RESEARCH PROJECT

ON

Performance Evaluation of Baseband Communication using Gaussian Pulse Train

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21 August 2016

EAST WEST UNIVERSITY

Letter of Transmittal

To
Dr. M.Ruhul Amin
Professor,
Department of Electronics and Communication Engineering,
East West University.

Subject: Submission of Project Report as (ETE-498)

Dear Sir,

We are pleased to let you know that we have completed our project on **"Performance Evaluation of Baseband Communication using Gaussian Pulse Train".** The attachment contain of the project that has been prepared for your evaluation and consideration. Working on this project has given us some new concepts. By applying those concepts we have tried to make something innovative by using our theoretical knowledge which we have acquired since last four years from you and the other honorable faculty members of EWU. This project would be a great help for us in future.

We are very grateful to you for your guidance, which helped us a lot to complete my project and acquire practical knowledge.

Thanking You.

Yours Sincerely

Mahabubur Rahman Likhon
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Declaration

This is certified that the project is done by us under the course "Project (ETE-498)". The project of "Performance Evaluation of Baseband Communication using Gaussian Pulse Train" has not been submitted elsewhere for the requirement of any degree or any other purpose except for publication.

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Acceptance

This Project paper is submitted to the **Department of Electronics and Communication Engineering, East West University** is submitted in partial fulfillment of the requirements for the degree of **Bachelor of Science** in **Electronics & Telecommunications Engineering (ETE)** under complete supervision of the undersigned.

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Dr. M. Ruhul Amin

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Abstract

According to Nyquist law of baseband communication, the received signal against a rectangular transmitted pulse will be a sinc pulse. Therefore rectangular pulse train will produce sinc pulse train at receiving end and the sampling instant will be at zero crossing points of the sinc pulse train. Only the desired pulse will provide the level of received signal since all other sinc pulses passed through zero but small jitter will produce huge error which is minimized by using raised cosine pulse instead of sinc pulse. Then raised cosine pulse alleviates the limitation of ideal LP characteristics of sinc pulse as well as the amplitude of tails of since pulse is heavily reduced in raised cosine pulse. Our objective is to make the amplitude of tail of sinc or raised cosine pulse to zero (hence jitter error is eliminated) at the same time the limitation of ideal LP characteristics of since pulse will be eliminated. The job is done Gaussian pulse at receiving end since Fourier transform of such pulse does not change the shape of the pulse at the same time amplitude of tail of the Gaussian pulse is zero. We measure the performance of three pulses at receiving end based on received SNR in dB varying jitter error and found the Gaussian as the best.

Content

CHAPTER 1: Introduction

Data transmission over band limited channels requires pulse shaping to eliminate or control Inter-Symbol Interference (ISI). A widely used filter for this purpose is the raised cosine filter which satisfies Nyquist's first criterion. There is a phase compensation technique for the square-root raised-cosine filter to design a Nyquist filter. In the special case of full raised-cosine spectrum, it is shown that the square-root filter satisfies Nyquist's criterion, provided appropriate time delay in the impulse response. Therefore, the square-root full raised-cosine transmitting filter can be used with or without a matched filter at the receiver [1].

Orthogonal frequency Division Multiplexing has several properties which makes it as an attractive alternative modulation technique for high speed data transmission. However one major disadvantage of OFDM is that the time domain OFDM signal which is a sum of several sinusoids leads to high Peak to Average Power Ratio (PAPR).It had been shown that using pulse shaping technique, it is possible to design a set of waveforms (raised cosine, root raised cosine) for the OFDM system that reduce the PAPR of transmitted signals [2].

QPSK and OQPSK are regarded as successful modulation formats in IS-95 CDMA, CDMA-2000 and WCDMA Mobile Communication Systems. The graphical representation of measured results of the analysis of different performance parameters such as EVM, Magnitude Error, Phase Error, Bandwidth efficiency and BER of the QPSK & OQPSK transmission systems, which are considered to be useful system metrics for any digital communication system and its critical analysis could be used by the system designers as a powerful tool for choosing a suitable modulation format with a proper Pulse-shaping filter along with its correct Roll-off factor [3].

The pulse shaping filter not only reduces inter-symbol interference (ISI), but it also reduces adjacent channel interference. The application of signal processing techniques to wireless communications is an emerging area that has recently achieved dramatic improvement in results and holds the potential for even greater results in the future [4].

The OFDM is a promising candidate for achieving high data rate transmission in mobile environment. A study on pulse shaped OFDM signal is made and the power spectral density (PSD) and MI of different pulse shaped OFDM signal is presented where the OFDM waveform is then analyzed for MI & frequency domain response and the PSD in each case. It is also possible to design a set of time domain waveforms that will reduce the PAPR of the OFDM transmitted signal and improve its power spectrum simultaneously. The effect of some of these sets of time waveform on the OFDM system performance in terms OfMI & power spectral density (PSD) is investigated and the data is tabulated to analyze and establish the superiority of a specific pulse shape over the other depending on the application or requirement [5].

16 QAM base band communication system using Raised Cosine and Root Raised Cosine pulse shaping filters is designed and implemented on the MATLAB platform. This is tested for different SNR conditions in the channel ranging from 0 db to 20db and the results are plotted for Raised Cosine and Root raised cosine communication systems. Results are also proved that ISI, Co-channel interference and additive noise are also suppressed over the band-limited channels by using these pulse-shaping filters [6].

To control ISI, control has to be exercised over the pulse shape in the overall system, known as pulse shaping. The mostly used pulse shaping filters to avoid ISI are Raised Cosine and Root Raised Cosine filters. 16 QAM base band communication system using Root Raised Cosine pulse shaping filter is designed and implemented on the MATLAB platform. Results shown above prove that ISI, Co-channel interference and additive noise are also suppressed over the band-limited channels by using these pulse-shaping filters [7].

Inter Symbol Interference (ISI) is an inevitable consequence that cannot be totally avoided. In order to reduce this error in communication the pulse shaping filters are designed. Among different pulse shaping filters the FIR Raised Cosine Filter is widely used in digital communication to reduce ISI, where the Raised Cosine Filter is the most advantageous due to the fact that it is easy for practical implementation [8].

Square root raised cosine filter is a FIR filter. Square root raised cosine filters are used in both transmitter and receiver for matching filter purpose. Shifting and addition method is proposed for designing of square root raised cosine filter. The FPGA implementation of Square root raised cosine filter for pulse shaping used in WCDMA system is done [9].

Analysis and evaluation of BER performance for WCDMA Network using OQPSK modulation technique with the Square Root Raised Cosine Filter has been carried out where the Stimulation has been studied at 960 Kbps data rate. Two constant roll off factors are considered for which bit error rate performance is evaluated for different group delays. The result shows the optimum value for Group Delay for minimum Bit error rate. It can be concluded that for a chip rate of 3.84 Mcps of WCDMA, best results are obtained for at a high data rate of 960 Kbps in terms of BER [10].

The digital filter for DDC in WiMAX application is designed using Gaussian and RRC pulse shaping filters in order to meet the spectral requirement of wireless communication. Filters designed by using two techniques are then compared. It is concluded that because of small transition bandwidth and more linear phase in magnitude and phase response of Gaussian filter as compared to RRC filter, Gaussian filter give significantly better BER performance than conventional RRC filter [11].

