A MODIFIED DIRECT-SEQUENCE SPREAD SPECTRUM MODULATION SCHEME FOR BURST TRANSMISSIONS

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ABSTRACT

The Direct-Sequence Spread Spectrum modulation (DSSS) is a very attractive modulation technique for military communication systems, mainly due to its resistance to interference and low probability of detection. However, the synchronisation to the spreading code phase in the receiver takes time, which makes this technique less interesting for sending very short messages. In this paper a novel direct-sequence modulation scheme for spread spectrum communication systems is presented, defined as the Delay-and-Add-Direct-Sequence (DADS) modulation. The DADS spread spectrum modulation is easy to implement and there is no more need for code alignment at the receiver side, which makes it very suitable for short burst transmissions. The performance of the DADS technique is studied in Additive White Gaussian Noise (AWGN) and flat Rayleigh channels, as well as the influence of some important parameters in the modulation scheme.

1. INTRODUCTION

Since the 1980s the spread spectrum technology has become increasingly popular, not only for military, but also for commercial applications. In a spread-spectrum communication system, the information signal is deliberately spread in the frequency domain to prevent detection and to increase resistance to jamming. One of the most popular spreading technique utilises a direct sequence spreading code, called Direct-Sequence Spread-Spectrum (DSSS). DSSS can be seen as a modulation which is applied on top of a conventional modulation such as Binary Phase Shift Keying (BPSK). The bi-polar data bits are multiplied with a spreading code prior to the PSK modulation. This spreading code is generally a pseudo-random bit sequence (PRBS), with a chip rate much higher than the bit rate of the data signal, thereby spreading the energy of the original data signal into a much wider band.

The main benefits of the DSSS modulation are the resistance to intended or unintended interference, the Low Probability of Detection and Interception and the fact that this technique allows the sharing of the same frequency band among multiple users by means of Code Division Multiple Access (CDMA). These nice properties make DSSS a popular spreading technique which is used in a lot of commercial wireless communication systems like Wireless LAN (IEEE802.11b), Wireless PAN (Bluetooth, IEEE802.15.4), satellite-positioning systems (GPS, Galileo), 3G mobile telecommunications (UMTS), and so on.

The main drawback of DSSS lies in the complexity of the receiver. Indeed, along with the usual carrier tracking as in a traditional digital receiver, a code alignment is needed for a correct despreading. The carrier and code tracking are respectively done by a Phase-Locked Loop (PLL) and a Delay-Locked Loop (DLL). The DLL is normally based on an early-late-prompt sampler. For the PLL, a Costas loop can be used. The particularity of the two loops however is that they are entwined, meaning that the output of one loop serves as input of the other and vice-versa. A second drawback, which is directly connected to the first, is the time needed to synchronise on the spreading code. In fact, to bring the DLL in its domain of operation, a first rough code phase estimation is needed. This coarse estimation is performed by an acquisition block, and is based on the evaluation of an ambiguity function. As a consequence, a long preamble is advisable, which makes the DSSS modulation less suitable for transmission of short messages.

In this paper, a novel direct-sequence modulation scheme, defined as the Delay-and-Add-Direct-Sequence (DADS), is presented. The DADS modulation scheme has lower complexity than the DSSS modulation scheme and is particularly suited for short burst transmissions. The transmitted signal is composed of an DSSS modulated signal added together with a delayed version of the spreading code, as in traditional *transmit reference* (TR) methods. This delayed version, embedded in the transmitted signal, is exploited for the despreading at the receiver. As a consequence, there is no more need for code synchronisation at the receiver side, resulting in a very simple receiver structure. DADS is inspired on non-coherent chaos based modulation schemes [1, 2], introduced in 2000 by Mikhail Sushchik. However, instead of a multilevel chaotic source, a PRBS is used for the spreading in the DADS modulation scheme.

The structure of the DADS based transmitter and receiver is detailed in section 2 of this paper. In section 3, a detailed performance analysis of the modulation technique in AWGN and flat Rayleigh channels is performed. An analytical expression of the BER is derived and compared with simulation results. In section 4, a parametric study is presented to show the influence of the delay and the code length on the BER performance.

