DROOP COMPENSATED CIC DECIMATION FILTER WITH IMPROVED PERFORMANCE

Dissertation submitted in the partial fulfillment of requirements for the award of degree

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Submitted by

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DECLARATION

I hereby declare that the work, which is being presented in the dissertation, entitled "**Droop compensated CIC Decimation filter with improved performance**" is an authentic record of my own work carried out as requirement for the award of degree of Master of Engineering in Wireless Communication Engineering submitted at Electronics and Communication Engineering department at Thapar University, Patiala under the guidance of **Dr. Sanjay Sharma, (Professor& Head)**, Electronics and Communication Engineering department and refers other research work which is duly listed in reference section. The matter presented in this dissertation has not been submitted in any other University/institute for award of degree.

Date : 22.06.2016.

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I certify that the above statement made by student is correct to the best of my knowledge and belief.

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Gaganjot Kaur

ABSTRACT

To fulfill the ever increasing demand of modern electronic devices operating at the different sampling rates, the interest as touched its zenith in the up-sampling based discrete-time signal processing techniques, which can be incorporated by using the efficient digital interpolators and decimators .

From the acquired results, it is evident that the CIC filters are efficient for the low cost applications because the multipliers are not required in its implementation. However, the pass-band droop present in the CIC filters confine the scope of its practical applications. By employing compensation and multi-stage techniques, the response of CIC filter in the pass-band can be significantly improved, but at the cost of escalated hardware requirement and the computational complexity. By employing different techniques the response of filters in passband is considerably improved. And various other sharpening techniques are present to improve filter passband response.

The main aim of this work is to improve the magnitude response of CIC filters and also solve the passband droop problem. For improving the passband and the transition band features of the CIC filter and for improving the performance of CIC filter there are many techniques such as compensation filter cascaded with CIC filter, sharpening technique, poly phase decimation FIR filter to achieve wide broadband compensation of the CIC filter and maximally-flat based compensator filter. The droop can be reduced by modifying the original CIC structure or by connecting an additional filter called CIC compensator in cascade with the CIC decimator. The former approach is based on a technique called sharpening. In recent years, several methods for the design of CIC compensators have been developed and now compensation is combined with a sharpening technique.

In this dessertation, firstly we implemented the CIC with two stage architecture which consist of comb based decimator and sharpened comb decimator in cascade. With this sharpened section operates at lower rate by decimation factor of second stage thus reducing the number of computation and making the filter simpler than previous one and we also studied another technique which improves the response of comb filter in stopband region by changing the position of zeros means optimally rotating the zeros the zeros to place in

stopband in such a way that stopband gets broader. Then based on this technique is new technique called sharpened Modified Comb filter(SMCF) which uses sharpening for passband improvement and rotated zeroes technique for stopband improvement. Also studied Compensation techniques which is also a passband improvement technique on which the very recent research in CIC area is based.

After that we proposed a new Compensated CIC decimation filter with improved response which is a cascade of Sharpened Modified Comb Filter by Laddomada and the Compensation techniques, both sine based and maximally flat compensation. So by using these techniques in cascade the passband response become more flat and the stopband response is wider than conventional CIC. Later, we compared the results of new design with already existing techniques and found that results are much better than earlier ones. So overall flat passband and wider stopband is achieved with this new compensated sharpened modified comb filter.

CONTENTS

LIST OF ACRONYMS

QN	Quantization Noise
FIR	Finite Impulse Response
CIC	Cascaded Integrator Comb
SRR	Software Radio Receiver
MCF	Modified Comb Filter
SRC	Sample rate Conversion
SNR	Signal to Noise Ratio
GCF	Generalized Comb Filter
SMCF	Sharpened Modified Comb Filter
CDMA	Code Division Multiplexing Access

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CHAPTER 1

INTRODUCTION

In the present scenario, digital domain has become the very prominent one as all the problems related to signal processing are resolved in this domain. Sampling is fundamental and very significant operation in discrete time signal processing. The discrete time signals are obtained from continuous time signals using sampling process based on Nyquist criteria [1].

In various applications, it becomes essential to transform from one sampling rate to another one. It may be up sampling or down sampling. In the present modern communication systems, a wide range of sample rates are required depending upon requisite quality, available channel bandwidth and data rate. To make their unification trouble free, adequate sample rate converters need to be used at each interface. Here multi rate signal processing plays a significant role and the vital part of it are digital filters. Digital filters are necessary to keep the spectrum band limited to prescribed bandwidth in accordance with actual sampling rate. In sample rate converters filters are used to tackle aliasing in case of decimation and imaging in case of interpolation. The performance of sampling rate converters is determined by filter characteristics as it is impossible to attain ideal frequency response.

If the existing sample rate is integer multiple of desired, then decimation is used and interpolation for other way round. But if the ratio of output to input sample rate is arbitrary number then the process becomes complicated. They are most used in the commercials like audio players is arbitrary sample rate but it adds unavoidable distortion, although in small amount even for well-designed converters. So we need to design well efficient filters for this. For higher rate change, multistage implementation is used over single stage, but it increases computational complexity and the implementation complexity as well. There are large number of tradeoffs and parameters required for design of filter. So it is necessary to find out techniques or methods by which complexity of filters can be managed within certain limits.

Hogenauer [2] proposed a CIC(cascaded integrator comb) filter, a class of FIR(finite impulse response) filter having linear phase for decimation and interpolation, which require no multiplier that is use limited storage making them a feasible alternate in the conventional implementation for certain applications. These comb filter structures are basically based on moving average filter. The filter characteristics are managed by three parameters, which are number of stages, the differential delay and the number of bits in input/output registers.

The CIC filters are multiplication free filters with limited storage requirements, which make them ideal for the high speed data converters. Hogenauer [2] presented an FIR structure, which consists of cascaded integrator stages working at the higher sampling rate and the same number of comb stages working at the low sampling rate. A number of cascaded integrator comb pairs are chosen to meet the design requirements for aliasing or imaging errors. Although, the CIC filters can implement decimation and interpolation efficiently in the hardware for a wide range of rate change factors, yet CIC filter response is lacking in a flat pass-band response and better transition bandwidth. To circumvent these problems, a compensation FIR filter can be employed in cascade with the CIC filter to provide frequency correction as well as spectrum shaping.

1.1 MOVING AVERAGE FILTER

This is a very inexpensive to realize filters as they do not need multipliers for their implementation. They can be realized just using delays, adders/ subtractors. All the weights are set to one. The Fig. 1.1 shows the eight weight moving average filter.

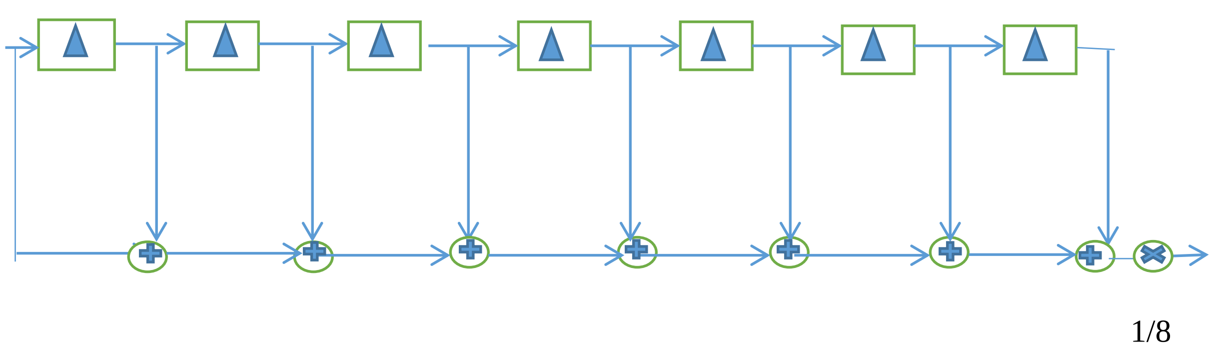


Fig.1.1. Structure of moving average filter with eight delays.

It has a low pass characteristic. For the filter with M weight, there are M-1 spectral zeros in the range 0 to f_s which means there are M/2 spectral zeros in the range 0 to f_s /2. So there are four evenly spaced zeros for eight weight filter the frequency range 0 to 5MHz with the sampling rate of 10 MHz

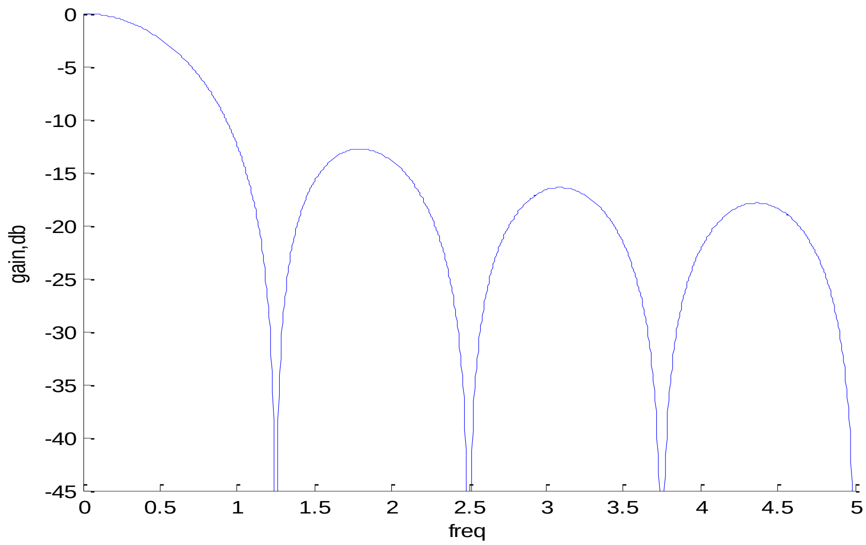


Fig.1.2. Response of moving average filter with eight weights

1.2 DIFFERENTIATOR

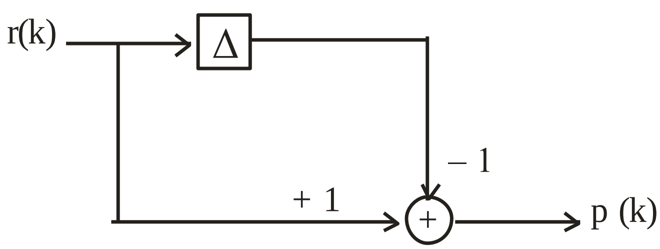


Fig.1.3. Basic structure of differentiator

It is a filter with weight value +1 and -1 only. It has a high pass magnitude response. It does not require any multiplier as the operation can be implemented as a direct connection to adder. Its output can be written as

$$p(k)=r(k)-r(k-1) \tag{1.1}$$

Output in z domain can be written as

$$P(z) = R(z) – R(z)z^{-1} \tag{1.2}$$

which can also be written as

$$P(z) = R(z)[1-z^{-1}] \tag{1.3}$$

Hence the transfer function of differentiator becomes

$$H(z) = P(z)/R(z) = [1-z^{-1}] \tag{1.4}$$

1.3 INTEGRATOR

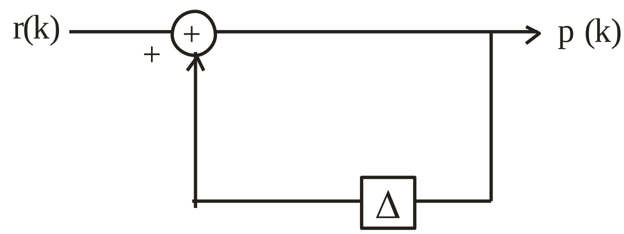


Fig. 1.4. Basic structure of integrator

It is a single weight IIR filter as shown in Fig.1.4. It does not require multiplier and has low pass characteristics. Its output can be written as

$$p(k) = r(k) + p(k+1) \tag{1.5}$$

In z domain it can be written as

$$P(z) = R(z) +P(z)z^{-1} \tag{1.6}$$

Hence the integrator transfer function

$$H(z) = P(z)/R(z) = 1/[1-z^{-1}] \tag{1.8}$$

1.4 COMB FILTER

This filter is derived from moving average filter and its architecture is similar to differentiator with the single delay in differentiator replaced by the cascading of N delays. Fig.1.5 shows the comb filter with 8 delays. It has N evenly spaced zeroes from 0 to f_s.

Response of eight delay comb filter is shown in Fig.1.5. Its response resembles a comb that's why it is called comb filter.

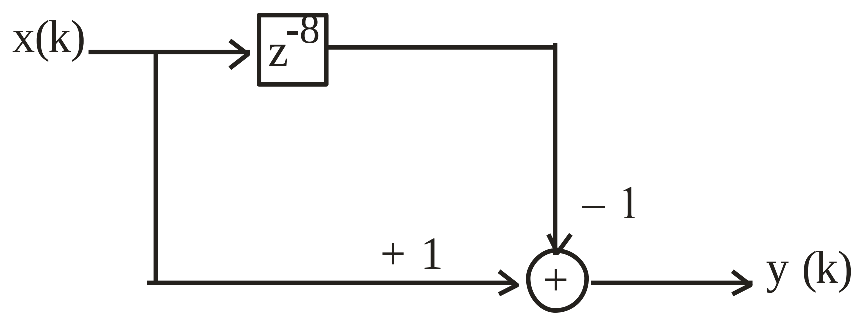


Fig.1.5. Structure of comb filter with eight delays

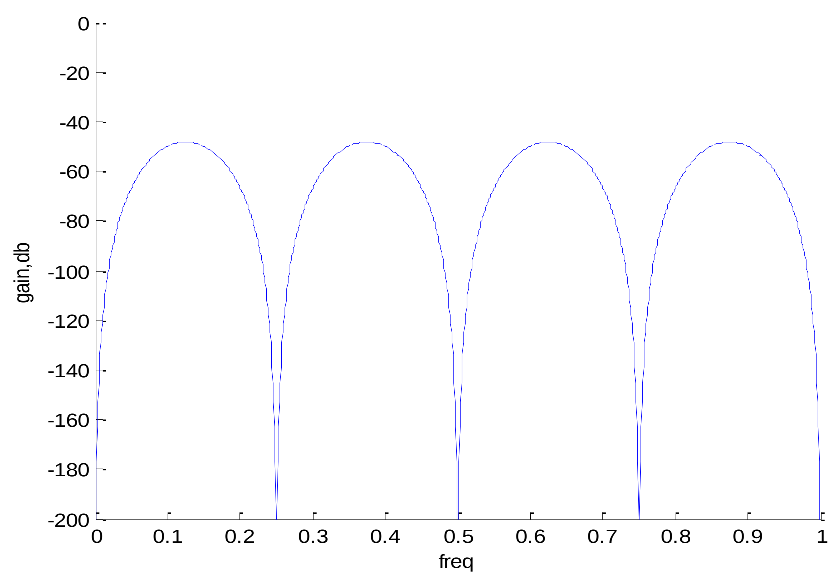


Fig.1.6. Frequency response of comb filter.

1.5 CASCADED INTEGRATOR COMB

It is a combination that is cascade of integrator and comb explained in previous sections. Hence called CIC (cascaded integrator comb). Fig.1.7 shows the structure of integrator comb with eight delay elements.

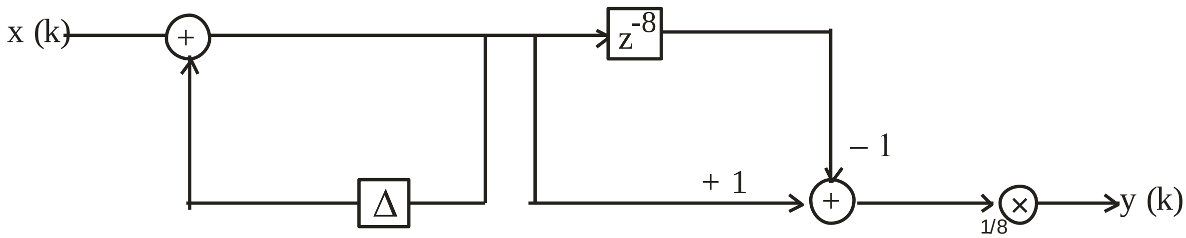


Fig.1.7. Integrator comb structure with eight delays.

Here transfer function of comb filter is written as

$$H(z) = \frac{1}{8}(1 - z^{-8})\left(\frac{1}{1-z^{-1}}\right) \quad (1.9)$$

It may be written as

$$H(z) = (\tfrac{1}{8})\,(1+z^{-1}+z^{-2}+z^{-3}+z^{-4}+z^{-5}+z^{-6}+z^{-7}) \quad (1.10)$$

As we can see M weight comb is similar to moving average filter with weight M-1. Fig. clearly shows that it has an advantage over moving average that it has only two additions compared to eight in moving average.

In the generalised form we can write transfer function as

$$H(z)=\frac{1-z^{-R}}{R(1-z^{-1})} \quad (1.11)$$

So according to equation H(z) can be implemented as a cascade of comb section $(1 - z^{-R})$ and integrator section$(1/\,1 - z^{-1})$ which leads to a efficient device that perform filtering operatin with only two additions irrespective of filter length R. The term CIC (cascaded integrator comb) is often used for this filter.

1.6 CHARACTERISTICS OF CASCADED INTEGRATOR COMB FILTER

Its frequency domain its response can be written as

$$H_c(e^{j\omega})=\frac{1}{R}\frac{\sin(\frac{\omega R}{2})}{\sin(\frac{\omega}{2})}\,e^{-j\omega[\frac{R-1}{2}]} \quad (1.12)$$

It is a linear phase low pass filter with sinRx/sin x amplitude characteristics. Due to this response it is also called sinc filter. CIC filter with transfer function $H_c(e^{j\omega})$ exhibits a comb like response. It has nulls at f=1/R, the response is shown in Fig. 1.8.

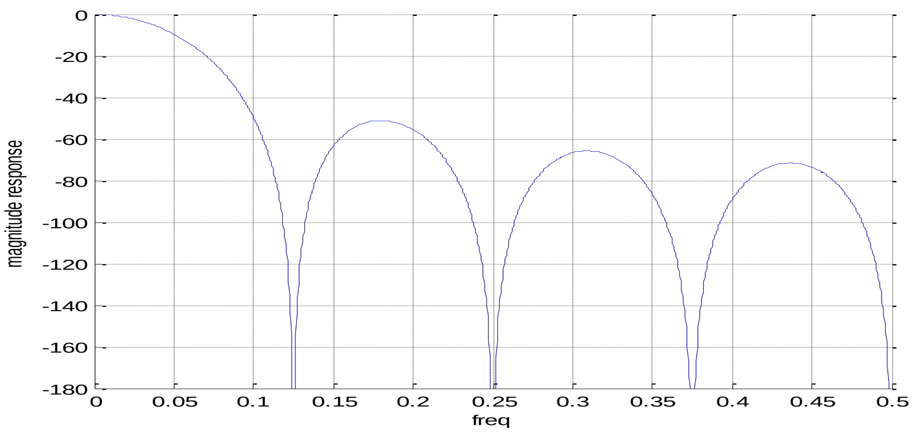


Fig.1.8. Frequency response of CIC filter

We can use only two parameters to modify the magnitude response of Comb filter which are R i.e filter order and K which is the number of comb filter section(explained in next section). We observe from the frequency response that firstly, nulls occur at exactly the integer multiple of F/R (F is the input sampling frequency) which means there is a maximum alias suppression at these frequencies.

