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Equalization for Computer Tape Recorders

M. R. Cannon
P. J. Seger

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Abstract

A two-volume update on magnetic recording, edited by Denis Mee and Eric Daniel, will be published in 1986 by McGraw-Hill, Inc., New York. The first volume covers magnetic recording principles, and the second book, "Magnetic Recording Applications," describes current magnetic recording technology. This report is taken from Chapter 4, "Data Recording—Tape," in the second volume.

"Data Recording—Tape," contains two major types of information, which will be discussed in two technical reports. This report describes the more technical portion of the chapter, which explains digital recording-channel signal processing. The processing of digital signals is different from the analog techniques used in other tape recorders. The increasing use of digital techniques for analog recording has resulted in the use of digital signal processing for applications other than computers. The techniques described in this report are useful for all digital magnetic recorders. The other technical report, "Introduction to Computer Tape Recorders," TR-82.0243, contains general descriptive information that is intended to inform novices about the applications and components of computer tape recorders.

ABOUT THE AUTHORS

Max Cannon is a Senior Technical Staff Member in the Magnetic Technology function. He has been with IBM for 35 years, and transferred from Boulder in 1978.

Paul Seger is an Advisory Engineer in the Magnetic Technology function. He has been with IBM for 17 years, and transferred from Owego, New York, in 1980.

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ODUCTION

The purpose of a digital magnetic recording system is to accept data from the user in a form convenient to the user, and to later return exactly that data in the same form. However, the nature of magnetic recording prohibits direct storage of the user's data. Several transformations occur in the process. There are unavoidable transformations due to the nature of magnetic recording, such as linear distortions, nonlinear distortions, and error generation. Also, there are transformations designed into the system to correct as many of the undesirable transformations as are necessary to meet the performance requirements of the system. The transformations introduced into the recording system that attempt to linearize its data-transmission qualities or shape the signals fall within the category of equalization. This report outlines the many aspects of equalization.

OG CHANNEL

The read and write heads plus the tape comprise the analog channel, as shown in Figure 1. Although digital signals are applied to the write head, analog signals are recovered from the read head.

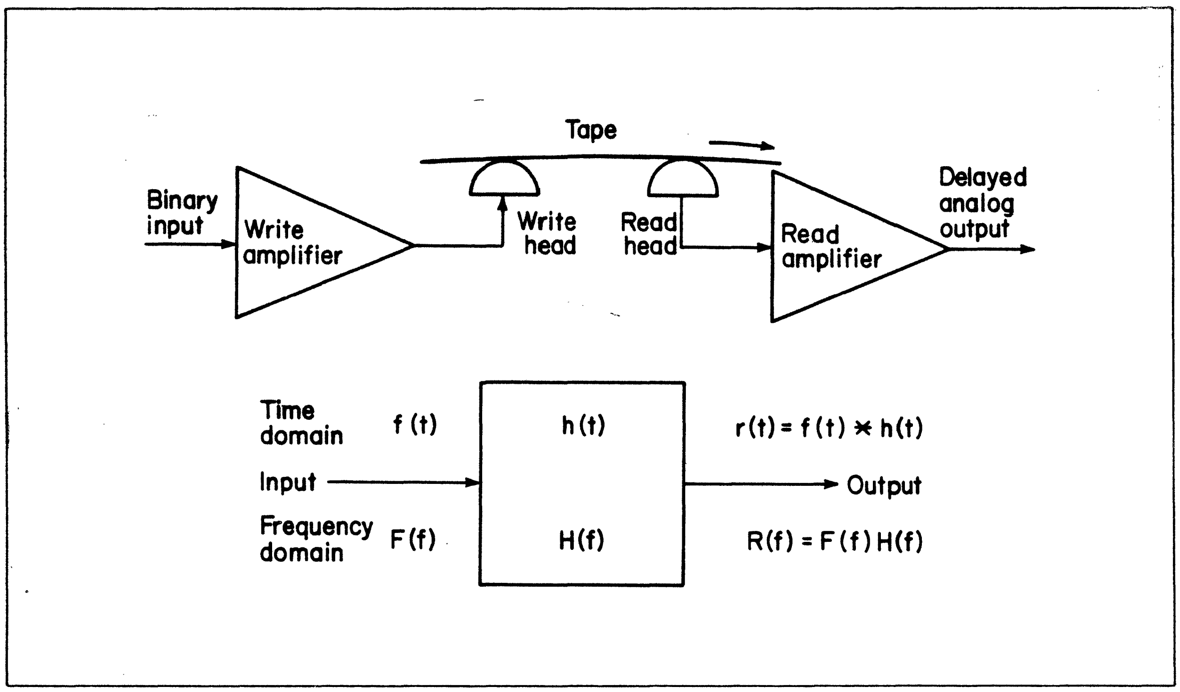


Figure 1. Analog Channel

Analysis of the analog-channel and signal-processing design requ
is simplified if the analog channel can be considered a "black b
linear characteristics. For the analog channel to be linear, th
response r(t) should change linearly with the driving function f
Symbolically,

$$\text{if} \quad f(t) \rightarrow r(t)$$

$$\text{then } k\, f(t) \rightarrow k\, r(t)$$

This is not the case for the typical, unbiased, magnetic-tape an
channel but, if only binary signals are applied so as to reverse
direction of magnetization at the tape surface, this may not be a
meaningful requirement. A second requirement for linearity is th
principle of superposition must apply, as discussed by Lathi [re
Symbolically,

$$\text{if} \quad f_1(t) \rightarrow r_1(t) \text{ and } f_2(t) \rightarrow r_2(t)$$

$$\text{then } f_1(t) + f_2(t) \rightarrow r_1(t) + r_2(t)$$

This property is shown by the typical computer-tape analog channe
binary input and partial penetration recording, as discussed by
Mallinson and Steele [ref.2] and Tjaden [ref.3]. Superposition i
valuable tool for analyzing computer-tape signal-processing needs
will assume that the analog channel is quasi-linear as used with
input signals having equal magnitudes of opposite polarity. The
due to this assumption is usually negligible. Under certain cond
a ternary signal satisfies the superposition requirement. Becaus
ternary signal cannot always be used, we will consider it illegit
for the purposes of this report.

COMPUTER TAPE HEADS

The read- and write-head elements are usually combined into one
read-write head assembly, as shown in Figure 2 on page 3. A full
dc erase head is generally included with partial penetration reco
tape devices to remove all previously recorded signals. It is
frequently separate from the read-write head. In most computer t
devices, the heads are stationary and the tape is moved. However
serpentine-format machine, the heads are also moved across the ta
various track positions. For a rotary-head device, the moving he
rapidly traverses a stationary tape that is stepped to the next t
the completion of the head pass. Rotary transformers couple the
and write electronics to the moving heads.

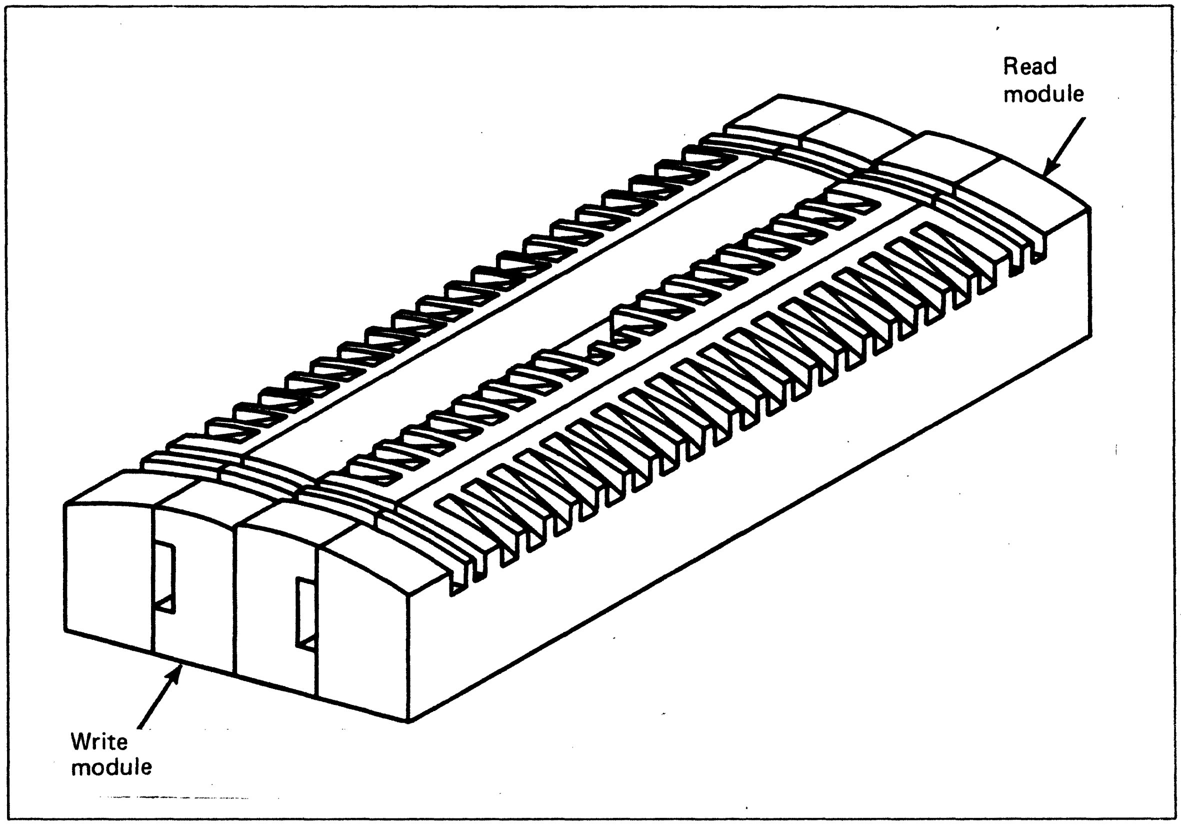


Figure 2. Two-Gap, 18-Track Tape Head

Separate read- and write-head elements are used to allow a verification of the written data, which is read just after it is written. The use of separate read- and write-head elements also permits optimization of the elements for their respective functions. The gap length, track width, throat height, and even the type of head may be different for the read and write elements. For example, a magneto-resistive head is a sensitive reader that does not write. Also, a read guard band is possible because different track widths are used for read and write.

In a multitrack head, the write-element gaps are in one plane with the read-element gaps in a parallel plane to avoid azimuth losses. A small write-to-read gap separation is important (particularly at high linear densities) to minimize the delay between data write and read verify. The verify operation requires sustained tape motion until the end of the block passes the read elements. By then the write elements are into the interblock gap at a distance equal to the write-to-read gap separation. Thus, for a fast start-stop machine, the interblock gap must be large enough to include the write-to-read gap distance plus the worst-case deceleration and acceleration times. Otherwise, a backhitch is required at the end of each data block (at least during writing). On the other hand, the interblock gap must be kept reasonably small to improve the tape utilization.

The write circuits and head elements act as transmitting antennas, and the read-head elements and amplifiers act as sensitive radio receivers. The radiated (and undelayed) write signal received by the read head is called feedthrough and, without proper shielding, it can be larger than the desired read signal from the tape. Feedthrough can result from either electromagnetic or electrostatic coupling, or a combination of both. Head wiring, as well as the read- and write-head elements, can cause feedthrough noise. The feedthrough must be small relative to the read signal to achieve an accurate read verification. As areal density increases, the feedthrough problem may be accentuated because the read signal generally decreases and the write signal may increase if a higher coercivity tape is used.

A multitrack machine can have crosstalk, which is the transfer of signal from one track to another. This crosstalk is usually not a significant problem if the system is adequately designed to eliminate feedthrough.

HEAD-TAPE INTERFACE

The recording-density capability of the head elements is largely dependent upon the head-tape interface. If the head-tape separation is large, the high-density or high-frequency components of the signal are attenuated and performance is adversely affected. If the separation is small, intermittent, or nonexistent, the head or tape may be prematurely degraded or destroyed due to wear. The protective hydrodynamic air bearing at the head-tape interface of a high-speed tape drive is produced by the relative motion between the head and tape. It can be controlled by the head contour and slots that maintain uniform head-tape separation in the region of all read and write gaps. Figure 2 on page 3 shows an 18-track tape head with air bleed slots to permit a small air bearing over each of the read- and write-head gaps.

Although modern computer tapes are quite smooth, their roughness is significant relative to the controlled physical separation of the tape from the head. The roughness of the head and tape, gap recession, flux-reversal transition width, and any magnetically dead layers on the surface of the tape or the head increase the effective magnetic separation (EMS) of the head-tape interface. The effective separation is typically about 0.2–0.3 µm larger than the physical separation between the head and tape, and even low-speed contact recording with a tape pressure pad typically has a finite magnetic separation.

ANALOG SIGNAL

The major analog-channel read-signal losses are due to excessive separation, thickness, and gap length. Figure 3 on page 5 shows the frequency-domain relationship between the physical parameters and the read signal. The low frequencies are attenuated at 6 dB/octave by the differentiating action of the read head. The high frequencies are attenuated by separation, recording thickness, and head-gap losses. The composite frequency-domain curve represents the transfer function of the quasi-linear analog channel. Figure 3 shows the effect on the analog signal of a ±50% change in read-gap length, recording thickness, and effective separation. The thickness loss can be adjusted within limits by varying the write current. The gap loss is not a significant factor if the read-head gap length is properly selected. Separation is the most critical loss in the high-linear-density operation. It includes physical separation, tape- or head-surface dead layers, and written transition length.

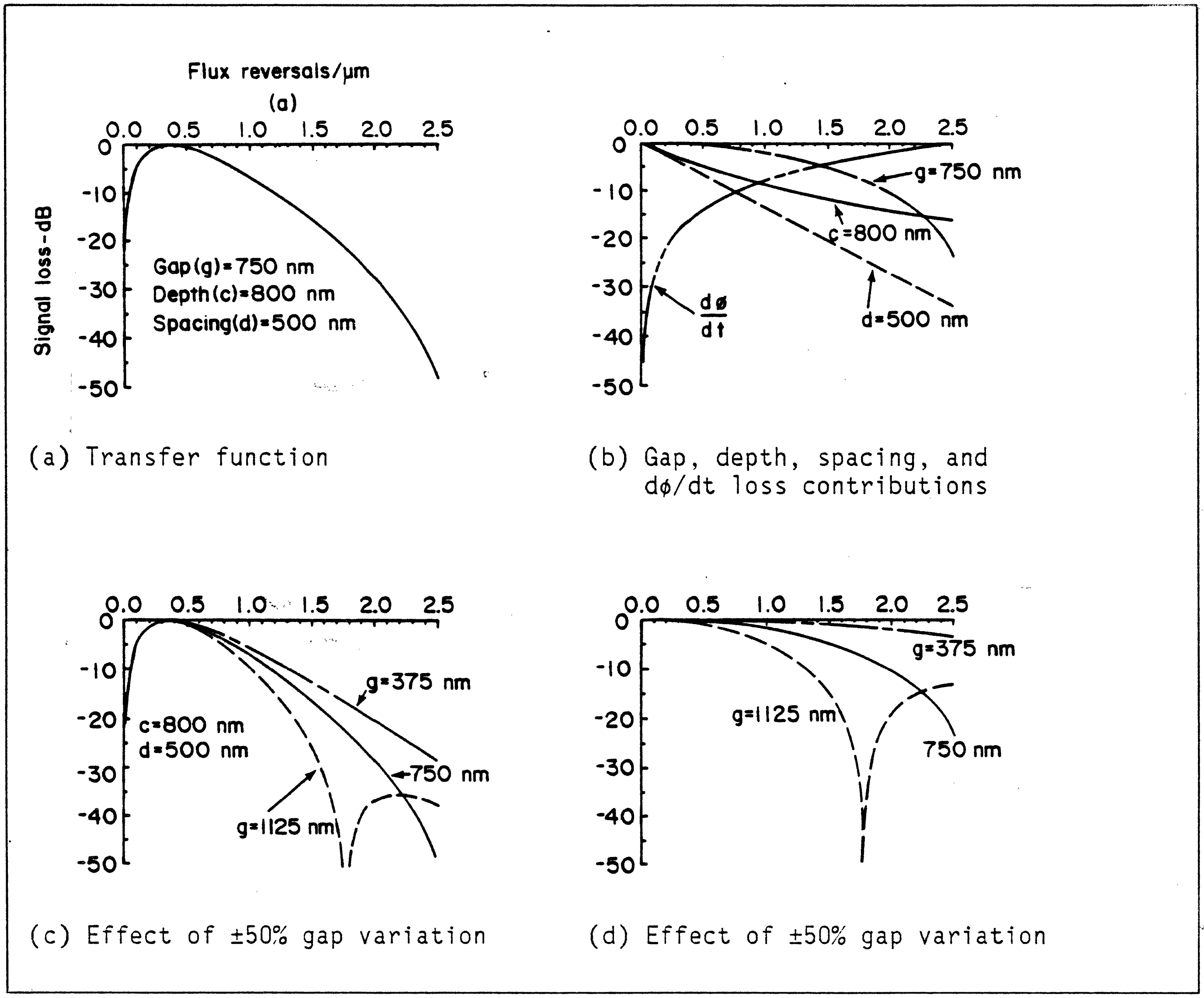
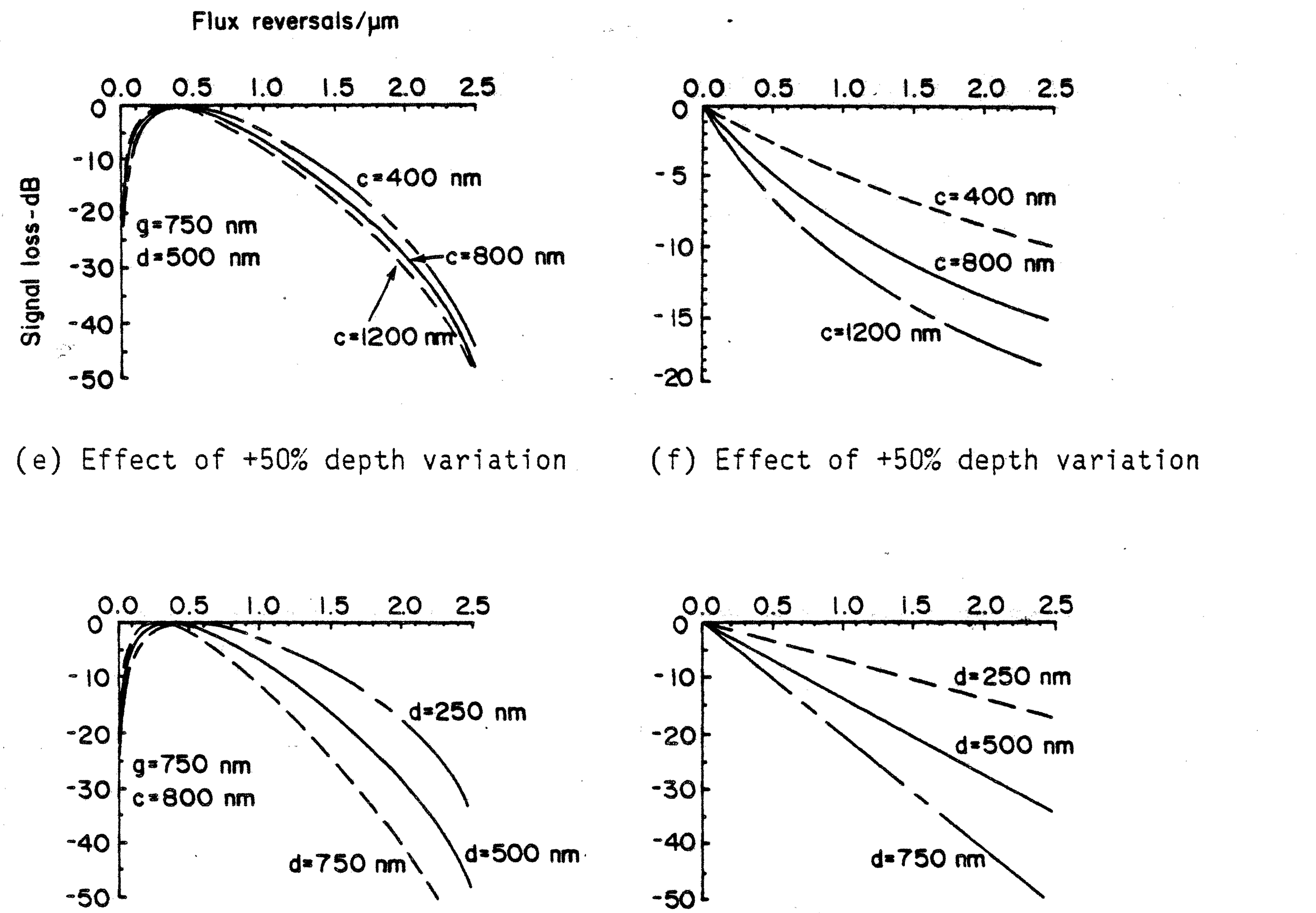


Figure 3 (Part 1 of 2). Analog-Channel Transfer Function



(e) Effect of +50% depth variation

(f) Effect of +50% depth variation

(g) Effect of ±50% spacing variation

(h) Effect of ±50% spacing variation

Figure 3 (Part 2 of 2). Analog-Channel Transfer Function. The effective values are gap = 750 nm, depth = 800 nm, and spacing = 500 nm.

The slope of the separation loss line in Figure 3 is the effective magnetic separation of the analog-channel head-tape interface, as discussed by Haynes [ref.4]. As a large surface defect passes the read gap, it increases the physical and magnetic separations, and results not only in a dropout, but also in a greater loss of high frequencies than low frequencies.

The transfer function shown in Figure 3a is calculated by using the physical parameters of a typical computer tape and head. It includes only the magnitude component of the transfer function, and represents the rms voltage from the read head. The peak-to-peak value of the signal is not attenuated at low frequencies or low densities as for the rms signal. It is possible to measure the transfer function of a specific head-tape combination with a modified pseudo-random binary-sequence input signal. The ratio of the output to input signal of the Fourier transforms provides the transfer function magnitude and phase, as discussed by Haynes [ref.5].

The design of proper signal processing for a high-density computer tape device requires an understanding of the analog-channel transfer function. This includes the nominal values plus the anticipated variations with manufacturing tolerances, degradation with time or environment, and tape interchange.

MODULATION CHANNEL

A computer tape recording system stores data, and then permits accurate retrieval of the information (sometimes years after it is recorded). Data integrity is a major requirement. The analog channel writes data onto the tape and reads a degraded signal back from the tape. To obtain reliable results from the heads and tape, it is necessary to modify the data signal before it is written and again after it returns from the read head. The electronic circuits that perform these operations, together with the analog channel, are called a modulation channel, as shown in Figure 4 on page 9. The portion of the modulation channel that determines the analog-signal characteristics is the equalization channel. The term recording channel is sometimes used to refer to the modulation channel plus a data flow that includes digital functions such as formatting and error control.

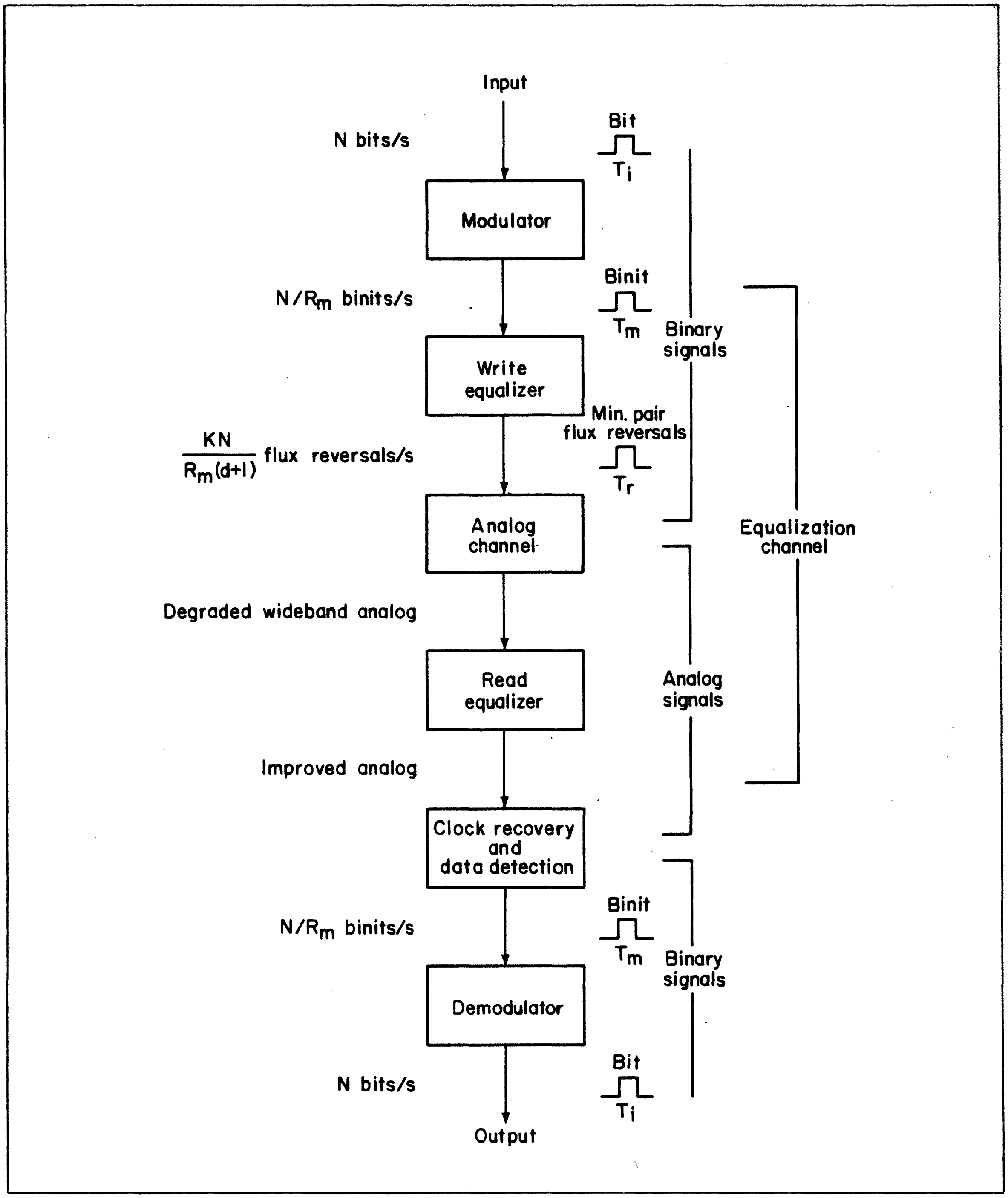


Figure 4. Modulation and Equalization Channels

The required signal-processing complexity depends upon the analog-channel transfer function, the linear density, and the desired data integrity. At low densities, little signal processing is needed and an inexpensive modulation channel performs well. As densities increase, more complex signal processing is necessary, and circuit costs increase.

The modulation channel does not operate error free, but it should be designed to minimize the frequency and duration of data errors within the available cost constraints. A computer tape device must reproduce accurately all data sequences, without exception. This means that it can distinguish between any possible code words, or that it will pass and detect the minimum difference waveform. In designing a modulation channel, the input is usually assumed to be random data, which contain all combinations of data sequences. The modulation-channel performance must be essentially the same, regardless of the data sequence. Therefore, worst-case patterns must be considered in the design, as well as the average signal characteristics.

Computer-tape modulation channels are sometimes compared in terms of data rate, cost, and accuracy. Data rate and cost are easily defined, but accuracy is somewhat ambiguous and described in terms of temporary or permanent errors, random or burst errors, and read or write errors. The probability of error designates statistically predictable random errors due to probabilistic channel impairments such as noise, which can be analytically determined. Most random errors involve a single binit or pair of binits, where a transition is shifted into the adjacent clock cell by noise or distortion. The modulation code may cause these errors to propagate to multiple data bits.

In most computer tape systems, random errors are overshadowed by the more frequent and more serious burst errors due to signal dropouts. Because this error rate cannot readily be predicted, the terms raw reliability or raw error rate (raw meaning before error correction) are used. Raw data reliability is the mean number of good data bits between error events. The error rate is the reciprocal of data reliability. There are various definitions for an error event. It may be a code word or a record containing an error, or a subsequent arbitrary number of good data bits can define the end of the error event. Raw reliability is empirically determined, and it varies considerably with tape quality. The measured raw reliability includes both random and burst errors, because they really cannot be separated. Raw reliability can be measured during a read-while-write, read-only, or read-retry (error-recovery) tape operation. This results in write, soft, and hard raw reliabilities (or error rates), respectively.

RECORDING SYSTEM LIMITATIONS

Because of the requirement for downward compatibility to earlier computer tape devices, signal-processing technologies have not advanced as rapidly as they might otherwise have done, as discussed by Hoagland [ref.6] and Lemke [ref.7]. Cost is the major limitation for small tape systems, and simple signal processing is necessary and adequate. Signal processing for many intermediate- and high-performance tape drives is controlled by compatibility constraints, and is similar to previous compatible devices. New-technology, high-performance products demand the highest practical data density from the available analog channel to achieve a large capacity and high data rate. The reasonable upper limit to data density is determined by the necessary data reliability, hardware cost, and operating speed. Because any error correction is possible, the practical limitation to user density is the affordable hardware cost and the required time to perform the necessary error correction. These limits usually require worst-case, end-of-life raw data reliability of approximately 10^6 bits per error for a high-performance tape system. The raw reliability of a high-density modulation channel is limited by dropouts, noise, interference, and distortion, as discussed by Thoma [ref.8].

A well-designed computer-tape modulation channel does not produce errors during nominal signal conditions. Errors are caused when the signal degrades, usually in a dropout. Even in a dropout, the signal can be read correctly if there are no other signal impairments, because it is the combination of low signal amplitude and noise or distortion that causes errors. During a severe dropout, the signal is so small relative to the noise and distortion that an error is inevitable. The signal in a moderate dropout is marginal, because it may be recoverable with proper signal processing; otherwise, it will result in an error. A slight dropout does not cause an error except in poor signal-processing circuits. The number and severity of tape dropouts depends upon tape quality. The bulk of the tape signal tends towards a Gaussian amplitude distribution, as expected from the central limit theorem. As shown in Figure 5 on page 12, dropouts also introduce a negative exponential distribution, as discussed by Haynes [ref.4]. The dropout depth at which errors begin to occur is a function of the noise and distortion in the signal.

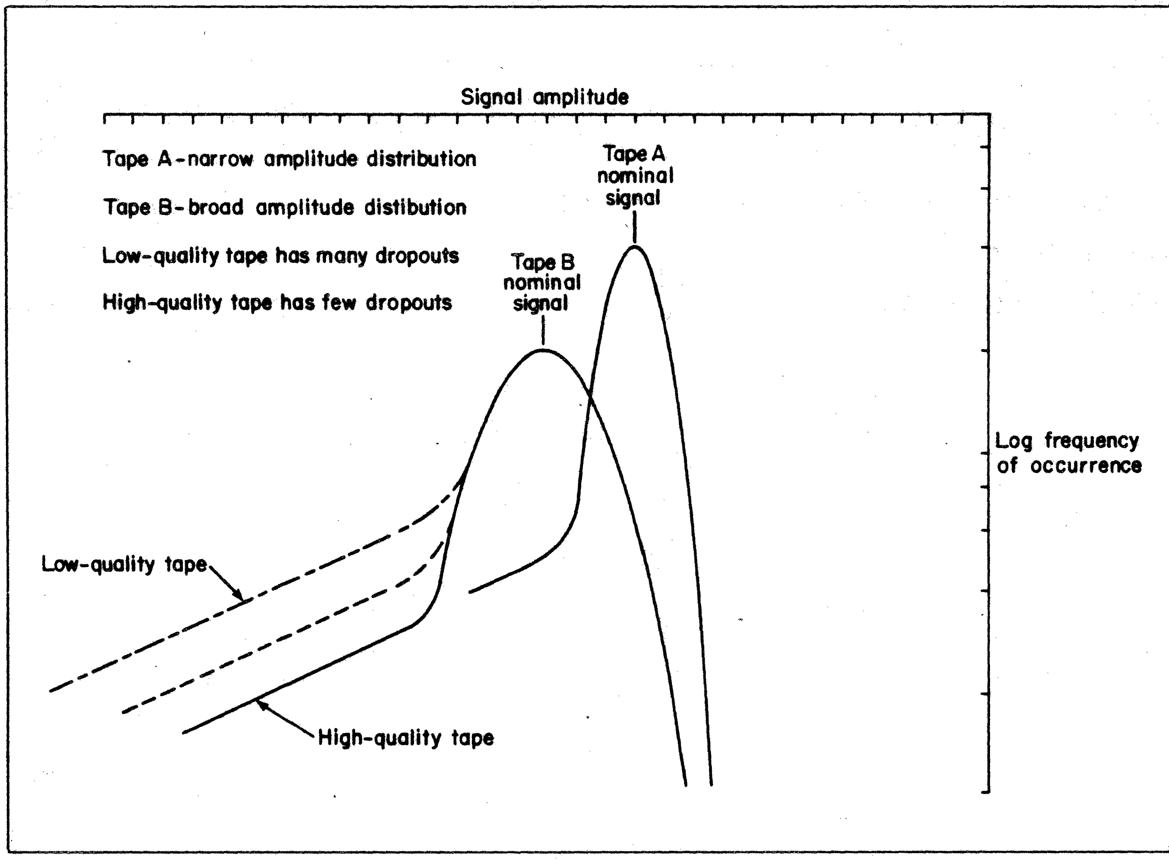


Figure 5. Amplitude Histograms for Typical Tapes

NOISE AND INTERFERENCE

A fundamental limitation in all recording channels is noise. The term noise is used to describe signal contamination, and signal-to-noise ratio (SNR) refers to the ratio of signal power to the contamination power. Extraneous signals of the same form as the desired signal are sometimes called interference, but unpredictable random signals generated by natural causes, whether internal or external to the system, are called noise, as discussed by Thoma [ref.8]. In the following pages, all signal contamination except distortion is included in the SNR, because separation of the noise components is virtually impossible. As the SNR decreases, the probability of error increases, as shown in Figure 6 on page 13. The exact probability of the error curve depends upon the modulation channel and dominant noise characteristics, and is typically several decibels to the right of the lower-bound curve in Figure 6. The actual curve can be approximated from experimental data.

