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Waveform Encoding Techniques Based on Differential and Adaptive Quantizing

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Dedicated to my Late Mother and my Father

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ABSTRACT

Waveform encoding technique based on pulse code modulation (PCM) and delta modulation (DM) allow the improvement signal-to- noise ratio, perform time division multiplexing (TDM) of signals from different sources over a single communication channel and provide a secure communication.

However, based on uniform quantizing charachteristic with fixed step-size approximation yield redundant of information in PCM and granular and slope-over load distortions in DM.

This thesis aims at analysing the method of adaptive PCM and DM techniques in which nonuniform-quantizing characteristics with controlled step-sizes are used.

The control of step-sizes of the quantizing characteristics is performed in accordance with the rate of variation of the input signal.

The suggested approach described within this thesis allows the decrease of redundant information in conventional PCM and granular and slope-over load distortions in DM.

CONTENTS

INTRODUCTION

The most widely used pulse modulation technique in the telecommunications industry is pulse code modulation (PCM) and delta modulation (DM). Currently PCM is the preferred method of transmission for public switched telephone network (PSTN).

PCM and DM are the methods of serially digital transmitting an analog signal.

PCM signal itself is a succession of discrete numerically encoded binary values derived from digitizing the analog signal.

DM is a simplified version of PCM in which the analog input is converted to a serial data stream of 1 and 0.

The objective of this thesis is a performance analysis of the PCM and DM systems and design adaptive time-varying step-size approximation strategy to improve signal-to-noise ratio of considered systems.

Chapter one presents the wave coding techniques, basic elements of pulse code modulation, sampling, quantizing, encoding, regeneration, decoding, filtering, multiplexing, synchronization, noise consideration in PCM systems, error threshold, virtues, limitations, and modifications of PCM, quantization noise and signal-to-noise ratio and idle channel noise.

Chapter two presents differential PCM techniques, processing gain, multiplexing of the PCM signals, digital multiplexers, T1 system, M12 multiplexer and light-wave transmission.

Chapter three discusses illustration of DM, DM system transmitter and receiver, quantization error in DM like slope overload and granular noise distortions and illustration of this quantization error and delta-sigma modulation system transmitter and receiver.

Chapter four describes adaptive pulse code modulation techniques, adaptive differential pulse code modulation, adaptive sub-band coding, subjective quality and adaptive delta modulation(ADM).

Chapter five is devoted to practical implementations, Matlab implementation, design of hardware layout and the result of investigation, block diagram of adaptive delta pulse code modulation (ADPCM) and its input and output and block diagram of continuously variable slope delta modulation (CVSDM) and its input and output.

In conclusion the important results obtained by the author of this thesis and the practical recommendations are provided.

1. WAVEFORM CODING TECHNIQUES

1.1 Overview

Pulse code modulation (PCM) was devised in 1937 at the Paris laboratories of AT &T by Alex H. Reeves. He conducted several successful transmissions across the English Channel using pulse width-band modulation (PWM), pulse amplitude modulation (PAM) and pulse position modulation (PPM). At that time, the circuitry involved was complex and expensive, so it was not until semiconductor industry evolved in 1960 that PCM becomes more prevalent. Almost all of the newer long-distance telephone lines carry voice signals in digital format using PCM [8].

Pulse code modulation (PCM) is one of the methods for digital transmission of analog signals. In this method of signal coding, the message signal is sampled and the amplitude of each sample is rounded off (approximated) to the nearest one of a finite set of discrete levels, so that both time and amplitude are represented in discrete form. This allows the message to be transmitted by means of a digital (coded) waveform, thereby distinguishing pulse code modulation from all analog modulation techniques [3].

In conceptual terms, PCM is simple to understand. Moreover, it was the first method to be developed for the digital coding of waveforms. Indeed, it is the most applied of all digital coding systems in use today. The use of digital representation of analog signals (e.g. voice, video) offers the following advantages:

1. Ruggedness to channel noise and interference.
2. Efficient regeneration of the coded signal along the transmission path.
3. Efficient exchange of increase channel bandwidth for improved signal-to-noise ratio, obeying all exponential rules.
4. A uniform format for the transmission of different kinds of base-band signals; hence their integration with other forms of digital data in a common network
5. Comparative ease with which message sources may be dropped or reinsert in a time-division multiplex system.
6. Secure communication through the use of special modulation schemes or encryption.

These advantages, however, are attained at the cost of increased transmission bandwidth requirement and increased system complexity. With the increasing availability of wide-band communication channels, coupled with the emergence of the requisite device technology, the use of PCM has indeed become a practical reality.

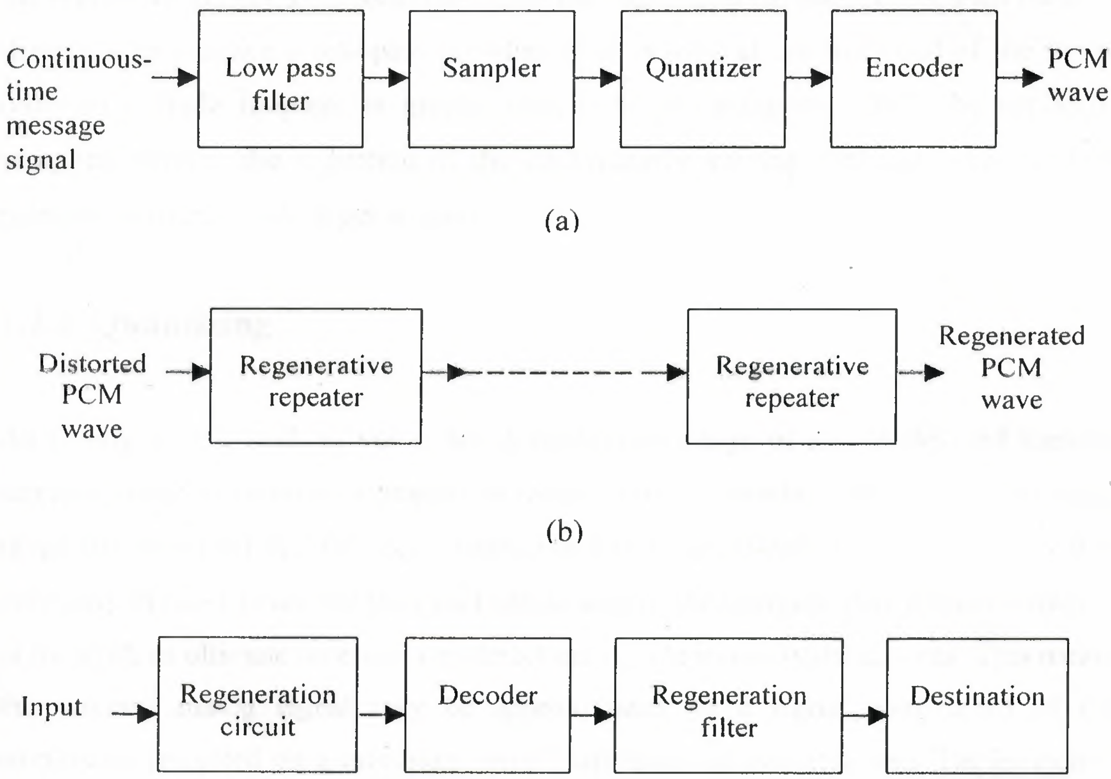
PCM belongs to a class of signal coders known as waveform coders, in which an analog signal is approximated by mimicking the amplitude-versus-time waveform; hence, the name. Waveform coders are (in principle) designed to be signal-independent. As such, they are basically different from source (e.g., linear predictive vocoders), which rely on a parameterization of the analog signal in accordance with an appropriate model for the generation of the signal.

1.2 Basic Elements of Pulse Code Modulation

Pulse code modulation systems are complex in that the message signal is subjected to a large number of operations. The essential operations in the transmitter of a PCM system are sampling, quantizing and encoding, as shown in the top part of figure 1.1. The sampling, quantizing, and encoding operations are usually performed in the same circuit, which is called an analog-to-digital converter. Regeneration of impaired signals occurs at intermediate points along the transmission path (channel) as indicated in the middle part of figure 1.1. At the receiver, the essential operations consist of one last stage of regeneration followed by decoding, then demodulation of the train of quantized samples, as in the bottom part of figure 1.1. The operations of decoding and reconstruction are usually performed in the same circuit, called a digital-to-analog converter. When time-division multiplexing is used, it becomes necessary to synchronize the receiver to the transmitter for the overall system to operate satisfactorily.

It is noteworthy that pulse code modulation is not modulation in the conventional sense.The term "modulation" usually refers to the variation of some characteristic of a carrier wave in accordance with an information-bearing signal. The only part of pulse code modulation that conforms to this definition is sampling. The subsequent use of quantization, which is basic to pulse code modulation, introduces a signal distortion that has no counterpart in conventional modulation.

In the sequel, the basic signal-processing operations involved in PCM are considered, one by one [12].



(c)

Figure 1.1 Basic Elements of a PCM System.

(a) Transmitter; (b) Transmission Path; (c) Receiver.

1.2.1 Sampling

The incoming message wave is sampled with a train of narrow rectangular pulses so as to closely approximate the instantaneous sampling process. In order to ensure perfect reconstruction of the message at the receiver, the sampling rate must be greater than twice the highest frequency component ω of the message wave (in accordance with the sampling theorem). In practice a low-pass pre-alias filter is used at the front end of the sampler in order to exclude frequencies greater than ω before sampling. Thus, the application of sampling permits the reduction of the continuously varying message wave to a limited number of discrete values per second.

1.2.2 Quantizing

An analog signal, such as voice, has a continuous range of amplitudes and therefore its samples cover a continuous amplitude range. In other words, within the finite amplitude range of the signal we find an infinite number of amplitude levels. However, it is not necessary in fact to transmit the exact amplitudes of the samples. Any human sense (the ear or the eye), as ultimate receiver, can detect only finite intensity differences. This means that the original analog signal may be approximated by a signal constructed of discrete amplitudes (selected on a minimum error basis from an available set). The existence of a finite number of discrete amplitude levels is a basic condition of PCM. Clearly, if we assign the discrete amplitude levels with sufficiently close spacing, we may make the approximated signal practically indistinguishable from the original analog signal[12].

The conversion of an analog (continuous) sample of the signal into a digital (discrete) form is called the quantizing process. Graphically, the quantizing process means that a straight line representing the relation between the input and the output of a linear analog system is replaced by a transfer characteristic that is staircase-like in appearance. Figure 1.2a depicts one such characteristic.

The quantizing process has a two-fold effect:

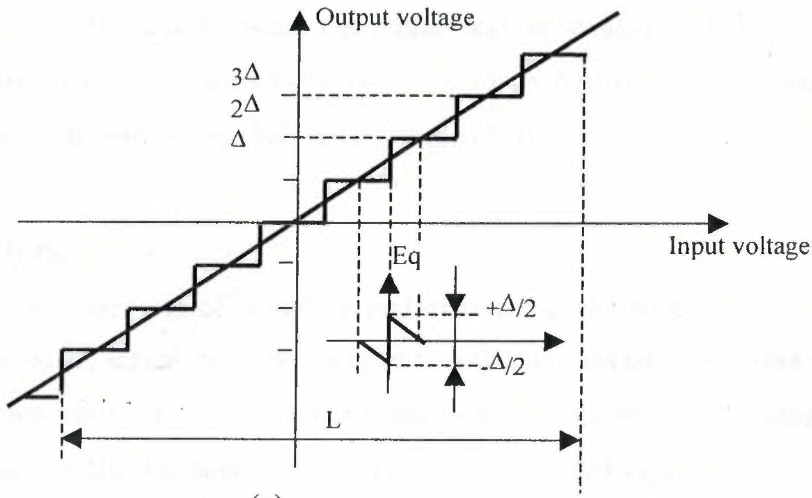
1.The peak-to-peak range of input sample values is subdivided into a finite set of decision levels or decision thresholds that are aligned with the "risers" of the staircase.

2.The output is assigned a discrete value selected from a finite set of representation levels or reconstruction values that are aligned with the "treads" of the staircase. In the case of a uniform quantizer, characterized as in figure 1.2a, the separation between the decision thresholds and the separation between the representation levels of the quantizer have a common value called the step size. According to the staircase-like transfer characteristic of figure 1.2a, the decision thresholds of the quantizer are located at $\pm\Delta/2$, $\pm 3\Delta/2$, $\pm 5\Delta/2$, ..., and the representation levels are located at 0, $\pm\Delta$, $\pm 2\Delta$, ..., where Δ is the stepsize. A uniform quantizer characterized in this way is referred to as a symmetric quantizer of the mid-tread type, because the origin lies in the middle of a tread of the staircase.

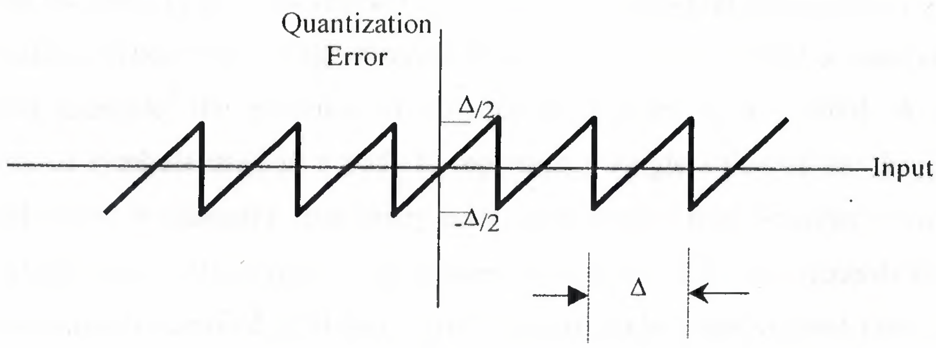
Figure 1.2b shows another staircase-like transfer characteristic, in which the decision thresholds of the quantizer are located at 0, $\pm\Delta$, $\pm 2\Delta$, ..., and the representation levels are located at $\pm\Delta/2$, $\pm 3\Delta/2$, $\pm 5\Delta/2$, ..., where Δ is again the step size. A uniform quantizer having this second characteristic is referred to as a symmetric quantizer of the mid-riser type, because in this case the origin lies in the middle of a riser of the staircase.

A quantizer of the mid-tread or mid-riser type, as defined, is memoryless in that the quantizer output is determined only by the value of a corresponding input sample, independently of earlier (or later) analog samples applied to the input (A memoryless quantizer is inefficient if the input sample are statistically dependent; such dependencies would have to be removed either prior to quantizing or as part of the quantizing process). The memoryless quantizer is the simplest and most often used quantizer.

In the transfer characteristics of figure 1.2a, we have included a parameter labeled the overload level, the absolute value of which is one half of the peak-to-peak range of input sample values [8]. Moreover, the number of intervals into which the peak-to-peak



(a)



(b)

Figure 1.2 (a) Transfer Characteristic of Quantizer of Mid-riser Type;
(b) Variation of The Quantization Error With Input.

excursion is divided, or equivalently the number of representation levels, is equal to twice the absolute value of the overload level divided by the step size. Accordingly, for an analog input sample that lies anywhere inside an interval of either transfer characteristic, the quantizer produces a discrete output equal to the mid-value of the pair of decision thresholds in question. In so doing, however, a quantization error is introduced, the value of which equals the difference between the output and input values of the quantizer. We see that the maximum instantaneous value of this error is half of one step size, and the total range of variation is from minus half a step to plus half a step.

1.2.3 Encoding

In combining the processes of sampling and quantizing, the specification of a continuous message (base-band) signal becomes limited to a discrete set of values, but not in the form best suited to transmission over a line or radio path. To exploit the advantages of sampling and quantizing for the purpose of making the transmitted signal more robust to noise, interference and other channel degradations, we require the use of an encoding process to translate the discrete set of sample values to a more appropriate form of signal. Any plan for representing each of this discrete set of values as a particular arrangement of discrete events is called a code. One of the discrete events in a code is called a code element or symbol. For example, the presence or absence of a pulse is a symbol. A particular arrangement of symbols used in a code to represent a single value of the discrete set is called a code word or character. In a binary code, each symbol may be either of two distinct values or kinds, such as the presence or absence of a pulse. The two symbols of a binary code are customarily denoted as 0 and 1. In a ternary code, each symbol may be one of three distinct values or kinds, and so on for other codes. However, the maximum advantage over the effects of noise in a transmission medium is obtained by using a binary code, because a binary symbol withstands a relatively high level of noise and is easy to regenerate. Suppose that, in a binary code, each code word consists of R bits: the bit is an acronym for binary digit; thus R denotes the number of bits per sample. Then, using such a code, we may represent a total of 2^R distinct numbers. For example, a sample quantized into one of 256 levels may be represented by an 8-bit code word.

There are several ways of establishing a one-to-one correspondence between representation levels and code words. A convenient method is to express ordinal number of the representation level as a binary number. In the binary number system, each digit has a place-value that is a power of 2.

There are several line codes that can be used for the electrical representation of binary symbols 1 and 0, as described here:

1. On-off signaling, in which symbol 1 is represented by transmitting a pulse of constant amplitude for the duration of the symbol, and symbol 0 is represented by switching off the pulse, as in figure 1.3 a.
2. Nonreturn-to-zero (NRZ) signaling, in which symbols 1 and 0 are represented by pulses of equal positive and negative amplitudes, as illustrated in figure 1.3 b.
3. Return-to-zero (RZ) signaling, in which symbol 1 is represented by a positive rectangular pulse of half-symbol width, and symbol 0 is represented by transmitting no pulse, as illustrated in figure 1.3 c.
4. Bipolar return-to-zero (BRZ) signaling, which uses three amplitude levels as, indicated in figure 1.3 d. Specifically, positive and negative pulses of' equal amplitude are used alternately for symbol 1, and no pulse is always used for symbol 0. A useful property of the BRZ signaling is that the power spectrum of the transmitted signal has no dc component and relatively insignificant low-frequency components for the case when symbols 1 and 0 occur with equal probability.
5. Split-phase (Manchester code), which is illustrated in figure 1.3e. In this method of signaling, symbol l is represented by a positive pulse followed by a negative pulse, with both pulses being of equal amplitude and half-symbol width. For Symbol 0, the polarities of these two pulses are reversed. The Manchester code suppresses the dc component and has relatively insignificant low-frequency components, regardless of the signal statistics. This property is essential in some applications.
6. Differential encoding; in which the information is encoded in terms of signal transitions, as illustrated in figure 1.3f. In the example of the binary PCM signal shown here, a transition is used to designate symbol 0, while no transition is used to

designate symbol 1. It is apparent that a differentially encoded signal may be inverted without affecting its interpretation. The original binary information is recovered by comparing the polarity of adjacent symbols to establish whether or not a transition has occurred.

The waveforms shown in figures. 1.3a to 1.3f are for the binary data stream 01101001 [3].

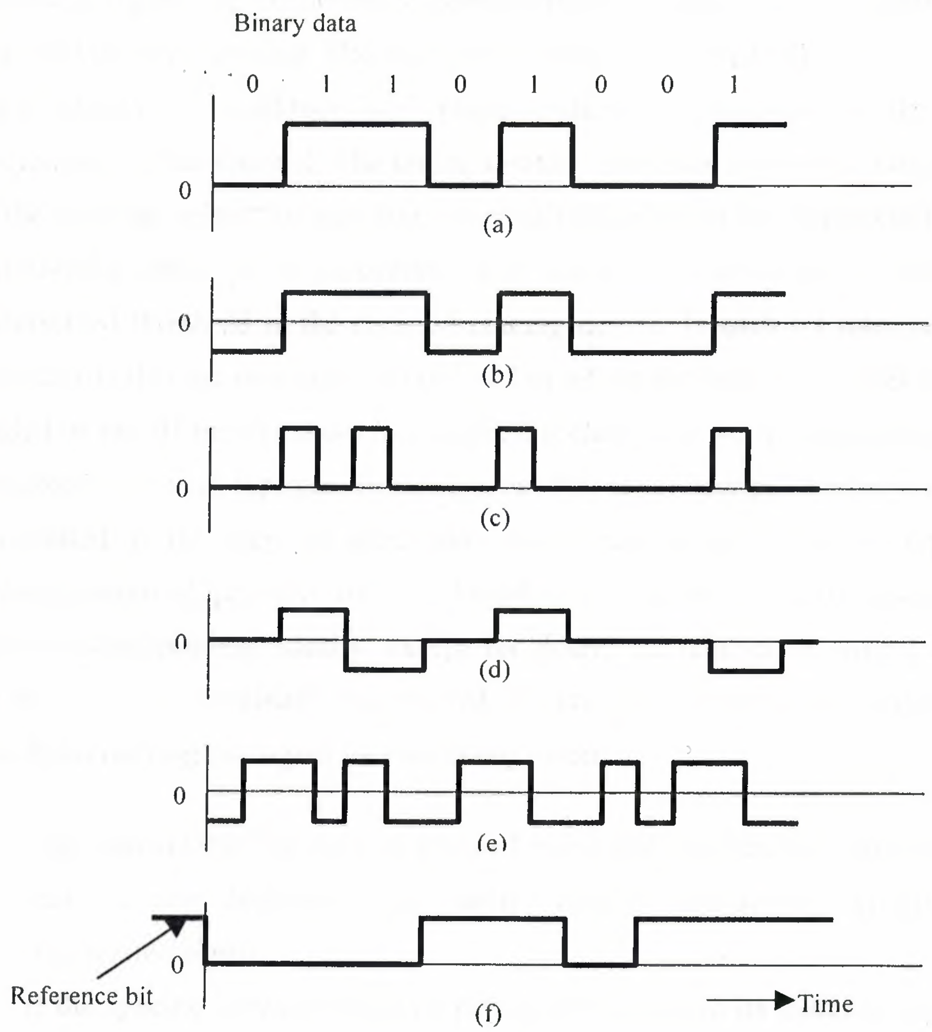


Figure 1.3 Electrical Representations of Binary Data.

(a) On-off Signaling; (b) Nonreturn-to-Zero Signaling; (c) Return-to-Zero Signaling; (d) Bipolar Signaling; (e) Split Phase or Manchester Code; (f) Differential Encoding.

1.2.4 Regeneration

The most important feature of PCM systems lies in the ability to control the effects of distortion and noise produced by transmitting a PCM signal through a channel. This capability is accomplished by reconstructing the PCM signal by means of a chain of regenerative repeaters located at sufficiently close spacing along the transmission route. As illustrated in figure 1.4, a regenerative repeater performs three basic functions: equalization, timing, and decision-making. The equalizer shapes the received pulses so as to compensate for the effects of amplitude and phase distortions produced by the transmission characteristics of the channel. The timing circuitry provides a periodic pulse train, derived from the received pulses, for sampling the equalized pulses at the instants of time where the signal-to-noise ratio is a maximum. The sample so extracted is compared to a predetermined threshold in the decision-making device. In each bit interval a decision is then made whether the received symbol is a 1 or a 0 on the basis of whether the threshold is exceeded or not. If the threshold is exceeded, a clean new pulse-representing symbol 1 is transmitted to the next repeater. Otherwise, another clean new pulse representing symbol 0 is transmitted. In this way, the accumulation of distortion and noise in a repeater span is completely removed, provided that the disturbance is not too large to cause an error in the decision-making process. Ideally, except for delay, the regenerated signal is exactly the same as the signal originally transmitted. In practice, however, the regenerated signal departs from the original signal for two main reasons:

1. The unavoidable presence of channel noise and interference cause the repeater to make wrong decisions occasionally, thereby introducing bit errors into the regenerated signal.
2. If the spacing between received pulses deviates from its assigned value, a jitter is introduced into the regenerated pulse position, thereby causing distortion.

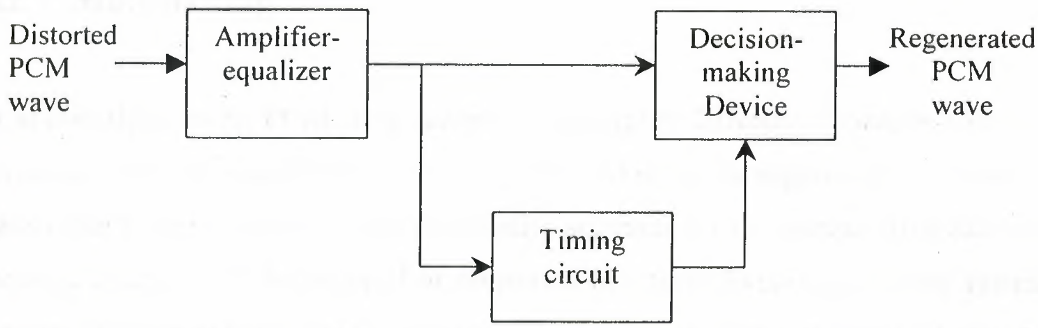


Figure 1.4 Block Diagram of A Regenerative Repeater.

