Radio Frequency Characteristics of High Power Common-Mode Chokes

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***Abstract*— Toroid EMC-ferrites are used to design chokes for low-pass filters. They provide high permeability up to a corner frequency in the range of kHz to MHz. At and above this corner frequency the damping effect is strongly supported by magnetic loss. Modelling this behaviour using complexe μ is a problem, because $\underline{\mu}$ is often known only for one exemplar of the material and only up to the corner frequency. The filter designer also has to deal with huge tolerances of the materials of up to 25%. In this paper measurements of $\underline{\mu}$ up to 100 MHz of MnZn-cores are presented and a modelling approach is shown to be accurate in the frequency domain. Besides parasitic capacitance depending mainly on wire distance to core, with EMC-ferrites magnetic loss is a predominant predicate at higher frequencies.**

I. Introduction

State-of-the-Art in EMI filter design is the determination of necessary insertion loss at the lowest frequency where the regarded interference usually has got its maximum. The desired operating temperature and its influence on the initial permeability is taken into account for the determination of nominal inductance of a common-mode choke. Leakage inductance is a concern in order to avoid saturation [1] and for attenuating differential-mode interference. But only low-frequency (10-200kHz) characterisics are taken into account and the filter performance at high frequencies up to 100MHz is measured and designed by a possibly time-consuming trial-and-error process of prototyping and measurement.
Especially when filters are to be integrated and applications be miniaturized the high frequency behaviour must be considered during the design phase. Besides common-mode chokes all other components have to be taken into account on the system level. As network based modeling of conducting structures is already state-of-the-art in system-level design[2], a design-flow including the circuits components is developed. High frequency behaviour of EMI filter capacitors is included in system-level design in [3]. The high frequency behaviour of common-mode chokes consists of the rapidly decreasing inductance above a corner frequency, the rise in loss, parasitic capacitance of coils and capacitive coupling among them. A common-mode choke's leakage flux may interfere with other parts of the system and vice versa it may also collect parasitic flux of other components. Couplings on system level are not discussed in this paper and will be part of future work.

II. Ferromagnetic Materials

Toroid cores are used to design chokes because they are made of one piece. Therefore they are much cheaper and provide a higher relative permeability than other core shapes. As the number of turns is small, the disadvantage of higher winding costs is not fatal and toroids are nearly always used for high power common-mode chokes.
In the frequency range above 10 kHz ferromagnetic materials with reduced conductivity provide high permeability up to very high frequencies. The main types of materials, nickel-zinc and manganese-zinc ferrites, iron-powder-cores and nanocristalline-iron-cores with resistivities higher seven orders of magnitude than that of iron, are compared in Table I.

TABLE I
Comparison of ferromagnetic materials for high frequency applications

	Ferrites (Ni-Zn, Mn-Zn)	Iron Powder	Nanocristalline Iron
Permeability	+	-	++
Saturation	–	++	+
Conductivity	+	+	+
Price	++	+	–

Although nanocristalline-iron-cores have the best performance they are very seldom used because of their high price. Applications with very high saturation currents are equipped with iron powder cores but the main material for common-mode chokes are the cheap ferrites.

III. Impedances of Common-Mode Chokes

In EMI filters design a common-mode choke is supposed to consist of a perfectly coupled inductor, a resistance accountung for loss and a leakage inductor per coil. The measurement of a coil's impedance with the impedance analyzer clearly shows the dependancy of inductance and loss on the frequency that is currently not taken into account.

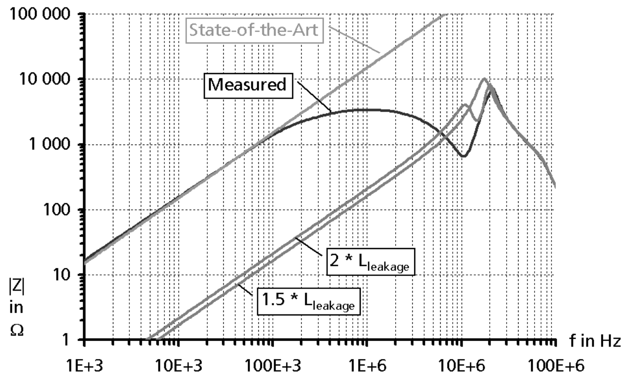

Fig. 1. Impedances of one coil of a three-phase common-mode choke with 15 turns per coil.

