Feasibility Analysis and Calculation of HTS Inductive Charging Technology

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Abstract — This work proposes an innovative large-current coupling technology by introducing high temperature superconducting (HTS) techniques, with loosely-coupled and air-coupled inductive charging schemes analyzed in the work. The proposed HTS inductive charging technology has less harmonic component, less energy loss, and can also transmit large power to the load.

I. INTRODUCTION

Currently the worldwide research and development concerning the applications of electric vehicles (EVs), e.g., pure electric vehicles, hybrid electric vehicles, have been progressed actively. The charging technology for on-board batteries is a key technology for EVs. The conventional contact-type charging technology has many practical problems, e.g., strict charging environment, possible safety risks, etc. A new focus has been formed recently in non-contact-type charging technology. The SAE electric vehicle inductively coupled charging standard (SAE J-1773) and other conventional inductive charging systems have adopted high-frequency coupling scheme. However high-frequency coupling will cause serious core loss and switching loss. Based on the worldwide developments of high temperature superconducting (HTS) air-core transformers [1,2], we propose an innovative large-current coupling technology by introducing HTS techniques in the work [1-4].

II. ANALYSIS ON LOOSELY-COUPLED INDUCTIVE CHARGING

The equivalent circuit of the primary side (inductance L_p, internal resistance R_p) with a compensation capacitor C_p is a series-resonance circuit, as shown in Fig. 1.

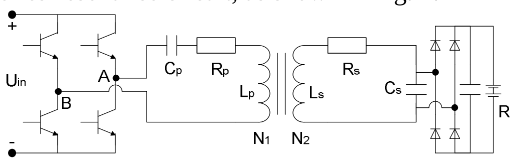

Fig. 1. The equivalent circuit of the inductive charging system

The AC voltage output of the inverter $U_{AB}(t)$ is square-wave voltage and can be expressed by

$$u_{AB}(t) = \sum_{n=1,3,5\ldots}^{\infty} \frac{4U_{in}}{n\pi} \sin(n\omega t) \tag{1}$$

where U_{in} is the amplitude of the square-wave voltage; n, the order of harmonics; ω, the angular frequency of the square-wave voltage.

According to Kirchhoff's current law (KCL), there is

$$L_p \frac{di_p(t)}{dt} + R_p i_p(t) + \int_0^t \frac{i_p}{C_p} dt = u_{AB}(t) \tag{2}$$

so the primary current $i_p(t)$ is obtained as follows

$$i_p(t) = \frac{-4\omega U_{in}}{\pi[(\frac{1}{C_p} - n^2\omega^2 L_p)^2 + (n\omega R_p)^2]} \times \sum_{n=1,3,5\ldots}^{\infty} [(\frac{1}{C_p} - n^2\omega^2 L_p)\cos(n\omega t) + n\omega R_p \sin(n\omega t)] \tag{3}$$

A. Harmonic Analysis

When the frequency of the square-wave voltage $f_0 = 1/(2\pi\sqrt{L_p C_p})$, the primary series-resonance circuit is operated at resonance status. The fundamental current component $i_{p1}(t)$ can be expressed by

$$i_{p1}(t) = I_{p1m} \sin\frac{t}{\sqrt{L_p C_p}} = \frac{4U_{in}}{\pi R_p} \sin\frac{t}{\sqrt{L_p C_p}} \tag{4}$$

Considering the quality factor Q [$=\omega L_p/R_p$] of the primary windings, the ratio of the amplitude I_{p1m} of the fundamental current component to the amplitude I_{pnm} of the n-order harmonic current component can be expressed by

$$\frac{I_{pnm}}{I_{p1m}} = \frac{1}{\sqrt{(1-n^2)^2 Q^2 + n^2}} \tag{5}$$

The ratio of I_{pnm}/I_{p1m} with different Q is shown in Fig. 2. I_{pnm}/I_{p1m} drops along with the increment of Q. So the HTS windings with approximate zero internal resistance ($R_p \approx 0$) with a high Q and restrain the harmonic ratio effectively.

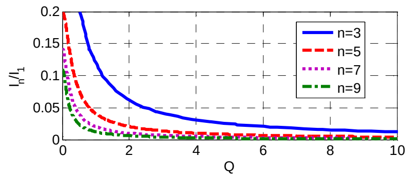

Fig. 2. The ratio of I_{p1m}/I_{pnm} with different Q.

