GENERALIZED LAPPED TRANSFORM (GLT) BASED HIGH-SPEED TRANSMISSION FOR WIRELESS MOBILE COMMUNICATIONS

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Abstract

In this paper, we describe a Generalized Lapped Transform (GLT) based high-speed transmission technique for wireless mobile communications over Rayleigh fading channel. In this technique, the high-rate data bits are serial-to-parallel converted into low-rate data streams which are then modulated by the GLT based signature sequences. Numerical results show that the GLT based system gives better results than the Walsh code, PN concatenated sequence based system [3].

I. INTRODUCTION

The ability to achieve high bit rates at low error rates over a wireless channel is limited by the propagation characteristics of the wireless environment. Typically, the transmitted signal arrives at the receiver via multiple propagation paths with different time delay, attenuation and phase. The net result is that the received signal undergoes random amplitude and phase perturbation. This multipath propagation results in intersymbol interference (ISI) which leads to an irreducible error floor [4].

One possible solution proposed is the multicarrier modulation or multitone modulation in which the transmitted high-rate data bits are divided into several low-rate data streams which are then used to modulate several sub-carriers [8]. In an alternative approach, low-rate data stream is modulated using Direct Sequence Spread Spectrum (DSSS) on a single carrier. This is refered to as 'multicode' modulation [3]. Obviously, the choice of the signature sequences or codes is crucial since these codes must be able to separate the interference between the low-rate bit streams and their multipath duplicates.

In this paper, we use the basis waveforms of the Generalized Lapped Transform (GLT) [12] as the signature sequences of the low-rate bit streams. Generalized Lapped Transform (GLT) is a type of linear phase perfect reconstruction FIR filter bank. It is biorthogonal in nature and more flexible than the Lapped Orthogonal Transform (LOT). Its structural parameters could be used to minimize the desirable properties such as the cross-correlation of the signature sequences. In addition, fast algorithms are available for its implementation [10],[11] since the GLT can be viewed as a generalization of the traditional discrete cosine transform (DCT).

The paper is organized as follows. In section II, the proposed GLT based high-speed transmission system is described. Section III describes the structure of the GLT and the design of the signature sequence. Section IV presents the numerical results which illustrate the potential of the proposed signature sequences. Conclusion is given in section V.

II. SYSTEM DESCRIPTION

The proposed GLT based high-speed transmission system is shown in Fig 2. The incomming data bits with bit duration T_b are serial-to-parallel converted into M parallel bit streams with symbol duration $T = MT_b$, similar to multitone or multicarrier modulation. The converted low-rate branch symbols are modulated using DS spread spectrum modulation. Here unlike the conventional system, each symbol is spreaded to the length of $2M$ and is overlapped with the adjacent spreaded bits. As a result, the bit rate is maintained as the conventional non-lapped DS spread spectrum system. Therefore, we can achieve high-rate DS spread spectrum modulation within the bandwidth of the original high-rate transmission stream while maintaining the advantages of DS spread spectrum such as multipath rejection. As each of the DS spread spectrum modulated low-rate streams passes through exactly the same wireless channel, it implies that the delay characteristics will be identical for all low-rate data streams. The receiver design is therefore very less intricate.

A. Transmitter model

The transmitted signal of the k th low-rate data stream of the Direct Sequence Spread Spectrum (DSSS) system with coherent BPSK modulation can be expressed as [2],

$$s_k(t) = \sqrt{2P}\, b_k(t) a_k(t) \cos(\omega_c t + \theta_k)\,, \qquad (2.1)$$

where $b_k(t)$ is the data sequence consisting of unit amplitude rectangular bipolar pulses of duration T, $a_k(t)$ is the spreading signal, which consists of a periodic sequence of duration T_c, P is the signal power, ω_c represents the common angular centre frequency, and θ_k is the phase of the k th carrier.

