{j=1}^n y P(R_j^y) \right]}{\sum_{y=1}^k \sum_{j=1}^n H \left[P(R_j^y) \right]} \tag{4}$$

TABLE 6.2.4-5 ENTROPY OF STANDARD SCENES

Scene	Entropy*					
	H ₁₄	H _{PE4}	H _{C4}	H ₁₃	H _{PE3}	H _{C3}
1	2.169	1.557	1.860	2.008	1.612	1.875
2	2.226	1.532	1.841	2.116	1.642	1.399
3	2.415	1.905	2.174	2.071	1.631	1.895
4	2.408	1.700	2.000	2.095	1.481	1.862
5	2.990	2.289	2.490	2.394	1.880	2.073
6	3.215	1.791	2.036	2.551	1.665	1.890
7	1.615	1.044	1.518	1.621	1.177	1.549
8	1.612	0.973	1.496	1.883	1.352	1.907
9	3.043	2.217	2.562	2.212	1.709	2.002
10	3.882	2.083	2.254	2.918	1.851	1.995
11	3.653	1.611	1.901	2.894	1.692	1.865
Average	2.657	1.700	2.012	2.251	1.608	1.892

*H₁₄ = First-order statistic entropy of 4-bit source.
H_{PE4} = PE and (1-PE) statistical entropy of 4-bit source.
H_{C4} = Flat-coded entropy of 4-bit source.
H₁₃ = First-order statistical entropy of 3-bit source.
H_{PE} = PE and (1-PE) statistical entropy of 3-bit source.
H_{C4} = Flat-coded statistical entropy of 3-bit source.

TABLE 6.2.4-6 COMPRESSION OF STANDARD SCENES

Parameter	Statistic					
	$\bar{C}_4$	$\bar{C}_{PE4}$	$\bar{C}_{C4}$	$\bar{C}_3$	$\bar{C}_{PE3}$	$\bar{C}_{C3}$
Average	2.258	3.529	2.982	2.665	3.731	3.171
Minimum	1.546	2.621	2.342	2.056	3.191	2.894
Maximum	3.722	6.166	4.011	3.701	5.098	3.873

TABLE 6.2.4-7 TRUNCATED RUN-LENGTH PROBABILITIES
FOR MOON SCENE AT 256 RESOLUTION

μ	j							
	0	1	2	3	4	5	6	7
1	0.0058	0.0378	0.0893	0.1131	0.0921	0.0397	0.0146	0.0012
2	0.0043	0.0258	0.0511	0.0577	0.0411	0.0208	0.0080	0.0001
3	0.0027	0.0159	0.0299	0.0344	0.0197	0.0110	0.0043	0.0001
4	0.0017	0.0101	0.0174	0.0197	0.0125	0.0060	0.0024	0
5	0.0014	0.0068	0.0113	0.0126	0.0072	0.0034	0.0014	0
6	0.0019	0.0052	0.0089	0.0078	0.0047	0.0018	0.0005	0
7	0.0019	0.0039	0.0056	0.0051	0.0035	0.0015	0.0003	0
8	0.0316	0.0184	0.0236	0.0249	0.0107	0.0030	0.0003	0

For the Moon scene the compression is 1.77:1.

Another form of run-length coding is PERL (previous-element-run-length) coding. For the Moon scene the PERL probabilities are tabulated in Table 6.2.4-8. The compression for the PERL coding technique is given by

$$C_{PERL} = \frac{(\log_2 n) \left\{ \sum_{j=1}^n P(j)' + \sum_{j=1}^n [1 - P(j)'] \sum_{y=1}^k y P(R_j^y) \right\}}{\sum_{j=1}^n \sum_{y=1}^k H(PERL)} \quad (5)$$

TABLE 6.2.4-8 PERL AND (1-PERL) PROBABILITIES FOR MOON SCENE
AT 256 RESOLUTION

Level	Probability	PE Run-Length	Probability
(1-PERL) 0	0.0078	1	0.2386
(1-PERL) 1	0.0385	2	0.1348
(1-PERL) 2	0.0776	3	0.0792
(1-PERL) 3	0.0906	4	0.0505
(1-PERL) 4	0.0649	5	0.0351
(1-PERL) 5	0.0304	6	0.0242
(1-PERL) 6	0.0112	7	0.0198
(1-PERL) 7	0.0005	8	0.0965

Substituting the probabilities into the compression expression, the PERL compression is 1.56:1. This compression is somewhat less than the compression for conventional run-length coding, but this was predicted. These compression figures do not appear to be too promising. However, another coding technique does have the potential for increasing the compression over PE coding.

The PEPL coding technique, which is an extension of PE coding in the vertical direction, does appear to have the potential for increasing the compression. Statistics were taken to obtain the PE, PL, and (1-PE - PL) probabilities for the Girl scene. Since a comparison between PEC and PEPL is desired, the PE and (1-PE) statistics were taken on the same picture and the results are tabulated in Tables 6.2.4-9 and 6.2.4-10.

TABLE 6.2.4-9 PE, PL, (1-PE-PL) PROBABILITIES
FOR "GIRL" SCENE AT 128 RESOLUTION

Event	Probability
PE	0.7820
PL	0.0988
(1-PE-PL) 0	0.0079
(1-PE-PL) 1	0.0203
(1-PE-PL) 2	0.0165
(1-PE-PL) 3	0.0191
(1-PE-PL) 4	0.0199
(1-PE-PL) 5	0.0222
(1-PE-PL) 6	0.0106
(1-PE-PL) 7	0.0030

TABLE 6.2.4.10 PE AND (1-PE) PROBABILITIES FOR
"GIRL" SCENE AT 128 RESOLUTION

Event	Probability
PE	0.7820
(1-PE) 0	0.0211
(1-PE) 1	0.0351
(1-PE) 2	0.0333
(1-PE) 3	0.0313
(1-PE) 4	0.0361
(1-PE) 5	0.0342
(1-PE) 6	0.0165
(1-PE) 7	0.0043

The compression for the PEPL and PE coding techniques is 1.92:1 and 1.86:1, respectively. Thus, experimental results verified the increase in compression by PEPL coding. The amount of the increase must be weighed with the increase in complexity.

These statistics are but a few of the many that could be taken on EDITS; however, they are representative and illustrate the remarks and conclusions in Section 2 on coding.

6.2.5 Photographs

Prints of the following scenes are presented in this section.

- Scene 1: APOLLO 1 (Variable Quantization)
- Scene 1: APOLLO 1 (Noise Effects)
- Scene 2: APOLLO 2 (Variable Quantization)
- Scene 2: APOLLO 2 (Noise Effects)
- Scene 3: Cooper 1 (Variable Quantization)
- Scene 3: Cooper 1 (Noise Effects)
- Scene 4: Cooper 2 (Variable Quantization)
- Scene 4: Cooper 2 (Noise Effects)
- Scene 5: Cooper 3 (Variable Quantization)
- Scene 5: Cooper 3 (Noise Effects)
- Scene 6: Laboratory (Variable Quantization)
- Scene 6: Laboratory (Noise Effects)
- Scene 7: Telescope (Variable Quantization)
- Scene 7: Telescope (Noise Effects)
- Scene 8: Emission Nebula (Variable Quantization)
- Scene 8: Emission Nebula (Noise Effects)

Scene 9: Moon Picture (Variable Quantization)
Scene 9: Moon Picture (Noise Effects)
Scene 10: Test Chart (Variable Quantization)
Scene 10: Test Chart (Noise Effects)
Scene 11: MIT Girl (Variable Quantization)
Scene 11: MIT Girl (Noise Effects)

Variation of PCM Parameters

An example of the use of the EMR-developed EDITS I to obtain comparative noise performance evaluation of 3 video modulation systems which exhibit gross compression ratios of 1, 2, and 3 (PCM, Roberts, and Delta).

