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NR-DCSK: A Noise Reduction Differential Chaos Shift Keying System

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***Abstract*—One of the major drawbacks of the conventional Differential Chaos Shift Keying (DCSK) system is the addition of channel noise to both the reference signal and the data bearing signal, which deteriorates its performance. In this paper, we propose a noise reduction DCSK (NR-DCSK) system as a solution to reduce the noise variance present in the received signal in order to improve performance. For each transmitted bit, instead of generating β different chaotic samples to be used as a reference sequence, β/P chaotic samples are generated and then duplicated P times in the signal. At the receiver, each P identical samples are averaged and the resultant filtered signal is correlated to its time delayed replica to recover the transmitted bit. This averaging operation of size P reduces the noise variance and enhances the performance of the system. Theoretical bit error rate expressions for AWGN and multipath fading channels are analytically studied and derived. Computer simulation results are compared to relevant theoretical findings to validate the accuracy of the proposed system and to demonstrate the performance improvement compared to the conventional DCSK, the Improved DCSK (I-DCSK) and the Differential Phase Shift Keying (DPSK) systems.**

***Index Terms*—Non-coherent chaos based communication system, NR-DCSK, Average filtering, Performance analysis.**

I. Introduction

OWING to their auspicious attributes, chaotic signals have lately become a seriously dominating concept for many research bodies in the field of wireless communications [1]–[6]. This is fundamentally relevant to the intrinsic wideband particularities of these signals that make them a strong fit for various spread spectrum modulation schemes [1], [7]. Chaotic modulation has similar advantages to other spread spectrum modulation schemes including jamming resistance and cutback of fading effects *exempli gratia*. Various chaos-based communication systems with coherent and non-coherent detection algorithms have been nominated and assessed in the last few years [3], [7], [8]. The utilization of chaos-based coherent detection tactics was limited because of the sensitive dependence of chaotic synchronization to the noise in the channel. Further, differential chaos shift keying (DCSK) is the mostly considered system amid all chaos-based communication systems [8]–[10]. The essential cause for this is the fact that DCSK does not require neither chaotic synchronization nor knowledge of channel state information at the receiver to recover the transmitted data [10], [11]. Moreover, DCSK systems are more vigorous to multipath fading environments than DPSK schemes and are more suitable for ultra wide band (UWB) applications [6], [10]–[12]. Nevertheless, the dominant hindrance of conventional DCSK is its inferior performance and capacity which limit its applicability. In fact, the reference and information bearing signals are corrupted by the channel noise and a noisy reference signal is correlated with a noisy information bearing signal at the receiver, which deteriorates the performance of the DCSK system. To overcome the acknowledged defects of the DCSK scheme and to reinforce its spectral efficiency, a large volume of research has lately been carried out, to come up with new non-coherent DCSK systems. A quadrature chaos shift keying (QCSK) approach that is proposed in [13] uses the Hilbert transform to produce an orthogonal basis of chaotic functions granting the transmission of larger number of bits in comparison to DCSK while demanding an equivalent bandwidth. Also, high efficiency HE-DCSK [14], M-DCSK [12], code-shifted CS-DCSK [15], [16] and differential DDCSK [17] are suggested to lift the data rate of DCSK systems at the cost of elevated system complexity. Eventually, multi-carrier (MC-DCSK) is proposed in [6] where chaotic reference sequences are transmitted over predefined subcarrier frequencies while multiple modulated data streams are transmitted over the remaining subcarriers. The indicated MC-DCSK scheme perks up energy efficiency and offers loftier data rates, but demands bandwidth.

