Devices for the Separation of the Common and Differential Mode Noise: Design and Realization

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***Abstract* - EMI measurements and the design of proper filter to mitigate the EMI noise is very important, since the EU Directive 336/89 has been published. For conduction emission noise, the separation of the common and the differential mode noise is important for determining whether the differential or common mode component is dominant. Therefore, it will be possible determine whether an anticipated change in a value of an element in the power supply filter will be effective on the considered component. Two devices for the separation of the conducted EMI noise into the common and the differential mode components have been realized and tested. In this paper the authors show the design considerations that have to be taken into account for the realization of the two devices.**

Keywords: EMI and EMC Issues.

I. Introduction

Electromagnetic interference (EMI) emission is an important matter for any electric and electronic equipment. When the noise emission of a part of equipment does not satisfy the standard limits, it is usually not easy to find the origin of the noise sources. The measured emissions are a mixture of common mode (CM) and differential mode (DM) noise. In the equipment, the sources of CM and DM noise are of different nature and have to be distinguished carefully [1], [2]. Furthermore, the design procedure for the filters is usually divided in common mode and differential mode filter design [3]. Therefore, it is quite important to discern the two modes to design a good filter.

The basic separation of CM and DM is shown in Fig.1. This scheme represents the equivalent circuit of a Line Impedance Stabilization Network (LISN) in the high frequency behavior and the Equipment under Test (E.u.T.) is shown as an electromagnetic noise source.

The differential mode voltage (V_{DM}) and current (I_{DM}), the common mode voltage (V_{CM}) and current (I_{CM}), depicted in Fig.1, are defined as follows:

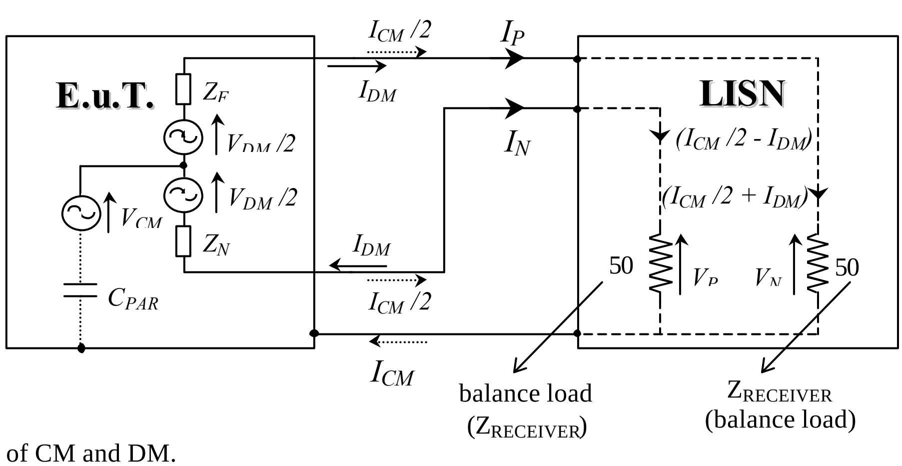

Fig.1: Definition of CM and DM.

$$V_{DM} = (V_P \quad V_N) = 50 \times \frac{I_{CM}}{2} = 25 \times I_{CM} \tag{1}$$

$$V_{CM} = \frac{(V_P + V_N)}{2} = 50 \times 2 I_{DM} = 100 \times I_{DM} \tag{2}$$

$$I_{DM} = \frac{(I_P \quad I_N)}{2} \tag{3}$$

$$I_{CM} = (I_P + I_N) \tag{4}$$

II. Requirements for the Common and the Differential Mode Measurements

The main requirements to measure the noise are:

- low damping of the signals;
- high damping of the suppressed signal components;
- linear amplitude response within the requested bandwidth for conducted emissions (150kHz - 30MHz);
- impedance matching (50);
- low distortion;
- no interference with device under test and the LISN.

III. Proposed Solutions

The first proposed solution, shown in Fig.2, allows to divide the common and the differential mode voltage components, respectively by the sum and the difference of the line and the neutral voltages [4].

From the given definition, it yields:

$$(V_P + V_N) = 2 \times V_{CM} \qquad (V_P \quad V_N) = V_{DM}$$

For practical reasons, two different devices have been realized, as shown in Fig.3. The first one makes the sum (common mode) and the other makes the difference (differential mode). The reason of this approach is due to the high parasitic coupling capacitance among the contacts introduced by the switch, which can produce an undesired unbalance on the two branches of the circuit.