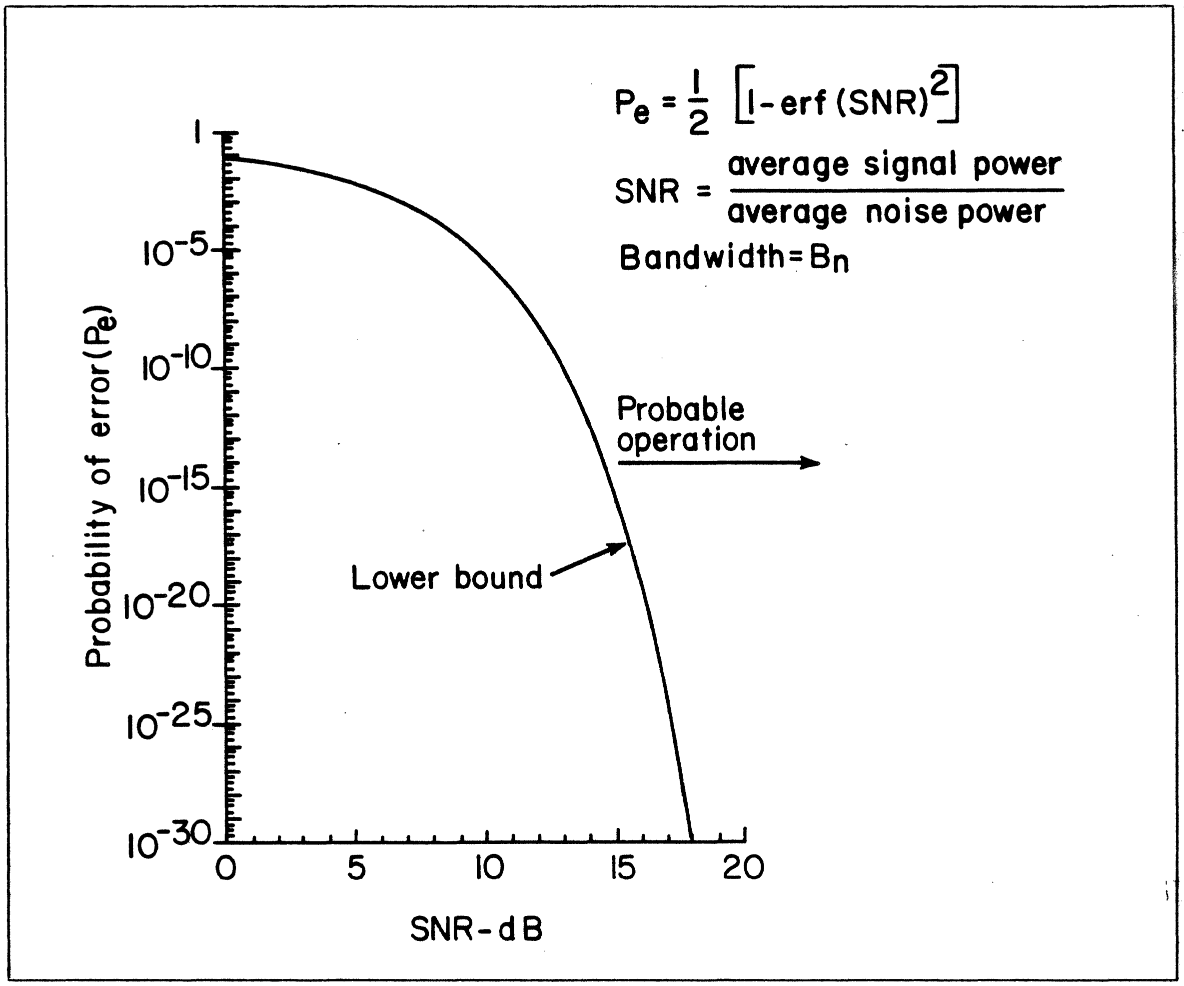


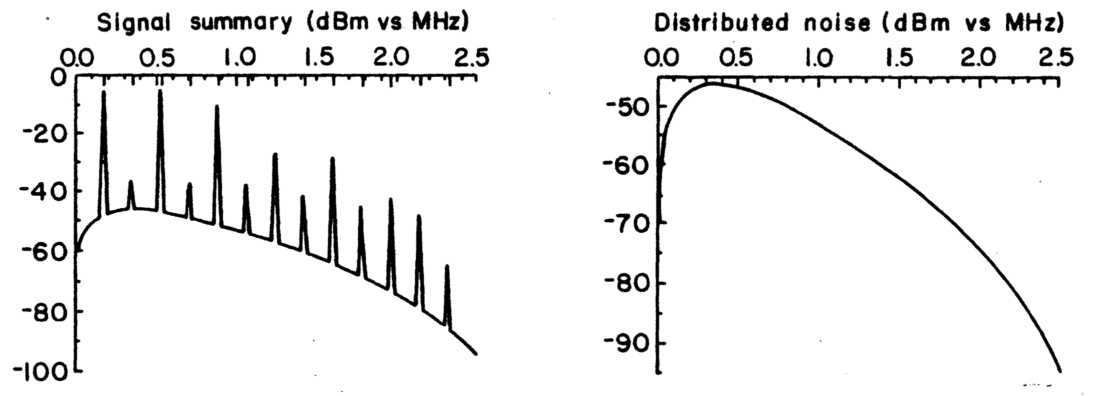
Figure 6. Lower-Bound Probability of Error. The typical tape modulation channel is several decibels to the right of this lower-bound curve.

Noise coupled with dropouts is detrimental to the raw reliability of a recording system. Because the signal amplitude decreases during a dropout, it is desirable to reduce all constant noise sources to significantly less than tape noise. If this is achieved, the tape signal and tape noise decrease simultaneously during the dropout. Ideally, the SNR is the same at the bottom of the dropout as in the nominal signal. Unfortunately, other noise sources usually predominate when tape signal and noise decrease more than a few decibels; therefore, the SNR in all significant dropouts is lower than during typical signal periods. This degradation increases the probability of an error due to noise during the dropout.

There are several tape-noise types and sources, as discussed by Arratia and Bertram [ref.9] and Thurlings [ref.10]. Particulate and off-track noises are wideband and additive. Erase or overwrite noise, and adjacent track noise due to read-head side sensitivity are low-frequency additive noises. There are also several sources of narrow-band, multiplicative, tape-dependent noises. Tape-surface irregularities, coating inhomogeneities, and variations in head-tape separation can amplitude modulate the signal. Therefore, narrow-band noise is a useful tool for evaluating tape quality, as discussed by Haynes [ref.4]. Tape-velocity variations frequency modulate the signal. Although some are insignificant, all of these undesirable effects degrade SNR. Relatively constant noise originates from the head, head-amplifier circuits, and other sources. Feedthrough can be a serious form of interference during a read-while-write operation, but print-through is seldom significant on computer tape devices.

Wideband SNR is an important consideration in the design of a modulation channel. It is the ratio of equalized pseudo-random-data signal power to noise power averaged over a reasonable length of tape. This signal contains a series of discrete frequencies. A computer-controlled spectrum analyzer is used to evaluate the power at the signal frequencies and the noise power between these frequencies. It is important to measure wideband noise with a recorded signal, as the noise floor is often higher for recorded tape than for unrecorded tape.

Wideband SNR is referred to as the input of the data detector, which converts the analog signal to a digital signal. The measurement is frequently made at the output of the analog channel, and SNR is predicted at the input to the data detector by using the equalizer characteristics. Figure 7 on page 15 shows the measurement of signal, wideband noise, and distortion at the output of the analog channel in Figure 3 on page 5; however, these values must be modified to include equalizer response. The signal measurement actually includes the odd harmonic distortion and some narrow-band noise, but this is not a significant error for a good analog channel.



(a) Measured response

(b) Noise floor

Discrete spectral peaks detected at:

kH	dBm	kHz	dBm
180	-5.24	1431	-42.17
359	-36.92	1610	-28.93
537	-5.01	1789	-45.22
716	-39.90	1967	-42.59
895	-10.56	2146	-48.25
1074	-38.14	2325	-64.64
1252	-27.62		

Discrete spectra power	-1.51 dBm
Odd harmonics	-1.51 dBm
Even harmonics	-32.48 dBm
Spurious lines	None
Distributed noise power	-26.31 dBm
Signal quality ratios	
SNR [odd÷(spurious + dist)]	24.79 dB
SDR [odd÷even]	30.97 dB

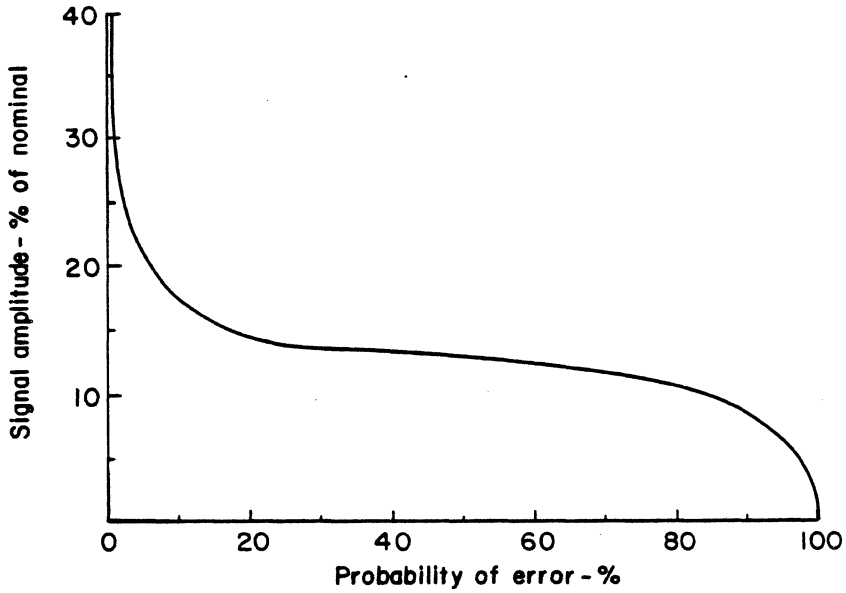
Figure 7. Measurement Technique for Signal, Wideband Noise, and Distortion

The shape of the noise spectrum in Figure 7b is similar to that of the analog-channel transfer function shown in Figure 3a on page 5, which indicates that the read losses filter the dominant tape noise in the same way that they filter the signal. If tape noise is not dominant over the entire operating bandwidth, the noise floor will rise at the frequencies where other noise sources exceed the tape noise. This often occurs at high frequencies where the filtered tape noise falls below the head or amplifier noise. If tape noise is not white, the noise floor will be different from the analog-channel transfer function. The ratio of signal power to noise power over the operating bandwidth has a spectral density as well as a single value—the integral over the required bandwidth.

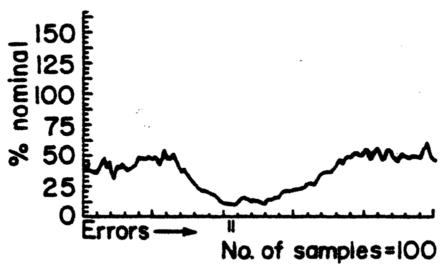
Bandwidth is an ambiguous term. Amoroso [ref.11] describes six different bandwidth definitions that are commonly used. Although we speak of bandlimited systems, they really only have rapidly attenuated frequency characteristics. Bandwidth depends upon the definition used, and the differences between various definitions can be large. The popular half-power or -3 dB bandwidth is not appropriate for this report. We will use the bounded power bandwidth, which is defined as the frequency range over which the equalization-channel response is within 40 dB of its maximum value.

For high-density recording, a nominal wideband SNR of at least 24 dB is desirable after modulation and equalization. This number is based upon experimental data, which indicates that raw reliability degrades rapidly for a nominal SNR of less than about 24 dB. Tape noise should dominate but, in a severe dropout, head or pre-amplifier noise is dominant. The signal amplitude may drop to about 25% of nominal (-12 dB) before errors become highly probable. This 12-dB signal loss typically lowers the SNR by several decibels, which will increase the probability of error by several orders of magnitude, as shown in Figure 6 on page 13.

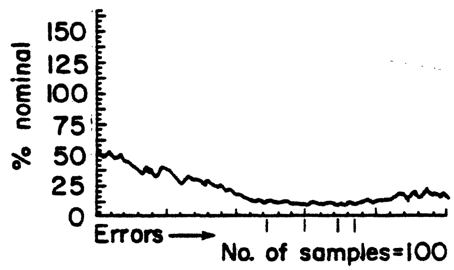
Approximately 90% of the total errors on a high-density experimental tape device occur when the signal amplitude falls below 25% of nominal. Figure 8 on page 17 shows the dropout signals for a typical tape system, including where the errors occur within a dropout, as discussed by Seymour [ref.12]. A well-designed modulation channel can tolerate signal losses of 10–12 dB without disastrous results. This implies that the operating density could be increased, and SNR reduced. Unfortunately, there are many more -6 dB dropouts than -12 dB dropouts. If errors occur with a 50% signal loss, the error rate will probably be intolerably high.



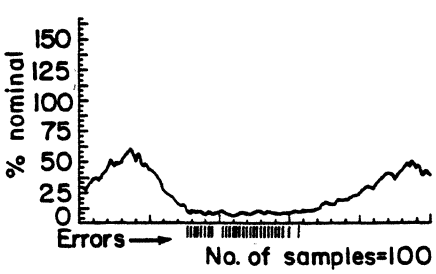
(a) Probability of error for low signal amplitudes



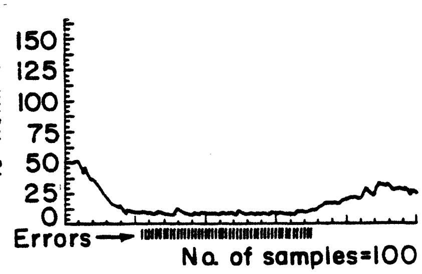
(b) Error locations (marks under axis) in dropouts



(c) Error locations (marks under axis) in dropouts



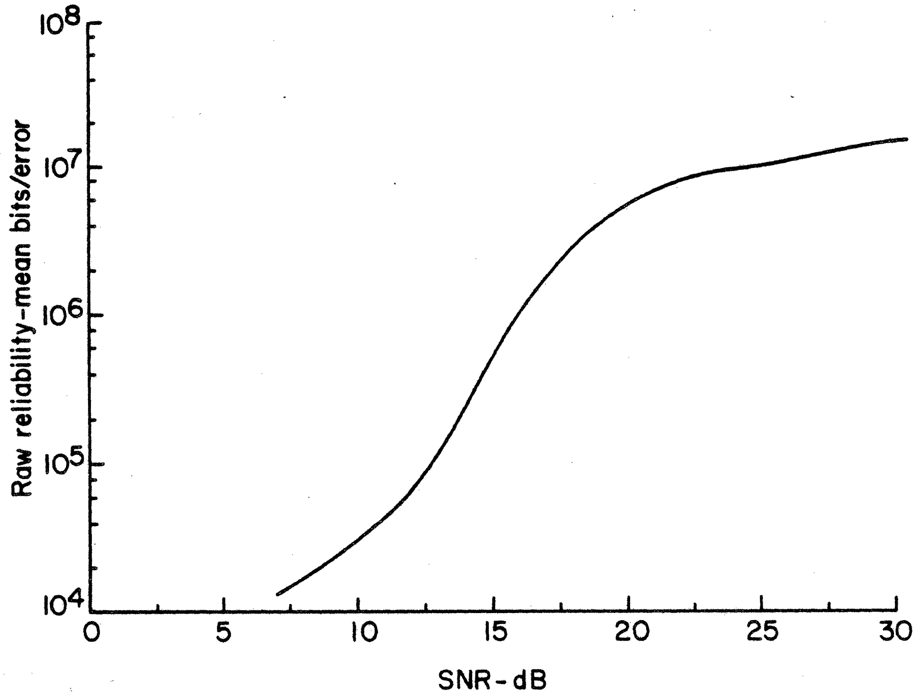
(d) Error locations (marks under axis) in dropouts



(e) Error locations (marks under axis) in dropouts

Figure 8. Errors in Dropouts

Experimental evaluation of raw reliability as a function of SNR verifies the relationship between noise and error rate. Figure 9 shows the effect of adding white Gaussian noise to the read side of an operating recording system as raw reliability is measured. The flat top is the result of deep dropouts that produce errors at any SNR. As more noise is added, the raw reliability degrades because less severe dropouts result in errors. This experiment includes only one type of noise, but the results indicate how noise can degrade raw reliability.



SNR = signal power/wideband noise power

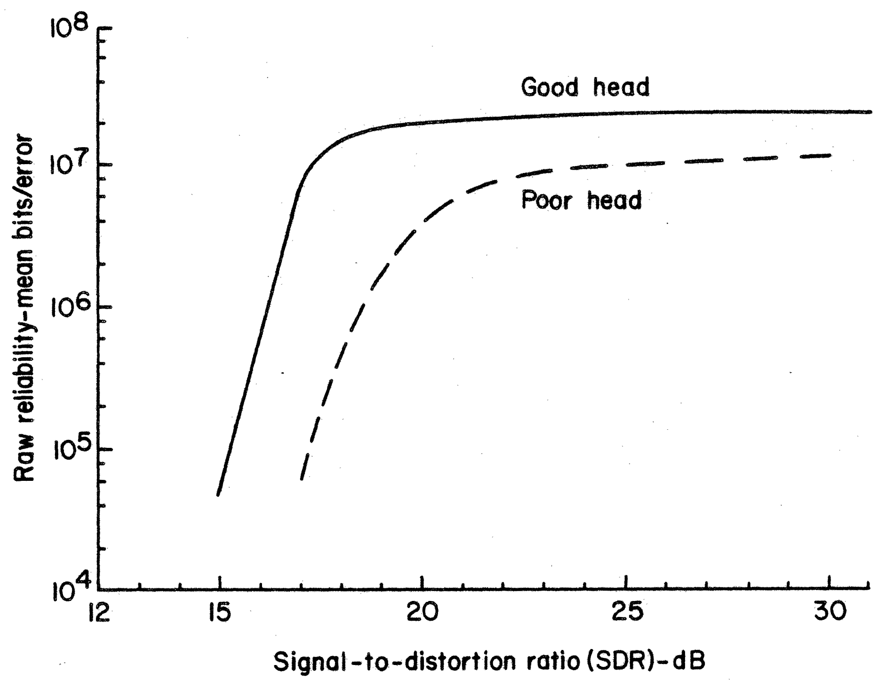
Figure 9. Raw Reliability versus SNR

Because of the complex nature of noise, including the multiplicity of types and sources, the probability of error cannot be accurately predicted for a specific analog channel. The more important noise statistics can be measured, but noise is not necessarily the dominant source of errors.

DISTORTION

Another source of errors from tape is distortion, which, like noise, alters the signal shape. Although we assume the analog channel is quasi-linear, the write process is nonlinear. This creates some nonlinear signal distortion, which is not significant at low densities but tends to increase with linear density, as discussed by Melbye and Chi [ref.13]. The use of a magneto-resistive read head can also introduce nonlinear distortion, as discussed by Shelledy and Brock [ref.14].

The effect of nonlinear distortion on data reliability is similar to the effect of noise. A signal-to-distortion ratio (SDR) can be measured by a similar technique to that used for measuring wideband SNR. A dc-balanced pseudo-random data sequence, which has no even harmonics, is used for the input signal. The SDR is the ratio of signal power to the even-order harmonic distortion power (including a small amount of narrow-band noise). Figure 10 on page 20 shows the effect of nonlinear distortion on raw reliability. This is an empirically derived relationship between raw reliability and SDR. Quadratic distortion is added to the read side of an operating channel as raw reliability and even harmonic distortion are measured. The location of the knee on the curve is determined by the analog-channel characteristics, including SNR, and signal-processing techniques. If nonlinear distortion is kept below a critical threshold, it is not a significant factor.



SDR = signal power/even harmonic distortion power

Figure 10. Raw Reliability versus SDR

Linear distortion, which is also present in the analog signal, results from the analog-channel characteristics. The restricted bandwidth, nonlinear phase, and deviation from flat amplitude versus frequency response of the analog channel cause the read-back signal to be a distorted replica of the write signal. Poor high-frequency response of the analog channel attenuates the higher signal frequencies. Bandwidth compression in the frequency domain results in time-domain expansion, as discussed by Lathi [ref.1], and the read pulses are broadened by the narrow bandwidth. The signal harmonic amplitude and phase are altered by the analog-channel frequency response, thus changing the shape of the read signal. This linear distortion is significant at high recording densities. Figure 11 on page 21 shows how linear distortion increases as the recording density is increased for the analog-channel transfer function shown in Figure 3 on page 5. This distortion is due to amplitude characteristics only, and linear phase response is assumed.

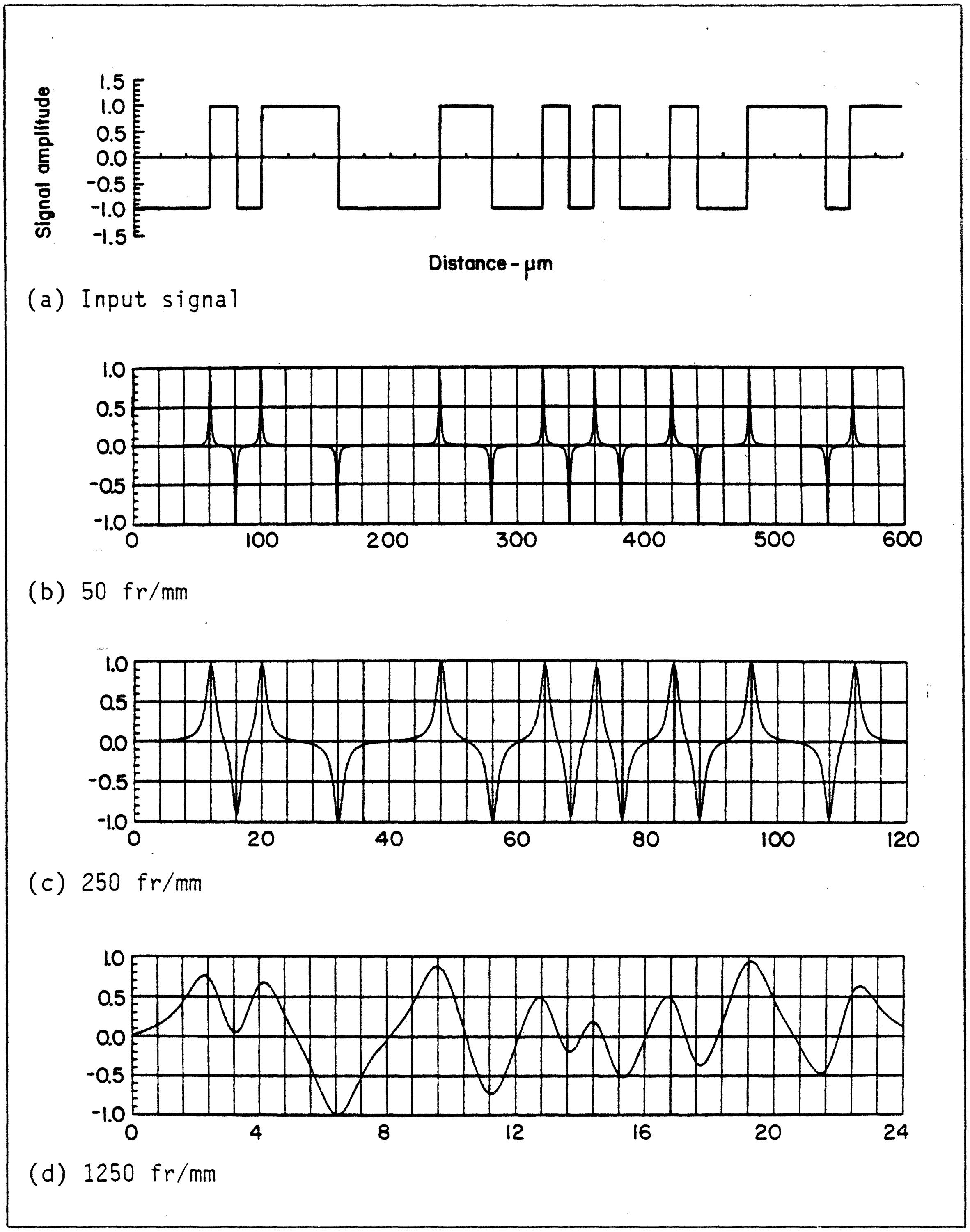


Figure 11. Intersymbol Interference. The analog-channel transfer function is shown in Figure 8 on page 17.

The time-dispersion characteristics of the analog channel cause a broad read-back pulse from the flux transition, as shown in Figure 11 on page 21. If the recorded flux transitions are close to each other, as for high recording densities, the broadened read-back pulses overlap. This linear distortion is called intersymbol interference (ISI). The degree of overlap depends upon the recording density and the analog channel. No intersymbol interference, which is a low recording density for this analog channel, is shown in Figure 11b. Figure 11c indicates some interference between pulses, and Figure 11d shows severe interference and a lack of amplitude and peak position uniformity at the highest density. If this interference is not corrected by signal processing, there is no obvious detection algorithm for recovering the data from the highest-density read signal. Interference with previous and future pulses is possible because of the time delay as the signal passes through the modulation channel, as discussed by Lathi [ref.1]. This bidirectional pulse crowding, even at the intermediate densities, can cause errors. Fortunately, this linear distortion can be at least partially compensated.

Linear distortion due to ISI increases in most dropouts because, as the head-tape separation increases, the high-frequency separation loss also increases. Thus, the distortion in a dropout is worse than in nominal data. The increased ISI, along with decreased SNR in a dropout, are the major causes of data errors.

For mechanical-design simplicity, high-performance computer tape recorders are designed to operate at high linear density and low track density, as discussed by Lemke [ref.7]. With wide tracks and high linear densities, noise may not be the major performance limitation, but intersymbol interference is frequently the limiting factor.

A minor limitation to recording digital data on a magnetic tape device is the poor low-frequency response in an analog channel. Most data sequences contain low frequencies and a dc component, which cannot pass through the analog channel directly. Another minor limitation is the loss of the data clock as the information is delayed by the computer tape recorder. There are practical techniques for handling these limitations.

Many factors affect raw data reliability. Some of these error-producing conditions can be beneficially influenced by signal processing, but others cannot be improved economically. The signal-processing architecture must make the optimum trade-off between performance improvement and increased cost. The following pages will describe some of the factors that minimize the effect of the recording-system limitations, and allow achievement of the desired results. Modulation-channel optimization involves equalization and data recovery as well as modulation. These are not independent variables, and must be optimized together rather than individually. Error control is briefly covered in a companion technical report, "Introduction to Computer Tape Recorders," TR-82.0243, and is not discussed in this report.

Modulation is the modification made to the data signal to facilitate its passage through the analog channel. As shown in Figure 12, a random data sequence has a (sin x)/x power spectral density and, although the signal alternates randomly between equal positive and negative values, the power spectral density has its maximum value at zero frequency, which cannot be directly transmitted through the analog channel.

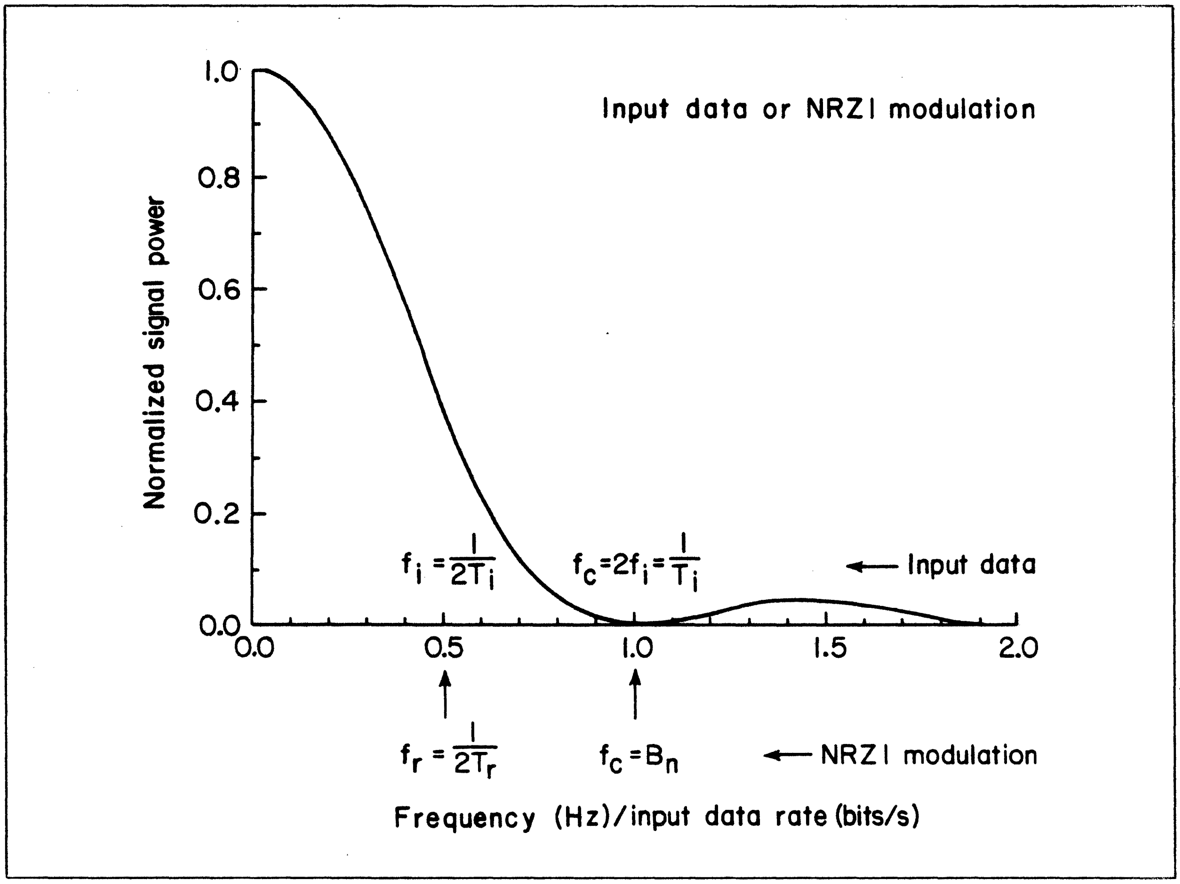


Figure 12. Power Spectral Density for Random NRZ or NRZI Data

Much of the signal power is at low frequencies and severely attenuated by the analog channel. The loss of the low-frequency components causes baseline wander, whereby the read signal drifts up and down according to the recorded pattern. This wander can complicate the equalization and data-recovery processes, as discussed by Bennett and Davey [ref.15] and Mallinson [ref.16]. Modulation permits the reduction of the low-frequency signal component, and even allows the conversion of the data to a dc-free signal. Modulation also simplifies the extraction of clock information from the data.

MODULATION CODES

Analog or digital modulation techniques can be used, but baseband digital modulation is almost universal for computer tape devices. Similarly, either binary or multilevel modulation techniques are possible, but binary modulation codes are generally used. Occasionally, a ternary modulation code is used; in this case, the three levels are plus, zero, and minus magnetization. An analog channel with a large SNR but narrow bandwidth might benefit from a higher-order multilevel modulation code. In general, the problems involved in linearizing the channel (as achieved with ac bias) and compensating for amplitude variations in dropouts are complications that make multilevel signaling with more than three levels undesirable. Multilevel modulation can be difficult if the tape is not completely degaussed, and the residual magnetization from previous recordings or dc erasure can present a problem even with ternary recording.

Digital modulation usually involves the conversion of a binary input data sequence to a different binary sequence with more desirable properties. The modulator converts m data bits to a code word with n modulation-code bits or binits. In most cases, $m < n$ and, thus, a higher frequency clock is required. Any binit in error may cause any or all of the m data bits within that code word to be in error. Virtually all modern computer tape devices use run-length-limited modulation codes, which are sometimes called group coded recording (GCR). These codes constrain the minimum and maximum number of contiguous binits without flux transitions, frequently designated by (d,k) respectively, as discussed by Franaszek [ref.17]. The minimum run-length constraint permits recording more information bits than flux reversals even if the ratio of modulator input bits to output bits is less than unity.