Secondly, aliasing bandwidth (bandwidth around nulls) is narrow which is too small for sufficient suppression of aliasing in whole baseband of signal. Thirdly, the pass band characteristics of a filter produce droop in pass band called pass band droop which needs to be compensated for many applications.

1.7 CASCADE INTEGRATOR COMB FILTER (CIC) AND CONCEPT OF PASSBAND DROOP

As CIC shows a poor magnitude characteristics so we improve this by cascade of number of stages. Multistage comb filter is composed of K number of stages and its transfer function is given by

$$H_{C(K)}(z) = \left[\frac{1-z^{-R}}{R(1-z^{-1})}\right]^{K} \quad (1.13)$$

A CIC filter consists of an equal number of stages of the ideal integrator filters and the comb filters. Its frequency response may be tuned by selecting the suitable number of cascaded integrator and comb filter pairs. The extremely symmetric structure of the CIC filter authorizes effective implementation in the hardware. The CIC filter can also be implemented very efficiently in hardware due to its symmetric structure. The CIC decimator would have K cascaded integrator stages clocked at fs, followed by a change in the rate by a factor R, followed by K cascaded comb stages consecutively running at fs/R. The Fig. manifests the multi stage decimating CIC filter. The CIC interpolator would have K cascaded comb stages running at fs/R, followed by interpolation function, followed by K cascaded integrator stages running at fs. The Fig. shows the multistage interpolating CIC filter. The numerator is the transfer function of a differentiator (1.4) and the denominator indicates the transfer function of an integrator (1.8).

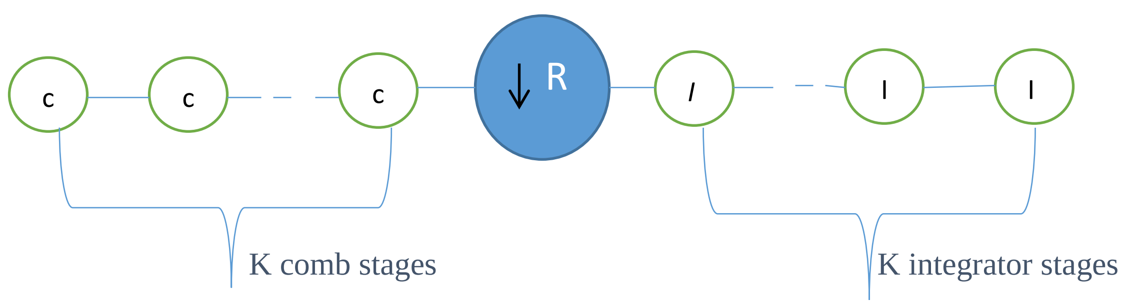


Fig.1.8. Decimating CIC filter with K stages

Next Fig. gives the response with increase in number of stages of CIC filter that how the pass band droop got worsened and the stop band got better. It shows that multi stage implementation improves stop band and selectivity of filter. Stop band attenuation in nulls is high as the filter possesses multiple nulls whose multiplicity is equal to number of stages.

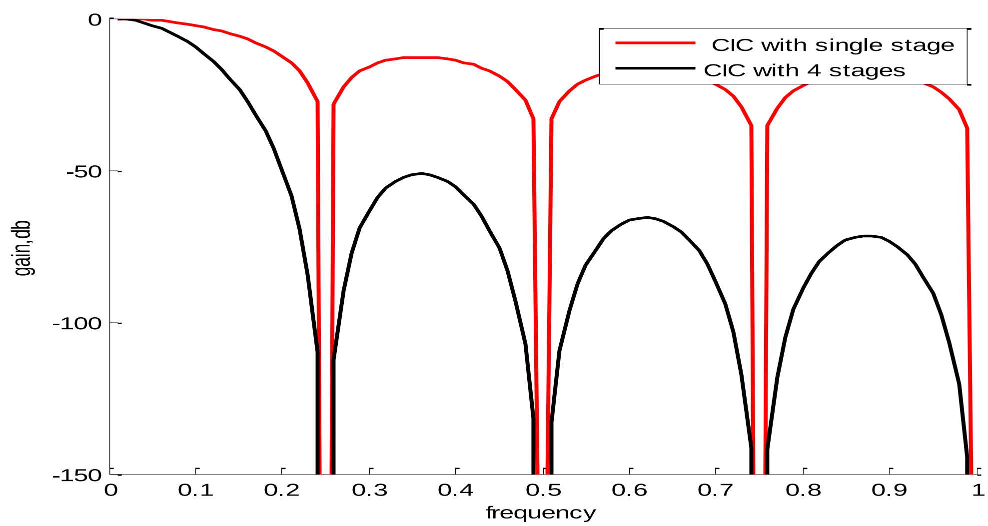


Fig.1.9. €ffect of increase in number of stages on comb filter response.

The decrease in the magnitude response of pass band is called pass band droop. As we see as we increase the number of stages, stop band attenuation increases but simultaneously droop at low frequencies becomes prominent and it creates problem as we are always interested in lower frequencies.

1.8 CIC FILTERS IN DECIMATION AND INTERPOLATION

Decimation is process of reducing the sampling rate i.e when the desired sampling rate is less than existing sampling rate and when the desired sampling rate is more than the existing one the process known as interpolation plays its role. These processes need to be used with care they give rise to aliasing and imaging.

Decimation gives rise to aliasing so here we need aliasing filter to get error free output whereas interpolation gives rise to imaging so we need anti imaging filter for that. Here CIC filters are used as antialiasing or anti imaging filter [1]. The configuration composed by cascading decimator and interpolator is called recursive realization structure.

The main advantage of comb filters is to perform decimation and interpolation without multiplying operation . This is of great advantage when we operate at high frequencies.

While considering implementation of these filters we can expect register growth since the integrator is having unity feedback coefficient. This can be overcome by using two's compliment arithmetic and using range of number system more than the maximum expected magnitude.

1.9 ADVANTAGES AND DISADVANTAGES OF CIC FILTERS

There are some advantages [2] of the CIC filters :-

1) No multipliers are required;

2) No storage .is required for filter coefficients;

3) Intermediate storage is reduced by integrating at the high sampling rate and comb filtering at the low sampling rate, compared to the equivalent implementation using cascaded uniform FIR filters;

4) The structure of CIC filters is very "regular" consisting of two basic building blocks;

5) Little external control or complicated local timing is required

Some disadvantages of CIC filters are:-

1) Register widths can become large for large rate change factors

2) The frequency response is fully determined by only integer parameters

(R, K and N) resulting in a limited range of filter characteristics.

1.10 MOTIVATION FOR THESIS

Wireless broad band communication is experiencing a fast development. To match up this noteworthy growth, new technologies are required to enhance system performance. With increase in cost pressure on equipment makers for wireless system there is a great need to limit both operating expenditures and capital expenditures attributes of communication system. In modern communication systems the basic building blocks are entities those increase or decrease the sampling rate, called interpolators or decimators.

In this thesis we use CIC filter which is a cost effective way to implement digital decimators as they do not require any multipliers and are very easy to implement. They are implemented using only delays and summations. The number of stages can be chosen to meet the requirements of aliasing and imaging and also they find applications in wide area of communication as software defined radios and many other areas. They possess a linear phase like FIR filters.

Although using CIC filters we can implement wide range of rate change factors, yet CIC filter is lacking flat pass band. Moreover it degrades as we increase the number the stages. So the main goal of this thesis is to improve response of CIC filters i.e. reduction in pass band with simultaneous broadening of stop band.

1.11 PROBLEM FORMULATION

Today modern digital communication is expanding rapidly. To meet stringent requirements posed by todays wireless communication system, a low cost and high performance architecture must be developed with significant reduction in cost of equipment. One basic function of modern digital communication is digital decimation. Decimation can be realized with help of cascade of CIC filters. However, disadvantage of CIC filter is that its passband is not flat which is undesirable and also stopband is not that wide to remove aliasing properly.

So we need techniques with which we can improve the performance of decimators to achieve various sampling rates. For high rate change we require a multistage filter which increases the complexity of system and also increases the pass band droop .So we need to develop a technique for efficient realization of decimators.

This paper extends the class of sharpened modified comb filter(SMCF) introduced by Missimilano Laddomada for $\sum\Delta$ A/D convertors in order to obtain more improved response i.e. lesser passband droop than SMCF and stopband response better than basic CIC with the use of compensation techniques along with SMCFs. The overall response improvement is targeted in this paper.

1.12 ORGANISATION OF DESSERTATION

The dissertation consists of six chapters including introduction as follows:

- *Chapter 2* deals with literature survey in which study on existing improvement methods of CIC filters is discussed. All the previous work done is summarized in this chapter.
- *Chapter 3* discusses about the passband improvement techniques such as sharpening, sharpened modified comb filter and compensators with their results. It also discusses the stopband improvement technique with rotated zeros.
- *Chapter 4* presents the new design which is obtained by the cascade of previous existing techniques that is sharpened modified comb filter with compensator to obtain sharpened modified comb filter.
- *Chapter 5* discusses the results obtained with the new cascade technique and compares it with already existing techniques
- *Chapter 6* highlights the conclusion of dissertation and tells about future scope.

CHAPTER 2

LITERATURE SURVEY

The literature review reviews, deduces, and evaluates various existing "literature" (or available material) in command to set up current understanding of a subject. The reason for doing so relates to in progress research to build up that knowledge. The literature review may determine a disagreement, establish the need for further research, and define a topic of inquisition.

E. B. Hogenauer et el. [2] proposed CIC filters which are an economical replacement for filters used in decimation and interpolation. This belongs to class of FIR filters and has linear phase. The main reason for their popularity was that they were very cost effective as they do not require any multiplier and use limited storage. CIC that is cascaded integrator comb consists of cascade of equal number of integrator stages working at higher sampling rate and equal number of integrator stages working at lower sampling rate. To meet the design requirements, the number of CIC pairs is chosen accordingly. The frequency response is fully controlled by only three parameters making the range of characteristics very limited. And also its passband suffers from the problem of passband droop then afterwards various scientists proposed ways to improve the passband droop. So the next discussion is mainly focused on ways to improve characteristic of CIC filter.

A. Y. Kwentus at el. [5] proposed a new architecture to improve response of CIC filters. This technique is mainly based on the use of sharpening technique given by Kaiser and Hamming [6]. He used the second order sharpening polynomial with CIC and the achieved sharpened CIC showed great characteristic improvement over traditional comb. The technique also reduced hardware requirement over existing ones as it allowed the second stage CIC(compensating stage) be followed by fixed coefficient filter rather than programmable one. This helped to improve overall throughput rate and architecture suits well for single chip VLSI implementation with high sample rate.

J. F. Kaiser et al. [6] presented a method of filter sharpening for symmetric non recursive filter. Filter sharpening is to make the filter response better. Better means less pass band error and more out of band attenuation. The main approach is to process the data through same filter repeatedly but while each pass increases the stop band attenuation but also increases the pass band error to undesirable level. The length (order) of filter of the equivalent filter also gets increased. It showed how to connect multiple instances of same filter to get better response. The method he described was based on amplitude change function or polynomial which is restricted to symmetric non-recursive FIR filter with piecewise constant passband and stopband. The proposed polynomial is one which passes through point (0,0) and (1,1) and has nth order tangency at zero and m^{th} order tangency at unity. It means the curve of amplitude change was tangent horizontally at both zero and unity. Hence giving more flat area.

T. Saramaki at el. [7] introduced a new structure for CIC implementation based on [8]. Comb filter in cascade is known to provide an effective first stage for multistage VLSI implementation of decimators. In this structure tapped interconnections were used instead of direct cascade. With this structure, the number of comb filters required were reduced considerably and also the silicon area was reduced for VLSI implementation. Data word length which is required for internal calculations was also considerably reduced. So a significant saving was achieved with this structure.

H. K. Yang at el. [9] proposed poly phase architecture for comb (CIC) decimation filter. The filters were then able to operate at very less sampling rate simultaneously achieving the same performance as conventional CIC. As this structure [10] used poly phase to further implement parallel processing which is used for high speed and low power consumption. So the speed of operation improved with low power consumption. After this partial poly-phase structure [11] was also proposed based on partial polyphase decomposition and parallel processing techniques. With their use complicated polyphase decomposition was avoided when filter order and decimation ration was high.

H. J. Oh at el. [12] used interpolated second order polynomial (ISOP) for to design efficient CIC. The use of ISOP reduced the passband droop which appears in conventional CIC filtering with simultaneous little degradation in stopband attenuation. ISOPs also simplified the half band filters which follows CIC filters. The new presented half band filters were much simpler than conventional ones. The filter given in paper is a cascade of CIC, ISOP, MHBF (modified half band filter) and programmable FIR filter. The complete design was given and this scheme was more efficient than existing ones for programmable down conversion. It was also observed that the use of ISOP's was effective in reducing conventional complexity.

J. O. Coleman at el. [13] proposed the use of Chebyshev polynomial to CIC filter for improvement in stop band. The conventional CIC filter has multiple stop bands but with the use of Chebyshev polynomial these multiple stop bands were merged to one but the pass band remained the narrow peak. For the Nth order system, the depth improved by 6(N-1) dB or roughly we can say 6(N-1) dB of stop band was added to Nth order system. The increase in computational complexity was modest that is it requires few low speed additions and multiplications by small integer coefficient which could be chosen as power of two. In arrays (1D and 2D) applications, it should be used where SNR is of little concern because Chebyshev sharpened response suffers from SNR taper.

L. L. Presti at el. [14] presented a scheme which consisted of two stages. The first one was obtained by the rotation of pole zero distribution of CIC filter in z plane and the second was allowed to design with relaxed specifications means any algorithm can be used to design it. The scheme was mainly designed for efficient multistage decimation filter to be used in sigma delta converters. In this the first block is very critical and it is within this block the quantization noise is eliminated.

The recursive structure proposed in this filter shows linear phase and is moreover very flexible as the nulls can be easily located in any place of null interval. The user can even choose the number of null points but it is with increase in number of multiplier by 2. Thus the performance improved with little increase in complexity. The second stage can be designed with any algorithm and it can further be divided in sub stage and RS filter can be used as first sub stage.

M. Laddomada at el. [15] presented efficient decimation sinc filter for software radio. The main idea used here was to free the position of zeros of classical filter to achieve high frequency selective behavior. This was obtained by rotating zeros in two opposite directions, one in clockwise and other in anticlockwise direction and then using cascade of them which results in recursive structure with more selective frequency behavior around noise fold frequency bands. The structure exhibits a low complexity and null interval can be chosen by varying value of one parameter that is angle of rotation.

A. Saud and G. L. Stuber at el. [16] gave a modified approach for sample rate conversion in software radio system. The conventional CIC are not suitable for SWR especially for factors close to unity. However they were attractive for SWR because they use only one addition/subtraction but have limited tuning parameters. This paper proposed an SWR receiver `which located high power channel and set the zeros of CIC filter close to these images and hence provide higher attenuation.

The modification was achieved by spreading the delays of CIC filter. Here the delays were either evenly distributed for uniform image attenuation or were set near about specific values for additional suppression of strong images. It provides high SNR and improved image attenuation than conventional one that is it gains improved performance with small increase in computations.

G. J. Dolecek and S.K. Mitra at el. [17] proposed a two stage architecture. The first stage consists of conventional comb decimator and second stage a sharpened comb decimator. The main advantage of using this was that sharpened section now operated at lower rate by the first stage decimation factor and also polyphase decomposition can be used at first section. This structure has much better aliasing rejection than conventional one and even sharpened comb. The passband droop was however similar to original sharpened but smaller than conventional one.

C. Zhang at el. [18] proposed non recursive structure for decimation filter. It was an alternative to conventional CIC with decimation ration of m^{th} power of 2 and m^{th} power of 3. It was a very power implementation but silicon area for non-recursive exceeded the area of recursive structure. However 70% of power saving was achieved with this new implementation.

G. J. Dolecek and M. Laddomada at el. [19] proposed a new scheme with design of GCF's (generalized comb filter) to quantize the multipliers in z transfer function with employment of power of two (PO2) terms. GCFs are very efficient anti-aliasing decimation filter with improvement in selectivity and QN rejection around folding band with respect to conventional comb filter. It is done in such a way to meet perfect pole zero cancellation. It totally avoids instability problem compared to classic comb filters. GCF's are very efficient as anti aliasing decimation filter as they have improved selectivity and better quantization noise rejection performance around the folding bands compared to conventional filter and also in paper the use of simple droop compensator is proposed to recover the passband droop with distortion in useful digital signal in baseband.

A. F. Vazquez and G. J. Dolecek at.el. [20] presented a design of compensation filter of GCF (Generalized Comb Filter). It is applied to different constraints i.e. maximally flat, least square and minimax and the coefficient for these three cases is obtained by solving linear equations. The proposed filter considerably reduces the pass band droop of GCF and also operates at low rate. Technique is based on 2R order compensation which after applying identity becomes second order filter. Only one multiplier is required for max flat and least square designs. So they are good choice for narrowband compensation and minimax design is best for wideband compensation which requires two multipliers.

Jimenez and G. J. Dolecek at el. [21] introduced the use of generalized sharpening technique for the improvement in amplitude characteristics of comb filter in the pass band and stop band region. They extended the use of two stage. At first stage operation can be performed at low rate by making use of poly phase decomposition. At second stage simple compensator is applied for improvement in pass band characteristics. Then the generalized sharpening technique was used for reduction in pass band droop due to CIC at first stage so the overall results obtained shows better characteristics than previously proposed ones. The increase in computational complexity is almost none.

M. Laddomada at el. [22] used the two stage architecture and employed modified sinc filter by Presti i.e. the one obtained by optimal rotation of zeros at first stage and sharpening at second stage to get class of sharpened modified comb filter (SMCF) which to aimed to increase the quantization noise rejection around folding bands with the simultaneous improvement in passband droop compared to conventional comb filter. The main focus was given to equivalent third order modified sinc filter with which noise suppression of 8db was achieved compared to classic third order filter. The scheme was given for design of delta sigma analog to digital converters.

K. S. Yueng and S. C. Chan at. el. [23] presents the design and multiplier free realization for software radio receiver (SRR) with the reduction in system delay. It employ a low-delay finite-impulse response (FIR) filter and digital all pass filters for effectively reducing the system delay of multistage decimator in software radio receiver . Optimal least square design and minimax designs of these low-delay finite impulse response and all pass-based filters are formulated as semi-definite programming (SDP) problem, which allow zero magnitude constraint at $\omega=\pi$ to incorporate readily as additional linear matrix inequalities (LMIs). With the implementation of sampling rate converter using a variable digital filter (VDF) immediately after integer decimator, the needs for an expensive programmable finite impulse response filter in the traditional SRR is neglected. A new method for the optimal minimax design of this VDF-based SRC using SDP is also proposed and compared with traditional weight least squares method. Other implementation issues including the multiplier-less and digital signal processor (DSP) realizations of the SRR and the generation of the clock signal in the SRC is also studied. Design results show that the system delay and implementation complexity of the proposed architecture is considerably reduced as compared to conventional approaches.

G. J. Dolecek at. el. [24] presented a simple second order sine based CIC. Depending on the number of stage K, there is only one design parameter which again depends on whether compensation is narrowband or wideband. It just uses three additions/subtractions to perform efficient compensation. The method requires very less computations and is less complex. CIC compensation is convenient in less than 3/5 of total band.