1.2.5 Decoding

The first operation in the receiver is to regenerate (i.e., reshape and clean up) the received pulses one last time. These clean pulses are then regrouped into code words and decoded (i.e., mapped back) into a quantized PAM signal. The decoding process involves generating a pulse the amplitude of which is the linear sum of all the pulses in the code word, with each pulse being weighted by its place value ($2^0, 2^1, 2^2, 2^3, \ldots, 2^{R-1}$) in the code, where R is the number of bits per sample.

1.2.6 Filtering

The final operation in the receiver is to recover the message signal wave by passing the decoder output through a low-pass reconstruction filter whose cutoff frequency is equal to the message bandwidth, assuming that the transmission path is error free, the recovered signal includes no noise with the exception of the initial distortion introduced by the quantization process.

1.2.7 Multiplexing

In applications using PCM, it is natural to multiplex different messages sources by time division, whereby each source keeps its individuality throughout the journey from the transmitter to the receiver. This individuality accounts for the comparative ease with which message sources may be dropped or reinserted in a time-division multiplex system. As the number of independent message sources is increased, the time interval that may be allotted to each source has to be reduced, since all of them must be accommodated into a time interval equal to the reciprocal of the sampling rate. This in turn means that the allowable duration of a code word representing a single sample is reduced. However, pulses tend to become more difficult to generate and to transmit as their duration is reduced. Furthermore, if the pulses become too short, impairments in the transmission medium begin to interfere with the proper operation of the system. Accordingly, in practice, it is necessary to restrict the number of independent message sources that can be included within a time-division group.

1.2.8 Synchronization

For a PCM system with time-division multiplexing to operate satisfactorily, it is necessary that the timing operations at the receiver, except for the time lost in transmission and regenerative repeating, follow closely the corresponding operations at the transmitter. In a general way, this amounts to requiring a local clock at the receiver to keep the same time as a distant standard clock at the transmitter, except that the local clock is somewhat slower by an amount corresponding to the time required to transport the message signals from the transmitter to the receiver. One possible procedure to synchronize the transmitter and receiver clocks is to set aside a code element or pulse at the end of a frame (consisting of a code word derived from each of the independent message sources in succession) and to transmit this pulse every other frame only. In such a case, the receiver includes a circuit that would search for the pattern of 1s and 0s alternating at half the frame rate, and thereby establish synchronization between the transmitter and receiver.

When the transmission path is interrupted, it is highly unlikely that transmitter and receiver clocks will continue to indicate the same time for long. Accordingly, in carrying out a synchronization process, we must set up an orderly procedure for detecting the synchronizing pulse. The procedure consists of observing the code elements one by one until the synchronizing pulse is detected. That is, after observing a particular code element long enough to establish the absence of the synchronizing pulse, the receiver clock is set back by one code element and the next code element is observed. This searching process is repeated until the synchronizing pulse is detected. clearly, the time required for synchronization depends on the epoch at which proper transmission is reestablished [3].

1.3 Noise Considerations in PCM Systems

The performance of a PCM system is influenced by two major sources of noise:

1. Channel noise, which is introduced anywhere between the transmitter output and the receiver input. Channel noise is always present, once the equipment is switched on.
2. Quantization noise, which is introduced in the transmitter and is carried all the way along to the receiver output. Unlike channel noise, quantization noise is signal-dependent in the sense that it disappears when the message signal is switched off.

Naturally, these two sources of noise appear simultaneously once the PCM system is in operation. However, the traditional practice is to consider them separately, so that we may develop insight into their individual effects on the system performance.

The main effect of channel noise is to introduce bit errors into the received signal. In the case of a binary PCM system, the presence of a bit error causes symbol 1 to be mistaken for symbol 0, or vice versa. Clearly, the more frequently bit errors occur, the more dissimilar the receiver output becomes compared to the original message signal. The fidelity of information transmission by PCM in the presence of channel noise may be measured in terms of the average probability of symbol error, which is defined as the probability that the reconstructed symbol at the receiver output differs from the transmitted binary symbol, on the average. The average probability of symbol error, also referred to as the error rate, assumes that all the bits in the received binary wave are of equal importance. When, however, there is more interest in reconstructing the analog waveform of the original message signal, different symbol errors may need to be weighted differently.

To optimize system performance in the presence of channel noise, we need to minimize the average probability of symbol error. For this evaluation, it is customary to model the channel noise, originating at the front end of the receiver, as additive, white, and Gaussian. The effect of channel noise can be made practically negligible by ensuring the use of an adequate signal energy-to-noise density ratio through the provision of proper spacing between the regenerative repeaters in the PCM system. In such a situation, the performance of the PCM system is essentially limited by quantization noise acting alone.

It can be made negligibly small through the use of an adequate number of representation levels in the quantizer and the selection of a companding strategy matched to the characteristics of the type of message signal being transmitted. We thus find that the use of PCM offers the possibility of building a communication system that is rugged with respect to channel noise on a scale that is beyond the capability of any codeword modulation or analog pulse modulation system [12].

1.3.1 Error Threshold

It suffices to say that the average probability of symbol error in a binary encoded PCM receiver due to additive white Gaussian noise depends solely on E_b/N_o, the ratio of the transmitted signal energy per bit, E_b to the noise spectral density, N_o. Note that the ratio E_b/N_o is dimensionless even though the quantities E_b and N_o have different physical meaning. In table 1.1 a summary of this dependence for the case of a binary PCM system using nonreturn-to-zero signaling is presented. The results presented in the last column of the table assume a bit rate of 10^5 b/s.

Table 1.1 Influence of E_b/N_o on The Probability of Error

E_b/N_o	Probability of Error P_e	For a Bit Rate of 10^5 b/s This is About One Error Every
4.3 dB	10^{-2}	10^{-3} second
8.4	10^{-4}	10^{-1} second
10.6	10^{-6}	10 second
12.0	10^{-8}	20 minutes
13.0	10^{-10}	1 day
14.0	10^{-12}	3 months

From table 1.1 it is clear that there is an error threshold (at about 11 dB). For E_b/N_o below the error threshold the receiver performance involves significant numbers of errors, and above it the effect of channel noise is practically negligible. In other words, provided that the ratio E_b/N_o exceeds the error threshold, channel noise has virtually no effect on the receiver performance, which is precisely the goal of PCM. When, however, E_b/N_o drops below the error threshold, there is a sharp increase in the rate at which errors occur in the receiver. Because decision errors result in the construction of incorrect code words, we find that when the errors are frequent, the reconstructed message at the receiver output bears little resemblance to the original message.

Comparing the figure of 11 dB for the error threshold in a PCM system using NRZ signaling with the 60-70 dB required for high-quality transmission of speech using amplitude modulation, we see that PCM requires much less power, even though the average noise power in the PCM system is increased by the R-fold increase in bandwidth, where R is the number of bits in a code word (i.e., bits per sample).

In most transmission systems, the effects of noise and distortion from the individual links accumulate. For a given quality of overall transmission, the longer the physical separation between the transmitter and the receiver, the more severe are the requirements on each link in the system. In a PCM system, however, because the signal can be regenerated as often as necessary, the effects of amplitude, phase, and nonlinear distortions in one link (if not too severe) have practically no effect on the regenerated input signal to the next link. We have also seen that the effect of channel noise can be made practically negligible by using a ratio E_b/N_o above threshold. For all practical purposes, then, the transmission requirements for a PCM link are almost independent of the physical length of the communication channel.

Another important characteristic of a PCM system is its ruggedness to interference, caused by stray impulses or cross talk. The combined presence of channel noise and interference causes the error threshold necessary for satisfactory operation of the PCM system to increase. If an adequate margin over the error threshold is provided in the first place, however, the system can withstand the presence of relatively large amounts of interference. In other words, a PCM system is quite rugged.

1.4 Virtues, Limitations, and Modifications of PCM

In a generic sense, PCM has emerged as the most favored modulation scheme for the transmission of analog information-bearing signals such as voice and video signals. The advantages of PCM may all be traced to the use of coded pulses for the digital representation of analog signals, a feature that distinguishes it from all other analog methods of modulation [12].

Although the use of PCM involves many complex operations, today they can all be implemented in a cost-effective fashion using commercially available and/or custom-made very-large-scale integrated (VLSI) chips. In other words, the requisite device technology for the implementation of a PCM system is already in place. Moreover, with continuing improvements in VLSI technology, we are likely to see an ever-expanding use of PCM for the transmission of analog signals.

If, however, the simplicity of implementation is a necessary requirement, then we may use delta modulation as an alternative to pulse code modulation. In delta modulation, the base-band signal is intentionally "over sampled" in order to permit the use of a simple quantizing strategy for constructing the encoded signal.

Turning next to the issue of bandwidth, we do recognize that the increased bandwidth requirement of PCM may have been a reason for justifiable concern in the past. Today, however, it is of no real concern for two different reasons. First, the increasing availability of wide-band communication channels means that bandwidth is no longer a system constraint in the traditional way it used to be. Liberation from the bandwidth constraint has been made possible by the deployment of communication satellites for broadcasting and the ever-increasing use of fiber optics for networking.

The second reason is that through the use of sophisticated data compression techniques, it is indeed possible to remove the redundancy inherently present in a PCM signal and thereby reduce the bit rate of the transmitted data without serious degradation in system performance. In effect, increased processing complexity (and therefore increased cost of implementation) is traded off for a reduced bit rate and therefore reduced bandwidth requirement. A major motivation for bit reduction is for secure communication over radio channels that are inherently of low capacity.

1.5 Quantization Noise and Signal-to-Noise Ratio

Quantization noise is produced in the transmitter end of a PCM system by rounding off sample values of an analog base-band signal to the nearest permissible representation levels of the quantizer. As such, quantization noise differs from channel noise in that it is signal dependent in this section; we evaluate statistical characteristics of quantization noise by making certain assumptions that permit a mathematical analysis of the problem [2].

Consider a memoryless quantizer that is both uniform and symmetric, with a total of L representation levels. Let x denote the quantizer input, and y denote the quantizer output. These two variables are related by the transfer characteristic of the quantizer, as shown by

$$y = Q(x) \tag{1.1}$$

which is a staircase function that befits the type of mid-tread or mid-riser quantizer of interest. Suppose that the input x lies inside the interval

$$P_k = \{x_k < x \leq x_{k+1}\} \qquad k = 1, 2, \ldots, L \tag{1.2}$$

where x_k and x_{k+1} are decision thresholds of the interval P_k as depicted in figure 1.5. Correspondingly, the quantizer output y takes on a discrete value y_k, k =1, 2, . . . , L. That is,

$$y = y_k, \qquad \text{if x lies in the interval } P_k \tag{1.3}$$

Let q denote the quantization error, with values in the range $-\Delta/2 \leq q \leq \Delta/2$. We may then write

$$y_k = x + q, \qquad \text{if x lies in the interval } P_k \tag{1.4}$$

We assume that the quantizer input x is the sample value of a random variable X of zero mean and variance σ^2_X. When the quantization is fine enough (say, the number of representation levels L is greater than 64), the distortion produced by quantization noise affects the performance of a PCM system as though it were an additive independent source of noise with zero mean and mean-square value determined by the quantizer step size Δ. The reason for this is that the power spectral density of the quantization noise in the receiver output is practically independent of that of the base-band signal over a wide range of input signal amplitudes. Furthermore, for a base-band signal of a root mean-square value that is large compared to a quantum step, it is found that the power spectral density of the

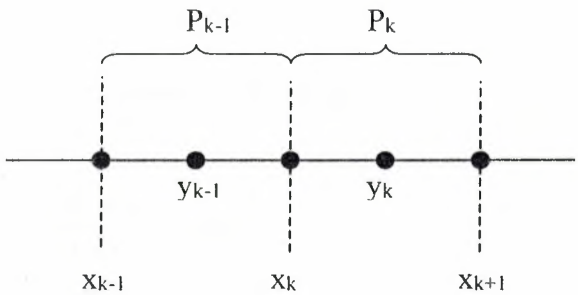


Figure 1.5 Decision Thresholds of The Quantizer.

quantization noise has a large bandwidth compared with the signal bandwidth. Thus, with the quantization noise uniformly distributed throughout the signal band, its interfering effect on a signal is similar to that of thermal noise.

Let the random variable Q denote the quantization error, and let q denote its sample value. (The symbol used for this random variable should not be confused with that for the transfer characteristic of the quantizer.) Lacking information to the contrary, we assume that the random variable Q is uniformly distributed over the possible range $-\Delta/2$ to $\Delta/2$, as shown by

$$f_Q(q)=\begin{cases}\dfrac{1}{\Delta} & -\dfrac{\Delta}{2}\leq q\leq\dfrac{\Delta}{2}\\ 0 & otherwise\end{cases} \tag{1.5}$$

where $f_Q(q)$ is the probability density function of the quantization error. For this to be justifiable, we must ensure that the incoming signal does not overload the quantizer. Then the mean of the quantization error is zero, and its variance σ^2_Q is the same as the mean-square value, as shown by

$$\sigma^2_Q = \mathrm{E}\left[Q^2\right]$$
$$= \int_{-\infty}^{\infty} q^2 f_Q(q)dq \qquad (1.6)$$

Substituting equation (1.5) in equation (1.6), we get

$$\sigma^2_Q = \frac{1}{\Delta}\int_{-\Delta/2}^{\Delta/2} q^2 dq$$
$$= \frac{\Delta^2}{12} \qquad (1.7)$$

Thus, the variance of the quantization noise, produced by a uniform quantizer, grows as the square of the step size. This is perhaps the most often used result in quantization.

Let the variance of the base-band signal x(t) at the quantizer input be denoted by σ^2_X. When the base-band signal is reconstructed at the receiver output, we obtain the original signal plus quantization noise. We may therefore define an output signal-to-quantization noise ratio (SNR) as

$$(SNR)_0 = \frac{\sigma^2_X}{\sigma^2_Q} = \frac{\sigma^2_X}{\Delta^2/12} \qquad (1.8)$$

Clearly, the smaller we make the step size Δ, the larger will the SNR be.

Equation (1.8) defines the performance of a quantizing noise-limited PCM system that uses a uniform quantizer.

1.5.1 Idle Channel Noise

A discussion of noise in PCM systems would be incomplete without a description of idle channel noise. As the name implies, idle channel noise is the coding noise measured at the receiver output with zero transmitter input. The zero-input condition arises, for example, during silences in speech. The average power of this form of noise depends on the type of quantizer used. In a quantizer of the mid-riser type, as in figure 1.2a, zero input amplitude is encoded into one of the two innermost representation levels $\pm\Delta/2$. Assuming that these two representation levels are equiprobable, the idle channel noise for mid-riser quantizer has zero mean and an average power of $\Delta^2/4$. On the other hand, in a quantizer of the mid-tread type, as in figure 1.2a, the output is zero for zero input, and the idle channel noise is correspondingly zero. In practice, however, the idle channel noise is never exactly zero due to the inevitable presence of background noise or interference. Moreover, the characterization of a quantizer exhibits deviations from its idealized form. Accordingly, we find that the average power of idle channel noise in a mid-tread quantizer is also in the order of, although less than, $\Delta^2/4$ [3].

1.6 Summary

PCM was the first method to be developed for the digital coding of waveforms. The use of digital representation of analog signals (e.g. voice, video) offers the following advantages:

1.Ruggedness to channel noise and interference.

2.Efficient regeneration of the coded signal along the transmission path.

3.Efficient exchange of increase channel bandwidth for improved signal-to-noise ratio, obeying all exponential rules.

4.A uniform format for the transmission of different kinds of base-band signals; hence their integration with other forms of digital data in a common network

5.Comparative ease with which message sources may be dropped or reinsert in a time-division multiplex system.

6.Secure communication through the use of special modulation schemes or encryption.

2. DIFFERENTIAL PULSE CODE MODULATION TECHNIQUE

2.1 Overview

When a voice or video signal is sampled at a rate slightly higher than the Nyquist rate, the resulting sampled signal is found to exhibit a high correlation between adjacent samples. The meaning of this high correlation is that, in an average sense, the signal does not change rapidly from one sample to the next, with the result that the difference between adjacent samples has a variance that is smaller than the variance of the signal itself. When these highly correlated samples are encoded, as in a standard PCM system, the resulting encoded signal contains redundant information. This means that symbols that are not absolutely essential to the transmission of information are generated as a result of the encoding process. By removing this redundancy before encoding, we obtain a more efficient coded signal [12].

Now, if we know a sufficient part of a redundant signal, we may infer the rest, or at least make the most probable estimate. In particular, if we know the past behavior of a signal up to a certain point in time, it is possible to make some inference about its future values; such a process is commonly called prediction. Suppose then a base-band signal m(t) is sampled at the rate $f_s = 1/T_s$ to produce a sequence of correlated samples T_s seconds apart; this sequence is denoted by $m(nT_s)$. The fact that it is possible to predict future values of the signal m(t) provides motivation for the differential quantization scheme shown in figure 2.1a.

In this scheme the input signal to the quantizer is defined by

$$e(nT_s) = m(nT_s) - \hat{m}(nT_s) \tag{2.1}$$

which is the difference between the unquantized input sample m(nTs) and a prediction of it, denoted by $\hat{m}$ (nTs). This predicted value is produced by using prediction filter whose input, consists of a quantized version of the input sample m(nTs). The difference signal e(nTs) is called the prediction error, since it is the amount by which the prediction filter fails to predict the input exactly. A simple and yet effective approach to implement the prediction filter is to use a tapped-delay-line filter, with the basic delay set equal to the sampling period. The block diagram of this filter is shown in figure 2.2, according to which the prediction m(nTs) is modeled as a linear combination of p past sample values of the quantized input where p is the prediction order.

By encoding the quantizer output, as in figure 2.1a, we obtain a variation of PCM,

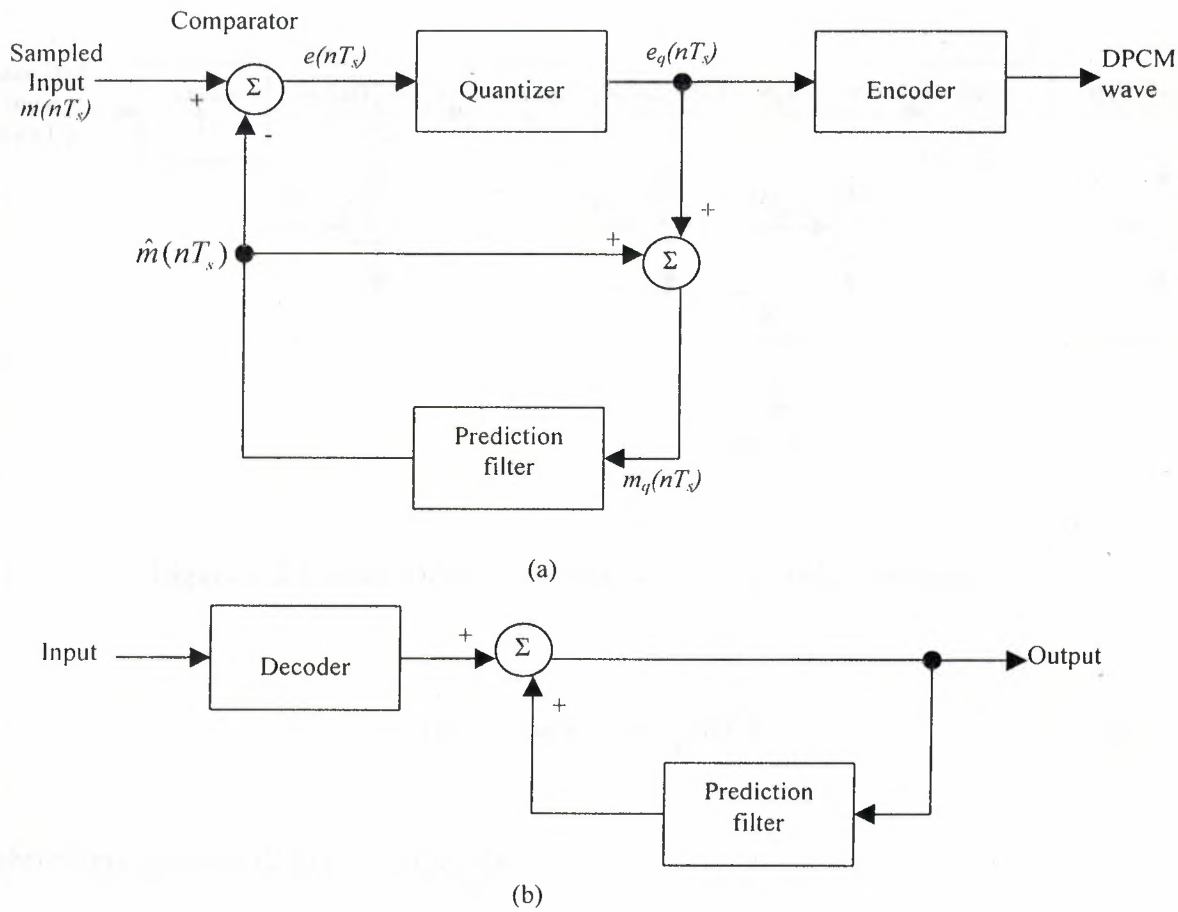


Figure 2.1 DPCM System. (a) Transmitter; (b) Receiver.

which is known as differential pulse code modulation (DPCM). It is this encoded signal that is used for transmission [5].

The quantizer output may be expressed as

$$e_q(nT_s) = e(nT_s) + q(nT_s) \tag{2.2}$$

where $q(nT_s)$ is the quantization error. According to figure 2.1a the quantizer output $e_q(nT_s)$ is added to the predicted value $\hat{m}(nT_s)$ to produce the prediction-filter input

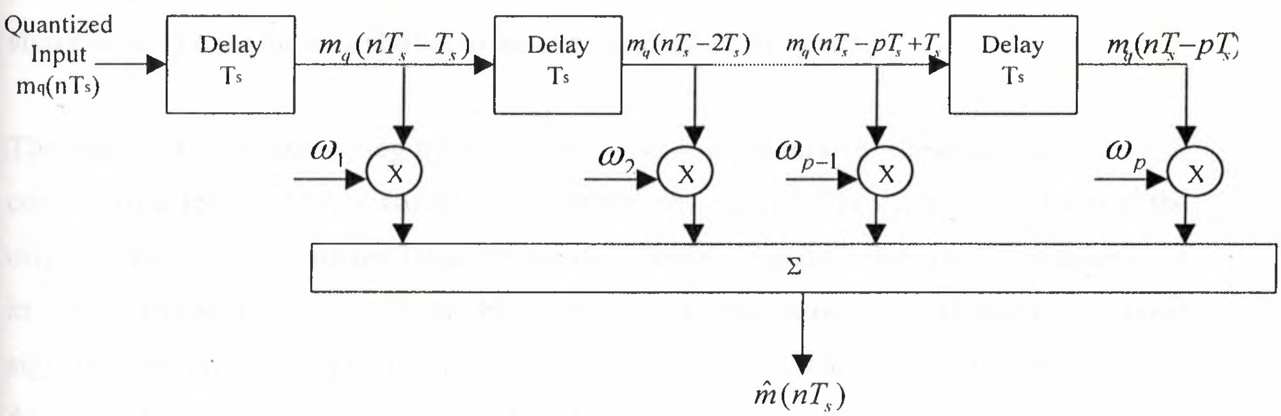


Figure 2.2 Tapped-Delay-Line Filter Used as a Prediction Filter.

$$m_q(nT_s) = \hat{m}(nT_s) + e_q(nT_s) \tag{2.3}$$

Substituting equation (2.2) in (2.3), we get

$$m_q(nT_s) = \hat{m}(nT_s) + e(nT_s) + q(nT_s) \tag{2.4}$$

However, from equation (2.1) we observe that the sum term $\hat{m}(nT_s) + e(nT_s)$ is equal to the input signal $m(nT_s)$. Therefore, we may rewrite equation (2.4) as

$$m_q(nT_s) = m(nT_s) + q(nT_s) \tag{2.5}$$

which represents a quantized version of the input signal $m(nT_s)$. That is, irrespective of the properties of the prediction filter, the quantized signal $m_q(nT_s)$ at the prediction filter input differs from the original input signal $m(nT_s)$ by the quantization error $q(nT_s)$. Accordingly, if the prediction is good, the variance of the prediction error $e(nT_s)$ will be smaller than the variance of $\hat{m}(nT_s)$, so that a quantizer with a given number of levels can be adjusted to produce a quantization error with a smaller variance than would be possible if the input signal $m(nT_s)$ were quantized directly as in a standard PCM system.

The receiver for reconstructing the quantized version of the input is shown in figure 2.1b. It consists of a decoder to reconstruct the quantized error signal. The quantized version of the original input is reconstructed from the decoder output using the same prediction filter used in the transmitter of figure 2.1a. In the absence of channel noise, we find that the encoded signal at the receiver input is identical to the encoded signal at the transmitter output. Accordingly, the corresponding receiver output is equal to $m_q(nT_s)$, which differs from the original input $m(nT_s)$ only by the quantization error $q(nT_s)$ incurred as a result of quantizing the prediction error $e(nT_s)$.

