System-level analysis of an Ultra-low power, low data-rate FHSS transceiver

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Abstract— **The market of wireless connectivity is growing rapidly. Various standards have been developed to satisfy the market. The area of ultra-low power, low data rate applications aims to become one of the most attractive areas in terms of revenue. Possible fields of application for ultra-low power low data rate networks can be the following: domotica, automotive, health care etc. Ultra-low power wireless links require a careful definition of radio parameters and a new architectural approach, compared to moderate and high-speed multimedia wireless link, in order to make battery lifetime and therefore battery replacement practical to the consumer. To achieve the ultra-low power target, the radio link budget must be examined and a careful study of the main system parameters is required. Furthermore this link budget is strictly related to the transmitted power, PTX, and the received signal power PRX. Starting from the definition of a radio link budget suitable for power constrained radio, all the basic radio parameters for a slow-hopping, Frequency Hopping Spread Spectrum (FHSS) transceiver are derived. These parameters include sensitivity, Noise Figure, spurious free dynamic range, phase noise, third order intercept point. The analysis demonstrates that the relaxed requirements on the radio side can allow the implementation of a single chip, ultra low power, low cost transceiver suitable for low data-rate applications, if reduced Quality of Service (QoS) can be tolerated.**

I. INTRODUCTION

The general focus in the next technology generations is squarely on the user. A new class of emerging technologies meant primarily for indoor use over very short distances is the class of the short-range wireless.

The market for wireless connectivity is growing rapidly. A number of standards have been developed to satisfy the requirements of various parts of this market. Examples include: IEEE 802.11x, ZigBee, HomeRF, Bluetooth for wireless links and networks, and ISO 15693 for RF ID tags.

While these active and passive solutions are well matched to the requirements of a wide range of applications, there exists a gap between them. This gap is in the area of low data-rate communications for applications such as ambient intelligence[1], sensor networking and control functions in the home of the consumer (domotica). What is required for these applications is simplified functionality and technology compared to Bluetooth, but at a cost level more related to passive tagging, and at a much reduced power levels.

Low power/low data-rate (LP/LBR) transceivers require a new architectural approach, compared to moderate and high-speed multi-media wireless links, in order to make battery lifetime and thus battery replacement practical for the consumer. Ultra-low power wireless transceivers may even extract all needed energy from their environment. Therefore a careful design of the air interface as well as of the system parameters of the radio are required in order to achieve the goals of low-cost and ultra-low power:this is the subject of this paper.

The paper is organized as follows. Section II describes basics choices in the system such as frequency bands or spreading techniques while Section III derives the basic system level parameters for the radio system (like NF, SFDR etc.) focusing on the 902-928 MHz Industrial-Scientific-Medical (ISM) band. Finally, concluding remarks are made in Section IV.

II. FREQUENCY HOPPING SPREAD SPECTRUM SYSTEMS

The use of the ISM band is mandatory for a consumer product that aims to be low-cost. The most attractive ISM band is the 2.45 GHz band due to its worldwide availability but also the 915 MHz band, which is available only in U.S., is a good field to test new ideas and concepts toward the developement of

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[1] One common definition of Ambient Intelligence environments is:"Electronic environments that are aware of and responsive to the presence of people". Such environments should be ubiquitous, transparent and intelligent (adapting to the people that live in the environment)[1].

an ultra-low power transceiver for low data-rate applications. Therefore the focus will be mainly on these two bands with a full system analysis for the 915 MHz band.

The ISM bands are generally open to different kinds of systems, therefore radio systems operating in this bands should be able to cope with a high level of intreferences (in the 2.45 GHz band the microwave oven represents the strongest source of interference). Moreover, the target applications of the aforementioned transceiver will be mainly indoor applications. This means that capability of the system to cope with severe fading is mandatory. Spread spectrum techniques are a powerful way to cope with such problems, and together with frequency and time diversity can make a link robust even under severe interferences or deep fading.

Two different spread spectrum techniques are largely used in radio systems:

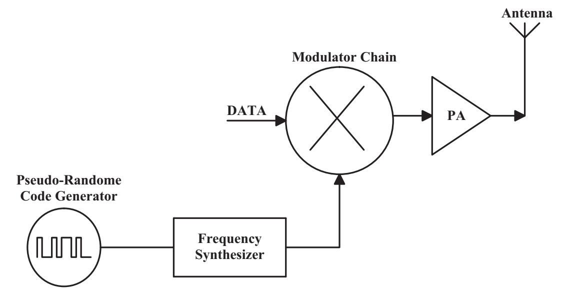
- Direct Sequence Spread Spectrum (DSSS).
- Frequency Hopping Spread Spectrum (FHSS).

