Energy-efficient Communication System Based on Nonlinear Scattering of Standard OFDM Signals

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Abstract—**Development of miniaturized communication systems for exchanging data over short distances with little power consumption is an actual problem. This problem is solved by such standards as Bluetooth, ZigBee etc. Nowadays for extra size devices with power supply from external electromagnetic field are considered. In the paper we propose using nonlinear scattering effect for reusing energy radiated with standard wireless devices such as mobile phones, access points etc. For proposed solution link budget estimation is given. Performance of different types of receivers in communications system is investigated with both theoretical analysis and simulation.**

I. Introduction

Communication system with speed of several kilobits per second and distances between transmitter and receiver of 5-10 meters are demanded by nowadays applications [1]. Usually such systems are used for transmit biometric parameters or for different tasks such a wireless and personal area networks (WPAN) . Battery time operation of each node should be more than several years. Thus place of this system in the map "energy consumption/frequency range" is given in fig. 1 in left upper corner [2].

One of the main energy consumers in network nodes is RF transmitter. That is why the reduction of energy consumption of transmitters is in the focus of research community.

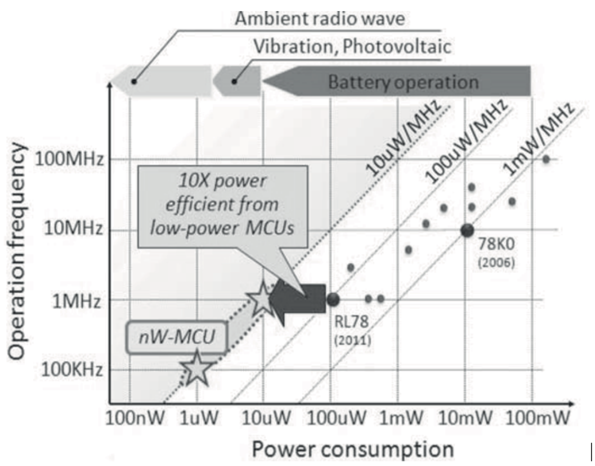

Fig. 1. Map "energy consumption/frequency range"

In modern communication systems using external electromagnetic energy for data transmission commonly a specialized external source is applied (e.g. like in NFC, ISO 14443 etc) [3]. This battery-less technique allows reusing radiated energy of different sources of electromagnetic energy. At the same time usage of non-specialized sources such as standard OFDM transmitters of communication devices (e.g. mobile phones, Wi-Fi access points etc.) is a field of active research (see e.g. [4]).

One of the prospective principles that can be used in such systems is a nonlinear scattering, which was initially described in [5]. In this paper authors discussed a digital communication system in which nonlinear scatterers as data transmitters had been used. The collected experimental data shows practical possibility of such systems.

In our paper we present enhancement of this research, analyzing potential characteristics of nonlinear scattering based transmitters in systems where OFDM signals are used as source of electromagnetic field.

Considered system configuration is present in fig. 2. Here OFDM signals are transmitted with standard communication devices (e.g. mobile phones, Wi-Fi access points etc.). The same communication device can be used for data gathering from sensor, equipped with nonlinear scatter.

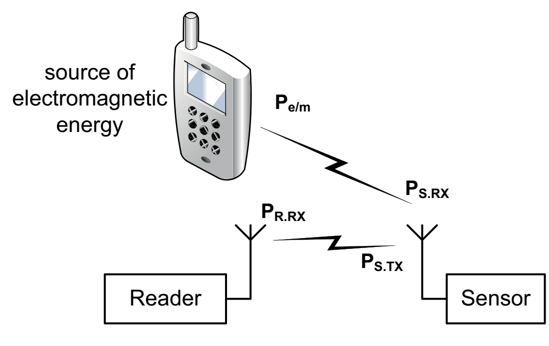

Fig. 2. System configuration

The rest of the paper is organized as follows: Section II describes basic principles of nonlinear scattering effect. Also in this section example of simple nonlinear scatter is shown. in Section III link budget of system is calculated. Basic principles of pulse position modulation and two schemes of such signals receivers (correlation receiver and radiometer-based receiver) are analyzed Section IV. In Section V the numerical example is given.

