

Experimenting Waveforms and Efficiency in RF Power Transfer

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Abstract— In this paper, we present a prototype test-bed for radio-frequency (RF) wireless power transfer (WPT) comprising a software-defined radio (SDR) transmitter and an energy harvesting receiver with a diode-based rectifier. The test-bed allows us to study the end-to-end efficiency of RF WPT when employing different co-phased multisine waveforms. In particular, we analyze the clipping and non-linear behaviours of the transmitter by experimentally evaluating how they affect the performance of waveforms. The experimental results indicate that transmitting impulse-like signals is actually not optimal for RF WPT in practice despite they would be ideal in terms of rectifier efficiency. Instead, the results highlight the superior performance of single-tone signals over co-phased multisines in terms of both end-to-end WPT efficiency and spectral purity.

Keywords— Wireless power transfer, RF energy harvesting.

I. INTRODUCTION

The buzz around 5G is escalating each passing day. Along with a ten-fold increase in data rates as compared to LTE-Advanced, 5G is expected to network millions of sensors into the so-called Internet of Things (IoT). Such an extreme-scale deployment would require the sensors to be equipped with replenishable energy sources. In recent years, far-field radio-frequency (RF) wireless power transfer (WPT) is being considered as a possible solution to this challenge. Conventionally, it has been believed that, due to large over-the-air propagation losses, RF WPT would be unable to deliver sufficient energy to the sensors for them to operate. However, the energy consumption of simple devices for sensing and performing computations has fallen drastically over the years in accordance to the Koomey's law [1], which has reignited the research interest in RF WPT of late.

Most of the recent scientific studies on RF WPT develop the theoretical foundations for the technology such as rectifier design, waveform design, beamforming algorithms, co-existence with information transfer, etc. A comprehensive survey of such research works is available in [2]. An important consideration in the design of a WPT system is its power efficiency. The end-to-end WPT efficiency is the ratio of the direct current (DC) power harvested at a receiver to the DC power supplied to a transmitter. This can be subdivided into DC-to-RF conversion efficiency at the transmitter, over-the-air RF-to-RF transmission efficiency, and RF-to-DC conversion efficiency at a rectifier. It is interesting to note that almost every study on RF WPT focuses on the RF-to-DC conversion efficiency at the receiver only, while the overall DC-to-DC efficiency of such systems has been conveniently neglected.

As reported in [2] and references therein, co-phased multisine waveforms exhibiting high peak-to-average power

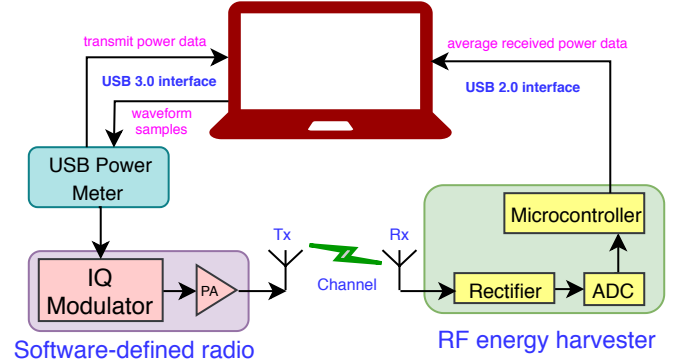


Fig. 1. Block diagram of the presented RF wireless power transfer system.

ratio (PAPR) are considered to be ideal candidates for RF WPT. However, the designs that lead to this intuitive result have ignored the presence of a non-linear power amplifier (PA) at any transmitter. Ideally, efficient RF-to-DC conversion would require the received signals to have as high PAPR as possible (with an impulse being the ultimate waveform). The problem with this is the non-linearity and that amplitude peaks drive the PA into saturation, thereby distorting the shape of the actual transmitted waveform. With their envelope distorted, the multisine waveforms might not remain optimal for WPT anymore. To the best of the authors' knowledge, there has not been any previous work that verifies this possibility.