2. SYSTEM DESCRIPTION

2.1. Structure of the transmitter

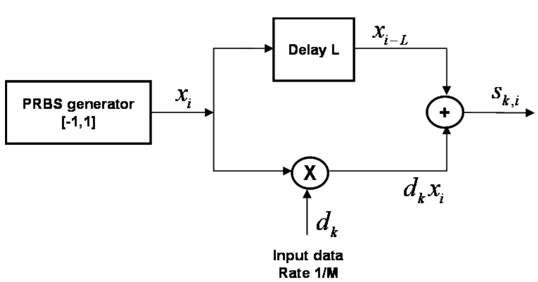

Figure 1: Structure of the DADS modulator

The generic block diagram of the DADS modulator is illustrated in Figure 1. In the lower branch, a bipolar spreading code $\boldsymbol{x} = \{x_i\}$ with length M is multiplied with the information signal $d_k \in \{-1, +1\}$ with data rate $\frac{1}{M}$. In the upper branch, the same spreading code is delayed by L, with $L \ll M$. The two signals are added together, resulting in a signal $\boldsymbol{s_k} = \{s_{k,i}\}$ with

$$s_{k,i} = d_k x_i + x_{i-L}. \quad (1)$$

The signal $\boldsymbol{s_k}$ will be up-converted by the RF front-end and send out.

The base band signal $\{s_{k,i}\}$ has two important characteristics. First of all, $s_{k,i}$ is a three level signal $\in \{+2, 0, -2\}$. The second characteristic is of essential importance for the receiver. Every 2 non-zero samples $s_{k,i}$, separated by L, will have the same sign if the data bit $d_k = +1$ and will have opposite signs if $d_k = -1$. This second property is illustrated in Figure 2. Figure 2(a) shows the first 15 samples of a modulated signal $\{s_{k,i}\}$ for $d_k = +1$ and a delay $L = 2$. Every non-zero sample will have, 2 samples to the left and to the right, a value that equals zero, or a value with the same sign, e.g. $s_{k,2} = s_{k,4}$ or $s_{k,9} = s_{k,11}$. Figure 2(b) shows the same for $L = 2$ and $d_k = -1$. In this case, every 2 non-zero samples, separated by 2, will always have opposite signs, e.g. $s_{k,1} = -s_{k,3}$. This important property will be used for the demodulation.

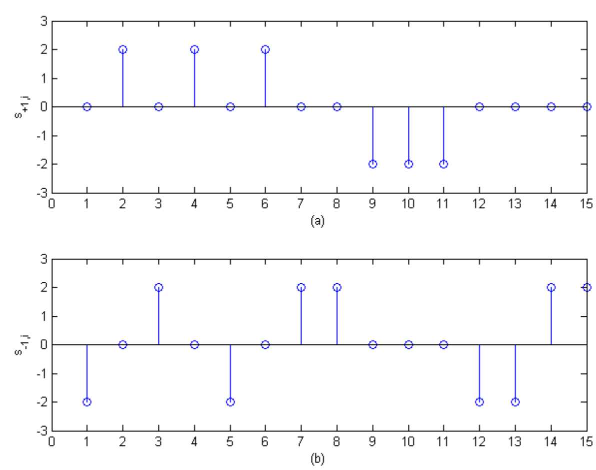

Figure 2: Modulated signal $\boldsymbol{s_k}$ with a delay $L = 2$: (a) structure of $\boldsymbol{s_k}$ for $d_k = +1$, (b) structure of $\boldsymbol{s_k}$ for $d_k = -1$

2.2. Structure of the receiver

The principle of the DADS demodulator is represented in Figure 3. In a first approach, the received baseband signal $\boldsymbol{r_k} = \{r_{k,i}\}$ is considered to be real-valued. The received signal $\boldsymbol{r_k}$ will be delayed by the same

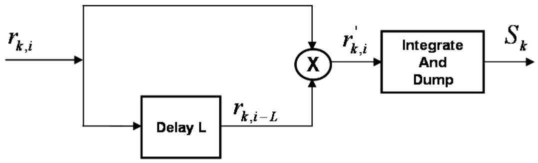

Figure 3: Principle of the DADS demodulator

delay L as in the transmitter and multiplied with its non-delayed version, resulting in a signal $\left\{r'_{k,i}\right\}$, with

$$r'_{k,i} = (d_k x_i + x_{i-L})(d_k x_{i-L} + x_{i-2L}) \tag{2}$$
$$= d_k \left(x_{i-L}^2 + x_i x_{i-2L}\right) + x_{i-L}(x_i + x_{i-2L})$$

