Contents lists available at ScienceDirect

INTEGRATION, the VLSI journal

journal homepage: www.elsevier.com/locate/vlsi

Impact of losses and mismatches on power and efficiency of Wireless Power Transfer Systems with controlled secondary-side rectifier

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ARTICLE INFO

Keywords:
Wireless Power Transfer System
Semiconductor losses
Parameters mismatches
Modeling

ABSTRACT

In this paper an effective model for the power and efficiency analysis of Wireless Power Transfer Systems (WPTSs) is proposed. Such enhanced model includes modulation of duty-cycle and phase-shift for the secondary side controlled rectifier, as well as power losses of semiconductor devices and parameters mismatches of resonant elements. The global influence of semiconductor devices parameters and passive elements on the overall WPTS performances is numerically determined by solving the non-linear equations of the system discussed in this paper. Simulations and experimental measurements referred to a 2 W@6.78 MHz WPTS fully demonstrate the validity of the proposed analysis.

1. Introduction

The research in the area of Wireless Power Transfer Systems (WPTSs) design has experienced a big expansion in recent years [1]. Indeed WPTSs and their relevant technologies can be successfully applied to wirelessly power small wearable or portable electronics [2,3], to contactless charging systems for electric vehicles [4–6], to factory automation systems [7,8], to medical and health care devices [9,10]. Many papers, standards, books and reports about WPTSs design have been published in the technical literature, each one with a special emphasis on well-defined applications and relevant issues and constraints, like different topologies and compensation networks [11–13], resonance frequency and soft-switching achievement [14,15], system efficiency *vs* load output power [16]. Recent trends in WPTS and its applications are given in [17] and an interesting comparison of the state-of-the-art in WPTS can be found in [18], with a complete efficiency analysis for a comparison among competing solutions. However many scientific works on this subject have only concentrated on the optimization of the transmitter-receiver modules [16,18]. As a consequence, several commercial WPTSs or prototypes are often characterized with reference efficiency values given as the ratio between the load power and the total power entering the transmitting loop. Under these conditions, the efficiency is typically obtained through experimental measurements, without a preliminary understanding of the silicon devices impact on the overall system performances. Additionally, previous studies on the impact of controlled rectifiers and/or post-regulators in WPTSs often adopt simplified loss models, representing the secondary side rectifier and its load as a linear resistors and neglecting the switching losses in the semiconductor devices of the inverter and rectifier stages [19]. Such simplifications yield too rough efficiency predictions. A correct efficiency analysis is needed in low-power high-frequency WPTSs design like wearable and portable applications, where the achievement of high efficiency and optimal energy management becomes more and more important. In such applications, indeed, the value of the coupling factor K between the primary transmitting coil and the secondary receiving coil is much lower than unity and it is affected by a large tolerance due to the high level of freedom required by new technology standards like Alliance for Wireless Power (A4WP) Rezence [20,21].

The goal of this paper is to discuss modeling issues about the analysis and the efficiency evaluation of WPTSs using controlled rectifiers in the secondary side. The model presented in the paper includes the switching losses of semiconductor devices, the mismatches of the resonant elements and of the coupling factor, for any given value of duty-cycle and phase-shift of the secondary side rectifier MOSFETs drives modulation [19]. Although the loss modeling issues discussed in this paper are focused on WPTSs for low-power applications relevant to wearable devices, the concepts, models and methods discussed herein are general. Indeed, the same loss investigation can be used for reliable design and optimization of other WPTSs, and extended to other applications like charging systems for electric vehicles [22]. Section 2 provides an overview of the state-of-the-art on WPTSs, with special emphasis to wearable and automotive applications. Section 3 presents a general non-linear analytical model

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http://dx.doi.org/10.1016/j.vlsi.2016.04.005
0167-9260/

for the calculation of the first harmonic solution in WPTS and illustrates a numerical method for the solution of the model. In Section 4, the results of the analysis of a 2 W@6.78 MHz WPTS are presented and validated with experimental measurements, and the impact of MOSFETs losses and mismatches for resonant elements and coupling factor is discussed.

2. State of the art for wearable and automotive applications

WPTSs have become an increasingly attractive technology in Electric Vehicles (EVs) because it is convenient, clean, and safe. In particular, the rise of EVs deployment has led researchers to investigate several aspects of EVs and charging technologies, including advanced battery technologies and on-board charging systems. The WPTS may mitigate the heavy dependence of an EV on battery, that can be heavy, expensive and with possible short operating time. A comprehensive review of existing technological solutions for WPTS used in EVs battery chargers is given in [4,5]. At present, many different prototypes for EVs charging systems have been designed to operate at a quite low frequency (e.g. at 20 kHz [18]). After the announcement of the SAE International J2954™ Task Force for WPT of Light Duty, Electric and Plug-in EVs, the operating frequency under study for EVs applications has been given at 85 kHz and the maximum input WPT power rating at 3.7 kW [23]. Therefore, because of the SAE International J2954™ resolutions, there are some challenges to overcome. Focusing on silicon devices, the high-power demands of the EVs and the constraint of operating at 85 kHz have to deal with the lack of semiconductor switches that can operate efficiently at these power levels and high frequencies [8,24]. Insulated Gate Bipolar Transistors (IGBTs) can handle the high power rating only at lower frequencies, whereas the latest Silicon Carbide (SiC) devices can handle the higher frequency only at reduced power ratings. As a consequence, it is important to understand what the impact of primary side inverters and secondary side rectifiers on WPTSs efficiency is. Moreover, the selection of passive components and semiconductor devices in WPTSs for electric vehicle chargers must be realized accounting for the combined effects of tolerances and nonidealities, in order to obtain a high efficiency operation. The models discussed in [22] allow the analysis of all losses in such kind of application and can support the identification of optimal components.

