INVESTIGATION OF A PAM TESTER USING PN WAVEFORMS

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Summary A PAM tester using a shift register generated sequence and a conventional data (PCM) bit synchronizer and detector for synchronization was investigated. It was found to function satisfactorily with and without predetection tape recording. This tester can be implemented by adding digital-to-analog converters (DAC's), an error gate, and a rms voltmeter to a PN PCM tester which uses a shift register to generate the PN sequence.

Introduction IRIG has standardized PCM end-to-end telemetry ground test procedures. The PCM test is based on a shift register generated PN sequence test pattern. The sequence generators and synchronizer used for the PCM test can be used for the PAM test by adding DAC's and an analog error measuring circuit.

The purpose of this paper is to provide system description, test results, and applications of the PN PAM tester.

Description of Test Equipment Figure 1 includes a simplified block diagram of a four stage shift register and a DAC connected to produce a PN PAM waveform as shown. The PN PAM waveform is inverted and balanced to zero. The PAM waveform is a 15 ary waveform with essentially the same autocorrelation properties as a binary PN reform and can be synchronized by either coherent or noncoherent methods .[1] The noncoherent method was used for these tests. The test data shows that noncoherent synchronization is satisfactory. This method of synchronization requires a replica of the binary PN sequence generated by the source pattern generator to be loaded in the receiver pattern generator. From Figure 1, it can be seen that the binary PN sequence can be obtained at the receiver by detecting the PAM data samples which include the most significant bit from the PN PCM generator. The output of the detector will be the binary PN sequence represented by the left-hand column of the word sequence shown in Figure 1.

Figure 2 is a system block diagram. The Schmitt trigger is set at zero threshold so that in the absence of noise, its output is the desired binary PN sequence except for the bit corresponding to the zero level PAM pulse. This bit is indeterminate and in the presence of balanced noise will be equally likely positive or negative. Thus, on the average, there is at least one error in thirty bits (two pattern sequences). The output of the Schmitt trigger is input to the bit synchronization and detector which outputs the binary PN sequence and a synchronized clock. The output of the bit synchronizer and detector is used to load and clock the receiver shift register. Note that only four correct bits of the binary PN sequence are required to load the receiver shift register. In the tests reported here, a load switch was activated manually while observing the rms meter or the oscilloscope for indication of synchronization of the received and local PN PAM signals. Loading can be automated by using the rms error as a criterion; the error is large when the receiver shift register is incorrectly loaded. The function of operational amplifier 2 is to subtract the locally generated PAM waveform from the received PAM waveform. The output of this amplifier is noise plus gliches. The gliches occur at the leading and trailing edges of each received PAM pulse, the gliches are due to misalignment, different rise and fall time, etc., of the input wavefoms. The function of the linear gate in the error detector is to sample the noise between the gliches. The sample period was 40 percent of the PAM pulse period. The requirements for this gate are

a. zero output when the gate is off,
b. zero output when the gate is on with zero input, and
c. reasonable linearity over the input voltage range.

The 6 diode balanced circuit[2] utilized satisfied these requirements.

Procedure In order to obtain a valid measurement of error due to distortion and noise, assuming synchronization lock, it is necessary to adjust the video gain of the receiver to match the reference waveform and to adjust the dc level of the two waveforms to a common level, nominally zero. These two adjustments are very easy to make by obtaining a minimum rms error reading on the error meter with a large input signal to the receiver.

The output from the DAC was balanced to ground by means of an operational amplifier (4 in Figure 2). When the video output of the receiver is in the ac mode, the receiver output signal is balanced about zero. In this case, the bias control on operational amplifier 2 in Figure 2 serves for the dc level adjustment. If the video output of the receiver is in the dc mode, then receiver tuning in the VFO mode can be used for dc level adjustment. If the video output is in the dc mode with AFC, then the bias control on operational amplifier 2 serves for dc adjustment. Whichever method is used, the adjustments can be made quickly and accurately.

