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DIGITAL FILTERS OPTIMISATION USING RESIDUAL LEARNING MODEL DIGITAL FILTERS FOR COHERENT OPTICAL RECEIVERS

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Abstract

Digital Filter Optimization (DFO) is one of the features that manages limiting the expansion of optical fiber transmission networks to exceedingly high speeds, large capacities, and great distances. DFO requires a great amount of computational effort, and this paper presents a novel optimisation technique that is based on the DFO approach. The optimisation approach that has been described has the potential to significantly cut down on the required quantity of mathematical effort that is essential for scenarios involving vast distances and enormous DFOs. When compared to the existing method that is the way that is most generally used, the proposed approach can reduce the complexity of the hardware implementation of commercial systems by more than 60%, and the optimisation impact is highlighted when the transmission distance is increased.

Keywords:

Residual, Filters, Digital, Optical Receivers, Learning Model

1. INTRODUCTION

When optical signals that have multiple frequency components or multiple model components are sent via an optical fibre, the signal pulses expand because of the varying group velocities that occur throughout the transmission. This is because optical signals have multiple frequency components and optical models have numerous model components [1]. The overlapping of the front optical pulse with the rear optical pulse creates inter-symbol interference (ISI), which in turn causes an increased bit error rate (BER). This happens as the CD builds up. As a result of this, the Chromatic dispersion (CD) is one of the features that manages limiting the expansion of optical fibre transmission networks to exceedingly high speeds, large capacities, and great distances [2].

The use of dispersion compensation fibres (DCF) or dispersion compensating modules (DCM) to enable CDE in the optical domain increases the level of technological complexity as well as the overall cost of deployment. Digital coherent receivers are typically used in high-speed coherent optical communication systems because to their capacity to conduct CDE in the electrical domain, recover pulse signal broadening and distortion, and conduct dynamic compensation [3].

Examples of digital filters that can be used for chromatic dispersion equalisation (CDE) include the time-domain CD filter and the frequency-domain chromatic dispersion equaliser (FD-CDE) filter [4]. Both filters are time-domain filters and frequency-domain filters, respectively.

The CDE module needs to be streamlined to ease improvements in the functionality of coherent receivers [5]. This is necessary because the CDE module has a large power consumption, and the development of coherent receivers requires a considerable number of calculation units. The CDE module places high memory demands on the Digital Signal Processor (DSP) [6].

Time aliasing and poor contact with other TD modules may prevent significant power savings from being realised [7], even though the FD-CDE approach is the best possibility for commercial coherent receivers. This is because time aliasing is a form of interference that occurs when two different time domains interact with each other. The reason for this is that the time and frequency of the input sequence need to be transformed before they can be used. Because the entire process is conducted in TD, the concerns of FFT block size and temporal aliasing are sidestepped by TD-CDE. The principal concentration of this investigation is directed on TD-CDE and the optimisation approaches that are used by it.

The conventional truncation approach for the infinite and non-causal TD impulse response function as a rectangular window function. This is one way to think about the method. It is possible to truncate the function by using this method.

It is difficult to say how much of an impact the proportion of the energy of the primary lobe of the window function to the total energy has on the filtering performance after the function has been truncated because this proportion is a ratio. Due to the high-spectrum side lobe that the rectangular window features, it is possible to quickly inject interference at high frequencies into the window transformation [8].

This is made possible by the fact that the window transformation itself is quite simple. It has been recommended that a variety of traditional window functions, such as the FIR window, the Blackman-Harris window, and the Rife-Vincent window, should be used to further improve the effectiveness of the window function [9].

These windows include the FIR window, the Blackman-Harris window, and the Rife-Vincent window. These window functions get extremely near to delivering the perfect combination of a narrow main lobe, tiny side lobes, and quick attenuation; but they do not yet provide this optimal combination of characteristics. Jia postulated that a triangular pulse may be produced by putting a rectangular pulse through the process of self-convolution [10].

