Non-Guard Interval based and Genetic Algorithm Assisted Frequency Domain Equalization for DS-UWB Systems

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ABSTRACT: In this work, a genetic algorithm (GA) based frequency domain equalization (FDE-GA) scheme was proposed for direct sequence ultra wideband (DS-UWB) wireless communication systems. The proposed FDE-GA scheme does not require a guard interval (GI) and the output of the RAKE receiver is used as the input to our GA. The scheme achieved much higher bandwidth efficiency than conventional FDE methods because of the removal of the inter block interference (IBI) within each block before the GA. The FDE-GA receiver was shown to significantly outperform the RAKE receiver and the RAKE-GA receiver proposed in a previous work, in terms of bit error rate (BER) at a similar complexity. An improvement in the mean square error (MSE) was observed from simulation results presented, as a result of increase in the number of pilot symbols.

KEYWORDS: frequency domain equalization, guard interval, ultra-wideband, genetic algorithm, inter block-interference

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I. INTRODUCTION

Ultra-wideband (UWB) wireless communication is a revolutionary technology for transmitting large amounts of digital data over a wide frequency spectrum using short-pulse, low-powered radio signals. UWB commonly refers to a system that either has a large relative bandwidth that exceeds 20% or a large absolute bandwidth of more than 500 MHz (Kshetrimayum, 2009). UWB technology is one of the promising solutions in terms of high-speed short-range wireless communication systems (Porcino and Hirt, 2003).

Genetic algorithm (GA) is an effective search technique, which works on the Darwinian principle of natural selection called "survival of the fittest" (Man *et al*., 1999). GA presumes that the potential solution of any problem is an individual, represented by genes of a chromosome and structured by a string of values in binary form.

In an impulse-based DS-UWB system, the transmitted data bit is spread over multiple consecutive pulses of very low power density and ultra-short duration. This introduces resolvable multipath components having differential delays in the order of nanoseconds. Thus the performance of a DS-UWB system is significantly degraded by the inter-chip interference (ICI) and inter-symbol interference (ISI) due to multipath propagation (Liu and Elmirghani, 2007).

The channel impulse response (CIR) of UWB has multipath delay spread which is greater than the reciprocal bandwidth of the transmitted message waveform and hence the received signal is distorted due to attenuated and delayed versions of the transmitted waveform. This poses a great challenge in the design of UWB and so lots of researches have been done to combat this problem.

Two adaptive detection schemes based on single-carrier frequency domain equalization (SC-FDE) for multiuser direct-sequence DS-UWB systems were proposed in (Li and Lamare, 2013). The scheme was termed structured channel estimation (SCE) and direct adaptation (DA).

Joint FD channel estimation and equalization for IR-UWB systems with short cyclic prefix (CP) and a novel iterative receiver employing soft IBI estimation and cancellation within both its FD channel estimator and FD equalizer components was considered by (Bahceci and Koca, 2010).

Sato and Ohtsuki, (2005), evaluated the computational complexity and the performance of the RAKE receivers for the DS-UWB considering the accuracy of channel estimation in a multipath channel. Also, the performance of a joint RAKE and minimum mean square error equalizer (MMSE) receiver for high data rate UWB communications, to combat the ISI by taking advantage of the RAKE and equalizer structure was proposed in (Eslami and Dong, 2005). (Li *et al*., 2003) combined a RAKE receiver and MMSE equalizer structure for UWB systems.

GA implementation in UWB and code division multiple access communication systems can be found in (Chang, *et al*., 2012), and (Hung and Chen, 2012). GA has also been applied to UWB communications systems in (Gezici *et al*., 2005) and (Wang *et al*., 2008). In our previous works, a GA based FDE, without a guard interval (GI), was proposed for DS-UWB wireless communication systems (Surajudeen-Bakinde, *et al*., 2011). Also in (Surajudeen-Bakinde, *et al*., 2009), GA was combined with a RAKE receiver to combat the ISI due to the frequency selective nature of UWB channels for high data rate transmission.

In (Kaligineedi and Bhargava, 2008) low-complexity frequency domain MMSE turbo equalization schemes for single-user binary phase shift keying (BPSK) and quaternary bi-orthogonal keying DS-UWB systems was derived. The frequency domain channel estimation problem for the SC-FDE over UWB channels was investigated by (Wang and

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Dong, 2006). In all the aforementioned works, GI has been used to cancel the IBI caused by the multipath channel.

In this work, we propose FDE-GA detection approach, which does not require any GI for DS-UWB wireless communication systems where equalization is performed in the frequency domain. This work is different from other research work on FDE because of the absence of GI and the use of GA for FDE and so it is more spectrum efficient than all other conventional FDE methods. At the time of presenting this work and to the best of our knowledge, no work has been done to implement GA based FDE for DS-UWB without GI.