B. Fading channel model

Experimental results indicate that the received envelope for mobile radio channels in urban and rural environments is Rayleigh and Rician distributed, respectively [2]. In this paper, we consider a multipath Rayleigh-fading channel having a slow fading rate compared to symbol rate, so that the channel random parameters do not change significantly over several consecutive symbol intervals. We also assume that the channel consists of a fixed number of paths. The lowpass equivalent impulse response of the pass band fading channel is given by [2],

$$h(t) = \sum_{l=1}^{L} \beta_l \delta(t - \tau_l) e^{j\phi_l} . \tag{2.2}$$

Where L is the number of paths and $j = \sqrt{-1}$. β_l, τ_l and ϕ_l are respectively the path gain, delay and phase of the l th path. Therefore the output signal $y_k(t)$ produced by the composite multipath fading channel model for input $s_k(t)$ will be a sum of delayed, phase shifted and attenuated replicas of the input signal. Using (2.1) and (2.2), we have [2],

$$y_k(t) = \sum_{l=1}^{L} \sqrt{2P}\beta_l b_k(t - \tau_l) a_k(t - \tau_l) \cos(\omega_c t + \psi_l) , \tag{2.3}$$

where the phase $\psi_l = \theta_k + \phi_l - \omega_c \tau_l$ is assumed to be independent and identically distributed uniform random variable over $[0,2\pi]$. The path gains β_l 's are assumed to be independent and identically distributed Rayleigh random variables.

In order to develop some general design guidelines for wireless systems, some parameters which grossly quantify the multipath channel are used. One of the key parameter is the rms delay spread. It is defined as the square root of the second central moment.

$$\tau = \left[\frac{\int (t - \mu)^2 p(t) dt}{\int p(t) dt} \right]^{1/2} , \quad \mu = \frac{\int t p(t) dt}{\int p(t) dt} \tag{2.4}$$

where function $p(t)$ is called the 'power delay profile' and μ is the mean delay spread. The irreducible BER for transmission through a multipath channel depends strongly on the normalized rms delay spread [4]

$$d = \frac{\tau}{T} , \tag{2.5}$$

where T is the symbol period. Throughout this paper, we consider an equal amplitude two ray profile with,

$$p(t) = \frac{1}{2}\left[\delta(t - \tau) + \delta(t + \tau) \right] . \tag{2.6}$$

In this case, $L = 2$ and we assume that the channel consists of two faded paths separated by 2τ . The resolution r is defined as

$$r = \frac{2\tau}{T_c} , \tag{2.7}$$

where T_c is the chip duration.

C. Receiver Model

Adding $y_k(t)$ together and assuming that the noise is additive, we obtain the received signal as follows,

$$y(t) = \sum_{k=1}^{M} \sum_{l=1}^{L} \sqrt{2P}\beta_l b_k(t - \tau_l) a_k(t - \tau_l) \cos(\omega_c t + \psi_l) + n(t) \tag{2.8}$$

Let $Z_{m,j}$ denote the output of the correlated receiver that is matched to the j th path of the m th data stream. $Z_{m,j}$ can be written as,

$$Z_{m,j} = \int_{\tau_j}^{2T+\tau_j} y(t) a_m(t - \tau_j) \cos(\omega_c t + \phi_j) dt . \tag{2.9}$$

If we use the Rake receiver with L correlators then the output of the Rake receiver for the m th data stream is given by [2],

$$\begin{aligned} Z_m &= \sum_{j=1}^{L} Z_{m,j} \\ &= \sum_{j=1}^{L} \sum_{l=1}^{L} \sum_{k=1}^{M} \int_{\tau_j}^{2T+\tau_j} \sqrt{2P}\beta_l \beta_j b_k(t - \tau_l) a_k(t - \tau_l) a_m(t - \tau_j) \cdot \\ &\qquad \cos(\omega_c t + \psi_l) \cos(\omega_c t + \phi_j) dt + \eta \\ &= 2\alpha \sqrt{\frac{P}{2}} b_m T + I_k + \eta \end{aligned} \tag{2.10}$$

where

$$\alpha = \sum_{l=1}^{L} \beta_l^2 \quad \text{and}$$

$$I_k = \sqrt{\frac{P}{2}} \sum_{j=1}^{L} \sum_{l=1, l \neq j}^{L} \beta_l \beta_j \sum_{k=1}^{M} \cos(\psi_l - \phi_j) \cdot \int_{\tau_j}^{2T+\tau_j} b_k(t - \tau_l) a_k(t - \tau_l) a_m(t - \tau_j) dt . \tag{2.11}$$

The noise term , η , is given by

$$\eta = \int_0^{2T} n(t) a_m(t - \tau_j) \cos(\omega_c t + \phi_j) dt . \tag{2.12}$$

It is assumed that $n(t)$ is white Gaussian noise with two-sided power spectral density of $(N_o/2)$. The mean of η is zero and the variance of η is $\sigma_\eta^2 = 2N_oT/4$.