From the foregoing analysis we observe that, in a noise-free environment, the prediction filters in the transmitter and receiver operate on the same sequence of samples, $m_q(nT_s)$. It is with this purpose in mind that a feedback path is added to the quantizer in the transmitter, as shown in figure 2.1a.

Differential pulse code modulation includes delta modulation as a special case. In particular, comparing the DPCM system of figure 2.1 with the DM system of figure 3.2, we see that they are basically similar, except for two important differences: the use of a one-bit (two-level) quantizer in the delta modulator, and the replacement of the prediction filter by a single delay element (i.e., zero prediction order). Simply put, DM is the 1-bit version of DPCM. Note that unlike a standard PCM system, the transmitters of both the DPCM and DM involve the use of feedback.

DPCM, like DM, is subject to slope-overload distortion whenever the input signal changes too rapidly for the prediction filter to track it. Also like PCM, DPCM suffers from quantization noise.

2.2 Processing Gain

The output signal-to-noise ratio of the DPCM system shown in figure 2.1 is, by definition.

$$(SNR)_O = \frac{\sigma^2_M}{\sigma^2_Q} \tag{2.6}$$

where σ^2_M is the variance of the original input $m(nT_s)$, assumed to be of zero mean, and σ^2_Q is the variance of the quantization error $q(nT_s)$. We may rewrite equation (2.6) as the product of two factors as follows:

$$\begin{aligned}(SNR)_O &= \left(\frac{\sigma^2_M}{\sigma^2_E}\right)\left(\frac{\sigma^2_E}{\sigma^2_Q}\right)\\ &= G_P(SNR)_Q \end{aligned} \tag{2.7}$$

where σ^2_E is the variance of the prediction error. The factor $(SNR)_Q$ is the signal-to-quantization noise ratio, defined by

$$(SNR)_Q = \frac{\sigma^2{}_E}{\sigma^2{}_Q} \tag{2.8}$$

The other factor G_P is the processing gain produced by the differential quantization scheme; it is defined by

$$G_P = \frac{\sigma^2{}_M}{\sigma^2{}_E} \tag{2.9}$$

The quantity G_p, when greater than unity, represents the gain in signal-to-noise ratio that is due to the differential quantization scheme of figure 2.1. Now, for a given base-band (message) signal, the variance σ^2_M is fixed, so that G_P is maximized by minimizing the variance σ^2_E of the prediction error $e(nT_s)$. Accordingly, our objective should be to design the prediction filter so as to minimize σ^2_E.

In the case of voice signals, it is found that the optimum signal-to-quantization noise advantage of DPCM over standard PCM is in the neighborhood of 4-11 dB. The greatest improvement occurs in going from no prediction to first-order prediction, with some additional gain resulting from increasing the order of the prediction filter up to 4 or 5, after which little additional gain is obtained. Since 6 dB of quantization noise is equivalent to 1 bit per sample, the advantage of DPCM may also be expressed in terms of bit rate. For a constant signal-to-quantization noise ratio, and assuming a sampling rate of 8 kHz, the use of DPCM may provide a saving of about 8-16 kb/s (i.e., 1-2 bits per sample) over standard PCM [12].

2.3 Multiplexing of The PCM Signals

In this section of the chapter, we describe two related applications:

1. Hierarchy of digital multiplexers, whereby digitized voice and video signals as well as digital data are combined into one final data stream.
2. Light wave transmission link that is well-suited for use in a long-haul telecommunication network [3].

2.3.1 Digital Multiplexers

In this section we consider the multiplexing of digital signals at different bit rates. This enables us to combine several digital signals, such as computer outputs, digitized voice signals, digitized facsimile and television signals, into a single data stream (at a considerably higher bit rate than any of the inputs). Figure 2.3 shows a conceptual diagram of the digital multiplexing-demultiplexing operation.

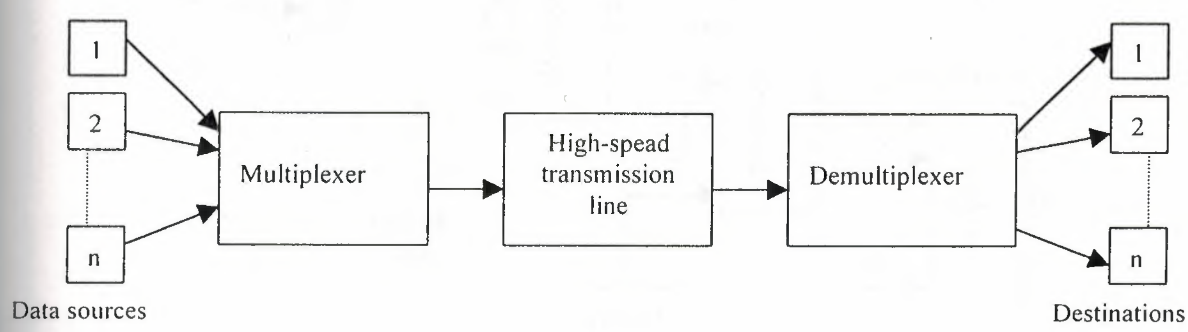


Figure 2.3 Conceptual Diagram of Multiplexing-demultiplexing.

The multiplexing of digital signals may be accomplished by using a bit-by-bit interleaving procedure with a selector switch that sequentially takes a bit from each incoming line and then applies it to the high-speed common line.

At the receiving end of the system the output of this common line is separated out into its low-speed individual components and then delivered to their respective destinations.

Two major groups of digital multiplexers are used in practice:

1. One group of multiplexers is designed to combine relatively low-speed digital signals, up to a maximum rate of 4800 bits per second, into a higher speed multiplexed signal with a rate of up to 9600 bits per second. These multiplexers are used primarily to transmit data over voice-grade channels of a telephone network. Their implementation requires the use of moderns in order to convert the digital format into an analog format suitable for transmission over telephone channels.

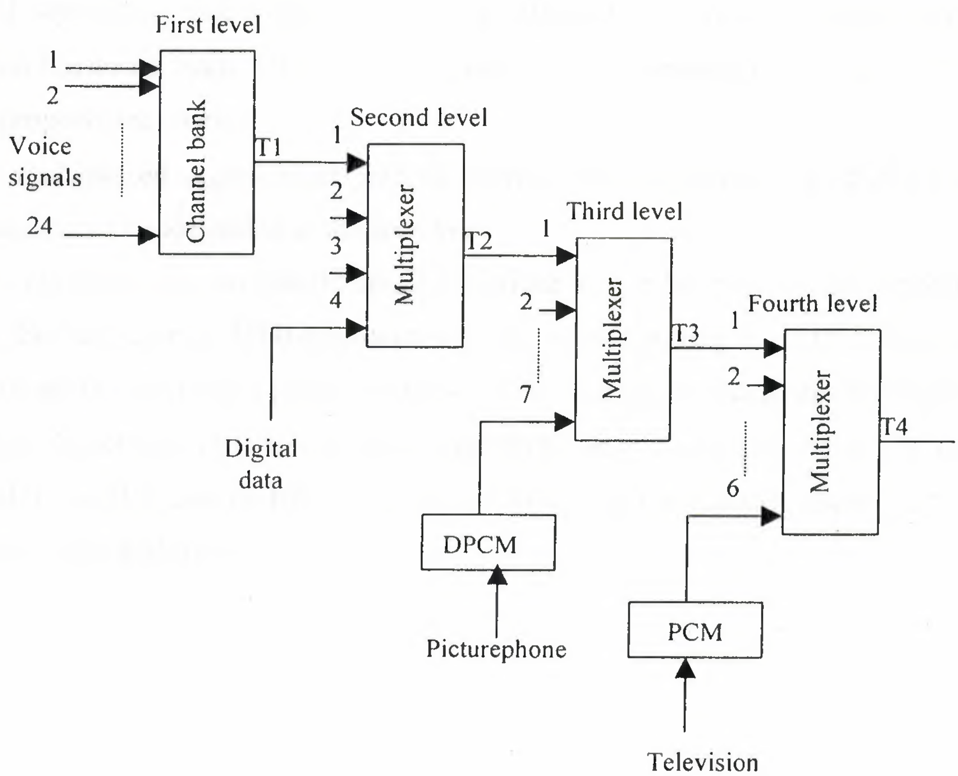


Figure 2.4 Digital Hierarchy, Bell System.

2. The second group of multiplexers, designed to operate at much higher bit rates, forms part of the data transmission service generally provided by communication carriers. For example, figure 2.4 shows a block diagram of the digital hierarchy based on the T1 carrier, which has been developed by the Bell System. The T1 carrier system, described below, is designed to operate at 1.544 megabits per second, the T2 at 6.312 megabits per second, the T3 at 44.736 megabits per second, and the T4 at 274.176 megabits per second The system is thus made up of various combinations of lower order T-carrier subsystems designed to accommodate the transmission of voice signals, picture-phone service, and television signals by using PCM, as well as (direct) digital signals from data terminal equipment.

There are some basic problems involved in the design of a digital multiplexer, irrespective of its grouping:

1. Digital signals cannot be directly interleaved into a format that allows for their eventual separation unless their bit rates are locked to a common clock. Accordingly, provision has to be made for synchronization of the incoming digital signals, so that they can be properly interleaved.
2. The multiplexed signal must include some form of framing, so that its individual components can be identified at the receiver.
3. The multiplexer has to handle small variations in the bit rates of the incoming digital signals. For example, a 1000-kilometer coaxial cable carrying 3×10^8 pulses per second will have about one million pulses in transit, with each pulse occupying about one meter of the cable. A percent variation in the propagation delay, produced by a 1^0F decrease in temperature, will result in 100 fewer pulses in the cable. Clearly, these pulses must be absorbed by the multiplexer.

In order to cater for the requirements of synchronization and rate adjustment to accommodate small variations in the input data rates, we may use a technique known as bit stuffing. The idea here is to have the outgoing bit rate of the multiplexer slightly higher than the sum of the maximum expected bit rates of the input channels by stuffing in additional non-information carrying pulses. All incoming digital signals are stuffed with a number of bits sufficient to raise each of their bit rates to equal that of a locally generated clock. To accomplish bit stuffing, each incoming digital signal or bit stream is fed into an elastic store at the multiplexer. The elastic store is a device that stores a bit stream in such a manner that the stream may be read out at a rate different from the rate at which it is read in. At the demultiplexer, the stuffed bits must obviously be removed from the multiplexed signal. This requires a method that can be used to identify the stuffed bits. To illustrate one such method, and also show one method of providing frame synchronization, we describe the signal format of the be11 system M12 multiplexer, which is designed to combine four T1 bit streams into one T2 bit stream. We begin the description by considering the T1 system first and then the M12 multiplexer.

2.3.1.1 T1 System

The T1-carrier system is designed to accommodate 24 voice channels primarily for short distance, heavy usage in metropolitan areas. The Bell System in the United States pioneered the T1 system in the early 1960s; with its introduction the shift to digital communication facilities started. The T1 system has been adopted for use throughout the United States, Canada, and Japan. It forms the basis for a complete hierarchy of higher order multiplexed systems that are used for either long-distance transmission or transmission in heavily populated urban centers.

A voice signal (male or female) is essentially limited to a band from 300 to 3400 Hz in that frequencies outside this band do not contribute much to articulation efficiency. Indeed, telephone circuits that respond to this range of frequencies give quite satisfactory service. Accordingly, it is customary to pass the voice signal through a low-pass filter with a cutoff frequency of about 3.4 kHz prior to sampling. Hence, with W = 3.4 kHz, the nominal value of the Nyquist rate is 6.8 kHz. The filtered voice signal is usually sampled at a slightly higher rate, namely, 8 kHz, which is the standard sampling rate in telephone systems.

Table 2.1 gives the projections of the segment-end points onto the horizontal axis, and the step sizes of the individual segments. The table is normalized to 8159, so that all values are represented as integer numbers. Segment 0 of the approximation is a colinear segment, passing through the origin; it contains a total of 32 uniform quantizing levels. Linear segments 1a, 2a, . . . , 7a lie above the horizontal axis, whereas linear segments 1b, 2b, . . . , 7b lie below the horizontal axis; each of these 14 segments contains 16 uniform representation levels. For colinear segment 0 the representation levels at the compressor input are 1, 3, . . . , 31, and the corresponding compressor output levels are 0, 1, . . . , 15. For linear segments 1a and 1b, the representation levels at the compressor input are 35, 39, . . . , 95, and the corresponding compressor output levels are 16, 17, . . . , 31, and soon for the other linear segments.

There is a total of 31 + 14 x 16 = 255 output levels associated with the 15-segment companding characteristic described above. To accommodate this number of output levels, each of the 24 voice channels uses a binary code with an 8-bit word. The first bit indicates whether the input voice sample is positive or negative; this bit is a 1 if positive and a 0 if negative. The next three bits of the code word identify the particular segment inside which the amplitude of the input voice sample lies, and the last four bits identify the actual quantizing step inside that segment.

Table 2.1 The 15-Segment Companding Characteristic ($\mu = 255$)

Linear segment number	Step size	Projections of segment-end point onto The horizontal axis
0	2	±31
1a, 1b	4	±95
2a, 2b	8	±223
3a, 3b	16	±479
4a, 4b	32	±991
5a, 5b	64	±2015
6a, 6b	128	±4063
7a, 7b	256	±8159

With a sampling rate of 8 kHz, each frame of the multiplexed signal occupies a period of 125 μs. In particular, it consists of twenty-four 8-bit words, plus a single bit that is added at the end of the frame for the purpose of synchronization. Hence, each frame consists of a total of 24 x 8 + 1 = 193 bits. Correspondingly, the duration of each bit equals 0.647 μs, and the corresponding bit rate is 1.544 megabits per second.

In addition to the voice signal, a telephone system must also pass special supervisory signals to the far end. This signaling information is needed to transmit dial pulses, as well as telephone off-hook/on-hook signals. In the T1 system this requirement is accomplished as follows. Every sixth frame, the least significant (that is, the eighth) bit of each voice channel is deleted and a signaling bit is inserted in its place, thereby yielding on average $7\frac{5}{6}$ *bit* operation for each voice input. The sequence of signaling bits is thus transmitted at a rate equal to the sampling rate divided by six, that is, 1.333 kilobits per second.

For two reasons, namely, the assignment of the eighth digit in every sixth frame to signaling and the need for two signaling paths for some switching systems, it is necessary to identify a super frame of 12 frames in which the sixth and twelfth frames contain two signaling paths. To accomplish this identification and still allow for rapid synchronization of the receiver framing circuitry, the frames are divided into odd and even frames. in the odd-numbered frames, the 193rd digit is made to alternate between 0 and 1. Accordingly, the framing circuit searches for the pattern 101010101010 . . . to establish frame synchronization. In the even-numbered frames, the 193rd digit is made to follow the pattern 000111000111 . . . . This makes it possible for the receiver to identify the sixth and twelfth frames as those that follow a 01 transition or 10 transition of this digit, respectively. Figure 2.5 depicts the signaling format of the T1 system.

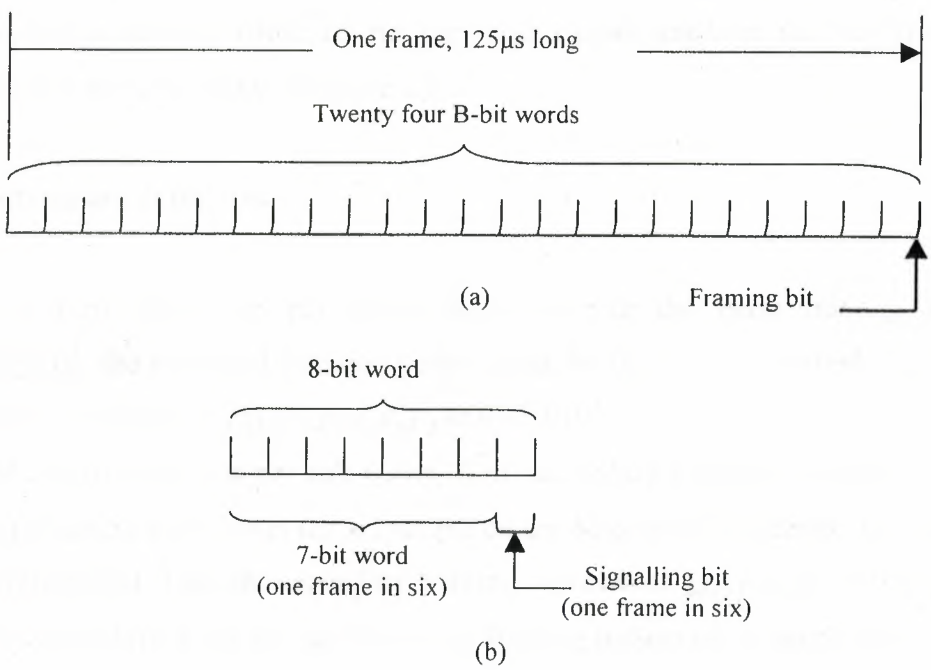


Figure 2.5 T1 Bit Stream Format. (a) Coarse Structure of a Frame; (b) Frame Structure of a Word.

2.3.1.2 M12 Multiplexer

Figure 2.6 illustrates the signal format of the M12 multiplexer. Each frame is subdivided into four sub-frames. The first sub-frame (first line in figure 2.6) is transmitted, then the second, the third, and the fourth, in that order.

Bit-by-bit interleaving of the incoming four T1 bit streams is used to accumulate a total of 48 bits, 12 from each input. A control bit is then inserted by the multiplexer. Each frame contains a total of 24 control bits, separated by sequences of 48 data bits. Three types of control bits are used in the M12 multiplexer to provide synchronization and frame indication, and to identify which of the four input signals has been stuffed. These control bits are labeled as F, M, and C in figure 2.6.

Their functions are as follows:

1. The F-control bits, two per sub-frame, constitute the main framing pulses.The subscription the F-control bits denote the actual bit (0 or 1) transmitted. Thus the main framing sequence is $F_0F_1F_0F_1F_0F_1F_0F_1$ or 01010101.
2. The M-control bits, one per sub-frame, form secondary framing pulses to identify the four sub-frames. Here again the subscripts on the M-control bits denote the actual bit (0 or 1) transmitted. Thus the secondary framing sequence is $M_0M_1M_1M_1$ or 0111.
3. The C-control bits, three per sub-frame are stuffing indicators. In particular, C_1 refers to input channel 1, C_{11} refers to input channel 11, and so forth. For example, the three C-control bits in the first sub-frame following M_0 in the first sub-frame are stuffing indicators for the first T1 signal. Setting all three C-control bits to 1 indicates the insertion of a stuffed bit in this T1 signal. To indicate no stuffing, all three are set to 0. If the three C-control bits indicate stuffing, the stuffed bit is located in the position of the first information bit associated with the first T1 signal that follows the F_1-control bit in the same sub-frame. In a similar way, the second, third, and fourth T1 signals may be

stuffed as required. By using majority logic decoding in the receiver, a single error in any of the three C-control bits can be detected.

M_0 [48]	C_I [48]	F_0 [48]	C_I [48]	C_I [48]	F_I [48]
M_I [48]	C_{II} [48]	F_0 [48]	C_{II} [48]	C_{II} [48]	F_I [48]
M_I [48]	C_{III} [48]	F_0 [48]	C_{III} [48]	C_{III} [48]	F_I [48]
M_I [48]	C_{IV} [48]	F_0 [48]	C_{IV} [48]	C_{IV} [48]	F_I [48]

Figure 2.6 Signal Format of Bell System M12 Multiplexer.

This form of decoding means simply that the majority of the C-control bits determine whether an all-one or all-zero sequence was transmitted. Thus three is or combinations of two 1s and a 0 indicate that a stuffed bit is present in the information sequence, following the control bit F_I in the pertinent sub-frame. On the other hand, three 0s or combinations of two 0s and a 1 indicate that no stuffing is used.

The demultiplexer at the receiving M12 unit first searches for the main framing sequence $F_0F_IF_0F_IF_0F_IF_0F_I$. This establishes identity for the four input T1 signals and also for the M- and C- control bits. From the $M_0M_IM_IM_I$ sequence, the correct framing of the C-control bits is verified. Finally, the four T1 signals are properly demultiplexed and destuffed.

The signal format just described has two safeguards:

1. It is possible, although unlikely, that with just the $F_0F_IF_0F_IF_0F_IF_0F_I$ sequence, one of the incoming T1 signals may contain a similar sequence. This could then cause the receiver to lock onto the wrong sequence. The presence of the $M_0M_IM_IM_I$ sequence provides verification of the genuine $F_0F_IF_0F_IF_0F_IF_0F_I$ sequence, thereby ensuring that the four T1 signals are properly demultiplexed.

2. The single-error correction capability built into the C-control bits ensures that the four T1 signals are properly destuffed.

The capacity of the M12 multiplexer to accommodate small variations in the input data rates can be calculated from the format of figure 2.6. In each M frame, defined as the interval containing one cycle of $M_0M_1M_1M_1$ bits, one bit can be stuffed into each of four input T1 signals. Each such signal has

$$12 \times 6 \times 4 = 288 \text{ Positions in each M frame}$$

Also the T1 signal has a bit rate equal to 1.544 megabits per second. Hence, each input can be incremented by

$$1.544 \times 10^6 \times \frac{1}{288} = 5.4 \text{ Kilobits/s}$$

This result is much larger than the expected change in the bit rate of the incoming T1 signal. It follows therefore that the use of only one stuffed bit per input channel in each frame is sufficient to accommodate expected variations in the input signal rate.

The local clock that determines the outgoing bit rate also determines the nominal stuffing rate S, defined as the average number of bits stuffed per channel in any frame. The M12 multiplexer is designed for S = 1/3. Accordingly, the nominal bit rate of the T2 line is

$$1.544 \times 4 \times \frac{49}{48} \times \frac{288}{288 - S} = 6.312 \text{ Megabits/s}$$

This also ensures that the nominal T2 clock frequency is a multiple of 8 kHz (the nominal-sampling rate of a voice signal), which is a desirable feature.

2.3.2 Light-wave Transmission

Optical fiber wave-guides have unique characteristics that make them highly attractive as a transmission medium. In particular, their low transmission losses and high bandwidths are important for long-haul, high-speed communications. Other advantages include small size, lightweight, and immunity to electromagnetic interference.

Consider the basic optical fiber link shown in figure 2.7 consisting of a transmitter, an optical fiber wave-guide, and a receiver. The binary data fed into the transmitter input may represent the output of a multiplexer in the digital hierarchy of figure 2.4. In any event, the transmitter emits pulses of optical power, with each pulse being "on" or "off " in accordance with the input data. Note that power is a base-band quantity that varies at the data (i.e., modulation) rate and not the optical frequency. For the light source, we may use a laser injection diode or a semiconductor light emitting diode (LED). In a system design, the choice of the light source determines the optical signal power available for transmission.

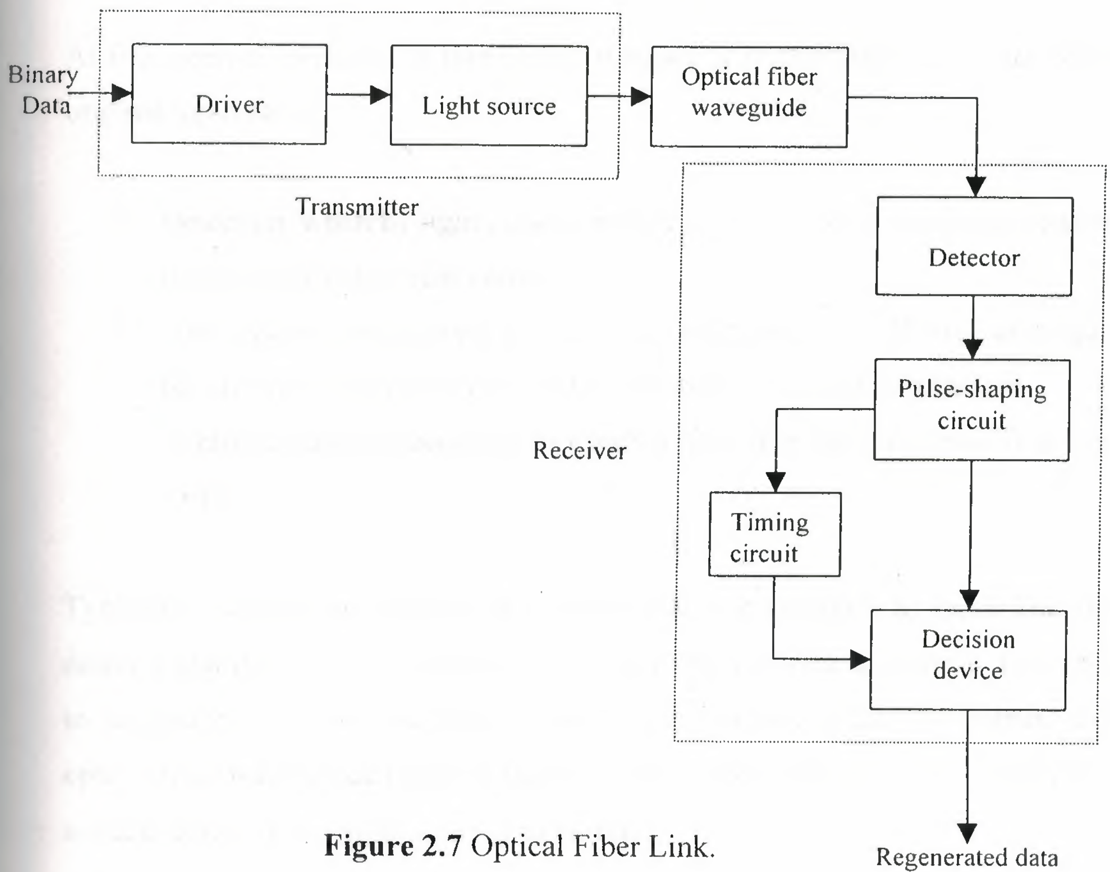


Figure 2.7 Optical Fiber Link.