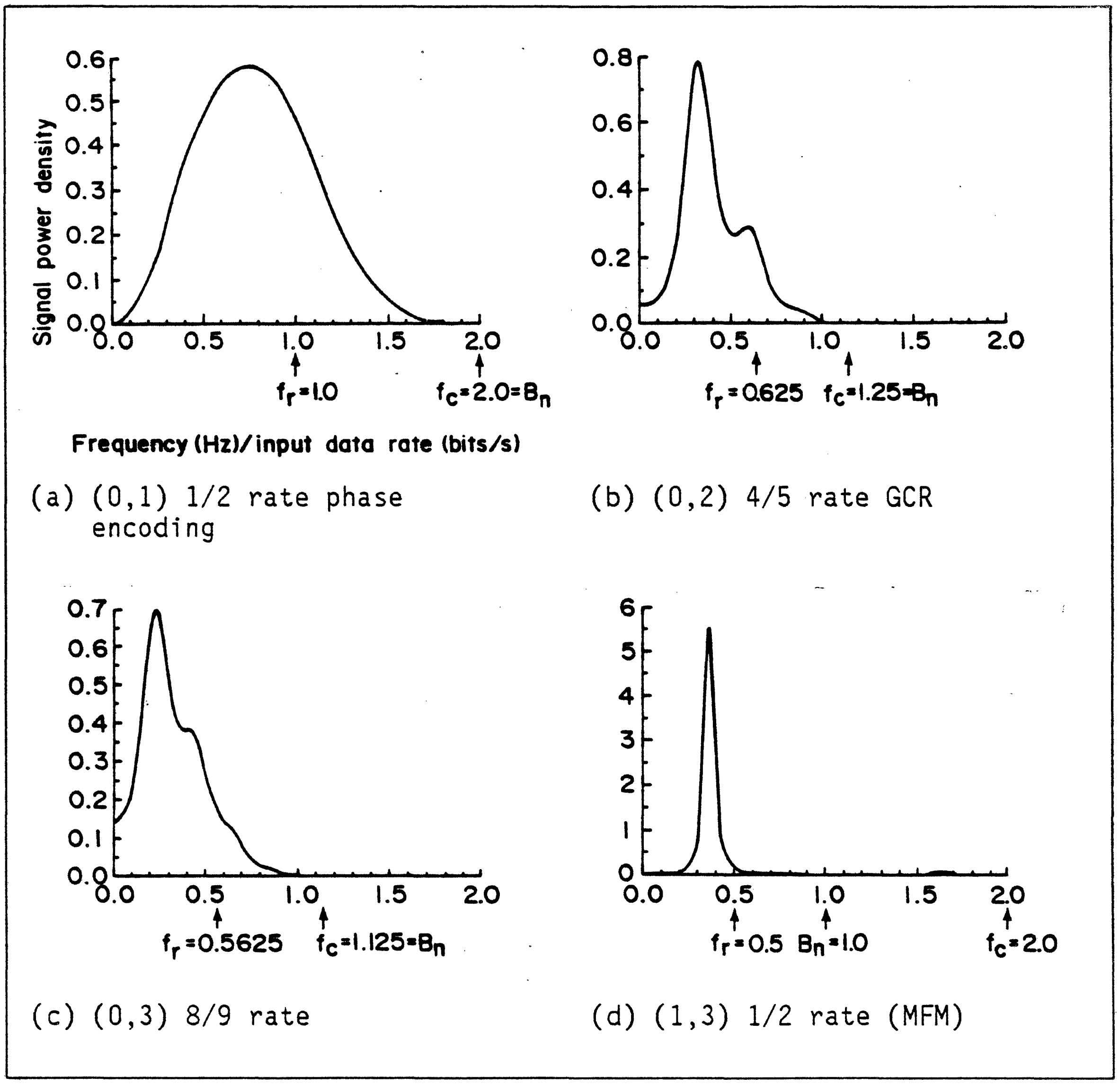
The data input signal to the modulation channel is assumed to be a random binary data sequence with bit period T_i seconds. If the signal polarity reverses for each bit, the input is a rectangular wave with a fundamental frequency of $f_i = 1/2\ T_i$ Hz. This is the highest fundamental signal frequency and, if it is directly written on tape, it produces the minimum fundamental wavelength, λ_{min}.

$$\lambda_{min} = f_i/v_t = 1/2\ T_i(v_t) \qquad (3)$$

where v_t is the tape velocity.

When the input signal contains random data, longer wavelengths are also recorded on tape—$\lambda_{min}(1,2,3,4,\ldots,n)$. These wavelengths represent frequencies $f_i(1,1/2,1/3,1/4,\ldots,1/n)$, but the signal frequencies also extend upward beyond the highest fundamental frequency as well as downward to zero Hz, as shown in Figure 12 on page 23. Frequencies beyond f_i may be needed to eliminate both timing and amplitude effects of intersymbol interference so that accurate clock and data may be recovered. In general, frequencies above $2f_i$ are not necessary for clock or data recovery.

The data output signal from the modulator is a binary sequence with a binit period T_r seconds. The power spectral densities for several modulation codes are shown in Figure 13; they are averages of long random data sequences. Although some of them seem to compress the bandwidth requirements, sufficient bandwidth is required to unambiguously transmit the worst-case code words, even though they do not occur often.



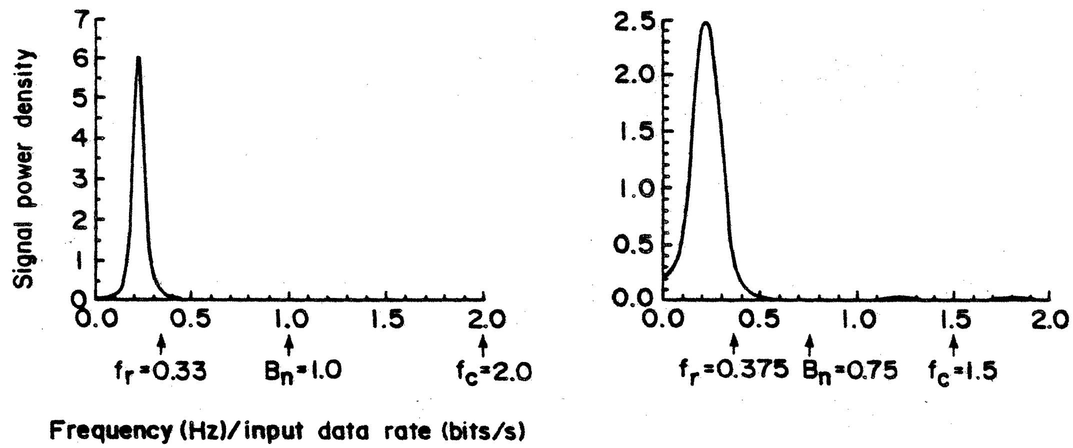
(a) (0,1) 1/2 rate phase encoding

(b) (0,2) 4/5 rate GCR

(c) (0,3) 8/9 rate

(d) (1,3) 1/2 rate (MFM)

Figure 13 (Part 1 of 2). Power Spectral Densities for Constant Signal Power



(e) (2,7) 1/2 rate

(f) (1,7) 2/3 rate

Figure 13 (Part 2 of 2). Power Spectral Densities for Constant Signal Power

The linear-power scale plots shown in Figure 13 can be deceiving, because the signal energy extends well beyond the apparent high-frequency limits. In these plots, $f_r = 1/2\ T_r$ is the highest fundamental recorded frequency (where T_r is the minimum recorded pulse duration). Substitution of f_r for f_i in Equation 3 gives the minimum recorded wavelength of a modulated signal. The binit period T_m determines the read-clock rate f_c; there is a read-clock cycle for each binit.

Early tape machines used NRZI modulation, where the flux direction was reversed for each 1, but not for 0. The parity track in the byte-wide parallel format ensured a transition in at least one track for each byte, which provided clocking information. As linear densities increased beyond 31.5 B/mm (800 BPI), static and dynamic skew prevented clock synchronization from other tracks, and a self-clocking modulation code was needed. Phase encoding (PE) is used for the 63 B/mm (1600 BPI) format. Although less efficient than NRZI, at least one flux transition is guaranteed in each bit cell of each track to provide clock information. The flux-transition direction defines the bit value, because an additional transition is required between similar adjacent bit values. Knoll [ref.18] compares the performance of these early tape modulation codes. For the 246 B/mm (6250 BPI) format, sometimes referred to as group coded recording, m = 4 data bits and n = 5 binits. This run-length-limited modulation code has sufficient flux transitions to synchronize a separate clock for each track. Further increases in the modulation rate have been made with the 1496 B/mm (38 KBPI) 18-track format, where m = 8 data bits and n = 9 binits. Typical modulation-code waveforms are shown in Figure 14 on page 27.

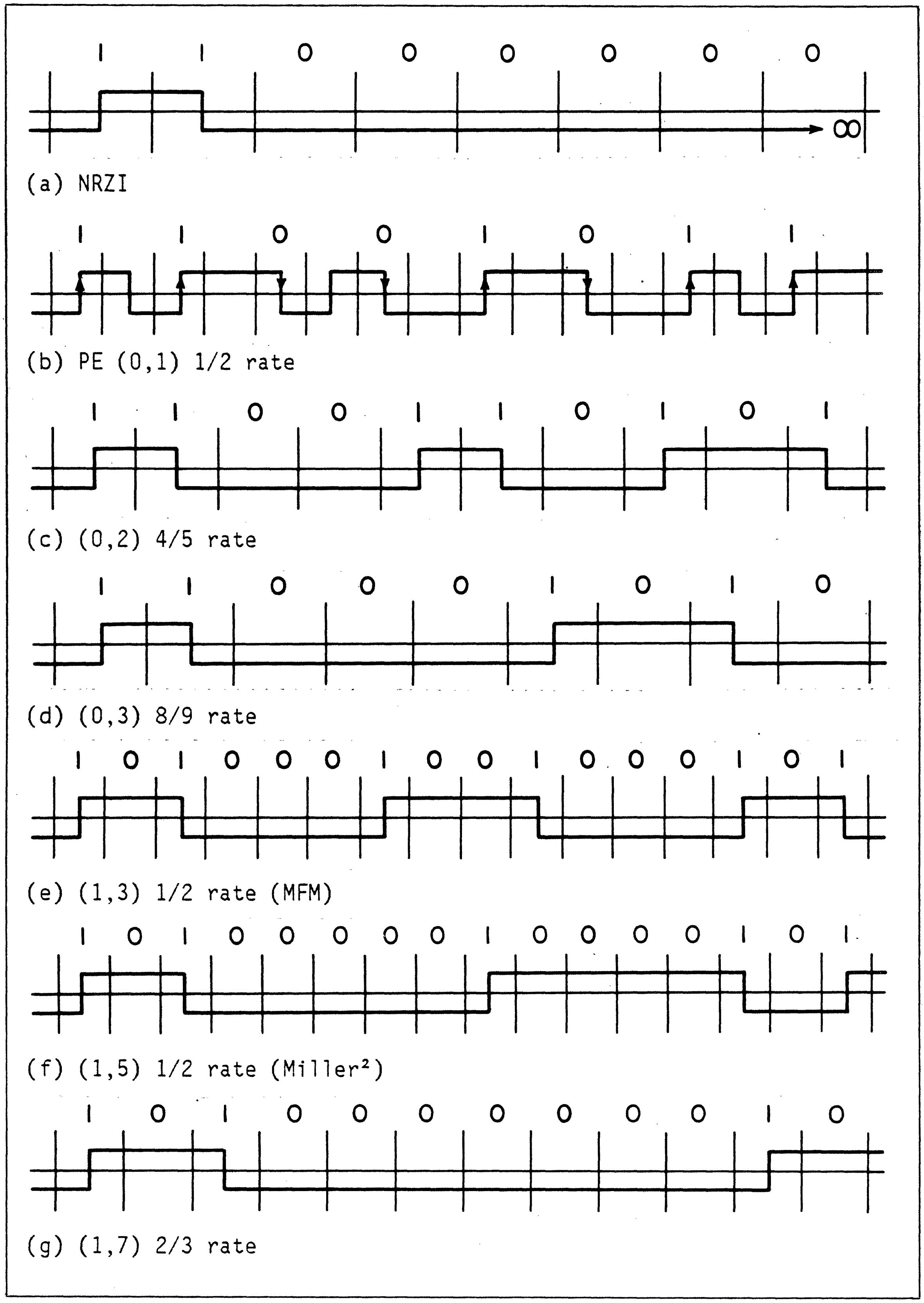


Figure 14 (Part 1 of 2). Typical Waveforms for Modulation Codes

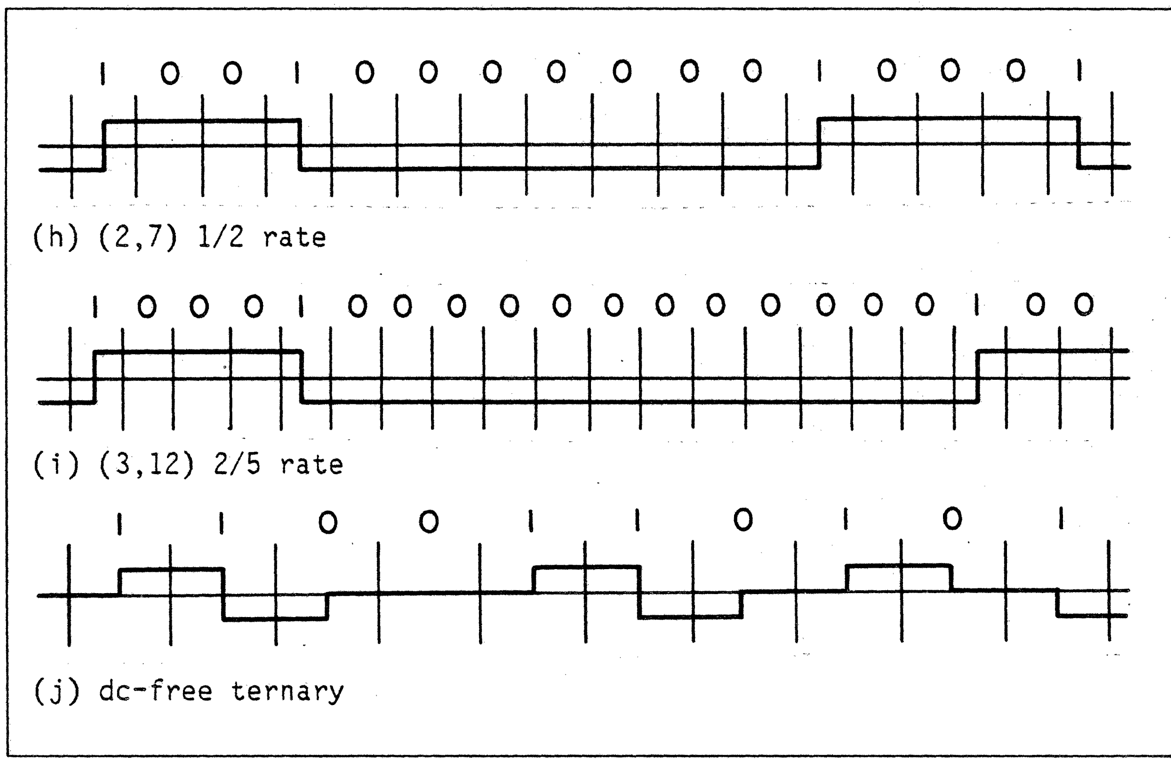


Figure 14 (Part 2 of 2). Typical Waveforms for Modulation Codes

Many digital modulation techniques produce a dc component that cannot pass through the analog channel, but can be seen from the power spectral density, as shown in Figure 12 on page 23. However, a linear power scale does not always make the dc component visible. For example, the (1,3) 1/2 rate code called MFM or Miller code has a dc component, which is not readily apparent in Figure 13d on page 25. A digital sum variation (DSV) indicates the maximum possible charge accumulation from a modulation code. An infinite digital sum variation indicates a dc component, and a finite value identifies a dc-free code.

The 1600 BPI phase-encoded signals used on many tape drives are dc-balanced with a DSV of one bit duration T_i. The IBM 3851 Mass Storage Facility uses a dc-free code called zero modulation (ZM), which has a digital sum variation of $3T_i/2$, as discussed by Patel [ref.19]. Other dc-free binary codes have been proposed by Mallinson and Miller [ref.20], Spitzer et al. [ref.21], and Yoshida et al. [ref.22]. Bipolar ternary codes can be dc-free. For example, any binary modulated sequence can have the dc removed by an additional ternary modulator, which has zero output except at input transitions. This produces alternating bipolar pulses, as shown in Figure 14j and discussed by Bennett and Davey [ref.15].

Although it is possible to include a clock-timing transition in every other binit of the modulated data sequence as with phase encoding, this is an inefficient utilization of the available bandwidth. When run-length-limited modulation codes are used with an adequate bandwidth equalization channel, the clock is recoverable from the data-flux reversal signals, which are submultiples of the clock frequency.

MODULATION CHARACTERISTICS

Several important characteristics should be considered in selecting a code for a magnetic-tape modulation channel. The modulation-code rate R_m is usually called the modulation rate, which is the ratio of modulator input bits to output binits. This is an important factor in determining the clock frequency, bandwidth requirement, and spectral efficiency. The run-length constraints d and k influence the recording ratio R_r (the ratio of information bits to flux reversals), and the wavelength ratio R_λ (the ratio of the maximum to the minimum recorded wavelength). The value of k also represents the maximum number of clock cycles or binits between synchronizing pulses. If this number is large, clock synchronization becomes more difficult, particularly if the tape instantaneous velocity varies significantly or if there is a large amount of intersymbol interference. A code with a dc component may require dc restoration to reduce baseline wander problems.

The nominal bandwidth B_n (to zero response or -40 dB) for no intersymbol interference at clock- or data-recovery times may be larger than the actual bandwidth, because equalization is selected for best overall performance rather than complete elimination of intersymbol-interference problems. The bandwidth is a compromise between interference, SNR, and hardware cost. A narrower bandwidth than B_n allows some intersymbol interference, but increases SNR. The modulation spectral efficiency η_s is an indication of bandwidth utilization, which is the ratio of the input-data bit rate to the nominal interference-free bandwidth B_n. It is sometimes used to compare linear density capabilities of various modulation codes. This should only be done with caution, however, as the other modulation-code characteristics are also important. For example, a modulation code with a large value of d might have a large recording ratio but low spectral efficiency. The improved SNR that results from the lower recording density allows use of a much narrower bandwidth than B_n. This code could conceivably outperform another code with higher spectral efficiency.

By assuming that the data-input sequence-bit width (T_i s/bit) or frequency (f_i Hz or $2f_i$ bits/s), modulation-code rate (R_m), and (d,k) constraints are known, the modulation code can be analyzed as follows:

$$R_r = R_m(d+1) \quad \text{information bit/flux reversal} \tag{4}$$

$$R\lambda = (k+1)/(d+1) \tag{5}$$

$$T_r = T_i\,R_r = T_i\,R_m(d+1) \quad \text{s/flux reversal} \tag{6}$$

$$f_r = 1/2T_r = f_i/R_m(d+1) \quad \text{Hz} \tag{6a}$$

$$T_m = T_i\,R_m \quad \text{s/binit} \tag{7}$$

$$f_c = 1/T_m = 2f_i/R_m \quad \text{binit/s} \tag{7a}$$

$$T_w = \pm T_c/2 = \pm T_i\,R_m/2 \quad \text{seconds} \tag{7b}$$

$$\begin{aligned} B_n &= 1/2T_m = f_i/R_m \quad \text{Hz} \quad \text{if } d > 0 \\ &= 1/T_m = 2f_i/R_m \quad \text{Hz} \quad \text{if } d = 0 \end{aligned} \tag{8}$$

$$\begin{aligned} \eta_s &= 2f_i/B_n = 2R_m \quad \text{bit/s/Hz} \quad d > 0 \\ &= R_m \quad \text{bit/s/Hz} \quad d = 0 \end{aligned} \tag{9}$$

where

T_r = minimum time between flux reversals
T_m = binit or clock-cycle duration
T_w = timing window duration

The importance of the modulation-code rate can be seen from these equations. A large R_m increases the recording ratio, signal amplitude, timing window, and spectral efficiency. The clock window becomes larger as the number of modulation-code binits decreases, thus creating fewer problems due to intersymbol interference or tape velocity variations. In general, large R_m and large d are desirable. A large d often requires the use of a small modulation-code rate. A good compromise must therefore be selected. A simple guideline for selecting a modulation code is to choose the largest value of R_m that provides all of the needed characteristics, the largest value of d with the selected modulation rate, and the smallest value of k within these constraints. A run-length-limited modulation code requires less bandwidth than the input data sequence only if $d > 0$ and $R_m > 1/2$. Pelchat and Geist [ref.23] observe that some of the bandwidth-compression modulation codes are "bandwidth expansion codes in disguise." Figure 15 on page 31 shows the characteristics of some run-length-limited modulation codes for digital recording.

R_m	d,k	R_r	R_λ	$B_n T_i$	η_s
1	0,∞	1	∞	1	1
1/2	0,1	1/2	2	2	0.5
4/5	0,2	4/5	3	1.25	0.8
8/9	0,3	8/9	4	1.125	0.889
1/2	1,3	1	2	1	1
1/2	1,5	1	3	1	1
2/3	1,7	4/3	4	0.75	1.33
1/2	2,7	3/2	8/3	1	1
2/5	3,k	8/5	(k+1)/4	1.25	0.8
2/5	4,k	2	(k+1)/5	1.25	0.8
$\frac{\text{bits}}{\text{binit}}$		$\frac{\text{bits}}{\text{flx rev}}$	$\frac{\text{max}}{\text{min}}$	$\frac{\text{cycles}}{\text{bit}}$	$\frac{\text{bits/sec}}{\text{Hz}}$

Figure 15. Comparison of Modulation Codes

Only power-limited modulation codes are practical for computer tape recorders because of tape saturation. These are restricted to bandwidths that provide 2 or less bits/s/Hz. More efficient modulation codes used in terrestrial microwave systems have spectral efficiencies between 3 and 6 bits/s/Hz, but they use high-energy pulses, as discussed by Feher [ref.24].

EQUALIZATION

Equalization involves the addition of linear filters to modify the equalization-channel transfer function so as to provide more reliable data detection. This technique compensates for some of the channel shortcomings. Most of the linear intersymbol interference in a computer tape recording system is caused by the limited bandwidth of the analog channel and its amplitude roll-off with increasing density. These undesirable characteristics are not a serious problem, because within the capability of the analog channel are several possible transfer functions that provide good performance. Typically, the equalizer filter is in the read-signal-processing circuitry but, under certain conditions, part of the filtering can be done on the write side of the analog channel.

Equalizers are designed to provide output data signals that conform to specified constraints at the data detector. Intersymbol interference is not eliminated, but it is controlled. As a result, linear distortion is not a serious problem except perhaps during dropouts when the analog-channel transfer function varies considerably. The ideal equalizer cannot be achieved, but the desired characteristics can be approximated. Read-equalizing filters modify the noise spectrum as well as the signal. An improvement in distortion is usually accompanied by a degradation in SNR. Channel design becomes a compromise attempt to minimize the performance degradation due to noise and distortion at an acceptable hardware cost.

The equalization process requires consideration of the entire recording channel. Because there is no specified sequence by which to approach the problem, a good place to begin is at the detector by determining what detectable parameters exist in a signal. These parameters may be magnitude, or points in the signal at which it crosses zero or has an inflection, or sample points with respect to time. The bandwidth and recording density are critical parameters that are traded against detection difficulty and the probability of error. The desired pulse at the detector input is determined as the result of these compromises. After the pulse is specified, the transfer function between the input-data write sequence and the input to the detector is determined with the aid of Fourier analysis by dividing the output-sequence spectrum by the input-sequence spectrum. This total input-to-output transfer function is the ideal that the equalization channel attempts to achieve. The products of all transfer functions within the channel should equal this ideal. After measuring and dividing those transfer functions that are given as fixed (for example, the head-tape interface), the remainder is the transfer function of the required equalization. This equalization may consist of a single equalizer on the read side or multiple equalizers with some equalization placed on the write side. How well the ideal is chosen and how well it is approximated are the challenges of equalization.

The equalizer-design goal is to provide the best possible signal to the data detector. The equalization-channel output signal obviously depends upon the input signal, and the transfer function that includes the analog channel and all equalizing filters. The binary input to a typical computer-tape equalization channel can be considered a series of bipolar rectangular pulses having unit amplitude and duration T_m, as shown in Figure 16a. As an alternative, the input signal can consist of a series of bidirectional step functions at each flux reversal location. The step amplitude is twice the pulse amplitude, as shown in Figure 16b. These are identical signals. Either technique may be used to analyze the input signal as it passes through the equalization channel. In the following pages, we assume an input consisting of binit pulses, as shown in Figure 16a.

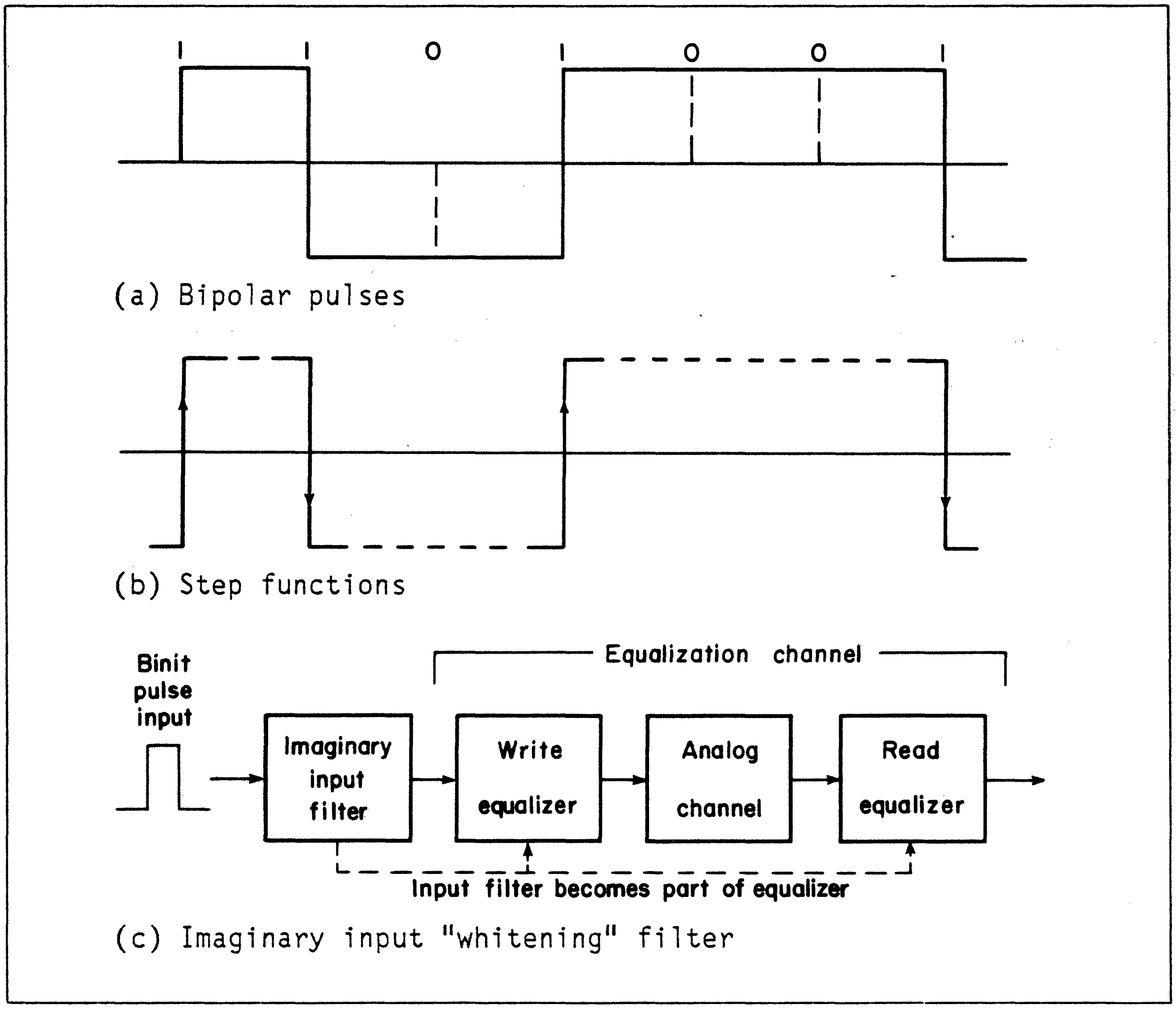
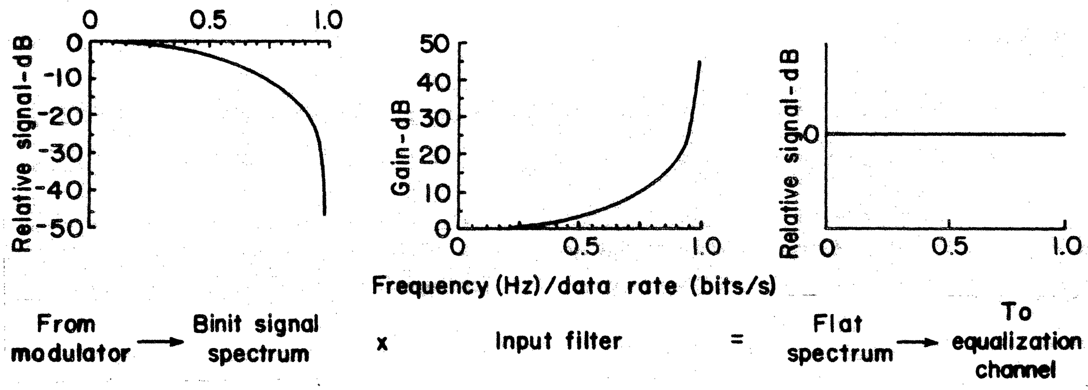


Figure 16 (Part 1 of 2). Equalization-Channel Input



(d) Input-filter frequency response

Figure 16 (Part 2 of 2). Equalization-Channel Input

A flat input spectrum, which simulates the input from a unit impulse or delta function, is useful for defining the equalizer characteristics. It is obtained by assuming an input filter that converts the (sin x)/x frequency spectrum of each input binit pulse to a flat spectrum. This input "whitening" filter has an x/(sin x) response that cannot be physically achieved because it requires infinite gain. Its only purpose is to simplify comparison of various equalizers, because it is constant for a given recording density. This hypothetical equalization-channel impulse response or transfer function is used to compare equalizers. Ultimately, the input filter response must be included in the equalizer filters. The binit-pulse response with the input filter included in the equalizer filter is called the equalization-channel pulse response. The equalization channel and input filter are shown in Figures 16c and 16d.

There are four principal classes of equalizers that can be used in a computer-tape modulation channel—pulse slimming, waveform restoration, derivative, and partial response. There are many possible variations within each class. Equalizer selection depends upon the amount of intersymbol interference to be compensated, the modulation code, the data-recovery technique used, the SNR, and the noise-spectrum shape.

Pulse slimming is usually a time-domain approach to equalization. The other equalization techniques typically define the equalizer in the frequency domain to achieve the desired output in the time domain. Whichever approach the designer takes, the equalizer transfer function determines the end result.

In the computer-simulated equalizer examples, the first two curves (a and b) represent the impulse response and pulse response of the equalization channel. The equalizer response required for the analog channel shown in Figure 3 on page 5, including the input filter, is shown at Figure 3f in the computer simulations. The output of the analog channel in Figure 3d is distorted by intersymbol interference because the recording density is high. The equalizer reshapes this signal in Figure 3e to allow reliable clock and data recovery.

In the laboratory, a convenient tool for evaluating equalizer performance is the eye pattern. A pseudo-random binary sequence is applied to the channel input, as the output is connected to an oscilloscope with the sweep speed adjusted to cover a few bits. The data clock triggers the sweep on the oscilloscope. The resulting eye pattern indicates the channel operating margin—the horizontal opening is the time margin and the vertical opening is the amplitude margin, as discussed by Feher [ref.24] and Lucky et al. [ref.25]. The computer-simulated eye patterns in Figure 3g, which are noise-free and dropout-free, demonstrate the distortion differences between various equalization alternatives.

PULSE SLIMMING

Many types of pulse-slimming filters have been used to reduce adjacent pulse interference, as discussed by Barbosa [ref.26], Jacoby [ref.27], Schneider [ref.28], and Sierra [ref.29]. Pulse slimming is achieved through the addition and subtraction of signal-derived compensation pulses. A good example of pulse slimming, as applied to a magnetic tape device, uses first and second derivatives of the signal to perform the pulse-slimming operation. The equalizer transfer function is

$$G(s) = (1+As)(1-Bs) = 1 + (A-B)s - ABs^2 \tag{10}$$

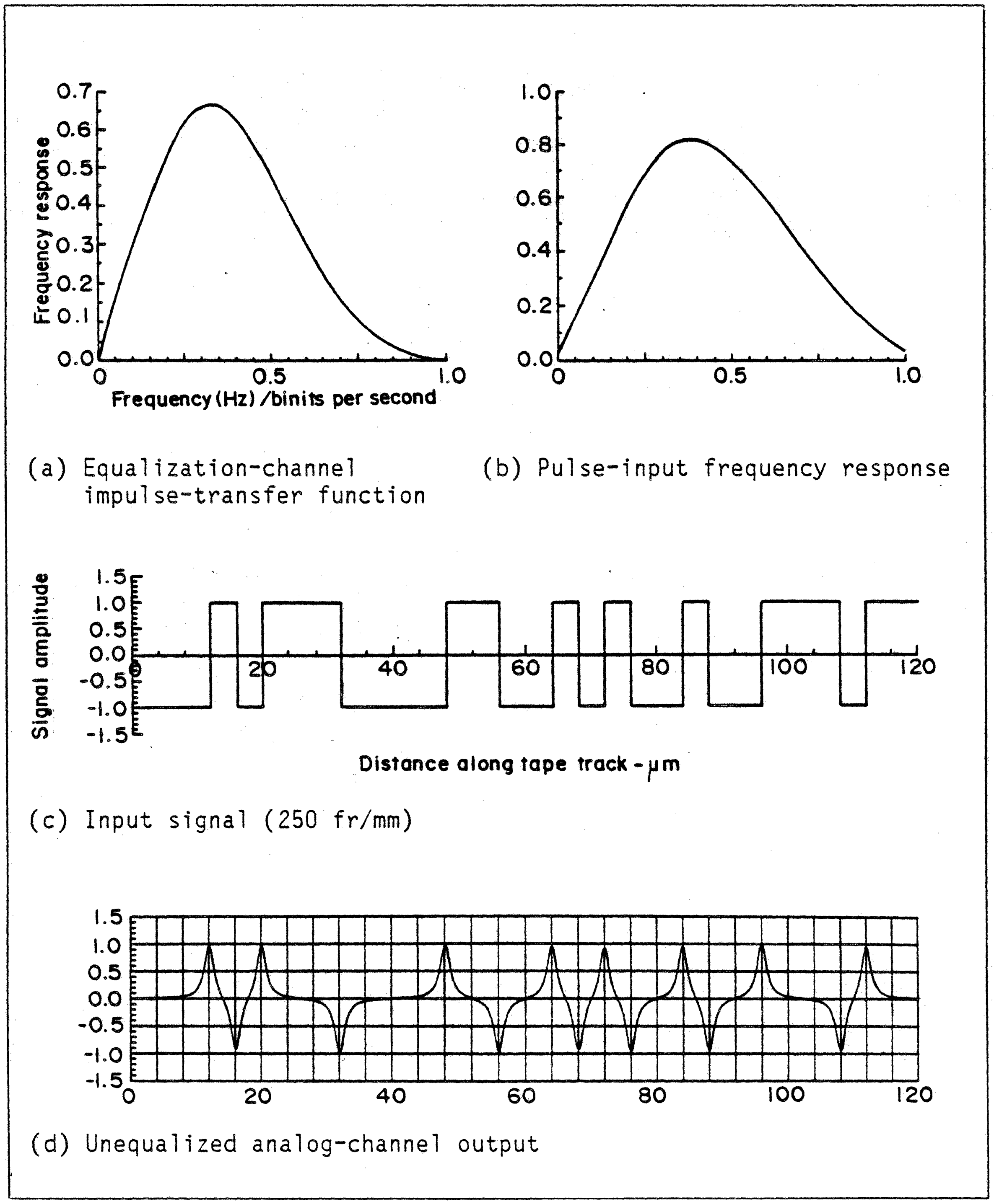
where A and B are the portions of the first derivative required to slim the trailing and leading edges.