CHAPTER 2: Pulse Shaping of Baseband Transmission

The bandwidth of a rectangular pulse is infinite hence if the baseband binary pulse is transmitted through a communication channel the received pulses are widened and create ISI. The communication channel (wired or wireless) works as a bandpass filter converts the rectangular pulse into smooth pulse like fig. below.

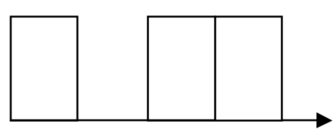

Fig. 1 (a): Transmitted pulse train

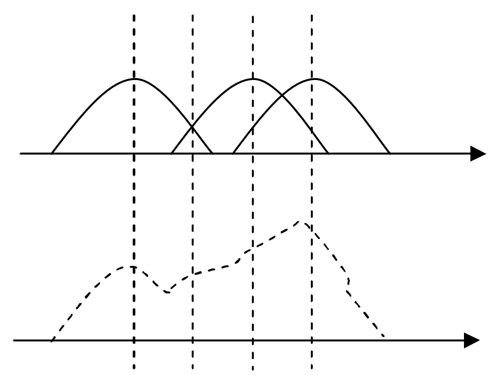

Fig. 1 (b): Received pulses

Nyquist established that if the rectangular pulse can be converted into *h(t)=sinc(t/T),* the pulses can be detected without ISI . Fig.4 shows three successive pulses *h(t), h(t-T)* and *h(t-2T).* Although *h(t)* has long tail but passes through zero amplitude at the sampling instant $t = kT$. This possible only if the sampling can be done exactly at the correct time but timing error introduces ISI for *sinc(t/T)* pulse. To overcome the situation *sinc(t/T)* pulse is modified to another Nyquist class pulse of zero ISI known as raised-cosine pulse.

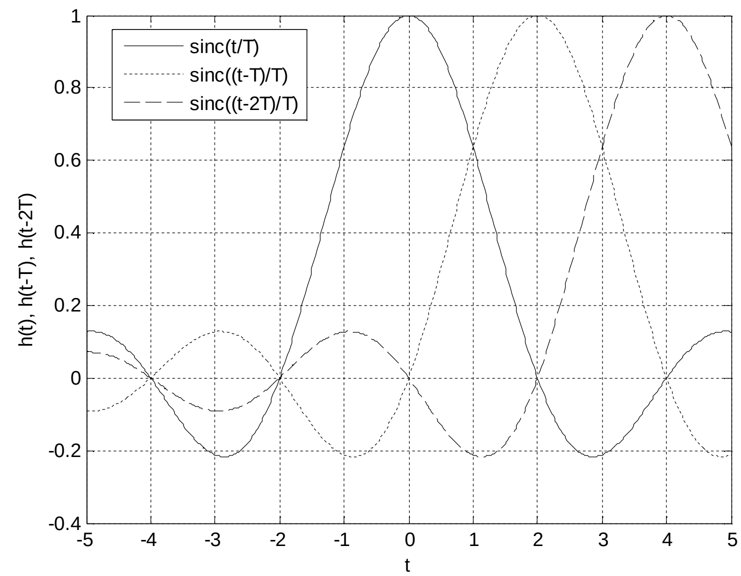

Fig.2:**Sinc pulse sequences**

The baseband communication system can be represented like Fig.3.

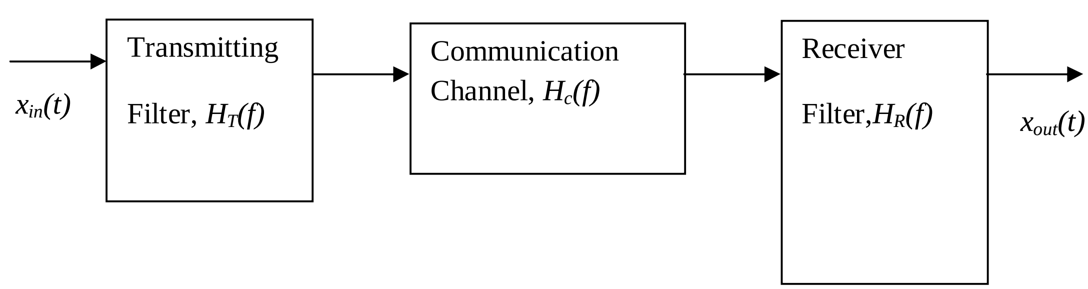

Fig.3 Baseband communication system

The overall transfer function of the system,

$H_{eq}(f) = H_T(f)H_c(f)H_R(f)$

$\Rightarrow h_{eq}(t) = h_t(t)*h_c(t)*h_r(t)$ (1)

The input pulse train,

$x_{in}(t) = \sum_n a_n h(t - nT_s)$; Where $a_n \in \{0, 1\}$ and $h(t) = \Pi\left(\frac{t}{T_s}\right)$ for binary pulse

$= \sum_n a_n \delta(t - nT_s) * h(t)$

$$= \left\{\sum_n a_n \delta(t - nT_s)\right\} * h(t) \qquad (2)$$

$; \because \delta(t) \leftrightarrow 1, \delta(t - nT_s) \leftrightarrow 1.e^{-nT_s} = e^{-nT_s}$ and $h(t) * \delta(t - nT_s) \leftrightarrow H(f)e^{-nT_s} \leftrightarrow h(t - nT_s)$

Considering (2) we can extend the fig.3 like fig.4 where the input signal is an impulse train.

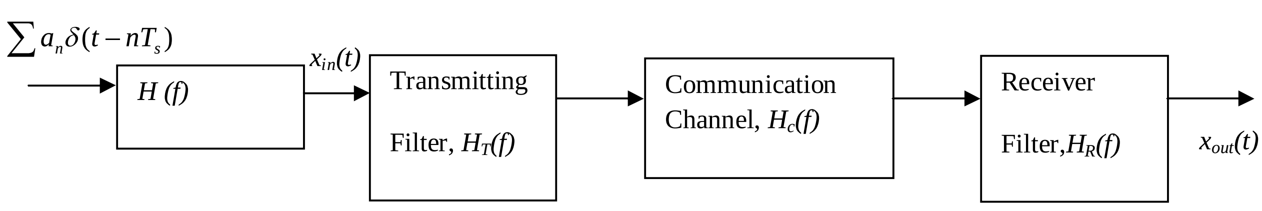

Fig.4: Extension of Fig 3.

Now

$$x_{out}(t)$$
$$= x_{in}(t) * h_{eq}(t)$$
$$= x_{in}(t) * h_t(t) * h_c(t) * h_r(t)$$
$$= \left\{ \sum_n a_n \delta(t - nT_s) \right\} * h(t) * h_t(t) * h_c(t) * h_r(t)$$
$$= \left\{ \sum_n a_n \delta(t - nT_s) \right\} * h_0(t)$$

Where $h_0(t) = h(t) * h_t(t) * h_c(t) * h_r(t)$ is the equivalent impulse response of the baseband communication system of Fig.4. To eliminate ISI Nyquist shows that the equivalent impulse response at $t = kT_s$,

$h_0(kT_s + \tau) = \begin{cases} C\,; & k = 0 \\ 0; & k \neq 0 \end{cases}$;where τ is the offset of sampling instant at receiving end provided $\tau = 0$ for ideal case.

