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Software Defined Channel Sounder for Power Line Communications

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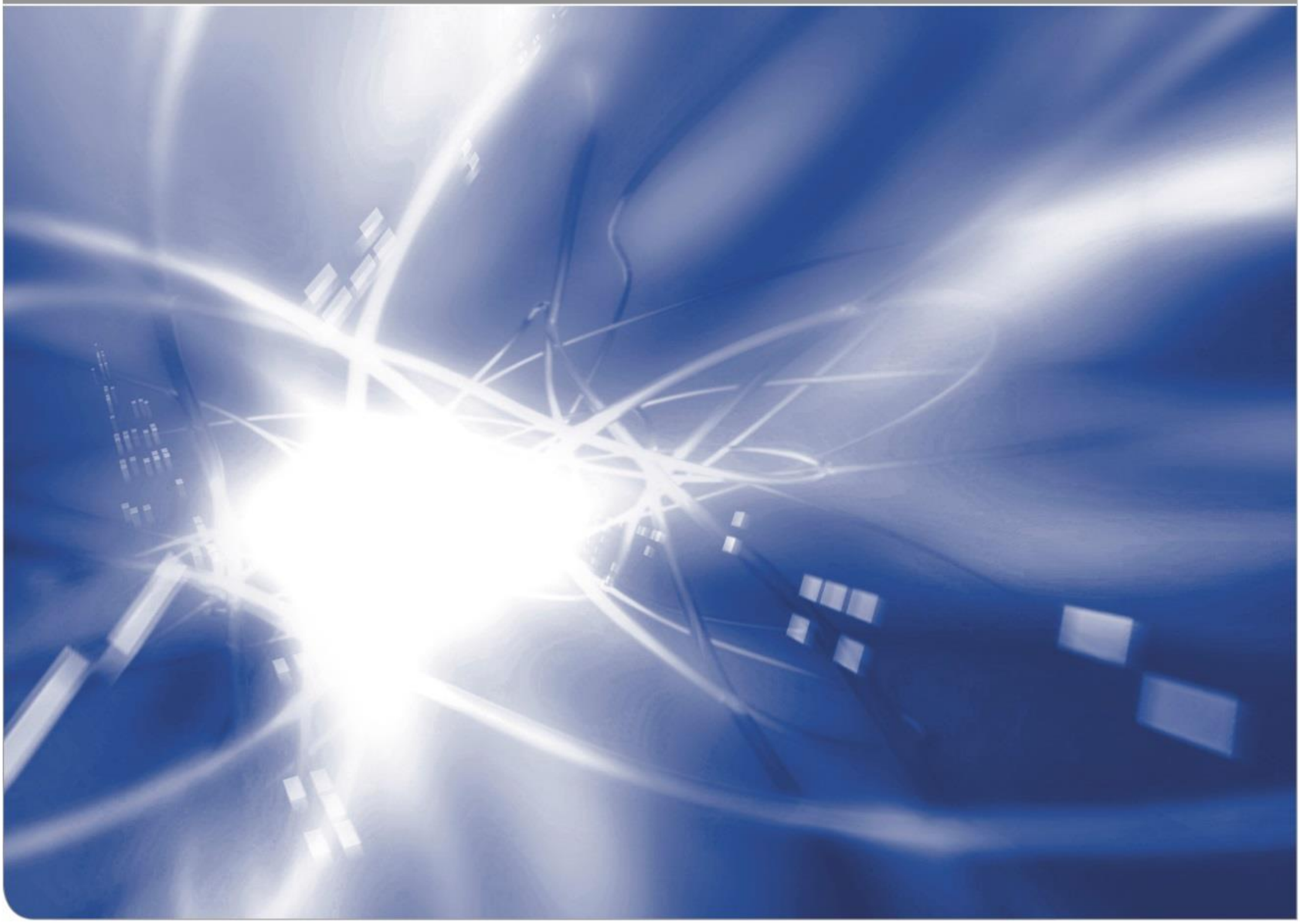

