A Novel method of Walsh-Hadamard Code Generation using Reconfigurable Lattice filter and its application in DS-CDMA system

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Abstract: - Walsh Hadamard codes are widely used as signature codes in the current wireless standards such as IS-95 CDMA, WCDMA, CDMA2000, and image transform applications. It is also used in conjunction with error correction algorithms for Reed-Muller codes and in dyadic invariant signal processing. In this paper, a novel method of generating Walsh Hadamard(WH) codes using lattice filter is proposed. The obtained results concur with the properties of WH code. The entire WH code set can be generated by changing the reflection coefficients without modifying the filter structure. This feature ensures reconfigurability and enables them to be used in software radio. Moreover, this reconfigurable structure can be used as a matched filter for multiuser detection in Direct Sequence Code division Multiple Access System (DS-CDMA). BER performance of a DS-CDMA system employing a correlator receiver and the proposed matched filter receiver for despreading is simulated in an AWGN channel and flat fading Rayleigh channel. Simulation results showing the generation of WH codes from the autoregressive model of first order Gauss-Markov process is also obtained.

Key-Words: - Software defined radio, Walsh Hadamard codes, Lattice filter, Matched filter, Autoregressive model, Gauss-Markov process.

1 Introduction

A reconfigurable transceiver should support multiple air interfaces such as CDMA, OFDM, MC-CDMA with dynamic reconfigurability in a single terminal [4]. This is achieved in Software defined Radio (SWR) by a change in the software without any change in the hardware element. As mentioned in [2], the size of the software to be downloaded should be small to reduce the reconfiguration time.
This can be optimised by parameterisation method, in which common aspects of different communication standards can be supported by the reconfigurable transceiver, which is identified and a common processing procedure is installed in the device. The two approaches in parameterisation method are Common Function technique (CF) and Common operator approach (CO). In [1], CO technique applied to the two operators namely Fast Fourier Transform (FFT) and Reconfigurable Linear Shift Register(R-LFSR) in digital domain is discussed. In CO technique, common elements with similar structural aspects that can be described by a general equation and independent of typical standards are identified. From the design and implementation point of view these selected CO functions should be reusable and reconfigurable so that they can be applied to any standards. In this paper, lattice filter module is used as a common element in realising WH code generator, Matched filter and Gauss Markov model representing the slowly varying signals. Generation of WH codes from the lattice filter structure of Gauss Markov model and the Bit Error Rate (BER) performance of the matched filter which is realised by the proposed lattice filter structure in a direct sequence code division multiple access system (DS-CDMA), is also discussed.

In CDMA systems, all the users transmit at the same frequency where individual users are identified by assigning distinct signature codes to them. Spreading codes used in CDMA systems include Walsh-Hadamard sequences, Orthogonal Variable Spreading Factor codes, Gold codes, Kasami codes, m-sequences etc. [6, 7]. Walsh Hadamard codes are widely used as signature codes for downlink transmission [13]. In general, these codes are generated using linear feedback shift register (LFSR). LFSR can be realised by a tapped delay line or finite impulse response (FIR) filter structure with binary inputs [5]. In the proposed method, the LFSR used to generate WH codes is replaced by the autoregressive (AR) or infinite impulse response (IIR) model realised using the lattice filter (LF) structure. Lattice filters possess several important properties such as a modular structure, low sensitivity to parameter quantization effects, and a simple method to ensure filter stability. Depending on the values of the reflection coefficients, the LF can be used to generate different WH code sequences. The required reflection coefficients can be stored in a look up table or it can be software downloaded. Accuracy can be increased by using higher order filters with less computational complexity since the modular structure of LF allows the filter order to be increased or decreased without having to recompute the reflection coefficients.

In section 2, relation between LFSR and AR modelling using Levinson Durbin algorithm is discussed. Section 3 describes the generation of WH codes using LF. In section 4, application of the proposed LF structure as a matched filter in DS-CDMA system is described and simulation results showing the BER performance of a DS-CDMA system using the conventional and matched filter method of despreading is presented. In section 5, relation between the AR model of WH codes and first order Gauss Markov process is discussed.

2 Relation between Linear Feedback Shift Register and Autoregressive (AR) modelling

In general, maximal length sequences like pseudo noise sequences are produced by using a LFSR [7]. For example, the block diagram to generate a pseudo noise sequence of length 15 using an LFSR of length 4 is shown in Fig. 1. In Fig.1, 'D' and '+' represents a delay element (z^{-1}) and exclusive OR operation respectively.

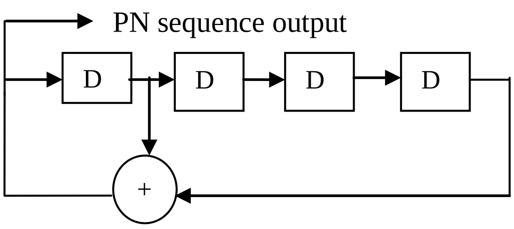


Fig.1 Block diagram of a binary LFSR

The problem of generating a given sequence $[c_0, c_1 \ldots c_{N-1}]$, of length 'N', using an LFSR is stated as follows:

- The feedback connection polynomial g(D) has to be determined.
- The initial register contents of the shortest LFSR that could produce the sequence c(D) has to be determined.

