Design of a Contactless Battery Charger for Cellular Phone

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Abstract - In this paper, the design of a contactless charger for a 3.3W lithium-ion battery of a cellular phone is presented. In this charger, the primary core of the transformer is in the charger unit and the secondary core is in the telephone. The gap(3mm) between them is the thickness of the two plastic cases. The transformer core design for the maximum coupling coefficient and the maximum magnetizing inductance with the size constraint on the secondary side is presented. A Half-Bridge Series Resonant Converter is used to compensate the leakage inductance and to achieve ZVS operation. Voltage gain, current gain analysis, and design procedure are presented. For the battery charging control, infrared LED is used and its performance is verified from the hardware experiments.

I. INTRODUCTION

One of the most frequent failures in chargers for rechargeable batteries of compact electronic devices such as the cellular phone is from the mechanical contact. Energy transfer utilizing inductive coupling[1-4], can overcome this contact failure problem. In this paper, the design of a contactless charger for a 3.3W lithium-ion battery of a cellular phone is presented. In this charger, the primary core of the transformer is in the charger unit and the secondary is in the telephone. The gap(3mm) between them is the thickness of the two plastic cases (Fig.1).

For a cellular phone application, there are two strict constraints in developing the charger. One is the minimum size for the secondary circuits and the other is the cost. Under these constraints, the following issues are covered in this paper.

• Optimum transformer core geometry for the maximum coupling coefficient k and the maximum magnetizing inductance
• Optimum converter topology for the maximum conversion efficiency with the minimum circuitry at the secondary side
• Controlling the battery current and voltage on the primary side

Input voltage is 85VAC ~ 270VAC. The output voltage regulation during the constant voltage mode is 4.1V±100mV, and the output current regulation during the constant current mode is 800mA.

II. TRANSFORMER CORE DESIGN

In order to maximize the coupling coefficient k, the cross-sectional area of the core should be maximized, and in order to maximize the magnetizing inductance Lm, the winding area should be maximized. However, when a core with the center pole is used, i.e. EE or pot core, the distance between the center pole and the outer pole is short, which increases the leakage flux. At the same time the winding area is restricted, resulting in a smaller magnetizing inductance. Therefore, UU and cylindrical core (Fig.2) are considered, which offer flexibility in the cross sectional area and the winding area design.

Using the MAXWELL 3-D field simulator[5-7], these two types of cores are simulated for various core geometrical variation. In the case of the cylinder type core, the coupling coefficient increases as the height and the radius increase. The radius r=6mm is found most cost effective from Δk/Δr trends. The height of the secondary side core is limited by the battery pack size. In the case of the UU type core, it is verified that the coupling coefficient increases as the cross

Fig.1 Arrangement of non-contact transformer

Fig.2 UU type core and cylinder type core geometry
Fig. 3 Coupling coefficient vs. volume of UU type and cylinder type cores

sectional area increases. In order to compare the two core types, the coupling coefficient, with respect to the volume, is plotted in Fig. 3. Higher coupling coefficients can be obtained from the UU type core than from the cylinder type. Considering the size and height restriction, two prototype cores are manufactured and verified by experiment. The experimental results of the two cores are listed in Table 1. The coupling coefficient is higher in experiment than in simulation because some of the flux generated in the primary windings can couple with the secondary windings without going through the core, and this phenomenon does not happen in simulation.

| Table 1 |
|---|---|
| Experimental results of UU type core and cylinder type core |
| | UU type | Cylinder type |
| secondary core volume (mm³) | 844 | 1020 |
| Primary turn (N1) | 420 turns | 471 turns |
| Secondary turn (N2) | 16 turns | 20 turns |
| primary inductance, Lp | 6.87mH | 6.50mH |
| coupling coefficient, k | 0.57 | 0.39 |

The dimensions of the two prototype cores in Fig. 2 are listed below.

UU type core : l₁=7mm, l₂=9.3mm, l₃=7.1mm, l₄=2.2mm, l₅=5.5mm, l₆=2.2mm
Cylinder type core : h₁=9.3mm, h₂=9.3mm, r=6mm

III. HALF-BRIDGE SERIES RESONANT CONVERTER

Using the UU core designed in the previous section, the coupling coefficient k is approximately 0.57. This makes the leakage inductance have a similar magnitude to the magnetizing inductance, resulting in a poor conversion efficiency. This problem can be alleviated if the impedance of the leakage inductance is lowered by employing a series resonant circuit on the secondary side or on both the primary and the secondary side[1-2]. In this application, however, the secondary side resonance is not possible because of the size of the resonant capacitor. Therefore, a Half-Bridge Series Resonant Converter using the primary side resonance has been chosen as shown in Fig. 4. This topology has additional merits that ZVS operation can be achieved, and the secondary side leakage inductance can be utilized as a filter inductor to reduce the size of the filter on the secondary side. Therefore, this converter requires only two rectifying diodes and an output filter capacitor on the secondary side.

Fig. 5 shows the switch drive and the primary voltage and current waveforms for each operating mode. To achieve ZVS operation, the energy of the leakage inductance must be
enough to charge and discharge the parallel capacitors of MOSFETs at the end of mode 3 and the gate signal of S2 must be applied before i1 changes to negative at the end of mode 0. (Fig. 5)

Voltage gain of the converter can be obtained using the frequency domain analysis (Fig.6).

\[ M = \frac{1}{2N} \left[ \frac{Q\omega_n}{k \left( \frac{1}{\omega_n^2} \right)^\frac{3}{2}} + \left( \frac{1}{k} - \frac{1}{\omega_n^2} \right)^\frac{3}{2} \right] \]

where, \( Q = \frac{\omega_n L_2}{N^2 \text{Re}} \), \( \omega_n = \frac{\omega}{\omega_n} \), \( \omega_n = \frac{1}{\sqrt{L_2 C_1}} \)
\( L_n = L_1 + L_2 / L_1 = (1 + k) L_1 \)

The reverse current gain is defined as I1/I2, where I1 is the peak value of i1 and I2 is the peak value of i2.

\[ \frac{I_1}{I_2} = \frac{1}{N} \left[ \left( \frac{1}{k} \right)^\frac{3}{2} + \left( \frac{1}{k} - \frac{1}{\omega_n^2} \right)^\frac{3}{2} \right] \]

In order to minimize the circulating current in the primary circuit, this current gain should be minimized for better efficiency.