Contributions and paper outline: The majority of proposed schemes suffer from high system complexity and fail to enhance the performance of non-coherent chaos based communication systems significantly. The main motivation of our paper is to propose a promising low complexity, non-coherent DCSK scheme that, besides simplicity, does not need a channel estimator to recover the data. To reach this goal, we propose a new frame architecture with a simple non-coherent receiver that increases the resultant SNR. In the proposed new design, the reference signal of the NR-DCSK system uses β/P different chaotic samples only instead of β samples as done in classical DCSK schemes. Therefore, each chaotic sample is repeated P successive times to make a reference signal of β samples in total. Like DCSK system, the reference signal or its inverted version depend on the transmitted bit. At the receiver, the received NR-DCSK is filtered with a moving average filter within a window of P samples. The resultant averaged signal is correlated with its delayed replica. Finally the decoded bits are recovered by comparing the correlator output to a zero threshold. In this scenario, the averaging operation assimilates

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* This work has been supported by the NSERC discovery grant 435243 – 2013.

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also P independent additive and identically distributed Gaussian noise samples present in measurements, which decreases the resultant variance by a factor of $1/P$ [18]. Consequently, the SNR of the averaged signal is elevated and the performance is enhanced. In addition, as in conventional DCSK systems, the receiver can perform without any need to complex channel estimators. In this paper, we derive the analytical bit error rate expressions over multipath fading and additive white Gaussian noise (AWGN) channels and we show the accuracy of our analysis by matching the numerical performance to simulation results. Eventually, we analyse and compare the performance of the proposed NR-DCSK to conventional DCSK, I-DCSK [19] and DPSK systems. Enduringly, the proposed noise reduction technique can be applied to any chaotic-based non-coherent transmit reference scheme.

The remainder of this paper is organized as follows: In section II, the conventional DCSK is briefly presented and the architecture of the proposed NR-DCSK system is explained. Performance analysis of NR-DCSK scheme is done in section III. Simulation results and discussions are presented in section IV and concluding remarks are put forward in section V.

II. NON COHERENT CHAOS-BASED COMMUNICATION SYSTEMS

A. DCSK Communication System

In the DCSK modulator, each bit $b_i = \{-1, +1\}$ is represented by two sets of chaotic signal samples; the first is allocated to the reference, and the second to the data-carrying symbols. If $+1$ is transmitted, the data-bearing sequence is equal to the reference sequence, and if -1 is transmitted, an inverted version of the reference sequence is used as the data-bearing sequence. The spreading factor in DCSK systems is defined as the number of chaotic samples used to spread each bit and is presented by 2β, where β is an integer. Moreover, $T_{DCSK} = 2T_b = 2\beta T_c$ is the DCSK bit duration and T_c is the chip duration. During the i^{th} bit time interval, the discrete baseband signal at the output of the transmitter $s_{i,k}$ can be given by

$$s_{i,k} = \begin{cases} x_{i,k}, & 0 < k \leq \beta, \\ b_i x_{i,k-\beta}, & \beta < k \leq 2\beta, \end{cases} \tag{1}$$

where $x_{i,k}$ is the chaotic sequence used as the reference signal and $x_{i,k-\beta}$ is its delayed version. In order to demodulate the transmitted bits, the received signal is correlated with its delayed version and summed over half of the bit duration that is equal to T_b where $T_b = \beta T_c$. The received bits are then estimated by computing the sign of the correlator output. This system suffers from the noise being added onto its two signal components, then a noisy reference signal is correlated with a noisy information bearing signal at the receiver, consequently, the performance of the DCSK system deteriorates. This poor performance makes such modulation strategies undesirable for many future applications in UWB area.

B. NR-DCSK System Architecture

A block diagram of the general structure of NR-DCSK communication system is shown in Fig.1. As illustrated, the chaotic generator generates β/P samples at a frequency of $f_s = \frac{1}{PT_c}$ per sample. Each chaotic sample is replicated P times to obtain the reference signal with length β. Similar to DCSK system, if $+1$ is transmitted, the data-bearing sequence becomes equal to the reference sequence, and if -1 is transmitted, an inverted version of the reference sequence is used as the data-bearing sequence. Therefore, during the i^{th} bit time interval, the transmitted discrete baseband signal $e_{i,k}$ of the NR-DCSK system would be given by

$$e_{i,k} = \begin{cases} x_{i,\lceil \frac{k}{P} \rceil}, & 0 < k \leq \beta, \\ b_i x_{i,\lceil \frac{k}{P} \rceil - \beta}, & \beta < k \leq 2\beta, \end{cases} \tag{2}$$

where $\lceil \epsilon \rceil$ is the ceiling operator which represents the smallest integer not less than ϵ. It is important to note that the use of this design introduces redundant samples in the frame which is averaged at the receiver side to enhance the performance of the system. In this paper a commonly used channel with