As an extension of this theory, convolution windows that are based on the FIR transform, triangular self-convolution windows, and hybrids incorporating rectangle and cosine windows have all been published in rapid succession [11]. As a possible extension of this idea, convolution windows that are derived from the FIR have also been proposed. The method that came before it does help in certain respects, but the TD-CDE has not yet been put to the test in actual practise.

The method that came before it does help. When there are more taps on the TD-CDE filter, the accumulative CD of the filter has a better chance of reaching a higher value. Achieve CDE in ultra-fast long-distance coherent optical communication systems, not only is a great amount of mathematical effort needed, but also a huge number of TD-CDE filter taps are essential components.

2. FINITE IMPULSE RESPONSE (FIR) DIGITAL FILTER

When the impulse response of a digital filter can be broken down into discrete portions of time, we say that the filter has a finite impulse response. The filter does not get any input, the impulse response will consistently be a limited value even if there is no feedback.

The illustration of the typical transfer function $H(z)$ for a one-dimensional FIR digital filter that follows is as follows:

$$H(z)=\sum_{n=0}^{N-1} h(n)\, z^{-n}. \quad (1)$$

$$H(z)=h(0)+h(1)z\text{-}1+\cdots+h(n)z^{-(N-1)} \quad (2)$$

The length N of the filter is what defines the impulse response of the filter, which is represented by $h(n)$. The constituent parts of a 1D FIR filter may be found in the attached file. In this situation, the result in the time domain, which is represented by the symbol $y(n)$, is.

$$y(n)=x(n)*h(n) \quad (3)$$

The results of the frequency domain are represented by $Y(z)$, and they are as follows:

$$Y(z)=X(z)H(z) \quad (4)$$

When input signals in the frequency domain are represented by $X(z)$ and time-domain signals are represented by the symbol $x(n)$, respectively. The following is an example of the frequency response of a 1D FIR filter:

$$H(\omega k) = \sum_{n=0}^{N} h(n)\, e^{-j\omega kn} \quad (5)$$

where $\omega k=2\pi kN$; $H(\omega k)$ is the complex vector that the Fourier transform produces.

As can be seen in the preceding section, the frequency response of the FIR filter is consistent with what was expected. Samples of the frequency are taken from a total of N distinct locations across the interval $[0, \pi]$.

2.1.1 2D FIR Filter:

In the interest of simplicity, we will refer to the impulse response of a two-dimensional FIR filter as $h(n_1,n_2)$ in the following discussion.

To be more explicit, in the scenario that the two-dimensional transfer function $0\leq n_1\leq N_1\text{-}1$ and $0\leq n_2\leq N_2\text{-}1$, the following statements are true with respect to the transfer function $H(z_1,z_2)$ will be,

$$H(z_1,z_2) = \sum_{n_1=0}^{N_1-1}\sum_{n_2=0}^{N_2-1} h(n_1,n_2)\, z^{-n_{11}} z^{-n_{22}}$$

A schematic representation of the operational architecture of a 2D FIR filter and the output $Y(z_1,z_2)$ is:

$$Y(z_1,z_2)=H(z_1,z_2)\ .\ X(z_1,z_2) \quad (7)$$

where

$X(z_1,z_2)$ – 2D input.

The input is in two dimensions is carried out by exchanging $z_1=\exp(j\omega_1)$ and $z_2=\exp(j\omega_2)$. After performing the multiplication, we now get the frequency response of a 2D FIR filter, which can be expressed as

$$H\left(e^{j\omega_1} e^{j\omega_2}\right) = \sum_{n_1=0}^{N_1-1}\sum_{n_2=0}^{N_2-1} h(n_1,n_2)\, e^{-j\omega_1 n_1} e^{-j\omega_2 n_2} \quad (8)$$

3. INFINITE IMPULSE RESPONSE FILTER

The conventional IIR filter can be represented with the help of the difference equation that is given below:

$$y(n) = \sum_{k=0}^{M} p_k x(n-k) - \sum_{k=0}^{N} q_k y(n-k) \quad (9)$$

The p_k and q_k refer to the filter coefficients. $x(n)$ and $y(n)$ are used to denote the filter input and output, respectively. Each of the filter coefficients has a value of M and N, with $N\geq M$.