Comparison on the EMR-developed EDITS I, 2 subjective reproduction compression systems with 6-bit PCM for 3 different scan resolutions with optical resolution held constant.



Quantization 4 bits
Resolution 256 lines



Quantization 5 bits
Resolution 256 lines



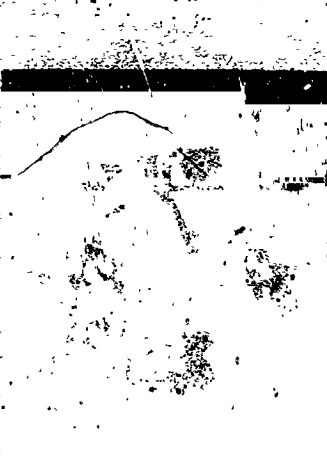
Quantization 3 bits
Resolution 256 lines



Quantization 6 bits
Resolution 256 lines



Quantization 2 bits
Resolution 256 lines

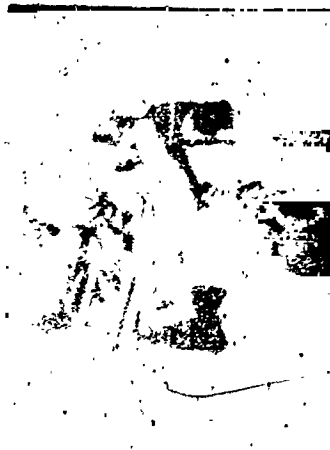


Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines

SCENE 1: APOLLO 1 (VARIABLE QUANTIZATION)



Quantization 4 bits
Resolution 512 lines



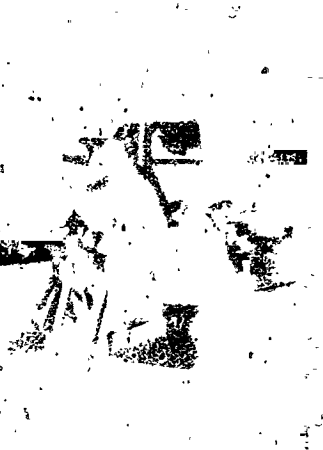
Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



Quantization 4 bits
Resolution 128 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}



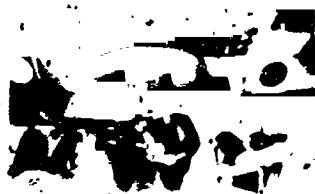
Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-3}

SCENE 1: APOLLO 1 (NOISE EFFECTS)





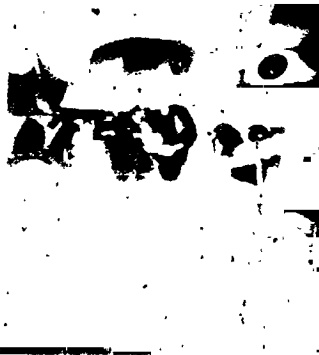
Quantization 4 bits
Resolution 256 lines



Quantization 5 bits
Resolution 256 lines



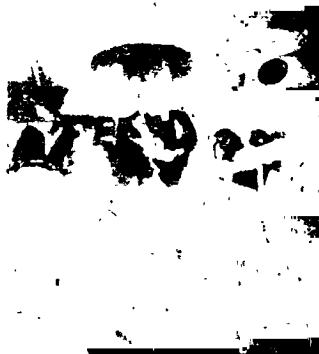
Quantization 3 bits
Resolution 256 lines



Quantization 6 bits
Resolution 256 lines



Quantization 2 bits
Resolution 256 lines



Quantization 5 bits
with 4 bits Roberts
Resolution 256 lines

SCENE 2: APOLLO 2 (VARIABLE QUANTIZATION)



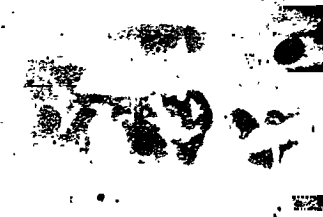
Quantization 4 bits
Resolution 512 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



Quantization 4 bits
Resolution 128 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}



Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-3}

SCENE 2: APOLLO 2 (NOISE EFFECTS)



Quantization 4 bits
Resolution 256 lines



Quantization 5 bits
Resolution 256 lines



Quantization 3 bits
Resolution 256 lines



Quantization 6 bits
Resolution 256 lines



Quantization 2 bits
Resolution 256 lines



Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines

SCENE 3: COOPER 1 (VARIABLE QUANTIZATION)



Quantization 4 bits
Resolution 512 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



Quantization 4 bits
Resolution 128 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}



Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-3}

SCENE 3: COOPER 1 (NOISE EFFECTS)



Quantization 4 bits
Resolution 256 lines



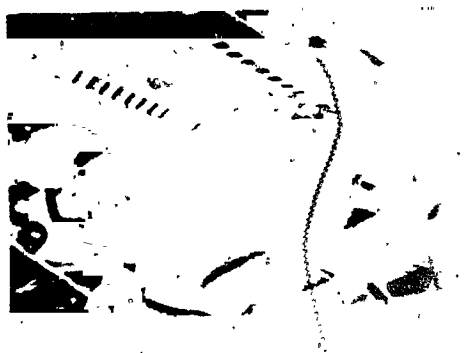
Quantization 5 bits
Resolution 256 lines



Quantization 3 bits
Resolution 256 lines



Quantization 6 bits
Resolution 256 lines



Quantization 2 bits
Resolution 256 lines



Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines

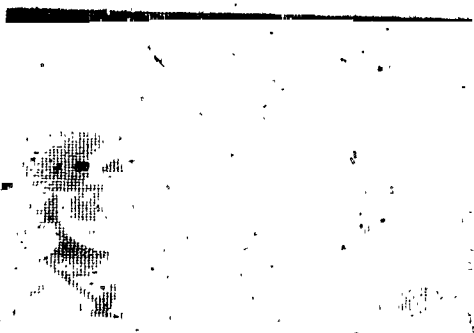
SCENE 4: COOPER 2 (VARIABLE QUANTIZATION)



Quantization 4 bits
Resolution 512 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



Quantization 4 bits
Resolution 128 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}



Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-3}

SCENE 4: COOPER 2 (NOISE EFFECTS)



Quantization 4 bits
Resolution 256 lines



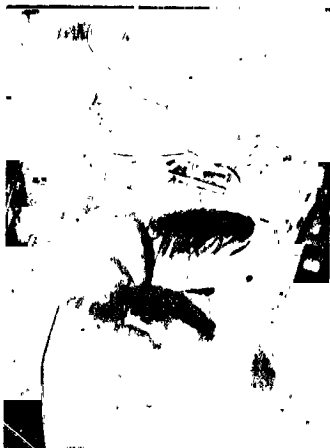
Quantization 5 bits
Resolution 256 lines



Quantization 3 bits
Resolution 256 lines



Quantization 6 bits
Resolution 256 lines



Quantization 2 bits
Resolution 256 lines



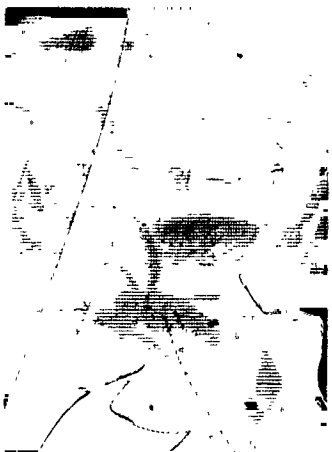
Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines



