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INTRODUCTION TO DIGITAL TELEPHONE WITH PULSE CODE MODULATION

Introduction

Just as the amplitude frequency, or phase of sinusoidal can be modulated with an information signal, so the amplitude, frequency or phase (or position) of pulses in pulse train can also be modulated. In the field of telecommunications the most widely used form is pulse code modulation, or PCM. This is a variant of pulse amplitude modulation in which the pulse amplitudes are transmitted in binary code.

Because of its great importance in telecommunication system PCM will be described in considerable detail in a later section. The basic of some of the other forms of pulse modulation is also covered.

ADVANTAGES OF DIGITAL COMMUNICATIONS

Digital communication systems usually represent an increase in complexity over the equivalent analogue systems. We therefore list here some of the reasons why digital communication has replaced existing analogue systems.

1- Increase demand for data transmission.

2- Increased scale of integration, sophistication and reliability of digital electrons for signal processing combined with decreased cost.

3- Facility to source code for data compression.

4- Possibility of channel coding (line, and error control, coding) to minimize the effects of noise and interference.

5-Ease with which bandwidth, power and time can be traded off in order to optimize the use of these limited resources.

6-Standardization of signals irrespective of their type origin or the service they support leading to an integrated services digital network (ISDN).

The increase in demand for voice and data connections is the principal driving force behind the growth in telecommunications.

PULSE MODULATION

In continuous wave (*CW*) modulation some parameter of a sinusoidal carrier wave is varied continuously in accordance with the message signal.

In pulse modulation some parameter of a pulse train is varied in accordance with the message signal. We may distinguish two families of pulse modulation: analog pulse modulation and digital pulse modulation.
In analog pulse modulation a periodic pulse train is used as the carrier wave and some characteristic feather of each pulse (e.g., amplitude duration or position) is varied in a continuous manner in accordance with the corresponding sample value of the message signal.
Thus in analog pulse modulation information is transmission takes place at discrete times.
In digital pulse modulation on the other hand the message signal is represented in a form that is discrete in both time and amplitude thereby permitting its transmission in digital form as a sequence of coded pulses;
This form of signal transmission has n *cw* counterpart.
The use of this is followed by a discussion of pulse-amplitude modulation, which is the simplest form of analog pulse modulation.
coded pulses for the transmission of analog information-bearing signals represents a basic ingredient in the application of digital communications.
The sampling process is basic to all pulse modulation systems.

PULSE-AMPLITUDE MODULATION

Now that we understand the essence of the sampling process, we are ready to formally define pulse-amplitude modulation, which is the simplest and most basic form of analog pulse modulation. In pulse amplitude modulation (PAM), the amplitudes of regularly spaced pulses are varied in proportion in the corresponding sample values of a continuous message signal, the pulses can be of a rectangular form or some other appropriate shape.

Pulse-amplitude modulation as defined here is somewhat similar to natural sampling where the message signal is multiplied by a periodic train of rectangular pulses.

However, in natural sampling the top of each modulated rectangular pulse varies with the message signal where as in PAM it is maintained flat.

The waveform of a PAM signal is illustrated in figure (1). The dashed curve in this figure depicts the waveform of a message signal *m(t)*, and the sequence of amplitude modulated rectangular pulses shown as solid lines represents the corresponding PAM signal *s(t)*.

There are two operations involved in generation of the PAM signal:

1-Instantaneous sampling of the message signal *m(t)* every T_s seconds, where the sampling rate $f_s=1/T_s$ is chosen in accordance with the sampling theorem.

2-Lengthening the duration of each sample so obtained to some constant value T.

In digital circuit technology these two operations are jointly referred to as (sample and hold). One important reason for intentionally lengthening the duration of each sample is to avoid the use of an excessive channel bandwidth, since bandwidth is inversely proportional to pulse duration.

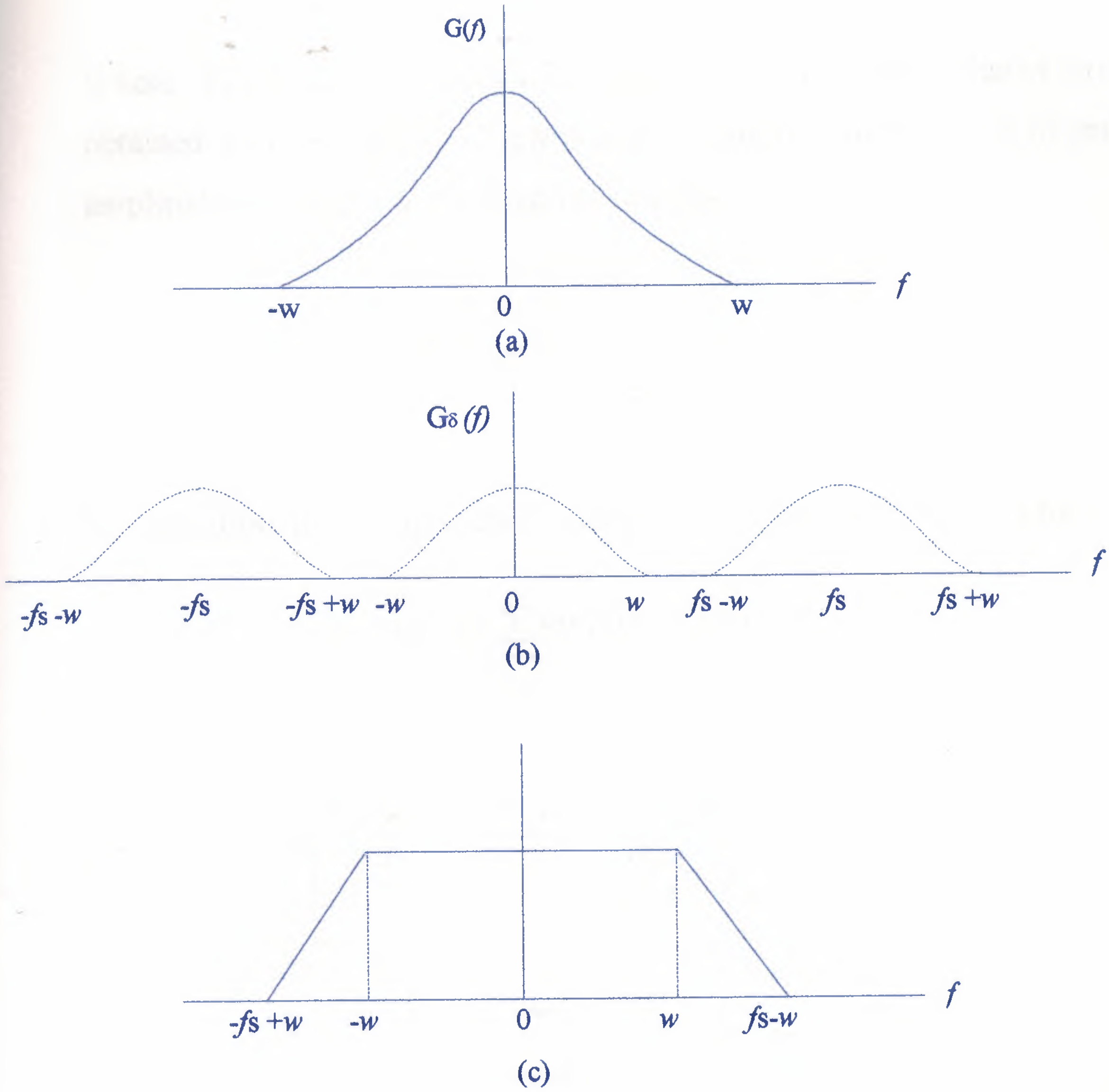


Figure 1. (a) Pre-alias filtered spectrum of an information-bearing signal. (b) Spectrum of instantaneously samples version of the signal, assuming the use of a sampling greater than the Nyquist rate. (c) Amplitude response of reconstruction filter.

However, care has to be exercised in how long we make the sample duration T, as the following analysis reveals.

Let $s\ (t)$ denote the sequence of flat-top pulses generated in the manner described in figure (1). Hence, we may express the PAM signal as:

$$s(t) = \sum_{n=-\infty}^{\infty} m(nT_s)\,h(t - nT_s)$$

Where *Ts* is the sampling period and m(*nTs)* is the sample value of m(t) obtained at time *t=nTs* . The *h(t)* is a standard rectangular pulse of unit amplitude and duration T, defined as follows

$$h(t) = \begin{cases} 1, & 0 < t < T \\ 1/2, & t = 0, t = T \\ 0, & other\ wise \end{cases}$$

By definition, the instantaneously sampled version of *m(t)* is given by

$$m_\delta(t) = \sum_{n=-\infty}^{\infty} m(nT_s)\,\delta(t - nT_s)$$

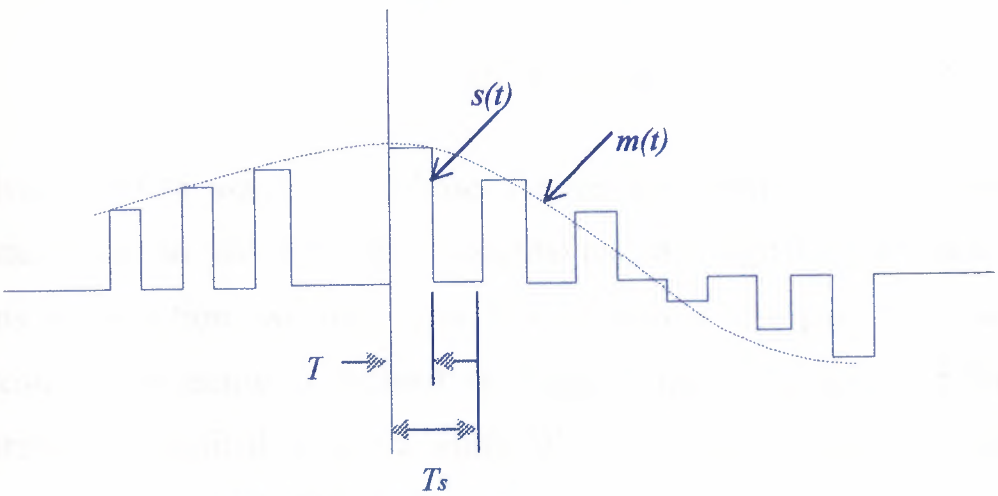


Figure 2. Flat-top samples.

Where δ*(t-nTs*) is a time shifted delta function. Therefore, convolving *m*δ *(t)* with the pulse *h(t)*,we get

$$m_\delta(t) * h(t) = \int_{-\infty}^{\infty} m_\delta(\tau)h(t-\tau)d\tau$$

$$= \int_{-\infty}^{\infty} \sum_{n=-\infty}^{\infty} m(nT_s)\,\delta\tau - nT_s)\,h(t-\tau)\,d\tau$$

$$= \sum_{n=-\infty}^{\infty} m(nT_s) \int_{-\infty}^{\infty} \delta(\tau - nT_s)\,h(t-\tau)\,d\tau$$

Using the sifting property of the delta function, we thus obtain:

$$m_{\delta}(t) * h(t) = \sum_{n=-\infty}^{\infty} m(nT_s)\delta(t - nT_s)$$

From Eqs. It thus follows that the PAM signal *s(t)* is mathematically equivalent to the convolution of ,the instantaneously sampled version of *m*δ *(t)*, the pulse *h(t)*, as shown by

$$S(t) = m_{\delta}(t) * h(t)$$

Taking the Fourier transform of both sides and recognizing that the convolution of two time functions is transformed into the multiplication of their respective Fourier transforms, we get

$$s(f) = M_{\delta}(f)H(f)$$

given a PAM signal s(t) whose Fourier transform S(*f*) is as defined in Equ. How do we recover the original message signal. As a first step in this construction, we may pass s(t) through a low pass filter whose frequency response is defined in figure 3 here it is assumed that the message is limited to band width W and the sampling rate *fs* is larger than the Nyquist rate 2W.

then we find that he spectrum of the resulting filter output is equal to M(*f*)H(*f*).