B. Transmitted Power Analysis

According to magnetic circuit law, the magnetic flux inside the iron core can be expressed by

$$\phi(t) = \phi_m \sin\frac{t}{\sqrt{L_p C_p}} = \frac{N_1 I_{p1m}}{2R_1 + R_2 + R_3} \sin\frac{t}{\sqrt{L_p C_p}} \tag{6}$$

where N_1 is the number of turns in primary windings; the magnetic reluctance R_1, R_2 and R_3 are $\delta/S\mu_0$, $l_1/S\mu_0\mu_r$, and $l_2/S\mu_0\mu_r$, as shown in Fig. 3; μ_0 and μ_r are air permeability and relative permeability of the iron core.

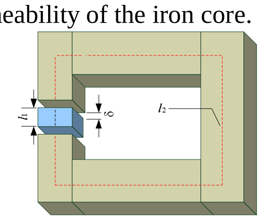

Fig. 3. The structure of the loosely-coupled transformer

According to Faraday's law, there is

$$\begin{aligned} E_2(t) &= -N_2\frac{d\phi(t)}{dt} = E_{2\mathrm{m}}\cos\frac{t}{\sqrt{L_\mathrm{p}C_\mathrm{p}}} \\ &= -\frac{\mu_0\mu_\mathrm{r}N_1N_2I_{\mathrm{p1m}}S}{(2\delta\mu_\mathrm{r}+l_1+l_2)\sqrt{L_\mathrm{p}C_\mathrm{p}}}\cos\frac{t}{\sqrt{L_\mathrm{p}C_\mathrm{p}}} \end{aligned} \tag{7}$$

then the KCL equation in the secondary side is

$$L_\mathrm{s}\frac{di_\mathrm{s}(t)}{dt} + R_\mathrm{s}i_\mathrm{s}(t) + \frac{1}{C_\mathrm{s}}\int_0^t i_\mathrm{s}(t)dt = E_2(t) \tag{8}$$

so the secondary current $i_\mathrm{s}(t)$ is obtained as follows

$$\begin{aligned} i_\mathrm{s}(t) &= \frac{E_{2\mathrm{m}}}{[(\frac{1}{C_\mathrm{s}} - \frac{L_\mathrm{s}}{L_\mathrm{p}C_\mathrm{p}})^2 + \frac{R_\mathrm{s}^2}{L_\mathrm{p}C_\mathrm{p}}]\sqrt{L_\mathrm{p}C_\mathrm{p}}} \\ &\times[(\frac{1}{C_\mathrm{s}} - \frac{L_\mathrm{s}}{L_\mathrm{p}C_\mathrm{p}})\sin\frac{t}{\sqrt{L_\mathrm{p}C_\mathrm{p}}} + \frac{R_\mathrm{s}}{\sqrt{L_\mathrm{p}C_\mathrm{p}}}\cos\frac{t}{\sqrt{L_\mathrm{p}C_\mathrm{p}}}] \end{aligned} \tag{9}$$

When f_0 in the primary side is equal to that in the secondary side, i.e., $L_\mathrm{p}C_\mathrm{p}=L_\mathrm{s}C_\mathrm{s}$, then (9) can be simplified as

$$i_\mathrm{s}(t) = \frac{E_{2\mathrm{m}}}{R_\mathrm{s}}\cos\frac{t}{\sqrt{L_\mathrm{p}C_\mathrm{p}}} \tag{10}$$

So the voltage across C_s can be expressed by

$$u_{\mathrm{Cs}}(t) = \frac{1}{C_\mathrm{s}}\int_0^t i_\mathrm{s}(t)dt = \frac{E_{2\mathrm{m}}\sqrt{L_\mathrm{p}C_\mathrm{p}}}{C_\mathrm{s}R_\mathrm{s}}\sin\frac{t}{\sqrt{L_\mathrm{p}C_\mathrm{p}}} \tag{11}$$

then the maximum transmitted power P_{Rmax} to the load R can be expressed by

$$P_{\mathrm{Rmax}} = \frac{32\pi^4 f_0^4}{R}[\frac{\mu_0\mu_\mathrm{r}N_1N_2I_{\mathrm{p1m}}SL_\mathrm{s}}{(2\delta\mu_\mathrm{r}+l_1+l_2)R_\mathrm{s}}]^2 \tag{12}$$

So the maximum transmitted power P_{Rmax} is proportional to the square of I_{p1m} and fourth power of f_0. For the current inductive charging devices with conventional copper windings, the effective scheme to improve the P_{Rmax} value is increasing f_0, e.g., 100 kHz, and achieving high-frequency coupling. However higher f_0 will cause more switching loss in the inverter and also generate more core loss. In addition, the design and production cost of such a high-frequency device should also be considered. For the proposed superconducting inductive charging scheme in the paper, the critical current density of HTS tapes can reach >10 kA/cm^2 (77K,0T), which is far larger than that of copper wires (about 300 A/cm^2), so P_{Rmax} can be improved by increasing I_{p1m} and achieving large-current coupling.

