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Design and Implementation of DWMT Transmission Systems using IIR Wavelet Filter Banks

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Abstract: Multicarrier modulation using orthogonal frequency division multiplexing (OFDM) is often recognized as an efficient solution for wireless communications. However, waveform bases other than sinusoidal functions can similarly be used for multicarrier systems in order to provide an alternative to OFDM. In this context, DWMT transmission systems are suggested. It is known that such systems suffer obviously from the two main problems of Inter Symbol Interference (ISI) and Inter-Carrier Interference (ICI). ISI may be considerably reduced via channel equalization. The effect of ICI can be reduced using high order Finite Impulse Response(FIR)wavelet filter banks, but at the expense of increased system complexity. Those FIR banks which are required to reduce effect the ICI should have autocorrelation functions approaching an impulse. On other hand, Infinite Impulse Response(IIR) filter banks can be used to replace those FIR counterparts to increase sub-band discrimination while reducing the required filter order, i.e., decreasing the complexity of system. In this paper, a special type of IIR filter banks is designed and implemented for such purpose. Bi-reciprocal lattice wave digital filters (BLWDFs) are utilized in an approximate linear phase design of 11th order IIR wavelet filter bank (FB) for discrete wavelet multi-tone (DWMT) transmission systems. The performances of the proposed DWMT for transmission over simulated wireless channels with additive white Gaussian noise (AWGN) are examined and compared with one of the recent FIR designs. Significant improvements over the classical DWMT approach are observed.

Keywords: OFDM, DWMT, BLWDF, IIR Wavelet filter, ICI, ISI, AWGN, SNR.

I. Introduction

It is known that, Single-carrier techniques are vulnerable for the problems of fading and multipath propagation. Recently, multicarrier modulation (MCM) or OFDM have received considerable attentions and have made a great deal of progress to get rid of such problems in the world of communication. Increasing the robustness against frequency selective fading and having narrowband interference are the main reasons for using OFDM or MCM. In a single carrier system, a single fade or interferer can cause the entire link to fail, but in a multicarrier system, only a small percentage of sub-carriers will be affected. The few erroneous affected sub-carriers can easily be corrected using error correction coding. MCM is the principle of transmitting data by dividing input stream into several sub- streams with much lower symbol rates.

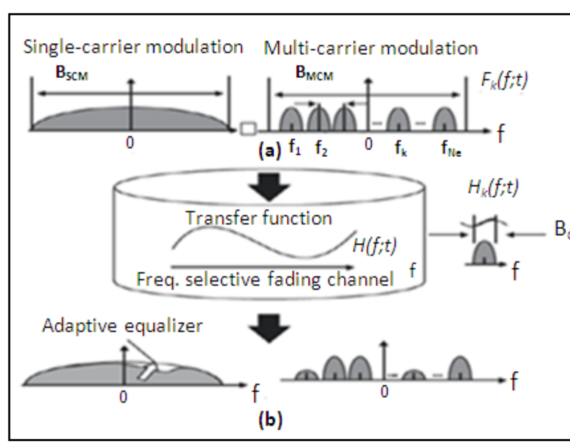


Fig. 1. Comparison of SCM and MCM: (a) frequency spectra of transmitted signals; and (b) frequency spectra of received signals, after [1].

Several sub-carriers are used to modulate these sub-streams. Figure 1 compares a single carrier modulation (SCM) and an MCM systems. In such figure, BSCM and BMCM denote the bandwidths of transmitted SCM and MCM signals, respectively [1].
From Fig. 1, it can be seen that for the recovery of the signals at the receiver, an adaptive equalizer is needed for SCM. While in the case of MCM, a simple amplifier can be used for the influenced band.
In wireless communication, the concept of parallel transmission of symbols is applied to achieve high throughput and better transmission quality. In early parallel transmission systems, few non-overlapping sub-channels shared the whole frequency band as shown in Fig. 2. Apparently the existence of a guard band between two adjacent sub-channels is to provide non-overlapping sub-channels so as to eliminate possible interference between adjacent sub-channels, which is known as Inter-Carrier Interference (ICI). This guard band constitutes a waste of spectrum. However in the mid-1960s, spectral efficiency was improved by overlapping the sub-channels as shown in Fig. 3, which saved up to 50% of the spectrum used and was developed using OFDM technology[2].OFDM is one of the powerful techniques for parallel transmission. The idea of OFDM is to split the total transmission bandwidth into a number of orthogonal sub-carriers in order to transmit the symbols using these sub-carriers in parallel [3].

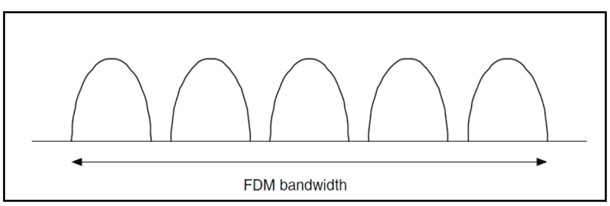


Fig.2. Conventional nonoverlapping multicarrier modulation, after [2].

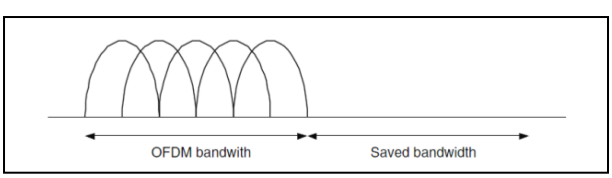


Fig.3. Overlapping multicarrier modulation, after [2].