The use of a few numbers of variables, which resulted in medium size of search space, for our algorithm is prompted, by the high complexity and simulation time of the GA, when large numbers of variables are used. Meanwhile, the use of GI, with a few number of variables lead to very high GI overhead, which is not desirable in wireless communication.

This paper is a revised and expanded version of a paper entitled "Genetic Algorithm based Frequency Domain Equalization for DS-UWB Systems without Guard Interval" presented at the IEEE International Conference on Communication, Kyoto, Japan. 5th – 9th June, 2011.

The remaining part of the paper is organised as follows. Section II is system model. We propose the FDE-GA equalization approach in Section III. Section IV presents the computational complexity of the FDE-GA receiver in comparison to the other receivers. Simulation setup and results are presented in Section V. Section VI draws the conclusion.

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II. SYSTEM MODEL

A. Transmitted Signal

A non-guard interval (GI) based DS-UWB system FDE over UWB channels is considered in this work. At the transmitter, each block consists of M symbols. Each symbol is first spread by the Ternary orthogonal spreading code sequence (Zhu and Murch, 2003), due to its orthogonality, and it is of length N_c. Let $d_{k,m} \in \{\pm 1\}$ denote $m\text{th}(m = 0,..., MN_c - 1)$ chip in the *kth* signal block.

B. Channel Model

The Saleh-Valenzuela channel model with a couple of slight modifications is implemented in this work. A log-normal distribution rather than a Rayleigh distribution for the multipath gain magnitude is recommended in this model, because the log-normal distribution seems to better fit the measurement data as stated in the Saleh-Valenzuela channel model. In addition independent fading is assumed for each cluster as well as each ray within the cluster (Foerster, 2003). The channel model recommended by the IEEE 802.15.3a channel modelling sub-committee is defined by the following channel impulse response (CIR):

$$h(t) = \sum_{i=0}^{C_L} \sum_{k=0}^{K} \alpha_{i,k} \delta(t - T_i - \tau_{i,k}) \tag{1}$$

where $\alpha_{i,k} = p_{i,k} \xi_{i,k}$ are the multipath gain coefficients, with $p_{i,k} \in \{\pm 1\}$ denoting the random polarity (the possible phases for real coefficients) with equal probability and the fading amplitude $\xi_{i,k}$ being log-normal-distributed. T_i is the delay of the i^{th} cluster and $\tau_{i,k}$ is the delay of the k^{th} path within the i^{th} cluster relative to T_i . T_i and $\tau_{i,k}$ are described with a double-Poisson process and all of them are rounded to integer times of the delay resolution T_c (Chu and Murch, 2006).

The total numbers of observed clusters and the number of multipath contribution within the i^{th} cluster are C_L and K respectively. Thus, with τ_{exc} denoting the multipath delay spread, $L_{tot} = \tau_{exc} / T_c$ denotes the total number of paths. Let h_l denotes the sum of all $\alpha_{i,k}$ at time index l , where $l = \lfloor (T_i + \tau_{i,k}) / T_c \rfloor$. Due to the clustering of multipath components, the channel does not necessarily have multipath arrivals within each delay bin. This is accounted for by setting $h_l = 0$ for any lT_c that has no path arrival (Chu and Murch, 2006). Therefore, the CIR in (1) can be simplified to:

$$h(t) = \sum_{l=0}^{L_{tot}-1} h_l \delta(t - lT_c) \tag{2}$$

where L_{tot} is the total number of paths, $\tau_l (= lT_c)$ is the delay of the l^{th} path component and h_l is the l^{th} path gain (Foerster, 2003). It is assumed that $L_{tot} = MN_c$ in our simulation, therefore the CIR is of order MN_c with taps $\mathrm{h} = [h_0, \ldots h_{MN_c - 1}]^T$.

C. Received Signal

Assuming timing is acquired, the received signal block, $\mathbf{r}_k$ can be expressed in matrix form as given by Wang and Dong, 2006.

$$\mathrm{r}_k = \mathrm{HS}_k + \mathrm{n}_k \tag{3}$$

where $\mathrm{H} = [\mathrm{h}_0^T, \ldots \mathrm{h}_{MN_c-1}^T]$, n_k is the AWGN whose elements have zero mean and single-sided power spectral density. S_k is the kth signal block which is a matrix expressed as:

$$\mathbf{S}_k = \begin{bmatrix} d_{k,0} & d_{k,1} & \ldots & d_{k,MN_c-1} \\ d_{k-1,0} & d_{k,0} & \ldots & d_{k,MN_c-2} \\ . & . & . & . \\ . & . & . & . \\ . & . & . & . \\ d_{k-1,MN_c} & d_{k-1,MN_c-1} & \ldots & d_{k,0} \end{bmatrix} \tag{4}.$$