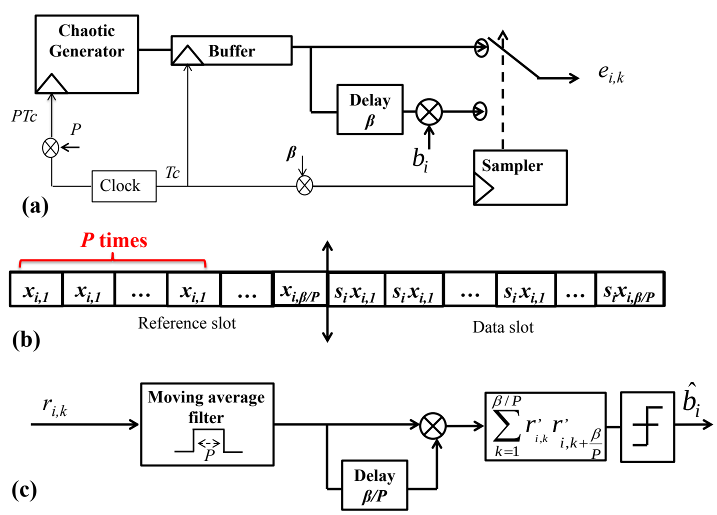

Fig. 1: Block diagram of the general structure of the NR-DCSK communication system; (a) transmitter, (b) frame structure and (c) receiver.

two independent paths called two-ray Rayleigh fading channel [11], [10] is considered. The channel coefficients and time delays between the two rays are denoted by α_1, α_2 and τ respectively. Besides, we assume that the channel coefficients are constant during a given NR-DCSK symbol. Therefore, the probability density function of the channel coefficient α in this case can be given as $f(\alpha|\sigma) = \frac{\alpha}{\sigma^2} e^{-\frac{\alpha^2}{2\sigma^2}}$, where $\sigma > 0$ is the scale parameter of the distribution representing the root mean square value of the received voltage signal before envelope detection. The received NR-DCSK signal corresponding to the transmission of the i^{th} bit could be written as

$$r_{i,k} = \alpha_1 e_{i,k} + \alpha_2 e_{i,k-\tau} + n_{i,k}, \tag{3}$$

where $e_{i,k}$ is the transmitted NR-DCSK signal and $n_{i,k}$ is additive white Gaussian noise with zero mean and variance $N_0/2$. As shown in Fig. 1, the received signal is first averaged using a moving average filter with a window size of P. In the second step, the resultant averaged signal is correlated with a replica that is time-delayed by a factor of $\frac{\beta}{P}$, then summed over the duration $\frac{\beta T_c}{P}$. The aim of using the moving

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average filtering operation at the receiver side is to reduce the noise variance such that a significant enhancement of the NR-DCSK performance is achieved. In our analysis, we assume that the delay τ is much shorter than the reference duration $0 < \tau << \frac{T_c\beta}{P}$, a case in which the inter-symbol interference (ISI) presence can be neglected [6], [10], [11]. For mathematical simplification, we consider a unity chip duration, i.e. $T_c = 1$. Based on this assumption, the decision variable at the output of the correlator of the i^{th} bit becomes