This output is equivalent to passing the original message signal m(t) through another low pass filter of transfer function H(*f*) .

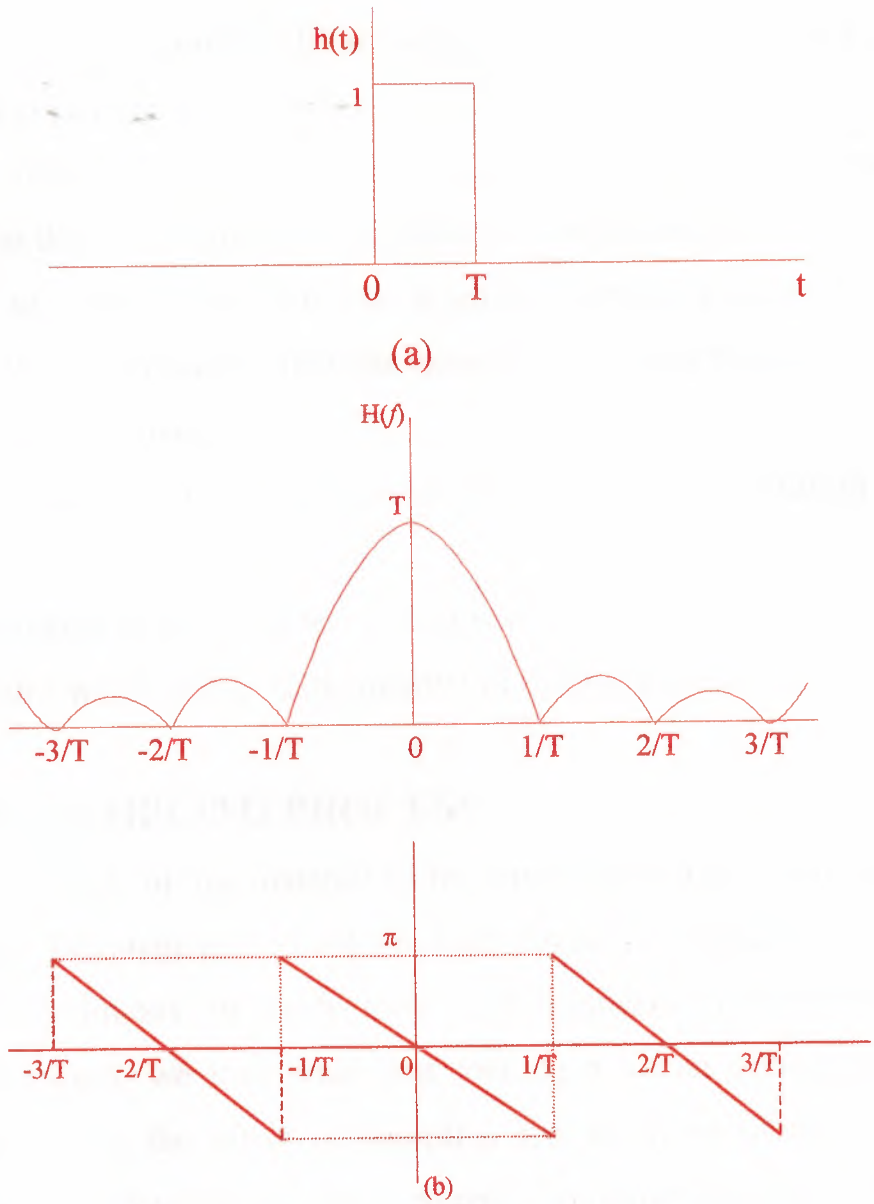


Figure 3. (a) rectangular pulse h(t).(b)spectrum H(f)

PULSE CODE MODULATION

Pulse code modulation (PCM) is the simplest and oldest wave-form by coding scheme. A pulse code modulator consists of three basic sections: a sampler, a quantizer, and an encoder. A functional block diagram of a PCM system.

The waveform entering the sampler is a band limited wave- form with bandwidth W.

Usually there exists a filter with bandwidth W prior to the sampler to prevent any component by nod W from entering the sampler.
This filter is called the pre-sampling filter. The sampling is done at a rate higher than the Nyquist rate to allow for some guard-band.
The sampled values then enter a scalar quantizer. The quantizer is either a uniform quantizer, which the results in a uniform PCM system, or non-uniform quantizer.
The choice of the quantizer is based on the characteristics of the source output.
The output of the quantizer is then encoded into a binary sequence of the length v where $N_0=2^n$ is the number of quantization levels.

THE SAMPLING PROCESS

Much of the material on the representation of signals and systems up to this stage in the book has been devoted to signals and systems that are continuous in both time and frequency. On the basis of this observation, we may state that making a signal periodic in the time domain has the effect of sampling the spectrum of the signal in the frequency domain. We may go one step further by invoking the duality property of the Fourier transform. And thus make the observation that sampling a signal in the time domain has the effect of making the spectrum of the signal periodic in the frequency domain.
The sampling process is usually described in the time domain. As such, it is an operation that is basic to digital signal processing and digital communications.

Through use of the sampling process an analog signal is converted into a corresponding sequence of samples that are usually spaced uniformly in time.

Clearly, for such a procedure to have practical utility, it is necessary that we choose the sampling rate properly, so that the sequence of samples uniquely defined the original analog signal. This is the essence of the sampling theorem, which is derived in what follows. Consider an arbitrary signal *g(t)* of finite energy, which is specified for all time. A segment of the signal *g(t)* is shown in figure. Suppose that we sample the signal *g(t)* instantaneously and at a uniform rate once every T_s seconds. Consequently, we obtain an infinite sequence of sample spaced T_s seconds apart and denoted by $g(nT_s)$, where n takes on all possible integer values.

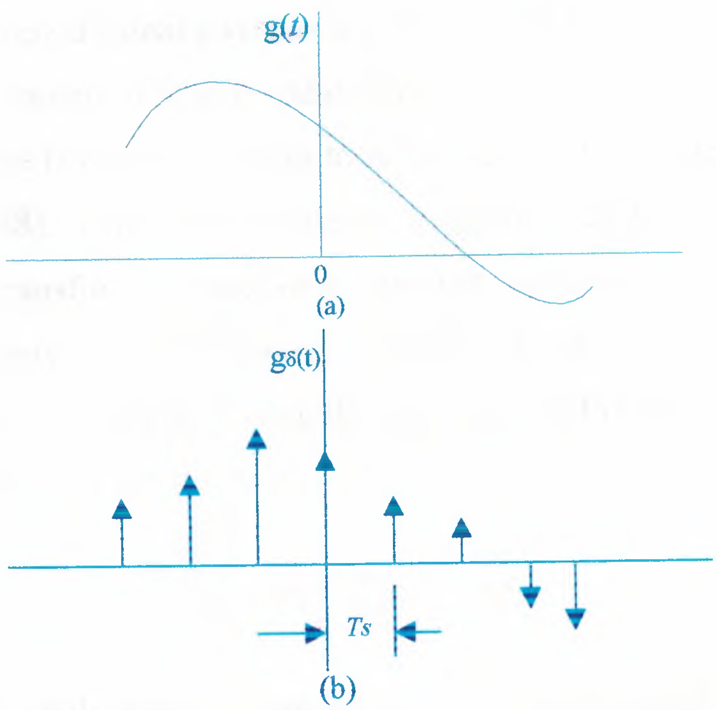


Figure (4). The sampling Process (a) analog signal. (b) instantaneously sampled version of the signal.

We refer to T_s as the sampling period, and to its reciprocal $f_s=1/T_s$ as the sampling rate. This ideal form of sampling is called instantaneous sampling. Let $g_\delta(t)$ denote the signal obtained by individually weighting the elements of a periodic sequence of Dirac delta functions spaced T_s seconds apart by the sequence of numbers $g(nT_s)$ as shown by figure (4). We refer to $g_\delta(t)$ as the ideal sampled signal. The term $\delta(t-nT_s)$ represents a delta function positioned at time $t=nT_s$.

From the definition of the delta function we recall that such an idealized function has unit area. We may therefore view the multiplying factor $g(nT_s)$ in Equation as a "mass" assigned to the delta function $\delta\ (t-nT_s)$.

A delta function weighted in this manner is closely approximated by a rectangular pulse of duration Δt and amplitude $g(nT_s)$; the smaller we make Δt the better will be the approximation.

The ideal sampled signal $g_\delta(t)$ has a mathematical from similar to that of the Fourier transform of a periodic signal. This is readily established by Equation for $g_\delta(t)$ with the Fourier transform of a periodic signal given in Equation (2.88). This correspondence suggests that we may determine the Fourier transform of the ideal sampled signal $g_\delta(t)$ by applying the duality property of the Fourier transform to the transform pair of Equation. By so doing and using the fact that a delta function is an even function of time, we get the desired result:

$$g_\delta(t) \Leftrightarrow f_s \sum_{m=-\infty}^{\infty} G(f - mf_s)$$

Where $G(f)$ is the Fourier transform of the original signal $g(t)$, and f_s is the sampling rate. States that the process of uniformly sampling a continuous time signal of finite energy results in a periodic equal to the sampling rate.

Another useful expression for the Fourier transform of the ideal sampled signal gδ(t) may be obtained by taking the Fourier transform of both sides of Equation and noting that the Fourier transform of the delta function $\delta(t-nT_s)$ is equal to $\exp(-j2f\pi T_s)$.

Let $G_\delta(f)$ denote the Fourier transform of $g_\delta(t)$. We may therefore write

$$G_\delta(f) = \sum_{n=-\infty}^{\infty} g(nT_s) \ \exp(-j2\pi f T_s)$$

This relation is called the discrete time Fourier the transform.

It may be viewed as a complex Fourier series representation of the periodic frequency function $G_\delta(f)$, with the sequence of samples $g(nT_s)$defining the coefficients of the expansion. The relations, as derived here apply to any continuous time signal g(*t*) of finite energy and infinite duration. Suppose, however, that the signal g(*t*) is *strictly band limited,* with no frequency components higher than W hertz. That is the Fourier transform G(*f*) of the signal g(*t*) has the property that G(*f*) is zero for $|f| \geq W$.

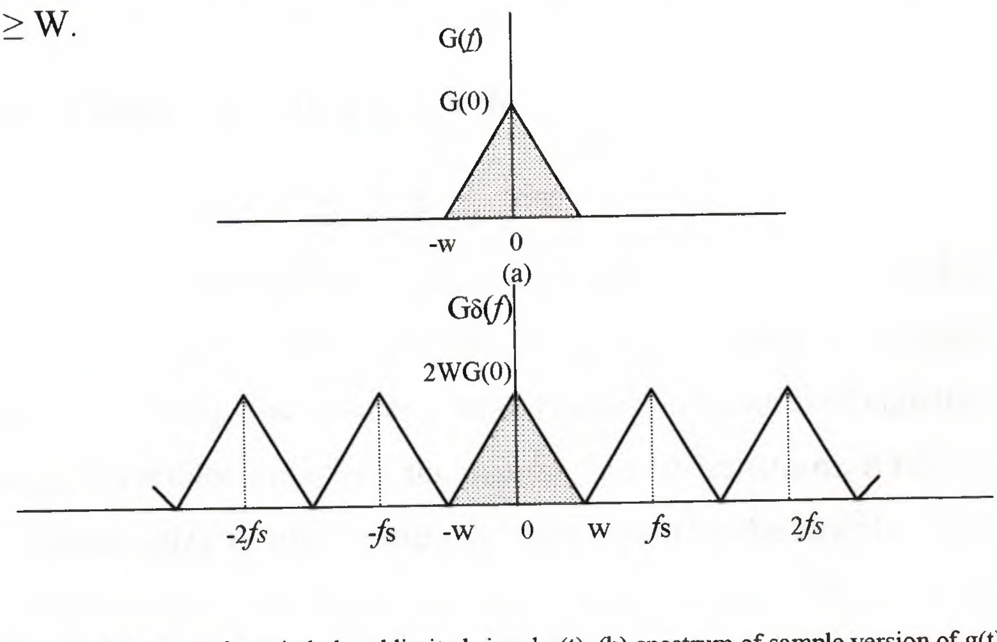


Figure (5). (a) Spectrum of a strictly band limited signal g(t). (b) spectrum of sample version of g(t) for a sampling period Ts=1/2w.