G. J. Dolecek and F. Harris at. el. [25] presented CIC compensation filter with sharpening technique. Sharpening is applied with a goal to improve pass band. The polynomial with m=1 and n=0 is used. The complexity is not increased. The new filter provide wideband and narrowband compensation in band ¾th of overall band. There is one design parameter b which depends on the no of cascaded CIC filters.

G. J. Dolecek and S. K.Mitra at. el. [26] proposed simple method to design multiplier less CIC decimation filter. It improves the gain response in pass band and also maintains simplicity of CIC filters. The parameter b which depends on number of stages K controls the droop and number of stages controls the alias rejection. Smaller value of b is used for wider pass bands. The scheme is also compared with already existing known methods that similar results are obtained with less computational complexity

G. J. Dolecek and F. Harris at el. [27] use the two stage given by Dolecek. He used the compensation filter in the two stage architecture. The first stage is implemented in non-recursive form but can also be done using polyphase implementation. Compensation and sharpening were applied to second section. Here the goal was two fold, one is to obtain lesser pass band droop of overall filter and to avoid the use of integrator at high sample rate. The proposed structure in this paper achieved this and also at first stage with the use of poly phase decomposition all the filtering is moved to lower rate than input one but the complexity is increased slightly.

G. Molnar at el. [28] proposed a filter called CIC compensator for reduction in droop of CIC filters. The high order finite impulse response filter based on maximally flat criteria is presented. The coefficients of compensator were obtained by solving linear equations. If implemented with double precision arithmetic it allows the compensator design up to 18 coefficients. They give efficient results with wideband system and in case of narrowband which use high order decimation filters.

S. Kim, W.C Lee, S. Ahm and S. Choi at. el. [29] presented a roll off compensation filter for reduction in roll off characteristics of CIC filter for W-CDMA digital IF receiver. The bit error rate is minimal with the compensation applied to roll off characteristics. The CIC roll off compensation filter is convolved with channel selection filter. Both have symmetric characteristics and roll off phenomena is compensated by using frequency response characteristics of received signal as weighting function. Also this method can be used in design of interpolation in transmitter and DAC (digital to analog) compensation filter. This method is also applicable to CDMA-2000.

F. J. Torres and G. J. Dolecek at el. [30] this paper presents the efficient modification of in CIC-cosine decimation filter. The second order compensating filter is applied at the last stage of decimation in order to improve pass band of interest. The coefficients of compensating filter are presented as canonical signed digits (CSD) and can also be implemented only with use of adders and shifts. Therefore, the resulting filter is a multiplier less filter and exhibits a high attenuation in the region of stop band, as well as a pass band has lesser droop.

G. J. Dolecek at. el. [31] presented a design for wideband compensation of CIC filter. The new proposed compensation is multiplier less and has linear phase finite impulse response and works at low sampling rate. It requires only three additions/subtractions. The structure is independent of R (decimation factor) and K (number of stages). However it requires K or K-1 stages. For $K \leq 3$ and $K>3$, compensator stages are K and K-1 respectively thus requires 3K and 3K-1 total number of adders. They present the comparison with known methods and show how the compensation is better than existing techniques and also less complex.

A. F. Vanquez and G. J. Dolecek at. el. [32] introduced design and implementation of maximal flat CIC compensation filter. Closed form equation is given for computation of filter coefficients for second and fourth order filter for narrow band and wide band compensation. Multiplier less implementation is considered and complexity depends on decimation factor and number of stages> decimation factor for narrowband must be power of two and for wideband it is of the form $2^{2R} - 1$, for R>0. Filter coefficients for second order filter are calculated by minimization of square error in passband, the design provides flat response in first half of passband. It has optimal implementation complexity which is 12 adders and 12 delays. However compensation filter does not adverse the attenuation in aliasing band of CIC filter.

G. J. Dolecek and L. Dolecek at. el. [33] presented a design of multiplier-less CIC compensation filter. The maximum deviation is less than 0.4 dB for pass band frequency of π/2R. By use of multi rate identity it becomes second order filter and also provides better droop compensation in wide band than existing techniques. It does not require multipliers and only maximum of five additions are required in this case.

C. Cagatay et al. [34] provided the optimization framework to select CIC sharpening polynomial and to implement it efficiently by Saramaki Ritoniemi structure [7]. He proposed the use of application specific polynomial or optimal sharpening polynomial compared to generic sharpening polynomial. These optimal polynomials gave high performance results eliminating the need of secondary compensation filter which was required in conventional systems. These optimally sharpened filters were then implemented with Saramaki Ritoniemi structure.

D. E. Tronsco Romero at el. [35] used optimization of Cagatey and presented optimal sharpening of compensated comb filter. They showed that compensated filter or sharpening compensated CIC filter needs low degree sharpening polynomial in comparison to sharpened CIC filter without compensation as far as magnitude specifications are concerned. It all resulted in complex solution. A low complexity scheme was introduced by the use of three addition compensator with optimization based derivation of sharpening polynomial. Sharpening coefficients were integers scaled by PO2 (power of two) terms which leads to low complexity structure. The results showed the great performance improvement in passband droop and selectivity as compared to other traditional methods like Kaiser sharpening and Chebyshev technique.

L. V. Haresh at el. [36] presented the generalized scheme for design of wide band comb based decimation filter. The design used maximally flat compensator and filter sharpening. It provided wideband compensation in passband region without any degradation in stopband attenuation. The multistage implementation is considered and each stage is implemented by maximally flat second order compensation filter. Sharpening is used at last stage and the last stage works at lower rate by decimation factor of previous stages. Poly phase decomposition is also applied to non-recursive form which resulted in lower power consumption

CHAPTER 3

SHARPENED AND MODIFIED COMB FILTERS

As we saw in earlier chapters that CIC filter suffers from passband droop and also the integrator section operates at higher sampling rate. So here in this chapter we discuss the methods to improve passband droop and reduction in sampling rate in detail.

Here we discuss two stage structure where in first section normal comb filter is used and second section uses sharpened comb. At the first section poly phase decomposition can be used to further reduce the rate. In this section we discuss sharpening and two stage structure. The method for improvement in stopband attenuation is also discussed.

3.1 SHARPENED CIC FILTER

It is a method to sharpen the magnitude response of a filter which uses multiple realizations of low-order filter. The approach is to process the data repeatedly through same filter. This is done to achieve better results, better means less passband error and more stopband attenuation. This whole process is named as sharpening. The idea of amplitude change function was restricted to symmetric non recursive filter. Amplitude change function tells that what amplitude in input transforms to what amplitude in output. The polynomial for the response output is given as

$$H_{nm}(f) = H^{n+1}(f)\sum_{k=0}^{m} \frac{(n+k)!}{n!k!} [1 - H(f)]^k \quad (3.1)$$

Where H (f) defines low-order basic filter. The integers n and m are non-negative and measures the number of the non-zero derivatives of $H_{nm}(f)$ at the points where $H_{nm}(f) = 0$ and $H_{nm}(f) = 1$, respectively. By choosing n and m, we select the order of tangency at the frequencies where $H_{nm}(f) = 0$ and $H_{nm}(f) = 1$. More tangency means more flat response.

The simplest case is given with n=m=1, with which the equation becomes

$$H_{11}(f) = H^2(f)[3 - 2H(f)] \quad (3.2)$$

This technique is applicable for linear-phase FIR filters. For H (z) being a causal linear-phase FIR filter having a group delay of D samples, the transfer function $H_{11}(z)$ that is for n=m=1 is given by

$$H_{11}(z) = H^2(z)[3z^{-D} - 2H(z)] \tag{3.3}$$

For the implementation of $H_{11}(z)$, we need 3 copies of H (z), an integer multiplier of value 3, a trivial multiplier of value –2, an adder, and a delay line of D samples. The block diagram that implements the sharpened filter $H_{11}(z)$ is shown in Fig.3.1.

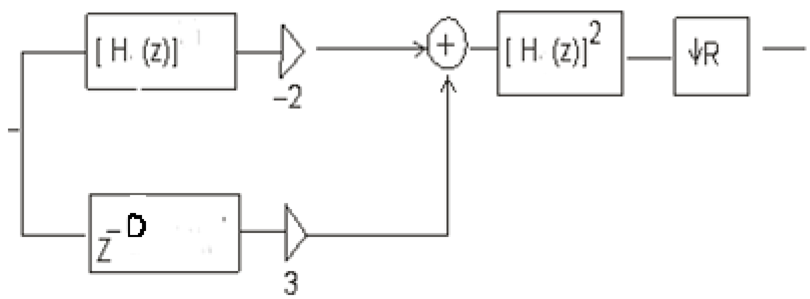


Fig.3.1. The block diagram of sharpened filter

If the original filter H(z) issued as a basic filter, the resulting filter will have reduced passband droop and even improved stopband.

The frequency response for sharpened CIC with H (z) as the building block and L stages is given by

$$H(e^{jw}) = 3\left(\frac{1}{R}\frac{\sin(\frac{\omega R}{2})}{\sin(\frac{\omega}{2})}\right)^{2L} - 2\left(\frac{1}{R}\frac{\sin(\frac{\omega R}{2})}{\sin(\frac{\omega}{2})}\right)^{3L} \tag{3.4}$$

Since the basic building block has linear phase so the overall filter will also have linear phase. If the filter stage which is following the sharpened CIC has decimation factor of 8 then the edge of frequency of passband of interest is given by normalized frequency 1/16R and the edge for the first aliasing band is given by 15/16R, here R is the decimation factor.

Next we show the effect of sharpening on the characteristics of filter response which is as given below.

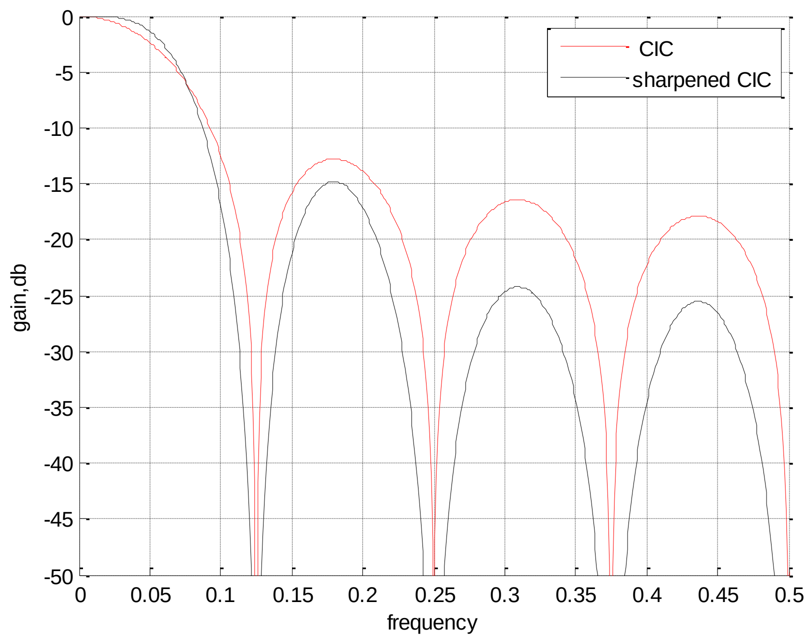


Fig.3.2. Comparison of sharpened CIC with conventional CIC

3.2 TWO STAGE IMPLEMENTATION

Various methods have been proposed till date by various scientists to solve the problems of CIC filters. The methods proposed solve either one of the problem but not both. The basic idea behind this two stage implementation is to integrate the advantages of various methods presented in [7]and [1] to get a new structure which while operating at lower sampling rate achieves the performance improvement over the original comb.

Here we are considering the case where the sampling rate conversion factor R is expressed as a product of two integers R = R1R2. In this case, the transfer function H (z) of decimation filter can be written as the product of two filter sections as given below

$$H(z) = [H_1(z^{R1})H_2(z)]^K \tag{3.5}$$

Where
$$H_1(z^{R1}) = \frac{1}{R_2}\left[\frac{1-z^{-R1R2}}{1-z^{-R1}}\right] \tag{3.6}$$

$$H_2(z) = \frac{1}{R_1}\left[\frac{1-z^{-R1}}{1-z^{-1}}\right] \tag{3.7}$$

The amplitude response are given as:

$$H(\omega R1) = \frac{1}{R_2}\left(\frac{\sin(\frac{\omega R}{2})}{\sin(\frac{\omega R1}{2})}\right) \tag{3.8}$$

$$H_2(\omega) = \frac{1}{R_1}\left(\frac{\sin\left(\frac{\omega R1}{2}\right)}{\sin\left(\frac{\omega}{2}\right)}\right) \tag{3.9}$$

The specific roles for filters $H_1(z^{R1})$ and $H_2(z)$ in the overall decimator are given as, Filter $H_1(z^{R1})$ is providing sharpening, and $H_2(z)$is for improvement in the stopband attenuation.

The pole zero plot for H(z), $H_1(z^{R1})$,$H_2(z)$ are given below

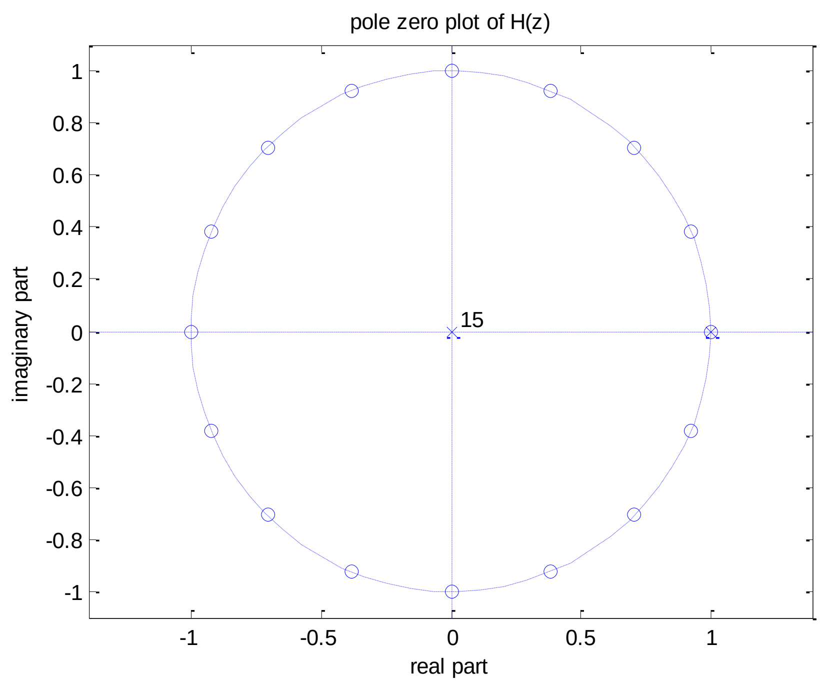


Fig.3.3. Pole zero plot of H(z) with R=16

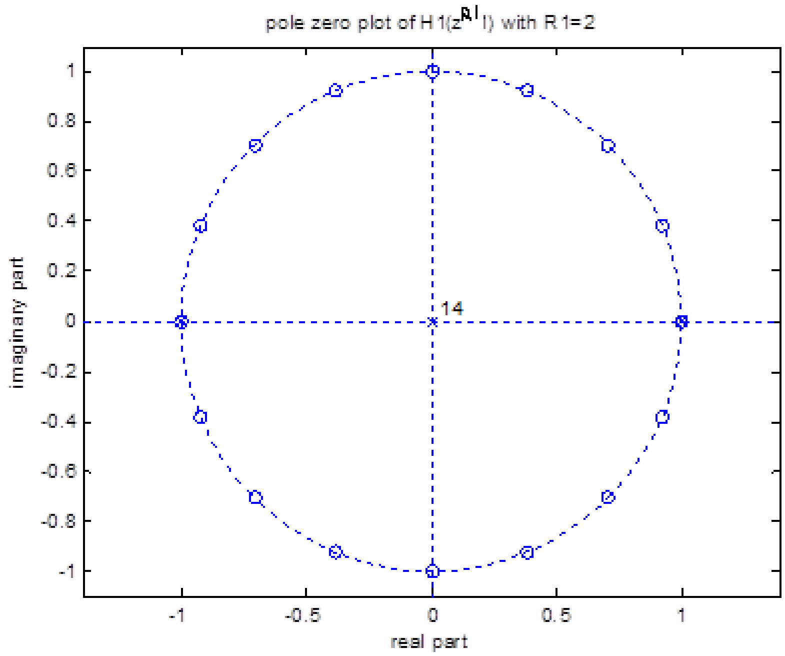


Fig.3.4. Pole zero plot of $H_1(z^{R1})$ with R1 =2 and R2 =8

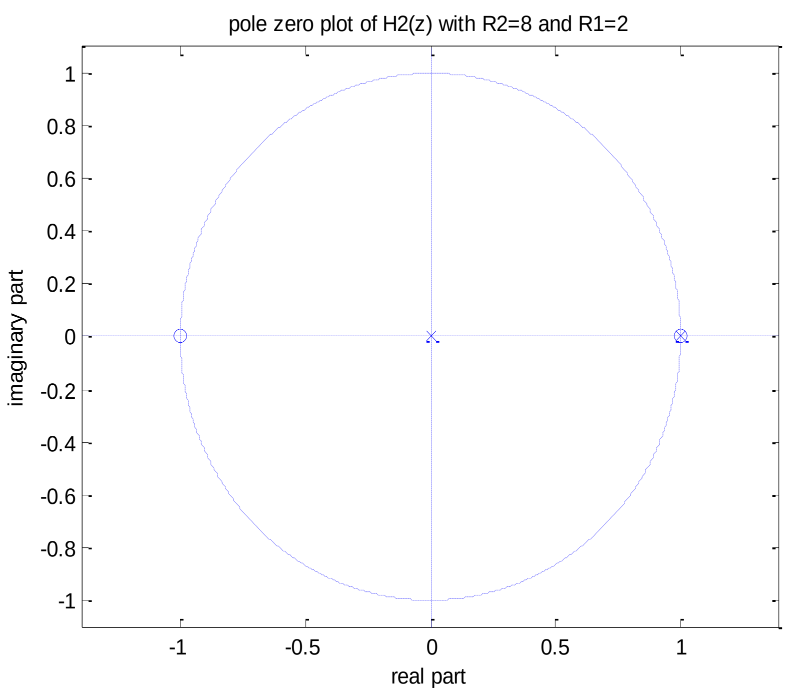


Fig.3.5. Pole zero pole of $H_2(z)$ with R1=2 and R2 = 8

As we can see from these pole zero plots that they show similarity in the position of zeros of $H1(z^{R1})$ and H(z) in the low pass region, which decrease with the increase in value of R1. This suggests that we can apply sharpening to one section and use other one for stopband section.