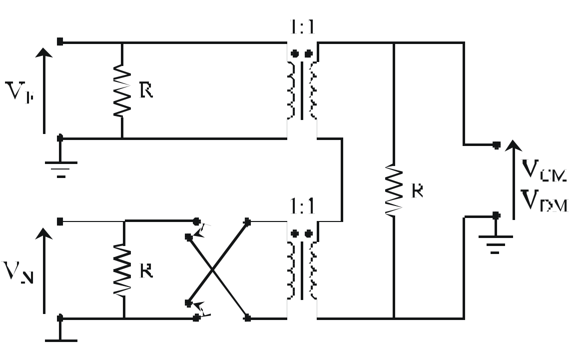

Fig.2: First proposed solution: the schematic.

The second proposed solution is shown in Fig.4. The common mode voltage is feed through primary side of the transformer to the resistor R_{CM}, while the differential voltage is blocked. At the output resistor R_{DM} only the differential mode is obtained [4], [5].

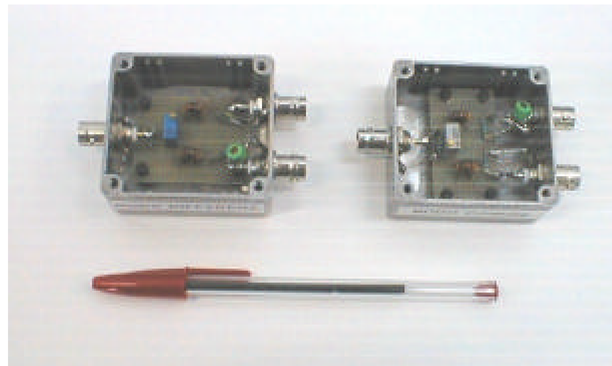

Fig.3: First proposed solution: the devices.

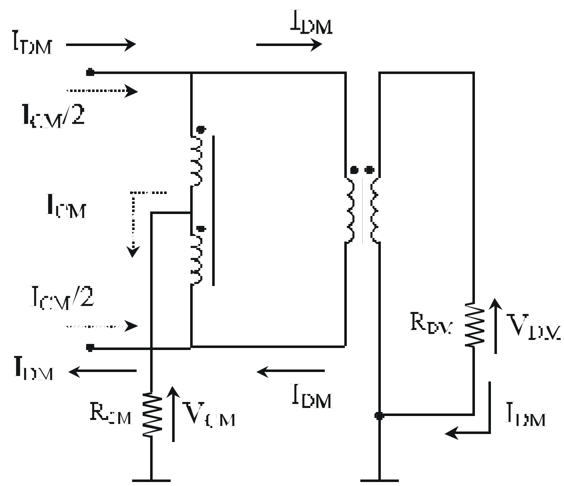

Fig.4: Second proposed solution: the schematic.

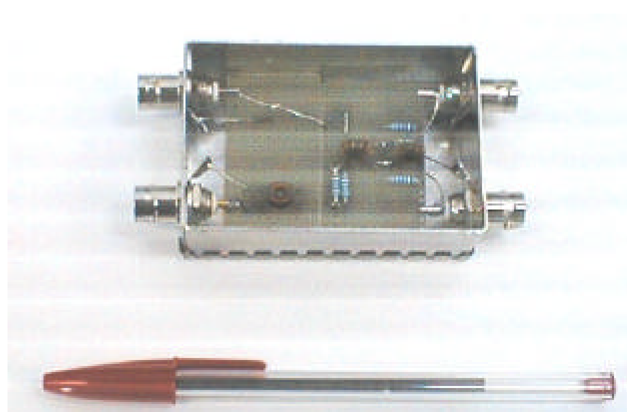

Fig.5: Second proposed solution: the device.

IV. Design Considerations

In this section the design considerations that have to be taken into account for the selection of the components and the realization of the devices are given.

The realization of the devices is quite difficult, because it is necessary to build transformers and common mode chokes able to work within the requested bandwidth for conducted emissions (150kHz - 30MHz). In fact, the parasitic elements of these components show their effects at the frequencies under consideration and the obtained results are often different from those expected. The main difficulty is to find some practical arrangements (placement of the components, layout, ect.) to minimize these effects and thus to improve the devices performance [6].

The first proposed solution has two wide band transformers, two resistors (R_1 and R_2) to match the impedance, and two capacitors and a trimmer (R_V) to compensate the possible differences between the two transformers ratio as shown on Fig.6.