Asymmetry can be improved by the proper selection of A and B. In general, odd derivatives are useful in correcting read-pulse asymmetry or phase errors, and even derivatives perform symmetrical pulse slimming.

If asymmetry compensation is not needed, then A = B and the transfer function is simplified to $1 - ks^2$. If a dc-free modulation code is used, an integrator can be used to improve the signal baseline.

A typical computer-tape analog channel produces significant amounts of intersymbol interference at high recording densities, as shown in Figure 11 on page 21. (This channel-transfer function magnitude is shown in Figure 3 on page 5.) The low-density signal (50 fr/mm) shown in Figure 11b has negligible distortion, and needs no equalization. The 250-fr/mm signal shown in Figure 11c shows some intersymbol interference that can be adequately reduced by pulse slimming. However, the high-density signal (1250 fr/mm) shown in Figure 11d needs more equalization than is available from conventional pulse slimming to reduce distortion to an acceptable level. Each of these pulses interferes with several nearby pulses.

Figure 17 on page 36 shows the use of pulse-slimming equalization on the 250-fr/mm signals from the analog channel shown in Figure 3 on page 5.



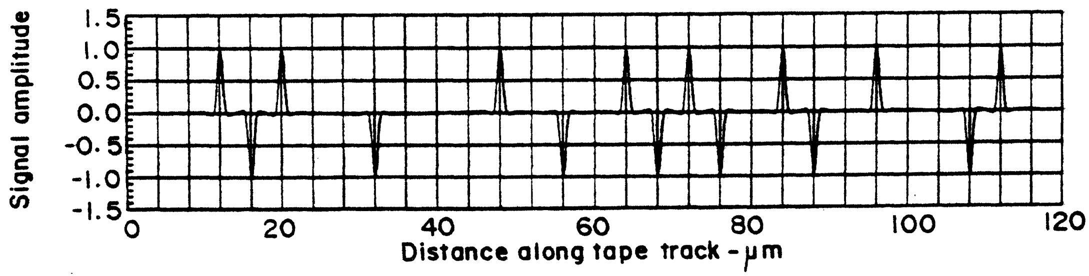
(a) Equalization-channel impulse-transfer function

(b) Pulse-input frequency response

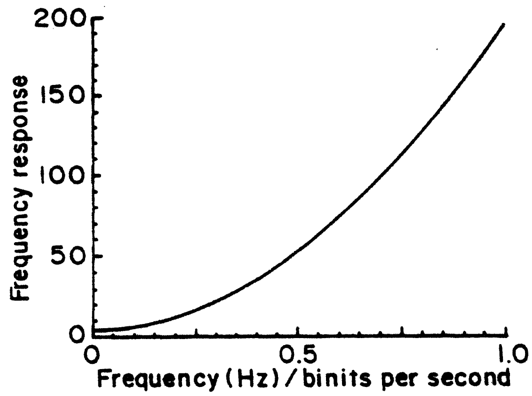
(c) Input signal (250 fr/mm)

(d) Unequalized analog-channel output

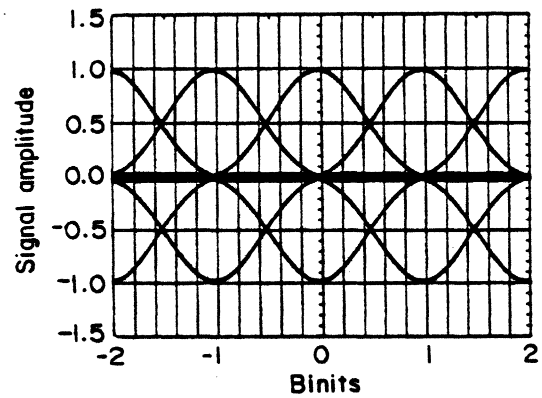
Figure 17 (Part 1 of 4). Pulse-Slimming Equalizer



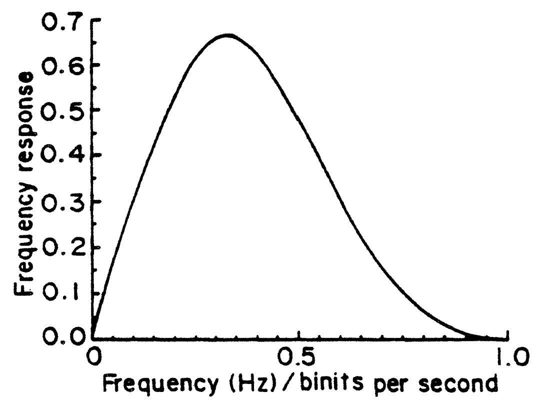
(e) Equalized output signal



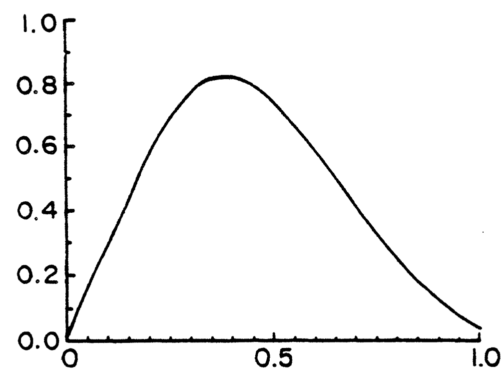
(f) Equalizer transfer function (30% of normalized second derivative)



(g) Eye pattern

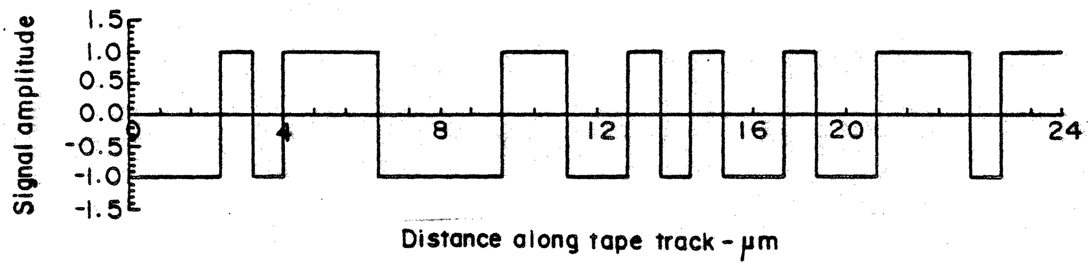


(h) Equalization-channel impulse-transfer function

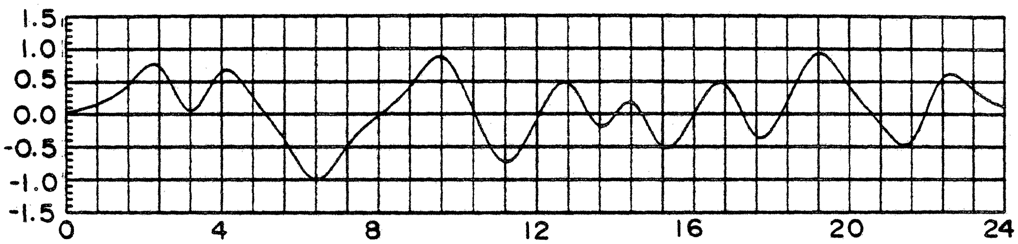


(i) Pulse-input frequency response

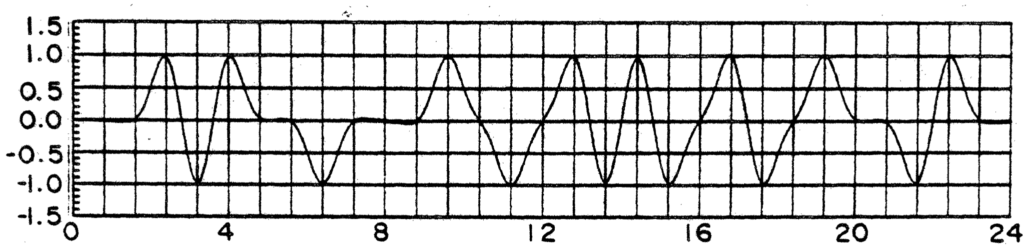
Figure 17 (Part 2 of 4). Pulse-Slimming Equalizer



(j) Input signal (1250 fr/mm)

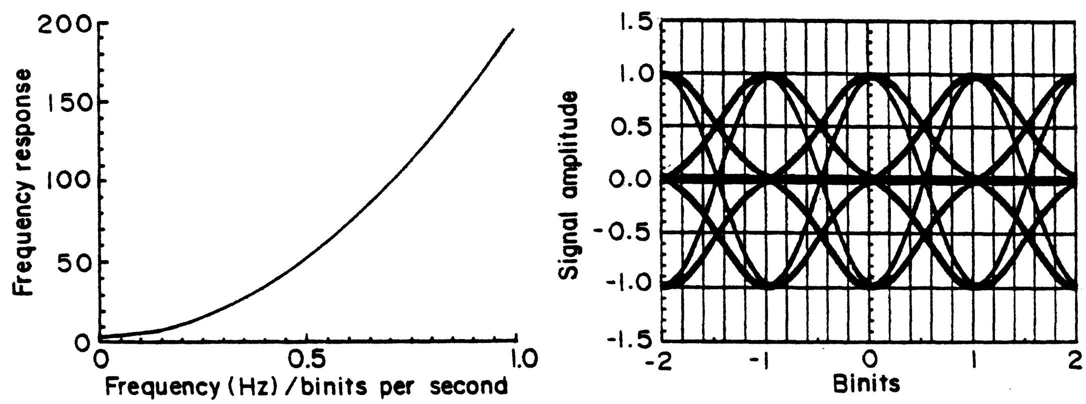


(k) Unequalized analog-channel output



(l) Equalized output signal

Figure 17 (Part 3 of 4). Pulse-Slimming Equalizer



(m) Equalizer transfer function (30% of normalized second derivative)

(n) Eye pattern

Figure 17 (Part 4 of 4). Pulse-Slimming Equalizer

Frequency- and time-domain representations of this equalizer are presented. Because this hypothetical channel has linear phase characteristics, no asymmetry correction is needed. The equalizer merely subtracts 30% of the second derivative normalized peak amplitude ($1 - 0.3s^2$). The bandwidth is limited to 2.5 MHz, although greater bandwidth has little effect on the noise-free signal. The equalized signals shown in Figure 17 can be peak detected with no significant intersymbol interference. This type of equalizer is marginal at the 1250 fr/mm density. A more sophisticated pulse slimmer, with higher-order derivatives, might eliminate distortion at the higher density. Because no organized method of defining the derivative requirements is known, the waveform-restoration or derivative-type equalizer may be preferable for high recording densities.

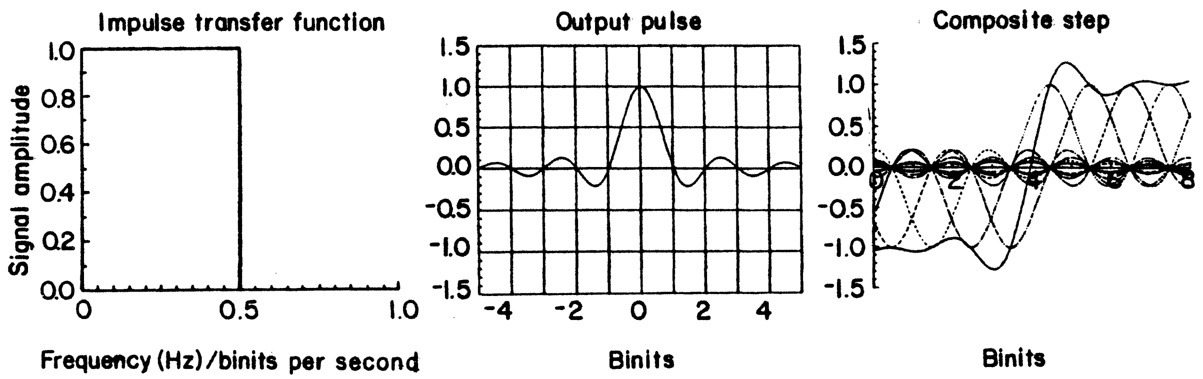
WAVEFORM RESTORATION

Waveform-restoration equalization produces a signal at the data detector that resembles the input waveform—a binary sequence. The corners of the rectangular pulses are rounded because the signal harmonics are attenuated in the channel and some output-signal amplitude variation can occur.

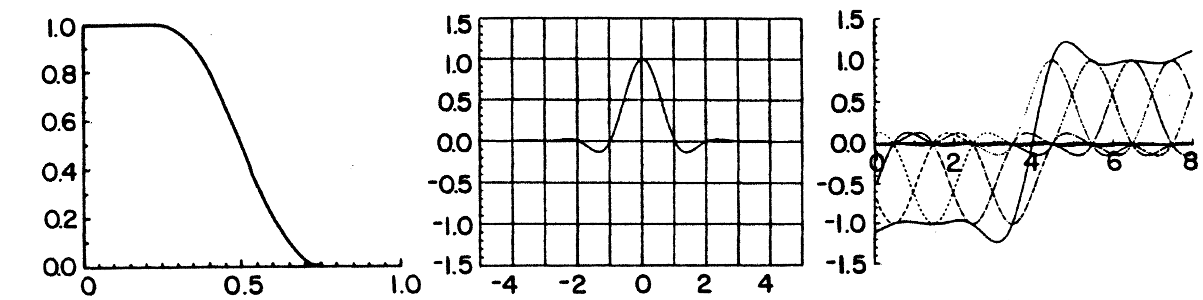
When the pulse input signal shown in Figure 16a on page 33 is applied to the equalization channel, the data detector must derive a clock from the output signal and determine the pulse polarity for each clock cell or binit. To achieve this result, an ideal waveform-restoration equalizer produces an output signal with mid-binit and binit-boundary values, which are similar to those of the input signal. The output signal zero-crossings occur at binit boundaries to regenerate a clock accurately, and a sample at the middle of each clock cell identifies the binit polarity. This actually provides redundant information because, if the zero-crossing time and direction are known, both clock and data can be extracted from the signal zero crossings.

If the equalization-channel transfer function has flat amplitude and linear phase response, as with a high-quality audio or instrumentation tape recorder, rectangular input pulses should pass through the channel with only loss of harmonics. The resultant signal will be similar to the input except for the rounded corners due to harmonic losses. If the upper-band edge frequency response (both magnitude and phase) is properly designed and if the correct data rate is used, intersymbol interference is virtually eliminated. This type of channel equalization has been used to restore the input waveform for digital recording, as discussed by Jacoby [ref.27], and Kiwimagi et al. [ref.30].

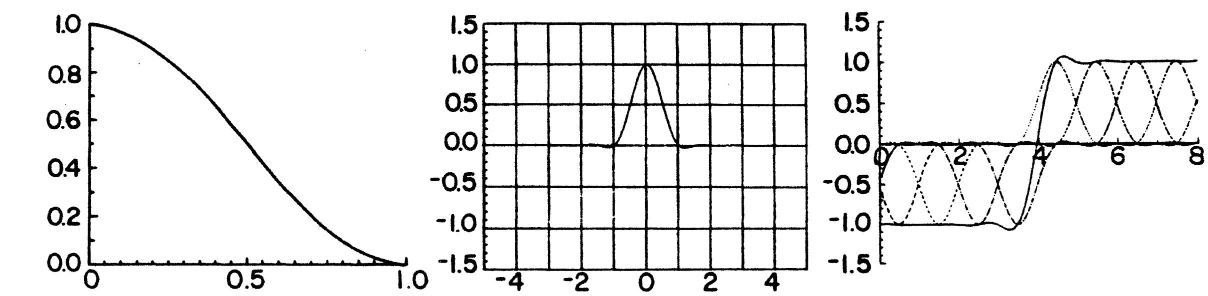
The minimum bandwidth waveform-restored output signal from a modulation code with $d = 0$ consists of a series of pulses with unit amplitude at the center of each binit cell, and zero amplitude at all multiple binit-cell periods $\pm nT_m$ away from the peak. Therefore, each successive input binit pulse creates an output pulse with unit amplitude when all other output pulses have zero response. This minimum bandwidth channel output response has the characteristics of a sinc pulse, as discussed by Bracewell [ref.31] and Ziemer and Tranter [ref.32], or the characteristics of the sampling function $(\sin x)/x$. Figure 18a on page 41 shows how the sinc pulses combine to represent an input-signal transition. In the absence of noise, the amplitude of the equalizer output is unity at the center of each binit or clock cell. The minimum bandwidth equalizer produces an output signal with unity amplitude at the center of each binit, but zero crossings do not occur at binit boundaries; therefore, accurate clock recovery is difficult.



(a) Minimum bandwidth equalizer



(b) Raised cosine $\alpha = 0.5$



(c) Raised cosine $\alpha = 1.0$

Figure 18. Waveform-Restoration Equalizers

The equalizer that produces this minimum bandwidth output signal is an ideal low-pass filter with unity response to the minimum cutoff frequency $f_m = 1/2T_m = f_c/2$, and no response at higher frequencies, as shown in Figure 18a. This low-pass filter is not physically realizable; the upper band edge must roll off more gradually if the filter is to be achievable. The speed sensitivity of this type of filter would prohibit its use even if the filter could be built. The Nyquist theorem on vestigial symmetry shows how the sharp cutoff minimum bandwidth filter can be modified and still retain output-pulse zero crossings at all mid-binit cell times, as discussed by Bennett and Davey [ref.15]. To achieve this desirable result, the high-frequency roll-off of the equalized channel must have symmetry about the half-amplitude point at the minimum bandwidth filter cutoff frequency, as shown in Figures 18b and 18c on page 41.

A raised cosine roll-off is frequently used, which leads to the term raised cosine equalizer. This type of transfer function is approximately realizable, and the response is an improvement over the minimum bandwidth filter. The output pulses still retain the zero values at times $\pm nT_m$, but the side-lobe damped oscillation amplitude is reduced, as shown in Figures 18b and 18c. The output zero crossings become more consistent, and linear phase characteristics are achieved more easily with a gradual roll-off transfer function. The improvement is obtained at the expense of increased bandwidth. The ratio of bandwidth extension to the minimum bandwidth f_m is sometimes referred to as the α of the raised cosine channel. Thus, in the case of a modulation code with $d = 0$, $\alpha = 0$ is the minimum bandwidth (but unrealizable) rectangular transfer function, and $\alpha = 1$ uses twice the bandwidth or B_n, as discussed by Bennett and Davey [ref.15], Feher [ref.24], Lucky et al. [ref.25], and Roden [ref.33].

The impulse transfer function of the raised cosine equalization channel (including the analog channel plus equalizer, but excluding the input filter) is:

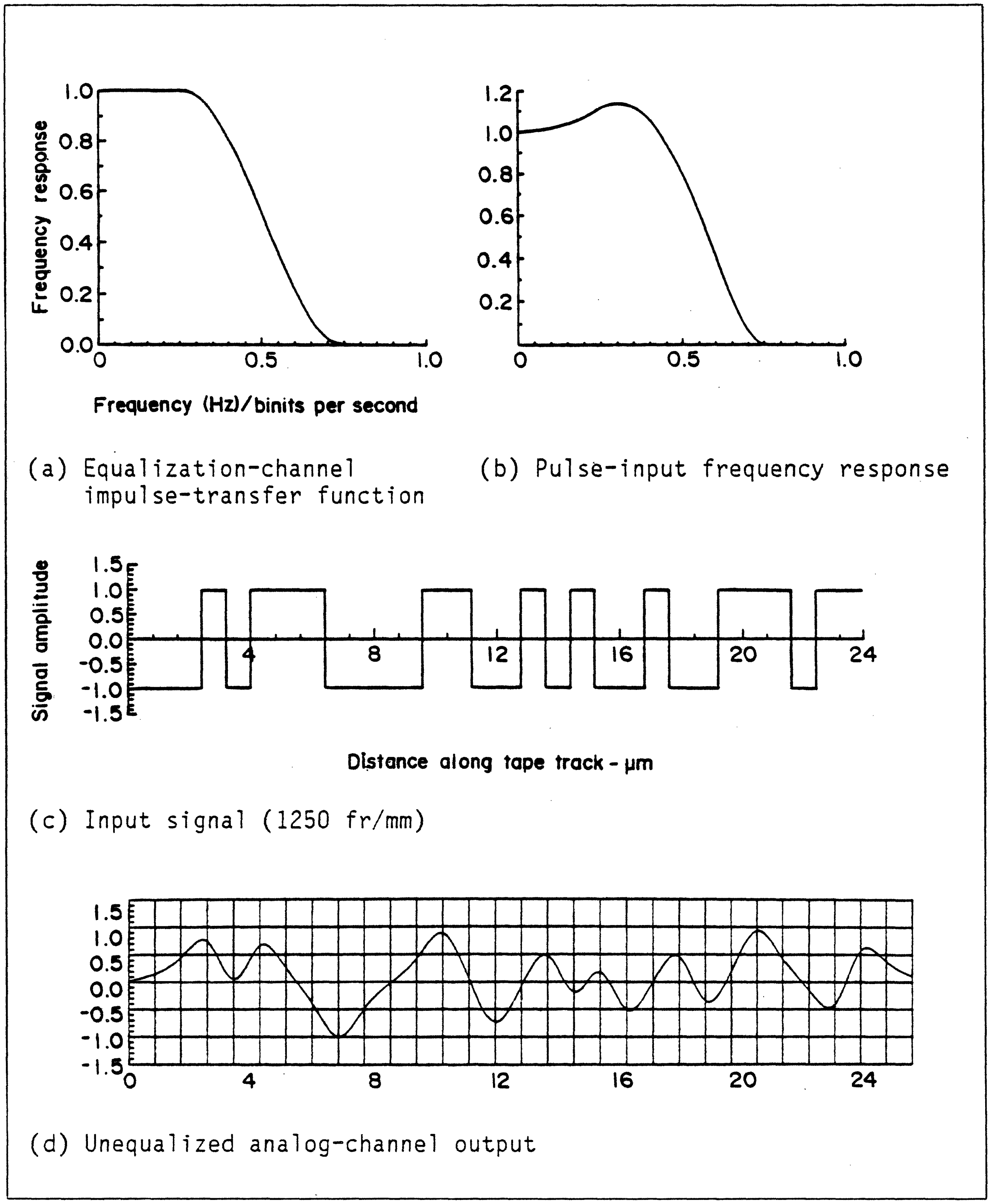
$$
\begin{aligned}
H(f) &= 1 && 0 < f < (1-\alpha)f_m \\
&= 1/2\{1+\cos(\pi/2\alpha f_m)\ [f-(1-\alpha)f_m]\} && (1-\alpha)f_m \le f \le (1\alpha)f_m \\
&= 0 && f > (1+\alpha)f_m
\end{aligned}
\qquad (11)
$$

$$\phi(f) = kf$$

where $\phi(f)$ is the phase and k is a constant.

This family will be called α equalizers in the following pages.

Figure 19 on page 43 shows the operation of the $\alpha = 0.5$ equalizer with the analog channel from Figure 3 on page 5 operated at 1250 fr/mm.



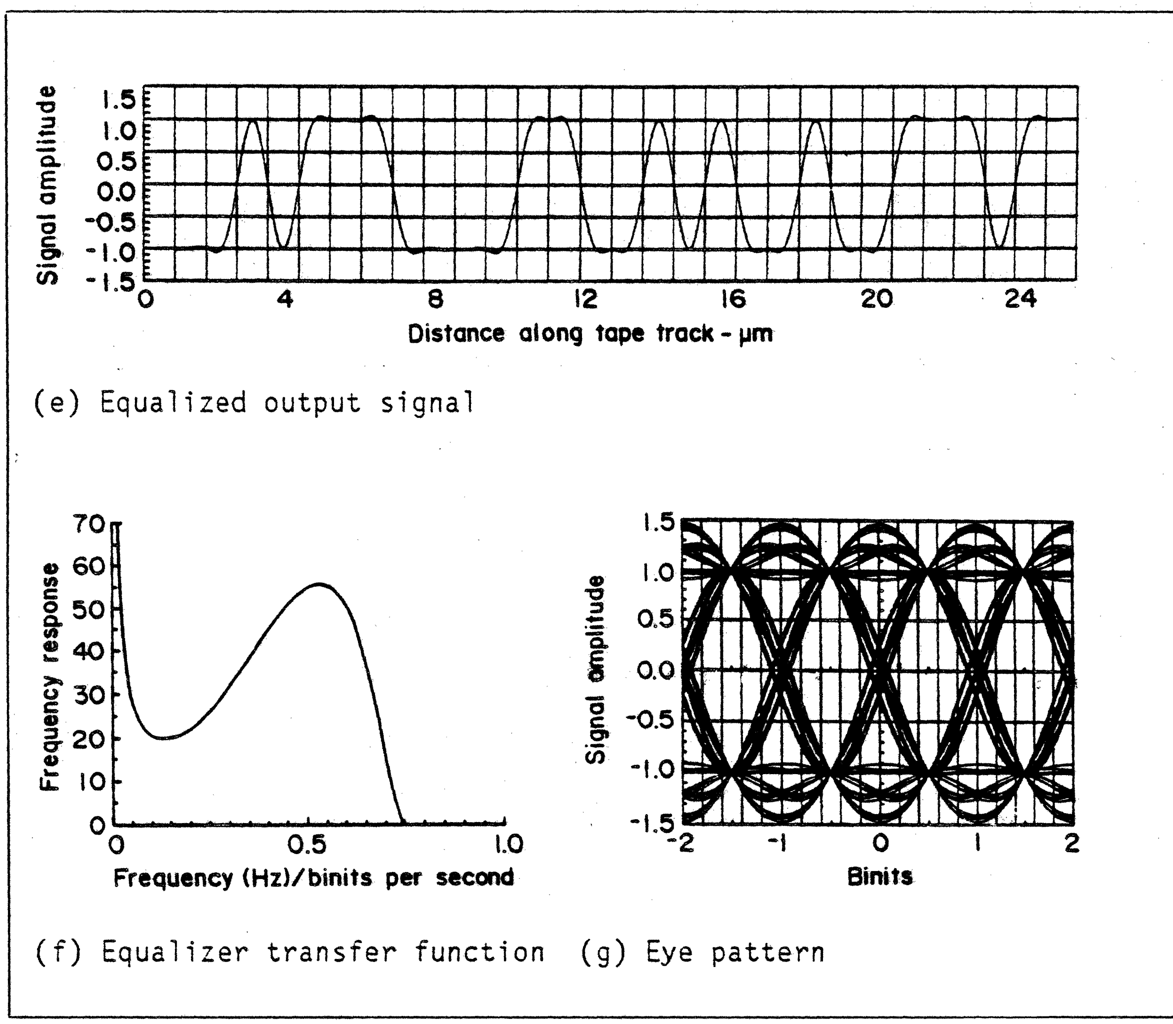
(a) Equalization-channel impulse-transfer function

(b) Pulse-input frequency response

(c) Input signal (1250 fr/mm)

(d) Unequalized analog-channel output

Figure 19 (Part 1 of 2). Raised Cosine α = 0.5 Equalizer



(e) Equalized output signal

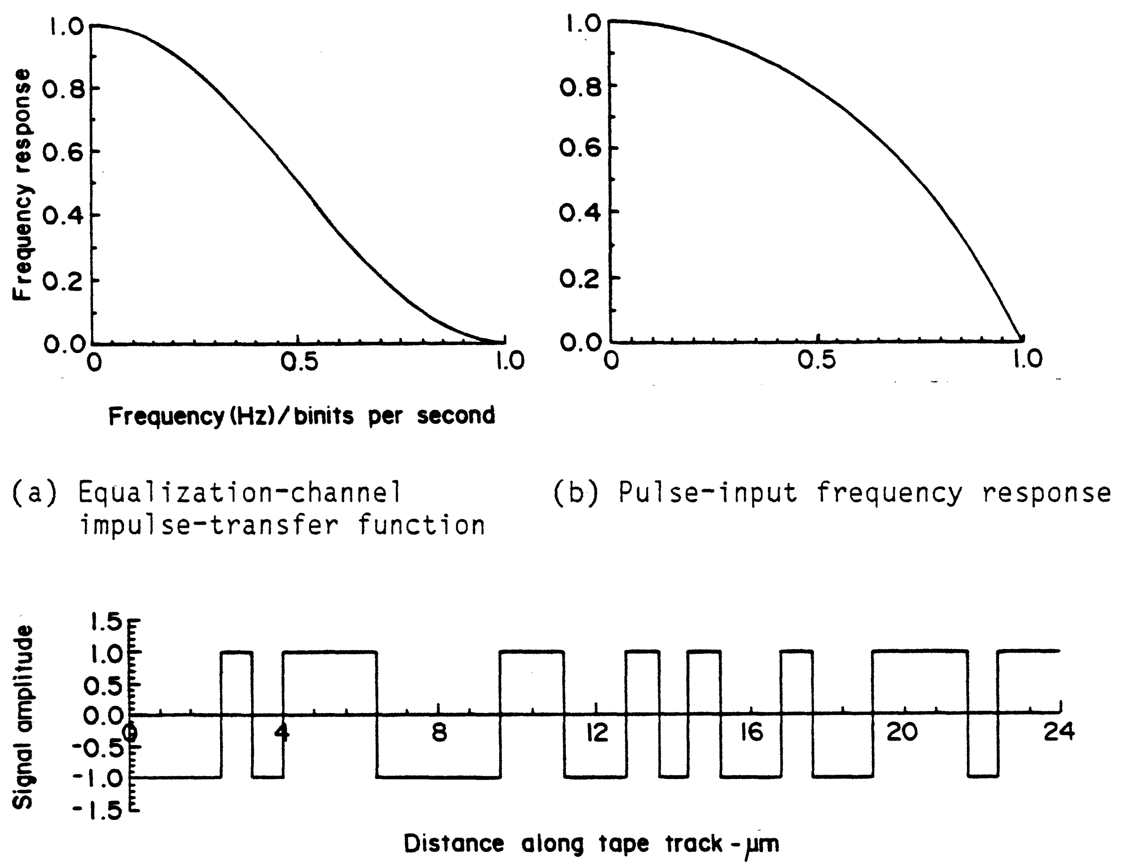
(f) Equalizer transfer function (g) Eye pattern

Figure 19 (Part 2 of 2). Raised Cosine $\alpha = 0.5$ Equalizer

The phase response is not shown, but the overall phase must be relatively linear for good results. Usually little phase compensation is required. Descriptions of phase distortion are available in the literature, as discussed by Hedeman [ref.34], Preis [ref.35], and Silver [ref.36]. The equalized signal passes through unity at exactly the center of each binit, as shown in Figures 19c and 19e. Although the signal zero crossings are near the binit boundaries, there is some intersymbol interference everywhere, except at the center of each binit. The amount of ISI is shown in Figure 18b on page 41.

The $\alpha = 1$ channel has the interesting property of having nulls at half-binit intervals as well as full-binit intervals, as shown in Figure 18c. This signal has no intersymbol interference at mid-binit or binit-boundary times, which are sample and signal zero-crossing times and thus allow accurate clock and data recovery. For this full-bandwidth equalizer, the roll-off starts at zero frequency and extends to f_c.

Raised cosine equalizers are discussed in many publications, and are in common use for many applications. They are capable of correcting extensive amounts of linear intersymbol interference if there is adequate SNR. Figure 20 shows the capability of an $\alpha = 1$ channel that operates at 1250 fr/mm. A large high-frequency boost is required to compensate for the separation, thickness, and head-gap losses. A lower fly height or shorter read gap would reduce this high-frequency requirement. A low-frequency boost may also be required to compensate for the differentiating action of the read head.

