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The characteristic analysis of contact-less power supply by 3D finite element method

Han-Seok Park† · Yun-Hyun Cho* · Hong-Sub Jung** · Kyung-il Woo*** · · Kyung-Ho Kim****

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Abstract : This paper proposes the calculation method of magnetic coupling coefficient of contact-less power supply by the 3D finite element method with a variation of the secondary core position. The primary, secondary self and leakage inductances and the capacitances of a resonant circuit are calculated by the finite element analysis results. The magnetic coupling coefficients are obtained also. The power factors are obtained by simulation for the magnetic coupling coefficients and compared.

Key words : Contact-less power supply, magnetic coupling coefficient, power factor, resonant circuit.

1. Introduction

In the last ten years, a new electric energy supplying technology, Contact-less Power Supply (CPS), has been developed and now thousands of these kinds of devices using this technology are working all around. CPS supplies electrical current by means of an electromagnetic induction, which is similar to the primary-to-secondary energy conversion of a transformer. However, the primary and the secondary windings of a conventional transformer are placed on a common ferromagnetic core. This geometry creates a high coupling coefficient but excludes any relative movement between the two windings. The CPS transformer, on the other hand, stretches the primary winding into a long loop and places the secondary winding on an open-end core with surrounds the primary conductors and allows relative movement between the two. The reduced coupling of the open core geometry is compensated by a higher primary frequency. So the CPS electric circuit has a high frequency intermediate circuit [1]-[5] . The exact

† 책임저자(부경대학교 전기제어계측공학과), E-mail : phanseok@pknu.ac.kr, T : 051)620-1436
* 동아대학교 전기공학과
** 한국컨테이너 부두공단 부산항 종합변전소 소장
*** 부경대학교 전기제어계측공학부
**** 동아대학교 전기공학과 대학원

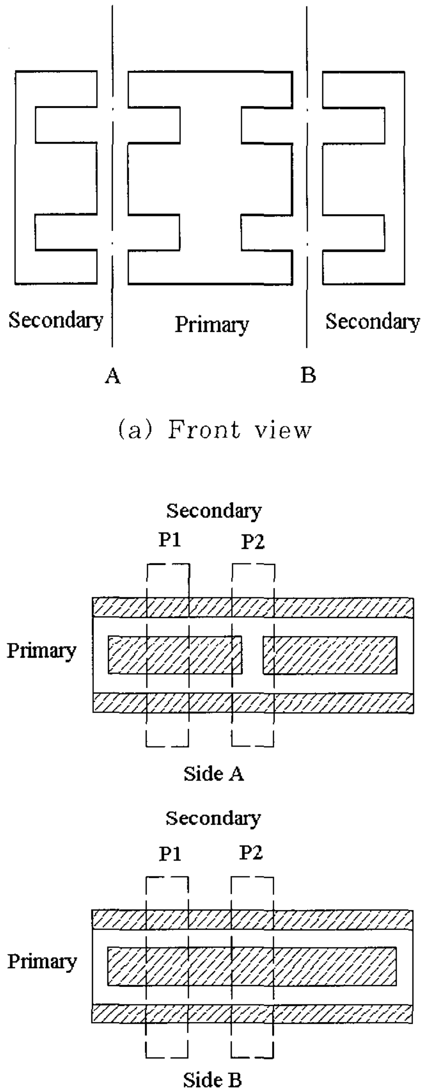
modeling of CPS transformer is very important. The papers dealt with the finite element method(FEM) in modeling the CPS electric circuit have not been reported.

This paper proposes the calculation method of magnetic coupling coefficient of CPS by the 3D FEM with the secondary core position change. The primary, secondary self and leakage inductances and the capacitances of a resonant circuit are obtained by the 3D finite element analysis and the magnetic coupling coefficients are calculated from these values and compared. The electrical scheme of CPS to check the effectiveness of the capacitances is designed. The transformer in the electrical scheme of CPS is designed by the inductances and magnetic coupling coefficient obtained by the 3D finite element analysis. The power factors of the transformer are compared by OrCAD simulation for the secondary core position change.

2. Characteristic Analysis

2.1 Analysis model

Fig. 1 shows a CPS model with alternative switching for the primary source adopted in this paper. The side A of the primary coil and core are segmented to reduce the leakage flux. We set P1, P2 as the secondary core positions to check the leakage flux effect. The specification of the analysis model is Table 1.



(a) Front view

(b) Side view

Fig. 1 CPS structure.