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Abstract

The knowledge of the communication channel properties is of high importance during the deployment of a communication system. One of the most important properties is the channel transfer function, which can be obtained by means of different channel sounding techniques. The topic of channel sounders for power line communications has been not so deeply investigated as for wireless communications. There are, however, some already published implementations, but most of them are based on very expensive lab instruments. The target of this work is to present self-implemented channel sounder techniques intended for power line channels based on software defined radio (SDR) platforms. These platforms can also be later employed as communication modems only by changing the implementation in the software layer. The re-usability of the hardware results in a decrease of cost and deployment times. We implemented two channel sounding techniques: frequency hopping (in frequency domain) and sliding correlator (in time domain). Along this work, we explain the complete implementation of the SDR-based channel sounder for power line communications. Finally, we present the performance test and comparison of both channel sounding techniques. The results suggest that the frequency hopping method might be a better candidate for the application in power line channels due to the expected low signal-to-noise ratio. In the future, an appropriate analog front-end must be designed, in order to operate this channel sounder in real power line channels.

Contents

1 Introduction

1.1 Power line channel sounder

For the deployment of any communication system, the knowledge of the communication channel is a key factor. Therefore, the design of channel sounders has been always a topic of great interest in the field of communications. These are devices capable of measuring the channel transfer function (CTF). They have been already extensively investigated in the past in the field of wireless communications and there is an extensive related work, e.g. [1]. However, the situation is not same for power line channel communications (PLC). Additionally, most of the related work traditionally employed very expensive lab instruments, such as vector network analysers, spectrum analysers or high frequency signal generators [2–4]. All these instruments are later also not part of the communication system. Hence, the duration of the deployment increases. A software defined platform can avoid these disadvantages.

1.2 Software defined platforms

The idea beyond software defined radio (SDR) is to implement many operations and signal processing involved in a radio communication system in software [5]. The same concept can be followed in the field of PLC. The so-called software defined power line communication (SD-PLC) has gained a lot of attention in recent years. SD-PLC can provide a rapid prototyping environment, pushing PLC development [6, 7]. Our proposal consists on employing these software defined platforms also as channel sounder. This would allow us to re-use the same hardware platform for this application only by modifying the software.

1.2.1 Strengths and challenges

The employment of SDR platforms as channel sounders offers several advantages. Firstly, the re-usability of the hardware platform would decrease the deployment costs, since the already mentioned lab instruments would not be longer necessary. Another advantage is the independence between the transmitter and receiver modules, in contrast to vector network analysers. This enables the measurement of long distance channels without problems. Furthermore, if the devices count with an internet connection, they can be operated remotely and store the measurement results online in real-time.

However, there are several challenges while measuring power line channels, whose properties differ from those of the wireless ones. For example, the high attenuation and high noise power lead to a high transmitter power in order to have an acceptable signal-to-noise ratio (SNR) at the receiver side [8]. Furthermore, the noise is not only additive white Gaussian noise (AWGN), and both the CTF and the access impedance are time-variant [4].

1.3 Related work and our contribution

Some work has been already published in the field of software defined channel sounders. Islam et al. [9] implemented similar channel sounding methods also with same SDR platforms that we employed. However, they were only tested on wireless channels and there is no comparison between both methods or with a reference instrument. Black et al. [10] implemented a tool with USRPs for the characterization of power line channels, but he did not analyzed its performance under different SNR environments. There are also many published FPGA-based implementations, as for example [11].

In this work, we implemented two different techniques, whose performance was evaluated and compared on emulated power line channels. The system is capable of measuring the channel until a frequency of 30 MHz, what covers both narrow-band (NB) and broad-band (BB) PLC.

1.4 Paper organization

The remaining of the paper is organized as follows: Section 2 explains the implementation of both channel sounding methods in the SDR platforms. The experimental validation of the system is performed in Section 3. Finally, Section 4 concludes the paper.

2 Software defined implementation of the channel sounder

We employed the universal software radio peripheral (USRP) devices from Ettus Research as SDR platforms, in particular the USRP N210 and X310 models, and the GNU Radio Toolkit. The employed daughter boards were the LFTX at the Tx side and LFRX at the Rx side. These are capable of working in the frequency range from 0 Hz to 30 MHz.

