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Reduction of CM Noise Emissions From PLC Modem Using Optically Coupled Signaling

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Abstract—This paper investigates the feasibility of implementing an optical coupler in the analog front-end (AFE) circuit of a power line communication (PLC) modem to reduce the common mode (CM) noise coupling onto the power lines. Based on a two-current-probe measurement approach, an equivalent CM circuit model is developed to predict the expected CM current generated by the PLC modem when it is plugged onto the power line network. The model provides insight into the CM noise coupling mechanism from any PLC modem onto the power line network, so that the effectiveness of using an optical coupling technique in reducing the CM current from the power line network can be properly quantified.

Index Terms—Electromagnetic analysis, electromagnetic interference, optical coupling, power system communication.

I. INTRODUCTION

THE power line network has been viewed as a convenient and inexpensive communication medium for data and voice transmission. Unfortunately, its unbalanced nature leads to significant common mode (CM) radiations, which might interfere with the existing wireless communications users [1]–[4]. In reality, one has no control over the unbalancing of the power lines and the loads connected to them. Hence, the only way to control the CM radiation is to ensure that the power line communication (PLC) modem (the RF source) is properly designed so that the CM current propagating through the power line network is kept at as low a level as possible. With the aim of reducing CM current in the power line network, the authors have initiated a feasibility study of implementing an optical coupler within the analog front-end (AFE) circuit of a PLC modem to reduce the CM current in the power line network. Some researchers [5] have recently conducted their own feasibility study of implementing an optical coupler in the AFE circuit of the PLC modem for broadband usage. The initial study has shown some promising results. However, no systematic analysis and reasoning are provided on how the optical coupler reduces the CM current in the power line network.

This paper aims to carry out an in-depth study on the effectiveness of using an optically coupled PLC modem in reducing the CM current in the power line network. For the study, the

Fig. 1. Equivalent circuit of a PLC modem and a low-voltage distribution network at the power outlet.

equivalent CM noise source impedance of the PLC modem and the unbalanced properties of the indoor low-voltage distribution network (LVDN) of a typical residential unit in Singapore are characterized based on a two-current-probe measurement approach. With the measured results, a CM equivalent circuit model of any PLC modem connected to any power outlet can be established to predict the level of the CM current generated in the power line network. This model would be useful for evaluating the CM suppression performance of a PLC modem with a built-in optical coupler.

II. PLC SYSTEM MODELING

A. CM Current Generation Mechanism

A PLC modem serves as an interfacing device to transfer data over low voltage power network. In a PLC network, the causes of the CM current are the differential mode (DM) RF signal source of the modem and the unbalance of the LVDN. The mechanism of the generation of the CM current can be explained further with Fig. 1.

Fig. 1 shows the equivalent circuit of the PLC modem, which is connected to the power line network through a power outlet. The PLC modem can be represented as a balanced DM RF signal source with impedances Z_1 and Z_2 , and CM impedance Z_3 . The power line network equivalent circuit, as seen by the modem, can be modeled by a T-network consisting of impedances Z_4 , Z_5 , and Z_6 . With this T-network, any unbalances in the power line network can be easily represented and taken into account. From the equivalent circuit shown in Fig. 1, the following equations can be obtained:

$$E_S = I_1(Z_1 + Z_4 + Z_3 + Z_6) - I_2(Z_3 + Z_6) \quad (1)$$

$$E_S = -I_1(Z_3 + Z_6) + I_2(Z_2 + Z_5 + Z_3 + Z_6). \quad (2)$$

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By combining (1) and (2), we have

$$I_1 - I_2 = \frac{(Z_2 - Z_1) + (Z_5 - Z_4)}{(Z_3 + Z_6)(Z_1 + Z_2 + Z_4 + Z_5) + (Z_4 + Z_1)(Z_5 + Z_2)} E_S \quad (3)$$

where $I_{CM} = I_1 - I_2$ is the CM current caused by the unbalance of the power line network. Most telecommunication devices, including PLC modems, are normally designed to be balanced symmetrical sources with $Z_1 = Z_2$. Thus, under such a condition, (3) can be simplified to

$$I_{CM} = \frac{\Delta Z}{(Z_3 + Z_6)(Z_0 + Z_4 + Z_5) + (Z_4 + 1/2 Z_0)(Z_5 + 1/2 Z_0)} E_S \quad (4)$$

where $Z_0 = Z_1 + Z_2$ is the DM impedance of the PLC modem, and $\Delta Z = |Z_5 - Z_4|$ is the unbalance of the LVDN. It can be seen that the factors that affect the magnitude of the CM current are the electrical unbalance of the LVDN ($Z_4 \neq Z_5$), which is responsible for the conversion of the transmitted DM signal E_s into CM voltage, and the CM impedances of the PLC modem Z_3 and the LVDN Z_6 , which are responsible for the conversion of the CM voltage into CM current.