Test Results Figures 3 and 4 give output (S/N) versus input carrier power in dBm for 500 kHz and 100 kHz linear IF filters in the ScientificAtlanta model 410A receiver. The video filter bandwidth was 50 kHz and the sample pulse rate was 25 x 10^3 per second in both cases and the carrier deviation was 240 kHz peak-to-peak with the 500 kHz IF for Figure 3 and 70 kHz peak-to-peak with the 100 kHz IF for Figure 4. The rms value of the output of the error gate as read on the rms meter with the output of the radio receiver removed was one volt rms. The ordinates of Figures 3 and 4 were obtained by dividing the rms error read from the rms meter into unity, i.e., by taking the reciprocal of the meter reading. Twenty times the log to the base ten of the reciprocal of the meter reading gives the S/N in dB for Figures 3 and 4.

The data points represented by dots are with no tape recorder and the data points represented by X's are for predetection recorded data on a Sangamo 3500 tape recorder. The data was recorded on Ampex 772-29Kll tape at a tape speed of 120 ips with the recorder in the servo control mode.

Figure 5 contains photographs of waveforms corresponding to data points in Figure 3 and Figure 6 contains photographs of waveforms corresponding to the X data points in Figure 4.

The bit synchronizer bandwidth settings for Figures 3 and 4 were 0.3 percent bandwidth and 1 percent tracking. The carrier power at which slippage occurred was nearly proportional to bit synchronizer bandwidth. For example, in Figure 3, slippage occurred at minus 106 dBm for 0.3 percent bandwidth and 1 percent tracking. Slippage occurred at minus 95 dBm for 3 percent bandwidth and 10 percent tracking. All data were taken with the bit detector in the filter and sample mode.

Data were obtained with a Microdyne model 1100-R(7) receiver and a DEI model TR 711 receiver with 500 kHz IF's. The data agreed within one dB or better to the data with the Scientific-Atlanta data using the same parameters. This shows that the PAM tester performs equally well with different receivers.

Discussion of Results Figures 3 and 4 show that synchronization can be maintained ratios well below carrier threshold, to levels so low that the output (S/N) in Figures 3 and 4 are nearly zero dB. These data are S/N for unprocessed baseband signals. In PAM practice, further processing of the baseband signals occurs. In order to interpret the unprocessed baseband S/N to obtain individual data channel output S/N, the improvement obtained by baseband processing must be added. In PAM practice, pulse integration is frequently used. It is shown in the Appendix that pulse integration gives about 10 dB improvement over the S/N ratios given in Figures 3 and 4. If, in addition, the individual data output channels are filtered by averaging over 5 samples, then an additional improvement of approximately 7 dB is obtained. Thus, the individual data channel output S/N relative to the S/N plotted in Figures 3 and 4 is obtained by adding the processing gain, 17 dB in the example, to the ordinates of Figures 3 and 4. This assumes that synchronization is maintained over the input carrier power range involved. Thus, slippage of the PAM tester synchronization occurs at corresponding output data S/N of less than 20 dB, i.e., more than 10 percent rms error for the example considered. This is believed to be more than adequate for PAM data.

The ratio of the IF bandwidths for Figures 3 and 4 is about 7 dB. Thus, the carrier powers corresponding to thresholds and slippage should be different by about 7 dB which is the case.

In the course of the measurements, it was observed that the receiver AFC became unstable at input signal strengths less than -98 dBm for both the 500 and 100 kHz IF bandwidths so that the PAM tester synchronization was lost at this power with the receiver in the AFC mode. Data for Figures 3 and 4 were obtained with the receiver in the VFO mode with manual tuning. In Figure 3, data for predetection recording for input power below -98 dBm were not obtained through an oversight. It is assumed that as in data with the 100 kHz IF bandwidth (Figure 4) synchronization would have been maintained down to -104 or -105 dBm.

In order to increase the confidence in the test data, the output S/N with the 500 kHz IF bandwidth is calculated in the Appendix. The calculated S/N is in good agreement with the data in Figure 3 for above threshold signal conditions. The corresponding calculation for the 100 kHz IF must take into account the IF gain characteristic and for this reason it was not done. Qualitatively, Figure 4 data are consistent with the data in Figure 3.

As can be seen from Figures 5 and 6, a noise floor existed in the system. It appeared to be of low frequency and, since it appeared to be about the same for several different makes of receivers, it was probably spurious frequency modulation in the Hewlett-Packard model 3205A Telemetry Signal Generator.

