SYNCHRONIZATION OF PSEUDO NOISE SEQUENCES FOR PCM TESTING

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Summary Coherent and noncoherent methods of synchronizing PN sequences for testing PCM telemetry receiving stations are compared. Test results are given for each method using a typical range S-band receiver, bit synchronizer and tape recorder. Effects of time-base-error from the tape are calculated and checked by test results. The laboratory tests indicated that for bit-error probabilities less than 0.01, the noncoherent synchronizer functioned satisfactorily.

Introduction The Telemetry Working Group of the IRIG is currently in the process of preparing standard procedures for telemetry receiving station testing. The procedures are more or less along the lines proposed by Nichols[1] and later modified to include the solar calibration proposed by Hedeman[2]. It is expected that test results based on these procedures will define a performance interface between the range and the user and will also serve as a performance figure-of-merit for individual receiving stations, thus removing a great deal of guess-work from mission planning.

Figure 1 is a block diagram of the PCM test being considered by IRIG. By means of the solar calibration data, the power output reading, P_s, from the RF signal generator can be expressed in terms of the equivalent power per unit area, J_a, of stated polarization incident on the receiving antenna. Power per unit area of stated polarization is the quantity delivered by the user. The purpose of the PCM test is to determine the bit error probability, P_e as a function of J_a, or its equivalent, P_s. As indicated in Figure 1 it is necessary to have available at the error detector an accurate and synchronized replica of the PCM test pattern which modulated the RF signal generator. The PCM test pattern should be completely defined and should be sufficiently random to simulate a data bit stream. In addition, particularly for the NRZ format, there should be periods of sufficient length with few bit transitions to show up poor low-frequency response and to exercise

the data bit synchronizer. A pseudo noise, PN, binary sequence satisfies these requirements, is easily generated, and has the added advantage that its autocorrelation properties permit easy synchronizatio[3]. This paper reports tests utilizing the PN method of generating and synchronizing the test pattern.

Figure 1 indicates a bit synchronizer and detector under test. As pointed out in reference 1, in a PCM system all operations downstream from the bit detector are purely digital and the setup can be completely rung out, if necessary, without the need for exercising the analog parts of the system. In some cases where the telemetry signal is recorded either pre-carrier-detection or post-carrier-detection, a bit synchronizer and detector may not be available. In this case, it may be desirable to equip the synchronizer with its own bit detector so that the system can be tested from RF input to tape output.

The tests reported in this paper were not intended to evaluate the signal generator, receiver, tape recorder, and bit synchronizer and detector used, but rather to obtain data on the coherent and noncoherent methods of PN synchronization.

The Coherent Method of Synchronization The coherent method of synchronization contains no nonlinear operations as long as the phase-lock loop, PLL, is above threshold. Figure 2 is a block diagram of a coherent synchronizer using a 7 position shift register. The balanced binary PN waveform input is generated by an identical 7 position shift register. Since the clock in the PLL is not initially synchronized, the two waveforms “run past” each other. The phase sensing circuit in Figure 2 provides an “SI’ shaped phase characteristic which permits the PLL to lock the two waveforms together in phase.

The functioning of the phase-sensing circuit is based on the autocorrelation property of FIN waveforms. Assuming that the input balanced NRZ PCM has ideal rectangular pulses (i.e., wide band), the cross-correlation between the input waveform and the locally generated replica is the same as the autocorrelation of the waveform. This is shown by the top line of Figure 3 (see reference 3, page 69). The width of the first triangular characteristic is two bit periods. The second triangular characteristic in Figure 3 is the same function delayed two bit periods and inverted. The sum of the two is shown in the third line. This phase characteristic permits the PLL to acquire and track (see reference 3, page 88). In order to obtain the cross-correlation, it is necessary, to multiply the two PCM waveforms together and to average at least over one sequence period. The averaging is done by the PLL. Thus, the closed loop bandwidth must be less than the repetition rate of the PN sequence. If the output PCM waveform has been filtered so that its edges are rounded, the phase characteristic is rounded but still retains the necessary “S” shape. Since the phase sensing circuit contains no nonlinear operations, the PLL threshold determines the system locking threshold.

In the test hardware used to obtain the results reported here, the multipliers consisted of linear quad-gates as shown in the block diagram in Figure 4. In Figure 4, x(t) is the locally generated PN waveform. When the input to the quad-gate is positive (i.e., a binary 1) the gate is turned on and when it is negative, it is turned off. A high gain PLL was used[4] with a second Bode plot breakpoint (integrator) to provide additional loop filtering of the waveform from the gates. After the desired undamped natural frequency, f_n, and damping ratio were obtained by setting the gain and the first breakpoint frequency, the second breakpoint frequency was set at a value which slightly reduced the damping in the transient response of the loop. In the test hardware shown in Figure 2, a sampling bit detector was used because the tests were run on narrow band PCM. Provision was made for adjusting the time of the sampling strobe (clock) so that it occurred at the peak of the bit pulses. In practice, this adjustment was made so as to minimize the error count.