3.1.1 1D FIR Filter:

The following is a 1D depiction of the transfer function of an IIR filter can be represented as follows:

$$H(z) = \sum_{k=0}^{M} p_k z^{-k_1} + \sum_{k=0}^{N} q_k z^{-k} \quad (10)$$

The schematic representation of a one-dimensional IIR filter component part. In the time domain, the input and output signals are denoted by $x(n)$ and $y(n)$, while in the frequency domain, the input and output signals are denoted by $X(z)$ and $Y(z)$.

3.1.2 2D IIR filter:

The transfer function of a 2D IIR filter can be found below. In $a(0,0)=1$, then the following represents the 2D transfer function:

$$H(z_1,z_2) = \sum_{k_1,k_2\in R_b}\sum b(k_1,k_2)\, z^{-k_{11}} z^{-k_{21}} + \sum_{k_1,k_2\in R_{a-(0,0)}}\sum a(k_1,k_2)\, z^{-k_{11}} z^{-k_{21}} \quad (11)$$

To find the values of the filter coefficients $a(k_1,k_2)$ and $b(k_1,k_2)$ to filter the data. The region of support for the function $a(k_1,k_2)$ that does not have the origin (0, 0) is denoted by $Ra\text{-}(0,0)$, while the region of support for the function $b(k_1,k_2)$ is denoted by R_b. The origin (0, 0) is not included in the region of support for the function $b(k_1,k_2)$.

3.2 WAVELET FILTER BANK

A filter bank is a collection of several types of filters that are used for the purpose of spectral decomposition and recomposition of one- or two-dimensional signals. These filters can be FIR or IIR filters, low-pass and high-pass filters, and other sorts of filters as well.

Analysis filters are applied to a piece of information signal at the point where it goes through a wavelet filter bank. This happens when the signal is conveying information. Filters are both H_0 and H_1, but H_0 is a low-pass filter and H_1 is a high-pass filter. While both H_0 and H_1 are filters, H_0 is a low-pass filter and H_1 is a high-pass filter.

After the application of each filter, a sample that is taken twice is done so that the results can be compared. Upsampling, filtering, and a summation of the subbands are utilised during the synthesis phase so that the original signal may be reconstructed. This phase also occurs when the signal is being synthesised.

The following are the findings of the analysis (after any down sampling that may have been needed):

$$S(z)=0.5(H_0(z_{12})X(z_{12})+H_0(-z_{12})X(-z_{12})) \quad (12)$$

$$D(z)=0.5(H_1(z_{12})X(z_{12})+H_1(-z_{12})X(-z_{12})) \quad (13)$$

The result of the computations $A(z)=12(H_0(z)F_0(z)+H_0(z)F_1(z))$ and $B(z)=12(H_0(-z)\ F_0(z)+H_0(-z)F_1(z))$, respectively, is denoted by the variable $Y(z)$.

$$Y(z)=A(z)X(z)+B(z)X(-z) \quad (14)$$

To accurately reconstruct $Y(z)$, it is needed for $X(z)=A(z)X(z)+B(z)X(-z)$, where $X(z)$ might have any value. This equation must be satisfied before one can go ahead with the reconstruction of $Y(z)$. The following are the prerequisites for an impeccable reconstruction:

$$H_0(-z)F_0(z)+H_1(-z)F_1(z)=0 \quad (15a)$$

$$H_0(z)F_0(z)+H_1(z)F_1(z)=2 \quad (15b)$$

Achieve a full output reconstruction, it is required to verify that the synthesis low-pass filter F_0 and the synthesis high-pass filter F_1 are compatible with the analysis filters H_0 and H_1.

