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Power and efficiency analysis of high-frequency Wireless Power Transfer Systems

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ABSTRACT

This paper discusses the losses analysis of low power high-frequency Wireless Power Transfer Systems (WPTSs). Ideal models for efficiency evaluation of WPTS can predict performances that are quite far from the real ones. The model proposed in this paper includes semiconductor devices losses, as well as modulation of duty-cycle and phase for the secondary side rectifier. The global influence of semiconductor devices and control parameters on the overall WPTS performances is numerically determined by solving the herein discussed non linear equations system. Experimental measurements realized on a 2 W@6.78 MHz WPTS demonstrate the validity of this analysis.

Introduction

In the last years, the research on Wireless Power Transfer Systems (WPTSs) experienced a big expansion [1]. WPTSs can be well 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, each one with a special emphasis on specific applications, different topologies and compensation networks [11–13], resonance frequency and soft-switching [14], system efficiency and power [15,16]. In this regard, WPTSs are often characterized with respect to reference efficiency values given as the ratio between the load power and the total power entering the transmitter loop. Many papers on this subject merely focus on the optimization of the transmitter-receiving modules and experimental efficiency values of WPTSs are only provided [16]. However, comparative discussions on global WPTS optimization, involving silicon devices and control issues impact, are missing. This problem is particular important in WPTSs design for wearable and portable applications, such as mobile phones or smart-watches chargers, where the achievement of high efficiency becomes more and more challenging. In such kind of applications, critical issues are:

– coupling factor between primary transmitting (Tx) coil and secondary receiving (Rx) coil, which is typically much lower than unity in charger application [17];
– load current, which is varying over the battery charge cycle [18];
– restricted and unlicensed lower Industrial Scientific Medical (ISM) band at 6.78 MHz [19];
– fire angles of the controlled rectifier switches used to modulate the rectifier duty-cycle and phase [20–23], in order to improve the WPTS efficiency.

Papers discussing the efficiency benefits of controlled rectifiers and/or post-regulators typically adopt simplified ideal models to characterize WPTSs. A first simplification consists in representing the secondary side rectifier and load simply as a linear resistor. A second simplification lies in neglecting switching losses in the semiconductor devices both of the primary side inverter and of the secondary side controlled rectifier and post-regulators. Such assumptions may yield too optimistic efficiency predictions, especially in high-frequency WPTSs. In this scenario, enhanced models and methods to predict the impact of semiconductor devices switching losses on the WPTS performance are required. Indeed,

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because of the ISM band restrictions, traditional MOSFET technology is approaching its capability limit and it is necessary to evaluate the opportunity to switch to other technologies, like enhancement mode Gallium Nitride (GaN) transistors [24,25]. The open problem is then to analyze the overall influence of semiconductor devices losses on performances of a WPTS. The problem becomes even more involved if the mismatch of the transmitter-to-receiver coupling factor is included, as this one strongly characterizes several applications of large interest beyond portable/wearable devices, such as electric vehicles.

Therefore, the main goal of this paper is to provide an effective model for the power and efficiency analysis of WPTSs. The investigation is focused on low power applications relevant to wearable devices. Nevertheless, concepts, models and methods herein discussed are quite general. Indeed, the same loss investigation can be used for reliable design and optimization of other WPTSs, and extended to other applications like the automotive ones. Section 'WPTS modeling' presents the general analytical model for the calculation of the first harmonic component solution in a WPTS and illustrates a numerical method for the solution of the proposed analytical model. In section 'Loss analysis and experimental verification', the results of the analysis of a 2 W@6.78 MHz WPTS are presented, and then validated with experimental measurements. Finally, section 'MOSFETs impact on WPTS output power and efficiency' discusses the impact of semiconductor devices on WPTS output power and efficiency.

WPTS modeling

All the parameters and the quantities adopted in the proposed modeling are summarized in Table 1. All superscript "*rec*" and "*inv*" are referred to rectifier and inverter elements, respectively.

The WPTS considered in this paper is based on the series compensation architecture shown in Fig. 1. The WPTS is supposed to work with a primary constant frequency and a secondary PWM control ensuring duty-cycle D and phase-shift modulation $\phi = \phi_{I2} - \phi_{V2}$ [26]. In these conditions, the phase modulation involves a phase lag (or a phase lead) with respect to the current zero crossing in the commutations of the rectifier switches. Figs. 2 and 3 show typical current and voltage waveforms for the primary and secondary side of a WPTS.

Such plots exemplify different commutation conditions for the inverter MOSFETs $M_1, \ldots, M_4$ and the rectifier MOSFETs $Q_1, \ldots, Q_4$, given different phase lag/lead ($\phi > 0$ or $\phi < 0$). In this regard, the key issues for the loss analysis of a WPTS are:

– identifying the type of commutations the MOSFETs undergo, depending on α, β and ϕ;
– formulate a specific set of loss formulae for all the possible combinations of commutations, depending on α, β and ϕ.

It is worth to remember that a MOSFET undergoes to a *hard commutation* whenever it changes its state while its drain-to-

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 harm. 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 harm. 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 harm. of RX current	L_1	[H]	TX coil inductances	g_{fs}	[S]	Transconductance
	L_2	[H]	RX coil inductances	C_{oss}	[F]	Output capacitance
V_2 = 1st harm. 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)	$M = K_{12}\sqrt{L_1 L_2}$	[H]	TX-RX mutual inductance	t_{dt}	[s]	Gate signal dead-time
ϕ (see Figs. 2 and 3)	$C_{s1} = (\omega_s^2 L_1)^{-1}$	[F]	TX resonant capacitor	V_{th}	[V]	Gate-source threshold voltage
$s_\alpha = \sin(\alpha)$	$C_{s2} = (\omega_s^2 L_2)^{-1}$	[F]	RX resonant capacitor	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	$R_{goff} = R_{gd,off} + R_g + R_{ext}$		

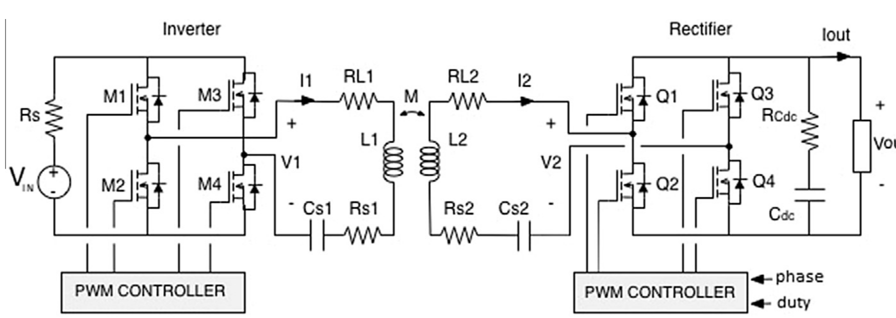

Fig. 1. Series–series resonant WPTS.