

APPENDIX A

FREQUENCY CONSIDERATIONS FOR TELEMETRY

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1.0 Purpose

This appendix was prepared with the cooperation and assistance of the Range Commanders Council (RCC) Frequency Management Group (FMG). This appendix provides guidance to telemetry users for the most effective use of the ultra high frequency (UHF) telemetry bands, 1435 to 1535 MHz, 2200 to 2290 MHz, and 2310 to 2390 MHz. Coordination with the frequency managers of the applicable test ranges and operating areas is recommended before a specific frequency band is selected for a given application. Government users should coordinate with the appropriate Area Frequency Coordinator and commercial users should coordinate with the Aerospace and Flight Test Radio Coordinating Council (AFTRCC). A list of the points of contact can be found in the National Telecommunications and Information Administration's (NTIA) Manual of Regulations and Procedures for Federal Radio Frequency Management.

2.0 Scope

This appendix is to be used as a guide by users of telemetry frequencies at Department of Defense (DOD)-related test ranges and contractor facilities. The goal of frequency management is to encourage maximal use and minimal interference among telemetry users and between telemetry users and other users of the electromagnetic spectrum.

2.1 Definitions. The following terminology is used in this appendix.

Allocation (of a Frequency Band). Entry of a frequency band into the Table of Frequency Allocations¹ for use by one or more radio communication services or the radio astronomy service under specified conditions.

Assignment (of a Radio Frequency or Radio Frequency Channel). Authorization given by an administration for a radio station to use a radio frequency or radio frequency channel under specified conditions.

Authorization. Permission to use a radio frequency or radio frequency channel under specified conditions.

¹ The definitions of the radio services that can be operated within certain frequency bands contained in the radio regulations as agreed to by the member nations of the International Telecommunications Union. This table is maintained in the United States by the Federal Communications Commission and the NTIA.

Certification. The Military Communications-Electronics Board's (MCEB) process of verifying that a proposed system complies with the appropriate rules, regulations, and technical standards.

J/F 12 Number. The identification number assigned to a system by the MCEB after the Application for Equipment Frequency Allocation (DD Form 1494) is approved; for example, J/F 12/6309 (sometimes called the J-12 number).

Resolution Bandwidth. The -3 dB bandwidth of the measurement device.

2.2 Other Notations. The following notations are used in this appendix. Other references may define these terms slightly differently.

$B_{99\%}$	Bandwidth containing 99 percent of the total power
$B_{-25\text{dBm}}$	Bandwidth containing all components larger than -25 dBm
$B_{-60\text{dBc}}$	Bandwidth containing all components larger than the power level that is 60 dB below the unmodulated carrier power
dBc	Decibels relative to the power level of the unmodulated carrier
f_c	Assigned center frequency

3.0 Authorization to Use a Telemetry System

All RF emitting devices must have approval to operate in the United States and Possessions (US&P) via a frequency assignment unless granted an exemption by the national authority. The NTIA is the President's designated national authority and spectrum manager. The NTIA manages and controls the use of radio frequency (RF) spectrum by federal agencies in US&P territory. Obtaining a frequency assignment involves a two step process: RF spectrum support certification of major RF systems design, and operational frequency assignment to the RF system user.

3.1 RF Spectrum Support Certification. All major RF systems used by federal agencies must be submitted to NTIA, via the Interdepartmental Radio Advisory Committee, for system review and spectrum support certification prior to committing funds for acquisition/procurement. During the system review process, compliance with applicable RF standards, and RF allocation tables, rules, and regulations is checked. For Department of Defense (DOD) agencies, and in support of DOD contracts, this is accomplished via the submission of a DD Form 1494 to the Military Communications and Electronics Board. Noncompliance with standards, the tables, rules, or regulations can result in denial of support, limited support, or support on an unprotected non-priority basis. All RF users must obtain frequency assignments for any RF system (even if not considered major). This is accomplished by submission of frequency use proposals through the appropriate frequency management offices. Frequency assignments may not be granted for major systems that have not obtained spectrum support certification.

3.1.1 Frequency Allocation. As stated before, telemetry systems must normally operate within the frequency bands designated for their use in the National Table of Frequency Allocations. With sufficient justification, use of other bands may at times be permitted, but the certification process is much more difficult, and the outcome is uncertain. Even if certification is granted (on a noninterference basis to other users), the frequency manager is often unable to grant assignments because of local users who will receive interference.

3.1.1.1 Telemetry Bands. Air- and space-to-ground telemetering is allocated in the UHF bands 1435 to 1535, 2200 to 2290, and 2310 to 2390 MHz, commonly known as the L band, the S band, and the upper S band.

3.1.1.2 VHF Telemetry. The very high frequency (VHF) band, 216-265 MHz, was used for telemetry operations in the past. Telemetry was moved to the UHF bands as of 1 January 1970 to prevent interference to critical government land mobile and military tactical communications. Telemetry operation in this band is strongly discouraged and is considered only on an exceptional case-by-case basis.

3.1.2 Technical Standards. The MCEB and the NTIA review proposed telemetry systems for compliance with applicable technical standards. For the UHF telemetry bands, the current revisions of the following standards are considered applicable:

RCC Document IRIG 106, Telemetry Standards

MIL-STD-461, Requirements for the Control of Electromagnetic Interference Emissions and Susceptibility

Manual of Regulations and Procedures for Federal Radio Frequency Management (NTIA).

Applications for certification are also thoroughly checked in many other ways including necessary and occupied bandwidths, modulation characteristics, reasonableness of output power, correlation between output power and amplifier type, and antenna type and characteristics. The associated receiver normally must be specified or referenced. The characteristics of the receiver are also verified.

3.2 Frequency Authorization. Spectrum certification of a telemetry system verifies that the system meets the technical requirements for successful operation in the electromagnetic environment. However, a user is not permitted to radiate with the telemetry system before requesting and receiving a specific frequency assignment. The assignment process takes into consideration when, where, and how the user plans to radiate. Use of the assignments is tightly scheduled by and among the individual ranges to make the most efficient use of the limited telemetry radio frequency (RF) spectrum and to ensure that one user does not interfere with other users.

4.0 Frequency Usage Guidance

Frequency uses are controlled by scheduling in the areas in which the tests will be conducted. The following recommendations are based on good engineering practice for such usage.

4.1 Frequency Assignments.

Frequency scheduling for simultaneous use at the same location typically will not be made for systems whose closest 99-percent power band edges are separated by less than the 99-percent bandwidth of the wider of the two. (The signals must also comply with the spectral mask presented in paragraph 6.1 of this appendix). Figure A-1 shows the radio frequency spectrum of two signals being transmitted simultaneously. The left signal

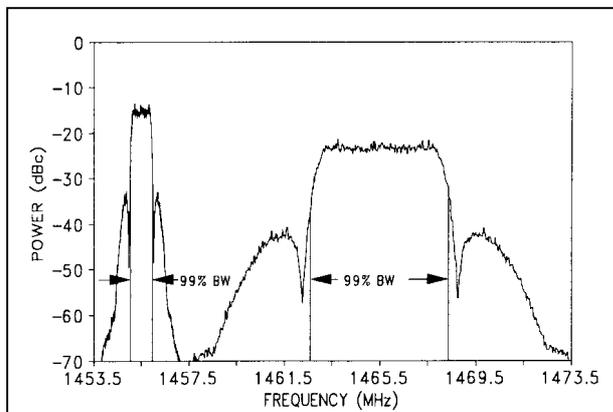


Figure A-1. 800-kb/s and 5-Mb/s RNRZ PCM/FM signals.

center frequency is 1455.5 MHz with 800-kb/s modulation. The right signal center frequency is 1465.5 MHz with 5-Mb/s modulation. The 99-percent power bandwidths of these two signals are approximately 930 kHz and 5800 kHz. (See first line of table A-1). Therefore, the minimum center frequency separation is $(930+5800)/2 + 5800 = 9165$ kHz. Because all telemetry signals are centered on xxxx.5 MHz, the separation must be an integer number of MHz. The smallest integer number of MHz larger than 9165 kHz is 10 MHz. Scheduling as stated here should ensure a desired signal-to-interfering-signal ratio of at least 40 dB for two signals of equal bandwidth and effective radiated power at the same distance from the receiving antenna when the receiver bandwidth includes less than 99 percent of the desired signal's energy. In instances when this ratio is insufficient to ensure desired data quality, the frequency separation must be increased.

4.2 Geographical Separation. Two or more telemetry systems operating in a given geographical area² should be separated in frequency such that overlap between spectra for each pair of signals is less than 0.5 percent of the power of either in the -20 dB receiver passband of the other. Overlap separation can be provided by a combination of frequency selection, power levels, antenna positioning and aiming, and geographical separation.

4.3 Simultaneous Operation. Standard practice for multiple emitters at the same location, power level, and transmitting antenna direction (if applicable) is to separate signals from one another by a "guard band" greater than or equal to the occupied bandwidth of the widest bandwidth signal in each pair of adjacent frequency transmitters. When more than one transmitter is used on the same host vehicle, frequency selection should be made to minimize spectrum overlap and RF interactions.

²The extent of a geographical area over which the frequency use must be protected varies with the nature of the usage. For airborne systems, such an area is specified by the actual vehicle flight path and its maximum altitude.

Multichannel operations should avoid channels separated by the IF frequencies of the receivers used, if possible.³

4.4 Multicarrier Operation. If two transmitters are operated simultaneously and send or receive through the same antenna system, interference due to intermodulation is likely at $(2f_1 - f_2)$ and $(2f_2 - f_1)$. Between three transmitters, the two-frequency possibilities exist, but intermodulation products may exist as well at $(f_1 + f_2 - f_3)$, $(f_1 + f_3 - f_2)$, and $(f_2 + f_3 - f_1)$, where f_1 , f_2 , and f_3 represent the output frequencies of the transmitters. Intermodulation products can arise from nonlinearities in the transmitter output circuitry that cause mixing products between a transmitter output signal and the fundamental signal coming from nearby transmitters. Intermodulation products also can arise from nonlinearities in the antenna systems. The generation of intermodulation products is inevitable, but the effects are generally of concern only when such products exceed -25 dBm. The general rule for avoiding third-order intermodulation interference is that in any group of transmitter frequencies, the separation between any pair of frequencies should not be equal to the separation between any other pair of frequencies. Because individual signals have sidebands, it should be noted that intermodulation products have sidebands spectrally wider than the sidebands of the individual signals that caused them.

4.5 Transmitter Antenna System Emission Testing. Radiated tests will be made in lieu of transmitter output tests only when the transmitter is inaccessible. Radiated tests may still be required if the antenna is intended to be part of the filtering of spurious products from the transmitter or is suspected of generating spurious products by itself or in interaction with the transmitter and feedlines. The tests should be made with normal modulation.

5.0 Bandwidth

The definitions of bandwidth in this section are universally applicable. The limits shown here are applicable for telemetry operations in the telemetry bands 1435 to 1535, 2200 to 2290, and 2310 to 2390 MHz. For the purposes of telemetry signal spectral occupancy, the bandwidths used are the 99-percent power bandwidth and the -25 dBm bandwidth. A power level of -25 dBm is exactly equivalent to an attenuation of the transmitter power by $55 + 10 \times \log(P)$ dB where P is the transmitter power expressed in watts. How bandwidth is actually measured and what the limits are, expressed in terms of that measuring system, are detailed in the following paragraphs.

5.1 Concept. The term "bandwidth" has an exact meaning in situations where an amplitude modulation (AM), double sideband (DSB), or single sideband (SSB) signal is produced with a band-limited modulating signal. In systems employing frequency modulation (FM) or phase modulation (PM), or any modulation system where the modulating signal is not band limited, bandwidth is infinite with energy extending toward zero and infinite frequency falling off from the peak value in some exponential fashion. In this more general case, bandwidth is defined as the band of frequencies in

³In theory, at least, J/F 12 data exist on all receivers as well as transmitters.

which most of the signal's energy is contained. The definition of "most" is imprecise. The following terms are applied to bandwidth.

5.1.1 Authorized Bandwidth. Authorized bandwidth is, for purposes of this document, the necessary bandwidth (bandwidth required for transmission and reception of intelligence) and does not include allowance for transmitter drift or Doppler shift.

5.1.2 Occupied Bandwidth. The width of a frequency band such that below the lower and above the upper frequency limits, the mean powers emitted are each equal to a specified percentage of the total mean power of a given emission. Unless otherwise specified by the International Telecommunication Union (ITU) for the appropriate class of emission, the specified percentage shall be 0.5 percent. The occupied bandwidth is also called the 99-percent power bandwidth in this document.

5.1.3 Necessary Bandwidth. For a given class of emission, the width of the frequency band which is just sufficient to ensure the transmission of information at the rate and with the quality required under specified conditions.

5.1.3.1 The NTIA Manual of Regulations and Procedures for Federal Radio Frequency Management states that "All reasonable effort shall be made in equipment design and operation by Government agencies to maintain the occupied bandwidth of the emission of any authorized transmission as closely to the necessary bandwidth as is reasonably practicable."

5.1.3.2 The NTIA's equation for calculating the necessary bandwidth of binary non-return-to-zero (NRZ) continuous phase frequency shift keying (CPFSK) is

$$B_n = 3.86\Delta f + 0.27R \quad \text{for } 0.03 < 2\Delta f/R < 1.0 \quad (\text{A-1})$$

where

B_n = necessary bandwidth
 Δf = peak frequency deviation
 R = bit rate

This modulation method is commonly called NRZ pulse-code modulation (PCM)/FM by the telemetry community. For example, assume the bit rate is 1000 kb/s and the peak deviation is 350 kHz, the necessary bandwidth is calculated to be 1621 kHz (using equation (A-1)). With this bit rate and peak deviation, the 99-percent power bandwidth with no filtering would be 1780 kHz and the 99-percent power bandwidth with a premodulation filter bandwidth of 700 kHz would be approximately 1160 kHz. Equations for other modulation methods are contained in the NTIA Manual of Regulations and Procedures for Federal Radio Frequency Management.

5.1.4 Received (or Receiver) Bandwidth. The received bandwidth is usually the -3 dB bandwidth of the receiver intermediate frequency (IF) section.

5.2 Bandwidth Estimation and Measurement. Various methods are used to estimate or measure the bandwidth of a signal that is not band limited. The bandwidth measurements are performed using a spectrum analyzer (or equivalent device) with the following settings: 10-kHz resolution bandwidth, 1-kHz video bandwidth, and max hold detector. The IRIG Document 106-86 and earlier versions of the Telemetry Standards specified a measurement bandwidth of 3 kHz but did not specify a video bandwidth or a detector type. Spectra measured with the new standard settings will be essentially the same as spectra measured with a 3-kHz resolution bandwidth, 10-kHz video bandwidth, and a max hold detector. However, for signals with random characteristics, the average spectral density measured with a 3-kHz resolution bandwidth without the max hold detector enabled will be approximately 10 dB lower than the spectral density measured with the new settings. Theoretical expressions for power spectral density typically assume random signals and calculate the average spectral density. The average power spectral density in a 10-kHz bandwidth for random signals is approximately 5 dB lower than the spectral density measured with the standard settings (the measured values for large, continuous, discrete spectral components will be the same with an average or a max hold detector). The most common measurement and estimation methods are described in the following paragraphs.

5.2.1 99-Percent Power Bandwidth. This bandwidth contains 99 percent of the total power. The 99-percent power bandwidth is typically measured using a spectrum analyzer or estimated using equations for the modulation type and bit rate used. If the two points that define the edges of the band are not symmetrical about the assigned center frequency, their actual frequencies should be noted as well as their difference. The 99-percent power band edges of randomized NRZ (RNRZ) PCM/FM signals are illustrated in figures A-1 and A-2. Table A-1 presents the 99-percent power bandwidth for several digital modulation methods as a function of the bit rate (R).

TABLE A-1. 99% POWER BANDWIDTHS FOR VARIOUS DIGITAL MODULATION METHODS⁴

Description	99% Power Bandwidth
NRZ PCM/FM, premod filter BW=0.7R, $\Delta f=0.35R$	1.16 R
NRZ PCM/FM, no premod filter, $\Delta f=0.25R$	1.18 R
NRZ PCM/FM, no premod filter, $\Delta f=0.35R$	1.78 R
NRZ PCM/FM, no premod filter, $\Delta f=0.40R$	1.93 R
NRZ PCM/FM, premod filter BW=0.7R, $\Delta f=0.40R$	1.57 R
Minimum shift keying (MSK), no filter	1.18 R
Feher's-patented quadrature phase shift keying (FQPSK-B)	0.78 R
Phase shift keying (PSK), no filter	19.30 R
Quadrature phase shift keying (QPSK), no filter	9.65 R
Offset QPSK (OQPSK), sinusoidal weighting	1.18 R

⁴I. Korn, Digital Communications, New York, Van Nostrand, 1985.

5.2.2 -25 dBm Bandwidth. The bandwidth beyond which all power levels are below -25 dBm. A power level of -25 dBm is exactly equivalent to an attenuation of the transmitter power by $55 + 10 \times \log(P)$ dB where P is the transmitter power expressed in watts. The -25 dBm bandwidth limits are shown in figure A-2. The -25 dBm bandwidth is primarily a function of the modulation method, transmitter power, and bit rate. The transmitter design and construction techniques also strongly influence the -25 dBm bandwidth. With a bit rate of 5 Mb/s and a transmitter power of 5 watts the -25 dBm bandwidth of an NRZ PCM/FM system with near optimum parameter settings is about 12.8 MHz, while the -25 dBm bandwidth of an equivalent FQPSK-B system is about 7.1 MHz.