In recent years, multicarrier modulation has attracted considerable attention as a practical and viable technology for high-speed data transmission over spectrally shaped noisy channels. The discrete multi-tone (DMT) is the most widely used technique for cost-effective realization of multicarrier transceivers. DMT-based modems have been accepted by standardization bodies in both wired and wireless channels [4]. DMT is a special multicarrier modulation which uses the properties of the discrete Fourier transform (DFT) in an elegant way to achieve a computationally efficient realization. However, owing to the poor quality (low stop-band attenuation) of the DFT filters, DMT transceivers are very susceptible to narrowband interferences as shown in the Fig. 4 (a).
A solution to this problem is to replace the DFT filters by a filter banks with higher stop-band attenuation. Trans-multiplexers using maximally decimated filter banks

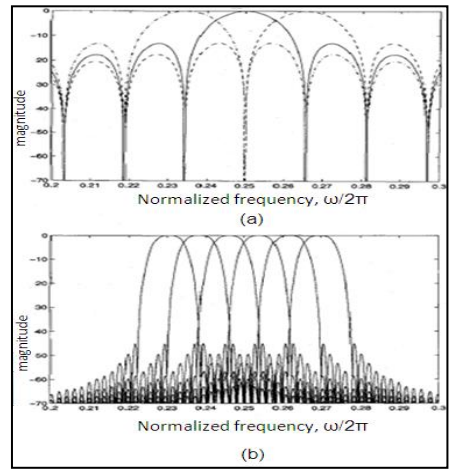


Fig.4. Frequency responses of six spectrally contiguous sub-channel pulse sequences: (a) DMT transmission; (b) DWMT transmission, after [4].

with perfect /near-perfect reconstruction (PR/NPR) property, known as discrete wavelet multi-tone (DWMT), are the most desirable candidate for this application[5] as shown in the Fig. 4 (b). The idea of using filter banks with PR/NPR to implement multicarrier modulation systems is supported by the following intuitions: (i) the inter channel interference in a filter bank is dominated by the interferences among neighboring channels; (ii) such interferences are minimized when the neighboring channels/bands undergo a similar (complex) gain. Ina traditional DWMT system, FIR filter banks are designed to minimize ICI by decreasing the side lob energy of the filter banks in the modulator and demodulator sides[6].For more ICI reduction and less complexity, IIR wavelet filters with lattice all-pass sections can be used as sub-filters in the filter banks of DWMT system. It should be noted that ISI is not considered in all those designs.
There were many applications for the multicarrier modulation (MCM). For example, in 2002, Daniel , et al. [6] proposed a power-line communication system using a type of DWMT modulation. Another example, in the same year, B. Farhang, et al. [7], proposed the cosine modulated filter bank (CMFB) as an MCM tool for wideband data transmission over wireless channels. In 2005, cosine modulated filter banks (CMFBs)were used by B. Farhang, et al. [8] for MCM in the application of very high-speed digital subscriber lines (VDSL) utilizing cosine modulated multi-tone transmission (CMT) as a modulating technique.CMT is fundamentally the same as

the DWMT except the receiver structure. With the modified receiver structure, CMT uses only two taps per sub-carrier for equalization. Wavelet packet-based multicarrier transceivers were applied in wireless communication, however their equalization remained as an open research area [9]. In 2008, K.Hung Chen, et al. [10] proposed cognitive radio (CR) system to operate in the GSM band and the discrete wavelet multi-tone was adopted as the modulation scheme. This modulation has stronger side-lobe attenuation than the popular OFDM modulation and thus generates much lower adjacent channel leakage, making it suitable for applications with dynamically allocated spectrum. In *Ref.*[11], a circular version of wavelet orthogonal frequency division multiplexing (Circular Wavelet-OFDM), which can be called also a circular discrete wavelet multi-tone (Circular-DWMT), was introduced and its application in power line communications (PLC) was presented and tested. More recently, A. Ikram , et al. [12], presented a new design for FIR wavelet filters for DWMT systems. The resulting filters were orthogonal and consisting of only two nonzero components for any filter order. In all those trials, FIR filter banks were utilized with their drawbacks of spread roll-off spectrum characteristics.

In this paper, BLWDFs are utilized in the design of some spatial type of IIR wavelet filter banks with linear phase processing for DWMT transmission systems. It is believed that such IIR filter banks can replace their FIR counterparts with less complexity and increased sub-band discrimination, which leads to some ICI reduction.

Besides this introductory section, this paper contains the following sections: Section II describes the bi-reciprocal lattice wave digital filters. The idea of IIR wavelet filter banks is presented in section III. Section IV contains the proposed design for such IIR perfect reconstruction wavelet filter banks. An example of 11th - order intermediate IIR filter design is given in section V. The resulting SNR values for the received signals from different DWMT systems and the autocorrelations of different tested filters are listed in section VI with discussions. Finally, section VII concludes this paper.

II. BI-RECIPROCAL LATTICE WAVE DIGITAL FILTER

Wave digital filters (WDFs) include a wide class of IIR digital filters that are compatible for implementation. WDFs are originally derived from analog reference filters from which they acquire several fundamental properties. Examples of WDF structures, suitable for implementation, are Richards' structures and ladder structures. However, a class of WDFs that is even more suitable for VLSI implementation is lattice WDFs (LWDFs). These filter structures are derived from analog lattice filters [13].

A lattice wave digital filter (LWDF) is a two-branch structure where each branch realizes an all-pass filter (see Fig. 5).The digital all-pass filter is a computationally efficient signal processing building block which is quite useful in many signal processing applications. The frequency response $A(e^{j\omega})$ of an M^{th} order all-pass filter exhibits unit magnitude at all frequencies, *i.e.*,

$$\left|A(e^{j\omega})\right|^2 = 1 \qquad \text{for all } \omega \tag{1}$$

In this case A(z) can be expressed as

$$A(z) = \frac{z^{-M}D(z^{-1})}{D(z)} \tag{2}$$

In effect, the numerator polynomial is obtained from the denominator polynomial by reversing the order of the coefficients. Thus, The numerator and denominator polynomials are said to form a mirror-image pair[14].For example,

$$A(z) = \frac{a_2 + a_1 z^{-1} + z^{-2}}{1 + a_1 z^{-1} + a_2 z^{-2}} \tag{3}$$

is a second-order all-pass function of the form of (2).

There exist several ways to realize LWDFs. One of the most attractive ways is to use cascaded first-order and second-order sections, as shown in Fig. 5. The first-order and second-order sections are realized using symmetric two-port series- or parallel adaptors with certain equivalence transformations. The second order sections can also be realized using three-port series- or parallel adaptors[15].

A bi-reciprocal (half-band) lattice WDF (BLWDF) is a special type of a LWDF. In this case, the even-indexed coefficients in Fig. 5become zeros which means less complexity for the same order. This results in the simplified structure shown in Fig.6.The transfer function of a bi-reciprocal lattice WDF can be written as

$$H(z) = \frac{1}{2}\left[A_0(z^2) + z^{-1}A_1(z^2)\right] \tag{4}$$

Where the transfer function $A_0(z^2)$ corresponds to the lower branch in Fig.6 and $A_1(z^2)$ is the upper.

