

Survey Paper of Performance Analysis of Class EF2 Inverter and Rectifier for Wireless Power Transfer Systems

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ABSTRACT: Recent technologies such as wireless power transfer (WPT) and small-size radio-frequency plasma sources demand efficient, powerful, and multi megahertz switching resonant inverters. The EF2 inverters and rectifiers are designed to have high efficiency reduced second harmonic component and lower total harmonic distortion (THD), higher power-output capability, reduction in magnetic core requirements and operation at higher frequencies in rectification compared to other circuit topologies. This paper presents the overview of modeling and analysis of Class EF2 Inverter and Rectifier for Multi-megahertz for two 25W wireless power transfer (WPT) systems, one operating at 6.78MHz and the other at 27.12MHz.

KEYWORDS: Wireless Power Transfer, Class EF inverters, Class EF rectifiers

I. INTRODUCTION

The Wireless power transfer (WPT), wireless energy transmission, or electromagnetic power transfer is the transmission of electrical energy from a power source to an electrical load, such as an electrical power grid or a consuming device, without the use of discrete human-made conductors. Wireless power is a generic term that refers to a number of different power transmission technologies that use time-varying electric, magnetic, or electromagnetic fields. In wireless power transfer, a wireless transmitter connected to a power source conveys the field energy across an intervening space to one or more receivers, where it is converted back to an electrical current and then used. Wireless transmission is useful to power electrical devices in cases where interconnecting wires are inconvenient, hazardous, or are not possible.

The increasing demands have imposed increased challenges in the design and construction of dc/ac inverters and rectifiers. At the circuit level, the switching frequency and output power of an inverter is limited by the currently available devices for a desired efficiency. For instance, the input gate charge, output capacitance, and on-state resistance of a certain MOSFET will have a significant effect of the performance of an inverter in terms of its efficiency and output power capability. One of the key interests in WPT technology is to increase the frequency of operation from the kilohertz range to the megahertz range to improve the tolerance to misalignment by allowing ferrite cores to be removed. A higher frequency also increases the reflected resistance seen by the primary coil driver, therefore, power can be transferred at reduced current stresses. Similar to switched-mode power supplies, the additional benefits of operating at multimegahertz switching frequencies such as the ISM bands 6.78, 13.56, and 27.12 MHz are reduction in the values of the passive components, increased power density, improved performance and transient response, and lower coupling with nearby objects and devices.