2.1 Channel sounding techniques

We implemented two channel sounding techniques. For both of them, we employed two USRPs, each of them commanded by one host computer, independently from the other one. One USRP works as transmitter (Tx). It excites the channel with a predefined signal. The second USRP acts as receiver (Rx), sensing the response of the channel to the excitation. As these SDR platforms have both input (Tx) and output (Rx) impedances matched at 50 Ω, what they actually measure is the S-parameter S_{21} of the channel.

2.1.1 Frequency domain technique - Frequency hopping

The frequency hopping (FH) technique obtains the channel transfer function (CTF) in the frequency domain. The Tx employs a variant narrowband signal as excitation: a sinusoidal signal with constant peak amplitude but whose frequency is modified discretely at predefined time intervals. The receiver also knows the sequence. It measures the amplitude of the received signal for each particular frequency and then, as the transmitted signal is previously measured (see Section 2.4) , the CTF along the frequency range can be calculated.

The desired tone is generated in the following way: firstly, a tone with a constant and predefined low frequency (e.g. 10 kHz) is generated in software (GNU Radio), i.e. in baseband. This signal is shifted (up-converting) to the desired frequency value along the frequency range of interest by re-tuning the USRP's center frequency. This is digitally done by the USRP's FPGA, since the LFTX and LFRX boards do not count with a analog mixer (up-covnerter) as the other USRP daughter boards. Finally, the analog signal is generated by USRP's digital to analog converter. In the Rx, the process is the opposite: the high frequency tone is digitalized by the USRP's analog to digital converter, down converted to the low frequency in baseband by the FPGA, and finally processed in software.

2.1.2 Time domain technique - Sliding correlator

The sliding correlator (SC) method employs a wideband signal and operates in time-domain. The Tx excites the channel with a known sequence, whose autocorrelation function has similar characteristics to the delta-dirac function, i.e. high peak at $\tau = 0$ and low sidelobes [1]. We employed a maximum length sequence (MLS) for this purpose. At the Tx side, we used a root-raised-cosine (RRC) filter to minimize the inter-symbol interference. At the Rx side, the receive filter is matched to the transmit filter and a polyphase filterbank clock is employed for timing synchronization. After that, the received signal is correlated with a copy of the sended MLS.

The channel impulse response (CIR) can be written in discrete time domain as following:

$$h_n = \sum_{l=0}^{L_h-1} \alpha_l \delta[n-l], \tag{1}$$

where h_n is the CIR, α_l are the CIR coefficients at the l-samples. L_h is the length of the CIR. The length of the MLS is $N = 2^q - 1$, where q is the degree of the MLS. N must be chosen equal or longer than L_h. Then, since the Tx and Rx are synchronized and their filters (RRC) are matched, the output signal of the correlator is, following the work of Islam et al. [9]:

$$R_{cy} = \sum_{l=0}^{L_h-1} \alpha_l \cdot R_{cx}[n-l] \tag{2}$$

where x is the signal sent by the Tx (a continuous repetition of the MLS) and c is the MLS, whose values are either +1 or -1. R_{cx} is the correlation function between c and x, which is known. Based on the MLS properties, it results:

$$R_{cx} = \begin{cases} N & n = \cdots, -2N, -N, 0, N, 2N, \cdots \\ -1 & \text{otherwise.} \end{cases} \tag{3}$$

The coefficients α_l are calculated by means of Equations (3) and (2), and the CIR is reconstructed with Equation (1).

2.2 Synchronization between nodes

Ensuring a highly accurate synchronization between the Tx and Rx is very important in channel sounders. There are two possibilities with the USRP devices: by means of one global positioning system (GPS) device at each communication node or by employing the MIMO cable from the manufacturer [12]. We employed the alternative based on GPS because in that way, Tx and Rx could be placed at long distance among them, as long as both of them are capable of receiving the GPS signal. The GPS provides timing and frequency synchronization. However, because of the USRP hardware characteristics, the carrier phase synchronization cannot be accomplished with the GPS [12]. There is always a phase offset between the center frequencies of both USRPs, and therefore, the system would measure actually the channel phase offset plus the phase offset between the center frequencies of the USRPs. The last phase offset changes randomly every time that the carrier frequency is set, therefore it cannot be permanently compensated easily. Hence, our system is currently only capable of measuring properly the absolute value of the CTF or of the CIR, but not the phase.