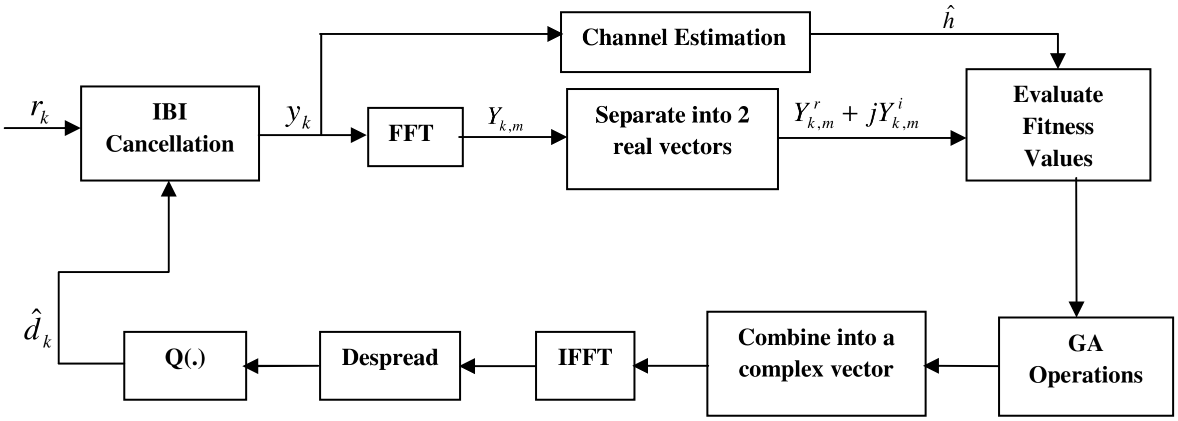

Figure 1: Block diagram of FDE-GA receiver for DS-UWB Systems.

III. FDE-GA WITHOUT GI FOR DS-UWB SYSTEMS

The FDE-GA receiver without GI is as illustrated in Figure 1. As shown in the block diagram, IBI cancellation process is carried out on the received signal with IBI as a result of the absence of GI, to increase the spectrum efficiency of the system. Also input to the IBI cancellation block is the hard estimate that is fed back into the system to be used in the cancellation of the IBI in the received signal. After the IBI cancellation, the output is passed through the fast Fourier transform (FFT) block for conversion into the frequency domain.

The GA does not accept complex functions and so the frequency domain signal is separated into two real vectors before sending them into the GA where the vectors are being converted into binary 1 and 0 before the fitness evaluation is done. After the fitness evaluation, the GA optimization takes place by the implementation of the GA operations which are later on enumerated and explained. At the end of the GA optimization, two vectors are returned which are then combined into a complex signal before passing it through the inverse fast Fourier transform (IFFT) block to convert into the time domain.

The time domain signal is now de-spread and quantised to obtain the hard estimates. The hard estimates is not the desired signal because of the absence of GI and resulting into IBI, the signal is then fed into the IBI cancellation block at the same time as the received signal with no GI. The input data is the current data block and the hard estimate, which is the previous data block, is used to remove the IBI in the current data block as explained later. The IBI free data block is now fed into the FFT block and all the other operations already explained take place and the desired signal is now obtained.

A. IBI Cancellation for FDE-GA Receiver

Equation (3) can be rewritten as the *k*th received signal block with IBI and is thus:

$$\mathrm{r}_k = \mathrm{H}\overline{S}_k + \mathrm{H}\mathrm{S}_k^{IBI} + \mathrm{n}_k \tag{5}$$

where $\overline{S}_k$ is the *k*th transmit signal block with the elements of the matrix of size $(MN_c \times MN_c)$ expressed as

$$\overline{S}_k = \begin{bmatrix} d_{k,0} & d_{k,1} & \dots & d_{k,MN_c-1} \\ 0 & d_{k,0} & \dots & d_{k,MN_c-2} \\ . & . & . & . \\ . & . & . & . \\ . & . & . & . \\ 0 & 0 & \dots & d_{k,0} \end{bmatrix} \tag{6}$$

and S_k^{IBI} is IBI to the *k*th transmit signal block, a matrix of size $(MN_c \times MN_c)$ and is derived as:

$$\mathbf{S}_k^{IBI} = \begin{bmatrix} 0 & 0 & \dots & 0 \\ d_{k-1,0} & 0 & \dots & 0 \\ . & . & \dots & . \\ . & . & \dots & . \\ . & . & \dots & . \\ d_{k-1,MN_c} & d_{k-1,MN_c-1} & \dots & 0 \end{bmatrix} \tag{7}$$