Therefore we can write,

$$h_0(t) = C \sin c(t/T_s)$$

$$\Rightarrow h(kT_s) = C \sin c(k) = \begin{cases} C\,; & k = 0 \\ 0; & k \neq 0 \end{cases}$$

The Nyquist's first formula of elimination ISI:

The necessary and sufficient condition of elimination ISI of a signal *x(t)* is,

$$x(kT_s) = \begin{cases} 1; & k = 0 \\ 0; & k \neq 0 \end{cases} \tag{3}$$

The Fourier transform $x(t) \leftrightarrow X(f)$ satisfy the condition,

$$\sum_{m=-\infty}^{\infty} X(f + mT_s) = T_s \tag{4}$$

Proof:

Inverse Fourier transform of *X(f)*,

$$x(t) = \int_{-\infty}^{\infty} X(f) e^{j2\pi ft} df$$

$$\Rightarrow x(kT_s) = \int_{-\infty}^{\infty} X(f) e^{j2\pi fkT_s} df$$

$$= \sum_{m=-\infty}^{\infty} \int_{(2m-1)/2T_s}^{(2m+1)/2T_s} X(f) e^{j2\pi fkT_s} df$$

; breaking the integral into integrals covering the finite range of $1/T_s$

$$= \sum_{m=-\infty}^{\infty} \int_{-1/2T_s}^{1/2T_s} X(f + m/T_s) e^{j2\pi fkT_s} df$$

$$= \int_{-1/2T_s}^{1/2T_s} \left\{ \sum_{m=-\infty}^{\infty} X(f + m/T_s) \right\} e^{j2\pi fkT_s} df$$

$$= \int_{-1/2T_s}^{1/2T_s} G(f) e^{j2\pi fkT_s} df \tag{5}$$

;Where $G(f) = \sum_{m=-\infty}^{m} X(f + m/T_s)$ (6)

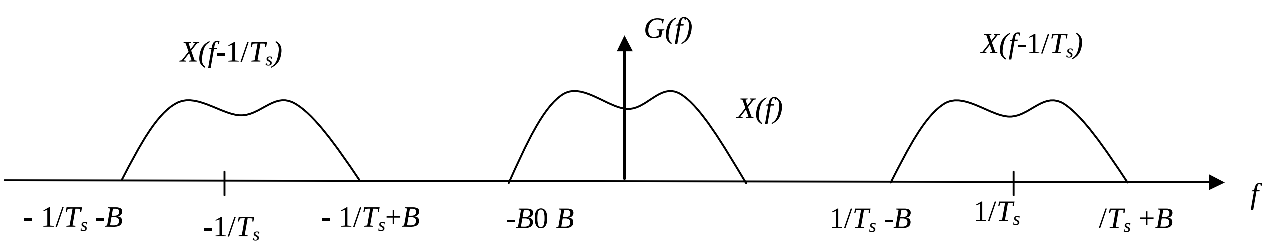

Fig.5: The graphical presentation of *G(f)*

Here $G(f)=\sum_{m=-\infty}^{\infty} X(f+m/T_s) = \ldots \quad \ldots \quad \ldots + X(f-1/T_s)+X(f)+X(f+1/T_s)+\ldots \quad \ldots \quad \ldots$

is a periodic function with period $1/T_s$ (shown in fig.5) can be expressed by Fourier series,

$$G(f)=\sum_{k=-\infty}^{\infty} a_k e^{j2\pi fkT_s} \qquad (7)$$

; Where $a_k = T_s\int_{-1/2T_s}^{1/2T_s} G(f)e^{-j2\pi fkT_s}df$ (8)

Comparing (5) and (7),

$a_k = T_s x(-kT_s)$

$\Rightarrow a_k = \begin{cases} T_s\,; & k=0 \\ 0; & k\neq 0 \end{cases}$; From (3) (9)

From (7) and (9),

$G(f)=T_s$

$$\Rightarrow \sum_{m=-\infty}^{m} X(f+m/T_s) = T_s \qquad (10)$$

This is the required condition of Nyquist.

Case-1

From Fig. 6 if the BW of *X(f)* is *B* and $T_s=1/2B$ then in the range of frequency [-∞,∞] there is only one type of *X(f)* (rectangular shape) to maintain $G(f)=T_s$.

Therefore we can write,

$$G(f)=\sum_{m=-\infty}^{m} X(f+m/T_s)$$
$$=\ldots\quad\ldots\quad\ldots+X(f-1/T_s)+X(f)+X(f+1/T_s)+\ldots\quad\ldots\quad\ldots$$
$$=\ldots\quad\ldots\quad\ldots+T_s\prod\left(\frac{t+1/T_s}{2B}\right)+T_s\prod\left(\frac{t}{2B}\right)+T_s\prod\left(\frac{t-1/T_s}{2B}\right)+\ldots\quad\ldots\quad\ldots=T_s$$

which results in,

$$X(f)=\begin{cases}T_s; & |f|<B\\ 0; & \text{Otherwise}\end{cases}.$$
$$\Rightarrow x(t)=\sin c\left(\frac{t}{T_s}\right)$$

The plot of *G (f)* for above condition is shown in Fig.6.

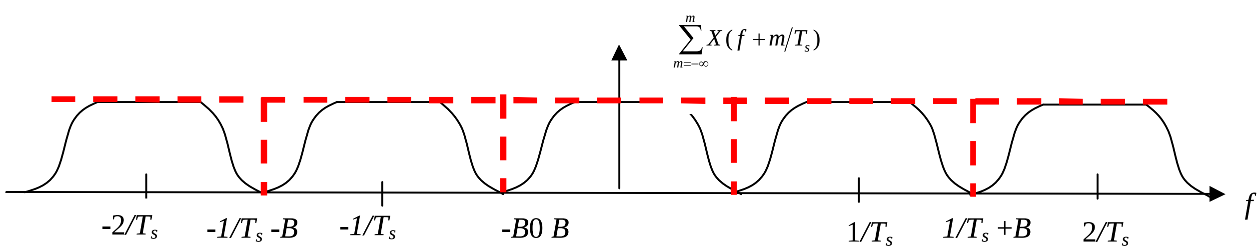

Fig.6: Plot of *G(f)* for T_s=1/2*B*

Case-2

If $T_s< 1/2B$ we get $X(f)$ in non overlapping fashion in $G(f)=\sum_{m=-\infty}^{m} X(f+m/T_s)$ hence we don't have any choice to take *G(f)* = T_s. The graphical presentation of *G(f)* is shown in Fig. 7.

Case-3

Again If $T_s > 1/2B$ we get $X(f)$ in overlapping fashion in $G(f)=\sum_{m=-\infty}^{m} X(f+m/T_s)$ hence we have several choices to satisfy the condition $G(f)=T_s$. The plot of above conditions is shown in Fig 8. In fig.8 if we choose the range of frequency [-B B] we have three components to satisfy (10) like,

$$X(f-m/T_s)+X(f)+X(f+m/T_s)=T_s \qquad (11)$$

Different band limited function *X(f)* can be chosen to satisfy (11), one of the chosen function to avoid ISI is raised cosine spectrum will be discussed later.