In this paper, we present a prototype test-bed (*cf.* Fig. 1) for measuring the overall end-to-end DC-to-DC power efficiency of RF WPT with different multisine waveforms. Towards the end, we report the optimal waveforms for different scenarios and the associated trade-offs. Thus, the rest of the paper is organized as follows. Section II surveys previous RF WPT test-beds. Section III presents the system model used in this work. We discuss the configurations of the test-bed and report the experimental results in Section IV. Finally, the conclusions of this work are presented in Section V.

II. RELATED WORK

The authors in [3] developed a test-bed that harvests energy from multiple sources (*viz.* RF energy, thermal energy, and solar energy). The performance metric was the time required to charge a 0.22 F super-capacitor. The three sources work in parallel for robust operation, each producing a 3.3 V output voltage charging the super-capacitor in about seven to eight minutes, much quicker than any of the individual sources. The work in [4] involved developing a WPT test-bed

for studying data acquisition from wireless sensor nodes in a smart city environment. An interesting observation is that the lowest amount of RF energy is harvested when the transmitter and receiver antennas are orthogonal to each other. A blind adaptive beamforming scheme for multiple-input-single-output WPT system was implemented on a Universal Software Radio Peripheral (USRP) test-bed in [5]. The beamformer weights at the transmitter are updated based on feedback from the energy harvester to enhance the energy output. It is observed that the usage of multiple antennas enhances the range of WPT.

A scenario with multiple power sources and multiple energy harvesters was envisioned in [6]. The authors proposed a charging protocol that caters to the charging efficiency of the sensors while another one to ensure that all the power-constrained sources have equal energy reserves. The first closed-loop WPT prototype based on pilot-based channel estimation and subsequent waveform optimization was presented in [7]. The authors designed a multisine waveform whose weighting coefficients are derived directly by a matched filter approach, thereby rendering low complexity. Based on channel estimates, more power is allocated to the sinusoids corresponding to higher frequency-domain channel gains. The authors also presented the design parameters for a rectenna fabricated for this prototype. The designed rectenna is claimed to achieve an RF-to-DC conversion efficiency of 12% at -20 dBm with a single-tone input.

Another test-bed based on beamforming was reported in [8], where multiple power transmitters charge a particular receiver. Here, the distributed beamforming technique aligns the phases of the various RF waves at the receiver so as to obtain a multisine waveform with high PAPR at the receiver input. The outright benefits of this technique involve concentrating energy only at the targeted receivers leading to power savings which indirectly also suppresses unwanted excessive RF energy exposure at other points in the network.

Addressing the problem of antenna directivity in WPT, the authors in [9] proposed a notion of omnidirectional charging whereby sensors with any orientation located within a predefined area are charged up to a certain threshold. The experiments conducted with a transmitter–receiver separation varying between 40 cm and 100 cm yielded results which lead to the conclusion that the sensor node can charge up to some threshold only if the orientation angle is between -60° and 60° . All the above reported test-beds operate in the 915 MHz or 2.4 GHz industrial, scientific, and medical (ISM) bands with USRP hardware, not so unlike our study.

III. SYSTEM MODEL

The main objective of our software-defined radio (SDR) test-bed is to study the overall end-to-end efficiency of RF WPT with different test waveforms. To do so, along with the average harvested power in the end, we also measure the total power supplied to the transmitter to begin with. A diagram of the RF energy transmission and harvesting system is shown in Fig. 1, and a photo of the test-bed is shown in Fig. 2.

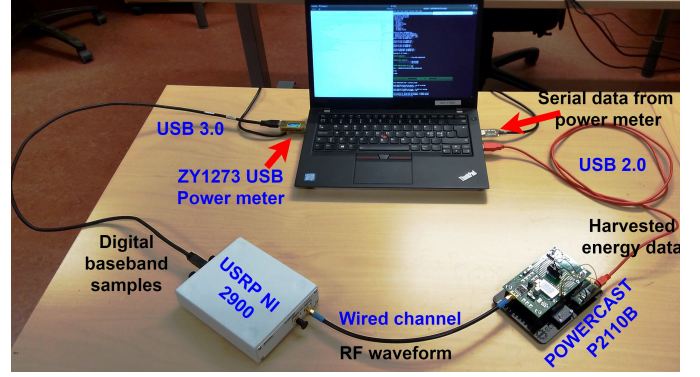


Fig. 2. The presented SDR-based prototype test-bed for RF power transfer.