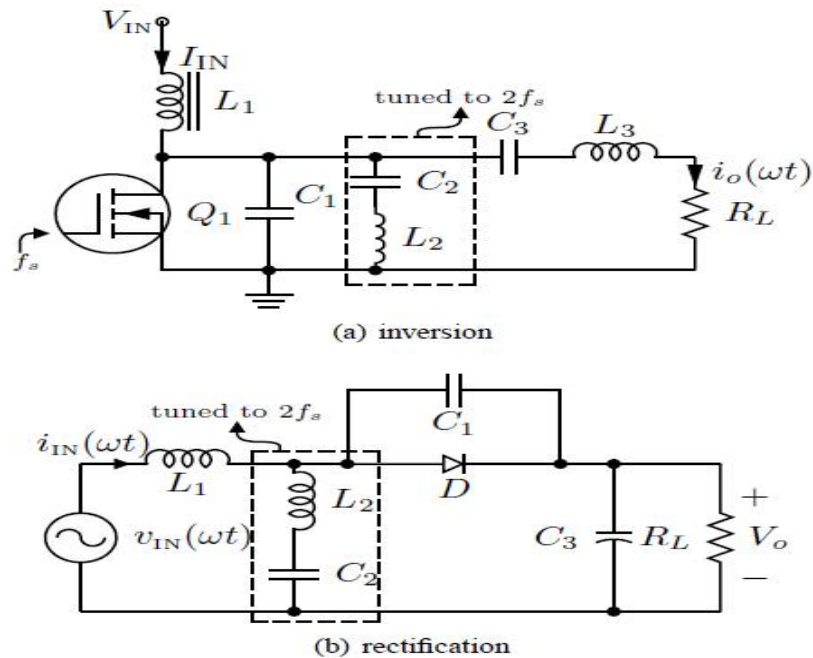


Figure.1. Circuit diagrams of the Class-EF2 inverter and rectifier

Class E network by connecting a series LC network, tuned to the second harmonic of the switching frequency, in parallel with the switch, referred to as class EF2. network. With this circuit can be designed at a higher operating dc voltage, which reduces the variation in the switch's nonlinear output capacitance. Also Class-EF2 topology is that the switch's drain voltage and the output load current does not contain a second-harmonic component, and therefore, has an improved EMI performance, whereas the Class-E topology contains a strong second-harmonic component, which may make it more difficult to meet EMI regulations. The Fig. 1 above shows the circuit diagrams of the Class-EF2 inverter and rectifier. The inductor can contribute to the majority of the losses in the circuit, especially when operating at frequencies in the megahertz range due to the losses associated with magnetic cores. In addition, the circulating dc current in the rectifier can cause the magnetic core to saturate, and therefore, the power that it can handle is limited.

II.LITERATURE SURVEY

The work in, for example, proposed a dual-frequency wireless power / wireless data system for implantable electronics. The two Tx coils, one for power and another for data, are placed orthogonal to each other in the transmitter in order to prevent interference between the two coils. Although this was effective for power and data transmission to one receiver, this cannot be applied to a multi-receiver use-case because the receiver would also need to be equipped with orthogonal coils, which is not feasible in most portable devices that have thin, planar geometries. Magnetic resonance coupling working at megahertz (MHz) is widely considered as a promising technology for the mid-range transfer of a medium amount of power. It is known that the soft-switching-based Class E rectifiers are suitable for high-frequency rectification, and thus potentially improve the overall efficiency of MHz wireless power transfer (WPT) systems. This paper reports new results on optimized parameter design of a MHz WPT system based on the analytical derivation of a Class E current-driven rectifier. 1.3 kW resonant power amplifier using a Gallium Nitride (GaN) device[3] at 13.56 MHz for wireless power transfer (WPT). The power amplifier driving the power transmitting coils is based on a Class Φ_2 inverter, a single switch topology with low switch voltage stress and fast transient response. This implementation utilizes a recently available GaN device in a low inductance package that is compatible with operation in the 10's of

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MHz switching frequency. These power GaN switching devices have low gate resistance R_G and low capacitance C_{GS} which greatly reduces the power requirements of the gate drive circuitry. The operation of traditional rectifiers such as half-wave and bridge rectifiers in wireless power transfer applications may be inefficient and can reduce the amount of power that is delivered to a load. An alternative is to use Class E resonant rectifiers that are known to operate efficiently at high resonant frequencies and at large input voltages. Class E rectifiers have a near sinusoidal input current which leads to an improved overall system performance and increased efficiency especially that of the transmitting coil driver. The concept of combining Class E and Class F or Class F-1 inverters was introduced by Kee in 2002 [12]. Since 2002, the research and contribution has been limited with fewer than ten journal papers published regarding Class EF and Class E/F inverters.

III.METHODOLOGY

1. Modeling and Analysis

As shown in Fig.1 for both the EF2 inverter and rectification circuits, the resonant network of L2C2 tuned to resonate at the second harmonic of the circuits' switching frequency. Both circuits shows dual nature, Both circuits are the dual of each other, that is the inverter circuit takes an input dc voltage (V_{IN}) and outputs ac current (I_o), and the rectifier circuit takes an input ac current (I_{IN}) and rectifies it into an output dc voltage (V_o).The parameters annotations for voltage and current are described in Table 1 below. The analysis of equivalent circuit is shown in Fig. 2 can be based on following assumptions:

- a) For both inversion and rectification the diode form an ideal switch for which ON resistance is zero, OFF resistance is infinity. The switching time is zero in inversion case whereas for rectification the forward voltage drop and reverse recovery time should be zero.
- b) The inductance value L_1 should be high such that the it keeps the input current I_{IN} to dc for inversion and the input current i_y to be sinusoidal with frequency same as switching frequency.