$$D_i = \sum_{k=1}^{\frac{\beta}{P}} \left(\alpha_{i,1} x_{i,k} + \alpha_{i,2} x_{i,k-\tau} + \frac{1}{P} \sum_{p=1}^{P} n_{i,p+k} \right) \left(\alpha_{i,1} b_i x_{i,k} + \alpha_{i,2} b_i x_{i,k-\tau} + \frac{1}{P} \sum_{p=1}^{P} n_{i,p+k+\frac{\beta}{P}} \right). \quad (4)$$

where $\frac{1}{P}\sum_{p=1}^{P} n_{i,p+k}$ is the noise signal after passing through the moving average filter. The summation over $\frac{\beta}{P}$ samples is due to the fact that after the averaging operation, the length of the reference and the data carrier signals is reduced from β to $\frac{\beta}{P}$. The decision variable may be further expanded as

$$\begin{aligned} D_i = &\sum_{k=1}^{\frac{\beta}{P}} \Bigg(\underbrace{\alpha_{i,1}^2 b_i x_{i,k}^2 + \alpha_{i,2}^2 b_i x_{i,k-\tau}^2}_{U} + \underbrace{2 b_i \alpha_{i,1}\alpha_{i,2} x_{i,k} x_{i,k-\tau}}_{I} \\ &\underbrace{+ (\alpha_{i,1} x_{i,k} + \alpha_{i,2} x_{i,k-\tau}) \frac{1}{P} \sum_{p=1}^{P} n_{i,p+k+\frac{\beta}{P}}}_{N_1} \\ &\underbrace{+ (\alpha_{i,1} b_i x_{i,k} + \alpha_{i,2} b_i x_{i,k-\tau}) \frac{1}{P} \sum_{p=1}^{P} n_{i,p+k}}_{N_2} \\ &\underbrace{+ \frac{1}{P} \sum_{p=1}^{P} n_{i,p+k+\frac{\beta}{P}} \frac{1}{P} \sum_{p=1}^{P} n_{i,p+k}}_{N_3} \Bigg), \end{aligned} \quad (5)$$

where U and I represent the useful and interference component signals generated from the multipath channel and the chaotic sequence, and N_1, N_2 and N_3 designate interference emerging from Gaussian noise.

Since the correlation between different chaotic samples is very low, the following expression is valid for large β/P values

$$\sum_{k=1}^{\beta/P} \left(x_{k-\tau} x_k \right) \approx 0 \cdot \quad (6)$$

Additionally, if the condition $\tau << \frac{T_c\beta}{P}$, i.e $P << \frac{T_c\beta}{\tau}$ is not respected, ISI becomes significant and the assumption given in (6) will no longer be valid. In effect, our performance analysis is based on this assumption to derive the lower bound BER performance of the system. In the simulation section, we discuss the limitations that hamper the validity of our assumption.

III. Performance Analysis of NR-DCSK System

In this section, the performance of NR-DCSK scheme is evaluated under AWGN and multipath fading channels. The decision variable scripted in (5) is decomposed into independent signal components U, I, N_1, N_2 and N_3. The independence is because channel coefficients are independent of each other, of the chaotic sequences and of the Gaussian noise. In addition, at any given time, the chaotic sequence is independent of its time delayed version and of the Gaussian noise as well [7]. In this work, the second-order Chebyshev polynomial function (CPF) is employed to generate chaotic sequences in the fashion shown below due to its easiness and good performance [7]

$$x_{k+1} = 1 - 2x_k^2 \cdot \quad (7)$$

The variance of the normalized chaotic map with zero mean is equal to one, i.e. $V[x] = \mathrm{E}[x^2] = 1$, where $\mathrm{E}[\cdot]$ denotes the expected value operator. Therefore, for the i^{th} bit, the instantaneous mean of the decision variable is the mean of the useful signal given by

$$E\left[D_i\right] = b_1(\alpha_1^2 + \alpha_2^2) E_r, \quad (8)$$

where $E_r = \sum_{k=1}^{\beta/P} E\left[x_{i,k}^2\right]$ represents the recovered bit energy at the output of the correlator. Equation (8) is significant because apart from the useful signal U, all other components described in (5) have zero mean. On the other hand, if the transmitted bit energy is $E_b = 2P \sum_{k=1}^{\beta/P} E\left[x_{i,k}^2\right]$, then $E_r = \frac{E_b}{2P}$. Based on (6) we can deduce that $I \approx 0$, so the conditional variance of the different remaining interference components present in the decision variable for the i^{th} bit might be expressed as

$$V\left[D_i\right] = V\left[N_1\right] + V\left[N_2\right] + V\left[N_3\right]. \quad (9)$$