2.3 Processes scheduling

The GPS also provides the devices with a very accurate and shared time reference. This can be employed in order to coordinate the scheduling of both nodes. In the case of the SC, the Tx sends constantly the same MLS. At the other side of the channel, the Rx stores the necessary amount of samples (according to its configuration) and then it calculates the CIR and CTF. Under this method is only important that the Tx starts to excite the channel before the Rx begins to sample it.

On the other hand, in the FH the process scheduling is more sophisticated. Every time that the USRP's f_c is changed in software, the USRP needs some time to retune its PLL and to recalculate a digital filter implemented in the USRP's FPGA [13]. Therefore, every time that f_c is changed, there is a preguard time in which the acquired samples are not stored for the measurements. We set empirically a pre-guard time of 20 ms. After that, there is a configurable time interval in which the samples are stored for CTF calculation. When this interval is finished, the USRP hardware driver (UHD) commands are used for retuning the USRP [14].

In the case of the FH, the Tx and Rx have a table with the timeslots at which the f_c must be changed and the corresponding frequency values in order to operate synchronously. Every time that the USRP's f_c changes or during the start-up, the USRP source block from GNU Radio, used to read at the host computer the signal sampled at the Rx input of the USRP, propagates together with the first sample value a tag message [15]. A tag is a metadata information that propagates between blocks in a flow graph in GNU Radio. In this case, the tag contains the current time information. Based on this and knowing the USRP's sampling rate, the host computer can always update the USRP local time when it receives a new set of samples. Since both host computers need to know the current local time of their USRPs, there is also a USRP source block at the Tx's host computer, which read samples only to update the local time.

2.4 Calibration

The USRPs are not designed as a measurement system, so they do not have a planar frequency response and they are also not calibrated. Therefore, we carried out an easy calibration process to compensate this issue. We connected the output of the Tx USRP to the input Rx USRP but by means of an attenuator (e.g. of 30 dB). Otherwise, a direct connection could damage their hardware [16]. Then, we used this measured CTF to compensate the next channel measurements.

2.5 Zero crossing detection

Our system is also capable of storing the power line zero crossings information. This would allow us to analyse the time-variant of the CTF based on the cyclostationary behaviour of the power line channels [4]. This is carried out by employing a circuit for zero-crossing detection, which has been described in [17]. This circuit converts the analogue mains voltage into a digital synchronization signal with a settable output voltage. This signal is applied to the second input of the receiver USRP, with the output voltage set to less than 1 V to avoid hardware damage.

3 Performance evaluation and comparison of the implemented sounding techniques

In this Section we validate and compare the implemented software defined channel sounding techniques under the effect of an emulated power line channel.

3.1 Testbed

Our target is to compare the measurement performance of the different channel sounding techniques. Therefore, we need a stable channel in order to get reproducible results. This is not the case of typical power line channel, which are strongly time-variant, as already mentioned. Therefore, we employed a low speed power line emulator (LS-PLE), developed in the past by our group [18–20]. This FPGA-based emulator can emulate customize power channels up to a frequency of 500 kHz (NB-PLC) and even emulate the typical power line noise types with a configurable noise power level. By employing this emulator, the channels can be set to time-invariant. In this work, we present the results of measuring one emulated power line channel with additive white Gaussian noise (AWGN).

As the LS-PLE does not emulate the mains voltage, the null detector has not been tested under this testbed.

3.2 Excitation signals and emulated noise

We set the output signal of the USRP Tx to the maximum possible value according to its dynamic range, in order to employ the maximum possible transmit power without clipping. We first connected a spectrum analyser (SA) at the output of the LS-PLE to measure the power spectrum density (PSD) at the Rx side. This was carried out while setting the emulating noise power to null. The results are shown in Fig. 1. It can be observed that the FH signal has a higher PSD than the SC. The reason is that as the FH employs a narrow band signal, the power density over each frequency point of interest is much higher than the one of the SC, which spreads the signal energy over a much wider frequency band.