In designing the power stage, the following steps are taken.
(1) Measure the transformer primary inductance (Lp) and coupling coefficient.
(2) Determine the resonant frequency considering conduction loss and gap loss.
(3) Determine the transformer turn ratio in order to minimize the current gain and primary current using Eq.(2).
(4) Determine the operating frequency range using Eq.(1).
(5) Verify that ZVS operation is satisfied.

The resonant frequency can be determined from the trade-off between the conduction loss in the primary side and the gap loss of the transformer. In designing the resonant frequency (step 2), the following steps are taken:
(a) Determine the lowest \( \omega_n \). For ZVS operation, \( \omega_n \) must be greater than one.
(b) For each resonant frequency, find possible N which can give the maximum output voltage in case of minimum input voltage.
(c) For a certain input voltage which should be optimized, calculate (N, \( \omega_n \)) set which can give the proper output voltage using Eq.(1). These are some (N, \( \omega_n \)) sets for one resonant frequency.
(d) Calculate minimum reverse current gain for each resonant frequency. Then there remains only one (N, \( \omega_n \), I1/I2) set for one resonant frequency.
(e) Calculate the conduction loss and gap loss for each resonant frequency.
(f) Determine the resonant frequency which minimizes the sum of the conduction loss and gap loss.

Using Eq.(3), the gap loss can be calculated[8]. Due to the large gap, the core loss of the transformer is small compared with the gap loss.

\[ P_g = K_i D I_g f B_m^2 [W] \]

where, \( K_i \) is a gap loss coefficient for different core, D is the strip or tongue width [cm], Ig is the gap length [cm], f is the switching frequency [Hz] and \( B_m \) is the maximum flux density [T].

**Design Result:**
- \( L_p = 6.87 \text{mH} \) (transformer primary inductance)
- \( k = 0.57 \) (transformer coupling coefficient)
- N1:N2 = 420:16 (transformer turns ratio)
- \( f_o = 48 \text{kHz} \) (resonant frequency in eq.(1))
- \( f_s = 50 \text{kHz} - 100 \text{kHz} \) (operating frequency)
- \( C_1 = 2.35 \text{nF} \) (resonant capacitor)

**IV. BATTERY CHARGING CONTROL**

In charging the lithium-ion battery, the output voltage regulation during the constant voltage mode is 4.1V±100mV, and the output current regulation during the constant current mode is 800mA. Since the output control must be done in the primary switching circuit, it is necessary to transfer the sensed information to the output to the primary side. For this, an infrared LED is used. The infrared LED with a current rating of 100mA can transfer information through the 3mm plastic block with a simple circuit as shown in Fig.7.

The diode OR circuit is used for an automatic mode change between the constant current mode and the constant voltage mode. The comparator drives the infrared LED. When the sensed output is higher than Vref1 of the comparator, the infrared LED is activated and the voltage of Vr goes low by the sensing transistor. Vr can have only two preset values (high and low) according to the On and Off of the LED. Vref2 of the integrator is set to a medium value of high and low voltage of Vr. The output voltage of the integrator (Vc) becomes high if Vr is low, and this makes the operating frequency increases, and the converter output voltage or current decreases.

The integrator of this control scheme provides the control...
voltage either to increase or to decrease the switching frequency depending on the output value of the comparator. In a charger application that does not require fast control speed, this simple control scheme can reduce the component number and size on the secondary side.

V. EXPERIMENTAL RESULTS

Fig. 8 shows the MOSFET voltage and current waveform for Vin=155VDC. The negative current makes the voltage zero and ZVS operation is achieved. Fig. 9–10 shows the efficiencies for the input voltage variation (Io = 800mA) and the load current variation (Vin = 230V). The maximum efficiency is 75% and the minimum efficiency is 72% for input voltage variation (the control circuit loss is not included in the efficiency data). If a synchronous rectifier is used, the efficiency will become much higher. The performance of the battery charging control is verified as shown in Fig. 11 for Vin=155VDC. Initially the charger operates in the constant current mode and is then switched to the constant voltage mode. The constant voltage regulation and the constant current regulation are performed well within the specified limit using the simple control circuit.

VI. CONCLUSION

The design of a contactless charger for a cellular phone is presented. The transformer core design for the maximum coupling coefficient and the maximum magnetizing inductance with a size constraint is presented. A Half-Bridge series resonant converter is used to compensate the leakage inductance, to achieve ZVS operation, and to reduce the secondary side volume. Voltage gain, current gain analysis, and design procedure are presented. For the battery charging control, the simple control scheme using infrared LED is used and its performance is verified from the hardware experiments.

REFERENCE

Fig. 8 MOSFET voltage and current waveform

- Upper: MOSFET voltage (100V/div)
- Lower: MOSFET current (100mA/div)

Fig. 9 Efficiency for input voltage variation (Io = 800mA)

Fig. 10 Efficiency for load current variation (Vin = 230V)

Fig. 11 Battery charging waveform

- Vb: battery voltage [1V/div]
- Ib: battery current [200mA/div]
- Time [1ks/div]