The overall frequency response can be written as

$$H\left(e^{j\omega T}\right) = \frac{1}{2}\left(e^{j\Phi_0(\omega T)} + e^{j\Phi_1(\omega T)}\right) \tag{5}$$

Where

$$\Phi_k(\omega T) = -k\omega T + \arg\{A_k(e^{j\omega T})\}, k = 0,1 \tag{6}$$

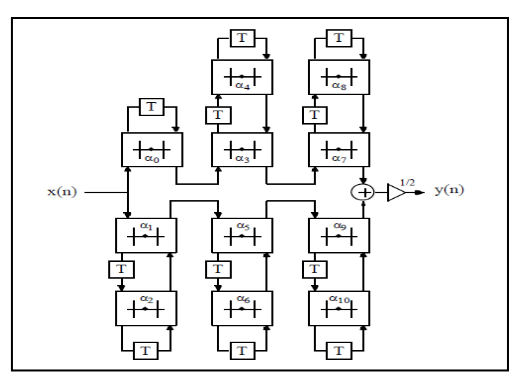


Fig. 5. An 11th-order lattice wave digital filter, after [15].

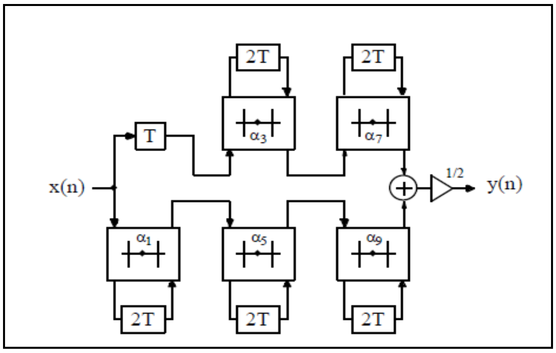


Fig.6. An 11th-order bi-reciprocal LWDF, after [15].

is the phase response of branch k and T is the sampling period. The magnitude of the overall filter is thus limited by

$$|H(e^{j\omega T})| \leq 1 \tag{7}$$

The transfer function of BLWDF and its complementary transfer function are power complementary functions. Therefore, BLWDF possesses the following relations[15]:

$$|H(e^{j\omega T})|^2 + |H(e^{j(\omega T-\pi)})|^2 = 1 \tag{8}$$

which means that the pass-band and stop-band edges (ω_c & ω_s, respectively) are related by $\omega_c T + \omega_s T = \pi$. Another consequence is that if the squared magnitude function approximates unity with the tolerance ε in the pass-band [0, ω_cT], then it will approximate zero with the same tolerance in the stop-band [ω_sT, π]. Thus, the pass-band ripple will be extremely small for practical requirements on the stop-band attenuation.

The concepts of attenuation zeros and transmission zeros can be used for the following discussions. These zeros always occur on the unit circle in the z-plane. An attenuation zero corresponds to an angle ω_0T at which the magnitude function reaches its maximum value. For BLWDFs this occur when

$$|H(e^{j\omega_0 T})| = 1 \tag{9}$$

A transmission zero corresponds to an angle ω_1T at which the magnitude function is zero, *i.e.*, when

$$|H(e^{j\omega_1 T})| = 0 \tag{10}$$

If the filter has an attenuation zero at ω_0T, then it follows from (8) that it also has a transmission zero at $\pi - \omega_0$T. To obtain a low-pass filter the phase responses of the overall branch transfer functions must approximate the phase behavior of an ideal poly-phase network. At an attenuation zero, the phase responses of the branches must have the same value, *i.e.*,

$$\Phi_0(\omega_0 T) = \Phi_1(\omega_0 T) \tag{11}$$

Hence, in the pass-band of the filter the phase responses must be approximately equal. At a transmission zero the difference in phase between the two branches must be

$$\Phi_0(\omega_1 T) - \Phi_1(\omega_1 T) = \pi \tag{12}$$

Thus, the difference in phase between the two branches must approximate π in the stop-band of the filter. To make sure that only one pass-band and one stop-band occur, the degrees of $H_0(z^2)$ and $H_1(z^2)$ must only differ by one[16].

III. IIR Wavelet Filter Banks

A very efficient way for representing the QMF bank can be obtained by using poly-phase structure. QMF banks, composed of two all-pass filters, are known to be the best circuits for building up a multichannel IIR filter banks. They can completely eliminate the aliasing error and amplitude distortion. Figure7 shows a two channel all-pass filter-based IIR QMF banks with lattice structure.

In Fig. 7, the lattice components are the 2nd order all-pass filters $A_0(z^2)$ and $A_1(z^2)$ with the following transfer functions [16]:

$$A_0(z^2) = \prod_{i=2,4}^{(N+1)/2} \frac{\alpha_i + z^{-2}}{1 + \alpha_i z^{-2}} \tag{13}$$

and

$$A_1(z^2) = \prod_{i=3,5}^{(N+1)/2} \frac{\alpha_i + z^{-2}}{1 + \alpha_i z^{-2}} \tag{14}$$

where $\alpha_i$is the value of the multiplier coefficient in the i^{th} all-pass section.

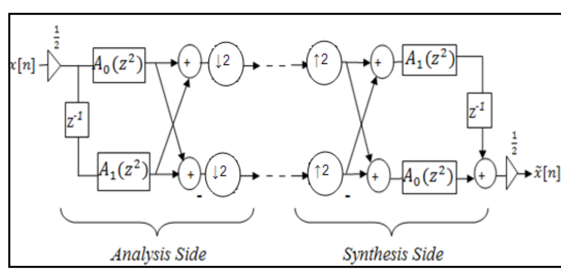


Fig.7. Lattice structure realization of the IIR wavelet filter bank, after [17].

Let $H_0(z)$ and $H_1(z)$ denote the transfer functions of the low-pass and high-pass filter of the analysis part of the two-channel quadrature-mirror filter (QMF) bank, and let $G_0(z)$ and $G_1(z)$ denote, respectively the transfer functions of the low-pass and high-pass filter of the synthesis part. By choosing transfer functions to satisfy the following conditions:

$H_1(z)=H_0(-z)$,$G_0(z)=H_0(z^{-1})$ and $G_1(z)=H_1(z^{-1})$

then, the filter bank will possess both perfect reconstruction and orthonormality properties [17].