The IBI cancellation process is carried out on the $\mathbf{S}_k^{IBI}$ which is the *k*th signal block of *M* transmitted symbols with IBI, generated from the input data block. Assuming perfect IBI removal then $\hat{S}_k^{IBI} = S_k^{IBI}$ and with perfect channel estimation, $\hat{H} = H$, the output received signal after the IBI has been removed is now obtained and expressed as :

$$\mathbf{y}_k = \mathbf{r}_k - \hat{H}\hat{S}_k^{IBI} = H\overline{S}_k + H\left(S_k^{IBI} - \hat{S}_k^{IBI}\right) + n_k \tag{8}$$

where $\hat{H}$ is an estimate of the CIR H, defined as $\hat{H} = \left[\hat{h}_0^T, \dots \hat{h}_{MN_c-1}^T\right]$, $\hat{S}_k^{IBI}$ is an estimate of S_k^{IBI} which is given as

$$\hat{S}_k^{IBI} = \begin{bmatrix} 0 & 0 & \dots & 0 \\ \hat{d}_{k-1,0} & 0 & \dots & 0 \\ . & . & \dots & . \\ . & . & \dots & . \\ . & . & \dots & . \\ \hat{d}_{k-1,MN_c} & \hat{d}_{k-1,MN_c-1} & \dots & 0 \end{bmatrix}$$

where $\hat{d}_{k,m}$ is an estimate of $d_{k,m}$. Equation (8) is passed through the FFT block to convert to the time domain signal into frequency domain. The frequency domain signal that resulted is:

$$Y_{k,m} = H_m D_{k,m} + N_{k,m} \tag{9}$$

where $H_m = \sum_{l=0}^{MN_c-1} h_l e^{-\frac{j2\pi lm}{MN_c}}$, h_l is the lth path gain of CIR already defined, $D_{k,m} = \sum_{l=0}^{MN_c-1} d_k e^{-\frac{j2\pi lm}{MN_c}}$ and $N_{k,m} = \sum_{l=0}^{MN_c-1} n_{k,i} e^{-\frac{j2\pi lm}{MN_c}}$.

B. Algorithm Description of the FDE-GA Receiver

The GA evaluates the fitness of individuals within the population of the receive signal using the objective function derived in the expression that follows:

$$\mathrm{J} = \sum_{m=0}^{MN_c-1} \left| Y_{k,m} - H_m \tilde{D}_{k,m} \right| \tag{10}$$

The GA does not accept complex functions like we have in (10) and so all the frequency components are first separated into two real vectors before the fitness evaluations, like we have in $Y_{k,m}$ and it is expressed as $Y_{k,m} = Y_{k,m}^r + Y_{k,m}^i$ where $Y_{k,m}^r$ is the real part and $Y_{k,m}^i$ is the imaginary part. The two vectors are passed to the GA where they are treated separately and then their elements are converted into binary 1 and 0. $\tilde{D}_{k,m}$ is the FFT of candidate solution of signals $\tilde{d}_{k,i}(i = 0 \dots MN_c - 1)$. At the end of the fitness function evaluation using the equation in (10), all the GA operations, proportional fitness scaling, stochastic selection, scattered crossover and Gaussian mutation are implemented (Surajudeen-Bakinde *et al*., 2011).

IV. COMPUTATIONAL COMPLEXITIES

A complexity analysis is provided for the FDE, FDE-GA, RAKE and RAKE-GA (Surajudeen-Bakinde, *et al*., 2011) receivers in terms of the floating point multiplication. The complexity of all the receivers depends on the derivation of the finger weights, the fitness function evaluation, FFT and IFFT operations and the IBI cancellation. The symbolic complexities for the FDE, FDE-GA, RAKE and RAKE-GA receivers are tabulated in Table 1.

The complexity of each operation for all the receivers is tabulated in Table 2 using the symbolic complexity derived and given in Table 1. All the complexity analysis is for CM3 with perfect channel state information (CSI) assumed. This shows that the complexity of the RAKE-GA receiver is highly influenced by the number of RAKE fingers.

In Table 3, the complexities of the RAKE and RAKE-GA receivers are normalised to the FDE-GA receiver since it is the least complex in this case. It is shown that the RAKE receiver at L = 10 is 1.14 times more complex than the FDE-GA receiver, while the RAKE-GA receiver at L = 2 is just about the same complexity as the FDE-GA receiver, while the RAKE-GA receiver at L = 10, is over five times more complex than the FDE-GA receiver.