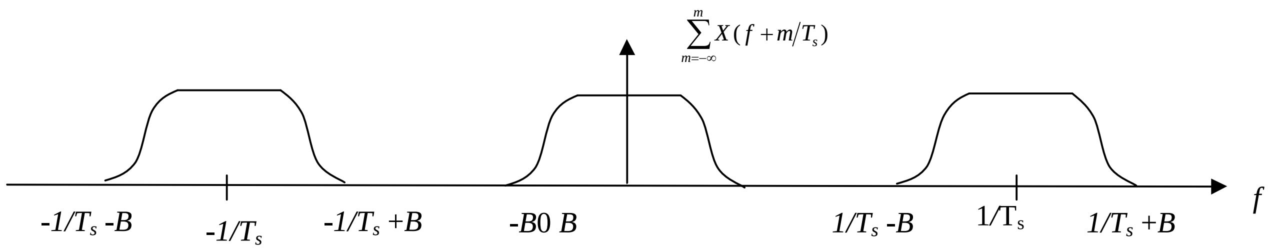

Fig.7: Plot of *G(f)* for $T_s < 1/2B$

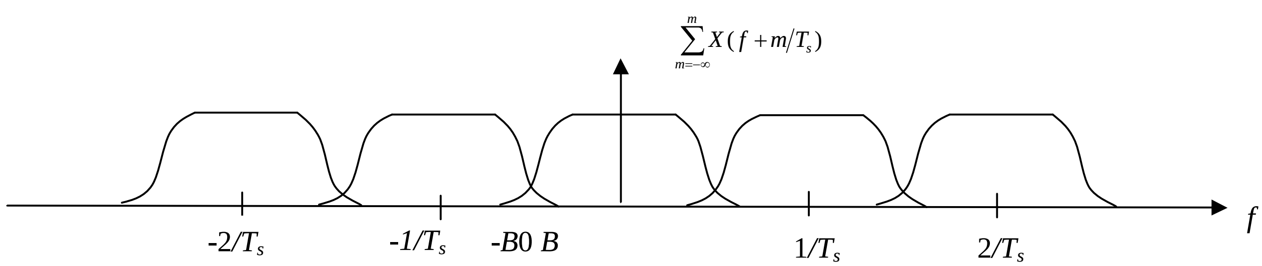

Fig.8: Plot of *G(f)* for $T_s > 1/2B$

Let us now concentrate on the case of fig.2 and consider overall impulse response of the baseband communication system $h_0(t)$ instead of *x(t)*. Taking IFT we get the overall transfer function of baseband communication system like, $h_0(t) = \sin c\left(\frac{t}{T_s}\right) \Leftrightarrow T_s \prod\left(fT_s\right)$. Such rectangular transfer function (ideal characteristics of LP filter) is not practically realizable because:

1. The cutoff at $B = 1/2T_s$ with infinite slope cannot be implemented practically. For example sharp rise or fall of signal in frequency domain provides huge ripple in time domain and vice-versa (Gibb's phenomena).
2. Sampling instant at both transmitting and receiving side must be extremely precise. In communication system jitter is unavoidable hence tails of adjacent pulses can create huge ISI.

To overcome above two difficulties we need transfer function of gradual fall (for example -60dB/dec) like practical LP filter with some wider BW so that can be implemented with hardware. Some additional factors should be used with $h_0(t) = \sin c\left(\frac{t}{T_s}\right)$ so that the tail of the pulse reduces more prominently therefore timing jitter can be tolerable. The raised cosine filter can satisfy both the above criteria. Actually the spectrum of raised cosine filter satisfies the condition of fig.8.

The impulse response of raised cosine filter is,

$$h_{rc}(t) = \frac{\omega_0}{\pi} \sin c\left(\frac{\omega_0 t}{\pi}\right) . \frac{\cos(\omega_0 t)}{1-\left(2\omega_0 t/\pi\right)^2}$$

The transfer function is,

$$H_{rc}(\omega) = \begin{cases} \frac{1}{2}\cos\left(1+\cos\frac{\pi\omega}{2\omega_0}\right); & |\omega| \le 2\omega_0 \\ 0; & \text{else where} \end{cases}$$

Fig.9 shows the transfer function of raised cosine filter.

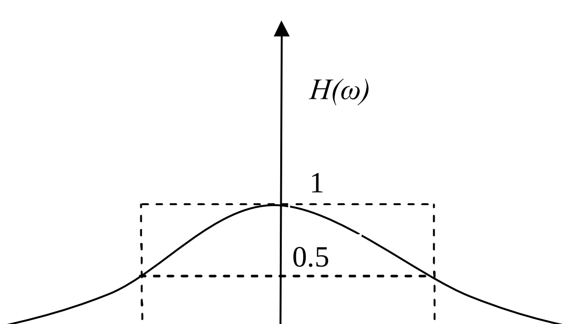

Fig.9: Transfer function of raised cosine filter

The frequency response of a raised-cosine filter is conventionally described with roll off factor α. The spectrum of raised-cosine filter of fig.10 has two parts: flat portion and nonlinearly falling part called roll-off portion. The response of the filter including the roll-off factor is defined as,

$$H_{rc}(f) = \begin{cases} 1; & |f| \le f_1 \\ \frac{1}{2}\left\{1 + \cos\left(\frac{\pi(|f| - f_1)}{2 f_\Delta}\right)\right\}; & f_1 < |f| \le B \\ 0 & |f| > B \end{cases}$$

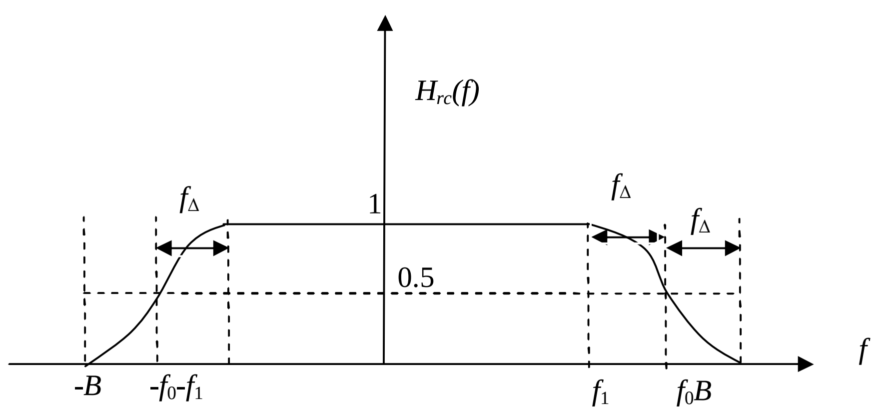

Fig.10: Raised cosine roll-off Nyquist filter

Where f_0 is the 6dB BW of the filter, $f_\Delta = B - f_0$, roll-off factor $\alpha = f_\Delta / f_0$. The roll-off factor α lies in the range of $0 \le \alpha \le 1$. The impact of roll-off factor is shown in fig.10.

The impulse response of the filter,

$$H_{rc}(f) \leftrightarrow h_{rc}(t) = 2 f_0 \sin c(2 f_0 t).\frac{\cos(2\pi f_\Delta t)}{1-(4 f_\Delta t)^2}$$

Example-1

Write Matlab code to plot the Nyquist sinc pulse, Gaussian pulse and their spectrum at receiving end.