The Noncoherent Method of Synchronization This method makes use of the clock from a data it synchronizer. Since data bit transitions occur at random, a nonlinear operation is required in the data bit synchronizer in order to generate discrete frequency components at the bit rate and its harmonics. A PLL is used to filter out the desired component for the bit clock. The bit clock is used to drive the bit detector and decommutator. Figure 5 is a block diagram of the noncoherent PN synchronizer. The clock from the bit synchronizer is used to drive the shift register and the detected bit stream out of the bit detector is used to load the shift register when the switch is in the load position. After the shift register is loaded, the switch is thrown to the feedback position and, if there are no errors in the loaded bits, the shift register is in lock with the input PN waveform. It will stay in lock until bit slippage occurs in the bit synchronizer. If one or more errors is loaded into the register, the shift register is not locked, and because of the autocorrelation properties of the FIN sequence the error rate is nearly 50 percent. For example, if the error probability is 0.01, it is easy to tell if the register is locked.

If it is not locked, the switch is returned to the load position and the process repeated until lock is obtained. In the test hardware, the switch was used in conjunction with logic elements to remove the switch bounce and to transfer conduction on the next clock pulse following switch closure. In operational hardware, the whole loading operation can be done automatically.

Test Results Figure 6 gives test results using the coherent method of synchronizing the local shift register. A 7 position register connected identically to that shown in Figure 2 was used to generate the PN waveform at 70 kilobits/second, kb/sec. The balanced PN waveform was fed to a Hewlett-Packard model 3205A RF generator operating in S-band with a peak-topeak deviation of 50 kHz. The output of the RF generator was fed to a Defense Electronic, Inc., model TR-711 S-band receiver with an IF bandwidth of

100 kHz. The video output filter was set for 50 kHz. The video output of the receiver was recorded at 60 ips on a Sangamo 3500 recorder using Ampex 772 tape. The natural frequency of the coherent synchronizer was set at approximately 120 Hz and the damping ratio was set at about 0.5. Both quantities were estimated from the transient response of the closed loop error voltage when the clock frequency of the input PN waveform was subjected to a step change.

In Figure 6, using predetection recording, the dots are the error probabilities as measured at the output of the bit detector shown in Figure 2. The coherent synchronizer showed no signs of slippage over the dBm range shown. The threshold for acquiring lock was about -114 dBm. By clamping the receiver AGC, the signal generator input power required to double the IF output power compared to the no-signal output power was -113 dBm. In Figure 6 the crosses are data obtained using an Electro-Mechanical Research, EMR, model 2721 bit synchronizer and detector. The detected bit stream out from the EMR synchronizer and detector was compared with the shift register output. The bit clock from the EMR synchronizer and detector lagged the clock of the coherent synchronizer by about 2 microseconds compared to the bit period of 15 microseconds. The error detector strobe was set at the center of the overlap of the two periods. The loop bandwidth of the EMR synchronizer and detector was set at 0.3 percent in the filter and sample mode. The bit frequency was set at 70 kb/sec. The principal reason that the EMR synchronizer and detector gave a higher bit error probability, under the narrow band test conditions stated above, appeared to be that the sample strobe did not occur at the peak of the bit pulses in the input waveform. Because the EMR synchronizer and detector was borrowed and the case sealed it was impossible to verify this directly. However, this hypothesis was tested by detuning the EMR VCO until sufficient phase error was developed to shift the strobe to minimize the bit error probability. This brought the EMR synchronizer and detector error probability into line with that from the coherently synchronized tester at -100 dBm, but set the EMR synchronizer at the edge of its tracking range. This result indicates the desirability of providing an adjustment on the bit synchronizer for varying the phase of the sample strobe, when operating with narrow band PCM.

Figure 7 gives test results for the noncoherent synchronizer using the same test conditions as for Figure 6, except that the test was run with and without the base band tape recorder. Although slippage was apparent at -108 dBm, it was possible to find intervals without slippage which could be used for evaluating P_e. For values of $P_e = 0.01$, or less, slippage did not occur over a number of minutes of observation.

Discussion of Results If the PCM telemetry mission support requirements are such that bit error probabilities greater than 0.01 produce useless data, then the results of Figure 7, taken under the conditions and with the hardware stated, indicate that the noncoherent synchronization method is satisfactory for testing PCM receiving stations

with and without tape recording. If more sophisticated coding is to be used, then it may be necessary to test to higher bit error probabilities in which case coherent synchronization may be required. If no bit synchronizer and detector is normally provided at the test site, then the coherent synchronizer may be more economical to acquire and to operate.

When using tape recorders, the time-base-error, TBE, is an important phenomenon in the sychronization process. Figure 8 shows the TBE spectrum of the Sangamo 3500 recorder. This plot was derived from data taken by a MICOM 830OW flutter meter. Figure 8 is plotted in dB below a second per Hz versus frequency in Hz. Over the frequency range 20 Hz to 8 kHz the characteristic can be approximated roughly by $G_n(f) = 2 \times 10^{-9}/f^3$. Assuming that the tape transport servo control is second order with f_n of 20 Hz, it may be assumed that $G_n(f) = 1.2 \times 10^{-14} f$ for $f < 20$ Hz. Using the approximate method of reference 5, the untracked rms TBE is given approximately in Figure 9 as a function of the f of the PLL. With $f_n = 120$ Hz used for the data in Figure 6, the rms untracked TBE from Figure 9 is 0.4 microseconds, approximately. Figure 10 is a photograph of the oscilloscope with expanded time scale of the output of the tape recorder with the sweep triggered by the coherent synchronizer clock. The exposure was for one second which is about 550 sweeps. The time scale is I microsecond per small division. The peak-to-peak untracked TBE is a little over 2 microseconds which is the width of the steeply rising part of the waveform from the tape. This is in agreement with an rms value of 0.4 microsecond rms.