Therefore, we can select the different number of stages for constructing the filter sections, and obtain the modified transfer function $H_m(z)$given by

$$H_m(\omega)=[H_1(z^{R1})]^k[H_2(z)]^L \quad (3.10)$$

Applying the sharpening to $H_1(z^{R1})$, which is explained next we get the transfer function of the modified sharpened filter $H_{sh,m}(z)$ as given below

$$H_{sh,m}(\omega)=[H_1(z^{R1})]^k[H_2(z)]^L\left[3z^{\frac{-R(R2-1)K}{2}} - 2[H_1(z^{R1})]^{\wedge}K\right] \quad (3.11)$$

For the sharpening function to follow the requirement is $L \geq 2K$

Introducing expressions (3.10) – (3.11), we obtain the magnitude response for the modified sharpened filter

$$H_{sh,m}(\omega) = \left[\left\{3\left(\frac{1}{R2}\frac{\sin(\frac{\omega R}{2})}{\sin(\frac{\omega R1}{2})}\right)^{2K} - 2\left(\frac{1}{R2}\frac{\sin(\frac{\omega R}{2})}{\sin(\frac{\omega R1}{2})}\right)^{3K}\right\}\left\{\left(\frac{1}{R1}\frac{\sin(\omega R1/2)}{\sin(\omega/2)}\right)^{L}\right\}\right] \quad (3.12)$$

Benefits of the modified sharpening technique proposed are found in the possibilities to exploit properly the two-stage decimation structure, and to reduce the sampling rate in the first decimation stage.

Also by applying polyphase implementation the power dissipation can be further reduced.

The magnitude response plots of original sharpened with two stage are shown below:

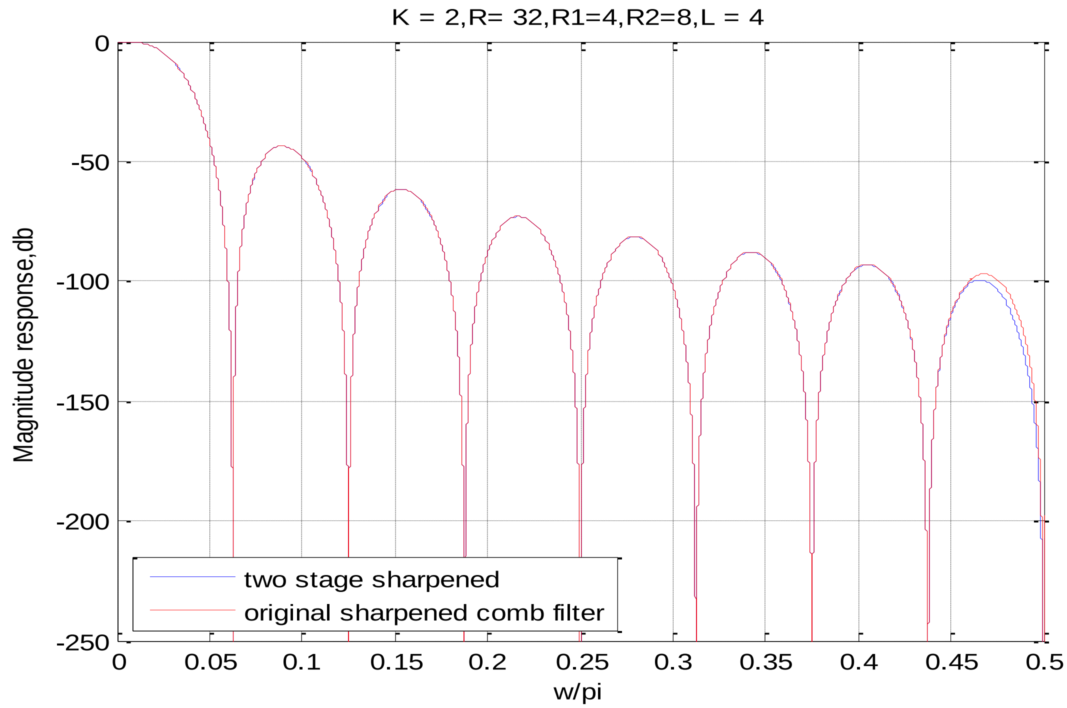


Fig.3.6. Magnitude response plot with R1=4, R2=8and L=4

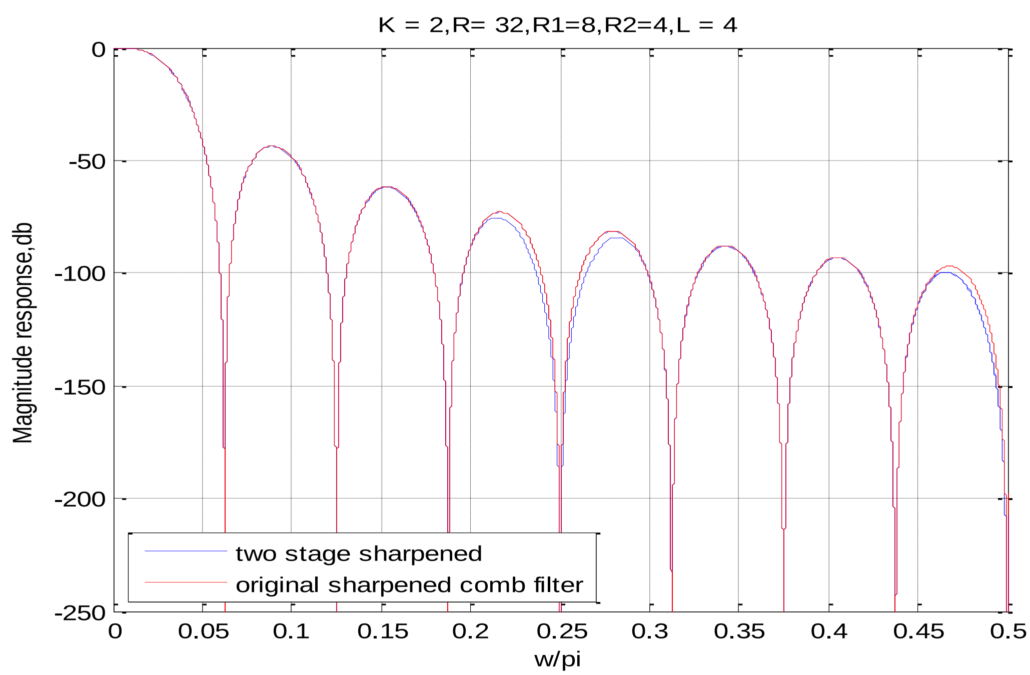


Fig.3.7. Magnitude response plot with R1=8, R2=4and L=4.

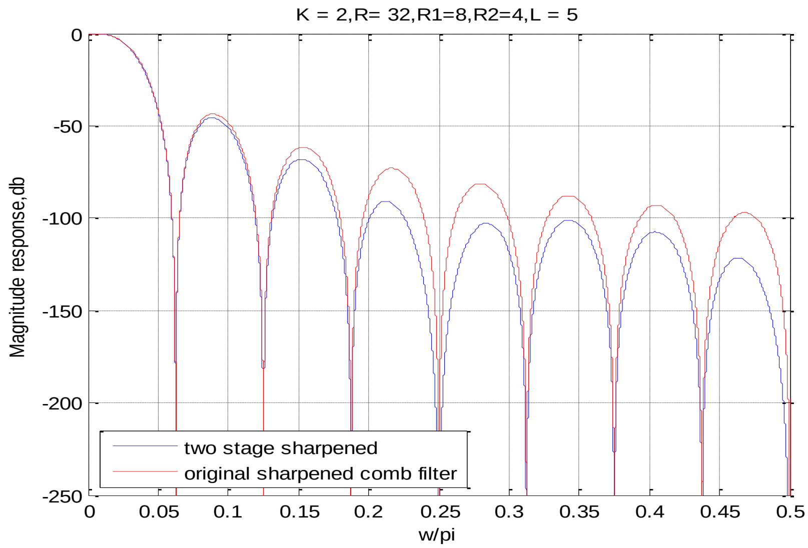


Fig.3.8. Magnitude response plot with R1=8, R2=4and L=5.

As it is clear from the above Fig. that the magnitude response of two stage structure is very similar to original sharpened structure. To study the worst case passband droop and worst case aliasing we need to study the response in low frequency region. The frequency f_c where the worst case passband distortion occurs and f_A where worst case aliasing occurs. In a filter with second stage factor to be 8 the frequency f_c is 1/8R and f_a is 15/16R. Where R is the decimation factor.

The passband droop is dependent on various factors like R1, number of stages K and L. Its effect can be analyzed from the Fig. given below

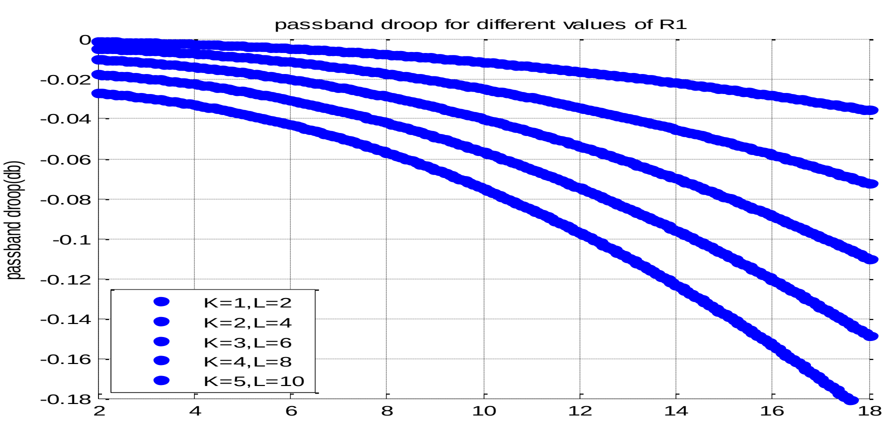


Fig.3.9. Variations in pass band droop with K and L values

3.2.1 STRUCTURE OF TWO STAGE

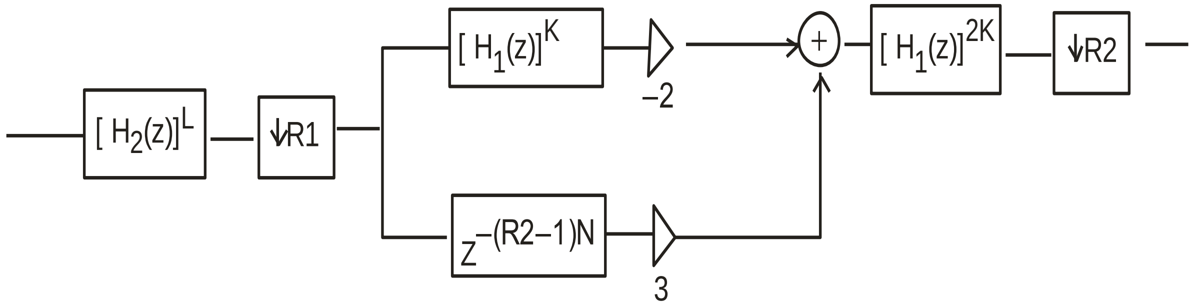


Fig.3.10. Basic structure of two stage implementation

The two stage structure and the original sharpened structure are shown above. The two stage structure is obtained using cascade equivalence. The group delay of $(H_1(z)^k)$ is

$$\text{Grpdelay}\ \{(H_1(z)^k)\} = \frac{R_2 - 1}{2} K \tag{3.13}$$

Therefore K must be K= 2N where N is an integer.

The first stage can be realized either in the recursive or non-recursive form and we can further use polyphase implementation to implement first stage which further reduces the computational complexity. The cascade of L comb filters of length R1 has transfer function given by

$$[H_2(z)]^L = (\frac{1}{R1})^L \sum_{n=0}^{L(R1-1)} h(n) z^{-n} = (\frac{1}{R1})^L H_2'(z) \tag{3.14}$$

Where

$$\sum_{n=0}^{L(R1-1)} h(n) z^{-n} = (H_2'(z)) \tag{3.15}$$

The filter coefficients are integer and symmetric h(n)= h(N-1-n) where N= $L(R_1$-1). Now if we apply poly phase decomposition to filter we get

$$H_2'(z) = E_0(z^{R1}) + z^{-1} E_1(z^{R1}) + \ldots\ldots\ldots\ldots\ldots\ldots\ldots + z^{-(R_1-1)} E_{R1-1}(z^{M1}) \tag{3.16}$$

Where $E_j(z^R)$, j=1,2,………R_1-1. denote the poly phase components. The down sampler can be moved before filter by using cascade equivalence. And as a result the filters in the first stage are now operated at lower rate which is R_1 times lower than input one.

The second stage is a normal sharpened filter with transfer function

$$H_1(z) = \frac{1 - z^{-R2}}{R2(1 - z^{-1})} \tag{3.17}$$

So now the sharpened section is operating at rate lower than than the input rate by R_1. So its length is R_1 times shorter than the conventional sharpened filter

3.3 MODIFIED SINC FILTER FOR STOPBAND IMPROVEMENT

A new filter aimed to improve the stop band attenuation in the aliasing bands in order to provide maximum suppression of the quantization noise in the first decimation stage. To achieve this, the concept of rotation of the natural nulls in the z-plane is applied to the comb filter sections. As a result, the new nulls placed in the intervals of the comb filter natural nulls are produced, and each pair of new nulls is located symmetrically around the natural comb filter nulls. In this way, a better distribution of the overall comb filter nulls may be achieved since they are not located one over the other anymore. This approach provides the possibility to modify regularly the stop band attenuation in a more desirable form.

The basic sinc filter is given as from equations (1.11) and (1.12)

$$H_c(z) = \frac{1-z^{-R}}{R(1-z^{-1})}$$

$$H_c(e^{j\omega}) = \frac{1}{R}\frac{\sin(\frac{\omega R}{2})}{\sin(\frac{\omega}{2})} e^{-j\omega[\frac{R-1}{2}]}$$

If we apply clockwise rotation by an angle α to any zero

$$z_i = e^{j\frac{2\pi}{R}i} \tag{3.18}$$

The transfer function becomes

$$H_u(z) = \frac{1-z^{-R}e^{j\alpha R}}{R(1-z^{-1}e^{j\alpha})} \tag{3.19}$$

By application of opposite rotation that is in counter clockwise direction the transfer function becomes

$$H_d(z) = \frac{1-z^{-R}e^{-j\alpha R}}{R(1-z^{-1}e^{-j\alpha})} \tag{3.20}$$

The two filters have complex coefficients but the resultant $H_r(z)$ which is obtained by the cascade of $H_u(z)H_d(z)$ which has real coefficients since the zeros of $H_r(z)$ is complex conjugate pair. Its transfer function is given as

$$H_r(z) = H_u(z)H_d(z) = \frac{1}{R^2}\frac{1-2\cos(\alpha R)z^{-R}+z^{-2R}}{1-2\cos(\alpha)z^{-1}+z^{-2}} \quad (3.21)$$

Above equation has linear phase and its frequency response is given by

$$H_r(f) = \frac{e^{-j\pi f(R-1)}}{R^2}\frac{\sin[(\pi f+\alpha/2)R]}{\sin[(\pi f+\alpha/2)]}\frac{\sin[(\pi f-\alpha/2)R]}{\sin[(\pi f-\alpha)]} \quad (3.22)$$

It has zeros at $\frac{i}{R}+\frac{\alpha}{2\pi}$ where i is integer value which is related to decimation factor R. the modified sinc is obtained by multiplying the basic comb with $H_r(z)$, then we obtain the transfer function of modified sink as

$$H_{ms}(z) = \frac{1}{R^3}\frac{1-2\cos(\alpha R)z^{-R}+z^{-2R}}{1-2\cos(\alpha)z^{-1}+z^{-2}}\frac{1-z^{-R}}{(1-z^{-1})} \quad (3.23)$$

The frequency response is the product of $H_r(f)$ and $H_c(f)$, therefore the product of three sinc functions.

$$H_{ms}(f) = \frac{e^{-j3\pi f(R-1)}}{R^2}\frac{\sin[(\pi f+\alpha/2)R]}{\sin[(\pi f+\alpha/2)]}\frac{\sin[(\pi f-\alpha/2)R]}{\sin[(\pi f-\alpha)]}\frac{\sin(\frac{\pi f R}{2})}{\sin(\frac{\pi f}{2})} \quad (3.24)$$

The comparison of the distribution of the z-plane zeros of the modified sinc $H_{ms}(f)$ with the zero distribution of the three stage-comb filter is shown by the following graph:

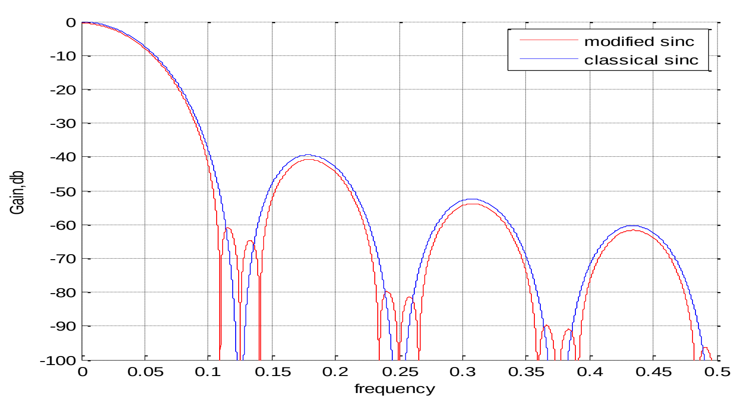


Fig.3.11. Magnitude response of modified sinc with with R=16 and alpha =0.03

The gain response of the two filters is shown in Fig. where the passband and the first aliasing band are emphasized. The various combinations of modified comb filters with classical comb filters can be used to construct the high-order filter structures. The zero-rotation approach considerably improves the ability of the modified comb filter to suppress aliasing in decimation, but the passband droop is slightly increased. Naturally, the passband characteristics should be compensated. So sharpening is applied to improve this. The characteristic for different values of alpha is shown below:

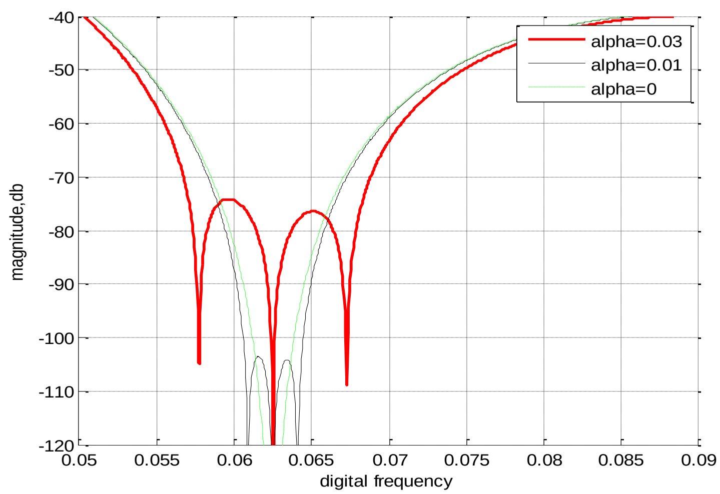


Fig.3.12. Variation in magnitude response with different alpha values

The rotation angle α has to be chosen in such a way as to place the zeros of the transfer function $H_{ms}(f)$ in the null intervals of the classical comb filter for the given value of N. Generally, the convenient value of α has to be selected in accordance with the signal bandwidth and with the bandwidth of the aliasing bands, i.e. the rotations of zeros for ±α have to place nulls of Go(f) within the aliasing bands. The value of α is expressible as

$$\text{alpha} = 2\, q\, \pi\, f_c \tag{3.25}$$

where q is a parameter which is less than 1 and f_c is the highest frequency of the input signal.

3.4 SHARPENED MODIFIED COMB FILTER

The novel decimation filtering technique tailor to ∑Δ A/D converters is given by Massimiliano Laddomada. The design is focused on third order modified sinc filter which leads to noise suppression of near order of 8db as compared to third order conventional comb filter when used with second order ∑Δ converters.

