Performance Improvement of QPSK Signal Predetection EGC Diversity Receiver

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Abstract. *This paper proposes a modification of quadrature phase-shift-keying (QPSK) signal diversity reception with predetection equal gain combiner (EGC). The EGC combining is realized by using the constant modulus algorithm (CMA). Carrier synchronization is performed by the phase locked loop (PLL). Comparative analysis of the modified and ordinary diversity receiver in the presence of carrier frequency offset in the additive white Gaussian noise (AWGN) channel, as well as in Rician fading channel is shown. The proposed diversity receiver allows significant frequency offset compared to the diversity receiver that uses only PLL, and the error probability of the proposed receiver is very close to the error probability of the receiver with only PLL and zero frequency offset.*

The functionality of the proposed diversity receiver, as well as its properties is experimentally verified on a system based on universal software radio peripheral (USRP) hardware. The performed comparison confirms the expected behavior of the system.

Keywords

Equal gain combiner, constant modulus algorithm, phase locked loop, adaptive filtering.

1. Introduction

The wireless communication channels are characterized by multipath propagation. The signal at the input of the receiver contains direct line-of-sight component, and a large number of reflected signals. Reflected signals interfere with the direct one and cause the variation of instantaneous value of the received signal, i.e. fading. Fading is one of the main causes of performance degradation in wireless communication systems [1], [2].

Diversity reception is a communication receiver technique that provides wireless link improvement at relatively low cost by combating the harmful effect of channel fading and increasing the communication reliability without enlarging either transmitting power or bandwidth of the channel. Diversity reception using spatially separated receiver antennas is certainly one of the most frequently used diversity techniques. Particular diversity methods and combining techniques are presented in [3]–[5]. In equal gain combiner (EGC), the received signals from all branches are co-phased in order to eliminate the random signal phase fluctuations occurring during transmission, equally weighted and then summed to form the decision variable. Since the EGC technique achieves performances comparable to the ones of the maximal-ratio combining (MRC) technique (which is an optimal technique under the condition of perfect co-phasing), but with lower implementation complexity, this technique has received a great interest in the literature [5]–[8]. In papers [6]–[8] performance analyses of EGC diversity receivers in various fading environments have been performed.

In practice, the equal gain combiner is often realized using constant modulus algorithm (CMA). Constant modulus (CM) based algorithms have been widely used for both blind beamforming, multiuser detection and equalization [9]–[14]. Many of the algorithms are based on the stochastic gradient method. However, this class of CM algorithms has a drawback in their sensitivity to stepsizes. With a large stepsize the algorithm convergence speed is faster, but there is a loss of the output signal-to-interference plus noise ratio and the stability of the algorithm is degraded. Many variations of the algorithm have been proposed to combat this problem, such as the least squares CMA (LSCMA) [15], [16], the one based on recursive least squares (RLS) [17]–[19] and those with variable step sizes [20]–[23].

In this paper we propose a modification of the EGC diversity receiver, using CMA for cophasing and phase locked loop (PLL) for carrier synchronization. The modification is in introducing a structure that performs the estimation with remodulation (ER). The proposed receiver can operate within a wider carrier frequency offset range with a very small variation of the performance. The simulation results will be verified with the experimental results on USRP (Universal Software Radio Peripheral) hardware.