Figure 2 further confirms the complexity values of the FDE-GA receiver in comparison to the RAKE-GA receiver where it is shown that the RAKE-GA receiver at L = 10 is much more complex than the FDE-GA receiver at P = 100. In this same figure, it is shown that FDE-GA receiver and RAKE-GA receiver at L = 2 are almost of the same complexity values.

Table 1: Symbolic Complexity for the Receivers.

Receivers	Order of Complexity
FDE	$O(MN_c \log MN_c)$
FDE-GA	$O(2Gen(MN_c \log MN_c + P \log P + MN_c)) + 2(MN_c \log MN_c + MN_c)$
RAKE	$O(LMN_c)$
RAKE-GA	$O(Gen(PLM + P \log P) + LM^2)$

Table 2: Symbolic Complexity of Operations for all Receivers (Gen = Number of generations, P = Populations size, L = RAKE finger size, M = Number of symbols per block, Nc = Spreading code length).

Operations	RAKE	RAKE-GA	FDE	FDE-GA
Fitness Evaluations	0	*Gen(PlogP+PLM)*	0	*2Gen(PlogP+ MN_c*
Finger Derivation	*LM*	*LM^2*	0	0
IBI Cancellation	0	0	0	*2(MN_c)*
FFT	0	0	*(MN_clogM N_c)*	*2(MN_clogMN_c)*
IFFT	0	0	*(MN_clogM N_c)*	*2(MN_clogMN_c)*

Table 3: Normalised Complexity for the Receivers at Nc = 24.

Receivers	Parameters	Normalised Complexity
FDE-GA	M = 10, P = 100, Gen = 10	1.00
RAKE	M = 100, L = 10	1.14
RAKE-GA	M = 10, P = 100, Gen = 10, L = 2	1.02
RAKE-GA	M = 10, P = 100, Gen = 10, L = 10	5.01

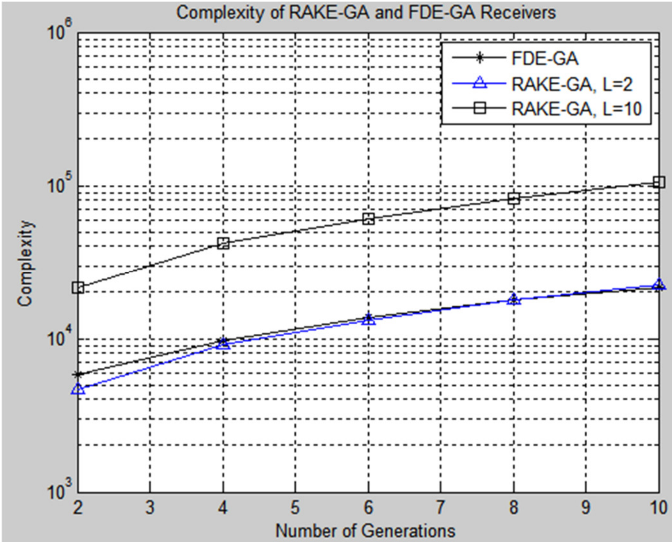

Figure 2: Complexity of RAKE-GA receiver and FDE-GA receiver for CM3.

V. SIMULATIONS

A. Setup

The performance of the proposed FDE-GA receiver, in comparison with the RAKE, RAKE-GA, and FDE-GA receivers was given in (Surajudeen-Bakinde, *et al*., 2011) and (Surajudeen-Bakinde, *et al*., 2011). All systems use BPSK modulation, with ternary orthogonal spreading code sequence of length N_c = 24, to spread the transmit symbol and a chip duration of T_c = 0.167ns. This results in a symbol duration of $T_s = N_c T_c$ = 4 ns, with a transmission rate of $R_s = 1/T_s$ = 250 Mbps (Sato and Ohtsuki, 2005).

The simulated IEEE 802.15.3a UWB multipath channel model for a single user scenario is employed for the simulation. The channel model 1 (CM1), a model based on line of sight for a distance of 0 to 4 m, mean excess delay of 5 ns, root mean square (RMS) delay of 5 ns and channel model 3 (CM3), a non-line-of-sight (NLOS) environment, with a distance of 4 to 10 m, mean excess delay of 15.9 ns and RMS delay of 15 ns are considered in this work (Foerster, 2003).