Table 1 Specification of CPS.

	Parameter	Value
Primary	Lamination depth	300[mm]
	Turn of coil	45
Secondary	Lamination depth	60[mm]
	Turn of coil	100
Frequency	1000[Hz]	
Air-gap	0.5 [mm]	

2.2 Governing Equation

The fundamental equation of the magnetic field using 3D finite element method with edge element can be written using the magnetic vector potential A as follows:

$$rot(\nu\ rotA) = J_0 + J_e \tag{1}$$

where ν is the reluctivity, J_0 is the current density and J_e is the eddy current density. J_e is given as follows:

$$J_e = -\sigma\left(\frac{\partial A}{\partial t} + \nabla \Phi\right) \tag{2}$$

where σ is the electrical conductivity and Φ is the electric scalar potential. From (2) and $J_e=0$, we can obtain the following equation.

$$\nabla \cdot \left\{-\sigma\left(\frac{\partial A}{\partial t} + \nabla \Phi\right)\right\} = 0 \tag{3}$$

The magnetic field can be calculated by coupling (1)-(3).
The following equations can be obtained by the Galerkin's method from (1) and (3).

$$G_i = \int_V rot\, N_i \cdot (\nu\ rot\ A)\ dV - \int_{V_e} \boldsymbol{N_i} \cdot J_0\ dV + \int_{V_e} \boldsymbol{N_i} \cdot \left\{\sigma\left(\frac{\partial A}{\partial t} + \nabla \Phi\right)\right\} dV - \int_S \boldsymbol{N_i} \cdot \{(\nu rot\ A) \times n\}\ dS = 0 \tag{4}$$

$$G_{di} = \int_{V_e} \nabla N_i \cdot \sigma\left(\frac{\partial A}{\partial t} + \nabla \Phi\right) dV + \int_{S_e} \boldsymbol{N_i}\left\{-\sigma\left(\frac{\partial A}{\partial t} + \nabla \Phi\right)\right\} \cdot \boldsymbol{n}\ dS = 0 \tag{5}$$

where $\boldsymbol{N_i}$ is the vector interpolation function for A, N_i is the scalar interpolation function for Φ, V is the analyzed region, V_e is the region of the conductor with the eddy current, S and S_e are the boundary of analyzed region and eddy current region, respectively. $\boldsymbol{n}$ is the unit outward normal vector on the surface S and S_e. The 3D region is divided into tetrahedral elements, and the matrix of finite element method is solved by the ICCG method and Newton-Raphson iteration technique is used for the non-linear characteristics. The circuit equations (6) and (7) should be considered to solve (4) and (5).

$$[U] = [R]\,[I] + [L_0]\frac{d}{dt}[I] + [E] \tag{6}$$

$$[E] = \frac{d}{dt}[\lambda_s] \tag{7}$$

where,
- $[U] = (U_a, U_b, U_c)^T$: phase voltages
- $[I] = (i_a, i_b, i_c)^T$: phase currents
- $[R] = diag(R_a, R_b, R_c)$: primary winding resistances
- $[L_0] = diag(L_a, L_b, L_c)$: end winding leakage inductances
- $[E] = (E_a, E_b, E_c)^T$: phase e.m.f.
- $[\lambda s] = (\lambda a, \lambda b, \lambda c)^T$: phase flux linkages

3. Analysis Result

3.1 Inductance profile and coupling coefficient

3D FEM is used to analyze the characteristic of CPS. Fig. 2 and Fig. 3 show the inductance profile and the coupling coefficient with variation of the secondary core position. In these figures, the displacement is the distance from the left position of the primary core to the secondary core position. P1 and P2 are the positions of the secondary core. Because in 3D analysis much longer calculation time is needed we divided the analysis model with coarse meshes. Fig. 4 shows the finite element analysis results at P1. It is known that there are lots of leakage fluxes. Table 2 shows the self inductance L_{11}, L_{22}, the mutual inductance L_{12}, L_{21} and the magnetic coupling coefficient k for two secondary core positions. Calculated by (8) the magnetic coupling coefficient of P1 is greater than that of P2 because of the leakage flux.

$$k = L_{12}/(\sqrt{L_{11}L_{22}}) \quad (8)$$

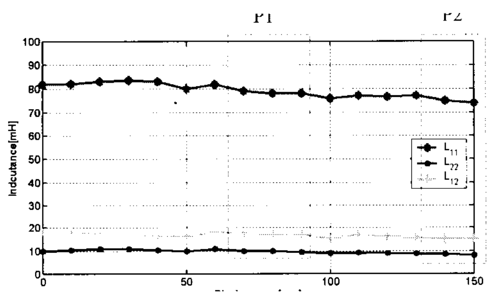


Fig. 2 Inductance profile.