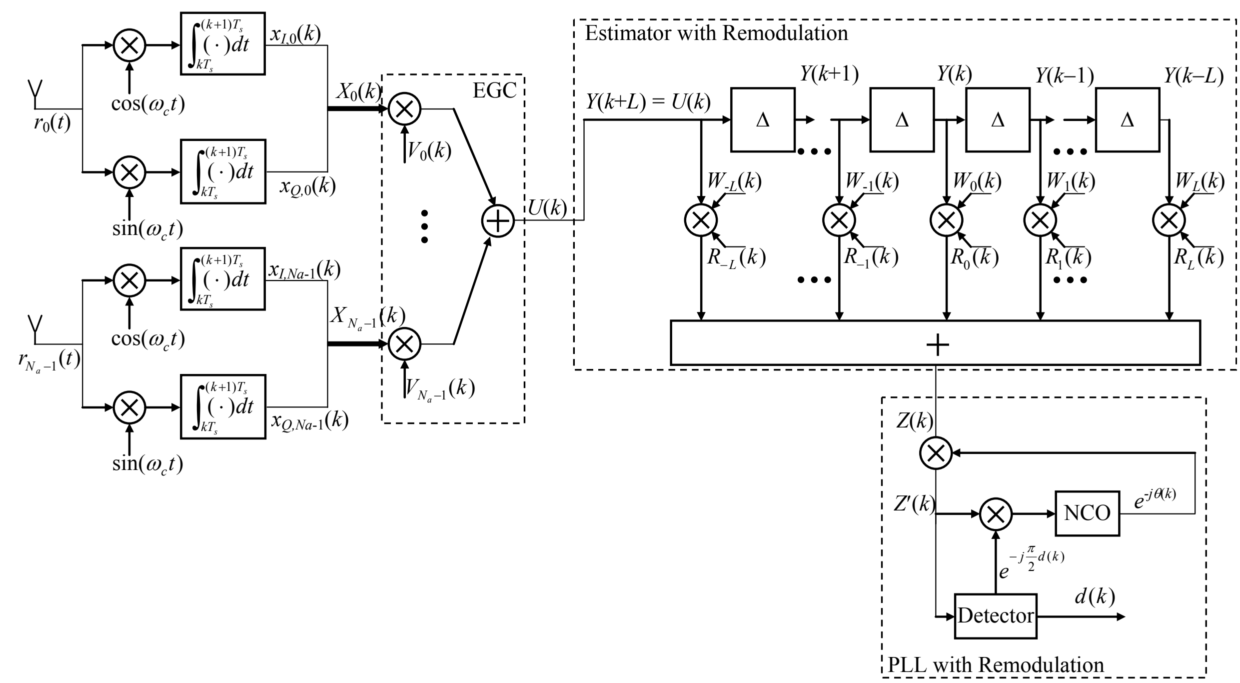

Fig. 1. Receiver block diagram.

2. System Model

A block diagram of the proposed quadrature phase shift keying (QPSK) signal EGC receiver with PLL detector is shown in Fig. 1. In case of low variance of the noise at the input of the PLL, the PLL bandwidth may be higher which provides lower sensitivity to the frequency offset. The basic idea is to process the signal after the EGC such that the noise variance is lowered and the useful signal is damaged as little as possible. In this paper we use the ER structure which fulfils the aforementioned criteria. ER is placed between EGC and PLL. In the proposed diversity receiver ER structure and PLL make a block which will be referred to as ER+PLL. If the ER structure is omitted, we get an ordinary EGC diversity receiver [1].

The signal at the output of each of the N_a receiving antennas is:

$$r_n(t) = s_n(t) + z_n(t),\ \ n = 0,1,...,N_a - 1 \tag{1}$$

where $z_n(t)$ is the white Gaussian noise, and $s_n(t)$ is the QPSK signal with a rectangular symbol pulse shape:

$$s_n(t) = e^{j\theta(t-\tau_n)} e^{j\hat{\omega}_c(t-\tau_n)}, \tag{2}$$

with

$$\theta(t) = \frac{\pi}{2} d(t) \tag{3}$$

where τ_n is the delay at the n-th path, and symbol $d(t)$ has one of the following four values:

$$d(t) \in \{0,1,2,3\},\ \ kT_s \le t < (k+1)T_s,\ \ k = 0,1,2... \tag{4}$$

The symbol interval is T_s, $\hat{\omega}_c = \omega_c + \Delta\omega$ is the carrier frequency at the input of the receiver, ω_c is the locally generated fixed reference carrier frequency, $\Delta\omega$ is the frequency offset.