A data-aided approach used in (Lottici, *et al*., 2002) is also implemented in estimating the CIR, and this implies that the transmitted signal is known to the receiver. The sliding window correlator method according to (Li *et al*., 2003) and (Mielczarek, *et al*., 2003) is used in estimating the channel gains and delays using pilot symbols of B = 10 to 100. B known pilot symbols are sent for the training $d^t_{k,m}, (k = 1,2\ldots,B)$ in order to estimate the channel. The estimated path gains of the channel vector $\hat{h} = \left[\hat{h}_0, \hat{h}_2, \ldots, \hat{h}_{L_{est}-1}\right]^T$, can be expressed as $\hat{h} = \frac{1}{B}\sum_{k=0}^{B-1} d^t_{k,m} y^{L_{est}}_{k,m}$ using the cross-correlation method, where L_{est} is the number of paths to be estimated and it is assumed that the receiver knows the optimal value of $L_{est}(i.e. L_{est} = MN_c)$. The mean square error (MSE) of the pilot aided channel estimation as given by (Sato and Ohtsuki, 2005), can be expressed as

$$MSE_{channel} = E\left[\frac{\left\|h-\hat{h}\right\|^2}{\sum_{l=0}^{MN_c-1}\left|h_l\right|^2}\right]$$

B. Results

The results obtained from this work are compared with previous works (Surajudeen-Bakinde, *et al*., 2011) and (Surajudeen-Bakinde *et al*., 2009), where the RAKE-GA receiver was implemented in time domain. The simulation of FDE-GA and RAKE-GA results in Figure 3 is a plot of BER against the number of generations at SNR of 20 dB. It is shown in the figure that the FDE-GA receiver converges at Gen = 8 while the RAKE-GA receiver converges at Gen = 10 for L = 2 & 10.

The FDE-GA receiver is of improved BER values at SNR = 20 dB for all the number of generations considered at Gen = 0 to 20 than the RAKE-GA receiver at L = 2. The FDE-GA receiver also converges earlier than the RAKE-GA receiver at L = 2. The RAKE-GA receiver at L = 10 is only of better BER than FDE-GA receiver at number of generations of Gen = 6 to 20.

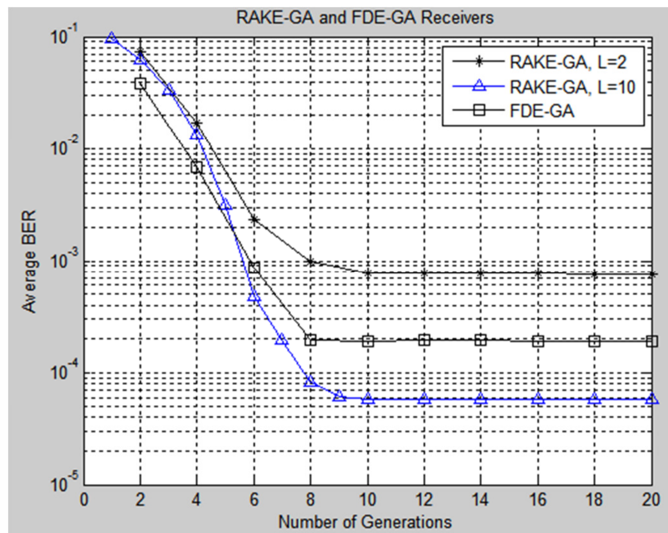

Figure 4: BER *vs*. Number of Generations for CM3.

In Figure 4, the BER performance of the proposed FDE-GA receiver, in a perfect CSI for CM3 scenario is compared to the FDE, RAKE and RAKE-GA receivers. The FDE receiver with GI overhead of 25% has the best performance, while the FDE-GA receiver without GI, at P = 100, Gen = 10 is of the same BER only at 0 to 5 dB. The FDE-GA receiver is better than the RAKE receiver at L = 10, at all Eb/No values and also of lower BER than RAKE-GA receiver at P = 100, Gen = 10, L = 10 at Eb/No = 0 to 12 dB. The FDE-GA receiver has an Eb/No gain of 4 dB at BER = 10^{-2} when compared to the RAKE receiver, whereas the RAKE-GA receiver at BER = 10^{-4}, has an Eb/No gain of only 2 dB when compared to the FDE-GA receiver. The proposed FDE-GA receiver without GI is more bandwidth efficient than the FDE receiver with GI inserted, and so the performance difference is highly justified.

Figure 5 compares the performance of the FDE-GA receiver to the FDE, RAKE and RAKE-GA receivers, all at the same simulation parameters as presented in the results shown in Figure 4. The UWB channel model considered in this figure is CM1 and so the GI overhead in this case is also 25%; though the number of GI is less than the CM3 case because the total number of paths in this channel model is less than CM3 used in the last results presented. The same performance trend observed in Figure 4 is also seen in this figure except that the BER performance is all better in CM1 for all the receivers considered. The FDE-GA receiver has an Eb/No gain of 4 dB at BER = 10^{-2} over the RAKE receiver, and the RAKE-GA receiver has an Eb/No gain of about 3 dB over the FDE-GA receiver at BER = 10^{-4}.