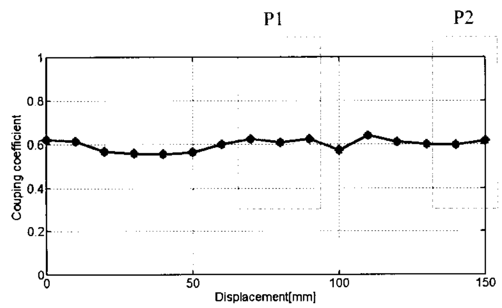
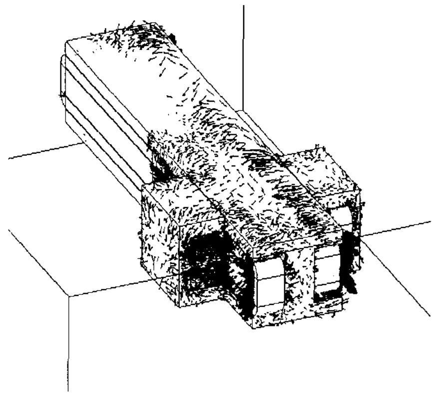


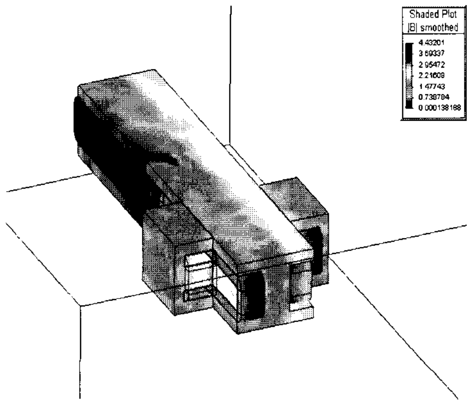
Fig. 3 Coupling coefficient.

Table 2 Characteristics of two CPSs.

	Primary		Secondary		Coupling coefficient
	Self inductance	Mutual inductance	Self inductance	Mutual inductance	
P1	76[mH]	22[mH]	7[mH]	18[mH]	0.59
P2	70[mH]	18[mH]	6[mH]	18[mH]	0.5



(a) Vector diagram of the flux



(b) Magnetic flux density

Fig. 4 Vector diagram and magnetic flux density.

3.2 Electrical scheme of CPS

Fig. 5 shows the electrical scheme of CPS to check the effectiveness of the capacitance. In this figure, the primary series capacitor C_p has been introduced to compensate the very high leakage inductance of primary and secondary coils. It is possible to calculate this capacitor from the condition of running at resonance by (9).

$$C_p = 1/(4\pi^2 f^2 (L_{1s} + L'_{2s})) \quad (9)$$

The secondary parallel capacitor C_s,

mounted in the secondary side, compensates the magnetizing inductance and generates an inductive current equal with the magnetizing current, so the magnetizing current is locally generated and is avoided to take it from the primary inverter. Its value can be also calculated from resonant condition of secondary circuit by (10).

$$C_s = 1/(4\pi^2 f^2 (L_m')) \tag{10}$$

In order to obtain high frequency an inverter is needed on the primary side. As resonant inverter is recommended because of the high efficiency, in this case, a series resonance inverter is used. Because of the resonance, in the primary circuits the primary voltage depends on the load currents, it is not constant and therefore it is very difficult to design such a transformer. On the secondary side, the energy is supplied as high frequency voltage, as d.c. current after rectifying or as a.c. current of industry frequency after another inverter. But, because of the primary voltage dependence on load current, it is necessary to provide a secondary voltage stabilization in a closed loop. Fig. 6 shows the output/

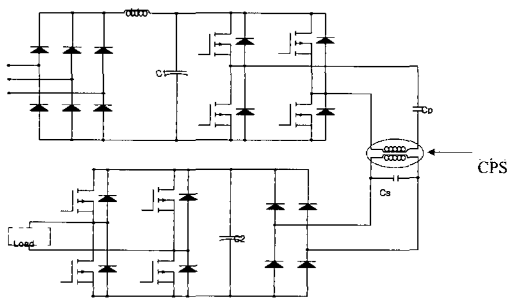


Fig. 5 The electrical scheme of CPS.