System analysis have shown that for a low data-rate, power constrained radio system, the FH technique is able to better cope with the aforementioned non-idealities still allowing, when implemented on silicon, to keep the power consumption low [2]. In addition an FHSS system has the ability to better cope with the near-far problem: a nearby jammer can be easily suppressed by the narrowband channel select filter as far as its spectrum does not coincide with the selected channel [2]. Therefore a full system analysis of a FHSS system will be carried out in the following sections of this article.

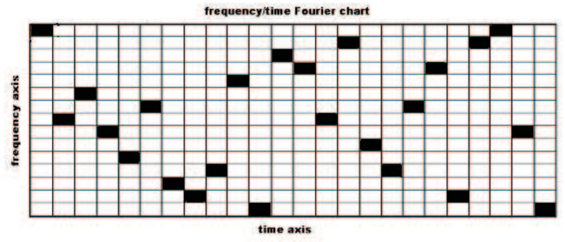
A simplified block diagram of an FHSS transmitter is shown in Fig.1(a). A Pseudo-random Code generator drives a Frequency Synthesizer, which synthesizes the desired hopping frequency. A mixer provide the upconversion to the desired band and the power amplifier drives the antenna which send the desired data stream through the air. Generally FSK modulation is used in FHSS transceiver, and due to the difficulty to maintain phase coherence between hops, a non-coherent demodulation scheme is used. The operation over time of a FHSS transceiver is depicted in Fig.1(b).

A. Frequencies Allocation

In this section as well as in the following section we will refer always to the 902-928 MHz ISM band. To use in an efficient way the spectrum available a calculation of the spectrum occupied by a FSK modulated signal is needed.



(a)Block Diagram of a Frequency Hopping System



(b)Frequency Hopping Signal

Fig. 1. FHSS System

As with any FM signal, the bandwidth of the FSK signal depends on the modulation index. Therefore the transmission bandwidth B_{T} of a FSK signal is given by the Carsons's rule as:

$$B_{\mathrm{T}} = 2\Delta f + 2B \tag{1}$$

where Δf is the frequency deviation and is related to the modulation index m by the following,

$$m = \frac{\Delta f}{f_{\mathrm{m}}} \tag{2}$$

where f_{m} is the frequency of the data signal and B is the bandwidth of the digital baseband signal. Assuming to use the first-null bandwidth, then the bandwidth of a rectangular pulse is $B = R$, and if a raised cosine pulse-shaping filter is used, then the transmission bandwidth becomes:

$$B_{\mathrm{T}} = 2\Delta f + (1 + \alpha)R \tag{3}$$

where α is the rolloff factor of the filter. Now supposing to use $\alpha = 1$, $m = 10$ and a data rate of 2 *kbps*, then the transmission bandwidth, in each frequency bin, is equal to 44 kHz. This is depicted in Fig.2.

Therefore, a channel bandwidth of 50 kHz has been chosen. The number of channels for the FHSS system

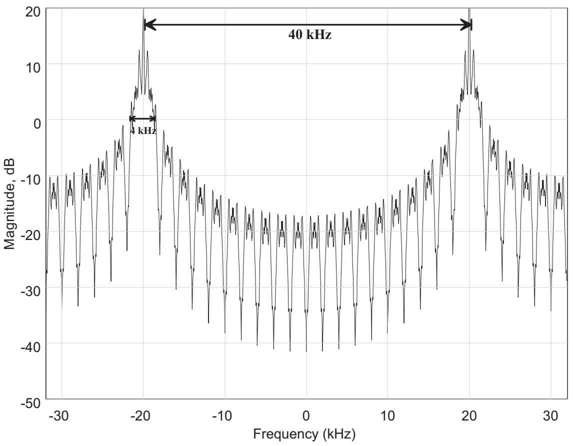


Fig. 2. Power Spectrum of a BFSK signal with data-rate equal to 2 *kbps* and m=10

has been chosen to fulfill FCC rules [3] and to have an acceptable processing gain, which can guarantee the reliability of the communication link in an indoor environment.