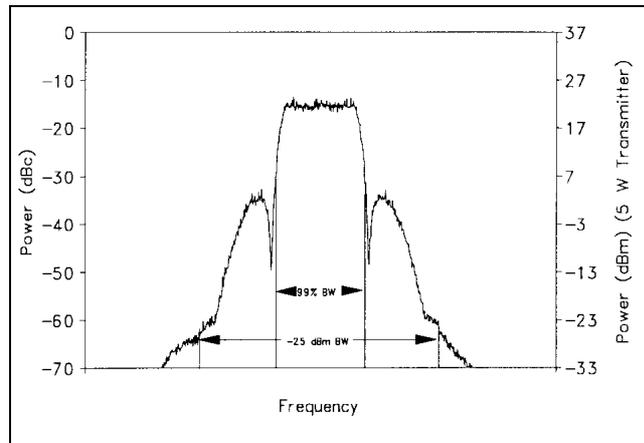


Figure A-2. RNRZ PCM/FM signal. The -25 dBm bandwidth of an NRZ PCM/FM system with near optimum parameter settings is about 12.8 MHz, while the -25 dBm bandwidth of an equivalent FQPSK-B system is about 7.1 MHz.

5.2.3 Other Bandwidth Measurement Methods. The previous methods are the standard methods for measuring the bandwidth of telemetry signals. The following methods are also sometimes used to measure or to estimate the bandwidth of telemetry signals.

5.2.3.1 Below Unmodulated Carrier. This method measures the power spectrum with respect to the unmodulated carrier power. To calibrate the measured spectrum on a spectrum analyzer, the unmodulated carrier power must be known. This power level is the 0-dB reference (commonly set to the top of the display). In AM systems, the carrier power never changes; in FM and PM systems, the carrier power is a function of the modulating signal. Since angle modulation (FM or PM) by its nature spreads the spectrum of a constant amount of power, a method to estimate the unmodulated carrier power is required if the modulation can not be turned off. For most practical angle modulated systems, the total carrier power at the spectrum analyzer input can be found by setting the spectrum analyzer's resolution and video bandwidths to their widest settings, setting the analyzer output to max hold, and allowing the analyzer to make several sweeps (see figure A-3). The maximum value of this trace will be a good approximation of the unmodulated carrier level. Figure A-3 shows the spectrum of a 5-Mb/s RNRZ PCM/FM signal measured using the standard spectrum analyzer settings discussed previously and the spectrum measured using 3-MHz resolution, video bandwidths, and max hold. The peak of the spectrum measured with the latter conditions is very close to 0 dBc and can be used to estimate the unmodulated carrier power (0 dBc) in the presence of frequency or phase modulation. In practice, the 0-dBc calibration would be performed first, and the display settings would then be adjusted to use the peak of the curve as the reference level (0-dBc level) to calibrate the spectrum measured using the standard spectrum analyzer settings. With the spectrum analyzer set for a specific resolution bandwidth, video bandwidth, and detector type, the bandwidth is taken as the distance between the two points outside of which the spectrum is thereafter some number (say, 60 dB) below the unmodulated carrier power

determined above. The -60 dBc bandwidth for the 5-Mb/s signal shown in figure A-3 is approximately 13 MHz.

The -60 dBc bandwidth of a random NRZ PCM/FM signal with a peak deviation of $0.35R$, a four-pole premodulation filter with -3 dB corner at $0.7R$, and a bit rate greater than or equal to 1 Mb/s can be approximated by

$$B_{-60\text{dBc}} = \{2.78 - 0.3 \times \log_{10}(R)\} \times R \quad (\text{A-2})$$

where B is in MHz and R is in Mb/s. Thus the -60 dBc bandwidth of a 5-Mb/s RNRZ signal under these conditions would be 12.85 MHz. The -60 dBc bandwidth will be greater if peak deviation is increased or the number of filter poles is decreased.

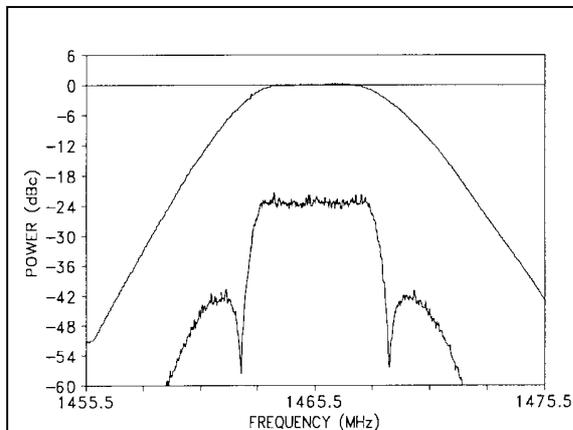


Figure A-3. Spectrum analyzer calibration of 0-dBc level.

5.2.3.2 Below Peak. This method is not recommended for measuring the bandwidth of telemetry signals. The modulated peak method is the least accurate measurement method, measuring between points where the spectrum is thereafter XX dB below the level of the highest point on the modulated spectrum. Figure A-4 shows the radio frequency spectrum of a 400-kb/s Bi ϕ -L PCM/PM signal with a peak deviation of 75° and a pre-modulation filter bandwidth of 800 kHz. The largest peak has a power level of -7 dBc. In comparison, the largest peak in figure A-3 had a power level of -22 dBc. This 15-dB difference would skew a bandwidth comparison that used the peak level in the measured spectrum as a common reference point. In the absence of an unmodulated carrier to use for calibration, the below peak measurement is often (erroneously) used and described as a below unmodulated carrier measurement. Using max hold exacerbates this effect still further. In all instances the bandwidth is overstated, but the amount varies.

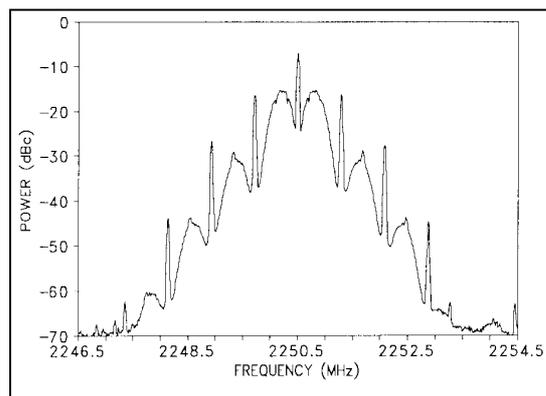


Figure A-4. Bi ϕ PCM/PM signal.

5.2.3.3 Carson's Rule. Carson's Rule is a method to estimate the bandwidth of an FM subcarrier system. Carson's Rule states that

$$B = 2 \times (\Delta f + f_{\max}) \quad (\text{A-3})$$

where B is the bandwidth, Δf is the peak deviation of the carrier frequency, and f_{\max} is the highest frequency in the modulating signal. Figure A-5 shows the spectrum that results when a 12-channel constant bandwidth multiplex with 6-dB/octave pre-emphasis frequency modulates an FM transmitter. The 99-percent power bandwidth and the bandwidth calculated using Carson's Rule are also shown. Carson's Rule will estimate a value greater than the 99-percent power bandwidth if little of the carrier deviation is due to high-frequency energy in the modulating signal.

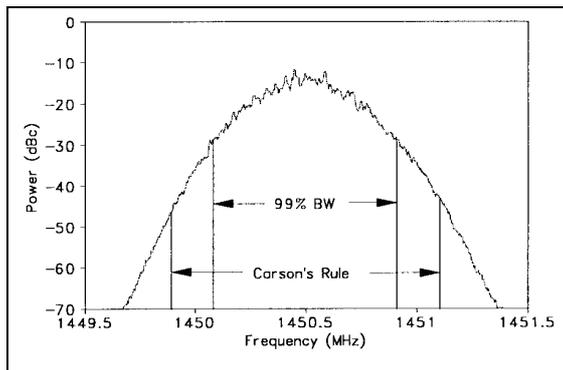


Figure A-5. FM/FM signal and Carson's Rule.

5.2.4 Spectral Equations. The following equations can be used to calculate the RF spectra for several digital modulation methods with unfiltered waveforms.^{5 6 7} These equations can be modified to include the effects of filtering.^{8 9}

Random NRZ PCM/FM (valid when $D \neq \text{integer}$, $D = 0.5$ gives MSK spectrum)

$$S(f) = \frac{4 B_{SA}}{R} \left(\frac{D}{\pi(D^2 - X^2)} \right)^2 \frac{(\cos \pi D - \cos \pi X)^2}{1 - 2 \cos \pi D \cos \pi X + \cos^2 \pi D}, \quad \cos \pi D < Q \quad (\text{A-4})$$

⁵I. Korn, Digital Communications, New York, Van Nostrand, 1985.

⁶M. G. Pelchat, "The Autocorrelation Function and Power Spectrum of PCM/FM with Random Binary Modulating Waveforms," IEEE Transactions, Vol. SET-10, No. 1, pp. 39-44, March 1964.

⁷W. M. Tey, and T. T. Tjhung, "Characteristics of Manchester-Coded FSK," IEEE Transactions on Communications, Vol. COM-27, pp. 209-216, January 1979.

⁸A. D. Watt, V. J. Zurick, and R. M. Coon, "Reduction of Adjacent-Channel Interference Components from Frequency-Shift-Keyed Carriers," IRE Transactions on Communication Systems, Vol. CS-6, pp. 39-47, December 1958.

⁹E. L. Law, "RF Spectral Characteristics of Random PCM/FM and PSK Signals," International Telemetry Conference Proceedings, pp. 71-80, 1991.

Random NRZ PSK

$$S(f) = \frac{B_{SA}}{R} \frac{\sin^2\left(\frac{\pi X}{2}\right)}{\left(\frac{\pi X}{2}\right)^2} \quad (\text{A-5})$$

Random NRZ QPSK and OQPSK

$$S(f) = \frac{2B_{SA}}{R} \frac{\sin^2(\pi X)}{(\pi X)^2} \quad (\text{A-6})$$

Random Bi ϕ PCM/FM

$$S(f) = \frac{B_{SA}}{4R} \left(\frac{\pi D}{2} \frac{\sin\left(\frac{\pi(X-D)}{4}\right)}{\frac{\pi(X-D)}{4}} \frac{\sin\left(\frac{\pi(X+D)}{4}\right)}{\frac{\pi(X+D)}{4}} \right)^2 + \left(\frac{D \sin\left(\frac{\pi D}{2}\right)}{\pi(X^2 - D^2)} \right)^2 \delta\{(f - f_c) - nR\} \quad (\text{A-7})$$

Random Bi ϕ PCM/PM

$$S(f) = \frac{B_{SA} \sin^2(\beta)}{R} \frac{\sin^4\left(\frac{\pi X}{4}\right)}{\left(\frac{\pi X}{4}\right)^2} + \cos^2(\beta) \delta(f - f_c), \quad \beta \leq \frac{\pi}{2} \quad (\text{A-8})$$

where

- $S(f)$ = power spectrum (dBc) at frequency f
- B_{SA} = spectrum analyzer resolution bandwidth
- R = bit rate
- D = $2\Delta f/R$
- X = $2(f-f_c)/R$
- Δf = peak deviation
- β = peak phase deviation in radians
- f_c = carrier frequency
- δ = Dirac delta function
- n = $0, \pm 1, \pm 2, \dots$
- Q = quantity related to narrow band spectral peaking when $D \approx 1, 2, 3, \dots$
- $Q \approx 0.99$ for $B_{SA} = 0.003 R$, $Q \approx 0.9$ for $B_{SA} = 0.03 R$

The spectrum analyzer resolution bandwidth term was added to the original equations.

5.2.5 Receiver Bandwidth. Receiver predetection bandwidth is measured at the points where the response to the carrier before demodulation is -3 dB from the center frequency response. The carrier bandwidth response of the receiver is, or is intended to be, symmetrical about the carrier in most instances. Figure A-6 shows the response of a typical telemetry receiver with a 1 MHz IF bandwidth selected. Outside the stated bandwidth, the response usually falls sharply with the response often 20 dB or more below the passband response at 1.5 to 2 times the passband response. The rapid falloff outside the passband is required to reduce interference from nearby channels and has no other effect on data.

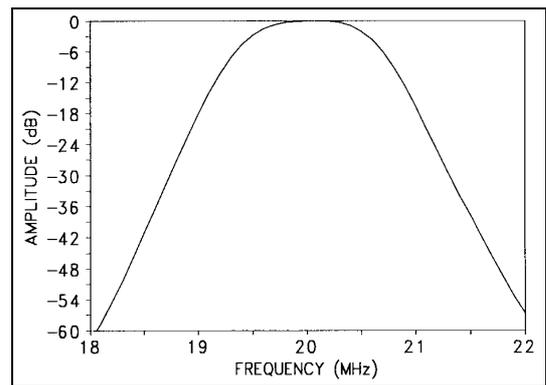


Figure A-6. Typical receiver IF filter response (-3 dB bandwidth = 1 MHz).

5.2.6 Receiver Noise Bandwidth. For the purpose of calculating noise in the receiver, the bandwidth must be integrated over the actual shape of the IF, which, in general, is not a square-sided function. Typically, the figure used for noise power calculations is the -3 dB bandwidth of the receiver.

5.3 Phase-Modulated Systems. Telemetry systems using phase modulation (PM) rather than frequency modulation (FM) produce spectra that may be considerably wider than the corresponding FM signal. This extra sideband energy is reduced in most systems by filtering at the modulation input, or the transmitter output, or both, and sideband energy is reconstructed in the receiving apparatus as part of the demodulation process. Phase-modulation systems, even with more than one data bit per symbol, are not necessarily more spectrally efficient than FM transmissions.

5.4 Symmetry. Many modulation methods produce a spectrum that is asymmetrical with respect to the carrier frequency. Exceptions include FM/FM systems, randomized NRZ PCM/FM systems, and randomized FQPSK-B systems. The most extreme case of asymmetry is due to single-sideband transmission, which places the carrier frequency at one edge of the occupied spectrum. If the spectrum is *not* symmetrical about the band center, the bandwidth and the extent of asymmetry must be noted for frequency management purposes.

5.5 FM Transmitters (ac coupled). The ac-coupled FM transmitters should not be used to transmit NRZ signals unless the signals to be transmitted are randomized because changes in the ratio of “ones” to “zeros” will increase the occupied bandwidth and may degrade the bit error rate. When ac-coupled transmitters are used with randomized NRZ signals, it is recommended that the lower -3 dB frequency response of the transmitter be no greater than the bit rate divided by 4000. For example, if a randomized 1-Mb/s NRZ signal is being transmitted, the lower -3 dB frequency response of the transmitter should be no larger than 250 Hz.

6.0 Spectral Occupancy Limits

Telemetry applications covered by this standard shall use 99-percent power bandwidth to define occupied bandwidth and -25 dBm bandwidth as the primary measure of spectral efficiency. The spectra are assumed symmetrical about the center frequency unless specified otherwise. The primary reason for controlling the spectral occupancy is to control adjacent channel interference, thereby allowing more users to be packed into a given amount of frequency spectrum. The adjacent channel interference is determined by the spectra of the signals and the filter characteristics of the receiver.

6.1 Spectral Mask. One common method of describing the spectral occupancy limits is a spectral mask. The aeronautical telemetry spectral mask is described below. Note that the mask in this standard is different than and in general narrower than the mask contained in the 1996 and 1999 versions of the Telemetry Standards. All spectral components larger than $-(55 + 10 \times \log(P))$ dBc, or in other words larger than -25 dBm, at the transmitter output must be within the spectral mask calculated using the following equation:

$$M(f) = K + 90 \log R - 100 \log |f - f_c|; \quad |f - f_c| \geq \frac{R}{m} \quad (\text{A-9})$$

where

- $M(f)$ = power (dBc) at frequency f (MHz)
- K = -20 for analog signals
- K = -28 for binary signals
- K = -63 for quaternary signals (e.g., FQPSK-B)
- f_c = transmitter center frequency (MHz)

- R = bit rate (Mb/s) for digital signals or
 $(\Delta f + f_{\max})(MHz)$ for analog FM signals
- m = number of states in modulating signal;
 m = 2 for binary signals
 m = 4 for quaternary signals and analog signals
- Δf = peak deviation
- f_{\max} = maximum modulation frequency

The -25 dBm bandwidth is not required to be narrower than 1 MHz. The first term in equation (A-9) accounts for bandwidth differences between modulation methods. Equation (A-9) can be rewritten as $M(f) = K - 10\log R - 100\log|(f-f_c)/R|$. When equation (A-9) is written this way, the $10\log R$ term accounts for the increased spectral spreading and decreased power per unit bandwidth as the modulation rate increases. The last term forces the spectral mask to roll off at 30-dB/octave (100-dB/decade). Any error detection or error correction bits, which are added to the data stream, are counted as bits for the purposes of this spectral mask. The quaternary signal spectral mask is based on the measured power spectrum of FQPSK-B. The binary signal spectral mask is primarily based on the power spectrum of a random binary NRZ PCM/FM signal with peak deviation equal to 0.35 times the bit rate and a multipole premodulation filter with a -3 dB frequency equal to 0.7 times the bit rate (see figure A-7). This peak deviation minimizes the bit error rate (BER) with an optimum receiver bandwidth while also providing a compact RF spectrum. The premodulation filter attenuates the RF sidebands while only degrading the BER by the equivalent of a few tenths of a dB of RF power. Further decreasing of the premodulation filter bandwidth will only result in a slightly narrower RF spectrum, but the BER will increase dramatically. Increasing the premodulation filter bandwidth will result in a wider RF spectrum, and the BER will only be decreased slightly. The recommended premodulation filter for NRZ PCM/FM signals is a multipole linear phase filter with a -3 dB frequency equal to 0.7 times the bit rate. The unfiltered NRZ PCM/FM signal rolls off at 12-dB/octave so at least a three-pole filter (filters with four or more poles are recommended) is required to achieve the 30-dB/octave slope of the spectral mask. The spectral mask includes the effects of reasonable component variations (unit-to-unit and temperature).