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II. Nonlinear Scattering

The nonlinear scatter is an antenna, loaded on nonlinear element, which can be either plugged or unplugged by means of external signal (see fig. 3).

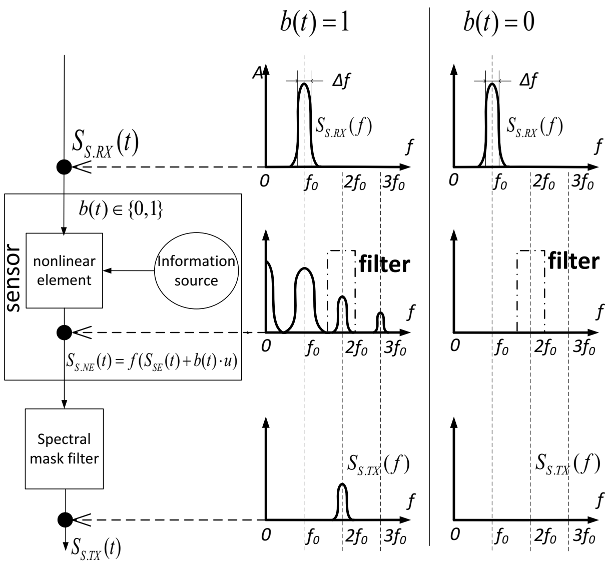

Fig. 3. Sensor structure

When the nonlinear element is plugged, combination tones of received signal appear in current going through antenna. These combination tones are reradiated with the same antenna. Usually just second tone is used, and others are rejected with filter [5]. In such a system a few modulation schemes can be realized: on/off keying, pulse position modulation (PPM) etc.

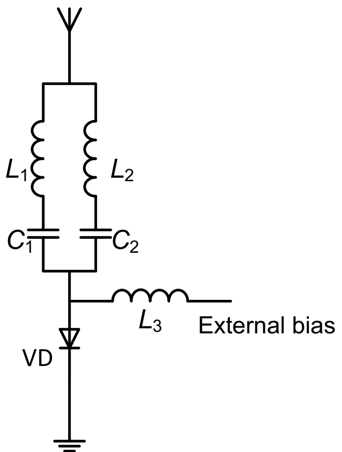

Fig. 4. Example of nonlinear scatter

The example of simple nonlinear scatter see in figure 4. Here, L_1, C_1 are to be chosen according to central frequency of external signal (f_0), L_2, C_2 according to second harmonic of external signal ($2f_0$). The external bias is to be chosen according to quadratic part of current-voltage characteristic of diode VD. Inductance L_3 stops high frequency signal from going to source of external bias. Thus, when the nonlinear scatter is introduced into electromagnetic field with central frequency f_0, a signal with central frequency $2f_0$ appears on the air. After switching off the bias, this reradiated signal disappears.

III. Link Budget of System

For estimation of achievable signal-to-noise ratio (SNR) for the proposed system the following notation is introduced:

- f_1 , f_2 – center frequency of the primary and scattered signals accordingly, in Hz;
- λ_1 , λ_2 – the wavelength of the primary and reflected signals, in meters;
- r_0 – a reference distance, in meters;
- r_1 – the distance between source of electromagnetic energy and sensor, in meters;
- r_2 – the distance between sensor and reader, in meters;
- $G_{e/m}$ – amplification of source of electromagnetic energy antenna, in times;
- G_S – amplification of sensor antenna, in times;
- G_R – amplification of reader antenna, in times;
- L – loss in the sensor;
- $P_{e/m}$ – maximum radiated power of source of electromagnetic energy;
- R – data rate, bit / s;
- k – Boltzmann constant $1.38 \cdot 10^{-23}$;
- K_n – noise figure of reader receiver;
- N – number of repetitions for coherent accumulation;
- q_s – signal-to-noise ratio in the reader receiver (as the ratio of the elementary pulse energy to noise power spectral density)
- β – path-loss factor.