On the other hand, the Gaussian noise signals present in N_1, N_2 and N_3 are the average of P independent and identically distributed Gaussian samples. Therefore, the variance of a random variable C_p with P i.i.d of c samples reduces [18] to

$$V[C_p] = \frac{V[c]}{P} \cdot \quad (10)$$

It is the P-fold reduction of the noise variance described in (10) that boosts the signal-to-noise ratio and enhances BER performance. Hence, based on (10), the variance of N_1, N_2 and N_3 will become

$$V[N_1] = V[N_2] = E_r \frac{N_0}{2P} \left(\alpha_1^2 + \alpha_2^2\right), \quad (11)$$

$$V[N_3] = \frac{\beta N_0^2}{4P^3} \cdot \quad (12)$$

Since bit energies (or chaotic chips) are deterministic variables, with the merit of central limit theorem the decision variable at the output of the correlator follows a Gaussian distribution. Therefore, the bit error probability may be represented by

$$\text{BER} = \frac{1}{2}\Pr(D_i < 0 | \, b_i = +1) + \frac{1}{2}\Pr(D_i > 0 | \, b_i = -1), \quad (13)$$

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which can be given as

$$\mathrm{BER} = \frac{1}{2}\mathrm{erfc}\left(\left[\frac{2\mathrm{Var}\left[D_i\right]}{\mathrm{E}\left[D_i\right]^2}\right]^{-\frac{1}{2}}\right), \tag{14}$$

where $\mathrm{erfc}(x)$ is the well-known complementary error function defined as $\mathrm{erfc}(x) \equiv \frac{2}{\sqrt{\pi}}\int_x^{\infty} e^{-u^2}du\cdot$ By virtue of equations (8), (11), and (12) above, the BER for the NR-DCSK scheme can be determined and stated as

$$\mathrm{BER} = \frac{1}{2}\mathrm{erfc}\left(\left[\frac{2N_0}{P\left(\alpha_1^2+\alpha_2^2\right)E_r} + \frac{\beta N_0^2}{2P^3\left(\alpha_1^2+\alpha_2^2\right)^2 E_r^2}\right]^{-\frac{1}{2}}\right). \tag{15}$$

After replacing E_r by E_b based on the relationship $E_r = \frac{E_b}{2P}$, and using $\gamma = \frac{(\alpha_1^2+\alpha_2^2)E_b}{N_0}$, the above equation reduces to

$$\mathrm{BER} = \frac{1}{2}\mathrm{erfc}\left(\left[\frac{4}{\gamma} + \frac{2\beta}{P\gamma^2}\right]^{-\frac{1}{2}}\right)\cdot \tag{16}$$

A high spreading factor is considered in this work, hence, the bit energy E_b can be assumed to be constant [20]. In such a case, and for L dissimilar Rayleigh-fading channels, the PDF of the instantaneous γ can be written as [21]

$$f(\gamma) = \sum_{l=1}^{L} \rho_l f\left(\gamma, \bar{\gamma}_l, 1\right), \tag{17}$$

where $\rho_l = \prod_{j=1, j\neq l}^{L} \frac{\bar{\alpha}_l}{\bar{\gamma}_l - \bar{\gamma}_j}$ and L is the number of paths. We define $\bar{\gamma}_l$ as the average value of the instantaneous SNR on the l^{th} channel that is given by $\gamma_l = \alpha_l^2 E_b/N_0$. Equation (16) gives the theoretical BER benchmark of NR-DCSK system over multipath fading channels. Based on this, it is essential to note that the channel keeps changing at every transmission instant, so the average BER expression becomes

$$\overline{BER} = \frac{1}{2}\int_0^{\infty} \mathrm{erfc}\left(\left[\frac{4}{\gamma} + \frac{2\beta}{P\gamma^2}\right]^{-\frac{1}{2}}\right) f(\gamma)d\gamma\cdot \tag{18}$$