Resistors

The resistors have to be selected to match the impedance to the value of 50 both for the inputs and the output. For the calculation of R_1 and R_2, it is necessary to replace the signal generators connected to the inputs with their internal impedance (50) and the output with the LISN impedance (50), as shown in Fig.7 for one input.

The equivalent circuit of Fig.7, can be simplified as shown in Fig.8.

The same considerations have to be taken into account for the other input and for the output. The following system can be derived:

$$\begin{aligned} R_1 //(R_2 //50 + R_1 // 50) &= 50 \\ R_2 //(R_1 //50 + R_1 // 50) &= 50 \end{aligned} \quad (5)$$

The solutions of this system are $R_1 = R_2 = 150$ and the used resistors for the devices have 1% of tolerance and 0.25W power.

Transformers

The most important parameter for the choice of the transformer is the insertion loss (IL), which is typically stated in dB. The IL_{dB} is defined by the following equation:

$$IL_{dB} = 10 \log_{10}\left(\frac{P_{L,wo}}{P_{L,w}}\right) = 20 \log_{10}\left(\frac{V_{L,wo}}{V_{L,w}}\right) \quad (6)$$

where $P_{L,wo}$ and $V_{L,wo}$ are the load power and the load voltage without the transformer; $P_{L,w}$ and $V_{L,w}$ are the load power and the load voltage with the transformer. The IL represents the fraction of power loss due to the insertion of the transformer in the circuit. These losses are due to the non ideal behavior of the transformer: core losses, copper losses, etc..

The equivalent circuit for the wide band transformer is shown in Fig.9.

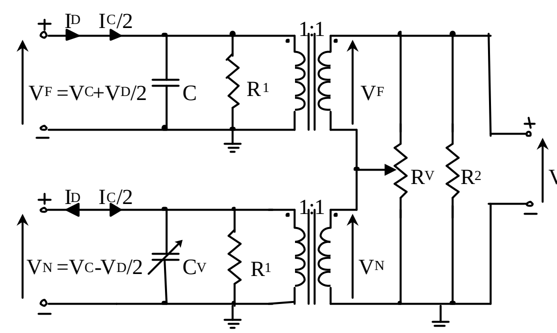

Fig.6: Schematic of the first proposed solution.

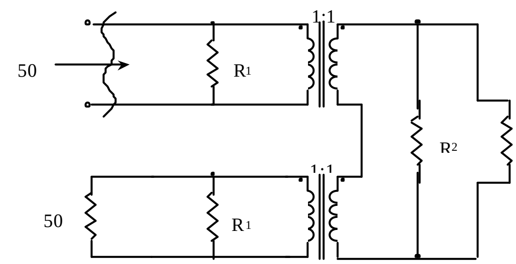

Fig.7: Schematic for the calculation of the input impedance.

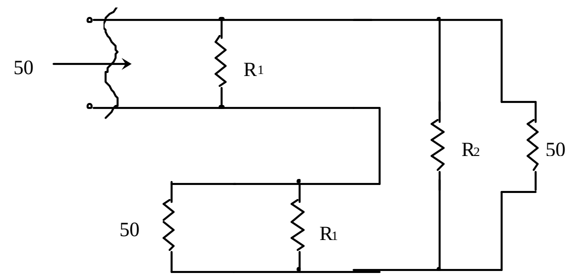

Fig.8: Equivalent circuit for the calculation of the input impedance.

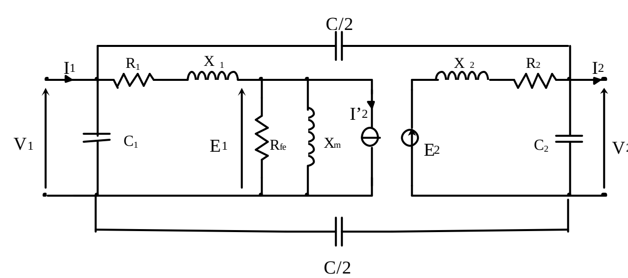

Fig.9: Wide band transformer: equivalent circuit.

The transformer parameters are:

- the resistances R_1 and R_2, that represent the windings copper losses;
- the reactances X_1 and X_2, that represent the leakage inductance due to the leakage flux;
- the resistance R_{fe} that represents the iron losses;
- the reactance X_m, that represents the core magnetization;

- the capacitances C_1 and C_2, that represent the capacitance among the turns of each winding;
- the capacitance C, that represents the capacitance between the primary and the secondary side of the transformer.

The most important parameters that have to be taken into account for the transformer core selection and the realization of the windings are the capacitors C_1, C_2, and C. These capacitances limit the transformer performance at high frequency and their value have been experimentally measured.