Since the unbalance of the power line network is beyond one's control, the CM current can only be reduced by increasing the CM noise source impedance Z_3 of the PLC modem. To quantify the CM noise source impedance of the PLC modem, a two-current-probe measurement methodology is employed to measure Z_3 .

B. Derivation and Estimation of CM Current

Using the equivalent circuit in Fig. 1, the CM current generated at the signal injection point of the LVDN can be estimated based on the knowledge of the values of the measured Z_3 and the T-network equivalent parameters (Z_4 , Z_5 , and Z_6) of the LVDN. Input impedances Z_{m1} (between live and neutral), Z_{m2} (between live and earth), and Z_{m3} (between neutral and earth) can be measured using the two-current-probe measurement approach at any power outlet of the LVDN, where the PLC modem is connected. With the impedances Z_{m1} , Z_{m2} , and Z_{m3} measured the equivalent T-network model of the LVDN, which consists of Z_4 , Z_5 , and Z_6 , can be derived as follows:

$$Z_4 = \frac{Z_{m1} + Z_{m2} - Z_{m3}}{2} \quad (5)$$

$$Z_5 = \frac{Z_{m1} - Z_{m2} + Z_{m3}}{2} \quad (6)$$

$$Z_6 = \frac{-Z_{m1} + Z_{m2} + Z_{m3}}{2}. \quad (7)$$

III. TWO-CURRENT-PROBE METHODOLOGY

The two-current-probe approach has been employed for DM [6], [7] and CM [8] impedance measurements of the power lines in the frequency range from 20 kHz to 500 MHz. This

Fig. 2. Two-current-probe measurement methodology.

section briefly describes the measurement methodology. Fig. 2 illustrates the basic setup of the two-current-probe approach to measure any unknown impedance Z_X . The measurement system consists of an injecting current probe, a monitoring current probe, and a network analyzer. The two current probes and a decoupling capacitor C form a high-frequency coupling circuit to avoid direct connection to the high-voltage power network. A pair of short wires with length d from reference plane b' to reference plane b connects the coupling circuit to Z_X . Port 1 of the network analyzer induces a continuous-wave (CW) signal in the closed loop through the injecting current probe. Port 2 of the network analyzer measures the resultant current in the closed loop with a monitoring current probe.

The injecting current probe can be represented by an equivalent transformer circuit as shown in Fig. 3. V_{p1} and Z_p are the source voltage and source impedance, respectively, of the injecting signal source from port 1 of the network analyzer. With the injected signal, the injecting probe induces a voltage V_w and results in a current I_w circulating in the closed loop. L_p , L_w , and M are the primary self-inductance of the probe, the self-inductance of the wire in the probe volume, and the mutual inductance between the probe and the wire, respectively. V_{p1} and V_w can be expressed as follows:

$$V_{p1} = (Z_p + j\omega L_p)I_p - j\omega M I_w \quad (8)$$

$$V_w = -j\omega M I_p + j\omega L_w I_w. \quad (9)$$

Combining (8) and (9) and eliminating I_p , we have

$$V_w = Z_{M1} I_w - V_{M1} \quad (10)$$

where

$$Z_{M1} = j\omega L_w + \left(\frac{(\omega M)^2}{Z_p + j\omega L_p} \right) \quad (11)$$

$$V_{M1} = V_{p1} \left(\frac{j\omega M}{Z_p + j\omega L_p} \right). \quad (12)$$

Equations (11) and (12) suggest that the injecting current probe can be represented at the reference plane a by a Thevenin equivalent circuit, as shown in Fig. 4, where Z_{M2} is the reflected impedance in the loop due to the monitoring current probe.

Fig. 3. Equivalent circuit of the injecting current probe.

Fig. 4. Equivalent circuit for the two-current-probe setup.

From (12), the ratio V_{M1}/V_{p1} , which is dependent on the probe properties and the operating frequency ω , is given by

$$K_R = \frac{V_{M1}}{V_{p1}} = \frac{j\omega M}{Z_P + j\omega L_P}. \quad (13)$$