A 7 position shift register, producing a sequence 127 bits long, was used for the data in Figure 6. At the rate of 70 kb/sec, the sequence frequency was 550/sec. Thus, with f_n = 120 Hz, sufficient averaging was provided to produce a smooth phase error characteristic. Also, a small untracked TBE was produced compared to the bit period. In order to determine the effects of poor low frequency response of tape recorders, etc., on P_e it is desirable to use a sequence length comparable to the actual frame length. For example, an 11 position shift register will produce a sequence 2047 bits long. At 70 kb/sec the sequence repetition frequency is 34/sec. This would require that $f_n < 34$ Hz in order to provide a smooth phase error characteristic. But according to Figure 8, this would produce about 1.2 microseconds rms untracked TBE. Thus, with the coherent synchronizer, there is an upper limit on sequence length, for a given bit rate, which can be used satisfactorily with a given TBE characteristic. Further tests are required to evaluate this limit. Since the noncoherent synchronizer tracks individual bit pulse zero crossings, the sequence length considerations do not apply directly to the TBE consideration. The sequence length does have an effect on the performance of the data bit synchronizers because in NRZ waveforms the number of bits with no transitions depends on sequence length and feedback connections.

With the noncoherent synchronizer, the shift register length enters into the probability of error, P_r, in the loading of the shift register. The probability of one or more errors is $P_r = 1 - (1 - P_e)N$, where N is the number of positions in the shift register and P_e is the bit error probability. When expanded, this gives

$$P_r = NP_e - N(N - 1)P_e^2/2! + N(N - 1)(N - 2)P_e^3/3! - \ldots$$

If T is the length of time to load and test for errors, the average time, $\overline{T}$, to get synchronization is $T/(1 - P_r)$. If it is assumed that the test for errors to determine if the register is correctly loaded extends over one sequence, then T is approximately one sequence length. Thus if $P_e = 0.01$ and N = 11, the probability of error in loading is 0.11 approximately and $\overline{T} \simeq 1.1T$. if $P_e = 0.05$, the probability of error is about 0.42 and $\overline{T} \simeq 1.6T$ which is acceptable. However, according to Figure 7, with the particular hardware used, bit slippage limits the P_e test values to less than 0.05.

Tests reported in this paper were run on NRZ waveforms because these present the greatest difficulty in bit synchronization since in periods of no bit transitions, a data bit synchronizer operates open loop. Split-phase waveforms present at least one transition per bit period. Also, the power spectral density of random split-phase waveforms is zero in the neighborhood of zero frequency. Therefore, the effects of lack of dc response in baseband tape recorders, radio links, etc., are eliminated. Of course, split-phase waveforms require more than twice the bandwidth required for NRZ waveforms for the same data rate so that other performance factors must be considered.

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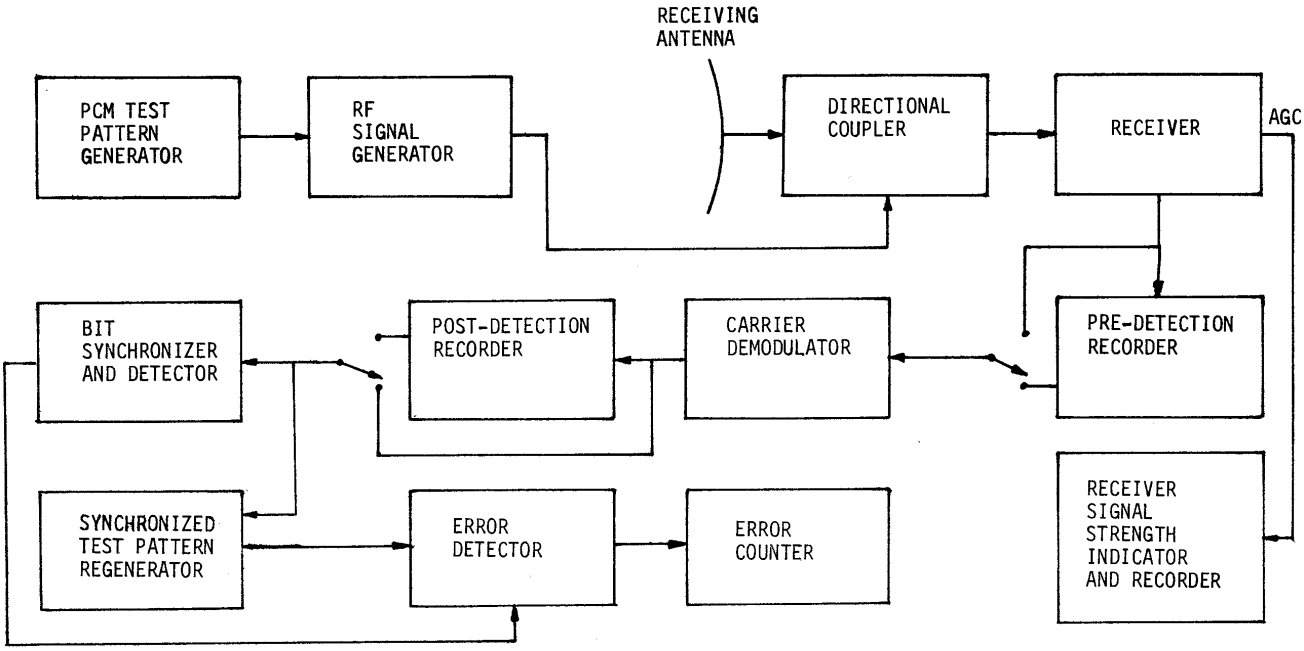


FIGURE 1. BLOCK DIAGRAM OF PCM RECEIVING STATION TESTER.