The magnetizing inductance is determined by the core selected. The characteristics required for their usage on the realized devices are:

- high value of A_L;
- high impedance all over the bandwidth;
- low harmonic distortion;
- high Curie temperature and high flux density saturation.
- low loss factor (tan) for high frequency.

Two cores have been selected: toroidal and with double cut as shown in Fig.10 and Fig.11.

For the toroidal core, two solutions have been realized: with crossed windings and with separated windings. Also for the double cut core, two solutions have been realized using two different type of core: a smaller and a larger model. These choices have been suggested by the experimental measurements.

The windings have been realized using twisted copper wires in order to improve coupling between the primary and the secondary side of the transformer. The number of turn chosen has been given by the measurement results as shown in the next section. Owing to the high and wide band frequency operation of the transformer, it is not possible to have a formula for this calculation.

Fig.10: Toroidal core transformer: (crossed and separated windings).

Fig.11: Double cut core transformer: (smaller and larger model).

Capacitors

The capacitors C_v and C have to compensate the phase differences introduced by the cable parasitic capacitance and the transformer parasitic capacitance. The value of C (4.7pF) has been selected to minimize its influence on the measurements. In fact, a capacitance of 4.7pF has an impedance of about 1120 at 30MHz, but it is placed in parallel to 50 , thus it has a negligible influence on the equivalent input impedance. Since the difference of the capacitance between the two inputs is very little, also the variable capacitor, after its setting, is close to 4.7pF. This means that the common and the differential mode measured noises have an error between 4% at the high frequency (30MHz) and 0.02% at low frequency (150kHz). Moreover, knowing the error, it is possible to correct it directly through the software of the EMI equipment used for the measurement. Although the capacitors are source of errors for the measurements, their presence is absolutely necessary for a good set up of the device.

V. Measurements Results

Several measurements have been realized to determine the behavior of the proposed device using the set up shown in Fig.12.

The Insertion Loss (IL) of the toroidal and double cut core transformers has been measured.

For practical reasons, Fig.13 and Fig.14 show the Voltage Transfer Ratio (VTR) instead of the IL. However, it is easy to demonstrate that the relationship between the VTRdB and the ILdB is given by the following equation:

$$VTR_{dB} = 20\log_{10}\left(\frac{V_{L,w}}{V_{L,wo}}\right) = \; IL_{dB} \qquad (7)$$

Different number of turns have been realized using both cores. The best results for the toroidal core have been reached with 10 turns: IL at 30MHz is only 0.38dB, as shown in Fig.13.

For the double cut core transformer different cores in size have been used. For the smaller core, because of the difficulty of folding the copper wires, the assembly has been very difficult and the only possible number of turns, without damaging the wires, was 3 (Fig.14).

With a larger core, it is possible to realized more turns and the results are shown in Fig.15. The number of turns is a trade off between the core magnetization and the negative effect of the parasitic capacitance at high frequency. In fact with only 4 turns the core is not sufficiently magnetized and with 10 turns the behavior is good up to 2MHz. From 2MHz to 30MHz, the effect of the parasitic capacitance between the primary and the secondary side of the transformer is predominant and the behavior is pretty bad. The best compromise in this case is 6 turns. However, as it can be seen in Fig.15, the larger core has a more negative influence in the behavior of the transformer than the smaller core: at high frequency the IL increases for any number of turns and consequently the voltage transfer ratio decreases.

To verify the quality of the measurements for the IL, the transformers impedance has been measured. If the impedance is about 50 , the measurements have been carried out correctly.

In Fig.16, the impedance for different transformers have been compared. It can be seen that the impedance is quite matched up to a few MHz. This behavior shows that the transformers, at low frequency works in a linear zone, but when the frequency increases, the influence of the inductance and the parasitic capacitance are not negligible any more.

The performance shown by the realized transformers are quite similar. For this reason, for the choice of the best transformer, it is necessary to make a comparison of the internal parameters of the transformers. In particular, it is important that the capacitance between the primary and the secondary winding (C in Fig.9) is as low as possible. In fact, at high frequency, this capacitor is the best way for the propagation of the noise.

The measurement of this capacitance is very difficult, because the influence of the windings inductance can produce errors. Several tests set up have been build as shown in Fig.17.