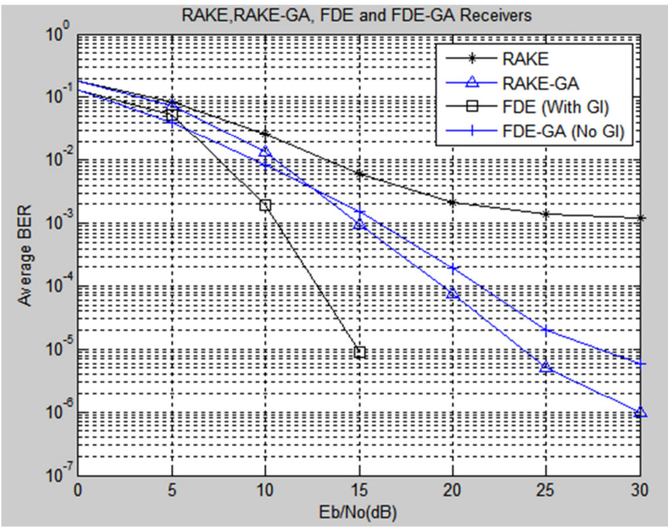

Figure 4: BER *vs*. Number of Generations for CM3.

In Figure 6, the complexity values of the FDE-GA receiver at P = 100, Gen = 8 and the RAKE-GA receiver at P = 100, Gen = 8, L = 2 are the same, but the FDE-GA receiver significantly outperforms the RAKE-GA receiver with a gain in power of 4 dB at BER = 10^{-3}. The FDE-GA receiver, at same complexity also has an enormous gain in power of about 15 dB over the RAKE receiver at BER = 10^{-2}. The FDE receiver with GI has only 2 dB gain in Eb/No over the FDE-GA receiver without GI at BER = 10^{-2}. The RAKE receiver at L = 2 has the lowest numerical complexity value of 4800 in this figure, followed by the FDE receiver which has a numerical complexity value of 8113. The FDE-GA receiver and the RAKE-GA in this figure are of the same numerical complexities of 17804 and 17800 respectively.

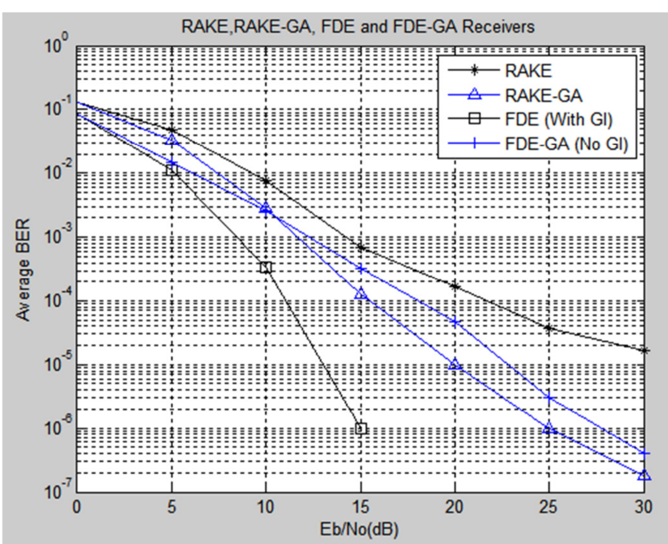

Figure 5: BER vs. Eb/No for all the receivers for CM1.

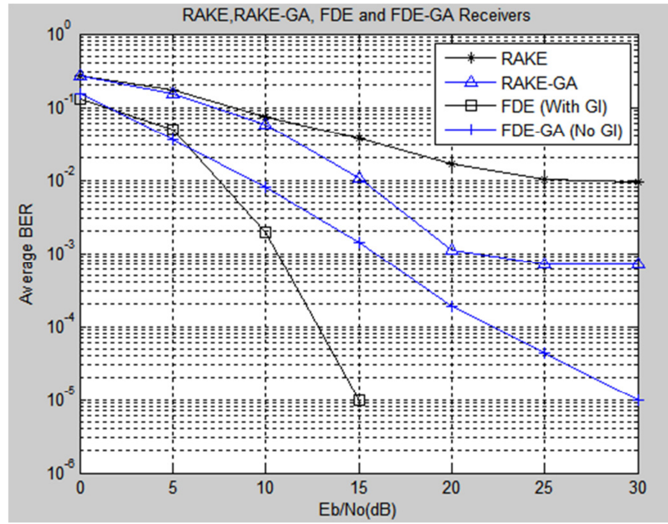

Figure 6: BER vs. Eb/No for all the receivers with similar complexity values for CM3.