According to radar range equation, power of signal received with sensor is [9], [10]:

$$P_{S.RX} = P_{e/m} \left(\frac{G_{e/m}}{4\pi r_0^2} \right) \left(\frac{G_S \lambda_1^2}{4\pi} \right) \left(\frac{r_0}{r_1} \right)^{\beta}$$

Taking into account losses in nonlinear scatter, sensor radiated power is:

$$P_{S.TX} = L \times P_{S.RX} =$$
$$= LP_{\mathrm{e/m}} \left(\frac{G_{\mathrm{e/m}}}{4\pi r_0^2} \right) \left(\frac{G_S \lambda_1^2}{4\pi} \right) \left(\frac{r_0}{r_1} \right)^{\beta}$$

Using the equation 1 for the second time, we find the power of signal received with reader:

$$P_{R.RX} = P_{S.TX}\left(\frac{G_S}{4\pi r_0^2}\right)\left(\frac{G_R\lambda_2^2}{4\pi}\right)\left(\frac{r_0}{r_2}\right)^{\beta} =$$
$$= P_{\mathrm{e/m}}\frac{LG_{\mathrm{e/m}}G_R\left(G_S\lambda_1\lambda_2\right)^2}{\left(4\pi r_0\right)^4}\left(\frac{r_0^2}{r_1 r_2}\right)^{\beta}$$

If the bit duration is equal to $\tau_b = {}^{1}\!/_{R}$ then the energy per bit is:

$$E_{R.TX} = \frac{1}{2}\frac{P_{\mathrm{e/m}}LG_{\mathrm{e/m}}G_R\left(G_S\lambda_1\lambda_2\right)^2}{R\left(4\pi r_0\right)^4}\left(\frac{r_0^2}{r_1 r_2}\right)^{\beta}$$

Here the factor 1/2 means that PPM modulation or on/off keying is used. Noise spectral density can be found with Boltzmann equation:

$$N_0 = kTK_n$$

Then the signal-to-noise ratio at the reader (as the ratio of energy per bit to noise power spectral density) is:

$$q_s = \frac{P_{\mathrm{e/m}}LG_{\mathrm{e/m}}G_R\left(G_S\lambda_1\lambda_2\right)^2}{2kTK_nR\left(4\pi r_0\right)^4}\left(\frac{r_0^2}{r_1 r_2}\right)^{\beta} \quad (1)$$

IV. Data transmission from sensor, equipped with nonlinear scatter

A. Pulse-position modulation

For data transmission we can only turn on or off signal. Ex facte, On-Off Keying is most common choice for this task. In this modulation the presence of a carrier for a specific duration represents a binary one, while its absence for the same duration represents a binary zero. However, if the signal-to-noise ratio is unknown, the task of choosing the threshold at the receiver can be quite complicated.

Hence, we propose to use pulse position-modulation (PPM) modulation. PPM is a form of signal modulation in which bits are encoded by transmitting a single pulse in one of possible time-shifts [7].

If bit value is 0 then the left subslot of the slot consists scattered signal, otherwise right subslot is used (see 5).

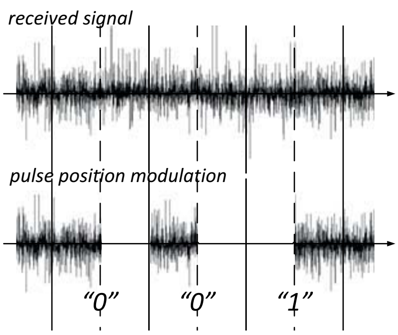

Fig. 5. Pulse Position-Modulation

Changing of the slot length leads to a changing of data rate fig. 6. Note that the greater the duration of the slots (and, therefore, less speed), the lower the Bit Error Rate.