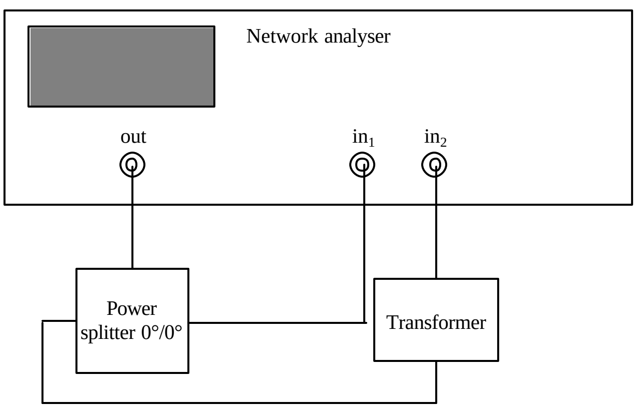

Fig. 12: Measurements set up.

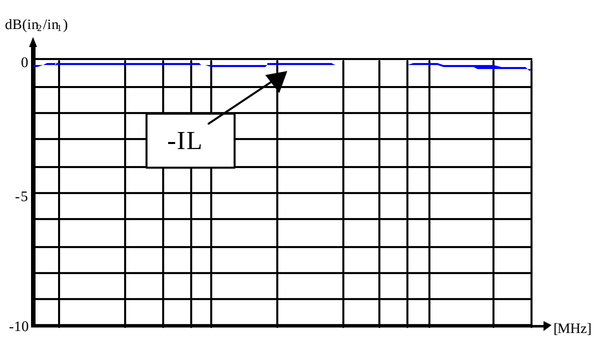

Fig.13: Toroidal core transformer (10 turns): -IL.

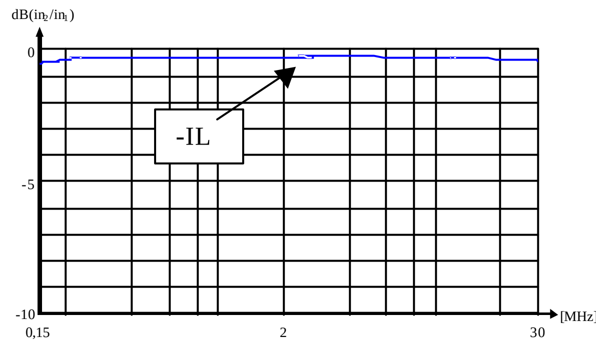

Fig.14: Double cut core transformer (3 turns): -IL.

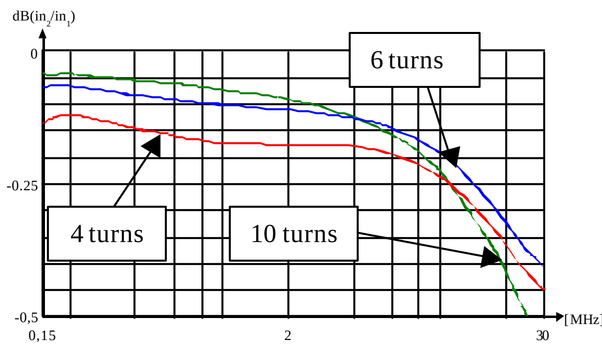

Fig.15: Double cut core transformer with different number of turns: -IL.

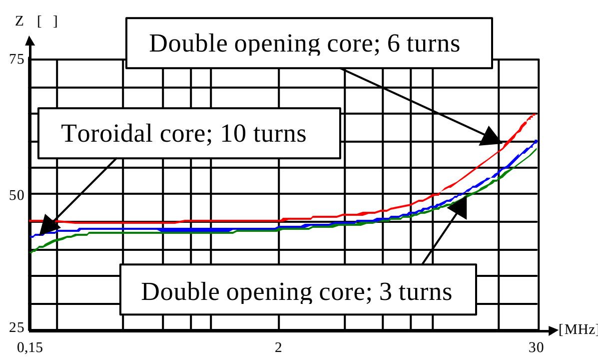

Fig.16: Impedance comparison for different transformers.

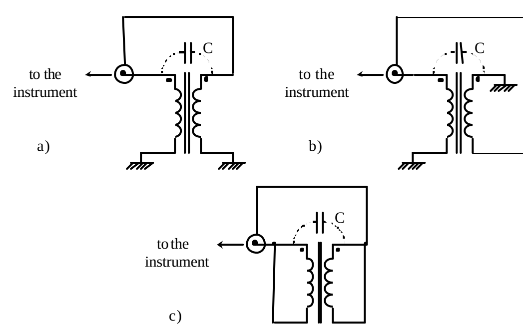

Fig.17: Different tests set up for the measurement of the coupling capacitance between primary and secondary.