Analysis filters $H_0(z)$ and $H_1(z)$ can respectively, be written for the low-pass side as

$$H_0(z) = \frac{1}{2}[A_0(z^2) + z^{-1}A_1(z^2)] \quad (15)$$

and for the high-pass side as

$$H_1(z) = \frac{1}{2}[A_0(z^2) - z^{-1}A_1(z^2)] \quad (16)$$

IV. THE PROPOSED DESIGN FOR IIR PERFECT RECONSTRUCTION WAVELET FILTER BANKS

In this section, how to drive the solutions for the design problems of intermediate filters (whose characteristics are between IIR Butterworth and Daubechies filters) are proposed. The half band filter's poles are placed on an imaginary axis of the complex z-plane, where one of them is placed at the origin and the remaining conjugate complex pole pairs are located between $e^{-j\pi}$ and $e^{j\pi}$. All zeros of such filters are located on the unit circle, where nine of them are placed at z = -1 for 11th order filter to meet the flatness condition, and remaining two of zeros placed at(-a+jb & -a-jb) to meet the good discrimination condition.

In applications of filter banks and wavelets, an important role is played by the regularity of the low-pass prototype filter; a feature which is closely related to the flatness on the magnitude response of the filter at the Nyquist frequency ω = π. In constructing orthonormal bases of wavelets from iterated filter banks, a greater number of zeros of the low-pass filter at ω = π results in more regular wavelets. It can be shown that having a maximum number of zeros at z= -1, implies a maximally flat characteristic for the filters involved. As an example, the Butterworth filter has a maximally flat magnitude response as it has all zeros at z = -1 with the highest possible regularity order, but it has the worst frequency selectivity[16].

V. 11TH ORDER INTERMEDIATE IIR FILTER DESIGN

The proposed 11th order intermediate IIR filter as shown in Fig. 8, has eleven zeros and eleven poles. Their values can be obtained depending on the desired magnitude response $|H_0(e^{j\omega})|$ that is shown in Fig. 9.The transfer function for the proposed design can be obtained as follows, depending on its positions of poles and zeros that are distributed as in Fig. 8.

$$H(z) = \frac{k[(z+1)^9\,[(z+a)^2 + b^2]]}{z(z^2+\alpha^2)(z^2+\beta^2)(z^2+\gamma^2)(z^2+\sigma^2)(z^2+ʒ^2)} \quad (17)$$

where a, b, α, β, γ, σ and ʒ are constants less than (1), as shown in Fig. 8 with k as a magnitude scaling factor.

Equation (17) can be simplified to the following general filter function:

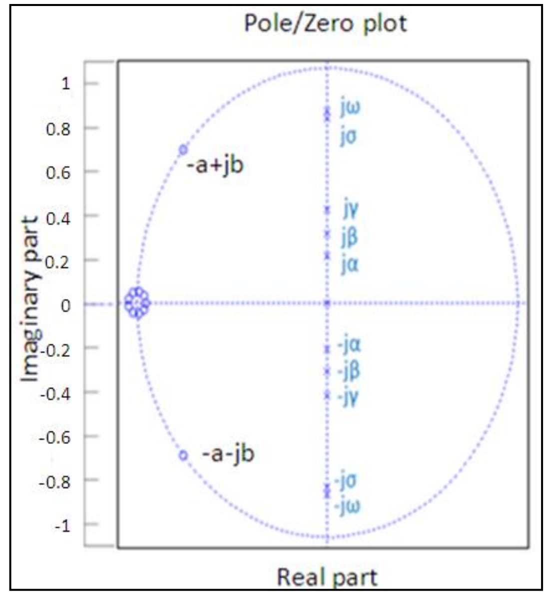


Fig.8. Pole-zero plots for 11th order intermediate IIR half-band filter.

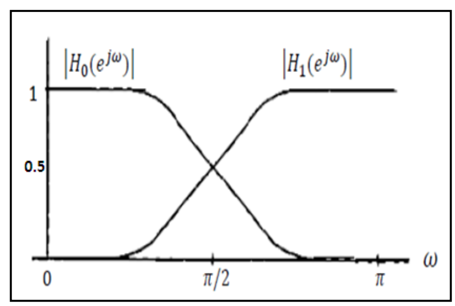


Fig.9. Magnitude responses for 11th order intermediate IIR half-band filters.

$$H(z) = k\left[\frac{\begin{gathered} z^{11} + z^{10}(2a+9) + z^9(a^2+18a \\ +b^2+36) + z^8(9a^2+72a+9b^2 \\ +84) + z^7(36a^2+168a+36b^2+ \\ 126) + z^6(84a^2+252a+84b^2+ \\ 126) + z^5(126a^2+252a+126b^2 \\ +84) + z^4(126a^2+168a+126b^2 \\ +36) + z^3(84a^2+72a+84b^2+9 \\ ) + z^2(36a^2+18a+36b^2+1) + \\ z(9a^2+2a+9b^2) + (a^2+b^2) \end{gathered}}{\begin{gathered} z(\alpha^2\beta^2\gamma^2\sigma^2ʒ^2) + z^3(\alpha^2\beta^2\gamma^2\sigma^2 + \alpha^2 \\ \beta^2\gamma^2ʒ^2 + \alpha^2\beta^2\sigma^2ʒ^2 + \alpha^2\gamma^2\sigma^2ʒ^2 + \beta^2 \\ \gamma^2\sigma^2ʒ^2) + z^5(\alpha^2\beta^2\gamma^2 + \alpha^2\beta^2\sigma^2 + \alpha^2 \\ \beta^2ʒ^2 + \alpha^2\gamma^2\sigma^2 + \alpha^2\gamma^2ʒ^2 + \alpha^2\sigma^2ʒ^2 + \\ \beta^2\gamma^2\sigma^2 + \beta^2\gamma^2ʒ^2 + \beta^2d^2ʒ^2 + \gamma^2\sigma^2ʒ^2) + \\ z^7(\alpha^2\beta^2 + \alpha^2\gamma^2 + \alpha^2\sigma^2 + \alpha^2ʒ^2 + \beta^2\gamma^2 \\ +\beta^2\sigma^2 + \beta^2ʒ^2 + \gamma^2\sigma^2 + \gamma^2ʒ^2 + \sigma^2ʒ^2) \\ +z^9(\alpha^2+\beta^2+\gamma^2+\sigma^2+ʒ^2) + z^{11} \end{gathered}}\right] \quad (18)$$