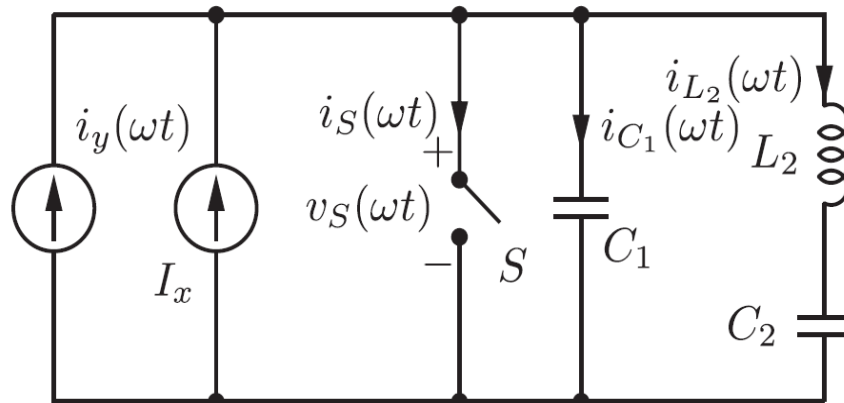


Figure 2. Class-EF2 equivalent circuit for inversion and rectification.

- c) The loaded quality factor of the L3C3RL should be high to output current i_y to be sinusoidal for inversion. For output voltage and current and voltage to be dc, the filter capacitor C_3 should be high for rectification.

- The shunt capacitance C_1 is assumed to be constant and independent of its voltage for both inversion and rectification.
- All the power should be delivered to load R_L without any losses.

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	Inversion	Rectification
S	Mosfet	Diode
V_x	Input DC voltage (V_{IN})	Output DC voltage (V_o)
I_x	Input DC current (I_{IN})	Output DC current (I_o)
i_y	Output AC current (i_o)	Input AC current (i_{IN})
v_s	Drain voltage (v_{DS})	Anode voltage (v_D)
i_s	Drain current (i_{DS})	Anode current (i_D)

Table I Description Of The Voltage And Current Annotations

The current i_y is sinusoidal and is given by

$$i_y(\omega t) = I_m \sin(\omega t + \phi) \quad (1)$$

By applying the KCL at the switch's drain node, the current in the switch is

$$i_S(\omega t) = I_x - i_{L_2}(\omega t) - i_y(\omega t), \text{ for } 0 \leq \omega t < 2\pi D. \quad (2)$$

The L_2C_2 network now is a source-free undamped circuit and its current (i_{L_2}) normalized with to the dc current (I_x) is given by

$$\frac{i_{L_2}}{I_x}(\omega t) = A_1 \cos(2\omega t) + B_1 \sin(2\omega t). \quad (3)$$

By applying the KCL at the drain node, the current in the series tuned L_2C_2 network is,

$$\begin{aligned} i_{L_2}(\omega t) &= I_x - i_y(\omega t) - i_{C_1}(\omega t) \\ &= I_x - I_m \sin(\omega t + \phi) - \omega C_1 \frac{dv_s(\omega t)}{d\omega t}. \end{aligned} \quad (4)$$

The switch's voltage is equal to the total voltage across the L_2C_2 network and is given by

$$\begin{aligned} v_S(\omega t) &= \omega L_2 \frac{di_{L_2}(\omega t)}{d\omega t} \\ &+ \frac{1}{\omega C_2} \int_{2\pi D}^{\omega t} i_{L_2}(\omega t) d\omega t + v_{C_2}(2\pi D). \end{aligned} \quad (5)$$

Differentiating the aforementioned equation gives,

$$\frac{dv_s(\omega t)}{d\omega t} = \omega L_2 \frac{d^2 i_{L_2}(\omega t)}{d\omega t^2} + \frac{1}{\omega C_2} i_{L_2}(\omega t). \quad (6)$$

Substituting (6) into (4) and normalizing with respect to the current I_x gives,

$$\begin{aligned} \frac{i_{L_2}}{I_x}(\omega t) &= 1 - \frac{I_m}{I_x} \sin(\omega t + \phi) \\ &- \omega^2 L_2 C_1 \frac{d^2 i_{L_2}}{I_x}(\omega t) - \frac{C_1}{C_2} \frac{i_{L_2}}{I_x}(\omega t). \end{aligned} \quad (7)$$

The aforementioned equation is a linear nonhomogeneous second-order differential equation, which has the following general solution:

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$$\frac{i_{L_2}}{I_x}(\omega t) = A_2 \cos(q_2 \omega t) + B_2 \sin(q_2 \omega t) - \frac{q_2^2 p}{q_2^2 - 1} \sin(\omega t + \phi) + \frac{1}{k + 1} \quad (8)$$

where

$$k = \frac{C_1}{C_2} \quad (9)$$

$$q_2 = \frac{1}{\omega} \sqrt{\frac{C_1 + C_2}{L_2 C_1 C_2}} = 2 \sqrt{\frac{k + 1}{k}} \quad (10)$$

$$p = \frac{C_2}{C_1 + C_2} \frac{I_m}{I_x} = \frac{1}{k + 1} \frac{I_m}{I_x} \quad (11)$$

and the coefficients A_2 and B_2 are to be determined based on the equation's boundary conditions. The boundary conditions are determined current and voltage continuity conditions when the switch turns ON and OFF, which can be described by,

$$i_{L_2}(2\pi D^-) = i_{L_2}(2\pi D^+) \quad (12)$$

$$i_{L_2}(0) = i_{L_2}(2\pi) \quad (13)$$

$$\left. \frac{di_{L_2}(\omega t)}{d\omega t} \right|_{\omega t=2\pi D^-} = \left. \frac{di_{L_2}(\omega t)}{d\omega t} \right|_{\omega t=2\pi D^+} \quad (14)$$

$$\left. \frac{di_{L_2}(\omega t)}{d\omega t} \right|_{\omega t=0} = \left. \frac{di_{L_2}(\omega t)}{d\omega t} \right|_{\omega t=2\pi} \quad (15)$$

Using (9), the normalized current through capacitor C_1 for the period $2\pi D \leq \omega t < 2\pi$ is,

$$\frac{i_{C_1}}{I_x}(\omega t) = 1 - p(k + 1) \sin(\omega t + \phi) - \frac{i_{L_2}}{I_x}(\omega t). \quad (16)$$

The drain voltage for the period $2\pi D \leq \omega t < 2\pi$ is,

$$v_S(\omega t) = \frac{I_x}{\omega C_1} \int_{2\pi D}^{\omega t} \frac{i_{C_1}}{I_x}(\tau) d\tau \quad (17)$$

where τ is a dummy variable. The drain voltage can also be written as,

$$v_S(\omega t) = \frac{I_x}{\omega C_1} \beta(\omega t) \quad (18)$$

Where

$$\beta(\omega t) = \int_{2\pi D}^{\omega t} \frac{i_{C_1}}{I_x}(\tau) d\tau. \quad (19)$$

The dc component (or average) of the drain voltage is equal to the dc voltage V_x , i.e.,

$$V_x = \frac{I_x}{2\pi\omega C_1} \int_{2\pi D}^{2\pi} \beta(\omega t) d\omega t. \quad (20)$$

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By substituting (20) into (18), the normalized drain voltage with respect to the voltage V_x can be written as,

$$\frac{v_S}{V_x}(\omega t) = \begin{cases} 0, & \text{for } 0 \leq \omega t < 2\pi D \\ \frac{2\pi\beta(\omega t)}{\int_{2\pi D}^{2\pi} \beta(\omega t) d\omega t}, & \text{for } 2\pi D \leq \omega t < 2\pi. \end{cases} \quad (21)$$

Using (11), the normalized switch current with respect to the dc current I_x can be written as,

$$\frac{i_S}{I_x}(\omega t) = \begin{cases} 1 - p(k+1)\sin(\omega t + \phi) - (A_1 \cos(2\omega t) + |B_1 \sin(2\omega t)|), & \text{for } 0 \leq \omega t < 2\pi D \\ 0, & \text{for } 2\pi D \leq \omega t < 2\pi. \end{cases} \quad (22)$$

