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Validating the power loss model of a transformer by measurement

# VALIDATION IS KEY

BY JEAN-PIERRE KERADEC

**T**ODAY, WHEN RUNNING CIRCUIT simulations to forecast behavior and power losses, the design engineer has, for involved components, several models at his disposal. Of course, the accuracy of the result is bound to that of component models and replacing one model by another leads sometime to large discrepancies. In such a case, the engineer often relies on measurements to choose between different models.

As soon as measured values agree with the estimates of one of a model's simulations, this one model—and all the individual models it includes—is considered correct. Is it so certain? To settle this matter, it is necessary to answer

two questions. First, are these measurements accurate enough to adopt one explanation and reject others? Second, considering all the measurements carried out, is there only one explanation that suits?

Because the main characteristic of power electronic circuits is their high power efficiency, lost power is always far smaller than average transferred power, which is the only one that can be directly measured using electronic apparatus. For this reason, in power electronics, loss measurements—either of whole systems or of single components—are often very difficult. As an example, measuring the instantaneous power dissipated in an insulated gate bipolar transistor (IGBT) working in a switch-mode

power supply seems easy; it is just the product of a voltage and a current. However, we spent seven years mastering this measurement [1]–[4].

Calorimetric measurements offer an attractive alternative to electronic ones because they directly access lost power. Unfortunately, they require a calorimeter, an apparatus that is neither ready to use nor fast in action. Accurate results are obtainable provided that impact of every physical property of the calorimeter has been correctly evaluated.

With all these difficulties in mind, it is clear that measurement is not automatically synonymous with truth. Obtaining reliable measured values requires meticulous attention, and this is often time consuming. A useful overview of different techniques suited for conducting power loss measurements is given in [5].

Our first example of power loss evaluation relies on the abovementioned work; it uses oscilloscopic measurements. During the study of a battery charger intended for electrical vehicles, we focused on a special 2.5-kW transformer. After its equivalent circuit was elaborated, we introduced it in a circuit simulation to evaluate its working power efficiency. However, equivalent circuit elaboration relied on various hypotheses, among which was the linear behavior of the component. This is why we decided to experimentally check the transformer power loss previsions using oscilloscopic measurements. Here we will show how this goal has been reached. We will list apparatus chosen for this purpose and we will detail experimental care that led to the required accuracy.

The second example illustrates the potential of calorimetric measurements. Some years ago, we were investigating high-frequency copper losses in transformers, especially skin and proximity effects. It is now well known that these effects both lead, on the high-frequency side, to series resistances proportional to the square root of the frequency. To represent this behavior in a way usable with circuit simulation software, we introduced a lumped element equivalent circuit and we asked: how can we validate it? Short-circuit impedance gives information about these phenomena but, on the high-frequency side, the real part of this impedance, which is the only one involved in losses, becomes masked by its imaginary part.

At the same time, we got the opportunity to use a calorimeter to carry out related loss measurements. Because these measurements directly access the losses, the imaginary part of the impedance does not matter and the accuracy is sometimes far better than that of the best impedance measurements. However, building a calorimeter is a specialist's task, calorimetric measurements are time consuming (weeks) and, during these experiments, the component is not in a natural cooling situation. We will detail the benefits and drawbacks of this type of measurement.

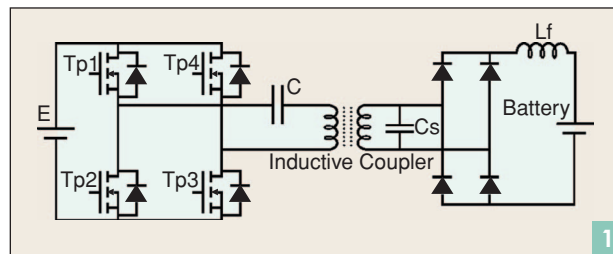
## Outstanding Accuracy for Oscilloscopic Power Measurements