The shape of the spectrum shown in this figure in intended for the purpose of illustration only

Suppose also that we choose the sampling period $Ts = 1/2W$. Then the corresponding of spectrum $G\delta(f)$ of the sampled signal $g\delta(t)$. Putting in equation below

$$G_\delta(f) = \sum_{n=-\infty}^{\infty} g\left(\frac{n}{2W}\right)\exp\left(\frac{j\pi nf}{W}\right)$$

We readily see that the Fourier transform of $g\delta(t)$ may also be expressed as

$$G_\delta(f) = f_s G(f) + f_s \sum_{m=-\infty}^{\infty} G(f - mf_s)$$

Hence, under the following two conditions:

1- $G(f)=0$ for $f \geq W$

2- $fs=2W$

We fined from equation that:

$$G(f) = \frac{1}{2W} G_\delta(f) \qquad -W < f < W$$

Substituting equations we may also write

$$G(f) = \frac{1}{2W} \sum_{n=-\infty}^{\infty} g\left(\frac{n}{2W}\right)\exp\left(-\frac{j\pi nf}{W}\right), \qquad -W < f < W$$

Therefore, if the sample values $g(n/2W)$ of a signal $g(t)$ are specified for all time, then the Fourier transform $G(f)$ of the signal is uniquely determined by using the discrete time Fourier transform of equation. Because $g(t)$ is related to $G(f)$ by the inverse Fourier transform, it follows that the signal $g(t)$ is itself uniquely determined by the sample values $g(n/2W)$ for $-\infty<n<\infty$. In other words, the sequence $g(n/2W)$ has all the information contained in $g(t)$.

theorem as describe herein, is based on the assumption that the signal $g(t)$ is strictly band limited. In practice, however, an information bearing signal is not strictly band limited, with the result that some degree of under sampling is encountered. Consequently, some aliasing is produce by the sampling process. Aliasing refers to the phenomenon of high frequency component in the spectrum of the signal seemingly talking on the identity of a lower frequency in the spectrum of its sampled version. As illustrated in figure (6.a) the aliased spectrum shown by the solid curve in figure (6.b) pertains to an "under sampled" version of the message signal represented by the
spectrum of figure (6.a). To combat the effects of aliasing in practice, we may use two corrective measures, as described here:

1. Prior to sampling, a low pass prealiase filter is used to attenuate those high frequency components of the signal that are not essential to the information being conveyed by the signal.
2. The filtered signal is sampled at a rate slightly higher then the Nyquist rate.

ALIASING

Figure (7) shows the spectrum of an under sampled base band signal ($f_s<2f_H$). The base band spectrum of $g(t)$ clearly cannot be recovered exactly, even with an ideal rectangular low pass filter. The best achievable, in terms of separating the base band spectrum from the adjacent replicas, would be to use a rectangular low pass filter with a cut off frequency of $f_s/2$. The filtered signal will then be, approximately, that of the original signal $g(t)$ but with the frequency above $f_s/2$ folded back so that they actually appear below $f_s/2$. (The approximation becomes better as the width of the sampling pulses gets smaller. In the limit of ideal (impulse) sampling the approximation becomes exact. The spectral components originally representing high frequencies now appear under the alias of lower frequencies. The sampling represents a high frequency To avoid aliasing a low pass anti-aliasing filter with a cut off frequency $f_s/2$ is often placed immediately before the sampling circuit.

Whilst this filter may remove high frequency energy from the information signal the resulting distortion is generally less than that introduced if the some energy aliased to incorrect frequencies by the sampling process.

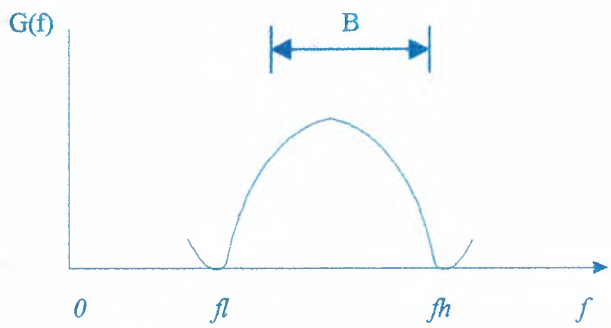


Figure (7.a) Definitions of base band and band pass signal

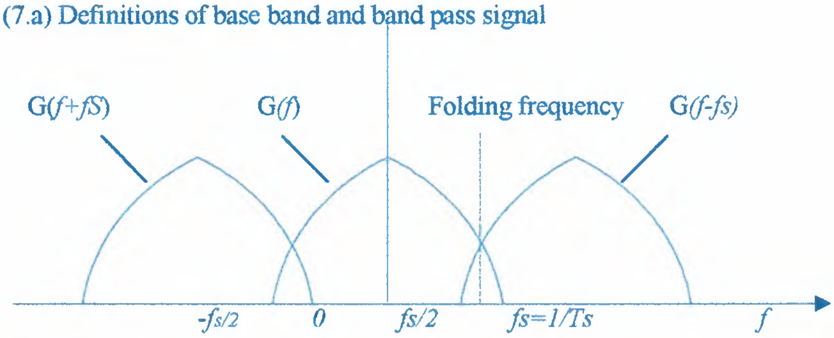
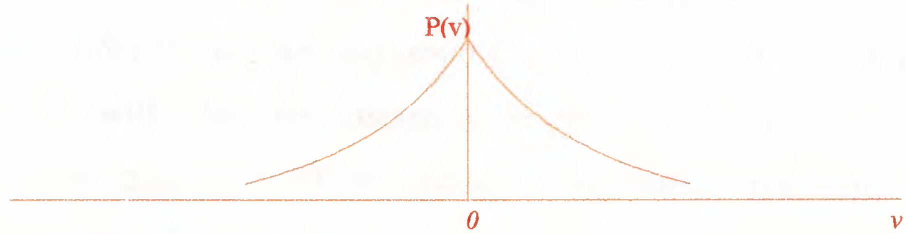


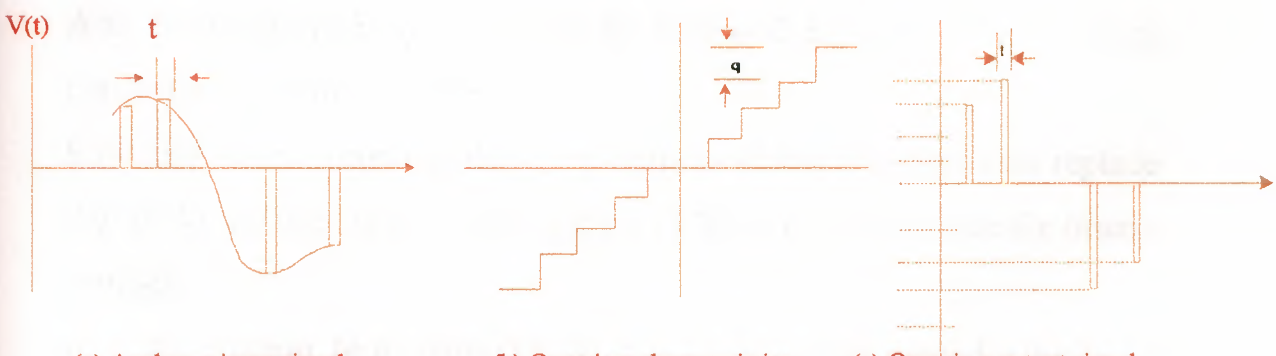
Figure (7.b) spectrum of under sampled information signal showing interference between spectral replicas and the folding frequency

QUANTISED PULSE AMPLITUDE MODULATION

An information signal which is the pulse amplitude modulated becomes discrete (in time) rather than continuous but nevertheless remains analogue in nature since all pulse amplitudes within a specified range are allowed . An alternative way of expressing the analogue property of a PAM signal is to say that the probability density function (pdf) of pulse amplitudes is continuous. If a PAM signal is quantised, i.e. each pulse is adjusted in amplitude to coincide with the nearest of a finite set of allowed amplitudes figure (8) then the resulting signal is no longer analogue but digital and as a consequence has a discrete pdf as illustrated in figure (10)



Figure(8) Continuous pdf of typical analogue PAM signal



(a) Analogue input signal (b) Quantiser characteristics (c) Quantiser output signal

Figure(9) Quantisation of a PAM signal.

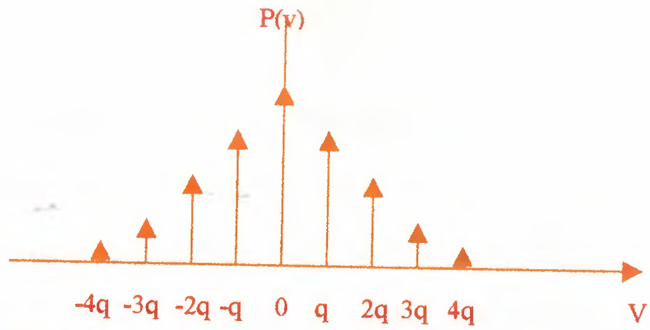


Figure (10) Discrete pdf of quantised PAM sgnal

This digital signal can be represented by a finite set of symbols-the obvious set consisting of one symbol for each quantisation level. Other symbol sets (or alphabets) can be conceived, however, and unique mappings from one set to another established.

Probably the simplest and most important alphabet is the binary set consisting of two symbols only, usually denoted by 0 and 1.

The subsystem of principle importance her are the quantizer. Pulse code modulation (PCM) encoder and source coder, although digital pulse multiplexing will also be discussed briefly. To some extent the separation of quantizer, PCM encoder and source coder might be misleading since the do not necessarily exist as identifiable separate pieces of hardware in all digital communications systems, for example, quantization and PCM encoding are implemented together as a binary A/D converter (which may than be followed by a parallel to series converter) in many systems.

Similarly some source coders (e.g. delta modulators) effectively replace the PCM encoder whilst others take a PCM signal and recode the binary symbols.

In some system (e.g. delta PCM) the source coder precedes the PCM encoder.

Quantizing PAM signals is usually a precursor to generating pulse code modulation (PCM) which has some significant advantages over other base band modulation types. The quantization process in it self, however, actually degrades the quality of the information signal. This is easy to see since the quantized PAM signal no longer exactly represents the original, continuous analogue, signal but a distorted version of it . The quantized signal can be decomposed into the some of the analogue signal and the difference between the quantized and the analogue signals. The difference signal is essentially random and can therefore be thought of as a special type of noise process. Like any other signal the power or RMS value of this quantization noise can be calculated or measured. This leads to the concept of a signal quantization noise ratio (SNqR).

COMPRESSION

With speech it is found that the peaks of the signal only infrequently extend over the full range of the input most of the time residing within a small range about zero.

Figure (11) addition of a compressor stage

In effect the signal does not have uniform probability density function and a result the $(S/N)_q$ ratio is lower than that given . To compensate for this further stage termed a compressor, this added. This is shown in figure (11)

In earlier design the compressor consisted of an analog amplifier that had variable gain characteristics with a lower gain at higher input so that, in effect the high level signal where 'compressed'. This compressed signal was passed on to a linear quantizer of the type described in the previous section. The number of quantized levels was chosen to gave the required

$(S/N)_q$ ratio for the low signal ranges and the peak signals swings where compressed to fit into these. This approach give rise to two compression characteristics which are now fairly will standardize. In North America and Japan and many other parts of the word low characteristics is used.

The compressions function are normally described in terms of normalized voltages. Let *V1* represent the input voltage and $V1_{max}$ its maximum value.