6.2 Spectral Mask Examples. Figures A-7 and A-8 show the binary spectral mask of equation (A-9) and the RF spectra of the 1-Mb/s randomized NRZ PCM/FM signals. The RF spectra were measured using a spectrum analyzer with 10-kHz resolution bandwidth, 1-kHz video bandwidth, and a max hold detector. The span of the frequency axis is 5 MHz. The transmitter power was 5 watts, and the peak deviation was 350 kHz. The modulation signal for figure A-7 was filtered with a 4-pole linear-phase filter with -3 dB frequency of 700 kHz. All spectral components in figure A-7 were contained within the spectral mask. The minimum value of the spectral mask was -62 dBc (equivalent to -25 dBm). The peak modulated signal power levels were about 15 dB below the unmodulated carrier level (-15 dBc), and the power levels near center frequency were about -17 dBc. Figure A-8 shows the same signal with no premodulation filtering. The signal was not contained within the spectral mask when a pre-modulation filter was not used.

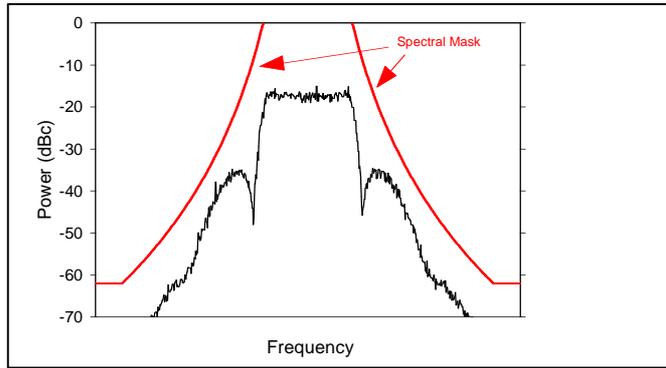


Figure A-7. Filtered 1-Mb/s RNRZ PCM/FM signal and spectral mask.

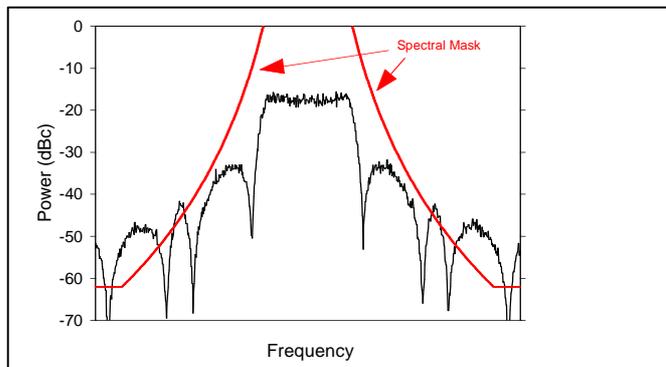


Figure A-8. Unfiltered 1-Mb/s RNRZ PCM/FM signal and spectral mask.

Figure A-9 shows the quaternary mask of equation (A-9) and the RF spectrum of a 1-Mb/s FQPSK-B signal. The transmitter power was assumed to be 10 watts in this example so the minimum value of the mask was -65 dBc. The peak value of the FQPSK-B signal was about -13 dBc.

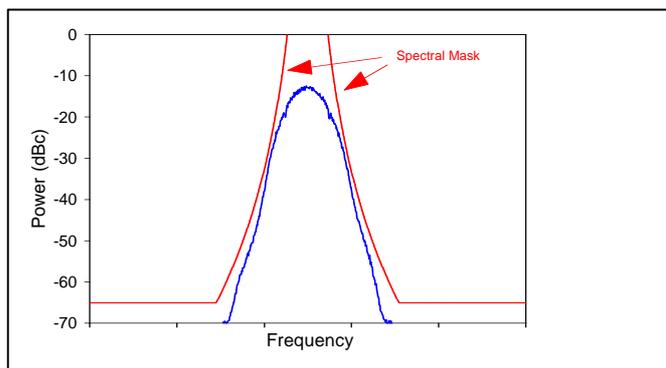


Figure A-9. Typical 1-Mb/s FQPSK-B signal and spectral mask.

7.0 FQPSK-B Characteristics

Feher's-patented quadrature phase shift keying^{10 11} (FQPSK-B) modulation is a variation of offset quadrature phase shift keying (OQPSK). OQPSK is described in most communications textbooks. A generic OQPSK (or quadrature or I & Q) modulator is shown in figure A-10. In general, the odd bits are applied to one channel (say Q), and the even bits are applied to the I channel.

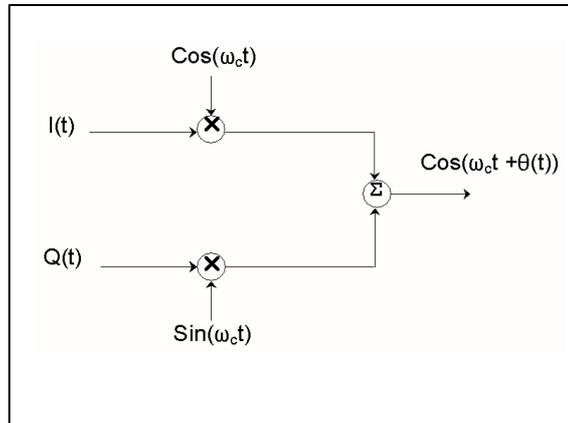


Figure A-10. OQPSK modulator.

If the values of I and Q are ± 1 , we get the diagram shown in figure A-11. For example, if $I=1$ and $Q=1$ then the phase angle is 45 degrees $\{(I,Q) = (1, 1)\}$. A constant envelope modulation method, such as minimum shift keying (MSK), would follow the circle indicated by the small dots in figure A-11 to go between the large dots. In general, bandlimited QPSK and OQPSK signals are not constant envelope and would not follow the path indicated by the small dots but rather would have a significant amount of amplitude variation.

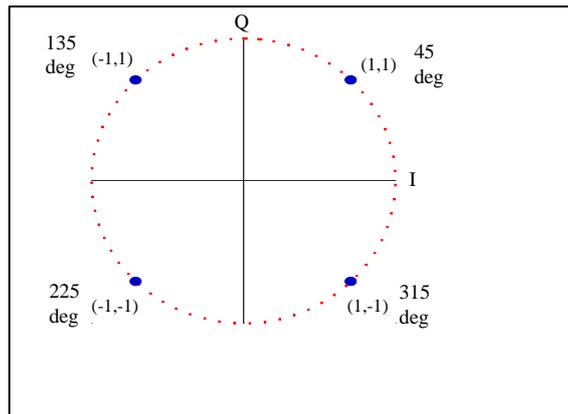


Figure A-11. I & Q constellation.

¹⁰ K. Feher et al.: US Patents 4,567,602; 4,644,565; 5,491,457; and 5,784,402, post-patent improvements and other U.S. and international patents pending.

¹¹ Kato, Shuzo and Kamilo Feher, "XPSK: A New Cross-Correlated Phase Shift Keying Modulation Technique," IEEE Trans. Comm., vol. COM-31, May 1983.

The typical implementation of FQPSK-B involves the application of data and a bit rate clock to the baseband processor of the quadrature modulator. The data are differentially encoded and converted to I and Q signals as described in Chapter 2. The FQPSK-B I and Q channels are then cross-correlated, and specialized wavelets are assembled that minimize the instantaneous variation of $(I^2(t) + Q^2(t))$. The FQPSK-B baseband wavelets are illustrated in figure A-12. The appropriate wavelet is assembled based on the current and immediate past states of I and Q. Q is delayed by one-half symbol (one bit) with respect to I as shown in figure A-13.

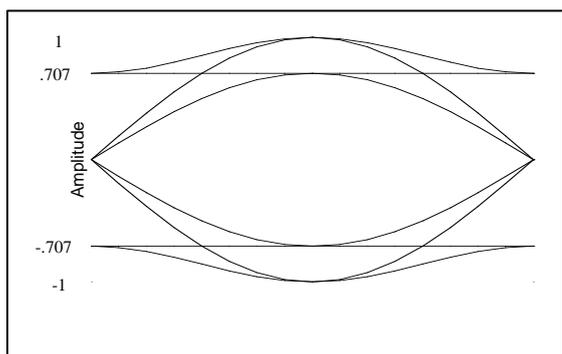


Figure A-12. FQPSK-B wavelets.

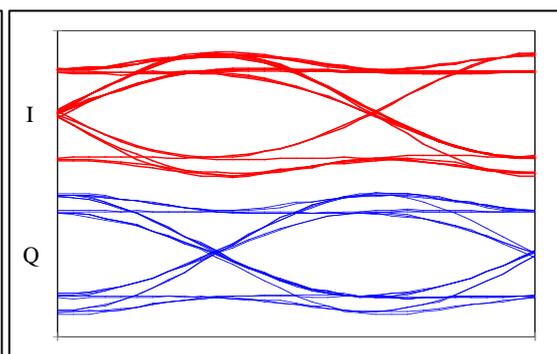


Figure A-13. FQPSK-B I & Q eye diagrams (at input to IQ modulator).

A common method of looking at I-Q modulation signals is called a vector diagram. One method of generating a vector diagram is to use an oscilloscope that has an XY mode. The vector diagram is generated by applying the I signal to the X input and the Q signal to the Y input. A sample vector diagram of FQPSK-B at the input terminals of an I-Q modulator is illustrated in figure A-14. Note that the vector diagram values are always within a few percent of being on a circle. The vector diagram of generalized filtered OQPSK would have more amplitude variations than FQPSK-B, and the vector diagram of QPSK would go through the origin. Any amplitude variations may cause spectral spreading at the output of a non-linear amplifier.

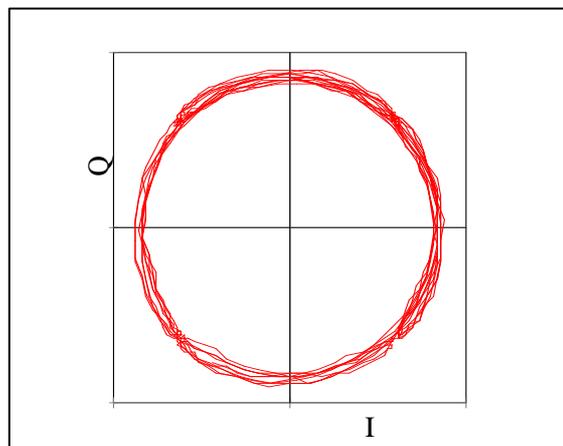


Figure A-14. FQPSK-B vector diagram.

A typical FQPSK-B spectrum measured at the output of a fully saturated RF non-linear amplifier with a random pattern of “1’s” and “0’s” applied to the input is illustrated in figure A-15. The bit rate for figure A-15 was 1 Mb/s. The peak of the spectrum will be approximately $-12-10\log(R)$, where R is in Mb/s. The 99-percent bandwidth of FQPSK-B is typically about 0.78 times the bit rate. Note that with a properly randomized data sequence and proper transmitter design, FQPSK-B does not have discernable sidebands.

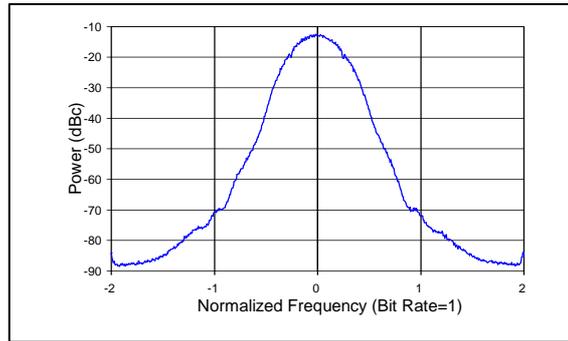


Figure A-15. FQPSK-B spectrum with random input data.

Figure A-16 illustrates FQPSK-B transmitter output with all “1’s” as the input signal. With an all “1’s” input the differential encoder, cross-correlator, and wavelet selector provide unity amplitude sine and cosine waves with a frequency equal to 0.25 times the bit rate to the I and Q modulator inputs. The resulting signal (from an ideal modulator) would be a single frequency component offset from the carrier frequency by exactly +0.25 times the bit rate. The amplitude of this component would be equal to 0 dBc. If modulator errors exist (they always will), additional frequencies will appear in the spectrum as shown in figure A-16. The spectral line at a normalized frequency of 0 (carrier frequency) is referred to as the remnant carrier. This component is largely caused by DC imbalances in the I and Q signals. The remnant carrier power in figure A-16 is approximately -31 dBc. Well designed FQPSK-B transmitters will have a remnant carrier level less than -25 dBc. The spectral component offset, 0.25 times the bit rate below the carrier frequency, is the other sideband. This component is largely caused by unequal amplitudes in I and Q and by a lack of quadrature between I and Q. The power in this component should be limited to -30 dBc or less for good system performance.

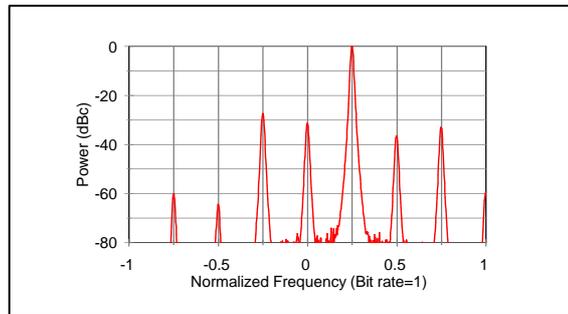


Figure A-16. FQPSK-B spectrum with all 1’s input and large modulator errors.

Figure A-17 shows the measured bit error probability (BEP) versus signal energy per bit/noise power per Hz (E_b/N_0) of one FQPSK-B modulator/demodulator combination including non-linear amplification and differential encoding/decoding in an additive white Gaussian noise environment (AWGN) with no fading. Other combinations of equipment may have different performance. Computer simulations have shown that a BEP of 10^{-5} may be achievable with an E_b/N_0 of slightly greater than 11 dB (with differential encoding/decoding). The purpose of the differential encoder/decoder is to resolve the phase detection ambiguities that are inherent in QPSK, OQPSK, and FQPSK-B modulation methods. The differential encoder/decoder used in this standard will cause one isolated symbol error to appear as two bits in error at the demodulator output. However, many aeronautical telemetry channels are dominated by fairly long burst error events, and the effect of the differential encoder/decoder will typically be masked by the error events.

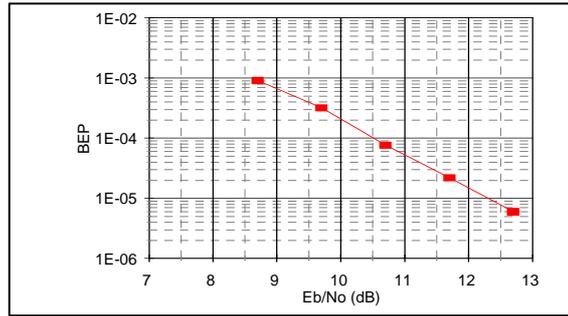


Figure A-17. FQPSK-B BEP vs E_b/N_0 .

APPENDIX B

USE CRITERIA FOR FREQUENCY DIVISION MULTIPLEXING

APPENDIX B

USE CRITERIA FOR FREQUENCY DIVISION MULTIPLEXING

1.0 General

Successful application of Frequency Division Multiplexing Telemetry Standards depends on recognition of performance limits and performance tradeoffs, which may be required in implementation of a system. The use criteria included in this appendix are offered in this context as a guide for orderly application of the standards, which are presented in chapter 3. It is the responsibility of the telemetry system designer to select the range of performance that will meet data measurement requirements and at the same time permit operation within the limits of the standards. A designer or user must also recognize the fact that even though the standards for FM/FM multiplexing encompass a broad range of performance limits, tradeoffs such as data accuracy for data bandwidth may be necessary. Nominal values for such parameters as frequency response and rise time are listed to indicate the majority of expected use and should not be interpreted as inflexible operational limits. It must be remembered that system performance is influenced by other considerations such as hardware performance capabilities. In summary, the scope of the standards together with the use criteria is intended to offer flexibility of operation and yet provide realistic limits.

2.0 FM Subcarrier Performance

The nominal and maximum frequency response of the subcarrier channels listed in tables 3-1 and 3-2 is 10 and 50 percent of the maximum allowable deviation bandwidth. The nominal frequency response of the channels employs a deviation ratio of five. The deviation ratio of a channel is one-half the defined deviation bandwidth divided by the cutoff frequency of the discriminator output filter.

2.1 The use of other deviation ratios for any of the subcarrier channels listed may be selected by the range users to conform with the specific data response requirements for the channel. As a rule, the rms signal-to-noise ratio (SNR) of a specific channel varies as the three-halves power of that subcarrier deviation ratio.

2.2 The nominal and minimum channel rise times indicated in tables 3-1 and 3-2 have been determined from the equation which states that rise time is equal to 0.35 divided by the frequency response for the nominal and maximum frequency response. The equation is normally employed to define 10 to 90 percent rise time for a step function of the channel input signal. However, deviations from these values may be encountered because of variations in subcarrier components in the system.

3.0 FM Subcarrier Performance Tradeoffs

The number of subcarrier channels which may be used simultaneously to modulate an RF carrier is limited by the RF channel bandwidth and by the output SNR that is acceptable for the application at hand. As channels are added, it is necessary to reduce the transmitter deviation allowed for each individual channel to keep the overall multiplex with the RF channel assignment. This reduction lowers the subcarrier-to-noise performance at the discriminator inputs. Thus, the system designer's problem is to determine acceptable tradeoffs between the number of subcarrier channels and acceptable subcarrier-to-noise ratios.

