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A Current-Comparator-Based HV R-L-C Bridge

Eddy So, *Fellow, IEEE*, and David Bennett

***Abstract*—The development of a current-comparator-based high-voltage impedance bridge at the National Research Council of Canada (NRC) for the accurate measurements of high-voltage resistive (R), Inductive (L), and capacitive (C) impedances is described. The bridge has ratio uncertainties of les than 10 x 10^{-6} in both magnitude and phase.**

***Index Terms*—Accurate measurements, capacitive, current-comparator-based bridge, high voltage, inductive impedance, resistive, uncertainty.**

I. Introduction

Transformer ratio bridges may be configured as voltage or current ratio bridges. In the voltage bridge, the transformer is used to determine the ratio of the voltages across the other two arms when the same current passes through them. In the current bridge the transformer is used to determine the ratio of the currents in the other two arms when the same voltage is applied to them. The voltage bridge operates with a finite flux in the magnetic core while in the current bridge the flux in the core is ideally at zero. The transformers in precision bridges are usually of the three-winding type, the extra winding being used to excite the core in the voltage bridge or to detect the zero flux, and hence the ampere-turn balance condition, in the current bridge.

Fundamentally there is no basic difference between the two types of bridge. One can be obtained from the other merely by exchanging source for detector. However in the voltage bridge there are limits on the flux magnitude owing to core saturation and this places some constraints on core size, the numbers of turns, and the operating frequency. For use with high-voltage impedances, the current bridge is essential.

Transformer-ratio-arm bridges of the current-comparator-type are capable of ratio accuracies of better than 10×10^{-6} for both magnitude and phase. Such bridges are therefore most suitable for accurate power loss measurements at very low power factors of less than 1 percent or 0.01. They are usually used as a high-voltage capacitance bridge to measure the losses of capacitive loads, such as high voltage capacitors and power cables.

This high-voltage capacitance bridge may be also adapted to the measurement of losses in inductive loads, such as shunt reactors and power transformers on short circuit test condition, by providing the means to reverse the phase of the inductive current and to measure what is essentially a negative tanδ balance [1]. Although, the bridge has high ratio accuracy, under certain test conditions when used for measuring low power factor losses in loads, the measurement results could be inaccurate/erroneous. To address this issue, a special current-comparator-based inductance bridge was developed at the National Research Council of Canada (NRC). Current-comparator-based (CCB) bridges have been developed at NRC, especially a low-voltage resistance bridge [2], a high-voltage capacitance bridge [3], and a high-voltage inductance bridge [4]. This paper describes an overview of the development of a current-comparator-based high-voltage impedance bridge for accurate measurements of high-voltage resistive, inductive, and capacitive impedances, that is a CCB high-voltage (HV) R-L-C bridge, which is an integration of previous CCB bridges in one CCB HV R-L-C Bridge.

II. CCB HV R-L-C Bridge

The basic circuit diagram of the CCB HV R-L-C Bridge is shown in Figure 1. It consists of a high-voltage low-loss gas-dielectric reference capacitor C_H, a low-voltage arm of a CCB HV Divider DIV followed by a unity-gain CCB Integrator INT, reference resistors R_{S1} and R_{S2}, a three-winding current comparator of N_X, N_{S1}, and N_{S2}, including a compensation and a detection winding (not shown) to detect an ampere-turn balance condition of the bridge. The current comparator is similar in design as that described in [1]. It is designed for operation with a current rating of 1 ampere-turn for the current comparator windings. In the actual bridge, the equivalent of six-digit resolution is obtained in the N_{S1} and N_{S2} windings with each winding having a total of 100 turns. These two windings have been designed to have a special feature that the