One method that provides a solution to this problem is Massey's algorithm [7]. This involves solving a set of system equations of the form shown in equation (1). Assuming a LFSR of length 4, the set of equations are given by,

$$\begin{bmatrix} c_3 & c_2 & c_1 & c_0 \\ c_4 & c_3 & c_2 & c_1 \\ c_5 & c_4 & c_3 & c_2 \\ c_6 & c_5 & c_4 & c_3 \end{bmatrix} \begin{bmatrix} g_1 \\ g_2 \\ g_3 \\ g_4 \end{bmatrix} = \begin{bmatrix} -c_4 \\ -c_5 \\ -c_6 \\ -c_7 \end{bmatrix} \qquad (1)$$

From equation (1), it can be inferred that the matrix on the left is a Toeplitz matrix, and these equations represent the Yule-Walker equations of an AR process in signal modelling. This implies that it may be possible to generate the maximal length sequences from the impulse response of an AR deterministic signal model [2]. In stochastic signal modelling, these equations can be solved by using the Levinson-Durbin algorithm. Levinson-Durbin recursion algorithm [6] is widely used to solve the Prony normal equations and autocorrelation normal equations. The autocorrelation method, a modified form of Prony's normal equations, is applied to a fixed length sequence (windowed sequence). The autocorrelation normal equations in matrix form for a real process **X** is a set of (p+1) linear equations in (p+1) unknowns a(1), a(2), …, a(p) and ε_{p+1} given by,

$$\begin{bmatrix} r(0) & r(1) & \cdots & r(p-1) & r(p) \\ r(1) & r(0) & \cdots & r(p-2) & r(p-1) \\ \vdots & \vdots & & \vdots & \vdots \\ r(p-1) & r(p-2) & \cdots & r(0) & r(p-2) \\ r(p) & r(p-1) & \cdots & r(1) & r(0) \end{bmatrix} \begin{bmatrix} 1 \\ a(1) \\ a(p) \end{bmatrix} = \varepsilon_{p+1} \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix} \qquad (2)$$

In matrix notation, eqn.2 can be written as,

$$\mathbf{R_{p+1} a_{p+1} = \varepsilon_{p+1} U_{p+1}} \qquad (3)$$

The autocorrelation sequence r(k) is given by,

$$r(k) = \sum_{n=k}^{N} x(n)x(n-k);\ k \geq 0 \qquad (4)$$

'N' represents the length of the sequence x(n) and the minimum modelling error ε_{p+1} is given by,

$$\varepsilon_{p+1} = r(0) + \sum_{k=1}^{k=p} a(k) r(k) \qquad (5)$$

Eqn.2 can be solved by using the Levinson Durbin Algorithm. It represents an all-pole model of the form given by,

$$H(z) = \frac{b_0}{1 + \sum_{k=1}^{p} a_p(k) z^{-k}} \qquad (6)$$

Eqn. (6) can be implemented using the direct form-II digital filter structure. The reflection coefficients, one of the by-products obtained by solving the autocorrelation normal equations can be used to implement an all-pole IIR lattice filter.

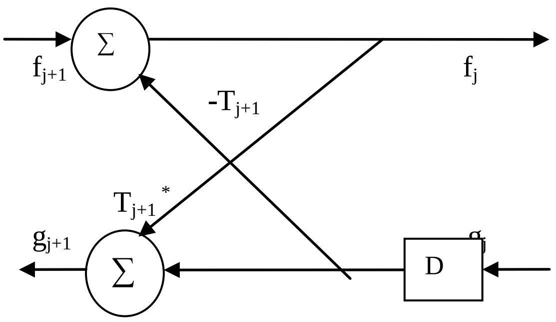


Fig.2a

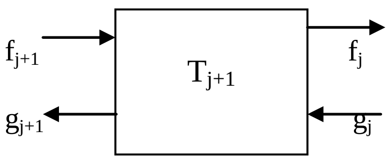


Fig.2b

Fig.2a, 2b Single stage of an all pole lattice filter

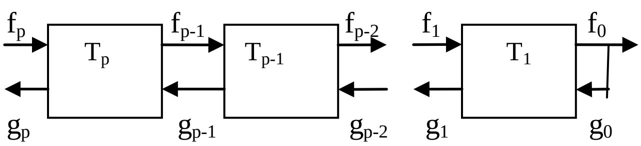


Fig.3 An order p all pole lattice filter

As discussed in [12], the poles and zeros of the transfer function in direct form structures are very sensitive to quantisation effects since these structures are implemented by using the quantised numerator and denominator coefficients directly. This drawback of sensitivity becomes worse as the number of crowded poles increases and its effect can be reduced by using cascade or parallel form of filter structures. However filters realised with lattice structures possess good numerical properties since it can be realised as a cascade of regular modules. Fig.2 and Fig.3 shows the single stage and p-th order all pole lattice filter respectively where 'T_p' is referred as p^{th} reflection coefficient and f, g stands for all pole and all-pass response respectively. This modular structure makes it more suitable for VLSI implementation. In this paper, generation of WH codes using LF obtained by solving the autocorrelation normal equations is discussed since the significance of these reflection coefficients is that they are all bounded by one in magnitude resulting in a stable all-pole model.