Figure 7 is a plot of BER against the Eb/No for population sizes of P = 20 to 100 at Gen = 8 for CM3 with perfect CSI assumed for the FDE-GA receiver. Gen = 8 is used because it is the number of generations where the FDE-GA receiver has been presented to converge in earlier result presented in Figure 3. There is improvement in the BER as the population size increases from P = 20 to P = 100 as shown in the graph. The average BER of the FDE-GA receiver at P = 80 and 100 is the same at Eb/No = 0 to 15 dB, with the FDE-GA receiver at P = 100 having a gain of about 2 dB over the FDE-GA receiver at P = 80 both at BER = 10^{-4}.

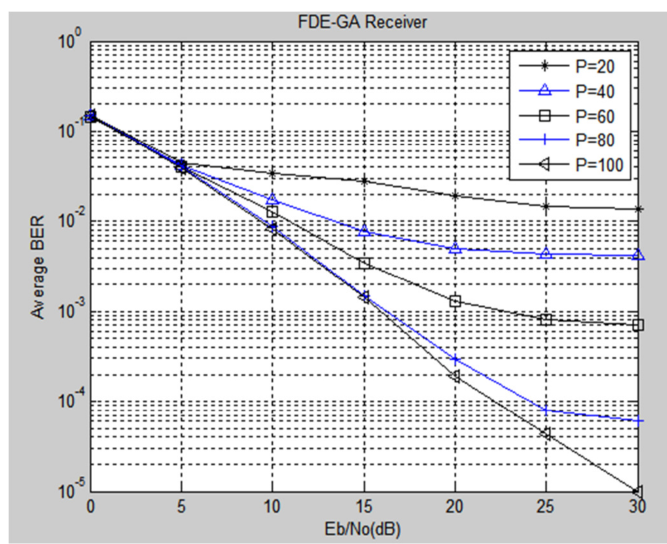

Figure 7: BER vs. Eb/No for FDE-GA receiver for CM3.

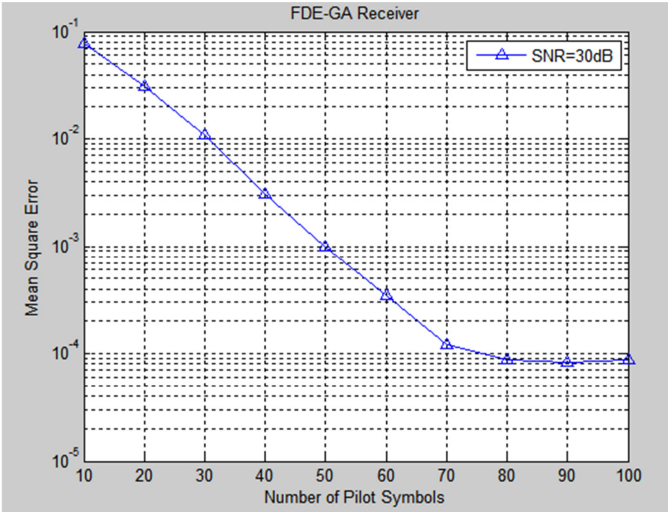

Figure 8: MSE vs. Pilot Symbols for FDE-GA receiver for CM3.

The MSE of the channel estimation for the FDE-GA receiver in terms of the number of pilot symbols at Eb/No = 30 dB is presented in Figure 8. The MSE channel for the channel estimation of FDE-GA receiver is derived by averaging the MSE more than 1000 channel realization. It is shown in the figure that the MSE improves as the number of pilot symbols is increased from B = 10 to 80; with the performance being the same at B = 80 to 100.

VI. Conclusion

Channel equalization of DS-UWB with GA based in the frequency domain without GI was carried out in this work. The FDE-GA receiver was compared to the RAKE-GA receiver in terms of the speed of convergence. The results obtained showed that FDE-GA receiver converges earlier than the RAKE-GA receiver. There was complexity reduction in the FDE-GA receiver in comparison to the previous RAKE-GA receiver at almost the same BER performance. It was equally observed from simulation results obtained and presented in this work that, FDE-GA receiver significantly outperformed the RAKE receiver and the RAKE-GA receiver at similar complexity. The proposed FDE-GA structure without a GI, achieved much higher bandwidth efficiency than the conventional FDE methods, because the resulting IBI was removed effectively, within each block before the GA. The performance of the two channel models of UWB, CM1 and CM3, considered were of the same BER performance, but better performance obtained for CM1. An increase in the population size for the FDE-GA also resulted in improved BER performance. The number of pilot symbols for the FDE-GA receiver for CM3 also gave us improved BER performance, though at increased computational complexity.

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