Quantization 4 bits
Resolution 512 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



Quantization 4 bits
Resolution 128 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}



Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-3}

SCENE 5: COOPER 3 (NOISE EFFECTS)



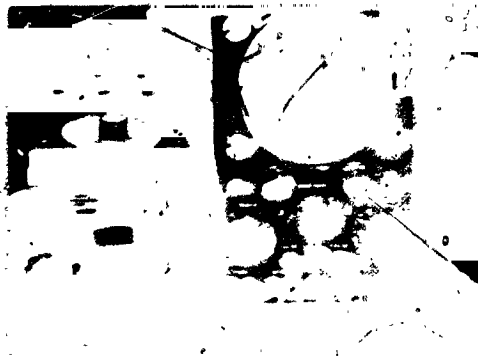
Quantization 4 bits
Resolution 256 lines



Quantization 5 bits
Resolution 256 lines



Quantization 3 bits
Resolution 256 lines



Quantization 6 bits
Resolution 256 lines

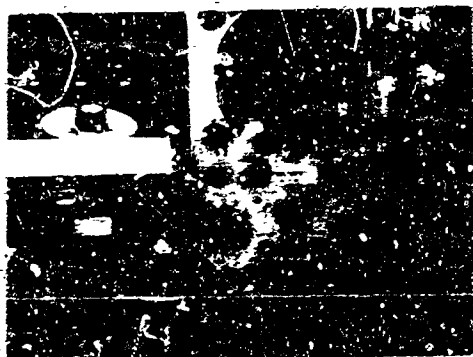


Quantization 2 bits
Resolution 256 lines



Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines

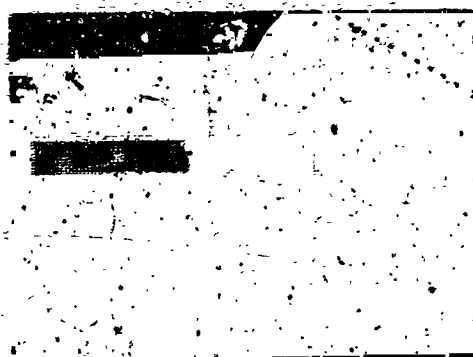
SCENE 6: LABORATORY (VARIABLE QUANTIZATION)



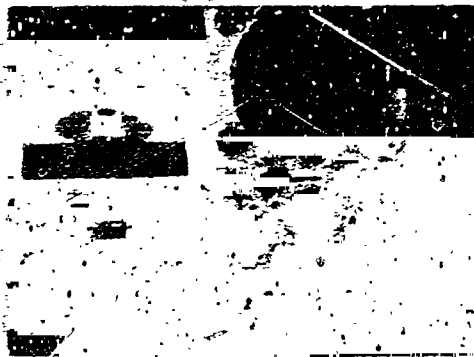
Quantization 4 bits
Resolution 512 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



Quantization 4 bits
Resolution 128 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}

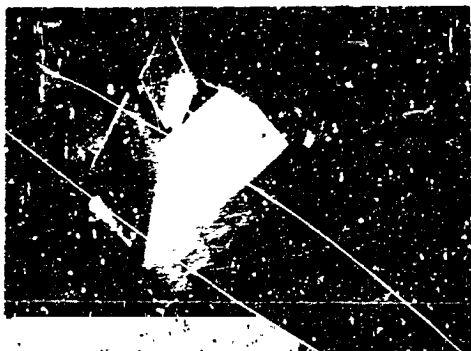


Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-3}

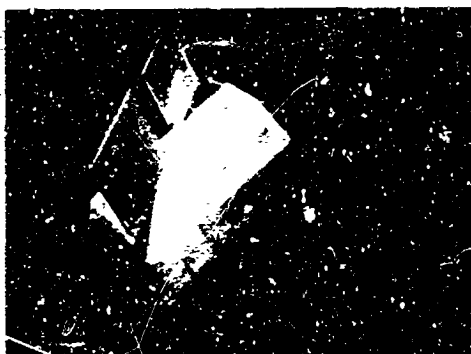
SCENE 8: LABORATORY (NOISE EFFECTS)



Quantization 4 bits
Resolution 256 lines



Quantization 5 bits
Resolution 256 lines



Quantization 3 bits
Resolution 256 lines



Quantization 6 bits
Resolution 256 lines



Quantization 2 bits
Resolution 256 lines



Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines

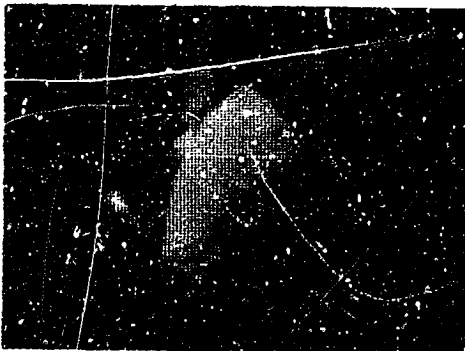
SCENE 7: CELESCOPE (VARIABLE QUANTIZATION)



Quantization 4 bits
Resolution 512 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



Quantization 4 bits
Resolution 128 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}

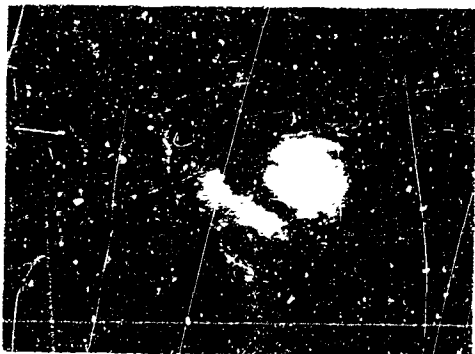


Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}

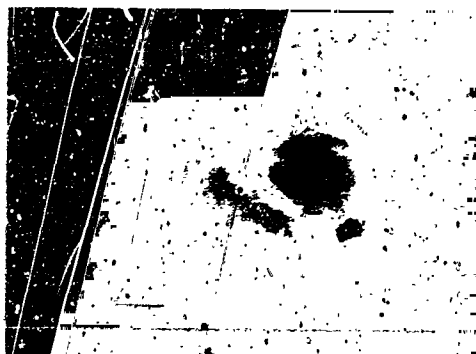


Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-3}

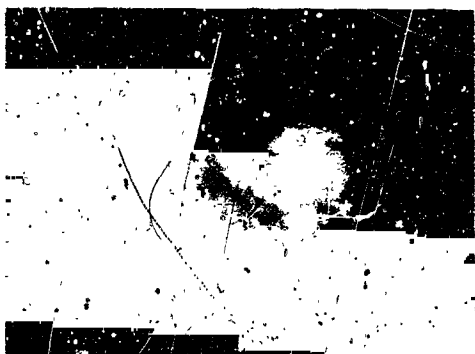
SCENE 11: CELESCOPE (NOISE EFFECTS)



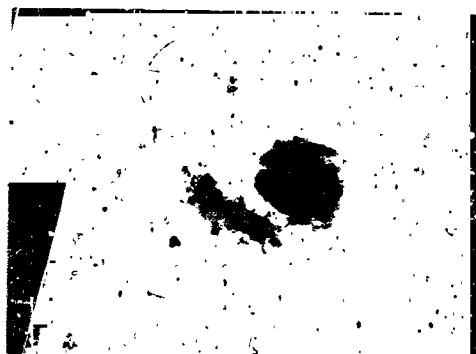
Quantization 4 bits
Resolution 256 lines