Application Since the synchronization is carried out by detecting the PAM data samples which include the most significant bit from the PN PCM generator, the test waveform from the DAC can be altered in any systematic way provided the most significant bit from the PN PCM generator is not changed. Thus, for example, it is possible to include a PAM frame synchronization pulse. This pulse can be inserted at any multiple of the frame period of the PAM tester by means of a counter and gating. If it is desired to simulate a PAM frame of 64 channels, then the frame synchronization pulse can be inserted every fourth sequence of the PAM tester pattern generator. With the four position pattern generator, this would provide a 60 channel frame. Or, a 6 position shift register could be used to generate a PAM test sequence of 63 channels. Only the 4 position shift register was investigated, but since it performed so well, it is assumed that a 6 position register would also perform satisfactorily.

An example of insertion of a frame synchronization pulse shape is shown in Figure 7. This is the same type of frame synchronization pulse used for the Navy PAM and is the synchronization pulse for 100 percent duty cycle PAM as defined in Telemetry Standards, IRIG 106-69.

The five PCM words corresponding to this waveform (inverted in sign) are 1111, 0001, 0001, 0001, and 1000. Reference to Figure I shows that this frame pulse can be obtained by substituting the two words 0001 and 0001 for the words 0111 and 0011 following the 1111 word. This can be done by simple logic.

The PAM tester can be used in different modes. If the mission is to provide a tape, the test pattern can be injected into a directional coupler and recorded. Then the tape is rewound and played back. The rms error versus input signal strength is measured and recorded. The frame synchronization word can be inserted into the test pattern and a section of the mission tape used to record the test pattern so that after tape dubbing, etc., the pattern can be used to check out the tape data quality together with the PAM data reduction equipment. The latter test can be carried out by the block diagram of Figure 2 plus coincidence (modulo two adders) detection for sample pulse clock and frame pulse synchronization. Or, the complete ground station can be tested by comparing the output of the individual data channels with the respective PAM samples generated by the local pattern generator.

It should be mentioned that in PCM decommutators, the frame synchronization word is detected digitally. Thus, frame synchronization acquisition can be tested, along with other sections of the decommutator, by use of a digital PCM simulator without the necessity of exercising the analog section of the system. In PAM, the frame synchronization may utilize an analog device and should be tested in connection with the analog test sequence.

References

1. Wade C. McClellan and M. H. Nichols, “Synchronization of Pseudo Noise Sequence for PCM Testing,” ITC Proceedings, vol. VI, p. 155; 1970.
2. Jacob Millman and Herbert Taub, “Pulse and Digital Circuits,” McGraw-Hill Book Co., Inc., New York, New York, p. 445; 1956.

APPENDIX

THEORETICAL CALCULATIONS

Pulse Integration In order to decrease the noise on the detected PAM samples, the pulse and noise are integrated over a window, cut out of the pulse, of Δt seconds in length. The window is suitable in length and spacing to reduce pulse overlap effects to the desired level. The output of the integrator is sampled at the end of each integrating period and fed to the decommutator (or computer).

The sampled integrator output can be modeled by a sampled aperture filter. The gain characteristic, A(f), of the aperture filter is given by

$$A(f) = \frac{\sin \pi \Delta t f}{\pi \Delta t f} \tag{1}$$

The ratio of the noise power, N_o, in the output of the aperture filter to the noise power, N_i, in the input is given by

$$\frac{N_o}{N_i} = \frac{\int_0^\infty A^2(f) G^2(f) S_{bb}(f)\,df}{\int_0^\infty S_{bb}(f) G^2(f)\,df} \quad , \tag{2}$$

where GM is the gain characteristics of the video filter, and $S_{bb}(f)$ is the baseband noise power spectrum. When well above threshold I the S_{bb} in the output of an FM receiver is given by kf where k is a constant.

Figure 8 gives the video gain characteristic, G(f), for the data in Figures 3 and 4 of the text. The numerical value of the integral in the degiminator of Equation (2), determined by graphical integration, is 1.2×10^{14}k (volts)^2. The integral in the numerator of Equation (2) is approximated by assuming G(f) = 1 from f = 0 to f = F_m the bandwidth of the video filter (50 kHz), and integrating over this range. For the test data in Figures 3 and 4 of the text, the sample period (Δt = 16 microseconds) was 40 percent of the PAM pulse period (40 microseconds). Substituting these parameters into Equation (2) gives

$$\frac{N_o}{N_i} \approx 0.1 \text{ or about } -10 \text{ dB} \quad . \tag{3}$$

Thus, the processing improvement is about 10 dB. In typical application the sample period for PAM data is 50 percent of the PAM pulse period. A 50 percent sample period would give approximately 2 dB more improvement over the 40 percent sample period used in the tests.