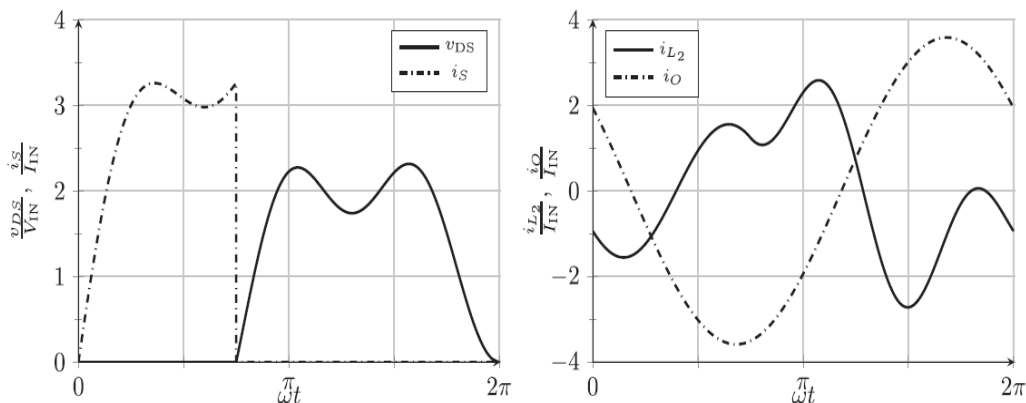


Figure. 3. Voltage and current waveforms for the Class-EF2 inverter (k = 0.867, D = 0.375).

2. Class EF2 Inversion

A. ANALYSIS

The analysis of EF2 is performed in cases of maximum power and frequency based on duty cycle and K. The maximum power operation can be described as

Voltage and Current Waveforms

Fig. 3 shows the voltage and current waveforms for the class EF2 inverter. the peak voltage and current stresses of the Class EF2 when k = 0.867 and D = 0.375 are

$$\frac{V_{DS \max}}{V_o} = 2.3162 \quad (23)$$

$$\frac{I_{DS \max}}{I_o} = 3.2632. \quad (24)$$

1) Input Resistance and Output Power:

The normalized value of I_m with respect to the input dc current,

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$$\frac{I_m}{I_x} = \frac{I_m}{I_{IN}} = \frac{2\pi(1 - D) + \frac{A_1}{2} \sin(4\pi D) + \frac{2B_1}{2} \sin^2(2\pi D)}{\cos(2\pi D + \phi) - \cos \phi} \quad (25)$$

and the input dc resistance seen by the source

$$\frac{R_{DC}}{R_L} = \frac{1}{2} \left(\frac{2\pi(1 - D) + \frac{A_1}{2} \sin(4\pi D) + \frac{2B_1}{2} \sin^2(2\pi D)}{\cos(2\pi D + \phi) - \cos \phi} \right)^2 \quad (26)$$

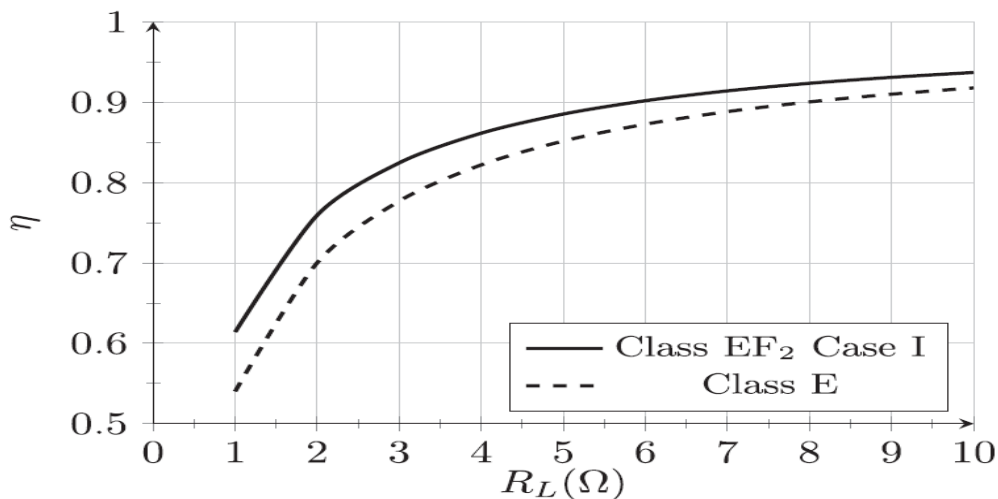


Figure. 4. Comparison between the efficiencies of the Class-EF2 inverter operating at maximum cp [8] and the Class-E inverter.