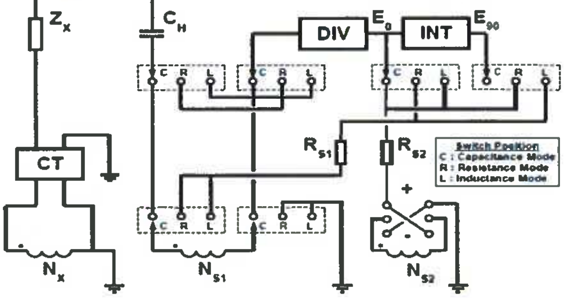

Fig.1. CCB HV R-L-C Bridge

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switches changing the six-digit of each winding could be mechanically coupled such that when a digit is switched in the N_{S1} winding, the same digit in the N_{S2} winding will automatically follow the same switch position as that of the N_{S1} winding. The N_X winding is configured for ratio multiplication, having taps at 1, 2. 5, 10, 20. 50 and 100 turns, allowing current operations at 1 A, 0.5 A, 0.2 A, 0.1 A, 0.05 A, 0.02 A, and 0.01 A, respectively. To measure large load/impedance currents, a two-stage reference current transformer CT can be cascaded to the current comparator winding N_X, as shown in Fig. 1.

Figure 2 shows the low voltage arm of the CCB HV divider DIV connected to the HV capacitor C_H. Capacitors C_H and C_L are high- and low-voltage low-loss gas-dielectric capacitors, respectively. The current comparator is used in a feedback loop to correct the magnitude and phase of the output voltage E_0 to an accuracy value of better than 10×10^{-6} and 10 μrad, respectively [5]. The ratio windings N_1 and N_2 have 100 turns and 1326 turns, respectively. Winding N_1 is subdivided to yield overall ratio multipliers of 1326/5, 1326/10, 1326/25, 1326/50, and 1326/100, corresponding to gain settings G of 1, 2, 5, 10, and 20 of amplifier A_2. The switches changing the winding ratio and the gain of amplifler A_2 are mechanically coupled to keep the current comparator in ampere-turn balance condition with a change in the gain of amplifier A_2.

The compensation winding N_3 also has 1326 turns and is connected in parallel with winding N_2 to reduce its leakage impedance. A 100-turn detection winding N_D is connected to a current-to-voltage converter to obtain a voltage proportional to, and in phase with, the unbalanced ampere-turns in the current comparator.

The capacitance of the low voltage gas-dielectric capacitor and that of the solid dielectric feedback capacitor are 1000 pF and approximately 0.265 μF respectively. The capacitance values of the low voltage gas-dielectric capacitor and that of the solid dielectric capacitor have been chosen such that at ampere-turn balance of the current comparator, the output voltage of the CCB low voltage arm DIV would be a nominal voltage of 100 V at a gain setting G corresponding to $(G) \times (E_H \times \omega \times C_H) \times 10$ mA and $(N_2/N_1) \times (G) \times 1326/5$, where ω is the angular frequency at 60 Hz. Thus, the effective capacitance of the feedback capacitor C_f corresponding to an ampere-turn balance in the current comparator, is 265200 pF. The output voltage E_0 can then be expressed as

$$E_0 = (C_H \times G \times E_H)/265200 \qquad (1)$$

where the capacitance of the low voltage gas-dielectric capacitor is not exactly 1000pF, the effective capacitance of C_f becomes $265200 \times C_L/1000$pF.

The unity-gain CCB Integrator INT circuit is basically a CCB HV divider circuit with a reference resistor R_S of nominal value of 10 kΩ in place of a HV capacitor C_H, as shown in Fig. 3. Thus, for a sinusoidal input voltage E_0, the output E_{90} of the integrator INT is a voltage signal shifted by 90 degrees with respect to the output voltage E_0 of the divider DIV and has a frequency characteristic of an inductor as shown in (2):

$$E_{90} = -jE_0 / (\omega \times C_f \times R_S) \qquad (2)$$

With an effective feedback capacitor C_f of 265200 pF, equivalent to a nominal capacitive reactance of 10 kΩ at a frequency of 60 Hz, a nominal unity gain of the integrator INT is achieved. The integrator may be designed to have adjustable gain; in this case, the equation above is modified to reflect this.