3.1 Background information relating to the level of performance and the tradeoffs that may be made is included in Telemetry FM/FM Baseband Structure Study, volumes I and II; which were completed under a contract administered by the Telemetry Working Group of IRIG. The Defense Technical Information Center (DTIC) access numbers for these documents dated 14 June 1965 are AD-621139 and AD-621140. The results of the study show that proportional bandwidth channels with center frequencies up to 165 kHz and constant bandwidth channels with center frequencies up to 176 kHz may be used within the constraints of these standards. The test criteria included the adjustment of the system components for approximately equal SNRs at all of the discriminator outputs with the receiver input near RF threshold. Intermodulation, caused by the radio-link components carrying the composite multiplex signal, limits the channel's performance under large signal conditions.

3.2 With subcarrier deviation ratios of four, channel data errors on the order of 2 percent rms were observed. Data channel errors on the order of 5 percent rms of full-scale bandwidth were observed when subcarrier deviation ratios of two were employed. When deviation ratios of one were used, it was observed that channel-data errors exceeded 5 percent. Some channels showed peak-to-peak errors as high as 30 percent. It must be emphasized, however, that the results of the tests performed in this study are based on specific methods of measurement on one system sample and that this system sample represents a unique configuration of components. Systems having different performance characteristics may not yield the same system performance.

3.3 System performance may be improved, in terms of better data accuracy, by sacrificing system data bandwidth; that is, if the user is willing to limit the number of subcarrier channels in the multiplex, particularly the higher frequency channels, the input level to the transmitter can be increased. The SNR of each subcarrier is then improved through the increased per-channel transmitter deviation. For example, the baseband structure study indicated that when the 165 kHz channel and the 93 kHz channel were not included in the proportional-bandwidth multiplex, performance improvement can be expected in the remaining channels equivalent to approximately 12 dB increased transmitter power.

3.4 Likewise, elimination of the five highest frequency channels in the constant bandwidth multiplex allowed a 6-dB increase in performance.

3.5 A general formula, which can be used to estimate the thermal noise performance of an FM/FM channel above threshold,¹ is as follows.

$$\left(\frac{S}{N}\right)_d = \left(\frac{S}{N}\right)_c \left(\frac{3}{4}\right)^{1/2} \left[\frac{B_c}{F_{ud}}\right]^{1/2} \left(\frac{f_{dc}}{f_s}\right) \left(\frac{f_{ds}}{F_{ud}}\right),$$

where

$\left(\frac{S}{N}\right)_d$ = discriminator output signal-to-noise ratio (rms voltage ratio),

$\left(\frac{S}{N}\right)_c$ = receiver carrier-to-noise ratio (rms voltage ratio),

B_c = carrier bandwidth (receiver IF bandwidth),

F_{ud} = subcarrier discriminator output filter: 3 dB frequency,

f_s = subcarrier center frequency,

f_{dc} = carrier peak deviation of the particular subcarrier of interest, and

f_{ds} = subcarrier peak deviation.

If the RF carrier power is such that the thermal noise is greater than the intermodulation noise, the above relation provides estimates accurate to within a few decibels. Additional information is contained in RCC document 119-88, *Telemetry Applications Handbook*.

3.6 The FM/FM composite-multiplex signal used to modulate the RF carrier may be a proportional-bandwidth format, a constant-bandwidth format, or a combination of the two types provided only that guard bands allowed for channels used in a mixed format be equal to or greater than the guard band allowed for the same channel in an unmixed format.

4.0 **FM System Component Considerations**

System performance is dependent on all components in the system. Neglecting the effects of the RF and recording system, data channel accuracy is primarily a function of the linearity and frequency response of the subcarrier oscillators and discriminators employed. Systems designed to transmit data frequencies up to the nominal frequency responses shown in tables 3-1 and 3-2 have generally well-known response capabilities, and reasonable data accuracy estimates can be easily made. For data-channel requirements approaching the maximum frequency response of tables 3-1 and 3-2,

¹ K. M. Uglow, *Noise and Bandwidth in FM/FM Radio Telemetry*, IRE Transaction on Telemetry and Remote Control, pp. 19-22 (May 1957).

oscillator and discriminator characteristics are less consistent and less well-defined, making data accuracy estimates less dependable.

4.1 The effect of the RF system on data accuracy is primarily in the form of noise because of intermodulation at high RF signal conditions well above threshold. Under low RF signal conditions, noise on the data channels is increased because of the degraded SNR existing in the receiver.

4.2 Intermodulation of the subcarriers in a system is caused by characteristics such as amplitude and phase nonlinearities of the transmitter, receiver, magnetic tape recorder/reproducer, or other system components required to handle the multiplex signal under the modulation conditions employed. In systems employing preemphasis of the upper subcarriers, the lower subcarriers may experience intermodulation interference because of the difference frequencies of the high-frequency and high-amplitude channels.

4.3 The use of magnetic tape recorders for recording a subcarrier multiplex may degrade the data channel accuracy because of the tape speed differences or variations between record and playback. These speed errors can normally be compensated for in present discriminator systems when the nominal response rating of the channels is employed and a reference frequency is recorded with the subcarrier multiplex.

5.0 Range Capability For FM Subcarrier Systems

See the following subparagraphs for additional range capabilities.

5.1 Receivers and Tape Recorders. The use of subcarrier frequencies greater than 2 MHz may require tape recorders of a greater capability than are in current use at some ranges. It is recommended that users, who anticipate employing any of the above channels at a range, check the range's capability at a sufficiently early date to allow procurement of necessary equipment.

5.2 Discriminator Channel Selection Filters. Inclusion of the higher frequency proportional-bandwidth channels and the constant-bandwidth channels may require the ranges to acquire additional band selection filters. In addition to referencing tables 3-1 and 3-2 for acquiring channel-selector filters, consideration should also be given to acquiring discriminators corresponding to the predetection carrier frequencies shown in table 6-6. In applications where minimum time delay variation within the filter is important, such as tape speed compensation or high-rate PAM or PCM, constant-delay filter designs are recommended.

APPENDIX C

PCM STANDARDS

APPENDIX C

PCM STANDARDS ADDITIONAL INFORMATION AND RECOMMENDATIONS

1.0 Bit Rate Versus Receiver Intermediate-Frequency Bandwidth

The following subparagraphs contain information about selection of receiver intermediate-frequency (IF) bandwidths. Additional information is contained in RCC document 119, Telemetry Applications Handbook.

1.1 The standard receiver IF bandwidth values are listed in table 2-1. Not all bandwidths are available on all receivers or at all test ranges. Additional bandwidths may be available at some test ranges. The IF bandwidth, for data receivers, should typically be selected so that 90 to 99 percent of the transmitted power spectrum is within the receiver 3 dB bandwidth.

1.2 For reference purposes, in a well-designed PCM/FM system (NRZ-L data code) with peak deviation equal to 0.35 times the bit rate and an IF bandwidth (3 dB) equal to the bit rate, a receiver IF signal-to-noise ratio (SNR) of approximately 13 dB will result in a bit error probability (BEP) of 10^{-6} . A 1 dB change in this SNR will result in approximately an order of magnitude change in the BEP. The relationship between BEP and IF SNR in a bandwidth equal to the bit rate is illustrated in figure C-1 for IF bandwidths equal to the bit rate and 1.5 times the bit rate. An approximate expression for the BEP is

$$\text{BEP} = 0.5 e^{(k \times \text{SNR})}$$

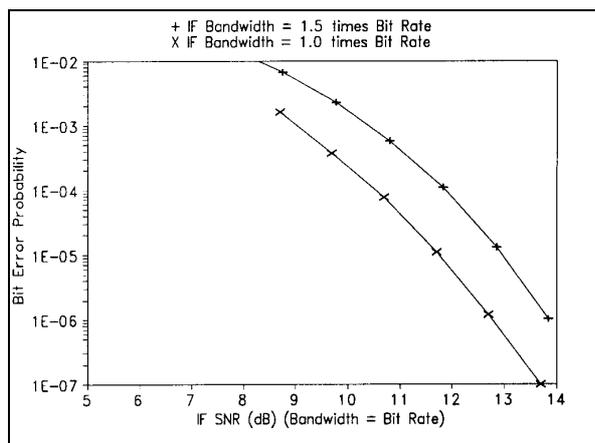


Figure C-1. BEP versus IF SNR in bandwidth = bit rate for NRZ-L PCM/FM.

where

$k \approx -0.7$ for IF bandwidth equal to bit rate

$k \approx -0.65$ for IF bandwidth equal to 1.2 times bit rate

$k \approx -0.55$ for IF bandwidth equal to 1.5 times bit rate

SNR = IF SNR \times IF bandwidth/bit rate.

Other data codes and modulation techniques have different BEP versus SNR performance characteristics.

1.3 It is recommended that the maximum period between bit transitions be 64-bit intervals to ensure adequate bit synchronization. Table C-1 contains recommended frame synchronization patterns for general use in PCM telemetry.

2.0 Recommended PCM Synchronization Patterns

Table C-1 contains recommended frame synchronization patterns for general use in PCM telemetry. Patterns are shown in the preferred order of transmission with "111" being the first bit sequence transmitted. This order is independent of data being LSB or MSB aligned. The technique used in the determination of the patterns for lengths 16 through 30 was essentially that of the patterns of 2^n binary patterns of a given length, n , for that pattern with the smallest total probability of false synchronization over the entire pattern overlap portion of the ground station frame synchronization¹. The patterns for lengths 31 through 33 were obtained from a second source².

3.0 Spectral and BEP Comparisons for NRZ and Bi ϕ ³

Figure C-2 shows the power spectral densities of baseband NRZ and Bi ϕ codes with random data. These curves were calculated using the equations presented below. Figure C-3 presents the theoretical bit error probabilities versus signal-to-noise ratio for the level, mark, and space versions of baseband NRZ and Bi ϕ codes and also for RNRZ-L. The noise is assumed to be additive white gaussian noise.

¹ A more detailed account of this investigation can be found in a paper by J. L. Maury, Jr. and J. Styles, "Development of Optimum Frame Synchronization Codes for Goddard Space Flight Center PCM Telemetry Standards," in Proceedings of the National Telemetering Conference, June 1964.

² The recommended synchronization patterns for lengths 31 through 33 are discussed more fully in a paper by E. R. Hill, "Techniques for Synchronizing Pulse-Code Modulated Telemetry," in Proceedings of the National Telemetering Conference, May 1963.

³ Material presented in paragraph 3.0 is taken from a study by W. C. Lindsey (University of Southern California), Bit Synchronization System Performance Characterization, Modeling and Tradeoff Study, Naval Missile Center Technical Publication.

$$NRZ \text{ SPECTRAL DENSITY} \propto \frac{\sin^2(\pi fT)}{(\pi fT)^2}$$

$$Bi\phi \text{ SPECTRAL DENSITY} \propto \frac{\sin^4(\pi fT/2)}{(\pi fT/2)^2}$$

where T is the bit period.

TABLE C-1. OPTIMUM FRAME SYNCHRONIZATION PATTERNS FOR PCM TELEMETRY

<u>Pattern Length</u>	<u>Patterns</u>											
16	111	010	111	001	000	0						
17	111	100	110	101	000	00						
18	111	100	110	101	000	000						
19	111	110	011	001	010	000	0					
20	111	011	011	110	001	000	00					
21	111	011	101	001	011	000	000					
22	111	100	110	110	101	000	000	0				
23	111	101	011	100	110	100	000	00				
24	111	110	101	111	001	100	100	000				
25	111	110	010	110	111	000	100	000	0			
26	111	110	100	110	101	100	110	000	00			
27	111	110	101	101	001	100	110	000	000			
28	111	101	011	110	010	110	011	000	000	0		
29	111	101	011	110	011	001	101	000	000	00		
30	111	110	101	111	001	100	110	100	000	000		
31	111	111	100	110	111	110	101	000	010	000	0	
32	111	111	100	110	101	100	101	000	010	000	00	
33	111	110	111	010	011	101	001	010	010	011	000	

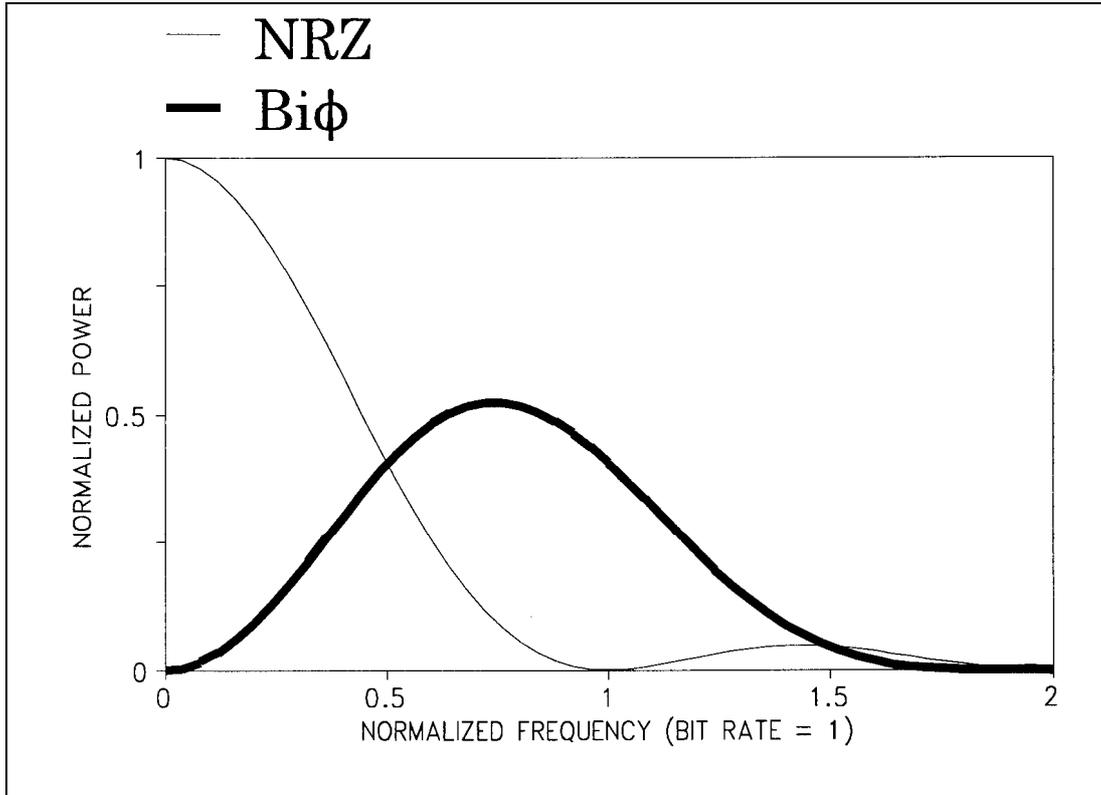


Figure C-2. Spectral densities of random NRZ and Biφ codes.

4.0 PCM Frame Structure Examples

Figures C-4, C-5, and C-6 show examples of allowable PCM frame structures. In each example, the Minor Frame Sync Pattern is counted as one word in the minor frame. The first word after the Minor Frame Sync Pattern is word 1.

Figures C-5 and C-6 show the preferred method of placing the subframe ID counter in the minor frame. The counter is placed before the parameters that are referenced to it.

Major Frame Length is as follows:

- Figure C-4 - Major Frame Length = Minor Frame Maximum Length.
- Figure C-5 - Major Frame Length = Minor Frame Maximum Length multiplied by Z.
- Figure C-6 - Major Frame Length = Minor Frame Maximum Length multiplied by Z.

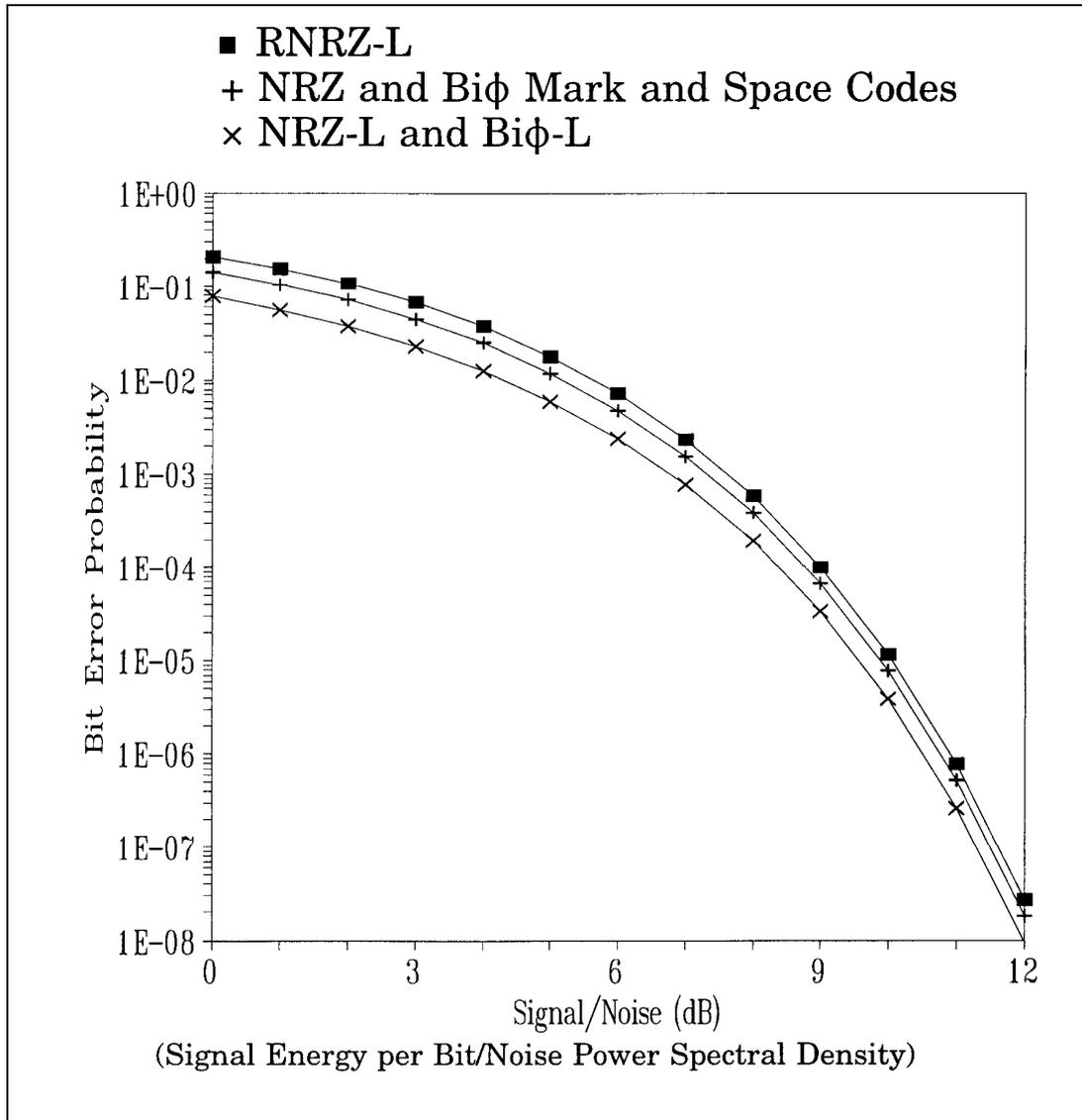


Figure C-3. Theoretical bit error probability performance for various baseband PCM signaling techniques (perfect bit synchronization assumed).