3 GENERATION OF WALSH-HADAMARD CODES USING THE RECONFIGURABLE LATTICE FILTER (RLF)

In this section, the proposed method of generation of WH codes from the impulse response of all-pole model realized using LF structure is described. A numerical example showing the generation of WH codes of length 8 is also discussed.

3.1 AR modelling of WH code generation

WH code generation using AR modelling is shown in Fig.4.

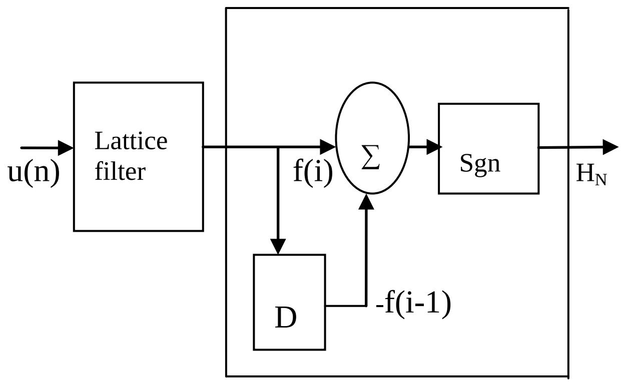


Fig.4 Walsh Hadamard code generation from all-pole response of Lattice filter

The algorithm to obtain the AR model of WH code is as follows:

a) The autocorrelation values for a Walsh Hadamard code sequence of length 'N' are calculated.

b) Levinson Durbin algorithm is applied to these autocorrelation values and the reflection coefficients T_1, T_2, $T_{N/2}$ are obtained.

c) An IIR all pole lattice filter is implemented using these reflection coefficients. To get highly precise results, the model order can be increased from (N/2) to (N/2)+1, …,N.

d) The all pole response $\{f_k : k=0,1,2,\ldots,(N-1)\}$ of this filter is obtained by giving a unit step sequence as input.

e) Since the first bit of all the Walsh Hadamard code sequences start with one, initialise the first bit of the code as H(1,1)=1. The remaining bits H(1,i), i=2,3,…,N are then given by,

$H_N(1,i)=\text{signum}\{[f(i)-f(i-1)]\};i=2,3,\ldots,N$

3.2 Numerical Example

As an example, consider a Walsh Hadamard code set of size and length 8 in which each row represents a code sequence.

$$\mathbf{H_8} = \begin{bmatrix} 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\ 1 & -1 & 1 & -1 & 1 & -1 & 1 & -1 \\ 1 & 1 & -1 & -1 & 1 & 1 & -1 & -1 \\ 1 & -1 & -1 & 1 & 1 & -1 & -1 & 1 \\ 1 & 1 & 1 & 1 & -1 & -1 & -1 & -1 \\ 1 & -1 & 1 & -1 & -1 & 1 & -1 & 1 \\ 1 & 1 & -1 & -1 & -1 & -1 & 1 & 1 \\ 1 & -1 & -1 & 1 & -1 & 1 & 1 & -1 \end{bmatrix} \quad (7)$$

Each row in the matrix H_8 represents a code sequence. The following pattern can be observed from the code sequences. The code sequence 2 can be generated from code sequence 1 by changing signs at bit positions 2,4,6,8. In the same manner, code sequences 4, 6, 8 can be obtained from code sequences 3, 5, 7 respectively. Table.1 and Table.2 shows the reflection coefficients for a Walsh Hadamard code of length 8 and 16 respectively generated by the proposed algorithm.

Table.1 Reflection Coefficients of Walsh-Hadamard code of length 8

Code 1	**Code 2**	**Code 3**	**Code 4**
-0.8750	0.8750	-0.1250	0.1250
0.0667	0.0667	0.7778	0.7778
0.0714	-0.0714	-0.3571	0.3571
0.0769	0.0769	0.3684	0.3684
Code 5	**Code 6**	**Code 7**	**Code 8**
-0.6250	0.6250	-0.3750	0.3750
0.2308	0.2308	0.4545	0.4545
0.3000	-0.3000	0.1000	-0.1000
0.4286	0.4286	0.5556	0.5556

The following observations can be noted from Table.1 and Table.2.

a) It can be observed that the numerical values of reflection coefficients are same for codes 1 and 2, 3 and 4, 5 and 6, 7 and 8, 9 and 10, 11 and 12, 13 and 14, 15 and 16 but only with a sign change in the odd bit positions. Hence it can be inferred that two different code sequences can be generated from the same set of reflection coefficients by changing the sign in odd positions which means that the code generation architecture is reconfigurable.

b) To generate a code of length 8, the number of reflection coefficients to be stored in memory is only 4. Hence to generate a code of length N, the number of memory elements needed is only N/2. This implies that the memory requirements are less.

c) The AR model order has to be increased to 2N to generate code sequences of length 4N. This can be accomplished by adding extra lattice filter modules and using the corresponding set of reflection coefficients.