The channel sounding methods was tested under 7 different emulated noise levels, which were denominated from NL0 to NL6. The selected noise type was always AWGN. A higher number implies a higher noise level (NL). NL0 corresponds to not emulated noise power. The noise presented in this case at the Rx is produced meanly by thermal and quantization noise. We measured the PSD of each these noise levels at the Rx side with the SA, setting the output signal of the Tx USRP to null. The results are also presented in Fig. 1. It can be observed, that the noise PSD behaves as

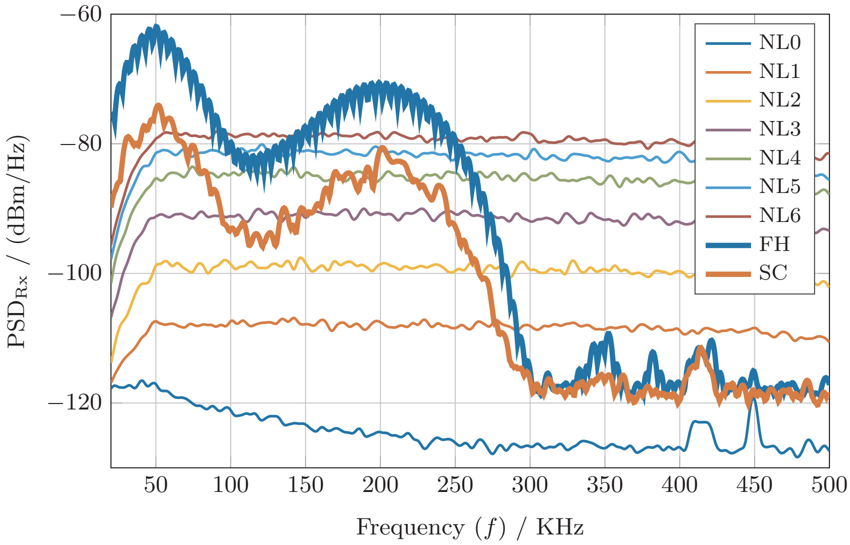


Figure 1: Measured PSD at Rx for the different NLs (with the Tx USRP output signal set to null), and with FH and SC methods when the emulated noise power is null (NL=NL0).

practically white noise only above 30 kHz. This is due to a limitation of the LS-PLE. Additionally, we calculated and depict the SNR at the Rx side with both CS techniques for each NL value in Fig. 2. There it is observable that the SNR with the FH method is 12.54 dB higher.

3.3 Comparison of the methods performance

We measured the CTF of the LS-PLE firstly with a vector network analyser (VNA) and without emulating any noise to have a reference, i.e. a ground truth. Then, we connected the Tx USRP to the input of the LS-PLE and the Rx USRP to the output. We measured the channel with our channel

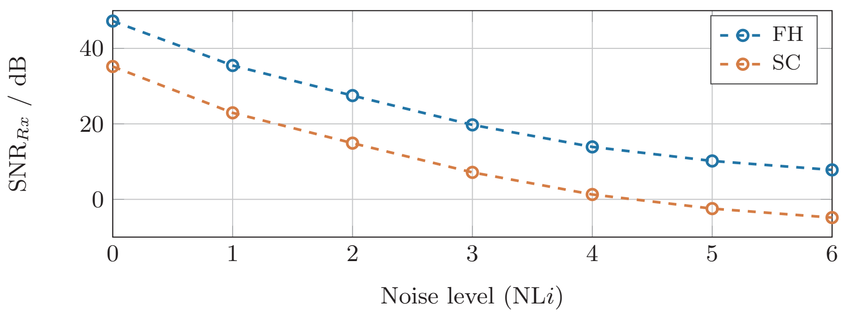
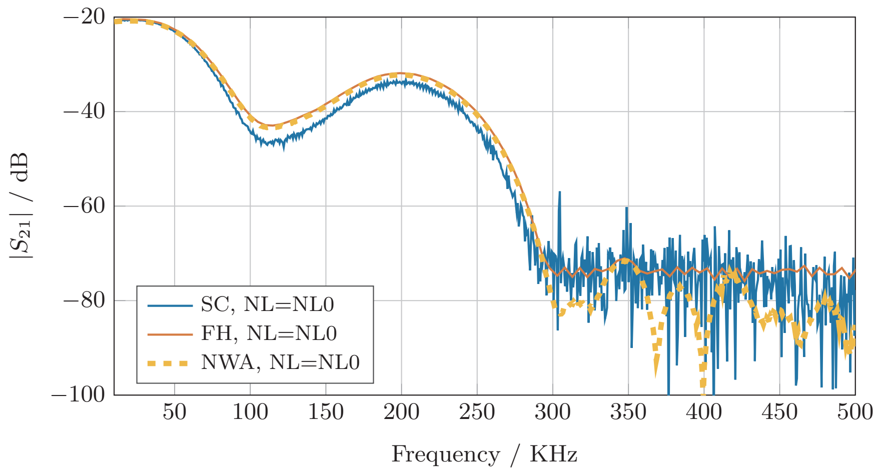