3.4.1 MODIFIED COMB DECIMATION FILTER (MCF)

Modified sinc filters are proposed by L.L Presti which are capable to increase attenuation of ∑Δ quantization noise(QN) around the folding band. The idea is to distribute the zeros of Rth order comb filter so as to achieve best noise suppression around folding band.

The second order modified sinc filter is given as(3.21)

$$H_r(f) = \frac{1}{R^2}\frac{1-2\cos(\alpha R)z^{-R}+z^{-2R}}{1-2\cos(\alpha)z^{-1}+z^{-2}} \tag{3.26}$$

The frequency response is given as

$$H_r(f) = \frac{e^{-j\pi f(R-1)}}{R^2}\frac{\sin[(\pi f+\alpha/2)R]}{\sin[(\pi f+\alpha/2)]}\frac{\sin[(\pi f-\alpha/2)R]}{\sin[(\pi f-\alpha)]} \tag{3.27}$$

It has zeros at $\frac{i}{R}+\frac{\alpha}{2\pi}$ where i is integer value which is related to decimation factor R. α is a parameter equals to q2πf_c and it is chosen in way to cover folding band ($\frac{i}{R}-$ fc to $\frac{i}{R}+f_c$) where f_c is highest frequency of interest or normalized maximum frequency in input signal. The above given filter is a linear phase filter and contain real coefficients .

The third order MCF is obtained by multiplication of first order comb filter with second order filter cell.

$$H_{MCF3}(z) = \frac{1}{R^3}\frac{1-2\cos(\alpha R)z^{-R}+z^{-2R}}{1-2\cos(\alpha)z^{-1}+z^{-2}}\ \frac{1-z^{-R}}{(1-z^{-1})} \tag{3.28}$$

Its frequency response is

$$H_{MCF3}(f) = \frac{e^{-j3\pi f(R-1)}}{R^2}\frac{\sin[(\pi f+\alpha/2)R]}{\sin[(\pi f+\alpha/2)]}\frac{\sin[(\pi f-\alpha/2)R]}{\sin[(\pi f-\alpha)]}\frac{\sin(\frac{\pi f R}{2})}{\sin(\frac{\pi f}{2})} \tag{3.29}$$

It has zeros at frequencies $\frac{i}{R}$ and $\frac{i}{R}\pm\frac{\alpha}{2\pi}$. For α=0, $H_{MCF3}(f)$ becomes classical third order filter.

The parameter α is fundamental to achieve a good noise suppression around folding band ,the optimal procedure to select the parameter α is as given. α is set to be $q2\pi f_c$ in order to spread rotated zeros all across the folding band . now zeros are located at $\frac{i}{R} \pm qf_c$.

q determines the position of zeros of $H_r(f)$ and hence the performance of filter for noise rejection in bands $\frac{i}{R} - fc$ to $\frac{i}{R} + f_c$. the value of q lies between 0 and 1(0<q<1) which makes the zeros to fall inside the folding band and itd value is independent of f and decimation factor R. this simplifies the design for A/D converters , best noise suppression around folding band is achieved for q=0.78. This extra attenuation of QN due to MCF is given as

$$G = \frac{\sum_i \int_{\frac{i}{R}-fc}^{\frac{i}{R}+fc} |H_{MCF3}(f)|^2 S_B(f)\, df}{\sum_i \int_{\frac{i}{R}-fc}^{\frac{i}{R}+fc} |H_{C3}(f)|^2 S_B(f)\, df} \tag{3.30}$$

Where $S_B(f)$ is the power spectral density of ∑Δ QN and is expressed as

$$S_B(f) = S_C(f)[2\sin(\pi f)]^{2B} \tag{3.31}$$

B is the order of ∑Δ modulation and $S_C(f) = \frac{\Delta^2}{12f_s}$ where f_s is the ∑Δ sampling rate.

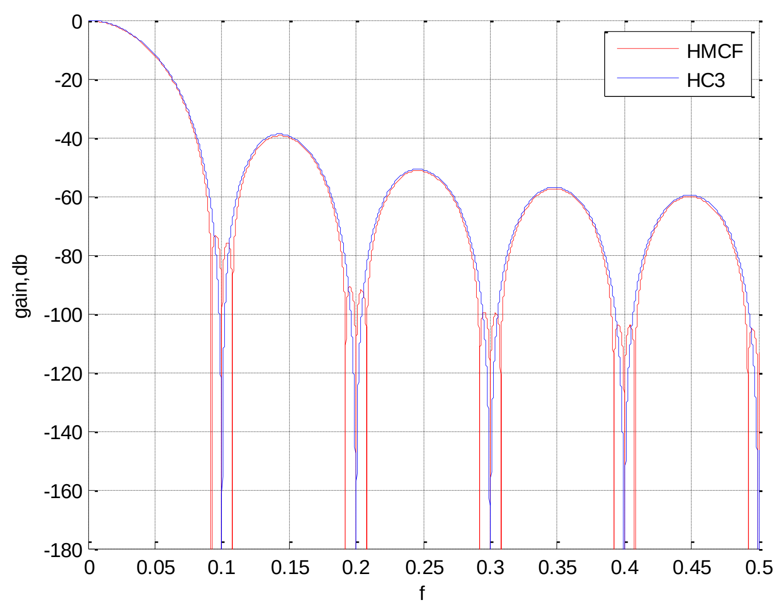


Fig.3.13.The response for MCF3 and HC3 for fc=0.01 ,R=10 and q=0.78

Folding bands are centered at $\frac{i}{R}$ and cover the areas $\frac{i}{R} - fc$ to $\frac{i}{R} +$ fc. This spanning of all folding band leads to extra noise suppression of 8db compared to conventional third order filter.

3.4.2 HIGHER ORDER MCFs

The more order MCFs are obtained by further cascade of basic filter cell with rotated zeros cell. Fourth order MCF is obtained by using 2nd order CIC filter $H_{C(2)}(z)$ with zeros at $\frac{i}{R}$ and cascading it with rotated zeros cell $H_r(f)$ to get $H_{MCF4}(f)$

$$H_{MCF4}(f) = H_{C(2)}(f).\, H_r(f) \tag{3.32a}$$

$$H_{MCF5}(f) = H_{C(2)}(f).\, H_{q1}(f).\, H_{q2}(f) \tag{3.32b}$$

$$H_{MCF6}(f) = H_{C(2)}(f).\, H_{q1}(f).\, H_{q2}(f) \tag{3.32c}$$

Where selection of q1 and q2 is optimally done based on selection procedure and are calculated considering fourth and fifth order ∑Δ modulator respectively.

TABLE 1 : OPTIMAL POSITION OF ZEROS

$\mathbf{H_{MCF5}(f)}$	**q1**	**q2**	**G[db]**
$H_{MCF3}(f)$	0.78	-	8
$H_{MCF4}(f)$	0.85	-	13
$H_{MCF5}(f)$	0.54	0.93	18
$H_{MCF6}(f)$	0.63	0.92	23

3.4.3 SHARPENED MODIFIED COMB

These are introduced with aim of reducing passband droop of filter with respect to classic comb filter with the simultaneous increase in rejection of ∑Δ QN around folding band. Sharpened modified comb filter refers to applying sharpening technique to MCFs proposed by Kaiser and Hamming (3.1)

$$H_{nm}(f) = H^{n+1}(f)\sum_{k=0}^{m} \frac{(n+k)!}{n!k!} [1 - H(f)]^k$$

Where H is a simple filter. The integers n and m are non-negative and measures the number of the non-zero derivatives of $H_{nm}(f)$ the points where $H_{nm}(f) = 0$ and $H_{nm}(f) = 1$, respectively.

The family of filters defined by above equation is applied to ∑Δ A/D converters for n and m=1 and 2 and then further modified to improve ∑Δ QN rejection around folding band.

$$H_{11}(f) = H^3(f)[3 - 2H(f)] \tag{3.33a}$$

$$H_{12}(f) = H^3(f)[6 - 8H(f) + 3H^2(f)] \tag{3.33b}$$

$$H_{21}(f) = H^3(f)[4 - 3H(f)] \tag{3.33c}$$

$$H_{22}(f) = H^3(f)[10 - 15H(f) + 6H^2(f)] \tag{3.33d}$$

- Taking causality into consideration, term $[3 - 2H(f)]$ should become $\left[3e^{-j\pi Kf(R-1)} - 2H(f)\right]$ and for having integer delay even number of taps are required. This applies to all the above equations.
- The terms in the square bracket are responsible for droop reduction since they are imposing m^{th} order tangency when $H_{nm}(f)$=1.
- Basic filter used is second order filter i.e. $H_{C(2)}(f)$.

SMCF mainly refers to substituting the second order filter cell which has zeros at frequencies $\frac{i}{R}$ with modified comb cell $H_r(f)$ to have zeros better distributed and then applying sharpening to it.

n=1 and m=1

$$H_{11}(f) = H_{C(2)}(f).\ H_{q1}(f).[3e^{j\emptyset(f)} - H_{C(2)}(f)] \tag{3.34a}$$

n=1 and m =2

$$H_{12}(f) = H_{C(2)}(f).\ H_{q1}(f).[6e^{2j\emptyset(f)} - 8e^{j\emptyset(f)}H_{C(2)}(f) + 3H^2_{C(2)}(f)\ ] \tag{3.34b}$$

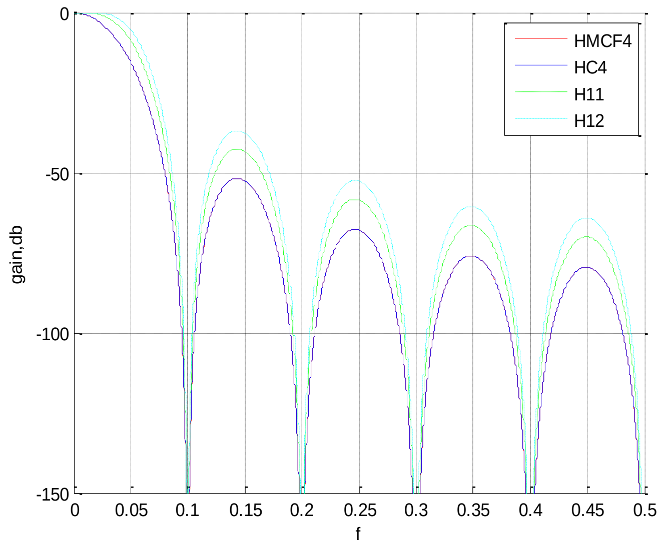
n =2 and m =1

$$H_{21}(f) = H_{C(2)}(f).\ H_{q1}(f)\ ).\ H_{q2}(f).\ [4e^{j\emptyset(f)} - 2H_{C(2)}(f)] \tag{3.34c}$$

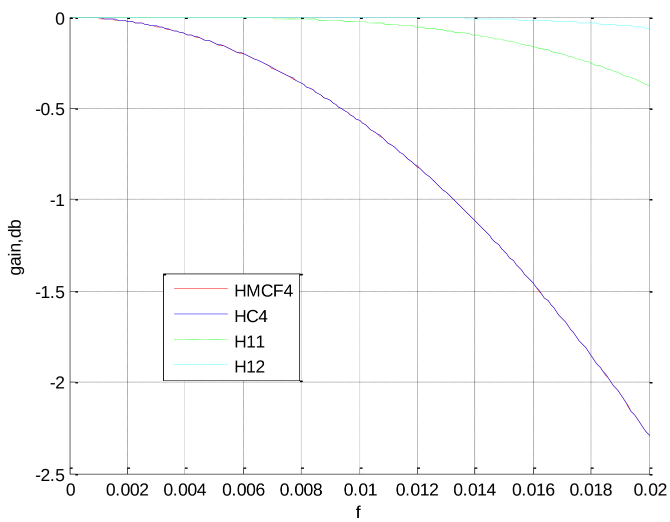
n =2 and m =2

$$H_{22}(f) = H_{C(2)}(f).\ H_{q1}(f)\ ).\ H_{q2}(f).[10e^{2j\phi(f)} - 15e^{j\phi(f)}H_{C\,2}(f) + 6H_{C\,2}^{2}(f)\ ]\ (3.34d)$$

The improvement results are shown in following figures for MCFs and SMCFs with R=10 and f=0.01

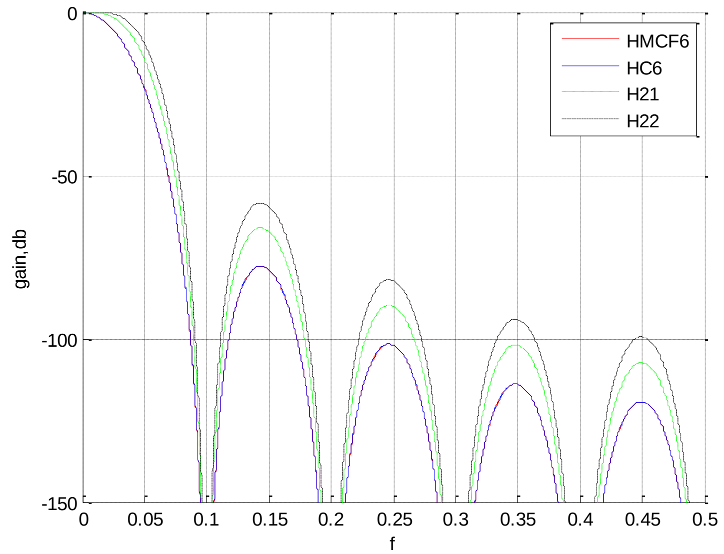


a. Overall magnitude response.

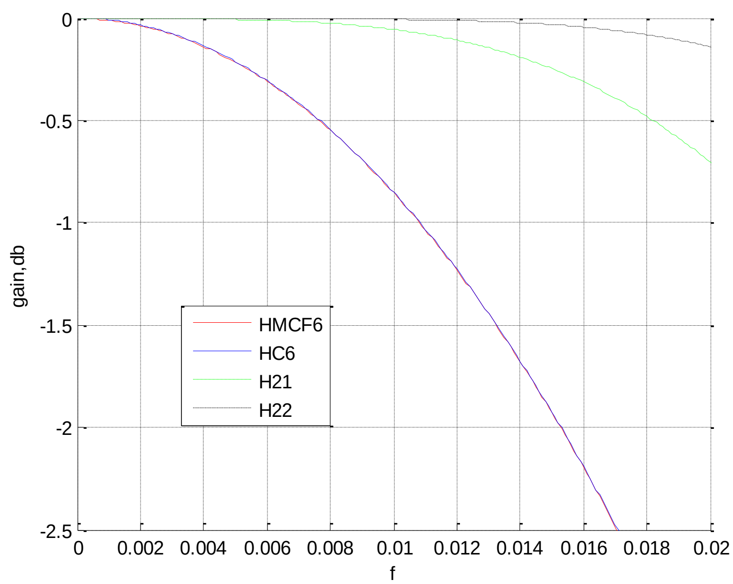


b. Passband zoom.

Fig.3.14. Frequency response of filter $H_{11}(f)$, $H_{12}(f)$,H_{MCF4},H_{C4} for f_c=0.01 and R=10.



a. Overall magnitude response.



b.Passband zoom

Fig.3.15. Frequency response of filter $H_{21}(f)$, $H_{22}(f)$,H_{MCF6},H_{C6} for f_c=0.01 and R=10.

- SMCFs have same order as repective SCFs.

- .Delays appearing in the brackets is to ensure proper group delay between different filter branches.
- Bracket terms are responsible for droop reduction.
- From Fig.3.14 and 3.15 it is clear that droops of SMCFs are smaller than in classic comb filter and MCFs.
- MCFs have droop almost similar to classic filter. At the edge of passband i.e. MCF and classic filter has distortion of 0.57dB and 0.85 dB in fig 4.1 and 4.2. While SMCF have less than 0.06 dB in both cases.

3.5 COMPENSATORS

Compensators are used in cascade with CIC filters for droop reduction in passband which appears in response of CIC filter

3.5.1 SINE BASED COMPENSATORS

Consider the filter with magnitude response

$$|G(e^{j\omega})| = |1+2^{-b}\sin^2(\omega R/2)| \tag{3.35}$$

Using relation

$$\sin^2(\alpha) = (1-\cos 2\alpha)/2$$

Corresponding transfer function is given as

$$G(z^R) = -2^{-(b+2)}[1-\left(2^{(b+2)}+2\right)z^{-R}+z^{2R}] \tag{3.36}$$

Or

$$G(z^R)=A[1+Bz^{-R}+z^{2R}] \tag{3.37}$$

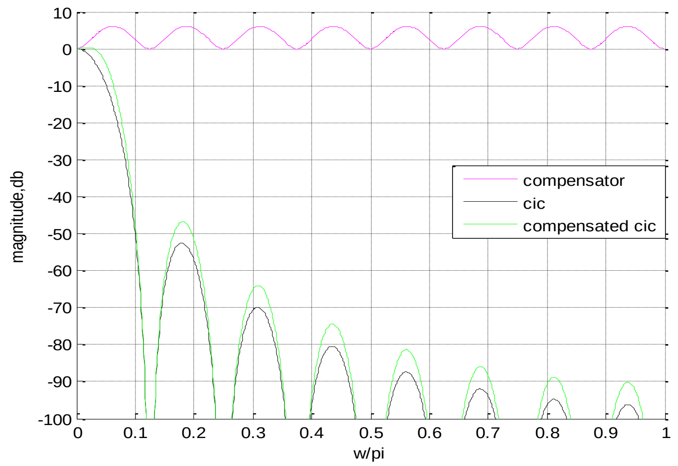
Where $A= -2^{-(b+2)}$ and $B= -\left(2^{(b+2)}+2\right)$

- It has a scaling factor A and one coefficient B and requires only one adder.
- With the use of multi rate identity compensator becomes of order two after down sampling by R
- This compensator is applied in cascade with CIC for droop reduction. Next given is the transfer function for compensated CIC.

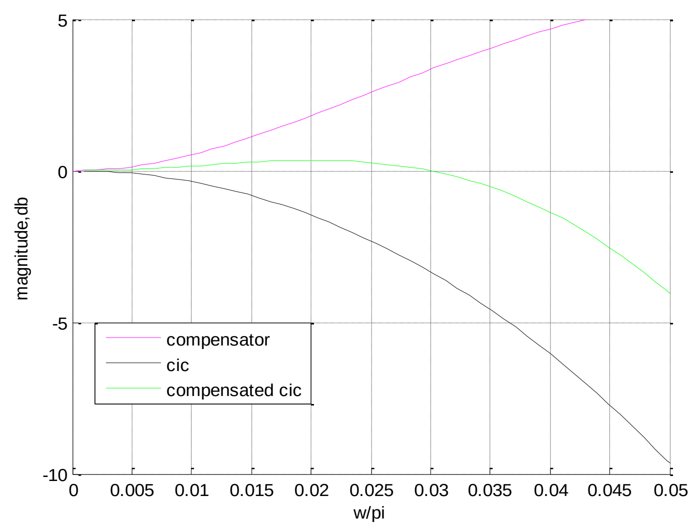
$$H_{CC}(f) = H_{C(K)}.\ G(z^R)$$

$$H_{CC}(f) = A.\left[\frac{1-z^{-R}}{R(1-z^{-1})}\right]^K.\ [1 + Bz^{-R} + z^{2R}] \tag{3.38}$$

Here $H_{CC}(f)$ refers to compensated CIC. Next Fig.3.16 shows the response of Compensator, CIC and that of compensated CIC.



a. Overall magnitude response.



b. Passband zoom.