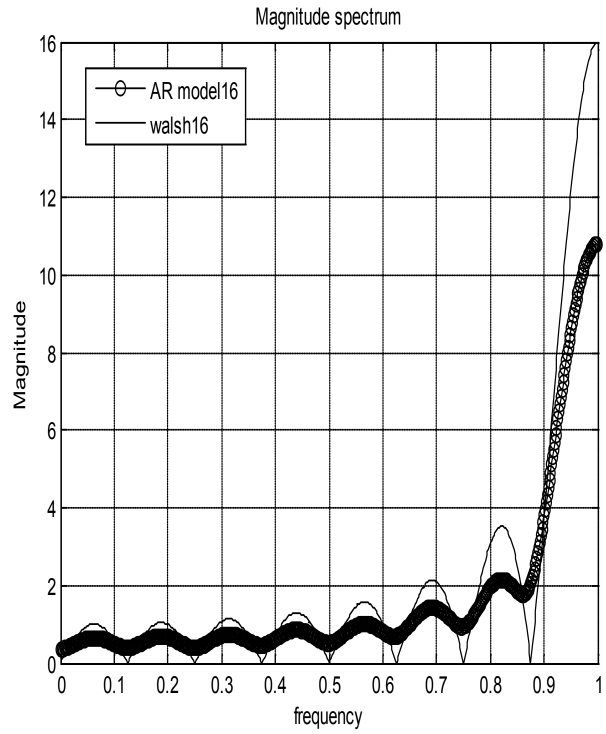
Fig.5 shows the magnitude spectrum for a sample code sequence in 8, 16 length WH code sets and their respective impulse responses from the WH-AR model. It is observed that the AR model and the WH code have nearly the same magnitude spectrum distribution. Hence the proposed AR model can be used to generate the WH codes.

Table.3 shows the all pole response, the first order difference of the all pole response and the WH code generation from this first order difference for a sample WH code sequence of length 16 by using the algorithm proposed in section 3.1. It can be observed that the WH code sequence generated by the proposed method is the same as the original WH code.

Table: 2 Reflection Coefficients of Walsh-Hadamard code of length 16

Code 1	Code 2	Code 3	Code 4
-0.9375	0.9375	-0.0625	0.0625
0.0323	0.0323	0.8824	0.8824
0.0333	-0.0333	-0.4333	0.4333
0.0345	0.0345	0.3488	0.3488
0.0357	-0.0357	-0.3116	0.3116
0.0370	0.0370	0.1831	0.1831
0.0385	-0.0385	-0.0897	0.0897
0.0400	0.0400	0.1529	0.1529
Code 5	**Code 6**	**Code 7**	**Code 8**
-0.5625	0.5625	-0.4375	0.4375
0.2800	0.2800	0.3913	0.3913
0.3889	-0.3889	0.2778	-0.2778
0.6364	0.6364	0.6923	0.6923
-0.6071	0.6071	-0.1981	0.1981
0.1556	0.1556	0.4634	0.4634
0.1842	-0.1842	-0.0649	0.0649
0.2258	0.2258	0.4069	0.4069

Code 9	Code 10	Code 11	Code 12
-0.8125	0.8125	-0.1875	0.1875
0.1034	0.1034	0.6842	0.6842
0.1154	-0.1154	-0.2692	0.2692
0.1304	0.1304	0.3939	0.3939
0.1500	-0.1500	-0.4978	0.4978
0.1765	0.1765	0.1321	0.1321
0.2143	-0.2143	0.2619	-0.2619
0.2727	0.2727	0.2258	0.2258
Code 13	**Code 14**	**Code 15**	**Code 16**
-0.6875	0.6875	-0.3125	0.3125
0.1852	0.1852	0.5238	0.5238
0.2273	-0.2273	-0.0455	0.0455
0.2941	0.2941	0.4783	0.4783
-0.1500	0.1500	-0.3912	0.3912
0.2174	0.2174	0.2093	0.2093
0.2778	-0.2778	0.2863	-0.2863
0.3846	0.3846	0.3772	0.3772