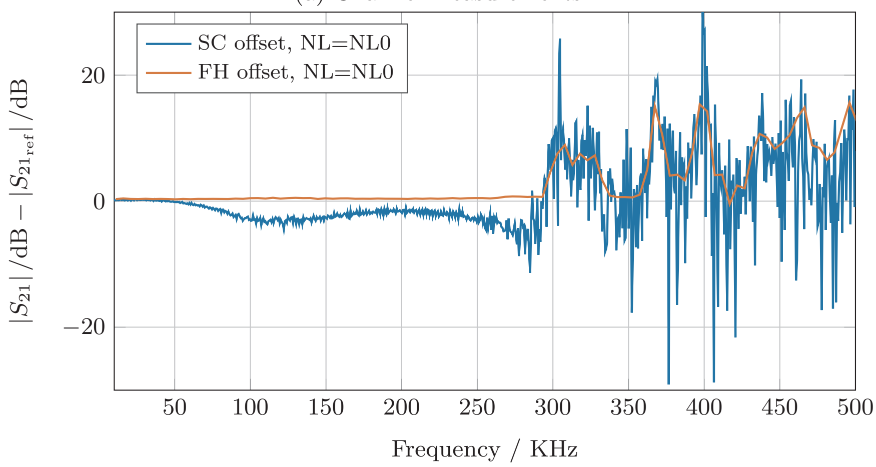
Figure 2: SNR at Rx for SC and FH as a function of the NL.

sounder employing both methods and varying the noise level, in order to analyze the influence of the SNR in the performance of the methods.

The CTF measured with the VNA, the SC and the FH methods without additional noise generated by the LS-PLE, i.e. NL=NL0, are shown in Fig. 3. In this situation, the FH has a very good performance: its result is very similar to the one of the VNA until an attenuation of around 70 dB. This situation is even clearer with the results shown in Fig. 3b, where the difference between the reference and each of the other two cases is illustrated. On the other hand, the SC's results presents a significant offset respect to the VNA measurement even at lower attenuation, due to its lower PSD in

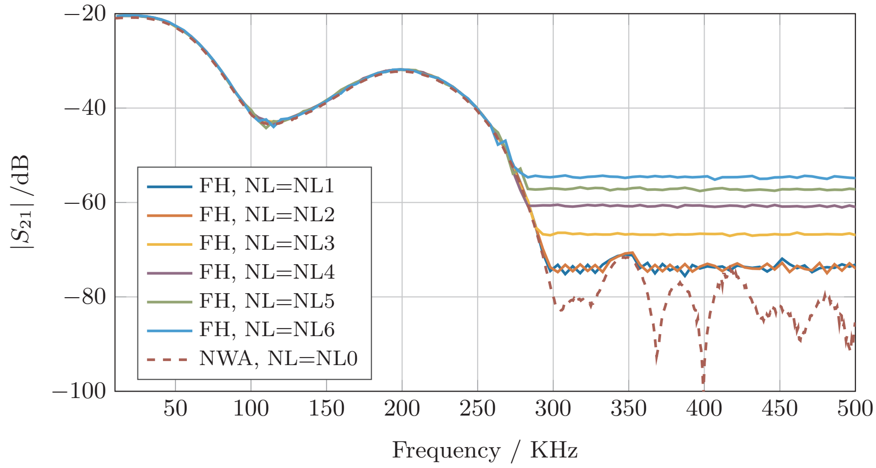


(a) Channel measurements.