Equation (18) is found as a function of z. In order to be used in terms of z^{-1}, the numerator and denominator of such equation are multiplied by z^{-11}, as follows:

$$H(z) = k\left[\frac{\begin{array}{c}1 + z^{-1}(2a+9) + z^{-2}(a^2+18a+b^2+\\ 36) + z^{-3}(9a^2+72a+9b^2+84) + z^{-4}\\ (36a^2+168a+36b^2+126) + z^{-5}(84\\ a^2+252a+84b^2+126) + z^{-6}(126a^2+\\ 252a+126b^2+84) + z^{-7}(126a^2+168\\ a+126b^2+36) + z^{-8}(84a^2+72a+84\\ 84b^2+9) + z^{-9}(36a^2+18a+36b^2+1)\\ +z^{-10}(9a^2+2a+9b^2) + z^{-11}(a^2+b^2)\end{array}}{\begin{array}{c}z^{-10}(\alpha^2\beta^2\gamma^2\sigma^2\xi^2) + z^{-8}(\alpha^2\beta^2\gamma^2\sigma^2+\alpha^2\beta^2\\ \gamma^2\xi^2+\alpha^2\beta^2\sigma^2\xi^2+\alpha^2\gamma^2\sigma^2\xi^2+\beta^2\gamma^2\sigma^2\xi^2)\\ +z^{-6}(\alpha^2\beta^2\gamma^2+\alpha^2\beta^2\sigma^2+\alpha^2\beta^2\xi^2+\alpha^2\gamma^2\sigma^2\\ +\alpha^2\gamma^2\xi^2+\alpha^2\sigma^2\xi^2+\beta^2\gamma^2\sigma^2+\beta^2\gamma^2\xi^2+\beta^2\\ d^2\xi^2+\gamma^2\sigma^2\xi^2) + z^{-4}(\alpha^2\beta^2+\alpha^2\gamma^2+\alpha^2\sigma^2\\ +\alpha^2\xi^2+\beta^2\gamma^2+\beta^2\sigma^2+\beta^2\xi^2+\gamma^2\sigma^2+\gamma^2\xi^2\\ +\sigma^2\xi^2) + z^{-2}(\alpha^2+\beta^2+\gamma^2+\sigma^2+\xi^2)+1\end{array}}\right] \quad (19)$$

Finding the transfer function of such 11th order filter means, an intermediate IIR filter has been concluded. The corresponding IIR filter can be implemented on BLWDF bases as a parallel connection of two all-pass IIR sections of the type shown in Fig. 7.

By Gazsi method [18] that uses an alternative pole technique as illustrated in Fig. 10, the transfer functions of the two all-pass sections $A_0(z^2)$ and $A_1(z^2)$ can then be derived.

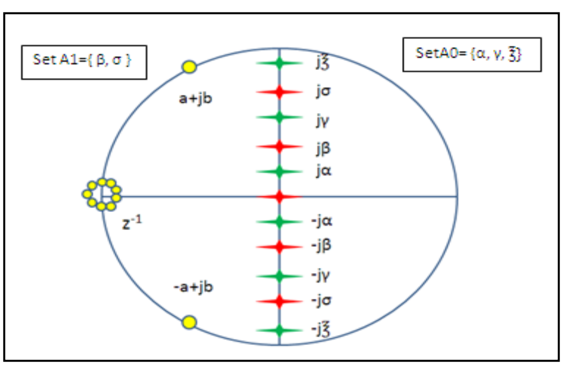


Fig.10. 11th order alternative pole technique.

In Fig. 10, the poles in set A_0can formulate the transfer function of the all-pass section $A_0(z^2)$ and the poles in set A_1 can formulate the transfer function of the all-pass section $A_1(z^2)$. The pole at the origin represents the delay element z^{-1}, then it can be written that:

$$A_0(z^2) = \frac{\alpha^2+z^{-2}}{1+\alpha^2 z^{-2}}\cdot\frac{\gamma^2+z^{-2}}{1+\gamma^2 z^{-2}}\cdot\frac{\xi^2+z^{-2}}{1+\xi^2 z^{-2}} \quad (20)$$

$$A_1(z^2) = \frac{\beta^2+z^{-2}}{1+\beta^2 z^{-2}}\cdot\frac{\sigma^2+z^{-2}}{1+\sigma^2 z^{-2}} \quad (21)$$

Substituting (20) and (21) in (15) to form the low-pass transfer function of BLWDF as:

$$H_0(z) = \frac{1}{2}\left[\frac{\begin{array}{c}(\alpha^2\gamma^2\xi^2+(\alpha^2\xi^2+\gamma^2\xi^2\alpha^2\gamma^2)\\ z^{-2}+(\xi^2+\alpha^2+\gamma^2)z^{-4}+z^{-6})\end{array}}{\begin{array}{c}(1+(\gamma^2+\alpha^2+\xi^2)z^{-2}+(\alpha^2\gamma^2\\ +\gamma^2\xi^2+\alpha^2\xi^2)z^{-4}+\alpha^2\gamma^2\xi^2z^{-6})\end{array}} + \frac{\begin{array}{c}z^{-1}(\beta^2\sigma^2+(\beta^2\\ +\sigma^2)z^{-2}+z^{-4})\end{array}}{\begin{array}{c}(1+(\sigma^2+\beta^2)\\ +\beta^2\sigma^2z^{-4})\end{array}}\right] \quad (22)$$

Equation (22) can further be simplified as follows:

$$H_0(z) = \frac{1}{2}\left[\frac{\begin{array}{c}(1+(\sigma^2+\beta^2)z^{-2}+\beta^2\sigma^2z^{-4})(\alpha^2\gamma^2\xi^2\\ +(\alpha^2\xi^2+\gamma^2\xi^2+\alpha^2\gamma^2)z^{-2}+(\xi^2+\alpha^2+\\ \gamma^2)z^{-4}+z^{-6})+(\beta^2\sigma^2z^{-1}+(\beta^2+\sigma^2)\\ z^{-3}+z^{-5})(1+(\gamma^2+\alpha^2+\xi^2)z^{-2}+(\alpha^2\\ \gamma^2+\gamma^2\xi^2+\alpha^2\xi^2)z^{-4}+\alpha^2\gamma^2\xi^2z^{-6})\end{array}}{\begin{array}{c}(1+(\gamma^2+\alpha^2+\xi^2)z^{-2}+(\alpha^2\gamma^2+\gamma^2\xi^2\\ +\alpha^2\xi^2)z^{-4}+\alpha^2\gamma^2\xi^2z^{-6})(1+(\sigma^2+\beta^2\\ )z^{-2}+\beta^2\sigma^2z^{-4})\end{array}}\right] \quad (23)$$