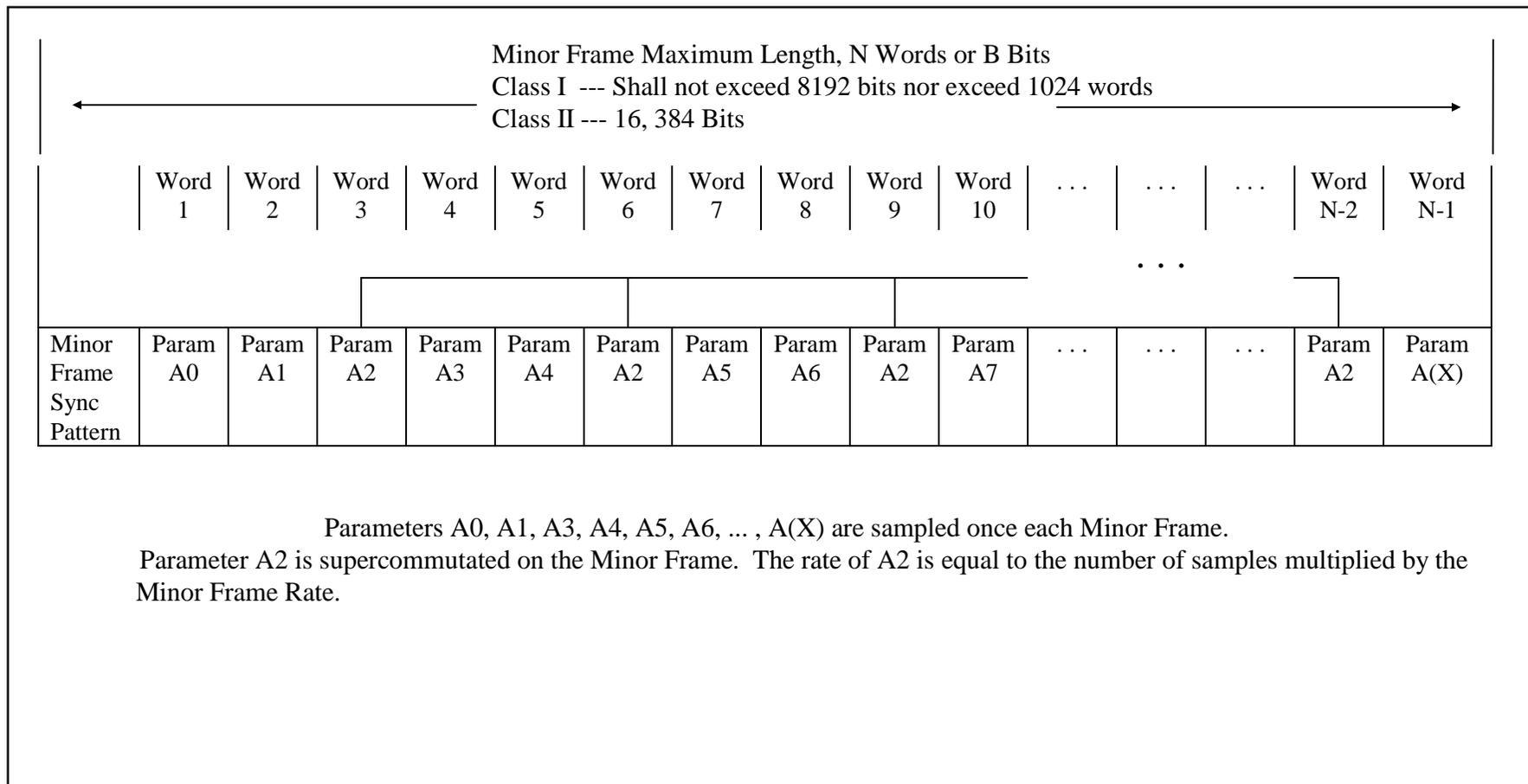


Figure C-4. Major Frame Length = Minor Frame Maximum Length.

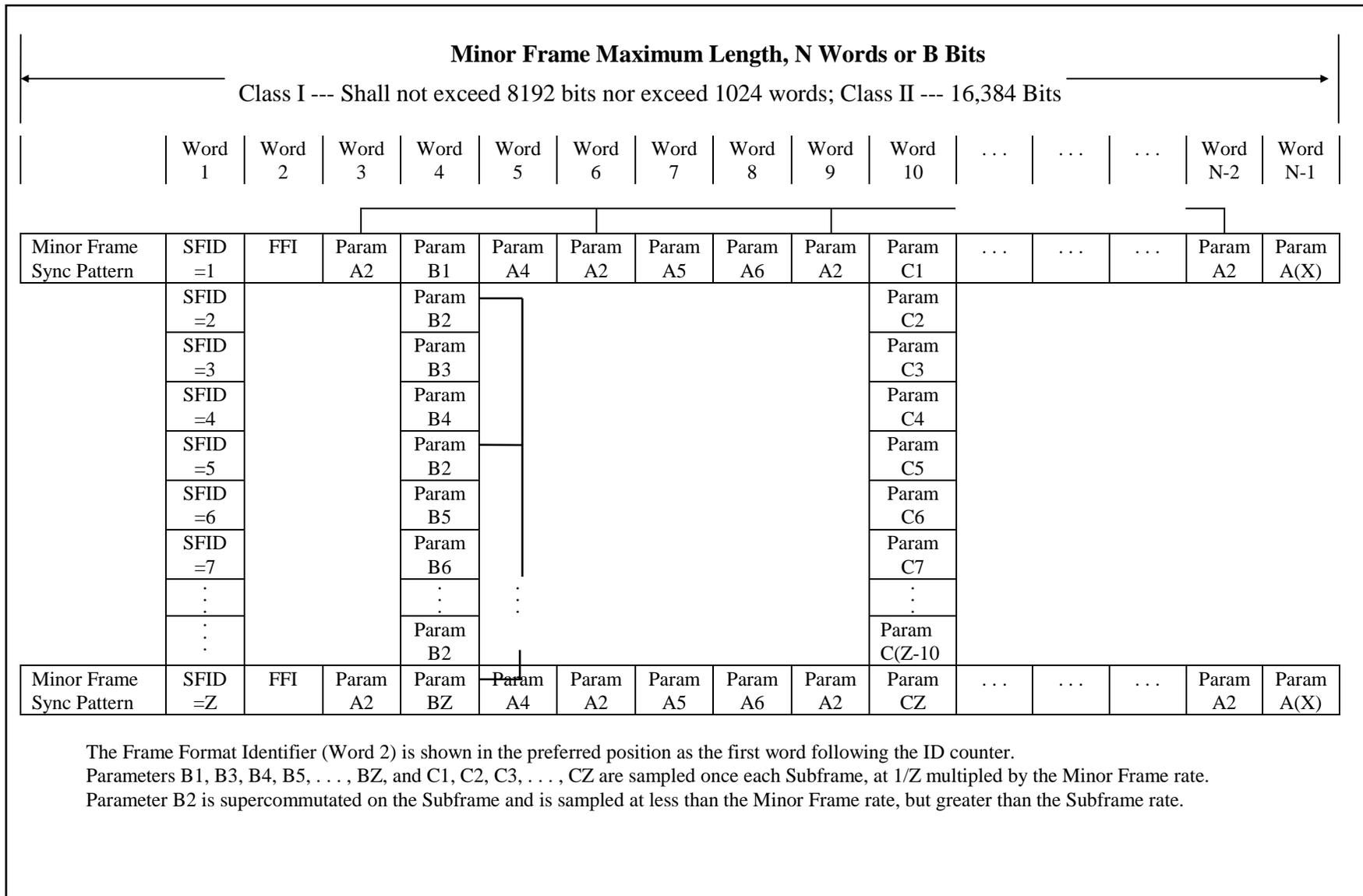


Figure C-5. Major Frame Length = Minor Frame Maximum Length multiplied by Z.

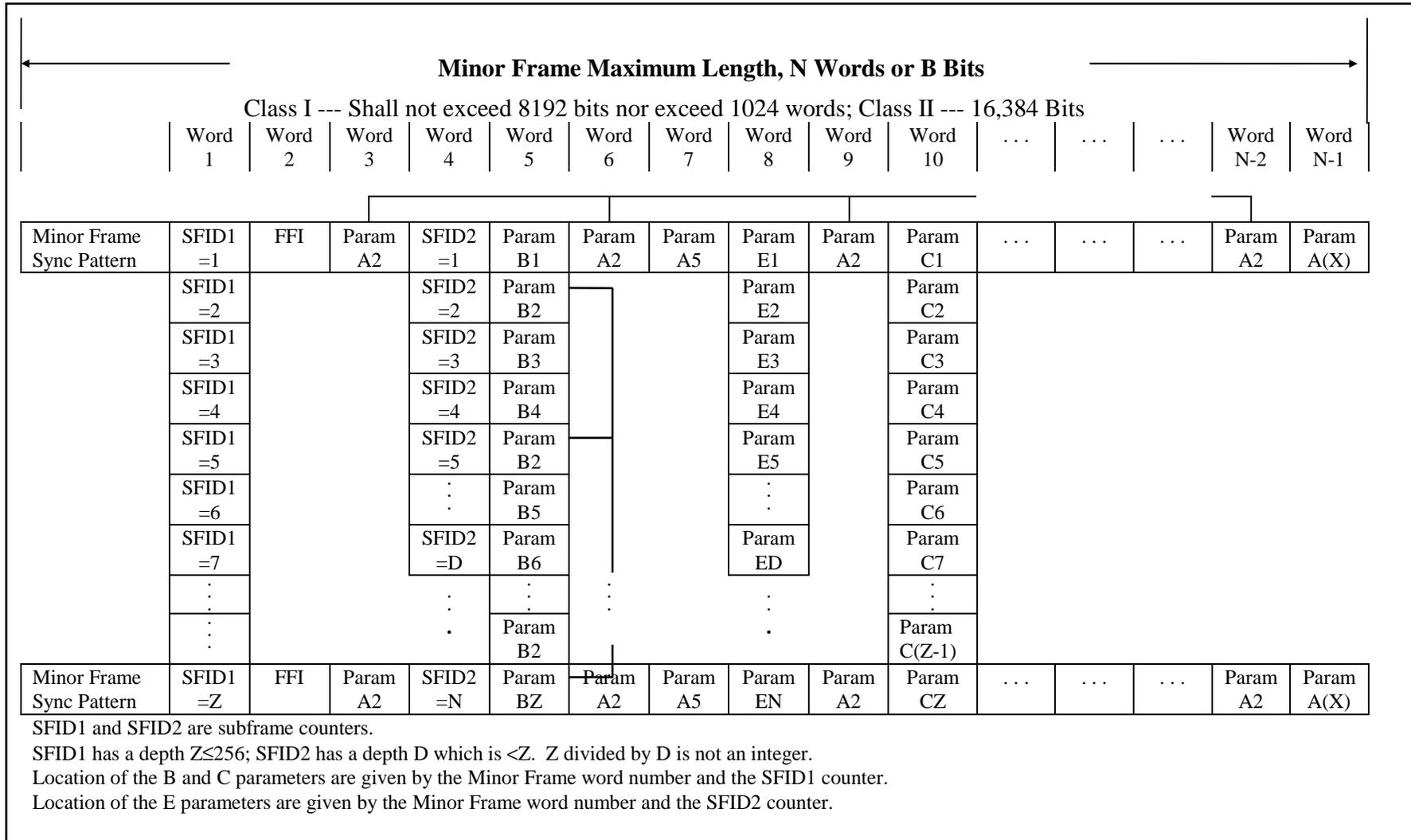


Figure C-6. Major Frame Length = Minor Frame Maximum Length multiplied by Z.

APPENDIX D

MAGNETIC TAPE RECORDER AND REPRODUCER INFORMATION AND USE CRITERIA

APPENDIX D

MAGNETIC TAPE RECORDER AND REPRODUCER INFORMATION AND USE CRITERIA

1.0 Other Instrumentation Magnetic Tape Recorder Standards

The X3B6 Committee of the American National Standards Institute and the International Standards Organizations have prepared several standards for instrumentation magnetic tape recording. Documents may be obtained by contacting the

American National Standards Institute, Inc.
1430 Broadway
New York, NY 10018
Telephone (212) 354-3300

The following documents may be of interest:

ISO 1860 Information Processing - Precision reels for magnetic tape used in interchange instrumentation applications.

ISO 6068 Information Processing - Telemetry systems (including the recording characteristics of instrumentation magnetic tape) - interchange practices and recommended test methods.

ISO 6371 Information Processing - Interchange requirements and test methods for unrecorded instrumentation magnetic tape.

ISO 8441/1 High Density Digital Recording (HDDR) - Part 1: Unrecorded magnetic tape for HDDR applications.

ISO 8441/2 High Density Digital Recording (HDDR) - Part 2: Interchange requirements and test methods for HDDR applications (including the characteristics of recorded magnetic tape).

ANSI X3.175-1990 19 mm Type 1D-1 Recorded Instrumentation - Digital Cassette Tape Format.

2.0 Double-Density Longitudinal Recording

Wide band double-density analog recording standards allowing recording of up to 4 MHz signals at 3048 mm/s (120 ips) are included in these standards. For interchange purposes, either narrow track widths 0.635 mm (25 mils) must be employed, or other special heads must be used. These requirements are necessary because of the difficulty in maintaining individual head-segment gap-azimuth alignment across a head close enough to keep each track's response within the ± 2 -dB variation allowed by the standards. Moreover, at the lower tape speeds employed in double-density recording, the 38-mm (1.5-in.) spacing employed in interlaced head assemblies results in interchannel time displacement variations between odd and even tracks that may be unacceptable for some applications. For those reasons, it was decided that a 14-track in-line configuration on 25.4-mm (1-inch) tape should be adopted as a standard. This configuration results in essentially the same format as head number one of the 28-track interlaced configuration in the standards.

2.1 The 14-track interlaced heads are not compatible with tapes produced on an in-line standard configuration, and if tapes must be interchanged, a cross-configuration dubbing may be required, or a change of head assemblies on the reproducing machine is necessary.

2.2 High energy magnetic tape is required for double-density systems. Such tapes are available but may require special testing for applications requiring a low number of dropouts per track.

2.2.1 Other Track Configurations. The previously referenced standards include configurations resulting in 7, 14, and 21 tracks in addition to the 14- and 28-track configurations listed in chapter 6. The HDDR standards also reference an 84-track configuration on 50.8-mm (2-inch) tape. Figure D-1 and table D-1 show the 7 track on 12.7-mm (1/2-inch) tape, table D-2 shows the 14 track on 12.7-mm (1/2-inch) tape, and table D-3 shows the 42 track on 25.4-mm (1-inch) tape configurations.

2.2.2 High-Density PCM Recording. High-density digital recording systems are available from most instrumentation recorder manufacturers. Such systems will record at linear packing densities of 33 000-bits-per-inch or more per track. Special systems are available for error detection and correction with overhead penalties depending on the type and the sophistication of the system employed. The HDDR documents listed in paragraph 1.0 of this appendix reference six different systems that have been produced; others are available.