Figure 1 shows the impedance of one coil of a three-phase common-mode choke with both secondary windings open, one secondary short and both other coils shortened. Deviating from the ideal inductance the impedance is lower at higher frequencies and resonances occur. The comparison of

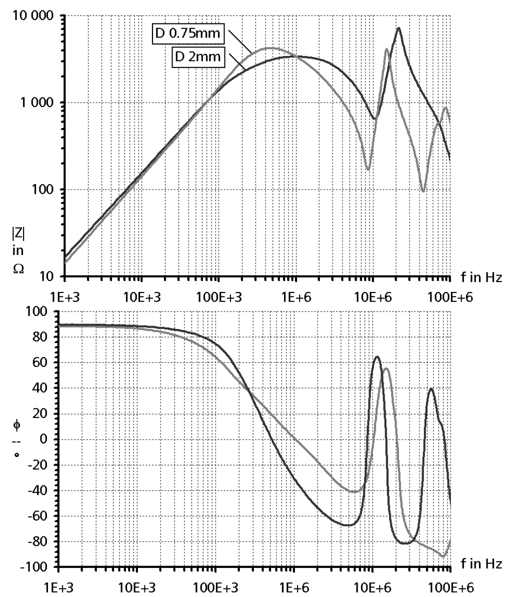

Fig. 2. Open impedance of the three-phase common-mode choke with different wire diameters.

the coil's impedance with 2mm and 0.75mm wire diameter with the same common-mode choke provides a big difference at higher frequencies. The impedance with 2mm becomes more capacitive and the resonances are also shifted. The difference in wire diameter has influence on the resistance of the wire, the leakage inductance and the capacitive coupling. As leakage inductance and copper loss has no influence on the open measurement the differences must be owed to capacitive coupling among the coils. A three-phase common-mode choke is to be modelled as a three-port network with frequency dependant parameters and coupling capacitances in order to achieve high accuracy.

IV. Frequency Dependant Material Properties

Core material properties are given in manufacturers datasheets in terms of complexe relative permeability and initial permeability. Common materials have high tolerances of 25% regarding initial permeability. Thus it is important if the high frequency characteristics also vary with different exemplars of cores. Complexe permeability $\underline{\mu} = \mu' - j\ \mu''$ of

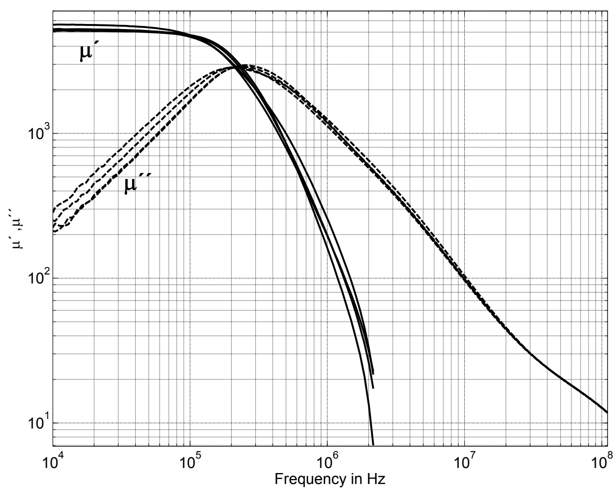

Fig. 3. Complexe permeability $\underline{\mu}$ of five exemplars delivered within 2 years by the same manufacturer.

five exemplars delivered within 2 years by the same manufacturer is shown in Figure 3. The impedance of a coil wound on the material results in a series equivalent circuit $\underline{Z} = R_s + j\omega\ L_s$ calculated from the air-core equivalent inductance L_0.

$$\begin{aligned}
\underline{Z} &= (\mu' - j\ \mu'')\ j\omega\ L_0 \\
Rs &= \mu''\ \omega\ L_0 \\
Ls &= \mu'\ L_0 \\
where\ \ L_0 &= \mu_0\, n^2\, \frac{A_{eff}}{leff}
\end{aligned}$$

The air-core equivalent inductance L_0 is calculated from μ_0, the number of turns n, effective area A_{eff} and path length l_{eff}, values given in the datasheet of the core. Therefore μ' is proportional to the inductance and μ'' times ω is proportional to the series resistance of the coil. Permeability decreases rapidly above the corner frequency at 200kHz and the impedance is predominated by loss at higher frequencies. Ferrites for EMI suppression have the special property of high loss at and above the corner frequency so μ'' is measured up to 100MHz while there is no significant inductance anymore. The differences between exemplars are small. Thus, it is possible to model the behaviour for different exemplars of a core.