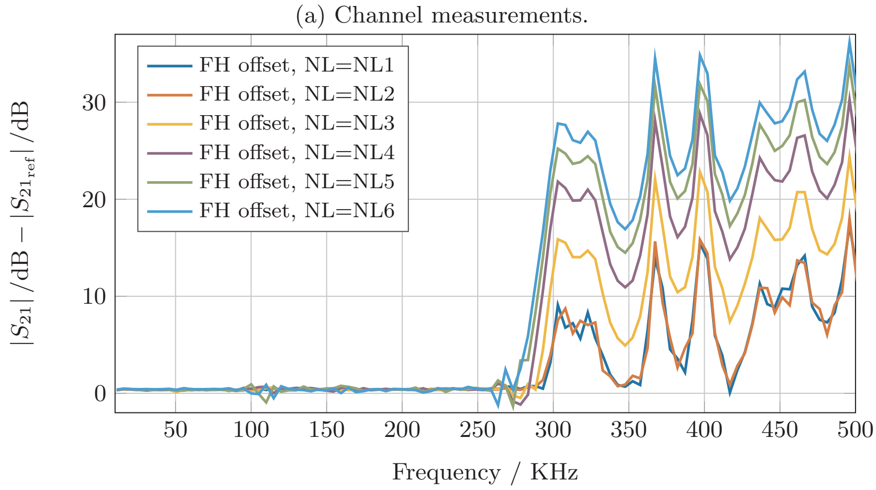


(b) Offset respect to the reference.

Figure 3: Measurement results with the VNA and both channel sounder methods in the LS-PLE, without additional emulated noise.



(a) Channel measurements.



(b) Offset respect to the reference.

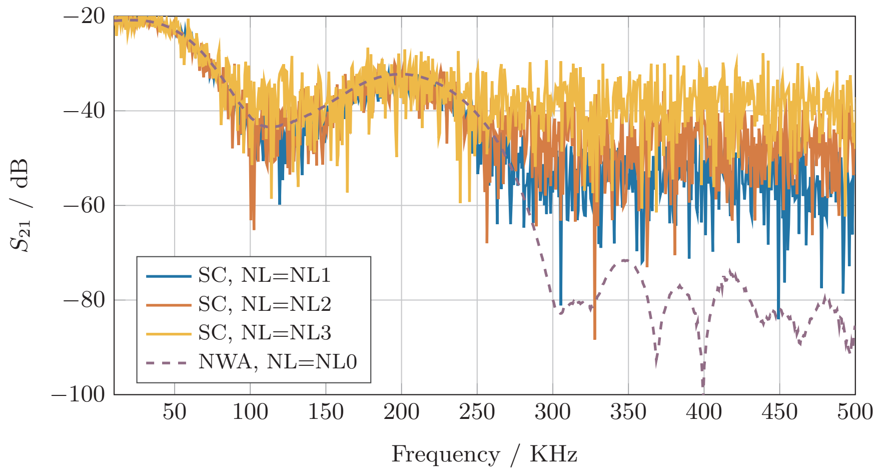
Figure 4: Measurement results of the FH method under different noise levels in the LS-PLE.

comparison with FH.