It can be verified that $r'_{k,i}$ has only two possible values $\in \{4d_k, 0\}$. This is a direct consequence of the second characteristic of $\boldsymbol{s_k}$, as described in the section 2.1. In case that $\{x_i\}$ is a PRBS, meaning an equal probability of $+1$ and -1, and no correlation between $\{x_i\}$ and $\{x_{i-L}\}$, it can be verified that

$$r'_{k,i} = \begin{cases} 4d_k & \text{in 25 \% of the cases} \\ 0 & \text{in 75 \% of the cases} \end{cases} \tag{3}$$

The signal S_k, after the integrate and dump equals

$$S_k = \sum_{i=1}^{M} r_{k,i} r_{k,i-L}$$
$$= d_k \sum_{i=1}^{M} x_{i-L}^2 + d_k \sum_{i=1}^{M} x_i x_{i-2L}$$
$$+ \sum_{i=1}^{M} x_{i-L}(x_i + x_{i-2L}) \tag{4}$$

The first term in (4) is the useful signal. The second and the third term tend to zero, as there is no correlation between x_i, x_{i-L} and x_{i-2L} for any i. By applying the correlation criterion of a PRBS [3], stating that the Periodic Autocorrelation function $R_{x,x}(s) = -1/M$ for $s \neq 0$ and M the code length, the expected value of S_k, for $M \gg 1$, can be further simplified to

$$E[S_k] = d_k M - d_k - 2$$
$$= d_k(M-1) - 2$$
$$\approx d_k M \tag{5}$$

This result confirms the statement in (3).

Let us now consider some practical implementation issues. First of all, it has to be noted that in this paper we consider a digital implementation of the receiver, based on SDR technology. Hence the received baseband signal $\boldsymbol{r_k} = \{r_{k,i}\}$ is a digitised signal with a given oversampling rate. In a classical digital receiver, the received signal will be down converted to baseband using an IQ-demodulator. The carrier phase recovery, needed for the coherent demodulation, is performed by a Costas Loop. In the DADS receiver, the same approach using a Costas Loop can be used. The Costas Loop will put all the energy of the down converted baseband signal in the I channel, resulting in real baseband signal $\boldsymbol{r_k}$. In this case, the demodulator as presented in Figure 3 can be used. However, with some minor modifications, the DADS receiver can also work without a carrier recovery circuit. Suppose the received signal is down converted to baseband using a non synchronised local carrier. In this case, the baseband signal $\boldsymbol{r_k}$ will be a complex signal modulated with a very low frequency component equal to the frequency difference between the received carrier and the local carrier. If the phase drift $\Delta\phi_L$ between the two carriers, due to this frequency offset, over the delay period L is small, a slightly modified demodulator structure as represented in Figure 4 can be used. In this demodulator, the complex conjugate of the delayed baseband signal is taken prior to the multiplier and only the real part of S_k is considered. Doing so, the useful signal as defined in (4) will slightly degrade by a factor $\cos(\Delta\phi_L)$. However, for $\Delta\phi_L \ll \pi/2$, this factor $\cos(\Delta\phi_L)$ is close to unity and (5) will still hold. In practice, the condition on the phase drift over the delay period L is not too hard to satisfy. For instance, for a chip rate of 2 Mchips/s and $L = 2$ chips, a frequency offset of $10kHz$ between the received carrier and the local carrier, will degrade the useful signal in (4) only by a factor $cos(\Delta\phi_L) = 0.998$. A

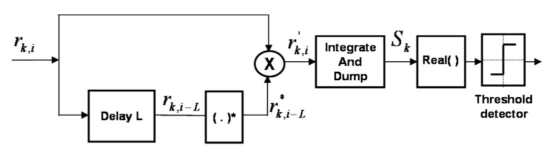

Figure 4: Practical implementation of the DADS demodulator

second practical consideration is on the delay L. In a digital implementation, the delay line implementation is straight forward using a shift register. The only concern could be the stability of the clocks yielding to slightly different delays in transmitter and receiver. To evaluate the influence of this difference between the two delays, some simulations were performed in Matlab. With a clock stability of 10^{-6} at both transmitter and receiver side, which is not too hard to meet, the simulation results revealed no significant degradation of the BER performance.