Subsequent discussion of the bridge differentiate between the feedback capacitor of the divider and that of the integrator as $C_{f(DIV)}$ and $C_{f(INT)}$, respectively.

The voltage signal E_{90} then drives a voltage-to-current converter, which is a simple reference resistor R_{S1} that would provide a reference inductive current [3] to winding N_{S1} in the *Inductance Mode*. See section "MODE OF OPERATION".

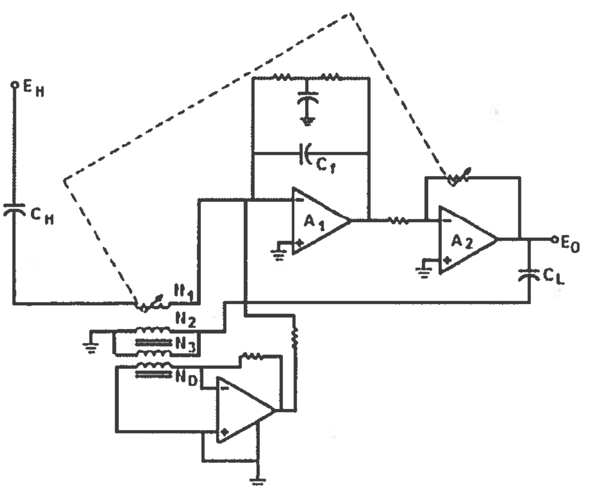

Fig. 2

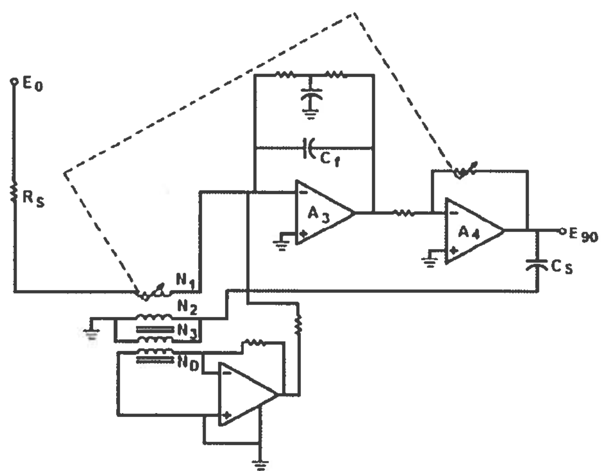

Fig. 3

Typically, the reference resistors R_{S1} and R_{S2} have resistances of 10 kΩ each. Optionally, a resistance of 100 kΩ may be selected for R_{S2} where the phase characteristics of the device under test (e.g. DF, Power Factor, etc.) are small. Elimination of the effects of lead and winding impedance at the low-voltage end of R_{S1} and R_{S2} is accomplished using lead compensation circuitry [6].

III. MODE OF OPERATION

Depending on the switch position, the bridge can function as either a high voltage resistance bridge, or a high voltage inductance bridge, or a high voltage capacitance bridge.

In practice, when measuring an unknown impedance, especially when used for high voltage applications, there are preferences, with regard to whether the equivalent resistive and reactive circuit parameters are in series or parallel. In the case of high voltage resistors and high value resistances, equivalent parallel resistance and reactance are preferred since high value resistors tend to have capacitive leakage currents in parallel with the resistance. For high voltage inductors and reactors, due to their large inductance and small resistance, the preference is to use the equivalent circuit of series combination of the inductor and resistor. For high voltage capacitors, the preference would be for equivalent series combination of the capacitor and a resistor.

The balance equations for the bridge in each mode are derived from the equivalent parallel parameters of parallel conductance G_X and susceptance B_X of the unknown impedance. From these, the series circuit parameters may be derived.