C. Power Loss Analysis

The coil loss can be sharply reduced by introducing HTS windings ($R_\mathrm{p}\approx 0$, $R_\mathrm{s}\approx 0$) to primary side, secondary side, or both sides. Moreover, HTS inductive charging system operated with very large primary current and relatively low resonant operation frequency will cause less core loss. In sum, the proposed HTS inductive charging system has very small energy loss and very high transmitted efficiency.

III. ANALYSIS ON AIR-COUPLED INDUCTIVE CHARGING

Since the practical relative permeability μ_r of the iron core is nonlinear and the magnetic flux $\Phi(t)$ inside the iron core will saturate according to the *B-H* curve of a certain magnetic material. It is very difficult to reach the magnetic flux density of 2 T or above. In the work, we propose an innovative HTS air-coupled inductive charging scheme based on the above analysis of the HTS loosely-coupled inductive charging scheme. A typical air-coupled schematic diagram is shown in Fig. 4.

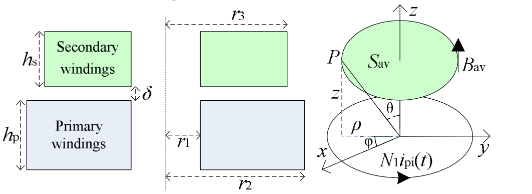

Fig. 4. The air-coupled transformer and its equivalent model

Define the average magnetic flux of the secondary windings $\Phi_{\mathrm{av}}=B_{\mathrm{av}}S_{\mathrm{av}}$, then the secondary induced voltage $E_2(t)$ can be expressed by

$$E_2(t) = -\frac{N_1N_2B_{\mathrm{av}}S_{\mathrm{av}}}{\sqrt{L_\mathrm{p}C_\mathrm{p}}}\cos\frac{t}{\sqrt{L_\mathrm{p}C_\mathrm{p}}} \tag{15}$$

Assume that the coil structure and current distribution are uniform, then the primary and secondary windings can be equivalent to two current loops, as shown in Fig. 4. The primary current loop is with the current $N_1i_{\mathrm{p1}}(t)$ and the radius $a=(r_1+r_2)/2$, the secondary current loop is with the current $N_2i_\mathrm{s}(t)$ and the radius $b=(r_1+r_3)/2$, the vertical distance between the primary current loop and secondary current loop $z=h_\mathrm{p}/2+\delta+h_\mathrm{s}/2$. So B_{av} is the axial magnetic flux density along the edge of the secondary current loop

$$B_{\mathrm{av}} = \frac{\mu_0 I_{\mathrm{p1m}}}{2\pi\rho\sqrt{(a+\rho)^2+z^2}}\times[\frac{a^2-\rho^2-z^2}{(a-\rho)^2+z^2}E+K] \tag{16}$$

and S_{av} is the area of the secondary current loop. So $E_2(t)$ can be calculated by

$$\begin{aligned} E_2(t) &= -\frac{\mu_0N_1N_2I_{\mathrm{p1m}}b^2}{4\rho\sqrt{L_\mathrm{p}C_\mathrm{p}[(a+\rho)^2+z^2]}}\cos\frac{t}{\sqrt{L_\mathrm{p}C_\mathrm{p}}} \\ &\times[\frac{a^2-\rho^2-z^2}{(a-\rho)^2+z^2}E+K] \end{aligned} \tag{17}$$

The maximum transmitted power P_{Rmax} to the load R can be expressed by

$$P_{\mathrm{Rmax}} = \frac{32\pi^4 f_0^4}{R}\{\frac{\mu_0N_1N_2I_{\mathrm{p1m}}b^2L_\mathrm{s}}{4\rho R_\mathrm{s}\sqrt{(a+\rho)^2+z^2}}[\frac{a^2-\rho^2-z^2}{(a-\rho)^2+z^2}E+K]\}^2 \tag{18}$$

IV. REFERENCES

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