As a conclusion we can say that the structure of the DADS receiver can be kept very simple. First of all, there is no more need for spreading code alignment, as the reference signal is integrated in the received signal. Further, it is shown that under certain conditions there is no more need for a Phase Locked Loop (PLL) to perform the carrier recovery. As a consequence, the implementation of the DADS receiver can be kept very simple, even simpler than the implementation of a BPSK receiver.

The fact that the receiver structure is simple can also be considered as a drawback. Although DADS is a Low Probability of Detection (LPD) modulation, it is very easy to demodulate the transmitted signal once it is detected, as the spreading code does not have to be known at the receiver side. The only parameter needed for the demodulation is the delay L, which is easy to extract from the captured signal by performing an autocorrelation. This disadvantage has to be considered when using this type of modulation in a military context.

3. PERFORMANCE IN AWGN AND FLAT RAYLEIGH CHANNELS

In this section, the performance of DADS is studied in the presence of noise and flat Rayleigh fading. The channel model is straight forward. The received signal $\boldsymbol{r_k} = \{r_{k,i}\}$ is given by

$$r_{k,i} = h_i(d_k x_i + x_{i-L}) + n_i, \tag{6}$$

with h_i the complex-valued channel attenuation and n_i a band limited Additive White Gaussian Noise (AWGN) with variance of $N_o/2$ per dimension. For simplicity, we consider no carrier frequency offset and we assume that the Rayleigh channel is a block fading Rayleigh channel, which can be represented mathematically as $h_i = h_k, \forall i$ during one data bit period.

The output of the Integrate and Dump can now be written as

$$\begin{aligned} S_k &= \sum_{i=1}^{M} r_{k,i} r^*_{k,i-L} \\ &= \underbrace{d_k |h_k|^2 \sum_{i=1}^{M} x^2_{i-L}}_{\text{useful part}} + \underbrace{d_k A + B}_{\text{noise and interference part}}, \end{aligned} \tag{7}$$

with

$$A = \sum_{i=1}^{M} \left(|h_k|^2 x_i x_{i-2L} + h_k x_i n^*_{i-L} + h^*_k x_{i-L} n_i\right)$$

and

$$\begin{aligned} B = \sum_{i=1}^{M} (&|h_k|^2 x_{i-L} (x_i + x_{i-2L}) \\ &+ h_k x_{i-L} n^*_{i-L} + h^*_k x_{i-2L} n_i + n_i n^*_{i-L}) \end{aligned}$$

As in (4), the first term in (7) represents the useful signal, whereas the second and third term represent zero-mean random quantities, interfering with the useful signal. From (7) and by calculating the variance of A and B, an analytical expression for the bit error rate can be derived.

3.1. Performance in AWGN channel

To evaluate the BER performance of DADS in an AWGN channel, the channel coefficients h_k in (7) are set to 1. The received signal and the Additive White Gaussian Noise, with variance $N_o/2$, can be considered as real-valued.

For $M \gg 1$, the distribution of the useful part in (7), as well as A and B approaches a zero-mean Gaussian distribution. For a given value of d_k, the useful part can be written as

$$\sum_{i=1}^{M} x^2_{i-L} = M E[x^2_{i-L}] = M \sigma^2_{x_i} = \frac{E_b}{2} \tag{8}$$

with E_b the energy in one data bit. Indeed, a data bit is composed of the sum of two spreading sequences with length M, hence the energy per bit is $2M$ times the energy per chip.

Because x_i is statistically independent from n_j for any (i, j) and n_i is statistically independent from its delayed versions, the cross-correlation of A and B is zero. Therefore, for a given value of d_k, the variance of the noise and interference part can be written as

$$\begin{aligned} \sigma^2_{A+B} &= \sigma^2_A + \sigma^2_B \\ &= 4 \frac{E_b}{2} \frac{N_o}{2} + \frac{3}{M} \left(\frac{E_b}{2}\right)^2 + M \left(\frac{N_o}{2}\right)^2 \end{aligned} \tag{9}$$

The bit error rate (BER) formula for a given value of

d_k is given by [4]

$$\begin{aligned} BER =& P\left((A+B) > \frac{E_b}{2}\right) \\ =& Q\left(\sqrt{\frac{(\frac{E_b}{2})^2}{\sigma^2_{A+B}}}\right) \\ =& \frac{1}{2}\text{erfc}\left(\sqrt{\frac{E_b}{8N_o}\left(1+\frac{MN_o}{4E_b}+\frac{3E_b}{4MN_o}\right)^{-1}}\right) \end{aligned} \quad (10)$$

with Q the Gaussian tail probability and $\text{erfc}(x)$ the complementary error function.