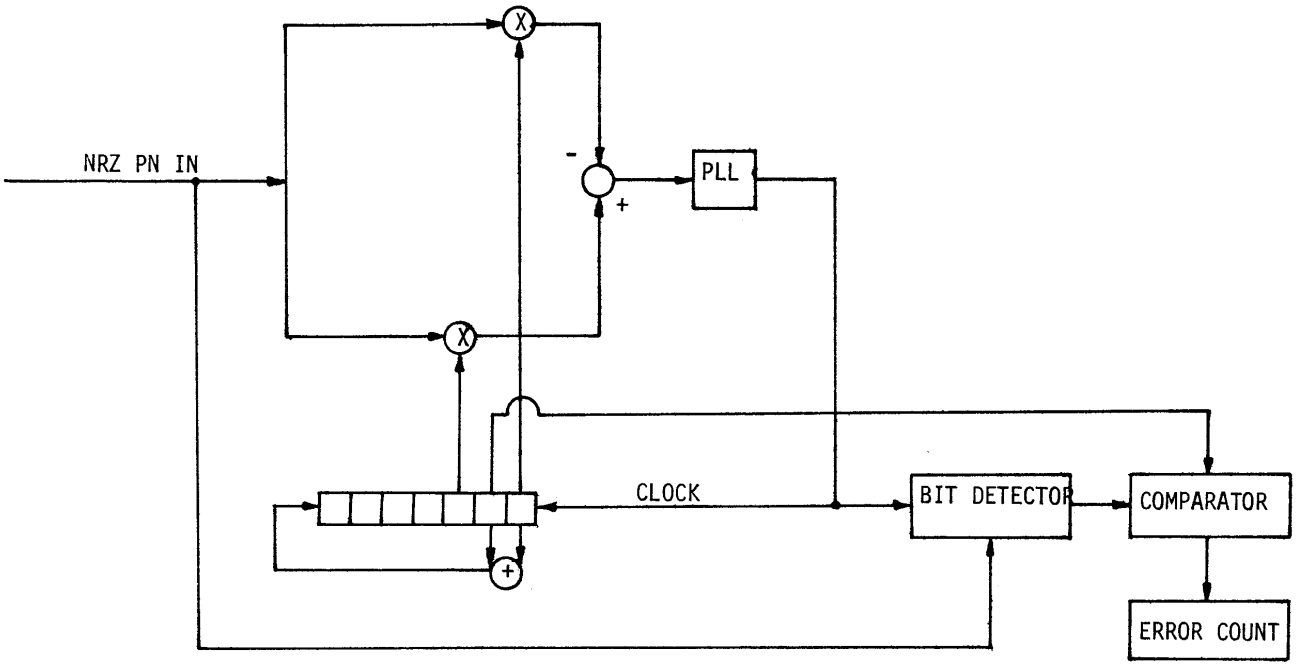


FIGURE 2. BLOCK DIAGRAM OF A COHERENTLY SYNCHRONIZED NRZ PCM TESTER USING A PN SEQUENCE TEST PATTERN. (THE SYMBOL ⊕ REPRESENTS MODULE 2 ADDITION.)

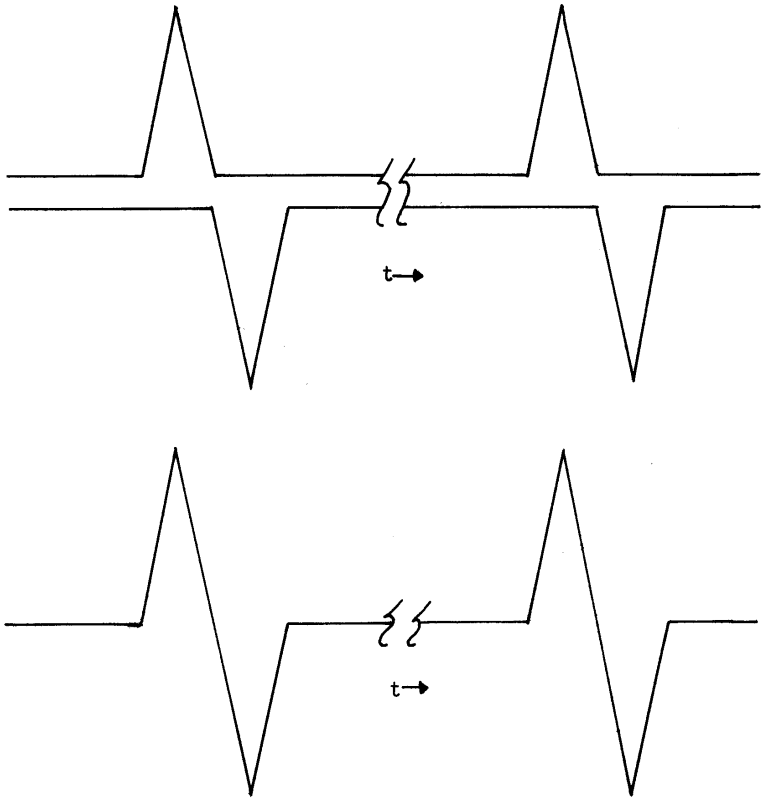


Figure 3. PN Autocorrelation Functions Showing Combination to Produce a Phase Characteristic Which Can be Tracked by a Phase-Lock Loop.