The driver for the light source, typically, consists of a high-current-low-voltage device. The light source is thus turned on and off by switching the drive current on and off in a corresponding manner.

The on-off light pulses produced by the transmitter are launched into the optical fiber wave guide. The collector efficiency of the fiber depends on its core diameter and acceptance angle. We thus have to account for a source-to-fiber coupling loss that varies over a wide range, depending on the particular combination of light source and optical fiber

During the course of propagation along the fiber, a light pulse suffers an additional loss or attenuation that increases exponentially with distance; we refer to this as fiber loss. Another important phenomenon that occurs during propagation is dispersion, which causes light originally concentrated into a short pulse to spread out into a broader pulse as it propagates along the optical fiber wave-guide.

At the receiver, performing three basic operations in the following order regenerates the original input data:

1. Detection, whereby light pulses impinging on the receiver input are converted back into pulses of electrical current.
2. Pulse shaping and timing that involves amplification, filtering, and equalization of the electrical pulses, as well as the extraction of timing information.
3. Decision making, according to which a particular received pulse is declared "on" or "off".

Typically, the detector consists of a photodiode that responds to light. The choice of the detector and its associated circuitry determines the receiver sensitivity. It is also important to recognize that since the optical power is modulated at the transmitter, and since the optical fiber wave-guide operates linearly on the propagation power the detector behaves as a linear device that converts power to current.

From this discussion, we see that a light-wave transmission link differs from its coaxial cable counterpart in that power, rather than current (both base-band quantities), propagates through the optical fiber wave-guide; otherwise, their individual block diagrams look very much alike. It differs from its microwave counterpart in that a light-wave receiver employs direct detection of the incoming signal rather than first performing a heterodyne down-conversion

In the design of a light-wave transmission link, two separate factors have to be considered: transmission bandwidth and signal losses. The transmission bandwidth of an optical fiber is determined by the dispersion phenomenon. This, in turn, limits the feasible data rate or the rate at which light pulses can propagate through the optical fiber wave-guide. The signal losses are contributed by source-to-fiber coupling loss, fiber loss, fiber-to-fiber loss (due to joining the fiber by a permanent splice or the use of a demountable connector), and fiber-to-detector coupling loss. The fiber loss is insensitive to the input data rate, but it varies with wavelength. in any case, knowing these losses, and knowing the optical power available from the light source, we can determine the power available at the detector.

2.4 Summary

advantage of DPCM over standard PCM is in the neighborhood of 4-11 dB. The greatest improvement occurs in going from no prediction to first-order prediction, with some additional gain resulting from increasing the order of the prediction filter up to 4 or 5, after which little additional gain is obtained. Since 6 dB of quantization noise is equivalent to 1 bit per sample, the advantage of DPCM may also be expressed in terms of bit rate. For a constant signal-to-quantization noise ratio, and assuming a sampling rate of 8 kHz, the use of DPCM may provide a saving of about 8-16 kb/s (i.e., 1-2 bits per sample) over standard PCM.

3. DELTA MODULATION

3.1 Overview

In delta modulation (DM), an incoming message signal is over sampled (i.e., at a rate much higher than the Nyquist rate) to purposely increase the correlation between adjacent samples of the signal. This is done to permit the use of a simple quantizing strategy for constructing the encoded signal [13].

In its basic form, DM provides a staircase approximation to the over-sampled version of the message signal, as illustrated in figure 3.1a. the difference between the input and the approximation is quantized into only two levels, namely, $\pm\Delta$, corresponding to positive and negative differences, respectively. Thus, if the approximation falls below the signal at any sampling epoch, it is increased by Δ. If, on the other hand, the approximation lies above the signal, it is diminished by Δ. Provided that the signal does not change too rapidly from sample to sample, We find that the staircase approximation remains within $\pm\Delta$ of the input signal.

Denoting the input signal as m(t) , and its staircase approximation as m_q (t), the basic principle of delta modulation may be formalized in the following set of discrete-time relations:

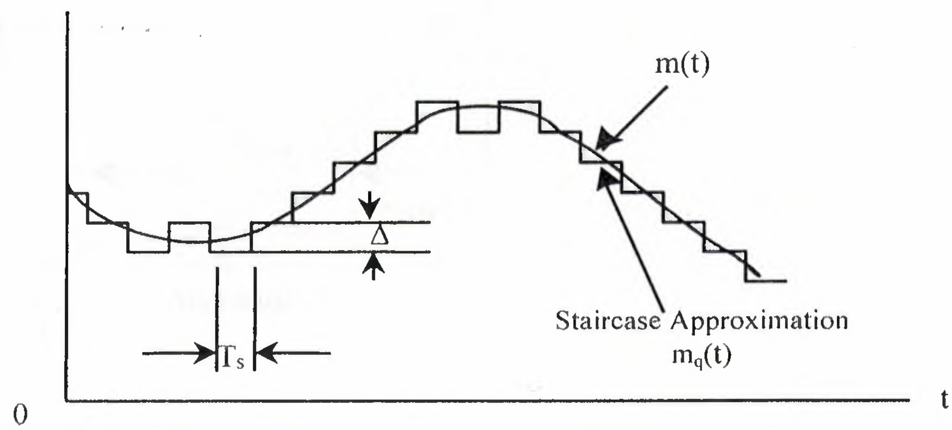
$$e(nT_s) = m(nT_s) - m_q(nT_s - T_s) \tag{3.1}$$

$$e_q(nT_s) = \Delta \,\mathrm{sgn}[e(nT_s)] \tag{3.2}$$

$$m_q(nT_s) = m_q(nT_s - T_s) + e_q(nT_s) \tag{3.3}$$

where T_s is the sampling period; $e(nT_s)$ is an error signal representing the difference between the present sample value $m(nT_s)$ of the input signal and the latest approximation to it, that is, $m(nT_s) - m_q(nT_s - T_s)$; and $e_q(nT_s)$ is the quantized version of $e\{nT_s)$. The quantizer output $e_q(nT_s)$ is finally coded to produce the desired DM signal.

Figure 3.1a illustrates the way in which the staircase approximation $m_q(t)$ follows variations in the input signal m(t) in accordance with equations (3.1) to (3.3), and figure 3.1b displays the corresponding binary sequence at the delta modulator output. It is apparent that in a delta modulation system the rate of information transmission is simply equal to the sampling rate $f_s = 1/T_s$.



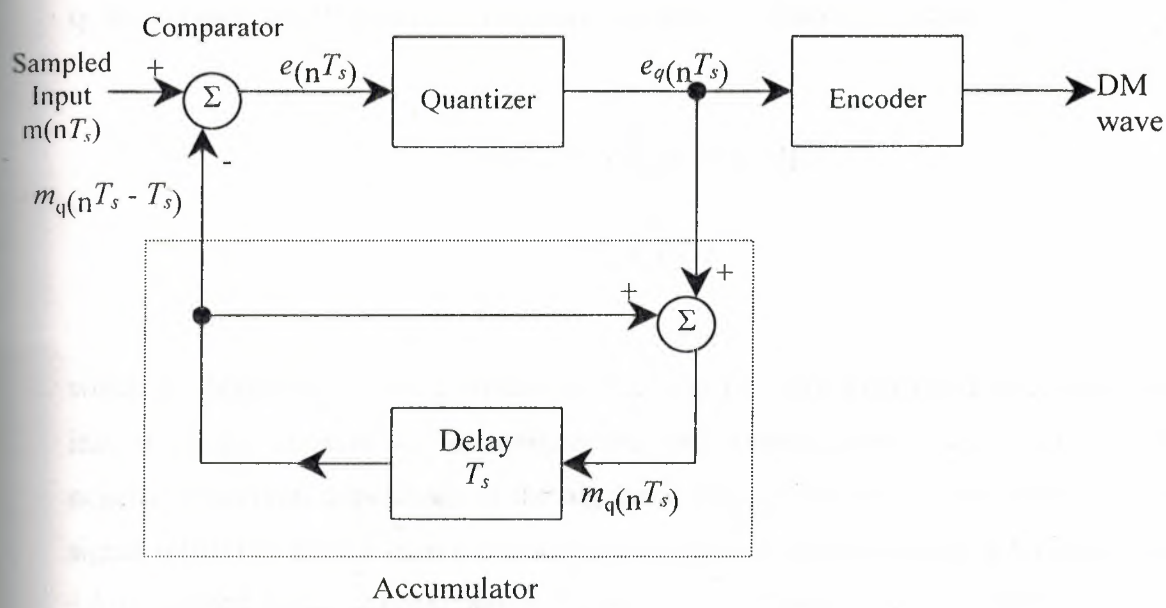
(a)

Binary Sequence at Modulator Output   0 0 1 0 1 1 1 1 1 0 1 0 0 0 0 0 0

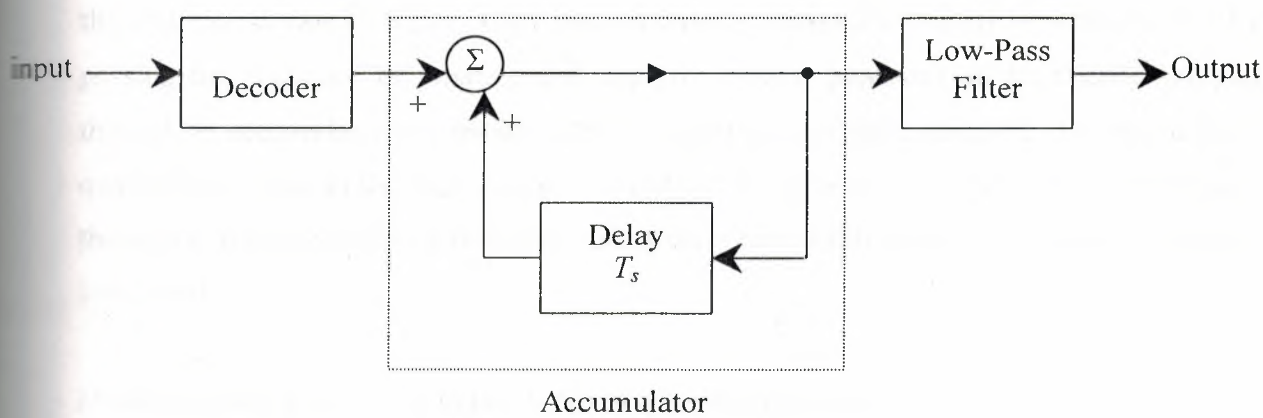
(b)

Figure 3.1 Illustration of Delta Modulation. (a) Staircase Approximation $m_q(t)$; (b) The Corresponding Binary Sequence at The Delta Modulator Output.

The principal virtue of delta modulation is its simplicity. It may be generated by applying the sampled version of the incoming message signal to a modulator that involves a comparator, quantizer, and accumulator interconnected as shown in figure 3.2a. Details of the modulator follow directly from equations (3.1) to (3.3).



(a)



(b)

Figure 3.2 DM System. (a) Transmitter; (b) Receiver.

The comparator computes the difference between its two inputs. The quantizer consists of a hard limiter with an input-output relation that is a scaled version of the signal function. The quantizer output is then applied to an accumulator, producing the result.

$$\begin{aligned} m\,(nT\,) &= \Delta \sum_{i=1}^{n} \operatorname{sgn}\left[e(iT_s)\right] \\ &= \sum_{i=1}^{n} e_q(iT_s) \end{aligned} \qquad (3.4)$$

which is obtained by solving equations (3.2) and (3.3) for $m_q(nT_s)$. Thus, at the sampling instant nT_s,the accumulator increments the approximation by a step Δ in a positive or negative direction, depending on the algebraic sign of the error signal $e(nT_s)$.If the input signal $m(nT_s)$ is greater than the most recent approximation $m_q(nT_s)$, a positive increment $+\Delta$ is applied to the approximation. If, on the other hand, the input signal is smaller, a negative increment $-\Delta$ is applied to the approximation. In this way, the accumulator does the best it can to track the input samples by one step (of amplitude $+\Delta$ or $-\Delta$) at a time. In the receiver shown in figure 3.2b, the staircase approximation $m_q(t)$ is reconstructed by passing the sequence of positive and negative pulses, produced at the decoder output, through an accumulator in a manner similar to that used in the transmitter. The out-of-band quantization noise in the high-frequency staircase waveform $m_q(t)$ is rejected by passing it through a low-pass filter, as in figure 3.2b, with a bandwidth equal to the original message bandwidth.

Delta modulation is subject to two types of quantization error:

1. slope overload distortion.
2. granular noise.

We first discuss the cause of slope overload distortion, and then granular noise.

We observe that equation (3.3) is the digital equivalent of integration in the sense that it represents the accumulation of positive and negative increments of magnitude Δ. Also, denoting the quantization error by $q(nT_s)$, as shown by,

$$m_q(nT_s) = m(nT_s) + q(nT_s) \tag{3.5}$$

we observe from equation (3.1) that the input to the quantizer is

$$e(nT_s) = m(nT_s) - m(nT_s - T_s) - q(nT_s - T_s) \tag{3.6}$$

Thus, except for the quantization error $q(nT_s - T_s)$, the quantizer input is a first backward difference of the input signal, which may be viewed as a digital approximation to the derivative of the input signal or, equivalently, as the inverse of the digital integration process. If we consider the maximum slope of the original input waveform m(t), it is clear that in order for the sequence of samples $[m_q(nT_s)]$ to increase as fast as the input sequence of samples $[m(nT_s)]$ in a region of maximum slope of m(t), we require that the condition

$$\frac{\Delta}{T_s} \geq \max\left|\frac{dm(t)}{dt}\right| \tag{3.7}$$

be satisfied. Otherwise, we find that the step-size Δ is too small for the staircase approximation $m_q(t)$ to follow a steep segment of the input waveform m(t), with the result that $m_q(t)$ falls behind m(t), as illustrated in figure 3.3. This condition is called slope-overload, and the resulting quantization error is called slope-overload distortion (noise). Note that since the maximum slope of the staircase approximation $m_q(t)$ is fixed by the step-size Δ, increases and decreases in $m_q(t)$ tend to occur along straight lines. For this reason, a delta modulation using a fixed step-size is often referred to as a linear delta modulalor.

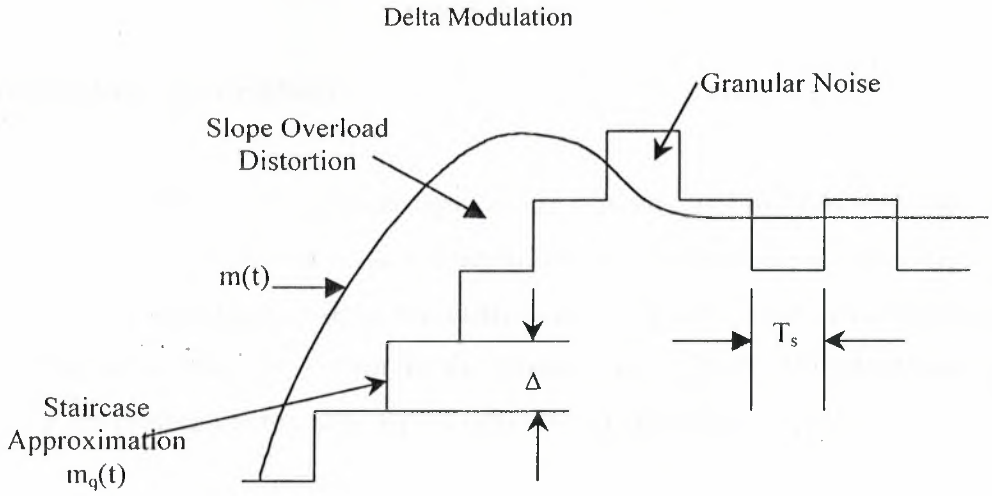


Figure 3.3 Illustration of Quantization Error in Delta Modulation.

In contrast to slope-overload distortion, granular noise occurs when the step-size Δ is too large relative to the local slope characteristics of the input waveform m(t), thereby causing the staircase approximation $m_q(t)$ to hunt around a relatively flat segment of the input waveform; this phenomenon is also illustrated in figure 3.3. Granular noise is analogous to quantization noise in a PCM system.

We thus see that there is a need to have a large step-size to accommodate a wide dynamic range, where as a small step-size is required for the accurate representation of relatively low-level signals. It is therefore clear that the choice of the optimum step-size that minimizes the mean-square value of the quantization error in a linear delta modulator will be the result of a compromise between slope-overload distortion and granular noise. To satisfy such a requirement, we need to make the delta modulator "adaptive," in the sense that the step-size is made to vary in accordance with the input signal.

3.2 Delta-sigma Modulation

As mentioned previously, the quantızer input in the conventional form of' delta modulation may be viewed as an approximation to the derivative of the incoming message signal. This behavior leads to a drawback of delta modulation in that transmission disturbance such as noise result in an accumulative error in the demodulated signal. This drawback can be overcome by integrating the message signal prior to delta modulation [12].

The use of integration in the manner described here has also the following beneficial effects:

1. The low frequency content of the input signal is pre-emphasized.
2. Correlation between adjacent samples of the delta modulator input is increased, which tends to improve overall system performance by reducing the variance of the error signal at the quantizer input.
3. Design of the receiver is simplified.

A delta modulation scheme that incorporates integration at its input is called delta-sigma modulation (Δ-Σ M). To be more precise, it should be called sigma-delta modulation. because the integration is in fact performed before the delta modulation. Nevertheless, the former terminology is the one commonly used in the literature.

Figure 3.4a shows the block diagram of a delta-sigma modulation system. In this diagram, the message signal m(t) is defined in its continuous-time form, which means that the pulse modulator now consists of a hard-limiter followed by a multiplier; the latter component is also fed from an external pulse generator (clock) to produce a 1-bit encoded signal. The use of integration at the transmitter input clearly requires an inverse signal emphasis, namely, differentiation, at the receiver. The need for this differentiation is, however, eliminated because of its cancellation by integration in the conventional DM receiver. Thus, the receiver of a delta-sigma modulation system consists simply of a low-pass filter, as indicated in figure 3.4a.

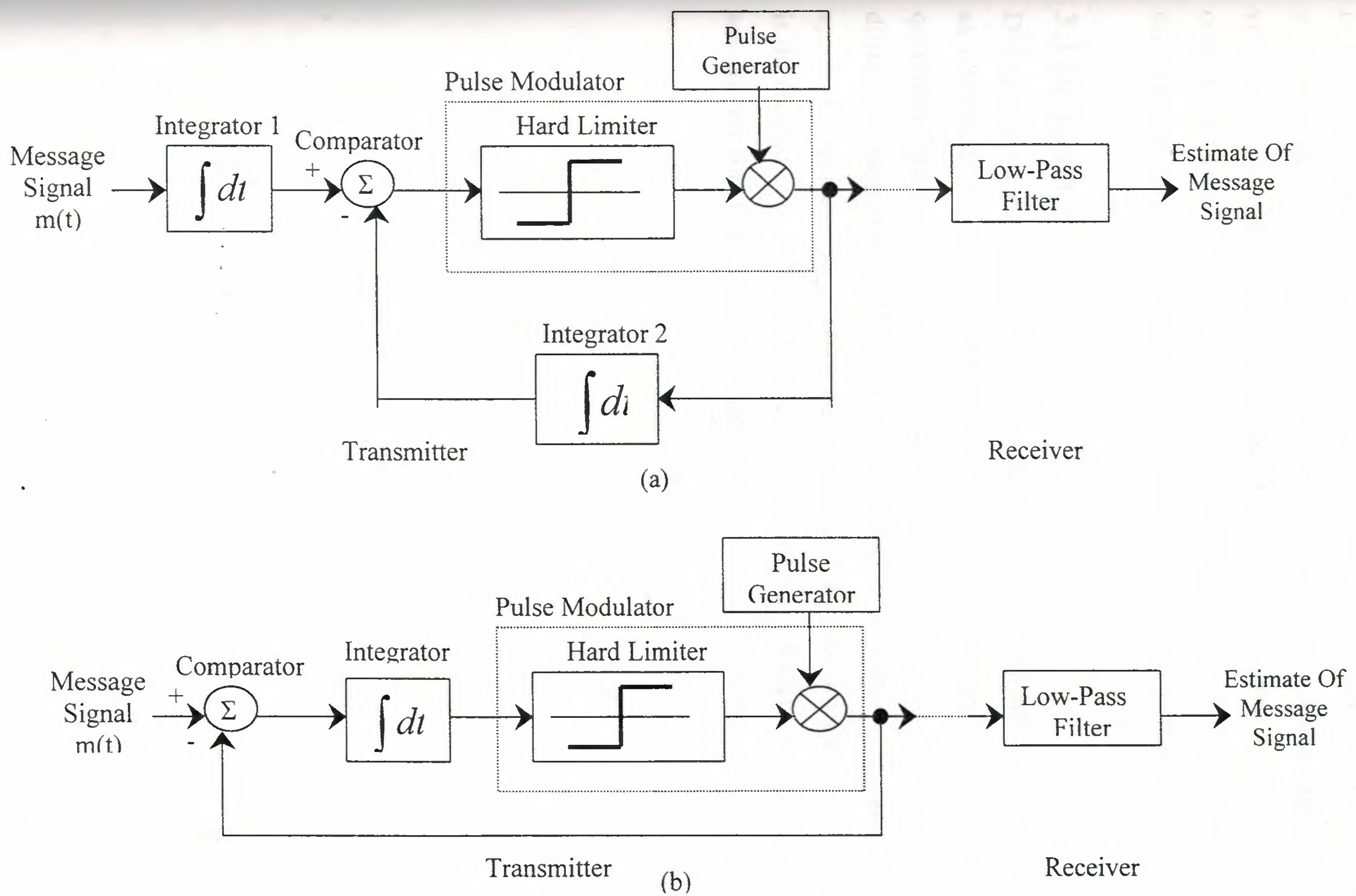


Figure 3.4 Two Equivalent Versions of Delta-sigma Modulation System.

Moreover, we note that integration is basically a linear operation. Accordingly, we may simplify the design of the transmitter by combining the two integrators 1 and 2 of figure 3.4a into a single integrator placed after the comparator, as shown in figure 3.4b. This latter form of the delta-sigma modulation system is not only simpler than that of figure 3.4a, but it also provides an interesting interpretation of delta-sigma modulation as a ''smoothed'' version of 1-bit pulse code modulation: The term smoothness refers to the fact that the comparator output is integrated prior to quantization, and the term 1-bit merely restates that the quantizer consists of a hard-limiter with only two representation levels.

3.3 Summary

DM provides a staircase approximation to the over-sampled version of the message signal, as illustrated in figure 3.1a. the difference between the input and the approximation is quantized into only two levels, namely, $\pm\Delta$, corresponding to positive and negative differences, respectively. Thus, if the approximation falls below the signal at any sampling epoch, it is increased by Δ. If, on the other hand, the approximation lies above the signal, it is diminished by Δ. Provided that the signal does not change too rapidly from sample to sample, We find that the staircase approximation remains within $\pm\Delta$ of the input signal.

4. ADAPTIVE PULSE CODE MODULATION TECHNIQUES

4.1 Overview

The use of PCM at the standard rate of 64 kb/s demands a high channel bandwidth for its transmission. Channel bandwidth is at a premium, in which case there is a definite need for speech coding at low bit rates, while maintaining acceptable fidelity or quality of reproduction. A major motivation for bit rate reduction is for secure transmission over radio channels that are inherently of low capacity. The fundamental limits on bit rate suggested by speech perception and information theory show that high-quality speech coding is possible at rates considerably less than 64 kb/s (the rate may actually be as low as 2 kb/s). The price that has to be paid for attaining this advantage is increased processing complexity (and therefore increased cost of implementation). Also, in many coding schemes, increased complexity translates into increased processing delay time. (Delay is of no concern in applications that involve voice storage, as in "voice mail").

For coding speech at low bit rates, a waveform coder of prescribed configuration is optimized by exploiting both statistical characterization of speech waveforms and properties of hearing.