### Description of the Tested Device

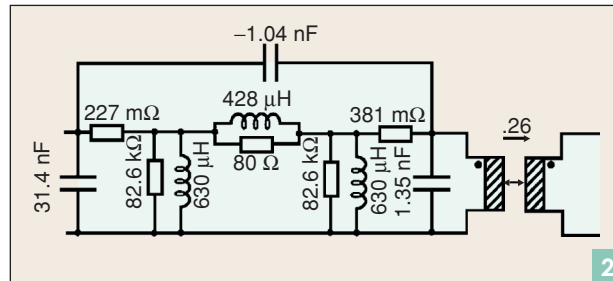
In 1998, we studied a 3-kW contact less charging system intended for an electrical vehicle [6]. After having

selected the circuit topology (Figure 1) to be used, we designed the transformer. Because of its availability, a Siemens PM114/93 magnetic core made of N27 ferrite was chosen to build our first prototype, despite the fact that it can sustain only 2.5 kW with a strong air flow cooling. According to an application constraint, a large air gap (5 mm) was introduced between its two halves.

To obtain simulation results that are as accurate as possible, we identified the equivalent circuit of the transformer prototype (Figure 2) following a method which has already been described [7], [8]. This identification, still used today, relies only on Bode plots of a set of measured impedances, and no extra information regarding size, turn numbers, physical data, and technology is required. Impedances have been measured with an HP4194A impedance analyzer, and, as a first check, measured curves were compared successfully up to 10 MHz to those deduced from the equivalent circuit.



Circuit structure of the contactless battery charging.



Equivalent circuit of the transformer.

A new set of circuit simulations, associated with different working frequencies, have been carried out for the entire circuit, which now includes the above representation of the transformer. First, as shown in Figure 3, all simulated waveforms agree well with observed ones. Notice that the current shown in Figure 3 becomes more and more distorted as working frequency grows.

Values shown in Table 1 are deduced from PSpice simulations using, for the transformer, the previously presented model (Figure 2). They are obtained at different working frequencies, and transformer power losses include all kinds of losses (series and parallel). These values teach us that global power efficiency of the converter is rather good and losses are mainly localized in the transformer.

### Measuring Apparatus and Care

Simulated results seem to be interesting, but they rely on a transformer equivalent circuit that is deduced from

**TABLE 1. INPUT AND OUTPUT CHARGER POWERS AND TRANSFORMER LOSSES.**

Working Frequency (kHz)	Input Power (W)	Output Power (W)	Transformer Losses (W)
28	1,126	960	136
27	1,286	1,040	106
26	1,255	1,110	100
21,4	2,574	2,300	224

**TABLE 2. TRANSFORMER LOSSES, MEASURED COMPARED TO SIMULATED VALUES.**

Working Frequency (kHz)	Measured Losses (W)	Simulated Losses (W)	Relative Deviation (%)
28	145	136	6,6
27	115	106	8,8
26	91	100	9,7
21,4	231	224	3,6

low-level measurements. To gain confidence in simulated results, it is desirable to measure transformer losses using our oscilloscopic measurement bench.

Because we want to check our transformer equivalent circuit, which is deduced from room temperature measurements, we kept transformer temperature close to room temperature; an extra fan was added to cool the transformer during operation in the converter and measurements were taken quickly after converter start.

#### Equipment Usually Used

In this bench, signals are acquired by Tektronix equipment: a DSA602 oscilloscope associated to various high-impedance voltage probes and active current probes. This oscilloscope is linked to a computer, and homemade software reminds the user how to proceed and helps him in applying some special calibration to the probes.

The selection of suitable apparatus resulted in an extensive market availability study. However, specifically when performing apparatus is required, numerous precautions must be taken to reach the finest accuracy. We now describe how we proceeded in this goal.

#### Usual Care

First, as reported in previous papers [3], [4], [9], let us remind readers of some of the precautions we generally take to obtain accurate results.