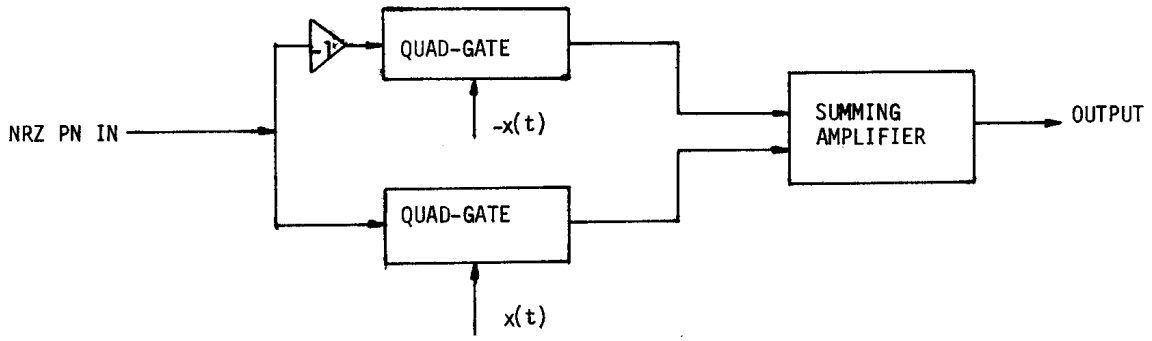


Figure 4. Block Diagram of the Linear Multipliers in the Coherent Synchronizer.

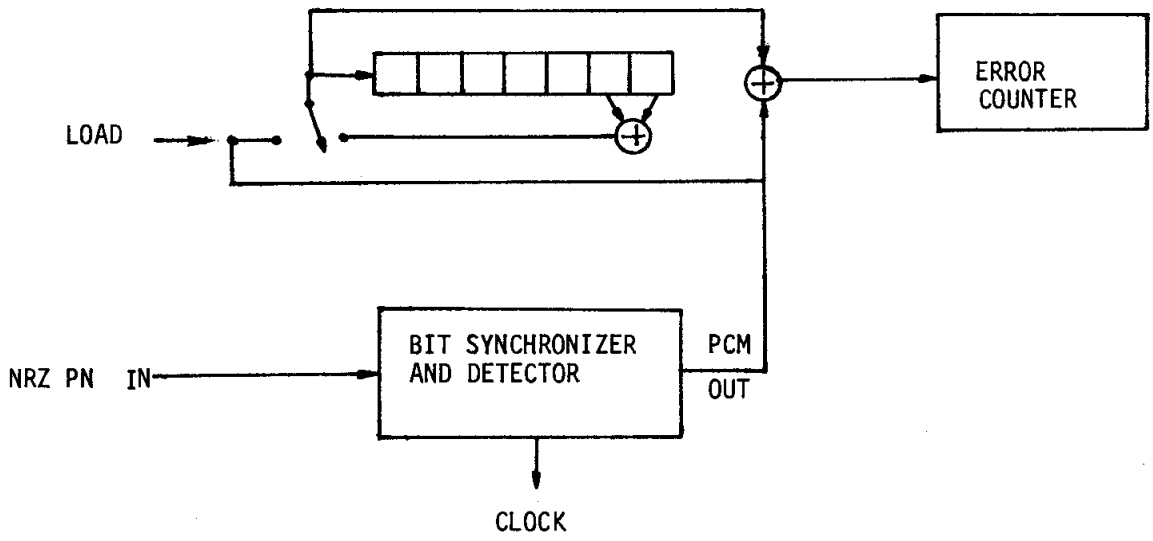


FIGURE 5. BLOCK DIAGRAM OF A NONCOHERENTLY SYNCHRONIZED NRZ PCM TESTER USING A PN SEQUENCE TEST PATTERN.