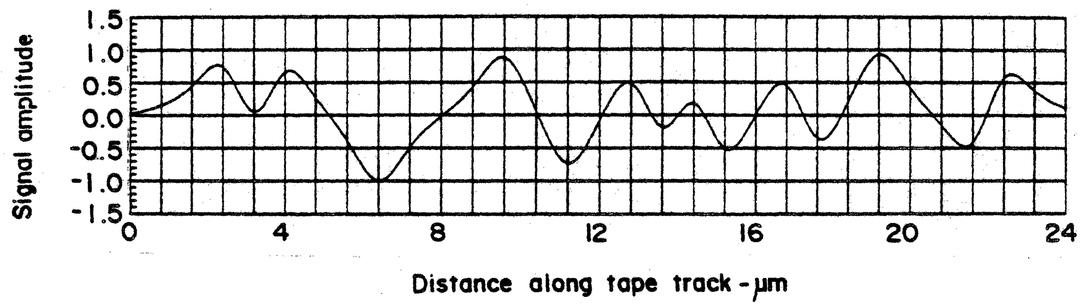


(a) Equalization-channel impulse-transfer function

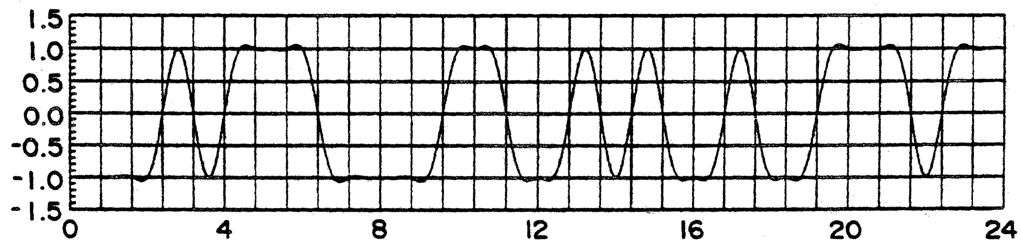
(b) Pulse-input frequency response

(c) Input signal (1250 fr/mm)

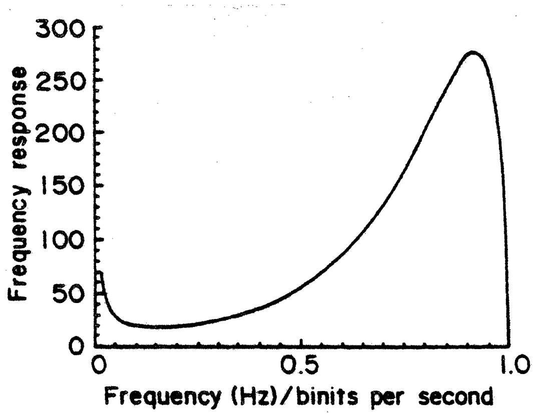
Figure 20 (Part 1 of 2). Raised Cosine $\alpha = 1$ or $\beta = 2$ Equalizer

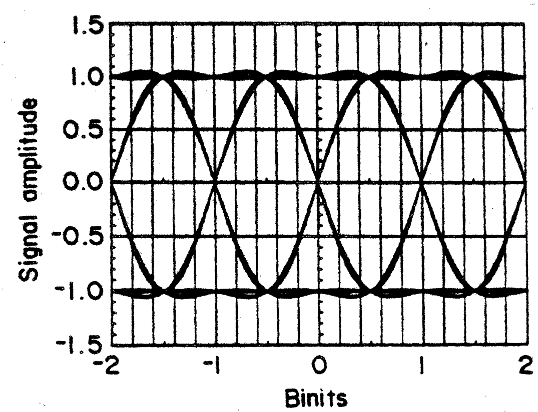


(d) Unequalized analog-channel output



(e) Equalized output signal



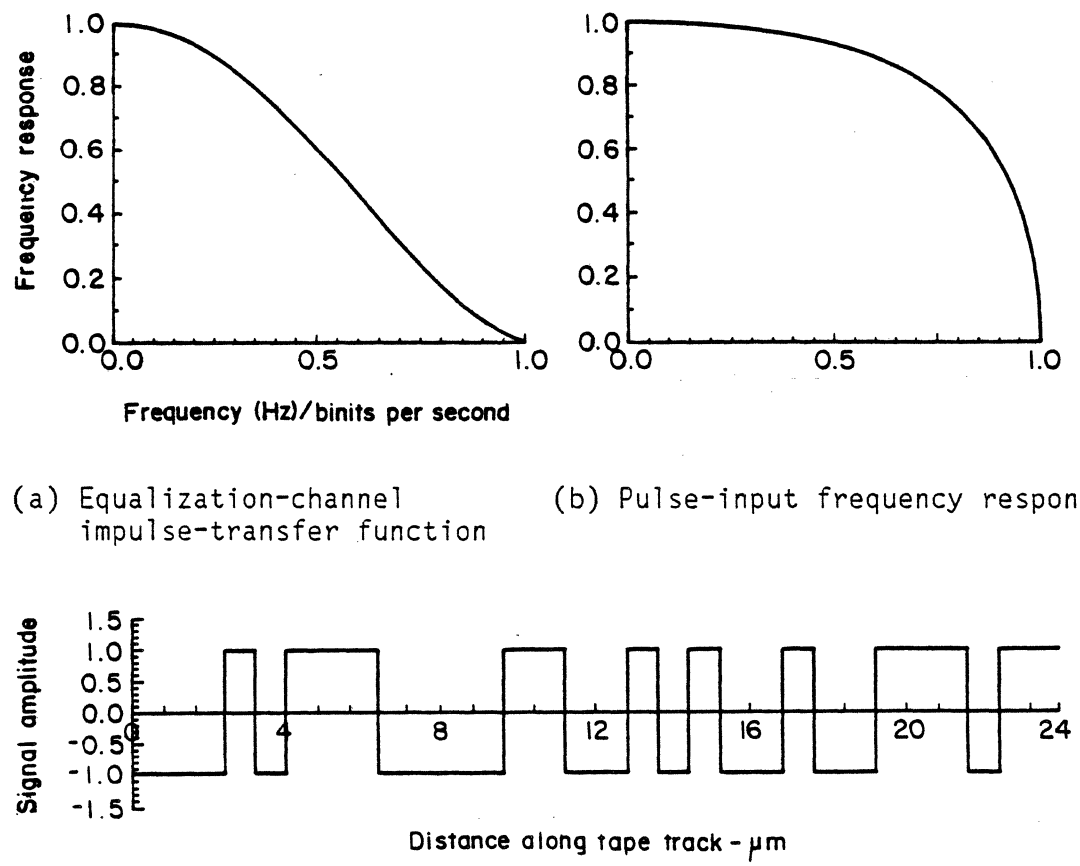


(f) Equalizer transfer function (g) Eye pattern

Figure 20 (Part 2 of 2). Raised Cosine $\alpha = 1$ or $\beta = 2$ Equalizer

A bandwidth equal to twice the minimum bandwidth (that is, $1/T_m$ rather than $1/2T_m$) is mandatory for complete elimination of linear intersymbol interference with a physically realizable channel passing a modulation code with d = 0. Unfortunately, this wide bandwidth reduces the SNR. This does not imply that $\alpha < 1$ cannot be used. The goal is to achieve the optimum compromise between distortion and noise. It may be desirable to narrow the bandwidth by using an $\alpha < 1$ transfer function to improve noise at the expense of added distortion in the form of clock jitter.

The raised cosine family of equalizers is not the only way to obtain waveform restoration. Another waveform-restoration equalizer is the cosine$^{\beta}$ response shown in Figure 21 and Figure 22 on page 49.

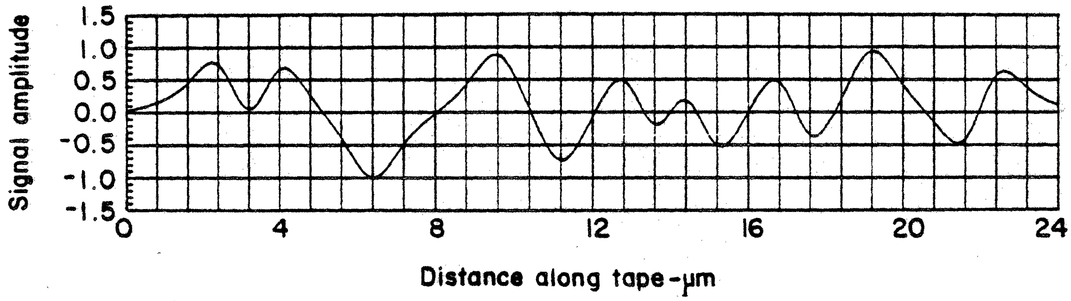


(a) Equalization-channel impulse-transfer function

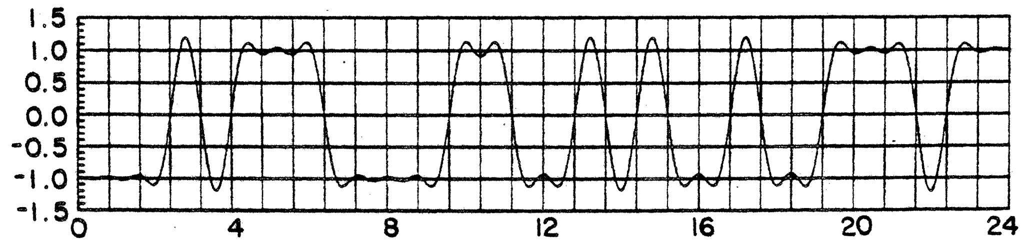
(b) Pulse-input frequency response

(c) Input signal (1250 fr/mm)

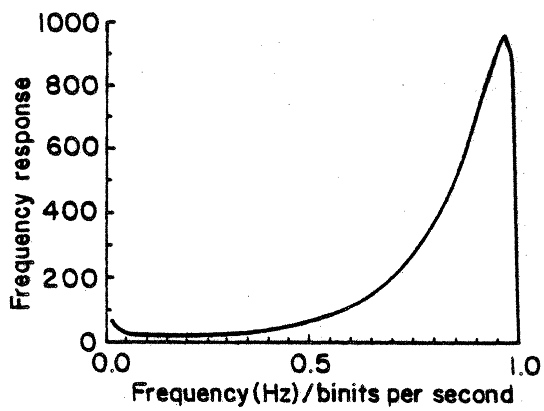
Figure 21 (Part 1 of 2). Cosine$^{\beta}$ β = 1.5 Equalizer

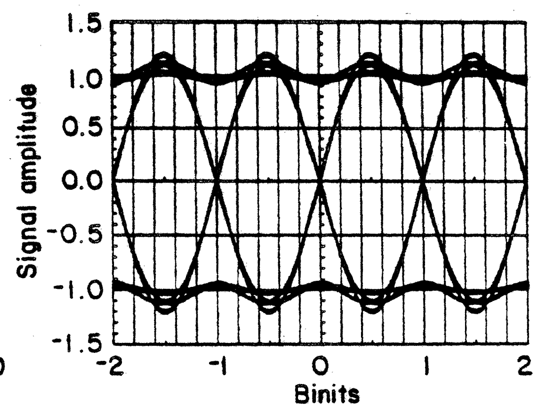


(d) Unequalized analog-channel output



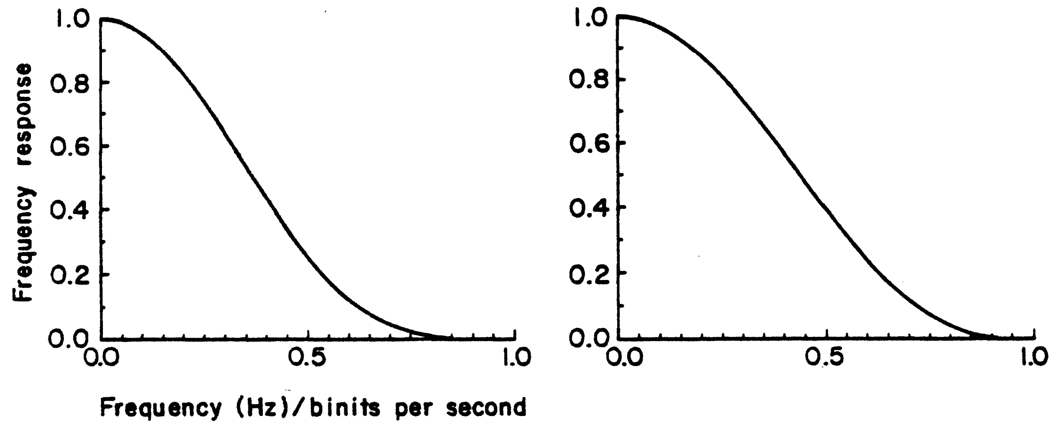
(e) Equalized output signal





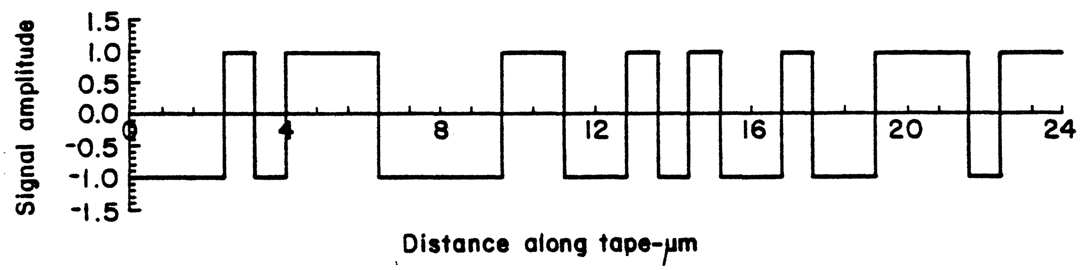
(f) Equalizer transfer function (g) Eye pattern

Figure 21 (Part 2 of 2). Cosine$^{\beta}$ β = 1.5 Equalizer

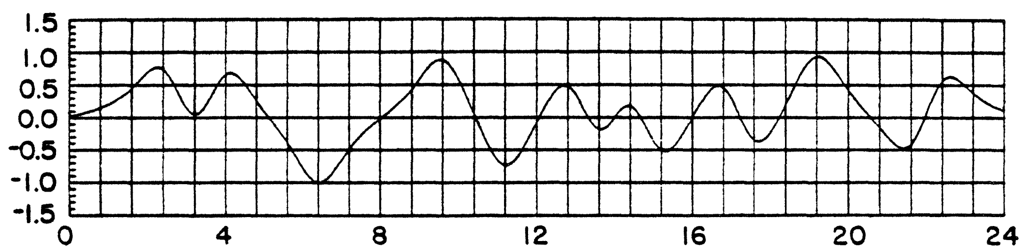


(a) Equalization-channel impulse-transfer function

(b) Pulse-input frequency response

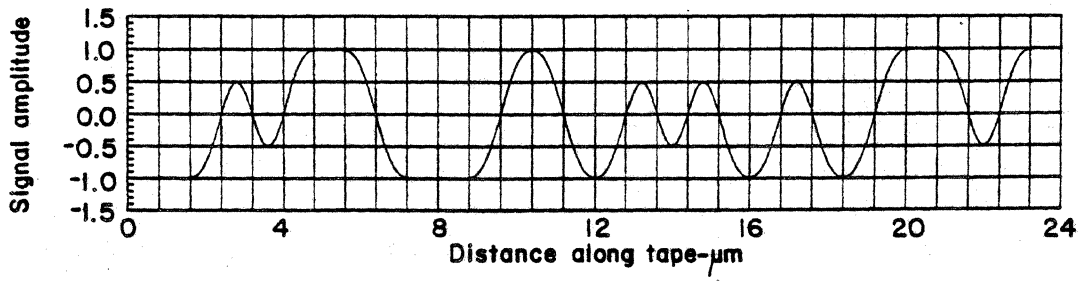


(c) Input signal (1250 fr/mm)

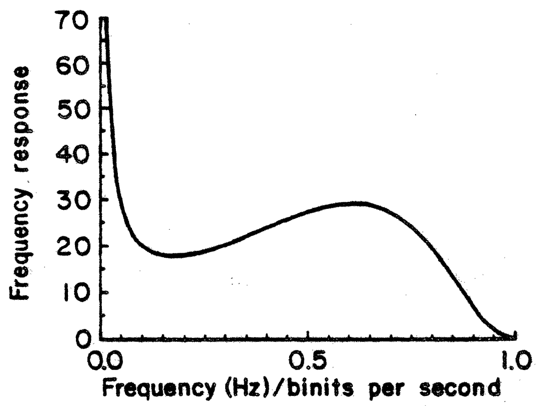


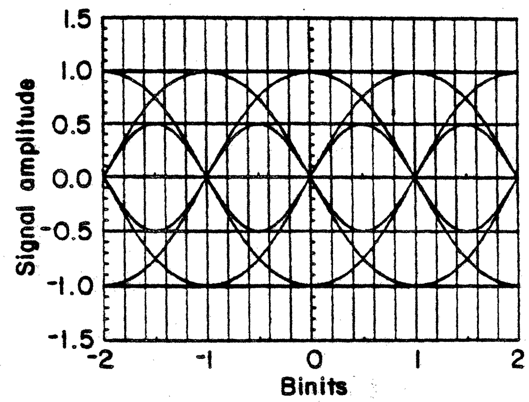
(d) Unequalized analog-channel output

Figure 22 (Part 1 of 2). Cosine$^{\beta}$ β = 4 Equalizer



(e) Equalized output signal





(f) Equalizer transfer function (g) Eye pattern

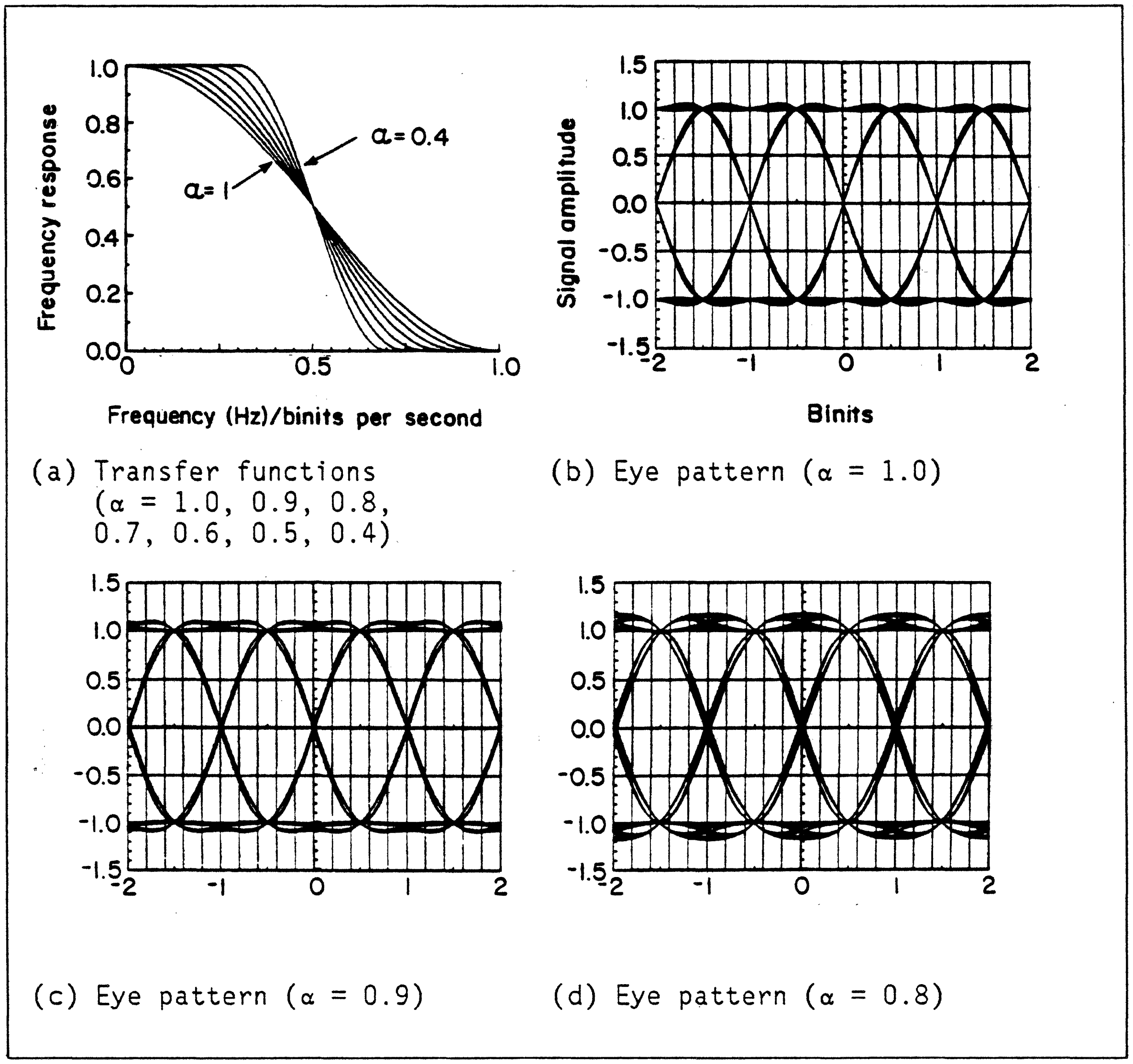
Figure 22 (Part 2 of 2). Cosine$^{\beta}$ β = 4 Equalizer

The impulse transfer function of a full-bandwidth β channel is:

$$H(f) = \begin{cases} \cos^{\beta} \pi f/(2f_c) & 0 \le f \le f_c \\ 0 & f > f_c \end{cases} \qquad (12)$$

Like the α equalizer family, there are numerous β equalizers. All of these full-bandwidth equalizers have a cutoff frequency of f_c and thus reduce clock jitter because there is little interference at binit boundaries. Bennett and Davey [ref.15] discuss techniques for optimizing these equalizing filters to achieve the minimum probability of error in various types of noise.

The selection of α permits the bandwidth to be narrowed, thus reducing noise at the expense of clock jitter or horizontal eye opening, as shown in Figure 23.



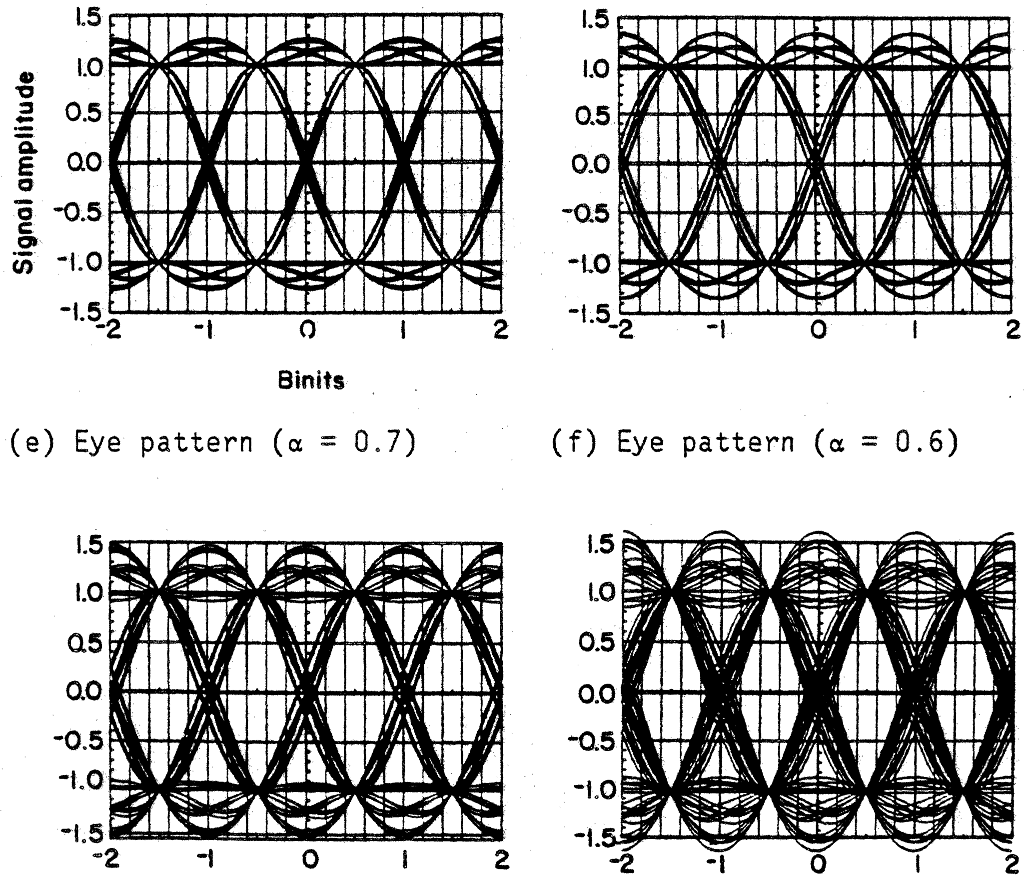
(a) Transfer functions (α = 1.0, 0.9, 0.8, 0.7, 0.6, 0.5, 0.4)

(b) Eye pattern (α = 1.0)

(c) Eye pattern (α = 0.9)

(d) Eye pattern (α = 0.8)

Figure 23 (Part 1 of 2). Comparison of Raised Cosine Equalizer



(e) Eye pattern (α = 0.7)

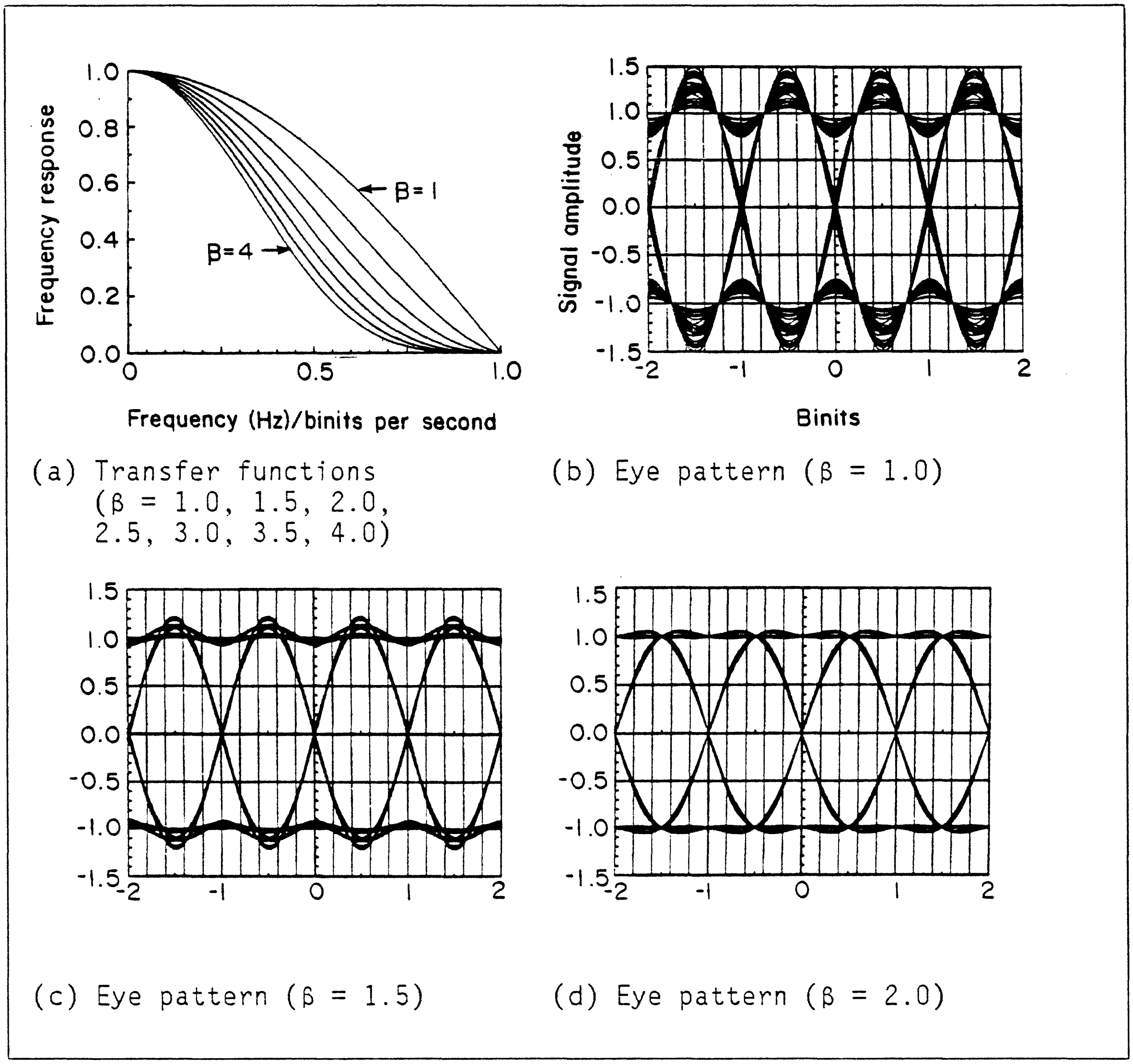
(f) Eye pattern (α = 0.6)

(g) Eye pattern (α = 0.5)

(h) Eye pattern (α = 0.4)

Figure 23 (Part 2 of 2). Comparison of Raised Cosine Equalizer

The selection of β permits SNR improvement by reducing the high-frequency boost without reducing the bandwidth. The penalty for this choice is a reduction in vertical eye opening, or an effective amplitude reduction, as shown in Figure 24.



(a) Transfer functions (β = 1.0, 1.5, 2.0, 2.5, 3.0, 3.5, 4.0)

(b) Eye pattern (β = 1.0)

(c) Eye pattern (β = 1.5)

(d) Eye pattern (β = 2.0)

Figure 24 (Part 1 of 2). Comparison of Cosine$^{\beta}$ Equalizer

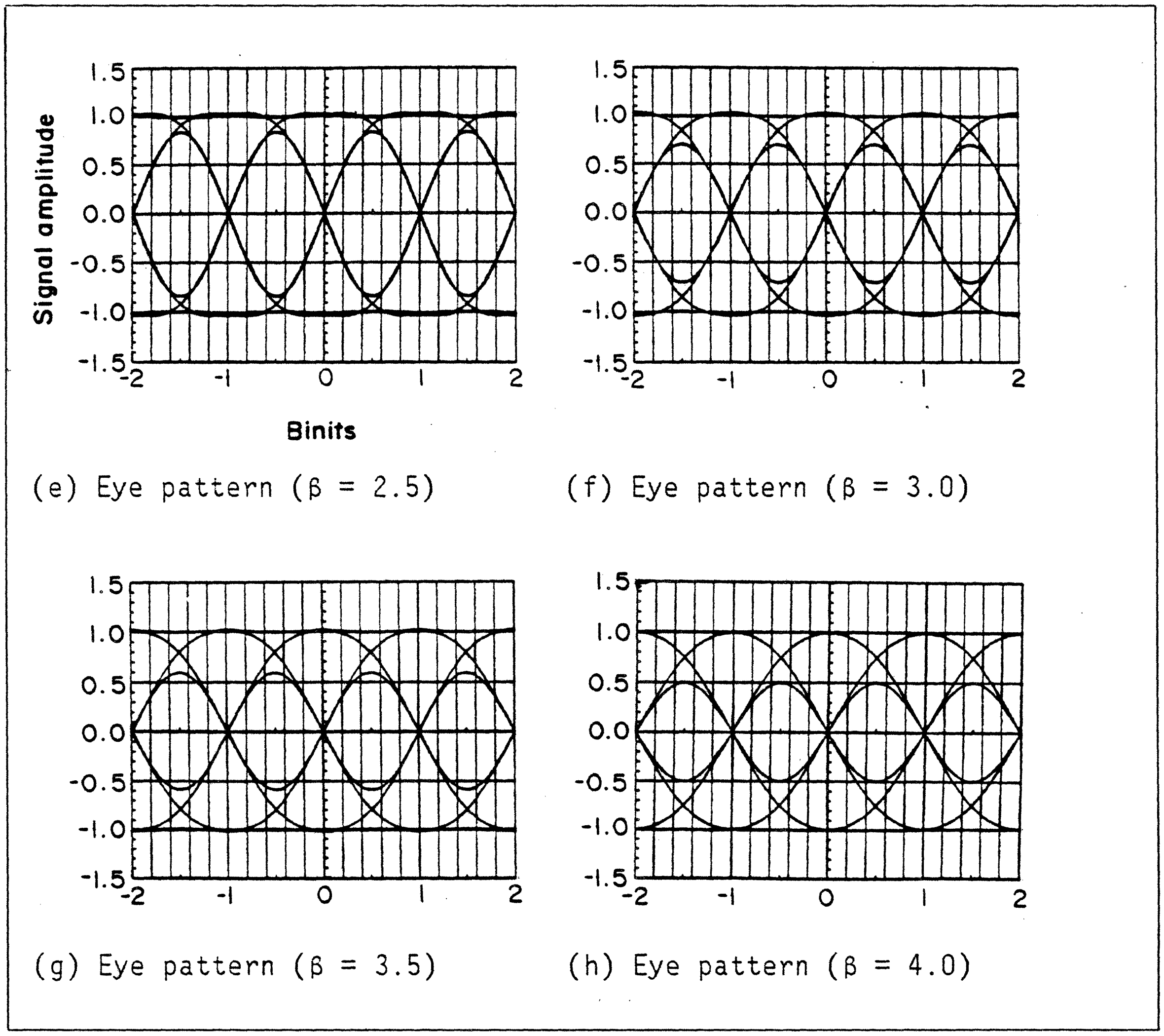
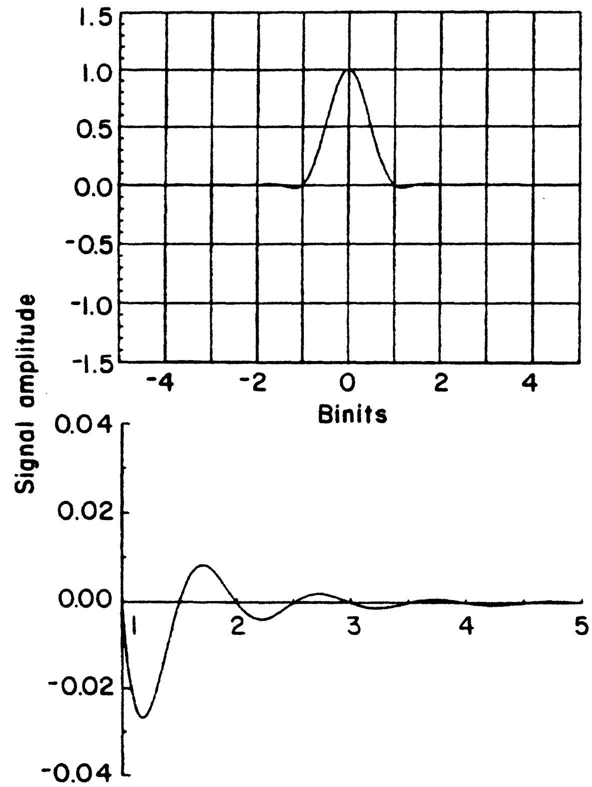


Figure 24 (Part 2 of 2). Comparison of Cosine$^{\beta}$ Equalizer

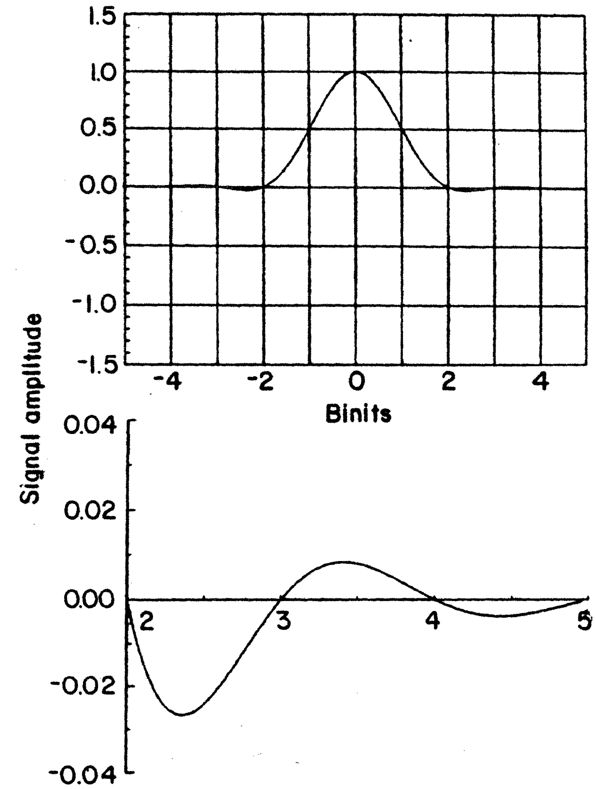
Channels $\alpha = 1$ and $\beta = 2$ are identical because $1 + \cos A = 2\cos^2 A/2$. The eye is wide open for this equalizer, as shown in Figure 20g on page 46.