- To increase the equivalent resolution of the analog-to-digital converter (ADC) and the signal-to-noise (SNR) ratio, we always use averaging mode (64 acquisitions or more).
- Each signal is acquired in a two-step process. The first step is the normal acquisition. The signal is acquired, along with a certain amount of noise. Averaging lowers random noise but not perturbations

generated by the working of the studied device itself. To remove this coherent noise as well as an eventual dc offset, these parasitic signals are recorded separately during the second step. During this step, probes are moved to acquire the zero references; the voltage probe is connected to ground and the current probe is closed beside the wire. Subtracting this acquisition from the previous one results in far cleaner signals.

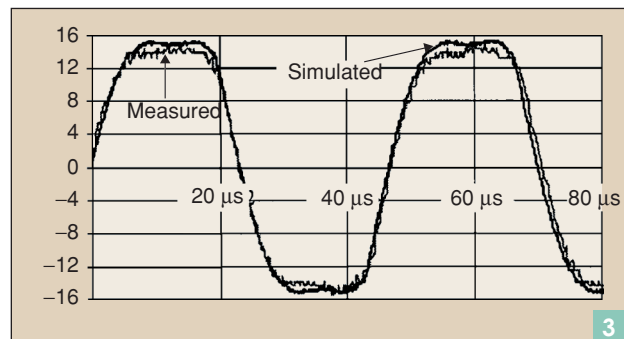
- To efficiently acquire zero references, it is mandatory to keep conditions as close as possible to previous ones. Don't touch anything except probes, and move them as little as possible. Have an external synchronization signal available during both steps. Connect the voltage probe to the closest ground without moving its low terminal and avoid a change in probe ground loop area and orientation. Place the current probe close to the wire with the same orientation as before (magnetic fields!).

A current transformer is often needed to adjust sensitivity. It is placed so as to be crossed by the current to measure, and the current probe embraces its secondary wire. In that situation, sensitivity to external field depends on the probe, the current transformer, and their orientations. To remove the noise due to that field, care must be taken to maintain the same sensitivity during the second step. To reach this goal, place, from the beginning, an identical current transformer not crossed by the current to measure, close to the former, with its axis orientated parallel to that of the former. Finally, simply move the current probe from one secondary to the other.

- For all our couples of Tektronix probes, we have measured the differential time delay, and we corrected it, within 1 ns, after the acquisition. The original deskew procedure is perfect for equal bandwidth probes but, when a 400-MHz voltage probe is used together with a 50-Hz current probe, a time lag of about 4 ns remains [10].

#### Special Choice of Probes and Precautions

To investigate the device presented here, we first choose appropriate probes. Transformer windings are both floating with respect to main ground so, due to common-mode currents, converter working can change and a slight shift in frequency may occur when connecting the voltage probe. To avoid this, we used optical insulation amplifier



**Primary current waveform: measured and simulated (current is in A, time in  $\mu$ s).**

Tek A6906S for voltage waveform acquisition and we linked together the two low terminals of the two windings (the transformer was characterized with this link).

To catch current waveforms, a couple of identical current transformers (dividing ratio: 1/30), only one of them having been gone through by the current, have been installed on each side of the transformer. Finally, current waveforms were input into the 11A16 amplifier of the oscilloscope through a Tek A6302 current probe.

At the beginning of the measurements, we observed that reference signals were far from quiet. So we first lowered magnetic field emissions by reducing all current loop area inside and in the vicinity of the converter. With the same goal, we also changed transformer orientation. To complete, we kept the A6906S receiver far enough from magnetic sources.

Thanks to the precautions described above, primary measurements appear as shown in Figure 4. Waveforms at the secondary side were acquired in the same way. Last but not the least, despite the fact that the transformer exhibits a 1/4 ratio, we kept the same vertical sensitivities to avoid consequences of subtle calibration differences. With the same goal, primary and secondary current transformers were exchanged so the same was used to measure both currents.