Figure 5 shows the theoretical BER curves given by (10) as a function of E_b/N_o for different values of the spreading factor M, together with results of a baseband simulation of the DADS modulation scheme performed in Matlab. It can be seen that the resemblance between the simulation results and the theoretical curves is very good. Note that the BER curves are not given as a function of E_c/N_o, with E_c the energy per chip, and hence do not take into account the processing gain $G[dB] = 10 \log M$ of the spread spectrum.

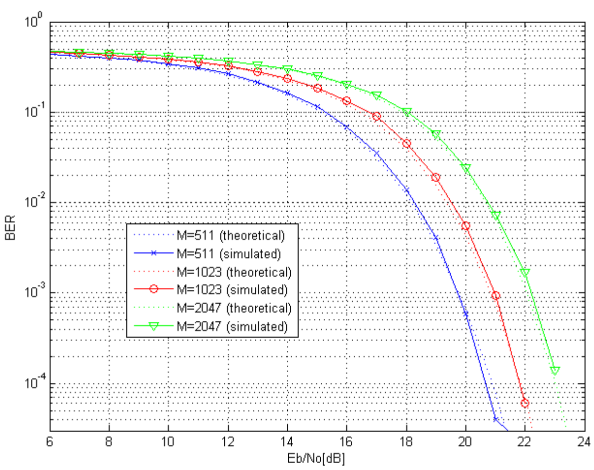

Figure 5: BER performance of DADS in an AWGN channel: comparison between theoretical curves and simulations

3.2. Performance in a flat Rayleigh fading channel

For the BER performance analysis of DADS in a Rayleigh channel, the channel coefficients h_k in (7) are considered to be complex valued. Both real and imaginary part of h_k are uncorrelated Gaussian random variables. For a given value of d_k and a given channel realisation h_k the useful part becomes

$$|h_k|^2 \sum_{i=1}^{M} x^2_{i-L} = |h_k|^2 M E[x^2_{i-L}] = |h_k|^2 \frac{E_b}{2} \quad (11)$$

The variance of the noise and interference part for a given value of d_k and a given channel realisation can be written as

$$\begin{aligned} \sigma^2_{A+B} =& \sigma^2_A + \sigma^2_B \\ =& 4|h_k|^2 \frac{E_b}{2}\frac{N_o}{2} + \frac{3}{M}|h_k|^4 \left(\frac{E_b}{2}\right)^2 + 2M\left(\frac{N_o}{2}\right)^2 \end{aligned} \quad (12)$$

A semi-analytical approach to evaluate the BER performance in a Rayleigh channel is to generate a large number of channel realisations and to estimate the average BER by averaging the conditional BER over these realisations [5]. The bit error rate (BER) formula can then be written as

$$\begin{aligned} BER =& E\left[P\left((A+B) > \frac{E_b}{2}\right)\right] \\ =& E\left[Q\left(\sqrt{\frac{|h_k|^4(\frac{E_b}{2})^2}{\sigma^2_{A+B}}}\right)\right] \\ =& \frac{1}{2}E\left[\text{erfc}\left(\sqrt{\frac{|h_k|^2 E_b}{8N_o}\left(1+\frac{MN_o}{2|h_k|^2 E_b}+\frac{3|h_k|^2 E_b}{4MN_o}\right)^{-1}}\right)\right] \end{aligned} \quad (13)$$

Figure 6 shows the theoretical and simulated performance of the DADS modulation in a flat Rayleigh channel for different values of the spreading factor M. It can be seen that the resemblance between the simulation results and the theoretical curves given by (13) is very good.

4. INFLUENCE OF DELAY L AND SPREADING FACTOR M

In the DADS modulation scheme, there are two parameters that can be tuned, e.g. the delay L and the spreading factor M. The influence of these parameters on the performance of the modulation scheme are studied in this section.

4.1. Influence of the Spreading Factor M

Just like in the classical spread spectrum techniques, the spreading code has a direct influence on the processing gain G, which is given by $G[dB] = 10 \log M$. However, contrary to DSSS, the spreading factor has also a negative influence on the BER performance.