Fig.3.16. Magnitude response of compensator, CIC and compensated CIC for K=4 and R=16.

- There are two design parameters K and b. K controls the stopband characteristics while parameter b compensated the corresponding passband droop of magnitude characteristics.
- The parameter b does not depend upon R
- For given value of K and parameter b, alias rejection is not affected by decimation factor
- b depends on K and a simple MATLAB program is used to calculate the values.

TABLE 2: VALUES OF PARAMETER b

K	B
2,3	2
4	1
5,6	0,-1

As b is a design parameter, the response of compensated CIC is affected with different b values and next is the response of compensated CIC with different b values.

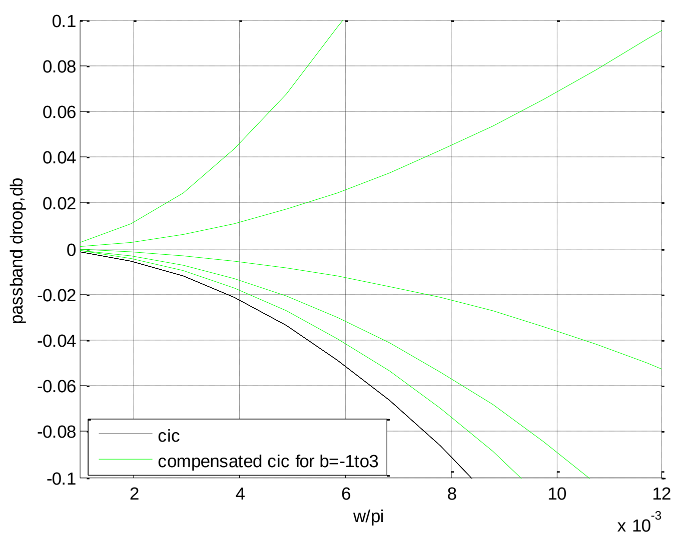


Fig.3.17. Passband view of compensated CIC for R=10 and K=4 for different values of b

- It is very clear from the above graph that b=1 provides best compensation.

3.5.2 MAXIMALLY FLAT COMPENSATION

Here discussion is on design of linear phase FIR compensation filter with magnitude as maximally flat

$$G(z)= H(z).Q(z^{R}) \tag{3.40}$$

Where H(z) from equation 1.13 is

$$H_{C(K)}(z)= \left[\frac{1-z^{-R}}{R(1-z^{-1})}\right]^{K}$$

And Q(z) is a filter with linear phase. Frequency response of H(z) is given as

$$H_{c}(e^{j\omega})=e^{-j\omega K[\frac{R-1}{2}]}.H(\omega)$$

$$\text{Where , } H(\omega) = \left[\frac{1}{R}\frac{\sin\left(\frac{\omega R}{2}\right)}{\sin\left(\frac{\omega}{2}\right)}\right]^{K}$$

For type 1 linear phase FIR compensation filter Q(z) as

$$Q(z) = \sum_{n=0}^{L} q_{n}\ z^{-n} \tag{3.41}$$

n=0,1………L are filter coefficients, L is an even integer and q satisfies $q_n= q_{L-n.}$

The corresponding frequency response is

$$Q(e^{j\omega}) = e^{-j\omega L/2}.Q(\omega) \tag{3.42}$$

Where Q(ω) is expressed as

$$Q(\omega)= q_{L/2} +2\sum_{n=0}^{\frac{L}{2}-1} q_{n} \cos(\omega(\tfrac{L}{2}-n)) \tag{3.43}$$

So frequency response of G(z) becomes

$$G(e^{j\omega}) = e^{-j\omega((R-1)N+LD)/2}.\ Q(\omega).\ H(\omega) \tag{3.44}$$

$G(e^{j\omega})$ has linear phase. Now we apply maximal flat condition on magnitude response.

Error function is defined as

$$E(\omega) = Q(\omega).\ H(\omega) -1 \tag{3.45}$$

E(ω) is maximal flat at ω=0 and has many derivatives as possible that are vanishing at ω =0[36]. The error function is even function so its odd indexed derivatives evaluated at ω =0 are also zero.

The conditions are

$$E(0) = 0 \quad (3.46a)$$

$$\frac{d^p E(\omega)}{d\omega^p}\Big|_{\omega=0} = 0 \quad (3.46b)$$

P is positive and even integer i.e. p=2r for r=1,2………L/2. Using(3.46) and (3.47a) implies

P(0)=1

Now we substitute (3.45) into (3.46b) and use of general Leibniz rule for p^{th} derivative of product, we get

$$\frac{d^p H(\omega)}{d\omega^p} + \sum_{s=1}^{P} \binom{p}{s} \left[\frac{d^s Q(\omega)}{d\omega^s} \; \frac{d^{p-s} H(\omega)}{d\omega^{p-s}}\right] \Big|_{\omega=0} = 0 \quad (3.47)$$

Where binomial coefficients are

$$\binom{p}{s} = \frac{p!}{s!(p-s)!} \quad (3.48)$$

The odd derivative of $Q(\omega)$ evaluated at $\omega = 0$ are zero .From (5.7) we get

$$\frac{d^s Q(\omega)}{d\omega^s} = \begin{cases} 2(-1)^s R^s \sum_{n=0}^{\frac{L}{2}-1} \left(\frac{L}{2} - n\right)^s q_n & , s \text{ even} \\ 0 & , s \text{ odd} \end{cases}, \quad (3.49)$$

Substituting (3.49) into (3.47) and (3.46b) we get

$$2\sum_{l=1}^{r} \binom{2r}{2s} 2(-1)^s R^{2s} \sum_{n=0}^{\frac{L}{2}-1} \left(\frac{L}{2} - n\right)^{2s} q_n \; \left[\frac{d^{2(r-s)} H(\omega)}{d\omega^{2(r-s)}}\right] |\omega = 0 \; = \frac{d^{2r} H(\omega)}{d\omega^{2r}} \quad (3.50)$$

For r= 1,2……………L/2

The coefficients for linear phase max flat compensation filter Q(z) as order L are obtained by solving equation (5.14)

3.5.3 SECOND ORDER MAX FLAT COMPENSATION

For L=2 ,transfer function of Q(z) is

$$Q(z) = q_0 + q_1 z^{-1} + q_0 z^{-2} \tag{3.51}$$

Applying max flat condition we get

$$q_1 + 2q_0 = 1 \tag{3.52a}$$

$$q_0 = \frac{1}{2R^2}\left[\frac{d^2H(\omega)}{d\omega^2}\right]\Big|_{\omega=0} \tag{3.52b}$$

Where

$$\frac{d^2H(\omega)}{d\omega^2} = \frac{-K(R^2-1)}{12}$$

Solving linear equations (5.16a) and(5.16b) we get

$$q_0 = \frac{-K(1-R^{-2})}{32(1-2^{-2})}$$

$$q_1 = 1\text{-}2\, q_0$$

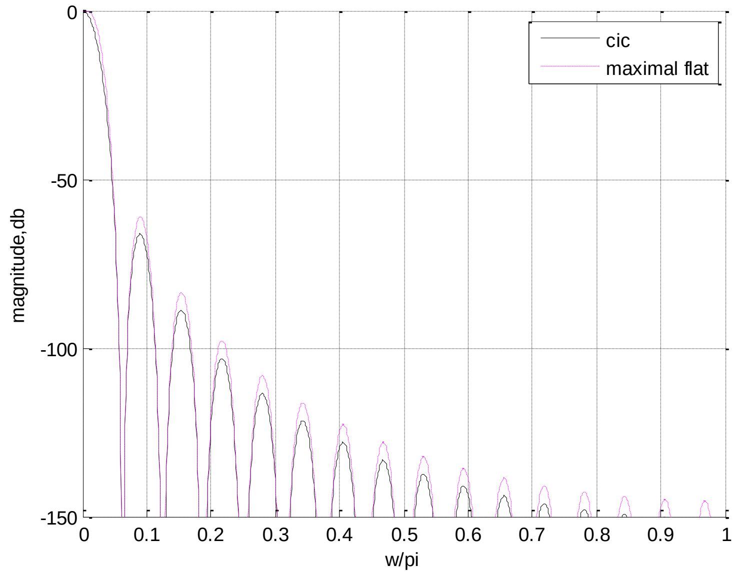
Defining

$$A = \frac{(1-R^{-2})}{(1-2^{-2})}$$

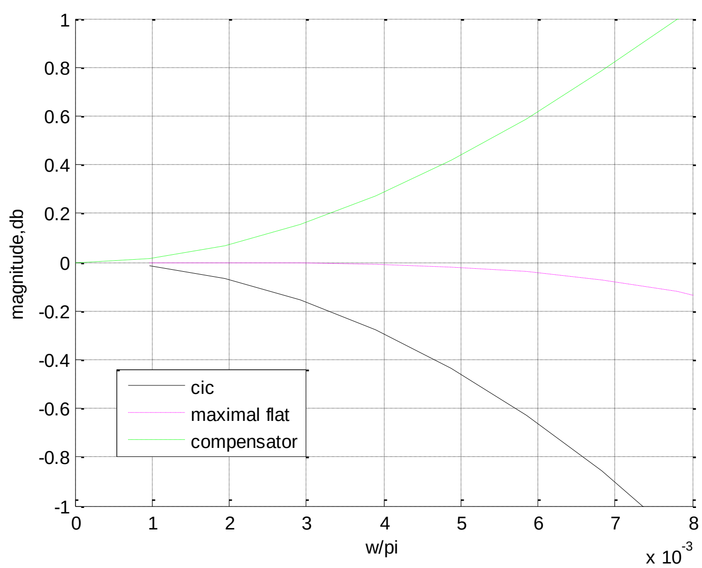
The filter coefficients are

$$q_0 = -2^{-5}.K.A\ .$$

This is a second order compensator filter and has only one multiplier. The magnitude response and its zoomed view are shown next in Fig.3.18 which makes the things about second order maximally flat compensators very clear.



a. Overall magnitude response.



b. Passband zoom.

Fig.3.18 Frequency response of CIC and maximally flat compensated CIC.

CHAPTER 4

COMPENSATED SHARPENED MODIFIED COMB FILTER

For the improvement in response of CIC filters, we are targeting the response improvement of SMCFs (3.4.3) already explained in chapter 3. As with SMCFs we are already getting wide stopband and for further improvement in passband we cascade it with the compensator i.e. those of sine based compensator and maximally flat compensator to achieve better passband. Hence the overall improvement in the response i.e. flat passband and wide stopband. From the equations of SMCFs (3.34) we have

$$H_{11}(f) = H_{C(2)}(f).\ H_{q1}(f).[3e^{j\emptyset(f)} - H_{C(2)}(f)]$$

$$H_{12}(f) = H_{C(2)}(f).\ H_{q1}(f).[6e^{2j\emptyset(f)} - 8e^{j\emptyset(f)}H_{C(2)}(f) + 3H^2_{C(2)}(f)\ ]$$

$$H_{21}(f) = H_{C(2)}(f).\ H_{q1}(f)\ ).\ H_{q2}(f).\ [4e^{j\emptyset(f)} - 2H_{C(2)}(f)]$$

$$H_{22}(f) = H_{C(2)}(f).\ H_{q1}(f)\ ).\ H_{q2}(f).[10e^{2j\emptyset(f)} - 15e^{j\emptyset(f)}H_{C(2)}(f) + 6H^2_{C(2)}(f)\ ]$$

Firstly we applying here sine based compensator in cascade with SMCFs then check the results. Equation for sine based compensator is given by (5.2). Using them in cascade we get

$$H_{CB}=H_{SMCF}.\ G(z^R)$$

For n=1 and m=1

$$H_{CS11}(f) = H_{C(2)}(f).\ H_{q1}(f).[3e^{j\emptyset(f)} - H_{C(2)}(f)].\ A[1 + Bz^{-R} + z^{-2R}]$$

$$H_{CS11}(f)=H_{C(2)}(f).H_{q1}(f).[3e^{j\emptyset(f)} - H_{C(2)}(f)].\ -2^{-(b+2)}[1 - \left(2^{(b+2)} + 2\right)z^{-R} + z^{-2R}] \quad (4.1a)$$

For n=1 and m =2

$$H_{CS12}(f) = H_{C(2)}(f).\ H_{q1}(f).[6e^{2j\emptyset(f)} - 8e^{j\emptyset(f)}H_{C(2)}(f) + 3H^2_{C(2)}(f)\ ].\ A[1 + Bz^{-R} + z^{-2R}]$$

$$H_{CS12}(f)= H_{C(2)}(f).\ H_{q1}(f).[6e^{2j\emptyset(f)} - 8e^{j\emptyset(f)}H_{C(2)}(f) + 3H^2_{C(2)}(f)\ ]$$

$$.-2^{-(b+2)}[1 - \left(2^{(b+2)} + 2\right)z^{-R} + z^{-2R} \quad (4.1b)$$

For n =2 and m =1

$$H_{CS21}(f) = H_{C(2)}(f).\ H_{q1}(f)\ ).\ H_{q2}(f).\ [4e^{j\emptyset(f)} - 2H_{C(2)}(f)].\ A[1 + Bz^{-R} + z^{-2R}]$$

$$H_{CS21}(f) = H_{C(2)}(f).\ H_{q1}(f)\ ).\ H_{q2}(f).\ [4e^{j\emptyset(f)} - 2H_{C(2)}(f)]$$

$$.\ -2^{-(b+2)}[1 - \left(2^{(b+2)} + 2\right)z^{-R} + z^{-2R}] \qquad (4.1c)$$

For n =2 and m =2

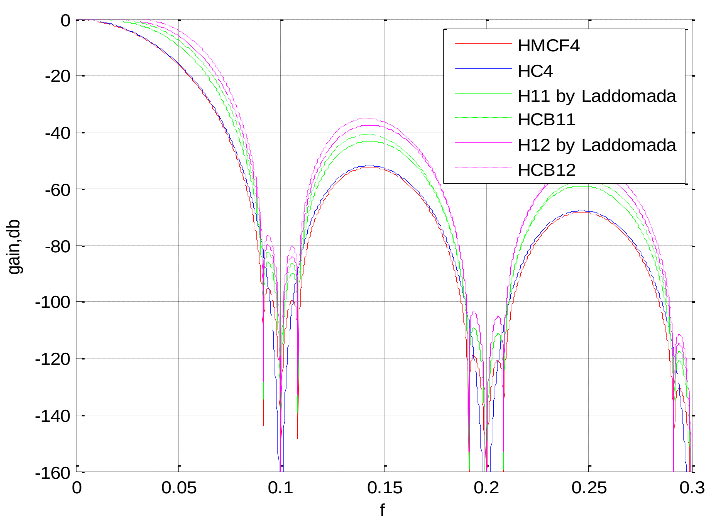
$$H_{CS22}(f) = H_{C(2)}(f).\ H_{q1}(f)\ ).\ H_{q2}(f).[10e^{2j\emptyset(f)} - 15e^{j\emptyset(f)}H_{C(2)}(f) + 6H^2_{C(2)}(f)\ ]$$

$$.\ A[1 + Bz^{-R} + z^{-2R}]$$

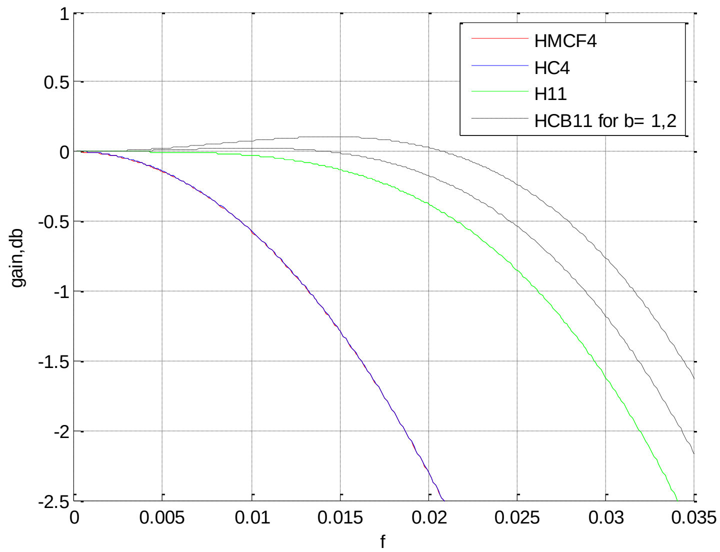
$$H_{CS22}(f) = H_{C(2)}(f).\ H_{q1}(f)\ ).\ H_{q2}(f).[10e^{2j\emptyset(f)} - 15e^{j\emptyset(f)}H_{C(2)}(f) + 6H^2_{C(2)}(f)\ ]$$

$$.\ -2^{-(b+2)}[1 - \left(2^{(b+2)} + 2\right)z^{-R} + z^{-2R} \qquad (4.1d)$$

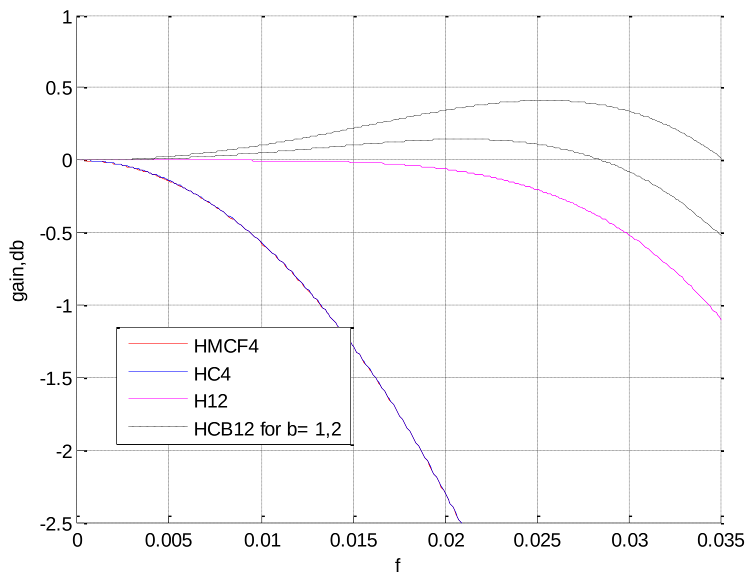
Following are the improvement results obtained after cascade . Figures shows the result of compensated SMCFs H_{CS11}, H_{CS12}, H_{CS21}, H_{CS22} and also their comparison with SMCFs H_{11}, H_{12}, H_{21}, H_{22}. Comparison shows the results improvement over already existing technique obtained by cascade. For sine based compensator results are shown with different values of b.



a. Overall Magnitude response.