In the case of the system here described, the 20 dB bandwidth of the hopping channel is less than 250 kHz and therefore a minimum number of 50 hopping frequencies shall be used. Due to the typical frequency selective nature of the fading, the more the channel are separated, the bigger the probability that if a channel n is jammed or experiments a very low signal-to-noise ratio (SNR), in the adjacent ones ($n+1$ *or* $n-1$) a reliable communication can be possible. Therefore, the separation between adjacent channels has been chosen equal to 100 kHz. This choice fulfills also the FCC rules, which demand a minimum separation of 25 kHz between adjacent channels. The 50 channels are allocated has depicted in Fig.3.

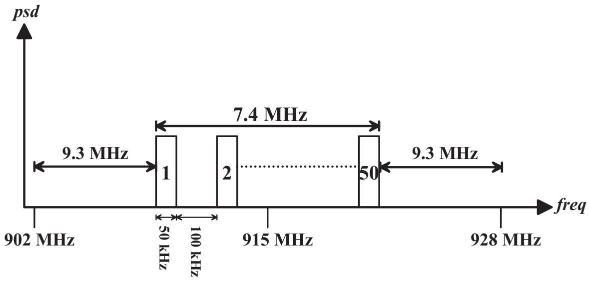


Fig. 3. Channels allocation in the 902-928 MHz ISM band

B. Interference Immunity

As mentioned in Section II, the ultra-low power radio must operate in a license-free band, where a large number of uncontrolled sources of interference is present. Robustness against these interferences is provided by the following features [2].

- Frequency diversity through hopping between channels.
- Forward Error Correction (FEC) code to protect the data packet.

The hopping rate of the system has been chosen equal to the data-rate in such a way that the radio will be a slow-hopping FHSS system, while as FEC code a Reed-Solomon error correction code can be used due to its inherent advantages in burst communications.

III. System-Level Considerations

Low data-rate applications demand, generally, much lower performances compared to high-end applications. Especially lower Quality of Service (QoS) requirements translate directly in relaxed specifications for the transceiver, allowing fully integration and therefore reduction of costs and power consumption.

The most demanding characteristics of an integrated transceiver are the sensitivity, selectivity, and output power while staying in an extremely tight power consumption budget. The initial range has been set to around 50 m and the transmitted power has been set to 0 dBm in order to not compromise the operational time of a battery-operated device or to allow self-recharging of the battery by harvesting energy from the ambient (solar or vibrational energy are two types of energy sources that can be used to recharge the internal battery of the radio).

A. Link Budget Analysis

To define the different radio parameters a link budget analysis is required. The link budget analysis is an estimation technique for evaluating or predicting communications system performance.

The propagating medium connecting the transmitter with the receiver is called the channel. Several parameters affect the signal on its path between the transmitter and the receiver. The signal is subjected first to attenuation loss L_{path} due to the propagation of the signal in the air, and the transmitter and receiver antennas have characteristics gain that can be denoted as G_{TX} and G_{RX}. Furthermore, especially in an indoor environment, the signal is subjected to multipath reflections, which can corrupt the received signal. Likewise, obstacles will greatly reduce the strength of the received signal.

Unity gain antennas are supposed at the transmitter as well as at the receiver. Therefore expressing all the parameters in decibel units, the received signal power is:

$$P_{\mathrm{RX}} = P_{\mathrm{TX}} + L_{\mathrm{path}} \tag{4}$$

where P_{RX} and P_{TX} are respectively the received and the transmitted power. The attenuation losses due to propagation and fading need to be calculated. When no objects are present between transmitter and receiver, a line-of-sight (LOS) is present while when there are objects in between the path is called a non-line-of-sight (NLOS) path. Generally all wireless links have both a LOS and NLOS propagation paths. To derive the attenuation when a NLOS condition occurs we first calculate the propagation loss in a LOS situation. The free space attenuation can be expressed as [4](in [4] a full derivation of the following formula can be found):

$$L_{\mathrm{path,LOS}} = 27.56dB - 20\log_{10}(f_{\mathrm{c}}) - 20\log_{10} \tag{5}$$

where f_{c} is the carrier frequency expressed in MHz and r is the communication distance between transmitter and receiver expressed in meter. The NLOS path loss can be approximated as [4]

$$L_{\mathrm{path}} = L_{\mathrm{path,LOS}} - 10 \cdot n \cdot \log 10(\frac{d}{d_0}) \tag{6}$$

where n is the path loss exponent, which indicates how fast the path loss increases with distance, d_0 is the reference distance for free-space (unobstructed) propagation, $L_{\mathrm{path,LOS}}$ is the corresponding propagation loss of the LOS path, and d is the distance between transmitter and receiver.