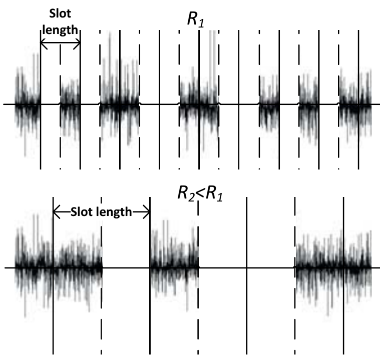

Fig. 6. Changing of data rate in PPM

B. Correlation receiver

Non-coherent correlation receiver is considered for signal reception from nonlinear scatter. For such receiver reference signal is required. Reader receives OFDM symbols from external source, detects and decodes them and use for reference signal generation (see fig. 7).

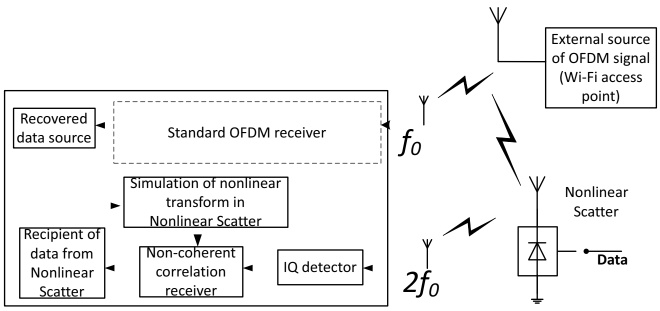

Fig. 7. Structure of correlation receiver

Used modulation scheme (PPM) is a scheme with orthogonal signals. Therefore, knowing the signal-to-noise ratio at the receiver reader we can estimate the bit error probability. It equals to:

$$P_e = \frac{1}{2}\exp\left(-\frac{q_s}{2}\right) \quad (2)$$

According to equation 1, 2, the achievable data rate is

$$R = -\frac{P_{\mathrm{e/m}}LG_{\mathrm{e/m}}G_R\left(G_S\lambda_1\lambda_2\right)^2}{2kTK_n\left(4\pi r_0\right)^4 2ln(2P_e)}\left(\frac{r_0^2}{r_1 r_2}\right)^{\beta}$$

C. Radiometer based receiver

Unfortunately it can be quite difficult to implement correlation receiver because it requires reference signals which may not be obtained in some cases (for example in case of unknown channel, impossibility of quality restoration of the primary signal, etc). Therefore, a simple receiver was considered using the approach of detecting energy in the slot [6]. Receiver for this scheme compares energies in the subslots and makes decision using the following equations.

$$E_0 = \sum_{i=1}^{N/2} |S_{R.RX}(i)|^2$$

$$E_1 = \sum_{i=N/2+1}^{N} |S_{R.RX}(i)|^2$$

$$\hat{b} = arg \max_i E_i$$

Here E_0 denotes energy in left subslot, E_1 - energy in right subslot and N - is slot duration (in samples). Let us call this scheme as radiometer-based receiver.

Let us consider the situation when the signal is transmitted in left sublots. Hence,

$$E_0 = \sum_{i=1}^{N/2} |S_{R.RX}(i)|^2 = \sum_{i=1}^{\frac{N}{2}} |s_i + n_i|^2$$

Here s_i and n_i are samples of signal and samples of noise respectively. These values are complex numbers.