Since the typical wire length d between the reference planes $b'-b$ is much shorter than the wavelength of the maximum frequency of interest (10 m at 30 MHz), its transmission line effect can be ignored. Let Z_{ip} be the impedance at the reference plane b as seen by the unknown impedance Z_X . Then

$$V_{M1} = (Z_{ip} + Z_X)I_W \quad (14)$$

where $Z_{ip} = Z_{M1} + Z_{M2} + Z_C$. Substituting V_{M1} from (14) into (13), Z_X can be evaluated by

$$Z_X = (K_R Z_{T2}) \left(\frac{V_{p1}}{V_{p2}} \right) - Z_{ip} \quad (15)$$

where $Z_{T2} = [V_{p2}/I_W]$ is the transfer impedance of the monitoring current probe, and V_{p2} is the voltage received by the monitoring probe. The ratio V_{p1}/V_{p2} can be obtained through S-parameters measurement using the network analyzer as follows:

$$\frac{V_{p1}}{V_{p2}} = \frac{S_{11} + 1}{S_{21}}. \quad (16)$$

The product $K_R Z_{T2}$ is a frequency-dependent coefficient that can be obtained by firstly removing Z_X and measuring Z_{ip} using an impedance analyzer. Based on the measured characteristics, Z_{ip} can be modeled as a resistor of 1.13 Ω , an inductor of 80.87 nH, and a capacitor of 0.09 μF connected in series. Then, Z_X is replaced with a known precision standard resistor R_{std} (100 Ω , carbon film $\pm 1\%$) and V_{p2}/V_{p1} is measured again with the network analyzer. Finally, $K_R Z_{T2}$ can be obtained by

$$K_R Z_{T2} = \frac{Z_{ip} + Z_{std}}{(V_{p1}/V_{p2})|_{Z_X=Z_{std}}}. \quad (17)$$

Once $K_R Z_{T2}$ and Z_{ip} are determined, the two-current-probe setup is ready to determine any unknown impedance Z_X using (15) and (16).

IV. DEVELOPMENT OF PLC TEST BOARD

A. Optically Coupled AFE

For the feasibility test, a PLC test board using an optically coupled AFE circuit is fabricated. Fig. 5 shows the block diagram of the optically coupled PLC test board. The test board performs PLC interface functions similar to the AFE circuit in a PLC modem, which provides electrical isolation and signal amplification for communication purposes. The test board is equipped with a test signal source E_S that generates periodic pulses of trapezoidal waveform with frequency of 1.063 MHz, rise and fall times of 3.11 ns, pulse width of 470.02 ns, and peak amplitude of 4.46 V. The signal source produces harmonics that sweep across the PLC operating frequency of 1–30 MHz.

A fiber optic link is used in the experiment to form an optical coupler for the test board. A short length of glass fiber optic cable with core/cladding diameter of 62.5/125 μm , specified for use with the fiber optic transmitter and receiver, is chosen as the transmission link, which allows the signal generator to be electrically isolated from the amplifier circuit. Jumper (JP) is implemented in the test board to allow the test personnel to enable or disable the optical coupling feature of the test board.

B. CM Impedance Measurement

Fig. 6 shows the setup for the measurement of the CM source impedance Z_3 of the test board using the two-current-probe approach. The Tektronic CT-1 (5 mV/mA, bandwidth 25 kHz to 1000 MHz) and CT-2 (1 mV/mA, bandwidth 1.2 kHz to 700 MHz) current probes are chosen as the injecting and monitoring current probes, respectively, while the Agilent 4395A Network Analyzer is employed for the S-parameters measurements. The two current probes and two coupling “Y” class 0.1- μF capacitors (one for phase to earth and the other for neutral to earth) form the CM coupling circuit to avoid any direct connection to the power mains. To ensure repeatability of the measurement, the coupling signal from the PLC test board is terminated with a line impedance stabilization network (LISN) as defined by the standards [10]–[12]. To facilitate the measurement of the CM noise source impedance of the PLC test board, a CM choke with suitable value is inserted between the LISN and the two-probe coupling circuit to provide sufficient CM isolation between the coupling circuit and the LISN, so that only the CM impedance of the PLC test board is being measured by the two-current-probe setup. As the rated current of the test board is 680 mA, a CM choke with current rating of 3 A is chosen to avoid possible core saturation. The measured CM choke inductance is 3.2 mH, which provides sufficiently high impedance across the measurement frequency range. Once the CM impedance of the test board is measured, the CM choke is removed to allow the propagation of the CM current. Another clamp-on type RF current probe is clamped onto the power line cables to measure the CM current.

Fig. 5. Block diagram of an optically coupled PLC test board.

Fig. 6. Two-current-probe setup for the measurement of CM source impedance Z_3 of the test board.