4. WINDOW FUNCTION WEIGHT OPTIMIZATION

The $HDes(e^{j\omega T})$ is used to show the response in the frequency domain. In a filter that is perfect, the amplitude-frequency characteristic of the passband does not change. To get the most out of the CDE effect, it is essential to focus most of the spectral energy that the window function produces on the primary lobe. This can be achieved by building the filter in such a way that its FD response is as close to $HDes(e^{j\omega T})$ as is practically possible. The filtered spectrum H(k) that is generated by truncating a rectangular window will, in contrast to the perfect FD response HDes($e^{j\omega T}$), exhibit extremely severe spectrum leakage. This is since the rectangular window has been truncated.

frequency-domain response HDes($ej\omega T$) of an ideal filter has a constant amplitude-frequency characteristic in the passband. When the FD response of the designed filter is close to HDes($ej\omega T$), the CDE effect is better, so the spectral energy of the window function is required to be concentrated on the main lobe as much as possible. The filtered spectrum $H(k)$ obtained by the rectangular window truncation will have serious spectrum leakage, which is quite different from the ideal FD response $HDes(ej\omega T)$

In this research, the second-order self-convolution window function is used since it owns the features of having the shortest side lobe as well as the fastest side lobe attenuation. This is since, in comparison to other window functions, the window with four terms and three orders has the quickest attenuation speed:

The following is a list of qualities that the TD has:

$$w_R(n) = \begin{cases} 1 & n = 1, 2, \ldots N \\ 0 & otherwise \end{cases} \quad (16)$$

Aside from the fact that its spectral characteristics are defined by,

$$W_R(\omega)=\sin(\omega^{N/2})\sin(\omega/2)\exp(-j\omega^{N-0.5}) \quad (17)$$

The window is a combination window because of the advantageous qualities possessed by its side lobes; the TD representation of this window is:

$$\omega_N(n) = \sum_{m=1}^{M-1} (-1)\, m_{cm} \cos\left(\frac{2\pi nm}{N}\right), n = 1, 2, \ldots, N-1 \quad (18)$$

The list of constraints M that follows, along with the length of the window function (denoted by N), the number of terms that are contained within the window function (denoted by C_m), and the total number of terms, are as follows:

$$\sum_{m=0}^{M-1} (-1)\, m_{cm} = 0, \sum_{m=0}^{M-1} cm = 1 \quad (19)$$

In this instance, the spectral function of the rectangular window is represented by W. The shape of the window is like that of a rectangle. These are the coefficients that have been picked for the window to be used.

This window is of the third order and has four terms and it is possible to buy the second-order self-convolution window function by performing a self-convolution operation on the four-term, third-order window.

$$w_{RN}(n)=w_N(n)*w_N(n) \quad (20)$$

The self-convolution window outperforms the rectangular window when it comes to the performance of the side lobes, as seen by the comparison between the two windows. It is possible to successfully cut the signal with the use of the self-convolution window, which is a handy technique that can be beneficial in reducing the amount of spectrum leakage.

Amplitude restoration is something that needs to be done because the signal original amplitude in the FD will be different after it has been shortened in the TD using the window function.

The amplitude-restoration coefficient, denoted by $h(n)$, has been given a definition that can be used in practise. The fact that the amplitude does not vary even when extra window functions are added to a window that has the same length as a rectangular window. This is referred to as the rectangular window property.

$$K_m = A_1/A_2 \quad (21)$$

where, A_1 stands for the amplitude of the zero-frequency point after the application of a rectangular window, and A_2 stands for the amplitude of the zero-frequency point after the application of further windows. The TD-CDE is only capable of producing a certain level of equalisation effect due to the limitations imposed by the constant modulus of the tap coefficient.

If there is no longer a constant value, the following strategy is equivalent to using a window function to optimise the weight of the CDE taps, which ultimately results in an increase in the efficiency of the CDE filter. If there is no longer a constant value, the following strategy is analogous to using a window function to optimise the weight of the CDE taps.

5. RESULTS AND DISCUSSION

The proposed algorithm is confirmed by first putting in place a coherent optical communication system using the programme

Optisystem 15, and then doing so through the implementation of DSP using components from MATLAB. In the graphic, you can see an example of how this procedure works.