The input signal is multiplied by the fixed frequency reference carrier and passed through the integrate and dump circuit. The complex baseband signal at the input of the EGC block can be expressed as

$$X_n(k) = x_{I,n}(k) + jx_{Q,n}(k)\,. \tag{5}$$

Signals at in-phase and quadrature branches are, respectively,

$$\begin{aligned} x_{I,n}(k) &= \int_{kT_s}^{(k+1)T_s} r_n(t)\cos(\omega_c t)dt, \\ x_{Q,n}(k) &= \int_{kT_s}^{(k+1)T_s} r_n(t)\sin(\omega_c t)dt, \end{aligned} \tag{6}$$

where k denotes the discrete time at the output of the integrate and dump circuit.

The signal at the output of the EGC combiner, with cophasing performed using CM algorithm, is

$$U(k) = \sum_{n=0}^{N_a-1} X_n(k)V_n(k) \tag{7}$$

where N_a is the number of diversity branches. $V_n(k)$ is the weight of the n-th branch obtained by the CMA and may be written as [9]

$$V_n(k) = V_n(k-1) + \mu_V\left(\frac{1}{|U(k-1)|} - 1\right)U(k-1)X_n^*(k-1) \tag{8}$$

where μ_V is the adaptation factor, and $(\cdot)^*$ denotes the complex conjugate.

The signal from the output of the EGC, $U(k)$, is led to the ER structure, which is an estimator with remodulation, and its operation is described with the following equations:

$$Z(k)=\sum_{l=-L}^{L}Y(k-l)R_l(k)W_l(k) \qquad (9)$$

where $2L$ is the ER structure length, and $Y(k+L)=U(k)$. The weights $W_l(k)$ are being adjusted by the Leaky LMS algorithm [24]:

$$W_0(k)=1,$$
$$W_l(k+1)=(1-\mu_W)W_l(k)+\mu_W Y(k)\cdot(Y(k-l)\cdot R_l(k))^*, \qquad (10)$$
$$l=-L,...,L, l\neq 0,$$

where μ_W is the adaptation factor, and $R_l(k)$ are remodulation weights which are determined as

$$R_0(k)=1,$$
$$R_l(k)=e^{j\frac{\pi}{2}m_l(k)}, l=-L,...,L, l\neq 0, \qquad (11)$$

where

$$m_l(k)=\underset{p\in\{0,1,2,3\}}{\arg\min}\left\{e^{j\frac{\pi}{2}p}Y^*(k-l)W_l^*(k)Y(k)\right\}. \qquad (12)$$

Carrier synchronization is performed by the second order remodualtion PLL [25]. In digital domain, remodulation PLL consists of the remodulation circuit and a numerically controlled oscillator (NCO). Phase difference in the NCO is equal to the argument of the remodulated signal:

$$\Delta\varphi(k)=\arg\left\{Z'(k)\cdot e^{-j\frac{\pi}{2}d(k)}\right\} \qquad (13)$$

where

$$Z'(k)=Z(k)\cdot e^{-j\theta(k)}, \qquad (14)$$

with $\theta(k)$ being the phase of the signal at the output of NCO for the k-th sampling interval. This parameter is going to be defined later.

The phase difference is filtered within the NCO using a first order low pass filter

$$e(k)=(1-A_{PLL})e(k-1)+A_{PLL}\cdot\Delta\varphi(k-1) \qquad (15)$$

where $e(k)$ is the filter output, and A_{PLL} is the filter constant.

The NCO frequency correction is obtained by:

$$\delta f(k)=\frac{1}{2\pi}\frac{K_{PLL}}{T_S}e(k) \qquad (16)$$

where K_{PLL} is the normalized gain used for the adjustment of the frequency correction value.