If an individual channel output low-pass filter which averages N-samples (where N is the sample rate per data cycle) were used, then there is an additional improvement of 10 log N dB. If N = 5, then the overall processing improvement would be (10 + 10 log 5) dB or about 17 dB.

Calculation of Signal To Noise (S/N) Theoretical S/N corresponding to the ordinate in Figure 3 of the text is calculated below. When well above threshold, the noise power in the output of an FM receiver is given by

$$N_o = \frac{H^2}{f_d^2 B} \left(\frac{N}{S}\right)_{carrier} \int_0^\infty f^2 G^2(f)\,df \quad , \tag{4}$$

where B = IF noise bandwidth, G(f) = relative gain characteristic of the video filter, and H = output voltage for peak deviation, f_d. From Figure 1 of the test, it is easy to show that for the PAM test sequence used, the mean square value, S_o , of the receiver output is $H^2/2.65$. Thus,

$$\frac{S_o}{N_o} = \frac{f_d^2 B}{2.65 \int_0^\infty f^2 G^2(f)\,df} \left(\frac{S}{N}\right)_{carrier} \quad . \tag{5}$$

For Figure 3 of the text, f_d = 120 kHz and the noise bandwidth of the receiver linear phase IF is 460 kHz. The relative gain characteristic of the video filter for data presented in Figure 3 is given in Figure 8. Numerical evaluation of the integral in Equation (4) gives a value of 1.2×10^{14} $(\mathrm{Hz})^3$. Substituting these values in Equation (5) gives

$$\frac{S_o}{N_o} = 21 \left(\frac{S}{N}\right)_{carrier} \quad , \tag{6}$$

or in dB

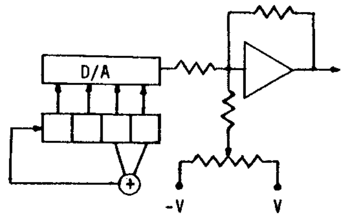
$$\left(\frac{S_o}{N_o}\right)_{dB} \simeq 13 + \left(\frac{S}{N}\right)_{dB\ carrier} \quad . \tag{7}$$

Assume that the noise figure of the receiver is 11 dB, then

$$\left(\frac{S}{N}\right)_{dB\ carrier} = S_{dBm} + 106 \quad , \tag{8}$$

where S_{dBm} is the input RF power corresponding to the abscissa of Figure 3 in the text. For S_{dBm} = -90 dBm, Equations (7) and (8) yield (S_o/N_o) = 29 dB. From Figure 3, the (S_o/N_o) is 28 dB, for input signal power of -90 dBm, which is in good agreement with the calculated (S_o/N_o).

For Figure 4 in the text, calculations cannot be made as simply since the IF bandwidth is only twice the video bandwidth and the IF is maximally linear phase. The gain characteristic of a maximally linear phase IF is not flat and would have to be taken into account.



WORD	LEVEL
1111	-7
0111	1
0011	5
0001	7
1000	0
0100	4
0010	6
1001	-1
1100	-4
0110	2
1011	-3
0101	3
1010	-2
1101	-5
1110	-6