$$\sum_{i=1}^{\frac{N}{2}} \left[\left(s_i^{(R)} + n_i^{(R)}\right)^2 + \left(s_i^{(I)} + n_i^{(I)}\right)^2 \right] =$$
$$= \sum_{i=1}^{\frac{N}{2}} \left(s_i^{(R)} + n_i^{(R)}\right)^2 + \sum_{i=1}^{\frac{N}{2}} \left(s_i^{(I)} + n_i^{(I)}\right)^2 =$$
$$= \sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(R)}\right)^2 + 2n_i^{(R)} s_i^{(R)} \right] + \sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(I)}\right)^2 + 2n_i^{(I)} s_i^{(I)} \right] +$$
$$+ \sum_{i=1}^{\frac{N}{2}} \left[\left(s_i^{(R)}\right)^2 + \left(s_i^{(I)}\right)^2 \right] = \sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(R)}\right)^2 + 2n_i^{(R)} s_i^{(R)} \right] +$$
$$+ \sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(I)}\right)^2 + 2n_i^{(I)} s_i^{(I)} \right] + \sum_{i=1}^{\frac{N}{2}} \left[\left(s_i^{(R)}\right)^2 + \left(s_i^{(I)}\right)^2 \right] =$$
$$= \sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(R)}\right)^2 + 2n_i^{(R)} s_i^{(R)} \right] + \sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(I)}\right)^2 + 2n_i^{(I)} s_i^{(I)} \right]$$
$$+ \frac{N}{2} P$$

Here P is power of received signal.

Variance and the mathematical expectation calculation is needed for BER analysis. Assume that noise is AWGN and n_i is gaussian random variable with variance equal to σ^2.

$$\mathbf{E}\left[\left(n_i^{(R)}\right)^2\right] = \mathbf{E}\left[\left(n_i^{(I)}\right)^2\right] = \frac{\sigma^2}{2}$$

$$\mathbf{Var}\left[\left(n_i^{(R)}\right)^2\right] = \mathbf{Var}\left[\left(n_i^{(I)}\right)^2\right] = \frac{\sigma^4}{2}$$

Let us calculate variance and the mathematical expectation of E_0. Assume that s_i are independent variables.

$$\mathbf{E}[E_0] = \mathbf{E}\left[\sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(R)}\right)^2 + 2n_i^{(R)} s_i^{(R)} \right] \right] +$$
$$+ \mathbf{E}\left[\sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(I)}\right)^2 + 2n_i^{(I)} s_i^{(I)} \right] \right] + \mathbf{E}\left[\frac{N}{2}P\right] =$$
$$= \frac{N}{2} \cdot \frac{\sigma^2}{2} + 0 + \frac{N}{2} \cdot \frac{\sigma^2}{2} + 0 + \frac{N}{2} P = \frac{N}{2} \left(\sigma^2 + P\right)$$

$$\mathbf{Var}[E_0] = \mathbf{Var}\left[\sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(R)}\right)^2 + 2n_i^{(R)} s_i^{(R)} \right] \right] +$$
$$+ \mathbf{Var}\left[\sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(I)}\right)^2 + 2n_i^{(I)} s_i^{(I)} \right] \right] + \mathbf{Var}\left[\frac{N}{2}P\right]$$

In this expression, the variance of the first two summands are equal.

$$\mathbf{Var}\left[\sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(R)}\right)^2 + 2n_i^{(R)} s_i^{(R)} \right] \right] =$$
$$= \mathbf{Var}\left[\sum_{i=1}^{\frac{N}{2}} \left[\left(n_i^{(I)}\right)^2 + 2n_i^{(I)} s_i^{(I)} \right] \right]$$

For simplicity, let us denote $n_i^{(R)}$, $n_i^{(I)}$ as n and $s_i^{(R)}$, $s_i^{(I)}$ as s.

$$\mathbf{Var}\left[n^2 + 2ns\right] = \iint_{\infty} \left[n^2 + 2ns - \frac{\sigma^2}{2} \right]^2 f(n, s) dn \cdot ds$$