III. GENERALIZED LAPPED TRANSFORM (GLT)

Fig 1. shows the structure of a M-channel uniform filter bank with $f_i(n)$ and $g_i(n)$, the analysis and synthesis filters, respectively. The incoming signal $x(n)$ is split into M frequency bands by filtering with the analysis filters. Each subband signal is then maximally decimated by a factor of M. After processing in the subband domain, the M decimated signals will be interpolated, filtered by the synthesis filters and added together to reconstruct the signal.

In a perfect reconstruction (PR) filter bank the input and output are equal except for a delay (i.e. $y(n) = x(n - n_d)$). For perfect reconstruction, $f_i(n)$ and $g_i(n)$ have to satisfy certain conditions. Let $F_{i,k}(z^M)$ and $G_{i,k}(z^M)$ be the type-1 and type-2 polyphase components of $F_i(z)$ and $G_i(z)$, respectively,

$$F_i(z) = \sum_{k=0}^{M-1} z^{-k} F_{i,k}(z^M) \text{ and } G_i(z) = \sum_{k=0}^{M-1} z^{-(M-k-1)} G_{i,k}(z^M)$$

The filter bank is PR if [7]:

$$R(z)E(z) = z^{-d} I \tag{3.1}$$

where d is a constant and $E(z), R(z)$ are the polyphase matrices of the analysis and synthesis filters and are given by :

$$[E(z)]_{i,k} = F_{i,k}(z); \qquad [R(z)]_{i,k} = G_{i,k}(z). \tag{3.2}$$

A biorthogonal PR filter bank or biorthogonal lapped transform of length $2M$ called the Generalized Lapped Transform (GLT) was introduced in [12]. The polyphase matrix of the GLT is given by:

$$E(z) = \frac{1}{2} P \begin{bmatrix} U_{00} & 0_{M/2} \\ 0_{M/2} & U_{11} \end{bmatrix} \begin{bmatrix} I_{M/2} & 0_{M/2} \\ 0_{M/2} & \left(C_{M/2}^{II} S_{M/2}^{IV}\right)^T \end{bmatrix} \cdot \begin{bmatrix} I_{M/2} & I_{M/2} \\ I_{M/2} & -I_{M/2} \end{bmatrix} \begin{bmatrix} I_{M/2} & 0_{M/2} \\ 0_{M/2} & z^{-1} I_{M/2} \end{bmatrix} R_1 \tag{3.4}$$

where

U_{00} and U_{11} are block diagonal invertible matrices,

$R_1 = P' diag\{B_2 \cdots\cdots B_2\} D C_M^{II} J_M$ with D a diagonal matrix,

C_M^k, S_M^k denote the type-k length- M discrete cosine and sine transform,

P is a permutation matrix which permutes the k and $(k+M/2)$ rows to the $2k$ and $(2k+1)$ $(k = 0, \ldots, M/2-1)$ rows, respectively,

P' is a permutation matrix which permutes the $2k$ and $(2k+1)$ rows to the k and $(k+M/2)$ $(k = 0, \ldots, M/2-1)$ rows, respectively.

We can parameterize U_{ii} by products of block diagonal (2×2) invertible matrices. U_{ii} can then be written as

$$U_{ii} = \prod_{k=1}^{M/2-1} v_k^i \tag{3.5}$$

where

$$v_k^i = \begin{bmatrix} I_{k-1} & & & 0 \\ & x_k^i & y_k^i & \\ & y_k^i & x_k^i & \\ 0 & & & I_{M-k-1} \end{bmatrix}$$

and

$$\left(v_k^i\right)^{-1} = \frac{1}{\left((x_k^i)^2 - (y_k^i)^2\right)} \begin{bmatrix} I_{k-1} & & & 0 \\ & x_k^i & -y_k^i & \\ & -y_k^i & x_k^i & \\ 0 & & & I_{M-k-1} \end{bmatrix}$$