Table 1 outlines some different measurement applications and the RLC bridge mode selected, and a summary of impedance equations to derive different equivalent series or parallel circuit parameters from G_P and B_P, when using the bridge.

A. Capacitance Mode

In the capacitance mode, the switch is in the "C" position, showing the high voltage reference capacitor C_H connected to the low voltage arm of the CCB Divider DIV through the variable current comparator winding N_{S1}, providing the in-phase balance of the unknown capacitive impedance C_X. The output of DIV is connected to the variable current comparator winding N_{S2} through the reference resistor R_{S2}, providing the quadrature (dissipation factor) balance of the unknown capacitive impedance C_X, as shown in Fig. 4. An alternate connection is also shown in Fig. 4, which is to have an inductive divider IVD connected to the output of DIV.

The balance equations for the capacitance mode, solving for G_P and B_P are:

$$G_P = (C_H \cdot G \cdot N_{S2})/(C_{f(DIV)} \cdot R_{S2} \cdot N_X) \quad (2)$$

$$B_P = (\omega \cdot C_H \cdot N_{S1})/N_X \quad (3)$$

Table I
Summary of RLC Bridge Modes and Impedance Equations Used for Different Measurements

Measurement	Mode	Equivalent Circuit
High Voltage Capacitors	Capacitance (C)	$Z_S = R_S - jX_S$
High Voltage Inductors and Reactors	Inductance (L)	$Z_S = R_S + jX_S$
High Voltage Resistors and High Value Resistances	Resistance (R)	$Z_P = R_P \pm jX_P$
Summary of Impedance Equations		
Series Impedance (Ω)	Z_S	$R_S \pm jX_S$
Parallel Impedance (Ω)	Z_P	$R_P \pm jX_P$
Series Resistance (Ω)	R_S	$G_P / (G_P^2 + B_P^2)$
Series Reactance (Ω)	X_S	$- B_P / (G_P^2 + B_P^2)$
Parallel Resistance (Ω)	R_P	$1 / G_P$
Parallel Reactance (Ω)	X_P	$- 1 / B_P$
Dissipation Factor (tan δ)	DF	(B_P / G_P)
Power Factor (cos □)	PF	$DF / (1 + DF^2)^{0.5}$

From the bridge balance equations, it could be shown that, for an equivalent parallel circuit,

$$C_P = C_H (N_{S1}/N_X) \quad (4)$$

$$DF = G_P/\omega C_P = (N_{S2}/N_{S1})(G/\omega C_f R_{S2}) \quad (5)$$

Where C_P is the equivalent parallel capacitance of C_X, DF is the dissipation factor of C_X, and G_P is the equivalent parallel conductance of C_X, representing its loss current component, and ω is the angular frequency. If divider gain G = 1 and the test frequency is 60 Hz, then since R_{S2} = 10 kΩ and $C_{f(DIV)}$ is equivalent to 265200 pF, which at a frequency of 60 Hz, provides a nominal capacitive reactance of 10 kΩ, then $(1/\omega C_f R_{S2}) \approx 1$, and Equation (5) becomes

$$DF = (N_{S2}/N_{S1})(60/\mathit{test\_frequency}) \quad (6)$$

Fig. 4 also shows an alternate configuration with an inductive divider IVD connected to the output of the low voltage arm of the divider DIV and the windings N_{S1} and N_{S2} are mechanically coupled. With divider gain G = 1, and

R_{S2} = 10 kΩ, the balance setting of the IVD becomes direct reading in DF at 60 Hz.