The end-to-end (overall DC-to-DC) efficiency of the RF power transfer system is defined as

$$\eta = \frac{P_{\text{out}}^{\text{DC}}}{P_{\text{in}}^{\text{DC}}}, \quad (1)$$

for which the system measures $P_{\text{in}}^{\text{DC}}$ and $P_{\text{out}}^{\text{DC}}$ as follows.

- The transmitter side comprises a computer that generates the digital baseband waveforms to be transmitted. The computer sends the digital samples of the waveforms to the SDR (namely, USRP) via a USB 3.0 interface, which also powers the radio. A USB power meter is placed in series between the computer and the USRP to measure the power supplied to the USRP for its operation ($P_{\text{in}}^{\text{DC}}$). The USRP transmits the modulated RF signal over a wired channel to a receiver.
- The receiver is an off-the-shelf RF energy harvester board. It is equipped with a diode-based rectifier that converts the incident RF energy into DC. The RF energy harvester board is accompanied by a discrete analog-to-digital converter (ADC) and a microcontroller to continuously measure the instantaneous DC voltage over a resistor and report the information to the computer in real time via a USB 2.0 interface. The computer logs this data to calculate the average harvested power ($P_{\text{out}}^{\text{DC}}$) over each transmission interval.

Let us then take a look at how the baseband test waveforms may get distorted before transmission from the USRP, and how it would affect the optimality of multisine waveforms. Consider the co-phased N -tone multisine signal

$$x(t) = A \sum_{n=1}^N \cos(2\pi f_n t), \quad (2)$$

where A is the amplitude of the multisine waveform and f_n is the baseband frequency of the n th tone. For simplicity, we only consider herein the case of $f_n = n f_0$, where f_0 is the fundamental frequency. Since it is not recommended to keep one of the branches of the IQ modulator idle, we feed the same signal to both the in-phase (I) and quadrature (Q) branches of the USRP. They would eventually merge before amplification by a PA and produce two multisine spectra mirrored around the carrier frequency and $2N$ tones in total.

A co-phased N -tone multisine has a peak power of $(NA)^2$ while an average power of $NA^2/2$ resulting in PAPR of $2N$. This would also be the PAPR of the transmitted RF waveform if the USRP was an ideal linear device. However, in practice, transmission distorts the waveform in the following two ways:

- (i) The digital-to-analog converter (DAC) in the USRP has a fixed input range of ± 1 for the digital IQ samples. As a result, the analog baseband waveform generated from the digital baseband samples is of the form

$$\bar{x}(t) = \begin{cases} +1, & x(t) \geq 1, \\ -1, & x(t) \leq -1, \\ x(t), & \text{otherwise,} \end{cases} \quad (3)$$

where the bar signifies the clipping of the original waveform which would generate non-linear distortion. The PAPR of the clipped baseband waveform $\bar{x}(t)$ can be easily computed by piecewise integration and is omitted here due to space constraints.

- (ii) The PA employed internally by the USRP is non-linear at higher gain settings. Operating the PA at lower gains (linear region) would mean lesser efficiency and hence lower transmit power, while operating it at higher gains (non-linear region) would give higher transmit power but at the cost of distorted waveform. Also, even in the case of linear operation, high amplitude peaks can still drive the PA into saturation thereby creating clipping distortion. So, only making PAPR high by employing a multisine signal may not be useful in RF WPT.

To avoid clipping, one solution is to reduce the amplitude A of the multisine. However, doing so would eventually result in lower transmit power and so would not help the larger cause. Most research works referenced in [2] have considered that the average power of all the multisine waveforms is kept equal for a fair comparison. This does not help much in practice either since digital waveforms having higher PAPR but same average power would be clipped by the DAC and the resultant analog baseband waveforms would neither have the same average power nor the same PAPR (as it reduces due to clipping).