The phase $\theta(k)$ is obtained by the integration of frequency correction values:

$$\theta(k)=\theta(0)+\frac{2\pi}{T_S}\sum_{m=1}^{k}\delta f(m). \qquad (17)$$

The decision is made using the following minimization:

$$d(k)=\underset{p\in\{0,1,2,3\}}{\arg\min}\left\{e^{j\frac{\pi}{2}p}Z'^*(k)\right\}. \qquad (18)$$

The performance of the algorithm described in this section is determined by the Monte-Carlo simulation in the computer, and by an experiment. The experiment will be described in the following section.

3. Hardware Experiments

In the experiment, the baseband processing is performed in a PC, up/down conversion is performed in USRP hardware, and the communication channel is real. The block diagram of the experimental setup is shown in Fig. 2.

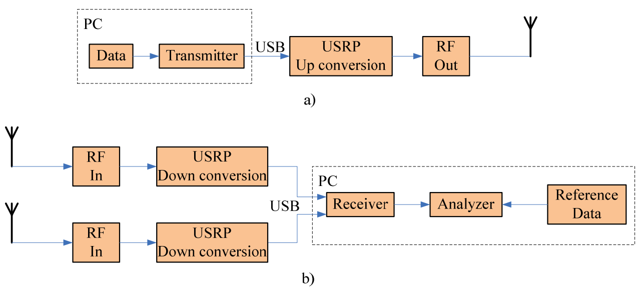

Fig. 2. Block diagram of experimental setup.

The PC part of the setup runs on Linux and is written in C++. In order to be able to compare the received and transmitted data bits, the block *Data* generates a pseudo-random sequence. The same sequence is generated within the *Reference Data* block. The block *Transmitter* performs baseband processing and generates QPSK modulated signal. QPSK signal is then transferred to USRP via USB interface. The communication at USB interface is performed using *libusb* library. USRP receives data from USB interface and performs digital to analog conversion and up-conversion to 2.4 GHz band. At the receiver chain, similar processing is performed. After down-conversion and analog to digital conversion in USRP, signal is transferred via USB interface to PC. The *Receiver* block performs demodulation and baseband processing. The received data are compared to the sent data in the *Analyzer* block.

4. Numerical Results

All results, which are presented here, are obtained for the detection of QPSK signals in AWGN and Rice fading channel. A diversity system with two branches ($N_a = 2$) is applied at the reception. CMA and ER adaptation factors, μ_V and μ_W, were kept constant for all simulations. The receiver has good performance for a wide range of the both adaptation factors values. We chose the values in the middle of that range, and the adaptation factors values are $\mu_V = \mu_W = 0.01$.

It is well known in the literature that there is always a compromise during the design of the PLL. The compromise is between the PLL performance in terms of the phase noise, and the ability of the PLL to quickly respond to the changes in the input signal. Therefore, if the PLL has good performance in steady state in case of zero frequency offset, then it will have poor performance in the presence of frequency offset, and vice versa. During the analysis we will determine the PLL parameters that make a balanced compromise between the error probability for zero frequency offset, and the ability to have satisfactory performance in a wide frequency offset range.

The value of the loop parameter $K_{PLL} = 0.2$ is chosen to support the ordinary operating mode of the PLL and Fig. 3 shows the performance curves for this value of K_{PLL}.

Fig. 3 shows the error probability as the function of the length of the proposed ER filter, for different values of amplification in PLL, A_{PLL}. There are two cases presented: with and without frequency offset. It may be noted that, in the absence of frequency offset, the existence of ER filter does not introduce any improvement in performance. However, in the case of non-zero frequency offset the proposed ER block brings a significant reduction in the error probability for filter lengths greater than 5. Only in the case of very narrow bandwidth of the loop ($A_{PLL} = 0.01$) the frequency offset disrupts its operation so much that performance is impaired, even when using the ER block.