The best test set up is c), because the influence of the windings inductance is minimum.

Thanks to the set up c) it has been possible to measure the parasitic capacitance between the primary and the secondary of each realized transformer.

The experimental results are shown in Fig.18, Fig.19 and Fig.20. It can be seen, that the measurements is pretty uncertain up to 700kHz, because the parasitic capacitance is very little, consequently the capacitor is substantially an open circuit and the instrument is not able to show any result. The results are summarized in Table I.

The best choice for the realization of the devices is to use the toroidal core transformer, because it shows the smaller parasitic capacitance between the primary and the secondary winding while the IL value is about the same for the three transformers.

The same considerations have been taken into account for the realization of the second proposed solution.

The common and differential mode transfer functions for both the proposed solutions are less than 0.5dB and this is a very good result. The common and differential mode rejection functions are better for the first proposed solution, but two devices have to be used. For the second proposed solution, the common and the differential mode rejection functions are a little worse than the first one, because of the parasitic effects due to the near montage of the two outputs (differential and common), but only one device has to be used [6], [7].

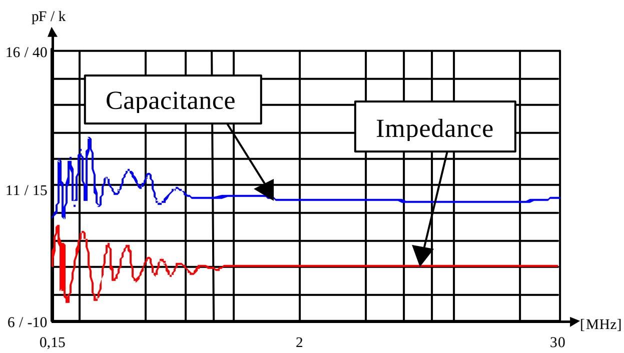

Fig.18: Parasitic capacitor of the double cut core transformer with 3 turns: capacitance and impedance.

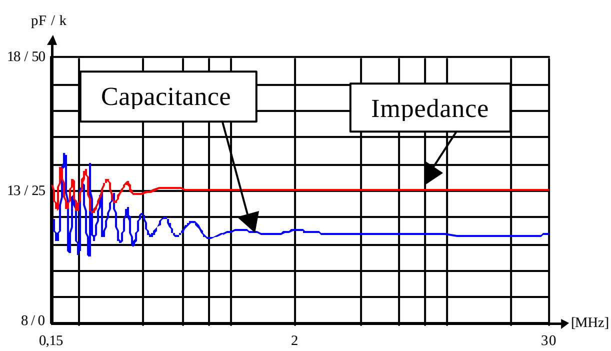

Fig.19: Parasitic capacitor of the double cut core transformer with 6 turns: capacitance and impedance.

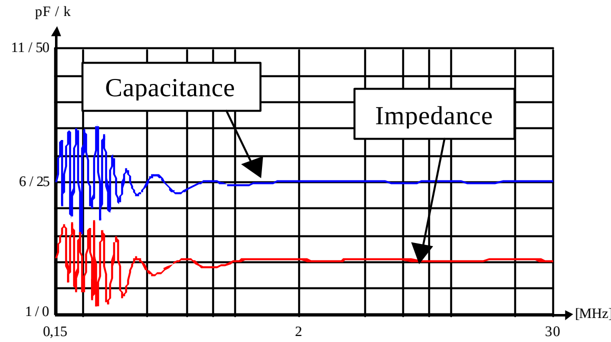

Fig.20: Parasitic capacitor of the toroidal core transformer with 10 turns: capacitance and impedance.

TABLE I: Parasitic capacitance for the realized transformers.

Type of transformers	Capacitance [pF]
Double opening core (3 turns)	10.4
Double opening core (6 turns)	11.5
Toroidal core	6

VI. CONCLUSIONS

The design considerations and the realization of the devices for the separation of the common mode and differential mode noise have been presented.

The choice of the components is only possible after measurement results, because at high frequency the influence of the parasitic parameters is predominant. For these reasons, several transformers have been realized and tested to show which are the critical elements that have to be considered to select the best solution. The found best solution is the toroidal core transformer, because it has the smaller parasitic capacitance.

The two proposed solutions have been realized and both show a common and differential mode transfer function less than 0.5dB and this is a very good result. For the common and the differential mode rejection functions, the second proposed solution is a little worse than the first one, because of the parasitic effects due to the near montage of the two outputs (differential and common).

VII. REFERENCES

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