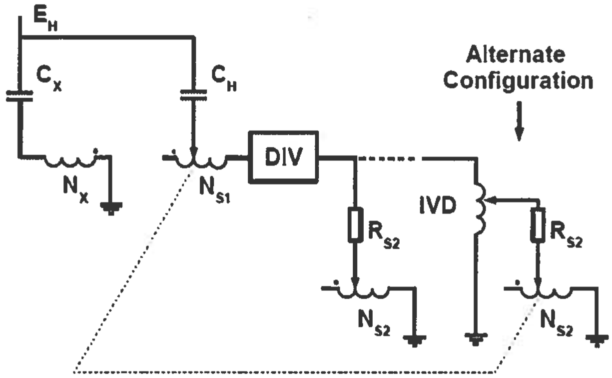

Fig. 4

B. Inductance Mode

In the inductance mode, the switch is in the "L" position, showing the high voltage reference capacitor C_H connected to the low voltage arm of the CCB Divider DIV, bypassing the variable current comparator winding N_{S1}. The output E_{90} of the integrator INT is connected to the variable current comparator winding N_{S1} through the reference resistor R_{S1}, providing the in-phase balance of the unknown inductive impedance. The output E_0 of the divider DIV is then connected to the current comparator winding N_{S2} through the reference resistor R_{S2}, providing the quadrature (loss angle) balance of the unknown inductive impedance L_X, as shown in Fig. 5.

The balance equation for the inductance mode, solving for G_P and B_P are:

$$G_P = (C_H \cdot G \cdot N_{S2})/(C_{f(DIV)} \cdot R_{S2} \cdot N_X) \quad (7)$$

$$B_P = (C_H \cdot G \cdot N_{S1})/(\square \cdot C_{f(DIV)} \cdot C_{f(INT)} \cdot R_S \cdot R_{S1} \cdot N_X) \quad (8)$$

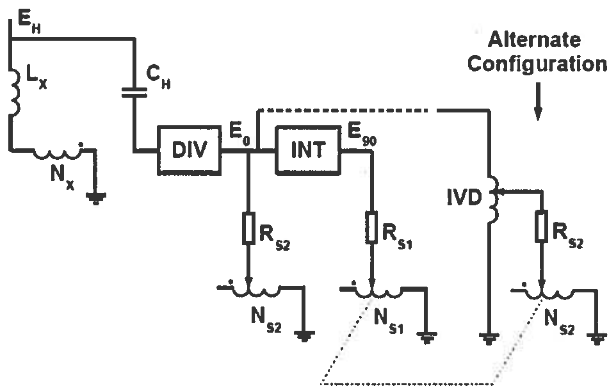

Fig. 5

C. Resistance Mode

In the resistance mode, the switch is in the "R" position, showing the high voltage reference capacitor C_H connected to the low voltage arm of the CCB Divider DIV bypassing the variable current comparator winding N_{S1}. Instead the output E_0 of DIV is then connected to the current comparator winding N_{S1} through the reference resistor R_{S1}, providing the in-phase balance of the unknown resistive impedance. The output E_{90} of the integrator INT is connected to the variable current comparator winding N_{S2} through the reference resistor R_{S2}, providing the quadrature (dissipation factor) balance of the unknown resistive impedance R_X, as shown in Fig. 6.

The balance equation for the resistance mode, solving for G_P and B_P are:

$$G_P = (C_H \cdot G \cdot N_{S1})/(C_{f(DIV)} \cdot R_{S1} \cdot N_X) \quad (9)$$

$$B_P = (C_H \cdot G \cdot N_{S2})/(\square \cdot C_{f(DIV)} \cdot C_{f(INT)} \cdot R_S \cdot R_{S2} \cdot N_X) \quad (10)$$

Where the N_{S1} and N_{S2} windings not mechanically coupled, and R_{S1} = R_{S2}, there is essentially no difference electrically, between the Inductance Mode and Resistance mode, and it can be seen that the equations are similar.

For high value resistances having capacitive leakage currents in parallel with the resistance, N_{S2} is reversed. However, in situations where the capacitive leakage currents are very large and the frequency of E_H is not stable, it may be necessary to measure high voltage resistors with the bridge in capacitance mode to minimize fluctuations in bridge readings.