Quantization 5 bits
Resolution 256 lines



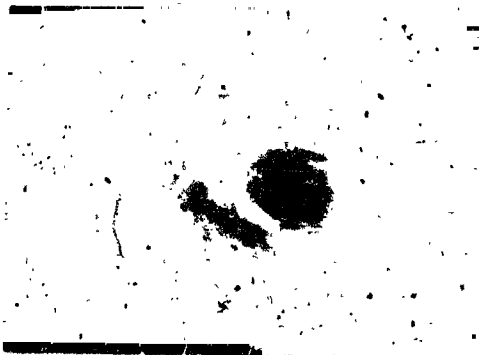
Quantization 3 bits
Resolution 256 lines



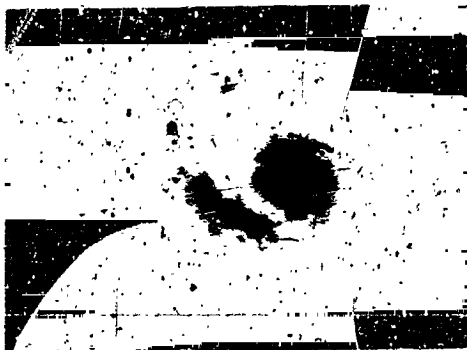
Quantization 6 bits
Resolution 256 lines



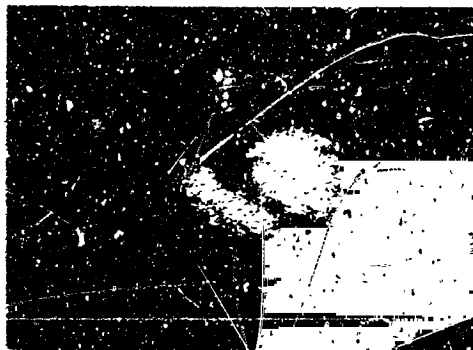
Quantization 2 bits
Resolution 256 lines



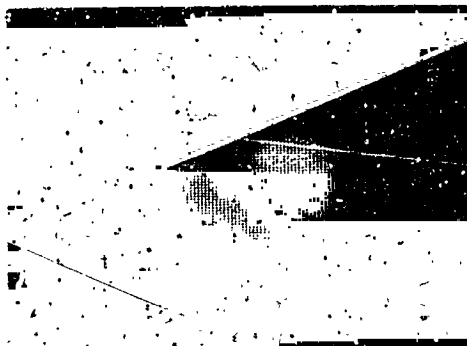
Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines



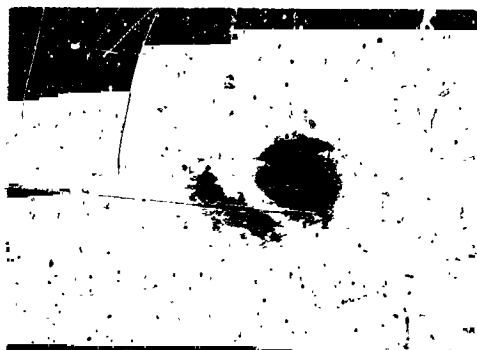
Quantization 4 bits
Resolution 512 lines



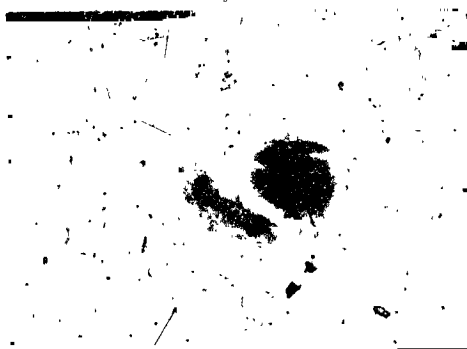
Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



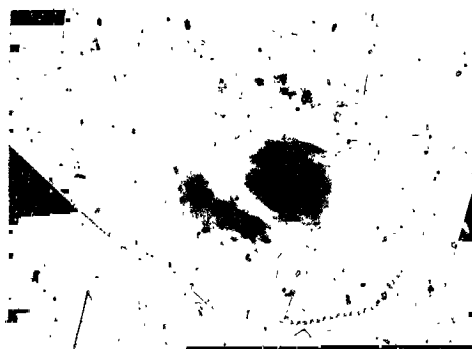
Quantization 4 bits
Resolution 128 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}

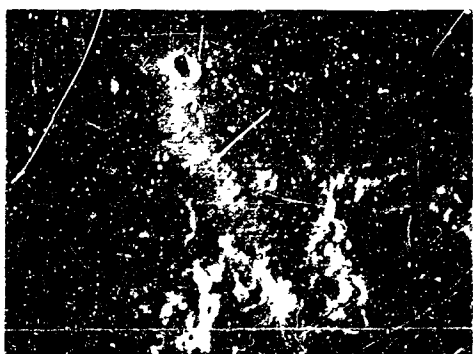


Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}

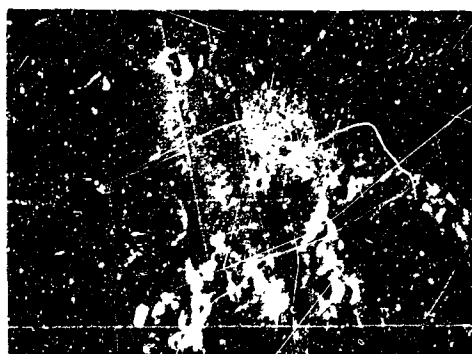


Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-3}

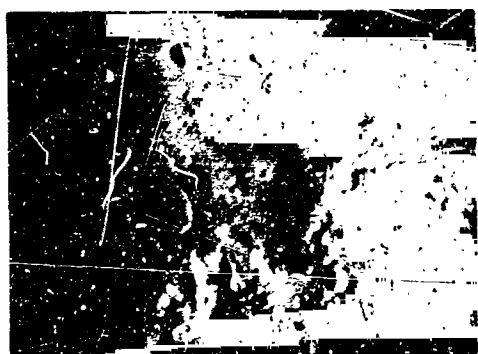
SCENE 8: EMISSION NEBULA (NOISE EFFECTS)



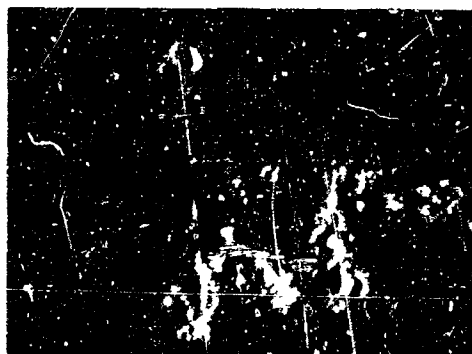
Quantization 4 bits
Resolution 256 lines



Quantization 5 bits
Resolution 256 lines



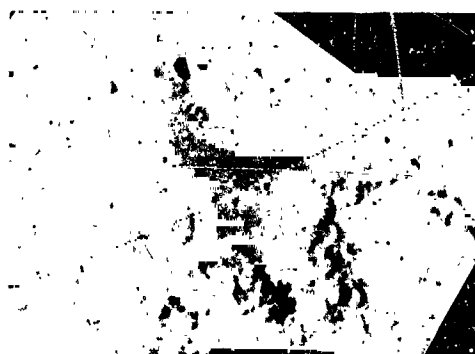
Quantization 3 bits
Resolution 256 lines