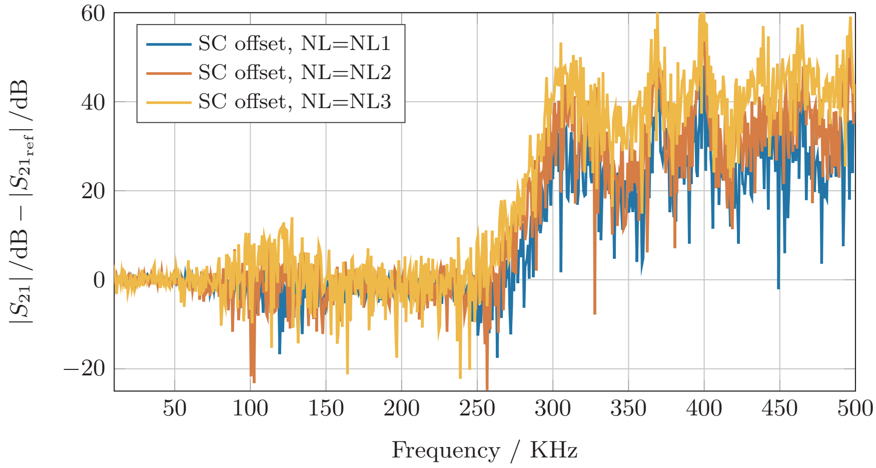
The next step is the performance comparison of both methods when the SNR decreases, since in PLC is normal to work under very low SNR. We measured the same channel on the LS-PLE but with the remaining six noise levels. The results with the FH method are displayed in Fig. 4a, and with the SC method in Fig. 5a. In the latter, only the results until NL3 are shown. The reason is that, under the higher noise levels, the effect of the noise of the measurements is too high, that the figure would be unintelligible if all the results were included.

The offset between the results of the reference and the other results are despicted in Fig. 4b for the FH and in Fig. 5b for the SC. As expected, with both methods the bigger the NL, the smaller the SNR at Rx, and consequently, the offset with the reference result increases, i.e. the measurement is less accurate. The bigger the noise at Rx, the little the maximum measurable attenuation.

The are also another important facts from both channel sounding techniques to compare. For example, the maximum measurable frequency range with each method. In the case of the SC, it measures the complete frequency interval in baseband, keeping the USRP's central frequency constant. That



(a) Channel measurements.



(b) Offset respect to the reference.

Figure 5: Measurement results of the SC method under different noise levels in the LS-PLE.

means, that the frequency range is limited by the bandwidth of our system. In the case of the USRP with the LFTX or LFRX daughterboards, the maximum bandwidth is even greater that the one of interest in PLC. However, the limitation is normally the processing speed of the USRP's host computer, i.e. the amount of samples per second (sps) that it is able to process. In our experience, we were capable of working until 5 MSps. On the other hand, since the FH operates modifying the USRP's central frequency, the limit of the measurable frequency interval is determined by the central frequency frequency range of the USRP's daughterboard. That means until 30 MHz with the LFTX and LFRX. That is a great advantange for the FH. Currently, there is even a daughterboard, the UBX, offered by the USRP manufacturer which allow the USRP to tune its central frequency between 10 MHz and 6 GHz, in case that we need to measure the CTF at higher frequencies.

However, the SC offers the advantage of measuring its complete frequency range much faster. It needs only some repetitions of the excitation signal sequence, which takes only some milliseconds. This is would be important in PLC channels in case we would want to analyze its time-variant behaviour. With the FH, it is in principle only possible to measure the long-term behaviour of the CTF.

4 Conclusions and further work

4.1 Conclusions

In this work, we presented the implementation with SDR platforms of two channel sounding techniques for power line channels. The developed techniques are frequency hopping and sliding correlator, which operate in the frequency and time domains, respectively. Both use the same hardware platforms.

In order to validate our system, we tested both channel sounding methods on the SDR platforms with a power line channel emulator, since an analog front-end for power line has not been developed yet. The results of both methods has been compared with a VNA measurement, chosen as ground truth. Then, we studied the performance degradation of the channel sounder when the SNR at Rx decreases. From these experiments, we proved that both techniques can performed well under high SNR. But under low SNR, which is a typical case in PLC, the frequency hopping performs clearly better. Therefore, the latter is a better candidate for the use on power line channels.

4.2 Further work

We will use this system to measure real power line channels, and compare the results with the system designed previously by our group [4] in the range of NB-PLC. In order to achieve this, an additional analog front-end to connect the USRPs to the powerline is currently under development. This front-end must be capable of coupling and decoupling the channel sounder signal at the transmitter and receiver side, respectively. It must provide a certain required gain in order to operate with an acceptable SNR at the receiver. And it must also ensure a safe connection between the USRP and the power line by blocking the main voltage.

Finally, regarding the zero crossing detection, we expect to carry out the signal processing of the detector output directly in the USRP's FPGA. The detector output signal should be injected in one of the FPGA general purpose input/output pins and the FPGA program modified.

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