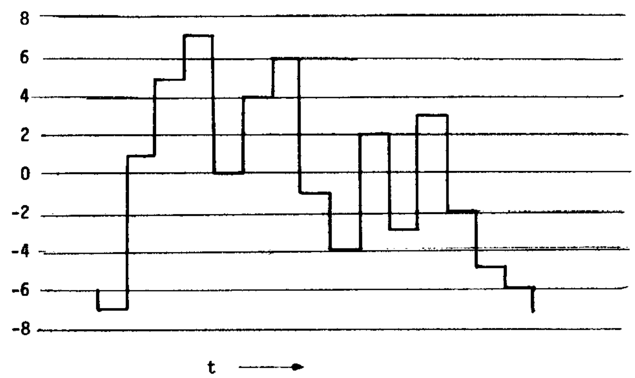


FIGURE 1. PAM TEST SEQUENCE FROM 4 POSITION SHIFT REGISTER

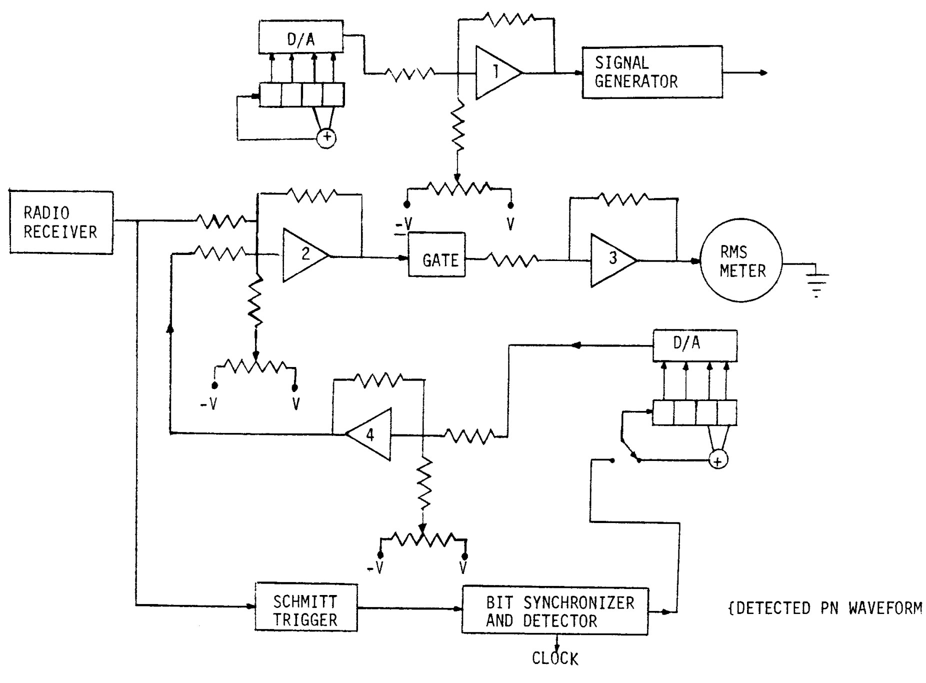


FIGURE 2. SYSTEM BLOCK DIAGRAM

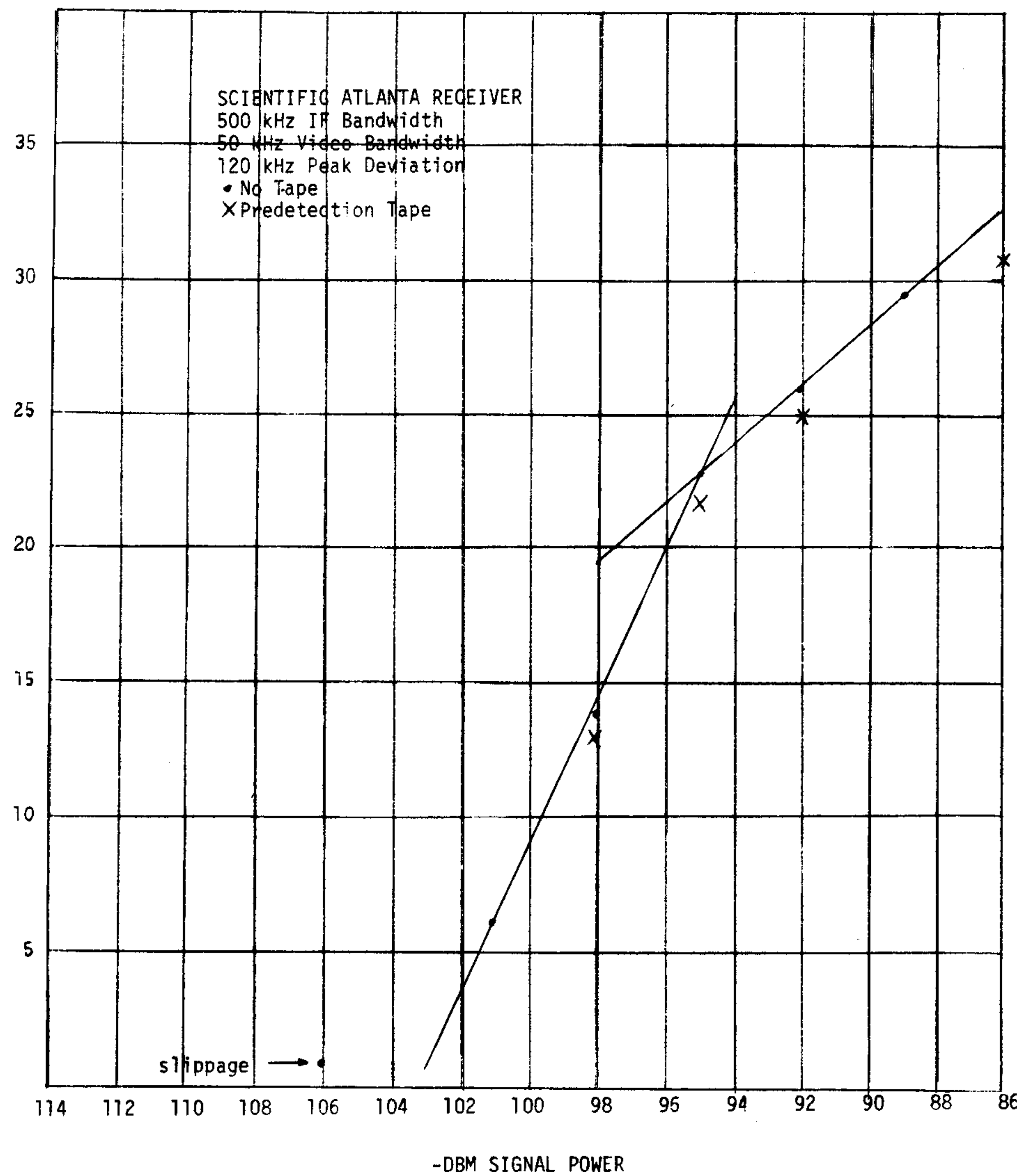


FIGURE 3. OUTPUT RMS (S/N) IN DB VERSUS INPUT SIGNAL POWER IN DBM