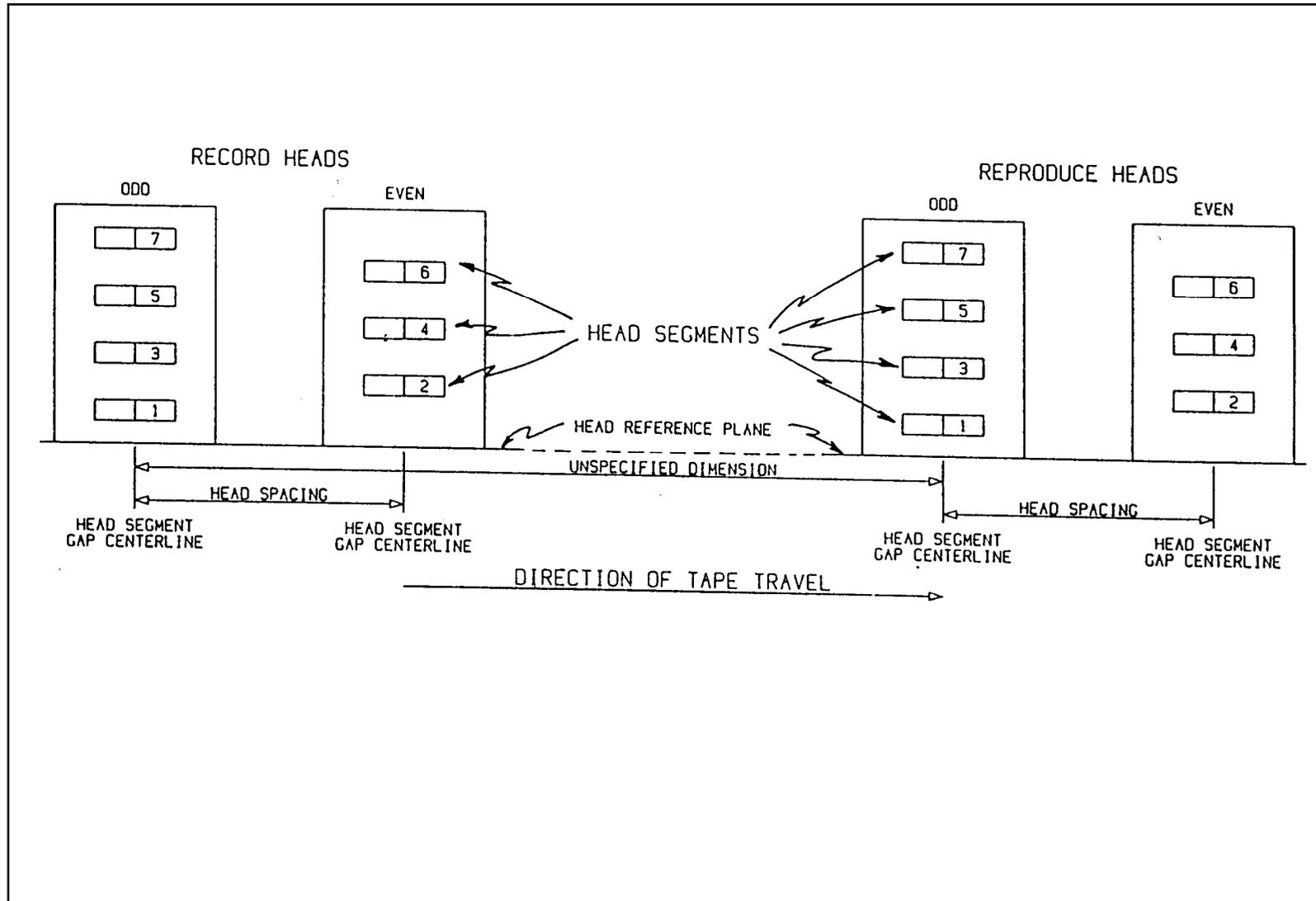


Figure D-1. Record and reproduce head and head segment identification and location (7-track interlaced system).

TABLE D-1. DIMENSIONS - RECORDED TAPE FORMAT, 7 TRACKS INTERLACED
ON 12.7-mm (1/2 in.) WIDE TAPE (REFER TO FIGURE 6-1).

Parameters	Millimeters		Inches	
	<u>Maximum</u>	<u>Minimum</u>		
Track Width	1.397	1.143	0.050	±0.005
Track Spacing		1.778	0.070	
Head Spacing:				
Fixed Heads	38.125	38.075	1.500	±0.001
Adjustable Heads	38.151	38.049	1.500	±0.002
Edge Margin, Minimum		0.127	0.005	
Reference Track				
Location	1.067	0.965	0.040	±0.002
Track Location				
Tolerance	0.051	-0.051		±0.002
		<u>Location of nth track</u>		
<u>Track Number</u>	<u>Millimeters</u>		<u>Inches</u>	
	<u>Maximum</u>	<u>Minimum</u>		
1 (Reference)	0.000	0.000	0.000	
2	1.829	1.727	0.070	
3	3.607	3.505	0.140	
4	5.385	5.283	0.210	
5	7.163	7.061	0.280	
6	8.941	8.839	0.350	
7	10.719	10.617	0.420	

**TABLE D-2. DIMENSIONS - RECORDED TAPE FORMAT, 14-TRACKS
INTERLACED ON 12.7-mm (1/2 in.) WIDE TAPE (REFER TO FIGURE
6-1).**

Parameters	Millimeters		Inches	
	<u>Maximum</u>	<u>Minimum</u>		
Track Width	0.660	0.610	0.025	±0.001
Track Spacing		0.889	0.035	
Head Spacing:				
Fixed Heads	38.125	38.075	1.500	±0.001
Adjustable Heads	38.151	38.049	1.500	±0.002
Edge Margin, Minimum		0.127	0.005	
Reference Track				
Location	0.546	0.470	0.0200	±0.001
Track Location				
Tolerance	0.038	-0.038		±0.0015
		<u>Location of nth track</u>		
<u>Track Number</u>	<u>Millimeters</u>		<u>Inches</u>	
	<u>Maximum</u>	<u>Minimum</u>		
1 (Reference)	0.000	0.000	0.000	
2	0.927	0.851	0.035	
3	1.816	1.740	0.070	
4	2.705	2.629	0.105	
5	3.594	3.518	0.140	
6	4.483	4.407	0.175	
7	5.372	5.292	0.210	
8	6.261	6.185	0.245	
9	7.150	7.074	0.280	
10	8.039	7.963	0.315	
11	8.928	8.852	0.350	
12	9.817	9.741	0.385	
13	10.706	10.630	0.420	
14	11.595	11.519	0.455	

TABLE D-3. DIMENSIONS - RECORDED TAPE FORMAT, 42-TRACKS
 INTERLACED ON 25.4-mm (1-in.) WIDE TAPE (REFER TO FIGURE
 6-1).

Parameters	Millimeters		Inches	
	<u>Maximum</u>	<u>Minimum</u>		
Track Width	0.483	0.432	0.018	±0.001
Track Spacing		0.584	0.023	
Head Spacing:				
Fixed Heads	38.125	38.075	1.500	±0.001
Adjustable Heads	38.151	38.049	1.500	±0.002
Edge Margin, Minimum		0.305	0.012	
Reference Track				
Location	0.737	0.660	0.0275	±0.015
Track Location				
Tolerance	0.025	-0.025		±0.0000
			<u>Location of nth track</u>	
<u>Track Number</u>	<u>Millimeters</u>		<u>Inches</u>	
	<u>Maximum</u>	<u>Minimum</u>		
1 (Reference)	0.000	0.000	0.000	
2	0.610	0.559	0.023	
3	1.194	1.143	0.046	
4	1.778	1.727	0.069	
5	2.362	2.311	0.092	
6	2.946	2.896	0.115	
7	3.531	3.480	0.138	
8	4.115	4.064	0.161	
9	4.699	4.648	0.184	
10	5.283	5.232	0.207	
11	5.867	5.817	0.230	
12	6.452	6.401	0.253	
13	7.036	6.985	0.276	
14	7.620	7.569	0.299	
15	8.204	8.153	0.322	
16	8.788	8.738	0.345	

TABLE D-3. (CONT'D) DIMENSIONS - RECORDED TAPE FORMAT, 42-TRACKS INTERLACED ON 25.4-mm (1-in.) WIDE TAPE (REFER TO FIGURE 6-1).

<u>Track Number</u>	<u>Location of nth track</u>		<u>Inches</u>
	<u>Millimeters</u>	<u>Millimeters</u>	
	<u>Maximum</u>	<u>Minimum</u>	
17	9.373	9.322	0.368
18	9.957	9.906	0.391
19	10.541	10.490	0.414
20	11.125	11.074	0.437
21	11.709	11.659	0.460
22	12.294	12.243	0.483
23	12.878	12.827	0.506
24	13.462	13.411	0.529
25	14.046	13.995	0.552
26	14.630	14.580	0.575
27	15.215	15.164	0.598
28	15.799	15.748	0.621
29	16.383	16.332	0.664
30	16.967	16.916	0.667
31	17.551	17.501	0.690
32	18.136	18.085	0.713
33	18.720	18.660	0.736
34	19.304	19.253	0.759
35	19.888	19.837	0.782
36	20.472	20.422	0.805
37	21.057	21.006	0.828
38	21.641	21.590	0.851
39	22.225	22.174	0.874
40	22.809	22.758	0.897
41	23.393	23.343	0.920
42	23.978	23.927	0.943

3.0 Serial HDDR

The following subparagraphs give some background for selecting the bi-phase and RNRZ-L systems specified in subparagraph 6.11.3, chapter 6 of this document.

3.1 Serial HDDR is a method of recording digital data on a magnetic tape where the digital data is applied to one track of the recording system as a bi-level signal. The codes recommended for serial HDDR recording of telemetry data are Bi ϕ -L and randomized NRZ-L (RNRZ-L) (refer to paragraph 6.11, chapter 6).

3.2 In preparing paragraph 6.11 of chapter 6, the following codes were considered: Delay Modulation (Miller Code), Miller Squared, Enhanced NRZ, NRZ Level, NRZ Mark, and NRZ Space. These codes are not recommended for interchange applications at the bit rates given in paragraph 6.11.

3.3 The properties of the Bi ϕ -L and RNRZ-L codes relevant to serial HDDR and the methods for generating and decoding RNRZ-L are described next. Recording with bias is required for interchange applications, because reproduce amplifier phase and amplitude equalization adjustments for tapes recorded without bias usually differ from those required for tapes recorded with bias.

3.4 The Bi ϕ -L and RNRZ-L codes were selected for this standard because the "level" versions are easier to generate and are usually available as outputs from bit synchronizers. "Mark" and "Space" codes also have about twice as many errors as the level codes for the same SNR. If polarity insensitivity is a major consideration, agreement between interchange parties should be obtained before these codes are used.

3.5 Some characteristics of the Bi ϕ -L code favorable to serial HDDR are listed in the following subparagraphs.

3.5.1 Only a small proportion of the total signal energy occurs near dc.

3.5.2 The maximum time between transitions is a 1-bit period.

3.5.3 The symbols for one and zero are antipodal; that is, the symbols are exact opposites of each other. Therefore, the bit error probability versus SNR performance is optimum.

3.5.4 The Bi ϕ -L can be decoded using existing bit synchronizers.

3.5.5 The Bi ϕ -L is less sensitive to misadjustments of bias and reproducer equalizers than most other codes.

3.5.6 The Bi ϕ -L performs well at low tape speeds and low bit rates.

3.6 The most unfavorable characteristic of the Bi ϕ -L code is that it requires approximately twice the bandwidth of NRZ. Consequently, the maximum bit packing density that can be recorded on magnetic tape is relatively low.

3.7 Characteristics of the RNRZ-L code which favor its use for serial HDDR are included in the following subparagraphs.

3.7.1 The RNRZ-L requires approximately one-half the bandwidth of Bi ϕ -L.

3.7.2 The symbols for one and zero are antipodal; therefore, the bit error probability versus SNR performance is optimum.

3.7.3 The RNRZ-L decoder is self-synchronizing.

3.7.4 The RNRZ-L data can be bit synchronized and signal conditioned using existing bit synchronizers with the input code selector set to NRZ-L.

3.7.5 The RNRZ-L code is easily generated and decoded.

3.7.6 The RNRZ-L data can be easily decoded in the reverse mode of tape playback.

3.7.7 The RNRZ-L data are bit detected and decoded using a clock at the bit rate. Therefore, the phase margin is much larger than that of codes that require a clock at twice the bit rate for bit detection.

3.7.8 The RNRZ-L code does not require overhead bits.

3.8 Unfavorable characteristics of the RNRZ-L code for serial HDDR are described next.

3.8.1 Long runs of bits without a transition are possible although the probability of occurrence is low, and the maximum run length can be limited by providing transitions in each data word.

3.8.2 Each isolated bit error that occurs after the data has been randomized causes 3-bit errors in the derandomized output data.

3.8.3 The decoder requires 15 consecutive error-free bits to establish and reestablish error-free operation.

3.8.4 The RNRZ-L bit stream can have a large low frequency content. Consequently, reproducing data at tape speeds which produce PCM bit rates less than 200 kb/s is not recommended unless a bit synchronizer with specially designed dc and low frequency restoration circuitry is available.

3.9 Randomizer for RNRZ-L

The randomizer is implemented with a network of shift registers and modulo-2 adders (exclusive-OR gates). The RNRZ-L bit stream is generated by adding (modulo-2) the reconstructed NRZ-L PCM data to the modulo-2 sum of the outputs of the 14th and 15th stages of a shift register. The output RNRZ-L stream is also the input to the shift register (see figure D-2).

3.9.1 The properties of an RNRZ-L bit stream are similar to the properties of a pseudo-random sequence. A 15-stage RNRZ-L encoder will generate a maximal length pseudo-random sequence of $2^{15}-1$ (32 767) bits if the input data consists only of zeros and there is at least a single one in the shift register. A maximal length pseudo-random sequence is also generated when the input data consists only of ones and the shift register contains at least a single zero. However, if the shift register contains all zeros at the moment that the input bit stream is all zeros, the RNRZ-L output bit stream will also be all zeros. The converse is also true: when the shift register is filled with ones and the input bit stream is all ones, the RNRZ-L output bit stream will contain only ones. In these two cases, the contents of the shift register does not change and the output data is not randomized.

However, the randomizer is not permanently locked-up in this state because a change in the input data will again produce a randomized output. In general, if the input bit stream contains runs of X bits without a transition with a probability of occurrence of $p(X)$, the output will contain runs having a length of up to $(X+15)$ bits with a probability of $(2^{-15} \times p(X))$. Therefore, the output can contain long runs of bits without a transition, but the probability of occurrence is low.

3.9.2 The RNRZ-L bit stream is decoded (derandomized) by adding (modulo-2) the reconstructed RNRZ-L bit stream to the modulo-2 sum of the outputs of the 14th and 15th stages of the shift register. The reconstructed RNRZ-L bit stream is the input to the shift register (see figure D-3). The RNRZ-L data which is reproduced using the reverse playback mode of operation is decoded by adding (modulo-2) the reconstructed RNRZ-L bit stream to the modulo-2 sum of the outputs of the 1st and 15th stages of the shift register (see figure D-3). The net effect is that the decoding shift register runs "backwards" with respect to the randomizing shift register.

3.9.3 Although the RNRZ-L decoder is self-synchronizing, 15 consecutive error-free bits must be loaded into the shift register before the output data will be valid. A bit slip will cause the decoder to lose synchronization, and 15 consecutive error-free data bits must again be loaded into the shift register before the output data is valid. The decoded output data, although correct, will contain the bit slip causing a shift in the data with respect to the frame synchronization pattern. Therefore, frame synchronization must be reacquired before the output provides meaningful data.

D-12

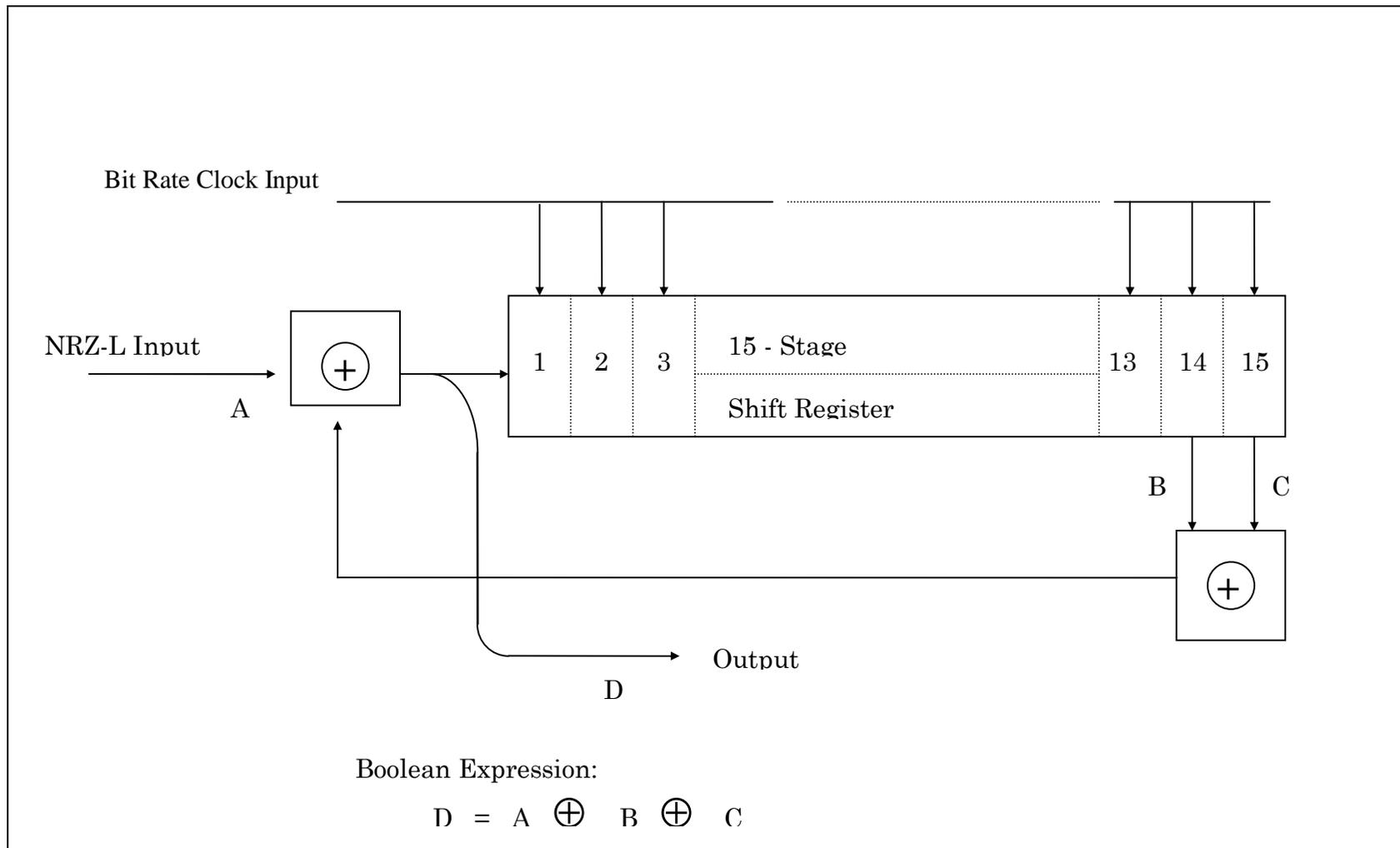


Figure D-2. Randomizer block diagram.