In particular, the design philosophy has two aims in mind:

1. To remove redundancies from the speech signal as far as possible.
2. To assign the available bits to code the nonredundant parts of the speech signal in a perceptually efficient manner.

As we strive to reduce the bit rate from 64 kb/s (used in standard PCM) to 32, 16, 8, and 4 kb/s, the algorithms for redundancy removal and bit assignment become increasingly more sophisticated. As a rule of thumb, in the 64 to 8 kb/s range, the computational complexity (measured in terms of multiply-add operations) required to code speech increases by an order of magnitude when the bit rate is halved, for approximately equal speech quality [3].

4.2 Adaptive Differential Pulse Code Modulation

Reduction in the number of bits per sample from 8 (as used in standard PCM) to 4 involves the combined use of adaptive quantization and adaptive prediction. In this context, the term "adaptive" means being responsive to changing level and spectrum of the input speech signal. The variation of performance with speakers and speech material, together with variations in signal level inherent in the speech communication process, make the combined use of adaptive quantization and adaptive prediction necessary to achieve best performance over a wide range of speakers and speaking situations. A digital coding scheme that uses both adaptive quantization and adaptive prediction is called adaptive differential pulse code modulation (ADPCM).

The term "adaptive quantization" refers to a quantizer that operates with a time-varying step size $\Delta(nT_s)$, where T_s is the sampling period. At any given time identified by the index n, the adaptive quantizer is assumed to have a uniform transfer characteristic. The step size $\Delta(nT_s)$ is varied so as to match the variance σ^2_x of the input signal $x(nT_s)$. In particular, we write

$$\Delta(nT)=\phi\hat{\sigma}_x(nT_s) \tag{4.1}$$

where ϕ is a constant, and $\hat{\sigma}_x(nT_s)$ is an estimate of the standard deviation $\sigma_x(nT_s)$ (i.e., square root of the variance σ^2_x). For a nonstationary input, $\sigma_x(nT_s)$ is time-variable. Hence, the problem of adaptive quantization, according to equation (4.1), is one of estimating $\hat{\sigma}_x(nT_s)$ continuously.

To proceed with the application of equation (4.1), we may compute the estimate $\hat{\sigma}_x(nT_s)$ in one of two ways:

1. Unquantized samples of the input signal are used to derive forward estimates of $\sigma_x(nT_s)$, as in figure 4.1a.
2. Samples of the quantizer output are used to derive backward estimates of $\sigma_x(nT_s)$, as in figure 4.1b.

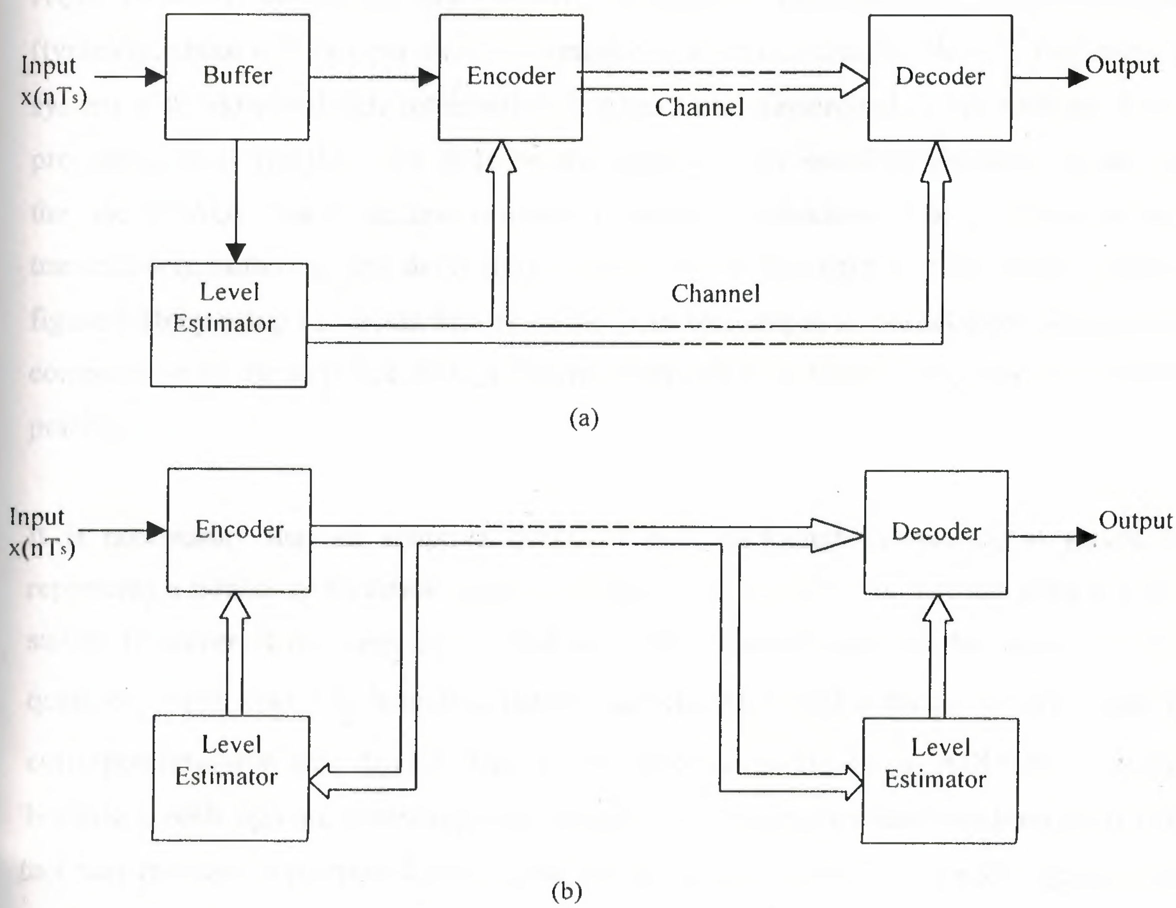


Figure 4.1 Adaptive Quantization.

(a) Adaptive Quantization with Forward Estimation (AQF);

(b) Adaptive Quantization with Backward Estimation (AQB).

The respective quantization schemes are referred to as adaptive quantization with forward estimation (AQF) and adaptive quantization with backward estimation (AQB).

The AQF scheme of figure 4.1a first goes through a learning period by buffering unquantized samples of the input speech signal. The samples are released after the estimate $\hat{\sigma}_x(nT_s)$ has been obtained. This estimate is obviously independent of quantizing noise. We therefore find that the step size $\Delta(nT_s)$ obtained from AQF is more reliable than that from AQB. However, the use of AQF requires the explicit transmission of level information (typically, about to 6 bits per step size sample) to a remote decoder, thereby burdening the system with additional side information that has to be transmitted to the receiver. Also, a processing delay (on the order of 16 ms for speech) in the encoding operation results from the use of AQF, which is unacceptable in some applications. The problems of level transmission, buffering, and delay intrinsic to AQF are all avoided in the AQB scheme of figure 4.1b by using the recent history of the quantizer output to extract information for the computation of the step size $\Delta(nT_s)$. Accordingly, AQB is usually preferred over AQF in practice.

It is noteworthy that an adaptive quantizer with backward estimation, as figure 4.1, represents a nonlinear feedback system. As such, it is not obvious that the system will be stable. However, it has been shown that the system indeed stable in the sense that if the quantizer input $x(nT_s)$ is bounded, then so are the backward estimate $\hat{\sigma}_x(nT_s)$ and the corresponding step size $\Delta(nT_s)$. The use of adaptive prediction in ADPCM is justified because speech sign are inherently nonstationary, a phenomenon that manifests itself in the fact that the auto correction function and power spectral density of speech signals time-varying functions of their respective variables. This implies that the sign of predictors for such inputs should likewise be time-varying, that is, adaptive.

As with adaptive quantization, there are two schemes for performing adaptive prediction:

1. Adaptive prediction with forward estimation (APF), in which unquantized samples of the input signal are used to derive estimates of the predictor coefficients.

2. Adaptive prediction with backward estimation (APB), in which samples of the quantizer output and the prediction error are used to derive estimates of the predictor coefficients.

In the APF scheme of figure 4.2a, N unquantized samples of the input speech are first buffered and then released after computation of M predictor coefficients that are optimized for the buffered segment of input samples. The choice of M involves a compromise between an adequate prediction gain and an acceptable amount of side information. Likewise, the choice of learning period or buffer length N involves a compromise between the rate at which statistics of the input speech signal change and the rate at which information on predictor coefficients must be updated and transmitted to the receiver. For speech, a good choice of N corresponds to a 16 ms buffer for a sampling rate of 8 kHz. and a choice of $M = 10$ ensures adequate use of the short-term predictability of speech.

However. APF suffers from the same intrinsic disadvantages (side information, buffering, and delay) as AQF. Using the APB scheme of figure 4.2b eliminates these disadvantages. Since, in the latter scheme, the optimum predictor coefficients are estimated on the basis of quantized and transmitted data, they can be updated as frequently as desired, for example, from sample to sample. Accordingly, APB is the preferred method of prediction. Continuing then with discussion of the adaptive prediction scheme shown in figure 4.2b, the box labeled 'logic for adaptive prediction" in figure 4.2b is intended to represent the mechanism for updating the predictor coefficients. Let $y(nT_s)$ denote the quantizer output, where T_s is the sampling period and n is the time index. Then, from figure 4.2b we deduce that the corresponding sample value of the predictor input is given by

$$u(nT_s) = \hat{x}(nT_s) + y(nT_s) \tag{4.2}$$

where $\hat{x}(nT_s)$ is the prediction of the speech input sample $x(nT_s)$.

We may rewrite equation (4.2) as

$$y(nT_s) = u(nT_s) - \hat{x}(nT_s) \tag{4.3}$$

The respective schemes are shown in figure 4.2.

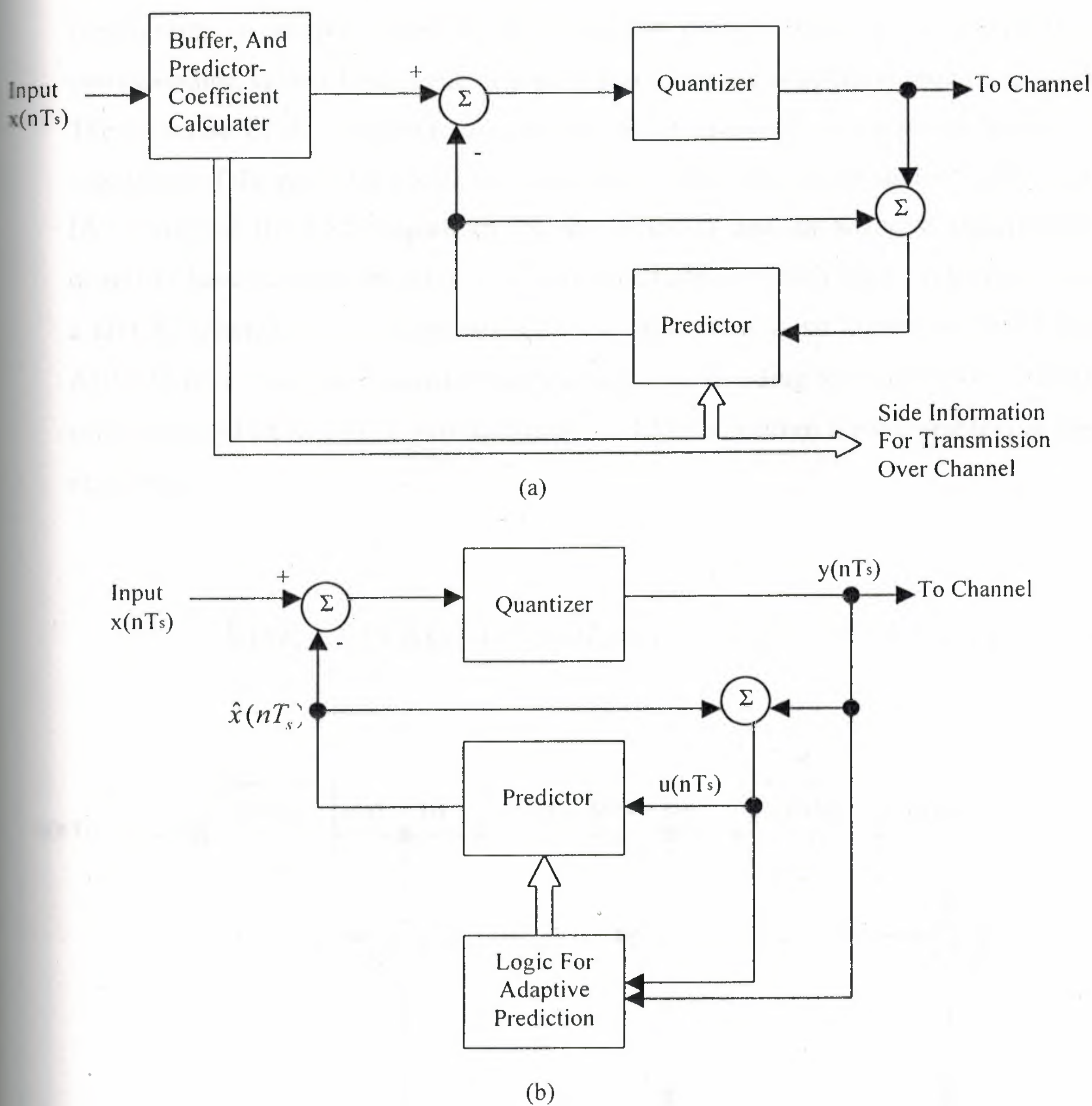


Figure 4.2 Adaptive Prediction.

(a) Adaptive Prediction with Forward Estimation (APF);

(b) Adaptive Prediction with Backward Estimation (APB).

Accordingly, with u(nT$_s$) representing a sample value of the predictor input, and $\hat{x}(nT_s)$ representing a sample value of the predictor output, we may view y(nT$_s$) as the corresponding value of the prediction error insofar as the adaptation process is concerned. The structure of the predictor, assumed to be of order M, is shown in figure 4.3. For adaptation of the predictor coefficients, we may use the least-mean-square (LMS) algorithm (A version of the LMS algorithm for the predictor and an adaptive algorithm for the quantizer have been combined in a synchronous fashion at both the encoder and decoder of a DPCM scheme. The performance of these algorithms is so impressive at 32 kb/s that ADPCM is now accepted internationally as a standard coding technique along with 64 kb/s using standard PCM). Thus, reformulating the LMS algorithm for the problem at hand, we may write

$$\hat{h}_k(nT_s + T_s) = \hat{h}_k(nT_s) + \mu y(nT_s)u(nT_s - kT_s) \qquad k = 1,2,, \ , \ , \ , M \qquad (4.4)$$

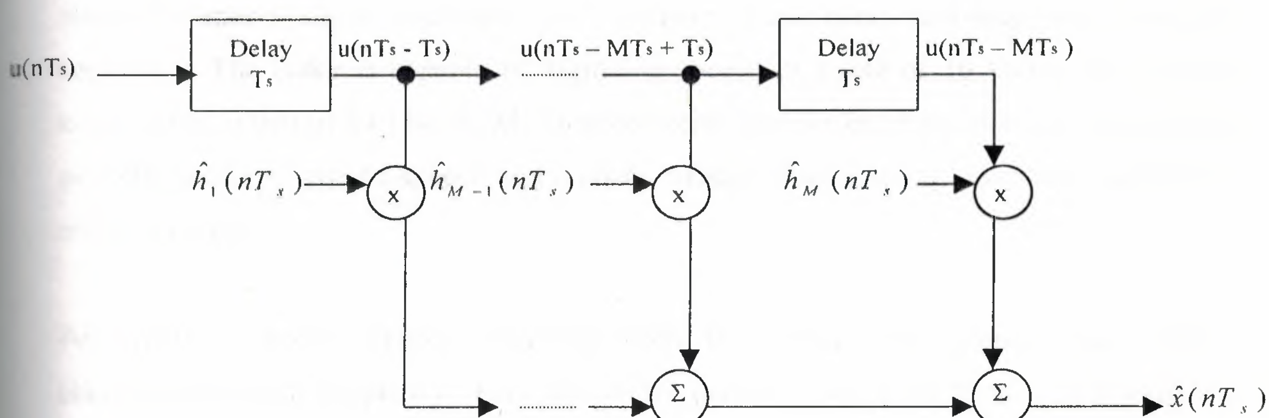


Figure 4.3 Predictor's Structure.

where μ is the adaptation constant. For the initial conditions, we set all of the predictor coefficients equal to zero at $n = 0$. The correction term in the update, equation (4.4), consists of the product $y(nT_s)u(nTs - kT_s)$, scaled by the adaptation constant μ. By using a small value for μ, the correction term will (on the average) decrease with the number of iterations n. Indeed, for stationary speech inputs and small quantization effects, the correction term may assume a small enough value in the steady-state condition for the average mean-squared error to closely approach the minimum possible value. For nonstationary inputs, on the other hand, the choice of the adaptation constant μ requires extra care in that it has to be small enough for the fluctuations in the predictor coefficients about their optimum values to be acceptable, and yet large enough for the adaptation algorithm to track variations in the input statistics [12].

4.3 Adaptive Sub-band Coding

PCM and ADPCM are both time-domain coders in that the speech signal is processed in the time-domain as a single full-band signal. We next describe a frequency-domain coder, in which the speech signal is divided into a number of sub-bands and each one is encoded separately. The coder is capable of digitizing speech at a rate of 16 kb/s with a quality comparable to that of 64 kb/s PCM. To accomplish this performance, it exploits the quasi-periodic nature of voiced speech and a characteristic of the hearing mechanism known as noise masking

Periodicity of voiced speech manifests itself in the fact that people speak with a characteristic pitch frequency. This periodicity permits pitch prediction, and therefore a further reduction in the level of the prediction error that requires quantization, compared to differential pulse code modulation without pitch prediction. The number of bits per sample that needs to be transmitted is thereby greatly reduced, without a serious degradation in speech quality.

The number of bits per sample can be reduced further by making use of the noise-masking phenomenon in perception That is, the human ear does not perceive noise in a given frequency band if the noise is about 15 dB below the signal level in that band. This means that a relatively large coding error (the equivalent of noise) can be tolerated near formants, and that the coding rate can be correspondingly reduced. In the context of speech production, the formant frequencies (or simply formants) are the resonance frequencies of the vocal tract tube. The formants depend on the shape and dimensions of the vocal tract.

In adaptive sub-band coding (ASBC), noise shaping is accomplished by adaptive bit assignment. In particular, the number of bits used to encode each sub-band is varied dynamically and shared with other sub-bands, such that the encoding accuracy is always placed where it is needed in the frequency-domain characterization of the signal. Indeed, sub-bands with little or no energy may not be encoded at all.

A block diagram of the adaptive sub-band-coding scheme is shown in figure 4.4. Specifically, the speech band is divided into a number of contiguous bands by a bank of band-pass filters (typically four to eight). The output of each band-pass filter is frequency-translated to assume a low-pass form by a modulation process equivalent to single-sideband modulation. It is then sampled (or resampled) at a rate slightly higher than its Nyquist rate (twice the width of the pertinent sub-band), and then digitally encoded by using an ADPCM with fixed (typically, first-order) prediction. A specific coding strategy is employed for each sub-band in accordance with perceptual criteria tailored to that band. Bit assignment information is transmitted to the receiver, enabling it to decode the sub-band signals individually and modulate them back to their original locations in the frequency band. Finally, they are summed to produce an output signal that provides a close replica of the original speech signal.

Let f_s denote the sampling rate for the (full band) input signal, and N the average number of bits used to encode a sample of the signal. The corresponding bit rate is therefore Nf_s bits per second. Clearly, we may write

$$Nf_s = (MN)\left(\frac{f_s}{M}\right) \qquad (4.5)$$

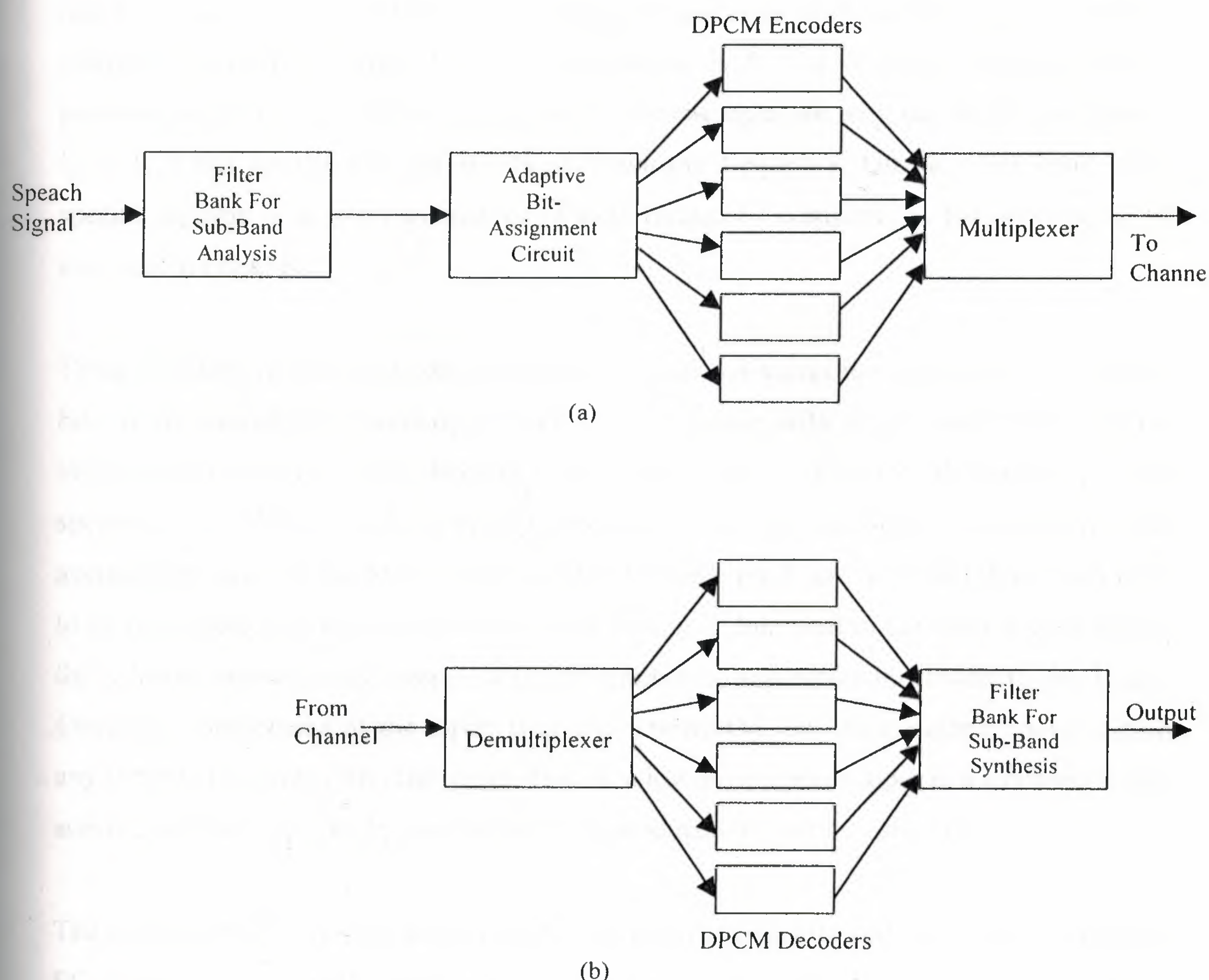


Figure 4.4 Adaptive Sub-band Coding Scheme. (a) Transmitter; (b) Receiver.

where M is the number of sub-bands, assumed to be of equal widths. The sampling rate for each sub-band is recognized to be f_s/M. The implication is a total number of MN bits per sample for the M sub-bands. To illustrate the significance of this result, consider a scheme with $M = 4$ equal-width sub-bands, a standard sampling rate $f_s = 8$ kHz for normal speech, and $N = 2$ bits per sample. Then, the sampling rate for each sub-band is 2 kHz, and the total number of bits per sample for the 4 sub-bands is 8. For a speech segment with a predominance of low-frequency components, for example, we may use the bit assignment 5, 2, 1, 0 bits for the four sub-bands in increasing frequency. On the other hand, for a speech segment with a predominance of high-frequency components, the appropriate bit assignment might be 1, 1, 3, 3.

Thus, the adaptive sub-band coding scheme of figure 4.4 varies the assignment of available bits to the various sub-bands dynamically in accordance with the spectral content of the input speech signal, thereby helping control the shape of the overall quantizing noise spectrum as a function of frequency. Specifically, more representation levels (on the average) are used for the lower frequency bands where pitch and formant information have to be preserved. If, however, high-frequency energy is dominant in the input speech signal, the scheme automatically assigns a larger number of representation levels to the higher frequency components of the input. It is also noteworthy that the quantizing noise within any sub-band is kept within that band. That is, a low-level speech input in a sub-band of the scheme cannot be hidden by quantizing noise produced in another sub-band.