The final form of (23) can be written as:

$$H_0(z) = \frac{1}{2}\left[\frac{\begin{array}{c}\alpha^2\gamma^2\xi^2+(\beta^2\sigma^2)z^{-1}+(\alpha^2\beta^2\gamma^2\xi^2+\alpha^2\gamma^2\sigma^2\xi^2+\alpha^2\gamma^2\\ +\alpha^2\xi^2+\gamma^2\xi^2)z^{-2}+(\alpha^2\beta^2\sigma^2+\beta^2\gamma^2\sigma^2+\beta^2\sigma^2\xi^2+\beta^2\\ +\sigma^2)z^{-3}+(\alpha^2\beta^2\gamma^2\sigma^2\xi^2+\alpha^2\beta^2\gamma^2+\alpha^2\beta^2\xi^2+\alpha^2\gamma^2\sigma^2\\ +\alpha^2\sigma^2\xi^2+\alpha^2+\beta^2\gamma^2\xi^2+\gamma^2\sigma^2\xi^2+\gamma^2+\xi^2)z^{-4}+(\alpha^2\\ \beta^2\gamma^2\sigma^2+\alpha^2\beta^2\sigma^2\xi^2+\alpha^2\beta^2+\alpha^2\sigma^2+\beta^2\gamma^2\sigma^2\xi^2+\beta^2\gamma^2\\ +\beta^2\xi^2+\gamma^2\sigma^2+\sigma^2\xi^2+1)z^{-5}+(\alpha^2\beta^2\gamma^2\sigma^2+\alpha^2\beta^2\sigma^2\xi^2\\ +\alpha^2\beta^2+\alpha^2\sigma^2+\beta^2\gamma^2\sigma^2\xi^2+\beta^2\gamma^2+\beta^2\xi^2+\gamma^2\sigma^2+\sigma^2\xi^2\\ +1)z^{-6}+(\alpha^2\beta^2\gamma^2\sigma^2\xi^2+\alpha^2\beta^2\gamma^2+\alpha^2\beta^2\xi^2+\alpha^2\gamma^2\sigma^2+\\ \alpha^2\sigma^2\xi^2+\alpha^2+\beta^2\gamma^2\xi^2+\gamma^2\sigma^2\xi^2+\gamma^2+\xi^2)z^{-7}+(\alpha^2\beta^2\sigma^2\\ +\beta^2\gamma^2\sigma^2+\beta^2\sigma^2\xi^2+\beta^2+\sigma^2)z^{-8}+(\alpha^2\beta^2\gamma^2\xi^2+\alpha^2\gamma^2\sigma^2\\ \xi^2+\alpha^2\gamma^2+\alpha^2\xi^2+\gamma^2\xi^2)z^{-9}+\beta^2\sigma^2z^{-10}+\alpha^2\gamma^2\xi^2z^{-11})\end{array}}{\begin{array}{c}1+(\alpha^2+\beta^2+\gamma^2+\sigma^2+\xi^2)z^{-2}+(\alpha^2\beta^2+\alpha^2\gamma^2\\ +\alpha^2\sigma^2+\alpha^2\xi^2+\beta^2\gamma^2+\beta^2\sigma^2+\beta^2\xi^2+\gamma^2\sigma^2+\gamma^2\xi^2\\ +\sigma^2\xi^2)z^{-4}+(\alpha^2\beta^2\gamma^2+\alpha^2\beta^2\sigma^2+\alpha^2\beta^2\xi^2+\alpha^2\gamma^2\sigma^2\\ +\alpha^2\gamma^2\xi^2+\alpha^2\sigma^2\xi^2+\beta^2\gamma^2\sigma^2+\beta^2\gamma^2\xi^2+\beta^2\sigma^2\xi^2+\\ \gamma^2\sigma^2\xi^2)z^{-6}+(\alpha^2\beta^2\gamma^2\sigma^2+\alpha^2\beta^2\gamma^2\xi^2+\alpha^2\beta^2\sigma^2\xi^2+\\ \alpha^2\gamma^2\sigma^2\xi^2+\beta^2\gamma^2\sigma^2\xi^2)z^{-8}+(\alpha^2\beta^2\gamma^2\sigma^2\xi^2)z^{-10}\end{array}}\right] \quad (24)$$

After finding the transfer function of BLWDF, the values of the coefficients α, β, γ, σ, and   can be calculated by equating the numerator of (24) to the numerator of the general filter function (19) and solving the resulting equations numerically using fixed point iteration method [19] to find out that:

α= 0.248772537,
β= 0.388817646,
γ= 0.588999485,
σ= 0.926657288,
ξ= 0.753419430,
a= 0.12,
b= 0.992773891.

Substituting the upper resulting coefficients in (24), the final form of the intermediate 11th order wavelet IIR BLWDF low-pass transfer function can be reduced to

$$H_0(z) = \frac{1}{2}\left[\frac{\begin{array}{c}0.004 + (0.037)z^{-1} + (0.156)z^{-2} + \\ (0.406)z^{-3} + (0.7286)z^{-4} + (0.961) \\ z^{-5} + (0.961)z^{-6}(0.7286)z^{-7} + \\ (0.406)z^{-8} + (\,0.156)z^{-9} + (0.037) \\ z^{-10} + (0.004)z^{-11}\end{array}}{\begin{array}{c}1 + (1.986)z^{-2} + (\,1.369)z^{-4} + \\ (0.395)z^{-6} + (0.045)z^{-8} + \\ (0.0016)z^{-10}\end{array}}\right] \tag{25}$$

The corresponding transfer function of the high-pass filter in (16)can then be given by

$$H_1(z) = \frac{1}{2}\left[\frac{\begin{array}{c}0.004 - (0.037)z^{-1} + (0.156)z^{-2} - \\ (0.406)z^{-3} + (0.728)z^{-4} - (0.961)z^{-5} + \\ (0.961)z^{-6} - (0.728)z^{-7} + (0.406)z^{-8} - \\ (0.156)z^{-9} + (0.037)z^{-10} - (0.004)z^{-11})\end{array}}{\begin{array}{c}1 + (1.986)z^{-2} + (\,1.369)z^{-4} + (0.395) \\ z^{-6} + (0.045)z^{-8} + (0.0016)z^{-10}\end{array}}\right] \tag{26}$$

The magnitude responses of $H0(e^{j\omega})$ and $H1(e^{j\omega})$ in (25) and (26) are shown in Fig. 11.
The Schematic diagrams for the proposed transmitter and receiver for DWMT transmission system with 3-level wavelet decomposition are shown in Fig.12a and b, respectively.