The equalized channel bandwidth is straightforward for modulation codes with d = 0, but codes with d > 0 sometimes cause confusion. The required bandwidth is not determined by the minimum recorded pulse width T_r (as might be expected), but rather by the binit width T_m. The data-recovery circuits must distinguish between pulses that differ by only one binit width, and time resolution is a function of bandwidth. The 0,k codes require a nominal bandwidth $B_n = 1/T_m = f_c$ that eliminates interference at the center and edge of each binit. As shown in Figure 25a on page 55, the pulse-response tails are zero at $nT_m/2$ and there is no pulse-to-pulse interference at these times.



(a) Single pulse response for d = 0 (with a full bandwidth and zero crossings at the center and edges of the binit)

Figure 25 (Part 1 of 2). Half-Bandwidth Operation (α = 1 or β = 2)

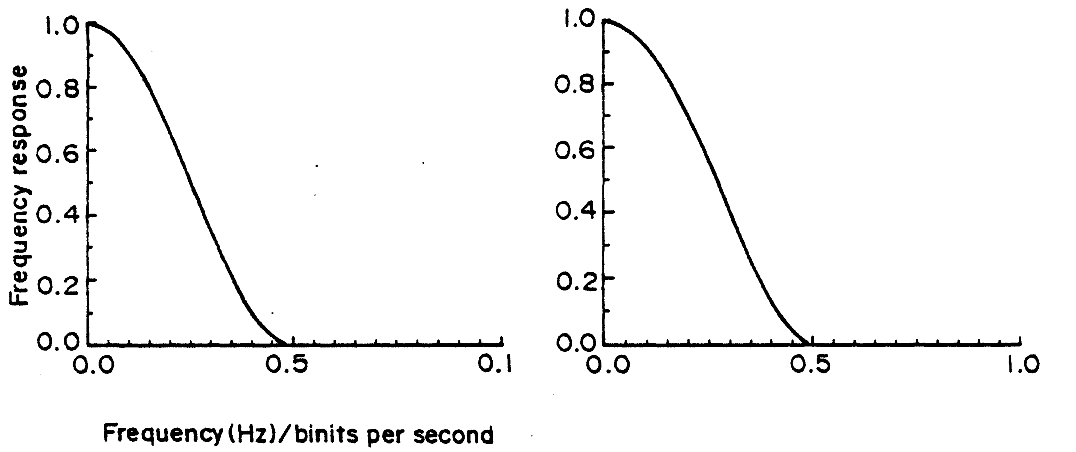


(b) Single pulse response for d > 0 (with a half bandwidth and zero crossings only at the binit edges = flux transitions)

Figure 25 (Part 2 of 2). Half-Bandwidth Operation (α = 1 or β = 2)

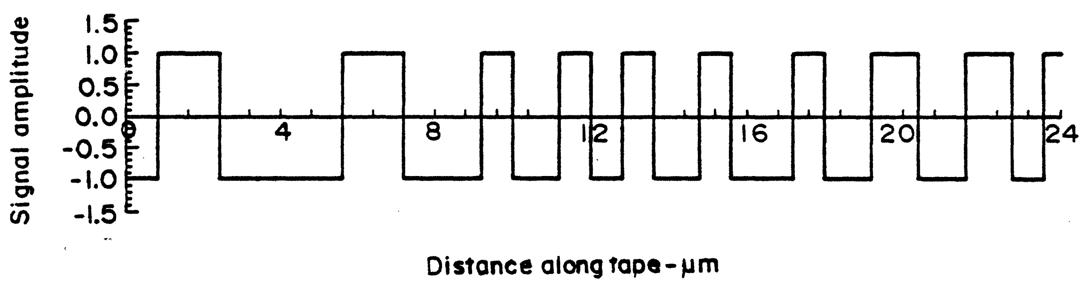
The important constraint is that there must be no intersymbol interference at binit boundaries; the mid-binit constraint is superfluous. For d > 0, interference can be eliminated at binit edges with a reduced bandwidth of $B_n = 1/2T_m = f_c/2$. The bandwidth reduction doubled the width of the pulse itself, which is only allowed by the d>0 constraint. All binit read pulses then have unit amplitude at a flux reversal, and the read pulse tails cross zero at flux transitions, as shown in Figure 25. This narrower bandwidth results in output signal zero crossings at a point of no interference, but binit centers are not constrained. This bandwidth reduction is obtained with a penalty of greater detection ambiguity in the presence of channel impairments. The signal zero-crossing slope is reduced with the bandwidth; this increases detection sensitivity to noise, tape speed variations, analog-channel differences, or improper equalization. Figure 25 shows the performance of a half-bandwidth β = 2 equalization channel with a 1,k 2/3 rate modulation code. The recording density is 1250 fr/mm, as shown in preceding figures, but the modulation density is increased to

2500 binits/mm and the linear density is 1667 bits/mm. Each binit is now only one half of a division in Figures 26c, 26d, and 26e.



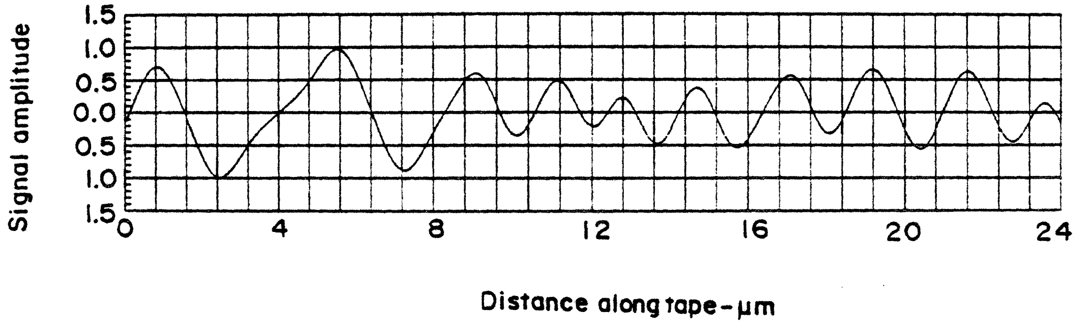
(a) Equalization-channel impulse-transfer function

(b) Pulse-input frequency response

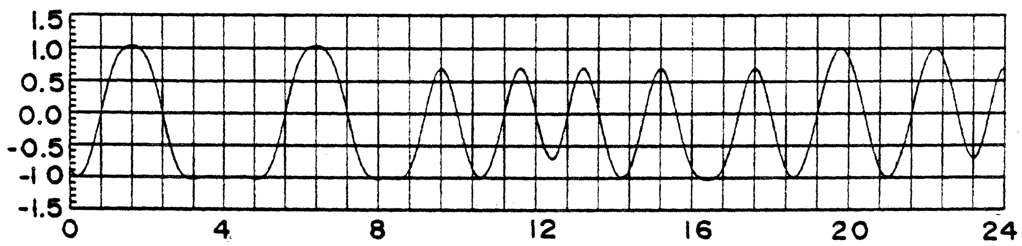


(c) Input signal (1250 fr/mm, 2500 binits/mm)

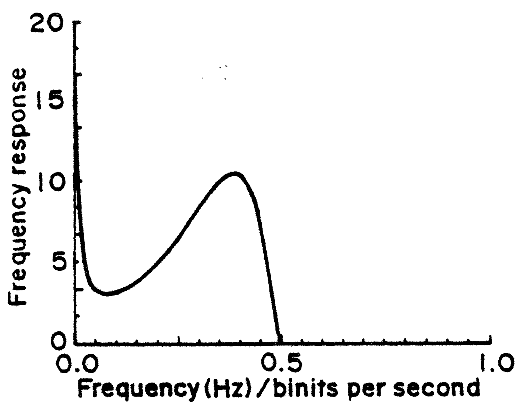
Figure 26 (Part 1 of 2). Half-Bandwidth Operation (1,7 2/3 Rate Code, β = 2)

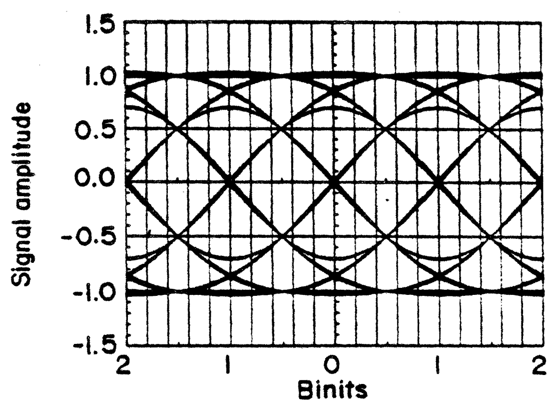


(d) Unequalized analog-channel output



(e) Equalized output signal





(f) Equalizer transfer function (g) Eye pattern

Figure 26 (Part 2 of 2). Half-Bandwidth Operation (1,7 2/3 Rate Code, $\beta = 2$)

There is no intersymbol interference at the signal zero crossings, but there is some amplitude variation between zero crossings. The bandwidth is less for NRZI modulation even though more information is recorded with NRZI (bandwidth = 0.75 and bit rate = 1.33 relative to NRZI). In this case, the reduced bandwidth compensates for the modulation-code rate loss.

The $\alpha = 1$ and β equalizers permit output zero crossings to occur at the equivalent of input pulse edges. Data detection can then be obtained by hard limiting the equalized signal. This produces an output that is similar to the input, but only if the equalizer response extends to dc, which is not the case for a magnetic tape channel. The low-frequency loss causes the signal to drift up and down, resulting in output binits that are lengthened or shortened according to the degree of amplitude offset at the detector. This problem can be reduced by the use of dc restoration (which adds cost) or with a dc-free modulation code.

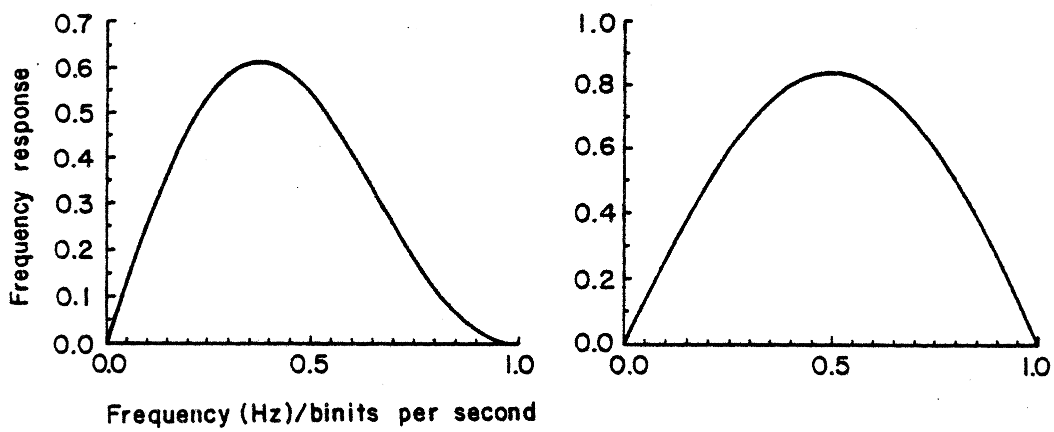
To achieve the desired low-frequency response for a waveform-restoration equalizer, the low-frequency signals may have to be amplified significantly. This can seriously degrade SNR under some conditions. If the tape is not erased, the low-frequency overwrite noise becomes a serious problem. Even an erased tape contains considerable low-frequency noise. The read-head side sensitivity can pick up low-frequency signals from an adjacent track that is then amplified. A magneto-resistive read head, which is also thermally sensitive, may pick up low-frequency noise from intermittent contact with the tape. If low-frequency noise is present in significant amounts, waveform-restoration equalization techniques may not be satisfactory unless a modulation code with no dc and little low-frequency content is used. The computer simulations are noise-free. A typical analog channel would produce a high noise level at the equalizer output. The waveform-restoration equalizer is sometimes outperformed by a derivative equalizer.

DERIVATIVE EQUALIZER

Waveform-restoration equalization does not match the analog-channel transfer function well because of the low-frequency discrepancy. Derivative equalization reduces the channel low-frequency requirement, and attempts to produce a signal that is the derivative of the input signal. This is automatically achieved at low densities, as shown in Figure 11b on page 21, because the reproduce head transfer function at low frequencies approximates a differentiator. But as recording densities increase, the read pulses overlap and interfere with each other, as shown in Figures 11c and 11d. If these higher density signals can be converted to pulses with well-defined peaks at the equivalent of flux reversal times, then both clock and data information can be extracted by peak detection, as with the unequalized low-density signal; there are several equalizers that will accomplish approximately this result. The low-frequency boost is significantly reduced, but more high-frequency boost is often required.

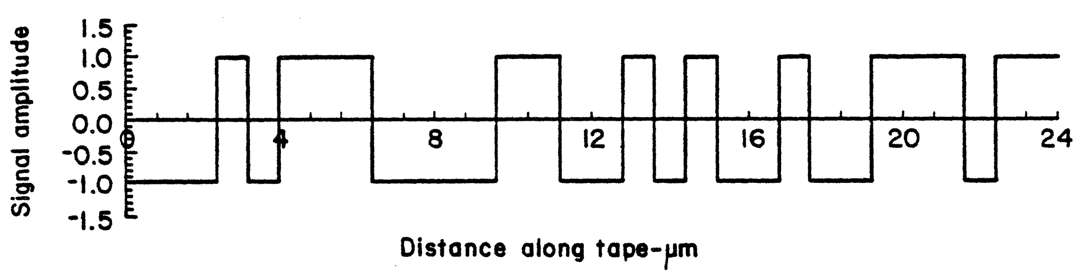
An ideal derivative equalizer produces alternating bipolar pulses with peaks that precisely correspond to the input steps, as shown in Figure 16b on page 33. A peak is desired when the waveform-restoration equalizer produces a zero crossing, which suggests that a time derivative of the waveform-restoration equalized signal might produce the desired output pulses. The time derivatives of many waveform-restoration equalizer signals do produce useful derivative equalizer signals. In addition, the frequency derivative of some β channel impulse responses produces a useful derivative equalizer impulse response.

Figure 27 shows the performance of a d/df β = 3 equalizer. Although the signal peaks are not all of equal amplitude, they all occur at binit-boundary time. This is sometimes called the $\cos^2$ equalizer.



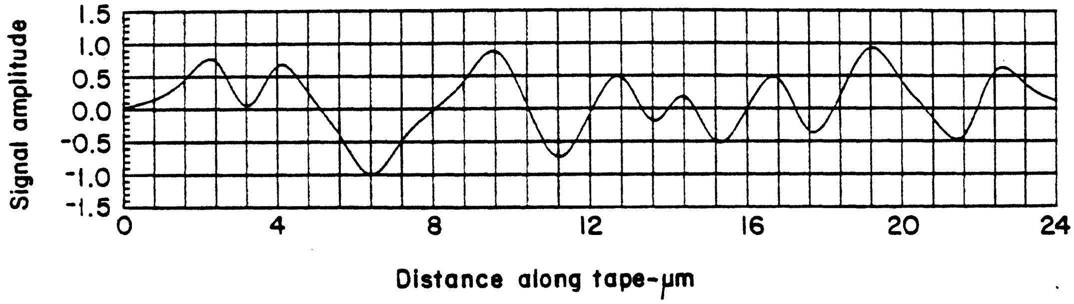
(a) Equalization-channel impulse-transfer function

(b) Pulse-input frequency response

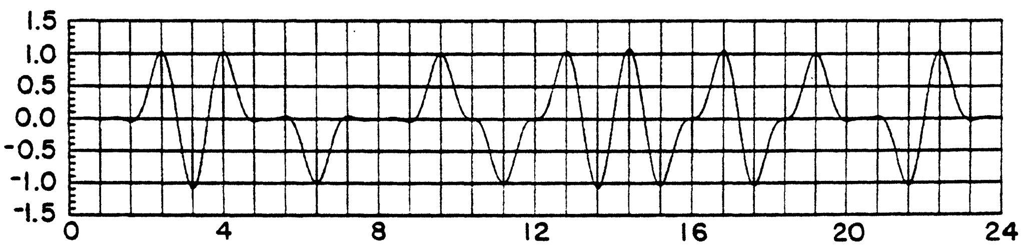


(c) Input signal (1250 fr/mm, 2500 binits/mm)

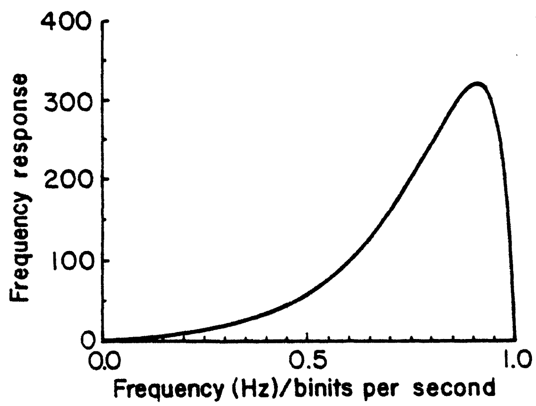
Figure 27 (Part 1 of 2). Derivative Equalizer (d/df β = 3)

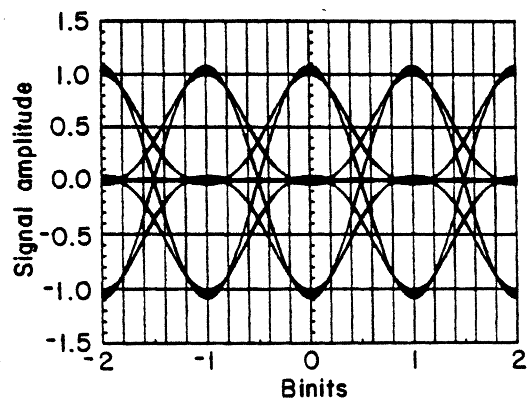


(d) Unequalized analog-channel output



(e) Equalized output signal





(f) Equalizer transfer function (g) Eye pattern

Figure 27 (Part 2 of 2). Derivative Equalizer (d/df $\beta = 3$)

Figure 28 shows the d/df β family characteristics, and Figure 29 on page 64 and Figure 30 on page 66 show the performance of the d/dt β and d/dt α equalizer families. These derivative equalizers all produce output signals with peaks that are approximately aligned with the flux transitions. The frequency derivative of a raised cosine $\alpha < 1$ equalizer does not produce useful results.

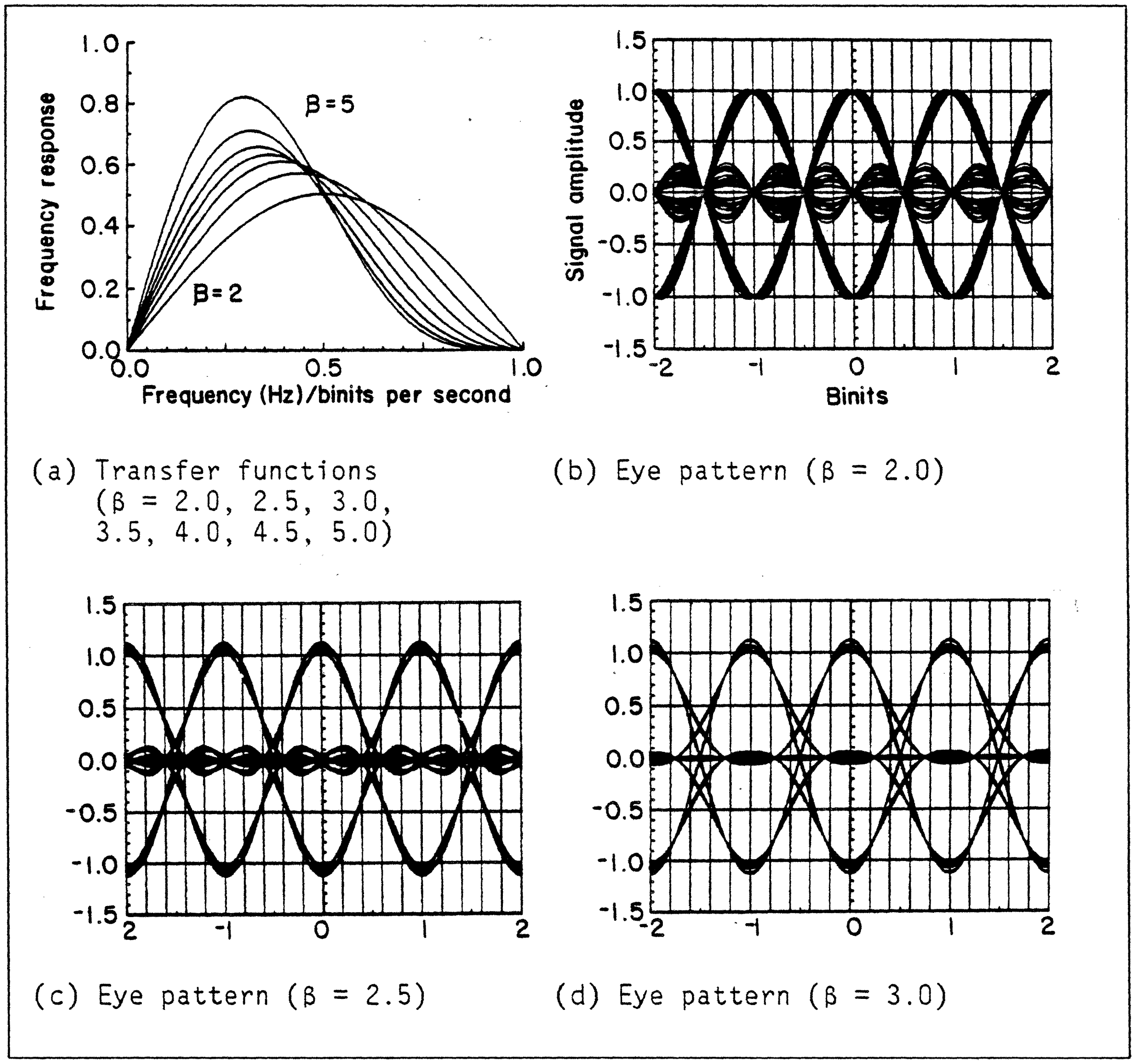
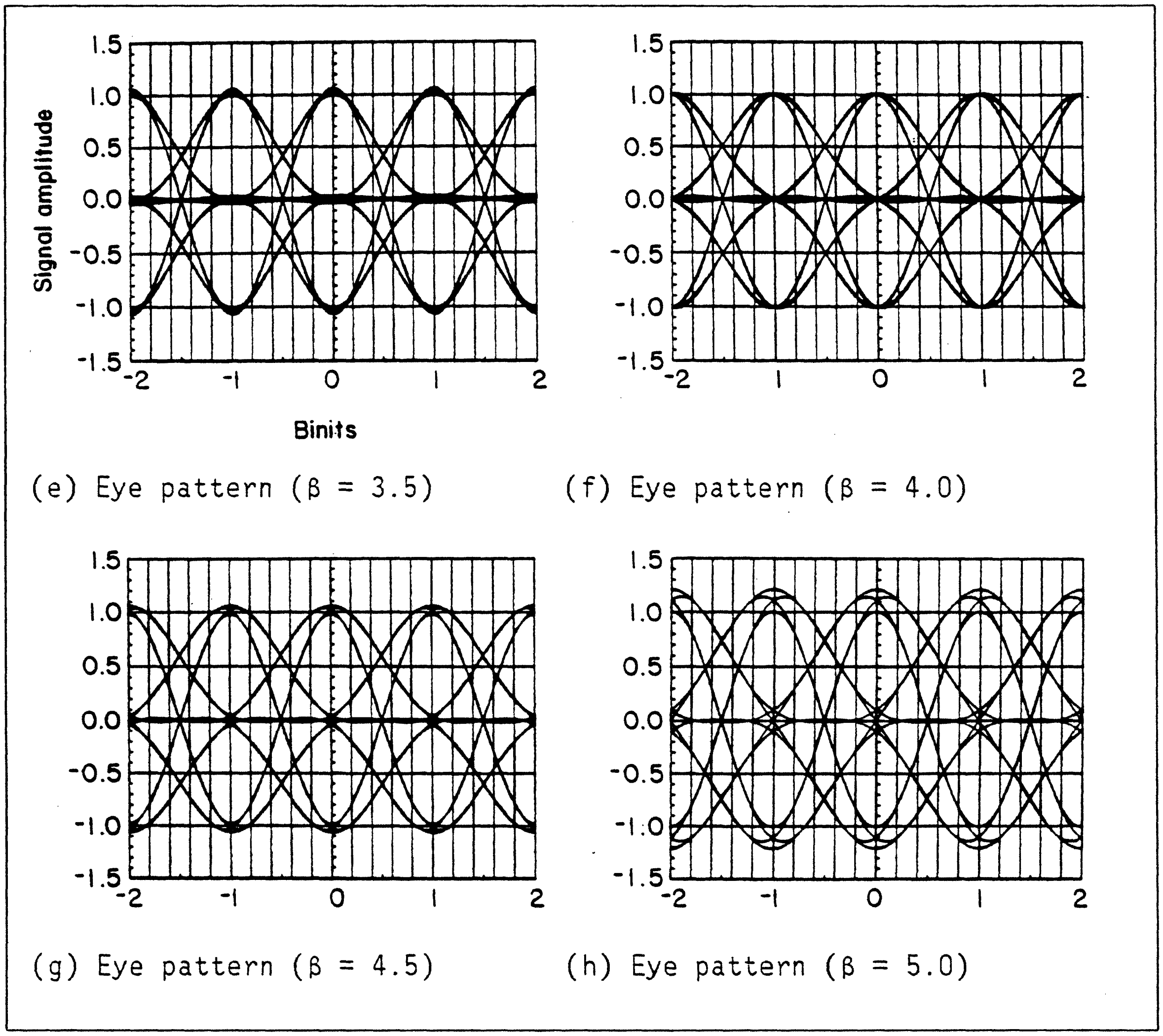


Figure 28 (Part 1 of 2). Comparison of d/df β Equalizer



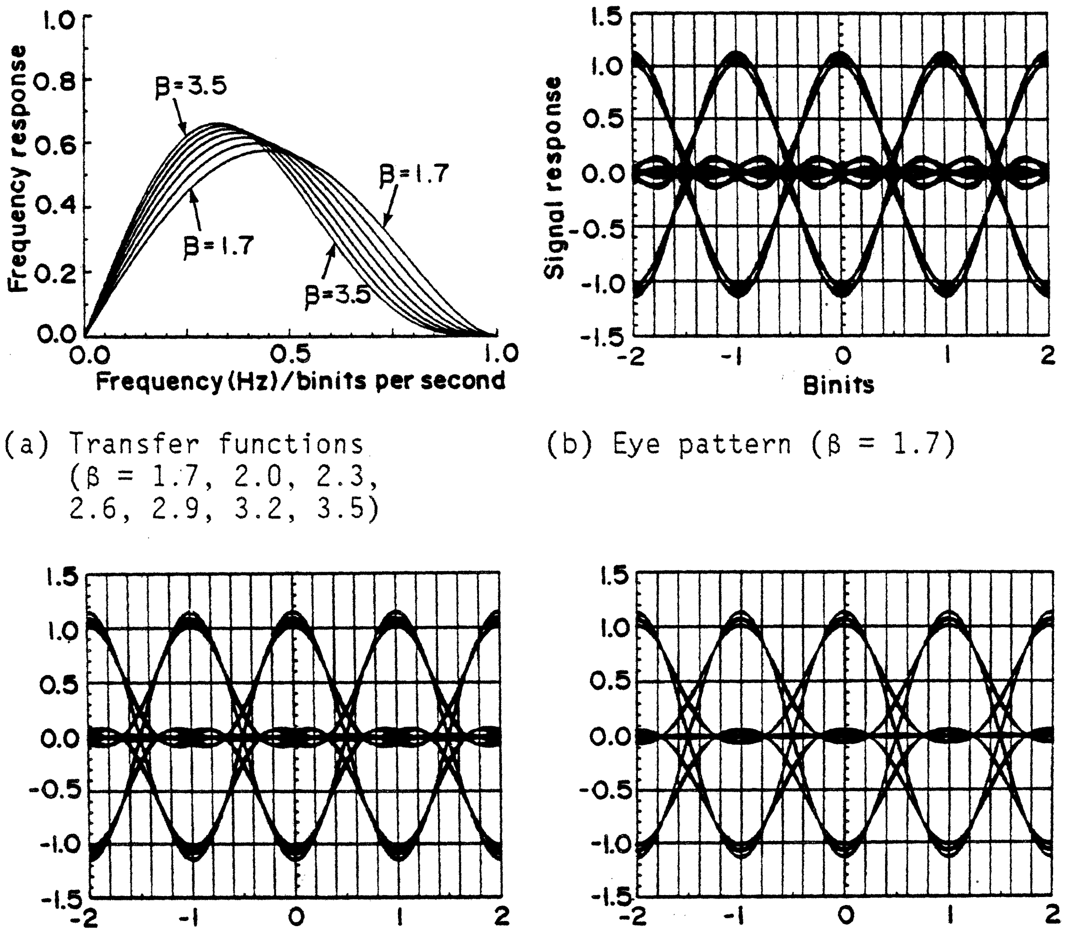
(e) Eye pattern (β = 3.5)

(f) Eye pattern (β = 4.0)

(g) Eye pattern (β = 4.5)

(h) Eye pattern (β = 5.0)

Figure 28 (Part 2 of 2). Comparison of d/df β Equalizer



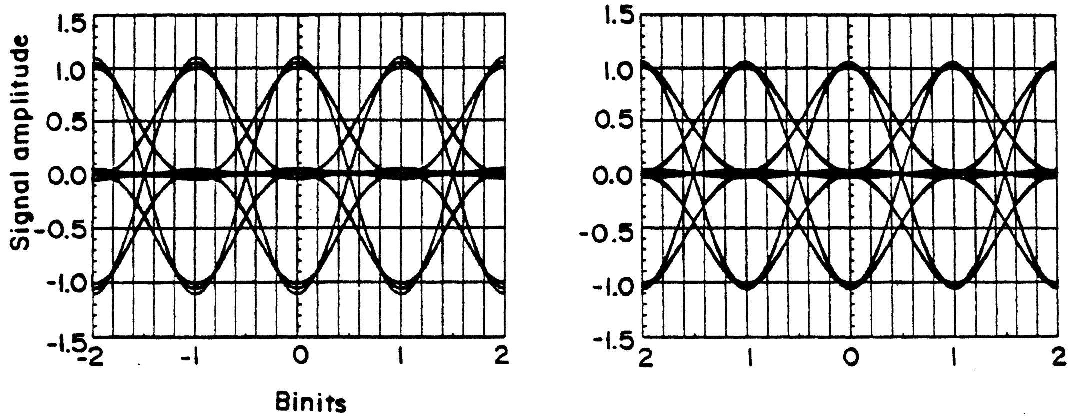
(a) Transfer functions (β = 1.7, 2.0, 2.3, 2.6, 2.9, 3.2, 3.5)

(b) Eye pattern (β = 1.7)

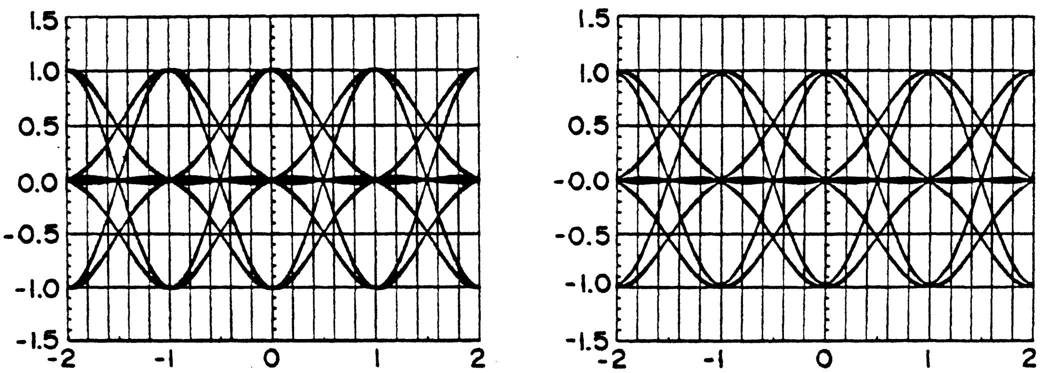
(c) Eye pattern (β = 2.0)