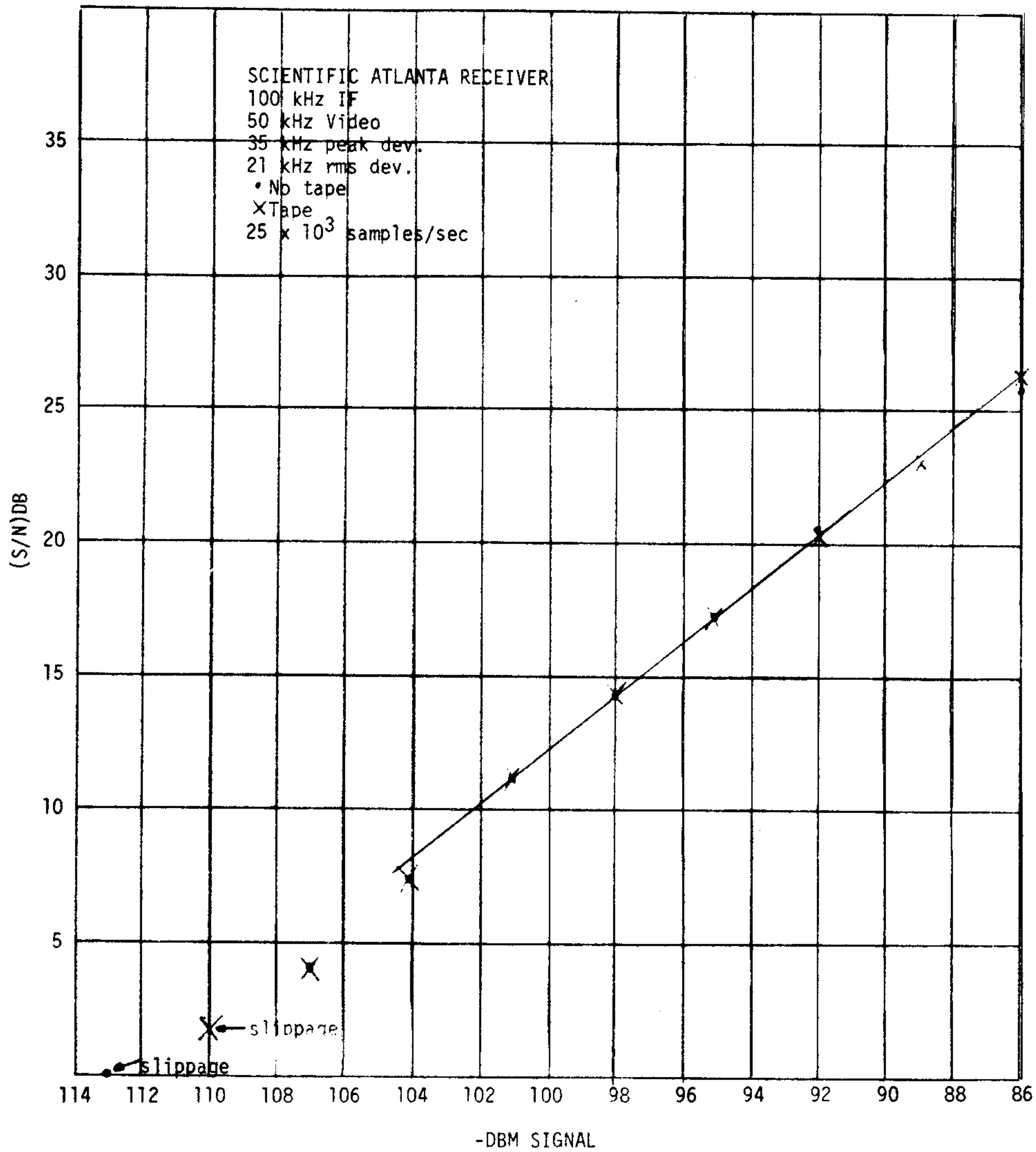


FIGURE 4. OUTPUT RMS (S/N) IN DB VERSUS INPUT SIGNAL POWER IN DBM

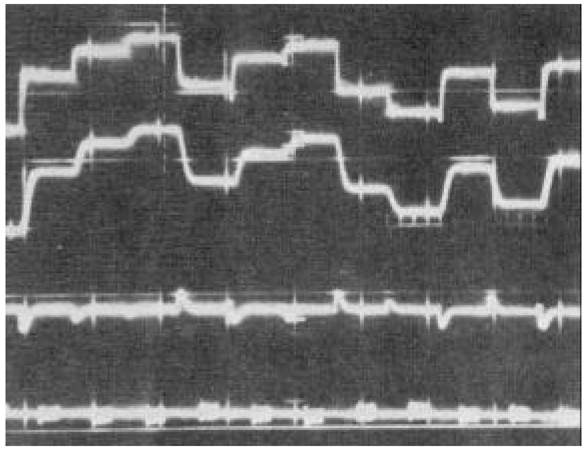


FIGURE 5A. INPUT TO THE RECEIVER AT -56 DBM

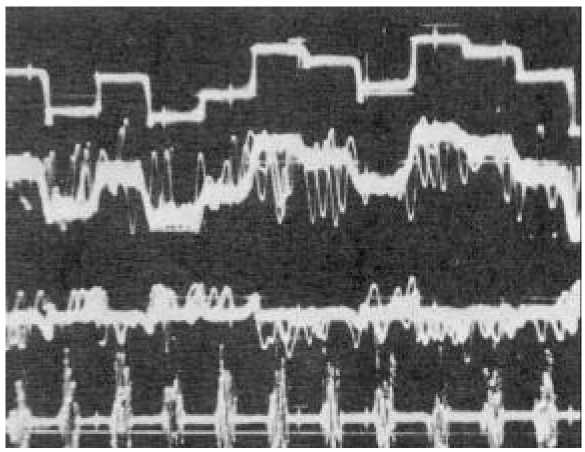