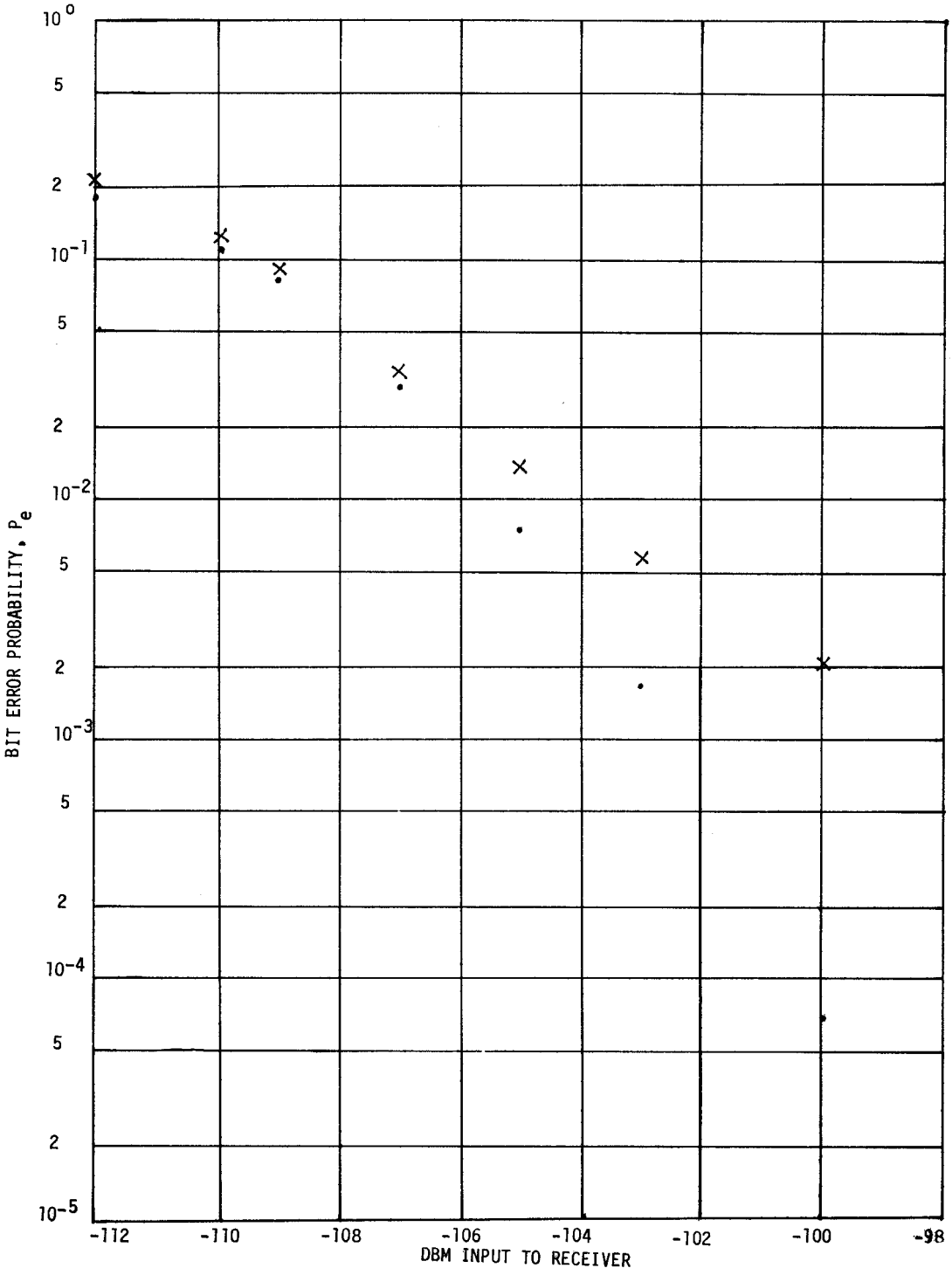
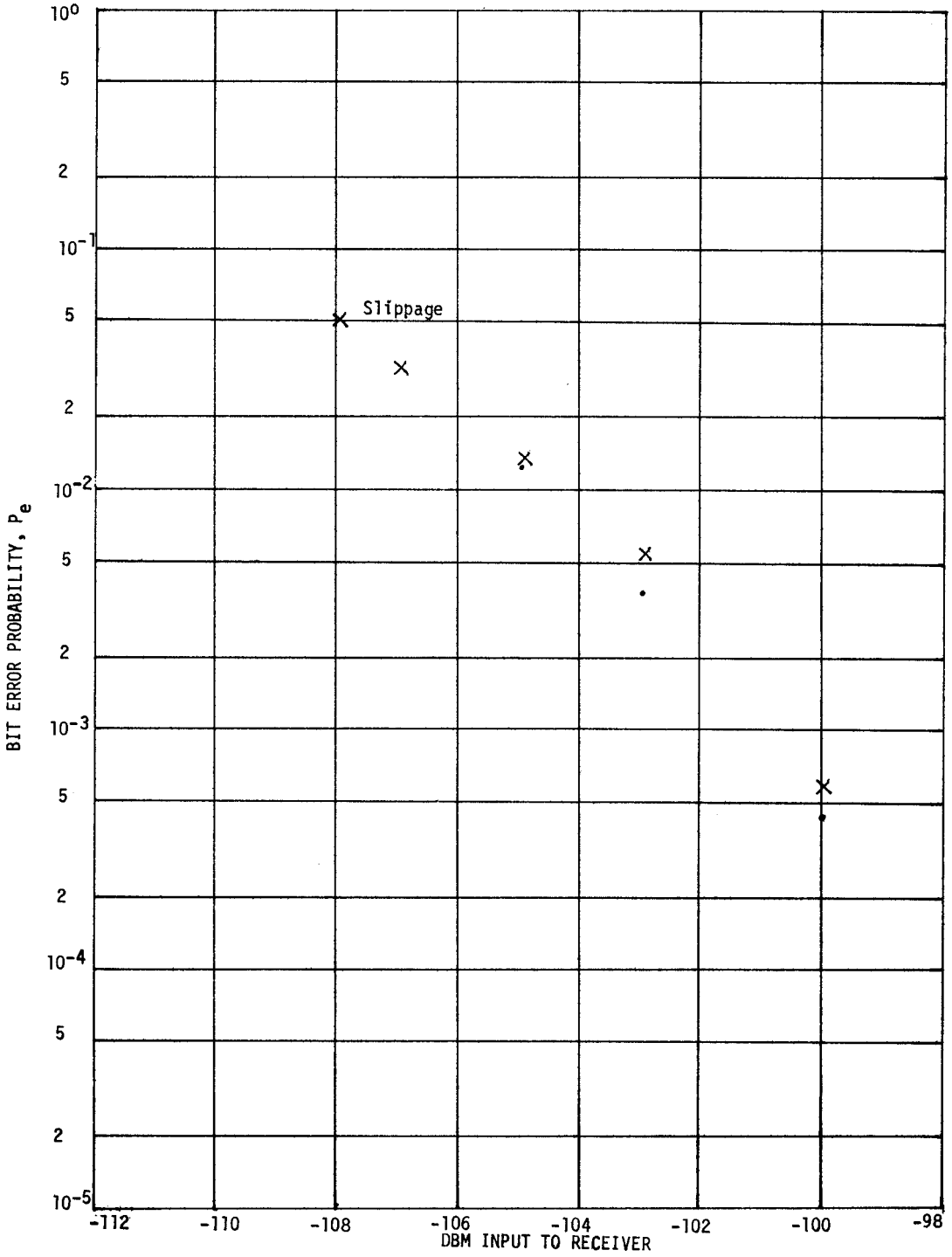