The integral of a sum of functions is equal to the sum of their integrals. Hence,

$$\iint_{\infty} n^4 \cdot f(n)f(s)dn \cdot ds + \iint_{\infty} 4n^3 s \cdot f(n)f(s)dn \cdot ds + \\ + \iint_{\infty} 4n^2 s^2 \cdot f(n)f(s)dn \cdot ds - \\ - \iint_{\infty} n^2 D \cdot f(n)f(s)dn \cdot ds - \\ -2\iint_{\infty} nsD \cdot f(n)f(s)dn \cdot ds + \\ + \iint_{\infty} \frac{\sigma^4}{4} \cdot f(n)f(s)dn \cdot ds = \\ = \frac{3\sigma^4}{4} + 0 + \sigma^2 P + \frac{\sigma^4}{2} + 0 + \frac{\sigma^4}{4} = \frac{\sigma^4}{2} + \sigma^2 P$$

Substitution of this expression into the equation 3 gives:

$$\mathbf{Var}\left[\sum_{i=1}^{\frac{N}{2}}\left[\left(n_i^{(R)}\right)^2 + 2n_i^{(R)}s_i^{(R)}\right]\right] + \\ +\mathbf{Var}\left[\sum_{i=1}^{\frac{N}{2}}\left[\left(n_i^{(I)}\right)^2 + 2n_i^{(I)}s_i^{(I)}\right]\right] = N\sigma^2\left(\frac{\sigma^2}{2} + P\right)$$

Analysis of E_1 is similar.

$$E_1 = \sum_{i=N/2+1}^{N} |S_{R.RX}(i)|^2 = \sum_{i=\frac{N}{2}+1}^{N} |n_i|^2 = \\ = \sum_{i=\frac{N}{2}+1}^{N}\left[\left(n_i^{(R)}\right)^2 + \left(n_i^{(I)}\right)^2\right]$$

As was mentioned above n_i is gaussian random variable.

$$\mathbf{E}\left[\sum_{i=1}^{\frac{N}{2}}\left(n_i^{(R)}\right)^2\right] = \mathbf{E}\left[\sum_{i=1}^{\frac{N}{2}}\left(n_i^{(I)}\right)^2\right] = \frac{N}{2}\cdot\frac{\sigma^2}{2} = \frac{N\sigma^2}{4}$$

$$\mathbf{E}\left[\sum_{i=1}^{\frac{N}{2}}\left(n_i^{(R)}\right)^2 + \sum_{i=1}^{\frac{N}{2}}\left(n_i^{(I)}\right)^2\right] = \frac{N\sigma^2}{2}$$

$$\mathbf{Var}\left[\sum_{i=1}^{\frac{N}{2}}\left(n_i^{(R)}\right)^2\right] = \mathbf{Var}\left[\sum_{i=1}^{\frac{N}{2}}\left(n_i^{(I)}\right)^2\right] = \frac{N}{2}\cdot\frac{\sigma^4}{2} = \frac{N\sigma^4}{4}$$

$$\mathbf{Var}\left[\sum_{i=1}^{\frac{N}{2}}\left(n_i^{(R)}\right)^2 + \sum_{i=1}^{\frac{N}{2}}\left(n_i^{(I)}\right)^2\right] = \frac{N\sigma^4}{2}$$

Bit Error Rate of this scheme is equal to probability $Pr\{E_0 > E_1\}$. For simplicity, transfer the variables on same side of inequality, and assume that the E_0 and E_1 are normal distribution variables. Then expected value of $\Psi = E_0 - E_1$ is $\mathbf{E}[\Psi] = \frac{N}{2}\left(\sigma^2 + P\right) - \frac{N\sigma^2}{2} = P$ and $\mathbf{Var}[\Psi] = N\sigma^2\left(\frac{\sigma^2}{2} + P\right) + \frac{N\sigma^4}{2} = N\sigma^2\left(\sigma^2 + P\right)$. Finally the BER is equal to probability $Pr\{\Psi > 0\}$.

As one can see in given equations all the items of the sums are considered to be independent, identically distributed random variables. Hence, variance of the sum is calculated as sum of the variances of all the items. However samples $s(i)$ can be correlated. Moreover if preliminary filtering is used in the front-end samples of $n(i)$ can be correlated either. Thus the obtained result can be treated as an estimation of bit-error ratio. For exact calculation of BER simulation was performed. This simulation allowed to indicate the accuracy of proposed estimation.