Denoting the normalized in voltage by X then

$$X = \frac{V1}{V2_{max}}$$

In a similar manner the normalization output voltage is defined as:

$$Y = \frac{V0}{V0_{max}}$$

In terms of normalized voltages the μ-law is described by

$$y = sign(V1)\frac{\ln(1+\mu|x|)}{\ln(1+\mu)}$$

Where sign(*V1*) is used to indicate the sign or polarity of *V1* and |x| is the magnitude of X. The compression parameter is μ which determines the degree of compression with μ=0, he limiting values of the logarithmic function must be used to show that *V0=V1* or no compression occurs. The value μ =255 is widely used and the characteristics for this value is shown in figure (12).

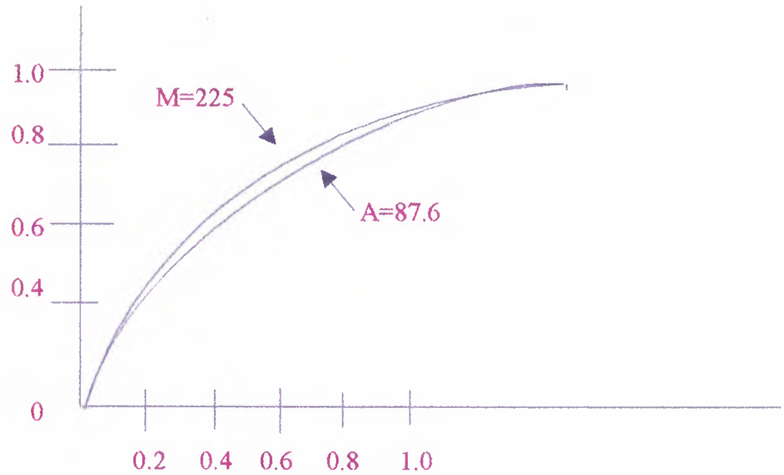
The A law is described by :

$$y = sign(V1)\frac{A|X|}{1+\ln(A)} \qquad for|X| \le \frac{1}{A}$$

Or

$$y = sign(V1)\frac{1+\ln(A|X|)}{1+\ln(A)} \qquad for\frac{1}{A} \le |X| \le 1$$

the law and A law system are incompatible and special conversion units are used where inter connections are required, such as might occur on international links.



Figure(12) compressor characteristics forμ=255 and A=87.6

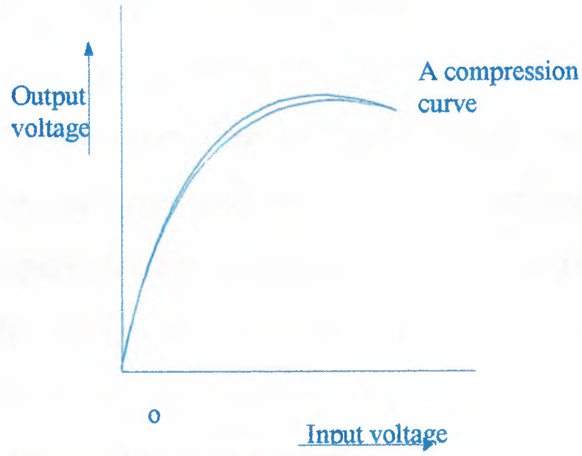
Although the compressor stage is shown as a separate block and in early PCM circuit it was implemented using analog techniques, in more recent equipment it is implemented as part of the A/D conversion.

Rather than having a continuous curve the compressor characteristics is approximated by linear segments as sketched in figure (13).

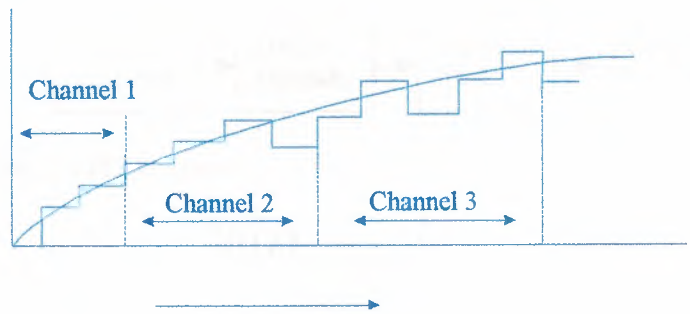
Each chord is made to cover the some number of input steps put the step size increases from chord to chord.

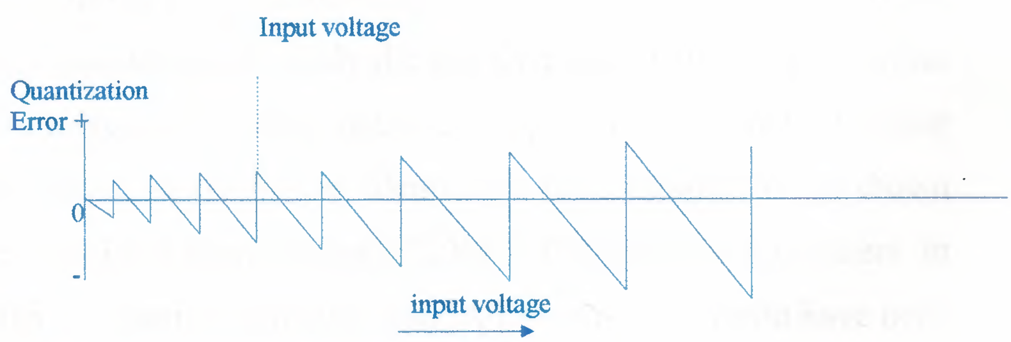
This is equivalent to having a nonlinear quantizing stage as illustrated in figure (14.a) which shows the first three chords of hypothetical compressor. Figure (14.b)shows the quantization error and this is seen to increases in amplitude for the larger steps.

Such a compressor might quantize the analog input voltage. As before the leading bit can be used to encode analog polarity. With four chords 2bits are required to encode these for example in ascending orders as 00011011. A further 2 bits are required to encode the step numbers within a chord, and again in ascending numbers these might be 00011011. Thus sample point A would be encoded as 11110 and sample point B as 01010



Figure(13) chorded approximation to a compression curve





Figure(14.a.b) Nonlinear quantization

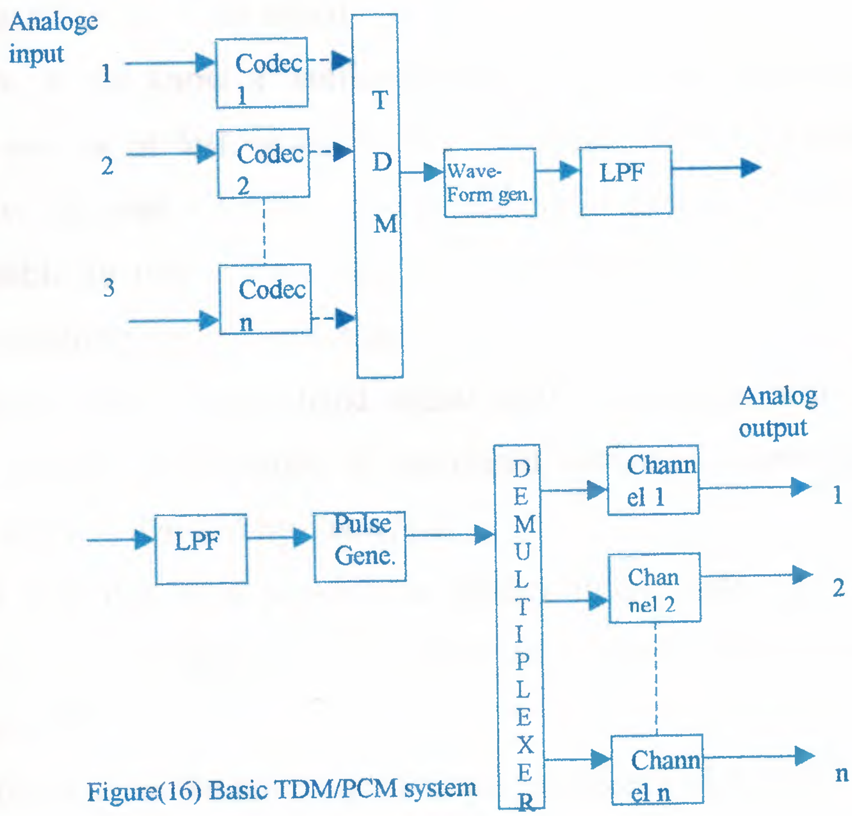
PCM RECEIVER

The receiving section of a coded must provide the inverse operation to those of the transmitter . figure (15) summarizes the main receiver blocks . the function of the input filter is to limit the noise bandwidth and to complete the waveform shaping required for the avoidance of ISI . a pulse regenerator is used to generate new pulses that are free of thermal noise (but note that quantization noise is always present and cannot be removed). The digital -to-analog converter (dac) converts the binary signal into flat-top samples and in the process provides the expansion necessary to compensate for the compression applied at the transmitter. the equalizer filter following the DAC compensates for the aperture distortion introduced by flat -top sampling as described . the magnitude of the equalizer filter response is given by

Figure(15) basic blocks in a PCM receiver

$$|H(f)|_{eq} = \frac{A}{\sin c \quad fT}$$

Where A is constant. The equalizer filter is followed by a low-pass filter often referred to as reconstruction filter, which essentially recovers the analog signal by passing only the low frequency part of the spectrum as shown however quantization noise is also present on the output so that the analog output m"(t) is not identical to m(t) .figure (15) , as shown how codes could be used in simple TDM / PCM transmitting system .in early design a common sampler and A/D converter would have been used for the multiplexed channels but width advances in integrated circuits the used of an individual coded for each channel is the most economical approach for digital telephony .thus time division multiplexed channels but width advances in integrated circuits the used of an individual coded for each channel is the most economical approach for digital telephony .Thus time division multiplexing (TDM) takes place in the multiplexer following the channel codes, as shown in figure (16). For illustration purposes a common line waveform generator is shown as a separate unit but the actual arrangement will depend on the facilities provided in the codes and multiplexer.



Figure(16) Basic TDM/PCM system

DIFFERENTIAL PULSE CODE MODULATION

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 dose 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 generated as a result of the encoding process. By removing this redundancy before encoding, we obtain a more efficient coded signal.

Now, if we know a sufficient part of a redundant signal, we may infer the rest or at lest make the most probable estimate. In particular, if we know the past behavior of a signal up to a certain points 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 (17).

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

$$e(nT_s)=m(nT_s)-m(nT_s)$$

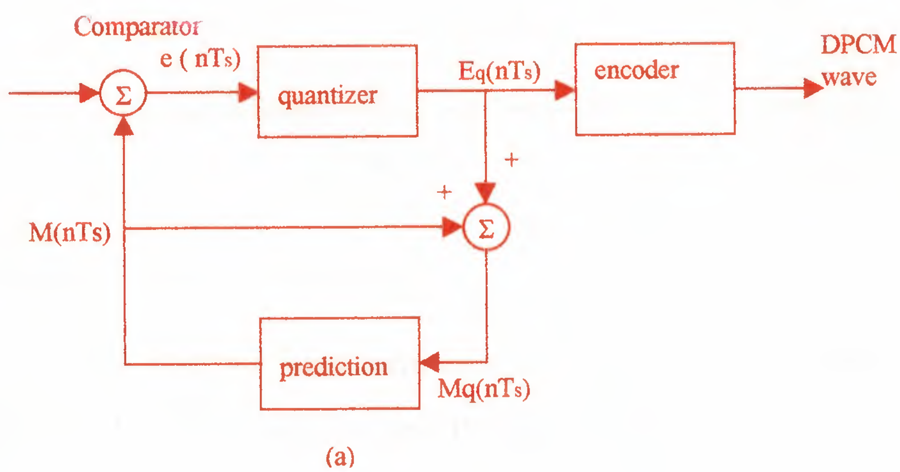
Which is the difference between the unquantized input sample $m(nT_s)$ and a prediction of it, denoted by $m(nT_s)$

This predicted value is produced by using a prediction filter whose input, as we will see, consists of a quantized version of the input sample $m(nT_s)$. The difference signal $e(nT_s)$ 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,* which the basic delay set equal to the sampling period.