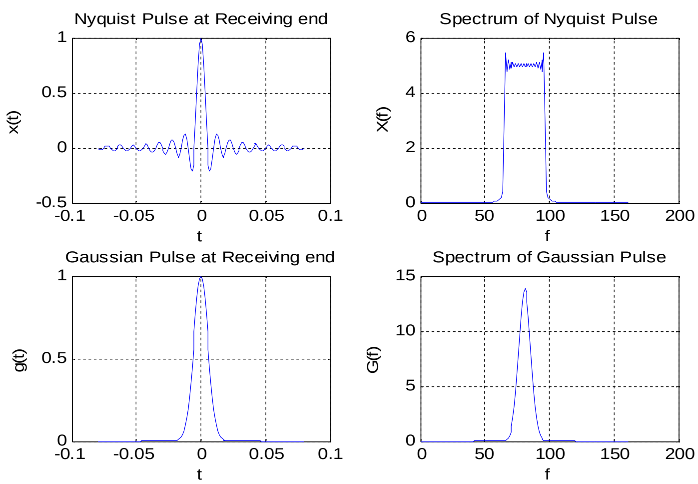

Fig.11: Pulses in time and frequency domain

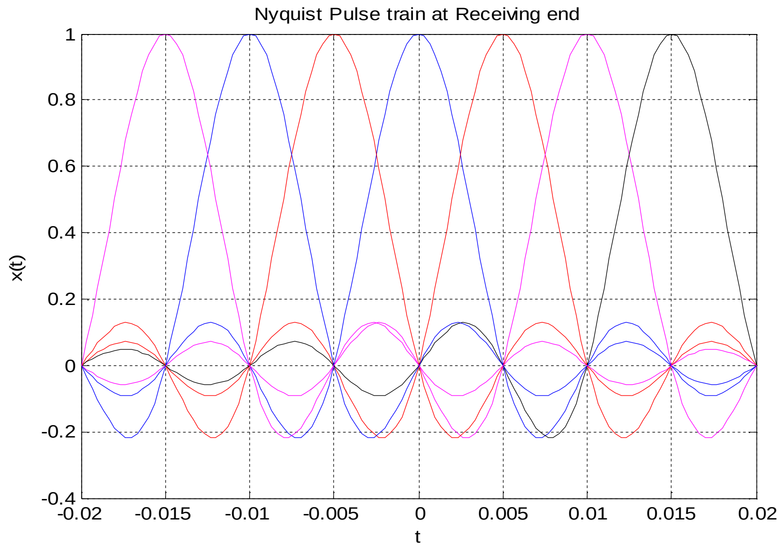

Fig.12: Zero crossing of sinc pulse train

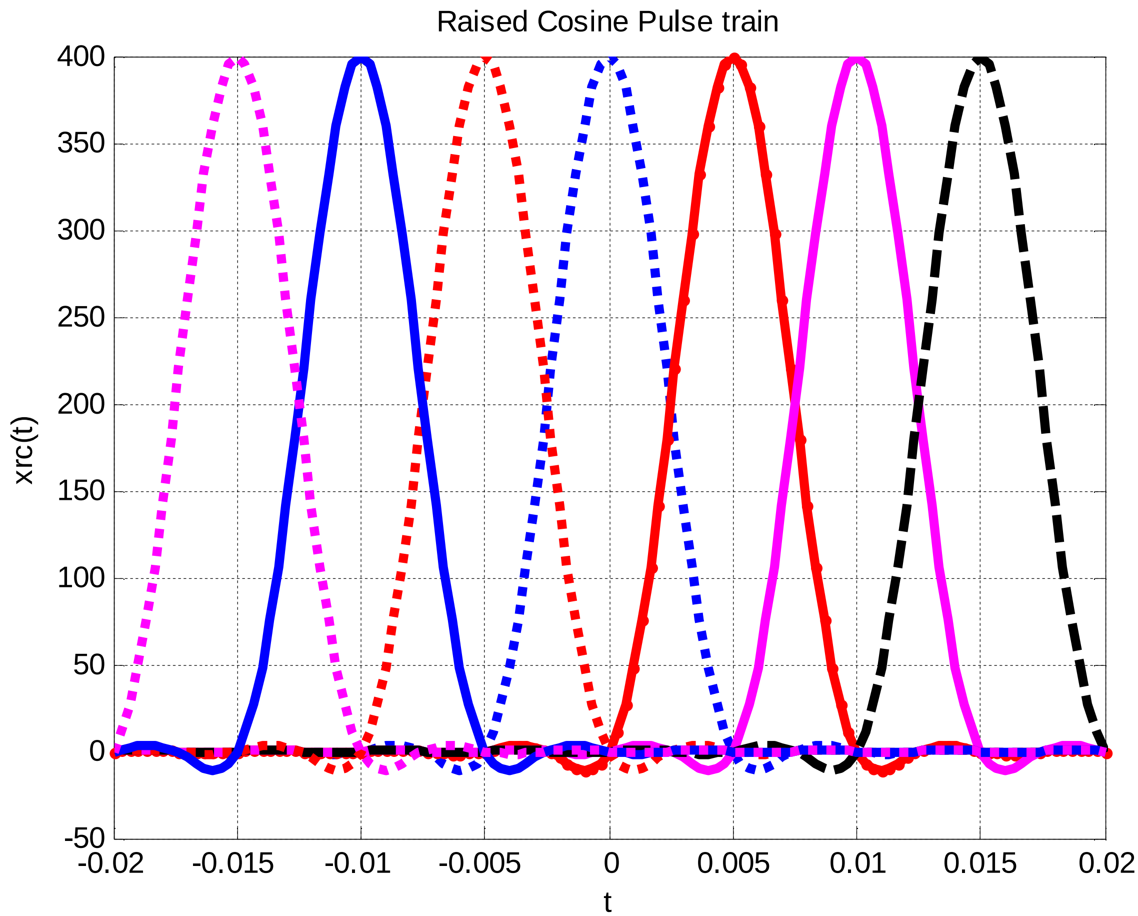

Fig.13： **Zero crossing of raised cosine pulse train**

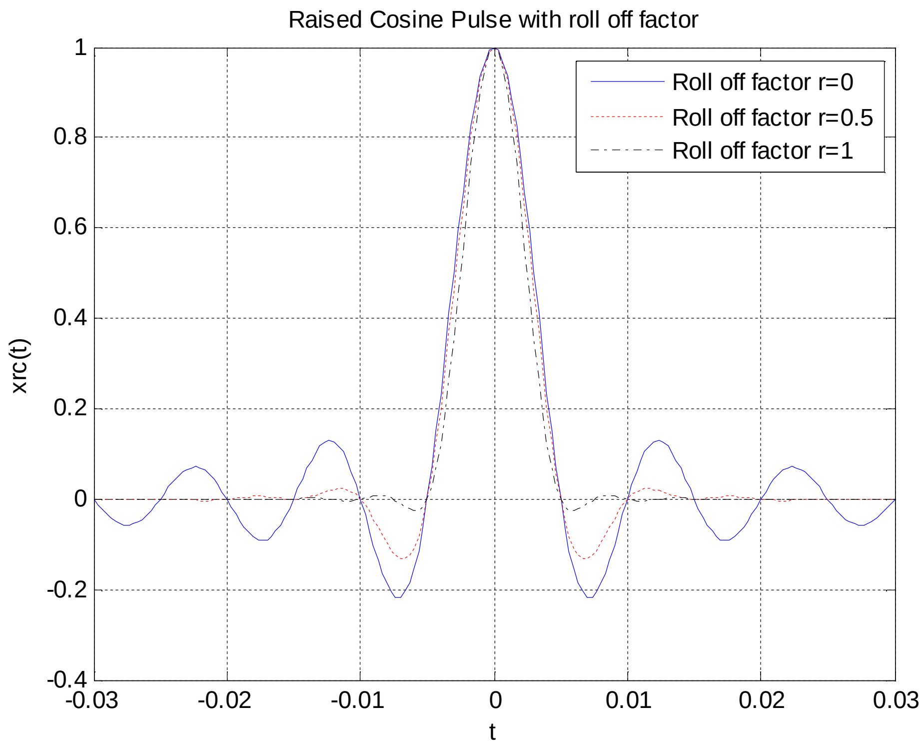

Fig.14: Raised cosine pulse under different roll-off factor

Chapter 3: Construction of raised cosine filter in a different approach

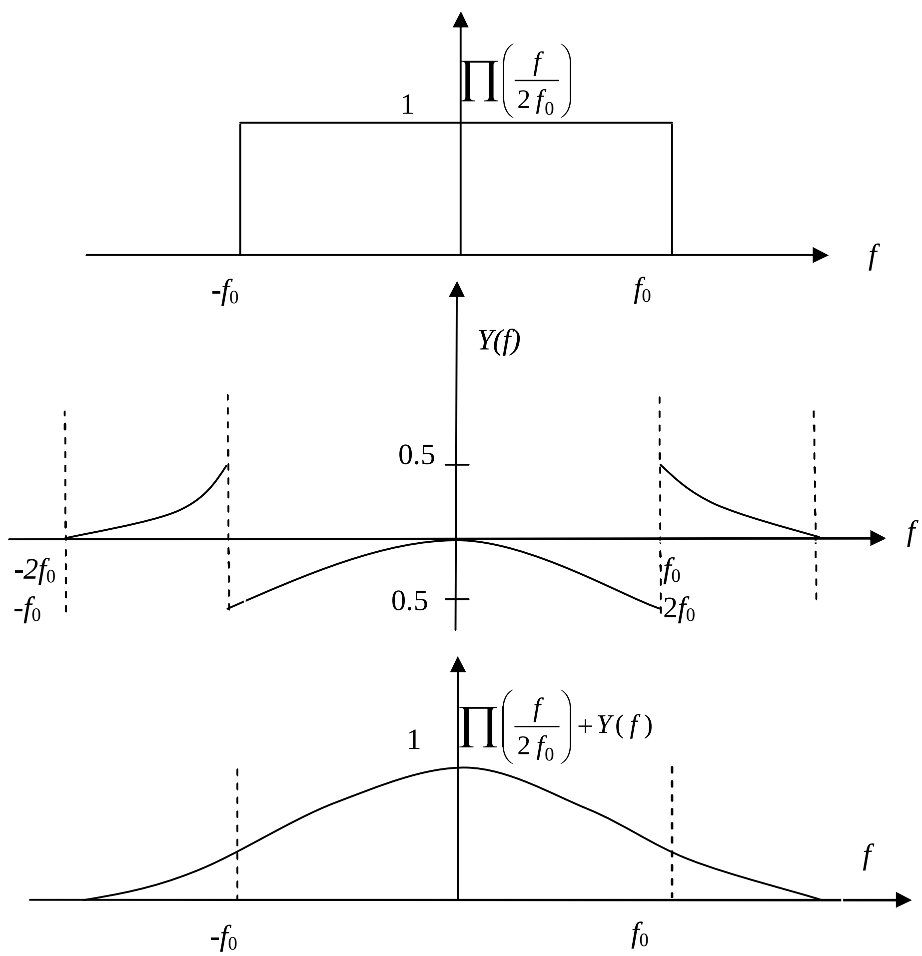

Fig.15: raised cosine filter

The raised cosine filter can be constructed using the equation,

$$H_e(f) = \begin{cases} \prod\left(\dfrac{f}{2f_0}\right) + Y(f); & |f| < 2f_0 \\ 0; & \text{elsewhere} \end{cases}$$

And the graphical presentation is shown in fig.15.

Here $Y(f)$ is a real function is even for $-2f_0 \le f \le 2f_0$ ($Y(-f) = Y(f)$) and odd function for two different range, $0 \le f \le 2f_0$ and $0 \ge f \ge -2f_0$ i.e. $Y(f_0+f) = -Y(f_0-f)$ or $Y(f+f_0) = -Y(-f-f_0)$ i.e $Y(f)$ is odd symmetric about $f = f_0$ and $f = -f_0$.

$$Y(f) = \begin{cases} \dfrac{1}{2}\left(\cos\dfrac{\pi}{2}\dfrac{f}{f_0} - 1\right); & |f| < f_0 \\ \dfrac{1}{2}\left(\cos\dfrac{\pi}{2}\dfrac{f}{f_0} + 1\right); & f_0 < |f| < 2f_0 \end{cases}$$

The impulse response of the filter is obtained taking IFT on $H_e(f)$.