where R_L here represents the reflected load of the inductive link and the rectifier. Assuming all the power supplied by the source will be consumed in the reflected load, thus

$$P_{IN} = P_o$$

$$V_{IN} I_{IN} = \frac{1}{2} I_m^2 R_L. \quad (27)$$

The normalized output power is, therefore

$$\frac{P_o R_L}{V_{IN}^2} = \frac{1}{2} \left(\frac{I_m}{I_{IN}} \frac{R_L}{R_{DC}} \right)^2 = 0.1556. \quad (28)$$

B. COMPARISON WITH CLASS-E INVERTER

Efficiency: The performance analysis of Class EF2 inverter is higher in all the cases of maximum power operation, maximum output capability. Fig. 4 compares the drain efficiency between the Class-E inverter and the Class-EF2 inverter and it can be noticed that the ClassEF2 inverter has a higher overall efficiency.

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Harmonics: The Fourier analysis is performed for the harmonic investigation of drain waveforms. Using (21), the Fourier components of the drain voltage are

$$A_n = \frac{1}{\pi} \int_{2\pi D}^{2\pi} \frac{v_{DS}}{V_{IN}}(\omega t) \sin(n\omega t + \phi) d\omega t \quad (29)$$

$$B_n = \frac{1}{\pi} \int_{2\pi D}^{2\pi} \frac{v_{DS}}{V_{IN}}(\omega t) \cos(n\omega t + \phi) d\omega t \quad (30)$$

$$C_n = \sqrt{a_n^2 + b_n^2} \quad (31)$$

The first five components are used to calculate the THD as follows as there is so little power in the higher harmonics

$$THD = \frac{\sqrt{\sum_{n=2}^6 C_n^2}}{C_1} \quad (32)$$

The comparisons of harmonic contents of Class EF2 and E inverter are shown in Fig. 5 above. It reveals that the Class-E inverter has a strong second-harmonic component as compared to class EF2 inverter therefore has a higher THD.

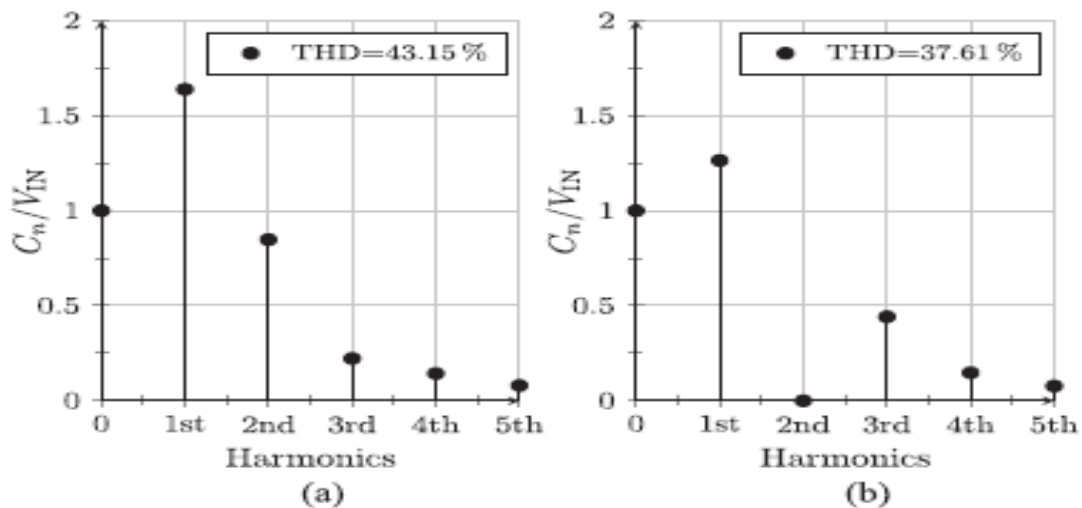


Figure. 5. Harmonic content of the switch's voltage.

3. Class-Ef2 Rectification

A. Analysis

For the analysis of Class EF2 rectification the diode voltage at the switching end is zero. It gives the following equations.