In the "*Resistance Mode*" the bridge has two modes of operations. That is: (1) direct reading in conductance and equivalent inductive/capacitive phase defect by connecting the unknown resistor R_X to the N_X winding and the reference resistor R_{S1} to the adjustable N_{S1} winding, as shown in Fig. 6; and (2) direct reading in resistance and equivalent capacitive/inductive phase defect by connecting the unknown resistor R_X to the adjustable N_{S1} winding and the reference resistor R_{S1} to the N_X winding.

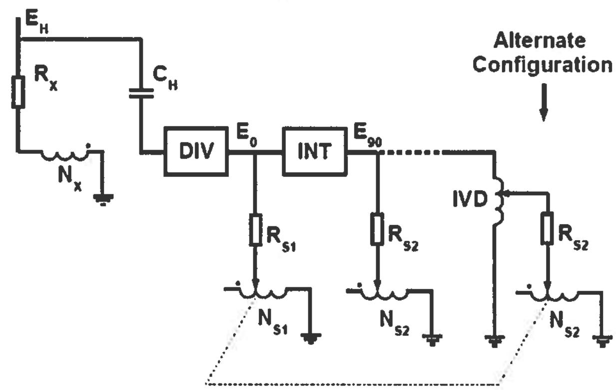

Fig. 6

IV. ERFORMANCE

The uncertainty of the measurements performed by the CCB HV R-L-C Bridge is determined by the performance characteristics of its main components of the CCB HV Divider DIV, which are the high-voltage reference capacitor, the integrator INT, the current comparator, including the ratio range extender CT, and the reference resistors. These main components are calibrated with known calibration techniques as described in [1].

Therefore, the overall magnitude and phase errors of the components of the bridge are known from their respective calibrations. These known errors are taken into account by either applying corrections or offsetting the actual turns of the ratio windings. The remaining uncertainties, consequently, are only due to the calibration uncertainties.

The overall uncertainties of the CCB HV R-L-C Bridge at each mode of operation are estimated to be less than 10×10^{-6} in both magnitude and phase for power frequencies.

The bridge operating voltage and current is determined by the voltage rating of the high voltage reference capacitor and current rating of the ratio range extender CT, respectively. Thus, the operating voltage could be up to 500 kV or higher, provided that the output voltage of divider DIV will be not more than 100 V. The operating current could be up to 1000 A or higher, provided that the secondary currrent output will not be more than 1 A. This in turn with the reference resistors will determine the measurement range of the impedance R, L, and C.

V. CONCLUSIONS

An overview of the development of a CCB HV R-L-C Bridge with an overall estimated uncertainty of less than 10×10^{-6} in both magnitude and phase for power frequencies is described.

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BIOGRAPHY

Eddy So (M'74-SM'84-F'90) received the M.Sc. and D.Sc. degrees in Electrical Engineering from George Washington University, Washington, D.C., USA. In 1977, he joined the National Research Council of Canada, Ottawa, Ontario, Canada. In 1991-2004 he was the Director of the Electromagnetic and Temperature Standards Section in the Institute for National Measurement Standards, where he is currently a Principle Research Officer and Leader of the High Voltage Power and Energy Measurements Project. His research interest includes the development of measurement techniques and instrumentation for accurate measurements of active/reactive power and energy under difficult operating conditions, and for assessing the operating conditions of different types of high voltage insulation. In 1979-1989 he was Adjunct Professor at the University of Ottawa and Carlton University. In 2002-2008, he was Chair of the Conference on Precision Electromagnetic Measurements (CPEM) Executive Committee. He is Past Chair of the IEEE Power Systems Instrumentation and Measurements Technical Committee, Power Engineering Society, Chair of its Subcommittee on Electricity Metering, Chair of its Working Group on Low-Power-Factor Power Measurements, and is also its Standards Coordinator. He is also a Registered Professional Engineer in the Province of Ontario. He is a Fellow of IEEE.

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