FIGURE 5B. INPUT TO THE RECEIVER AT -101 DBM

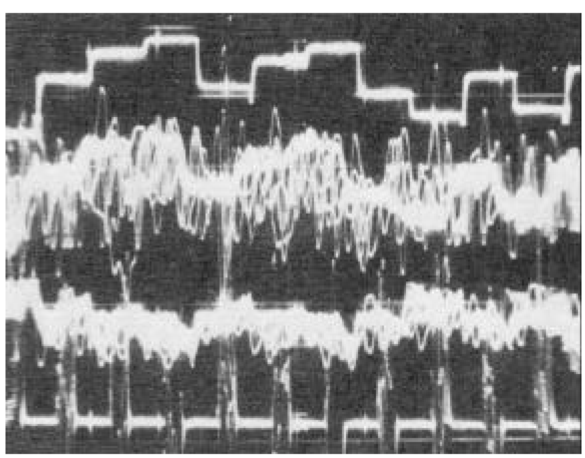


FIGURE 5C. INPUT TO THE RECEIVER AT -106 DBM

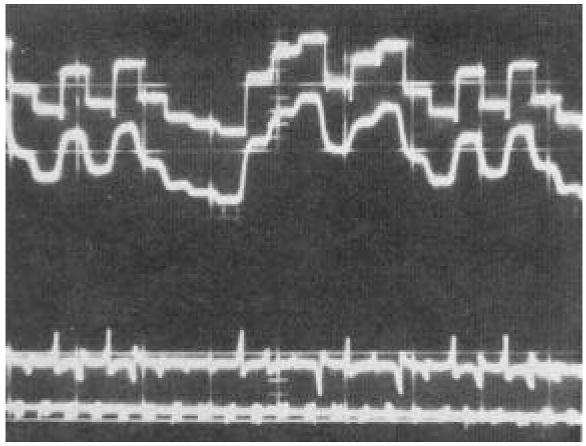


FIGURE 6A. TAPE OUTPUT WITH -56 DBM INPUT TO RECEIVER

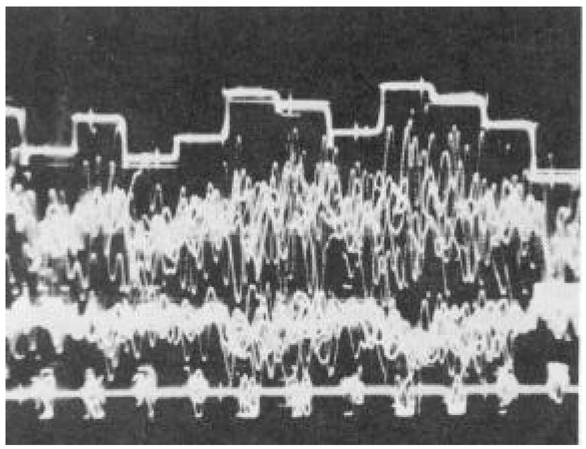