(d) Eye pattern (β = 2.3)

Figure 29 (Part 1 of 2). Comparison of d/dt β Equalizer

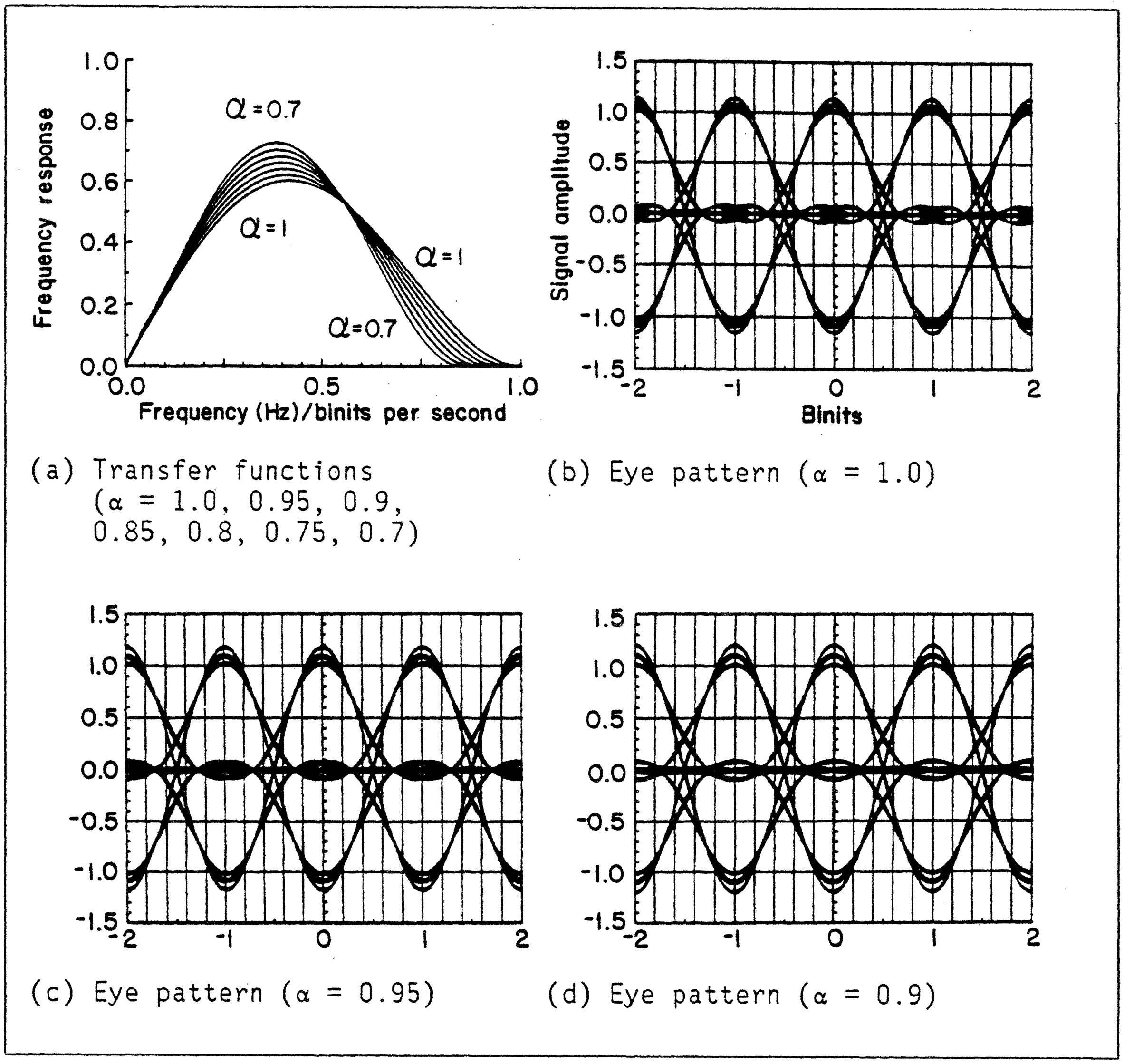


(e) Eye pattern (β = 2.6) (f) Eye pattern (β = 2.9)



(g) Eye pattern (β = 3.2) (h) Eye pattern (β = 3.5)

Figure 29 (Part 2 of 2). Comparison of d/dt β Equalizer



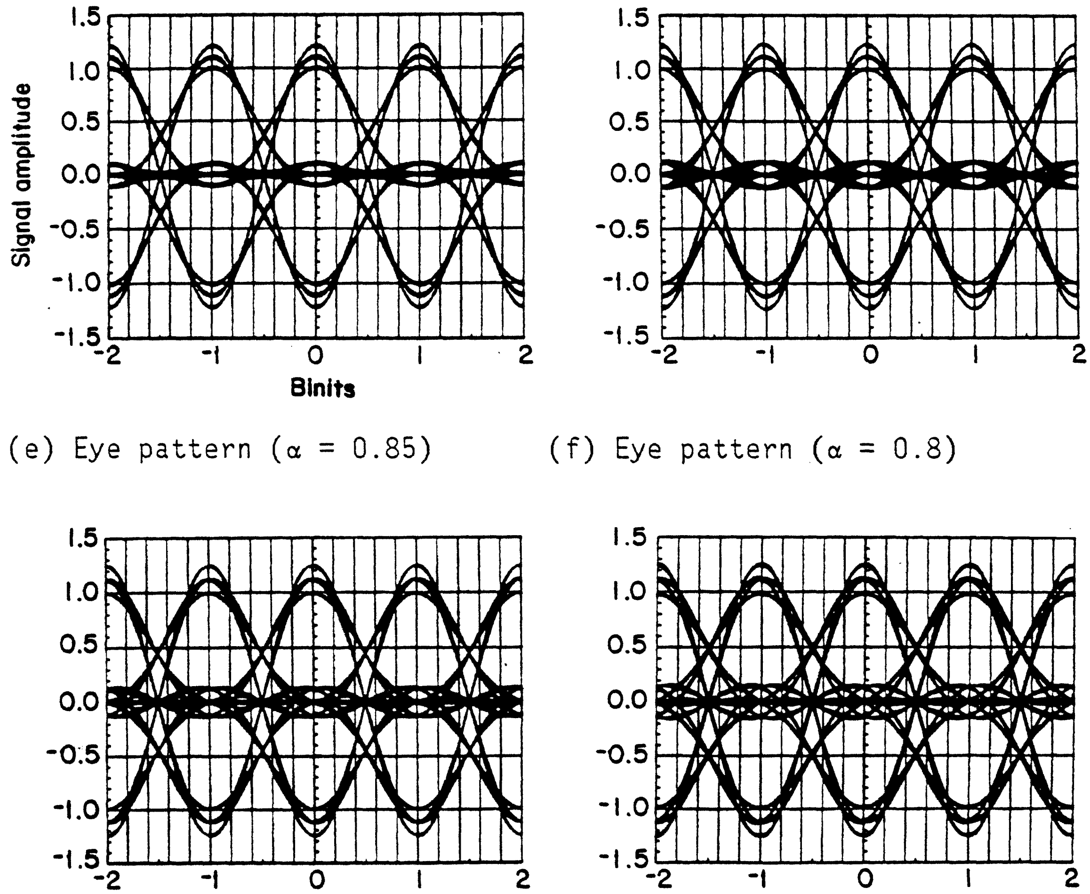
(a) Transfer functions (α = 1.0, 0.95, 0.9, 0.85, 0.8, 0.75, 0.7)

(b) Eye pattern (α = 1.0)

(c) Eye pattern (α = 0.95)

(d) Eye pattern (α = 0.9)

Figure 30 (Part 1 of 2). Comparison of d/dt α Equalizer



(e) Eye pattern (α = 0.85) (f) Eye pattern (α = 0.8)

(g) Eye pattern (α = 0.75) (h) Eye pattern (α = 0.7)

Figure 30 (Part 2 of 2). Comparison of d/dt α Equalizer

Differentiation of a waveform-restoration channel is not the only way to obtain a useful derivative equalization channel. Another equalizer family that approximates the desired result is the γ equalizer, which has a transfer function:

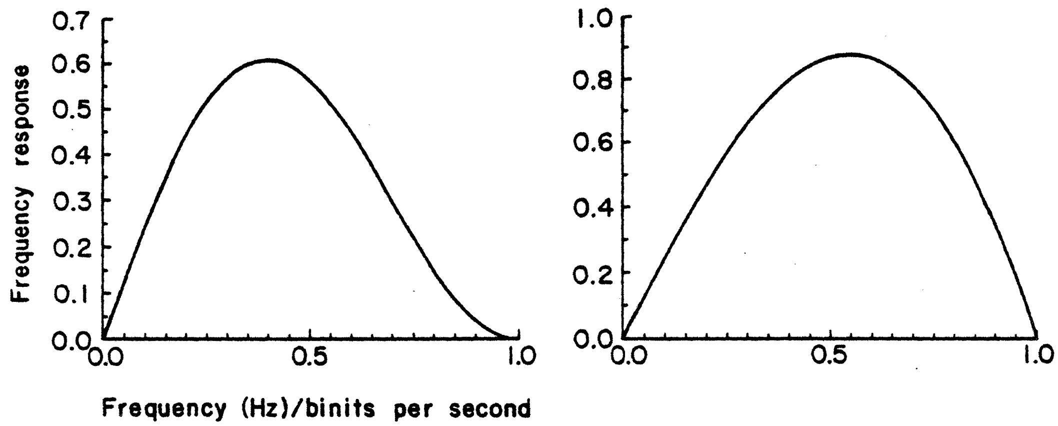
$$H(f) = x(\sin x/x)^{\gamma} \qquad 0 \le x \le \pi \tag{13}$$

$$= 0 \qquad x > \pi$$

where

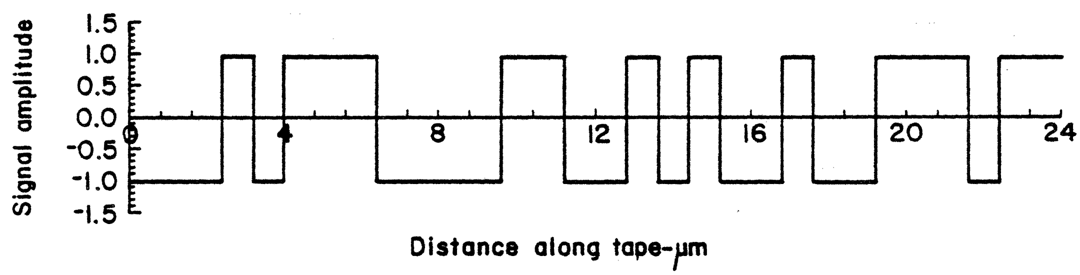
$x = \pi\, f/f_c$

Figure 31 shows the γ = 2 equalizer or sine equalizer.

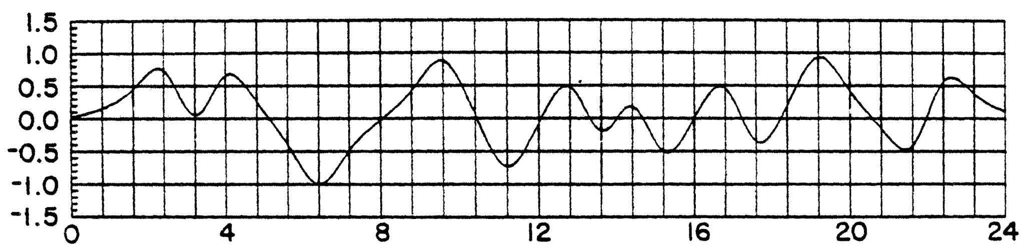


(a) Equalization-channel impulse-transfer function

(b) Pulse-input frequency response

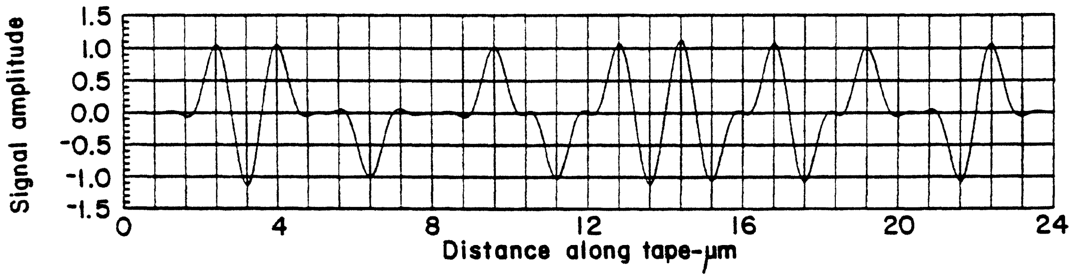


(c) Input signal (1250 fr/mm)

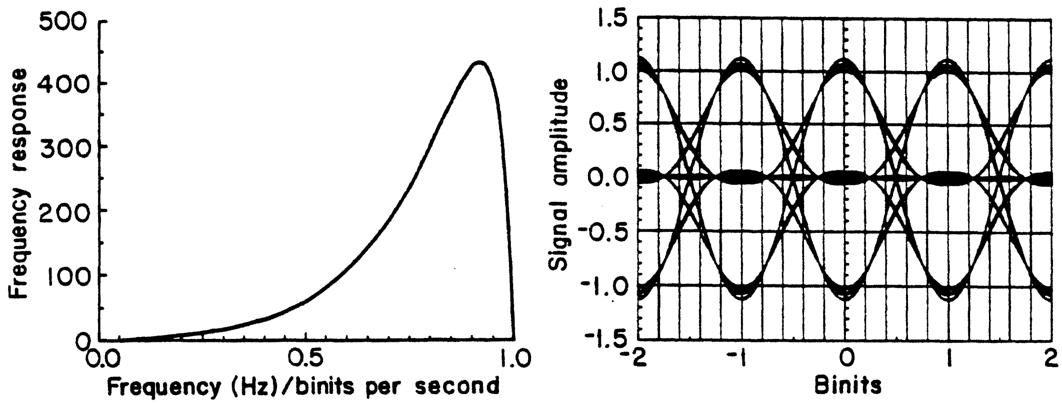


(d) Unequalized analog-channel output

Figure 31 (Part 1 of 2). Derivative Equalizer (γ = 2)



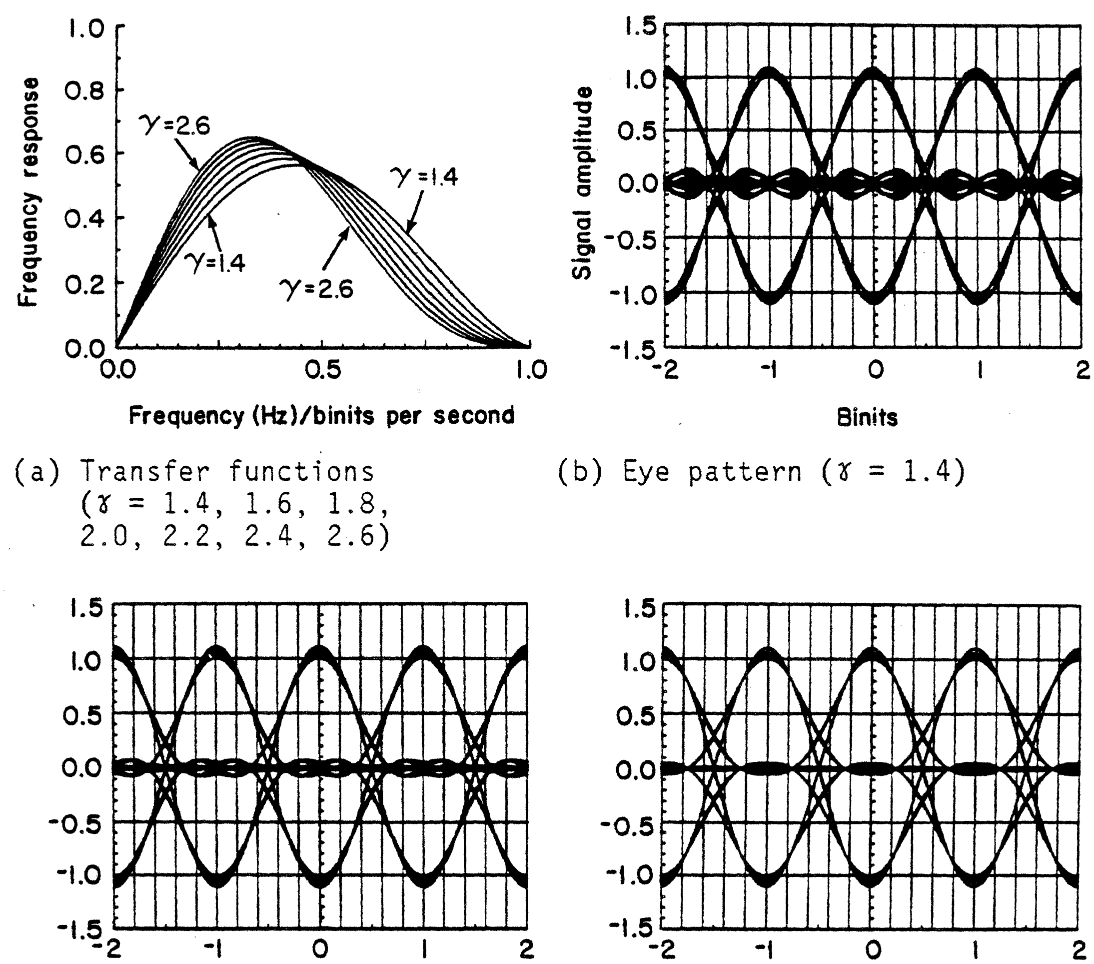
(e) Equalized output signal



(f) Equalizer transfer function (g) Eye pattern

Figure 31 (Part 2 of 2). Derivative Equalizer ($\gamma = 2$)

Figure 32 shows characteristics of the γ equalizer family. As with the α and β equalizers, the half-bandwidth operation is possible if $d > 0$, but analog-channel perturbations degrade performance more than with a full-bandwidth channel.



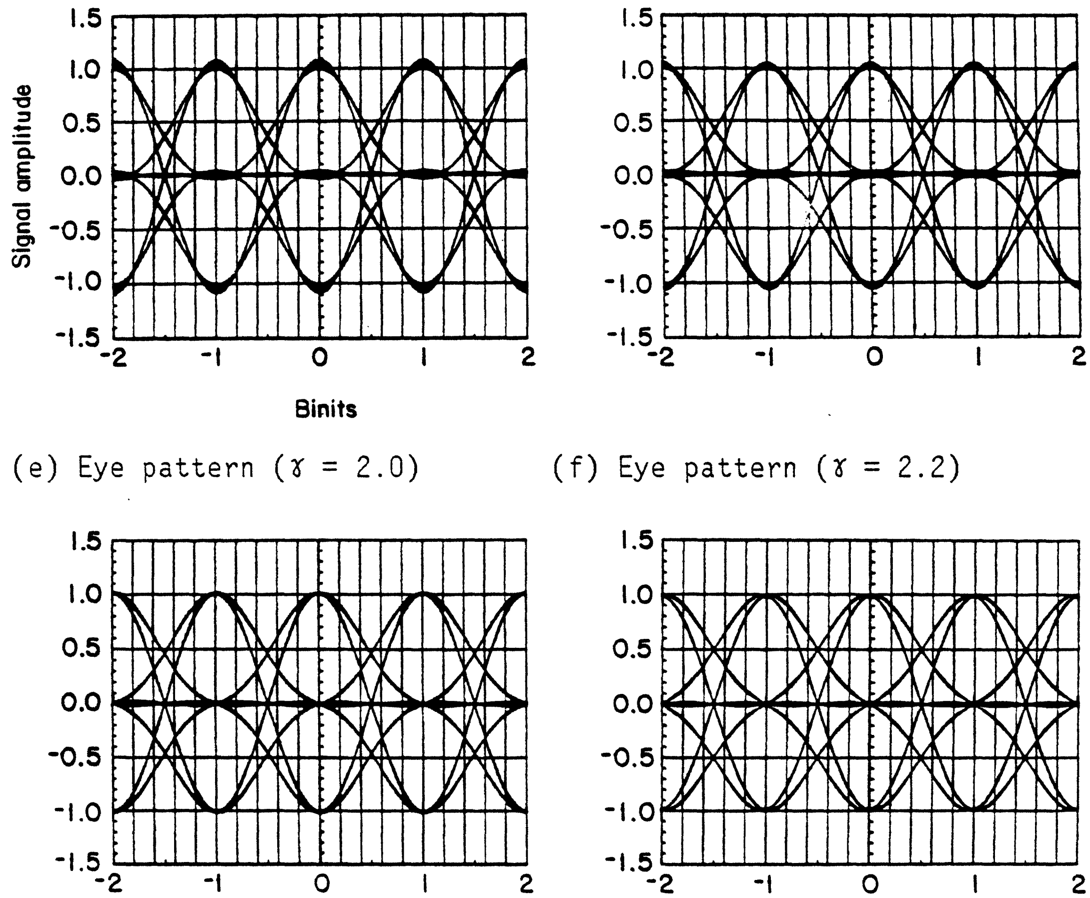
(a) Transfer functions (γ = 1.4, 1.6, 1.8, 2.0, 2.2, 2.4, 2.6)

(b) Eye pattern (γ = 1.4)

(c) Eye pattern (γ = 1.6)

(d) Eye pattern (γ = 1.8)

Figure 32 (Part 1 of 2). Comparison of -γ Equalizer



(e) Eye pattern (γ = 2.0)

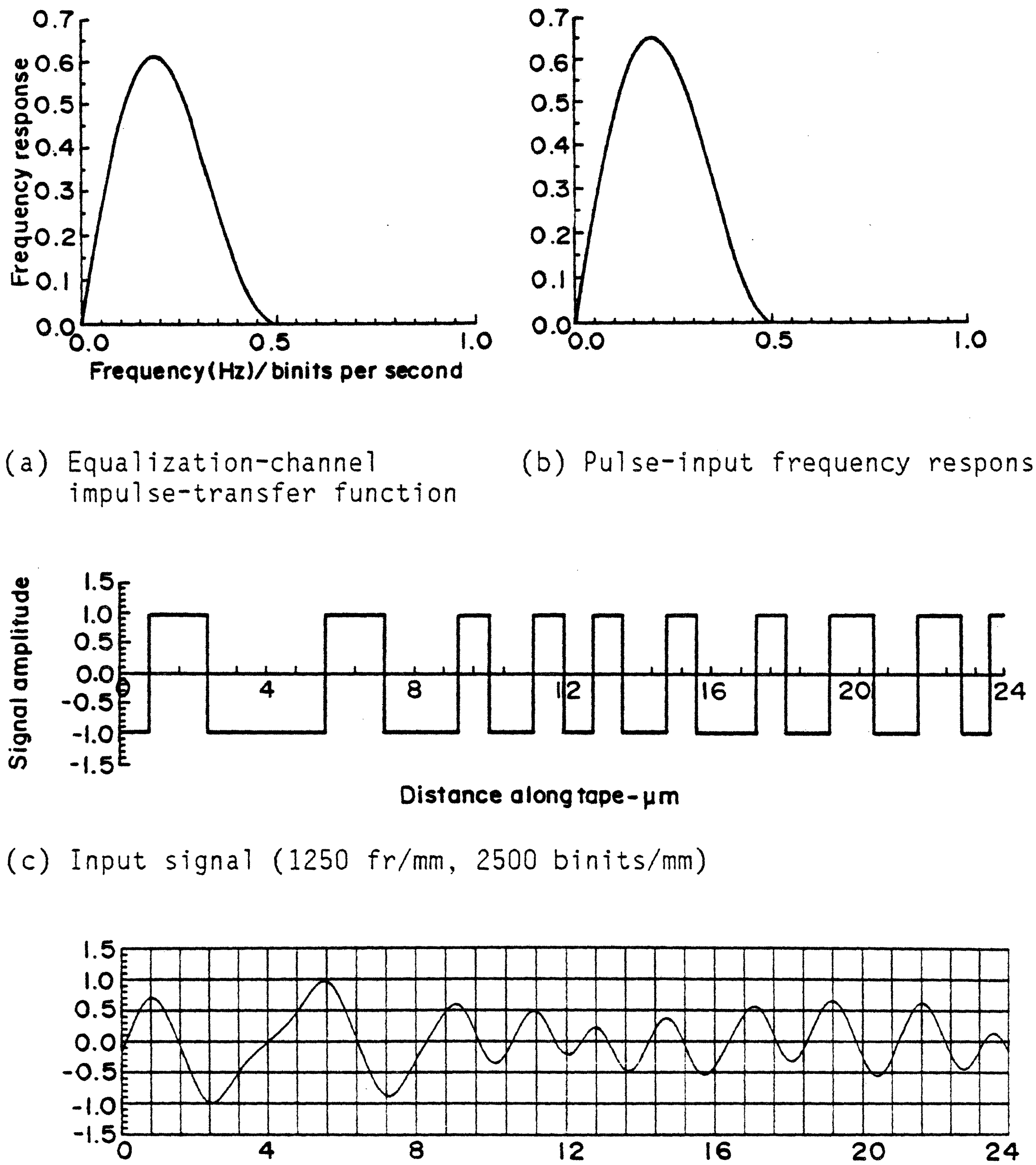
(f) Eye pattern (γ = 2.2)

(g) Eye pattern (γ = 2.4)

(h) Eye pattern (γ = 2.6)

Figure 32 (Part 2 of 2). Comparison of -γ Equalizer

Figure 33 shows the half-bandwidth operation with a $\gamma = 2$ channel and a 1,7 2/3 rate modulation code. If more waveform-restoration equalizers are needed, the integral of some γ equalizers offers more options.



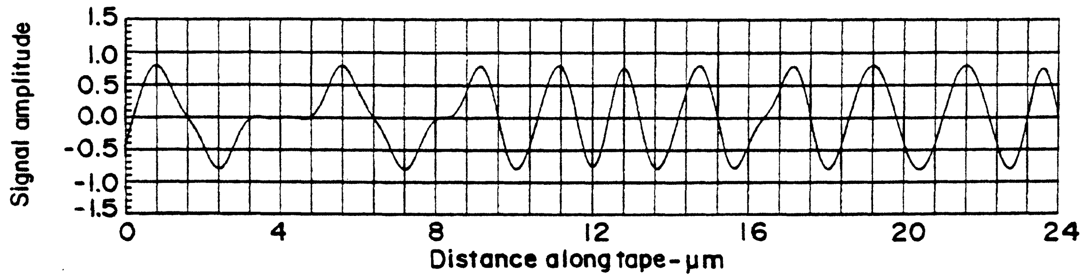
(a) Equalization-channel impulse-transfer function

(b) Pulse-input frequency response

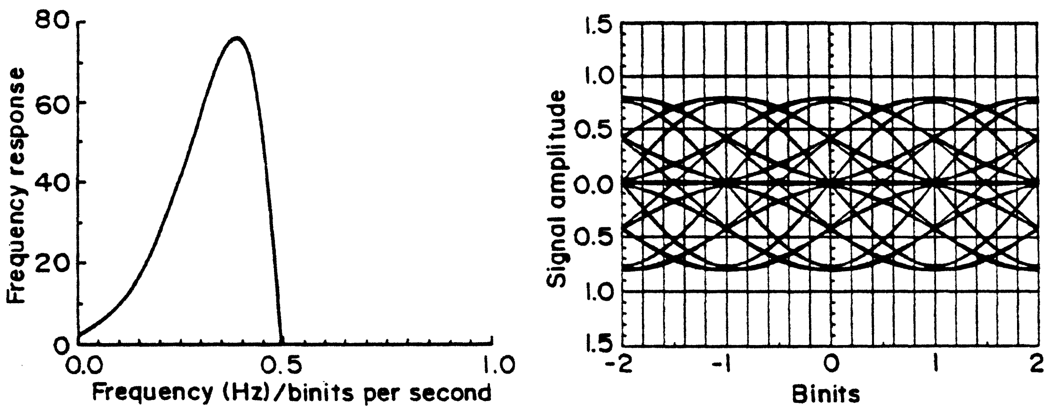
(c) Input signal (1250 fr/mm, 2500 binits/mm)

(d) Unequalized analog-channel output

Figure 33 (Part 1 of 2). Half-Bandwidth Operation (1,7 2/3 Rate Code, $\gamma = 2$)



(e) Equalized output signal



(f) Equalizer transfer function (g) Eye pattern

Figure 33 (Part 2 of 2). Half-Bandwidth Operation (1,7 2/3 Rate Code, $\gamma = 2$)

Derivative equalizers produce a ternary signal with peaks that represent the flux transitions, and a near-zero signal at no-transition binit boundaries and during binits, which are not adjacent to the flux transitions, as shown in Figure 31e on page 69. Because noise and signal overshoots produce small signal peaks near the baseline, the detector must only sense the peaks above a threshold.

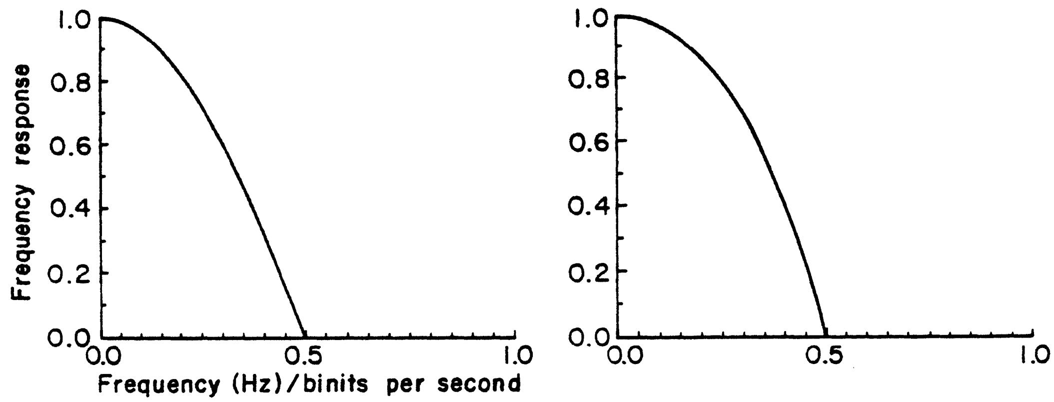
The derivative equalization-channel transfer functions do not require dc response, and little low-frequency response. They are less susceptible to low-frequency noise from poor erase, overwrite, read-head side sensitivity, or magneto-resistive head thermal sensitivity, but they typically require more high-frequency gain than a waveform-restoration channel. An analog channel with more low-frequency noise than high-frequency noise may benefit from the use of a derivative equalizer. If high-frequency noise predominates, then waveform restoration may be preferable.

Equalizer design involves a compromise between SNR and intersymbol interference. With a bandwidth B_n, the overall transfer function can be modified somewhat without disastrous effect on the output signal. As the bandwidth is narrowed, channel variations have greater effect on performance. An optimum equalizer for nominal signal is probably suboptimal when the transfer function changes during a dropout. A common design procedure is to characterize the analog-channel transfer functions for a typical range of heads and tapes, and then design a reasonable equalizing filter. The final selection of equalizer components is empirically determined by the best results achieved from extreme case heads and tapes with dropouts. An alternative procedure requires a definition of the "typical error producing dropout" transfer function. The equalizer can be designed for this situation, but some experimental optimization is usually required.

PARTIAL RESPONSE EQUALIZER

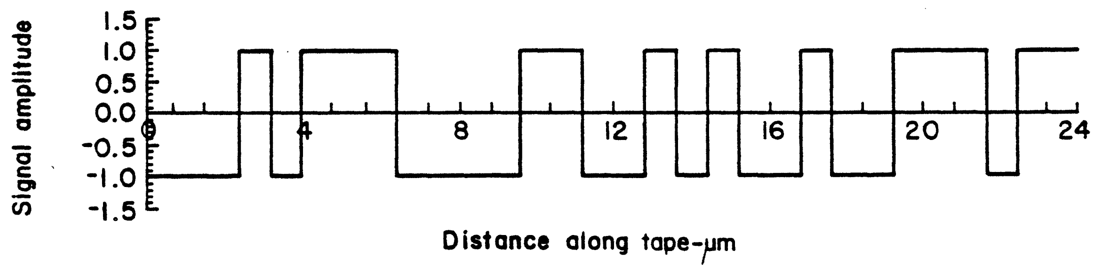
The waveform restoration and derivative equalizers eliminate the read-signal intersymbol interference at critical times. An interference-free clock can then be extracted, and there will be no distortion at data sample times. This performance is achieved at the expense of bandwidth. A partial response equalizer permits additional controlled intersymbol interference, and compensates for the known effect at the data detector. The advantage is a reduced bandwidth requirement that can be achieved with a practical filter. The disadvantages are increased hardware complexity and sensitivity to signal perturbations.

Many partial response equalizers are possible. The most popular is a half-bandwidth $\beta = 1$ channel equalizer, which is called duobinary by the inventor, although other terms have also been applied, as discussed by Lender [ref.37]. The duobinary equalizer is shown in Figure 34 on page 75.