WPTSs for wearable and portable applications are gaining popularity in many commodity products, such as mobile phones or smart-watches chargers. The achievement of high efficiency in wearable and portable applications has become more and more challenging. Indeed, in these applications the typical coupling factor between primary transmitting (TX) coil and secondary receiving (RX) coil can be lower than 0.1. Moreover, the large number of battery-operated consumer electronics has generated an increasing interest in designing charging platforms for multiple devices [3]. Therefore, effective models for the analysis, design and optimization of such kind of WPTS have become indispensable for designers. Recently, for such kind of applications there has been a push for operation in the restricted and unlicensed lower Industrial Scientific Medical (ISM) band at 6.78 MHz ($\pm$ 15 kHz) [25], where traditional MOSFET technology is approaching its capability limit. Enhancement mode Gallium Nitride (GaN) transistors offer an alternative to MOSFETs, as they can switch fast enough to be ideal for wireless power applications [24,26]. High-frequency magnetic WPTSs have some clear advantages over lower-frequency technologies. The two main advantages are reduced form factors and reduced losses in metal objects unintentionally placed in proximity of the magnetic field. If the frequency of operation is 6.78 MHz, TX and RX coils with quality factors between 100 and 300 can be easily manufactured with low-cost single layer printed circuit boards. At this frequency, losses in foreign metal objects can be three to ten times lower than they would be in systems operating in the kHz range. However, high-frequency WPTSs are naturally subject to higher losses, like switching and proximity losses, and higher dependence on parasitic components and on uncertainties of parameters characterizing the resonant elements of transmitting and receiving loops. Typically, semiconductor devices of a controlled bridge rectifier are turned on when their current has just crossed the zero, so that their switching losses are really minimized. Though, recent studies have been published where the fire angles of the controlled rectifier switches are used to modulate the rectifier duty-cycle and phase [19,27], in order to improve efficiency and power factor, as well as for the possibility to develop new contactless battery chargers with bidirectional energy transfer.

Table 1
Main parameters for WPTS and MOSFETs.

Symbols	WPTS main parameters			MOSFETs main parameters		
TX=transmitter	f_s	[Hz]	Switching frequency	R_{dson}	[Ω]	Channel resistance
RX=receiver	$\omega_s=2\pi f_s$	[rad/s]	Angular frequency	R_g	[Ω]	Gate resistance
$I_{1=}$1st harmonic of TX current	V_{in}	[V]	Input voltage	$R_{gd,on}$	[Ω]	Turn-on gate driver resistance
	V_{bat}	[V]	Battery voltage	$R_{gd,off}$	[Ω]	Turn-off gate driver resistance
$V_{1=}$1st harmonic of TX voltage	R_{bat}	[Ω]	Battery series resistance	R_{sns}	[Ω]	Series sensing resistance
	$V_{out=}V_{bat+}R_{bat}I_{out}$	[V]	Output voltage	R_{ext}	[Ω]	External series gate resistance
$I_{2=}$1st harmonic of RX current	L_1	[H]	TX coil inductance	g_{fs}	[S]	Transconductance
	L_2	[H]	RX coil inductance	C_{oss}	[F]	Output capacitance
$V_{2=}$1st harmonic of RX voltage	Q_{L1}	–	TX coil quality factor at ω_s	Q_g	[C]	Total gate charge
	Q_{L2}	–	RX coil quality factor at ω_s	Q_{gs}	[C]	Gate–source charge
Z_1=TX coil impedance	$R_{L_1}=\omega_s L_1 Q_{L_1}^{-1}$	[Ω]	TX coil resistance	Q_{gd}	[C]	Gate–drain charge
Z_2=RX coil impedance	$R_{L_2}=\omega_s L_2 Q_{L_2}^{-1}$	[Ω]	RX coil resistance	Q_{gsw}	[C]	Switching gate charge
α (see Figs. 2 and 3)	K_{12}	–	TX-RX coupling coefficient	Q_{rr}	[C]	Body diode recovery charge
β (see Figs. 2 and 3)	$M=K_{12}\sqrt{L_1L_2}$	[H]	TX-RX mutual inductance	t_{dt}	[s]	Gate signal dead-time
ϕ (see Figs. 2 and 3)	$C_{s1}=(\omega_s^2L_1)^{-1}$	[F]	TX resonant capacitance	V_{th}	[V]	Gate–source threshold voltage
$s_\alpha=\sin(\alpha)$	$C_{s2}=(\omega_s^2L_2)^{-1}$	[F]	RX resonant capacitance	V_{sd}	[V]	Body diode forward voltage
$c_\alpha=\cos(\alpha)$	Q_{C1}	–	TX capacitor quality factor at ω_s	Q_{rr}	[C]	Body diode recovery charge
$s_\beta=\sin(\beta)$	Q_{C2}	–	RX capacitor quality factor at ω_s	V_{dr}	[V]	Driver voltage
$c_\beta=\cos(\beta)$	$R_{C_1}=(\omega_s C_{s1}\ Q_{C1})^{-1}$	[Ω]	Series resistance of TX capacitor	$R_{gon=}R_{gd,on}+R_g+R_{ext}$		
$s_\phi=\lvert\sin(\phi_{V1}-\phi_{I1})\rvert$	$R_{C_2}=(\omega_s C_{s2}\ Q_{C2})^{-1}$	[Ω]	Series resistance of RX capacitor	$R_{goff=}R_{gd,off}+R_g+R_{ext}$		