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CHARACTERISATION OF PARTIAL DISCHARGE CURRENT TRANSFORMER BEHAVIOUR BASED ON SECONDARY-SIDE PARAMETERS AND MODELLING

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Abstract

High-frequency current transformers (HFCTs) are widely used for partial discharge (PD) measurements. Accurate analytical modelling is often hindered by design information and the presence of parasitic components. This study investigates the characteristics of HFCT behaviours through modelling and proposes a simplified small-signal equivalent circuit to address measurable secondary-side parameters. The model introduces analytical relationships between the physical sensor parameters and the resulting frequency response, including cut-off regions and flat-band behaviour; and explains the influence of burden resistance. A practical method is proposed for extracting key model parameters from secondary-side impedance measurements without prior knowledge of the sensor construction. The approach is experimentally validated using five HFCT sensors of different designs through secondary-side characterisation. The measured and simulated responses show good agreement, confirming that secondary-side-only parameterisation is sufficient to estimate the usable bandwidth and frequency behaviour of unknown HFCTs.

Keywords: *High-frequency current transformer (HFCT), partial discharge, equivalent circuit modelling, frequency response, secondary-side characterisation, parameter extraction, bandwidth*

1. INTRODUCTION

High-frequency current transformers (HFCTs) are widely employed for the detection and measurement of partial discharge (PD) activity in high-voltage (HV) equipment due to their non-intrusive design, galvanic isolation, wide frequency bandwidth, and suitability for fast-transient current sensing. The importance of accurate PD measurement and interpretation is well established, especially for the condition assessment of HV insulation, diagnostics, preventive maintenance, and asset management [1,2]. Standardised approaches, such as those recommended by the international standard IEC 60270 (2000) [3], provide a reference framework for PD quantification, also at higher frequencies where HFCT-based methods are used.

Various research efforts have been directed towards understanding and modelling the frequency-dependent behaviour of HFCT sensors and their influence on measured PD signals. HFCT modelling can be based on several methods, depending on the selected approach and the required operating frequency ranges, such as Laplace-domain models [4,5], broadband transfer [6], and formalised transfer function characterisation [7]. Studies have also focused on charge estimation of PD pulses and their dependence on sensor characteristics, which is crucial for PD measurements and diagnostics in HV equipment. Laplace-domain formulations [8] approaches demonstrate that accurate charge reconstruction is highly sensitive to deviations from flat-band response, which is always difficult to achieve in real devices. In parallel, optimisation and design-oriented studies have explored the influence of properties of the magnetic core, winding configuration, and parasitic elements on HFCT performance [9,10]. Design-focused studies on practical design considerations, including shielding, positioning, and transfer impedance measurement, are addressed [11,12]. The HFCT application beyond 1 MHz has also

been examined, highlighting measurement limitations and the need for improved wideband calibration [13].

Despite these advances, from industrial applicability perspective, a gap remains between analytical modelling approaches, simulation-based predictions, and experimentally measured HFCT responses. Existing models rely on transfer function representations or simplified equivalent circuits without explicitly linking individual physical parameters to observable changes in bandwidth, gain, and cut-off frequencies. Limited studies provide a concept combining analytical interpretation, circuit simulation, and direct comparison with measured responses of complete HFCT sensor samples. The relationship between parameter variation and measurable frequency response shifts is often partially demonstrated and rarely across multiple sensor implementations. It is useful to explore as parameters differ for individual HFCT designs.

An additional difficulty in HFCT analysis lies in the fact that manufacturers of such sensors rarely specify all the parameters that would be necessary for accurate reconstruction of all parasitic influences. The combined equations are too complex to be useful [5], and therefore simplified equivalent models are a necessity for such analysis.