$$\begin{aligned} h_e(t) &= \int_{-2f_0}^{2f_0} H_e(f)e^{j\omega t}df \\ &= \int_{-2f_0}^{2f_0} Y(f)e^{j\omega t}df + \int_{-f_0}^{f_0} 1e^{j\omega t}df \\ &= \int_{0}^{2f_0} Y(f)e^{j\omega t}df + \int_{-2f_0}^{0} Y(f)e^{j\omega t}df + 2f_0 \sin c(2f_0 t) \\ &= I_1 + I_2 + 2f_0 \sin c(2f_0 t) \end{aligned}$$

Putting $g = f-f_0$ in I_1 and $g = f+f_0$ in I_2 we get,

$$h_e(t) = \int_{-f_0}^{f_0} Y(g+f_0)e^{j2\pi(g+f_0)t}dg + \int_{-f_0}^{f_0} Y(g-f_0)e^{j2\pi(g-f_0)t}dg + 2f_0 \sin c(2f_0 t)$$

$$= \int_{-f_0}^{f_0} Y(g+f_0)e^{j2\pi(g+f_0)t}dg - \int_{-f_0}^{f_0} Y(g+f_0)e^{j2\pi(g-f_0)t}dg + 2f_0 \sin c(2f_0 t)$$

; since $Y(g+f_0) = -\ Y(g-f_0)$

$$= \left\{e^{j2\pi f_0 t} - e^{-j2\pi f_0 t}\right\} \int_{-f_0}^{f_0} Y(g+f_0)e^{j2\pi g t}dg + 2f_0 \sin c(2f_0 t)$$

$$= j2\sin(2\pi f_0 t)\int_{-f_0}^{f_0} Y(g+f_0)e^{j2\pi gt}dg + 2f_0 \sin c(2f_0 t)$$

Here transmission BW is $2f_0$ $\therefore$ sampling interval $T_s = 1/2f_0$. At $t = \frac{n}{2f_0} = nT_s \,;\, n \neq 0$,

$$h_e\left(\frac{n}{2f_0}\right) = j2\sin\left(2\pi f_0 \frac{n}{2f_0}\right)\int_{-f_0}^{f_0} Y(g+f_0)e^{j2\pi gt}dg + 2f_0 \sin c\left(2f_0 \frac{n}{2f_0}\right)$$

$$= j2\sin(n\pi)\int_{-f_0}^{f_0} Y(g+f_0)e^{j2\pi gt}dg + 2f_0 \sin c(n)$$

$$= j2.0\int_{-f_0}^{f_0} Y(g+f_0)e^{j2\pi gt}dg + 2f_0.0 = 0$$

The characteristics of raised cosine filter is plotted in fig.15 in frequency and time domain respectively for three different values of α and T = 0.5.