The analytical solution to equation (18) is difficult to derive for multipath Rayleigh fading channels, leaving numerical integration as the only way to perform average BER computation. Equation (16) can be extended to AWGN channel case by choosing $\alpha_1 = 1$, $\alpha_2 = 0$ and $\gamma = \frac{E_b}{N_0}$. By doing so, the BER expression under AWGN simplifies to

$$\mathrm{BER} = \frac{1}{2}\mathrm{erfc}\left(\left[\frac{4N_0}{E_b} + \frac{2\beta N_0^2}{PE_b^2}\right]^{-\frac{1}{2}}\right)\cdot \tag{19}$$

IV. Discussions and Simulation Results

In order to validate the BER performance of the NR-DCSK scheme proposed in this work and compare it to existing DCSK, I-DCSK and DPSK schemes, the computed BER expressions are verified and well justified by simulation results under AWGN and multipath fading channels. Fig. 2 shows the BER performance of the proposed NR-DCSK system for different values of P as well as for DCSK, I-DCSK and DPSK schemes. A careful examination of the performance curves shown in Fig. 2 reveals the extent to which simulation results perfectly validate the analytical BER expression provided in (19). As expected, the achieved enhancement in the performance of NR-DCSK is well proportional to the value of P,

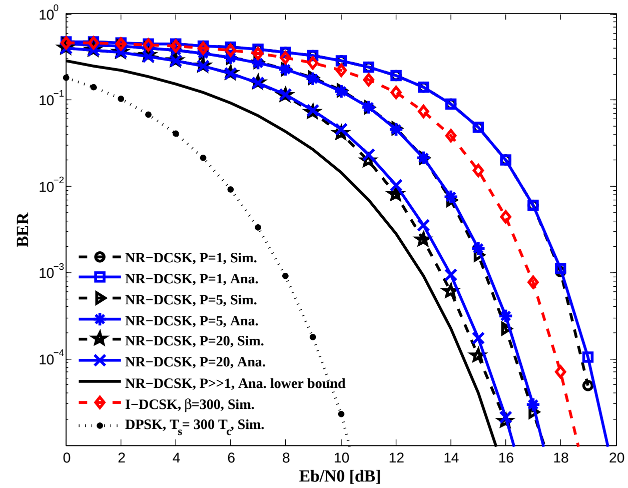

Fig. 2: Simulation, analytical and lower bound BER in AWGN channel of DPSK, DCSK (i.e $P = 1$), I-DCSK and NR-DCSK with a spreading factor of $\beta = 300$ for $P = 1$, 5 and 20.

i.e. the higher the value of P the better the performance of NR-DCSK. Note that the NR-DCSK with $P = 1$ reduces to the conventional DCSK. This enhancement is due to the averaging operation applied to the received signal which reduces the noise power as narrated in (10). Therefore, for higher values of P the performance of NR-DCSK tends to the lower bound BER performance given by $\mathrm{BER} = \frac{1}{2}\mathrm{erfc}([4N_0/E_b]^{-\frac{1}{2}})$. Hence, from the lower bound performance shown in Fig. 2, the maximum gain that can be achieved by NR-DCSK is 10 dB. Moreover, as predicted, the DPSK system results in a better performance over AWGN channel in comparison to the other rivals because such a *transmit reference* modulation class spends half of the bit energy on transmitting the reference signal, and the additive noise influences both reference and data carrier signals. In addition, the I-DCSK barely outperforms the NR-DCSK system for $P = 1$, (i.e $P = 1$ is equivalent to DCSK system) [19]. Based on these results, the NR-DCSK turns into a powerful tool and a potential candidate for many future UWB applications.