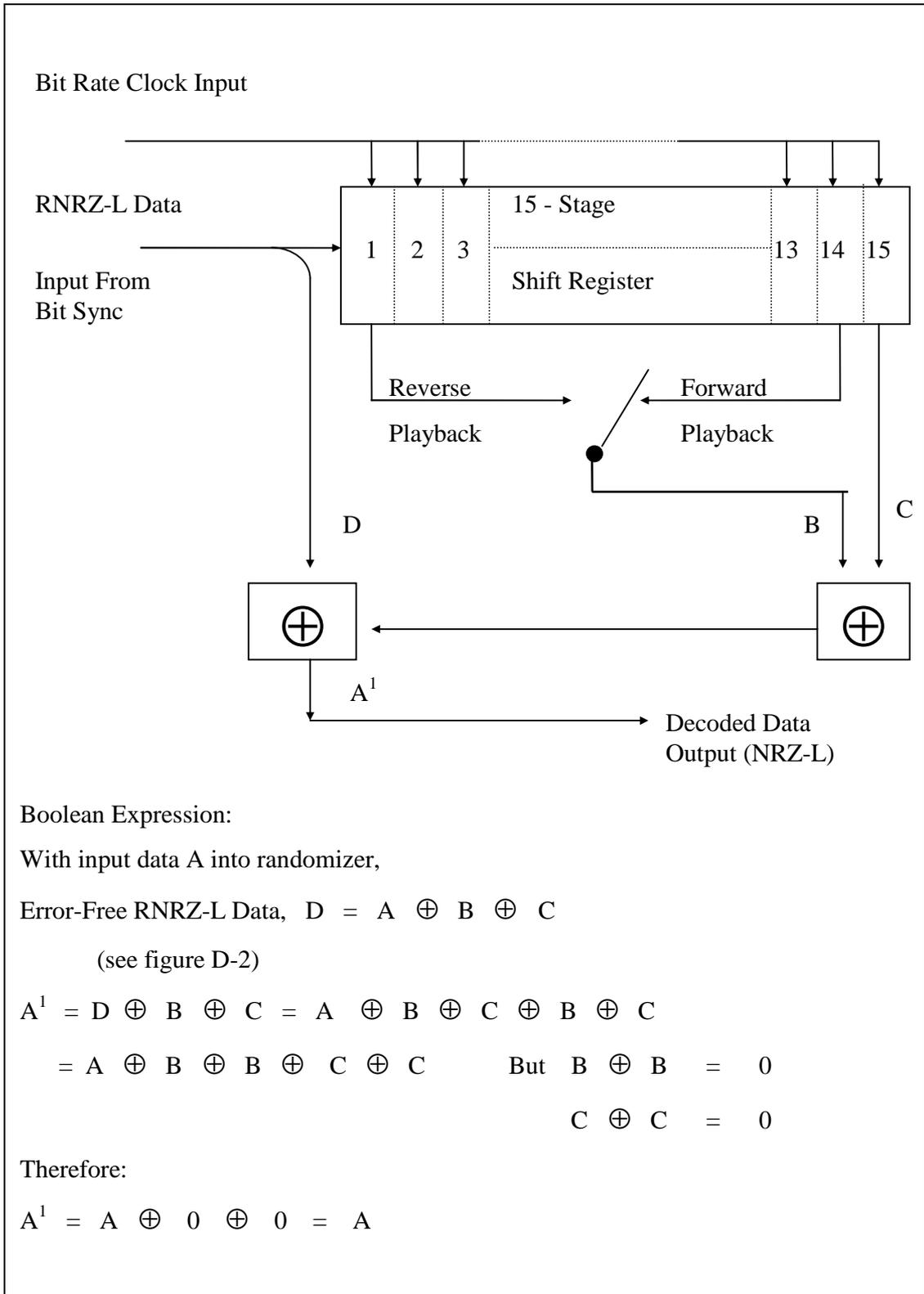


Figure D-3. Randomized NRZ-L decoder block diagram.

3.9.4 The RNRZ-L decoding system has an error multiplication factor of 3 for isolated bit errors (separated from adjacent bit errors by at least 15 bits). An isolated bit error introduced after randomization will produce 3 errors in the output data; the original bit in error, plus 2 additional errors 14 and 15 bits later. In addition, a burst of errors occurring after the data has been randomized will produce a burst of errors in the derandomized output. The number of errors in the output depends on the distribution of errors in the burst and can be greater than, equal to, or less than the number of errors in the input to the derandomizer. However, the derandomization process always increases the number of bits between the first and last error in the burst by 15. Errors introduced prior to randomization are not affected by either the randomizer or the derandomizer. The reverse decoder has the same bit error properties as the forward decoder.

3.9.5 Input data containing frequent long runs of bits without transitions creates potential dc and low frequency restoration problems in PCM bit synchronizers because of the low frequency cutoff of direct recorder and reproducer systems. The restoration problem can be minimized by reproducing the data at tape speeds that produce a bit rate for which the maximum time between transitions is less than 100 microseconds. Additional methods of minimizing these effects include selecting bit synchronizers containing special dc and low frequency restoration circuitry or recording data using Bi ϕ -L code.

3.9.6 The power spectra of the RNRZ-L and Bi ϕ -L codes are shown in figure D-4. The power spectral density of RNRZ-L is concentrated at frequencies, which are less than one-half the bit rate. The power spectral density of Bi ϕ -L is concentrated at frequencies in a region around 0.75 times the bit rate. The concentration of energy in the low-frequency region (when using the RNRZ-L code) has the effect of reducing the SNR as well as creating baseline wander, which the bit synchronizer must follow. Therefore, reproducing data at tape speeds which produce PCM bit rates of less than 200 kb/s is not recommended when using RNRZ-L unless a bit synchronizer with specially designed dc and low frequency restoration circuitry is available.

3.9.7 Alignment of the reproducer system is very important to reproducing high quality PCM data, that is, with the lowest possible bit error probability. A PCM signature using the standard 2047-bit pseudo-random pattern, recorded on the leader or the trailer tape, provides a good method for reproducer alignment. When a pseudo-random bit error detection system is not available or when a PCM signature signal is not recorded, the recommended procedure for reproducer alignment involves the use of the eye pattern technique. The eye pattern is the result of superpositioning the zeros and ones in the PCM bit stream. The eye pattern is displayed on an oscilloscope by inserting the raw reproduced bit stream into the vertical input and the reconstructed bit-rate clock into the external synchronization input of the oscilloscope. The reproducer head azimuth, amplitude equalizers, and phase equalizers are then adjusted to produce the eye pattern with the maximum height and width opening.

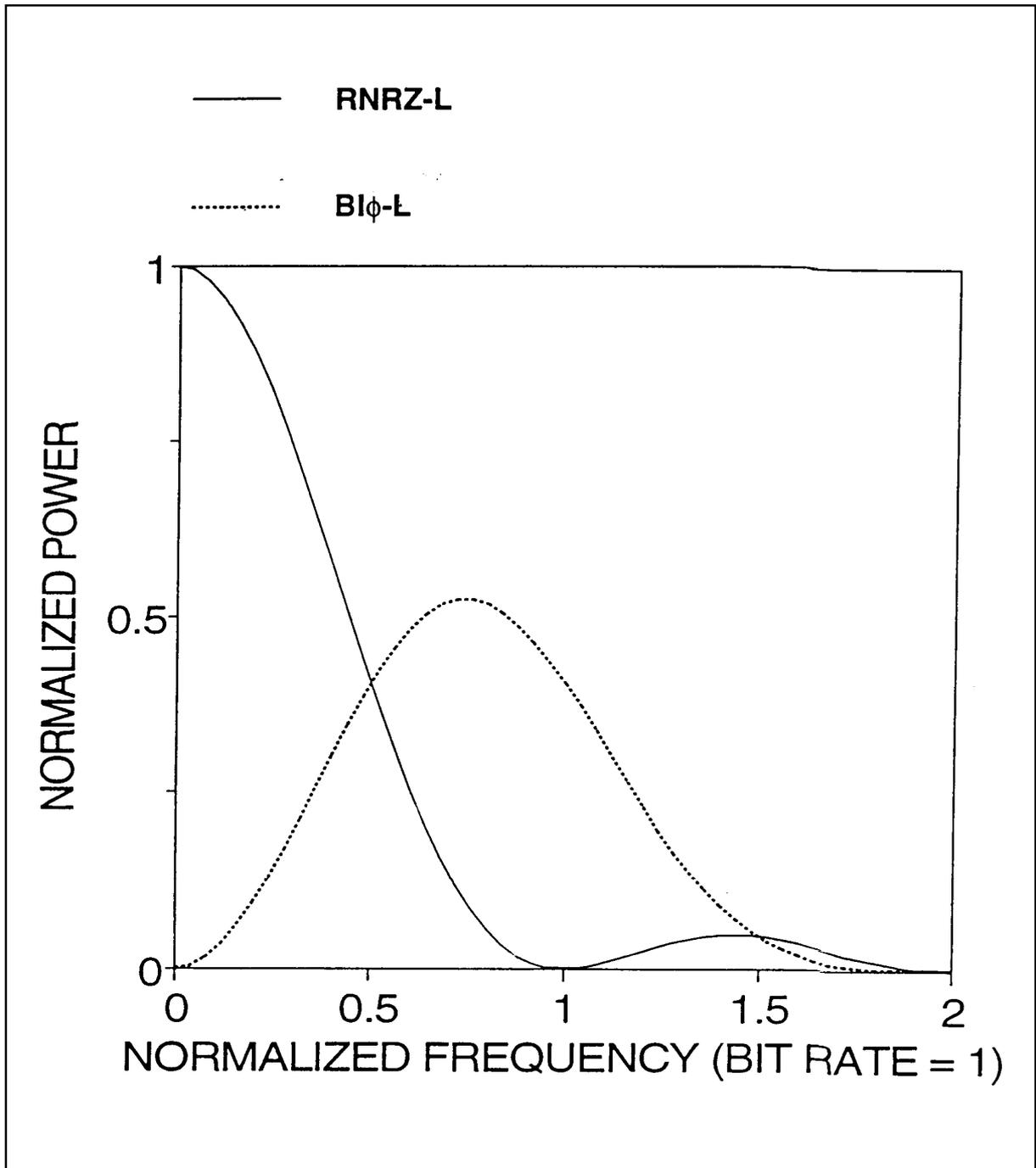


Figure D-4. Random PCM power spectra.

3.9.8 Sample eye patterns are shown in figure D-5. Figure D-5(a) shows a Bi ϕ -L eye pattern at a recorded bit packing density of 15 kb/in (450 kb/s at 30 ips). Figure D-5(b) shows an RNRZ-L eye pattern at a recorded bit packing density of 25 kb/in (750 kb/s at 30 ips).

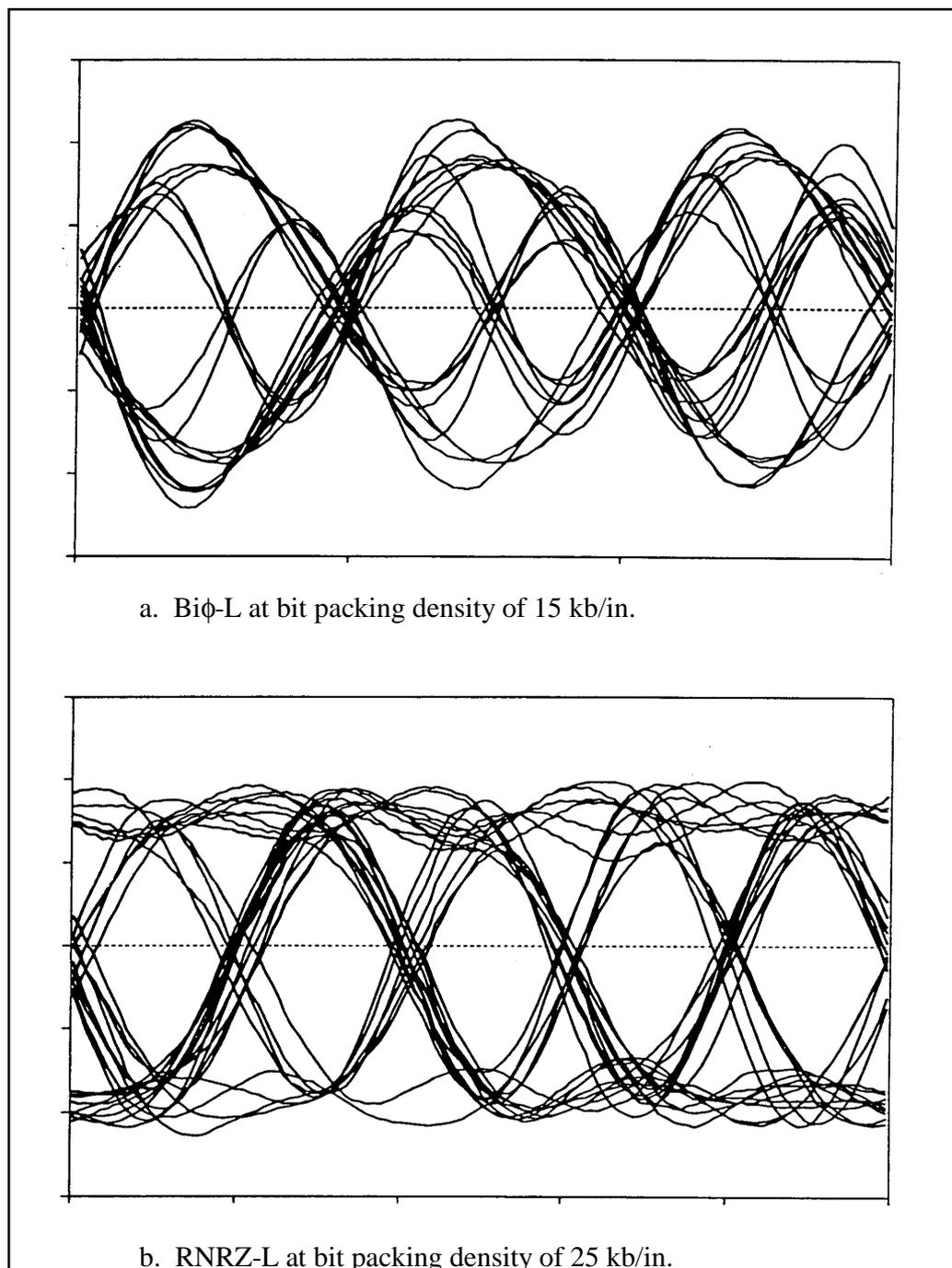


Figure D-5. Sample eye patterns at output of recorder/reproducer.

4.0 Head Parameters

The following subparagraphs describe the head parameters.

4.1 Gap Scatter. Refer to the definitions in subparagraphs 6.2.30 and 6.2.31 in chapter 6. Gap scatter contains components of azimuth misalignment and deviations from the average line defining the azimuth. Since both components affect data simultaneity from record to reproduce, the measurement is the inclusive distance containing the combined errors. Because azimuth adjustment affects the output of wide band systems, a 5.08- μm (0.0002-in.) gap scatter is allowed for such recorders and reproducers. A 2.54- μm (0.0001-in.) gap scatter is recommended for fixed-head systems (see upper illustration in figure 6-3).

4.2 Head Polarity. The requirement that a positive pulse at a record amplifier input generate a south-north-north-south magnetic sequence and that a south-north-north-south magnetic sequence on tape produce a positive pulse at the reproduce amplifier output, still leaves two interdependent parameters unspecified. These parameters are (1) polarity inversion or noninversion in record and playback amplifiers and (2) record or playback head winding sense. For the purpose of head replacement, it is necessary that these parameters be determined by the user so that an unsuspected polarity inversion, on tape or off tape, will not occur after heads are replaced.

5.0 Record Level

The standard record level is established as the input level of a sinusoidal signal set at the record level set frequency which, when recorded, produces a signal containing 1 percent third harmonic distortion at the output of a properly terminated reproduce amplifier (see subparagraph 4.1.3.3 of volume III, RCC document 118). A 1 percent harmonic distortion content is achieved when the level of the third harmonic component of the record level set frequency is 40 ± 1 dB below the level of a sinusoidal signal of 0.3 UBE which is recorded at the standard record level. Standard test and operating practice is to record and reproduce sinusoidal signals at 0.1 and 0.3 UBE and adjust the equalizers as necessary to establish the reproduced output at 0.3 UBE to within ± 1.0 dB of the output at 0.1 UBE. Then a 1 V rms signal at the record level set frequency is applied to the record amplifier input and the record and reproduce level controls are adjusted until the reproduced output contains 1 percent third harmonic distortion at a level of 1 V rms.

The optimum level for recording data will seldom be equal to the standard record level. Signals having noise-like spectral distribution such as baseband multiplexes of FM subcarriers contain high crest factors so that it may be necessary (as determined in paragraph 1.1, Noise Power Ratio (NPR) Test, volume IV, RCC document 118, Test Methods for Data Multiplex Equipment) to record at levels below the standard record level. On the other hand, for predetection and HDDR recording, signals may have to be recorded above the standard record level to give optimum performance in the data system.