### Results and Conclusions

Finally, while using the same equipment with normal care generally leads to power efficiency over 1, we obtained the results presented in Table 2. This table shows that, for this transformer,

small signal characterization leads to reliable forecasting at a working level. According to my own experience, this conclusion is widely applicable to a ferrite core transformer working at standard induction levels, provided that small signal characterization has been carried out at the same temperature.

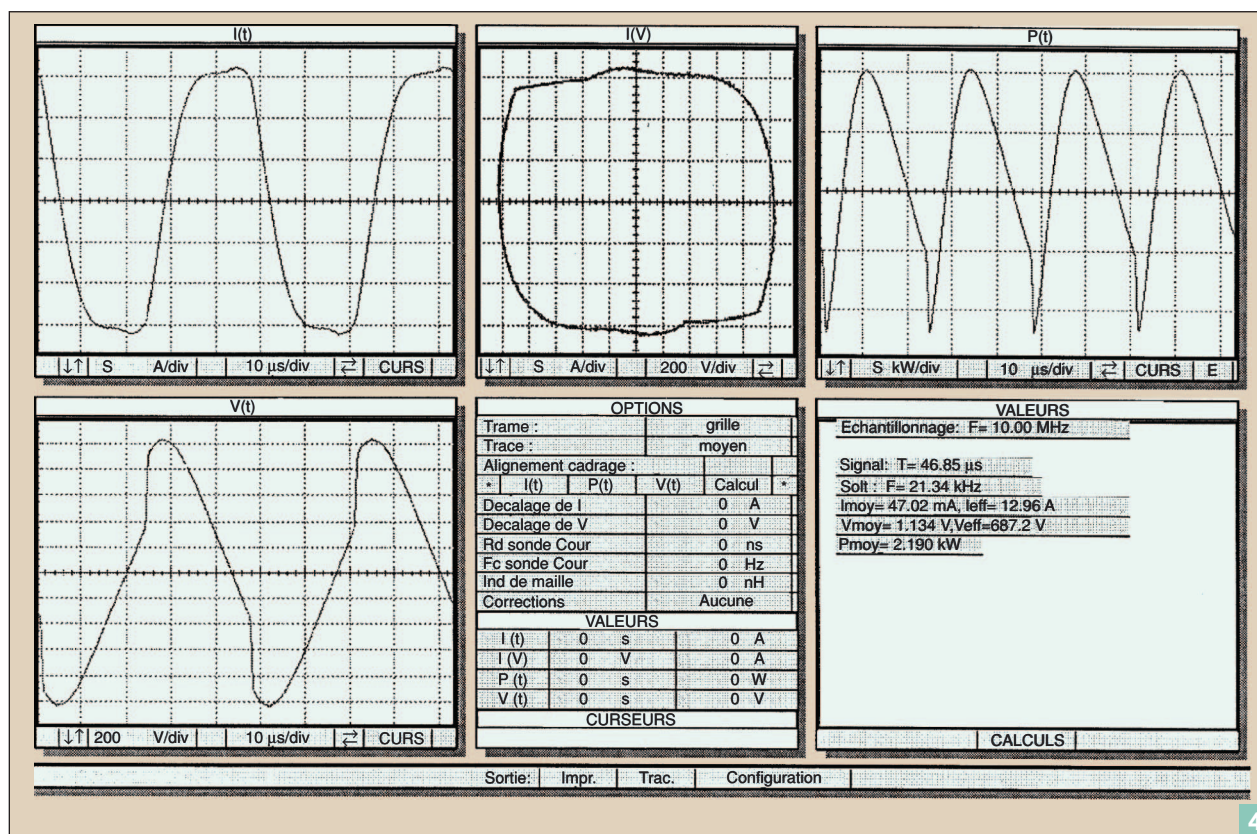
Remaining discrepancies can be attributed to experimental errors as well as to nonlinear behavior of the component. One of the main causes of deviation from this linear behavior is nonuniform temperature inside the core and/or windings. Even assuming that experimental errors are the only cause of discrepancies, these results show that input and output powers have been measured within 0.4%.