Other product forms for U_{ii} can be used but the present choice has the advantage that U_{ii} will be diagonal dominance. Parameters x_k^i, y_k^i, d_{ii} can be used to minimise different objective functions. For example, in source coding or compression applications, the coding gain is maximised. But in spread spectrum applications, cross-correlation between the basis functions should be minimised for all relative shifts. At the same time, auto correlation should be constrained to a delta function. Here, the objective function can be written as,

$$y = \sum_{m=1}^{2M-1} r(m) \tag{3.6}$$

where

$$r(m) = \sum_{k=1}^{M} \sum_{l=1, l\neq k}^{M} \sum_{n=0}^{2M-1} s_k(n) h_l(n+m). \tag{3.7}$$

$s_k(n)$ and $h_k(n)$ are the impulse responses of the k th analysis and synthesis filters of length $2M$ which are used as the spreading and despreading signature sequences. The signal flow graphs of the forward Generalized Lapped Transform and inverse Generalized Lapped Transform are shown in Fig 3a. and Fig 3b.

IV. NUMERICAL RESULTS

In this application, low-rate data streams are transmitted synchronously. However, multipath delays can introduce significant non-zero cross-correlations between the orthogonal codes. The performance of the system over multipath fading channel can be significantly improved by employing some type of diversity or maximal ratio combining [2]. In our simulations, Rake receiver with maximal ratio combining

(MRC) is considered. We have also restricted our considerations to the high signal-to-noise ratio (SNR) or the irreducible BER performance. These irreducible bit errors typically occur because of signal fading and ISI caused by multipath delay spread.

Fig 4. shows BER performance against resolution of the Walsh code based multicode modulation system (Walsh sequences with $M = N = 64$) and the GLT based system with $M = 64$ and $N = 128$. The (E_b / N_o) is 40 dB. Fig 5 shows the BER performance of the concatenated code based multicode system and the GLT based system. The concatenated codes are obtained by multiplying the Walsh code with a PN code which is identical for all data streams. More precisely, the Walsh codes with $M = 64$ are multiplied with a 64 bits of a m-sequence which is generated by appending one extra bit "0" to a 63 bits PN sequence. This concept of concatenated codes has been successfully applied to the down-link transmission of the Qualcomm Code Division Multiple Access (CDMA) system.

From Fig 4. and Fig 5., we can see that the BER performance of the GLT based system is better than the Walsh code or concatenated code based system. The irreducible BER reach the ceiling when r takes on integer values. On the other hand, the irreducible BER drop to the floor when $r = 1.5, 2.5, 3.5, \ldots\ldots$ This is because the cross-correlation function reaches the maximum and minimum values when the separation between the two sequences is integer and integer-plus-half multiples of the chip duration, respectively.

V. CONCLUSION

This paper presents a GLT based high-speed transmission system to transmit high bit rate data over a wireless mobile radio environment which is characterized by a frequency selective fading channel. Numerical results show that the GLT based system gives low BER than the walsh code or concatenated code based system. In designing the GLT based signature sequences we formulated the objective function as the cross-correlation of the signature sequences. The structural parameters of the GLT are obtained by minimizing the cross-correlation. Fast Algorithms are also available for its implementation since the GLT can be viewed as a generalization of the traditional discrete cosine transform (DCT).

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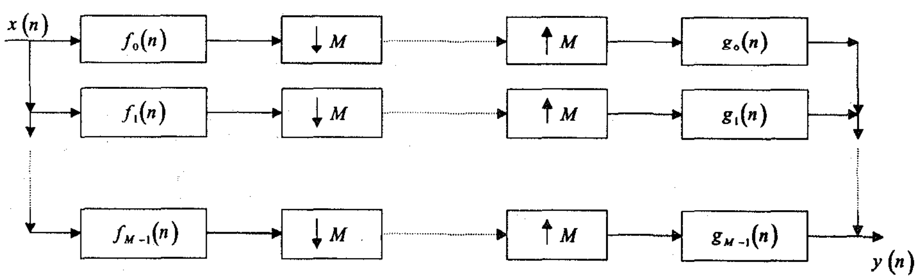