V. Numerical example

A. System parameters

For numerical analysis the following parameters of the system were chosen:

- $r_0 = 75$ cm (it is 6 wavelengths of WiFi signal);
- $G_{e/m} = 2.15$ dB (corresponds to half-wave dipole antenna);
- $G_S = 2.15$ dB (corresponds to half-wave dipole antenna);
- $G_R = 2.15$ dB (corresponds to half-wave dipole antenna);
- $L = -60$ dB (this estimation is given in [5]);
- $K_n = 10$ dB (corresponds to typical receiver);
- $P_{e/m} = 100$ mW (corresponds to WiFi signal);
- $f_1 = 2.4$ GHz (corresponds to central frequency of WiFi signal);
- $f_2 = 4.8$ GHz (corresponds to second tone of nonlinearly transformed WiFi signal);

B. Signal-to-noise ratio

For the system with described parameters the dependency of SNR on used data rate and distance between source of e/m energy (standard OFDM transmitter) and nonlinear scatter was calculated (see fig. 8-9).

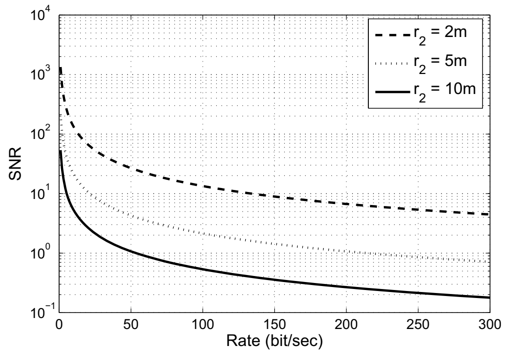

Fig. 8. SNR analysis $\beta = 2$

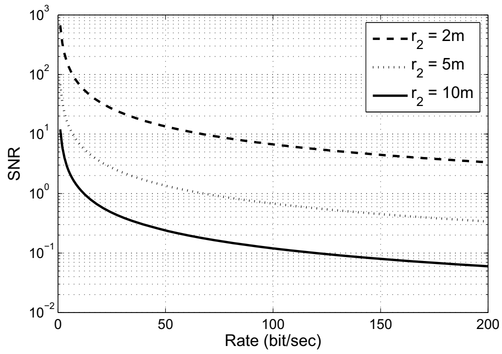

Fig. 9. SNR analysis $\beta = 2.5$

Note that at small distances between the sensor and the reader, the main LOS path of the channel response is dominant. So $\beta = 2$ and $\beta = 2.5$ are taken [8].

C. Correlation receiver

For the system with described parameters the dependency of bit-error ratio (BER) on used data rate and distance between source of e/m energy (standard OFDM transmitter) and nonlinear scatter was calculated (see fig. 10-11).

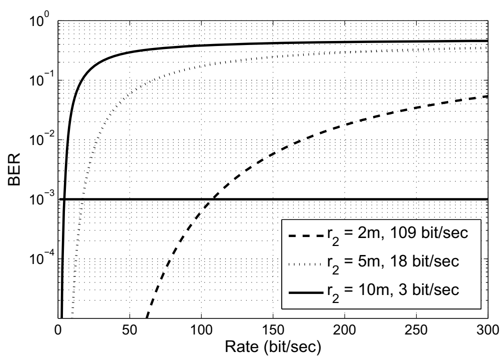

Fig. 10. BER analysis of correlation receiver, $\beta = 2$

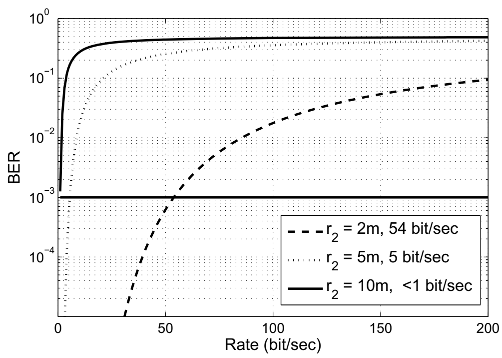

Fig. 11. BER analysis of correlation receiver, $\beta = 2.5$

As one can see if required bit-error ratio is 10^{-3} then up to a few tens bits-per-second can be achieved. This data rate is enough to update all the main biometric parameters (heart rate, temperature, pressure) of sensor user every second.