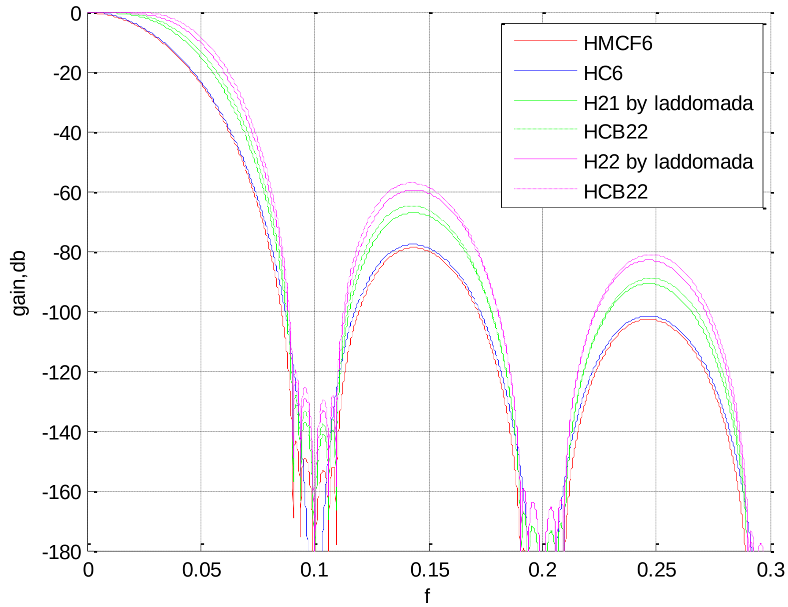


b. Passband zoom for different values of b(1 and 2) for H_{CB11} over H_{11} by Laddomada

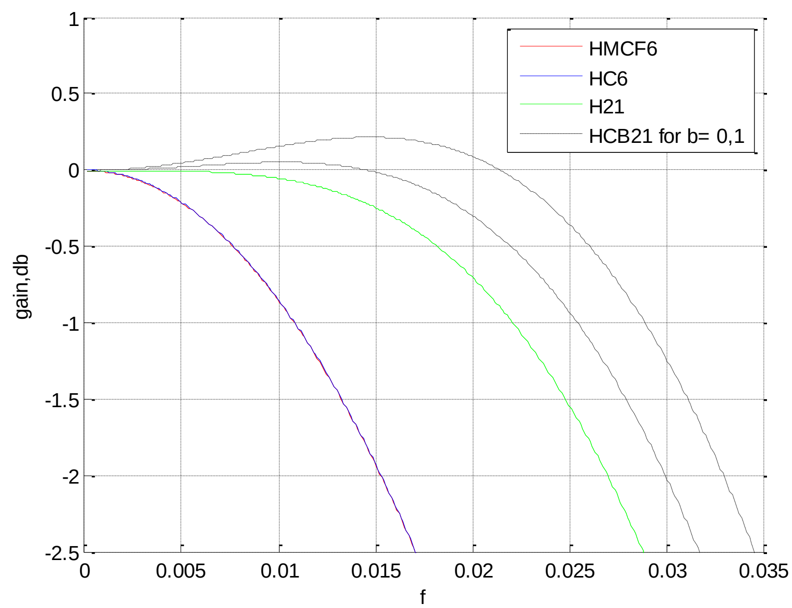


c. Passband zoom for different values of b(1,2) for H_{CB12} over H_{12} by Laddomada

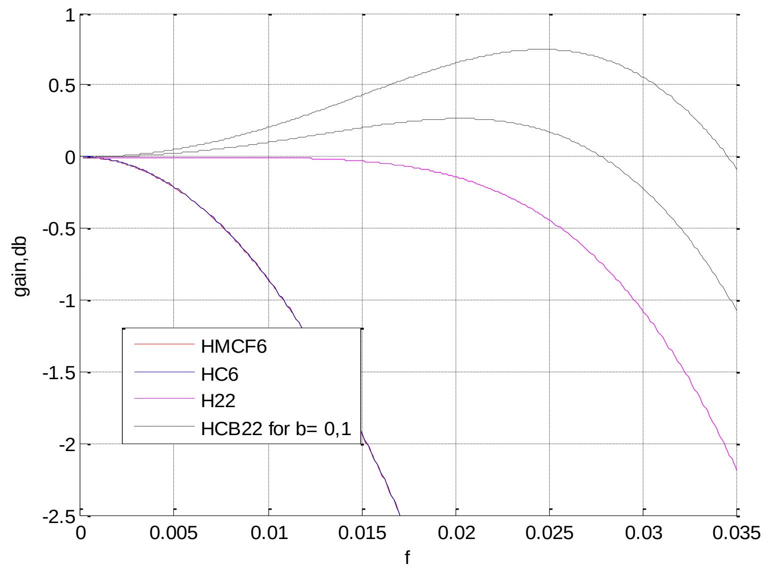
Fig.4.1. Frequency response of filter H_{11} and H_{12} by Laddomada and their cascade with basic compensator H_{CB11} and H_{CB12} for R =10 ,K=4, b=1,2 and q=0.85 and f_c=0.01.



a. Overall magnitude response.



b. Passband zoom for different values of b(0,1) for H_{CB21} over H_{21} by Laddomada.



c. Passband zoom in for different values of b(0,1) for H_{CB22} over H_{22} by Laddomada.

Fig.4.2 Frequency response of filter H_{21} and H_{22} by Laddomada and their cascade with basic compensator H_{CB11} and H_{CB12} for R =10 ,K=6, b=1 , q1=0.64, q2=0.94 and f_c=0.01.

Now we cascade the SMCFs with the maximally flat second order compensator given by equation (5.15) and get the following equations. The coefficients q_0 and q_1 are given in section5.3.

$H_{CM} = H_{SMCF}.Q(z^R)$

For n=1 and m=1

$$H_{CM11}(f) = H_{C(2)}(f).\ H_{q1}(f).[3e^{j\emptyset(f)} - H_{C(2)}(f)].\ q_0\ [1 + q_1 z^{-R} + z^{-2R}] \qquad (4.2a)$$

For n=1 and m =2

$$H_{CM12}(f) = H_{C(2)}(f).\ H_{q1}(f).[6e^{2j\emptyset(f)} - 8e^{j\emptyset(f)}H_{C(2)}(f) + 3H^2_{C(2)}(f)]$$

$$.\ q_0\ [1 + q_1 z^{-R} + z^{-2R}] \qquad (4.2b)$$

For n =2 and m =1

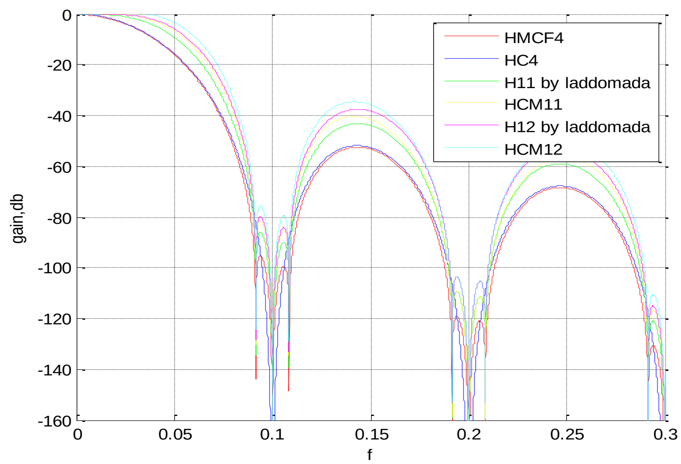
$$H_{CM21}(f) = H_{C(2)}(f).\ H_{q1}(f)).\ H_{q2}(f).\ [4e^{j\emptyset(f)} - 2H_{C(2)}(f)].\ q_0\ [1 + q_1 z^{-R} + z^{-2R}] \qquad (4.2c)$$

For n =2 and m =2

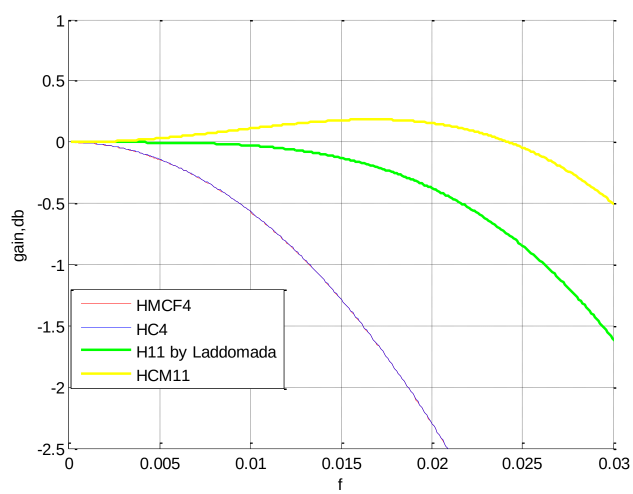
$$H_{CM22}(f) = H_{C(2)}(f).\ H_{q1}(f)\ ).\ H_{q2}(f).[10e^{2j\emptyset(f)} - 15e^{j\emptyset(f)}H_{C(2)}(f) + 6H^2_{C(2)}(f)\ ]$$

$$.\ q_0\ [1 + q_1 z^{-R} + z^{-2R}\ ] \quad (4.2d)$$

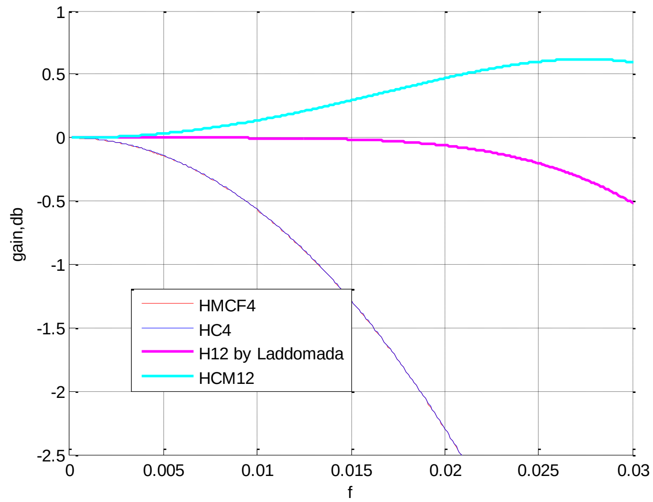
Following are the improvement results obtained after cascade. Figures shows the result of compensated SMCFs H_{CM11}, H_{CM12}, H_{CM21}, H_{CM22} and also their comparison with SMCFs H_{11}, H_{12}, H_{21}, H_{22}. Comparison shows the results improvement over already existing technique obtained by cascade.



a. Overall magnitude response.

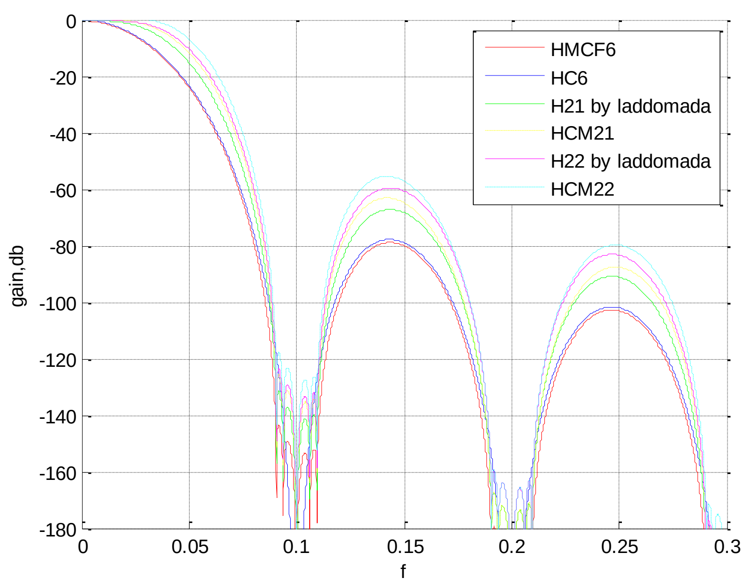


b. Passband zoom showing improvement in droop for H_{CM11} over SMCF H_{11}.

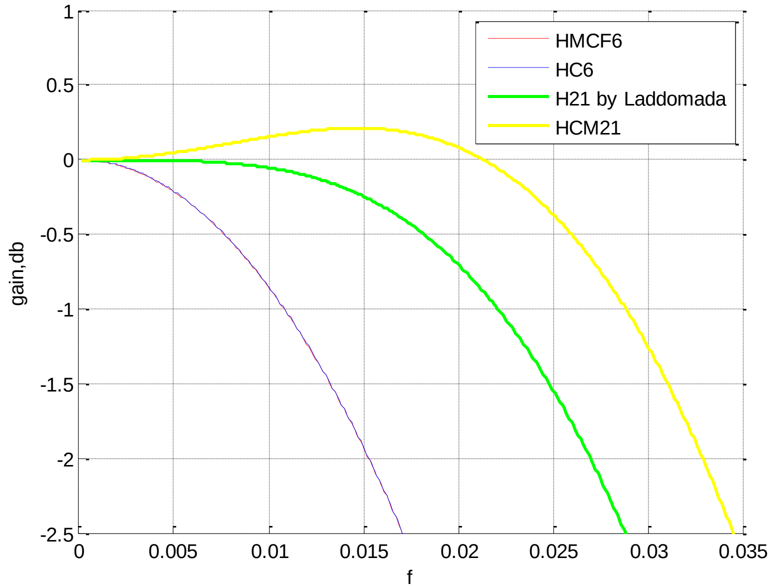


c. Passband zoom showing improvement in droop for H_{CM12} over SMCF H_{12}.

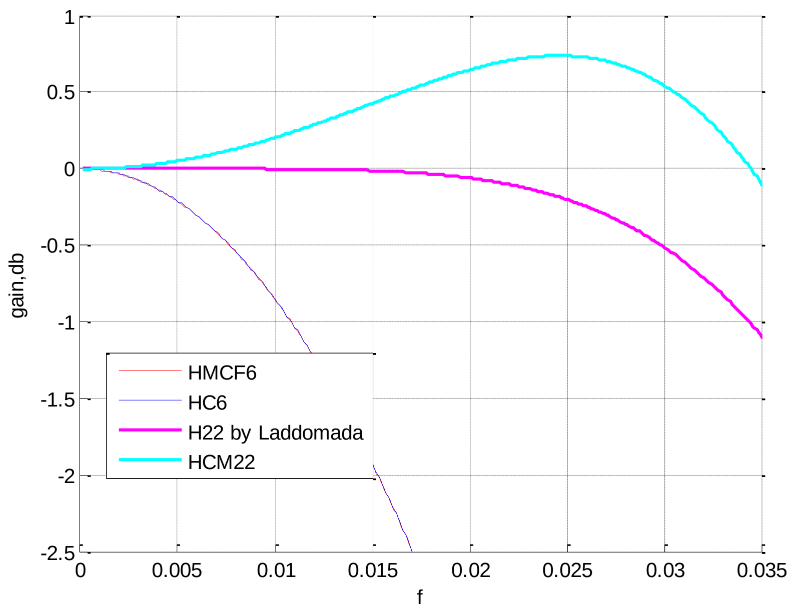
Fig.4.3. Frequency response of filter H_{11} and H_{12} by Laddomada and their cascade with basic compensator H_{CM11} and H_{CM12} for R =10 ,K=4, b=1 and q=0.85 and f_c=0.01.



a. Overall magnitude response



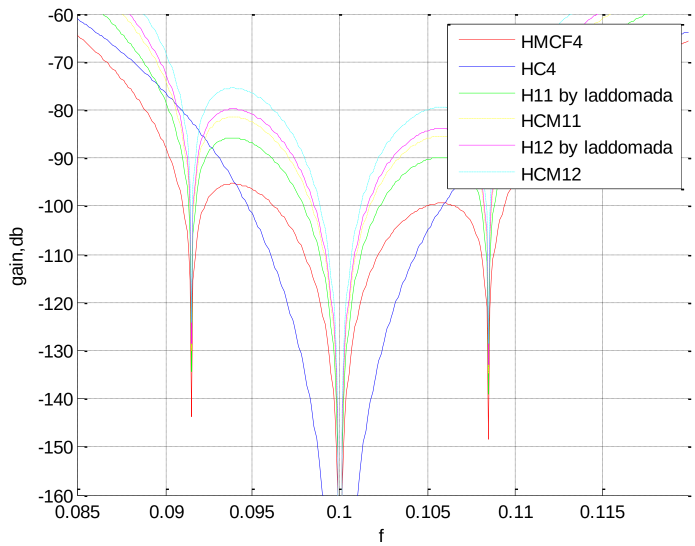
b. Passband zoom showing improvement for H_{CM21} over SMCF H_{21} by Laddomada.



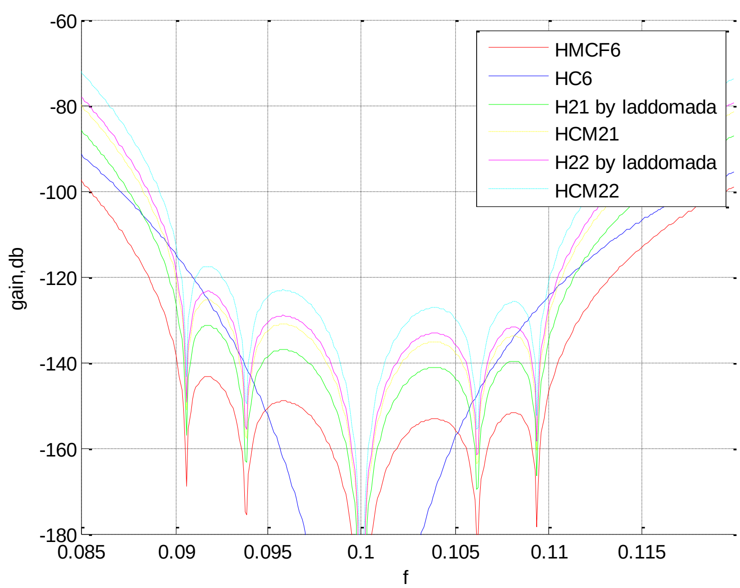
c. Passband zoom showing improvement for H_{CB22} over SMCF H_{22} by Laddomada.

Fig.4.4. Frequency response of filter H_{21} and H_{22} by Laddomada and their cascade with maximally flat compensator H_{CM21} and H_{CM22} for R =10, K=6, b=1, q1=0.64, q2=0.94 and f_c=0.01.

The above discussion was all about the pass band and shows the improvement in pass band . Now following is results for stopband attenuation for H_{11}, H_{12} and H_{21}, H_{22}



a. Stopband zoom for H_{CB11}, $H_{CM11}$and H_{CB12} and H_{CM12}.



a. Stopband zoom for H_{CB21}, H_{CM21} and H_{CB22} and H_{CM22}.

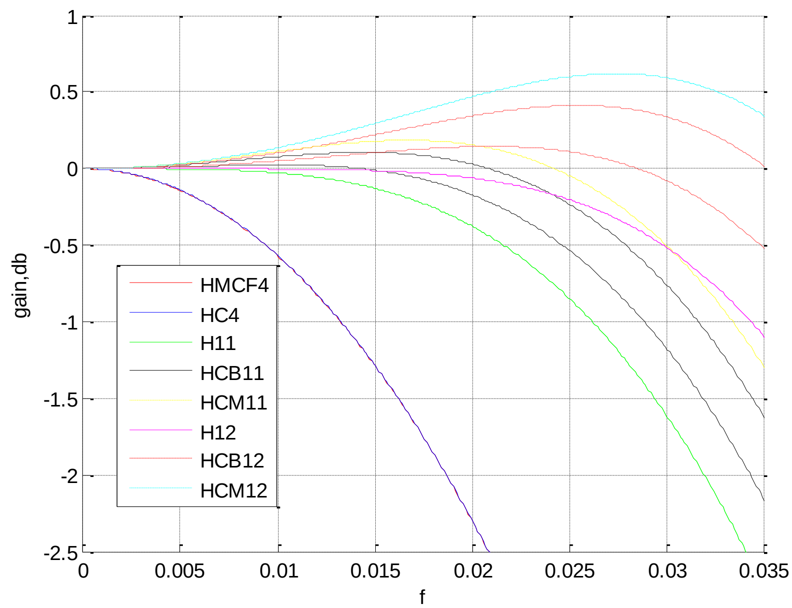
Fig.4.5. Stopband zoom showing high attenuation around nulls compared to basic Comb filter for both sine based compensator and maximal flat compensator.