Figure 6. Bit Error Test Data Obtained With the Coherently Synchronized Tester. (The dots are data from the tester of Figure 2 and the crosses are from a data bit synchronizer and detector under test. The PN output of the telemetry receiver was tape recorded and played back under servo control into the tester.)



Bit Error Test Data Obtained With the Noncoherently Synchronized Tester. (The dots were measured with the PN waveform directly out of the telemetry receiver and the crosses were measured with the output of the receiver tape recorded and played back under servo control into the tester.)

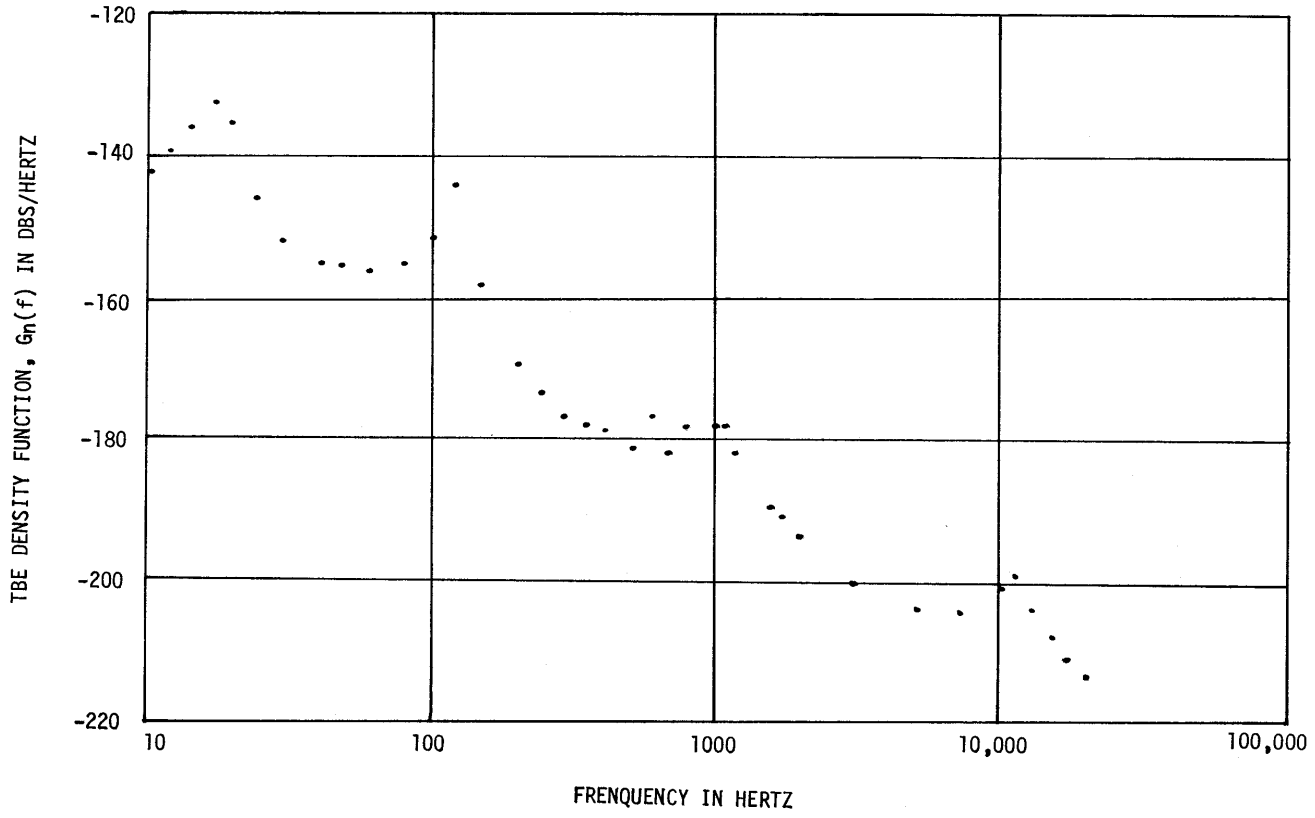


FIGURE 8. MEASURED TIME BASE ERROR POWER SPECTRAL DENSITY. (THE ORDINATE IS POWER SPECTRAL DENSITY IN UNITS OF dB BELOW A SECOND PER Hz-- i. e., 20 LOG_{10} (SEC) PER Hz.)