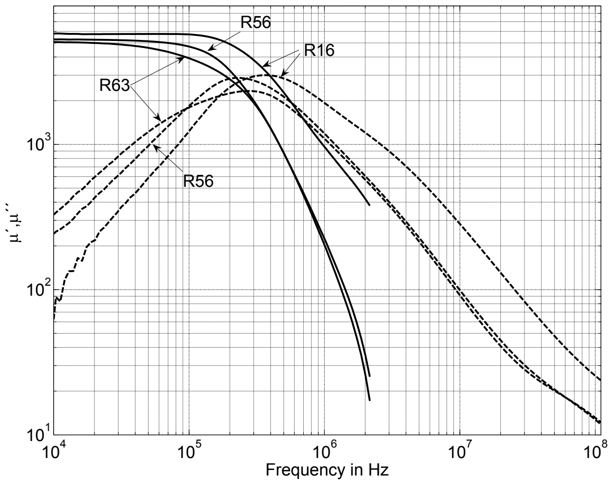

Fig. 4. Differences between cores of the same material with different size. Mean values of 5 exemplars

Another thing to consider is, that the material properties of ferrite toroid cores depend on the size of the core due to different pressure and temperature processing during production which is often not mentioned in the datasheet. Material properties of one core size may not be applicable on other core sizes from the same material as shown in Figure 4. The bigger cores (R56, R63) have similar properties but different from the small core(R16).

V. Modeling Complexe μ by Analytic Functions

Relative permeability's relation to the frequency is approximately $\mu_{initial}$ up to the corner frequency f_c. Therefore it can be calculated similar to the transfer function of Bessel low-pass filters with filter coefficients a and b. Figure 5 shows Rs and Ls, where $f_c = 270\,kHz$.

$$\mu' = \left|\frac{\mu_{initial}}{1 + ja\; f_n + j^2 b\; f_n^2}\right| \qquad f_n = \frac{f}{f_c} \quad (1)$$

As Rs behaves like the transfer function of a high-pass filter μ'' is approximated with the same function but f_n^{-1}:

$$\mu'' = \frac{1}{\omega L_0}\left|\frac{R_{smax}}{1 + ja\; f_n + j^2 b\; f_n^2}\right| \qquad f_n = \frac{f_c}{f} \quad (2)$$

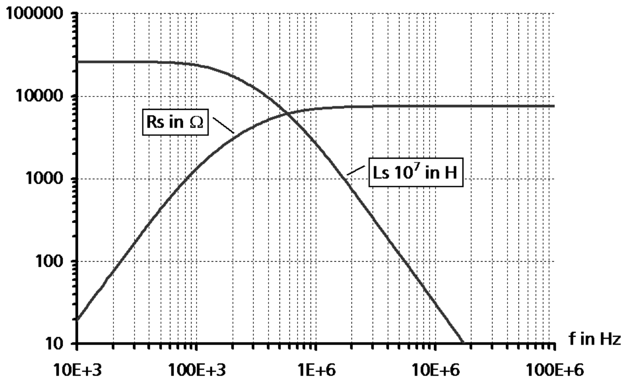

Fig. 5. Serial resistance and serial inductance modeled with analytic functions: Equation 1 and 2.

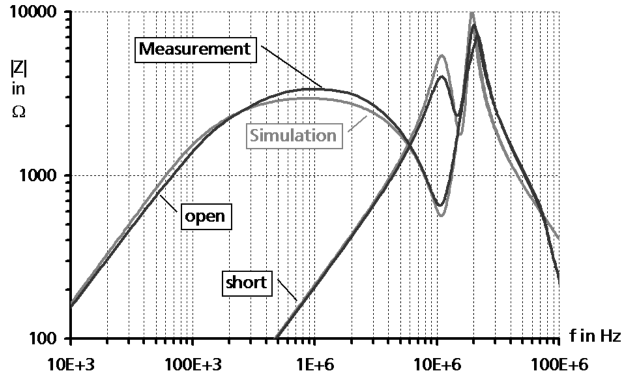

Fig. 6. Comparison of model data with the measurement.