The complexity of a 16 kb/s adaptive sub-band coder is typically 100 times that of a 64 kb/s PCM coder for about the same reproduction quality. However, as a result of the large number of arithmetic operations involved in designing the adaptive sub-band coder, there is a processing delay of 25 ms, on the other hand, no such delay is encountered in the PCM coder.

4.4 Subjective Quality

In coding speech, it is normal practice to supplement objective measures of performance such as signal-to-noise ratio (SNR) with subjective measures of quality, based on a mean opinion score (MOS). Indeed, in assessing the reproduction quality of digital speech coders (particularly at low bit rates), MOS ratings are often found to be more revealing than SNRs.

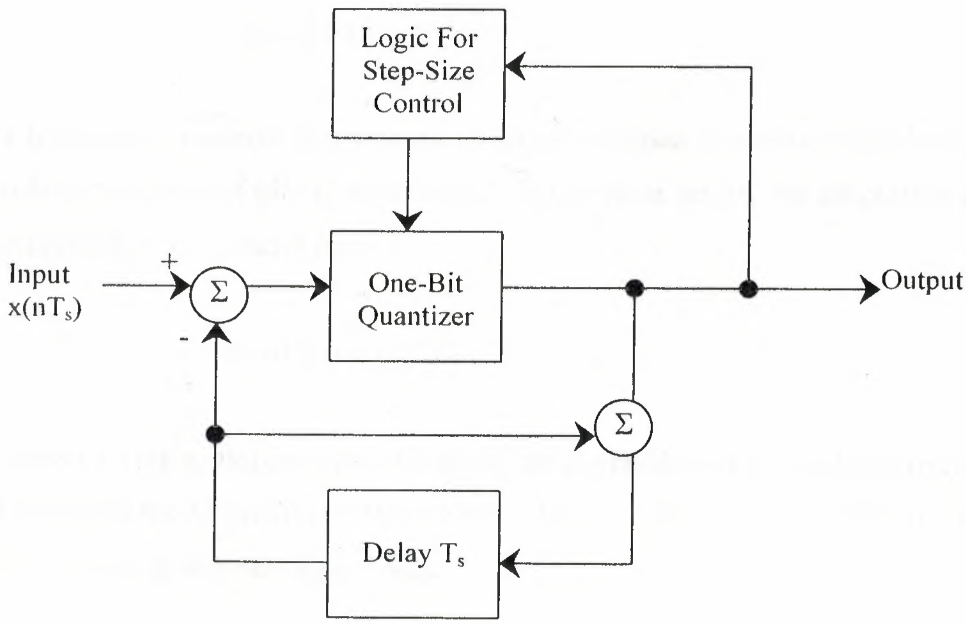
An MOS is obtained by conducting formal tests with human subjects. An MOS of 5 represents perfect quality; however, such a score is hardly ever attained. A score of 4 or more means high quality. (In standard waveform coders, high quality is referred to by telephone engineers as "toll quality," when certain transmission specifications are also met.) An MOS exceeding 4 indicates that the reproduced speech is as intelligible to test subjects as the original and also free of distortion. A score between 3 and 4 represents communication quality, implying that intelligibility is still very high and that distortion is present but not obvious.

In subjective measurements, it is found that 64 kb/s PCM and 32 kb/s DPCM coders rate "high" for quality, and that the best 16 kb/s adaptive sub-band coders approach the higher bit-rate PCM coders in quality, attaining MOS ratings very close to 4. If, however, the comparison was to be made on the basis of SNR measurements, the adaptive sub-band coders (no matter how complex) perform poorly compared to the higher bit-rate PCM coders; this supports the earlier statement that SNR ratings are not always as revealing as MOS ratings.

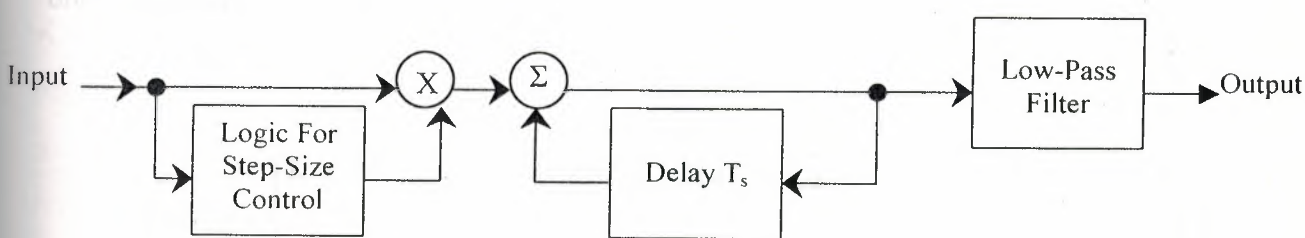
In one respect, 16 kb/s adaptive sub-band coders fall short of 64 kb/s PCM and 32 kb/s ADPCM coders in that their quality drops sharply with tandem codings (successive encoding-decoding stages). On the other hand, the higher bit-rate coders maintain a high quality after as many as eight coding-decoding stages. This issue is of particular concern in a combined analog-digital transmission path. In an all-digital link, the multistage advantage of the higher bit-rate coders is not significant since the coded signals are decoded into analog form only once.

4.5 Adaptive Delta Modulation

The performance of a delta modulator can be improved significantly by making the step size of the modulator assume a time-varying form. In particular, during a steep segment of the input signal the step size is increased. Conversely, when the input signal is varying slowly, the step size is reduced. In this way, the step size is adapted to the level of the input signal. The resulting method is called adaptive delta modulation (ADM) [3].



(a)



(b)

Figure 4.5 Adaptive Delta Modulator. (a) Transmitter; (b) Receiver.

There are several types of ADM, depending on the type of scheme used for adjusting the step size. In one type, a discrete set of values is provided for the step size. In another type, a continuous range for step-size variation is provided. There are also other types. In the sequel, we describe an example of the first type of ADM.

Figure 4.5 shows the block diagrams of the transmitter and receiver of an ADM system. In practical implementations of the system, the step size $\Delta\,(nT_s)$ or $2\delta(nT_s)$ is constrained to lie between minimum and maximum values. In particular, we write

$$\delta_{min} \le \delta\,(nT_s) \le \delta_{max} \tag{4.6}$$

The upper limit, δ_{max}, controls the amount of slope-overload distortion. The lower limit, δ_{min}, controls the amount of idle channel noise. Inside these limits, the adaptation rule for $\delta(nT_s)$ is expressed in the general form

$$\delta\,(nT_s) = g\,(nT_s)\,\delta\,(nT_s - T_s) \tag{4.7}$$

where the time-varying multiplier $g(nT_s)$ depends on the present binary output $b(nT_s)$ of the delta modulator and the M previous values $b(nT_s - T_s), \ldots, b(nT_s - MT_s)$. The algorithm is initiated with a starting step size $\delta_{start} = \delta_{min}$,

A simple version of the formula in equation (4.7) involves the use of $b(nT_s)$ and $b(nT_s - T_s)$ only, as shown by

$$g(nT_s) = \begin{cases} K & \text{if } b(nT_s) = b(nT_s - T_s) \\ K^{-1} & \text{if } b(T_s) \neq b(nT_s - T_s) \end{cases} \tag{4.8}$$

This adaptation algorithm is called a constant factor ADM with one-bit memory, where the term “one-bit memory” refers to the explicit utilization of the single previous bit $b(nT_s - T_s)$. The algorithm of equation (4.8), with K = 1.5, has been found to be well-matched to typical speech and image inputs alike, for a wide range of bit rates. In particular, at bit rates of 20 to 60 kb/s, the use of the algorithm for speech coding realizes gains in SNR equal to 5 to 10 dB over the optimum LDM for which the constant K equals 1.

Figure 4.6 illustrates the adaptive delta modulation (ADM).

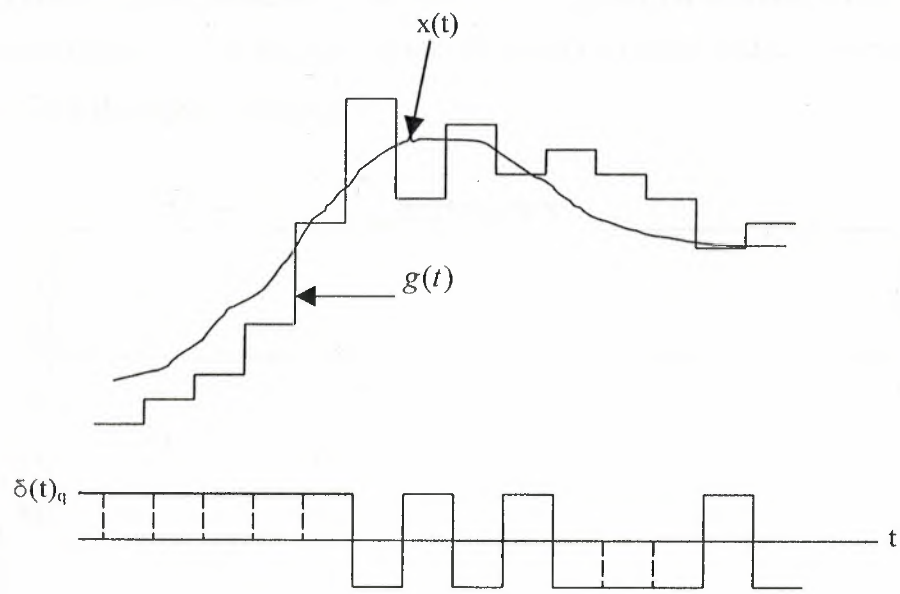


Figure 4.6 Adaptive Delta Modulations Waveforms.

4.6 Summary

For secure transmission over radio channels, we need to reduce the rate of bits and we can get more quality signal and we can decrease the rate from 64 kb/s as in PCM to 32, 16, 8 and 4. "Adaptive" means being responsive to changing level and spectrum of the input speech signal. Adaptive quantization and adaptive prediction is called adaptive differential pulse code modulation (ADPCM). The term "adaptive quantization" refers to a quantizer that operates with a time-varying step size $\Delta(nT_s)$, where T_s is the sampling period.

5. PRACTICAL IMPLEMENTATIONS

5.1 Overview

This chapter presents design procedure of Waveform Encoding Techniques Based on Differential and Adaptive Quantizing using MATLAB.

5.2 MATLAB Implementation

We first consider the operation for the case of a constant input signal using MATLAB, as shown in figure 5.1. To this we can use the program below, and its output is shown in figure 5.2, for the input in figure 5.1.

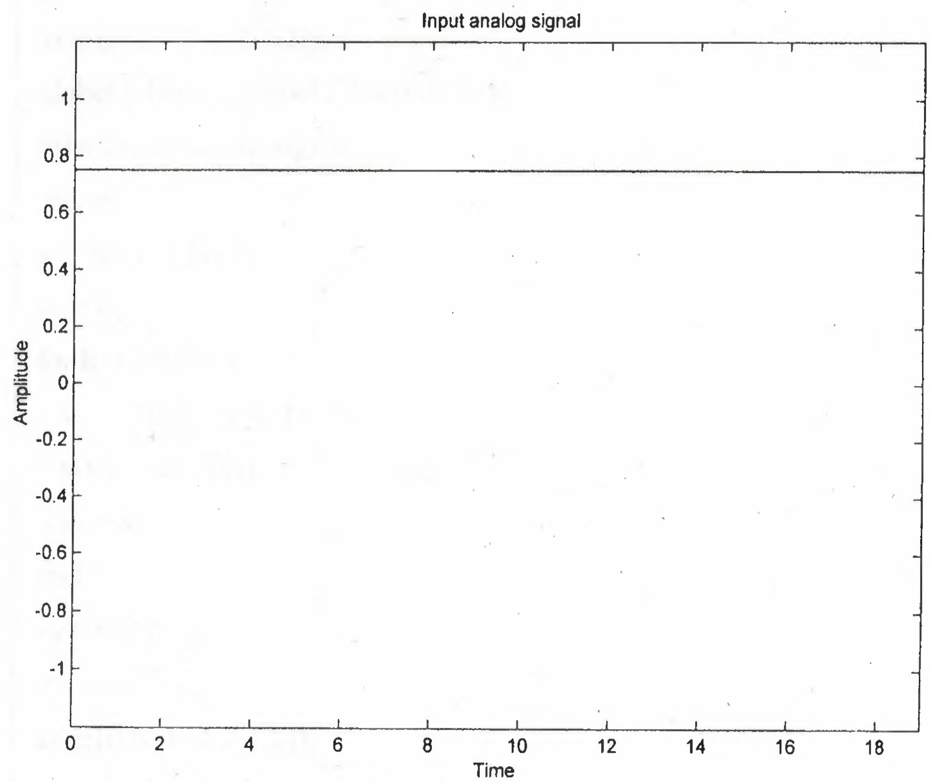


Figure 5.1 Constant Input Waveform of The Sigma-delta Quantizer.

Applying the input signal to the quantizer and the output is shown in figure 5.2

```
%N = input('Type in the length of input sequence = ');
N=20;
n = 1:1:N;
m = n-1;
%A = input('Type in the input amplitude = ');
A=0.75;
x = A*ones(1,N);
plot(m,x);
axis ([0 N-1 -1.2 1.2]);
xlabel ('Time'); ylabel ('Amplitude');
title('Input analog signal');
pause
y = zeros(1,N+1);
v0 = 0;
for k = 2:1:N+1;
   v1 = x(k-1) - y(k-1) + v0;
   y(k) =  sign(v1);
   v0 = v1;
end
yn = y(2:N+1);
stem(m, yn);
axis([0 N-1 -1.2 1.2]);
xlabel('Time'); ylabel('Amplitude');
title('Output of sigma-delta modulator');
ave = sum(yn) /N;
disp('Average value of output is = ');
disp(ave)
```

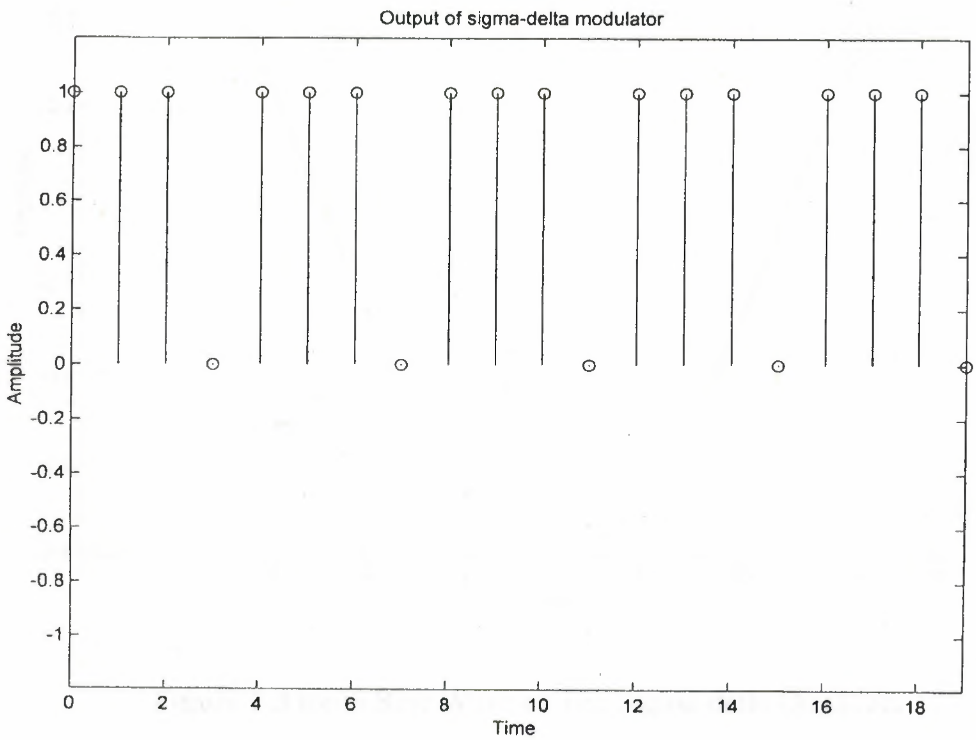


Figure 5.2 Output Waveform of The Sigma-delta Quantizer.

We now verify the operation of the sigma-delta A/D converter for a sinusoidal input of frequency 0.01 Hz using MATLAB. Because of the short length of the input sequence, the filtering operation is performed here.

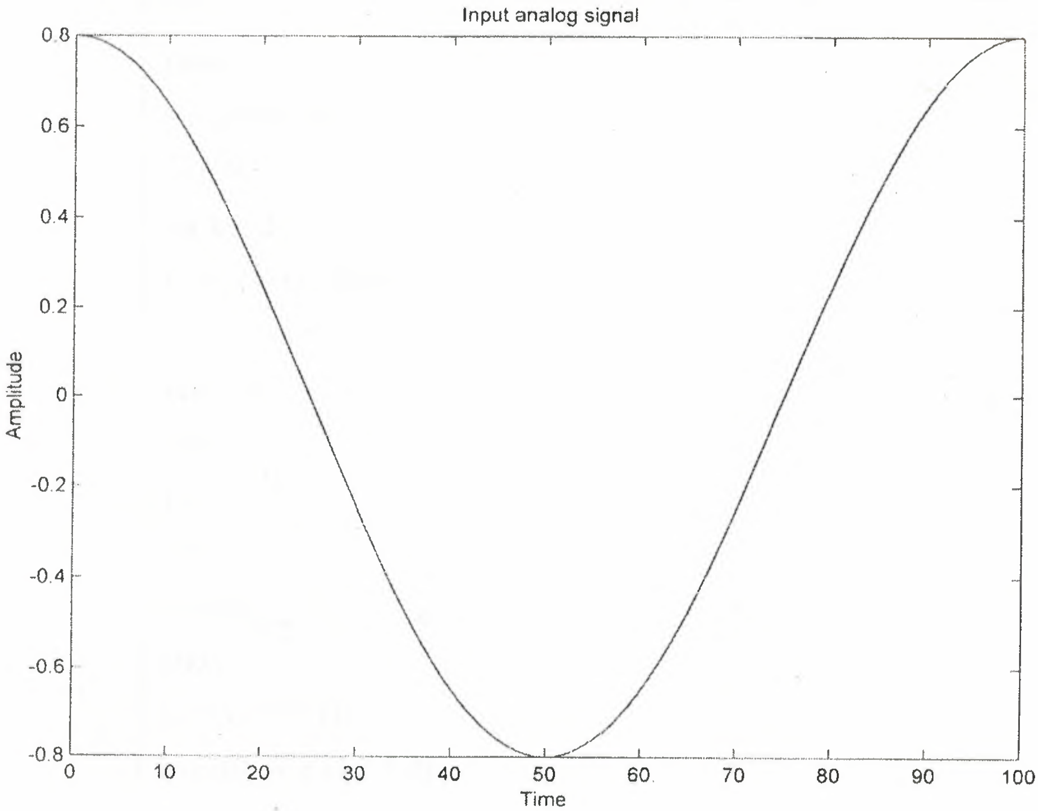


Figure 5.3 Input Sine Wave of The Sigma-delta Quantizer.

```
W0 = 2*pi*0.01;
%N = input('Type in the length of input sequence = ');
N=100;
n = 1:1:N;
m = n-1;
%A = input('Type in the amplitude of the input = ');
A=0.8
x = A*cos(w0*m);
axis([0 N-1 -1.2 1.2]);
plot (m,x);
xlabel('Time'); ylabel('Amplitude');
title('Input analog signal');
```

```
pause
y = zeros(1,N+1);
v0 = 0;
for k = 2:1:N+1;
v1 = x(k-1) - y(k-1) + v0;
if v1 >= 0;
y(k) = 1;
else
y(k) = -1;
end
v0 = v1;
end
yn = y(2:N+1);
axis([0  N-1  -1.2 1.2]);
stairs(m, yn);
xlabel('Time'); ylabel('Amplit.ude');
title('Output of sigma-delta quantizer');
Y = fft(yn);
pause
H = [1  1 0.5 zeros(1,N-5) 0.5 1];
YF = Y.*H;
out = ifft(YF);
axis([0 N-1  -1.2 1.2]);
plot(m,out);
xlabel('Time'); ylabel('Amplitude');
title('Lowpass filtered output');
```

% Sigma-delta A/D Converter Operation

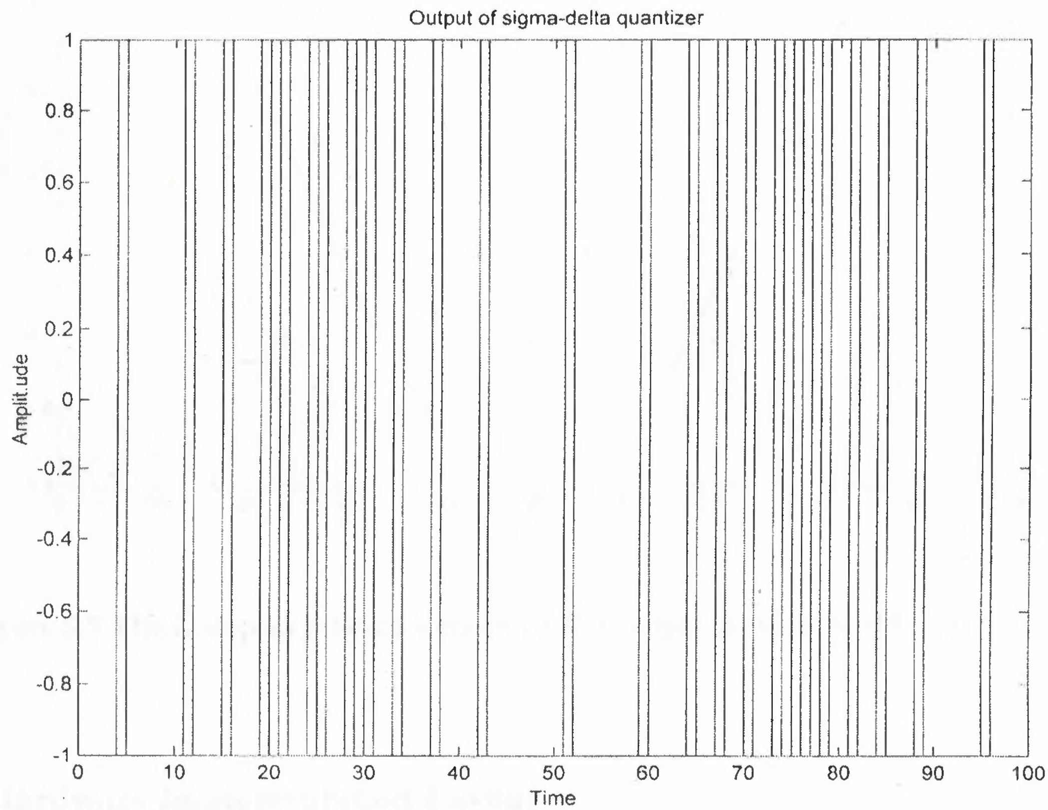


Figure 5.4 Output Waveform of The Sigma-delta Quantizer.

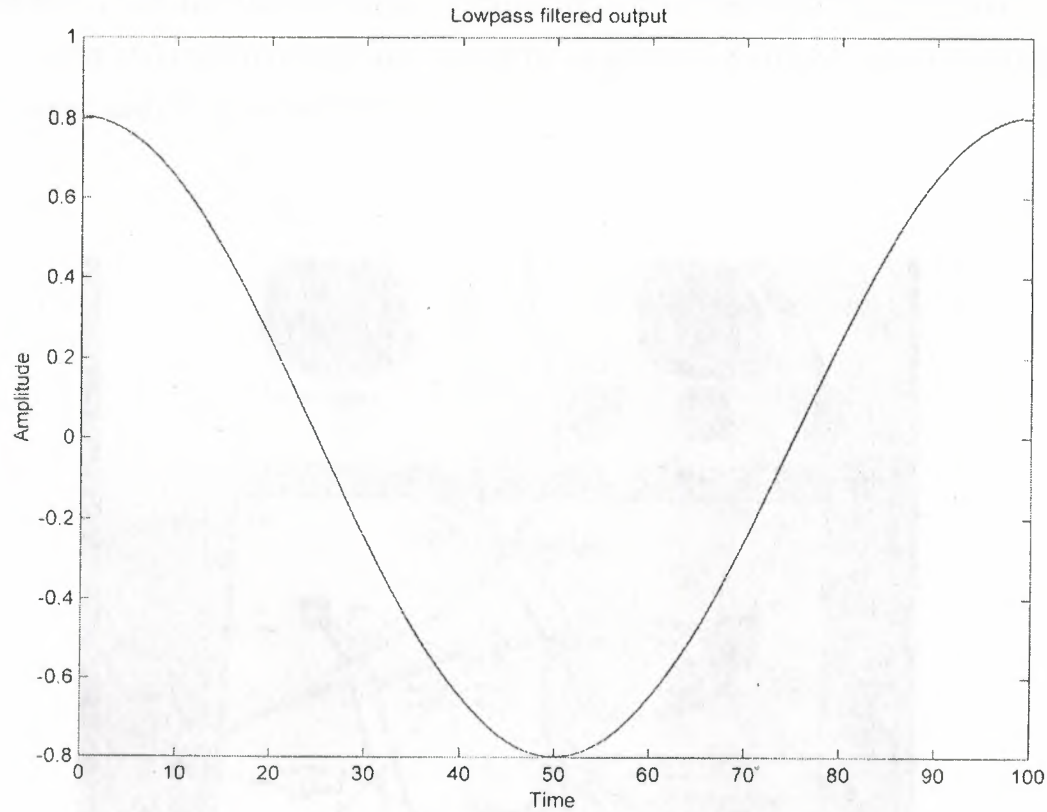


Figure 5.5 The Low-pass Filtered Version of The Output Waveform of Figure 5.4.