The block diagram of this filter is shown in figure (18) according to which the prediction $m(nT_s)$ 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 (17.a), we obtain a variation of PCM, which is known as differential pulse-code modulation (DPCM). It is this encoded signal that6 is used for transmission.



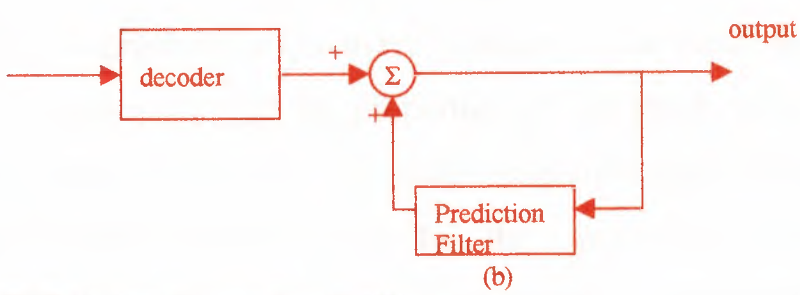


Figure (17) DPCM system. (a)transmitter. (b)receiver.

The quantizer output may be expressed as,

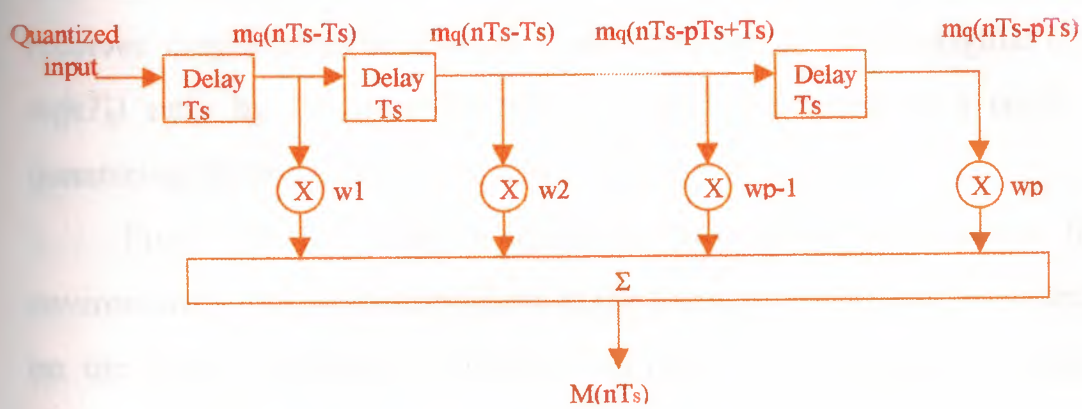
$$eq(nT_s) = e(nT_s) + q\ (nT_s)$$

Where $q(nT_s)$ is the quantization error. The quantizer output $e_q(nT_s)$ is added to the predicted value $m(nT_s)$ to produce the prediction filter input

$$m_q(nT_s)=m(nT_s)+e_q(nT_s)$$

Substituting equations, we get

$$m_q(nT_s)=m(nT_s)+e(nT_s)+q(nT_s)$$



Figure(18). Tapped-delay-line filter used as a prediction filter

However, we observe that the sum term $m(nT_s)+e(nT_s)$ is equal to the input signal $m(nT_s)$. Therefore, we may rewrite equation as:

$$m_q(nT_s)= m(nT_s)+q(nT_s)$$

Which represents a quantized version of the input signal $m(nT_s)$. That is, irrespective of the properties of the predication filter, the quantized signal $m_q(nT_s)$ at the predication filter input difference from the signal input signal $m(nT_s)$ by the quantaizaion error $q(nT_s)$. According, if the prediction is good, the variance, of the prediction error $e(nT_s)$ will be smaller than the variance of $m(nT_s)$ 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 standard PCM system.

The receiver for reconstructing the quantaized version of input is shown in figure (17.b). It consists of a decoder to reconstruct the quantaized error signal. The quantized version of the signal input is reconstructed from the decoder output using the same predication filter used in transmitter of figure (17.a). 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 quantizetion 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 simples, $m_q\ (nT_s)$. It is with this purpose in mined that a feedback path is added to the quantizer in the transmitter, as shown in figure (17.a).

Differential pulse code modulation includes delta modulation as a special case. In particular, comparing that DPCM system of figure (17) with the DM system, we see that they are basically similar, except for two important difference, the use of a one pit (two level) quantizer in the delta modulator, and the replacement of the prediction filter by a signal delay element (i.e., zero prediction order). Simply put, DM is the 1-bit version of DPCM and DM involves 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 quantizetion noise.

DELTA MODULAION

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 sample quantizing strategy for constructing the encoded signal.

In its basic form, DM provides a *staircase approximating* to the over sampled version of the message signal, as illustrated in figure (19). The difference between the input and the approximation is quantized into only two levels, namely, $\pm\Delta$, corresponding to positive and negative difference, respectively. Thus, if the approximation falls bellow 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 dose 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:

$$e(nT_s) = m(nT_s) - m_q(nT_s - T_s)$$
$$e_q(nT_s) = \Delta \operatorname{sgn}[e(nT_s)]$$
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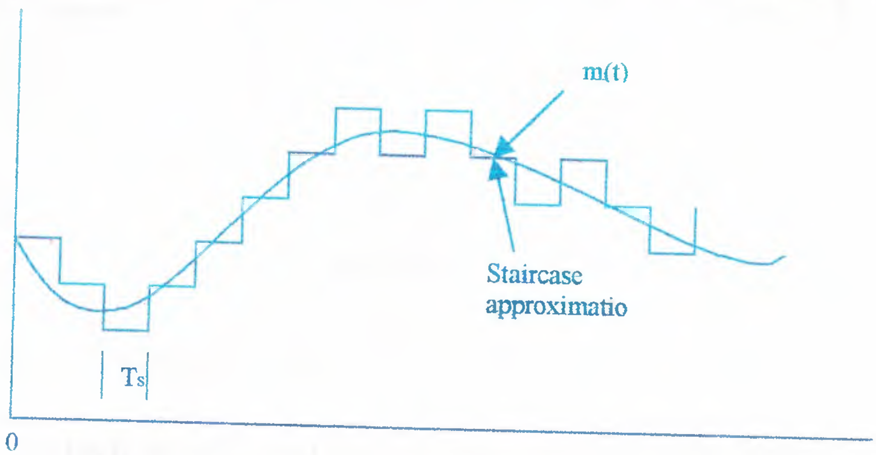
Where T_s is the sampling period; $e(nT_s)$, is an error signal representing the difference between the present sample value $m(nT_s)$ of the input signal and the latest Approximation to it, that is, $m(nT_s)-m_q(nT_s-T_s)$; and $e_q(nT_s)$ is the quantized version of $e_q(nT_s)$.

The quantizer output $e_q(nT_s)$ is finally coded to produce the desired DM signal.

Figure (19) illustrates the way in which the approximation $m_q(t)$ follows variations in the input signal $m(t)$ in accordance with equations and figure (19) displays the corresponding binary sequence at the delta modulator output. It is apparent that in delta modulation system the rate of information transmission is simply equal to the sampling rate $f_s=1/T_s$.



Figure(19). Illustration of delta modulation

The principal virtue of delta modulation is its simplicity. It may be generated by applying the sampled version of incoming message signal to a modulator that involves a *comparator, quantizer,* and *accumulator* interconnecting as shown in figure (20.a). Details of the modulator follow directly from equation .The comparator computes the difference between its tow inputs. The quantizer consists of a *hard limiter* with an input-output relation that is a scaled version of the signup function. The quantizer output is then applied to an accumulator, producing the result

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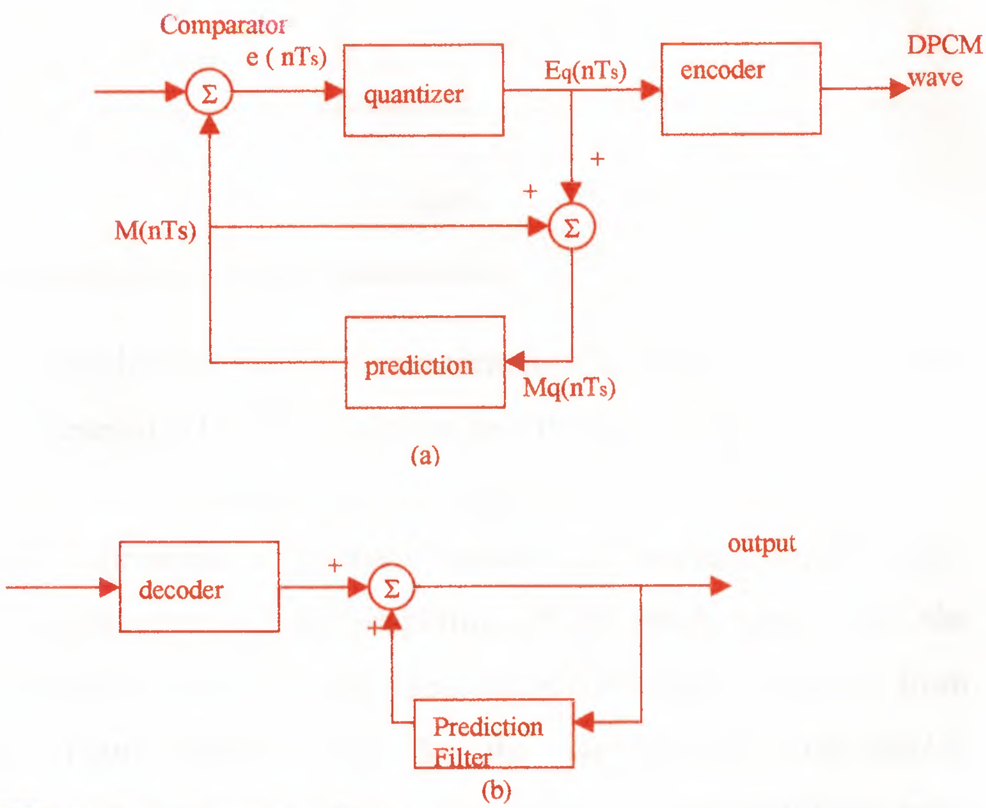

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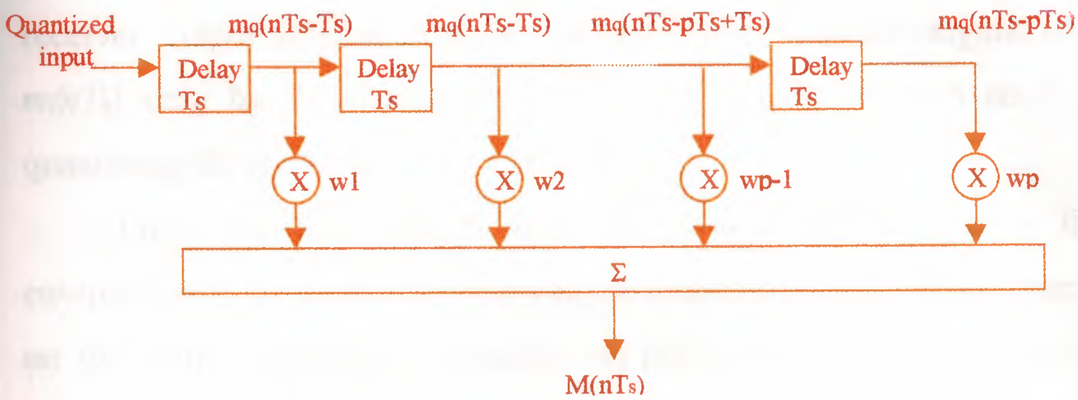