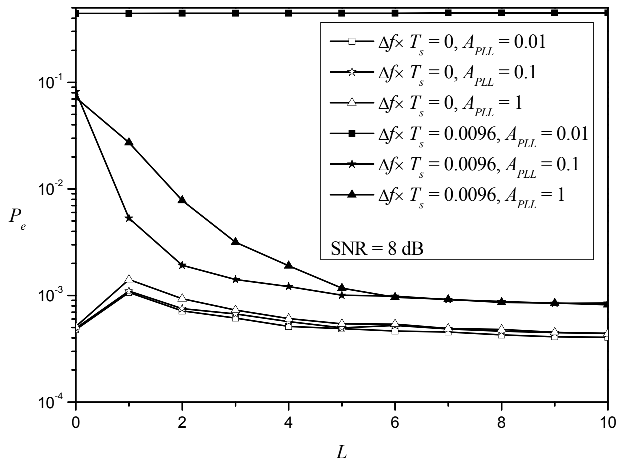

Fig. 3. Error probability as a function of filter length for $K_{PLL} = 0.2$ in AWGN channel.

The fact that the ER structure lowers the noise variance at its output, allows the use of a large value for K_{PLL}, as shown in Fig. 4 that depicts the error probability for $K_{PLL} = 1$. In Fig. 4 the error probability, as a function of the length of the proposed ER filter, is presented for different A_{PLL} values. Again, both cases, with and without frequency offset, are considered. It should be noted that for $L = 0$ the ER+PLL receiver comes down to PLL only receiver. As one can see, the PLL circuit itself has problems to operate in these conditions. By adding ER block, the performance is significantly improved. When $L > 5$, the values of error probability in the presence of frequency offset are approaching to the values without offset, and this error probability is very close to the error probability for $K_{PLL} = 0.2$ and $\Delta f \times T_s = 0$ (Fig. 3). Even in the case $A_{PLL} = 1$, when the operation of PLL without ER block is practically not possible under the given conditions, the error probability drops by more than one order of magnitude if using ER block with length $L \geq 6$.

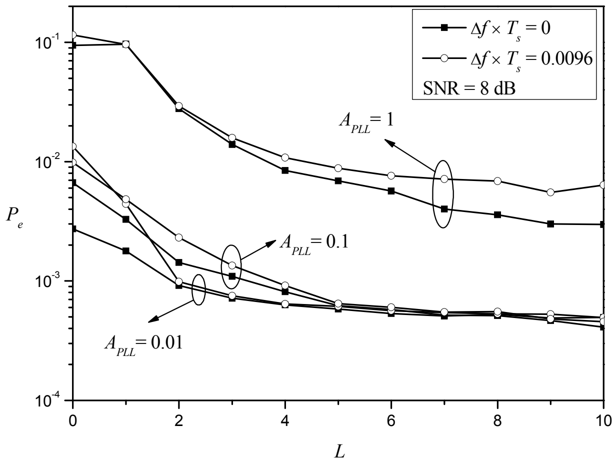

Fig. 4. Error probability as a function of filter length for $K_{PLL} = 1$ in AWGN channel.

From Fig. 4 it can be concluded that the increase of the length of ER structure beyond value $L = 6$ does not lead to a reduction of error probability. Also, for this length of ER structure the system shows approximately the same performance in the case of $A_{PLL} = 0.1$ as in the case of $A_{PLL} = 0.01$ with the performance being very close to the performance of the receiver with PLL only for zero frequency offset and parameter $K_{PLL} = 0.2$.

In Figs. 5 and 6 the error probability as a function of normalized frequency offset $\Delta f \times T_s$ can be traced, for the case of $K_{PLL} = 0.2$ and $K_{PLL} = 1$, respectively. In Fig. 5, which corresponds to the optimal operation parameters of a PLL block without ER, it can be seen that the sensitivity to the frequency offset is high. The values of frequency offset $\Delta f \times T_s \geq 0.005$ lead to the rapid increase of error probability for all values of the parameter A_{PLL}. Inclusion of ER block ($L = 6$, $A_{PLL} = 1$ and $A_{PLL} = 0.1$, $K_{PLL} = 0.2$) reduces sensitivity to frequency offset, so that the performance of the receiver changes only slightly for up to 2 times larger offset. This does not apply to the parameter value $A_{PLL} = 0.01$, which corresponds to a small PLL bandwidth, thus the introduction of ER block cannot improve the performance.