Both the above distortions would alter the shape (and hence the PAPR) of the modulated RF waveform due to which it may not remain ideal for RF WPT. Also distortions would inevitably lead to spectral regrowth and interference to adjacent channels. We shall shortly observe both the effects.

IV. EXPERIMENTAL SETUP AND RESULTS

In this section, we introduce the configuration of our RF power transfer test-bed in Table 1 followed by the explanation of the various experimental parameters and the results. The setup used in the experiments is shown in Fig. 2.

The waveforms employed for the experiments were co-phased multisines with different PAPRs. When $N \geq 2$, we generated 11 versions of each N -tone multisine by varying the amplitude uniformly from $1/N$ to one. The multisines generated with $A = 1/N$ would not get clipped by the DAC and so their PAPR is preserved. For the remaining 10

Table 1. Hardware configuration for the test-bed.

Component	Product details
Computer	Lenovo Thinkpad T470p, 32GB RAM, Core i7 processor, Linux OS, GNU C++ compiler
SDR	National Instruments USRP-2900
USB power meter	YZXStudio ZY1273
RF energy harvester	Powercast P2110B

Table 2. Operational parameters for the experiments.

Parameter	Details
Frequency band	902–928 MHz (ISM)
Carrier frequency	$f_c = 915$ MHz
Sampling rate	40 MHz
Baseband fundamental frequency	$f_0 = 1$ MHz
Number of tones in multisine	$N \in \{1, 2, 3, 4, 5, 6, 7, 8\}$
Amplitude	$A \in [1/N, 1]$ ($\frac{N-1}{10N}$ steps)
USRP gain setting	$G \in [60, 90]$ dB (0.5 dB steps)

versions, the peak amplitude would be greater than one which would lead to their peaks being clipped by the DAC and the consequent reduction in PAPR. In this way, we tested a total of 78 multisine waveforms, for which we varied the gain setting of the USRP from 60 dB to 90 dB. Each transmission lasted for around 10 seconds and we computed the average efficiency of each waveform from four such transmissions. The parameters for the experiments are summarized in Table 2.

We have employed a coaxial cable as the channel between the transmitter and the harvester, because the main scope of this paper is to compare different multisine waveforms. However, the cable can be easily replaced by antennas at both ends to evaluate real WPT, although an external PA would be required then. The measurement results are shown in Fig. 3.

As seen in Fig. 3(a), for $A = 1$, the number of co-phased tones (N) does not matter much as all the multisine signals yield almost similar efficiency. Figure 3(b) illustrates that for a given multisine signal (here $N = 4$), as we reduce A from one to $1/N$, the end-to-end efficiency goes on degrading with $A = 1/N$ giving the lowest efficiency. These observations lead us to the conclusion that adding more sinusoids does not enhance the end-to-end efficiency of RF power transfer.

In addition, we observe that increasing the PAPR may actually decrease the efficiency of RF WPT in contrast to common belief. This again implies that adding sinusoids to gain higher PAPR is not very productive. Especially, Fig. 3(c) shows the efficiency curves for different multisine signals when their PAPR is preserved at $2N$ by setting $A = 1/N$. It is clear from the plot that trying to preserve the shape of the waveform by scaling it down to avoid getting clipped by USRP's DAC takes a toll on the efficiency of the system. Even this observation leads to the conclusion that a single sinusoid waveform is better off than multisines for RF WPT.

Finally, the spectral effect of the two forms of distortion caused by the USRP is shown by Table 3. The aim was to verify whether the waveforms adhere to the FCC guidelines for spectral regrowth in the ISM band [10], which state that the power levels of the harmonics should be 20 dB lower than that