Quantization 6 bits
Resolution 256 lines

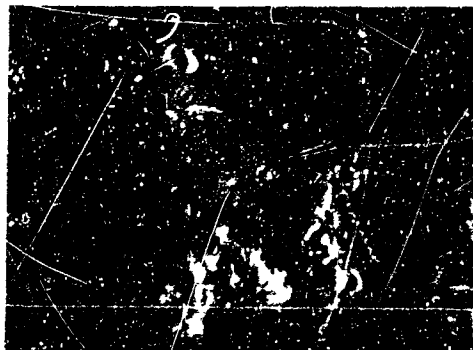


Quantization 2 bits
Resolution 256 lines

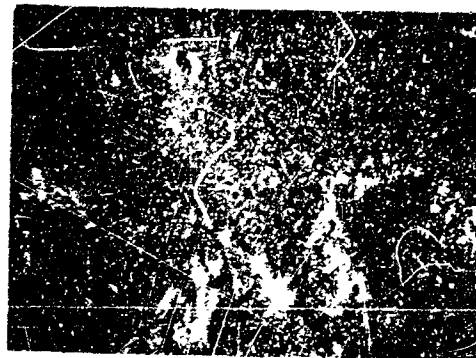


Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines

SELECTION OF PICTURE (VARIABLE QUANTIZATION)



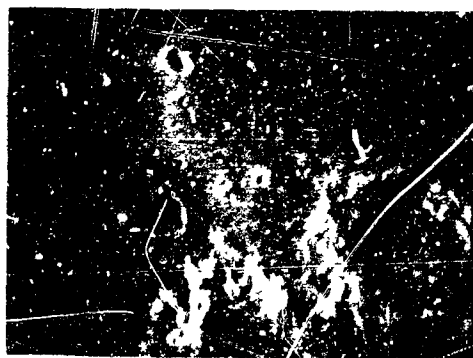
Quantization 4 bits
Resolution 512 lines



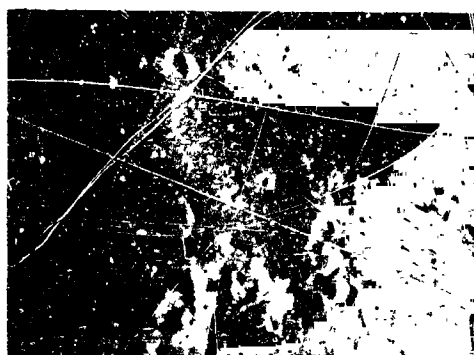
Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



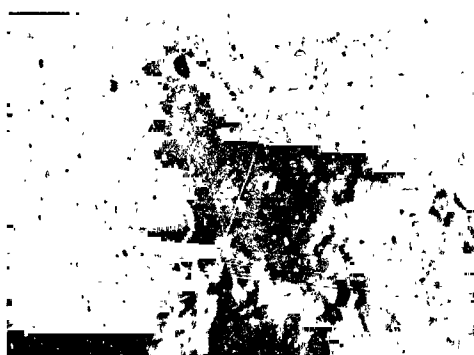
Quantization 4 bits
Resolution 128 lines



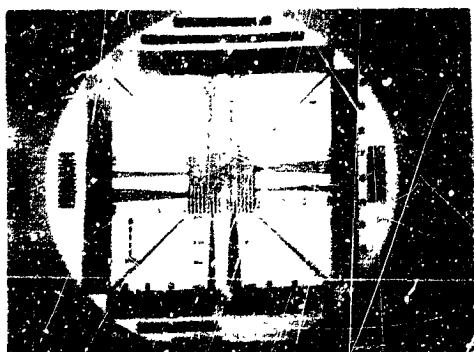
Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}



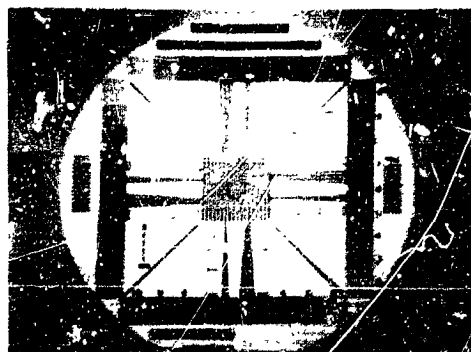
Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}



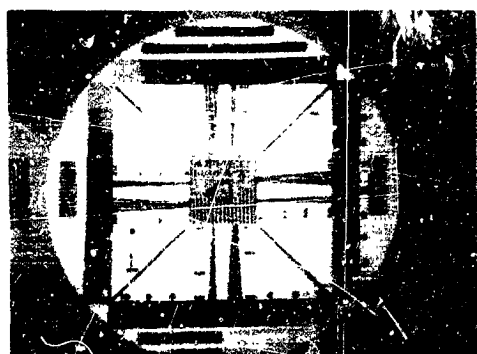
Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



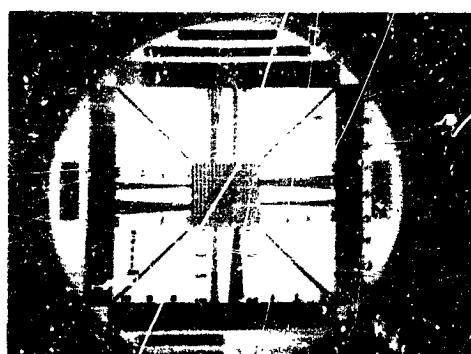
Quantization 4 bits
Resolution 256 lines



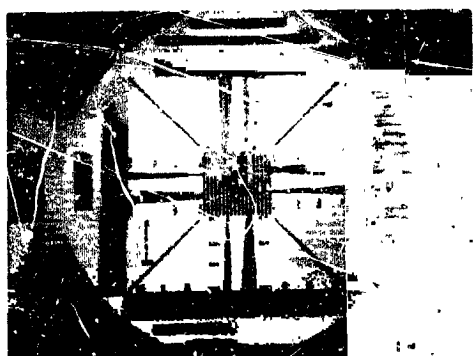
Quantization 5 bits
Resolution 256 lines



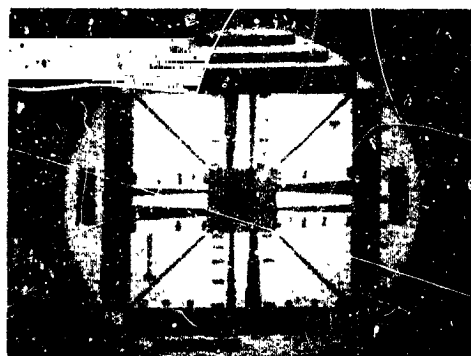
Quantization 3 bits
Resolution 256 lines