5.3 Hardware Implementation Layout

The experiment was connected as shown in figure 5.6, the input signal from the function generator which is shown in figure 5.7, the output of the experiment and the input were connected to the oscilloscope as shown in figure 5.8, which shows the input signal and according to the reading which is appearing in the monitor of the oscilloscope each division is equivalent to 2mV, that is the input signal applied is equivalent to 17.5 divisions, i.e. 35mV, and according to figure 5.9 appears the digital output signal of the delta modulation technique of converting the analog into digital signal, and according to what appears in the oscilloscope that each division is equivalent to 5mV, this means that 8 divisions that equivalent to 40mV. Also as shown in figure 5.6, this shows that the signal is applied to a filter.

Because the input signal is applied to differential amplifier which is the input to the comparator, and also the comparator is the input of pulse generator in the receiver and the receiver and transmitter are also appearing in figures 5.8 and 5.9 the signal while it was analog and the signal while it became digital.

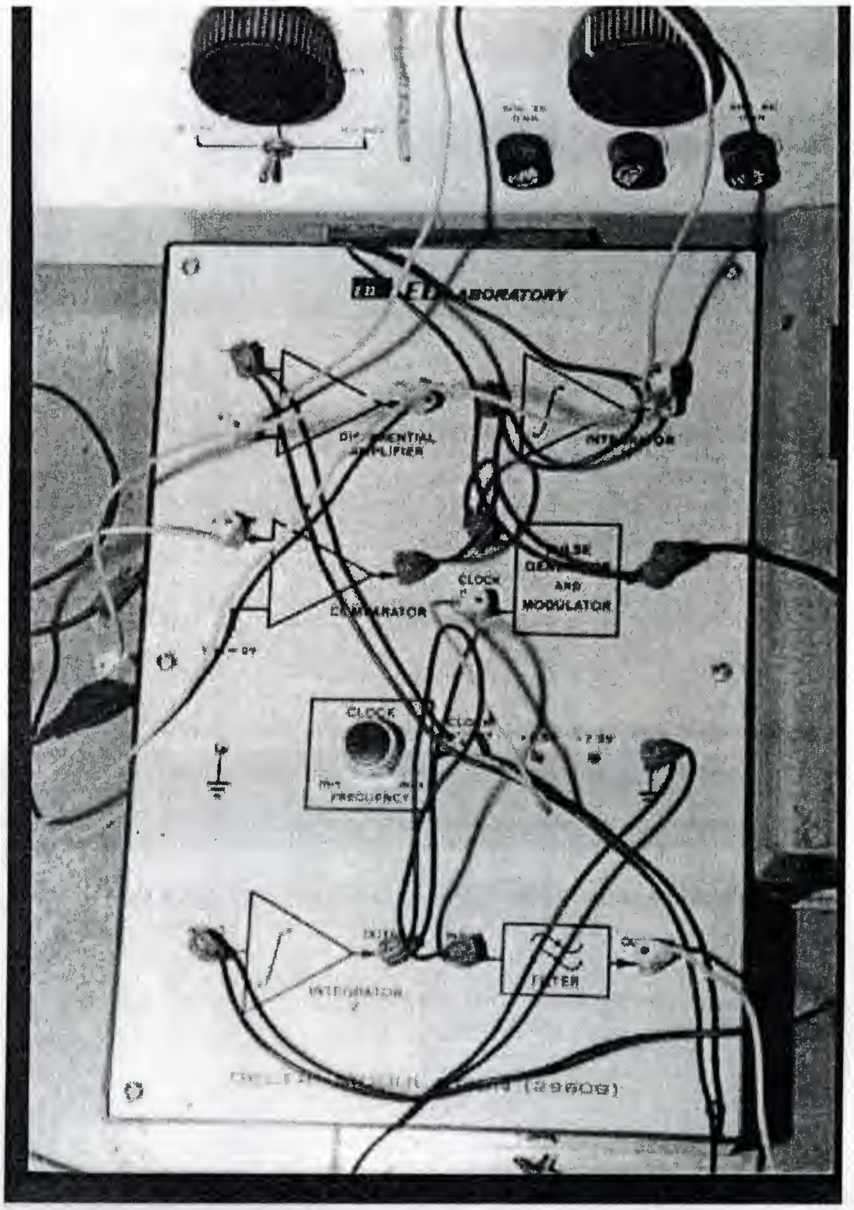


Figure 5.6

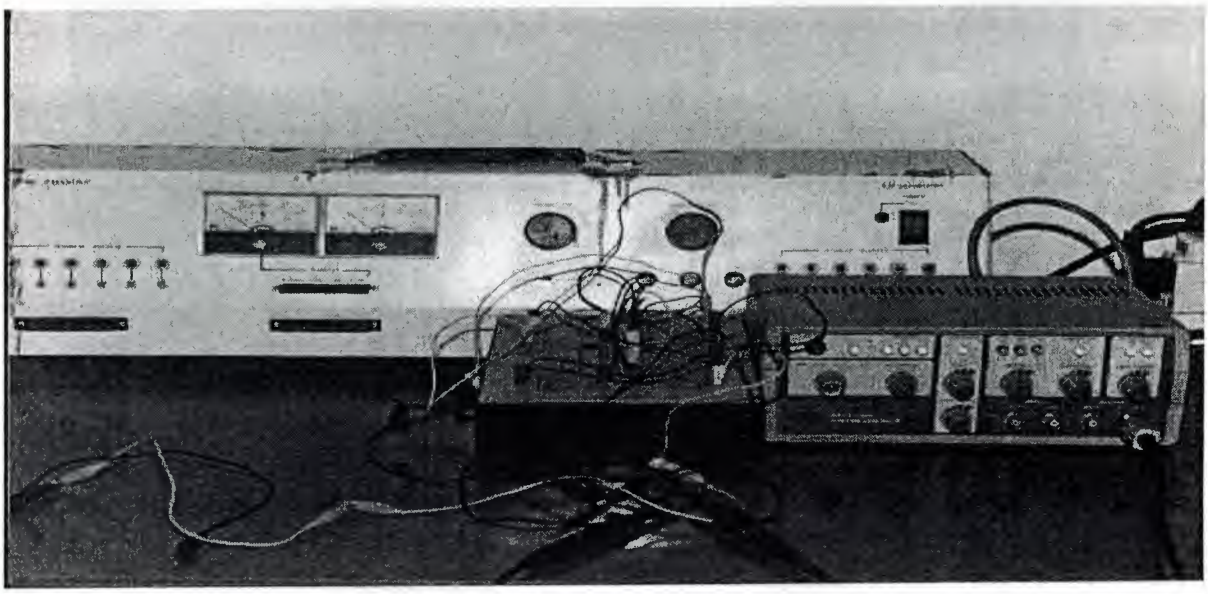


Figure 5.7

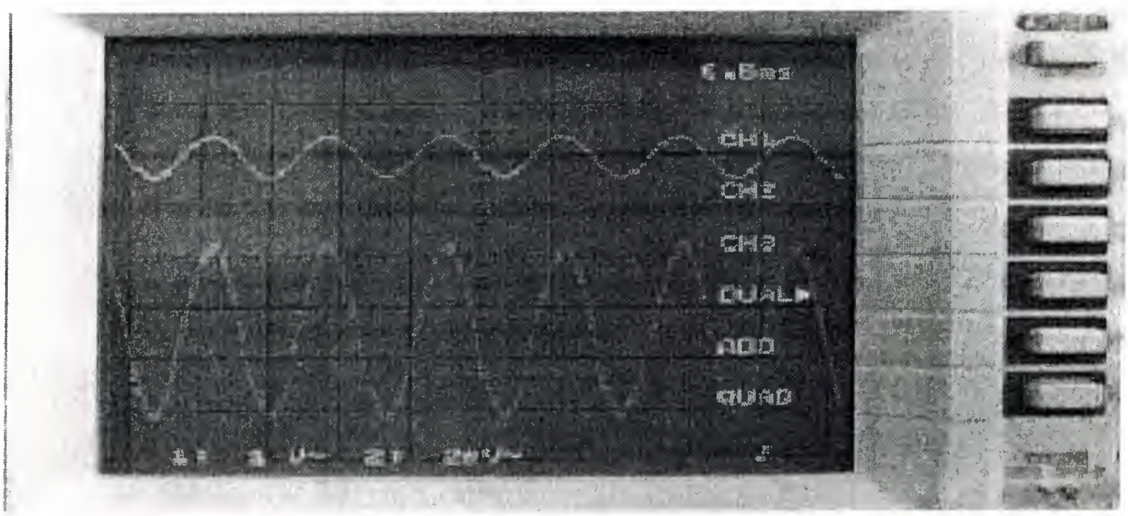


Figure 5.8

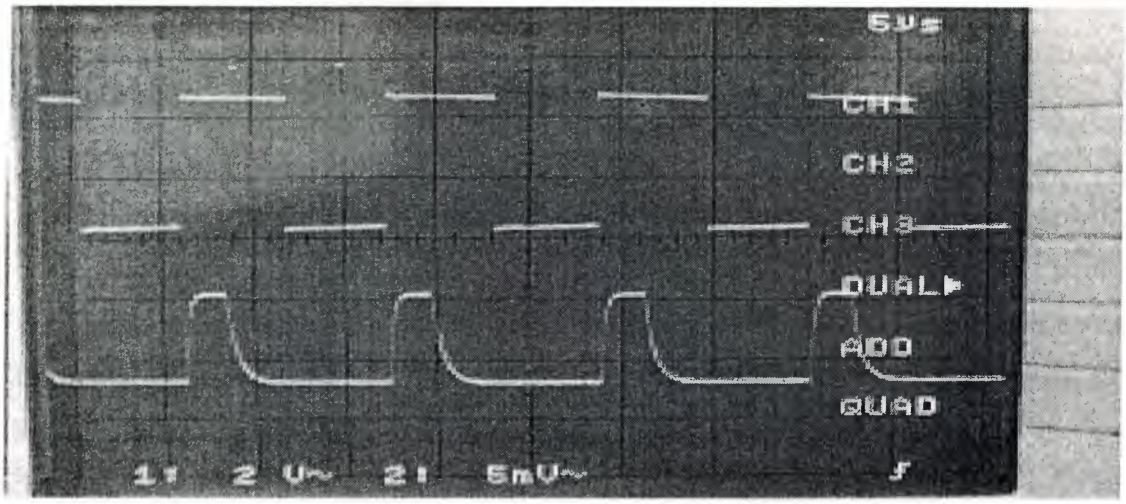


Figure 5.9

Table 5.1 shows the input analog signal and the output digital signal voltages according to the readings in our experiment.

Table 5.1 Input Analog Signal and Output Digital Signal Voltages

Input analog signal, Volt	Output digital signal, Volt
0	0.021
0.035	0.040
1	1
2	2.1
3	3
4	4
5	5
6	5.2
7	5.4
8	5.4
9	5.4
10	5.4

The block diagram of adaptive delta pulse code modulation (ADPCM) system as in figure 5.10 consists of three parts: transmitter, transmission channel and receiver. In transmitter the analog signal will be converting to digital by using A/D converter and then pass to the encoder and this encoded signal will be sent by transmission channel. In receiver we have encoded signal and after decoding the output of decoder will connect with the original signal together to see the same signal, and in the monitor we can see the input analog signal as in figure 5.11a, the original and reconstructed signals as in figure 5.11b, step size signal as in figure 5.11c and the encoded signal as in figure 5.11d.

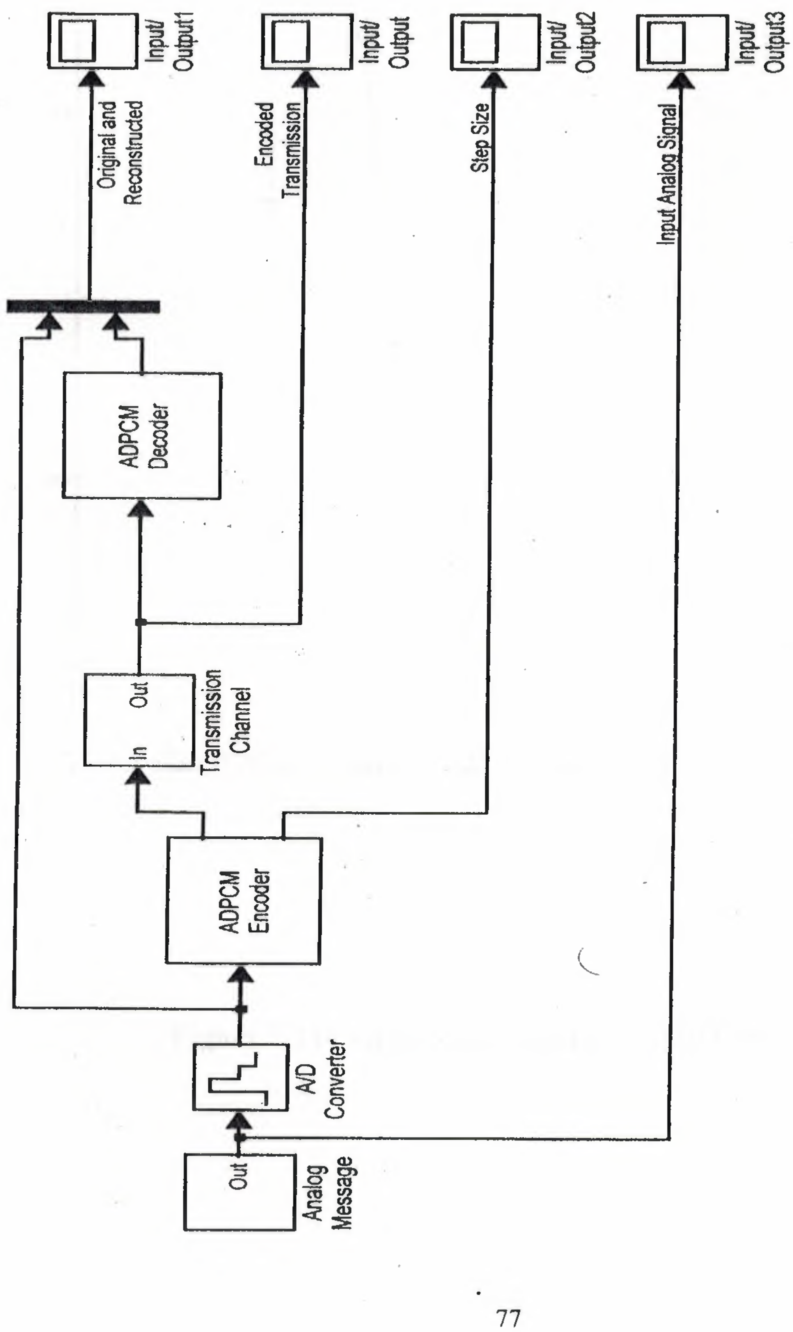


Figure 5.10 Adaptive Delta Pulse Code Modulation (ADPCM) System.

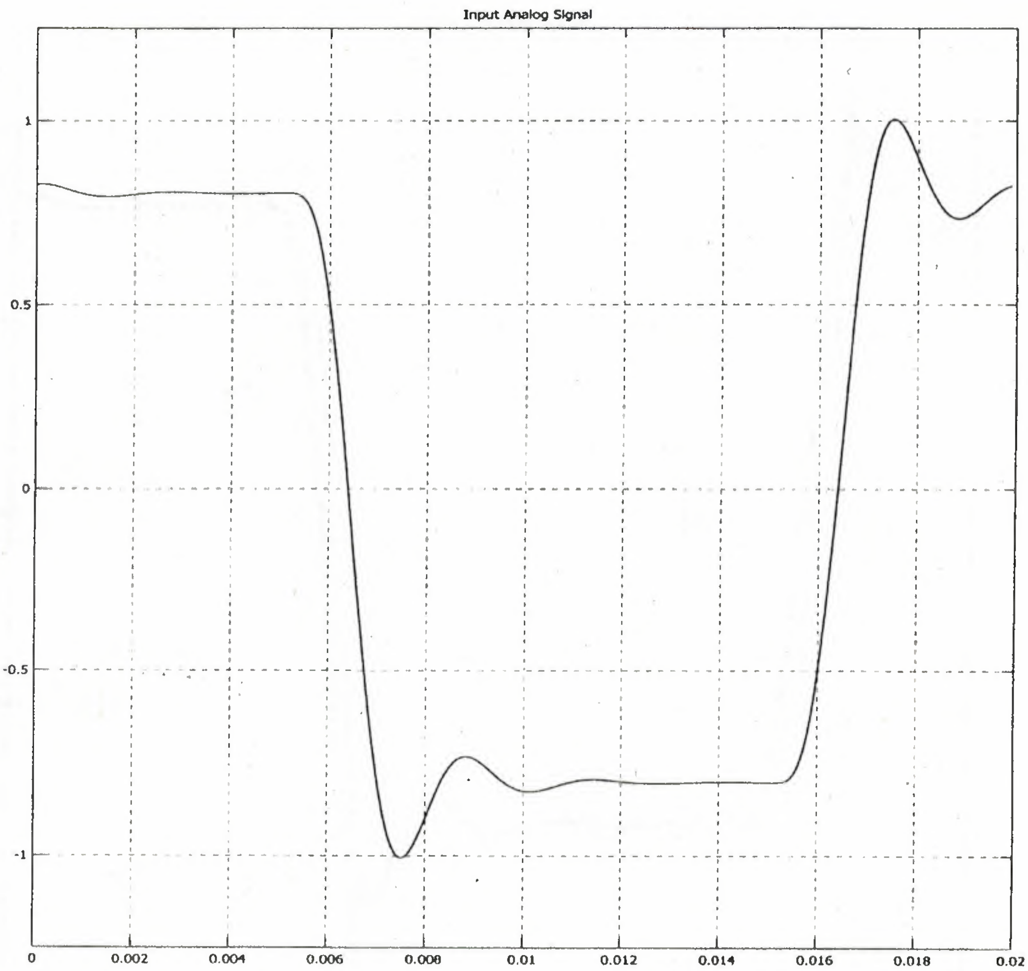


Figure 5.11a Input Analog Signal of ADPCM System.

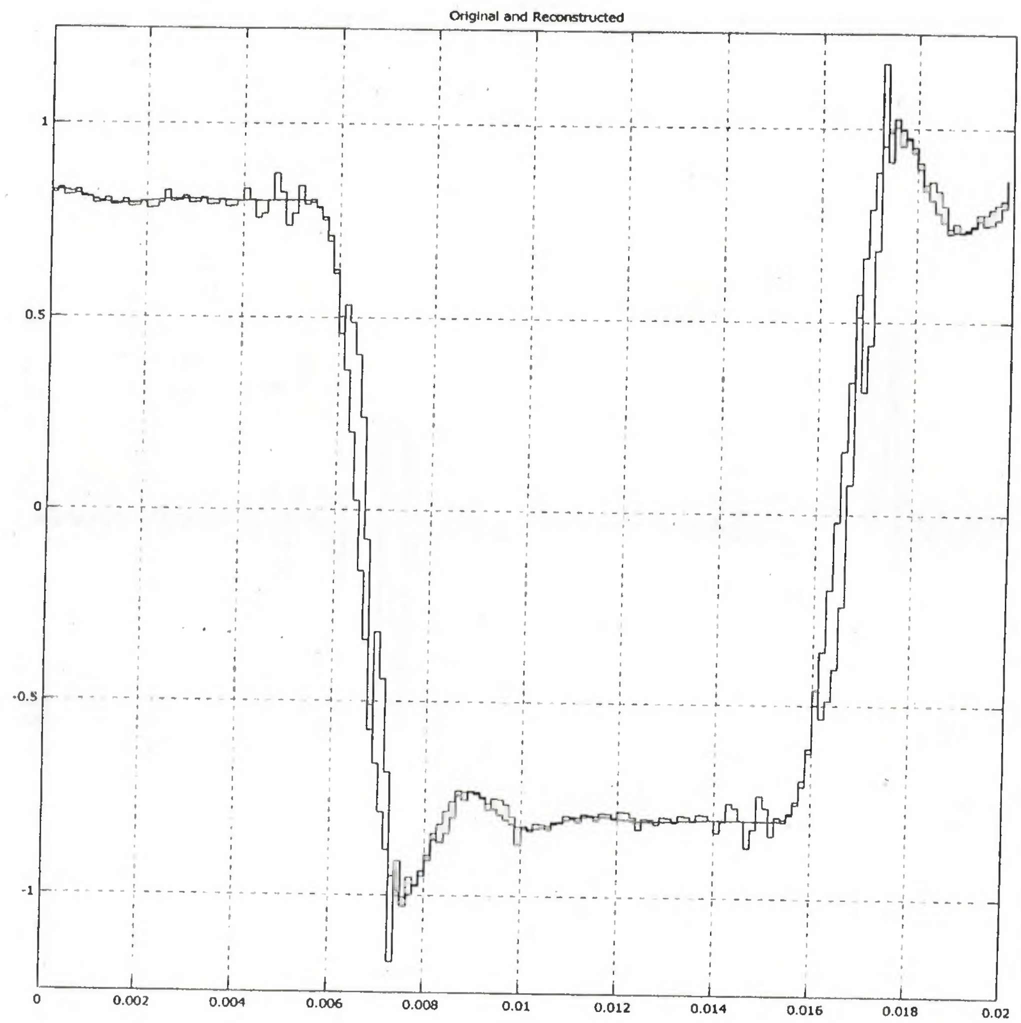


Figure 5.11b Original and Reconstructed Signals of ADPCM System.

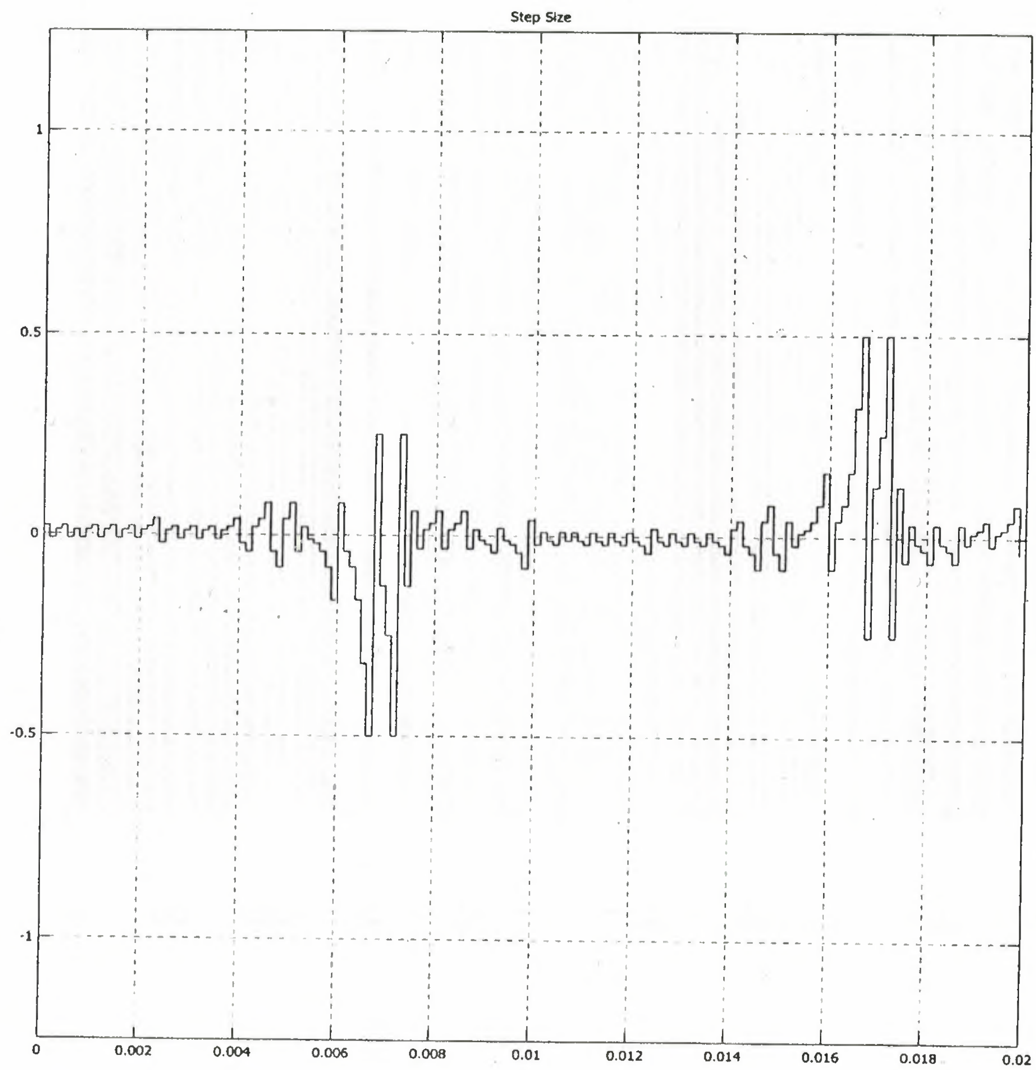


Figure 5.11c Step Size of ADPCM System.