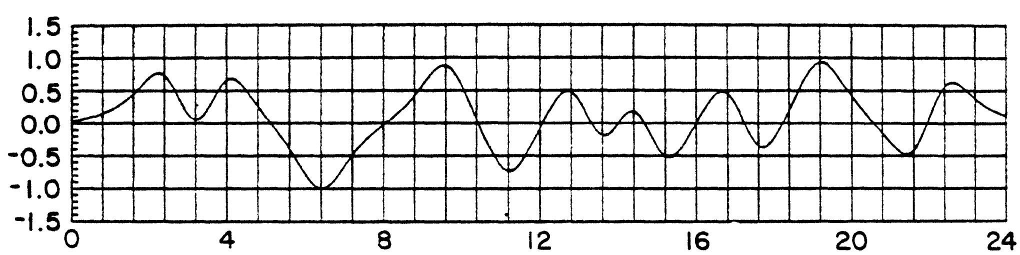


(a) Equalization-channel impulse-transfer function

(b) Pulse-input frequency response

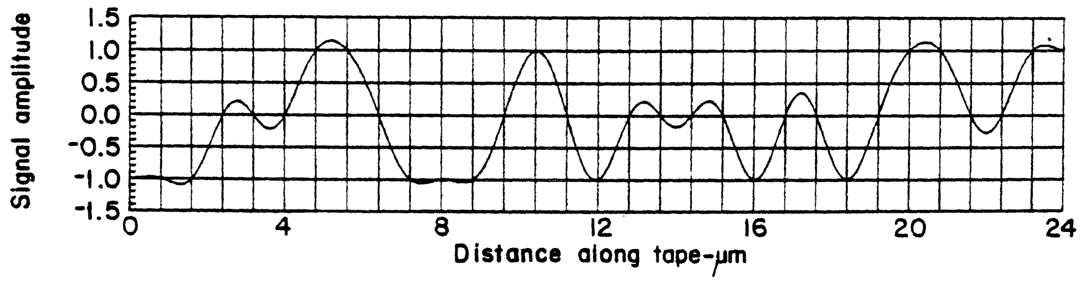


(c) Input signal (1250 fr/mm)

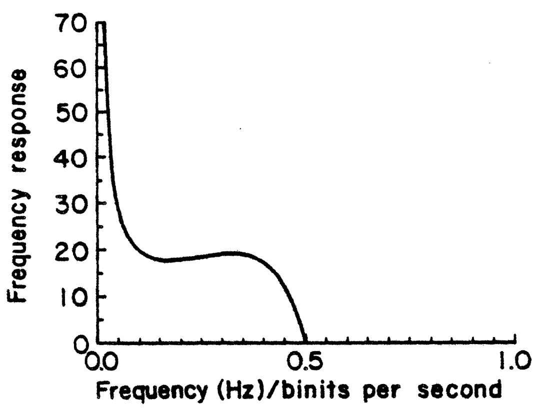


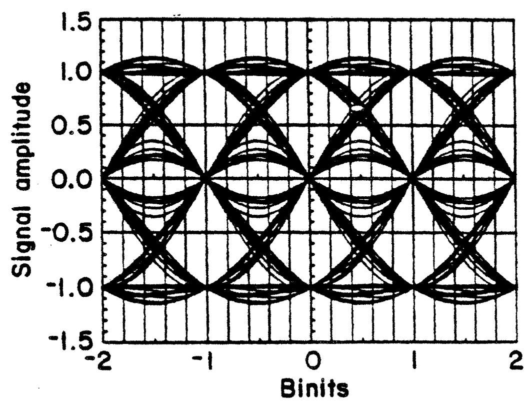
(d) Unequalized analog-channel output

Figure 34 (Part 1 of 2). Partial Response for PR1 Equalizer (Duobinary)



(e) Equalized output signal

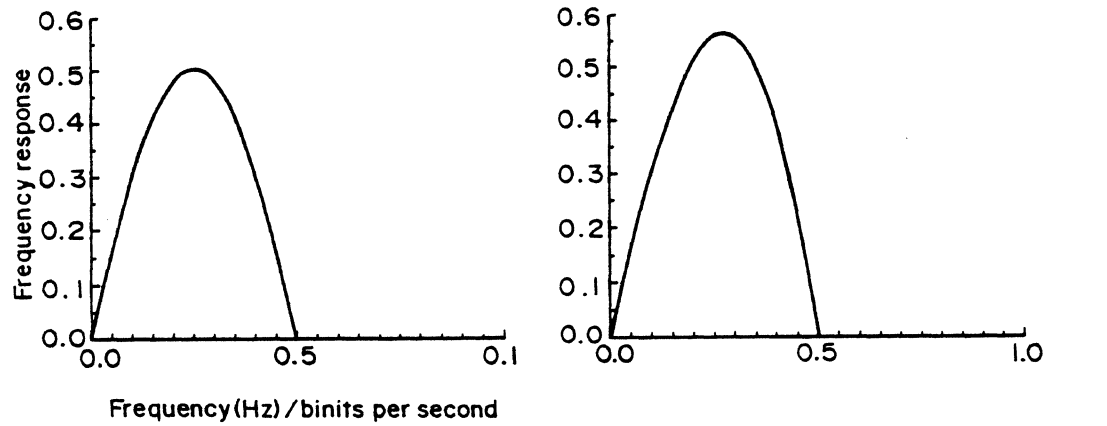




(f) Equalizer transfer function (g) Eye pattern

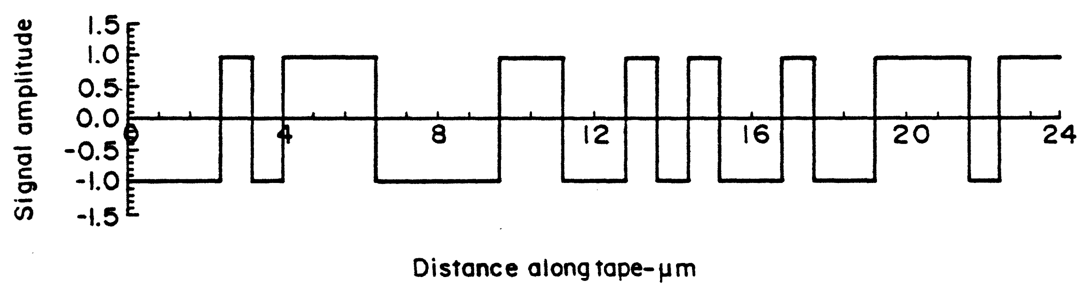
Figure 34 (Part 2 of 2). Partial Response for PR1 Equalizer (Duobinary)

Kretzmer [ref.38] categorized some characteristics of various partial response equalizers. The duobinary technique is called partial response class 1 (PR1). The other type of partial response equalizer that is of some interest for magnetic tape is called modified duobinary or class 4 (PR4). The PR4 equalization-channel characteristics are shown in Figure 35 on page 77.

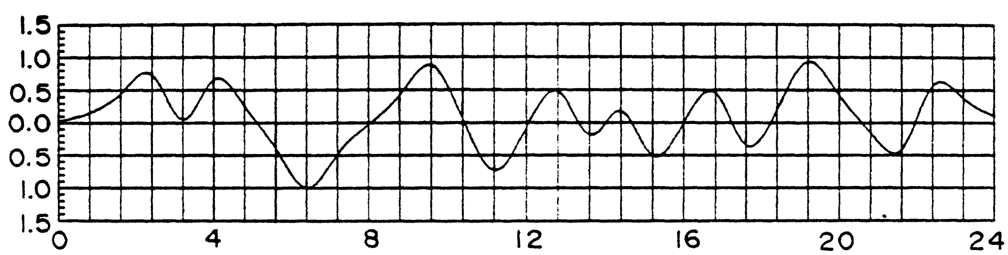


(a) Equalization-channel impulse-transfer function

(b) Pulse-input frequency response

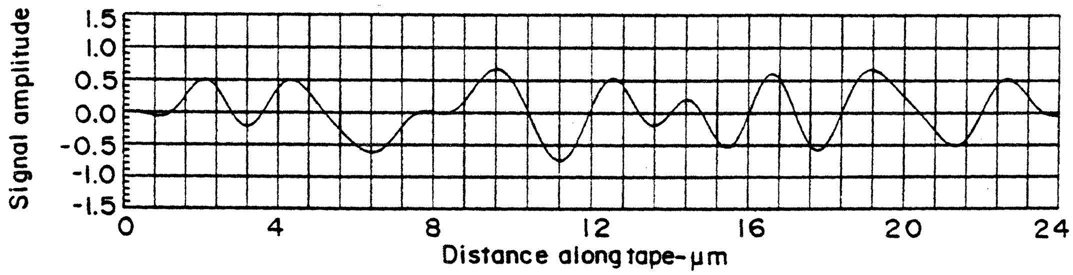


(c) Input signal (1250 fr/mm)

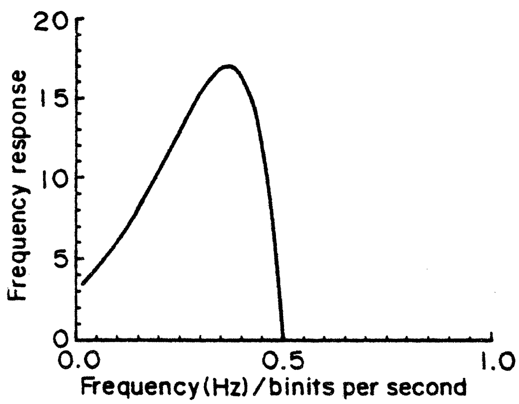


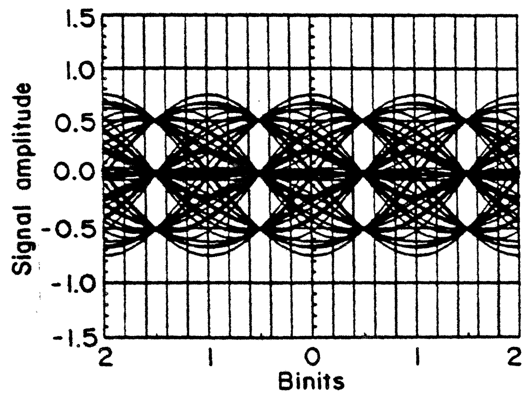
(d) Unequalized analog-channel output

Figure 35 (Part 1 of 2). Partial Response for PR4 Equalizer (Modified Duobinary)



(e) Equalized output signal





(f) Equalizer transfer function (g) Eye pattern

Figure 35 (Part 2 of 2). Partial Response for PR4 Equalizer (Modified Duobinary)

The other classifications have not been recognized as useful for magnetic recording purposes, as discussed by Kretzmer [ref.38].

The partial response channel characteristics can be described in terms of linear digital filters composed of delay operators F(D). Kabal and Pasupathy [ref.39] define nine different partial response channels in terms of the system polynomial F(D), transfer function H(ω), impulse response h(t), and number of output-signal levels. They also describe performance characteristics. Their detailed analysis of partial response capabilities concludes that duobinary and modified duobinary, PR1 and PR4, have the best performance potential. The system polynomials for these responses are 1+D and $(1+D)(1-D) = 1-D^2$, respectively, as discussed by Kabal and Pasupathy [ref.39]. Partial response channels usually require precoders to transform user data sequences into sequences suitable to the partial response channel. These can be digital filters with the system polynomial response, but these filters produce ternary signals that violate the saturated tape recording constraint that the write signal be binary. Equivalent binary precoders are also possible.

The duobinary or PR1 channel shown in Figure 34 on page 75 uses a half-bandwidth $\beta = 1$ equalizer with a $d = 0$ modulation code. This bandwidth is equal to that of a raised cosine $\alpha = 0$ equalizer; however, the cosine response is achievable but the abrupt cutoff filter is not. Because the narrow bandwidth does not allow the signal to slew from one extreme to the other in a single binit period, this equalizer produces a quasi-ternary output from a binary input. This is a degenerate form of waveform-restoration equalizer. In the absence of noise, hard limiting would restore the input signal. This is an impractical way to detect signals in an actual channel that contains noise and other signal impairments.

Equation 14 shows partial response 1 precoding, detection, and decoding. The partial-response 1 precoder adds each incoming binit i_k to the previous prerecorded binit r_{k-1} in a modulo-2 fashion to generate the current recorded binit r_k.

$$r_k = i_k \oplus r_{k-1} \tag{14}$$

where

i_k = the kth incoming binit

r_k = the kth recorded binit

0 = modulo-2 addition

The incoming binit sequence shown in Figure 34c on page 75 gives:

i_k = ?00111001000101011110110100110

r_k = 0001011100000110010100100111011

Figure 34e on page 76 shows the equalized analog reproduce signal resulting from the recorded data sequence r_k. Let us assume that a clock can be derived from the signal and that this clock permits sampling of the equalized signal at binit boundaries (vertical lines). Then, there are three possible sample values: -, 0 or +. The sample values s_k are:

s_k = x--000++0---0+0-0000-00-0++00+

The s_k + and - values are seen to represent the i_k 0 values and the s_k 0 values represent the i_k 1 values. Thus, the partial response channel, with a simple algorithm, reproduces the input sequence.

The precoder could be in the read circuitry or in the write circuitry. However, because the precoder uses previous binit information to generate the next binit, an error in any r_k could propagate indefinitely when this r_k is fed back as the next r_{k-1}. For this reason, the precoder is placed at a point in the system where errors are unlikely to occur in the write-side logic.

The PR4 channel shown in Figure 35 on page 77 can be analyzed in a similar manner to that used for the PR1 channel. The precoder equation is:

$$r_k = i_k \oplus r_{k-2} \tag{15}$$

As shown in Figure 35, the input sequences, recorded sequences and read samples are:

i_k = ??01001011001111110001101110101

r_k = 0001011100001100101001001111011

s_k = xx0+00+0--00++--+000-+0-++0-0+0

The detection algorithm is opposite to that for PR1; a 0 sample represents a 0 and a + or - sample represents a 1.

Although experimental and analytical effort has been expended on partial response channels for magnetic recording, no commercial computer tape product using duobinary or modified duobinary is known at the present time. If partial response techniques allow approximately double capacity and data rate, why are they not universally used? Although partial response is an effective bandwidth-compression technique, it has some shortcomings. An ISI-free clock cannot easily be recovered from the signal, and the multiple detection levels reduce data reliability even if an accurate clock exists. The clock frequency is beyond the partial response passband. Attempts have been made to utilize a parallel signal path for clock extraction, but this is expensive. Other implementations have involved nonlinear clock signal processing, such as full wave rectification or squaring circuits, but these have usually been sensitive to channel parameter variations. Clock stability and signal detection during dropouts are often difficult problems to solve. The best performance is achieved from the use of a maximum likelihood detector, which is too expensive for most applications, as discussed by Kabal and Pasupathy [ref.39]. Both duobinary and modified duobinary technique are used in communications applications, but they have not found acceptance for tape recording.

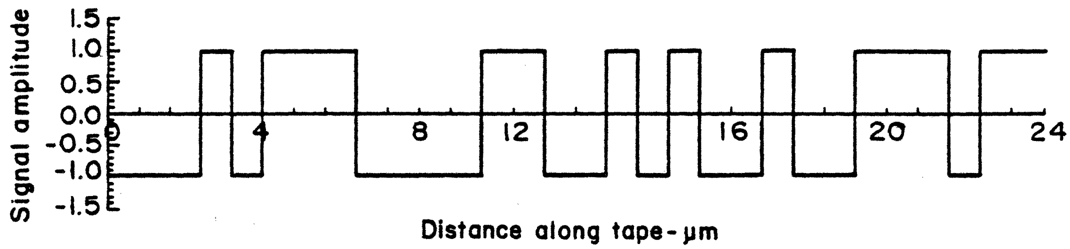
Partial response signaling offers interesting possibilities that may be used in future computer tape recorders, but there are other alternatives. For example, the half-bandwidth 1,7 2/3 rate modulation channels shown in Figure 26 on page 57 and Figure 33 on page 72 require only 13% more bandwidth for a given data rate than the partial response channels shown in Figure 34 on page 75 and Figure 35 on page 77, but they are much less complex to implement. A large value of β or γ can improve SNR with the penalty of increased ISI. The optimum selection depends upon the operating conditions and the analog channel.

WRITE EQUALIZER

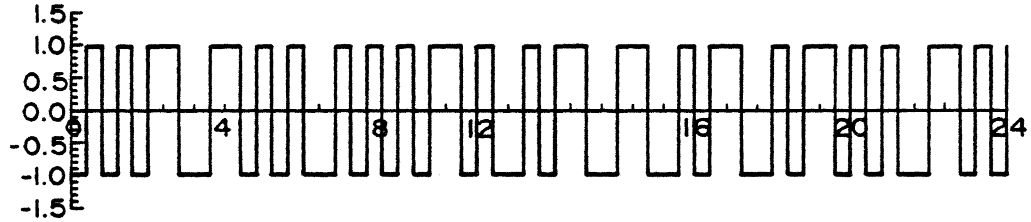
Write equalization (transformations performed before the write head) has been used in tape recording channels for many years. Some of these implementations involve empirically derived nonlinear techniques; however, others assume channel linearity and merely move some of the analog read equalization to the write side of the analog channel. Amplitude write equalization varies the write pulse amplitude to achieve high-frequency boost. A common form of this write equalizer increases the magnitude of the write current at the leading edge of each flux reversal, and for a short time thereafter, as discussed by Ambrico [ref.40]. This is called single-step write equalization. There is also a double-step write equalization, with an additional write-current increase just prior to transition time. Linear filters prior to the write head cause current reversals to overshoot with results somewhat similar to the single-step amplitude write equalization, but linearity is lost as the write current is already at the saturation level. The cosine response that is sometimes used for read equalization is suggested as a write equalizer to shape the write current, as discussed by Jacoby [ref.41]. Time write equalization shifts the transition positions so as to compensate for pulse crowding or peak shifts. Because of hardware complexity, it has not found wide acceptance for computer tape applications.

All of the linear equalizers in the preceding pages describe an overall transfer function, which is the product of the analog-channel and equalizer transfer functions. The equalizer transfer function may be achieved from a single filter or multiple filters. If the analog channel is truly quasi-linear as assumed, then the equalizer filters may be divided between the write and read side of the analog channel. As long as the analog-channel input is a binary signal, the required linearity exists. Thus, a linear digital filter with binary output that is inserted prior to the write head will have the same effect as a similar filter transfer function on the read side of the analog channel.

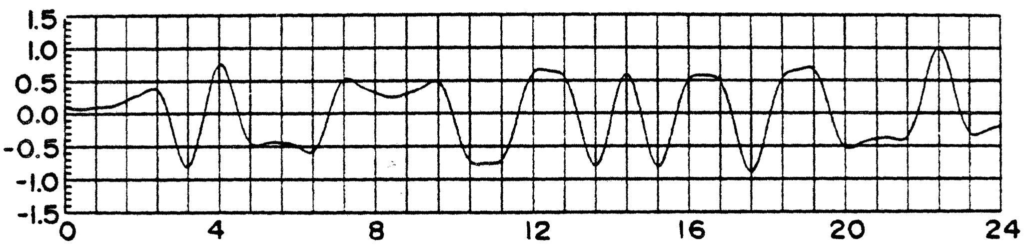
The write process is capable of recording high densities on magnetic tape, but the read-head resolution is limited by spacing and gap losses. High-density flux transitions may be written on tape, and then filtered out of the signal by the read process and bandlimiting read filters. For example, a pair of double density flux reversals can be inserted for 0 to improve tape performance, as discussed by Schneider [ref.42]. At first glance, it seems absurd to double the recording density, but the high-frequency transitions are severely attenuated in the equalized signal applied to the data detector. The beneficial result of these additional transitions is a boost in the high-frequency signal components and attenuation of the low frequencies. Thus, some of the necessary derivative-type equalization is accomplished in the write process. The reduction in the read-equalizer signal boost also reduces the amplification of high-frequency noise and improves SNR. Figure 36 on page 82 shows the use of this write equalizer on the analog channel shown in Figure 3 on page 5.



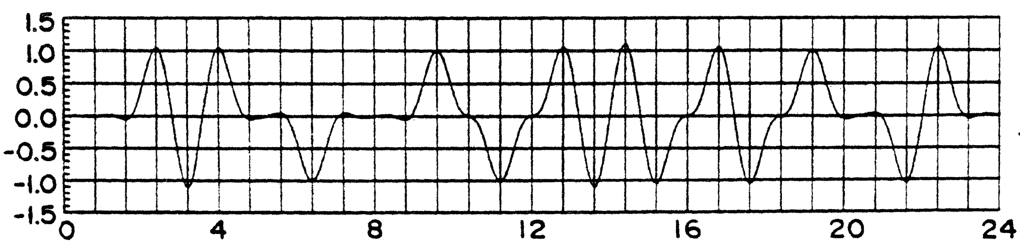
(a) Input signal (1250 transitions/mm)



(b) Write-equalized write signal (2500 fr/mm)



(c) Analog-channel output (write-equalized but not read-equalized)



(d) Read-equalized output signal

Figure 36 (Part 1 of 2). Digital Write Equalizer

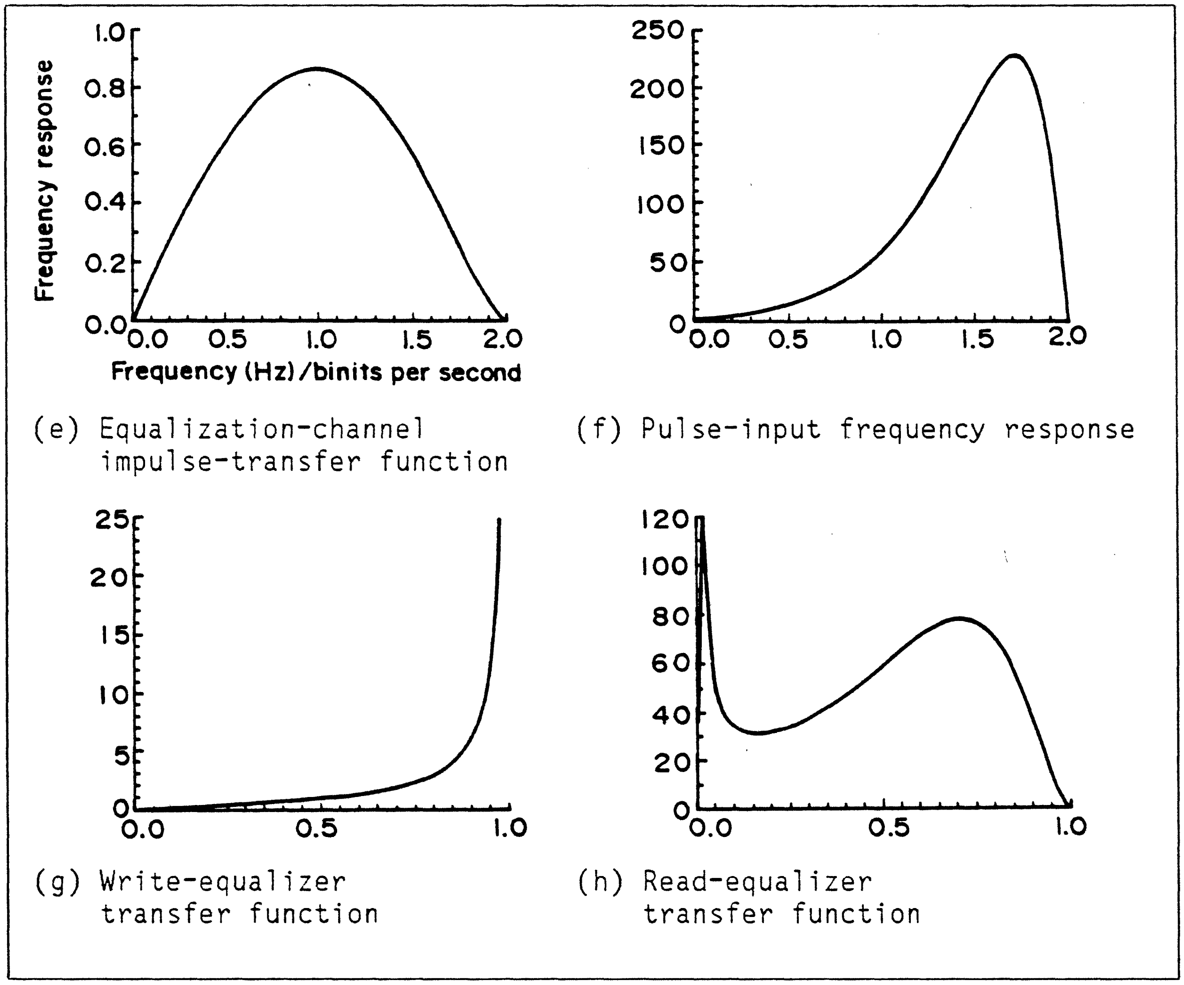


Figure 36 (Part 2 of 2). Digital Write Equalizer

Because this write equalizer produces a dc-balanced and run-length-limited signal, it appears that no modulation code is required. The run-length limitation comes from the extra 0 transitions that never reach the data detector. With random data input, it is possible for the output signal to remain at the zero level for prolonged periods of time; a clock cannot be recovered during this time. Therefore, a run-length-limited modulation code is used with this write equalizer to ensure the presence of sufficient clock synchronization pulses.

Schneider [ref.42] shows that this type of filter is linear, and how to calculate the transfer functions. Veillard [ref.43] describes a different implementation that achieves the same result. Many filters of this type are possible. For example, the dc-balanced ternary waveform shown in Figure 14j on page 28 is achieved with a 1-D write equalizer, which boosts the high-frequency response like a differentiator; of course, we have restricted the input to binary signals. Several similar schemes have been described without calling them digital write equalizers, as discussed by Behr and Blessum [ref.44], Heidecker [ref.45], and Schneider [refs.46,47].

Many write equalizers that have binary inputs and binary outputs may be described by a subclass of digital filters. They are specified completely and uniquely by the ratio of two polynomials in the z domain and they are linear. Equation 17 represents the general digital-filter write-equalizer transfer function H(z):

$$H(z) = (a_n z^{-n} + \ldots + a_2 z^{-2} + a_1 z^{-1} + a_0 z^0) \\ /(b_n z^{-n} + \ldots + b_2 z^{-2} + b_1 z^{-1} + b_0 z^0) \qquad (17)$$

The frequency-domain representation may be obtained by substituting the sine and cosine identities for the z variable. The trigonometric form of the resultant equation is not unique for it depends upon the applied sets of identities. The response, however, will be identical. The fact that transfer functions specified in trigonometric form are not unique in form and the digital filter nature of equalization has not been generally recognized and has hindered the development of digital filter applications in magnetic recording.

The write equalizer is typically only a portion of the overall equalizer filter. The product of the write and read equalizer and analog-channel transfer functions provides the desired equalization-channel frequency response, as described in the preceding pages. Dividing the total equalizing filter between the write and read sides of the analog channel can be beneficial at high recording densities. The difference between maximum and minimum wavelength read-signal amplitudes is reduced and SNR can be improved. Phase equalization is possible by proper selection of equalizer-pulse position or width. Although a higher-frequency write clock may be required for this type of write equalization, the read-clock frequency is not necessarily affected by the write equalizer.

As a final thought on equalizers, a variable read filter might compensate for analog-channel differences. A computer tape device often reads tapes that were written on other drives. Even new heads and tapes have some performance differences; after extensive use, this variability can be accentuated. This is not a serious problem at low densities, because little equalization is used. A high-recording-density tape drive that utilizes extensive signal equalization must accommodate the variations in analog-channel transfer function that will be encountered. Eventually, as recording densities increase, automatic or adaptive equalization techniques may be required. Automatic equalization uses a "training sequence" that is written on the tape at the beginning of the data. Prior to reading data, the equalizing filter is adjusted to optimize the training-sequence performance for the specific read and write heads and the tape that is being used. An adaptive equalizer continuously makes this adjustment from the recorded data. Obviously, unknown data is more difficult than a known sequence for determining the filter adjustment. Neither of these techniques is required unless fairly large differences in heads and tapes exist.

DATA RECOVERY

Modulation and equalization are designed to simplify or improve the reliability of the data-recovery process. The two major functions performed by the data-recovery circuits are:

- The clock frequency and phase must be recovered from the recorded signal.

- The equalized analog signal must be converted back to the original digital modulated signal.

The optimum technique for accomplishing these operations depends upon the modulation code and equalization used. The various equalizers described in the preceding pages have different performance characteristics, even under ideal conditions of no noise or dropouts. Figure 37 shows the maximum zero crossing or peak timing errors for ideal operation. Realistic operating conditions produce greater timing errors.

Waveform-restoration channel zero shift				
α	%ZS	β	%ZS	PR1 %ZS
1	0	1	2.5	0.1
0.9	1.5	1.5	1.1	
0.8	3.0	2.0	0	
0.7	5.0	2.5	0.7	
0.6	7.0	3.0	0.8	
0.5	10.3	3.5	0.6	
0.4	14.1	4.0	0	

Derivative channel peak shift							
d/dt α	%PS	d/dt β	%PS	d/df β	%PS	γ	%PS
1	2.8	1.7	4.5	2.0	6.6	1.4	4.2
0.95	2.4	2.0	2.8	2.5	4.2	1.6	2.9
0.90	2.2	2.3	1.2	3.0	1.4	1.8	1.4
0.85	1.8	2.6	0.7	3.5	1.6	2.0	0.1
0.80	2.1	2.9	2.5	4.0	4.5	2.2	1.5
0.75	3.0	3.2	4.2	4.5	7.3	2.9	3.0
0.70	4.4	3.5	5.9	5.0	10.0	2.6	4.5
Note: The half-bandwidth operation with d>0 modulation doubles these values.							

Figure 37. Comparison of Equalizer Timing Errors. The maximum percentage of deviation is from the ideal location.

Because of tape-velocity perturbations, the read-clock frequency is not constant. The clock is usually recovered by a voltage-controlled oscillator and phase-locked loop. The two major sources of clock timing pulses are signal peaks (for pulse-slimming and derivative equalization channels), and zero crossings (for waveform-restoration channels). Although these timing pulses are intermittent, they occur at the proper time to phase lock the clock. Ideally, the selection of modulation-code d and k values provides a large number of write-signal transitions for clock synchronization. The clock-recovery circuits must have sufficient bandwidth to follow the tape velocity variations. Allowable tape accelerations during typical full-speed operation are usually specified to permit proper clock design. The apparent acceleration at the clock may be twice the specified amount, if the velocity changes in opposite directions during read and write. This can occur in the case of longitudinal tape vibrations. Accurate clock timing is extremely important; a phase error decreases the available clocking window and increases the probability of error.

There are four types of data detection that are used with computer tape devices—sampling, zero crossing, peak, and integration detection. Sampling detection requires a clock to identify sample times. A properly equalized waveform-restoration channel produces accurate zero crossings for clock synchronization. If the signal is sampled at the center of each clock period, there should be no intersymbol interference. Sampling detection is usually used for partial response channels, but accurate and stable clock recovery can be difficult—particularly in the presence of signal impairments. Sampling is not really required for waveform-restoration equalization. Because the zero crossings are detected to synchronize the clock, they may be used also to identify the data; typically, each zero crossing represents 1 and each clock cell between zero crossings represents 0.

Derivative and pulse-slimming equalizers generally use a peak detector to define both clock and data. Some form of threshold is used to prevent erroneous detection of noise peaks. The threshold may be variable to allow for signal loss during a signal dropout. The peak detected signal is essentially the derivative of the waveform-restored signal, so waveform-restored zero crossings become derivative channel peaks.

An integration detector is used with a zero-crossing clock synchronization circuit; it integrates the signal over each clock cell to determine whether the average signal value is positive or negative. Noise tends to integrate to zero, but the signal has either a plus or minus integral. This is sometimes accomplished with a pair of integrators to allow for integrator sample and reset time.

Data recovery would be simple if there were no signal impairments. The recovered clock is not precise, particularly during the most critical periods of signal dropouts. Conditions are rarely ideal. Tape velocity fluctuates, and the signal is degraded by noise and intersymbol interference. To add to the problem, dropouts cause the signal amplitude to decrease and intersymbol interference to vary, further reducing the detection margin. Figure 38 on page 87 shows the maximum timing variation in zero crossings or peaks that results from a 50% increase in effective separation (+200 nm) for the channel shown in Figure 3 on page 5, if we assume proper equalization for typical conditions.

Equalizer	Bandwidth	Timing error (% of binit)	
		Ideal	+50% EMS
Waveform-restoration channels			
$\alpha = 1$ or $\beta = 2$	Full	0	12.4
$\beta = 2$	Half	2.3	24.7
PR1	Half	0.1	Failure
Derivative channels			
d/dt $\alpha = 0.85$	Full	1.8	7.1
d/dt $\beta = 2.6$	Full	0.7	8.2
d/df $\beta = 3.0$	Full	1.4	5.9
$\gamma = 2.0$	Full	0.1	7.6
$\gamma = 2.0$	Half	14.2	Failure
Failure = over 50% data errors with no noise			

Figure 38. Comparison of Equalizer Timing Sensitivities

The half-bandwidth channels are sensitive to channel perturbations such as tape speed or equalization variations. None of them will operate well with a 50% change in magnetic separation, which is a common occurrence with high-density tape devices. A well-designed system can tolerate the noise and distortion until a severe dropout occurs.

The error-correction circuits are more effective if a pointer is available to indicate the error location. An analog pointer is sometimes generated when the signal amplitude drops below a fixed threshold. This is an indication that the data is questionable at this point. Other pointers are available from specific signal requirements such as phase. Illegal modulation code words can provide pointers. Because binary transitions must alternate, two derivative channel peaks of the same polarity are obviously incorrect. Various pointers have been used to indicate probable error locations.

The detected signal must be demodulated and decoded. A random binit in error can propagate to several bits after demodulation of some codes. A parallel-track machine will produce read signals that are not in synchronism with each other; therefore, they must be deskewed. A skew buffer accepts the multiple track signals and resynchronizes them at the output. During a deep dropout, noise and distortion can cause erroneous clock-synchronization pulses. Sometimes these pulses are disconnected from the clock during dropouts, allowing the clock to coast. A periodic resynchronization character permits the system to regain synchronism after a severe dropout. These techniques allow the data to be reorganized into the original format, but it will contain errors. Proper error control techniques must also be included.

SUMMARY

The design of equalization for computer tape recorders requires that equalization and detection be jointly considered. Equalization, in turn, may be separated into read and write equalization. The many variables in a recording system, which include some unknown and unrestrained variables, and the changing nature of the head and the tape do not allow a unique solution to the design problem. Advances in instrumentation and modelling permit greater insight and offer the ability to sift through the seemingly limitless combinations much more rapidly than in the past. However, it is still in the laboratory in which the candidate systems are proven.

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