Quantization 6 bits
Resolution 256 lines

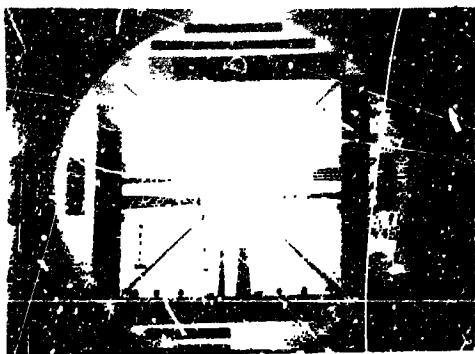


Quantization 2 bits
Resolution 256 lines

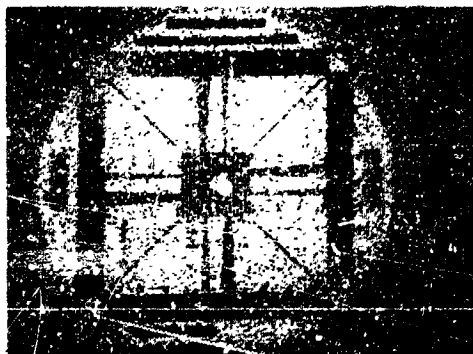


Quantization 3 bits
with 4 bits Robert's
Resolution 256 lines

SCENE 10. TEST CHART (VARIABLE QUANTIZATION)



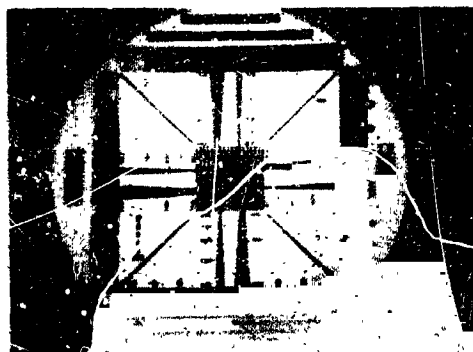
Quantization 4 bits
Resolution 512 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



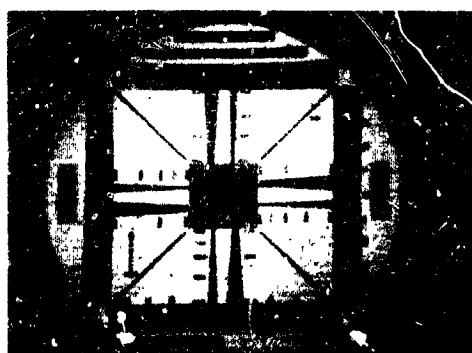
Quantization 4 bits
Resolution 128 lines



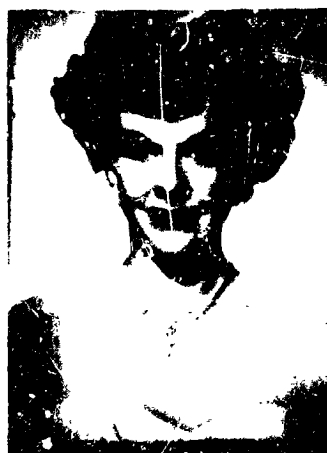
Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}



Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-3}





Quantization 4 bits
Resolution 512 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-1}



Quantization 4 bits
Resolution 128 lines



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-2}



Quantization 3 bits
with 4 bits Roberts
Resolution 256 lines
Error Rate 10^{-2}



Quantization 4 bits
Resolution 256 lines
Error Rate 10^{-3}

SCENE 11: MIT GIRL (NOISE EFFECTS)



Quantization: 6 bits
Resolution: 512 lines



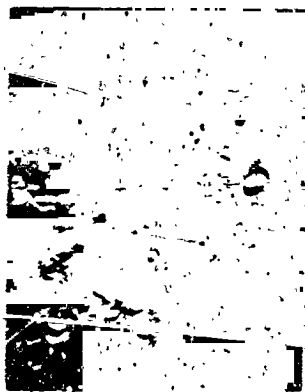
Quantization: 6 bits
Resolution: 256 lines



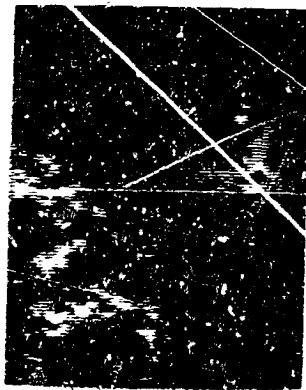
Quantization: 6 bits
Resolution: 128 lines



Quantization: 4 bits
Resolution: 512 lines



Quantization: 4 bits
Resolution: 256 lines



Quantization: 4 bits
Resolution: 128 lines



Quantization: 2 bits
Resolution: 2 lines



Quantization: 2 bits
Resolution: 256 lines

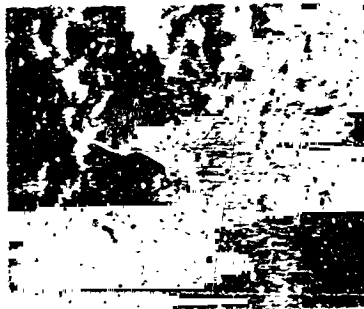


Quantization: 2 bits
Resolution: 128 lines

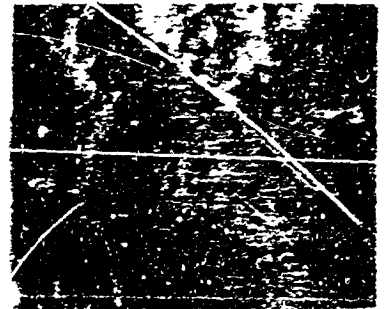
Variation of PCM Parameters



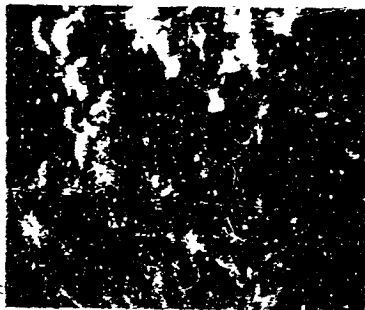
2-Bit Delta Modulation



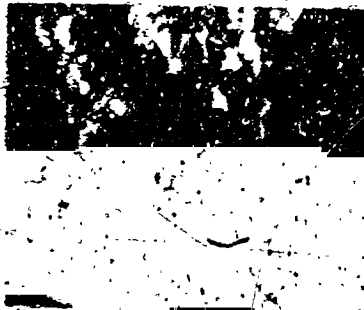
7-Bit Delta Modulation



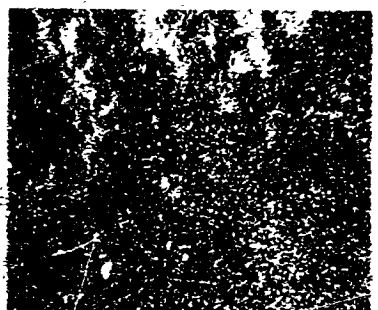
2 Bit Delta Modulation



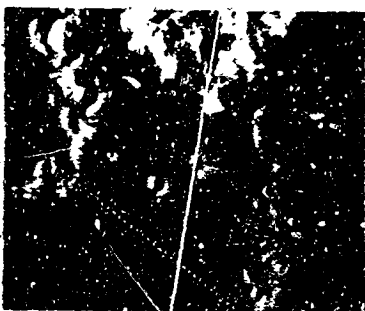
3-Bit PCM



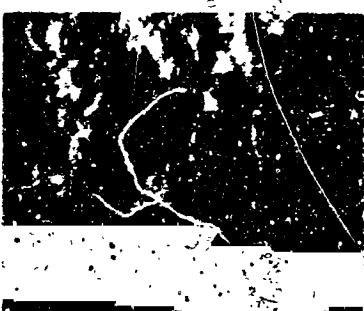
6-Bit PCM



6-Bit PCM



3-Bit PCM
with 4-Bit Roberts



3-Bit PCM
with 4-Bit Roberts



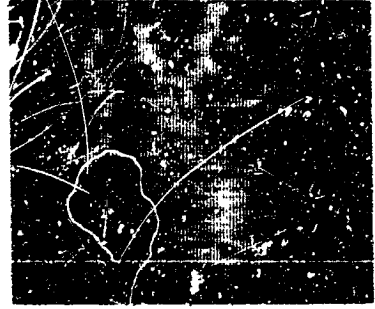
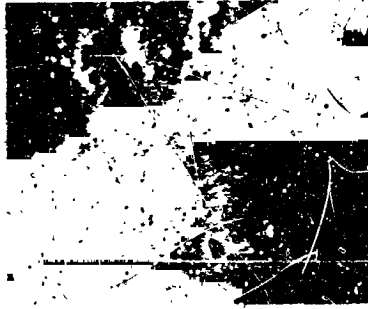
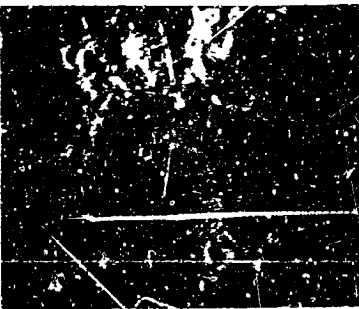
3-Bit PCM
with 4-Bit Roberts