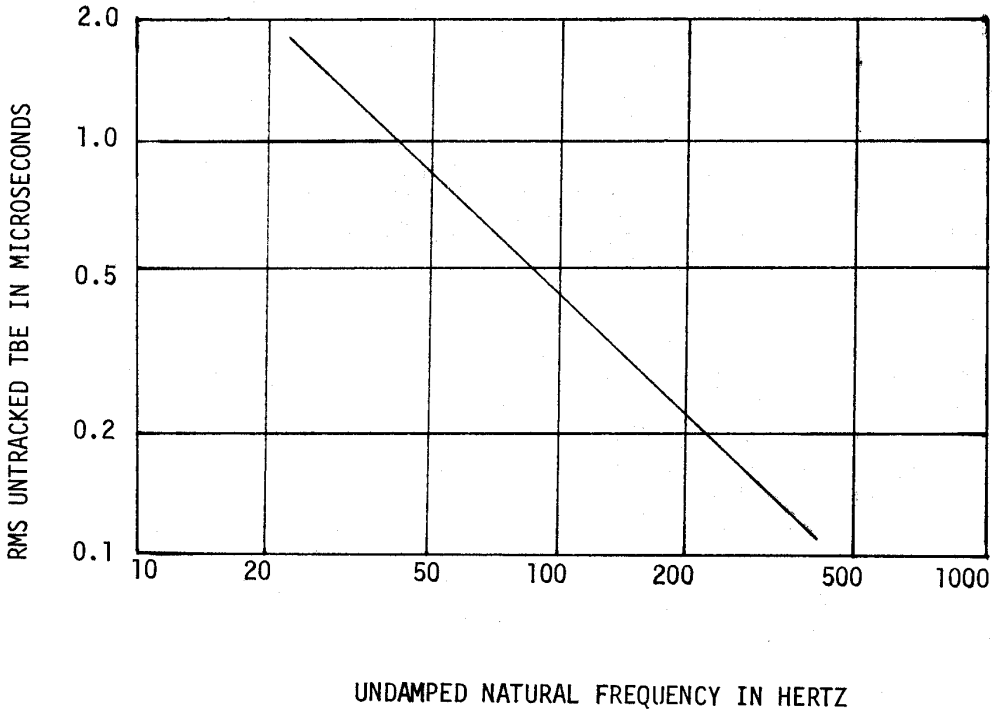


Figure 9. Calculated Untracked RMS Time Base Error as a Function of PLL Undamped Natural Frequency f_n.

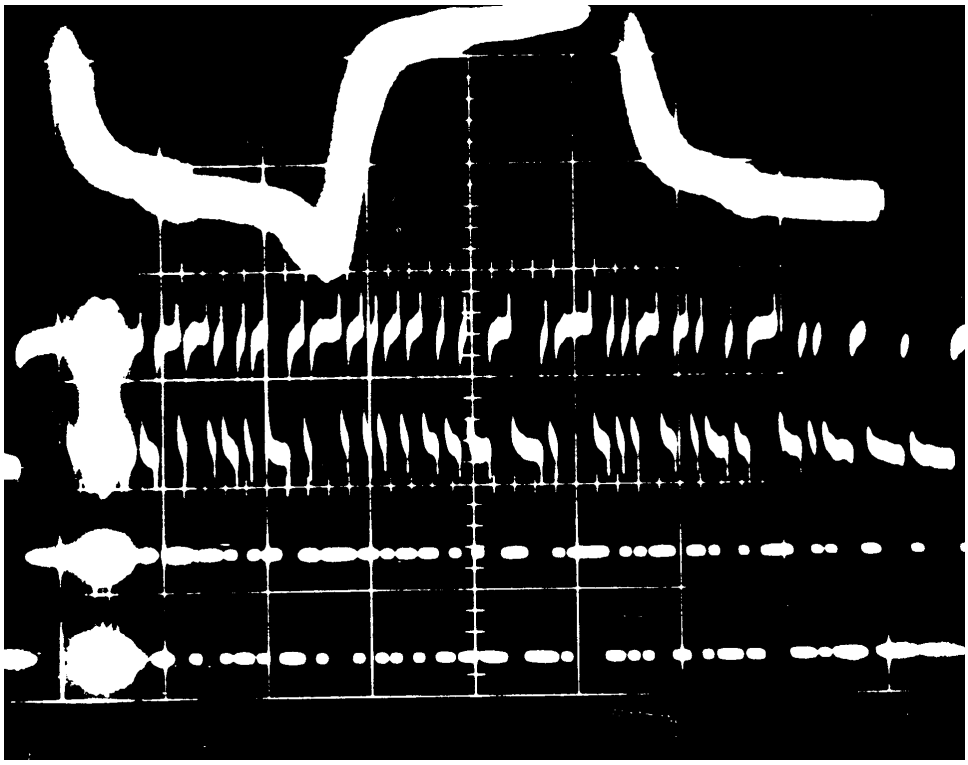


FIGURE 10. PHOTOGRAPH OF PN WAVEFORM FROM THE TAPE PLAYBACK. (THE HORIZONTAL THICKENING OF THE TRACE IS A MEASURE OF UNTRACKED TIME-BASE-ERROR.)