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Design and verification of large-moment transmitter loops for geophysical applications

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ABSTRACT

In this paper we discuss the modeling, design and verification of large-moment transmitter (TX) loops for geophysical applications. We first develop two equivalent circuit models for TX loops. We show that the equivalent inductance can be predicted using one of two empirical formulas. The stray capacitance of the loop is then calculated using the measured self-resonant frequency and the loop inductance. We model the losses associated with both the skin effect and the dissipation factor in both of these equivalent circuits. We find that the two equivalent circuit models produce the same results provided that the dissipation factor is small.

Next we compare the measured input impedances for three TX loops that were constructed with different wire configurations with the equivalent circuit model. We found excellent agreement between the measured and simulated results after adjusting the dissipation factor. Since the skin effect and dissipation factor yield good agreement with measurements, the proximity effect is negligible in the three TX loops that we tested. We found that the effects of the dissipation factor dominated those of the skin effect when the wires were relatively close together. When the wires were widely separated, then the skin effect was the dominant loss mechanism. We also found that loops with wider wire separations exhibited higher self-resonant frequencies and better high-frequency performance.

1. Background

Large-diameter, multiple-turn transmitting loops, with large drive current, are often used for geophysical prospecting. For example, Ward et al. (1974) use a ground-based transmitting loop with a diameter of 9 m and up to 38 turns. Sorenson and Auken (2004) use an airborne transmitting loop that is 12.5 m × 12.5 m square, with up to 4 turns. Despite the wide use of these large-diameter, multiple-turn transmitting loops, there is very little discussion about the details of the design and construction of these loops in the open literature. Our goal in this paper is to thoroughly document the details of the modeling, design and verification of large-moment transmitter (TX) loops for geophysical applications. The desired characteristics for these loops include large moment, wide bandwidth, and high efficiency.

2. Introduction

We discuss the design, simulation, and testing of high-power transmitter (TX) loops for geophysical applications. In typical frequency-domain geophysical applications with large-moment TX loops, a tuning capacitor is added in series with the TX loop in order to cancel out the large inductive reactance of the loop. Capacitive tuning reduces the input impedance of the tuned TX loop so that it is approximately equal to the Alternating Current (AC) resistance of the wire. This allows for larger currents to be driven in the TX loop for a given source voltage, thereby increasing the TX moment. By minimizing the AC loop resistance, we also reduce the power requirements for the TX amplifier, which is crucial in the design of tuned, high-moment TX loops, especially at higher frequencies.

In order to minimize the AC resistance of the TX loop, it is important to understand the loss mechanisms in TX loops. This includes an understanding of stray capacitance and the self-resonant frequency of the loop, the dissipation factor associated with the stray capacitance, and the added resistance associated with the skin effect. Therefore, in this paper we first show two different equivalent circuit models that are commonly used to model air-core inductors and can be applied to geophysical TX loops. We then derive analytical expressions for the input resistance and input reactance associated with each of these circuits. These analytical expressions are first used to find expressions for the self-resonant frequencies for TX loops. The measured self-resonant frequency can then be used to calculate the stray capacitance of the loop. We then outline a skin effect model that can be used to predict the series resistance terms in the equivalent circuit models. Finally, we show how the dissipation factor can be incorporated into the circuit models.

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After developing a theoretical analysis for the input impedance associated with air-core inductors, we then apply these theoretical models to three different geophysical TX loops with different wire spacings. We provide two different empirical formulas from the literature that are shown to provide accurate estimates of the inductance for the loops. We also show that the theoretical model provides a good fit for the measured impedance data for all three TX loops. Finally, we use the theoretical and measured results to better understand the characteristics of the three TX loops and discuss the advantages and limitations of each loop.

3. Equivalent circuit models for TX loops

3.1. Standard equivalent circuit model for inductors

As discussed by Besser and Gilmore (2003), the standard equivalent circuit model for a single-layer, air-core inductor can be represented as shown in Fig. 1. The series resistance R_S models the conductive resistance and the parallel capacitor C_P models the self-capacitance associated with the windings. The parallel resistance R_P models the losses associated with the wire insulation and any coating or low-loss core material that is used to support the inductor. Agilent (2013) has one model with a parallel resistor and capacitor where they place the series resistance in front of the parallel circuit. Other authors, such as Massarinil and Kazimierczuk (1997), Eroglu (2013), and Agilent (2013) use a simpler model for inductors that neglects the parallel resistor.

Since a TX loop is like a large inductor, geophysicists have also employed the equivalent circuit in Fig. 1 to model their air-core TX loops (Butler, 2015, and Kozhevnikov, 2009). Note that Kozhevnikov (2009) actually attached a damping resistor in parallel with the capacitor shown in Fig. 1 to damp the ringing in their transient application. Rosthal and Zhang (2013) also employ an equivalent circuit like the one shown in Fig. 1, where the parallel resistor is used to account for core losses. They state that this resistor can be neglected when modeling an air-core coil. Sternberg (1999) neglects the parallel resistor in Fig. 1 and includes all of the loss terms for the TX loop into the series resistance.