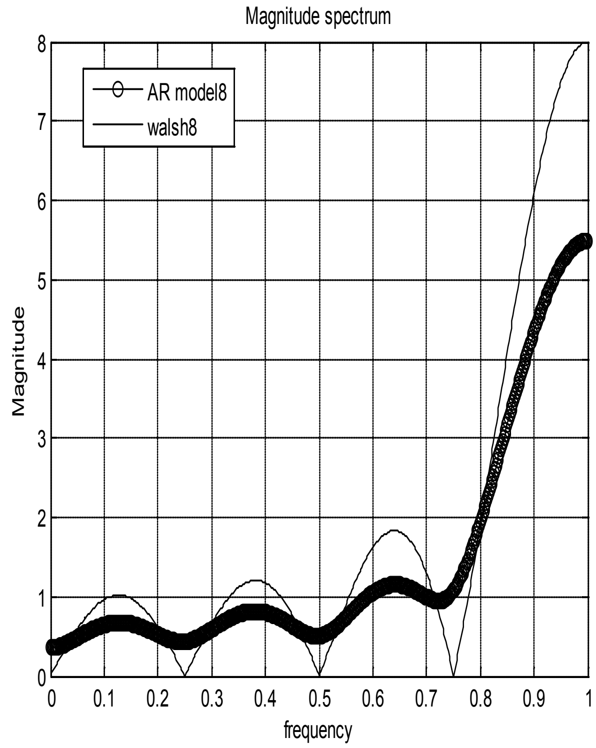


Fig.5 Magnitude spectrum of a sample sequence of WH-code set of length 8, 16 and their respective AR models

Table.3 Proposed method of WH code generation for a sample sequence of length 16

Reflection Coefficients	f	difof	Signum (difof)	WH code
0.3125	1		1	1
0.5238	0.3772	-0.6228	-1	-1
0.0455	-0.1345	-0.5117	-1	-1
0.4783	0.1802	0.3147	1	1
0.3912	-0.1223	-0.3025	-1	-1
0.2093	0.2284	0.3506	1	1
-0.2863	0.5083	0.2799	1	1
0.3772	0.3256	-0.1827	-1	-1
	-0.0809	-0.4065	-1	-1
	0.1096	0.1905	1	1
	0.3517	0.2421	1	1
	0.0747	-0.277	-1	-1
	0.3979	0.3233	1	1
	0.2331	-0.1649	-1	-1
	-0.0757	-0.3088	-1	-1
	0.2391	0.3148	1	1

4 RLF as a Matched filter in DS-CDMA system

In communication systems, binary messages transmitted across a noisy channel are detected by using a matched filter at the receiving end [6], [7]. Matched filter is an optimum filter that maximises the signal to noise ratio of a signal in the presence of additive white Gaussian noise. A matched filter operates by correlating a known signal with an unknown signal and detects the presence of the known signal pattern. This is equivalent to convolving the incoming unknown signal with the known signal pattern.

In the traditional method of direct sequence spread spectrum technique, spreading is accomplished by multiplying the information bearing signal b(t) with the spreading codes at the transmitting end. At the receiving end, the original message is recovered by correlating the received spreaded signals with a replica of the signature sequence used at the transmitting end. This process is called as despreading[8],[9]. In CDMA systems, despreading is achieved by sliding correlators, surface acoustic wave (SAW) matched filters and digital matched filters [7]. As mentioned in [3], the matched filter used in CDMA systems is tuned to match a code sequence. It is in general realized using a tapped delay line filter structure i.e, FIR structure or moving average model where the input data pattern is represented by digital words and the spreading code pattern of +1 or -1 are stored as the multiplication coefficients. As discussed in section 3, WH code generator can also be realized using the AR model. In this paper, despreading is done at the receiving end by the AR model of WH code generator realized by a lattice filter instead of the digital matched filter.

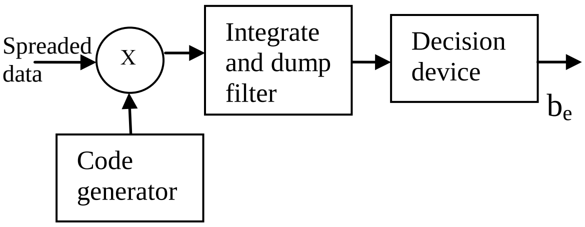


Fig.6 Conventional method of despreading

Fig.6 shows the conventional method of despreading. DS-CDMA system employing the proposed reconfigurable matched LF is shown in Fig.7 in appendix.

The modified algorithm of despreading is discussed below:

a) Multiplier and code generator are replaced by the AR model WH (z) of the WH code.

b) The received signal after demodulation is converted into NRZ polar format represented as R(z). It is then passed through the AR lattice filter with transfer function WH (z) and the filter output response f(n) is obtained which is given by,

$$f(n) = F^{-1}[R(z)WH(z)] \quad \text{----(7)}$$

Where F^{-1} indicates inverse Fourier transform.

c) This output f(n) is then passed through an integrator that integrates the lattice filter response over a period equal to the length of the spreading code and outputs one value during each period.

d) Output of integrator is then passed through a decision device and an estimate $b_e(n)$ of the transmitted data is obtained. It is given by,

$$b_e = \operatorname{sgn}\left[\sum_{k=1}^{N} f(k)\right] \qquad (8)$$

where N is the length of the spreading code.

e) The proposed MF receiver can be reconfigured to match the users spreading codes by using their respective reflection coefficients stored in a look-up table.