Column 1
No Noise

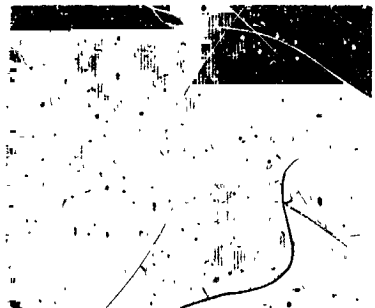
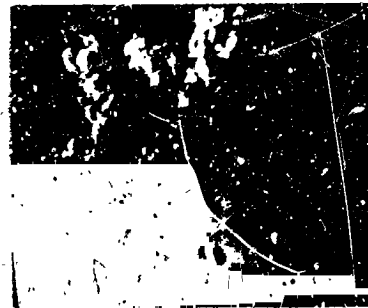
Column 2
 10^{-2} Bit Error Rate

Column 3
 10^{-1} Bit Error Rate

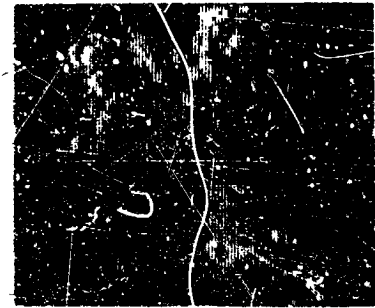
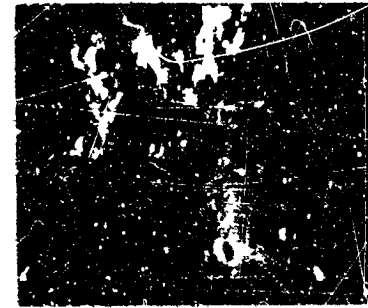
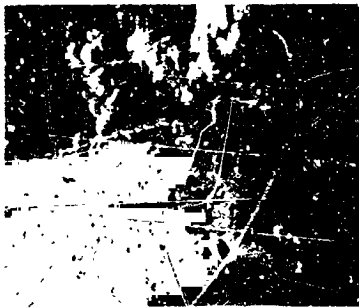
AN EXAMPLE OF THE USE OF THE EMR-DEVELOPED EDITS 1 TO OBTAIN COMPARATIVE NOISE PERFORMANCE EVALUATION OF 3 VIDEO MODULATION SYSTEMS WHICH EXHIBIT CROSS COMPRESSION RATIOS OF 1, 2, AND 5 (PCM, ROBERTS, AND DELTA)



2-Bit Delta Modulation
Gross Compression 3:1



6-Bit PCM
Gross Compression 1:1



3-Bit PCM with 4-Bit Roberts
Gross Compression 2:1

Column 1
128 Line Resolution

Column 2
256 Line Resolution

Column 3
512 Line Resolution

COMPARISON ON THE LMR-DEVELOPED EDITS 1 OF 2 SUBJECTIVE REPERCUSSION ON
COMPRESSION SYSTEMS WITH 6-BIT PCM FOR 3 DIFFERENT SCAN RESOLUTIONS
WITH OPTICAL RESOLUTION FIELD CONSTANT

SECTION 7

CONCLUSIONS AND RECOMMENDATIONS

7.0 EXECUTIVE SUMMARY

This section is omitted due to the brevity of the succeeding sections.

7.1 CONCLUSIONS

1. It is concluded that a compressed digital television system of the following characteristics is feasible of implementation, compatible with future world-wide ranges and will transmit television pictures from deep space to earth with minimum possible power drain from the spacecraft and still maintain an acceptable error rate at the earth receiving terminal.

System Characteristics

Sensor:	Slow scan electrostatic vidicon	
Resolution:	256 lines	(512 and 1024 lines future)
Quantization:	3 bits of PCM plus 4 bits of Roberts encoding	
Frame rate:	15 frames per second	(Less for 512 and 1024 lines)
Redundancy elimination:	Element-to-element with Previous Element Coder (PEC)	
Coding:	Maximally efficient coding with prefix property	(Modified Huffman code)
Buffer store:	None - efficient coder controls information source rate	
Input picture element rate	10^6 picture elements per second, long-term average	

Output bit rate:	Constant at 2 megabits per second
Spacecraft store:	Videotape capable of 2 megabit read-in and 1.2 or 2 megabit read-out compatible with future world-wide ground ranges
Ground scan conversion:	TV-display to TV-camera
Construction:	Semi-conductor integrated circuitry or discrete miniaturized parts depending on reliability vs size, weight and power trade-offs

2. It is concluded that the size, weight, power and complexity of the referenced compressed digital television system is less by a factor of 2 than the size, weight, power and complexity of a comparably miniaturized standard analog system of equal frame rate and equal horizontal and vertical resolution.
3. It is concluded that useful compressions (3:1) can be obtained by considering only digital modulation systems.
4. It is concluded that useful compressions (3:1) may be obtained by utilizing no more than the redundancy inherent in element-to-element correlations, and that high compressions ($\geq 6:1$) may be obtained with further study of the utilization of the redundancy inherent in higher order correlations.
5. It is concluded that the use of an efficient (modified Huffman) coding technique is necessary to utilize the maximum benefits afforded by the reduction of element-to-element redundancy.
6. It is concluded that a differentially coherent phase shift keying rf modulation system imposes no restrictions on the referenced compressed digital television system.
7. It is concluded that for Condition 1 (of the RFP) the horizontal and vertical resolutions can be 525 lines and that the frame rate with 3:1 compression is then 42.45 frames per second for real-time transmission.
8. It is concluded that for the path described as Condition 2 (of the RFP) frames of 525 by 525 line resolution can be transmitted with 3:1 compression from a storage medium at the rate of 11.72 frames per second.

9. It is concluded that for Condition 2 videotape recording is the preferred storage system when compared with rapidly and simply developed motion picture film.
10. It is concluded that no work was required on the rf modulation portion of the system beyond incorporating the information in the RFP.
11. It is concluded that the earth-based scan conversion portion of the system, being based on TV broadcast practice, poses no exceptional problems in receiving and converting compressed digital television to commercial analog TV standards.
12. It is finally concluded that the immediate development of compressed digital TV laboratory engineering hardware is warranted based on the findings of this study.

7.2 RECOMMENDATIONS

The recommendations of this study deal with two areas: (1) Follow-on development of the system originated, and (2) further study of more efficient systems which came to light during the current study.