Chapter 4: Results

Fig.16 shows 8 sinc pulse train of receiving end where the sampling instants are 0, 0.005, 0.01, 0.15, 0.2, 0.25, 0.03 and 0.35 respectively. These points are also the zero crossing points visualized from the profile of sum of pulses of fig.16. Since the amplitude of sum of pulses varies with time hence a fixed timing error produce different amount of error at different sampling point. The variation of SNR in dB at receiving end is shown in fig.17 for 3 different timing errors. The relation between the profile of SNR of fig.17 and that of sum of pulses at fig.16 can be explained as: the 'sum of pulse' at the sampling instant is 1 which is same as the amplitude of corresponding *sinc* pulse but variation is found on left or right of the sampling instant. Higher the amplitude of the 'sum of pulses' larger the error signal for timing error/jitter. Similar results are found for raised cosine pulse but the variation of 'sum of pulses' are smaller compared to the case of *sinc* pulse. The similar results for raised cosine pulse are shown in fig.18 and 19. Finally for Gaussian pulse train the profile of 'sum of pulse' resembles the pulse train hence the variation of SNR for a fixed timing error is found flat. The results of Gaussian pulse train are shown in fig.20 and 21. The relative performances of three types of pulse train are: Gaussian pulse train> Raised cosine pulse train <*sinc* pulse train.

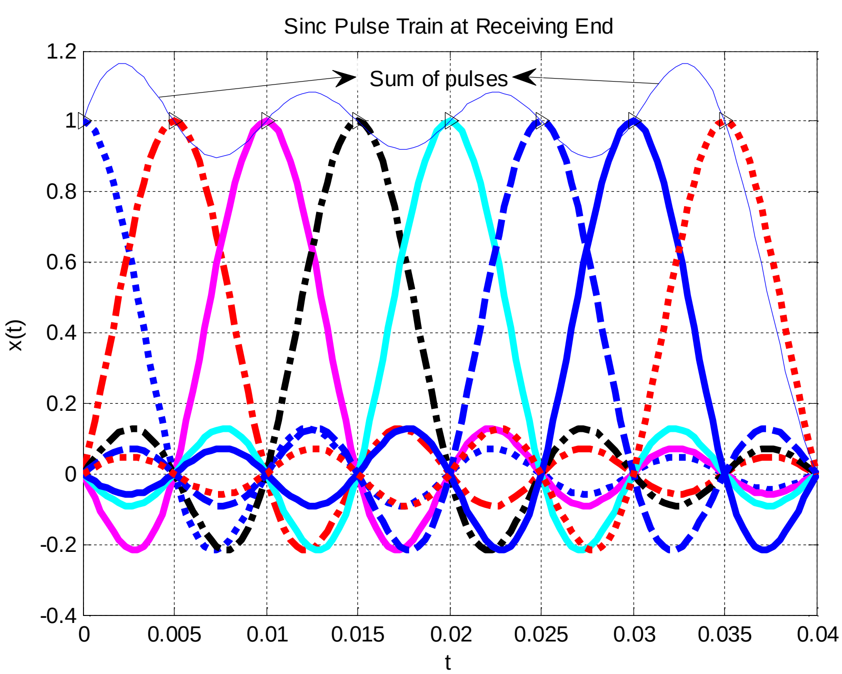

Fig.16: Sinc pulse train at receiving end

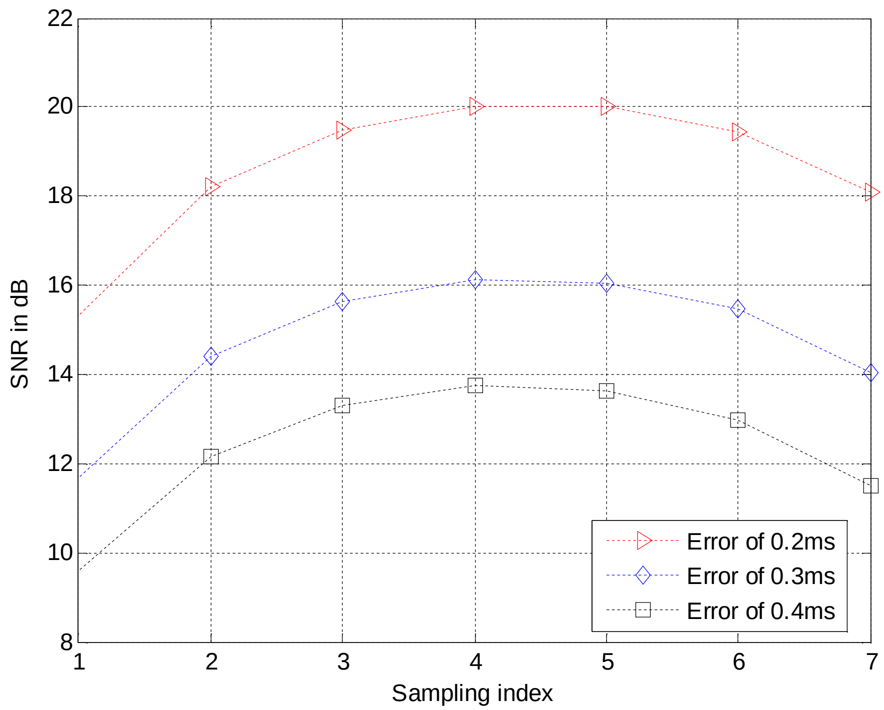

Fig.17: Variation of SNR with variation of timing error at consecutive 7 sampling points for sinc pulse train