VI. Results and Disscussions

In this section, SNR values for two-channel DWMT transmission systems at the receiving end are calculated after adding different levels of AWGN at the transmission channel. Figure 13 shows the resulting SNR values for the received signals from different DWMT systems with 5th& 7th order IIR filter banks of Ref. [16], 9th order IIR filter bank of *Ref.* [17], the proposed 11th order wavelet IIR intermediate filter bank using BLWDF and those from DWMT system with 11th order FIR filter of *Ref.* [12]. It can be seen that the proposed DWMT system with 11th order IIR filter outperforms all other tested systems for all levels of AWGN.
The autocorrelation functions for the same (5^{th}, 7^{th}, 9^{th}, 11^{th} order IIR and 11^{th} order FIR) wavelet filters are computed to find the best filter for sub-band processing in DWMT transmission systems. The best filter is the filter that have an autocorrelation function which can best-approximate an impulse function. The autocorrelation function of a filter Hi(z) is the inverse Fourier transform of $\{H_i(z)\ H_i(z^{-1})\}$ at $z = e^{j\omega}$. Different computed autocorrelation functions are shown in Fig. 14. From this figure, it can be seen that the two impulse-best-fitting autocorrelation functions are those which belong

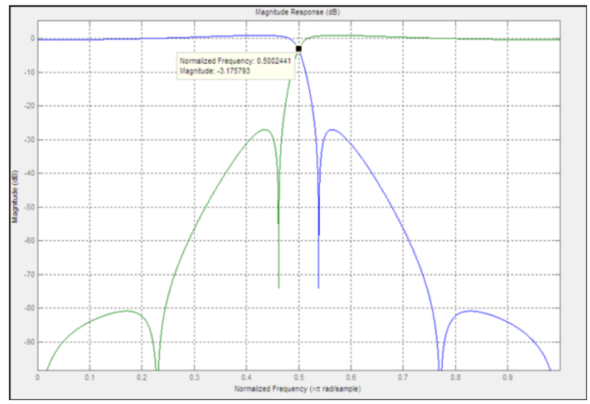
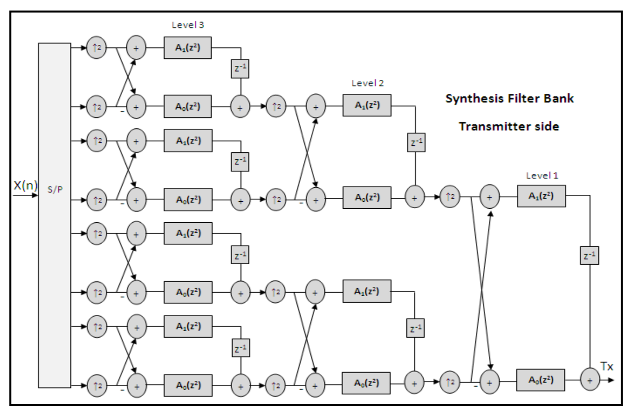
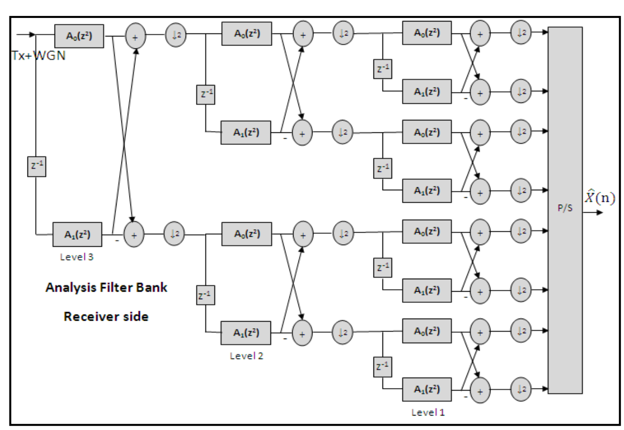


Fig.11. Magnitude responses of 11th order low-pass and high-pass BLWDFs.



(a)



(b)

Fig.12. Block diagram of (a) the proposed DWMT transmitter, (b) the proposed DWMT receiver.

to 11th order IIR wavelet and 11thorder FIR wavelet filters. However, IIR filter is the best because of its lower forward delay for the same FIR filter order as well as its best roll-off characteristic which means minimum ICI.

VII. CONCLUTIONS

A new DWMT transmission system has been designed and implemented using 11th -order IIR wavelet filter banks. It has been shown that the proposed system can outperform the existing and recent systems in its performance. Such performance includes best magnitude response discrimination, maximum SNR values for the received signals and impulse-best-fitting autocorrelation function.

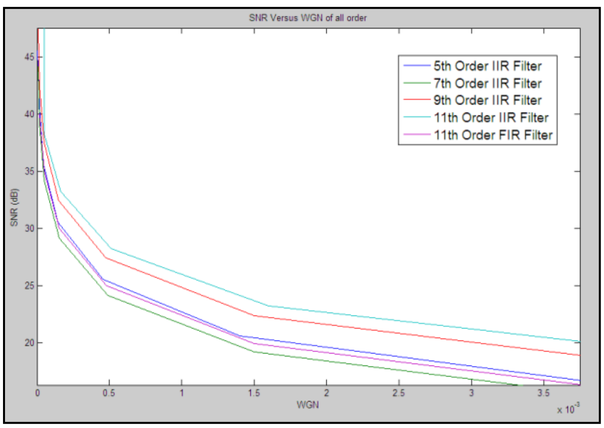


Fig.13. SNR Versus AWGN for 5th, 7th, 9th, 11th FIR and 11th IIR filter order for BLWDFBs.