Fig 1. M-channel uniform filter bank

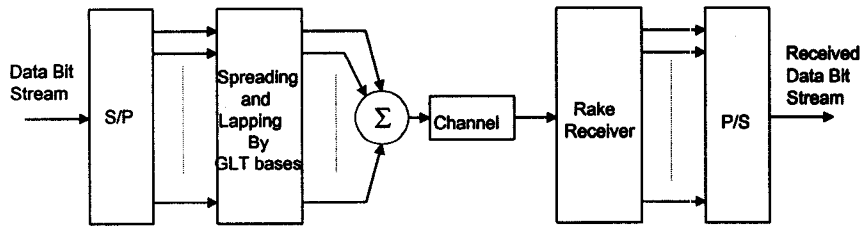

Fig 2. GLT based High-Speed Transmission system model

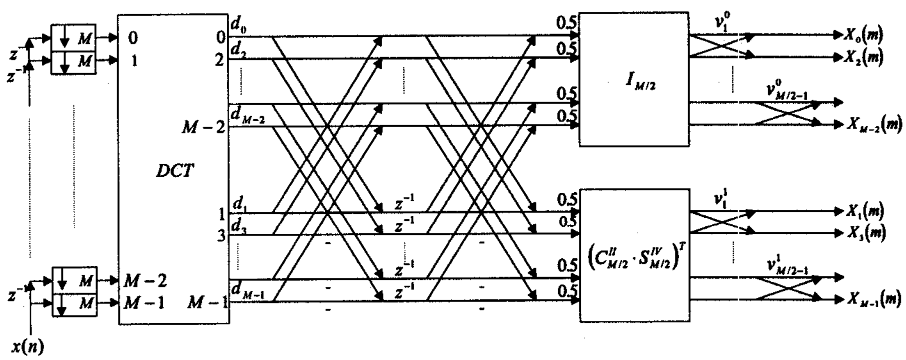

Fig 3a. Flow graph of Forward Generalized Lapped Transform

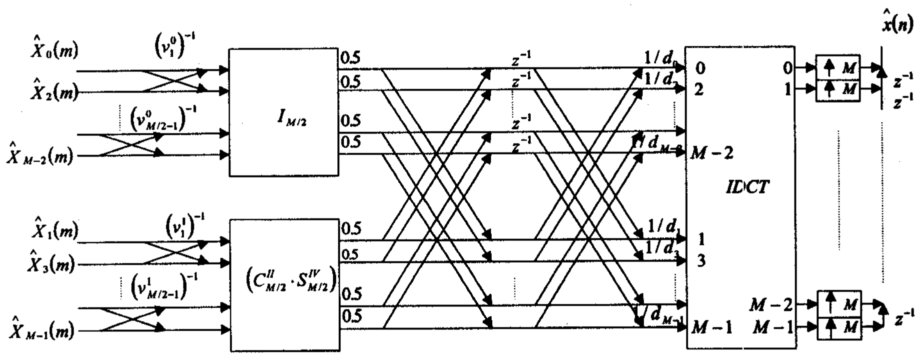

Fig 3b. Flow graph of Inverse Generalized Lapped Transform

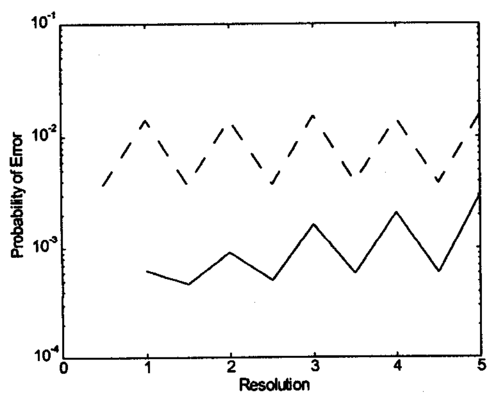

Fig 4. Bit error against Resolution for GLT based codes (solid line) and Walsh codes (dashed line)

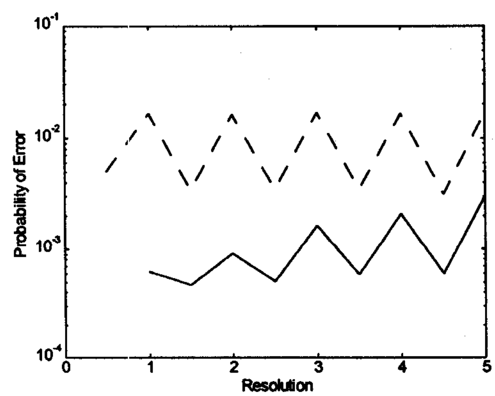

Fig 5. Bit error against Resolution for GLT based codes (solid line) and concatenated codes (dashed line)