4.1 Simulation Results

In this section, BER performance of a DS-CDMA system employing conventional method and the proposed MF method of despreading is investigated. WH codes are used as signature codes to distinguish different users and spreading at the transmitting end is accomplished using the traditional approach. BPSK modulation is used. The signal at the receiving end is given by,

$$rx = \sum_{i=1}^{K} H_i s_i + W \qquad (9)$$

i=1,2,…,K represents the number of users,

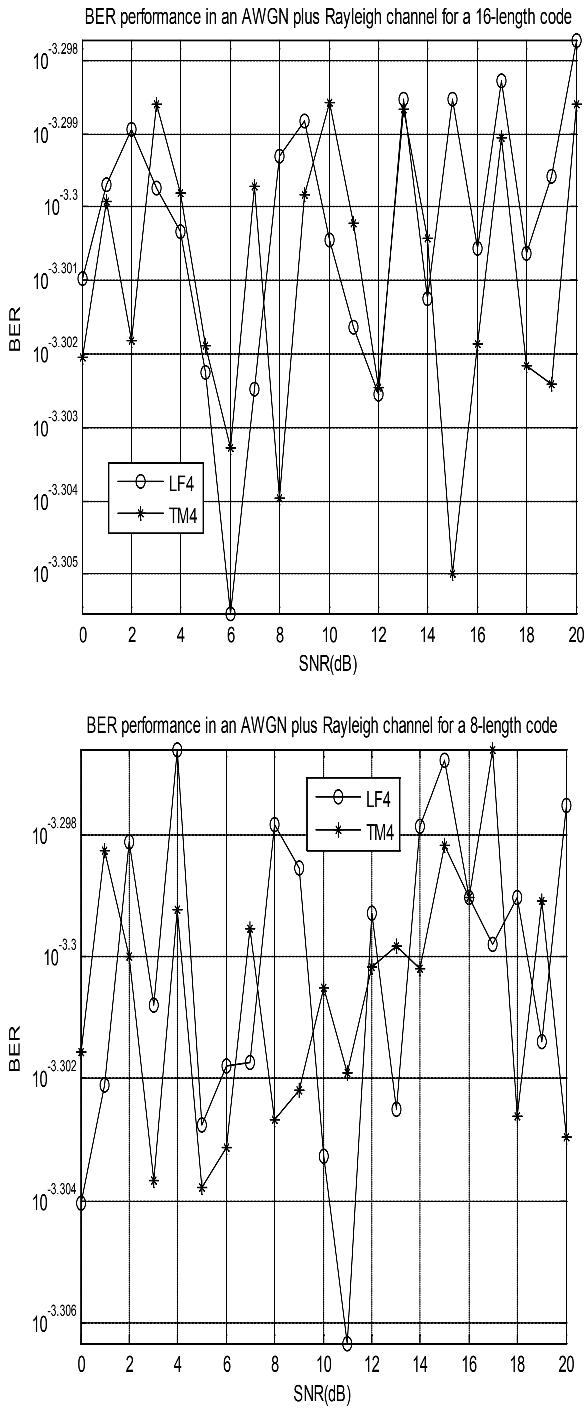
s_i is the transmitted spreaded signal of user i.

W is the additive white Gaussian noise,

H_i represents the flat fading Rayleigh channel of user i.

Fig.8 shows the BER performances in a DS-CDMA system employing conventional method and proposed matched filter method of despreading for 8 and 16 lengths Walsh codes respectively in an AWGN plus Rayleigh flat fading channel with four user scenario. It can be observed that the output of the proposed model and conventional method are marginally the same.

Fig.9 shows the multi-user BER performances employing conventional method and proposed matched filter method of despreading of 16-length Walsh codes in an AWGN plus Rayleigh flat fading channel for SNR=6dB and SNR=12dB as a function of number of users in a DS-CDMA system.