When the frequency is low (i.e., less than about 20% of the Self-Resonance Frequency (SRF) according to Besser and Gilmore (2003)), the capacitive reactance is large and the capacitor and parallel resistor in Fig. 1 can be ignored. Therefore, many researchers use a simple series R and L circuit to model TX loops at low frequencies (Aldridge, 1973).

Circuit analysis can be used to obtain an expression for the input impedance associated with the equivalent circuit in Fig. 1. By summing the admittances for the three parallel legs, we find that

$$Z_{In}^{P} = R_{In}^{P} + jX_{In}^{P} = \left[\frac{1}{R_S + j\omega L_S} + j\omega C_P + \frac{1}{R_P}\right]^{-1}. \tag{1}$$

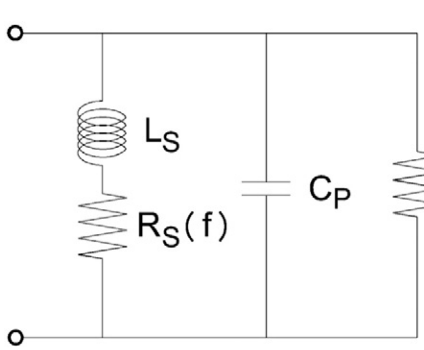

Fig. 1. Equivalent circuit model for a single-layer, air-core inductor. This model includes a resistor in parallel with the capacitor.

Note that the superscript P is used to denote that a parallel arrangement is used for the capacitor and resistor as shown in Fig. 1. Algebra can then be used to find the following expressions for the input resistance and input reactance:

$$R_{In}^{P} = \frac{R_S\left[1 + \frac{R_S}{R_P} + \frac{(\omega L_S)^2}{R_S R_P}\right]}{\left(1 + \frac{R_S}{R_P} - \omega^2 L_S C_P\right)^2 + \left(\frac{\omega L_S}{R_P} + \omega C_P R_S\right)^2} \tag{2}$$

$$X_{In}^{P} = \frac{\omega L_S(1 - \omega^2 L_S C_P) - R_S^2 \omega C_P}{\left(1 + \frac{R_S}{R_P} - \omega^2 L_S C_P\right)^2 + \left(\frac{\omega L_S}{R_P} + \omega C_P R_S\right)^2}. \tag{3}$$

First note that the SRF occurs when $X_{In}^{P} = 0$, i.e.,

$$f_{SRF} = \frac{1}{2\pi}\sqrt{\frac{L_S - R_S^2 C_P}{L_S^2 C_P}}. \tag{4}$$

Note that Eq. (4) can be simplified for low-loss inductors as

$$f_{SRF} \approx \frac{1}{2\pi\sqrt{L_S C_P}}; \quad \text{if } R_S << \sqrt{\frac{L_S}{C_P}} = \omega_{SRF} L_S. \tag{5}$$

3.2. Alternate equivalent circuit model for inductors

We have found that the equivalent circuit in Fig. 2 also provides a good model for TX loops. The inspiration for this equivalent circuit model came from Coilcraft (2001) who provides a Radio Frequency (RF) inductor model like the one shown in Fig. 2. Their model also includes an additional resistor in series with the circuit shown in Fig. 2, but this resistor isn't necessary for our application. Once again the frequency-dependent resistor that is in series with the ideal inductor can be used to model the skin effect and conduction losses of the wire. This time the Equivalent Series Resistance (ESR) can be used to model the dielectric (insulation) losses associated with the self-capacitance of the inductor.

Circuit analysis can once again be used to find the input resistance and reactance looking into the equivalent circuit shown in Fig. 2:

$$R_{In}^{S} = \frac{R_S + (\omega C_P)^2 R_S \cdot ESR(R_S + ESR) + ESR(\omega^2 L_S C_P)^2}{(1 - \omega^2 L_S C_P)^2 + (\omega C_P)^2 (R_S + ESR)^2} \tag{6}$$

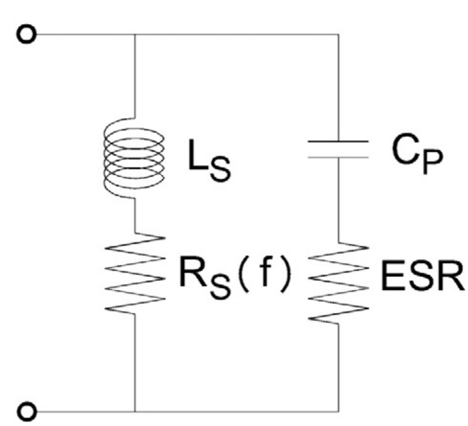

Fig. 2. Equivalent circuit model for the air-core loop TX. This model includes a resistor in series with the capacitor.