### Calorimetric Measurements

#### Aim of the Study

From a transformer manufacturer point of view, understanding the electrical behavior of components does not suffice; the design of any component intended for power application requires temperature impact evaluation. Evaluation in real conditions is difficult because heat goes through inhomogeneous materials (insulators, varnish, air, copper wire, ferrite) and temperature does not remain homogeneous. As a first step toward this evaluation, we decided, along with our industrial partner, to study power dissipation under uniform temperature conditions [11].

A small transformer was built around an ETD44 core made of 3C85 ferrite, and both windings used cylindrical copper wire 1 mm in diameter. Its equivalent circuit (Figure



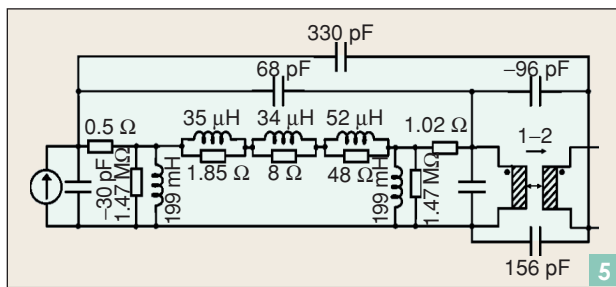
Acquisition at the primary side: current, voltage, and instantaneous power waveforms. Calculated average and root mean square values are in the lower-right-hand-side frame.

5) was identified in the same manner as for the previous one. The three Rp-Lp cells connected in series aimed to represent skin and proximity effects above 10 kHz, where series resistance is proportional to the square root of the frequency [12].

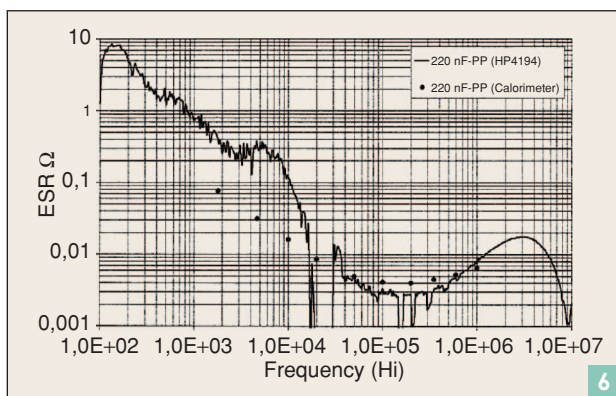
This manner of representing a distributed constant phenomenon by a lumped constant circuit was attractive, but it needed to be validated. To experimentally observe this phenomenon, input impedance in a short-circuit condition is the most direct way. Unfortunately, serial resistance is masked, on the low-frequency side, by magnetizing inductance and, on the high-frequency side, by leakage inductance and stray capacitance. Finally, serial resistance appears only in a narrow frequency range. So, the only way to approach it on a wide frequency range seems to be calorimetric measurements.

### Calorimetric Measurements

Our team collaborated with two physics labs—the Laboratoire d'Electrostatique des Matériaux Dielectriques (LEMD) and the Centre de Recherche sur les Très Basses Températures (CRTBT), both located in the Centre National de la Recherche Scientifique (CNRS) of Grenoble. This collaboration aimed to build a calorimeter intended to measure power losses in good capacitors.



Equivalent circuit of the DUT at 22 °C, as deduced from impedance measurements.



Comparison of electrical and calorimetric ESR measurements. Polypropylene capacitor,  $\delta = 210^{-4}$ .

Because such an apparatus accesses the losses directly, reached accuracy does not depend on the reactive part of the impedance. That is very well suited for measuring the equivalent serial resistor (ESR) of capacitors. Indeed, reached accuracy is far better than that resulting from the

best impedance measurements (Figure 6), especially for low-loss components like polypropylene capacitors [13], [14].