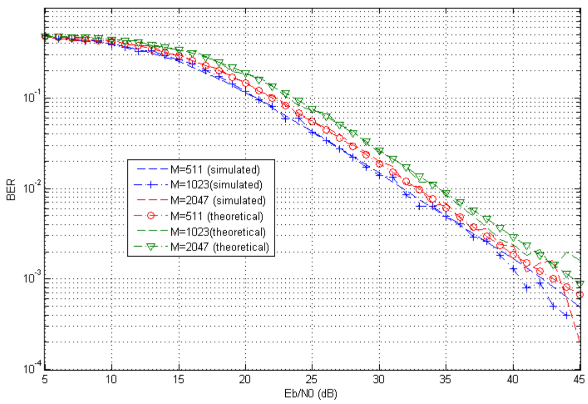

Figure 6: BER performance of DADS in a flat Rayleigh channel: comparison between theoretical curves and simulations

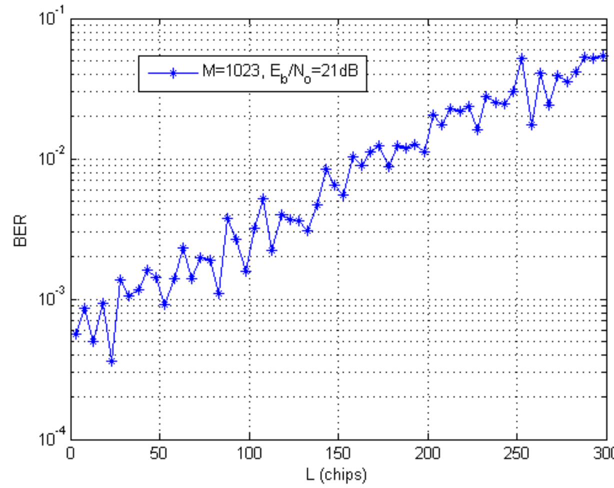

Figure 7: Influence of the delay L on the BER performance for $M = 1023$ and $E_b/N_o = 21dB$

This is clearly visible in Figure 5 and 6. If the value of M doubles, the BER curve is shifted about 1dB to the right. This degradation however is compensated by the increase of the processing gain by 3dB, meaning that a gain of about 2.0 dB can still be achieved by increasing the length of M by a factor 2. The performance degradation can intuitively be explained by the fact that the despreading is done using a delayed copy of the spreading code, which is also affected by the noise. It is mainly the last term in (12), which is proportional to M, that will increase the variance of the noise part and hence decrease the BER performance. Simulations showed that the second term in (12), proportional to $1/M$, only influences the BER floor. As the BER floor is far below 10^{-5}, this can not be seen on Figure 5 and 6.

4.2. Influence of the delay L

An increase of L will increase the overlap between two adjacent data bits in the receiver. Up till now, the influence of this inevitable inter symbol interference has been neglected in all previous calculations, as $L \ll M$. For larger values of L however, the impact on the performance of the modulation scheme becomes obvious. Figure 7 shows the influence of the delay L on the BER performance for $M = 1023$ and $E_b/N_o = 21dB$. It can be seen that the BER performance degrades with the delay L. Therefore, it is better to keep L as low as possible in a practical implementation. Note that the DADS modulation scheme can also be used as a multiple access technique. In this case, each user pair has to have a different delay L to demodulate its data information.

5. CONCLUSION

In this paper, a modified direct-sequence modulation scheme for spread spectrum burst transmissions, the Delay-and-Add-Direct-Sequence, is proposed. The main drawback of DSSS for burst transmissions is the time needed to perform the code synchronisation. In DADS, a copy of the spreading code, delayed by L, is embedded in the transmitted signal. At the receiver, the signal is again delayed by L and multiplied with the non-delayed signal for despreading, without the need for code synchronisation. This results in a very simple receiver structure, without loosing the benefits of a spread spectrum modulation, like e.g. the processing gain, Low Probability of Detection and interference resistance. An analytical expression of the BER performance in an AWGN and flat Rayleigh channel is derived and compared to simulation results. The resemblance between the simulation results and the theoretical curves is very good. In the last part, the influence of the two parameters, M and L, are discussed. Doubling the code length M will increase the processing gain but in the mean time increase the noise in the demodulator, resulting in a 2 dB gain. It is also shown that the BER performance degrades with the delay L. Therefore, it is better to choose L as low as possible. Future studies will evaluate the performance of the DADS modulation scheme with multiple users in a multiple access scheme considering the length of the spreading code M and the delay L.

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