Supposing a reference distance d_0=3m, a path loss exponent n=3 and a distance between transmitter and receiver d=50m the LOS path loss is: $L_{\mathrm{path,LOS}} \approxeq$ -41.2dB. Therefore, substituting this value in (6) the path loss in the NLOS condition becomes: $L_{\mathrm{path}} \approxeq$ -78dB. Considering 0 dBm transmitted power and substituting the value found from (6) in (4) we have a minimum received signal strength in absence of fading equal to: $P_{\mathrm{RX,min}}$=-78 dBm. This minimum received signal strength is the minimum signal the receiver should be able to detect and is called receiver sensitivity RX_{sens}.

Furthermore, we can relate the receiver sensitivity to the Noise Factor (NF) of the front-end through the following:

$$RX_{\mathrm{sens}} = N_0 + BW + SNR_{\mathrm{demod}} + NF \tag{7}$$

where N_0 is the thermal noise power spectral density, BW is the receiver noise bandwidth and SNR_{demod} is the required Signal-to-Noise ratio at the input of the demodulator (output of the front-end) to obtain a certain Bit Error Rate (BER). Now, the quantity SNR_{demod} should be calculated. The system should be able to achieve a BER=0.1% when no FEC is applied. If a non-coherent FSK demodulation scheme is applied at the receiver side, then it is well known from communication theory that:

$$P_{\mathrm{e,NCFSK}} = \frac{1}{2}e^{(-\frac{E_{\mathrm{b}}}{2N_0})} \tag{8}$$

where E_{b} is the energy per bit and $P_{\mathrm{e,NCFSK}}$ is the error probability. Using (8) we obtain $\frac{E_{\mathrm{b}}}{N_0} \approxeq 11$ dB. Equation (8) does not take in account the fading. When fading is considered, it can be proved [4] that the probability of error for a non-coherent BFSK becomes

$$P_{\mathrm{e,NCFSK}} = \frac{1}{2+\Gamma} \tag{9}$$

where

$$\Gamma = \frac{E_{\mathrm{b}}}{N_0}\overline{\alpha^2} \tag{10}$$

where α is the gain of the channel with Rayleigh distribution. The term Γ represents, indeed, the average value of the normalized SNR. Therefore to have a raw BER of 0.1% a Γ=30 dB is required.

Furthermore, SNR is related to $\frac{E_{\mathrm{b}}}{N_0}$ by the following

$$SNR = \frac{E_{\mathrm{b}}}{N_0} \cdot \frac{D}{W} \tag{11}$$

where D is the data rate and W is the bandwidth occupied by the signal in baseband. In the case of the system described in this article, the data rate is equal to 2 kbps while the occupied data bandwidth is 8 KHz (when first-null bandwidth is taken into account). Therefore the required SNR expressed in dB becomes

$$SNR = \frac{E_{\mathrm{b}}}{N_0} - 6dB \tag{12}$$

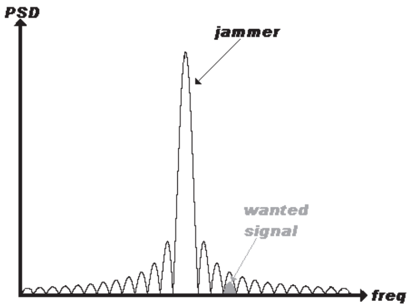
From (12),(10) and (7) the NF can be derived from the following equation

$$
\begin{aligned}
-78[dBm] = -174[dBm/Hz] + 47[dBHz] \\
+30[dB] - 6[dB] + NF + 6[dB] \qquad (13)
\end{aligned}
$$

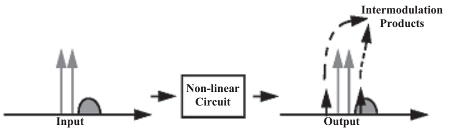
where 6dB extra losses due to antenna matching, filters, connectors etc. have been considered. From (13) a NF=19dB has been derived.