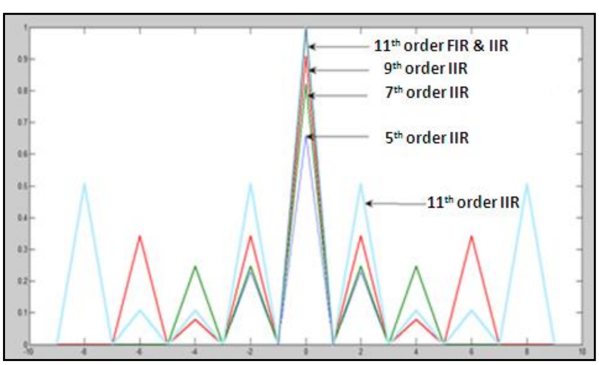


Fig.14. Autocorrelation functions for 5th, 7th, 9th, 11th order IIR -BLWDF and 11th order FIR filters.

REFERENCES

[1] A. A. Khader, "Matlab Simulation Design and performance Comparison of OFDM & FH-OFDM System", Alrafidain Engineering Journal, Collage of Engineering,University of Mosul, Mosul, Iraq, Vol. 20, No. 1, Feb.2012, pp. 38-48.

[2] R. Mohamad, W. M. Hafeez,. W.M. Salleh and N. M. Anas, "Multiband OFDM and OFDM Simulation Software Using MATLAB® Graphical User Interface", Computer Science and Convergence, Lecture Notes in Electrical Engineering, Vol. 114, 2012, pp 305-313.

[3] J. N. Patel, U. D. .Dalal "A Comparative Performance Analysis of OFDM using MATLAB Simulation with M-PSK and M-QAM Mapping", Proceeding Proceedings of the International Conference on Computational Intelligence and Multimedia Applications (ICCIMA 2007) – Vol. 04, pp. 406-410.

[4] S. D. Sandberg, M. A. Tzannes, "Overlapped Discrete Multitone Modulation for High Speed Copper Wire Communications", IEEE Journal on Selected Areas in Communications, Vol. 13, No. 9, Dec 1995, pp. 1571 – 1585.

[5] E. Lavallk, J. Bcauvais and D. Drouin, "Time Domain Equaliser Design for DWMT Multicarrier Transceivers", Electronics Letters, Vol. 36, No. 18, Aug. 2000, pp. 1590-1592.

[6] M. C. Sun and D. P.K. Lun, "Power-Line Communications using DWMT Modulation", Proceedings of IEEE International Symposium on Circuits and Systems, 2002 (ISCAS 2002), Vol.4, pp. IV-493 - IV-496.

[7] B. Farhang-Boroujeny,"Discrete Multitone Modulation with Blind Detection Capability", Proceedings of Vehicular Technology Conference, 2002 (VTC 2002), Vol. 1, pp. 376-380.

[8] B. Farhang-Boroujeny and L. Lin, "Cosine-Modulated Multitonefor Very-High-Speed Digital Subscriber Lines," Proceedings of IEEE International Conference on Acoustics, Speech, and Signal Processing, 2005 (ICASSP '05), 18-23 March 2005, Vol. 3, pp. iii/345 - iii/348.

[9] S. Baig and M. Junaid Mughal, "A Frequency Domain Equalizer in DiscreteWavelet Packet Multitone Transceiver forIn-Home PLC Lans", IEEE International Symposium on Power Line Communications and Its Applications, 2007 (ISPLC '07), pp. 181-186.

[10] K. Chen and T. Chiueh, "A Cognitive Radio System using DiscreteWavelet Multitone Modulation",IEEE Transactions onCircuits and Systems I, Vol. 55, No. 10, Nov. 2008,pp. 3246-3258.

[11] L. Zbydniewski, T. Zieliński, "Performance of Wavelet-OFDM and Circular Wavelet-OFDM in Power Line Communications",proceedings of Conference on Signal Processing Algorithms, Architectures, Arrangements, and Applications (SPA 2008), 25-27 Sept. 2008, pp. 115-120.

[12] A. I. Siddiq, A. A. Kadhim, "Design of Wavelet Filters for DWMT Systems",2012 IEEE International Conference on Future Communication Networks (ICFCN'12), Al-Nahrain University- Baghdad,9-12 April 2012,pp. 7 - 11.

[13] H. Ohlsson, "Studies on Implementation of Digital Filters with High Throughput and Low Power

Consumption",Thesis No. 1031, Linköping Studies in Science and Technology,Sweden,June 2003.

[14] P. A. Regalia, S. K. Mitra and P. P. Vadianathan, "The Digital All-Pass Filter: A Versatile Signal Processing Building Block", Proceedings of IEEE, Vol. 76, No. 1, Jan. 1988, pp. 19-37.

[15] H. Johansson and L. Wanhammar, "Design of Bireciprocal Linearphase Lattice Wave Digital Filters", Report LiTH-ISY-R-1877, Linköping University, Sweden.

[16] J. M. Abdul-Jabbar and S. N. M. Al-Faydi, "Design and Realization of Bireciprocal Lattice Wave Discrete Wavelet Filter Banks", Alrafidain Engineering Journal, Collage of Engineering, University of Mosul, Mosul, Iraq, Vol. 20, No. 1, Feb.2012, pp. 38-48.

[17] J. M. Abdul-Jabbar and R.W. Hmad,"Allpass-Based Design, Multiplierless Realization and Implementation of IIR Wavelet Filter Banks with Approximate Linear Phase", The Fourth IEEE International Symposium on Innovation in Information & Communication Technology (ISIICT 2011), 29 Nov.- 1 Dec. 2011, Amman, Jordan, pp. 118 - 123.

[18] L. Gazsi, "Explicit Formulas for Lattice Wave Digital Filters", IEEE Transaction on Circuits and Systems, Vol. CAS-32, No. 1, 1985, pp. 68-88.

[19] S. Salleh, A. Y. Zomaya, and S. A. Bakar, "*Computing for Numerical Methods Using Visual C++‖*", John Wiley &Sons, Canada, ISBN 978-0-470-12795-7, 2007.

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