FIGURE 6B. RECEIVER OUTPUT WITH -113 DBM INPUT TO RECEIVER

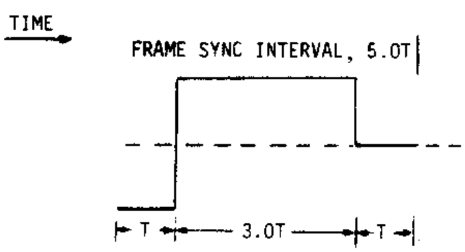


FIGURE 7. PAM FRAME SYNCHRONIZATION PULSE

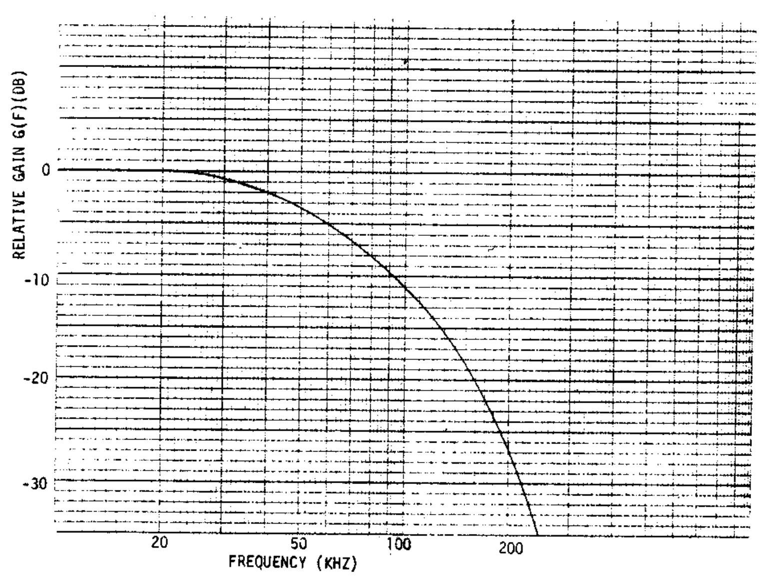


FIGURE 8. VIDEO OUTPUT RELATIVE GAIN CHARACTERISTIC