### Calorimeter Preliminary Specification

Our apparatus was designed to operate at ambient temperatures ranging from  $-50$  °C to  $+100$  °C, receive components that enter a 10-cm side cube, measure heat flow from 10 mW to 10 W within a few percent, and allow component supply up to 10 A and 300 V, even at frequencies beyond 1 MHz. The chosen heat flow range has been found consistent, for various devices under test (DUTs), with the uniform temperature hypothesis. However, for our transformer, the upper limit was located at about 100 mW. When such a simple transformer is in a vacuum, thermal exchange between parallel wires is drastically reduced. For this reason, low injected power leads to significant temperature elevation. Relative limit is easily located by measuring the dc resistance of a winding versus input power. As long as temperature remains constant, dc resistance also does so.

### Description of the Calorimeter

Generally speaking, a warm device exchanges heat through its entire external surface, and three physical phenomena contribute to this: conduction, convection, and radiation. In a calorimeter, dissipated power is deduced from the temperature difference that is measured between the two ends of a calibrated thermal leakage resistance. This requires all the heat flow to be channelled in that resistance. This general principle applies to our calorimeter (Figure 7).

More specifically, in our apparatus, power dissipated in the DUT is directly deduced from two powers. A heating resistor (Figure 8) is fixed against the DUT and, when the DUT is supplied, it injects a small power  $P_0$  (slightly smaller than the researched one). Then, the DUT supply is disconnected and the resistor power is increased to  $P_1$  in order to reach the same total power. Power dissipated in the DUT is then calculated as  $P_1 - P_0$ .  $P_1$  is controlled by a regulating loop which keeps constant the temperature difference between the edges of the thermal resistance. Thanks to this principle, DUT temperature is constant during the measurement and thermal resistance does not require careful calibration.

Principle is simple; realization (Figure 7) is not. First, to sweep over the entire temperature range, a constant temperature cold source—a 1) liquid nitrogen tank—is associated with a heating system located on the top of the measuring chamber. Second, to avoid convection exchange and to null conduction exchange except through the chosen path, the DUT is in a vacuum ( $10^{-2}$  P) and so is the nitrogen tank. Radiation exchanges are reduced owing to aluminium plating of the 7) double envelope measurement chamber.

The calibrated thermal resistance is located between the 8) DUT and the 6) top cover of the 7) measurement chamber. Temperature difference between its edges is regulated between 1 °C and 10 °C. This supplies a ten ratio of power flow dynamic range. Several thermal resistances are available to give that dynamic range two decades more. The temperature of the external envelop of the chamber can vary between  $-50$  °C and  $+100$  °C, according to the power injected into the heater located on the top of the chamber.

Every temperature is measured using a standard Pt100 probe, supplied by a low current (10–100  $\mu\text{A}$ ) to avoid disturbing heating. The probe resistor is connected in a four-wire layout, and its voltage is acquired by a phase lock amplifier. Temperature uncertainty remains below  $.05^\circ\text{C}$ .

In the chamber, the DUT is fixed with the heating resistor by a wide copper clamp, 1-mm thick, which drains heat. This clamp is electrically open to avoid magnetic coupling with the device. It is insulated from the external winding by a polypropylene film, coated on both sides with thermally conductive silicon grease. Due to this special cooling, if the component is supplied by a constant current source, hot points probably appear along the copper wire, so voltage and power do not stabilize. This effect disappears when injected power decreases below 500 mW. A constant voltage source leads to natural power regulation, but this does not change the temperature profile along the wire. For this reason, all measurements have been carried out with an injected power ranging from 50–100 mW.