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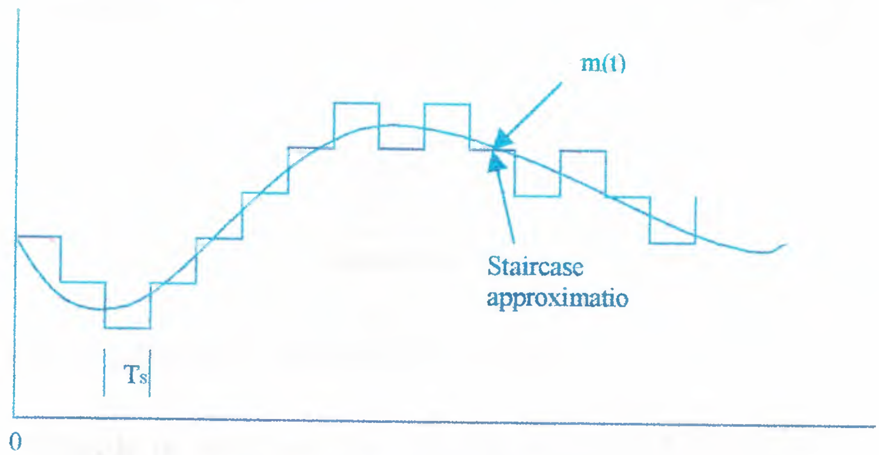

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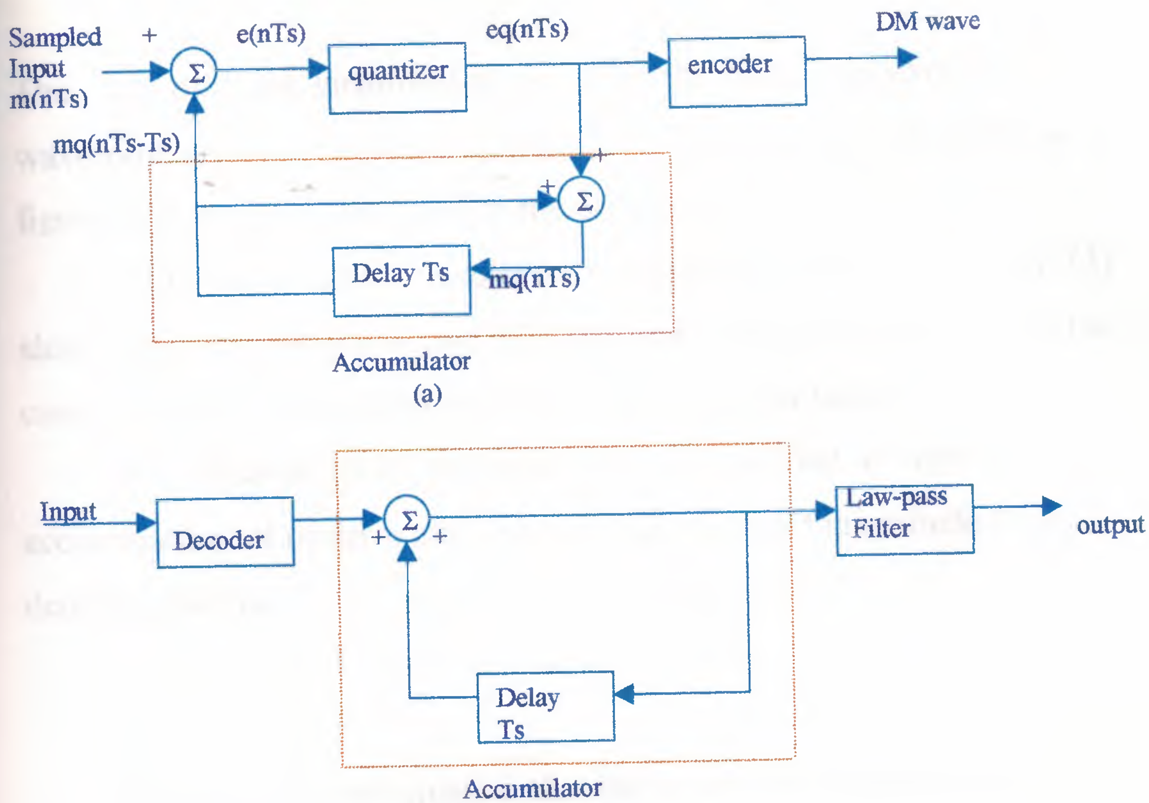


(a)

Figure(20). DM system. (a)Transmitter(b) Receiver

Which is obtained by solving equations 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 a 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 dose the best it can to track the input samples by one step at a time. In the receiver shown in figure (20.b), 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 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 *mq(t)* is rejected by passing it through a law pass filter, as in figure, with a bandwidth equal to the original message bandwidth.

Delta modulation is subject to two types of quantization error: (1) slope overload distortion and (2) granular noise. We first discuss the cause of slope overload distortion, and than granular noise.

We observe that equation is 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,

$$m_q(nT_s) = m(nT_s) + q(nT_s)$$

We observe from equation that the input to the quantizer is:

$$e(nT_s) = m(nT_s) - m(nT_s - T_s) - q(nT_s - T_s)$$

Thus, except for the quantization error $q(nT_s - T_s)$, the quantization 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 increases 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

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

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 $m(t)$. This condition is called *slope over load,* 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 modulator using a fixed step-size is often referred to as a *linear delta modulator.*

In contrast to slope-overload distortion, *granular noise* occurs when the step-size Δ it 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. 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, whereas 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 set-size that 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.

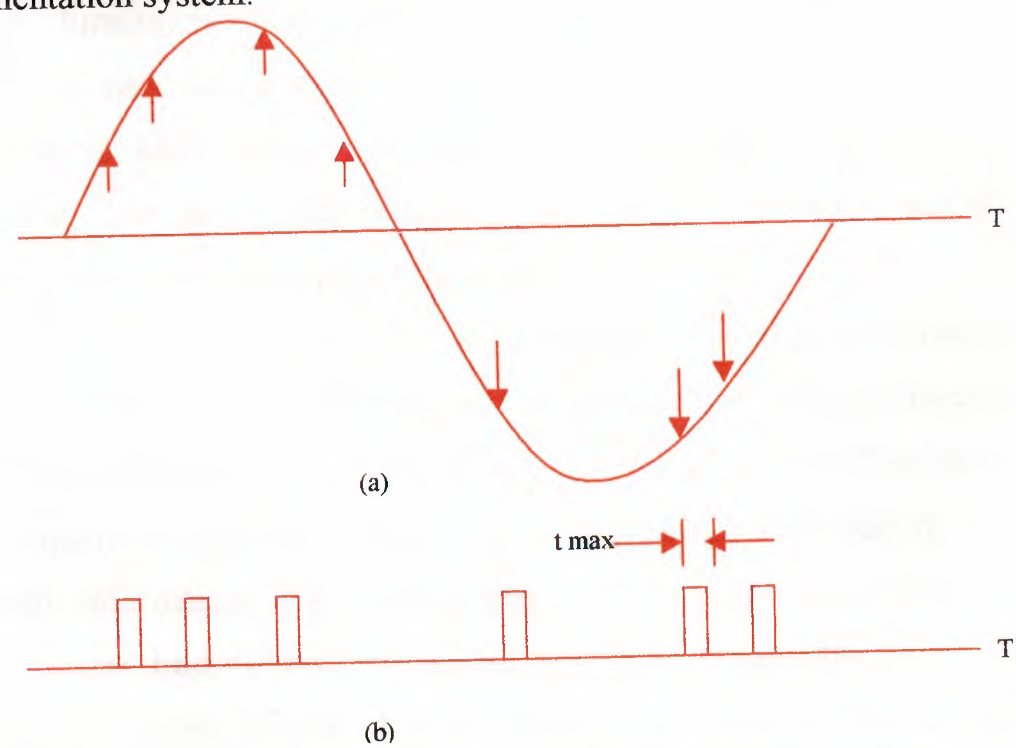
PULSE FREQUENCY MODULATION (PFM)

A train of rectangular pulses is frequency modulated if their amplitude is kept constant, and the pulses period T_c and pulse duration τ are both made proportional to the modulating signal so that the duty cycle (τ/T_c) of the pulses train remains constant.

The modulating wave is sampled at fixed intervals as was the case for PAM, but is sampled at the time of occurrence of the modulated pulses, as shown in figure (22.a).

The samples taken is used to adjust the following pulse period. The resulting frequency modulated pulse train is illustrated in figure (22.b), which contains a fixed dc level and frequency modulated of carrier and side-bands at each harmonic of the un-modulated carrier frequency $fc=1/Tc$. The amplitudes of the harmonic of the un-modulated carrier are constrained by the envelope of (sin x)/x, where $X = n\pi\tau / T$.

The spectrum dose not contains any of the base band frequency components, so the modulating signal cannot be recovered by simple low pass filtering. A frequency demodulator must be used. However the signal is easy to generate using a mixture of digital and analog components, which has made PFM popular for some types of analog instrumentation system.



Figure(22.a.b) pulse frequency modulation PFM(a) modulating signal.(b)modulated pulse train

PULSE TIME MODULATION (PTM)

Pulse time modulation includes pulse modulation (PPM) and pulse width modulation (PWM). Both of these produce a form of pulse phase modulation and are sometimes called by that name. Pulse frequency modulation (PFM) is also included, although it is not strictly a time modulation.

PULSE POSITION MODULATION

In a pulse modulation system we may use increased bandwidth consumed by pulses to obtain an improvement in noise performance by representing the sample values of the message signal by some property of the pulse other than amplitude.

In pulse duration modulation (PDM), the samples of the message signal are used to vary the duration of the individual pulses. This form of modulation is also referred to as pulse width modulation or pulse length modulation. The modulating signal may vary the time of occurrence of the leading edge, or both edges of the pulse.

In figure (23.c) the trailing edge of each pulse is varied in accordance with the message signal, assumed to be sinusoidal as shown in figure (23.a). The periodic pulse carrier is shown in figure (23.b). In PDM, long pulses expend considerable power during the pulse while bearing no additional information. If this unused power is subtracted from PDM, so that only time transitions are preserved, we obtain a more efficient type of pulse modulation known as *pulse position modulation* (PPM) the position of a pulse relative to its unmodulated time of occurrence is varied in accordance with the message signal, as illustrated in figure (23.d) for the case of sinusoidal modulation. Let T_s denote the sample duration. Using the sample $m(nT_s)$ of a message signal $m(t)$ to modulate the position of the nth pulse, we obtain the PPM signal.

$$s(t) = \sum_{n=-\infty}^{\infty} g(t - nT_s - k_p m(nT_s))$$

Where k_p is the *sensitivity* of the pulse-position modulator and $g(t)$ denotes a standard pulse of interest. Clearly, the different pulses constituting the PCM signal $s(t)$ must be *strictly non over lapping,* a sufficient condition for this requirement to be satisfied is to have

$$g(t) = 0, \qquad |t| \succ \frac{T_s}{2} - k_p |m(t)|_{\max}$$

Which, in turn, requires that

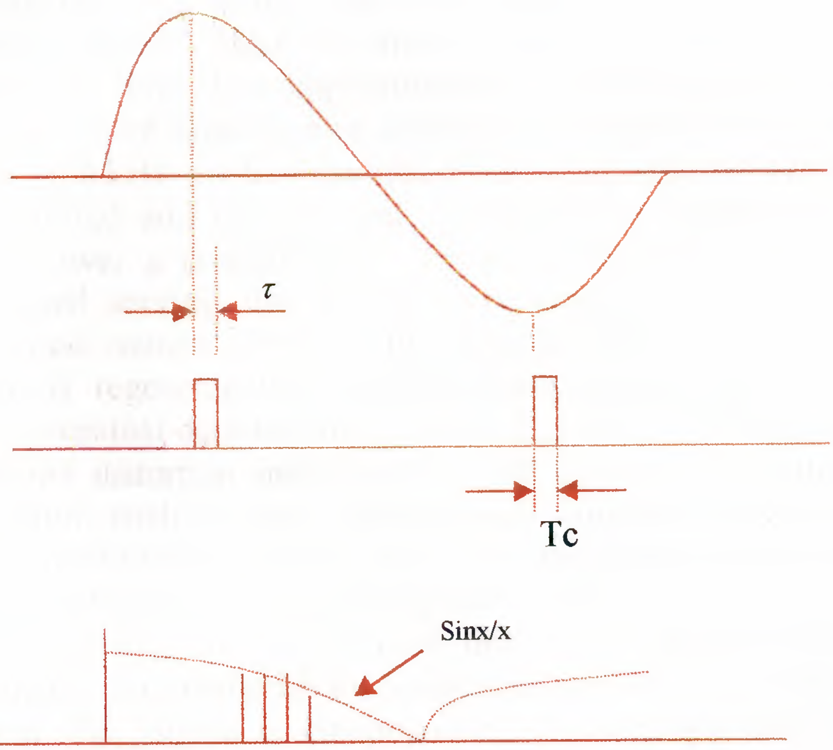
$$k_p |m(t)|_{\max} \prec \frac{T_s}{2}$$

The closer $k_p |m(t)|_{\max}$ is to one half the sampling duration T_s, the must the standard pulse $g(t)$ be in order to ensure that the individual pulses of the PPM signal $s(t)$ do not interfere with each other, and the wider will the bandwidth occupied by the PPM signal be. Then the signal samples $m(nT_s)$can be recovered perfectly. Furthermore, if the message signal $m(t)$is strictly band limited, it follows from the sampling theorem that the original message signal $m(t)$ can be recovered from the PPM signal $s(t)$ without distortion.