Two CM source impedance measurements are carried out. The first measurement is conducted with JP connected, which bypasses the optical coupler. The source signal from the test board is now coupled onto the power line network through the isolation transformer (which is the usual practice for most commercially available PLC modems). Fig. 7 shows that the measured CM impedance of the test board is capacitive in nature with an estimated capacitance of 48 pF, which is contributed primarily by the parasitic capacitance between the primary and secondary windings of the isolation transformer. In the second measurement, JP is removed to activate the additional electrical isolation provided by the optical coupler. The measured result shown in Fig. 7 indicates an increase in CM impedance, which is still very much capacitive in nature with an estimated capacitance of 17 pF, and which is contributed mainly by the parasitic capacitances between the amplifier circuits and the chassis ground [9] in the power line side of the test board. Thus, the optical coupler has effectively increased the CM impedance of the test board due to much lower parasitic capacitance in the CM signal path.

Fig. 8 shows the measured CM current in the frequency domain. The measurement results clearly indicate that the optical coupler in the test board has reduced the CM current level by nearly 10–30 dB up to 30 MHz. The reduction has brought the CM current below the CISPR 22 limit [12].

Fig. 7. CM source impedance Z_3 before and after the optical coupler is enabled.

Fig. 8. CM current before and after the optical coupler is enabled.

V. ESTIMATION OF CM CURRENT

To quantify the performance of the optical coupler in an actual power line environment, an equivalent circuit that represents the power line network is needed so that the CM current generated

Fig. 9. Two-current-probe setup for the measurement of input impedances Z_{m1} , Z_{m2} , and Z_{m3} of the network under test.

by a PLC modem with a built-in optical coupler can be estimated with reasonable accuracy.

A. Validation of CM Current Estimation Method

Based on the two-current-probe measurement approach, the input impedances Z_{m1} , Z_{m2} , and Z_{m3} of the LISN are measured, as shown in Fig. 9, and converted to the equivalent T-network parameters Z_4 , Z_5 , and Z_6 , using (5)–(7), respectively. The T-equivalent model is simply derived by fitting equivalent resonant circuits [13], [14] to the measured T-network parameters. Since the model parameters depend mainly on measurement results, the model derived would be able to estimate the CM current generated due to the unbalance of the LISN or any ac mains accurately.

With the known transmitted signal E_S , DM impedance Z_0 , and CM source impedance Z_3 of the PLC modem, the CM current generated by the PLC modem can be calculated. For validation purpose, the same PLC test board with optical coupler is employed to emulate a PLC modem. Based on the previous characterization of the optically coupled PLC test board, the CM source impedance Z_3 of the test board can be represented by a capacitance of 17 pF. The DM impedance Z_0 of the PLC modem can be represented as a 10- Ω resistor connected in series with a 1- μH inductor. The full equivalent circuit model of the optically coupled PLC test board terminated by the LISN is shown in Fig. 10. Fig. 11 compares the measured CM current generated by the PLC test board and the calculated CM current based on the equivalent electrical model given in Fig. 10. Close agreement between the measured and the calculated CM currents is demonstrated.

B. Estimation of CM Current in Actual AC Mains Line

In this section, the CM current generated by the same optically coupled PLC test board in the LVDN of a four-room residential unit in Singapore is estimated. Fig. 12 illustrates the typical layout of the residential unit in Singapore. The equivalent electrical models of three different power outlets at points A–C, as indicated in the layout, are established using the two-current-probe measurement. All of the power outlets in this residential unit belong to a single-phase network.

Fig. 10. RF equivalent circuit of the PLC test board and the LISN.

Fig. 11. Comparison of the measured and calculated CM currents.

Fig. 12. Measurement points for input impedances of indoor ac mains line in a four-room residential unit.

Based on the two-current-probe technique, the T-network electrical models at power outlets A–C are developed. The CM currents due to the PLC test board at the three outlets are then estimated. Fig. 13 shows the estimated CM currents generated at the three different power outlets. It clearly indicates that the CM current at power outlet C is found to be higher than those of the other two power outlets A and B.

Fig. 13. Estimated CM currents at the indoor power outlets A–C.

VI. CONCLUSION

Based on the preliminary measurement results of a PLC test board, the authors have shown that the inherent low-parasitic capacitance of the optical coupler provides much higher CM impedance over the conventional transformer coupler, where the winding-to-winding capacitance is generally higher. Such an inherent feature is desirable because it results in superior CM attenuation and therefore, lowers the CM current in the power network. Since the unbalanced characteristic of the power line network varies from place to place, a T-network electrical model is derived based on a two-current-probe approach to reflect such variation. With the same two-current-probe approach, the source model of any PLC modem could also be developed. The full equivalent circuit model, which emulates the connection between the PLC modem and the power line network, allows one to quantify the CM suppression performance of a built-in optical coupler. Further research work will be carried out to reduce the parasitic capacitance of the optical coupler. Also, more measurements will be conducted at different residential and commercial buildings to obtain a good statistical power line network model in Singapore. These additional efforts would allow more conclusive findings on using optical coupling technique in PLC modems to suppress CM current in the power line network.

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