All copper wires—especially the thickest that connects the component to its external supply—conduct heat. To account for this, we take advantage of the two-step measurement. If the losses these wires cause are equal during the two steps, no error remains after the subtraction. This is guaranteed by keeping, on each wire, one point at the DUT temperature and another at the chamber temperature. Remember that this temperature difference does not vary from one step to the other. Technically, the job is done by soldering the wires on thermally conductive ceramic tablets. This is what we call “thermalization.” Every impedance measurement uses a four-wire technique. This greatly reduces long leads incidence.

To end this description, let us mention that, to reach high frequencies, electromagnetic compatibility (EMC) aspects have been carefully checked. Some screenings and filters are mandatory for proper working beyond 100 kHz.

### Apparatus Final Performances

The apparatus described above reached almost all our initial goals. In the  $-45^\circ\text{C}$  to  $+85^\circ\text{C}$  range, it allows power from 10 mW to 10 W to be measured within 1.3%. Above  $85^\circ\text{C}$ , accuracy decreases quickly due to excessive radiation losses. While a  $10^\circ\text{C}$  temperature difference across the thermal resistance is desirable to reach a good relative accuracy of the controlled heat, this temperature difference leads to excessive radiation exchange between the two chambers when temperature exceeds  $85^\circ\text{C}$ .

Measurements are reliable up to 1 MHz. Beyond this frequency, accuracy deteriorates due mainly to currents induced in metallic parts of the calorimeter and stray capacitances between supply wires. Note that the above-mentioned 500-mW limit is linked to our component and, especially, to its bad heat draining. It is not an intrinsic limit of the apparatus.

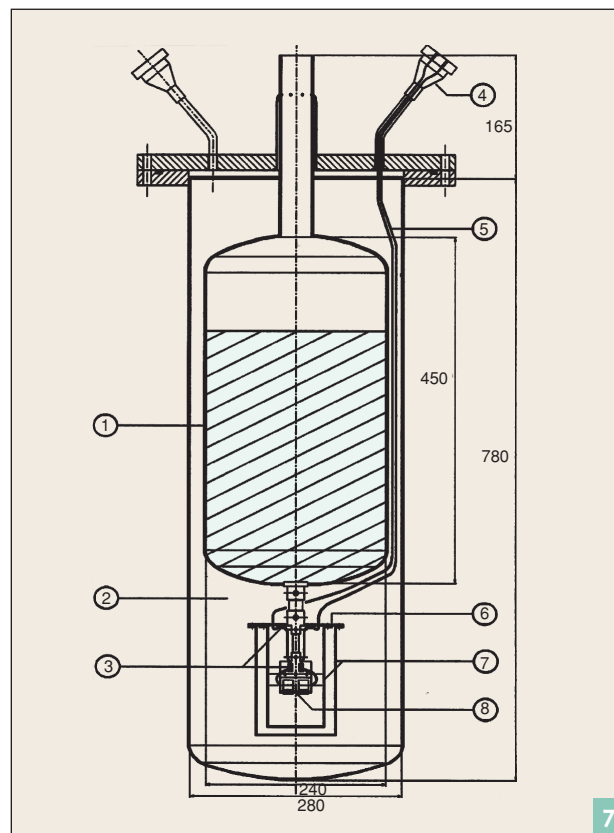
This apparatus can be improved in several aspects but, unfortunately, its main drawback—the time to spend—is not easy to shorten. To give some orders of magnitude, temperature stabilization lasts 35 min for P0 and 20 min for P1 when working close to  $20^\circ\text{C}$ . At every frequency change, if the DUT impedance is known, applied voltage

can be adjusted to reach about the same power. This efficiently reduces the time needed to reach thermal equilibrium. Raising the temperature to  $85^\circ\text{C}$  requires three to four hours as well as installation time; pumping and preliminary temperature stabilization of the entire apparatus must not be neglected! This slowness is mainly due to the lack of air, which usually helps temperature to become uniform. The weak thermal conduction of ferrite is another parameter that is responsible for this. Independently, because of the special heat draining adopted here (copper clamp), measures under large power injection, even if thermal stability was possible, should not be representative of real working conditions.