D. Radiometer based receiver

Real model of scatter and reader structure is shown in fig. 12, a. However, this model is complicate for simulation. Consequently the modified model (see fig. 12, b) was developed in order to simplify the simulation. In this model first of all signal is generated, and then it is divided into slots. The only difference between this model and real model is the location of key on the reader. This simplification eliminates the need for multiple simulation of signal passing through a nonlinear element process.

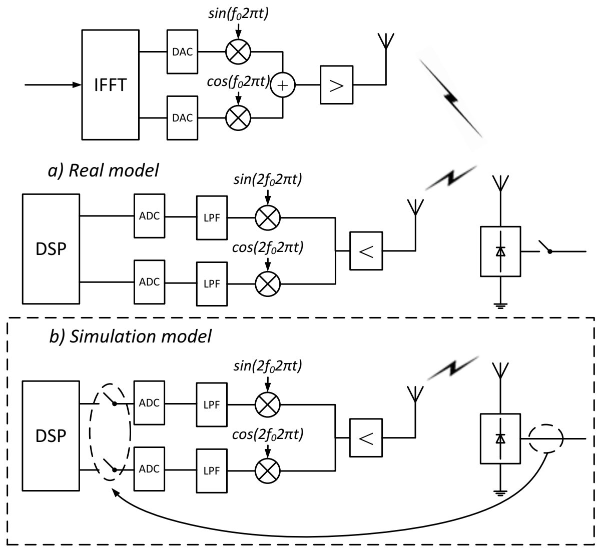

Fig. 12. Real and simulation models

Simulation result and analytical calculation are presented in fig. 13

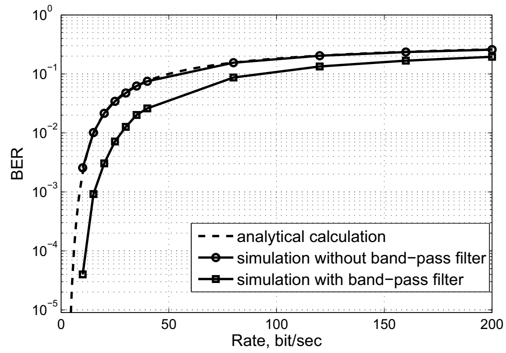

Fig. 13. BER analysis of radiometer-based receiver, $\beta = 2$, $r_1 = r_2 = 0.8$

VI. CONCLUSION

It was found that nonlinear scattering effect allows building low-rate communication system with significantly reduced hardware complexity and energy consumption.

We have proposed to use OFDM transmitters of standard communication devices as external source of electromagnetic field. Analysis allowed estimating of signal-to-noise ratio in such system. For example if distance between source of e/m energy equal to 2 meter expected SNR is 14 dB for data rate 100 bit/sec.

BER characteristics of nonlinear scatter system were calculated. By using correlation receiver if required bit-error ratio is 10^3 than data rate up to 109 bit/sec can be achieved for distance between source of e/m energy equal to 2 meter. This is enough to update all the main biometric parameters of nonlinear scatters user every second.

Since correlation receiver is difficult to implement, radiometer-based receiver was also considered. This receiver is simple to implement, but it works at sufficiently shorter distances (less than one meter).

In further studies developments of multiple user access scheme for nonlinear scattering based transmitters and development of the synchronization scheme are planed.

VII. Acknowledgements

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