The obtained BER performance under Rayleigh multipath fading channels for a spreading factor of $\beta = 300$ is shown in Fig. 3. The average power gains of the first and second paths are $E[\alpha_1^2] = 0.7$ and $E[\alpha_2^2] = 0.3$ respectively with a delay of $\tau = 5T_c$ between the two paths. Likewise, simulation results warrant the accuracy of the BER expression given in (18) for lower and average values of P. Therefore, for higher values of P a gap is observed between the computed and simulated BER performance. This gap is due to the fact that the $P << \frac{\beta T_c}{\tau}$ is not respected which renders the assumption given in equation (6) unreasonable. Therefore, as observed in Fig. 3, the proposed NR-DCSK system outperforms the conventional DCSK even for higher values of P, for instance for $P = 20$ which indicates the presence of interference. In fact, when P increases the noise variance decreases and the performance of NR-DCSK enhances. Furthermore, without any additional receiver complexity, NR-DCSK gives better performance than the conventional system. Finally, the performance of I-DCSK

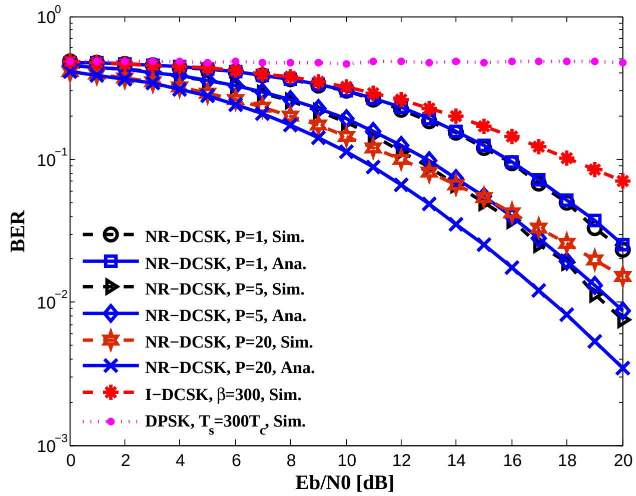

Fig. 3: Simulation and computed BER under multipath Rayleigh fading channels of DPSK, DCSK, I-DCSK and NR-DCSK with $\beta = 300$ for $P = 1,\ 5$ and 20.

and DPSK systems over the same channel profile are plotted. Results show that the I-DCSK system performs as good as the DCSK but with double spectral efficiency than this latter [19]. In fact, the multipath channel significantly increases the number of cross terms interference in I-DCSK which degrades its overall signal-to-noise ratio. Moreover, we can clearly see that when the signals from the two paths are out of phase, the NR-DCSK system shows higher resistance for this propagation environment and outperforms the conventional DPSK. A detailed analysis of the performance behaviour of DPSK system is given in [11].

V. Conclusions

A noise reduction DCSK system architecture has been proposed in this paper. This new architecture allows the application of moving average filter to increase the SNR of the received signal and enhance the system performance. Instead of generating β chaotic samples of each transmitted bit, to be used as reference and data carrier sequences, in the new system proposed here, only β/P samples are generated and then each sample is duplicated P times. The value of P represents the number of points in the averaging process.

The collected signal at the receiver is filtered, then the resultant signal is correlated with its delayed replica, located to recover the transmitted data.

Moreover, the performance of the NR-DCSK is analytically studied and the general bit error rate expression under multipath fading channels is derived. The analytical BER expression is then simplified to suit the AWGN channel scenario. Computer simulations are carried out to confirm the derived analytical results. Simulation results show that the proposed system outperforms I-DCSK, and DCSK for $P > 1$. Finally, motivated by the ambitious results of the promising new architecture obtained in this work, NR-DCSK is expected to be considered a serious candidate for UWB platforms.

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