B. Dynamic Range

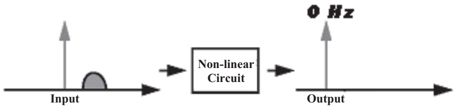
For the derivation of the Spurious-Free Dynamic Range (SFDR) all the sources of interference should be considered.



(a)Effect of a strong adjacent jammer on the wanted signal



(b)Effect of intermodulation products



(c)Effect of a strong blocker on the receiver

Fig. 4. Various sources of interference at the input of the RF front-end

There are three major sources of interferences. One is when a nearby transmitter is using a frequency close to the desired signal. Then the spectrum of the unwanted signal can contain enough power to jam the communication in the adjacent channel. The situation is depicted in Fig.4(a). Another source of interference comes from intermodulation distortion from mixing of two signals. If the distance between the two interferers and the the distance between the closest interferer and the wanted signal is the same, then the third-order intermodulation product (IM_3) will overlap with the desired signal. This situation is shown in Fig.4(b).
The last source of interference results when a nearby transmitter has enough power to saturate the receiver. Then the receiver is completely blocked and the desired signal cannot be detected. This situation is depicted in Fig.4(c). The most dangerous source of interference is the IM_3. The amplitude of the intermodulation product is proportional to the third power of the interference level. Therefore, if the interference is strong enough, then the intermodulation product can completely jam the communication.

To quantify the linearity of a radio the third-order intercept point (IP_3) has been defined. Supposing that the two interferers have the same power level, the SFDR is defined as the point at which the level of interferences produces an intermodulation product, which equals the receiver noise floor (N_{RX}). Assuming decibel units we get

$$
\begin{aligned}
SFDR &= \frac{2}{3}(IP_3 - N_{RX}) \\
&= \frac{2}{3}(IP_3(RX_{\mathrm{sens}} - SNR))
\end{aligned} \qquad (14)
$$

assuming an indoor environment scenarios, from (5) and (6) two 0-dBm transmitters at 3 meter give an interference level of -41.2 dBm (P_{I}). Considering f_0 the central frequency of the desired radio channel, two interferers place at $f_0+\Delta f$ and $f_0+2\Delta f$, will generate an IM_3 product, which will fall in the desired channel. In decibel units the power level of the IM_3 product will be

$$
P_{IM3} = IP_3 + 3(P_I - IP3) \qquad (15)
$$

When the IM_3 is at least the required SNR below the desired signal, then the BER will not be spoiled. Indeed, a desired signal weaker than the receiver sensitivity will be noise limited rather than interference limited.

Therefore, considering the minimum detectable signal (-78 dBm), from (15) and (11) we obtain $IP_3 \geq$ -20 dBm for the BER to be 0.1% or lower. Therefore, the required SFDR will be from (14) 42 dB, which seems reasonable from a low-power implementation point of view.

C. Selectivity and Phase Noise considerations

Two of the main characteristics of the radio system described in this paper are low-cost implementation and low-power. To achieve the aforementioned goals external filters should be avoided. Unfortunately it is not possible to achieve sufficient selectivity at high IF. Therefore, a zero-IF seems the way to achieve low-cost implementation as well as low-power consumption. Well known problems such as self-mixing of the local oscillator, or flicker noise can be greatly alleviated by using a large modulation index together with AC coupling [5].

At the receiver side, as well as at the transmitter side, the phase noise of the oscillators is of major concern. At the transmitter side the leakage of the LO in the adjacent channels can corrupt a signal transmitted by another device, while at the receiver side, the main concern is the reciprocal mixing, which can downconvert an unwanted signal in the band of interest (in this case at DC due to the zero-IF architecture chosen).

Looking at Fig.5, the wanted signal is downconverted to zero-IF by the local oscillator. Unfortunately, if a strong unwanted signal is present somewhere in the sidebands of the noisy local oscillator, then a band of phase noise equivalent to the signal bandwidth will be downconverted at zero-IF and will overlap with the wanted signal. This can seriously degrade the capability of the receiver to correctly demodulate the wanted signal.