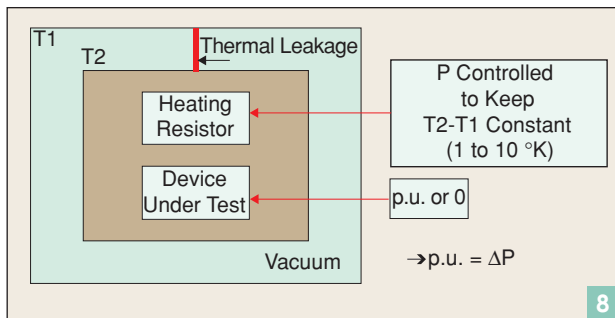
### Electrical Results

As usual, we extracted the equivalent circuit parameters of our transformer from impedance measurements. This technique has been recently updated [15] and an interesting study of impedance measurements can be found in [16]. The three  $R_p$ - $L_p$  cells connected in series (Figure 7) aim to represent skin and proximity effects above 10 kHz, where series resistance is proportional to the square root of the frequency [14].

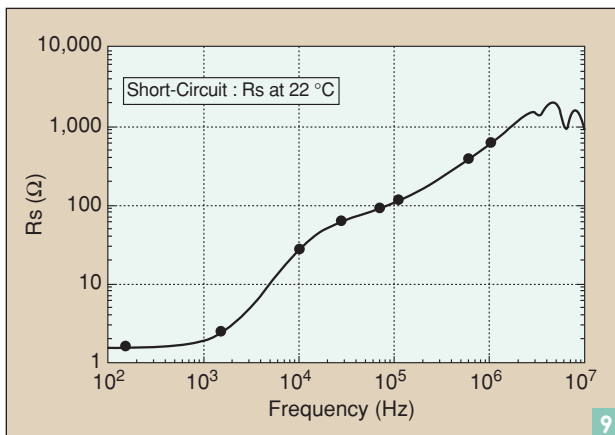
Obviously, we were curious to compare the losses forecasted with that circuit to those directly measured. Results are presented in Figure 9; discrepancies are within 1.5%.



**Calorimeter diagram (sizes in mm): 1) liquid nitrogen tank, 2) vacuum chamber, 3) wire thermalization, 4) connector, 5) supply wires, 6) top cover of the chamber, 7) double envelop, and 8) DUT+ heating resistor.**



Principle of calorimetric power measurement.



Comparison of losses deduced (secondary shorted) from the equivalent circuit (solid line) and from calorimeter measurements (points).

To ease the comparison, we represented the global parallel resistance variation versus frequency. This good agreement has been confirmed at other temperatures ranging from  $-40\text{ }^{\circ}\text{C}$  to  $+50\text{ }^{\circ}\text{C}$ .

Finally, comparison of accuracies obtained using the two methods of measuring power losses shows why the calorimetric approach becomes so topical [17]. However, there is still room for investigation before it enables easy and fast operation, provides an acceptable accuracy with an acceptable disturbance of the DUT, and gives results over a suitably wide temperature range.

## Conclusions

Attempts to verify power losses deduced from simulation by direct measurement are highly respectable. It is the only means to become confident in models that limit simulation accuracy.

Unfortunately, in the power electronics area, power measurements are often very difficult. If we rely on them to validate models, we must expect to pay not only for high-performance equipment but also for the time to use it at its best.

Despite the fact that it is difficult to build, costly, and time consuming, calorimetric apparatus remains the only way to accurately measure the losses of low-loss ( $\tan \delta < 10^3$ ) components such as some capacitors.

## Acknowledgments

I would like to warmly thank Microspire, who has financially supported us since 1991; the Laboratoire d'Electrostatique des Matériaux Dielectriques (LEMD), who designed and built the calorimeter; and all the Ph.D. students mentioned in the references, who did the job.

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