This is achieved by using a square root raised cosine (SRRC) pulse signal generator at the transmitter and an SRRC filter at the receiver. Both components are in the receiver. The method of matched filtering is utilised by the system. The fibre link is formed of a loop controller, an optical filter, and optical fibre that extends for 100 kilometres.

In addition, there is twenty decibels of gain correction and six decibels of noise that are present in the system. Moving the sequence from the ADC, which samples it at twice the symbol rate, to the MATLAB part is the final stage in the development of the DSP. The study considers multiple quantization levels *M*, investigates how the bit error rate (BER) changes with transmission distance for a variety of different filters, and compare the chromatic dispersion equalisation (CDE) performance of the filter.

In an actual connection, the signal OSNR changes depending on the link length. The minimum OSNR that must be achieved to reach the target bit error rate (BER) is computed. The rate at which bits are dropped during transmission is referred to as the bit error rate (BER).

Table.1. Parameters

Parameter	**Value**
Wavelength	1450 nm
Sequence Length	131000
Symbol Rate	26 GBaud
Modulation Format	QPSK/32QAM
Roll-off Factor	0.35/1
Chromatic Dispersion Coefficient	16.5 ps/nm/km
Group Velocity Dispersion Coefficient	0.2 ps/km
Width	0.1 MHz
Attenuation Coefficient	0.2 db/km
Signal Power	10 dbm

In the simulation experiment that is discussed in this paper, we hold the OSNR constant while taking into consideration the natural variation in BER. To guarantee that we would be able to generate a particular amount of noise within the simulation, we arranged the setup OSNR component in such a way that it would run after the fibre link transmission. This was done to ensure that we would be able to do so.

The results of the four filters that used DP-QPSK and DP-16QAM modulations and had an OSNR of 15 dB and 21 dB, respectively. The quality factor Q so that an objective investigation of the performance reduction brought on by quantization may be carried out. This factor is defined as the ratio of the difference in BER before and after quantization to the BER before quantization. This ratio is referred to as the BER factor.

Table.2. degree of correlation

L/km	***N***	***M***	**mul**	**add**
100	200	8	95.30	41.00
		12	94.03	40.51
		16	92.76	39.92
200	400	8	96.58	41.49
		12	95.99	41.19
		16	95.30	41.00
300	600	8	96.97	41.68
		12	96.58	41.29
		16	96.18	41.29

The Table.2 illustrates that anytime the number of quantization stages is raised, the computational complexity of the optimised approach will, on average, experience a slight increase. This is the case even when the fibre transmission distance is held constant at the value shown in the table. Using computer optimisation, the complexity of the operation known as real-number multiplication (mul) can still be reduced by more than 91%.

Increasing the quantization order is one method by which the performance of the approach can be considerably enhanced. Other methods are also discussed. There is an inverse relationship between the transmission distance and the reduction in the computation cost of the optimised technique when the quantization step *M* is held constant. This relationship exists because there is a correlation between the two variables.

In conclusion, the optimisation technique that was described has the potential to significantly cut down on the required quantity of computational effort that is essential for scenarios involving vast distances and enormous CDs. Even though it was essential to make a performance sacrifice and *M* = 16 was used for the quantization phases, the optimisation indices mul and add nevertheless show effective optimisation. This is the case even though M = 16 was used for the quantization phases.

6. CONCLUSION

There is a 90% decrease in the real-number multiplication operation and a 40% decrease in the real-number addition operation when the quantization coefficient M is equal to 16. Despite this, there is a 4% decrease in performance when the quantization coefficient M = 16. When compared to the proposed method, which is the way that is most generally used, the proposed approach that has been presented can reduce the complexity of the hardware implementation of commercial systems by more than 55%, and the optimisation impact is highlighted when the transmission distance is increased. This is because the method that was proposed considers a certain equilibrium performance loss in the long-distance optical fibre link.

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