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$$\begin{aligned}
 &2\pi \frac{k}{k+1} (1-D) + p(\cos(2\pi D + \phi) - \cos \phi) \left(\frac{q_2^2}{q_2^2 - 1} - (k+1) \right) \\
 &+ \frac{A_2}{q_2} (\sin(2\pi D q_2) - \sin(2\pi q_2)) \\
 &+ \frac{B_2}{q_2} (\cos(2\pi q_2) - \cos(2\pi D q_2)) = 0. \quad (33)
 \end{aligned}$$

The ZDS in rectification occurs at $\omega t = 2\pi D$ ($iC1(2\pi D) = 0$), giving the following equation:

$$\begin{aligned}
 &1 - A_2 \cos(2\pi D q_2) - B_2 \sin(2\pi D q_2) - \frac{1}{k+1} \\
 &- p \sin(2\pi D + \phi)(k+1) + \frac{q_2^2 p}{q_2^2 - 1} \sin(2\pi D + \phi). \quad (34)
 \end{aligned}$$

For the rectification circuit the output load current I_o and k can be set as the independent parameters for diode operation and duty cycle need to be calculated. From (11), the ratio of the amplitude of input ac current I_m to the output dc current I_o can be written as

$$\frac{I_m}{I_o} = p(k+1). \quad (35)$$

This ratio depends on load resistance. The duty cycle D as a function of normalized load current and k .

Voltage and Current Waveforms: By using the numerical solutions for A_1 , A_2 , B_1 , B_2 , p , and ϕ , the voltage and current waveforms throughout the rectifier can be plotted for given values of I_m , I_o and k . Fig. 7 shows the diode's voltage and current waveforms and the current of inductor L_2 for selected values of normalized load current and k .

Maximum Stresses and Power-Output Capability: Similar to the inversion case, the maximum cp is 0.13231, which occurs when $k = 0.867$ and $I_m I_o = 3.5853$ corresponding to the following voltage and current stresses:

$$\frac{V_{D \max}}{V_o} = 2.3162 \quad (36)$$

$$\frac{I_{D \max}}{I_o} = 3.2632. \quad (37)$$

Input Impedance: The equivalent circuit for calculating the input impedance is shown in Fig. 6 above. The fundamental Fourier component of the diode voltage is given by

$$\begin{aligned}
 \frac{v_{D1}}{V_o}(\omega t) &= V_{RAC} \sin(\omega t + \phi) + V_{Cx} \cos(\omega t + \phi) \\
 &= I_m R_{AC} \sin(\omega t + \phi) + \frac{I_m}{\omega C_x} \cos(\omega t + \phi) \quad (38)
 \end{aligned}$$

where $VRAC$ and $VCAC$ are the Fourier coefficients.

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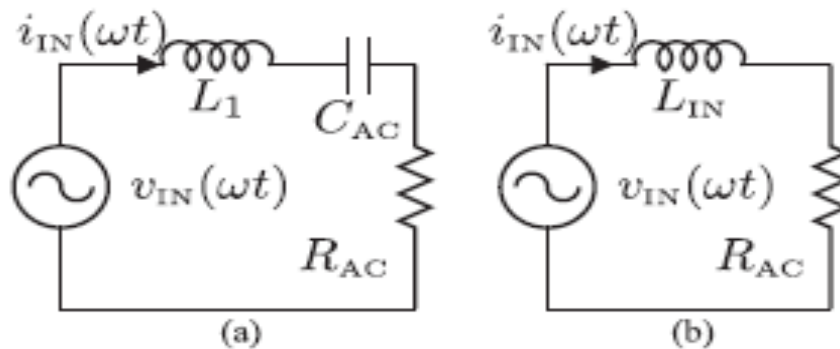


Fig 6. Class-EF2 rectifier equivalent circuit and input impedance at the fundamental switching frequency. (a) equivalent circuit seen by v_{IN} and L_1 (b) input impedance circuit

IV. CONCLUSION

This paper presents the performance analysis of Class-EF2 inverter and a Class-EF2 rectifier topology for WPT systems that operate at multimegahertz frequencies. Detailed mathematical analysis was performed on the Class-EF2 topology to derive its component parameters for optimum switching conditions and other performance parameters such as voltage and current stresses, power-output capability, and harmonic content. Further will be presenting the design and implementation of Class EF2 inverter and rectifier.

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