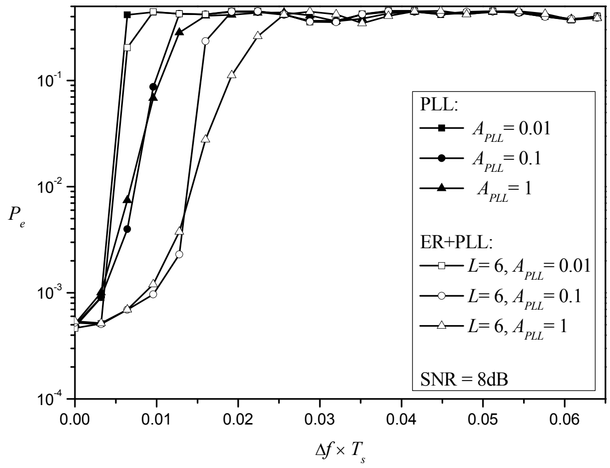

Fig. 5. Error probability as a function of carrier frequency offset for $K_{PLL} = 0.2$ in AWGN channel.

Fig. 6 shows the dependence of the error probability on frequency offset ($\Delta f\times T_s$) in the case of $K_{PLL} = 1$. Two cases were observed, namely: $A_{PLL} = 0.1$ and $A_{PLL} = 0.01$. For a frequency offset $\Delta f\times T_s \leq 0.0125$ both A_{PLL} values give approximately the same performance that are very close to the one that single PLL ($K_{PLL} = 0.2$) achieves at zero frequency offset. In the figure it can be noticed that the use of value $A_{PLL} = 0.1$ allows more than 2 times larger frequency offsets ($\Delta f\times T_s \leq 0.035$) in comparison with the case when $A_{PLL} = 0.01$, while system performance is only slightly impaired. For this reason, we retain for further analysis the following parameters: $K_{PLL} = 1$, $A_{PLL} = 0.1$ and $L = 6$.

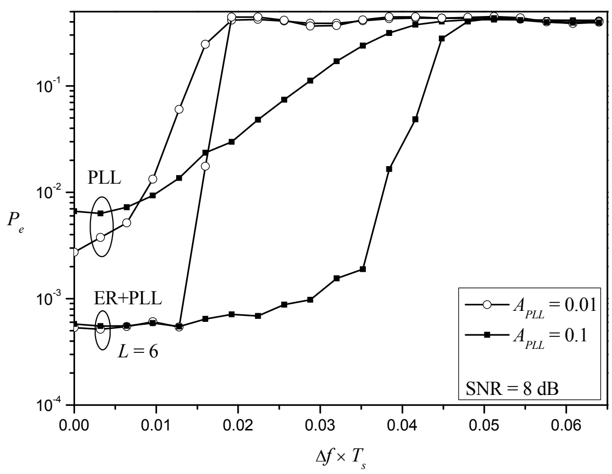

Fig. 6. Error probability as a function of carrier frequency offset for $K_{PLL} = 1$ in AWGN channel.

In Fig. 7 the dependence of error probability on the SNR is shown. Solid lines indicate the case when there is no fading, and dashed lines correspond to the case of Rician fading channel, characterized by a fading parameter $K = 10$ dB. The parameter K is the ratio of the power received via the line-of-sight (LOS) path to the power contribution of the non-LOS paths. It can be seen that for all the observed values of SNR, regardless of whether there is a frequency offset $\Delta f\times T_s = 0.01$ or not, the proposed receiver ER+PLL ($L = 6$) retains approximately the same value of error probability as in the case when there is no frequency offset and only PLL circuit is applied (with parameter $K_{PLL} = 0.2$). These curves belong to group I in Fig. 7. The same conclusion can be applied in case of signal propagation over a Rician fading channel (group II).