PULSE WIDTH MODULATION (PWM)

If the frequency and amplitude of a pulse train are kept constant and the width of the pluses is varied with modulating signal, then the result is a pulse width modulated (PWM) signal. The variations are possible.

The first pulse center may be fixed in the center of the repeating time window Tc and both edges of the pulse moved to compress or expand the width τ. Second the lead edge can be held at the lead edge of the window and the tail edge modulated. Third, the tail edge can be fixed and the lead edge modulated. They each contain a DC component and a base side-band containing the modulating signal as well as phase modulated carrier at each harmonic of the pulse frequency.



Figure(23) pulse position modulated signal

The amplitude of the harmonic groups are constrained by a (sin (x))/x is envelope and extend to infinity. Since the base band information appears in the signal and is not distorted by any modulation effects it may be recovered using a simple low pass filter to remove the carrier and its harmonic and high pass filter to remove the DC component.

MULTIPEXING PCM SIGNALS

We have already noted the advantage of converting an analog signal into a PCM waveform when the signal must be transmitted over a noisy channel. When a large number of such PCM signals must be transmitted over a common channel, multiplexing of these PCM signals is required.

In this section we discuss the multiplexing methods for PCM waveforms used in the United States by the common carriers.

THE T1 DIGITAL SYSTEM

The T1 digital system, which is used to convey multiple signals over telephone lines using wide band coaxial cable. It accommodates (24) analog signals, which we shall refer to as s1 through S24. Each signal is band limited to approximately 3.3 kHz and is sampled at the rate 8 kHz which exceeds, by a comfortable margin, the Nyquist rate of 2 X 3.3 = 6.6 kHz. Each of the time division multiplexed signals is next A/D converted and compounded. The resulting digital waveform is transmitted over a coaxial cable, the cable serving to minimize signal distortion and serving also to suppress signal corruption due to noises from external sources. Periodically, at approximately 6000 ft intervals, the signal is regenerated by amplifiers called repeaters and then sent on toward its eventual destination. The repeater eliminates from each bit the effect of the distortion introduced by the channel. Also, the repeater removes from each bit any superimposed noise and thus, even having received a distorted and noisy signal, it retransmits un-distortion and noise-free duplicate of the signal originally sent.

Such is, of course, the case for all bits except those infrequent bits, which arrive so corrupted by noise that the bit is misread. At the destination the signal is compounded, decoded, and demultiplexed, making available, individually, the 24 original signals.

BITS/FRAME

The commutations sweep continuously from S_1 to S_{24} and back to S_1, etc. at the rate of 8000 revolutions per second thereby providing 8000 samples per second of each signal. Each sample is encoded into eight bits (corresponding to 2^8=256 quantization levels). The digital signal generated during the course of one complete sweep of the commutation is therefore 24x8 =192 bits.

FRAME SYNCHRONIZATION

It is necessary to make available at the receiver not only the bits into which the signal have been encoded but also some synchronizing information. Without such synchronization information, the receiver cannot know which bits correspond to which of the original signal. To provide such synchronization an extra single bit is made available immediately preceding the 192 bits that carry the encoded signals. The 192 bit slots assigned to the encoded signal together with the one extra timing bit, for a total of 193 bits, is called a frame. The time slots for the 24 signals together width the extra frame-synchronizing bit F.

Twelve successive of F slots are used to transmit a 12-bit code. The code happens to be 110111001000. This code is transmitted repetitively once every 12 frames and is used at the receiver to establish synchronization.

BITE RATE

Each signal is sampled 8000 times per second so that a complete frame occupies a time

$$T_p = 1/8000 = 125\ \mu s$$

This time T_p accommodates 193 bits so that the bit rate on a T1 channel is

$$f_b(T1) = \frac{193}{125}\ Mb/s = 1.544\ Mb/s$$

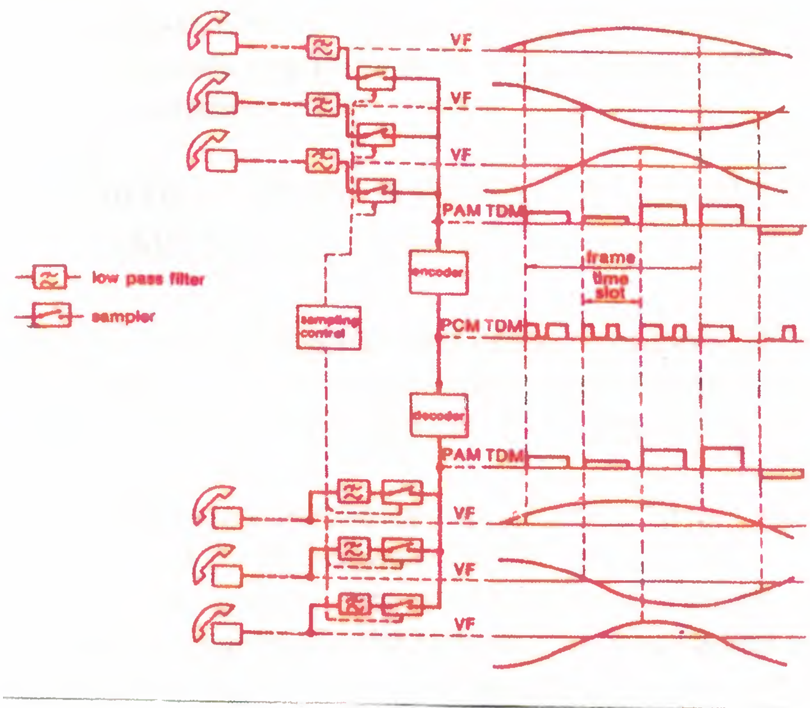
Bits in the frame synchronizing code occur once per frame or every 125 µs. hence the frame synchronizing code repeats every 1.5 ms and the frame rate is 667 frames/s.

PCM TRANSMISSION SYSTEM

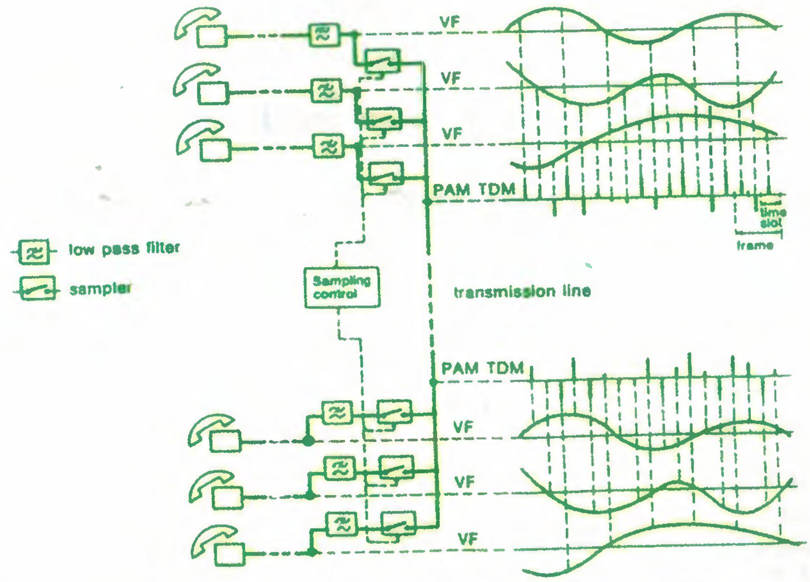
We have now delta with the fundamental principles of PCM. In this part we shall describe how these principles are used to build up practical PCM transmission systems. However, we shall start by explaining the time division multiplex principle as this makes PCM transmission system for telephony economically attractive.

FIRST ORDER PCM SYSTEMS

Several signals in pulse from can use a common transmission path if the signals have different phases. Figure (24) shows how three PAM-signals are time division multiplexed on the same transmission line. The pulses of the three signals are interleaved by opening the sampling gates one by one cyclically. During one cycle the transmission line receives one PAM pulse from each of the participating signals.
Such a set of pulses is called one frame. The time interval that each of these pulses occupies is called a time slot. In this example each has three time slots. On the receiving side the pulses are distributed again. This is done by opening the sampling gates cyclically in the same manner as on the transmitting side does this. Due regard must of course be paid to the transmission delay. This delay has been omitted in figure (24) for clarity.
In the case of PCM signals the time division multiplexing is most after carried out before the samples are pulse coded, i.e., the samples from the participating analogue signals are combined on a common PAM transmission line. See figure (24).



Figure(24) A PAM transmission system using time division multiplex (TDM)



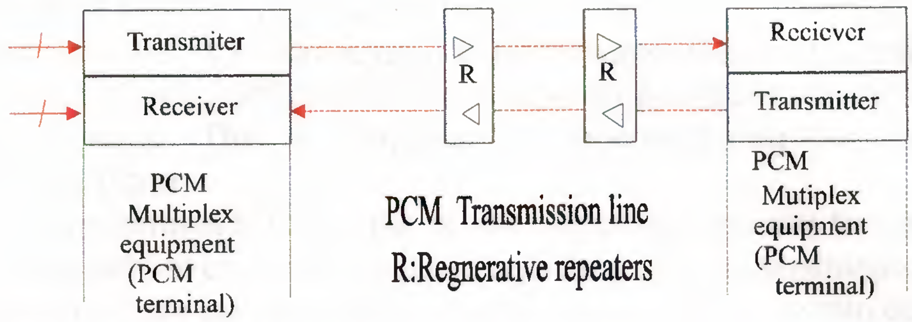
Figure(25) A PCM-TDM transmission system. Attenuation and delay on the PAM and PCM transmission lines are not shown

In this way the coding equipment can be used in time division multiplex. We see from the figure that the PCM pulses are not interleaved pulses by pulse, but PCM word by PCM word. This is often called time slot interleaving. PCM systems used in
Telephony are most often TDM systems, so when we read or hear the term " it is almost always referring to a PCM-TDM system. However it must not be forgotten that PCM in itself can be used, and is used in some cases, on a one- channel basis.