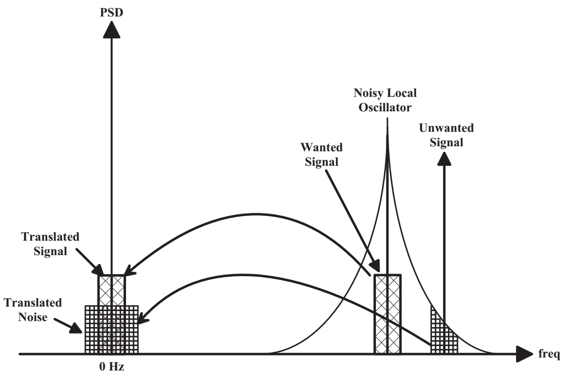


Fig. 5. Reciprocal Mixing

In [6] it is shown that for the narrowband case, the phase noise slope can be neglected across the bandwidth. Therefore, in the wanted bandwidth a constant value of the Power Spectral Density (PSD) of the Local Oscillator (LO) will be used (which can be reasonably chosen as the value in the middle of the band). Furthermore, to simplify the calculations, we will assume an ideal transmitter and we will degrade the receiver performance by 3 dB: this assume equal phase noise performances at the receiver as well as the transmitter [7]. Therefore, we can calculate the Phase Noise requirements for the LO in decibel units by the following:

$$L_{\Delta \mathrm{f}} \leq P_{\mathrm{WS}} - P_{\mathrm{US}} - BW - SNR - NF - 3dB \quad (16)$$

where P_{WS} is the power of the wanted signal, P_{US} is the power of the unwanted signal, $L_{\Delta \mathrm{f}}$ is the phase noise expressed in dBc/Hz at a certain frequency offset from the carrier, BW is the noise bandwidth and the 3 dB takes in account the degradation due to the transmitter.

TABLE I

Power difference between wanted and unwanted signals for Bluetooth

Channel	$\mathbf{P_{unwanted}} - \mathbf{P_{wanted}}$
Adjacent	0 *dB*
Alternate	30 *dB*
Third and beyond	40 *dB*

Now a reasonable level for the wanted and unwanted signal should be chosen. Looking at Bluetooth specifications, the power level difference between the unwanted and the wanted signals has been chosen to be as shown in Table I. In the Bluetooth standard, system simulations have proved that these protection ratios, result in a very little degradation in the throughput. Therefore, it is reasonable to suppose that, when the unwanted signal is in the adjacent channel (150 kHz apart), then wanted and unwanted signal can have the same power (it means that they can be at the same distance). In this case the allowed phase-noise from (16), given NF=15 dB, is $L_{\Delta \mathrm{f}}$=-70 dBc/Hz.

In the alternate channel (it means 300 kHz apart), we suppose that the unwanted signal can have a power 20 dB higher than the wanted one. From (8) and (9), given n∈[2,3], this corresponds to a relative distance between wanted and unwanted transmitters between 5 and 10. In this case, $L_{\Delta \mathrm{f}}$=-90 dBc/Hz.

Finally, in the third and beyond channels (it means more than 450 KHz apart), it is reasonable to set the power of the unwanted signal 30 dB higher than the wanted one. This corresponds to a relative distance between wanted and unwanted transmitters between 10 and 30 and translates in $L_{\Delta \mathrm{f}}$=-100 dBc/Hz.

The most demanding phase noise requirement is therefore at 450 KHz far from the carrier. Of course it is possible to trade NF for phase noise (see equation

16) to relax as much as possible the radio specifications. The phase noise requirement will be at 3 MHz far from the carrier around -116 dBc/Hz, that is less demanding than what is found in typical Bluetooth systems. If resonators with a quality factor Q of 10 are available, then an ultra-low power VCO can be designed.

IV. Conclusion

A complete system-level analysis for a FHSS, low data-rate radio system has been presented. It has been shown that correct choices in the design of the air interface yield to a relaxed requirements on the radio side.

The most important parameters for a radio system working in the 902-928 MHz band have been derived and the procedure can be easily extended to the 2.45 GHz ISM band. It has been demonstrated that the radio system requirements are less tight than those commonly found in Bluetooth systems. This can directly translate in a single-chip, low-cost, ultra-low power transceiver allowing, therefore, the implementation of an autonomous node with a very-long battery lifetime or which, can even harvest the energy required to transmit and receive data, from the ambient.

Acknowledgment

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