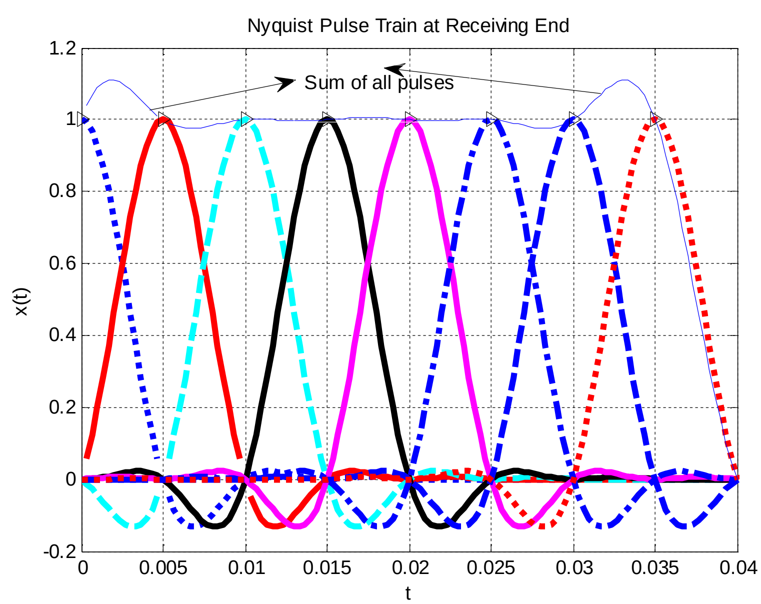

Fig.18: **Raised cosine pulse at receiving end**

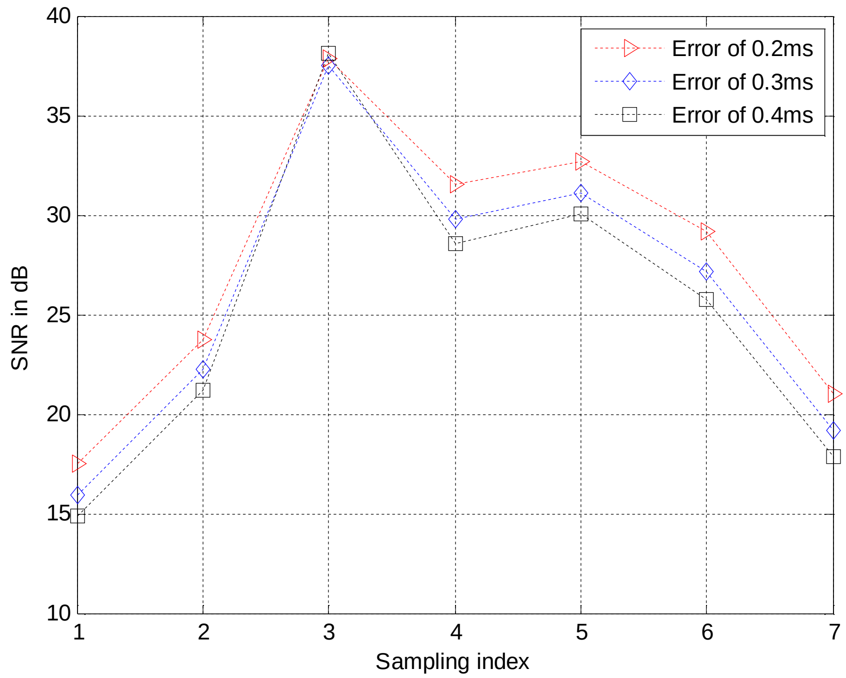

Fig.19: **Variation of SNR with variation of timing error at consecutive 7 sampling points for raised cosine pulse train**

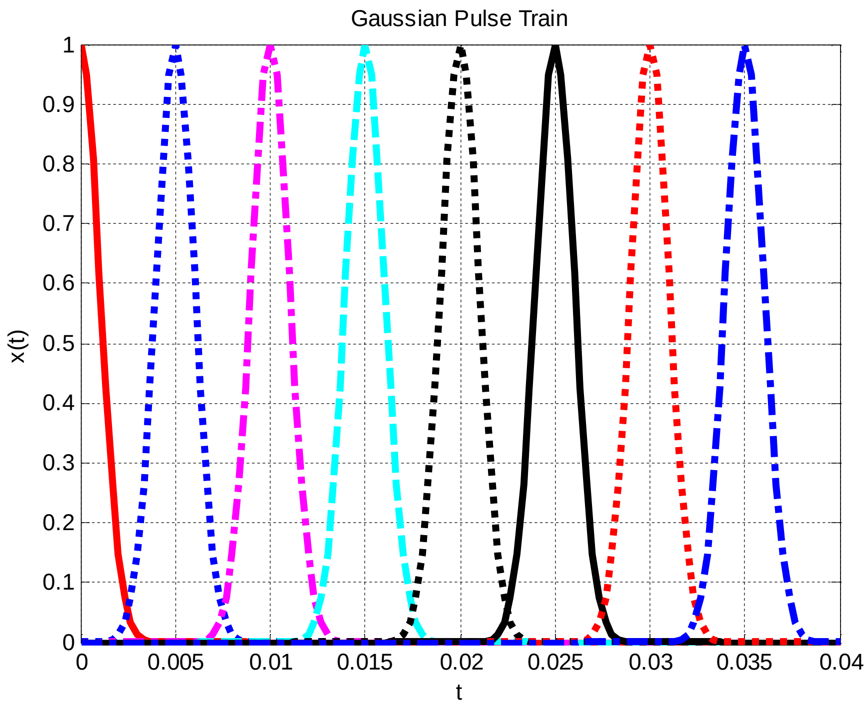

Fig.20: Gaussian pulse at receiving end

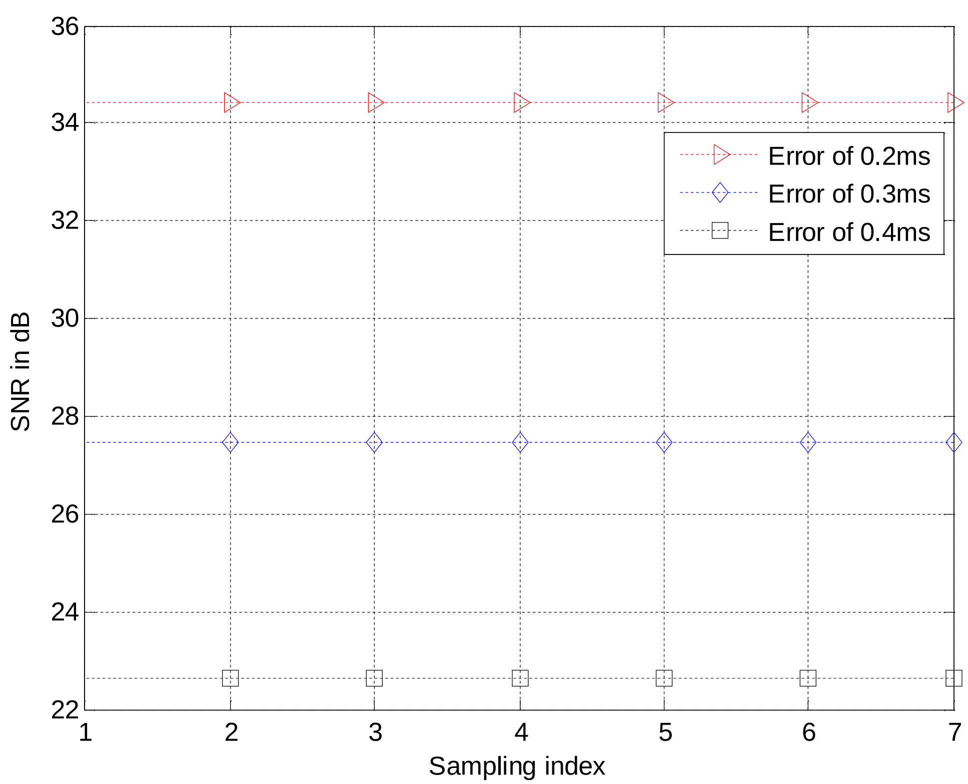

Fig.21: Variation of SNR with variation of timing error at consecutive 7 sampling points for Gaussian pulse train

Chapter 5: Conclusions

The finding of the project work is that Gaussian pulse at receiving end is better than sinc or raised cosine pulse to combat the jitter affect. Still we have the scope to use different smooth window functions (Hanning, hamming, Kaiser, Blackman used in FIR filter) instead of Gaussian pulse to observe the relative performance. In this project work we only consider jitter error but we can consider awgn and fading of signal on communication link to observe the variation of performance.

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List Of Figures