FIRST ORDER PCM SYSTEM RECOMMENDED BY CCITT FRAME STRUCTURES

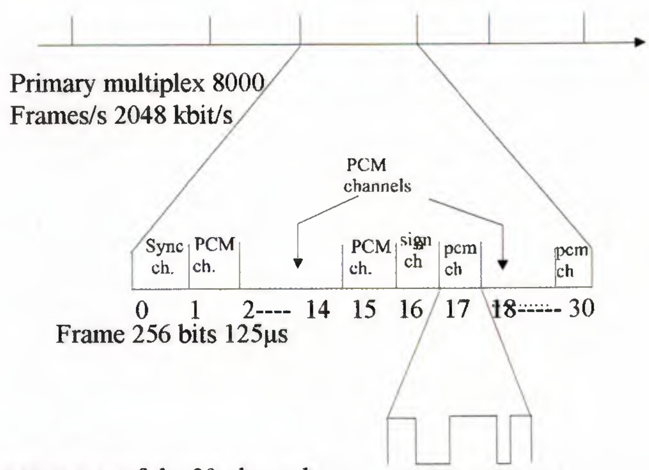
The CCITT recommends two first order, or primary, PCM systems for use in telephony; the 30-channel system, proposed by AT & T. The first order system will form the basis for hierarchies of digital transmission systems.
We have to distinguish between the PCM multiplex equipment, or the PCM terminal, and the PCM transmission line.
The multiplex equipment converts a number of analogue signals (30 or 24) to a digital signal on the transmitting side and carries out the inverse functions on the receiving side. The transmission line conveys the digital signals between two multiplex equipment units. See figure (25).
In the following, the 30-channel multiplex will be treated in some detail, as the multiplex forms the basis for the subsequent presentation of digital telephony. A summary of the most important data on the 24-channel multiplex is also given.

telephony. A summary of the most important data on the 24-channel multiplex is also given.
The presentation of first order PCM systems will end with a brief description of the transmission lines.



Figure(26) Two multiplex equipment units

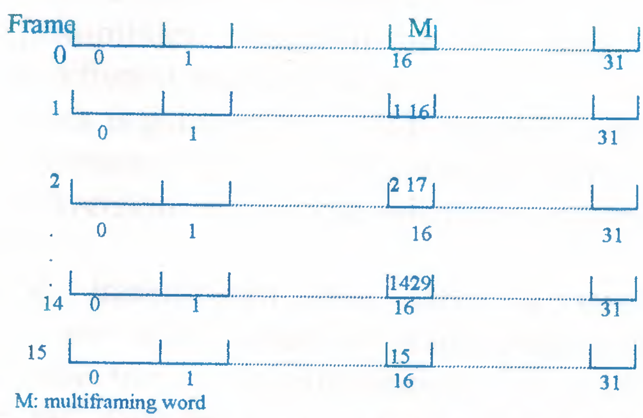
Thirty analogue speech channels together with associated signaling are converted to one digital signal by the 30-channel system. The structure of this digital signal is shown in figure (27)



Figure(27) frame structure of the 30-channel Primary multiplex

The digital signal is divided into frames, with a repetition rate of 8000 frames/sec.
This is of course because the sampling frequency is 8000 Hz and the fact that the frame contains one binary coded sample from each of the analogue signals. Each frame consists of 32 eight-bit time slots.
Of these, 30 time slots are used for PCM channels and the remaining two for synchronization and signaling.
The PCM channels carry analogue signals within the frequency band 300-3400 Hz, coded according to the a-law.

The synchronization time slot, time slot 0 in each frame, contains 8 bits whose purpose is to form a recognition signal to the receiver in order to keep this synchronized to the transmitter so that each PCM channel can be correctly identified.
This function is the same as the function indicated by the sampling control block in figures (24) and(25).
The signaling time slot can be used in many ways. The great signaling capacity, (64) Kbit/s, offers flexibility in choosing suitable schemes for different purposes. This is important when considering the digital network of the future.
CCITT has recommended the use of the signaling time slot for either common channel or channel associated signaling. The arrangements for common channel are not yet specified, so here we can only go into detail concerning the channel associated signaling scheme.
The signaling scheme is the one used today when introducing PCM primary systems into the existing network. The scheme uses the time slots 16 in sequences of 16 frames, referred to as multi-frames, as shown in to figure (28).



Figure(28). Structure of the channel associated signaling scheme for the 30-channel PCM system

In the first frame of the sequence, frame 0, time slot 16 carries a multiframing word, i.e. a recognition signal that tells the receiver that a new multiframe has started. The eight bits of time slot 16 in the next frame, frame 1, are divided so that the first four bits carry of signaling information associated with PCM channel 16. In frame 2, time slot 16 carries signaling information associated with PCM channel 16.
In frame 2, time slot 16 carries signaling information for the channels 2 and 17 and so on up to frame 15, the last frame in the multiframe, which carries signaling information for the channels 15 and 30. Then the next frame is frame 0 in the next multiframe.

Thus four signaling bits are associated with each PCM channel. Each bit can be used to reproduce the state of a signaling relay in a juncture connected to the PCM terminal, i.e. the scheme provides four signaling channels per PCM channel; normally, for conveying conventional signals, only one or two of the channels are used. The 24- channel PCM multiplex has a somewhat different structure as can be seen from figure (29).

Figure(29). Frame structure of the 24-channel system.

No special time slot is assigned to signaling. A channel associated signaling scheme is achieved by taking the least significant bit in every PCM channel for signaling purposes every 6th frame. For synchronization one extra bit is inserted in the frame.

This bit can also be used for common channel signaling. A summary of important technical data for the primary multiplexes are not compatible; they have, for example, a different number of time slots and different signaling possibilities. Not even the time slots are compatible as the systems use different encoding laws.

However, work is going on at CCITT aimed at finding digital method for conversion between PCM words using different encoding laws in order to avoid conversions to analogue when connecting different PCM stems to each other.

The PCM transmission lines used for interconnecting primary multiplexes are most often all ready existing pair cables used for analogue voice frequency transmission. For a PCM line we need two pairs, one for each direction see figure (29). The line must be equipped with regenerative repeaters every 1.5-2.5 km, depending on the type of cable. This distance is about the same as the distance between loading coils on loaded cable circuits, and conversion of loaded pairs into PCM transmission lines can be carried out merely by replacing the loading coils with PCM regenerative repeaters on specially selected pairs.

The transmission lines are designed to convey digital signals of specified rates 2048 0r 1544 Kbit/s. the digital signals must fulfil requirements concerning not only pulse polarities, but also pulse distribution. The timing circuits in the repeater need, for their operation, digital signal with alternating pulses. Long sequences of zero- bits must be avoided. This is done by a special line code that coverts some of the zero-bits to pulses in such a way that these extra bits can be removed before arriving at the decoder.

SECOND ORDER PCM SYSTEM

The primary PCM systems are intended for short-distance applications. In the medium and long distance network, where high cannel capacity is demanded, it is more economical and practical to group together larger number of PCM channels to one common transmission line, thus forming higher order systems than to use several primary PCM systems.
In general multiplexes can be of types:

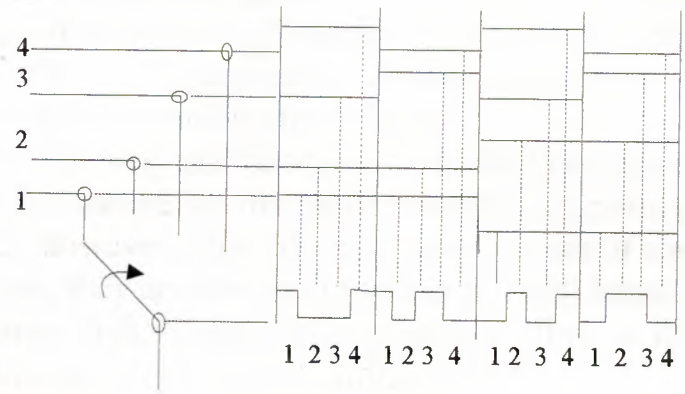
PCM multiplexes and digital multiplexes
PCM multiplexes drive one single from a number of analogue signals by a combination of pulse code modulation and time division multiplexes described earlier are of this type.

Digital multiplexes derive one single digital signal by combining a number of digital signals by time division multiplexing and also carry out the inverse function.

Digital transmission lines convey digital signals between multiplex equipment units. These lines are designed to carry digital signals of specified rates, but they are not dependent on what type of original signal is conveyed in digital form. That is the same transmission line can be used for both PCM multiplexes and digital multiplexes if these have the same rate on the multiplexed digital signals. Even other types of digital signals, for example digitally coded visual telephone signals or data signals, can use the transmission lines if their digit rates are correct.

TDM HIERARCHY

Two-second order systems have been recommended by CCITT. These have digital multiplexes and are based on each of the two primary multiplexes. They both combine four primary PCM signals to one digital signal.
The signals are multiplexed by bit interleaving, which is the participating signals are combined time slot by time, as in this latter case it is necessary to collect the bits of the time slots in buffers before interleaving can be carried out.



Figure(30). Bit interleaving

The digital multiplexes must accept that the primary signals, for practical reasons, have bit rates slightly differing from the ideal but rate. In the systems recommended by the CCITT this is accomplished by having the second order bit rates somewhat higher than four times the ideal primary bit rates, there by ensuring that, even "fast" primary multiplexes can be treated in a proper ways.

A second order PCM multiplex is all readies are use in the United States, this system combines 96 analogue channels to a 6312 Kbit/s digital signal, i.e. it uses the same transmission line as the digital multiplex. In Europe a second order PCM multiplex based on the parameters of the 30-channel system and the 8448 kbit/s transmission line has been discussed. This system is proposed to be time slot interleaved with 132 time slots. Of these, 128 can be used for speech channels, 2for synchronization and 2 for signaling.

TDM HIERARCHY BASED ON 24 CHANNEL SYSTEM

To take further advantages of the merits of TDM and digital transmission the common carries employ a hierarchy of further multiplexing. Four T1 lines are multiplexed in an M12 multiplexer to generate a T2 transmission system, seven T2 lines convert to a T3 line in an M23 multiplexer and six T3 lines convert to a T4 line in an M34 multiplexer. At each stage additional frame synchronizing bits must be added as with the first order multiplexing so that at each multiplexer output it would be possible to distinguish which bits belong to which input.

There is a problem that arises in connection with the higher orders of multiplexing that does not occur at the first order. In the first order multiplexing there is just a single clock to contend with, that is the clock that drives the commutation. However, the four input lines in the M12 multiplexer come from physically widely separated locations and employ four separate unsynchronized clocks.

These clocks are very stable crystal controlled oscillators, which are, of course, set to operate at the same nominal frequency as nearly as possible. Since, however, they do not have means of communicating with one another, they are not synchronized and will hence experience a relative frequency drift. Design specifications allow a drift from the nominal set frequency of ±130 parts/million.

As may be verified, if then two clocks have frequencies which differ by (2 x 130)=260 parts/million, the faster clock will have generated one more time slot than the slower clock in the course of just 20 frames. For proper interleaving of bits at the M12 level it is necessary that all input bit streams have or be made to appear to have the same rate.

To put the matter most simplistically, the process of adjusting bit rates to make them equal involves adding bits to the slower bit stream in an operation referred to as ‘pulse stuffing.

Further bits must then be added to all bit streams to allow the receiver of the composite signal to distinguish time slots, which carry information from slots, which carry the 'stuffed' bits.
The M12 multiplexer adds nominally 17 bits for frame synchronization and pulse stuffing. Hence the number of bits per frame is

$$193 \times 4 + 17 = 789 \text{ bits/frame}$$

The T2 line bit rate is therefore

$$f_b(TT2) = 789 \text{bits/frame} \times 8000 \text{frames/s} = 6.312 \text{Mb/s}$$

The M23 multiplexer adds nominally (69) bits for synchronization and pulse stuffing, hence the number of bits per frame for a T3 line is

$$(789 \times 7) + 69 = 5592 \text{ bits/frame}$$

and

$$f_b(T3) = 5592 \times 8000 = 44.736 \text{Mb/s}$$

The M34 multiplexer adds nominally (720) bits for synchronization and pulse stuffing and therefore the T4 system has a bit rate.

fb(T4)=274.176 Mb/s

A detailed analysis of the architecture and operation of these digital systems it be found in Ref.3.

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