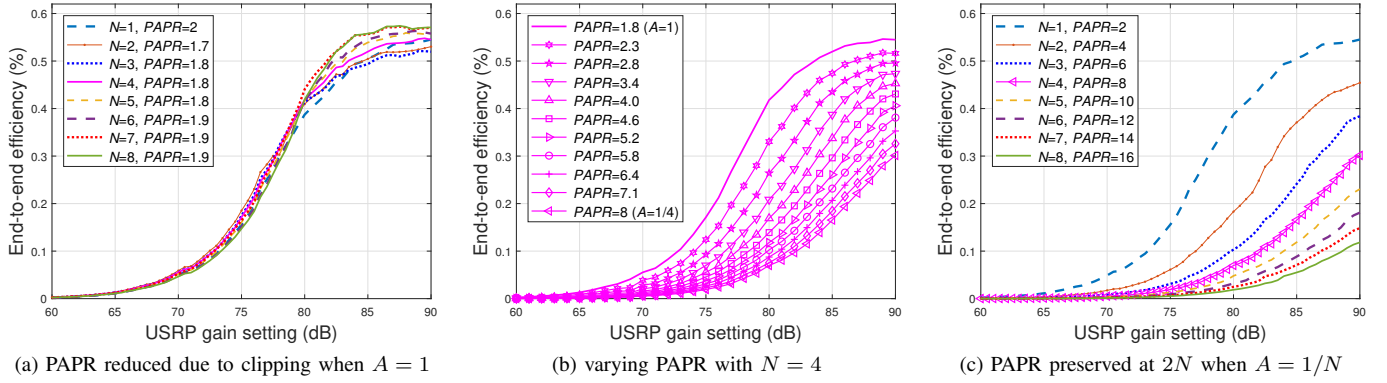


Fig. 3. Measured end-to-end efficiency of wireless power transfer in terms of the gain value in the USRP.

of the fundamental frequency. Thus, we assume the definition of bandwidth to comprise the frequencies below $f_c - f_N$ and above $f_c + f_N$ for which the power in the spectrum is at most 20 dB below the level at f_n , $n = 1, 2, \dots, N$, and the bandwidth was measured with coarse 1 MHz resolution.

For example, with the four-tone multisine for which PAPR is preserved at $2N = 8$, the transmitted RF signal bandwidth increases from 4 MHz at low gain to 7 MHz at 84 dB gain and 9 MHz at 90 dB gain. An interesting observation here is that, for $N \geq 4$ in the case of $A = 1$, the transmitted RF waveform interferes beyond the ISM band even for low 60 dB gain setting at the USRP (thus, PA operating in linear region). This implies that, in these cases, the clipping at the DAC of the USRP is very severe resulting in very high distortion. It can also be observed from the table that the RF signals corresponding to $N \leq 4$, for which PAPR is preserved at $2N$, leak into adjacent channels only in the saturation region of the PA and never leak outside the ISM band.

V. CONCLUSION

A test-bed system for RF power transfer was presented. We then analyzed the end-to-end efficiency of the system. The implications of the dynamic range of the USRP's DAC and the non-linearity of the USRP's PA on the performance of transmitted RF waveforms were observed. It was noticed that achieving higher PAPR with co-phased multisine waveforms is not very efficient. This is because the DAC clips high-PAPR signals, thereby reducing their PAPR and introducing distortion. Scaling down the input signal to the DAC does not help either, because then the transmitted RF power and hence the efficiency is low. Also high-PAPR signals tend to drive the PA into saturation resulting in more distortion to the signal shape. Both these distortions lead to spectral regrowth in the adjacent channels and out-of-band radiation in some cases. These observations hint towards single sinusoid waveform being an appropriate waveform for RF WPT.

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Table 3. Spectral regrowth due to the USRP transmitter's non-linearity.

N	PAPR	Measured bandwidth in terms of USRP gain setting		
1	2	2 MHz (≤ 77 dB)	6 MHz (84 dB)	6 MHz (90 dB)
2	4	4 MHz	6 MHz	10 MHz
4	8	8 MHz	14 MHz	18 MHz
2	1.7	10 MHz	12 MHz	16 MHz
7	14	14 MHz	24 MHz	28 MHz
8	16	16 MHz	28 MHz	28 MHz
3	1.8	20 MHz	20 MHz	28 MHz
4	1.8	28 MHz	28 MHz	28 MHz
7	1.9	28 MHz	28 MHz	28 MHz
8	1.9	28 MHz	28 MHz	28 MHz

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