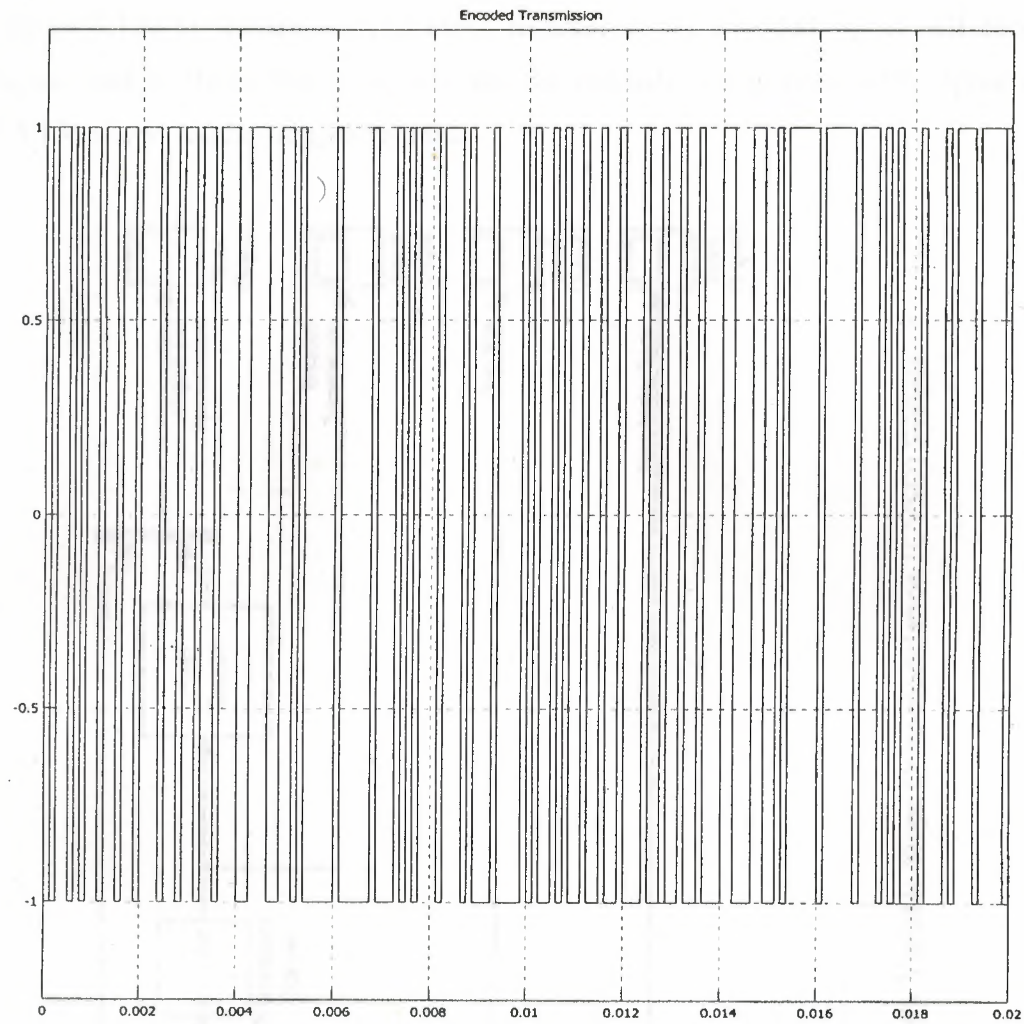


Figure 5.11d Encoded Signal of ADPCM System.

In figure 5.12 we see the block diagram of continuously variable slope delta modulation (CVSD) system that consists of three parts, transmitter, transmission channel and receiver. In transmitter an analog signal as in figure 5.13a pass to A/D converter and converted to digital and this signal pass to encoder and the encoded signal will be sent as in figure 5.13d by transmission channel. In receiver the encoded signal will decode by decoder and in figure 5.13b we will see the original and reconstructed signal and figure 5.13c shows us the step size signal.

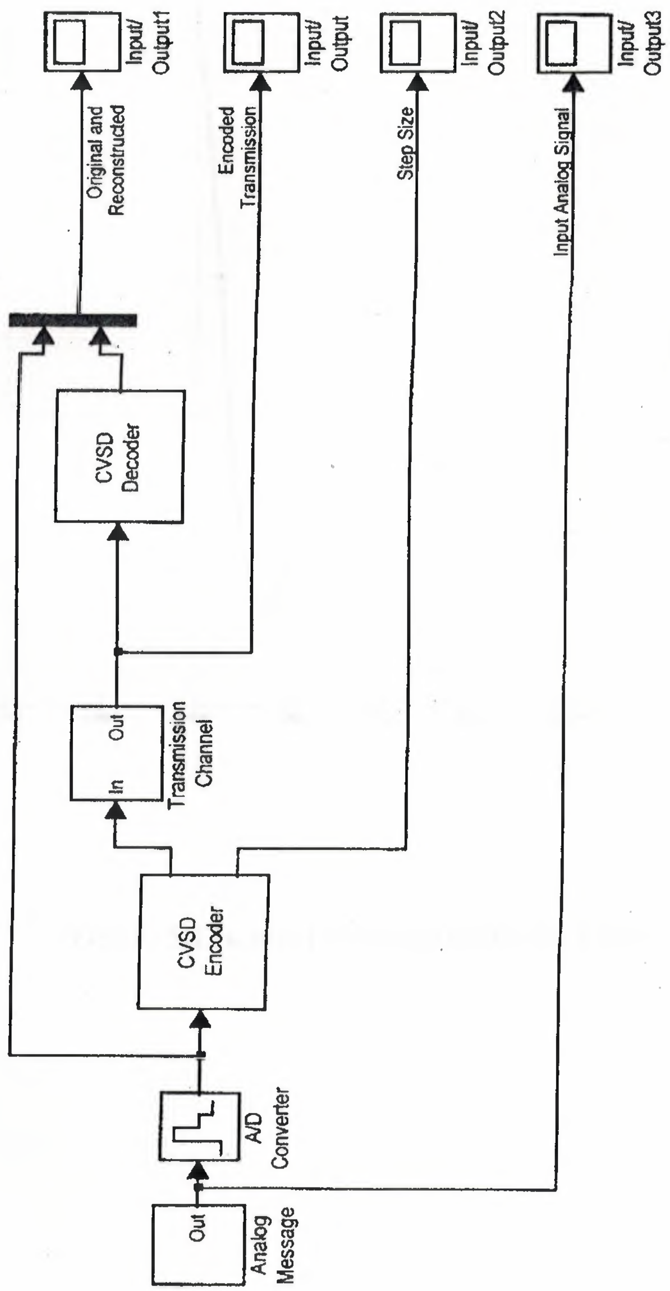


Figure 5.12 Continuously Variable Slope Delta-modulation (CVSD) System.

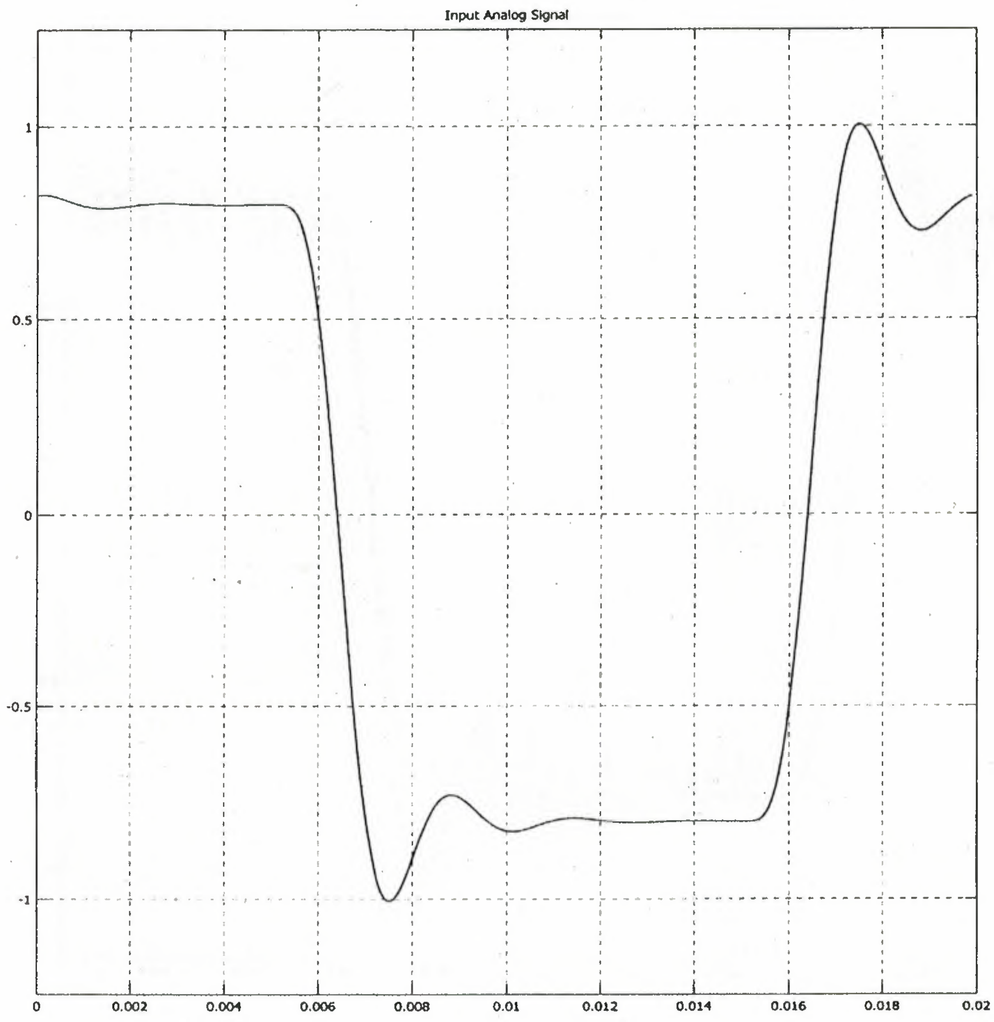


Figure 5.13a Input Analog Signal of CVSD System.

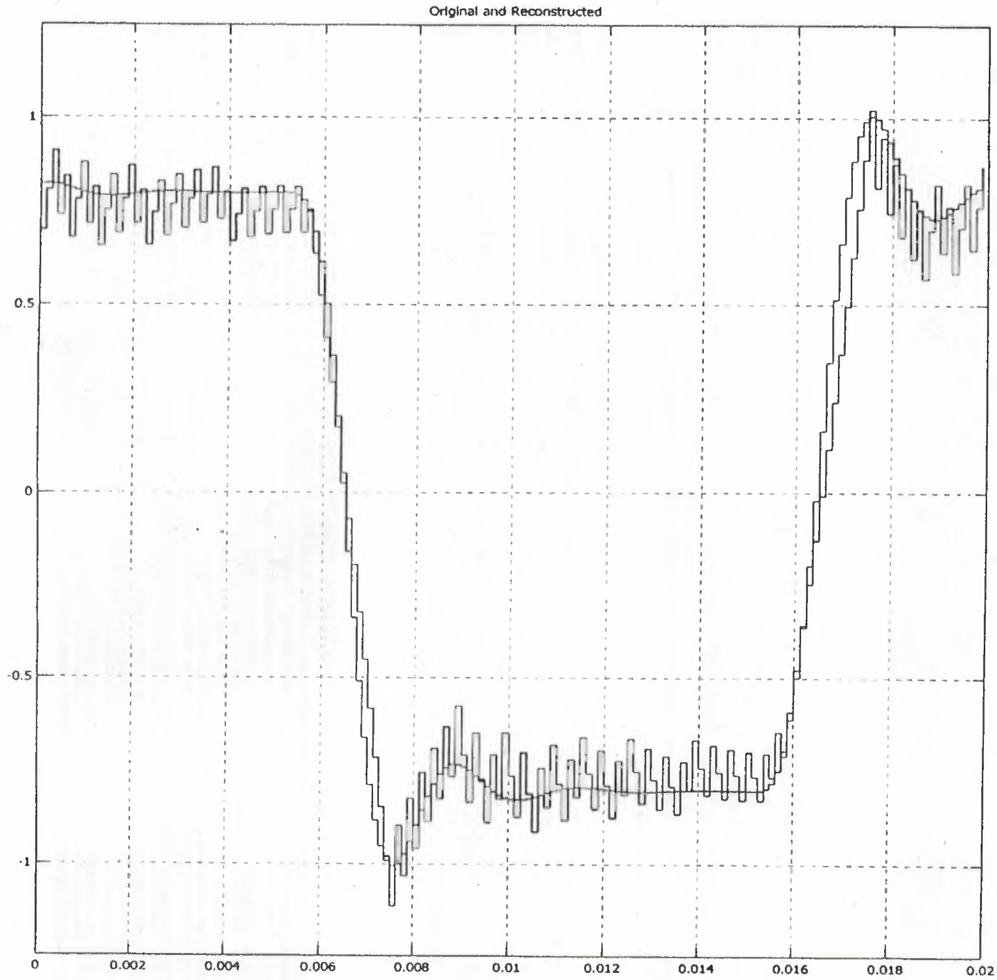


Figure 5.13b Original and Reconstructed Signals of CVSD System.

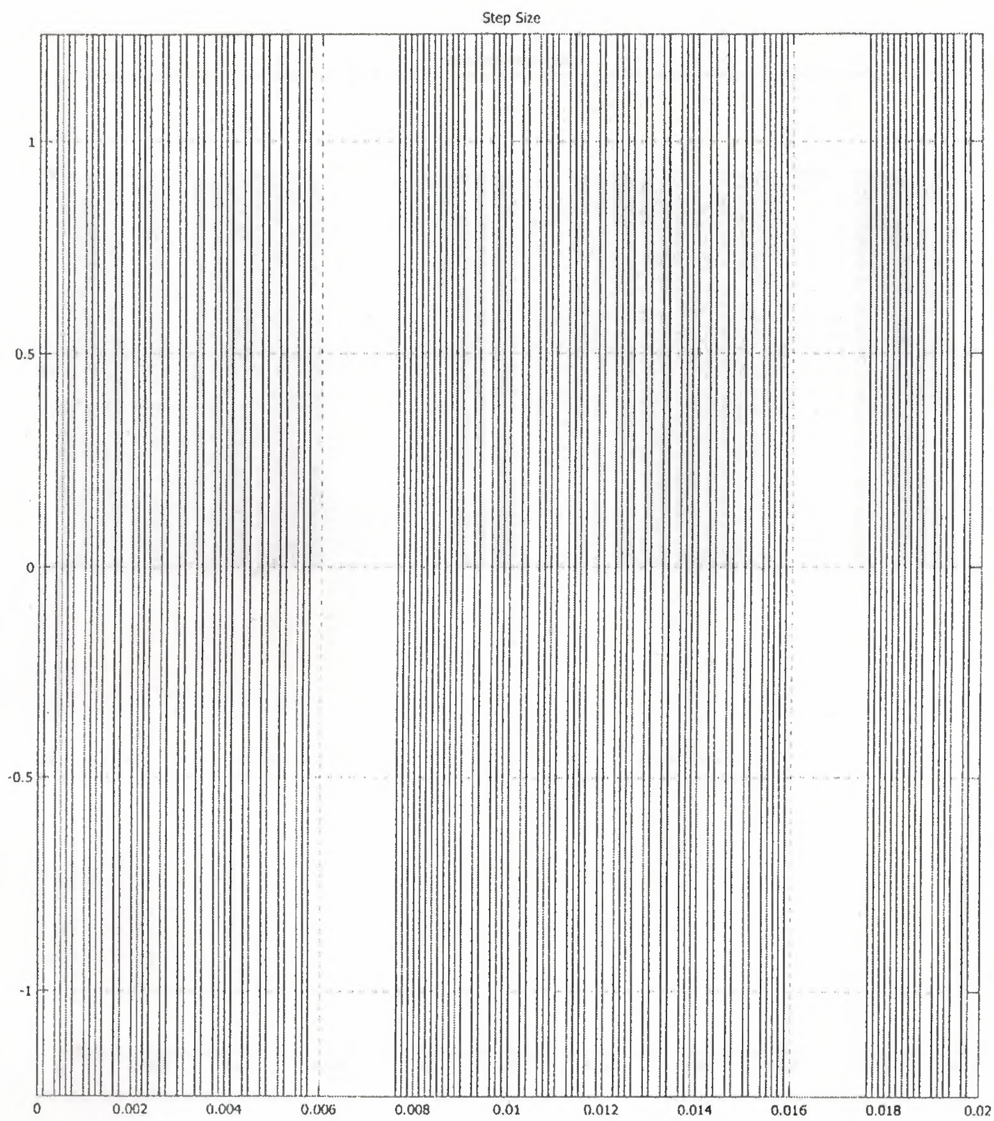


Figure 5.13c Step Size of CVSD System.

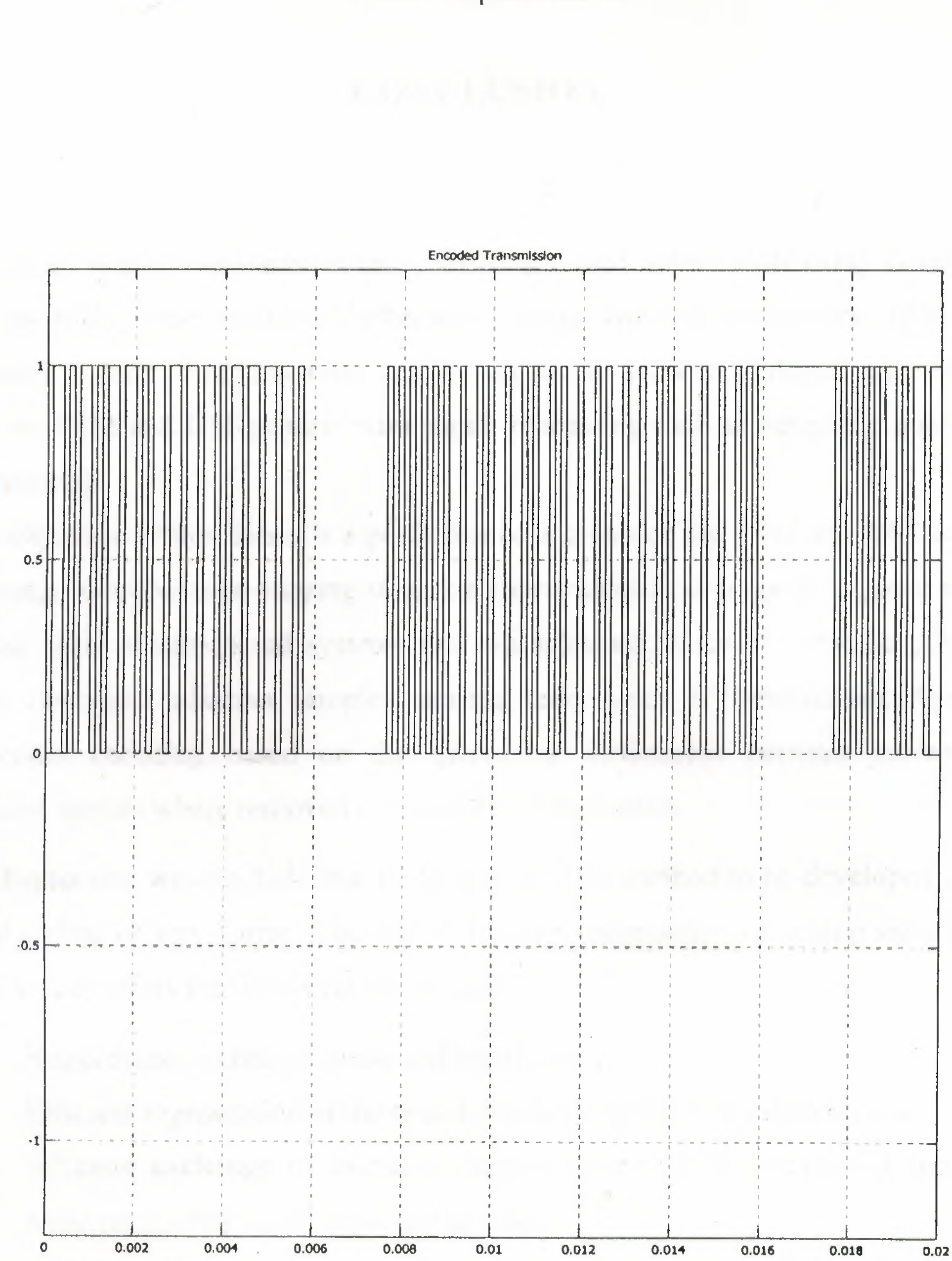


Figure 5.13d Encoded Signal of CVSD System.

CONCLUSION

Analysis of digital transmission using PCM and DM shows that, many advantages such as high noise immunity effeciency using channel band-width (BW) and providing secure communication can be achieved. However, digital transmission based on PCM and DM require extra signal processing such as sampling, quantizing and encoding.

The objective of this thesis is a performance analysis of the PCM and DM systems and design adaptive time-varying step-size approximation strategy to improve signal-to-noise ratio of considered systems that we achieved, A small step-size generates highly correlated adjacent samples causing redundancy of information. Applying differential encoding based on the quantizing differences between current and predicted sample where removed redundant of information.

In chapter one we conclude that PCM was the first method to be developed for the digital coding of waveforms. The use of digital representation of analog signals (e.g. voice, video) offers the following advantages:

1. Ruggedness to channel noise and interference.
2. Efficient regeneration of the coded signal along the transmission path.
3. Efficient exchange of increase channel bandwidth for improved signal-to-noise ratio, obeying all exponential rules.
4. A uniform format for the transmission of different kinds of base-band signals; hence their integration with other forms of digital data in a common network
5. Comparative ease with which message sources may be dropped or reinsert in a time-division multiplex system.
6. Secure communication through the use of special modulation schemes or encryption.

The performance of a PCM system is influenced by two major sources of noise:

1. Channel noise, which is introduced anywhere between the transmitter output and the receiver input. Channel noise is always present, once the equipment is switched on.

2. Quantization noise, which is introduced in the transmitter and is carried all the way along to the receiver output. Unlike channel noise, quantization noise is signal-dependent in the sense that it disappears when the message signal is switched off.

In chapter two we conclude that the advantage of DPCM over standard PCM is in the neighborhood of 4-11 dB. The greatest improvement occurs in going from no prediction to first-order prediction, with some additional gain resulting from increasing the order of the prediction filter up to 4 or 5, after which little additional gain is obtained. Since 6 dB of quantization noise is equivalent to 1 bit per sample, the advantage of DPCM may also be expressed in terms of bit rate. For a constant signal-to-quantization noise ratio, and assuming a sampling rate of 8 kHz, the use of DPCM may provide a saving of about 8-16 kb/s (i.e., 1-2 bits per sample) over standard PCM.

In chapter three we designed DM system from DPCM by replacing predictor with time delay element and DM is the 1-bit version of DPCM, but here we have problem that we could not cancel the quantization errors: slope over load and granular noise distortions. We simplified the design of the transmitter in block diagram of delta-sigma modulation system by combining the two integrators 1 and 2 of figure 3.4a into a single integrator placed after the comparator, as shown in figure 3.4b and this latter form of the delta-sigma modulation system is not only simpler than that of figure 3.4a, but it also provides an interesting interpretation of delta-sigma modulation as a "smoothed" version of 1-bit pulse code modulation.

İn chapter four we conclude that a performance characteristic of adaptor quantizer and predictor with backforward and forward estimations were analysed, and supported with examples. Time varying non-uniform characteristics were achieved by controlling step-sizes in term of rate of variation of input signal. For secure transmission over radio channels, we need to reduce the rate of bits and we can get more quality signal and we can decrease the rate from 64 kb/s as in PCM to 32, 16, 8 and 4.

In chapter five, practical implementation using feedback training models and computer simulation using matlab files have shown the adequancy of theoretical novel approach.

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