6.0 Tape Crossplay Considerations

Figure D-6 illustrates the typical departure from optimum frequency response that may result when crossplaying wide band tapes which were recorded with heads employing different record-head gap lengths. Line AA is the idealized output-versus-frequency plot of a machine with record bias and record level, set upper IRIG standards, using a 3.05- μm (120-microinch) record-head gap length and a 1.02- μm (40-microinch) reproduce-head gap length. Lines BB and CC represent the output response curves of the same tapes recorded on machines with 5.08- μm (200-microinch) and 1.27- μm (50-microinch) record-head gap lengths. Each of these recorders was set up individually per IRIG requirements. The tapes were then reproduced on the machine having a 1.02- μm (40-microinch) reproduce-head gap length without readjusting its reproduce equalization.

6.1 The output curves have been normalized to 0 dB at the 0.1 UBE frequency for the purpose of clarity. The normalized curves may be expected to exhibit a ± 2.0 dB variance in relative output over the passband. The tape recorded with the shortest head segment gap length will provide the greatest relative output at the UBE.

6.2 While the examples shown are from older equipment with record gap lengths outside the limits recommended in subparagraph 6.5.4, chapter 6, they illustrate the importance of the record gap length in tape interchange applications.

7.0 Standard Tape Signature Procedures

The following subparagraphs describe the PCM signature and the swept-frequency signature.

7.1 PCM Signature Recording Procedures. Configure test equipment as described in paragraph 2.1, volume IV, RCC document 118. The configuration should simulate the operational link as closely as possible, for example, same RF frequency, deviation, bit rate, code type, predetection frequency, receiver bandwidth, and recorder speed.

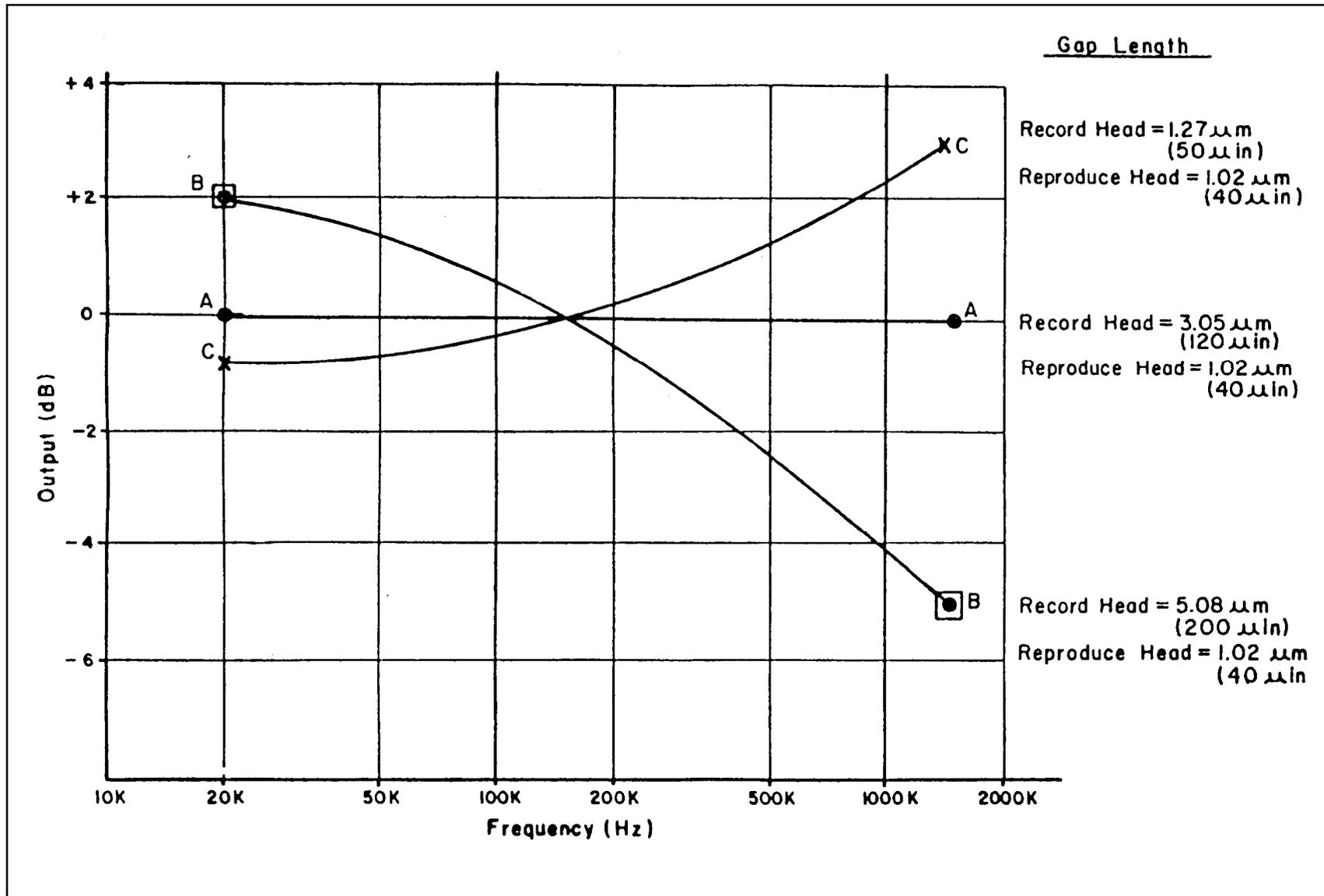


Figure D-6. Tape crossplay.

7.1.1 While recording the pseudo-random data at standard record level, adjust the signal generator output level until approximately one error per 10^5 bits is obtained on the error counter.

7.1.2 Record 30 seconds of the pseudo-random data at the beginning or end of the tape for each data track. A separate 30-second tape signature is recommended for each different data format.

7.1.3 The content, track assignments, and location on the tape leader and trailer of signature signals should be noted on the tape label.

7.2 PCM Signature Playback Procedure. The following subparagraphs explain the playback procedure.

7.2.1 Optimize playback equipment such as receiver tuning and bit synchronizer setup for data being reproduced.

7.2.2 Reproduce the tape signature and observe the error rate on the error counter.

7.2.3 Optimize head azimuth for maximum signal output and minimum errors.

7.2.4 If more than one error per 10^4 bits is obtained, initiate corrective action.

7.2.5 Repeat for each data track.

7.3 Swept Frequency Signature Recording Procedure. The following subparagraphs describe the recording procedure for the swept-frequency signature.

7.3.1 Patch a sweep-frequency oscillator output to all prime data tracks (up to 6 on 7-track recorders or up to 13 on 14-track recorders) (see appendix A, volume III of RCC document 118). As a minimum, patch the sweep oscillator to one odd and one even track.

7.3.2 Connect the sync output of the sweep oscillator to a track not used for sweep signals, preferably an outside track.

7.3.3 Record the signature signals for a minimum of 30 seconds at standard record level.



Record levels may be either preadjusted or quickly adjusted in all tracks during the first few seconds of the signature recording.

7.3.4 The content, track assignments, and location on the leader or trailer tape of signature signals should be noted on the tape label.

7.4 Swept-Frequency Signature Playback Procedure. The following subparagraphs define the steps for the playback procedure.

7.4.1 Connect the sync track output of the reproducer to the sync input of the scope.

7.4.2 Select an odd-numbered sweep-signal track and connect the output of the reproducer to the vertical input of the scope. Playback the sweep signal and adjust the scope gain for an amplitude of approximately ± 10 minor vertical divisions about the center baseline. Adjust the odd-track azimuth for maximum amplitude of the highest frequency segment (extreme right of the sweep pattern).

7.4.3 Observe amplitude variations through the sweep pattern and adjust the equalization, if necessary, to maintain the amplitude within the required tolerance over the required frequency range.

 <p>NOTE</p>	A decrease of sweep signal amplitude to about 0.7 represents a 3 dB loss.
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7.4.4 Repeat the playback procedure in subparagraphs 7.4.2 and 7.4.3 for azimuth and equalization adjustments of an even-numbered tape track.

7.4.5 Repeat the procedure in subparagraph 7.4.3 for equalization only of other selected prime data tracks, as required.

8.0 Equipment Required for Swept-Frequency Procedures

Equipment required at the recording site consists of a sweep-frequency oscillator having a constant amplitude sweep range of approximately 400 Hz through 4.4 MHz with frequency markers at 62.5, 125, 250, and 500 kHz and 1.0, 2.0, and 4.0 MHz. The sweep range to 4.4 MHz may be used for all tape speeds because the bandwidth of the recorder and reproducer will attenuate those signal frequencies beyond its range. The sweep rate should be approximately 25 Hz. Care should be exercised in the installation of the sweep generator to ensure a flat response of the sweep signal at the input terminals of the recorder. Appropriate line-driver amplifiers may be required for long cable runs or the low impedance of paralleled inputs.

8.1 A stepped-frequency oscillator could be substituted for the sweep-frequency generator at the recording location. Recommended oscillator wavelengths at the mission tape speed are 7.62 mm (300 mils), 3.81 mm (150 mils), 0.254 mm (10 mils), 0.0254 mm (1 mil), 0.0127 mm (0.5 mil), 0.0064 mm (0.25 mil), 0.0032 mm (0.125 mil), 0.0025 mm (0.1 mil), 0.0020 mm (0.08 mil), and 0.0015 mm (0.06 mil).

8.2 Equipment required at the playback site consists of an ordinary oscilloscope having a flat frequency response from 400 Hz through 4.4 MHz.

9.0 Fixed-Frequency Plus White Noise Procedure

The signature used in this method is the same for all applications. For direct recording of subcarrier multiplexes, only static nonlinearity (nonlinearity which is independent of frequency) is important for crosstalk control. Subparagraph 6.8.2 in chapter 6 provides a reference level for static nonlinearity. All formats of data recording are sensitive to SNR. Predetection recording and HDDR are sensitive to equalization. The following signature procedure satisfies all the above requirements.

9.1 Record a sine-wave frequency of 0.1 UBE (see table 6-3) with the following amplitudes.

9.1.1 Equal to the standard record level for direct recording of subcarrier multiplexes and HDDR (see subparagraph 6.8.2, chapter 6).

9.1.2 Equal to the carrier amplitude to be recorded for pre-detection recording of PCM/FM, PCM/PM, FM/FM, and PAM/FM.

9.2 Record flat band-limited white noise of amplitude 0.7 of the true rms value of the 0 dB standard record level as described in subparagraph 6.8.2, chapter 6. Noise must be limited by a low-pass filter just above the UBE.

9.3 Record with zero input (input terminated in 75 ohms). The three record steps previously described can consist of 10 seconds each. The spectra can be obtained with three manually initiated sweeps of less than a second each, because no great frequency resolution is required. All of the spectrum analyzer parameters can be standardized and set in prior to running the mission tape.

10.0 Signature Playback and Analysis

Before analyzing the signature, the reproducer azimuth should be adjusted. With the short signature, it is probably more convenient to use the data part of the recording for this purpose. If predetection recording is used, the azimuth can be adjusted to maximize the output as observed on the spectrum analyzer or on a voltmeter connected to the output. If baseband recording is used, the azimuth can be adjusted to maximize the spectrum at the upper end of the band. Using a spectrum analyzer, reproduce, store, and photograph the spectra obtained from paragraphs 9.1, 9.2, and 9.3 in this appendix. Store and photograph the spectrum analyzer input level of zero.

10.1 It is evident that any maladjustment of the recorder and reproducer or magnetization of the heads will result in the decrease of SNR across the band and will be seen from the stored spectra or photograph.

10.2 By having a photograph of the spectra, amplitude equalization can be accomplished without shuttling the mission tape as follows.

10.2.1 Use an auxiliary tape (not the mission tape, but preferably the same type tape). With a white-noise input signal band limited, adjust the amplitude equalization of the recorder and reproducer at the tape dubbing or data reduction site and photograph the output spectrum (see paragraph 9.0 of this appendix).

10.2.2 Compare this photo with the photo made from the signature. Note the difference at several points across the band.

10.2.3 Using the auxiliary tape, adjust the amplitude equalization to compensate for the differences noted.

10.2.4 Recheck with the mission tape to verify that the desired amplitude equalization has been achieved.

10.3 If the phase equalization is to be checked, a square wave signal can be added to the signature in accordance with the manufacturer's specification (see subparagraph 4.9.2.3.3 of volume III, RCC document 118). The same procedure as that recommended for amplitude equalization can be used, except based on oscillograms.

11.0 Recording and Playback Alignment Procedures

When using standard preamble (or postamble), see paragraph 6.12, chapter 6.

11.1 Recording of Preamble for Direct Electronics Alignment.

11.1.1 Patch a square wave generator output set to 1/11 band edge to all tracks having direct electronics or initiate procedure for recording internally generated 1/11 band edge square wave according to manufacturer's instructions.

11.1.2 If the preamble will be used for a manual adjustment, record for a minimum of 30 seconds at the standard record level and tape speed to be used for data recording.

11.1.3 If the preamble will be used only for automatic alignment, record at the standard record level and tape speed to be used for data recording for a sufficient time as specified by the manufacturer of the playback recorder reproducer or as agreed by the interchange parties.

11.2 Playback of Preamble for Direct Electronics Alignment.

For systems so equipped, initiate automatic alignment procedure per manufacturer's instructions. The procedure for manual adjustment is described in the following subparagraphs.

11.2.1 Display fundamental and odd harmonics of the square wave (third through eleventh) of selected odd numbered direct track near center of head stack on the spectrum analyzer. Adjust azimuth by peaking output amplitude of the third through eleventh harmonic. Final adjustment should peak the eleventh harmonic.

11.2.2 Repeat the above subparagraph for even numbered direct track. (Only one track is necessary for double density, 14-track, in-line system).

11.2.3 Observe frequency response across the band pass on selected track and correct if necessary. For a flat response, the third harmonic will be 1/3 of the amplitude of the fundamental, fifth harmonic 1/5 the amplitude, and so on. A convenient method is to compare the recorder/reproducer output with that of a square wave generator patched directly to the spectrum analyzer.

	An alternate but less accurate method is to optimize the square wave as displayed on an oscilloscope rather than a spectrum analyzer.
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11.2.4 Repeat the previous subparagraph for each direct track.

11.2.5 Display square wave on oscilloscope. Adjust phase for best square wave response as shown in figure D-7.

11.2.6 Repeat the previous subparagraph for each direct track.

11.3 Recording of Preamble for FM Electronics Alignment.

If available, initiate procedure for recording internally generated 1/11 band edge square wave and ± 1.414 Vdc per manufacturer's instructions. Otherwise, patch a square wave generator output to all tracks having FM electronics. A near dc signal may be obtained by setting the square wave generator to 0.05 Hz and ± 1.414 V or by using a separate dc source.

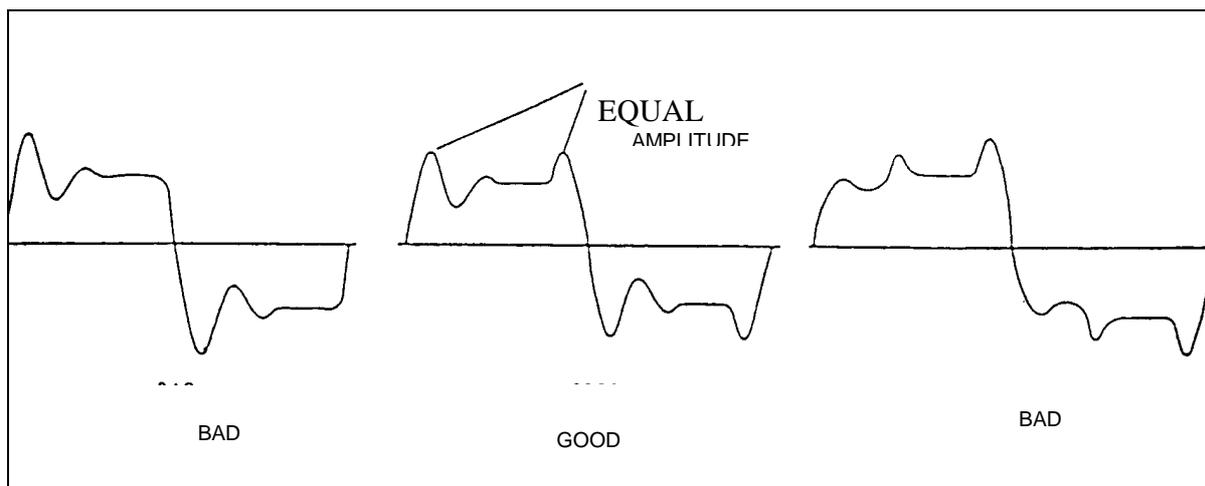


Figure D-7. Square wave responses.

11.3.1 If the preamble will be used for manual alignment, record at least one cycle of the 0.05 Hz square wave at ± 1.414 V or a positive and negative 1.414 Vdc for a minimum of 10 seconds each at the tape speed to be used for data recording. Next, record a 1/11 band edge square wave for a minimum of 20 seconds.

11.3.2 If the preamble will be used only for automatic alignment, record the above sequence for a sufficient time as specified by the manufacturer of the playback recorder/reproducer or as agreed by the interchange parties.

11.4 Playback of Preamble for FM Electronics Alignment.

For systems so equipped, initiate automatic alignment procedure per manufacturer's instructions. The procedure for manual adjustment is described in the next subparagraphs.

11.4.1 Check and adjust for 0 V output at center frequency per RCC document 118, Test Methods for Telemetry Systems and Subsystems, volume III, Test Methods for Recorder/Reproducer Systems and Magnetic Tape.

11.4.2 Use dc voltmeter to verify a full positive and negative output voltage on the selected track and correct if necessary.

11.4.3 Display fundamental and odd harmonics of the square wave (third through eleventh) on the spectrum analyzer.

11.4.4 Observe frequency response per subparagraph 11.2.3.

11.4.5 Repeat subparagraphs 11.4.1 through 11.4.3 for each FM track.