From the graphs of stopband attenuation it is very clear that alias rejection remains same as that for MCF i.e. modified comb filter.so overall passband droop and reduced and stop band is already wide with the use of rotation of nulls. So the overall response has improved over the conventional CIC filter and for the case of ∑Δ A/D converters the droop has reduced and hence the gain improvement will occur.

CHAPTER 5

RESULTS AND DISCUSSION

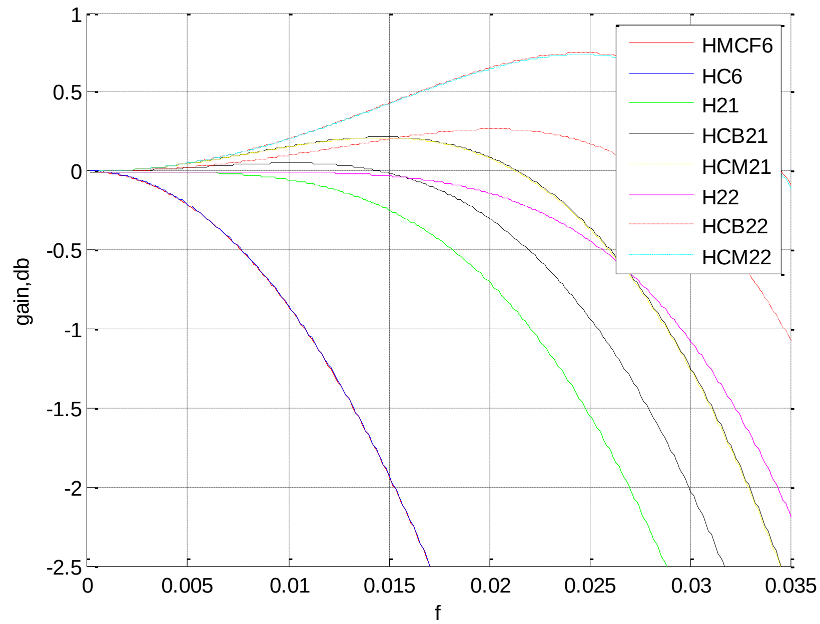
Here is the discussion in frequency response improvement which is obtained by cascade of SMCF with compensation techniques. Here we will discuss the two cases i.e. for f_c=0.01 and f_c=0.02. The discussion given in paper by Laddomada was with f_c=0.01. We will first discuss the case for f_c=0.01 and then the other one.



a. Passband zoom for b (1 and 2) for H_{CB11}, H_{CM11}, H_{CB12} and H_{CM12}.

From these figure we can clearly make inference about the droop reduction or we can droop compensation. Fig.5.1a and 5.1b shows the zoom in passband response for H_{CB11} and H_{CB21} respectively.

We can observe for H_{CB11} that at fc = 0.01 the distortion is zero for b=2 while SMCFs are giving that of 0.06 db. It is much greater for classic comb filter and MCF both overlapping as shown in fig nearly 0.57db and there is a gain of 0.07 dB for b=1 and H_{CM11}.



b. Passband zoom for b(0 and 1) for H_{CB21}, H_{CM21} , H_{CB22} and H_{CM22}.

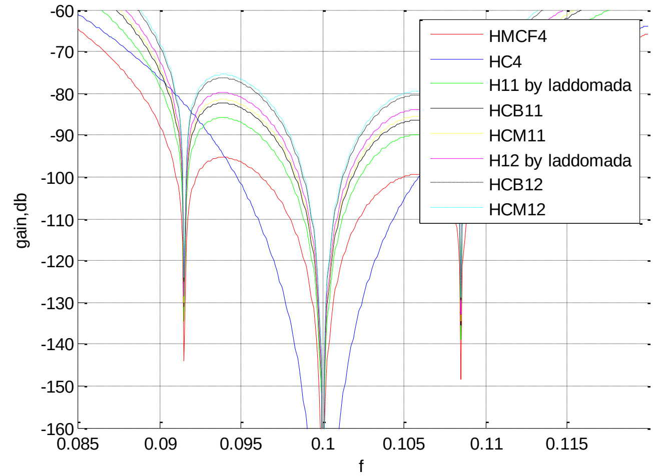
Fig.5.1. Passband zoom for different values of b for SMCFs and compensated SMCFs for R=10 and f_c=0.01

Similarly for H_{CB21} at fc=0.01 droop is zero (slight gain 0.05) for H_{CB21} for b= 1, while for SMCF it is 0.02 dB and there is a gain of 0.15db for b=0 and H_{CM21}.

Both the figures also show the zoom in pass band response for H_{CB12} and H_{CB22} respectively. H_{CB12} gives slight gain of 0.05dB for b=2 and 0.1 for b=1 while there is a gain 0.14 dB for H_{CM12}.

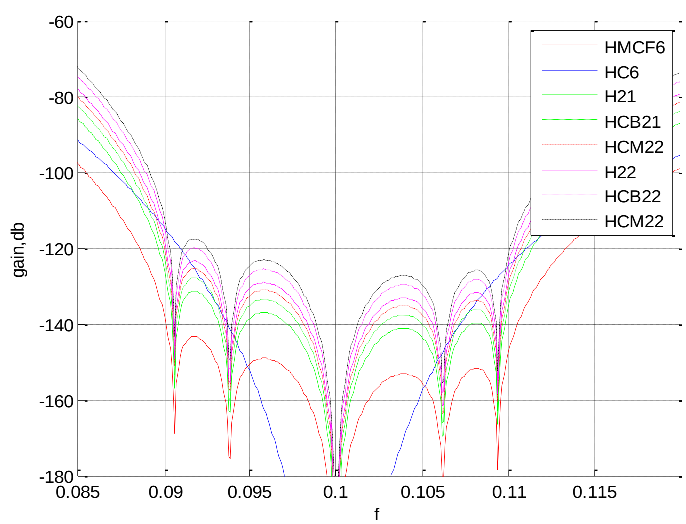
Similarly for H_{CB22} at f_c=0.01, droop is 0.01db for H_{22} and there is gain of 0.1db for b=1 and 0.2db for b=0 and H_{CM22}. So there is significant improvement in both the cases which is clearly visible

So it is clear that that the roll off response presented by CIC in passband is reduced with SMCF and now further with compensated SMCF and results are confirmed by the figures. The alias rejection in the stopband of CIC is also improved with L. Presti technique shown with the following figure.



a. Stopband zoom for H_{CB11},H_{CM11},H_{CB12} and H_{CM12}.

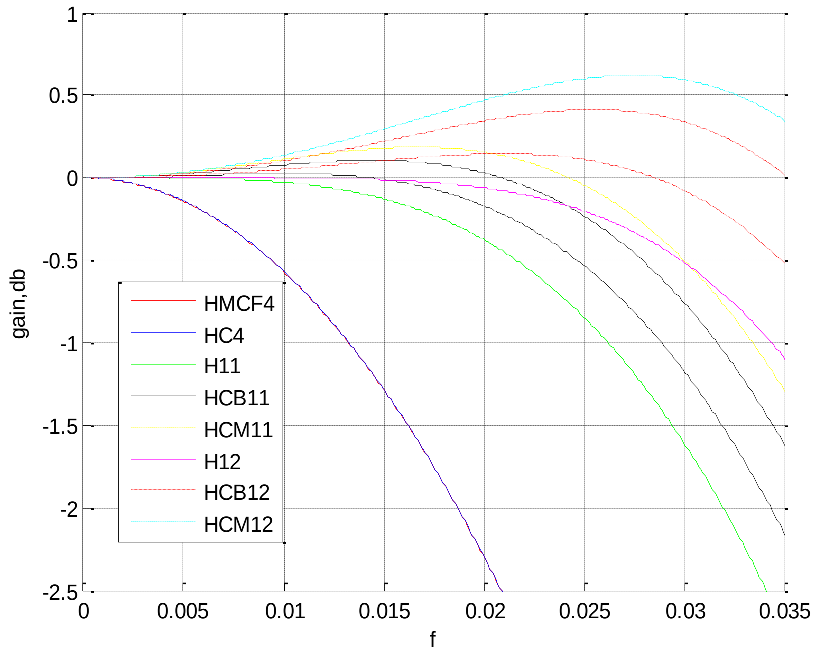
The stopband lies in the range $\frac{1}{R} - q f_c$ to $\frac{1}{R} + q f_c$. Here q=0.85 for H_{11} and H_{12} and hence stopband range is (o.1± 0.85*f_c). It is very clear from figure that stopband lies in this range only. Similarly the results are shown for H_{21} and H_{22} in the below figure. Stopband rejection of SMCF and compensated SMCF is same. Hence there is reduction in droop and stopband remains same so the overall response improvement is achieved through this compensated technique.



c. Stopband zoom for H_{CB21},H_{CB22}, H_{CM21} and H_{CM22} for K=6, q1=0.64 & q2 =0.94.

Fig.5.2. Stopband zoom for SMCF and compensated SMCFs for R=10 and f_c=0.01.

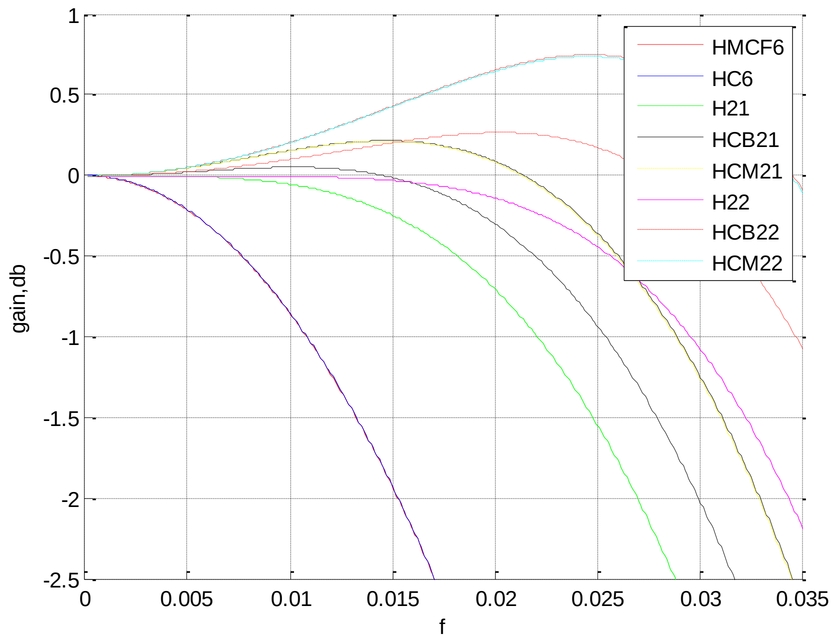
Now we consider another case for f_c = 0.02. The result is as shown in figure for the passband then the stopband. Passband characteristics are almost same but now we will take the readings at 0.02 and compare SMCF with compensated SMCFs for both the compensation techniques i.e. sine based compensator and maximally flat compensator.



a. Passband zoom for b (1 and 2) for H_{CB11}, H_{CM11}, H_{CB12} and H_{CM12}.

From Fig 5.3a and 5.3b we can make the inferences about droop at frequency 0.02. Fig 5.3a show that SMCF has droop of 0.38 dB while compensated SMCF H_{CB11} for b=2 gives droop of 0.18 dB and it is zero for b=1. While H_{12} gives droop of 0.03 dB while there is gain of 0.12db and 0.32db for b=2 and 1 respectively. So b=1 is giving the best desired results i.e. nearly zero droop. While H_{CM} is giving gain of 0.18db and 0.48db for H_{11} and H_{12} respectively.

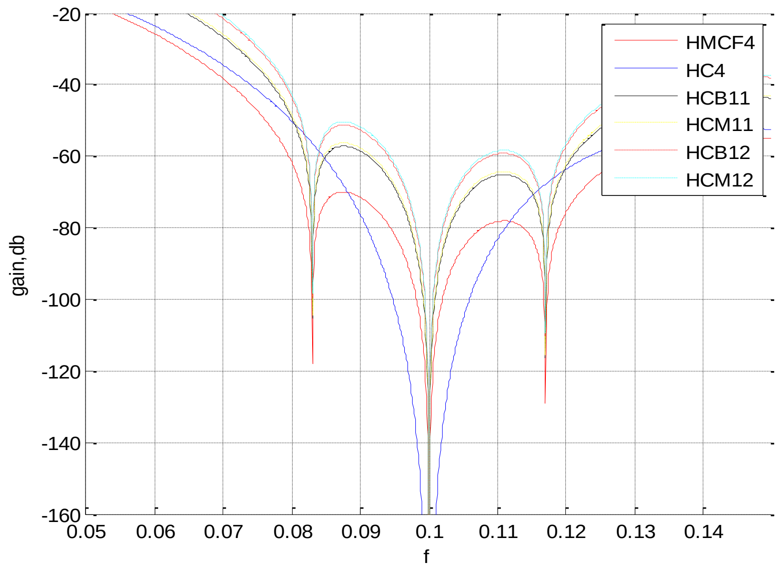
Fig.5.3b shows that H_{21} is giving droop of 0.71 dB while it is 0.3 dB for b= 1 and it has gain of 0.05db for both b=0 and H_{CM21}. H_{22} has droop of 0.15 dB while compensate SMCF gives gain of 0.2 dB and same for H_{CM22} and H_{CB22} for b=0. Compensated SMCF shows significant improvement as clear from graphs.



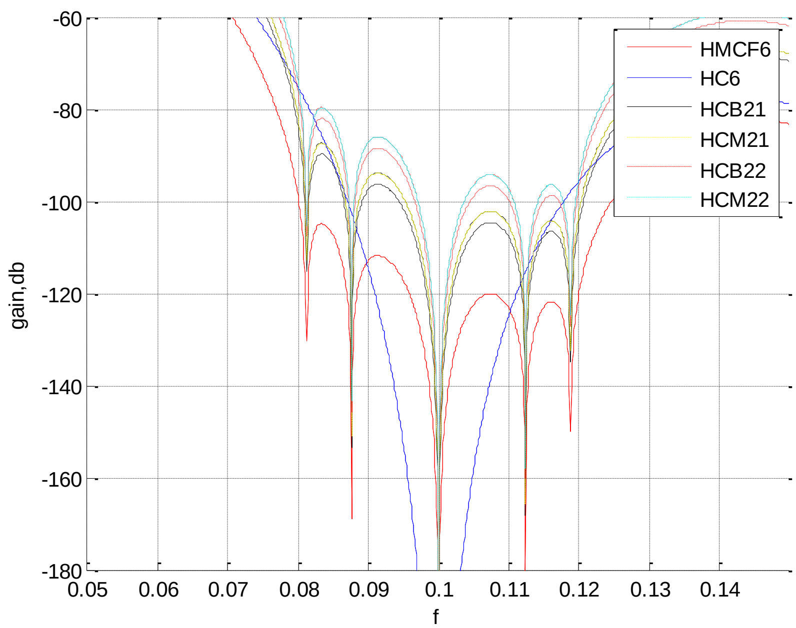
b. Passband zoom for b(0 and 1) for H_{CB21}, H_{CM21} , H_{CB22} and H_{CM22}.

Fig.5.3. Passband zoom for different values of b for SMCFs and compensated SMCFs for R=10 and f_c=0.02

Next are the figures showing alias rejection that stopband has widened with f=0.02 as compared to case with 0.01 since we are having q1=0.64 and q2 = 0.94 making the alias rejection much better than previous case. As the stopband lies in the range $\frac{1}{R} - q\mathrm{f_c}$ to $\frac{1}{R} + q\mathrm{f_c}$. Hence stopbands range is (o.1± 0.94*f_c) . Compared to previous case here f_c is 0.02 than 0.01 in previous case, so the stopband has widened accordingly. From the whole discussion it is very clear that response has improved significantly compared to already existing techniques. So overall improvement which was targeted is achieved with the use of sine based and maximally flat compensation techniques as is clear from the graphs. The graphs shown next shows the stopband of compensated Comb filter.



a. Stopband zoom for H_{CB11},H_{CM11},H_{CB12} and H_{CM12}.



b. Stopband zoom for H_{CB21},H_{CM21},H_{CB22} and H_{CM22}.

Fig.5.4. Stopband zoom for SMCF and compensated SMCFs for R=10, fc=0.02.

CHAPTER6

CONCLUSION AND FUTURE SCOPE

6.1 CONCLUSION

CIC filter is a simple and very economical decimation filter but its response suffers from the problem of passband droop with the increase in number of stages of filter. In practical application it is required to have a flat passband to preserve the original signal but CIC suffers from droop in passband which for various applications is not accepted. So this is great area of interest to improve the passband of CIC filters. Various methods are presented by various scientists till date for the response improvement of CIC filters. Sharpening technique, poly phase decomposition, two stage implementation, modification with rotated zeros, compensations and their cascade arrangements are employed for the droop reduction or response improvement of CIC filters. Sine based compensator and maximally flat compensators are the few latest techniques used in this area.

The droop can be reduced by modifying the structure of CIC filter, this technique is called sharpening or it can also be done by cascading the CIC with the additional filter called compensator which compensates the droop. The compensator filter takes the form of inverse of CIC filter frequency response. Stopband response can be improved by rotating the zeros of filter in both clockwise and anticlockwise direction to achieve wide stop band or more alias rejection.in the recent years CIC compensators are used to approximate inverse amplitude response of CIC filter. Several methods for the design of compensators are proposed and are now combined with already existing techniques to achieve better results.

This brief shows the response obtained by cascade of sharpened modified comb filter (explained in chapter 3) which are used in delta sigma A/D converters with the compensator both sine based compensator and maximally flat compensator. It also compares the results obtained with already existing results. Sharpened modified comb filter is basically an overall response improvement technique in which passband droop reduction is done using sharpening technique and for stopband improvement modified sinc is used and the cascade of two techniques gives better overall response of cascaded integrator comb (CIC) filter. Compensators are the filters which take the form of inverse of CIC filter response and are used for droop reduction. And this brief combines both the techniques to further reduce the droop in sharpened modified comb filter and with this new cascade the droop at cut off frequency is zero. So we are getting zero droop and wide passband which indicates the overall response improvement in CIC.

6.2 FUTURE SCOPE

This research work is based on improving characteristic of CIC filter especially improving the passband with simultaneous improvement in stopband. Improvement of SMCF is targeted here. The future work includes the proposal of new algorithms which improves the pass band, reduces hardware complexity and cost of filters. The new methods to make CIC filters application specific i.e. to get desired amount of pass band and stop band. The two stage architecture can also be employed to this compensated Sharpened modified comb filter so reduce the complexity of the filter design and reducing the number of computations required. Optimization techniques can also be used to get the optimal response.

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