Figure 6 shows measurements and model data of the impedance of one coil of the common-mode choke with the secondary windings open and short. The good approximation of the frequency dependent parameters with the discribed functions enables high accuracy design in the frequency domain.

VI. Parasitic Capacitance

The impedance measurements showed a great sensibility of parasitic capacitances on the design of the coil. In Figure 2 the bigger wire diameter generates the higher parasitic capacitance. In order to take into account the parasitic capacitances for design at the system-level, the relationship between parasitic capacitance and the number of turns and the wire diameter is investigated.

Measured parasitic capacitances of 16 turns single layered by hand on the same core are shown in grey bars in Figure 7. The capacitance increases slightly with the wire diameter from 0.2 to 1mm as expected. At 2mm wire diameter a smaller capacitance is measured because with thick wires the coil wounded with the same forces like the smaller wires have a bigger distance to the core.

The parasitic capacitance of single-layer coils is

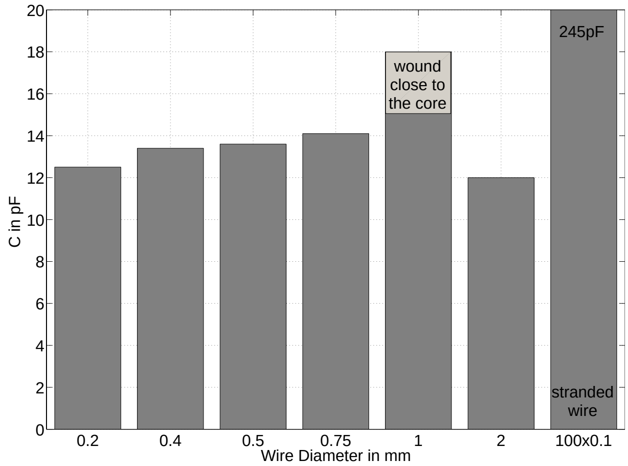

Fig. 7. Measured capacitance of 20 turns with different wire diameter.

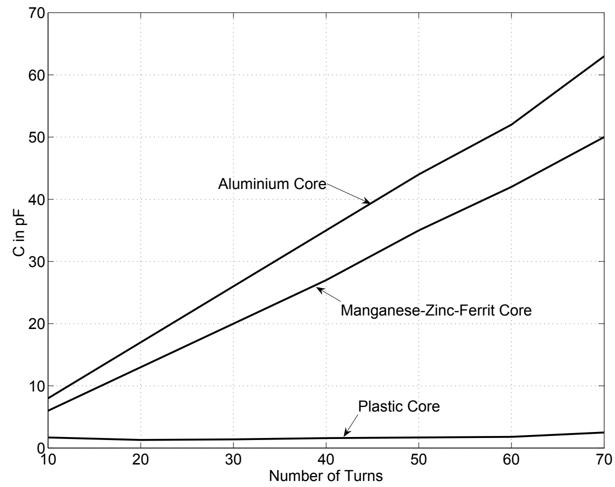

Fig. 8. Measured coupling capacitance between two coils related to the number of turns on an aluminum, a polyethylen and a ferrite core.

dominated by capacitive coupling to the core which is clearly demonstrated by repeating the measurement with 1mm diameter wounded close to the core. The capacitance of a flexible stranded wire wound smoothly on the core is two orders of magnitude higher than the capacitances of solid wires.

Capacitive coupling to the core lets one expect values of coupling capacitances between 2 coils as high as parasitic capacitances of the single coil. Figure 8 shows the relation between coupling capacitance and the number of turns of two coils on three different cores consisting of aluminum with perfect conductivity, polyethylen with no conductivity and a ferrite core. Coupling capacitance of the single layered coil on the plastic core is in the range of pF, while the capacitances on the ferrite core are almost as high as with an aluminum core. Although the conductivity of ferrites is much lower than that of conductors it acts as an electrode for parasitic capacitance almost in the same way.

VII. Conclusion

The influence of the wire diameter and the distance to the core on the parasitic capacitance of windings on toroid cores was presented. Analytic functions are used to model frequency dependent characteristics of common-mode chokes with high accuracy. Investigating the parasitic capacitance has not yet come to an end where it will be predicted on toroid cores. This is our task for future work.

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