It is recommended that

1. Manned Spacecraft Center have fabricated as the next phase of this work a complete laboratory engineering model of a 3-bit PCM - 4-bit Roberts compressed digital television system employing previous-element redundancy elimination and efficient (modified Huffman) encoding.
2. Manned Spacecraft Center initiate a comprehensive, one-year, theoretical and experimental study of those video compression techniques expected to yield large ($\geq 6:1$) compressions, such study to be conducted from the source-entropy point of view and to run concurrently with (1) above.

APPENDIX A

RFP EXHIBIT "A"

STATEMENT OF WORK

FOR

MANNED SPACECRAFT ADVANCED TELEVISION STUDY

I. Intent

This Statement of Work describes a study which is required for the development of an advanced and highly efficient modulation system for transmitting television information from manned spacecraft in deep space to earth. This task is a study only; no deliverable equipment is required. However, all important assumptions made in the study must be proved by results obtained from tests made on "breadboard" models of circuits in question and must be well documented by test results, written discussions and schematic diagrams (where applicable).

II. Specific Requirements

A. The prime requirement of this minimum weight and volume system is that it transmit television pictures from deep space to earth with the minimum possible power drain from the spacecraft and still maintain an acceptable error rate at the earth receiving terminal. This system would not be considered to have an advantage over the standard U. S. broadcasting television system, for example, if:

1. Its bandwidth requirement for real time transmission were markedly reduced from the 4 mc of commercial television but its sensitivity to errors were increased so that approximately the same transmitter output power is required.

2. The transmitter power and weight can be reduced only to have this advantage offset by a comparable increase in the power and weight of the television terminal equipment inside the spacecraft.

B. The study is to provide trade-off comparisons on reliability, size, weight, power and complexity of this system vs. a comparably miniaturized standard analog system of equal frame rate and equal horizontal and vertical resolution.

C. Only digital modulation systems are to be considered in this study.

D. The study is to consider elimination of redundancy of element-to-element, line-to-line, and frame-to-frame information. Any special coding techniques which the contractor may propose are also to be considered.

E. The rf modulation to be used in the spacecraft may be assumed to be differentially coherent phase shift keying for the purpose of this study. A plot of the probability of error vs. input signal-to-noise ratio is included as Appendix I of this outline and was extracted from Report RADC-TRI-58-91, AD 148 803 prepared by the Cornell Aeronautical Laboratory. Appendix II contains two conditions for which information on the system under consideration is required. For Condition 1, the contractor is to provide information as to the frame rate and vertical horizontal resolutions which can be obtained for real time transmission. (The term "real time" does not prevent the contractor from electrically storing several frames of information before actually starting its transmission). For the path described as Condition 2, the contractor is to provide information as to the rate of transmission which can be obtained for 525 line x 525 line resolution frames from a storage medium. This information should be estimated in the proposals, and the final report is to contain a comparison of the expected frame rates and resolutions of the proposal with those which are to be expected from the finalized system.

F. The storage mediums to be considered for Condition 2 of Appendix II are rapidly and simply developed motion picture film (such as the saturated web process) and the various types of video tape recorder. The contractor is to investigate both processes in this study and is to adequately document his choice by comparison of weights, volumes, power requirements and complexities.

G. No work on the rf portion of this television system is required other than incorporating the information given in Paragraph E. in the necessary computations.

H. The earth-based portion of the system is the picture reconstruction equipment and is to include the necessary post detection demodulation equipment, and 8 inch monitor scope and a translator which converts the signals received from the spacecraft to a regular commercial television 525 line, 30 frames per second format for recording on an Ampex video tape recorder and later broadcasting by commercial television stations. The size, weight, and power consumption of this equipment are not important factors.

I. Ten copies of the final report are required. The report is to contain, in addition to the information required in the previous paragraphs, a description of the operation of each stage of the system and a presentation of the design information for each unusual stage involved. Sufficient information is to be presented to enable a complete engineering model of the system to be fabricated as the next phase of this work without the necessity of developing new concepts or circuits.

J. All equipment fabricated for this contract or purchased and charged to the contract is to be the property of the Manned Spacecraft Center.

APPENDIX I

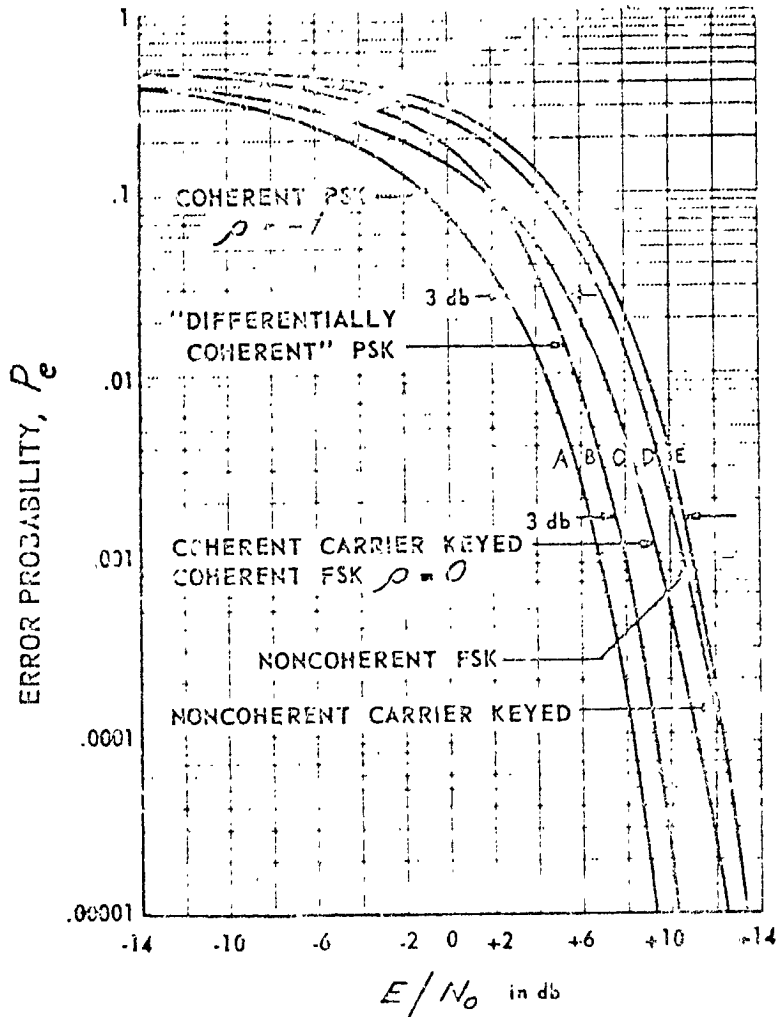


Figure 3-2 PROBABILITY OF ERROR vs SIGNAL ENERGY/NOISE POWER DENSITY

APPENDIX II

Factors to be Considered in Information Rate Determination

Range	250,000 S. Mi.	53,000,000 S. Mi.
Spacecraft transmitter power	25 watts	1 KW
Spacecraft transmission line loss	1.5 db	1.5 db
Frequency	2.3 KMC	2.3 KMC
Spacecraft antenna	5 ft parabola	10 ft parabola
Spacecraft antenna pointing loss	1.0 db	1.0 db
Ground station antenna	85 ft parabola	225 ft parabola
Ground station receiver bandwidth	As required for system under consid.	As required for system under consid.
Ground station excess noise temp. (Includes temperature of distant body, antenna, feed line and pre- amplifier)	330°K	30°K