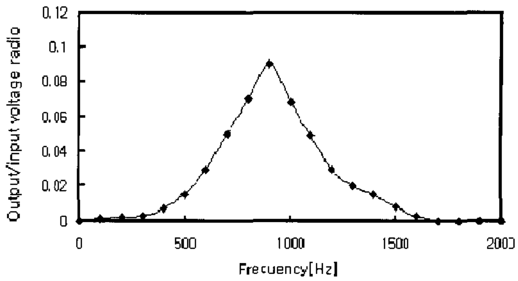
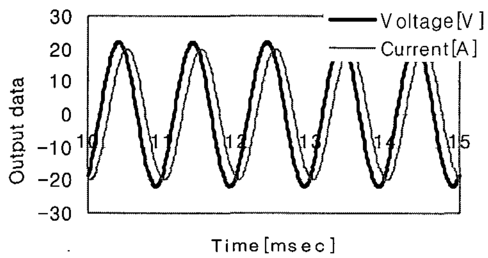
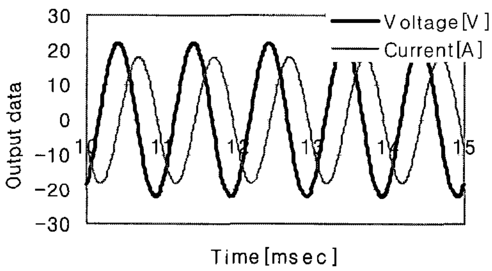


Fig. 6 Output/input voltage radio dependence of the CPS on the running frequency at C_p = 0.09 μF, C_s = 0.18 μF



(a) Secondary voltage and current at P1



(b) Secondary voltage and current at P2

Fig. 7 Simulation result of transformer with a resonant circuit.

input voltage dependence of the CPS on the running frequency for C_p and C_s. We use the frequency between 0.5 and 1.2kHz in order to stabilize the output voltage. Fig. 7 shows the secondary voltage and current of CPS at P1 and P2. It is shown that the power factor of P1 is better than that of P2 because of the large leakage fluxes. We used OrCAD simulation and switching devices are MOSFET, sampling time is 0.1[msec]. $L_1=50[\mu H]$, $C_1=1[\mu F]$, $C_2=1[\mu F]$.

4. Conclusion

This paper proposes the calculation method of magnetic coupling coefficient of CPS by the 3D FEM. The primary, secondary self and leakage inductances and the capacitances of a resonant circuit are obtained by the 3D finite element analysis. The magnetic coupling coefficients are calculated from these obtained values and compared. The electrical scheme of CPS to check the effectiveness of the capacitances is designed and it is shown that the power factor of P1 is better than that of P2 because of the large leakage fluxes. As a result, we know that the combination of 3D FEM modeling with circuit analysis is useful to obtain a higher modeling accuracy.

Acknowledgment

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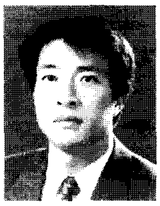


박한석 (朴瀚錫)

1955년 3월 23일생. 중앙대학교 전기공학과 공학사(1981년), 공학석사(1983년), 한국해양대학교 전기공학과 공학박사(1996년), 부경대학교 전기제어계측공학부 교수(1986년-현재)

조윤현 (曺允鉉)

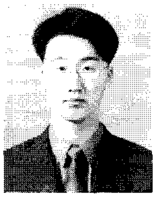
1960년 2월 27일생. 한양대학교 전기공학과 공학사(1984년), 공학석사(1986년), 공학박사(1991년), 삼성종합기술원 선임연구원(1986년-1988년), 한국전기연구소 선임연구원(1990년-1995년). 동아대학교 전기공학과 교수(1995년-현재)

정홍섭 (鄭洪燮)

1959년 12월 25일생. 부경대학교 전기공학과 졸업(1997년). 부경대학교 전기공학과 설사졸업(1999년). 동대학원 박사과정(2003-현재) 현재 한국컨테이너 부두공단 부산항 종합변전소 소장.

우경일 (禹炅一)

1969년 9월 6일생. 한양대학교 전기공학과 공학사(1995년), 공학석사(1997년), 공학박사(2001년), 수원대학교 초빙교수(2001년), 부경대학교 전기제어계측공학부 선임강사(2002년-현재)



김경호 (金景浩)

1976년 12월 6일생. 동아대학교 전기공학과 공학사(1999년), 공학석사(2001년), 동대학원 박사과정(2001년-현재)