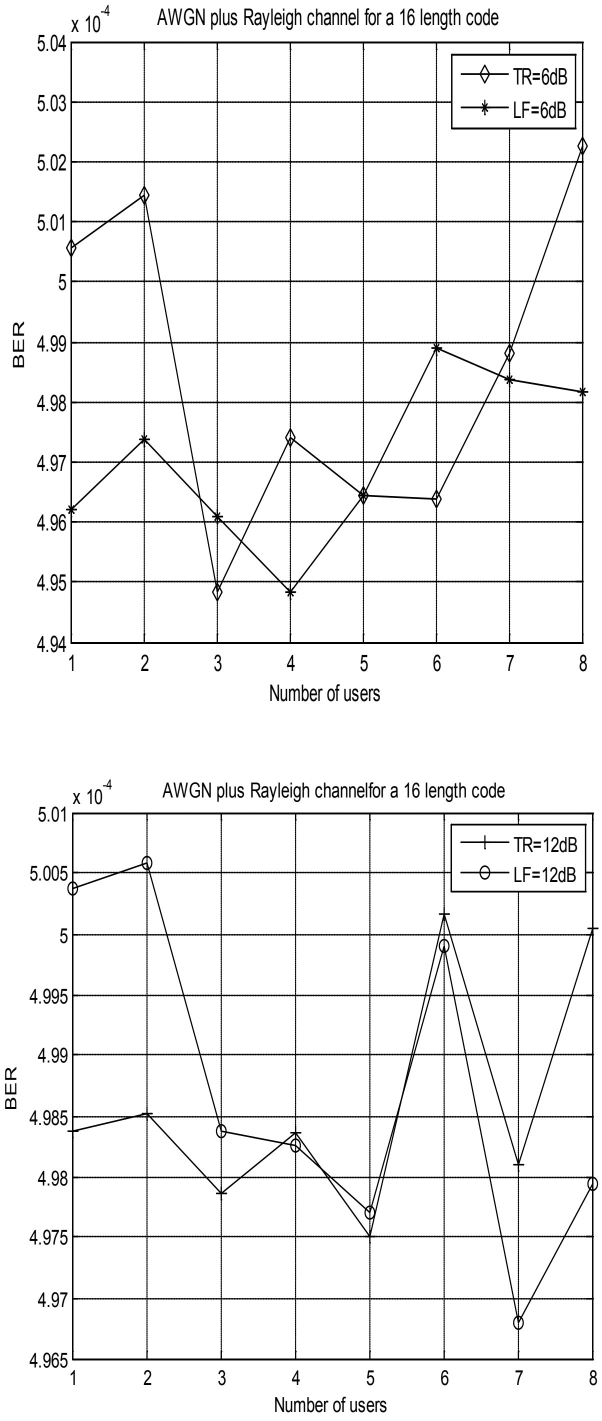


Fig.8 BER performances employing conventional method and matched filter AR signal model of 8, 16-length Walsh codes in an AWGN plus Rayleigh flat fading channel with four user scenario in a DS-CDMA system.

Fig.9 Multi-user performances employing conventional method and matched filter AR signal model of 16-length Walsh codes in a DS-CDMA system.

Simulation results show that the BER performance of the AR lattice model is slightly better than the conventional method of despreading. This slight difference is due to round off noise effects that occur in lattice filter structures. Effect of round off noise can be reduced if the proposed LF structure is replaced by scaled normalised LF based on Schur algorithm. These filter structures provide the added advantages of pipelining, parallel processing, Givens rotation implementation using CORDIC algorithm and low power VLSI implementation.

5 Relation between AR model of WH code and Gauss-Markov process

In transform coding, slowly varying signals are represented by Gauss Markov model [11]. In this type of modelling, if the distribution of observation of the current random variable is dependent only on one previous observation, then it represents an autoregressive process (AR) signal model. For a wide sense stationary input random vector $\mathbf{X} = (X_1, X_2, \ldots, X_n)^T$, of unit variance, the covariance matrix for this single step correlation is given by,

$$\mathbf{K_X} = \begin{bmatrix} 1 & \rho & \rho^2 & \cdots & \rho^{n-1} \\ \rho & 1 & \rho & & \rho^{n-2} \\ \vdots & & & \ddots & \vdots \\ \rho^{n-1} & \rho^{n-2} & \rho^{n-3} & \cdots & 1 \end{bmatrix} \quad (10)$$

where ρ represents the correlation coefficient.

Principal Component analysis of this covariance matrix (i.e.,) Karhunen-Loeve transform gives the eigen values and their corresponding eigen vectors respectively. These eigen vectors are in general arranged in the order of increasing eigen values. The resulting eigen vectors represent unitary spreading codes. In [10], KLT bases generated for low values of correlation coefficient from this covariance matrix can be used in the design of spreading codes. It has been shown that these varying power spreading codes, when applied in DS-CDMA communications, performs comparable to or marginally better than the widely used Gold codes and outperforms Walsh codes in AWGN and Rayleigh flat fading channel conditions.

By applying the same procedure as discussed in Section 4 to these eigen vectors, a set of binary spreading codes can be generated. In this paper, the autocorrelation values of these eigen vectors are calculated and used to find the reflection coefficients of the lattice filter structure. It is observed that the binary code set generated from the KLT bases of the Gauss Markov signal model realised by the proposed lattice filter structure are nearly the same as the WH codes.

5.1 Numerical Example

Table.4 gives the reflection coefficients generated for a Gauss Markov process with correlation coefficient ρ=0.91. Table.5 shows the decimal equivalent of the WH code set of size 8, and KLT codes generated from a covariance matrix of size 8 by 8 for the correlation coefficients taking the values of ρ=0.2 and ρ=0.91.