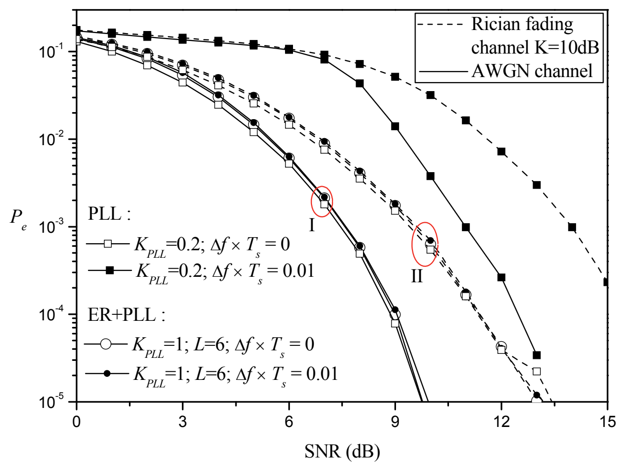

Fig. 7. Error probability as a function of signal to noise ratio for $A_{PLL} = 0.1$ in Rician fading and AWGN channel.

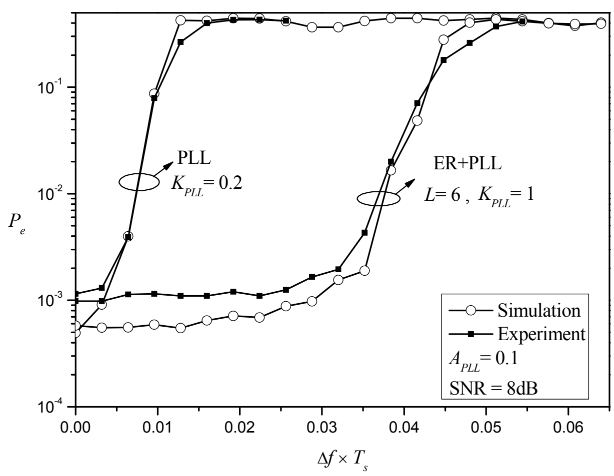

Fig. 8. Comparison of the experimental and simulation results for the error probability as a function of carrier frequency offset in AWGN channel.

The verification of the simulation results by an experiment, performed using USRP platform (as described in Section 3), is presented in Fig. 8. There are two typical cases shown:

1. when the PLL only receiver shows good performance ($K_{PLL} = 0.2$ and $A_{PLL} = 0.1$), and

2. when the whole block ER+PLL has good performance ($K_{PLL} = 1$, $A_{PLL} = 0.1$ and $L = 6$).

In both cases there is a satisfactory match between experimental and simulation results. The proposed receiver allows more than 7 times larger frequency offset, while the performance remains very close to the values obtained using only the PLL at zero frequency offset.

5. Conclusion

A modification of QPSK signal diversity reception with predetection EGC is proposed in this paper. EGC is realized by using the CM algorithm, and carrier synchronization is performed by the PLL. The modification is in introducing a structure that performs the estimation with remodulation (ER). By the adequate choice of the parameters for ER+PLL block we get the diversity receiver which allows more than 7 times larger frequency offset than the diversity receiver using only PLL. In this frequency offset range the error probability of the proposed receiver is very close to the error probability of the receiver with PLL and zero frequency offset.

The functionality of the proposed diversity receiver, as well as its properties are experimentally verified on a system based on universal software radio peripheral (USRP) hardware.

The performed comparison has confirmed the expected behavior of the system.

Acknowledgements

This work was supported in part by the Serbian Ministry of Education and Science within the project "Development and implementation of next-generation systems, devices and software based on software radio for radio and radar networks" (TR-32051).

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