Table.4 Reflection Coefficients of KLT binary code of length 8

Code 1	**Code 8**	**Code 4**	**Code 5**
-0.8925	-0.6992	-0.4871	-0.2001
0.1339	0.3616	0.5773	0.7092
0.1162	0.2522	0.3036	0.321
0.1004	0.1647	0.1482	0.2234
Code 2	**Code 7**	**Code 3**	**Code 6**
0.1332	0.4653	0.7464	0.9341
0.7814	0.8212	0.8432	0.8544
0.3586	0.4296	0.5158	0.5857
0.3691	0.4555	0.3948	0.2278

The order of KLT code sequence in the table corresponds to the code generated from the eigen vector that corresponds to the ascending eigen value. It can be noted from the table that only one code sequence is different between the WH codes and the KLT codes. It can be observed that by increasing the values of the correlation coefficient from 0.2 to 0.91, the number of bits in which they differ from the original Walsh Hadamard code set is only 1 or 2 bits. Hence the lattice filter structure representing an AR Gauss Markov process can be used to generate Walsh-Hadamard codes. Since the lattice filter transfer function represents a Gauss Markov process, this lattice filter structure can be interpreted as matched with a slowly varying transmitted signal or slowly varying channels. Thus this structure can be used as a matched filter receiver or as an equaliser at the receiving end.

Table.5 Decimal values of Walsh Hadamard Codes and KLT codes of length 8

Walsh-Hadamard code set	**Proposed KLT code set for ρ=0.2**	**Proposed KLT code set for ρ=0.91**
255	255	255
170	240	240
204	195	195
153	204	206
240	153	153
165	**178**	**178**
195	165	165
150	170	170

4 Conclusion

In software radio (SWR), the main aspect is to use flexible components suitable for all types of communication systems. The main significance of the proposed Walsh Hadamard Code generation is that a new code can be generated by just changing the reflection coefficients. The proposed lattice filter AR signal model can also be used as a matched filter for multiuser detection in a DS-CDMA system. Simulation results show that the BER performances of conventional method and matched filter method of despreading are marginally the same. The AR lattice filter representing the Gauss Markov model can be used to generate the Walsh-Hadamard codes when the reflection coefficients are chosen according to the signature code of the particular user. In general, interference and channel noise effects in wireless communication receivers are reduced by using equalisers implemented using filter structures. Reconfiguring the LF to act as an equaliser for other wireless standards such as OFDM, MC-CDMA etc. and integrating the equaliser and spreading operations in the LF structure by optimising the reflection coefficients at the receiving end in a DS-CDMA system is a further extension of this work. This work can be extended to implement the generation of the various spread spectrum codes such as Gold code, Kasami code and multilevel integer codes using LF.

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APPENDIX

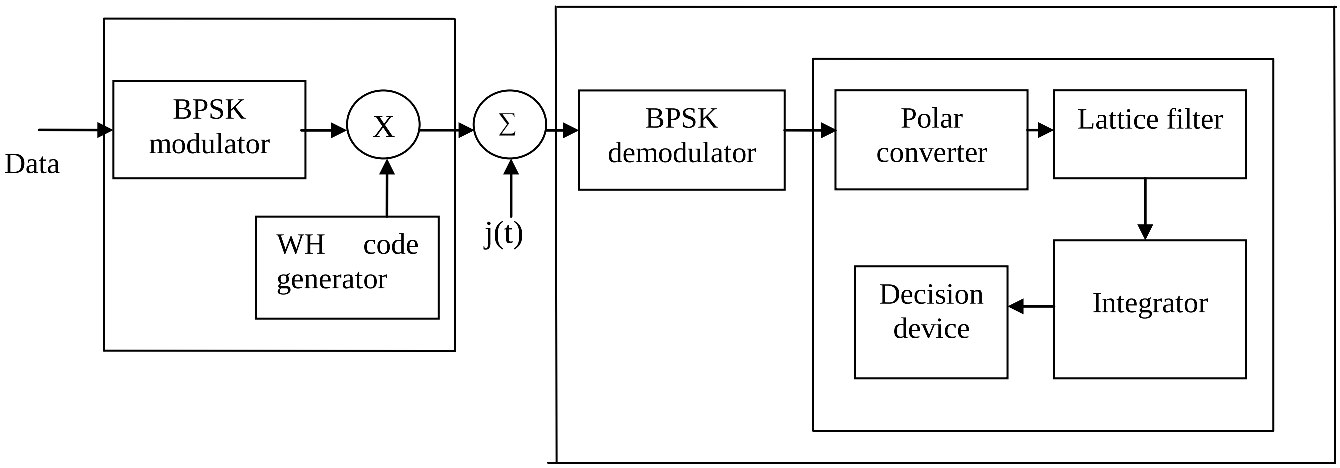

Fig.7 Proposed DS-CDMA system