This paper studies the limitations arising from the simplification of the complex model by presenting a parameterised equivalent circuit model that connects HFCT physical parameters to its frequency response characteristics. The focus is on the parameters that are measurable on already-constructed devices, without a priori knowledge about their design. The work shows that for a simplified model, it is sufficient to analyse the yet-unknown HFCT just from the secondary side, without (easier) or with termination (more difficult). The simplified model is then used for the understanding of analytical expressions for bandwidth limits and mid-band behaviour, which are validated through circuit simulation and compared with experimentally measured responses of multiple HFCT sensors. This study also focuses on characterising the frequency behaviour of HFCT sensors by measuring secondary side frequency response and proving that the obtained results are comparable to those obtained by energising from the primary side. This integrated approach provides a consistent interpretation of HFCT behaviour, even with limited sensor design information, therefore supporting more reliable application in PD measurement systems. Moreover, it is possible to characterise the key parameters of an unknown sensor, so that analytical modelling becomes possible.

2. THEORETICAL MODEL

2.1. Operating Principle

On a fundamental level, HFCTs operate in the same way as conventional current transformers (CT), governed by Maxwell's equations. However, conventional CTs are typically designed for higher currents in nominal operation. Sensitivity is always important, but detection of the higher nominal currents is relatively easy in such sensing applications. Protection against magnetic saturation is a real concern for HFCTs. Thus, the magnetic circuit and the CT performance must be designed accordingly. This typically results in a relatively high turn ratio, with the typical nominal current output of 1 A [14] or 5 A [15, 16]. Bandwidth is of limited concern, especially in power-frequency applications (50/60 Hz). A suitable alternative for measuring high currents is, for example, sensors based on the Rogowski coil principle, which do not comprise a magnetic core and are thus inherently immune to magnetic saturation [17]. However, their sensitivity is inferior for the lower frequencies, and even more so for the higher bandwidth, which calls for significantly fewer sensing turns.

On the other hand, HFCTs are optimized to operate over a wideband range (or at least at high frequencies), typically spanning from several kHz to hundreds of MHz, depending on the exact

application. Specifically for PD, the current impulses of interest are very short (typically $< 1 \mu\text{s}$) and of relatively small amplitude, so high sensitivity is required of the HFCT, which means that the Rogowski-coil-type approach is not suitable for ordinary PD applications. So, HFCTs rely on electromagnetic induction between a single-turn primary conductor and a multi-turn secondary winding wound on a high-permeability magnetic core, typically with additional electrostatic shielding [8]. A rapidly changing primary current drives variation in the core magnetic flux, and thus a secondary voltage is induced proportional to the flux changes. Assuming linear magnetic permeability, this secondary voltage would be directly proportional to the derivative of the primary current. However, no magnetic materials have perfectly linear permeability, and thus, such direct proportionality cannot be assumed for a realistic sensor. Consequently, a perfectly flat bandwidth is impossible to achieve. With the classical approach, the secondary winding should be loaded with a suitable burden resistance, which provides a path for the secondary current, opposing changes in the magnetic flux in the core. This linearises the response of the magnetic circuit and provides the possibility of flattening the response versus frequency. Additionally, if the ratio of impedances is at the desired level, then the secondary current will become an integral of the induced secondary voltage, and thus the secondary current will become a facsimile of the primary current, scaled by the turn ratio (secondary N_2 to primary N_1 turns). In a typical application, $N_1 = 1$, so N_2 dictates the turn ratio. However, some HFCT can be used without a defined resistive termination, so that they are not operated in the *current transformer mode*, but rather in a *voltage transformer mode*, also acting as a narrow-band sensor [2,3]. For specific PD applications, this is permissible because the proportionality factor (primary to secondary) can be calibrated in situ, through dedicated external equipment, at the frequency of interest [18]. Ideal HFCT should operate completely in the current transformer mode over the full required bandwidth, with a flat frequency response.

It is not possible to provide a truly *full* equivalent model of an HFCT because of the distributed nature of all the parasitic parameters. In practice, these are not measurable, and thus cannot be known, and the lumped model (even very detailed) cannot represent the true behaviour, especially at higher frequencies. Models based on trans-resistance become very convoluted, as evident from Eq. (1) and Fig. 1. The complexity is prohibitive for extracting the key performance components, such as cut-off frequency limits [5].

Fig. 1. Detailed model of HFCT with parasitic components, where: R_{PRI} - primary winding resistance; L_{P_LEAK} - primary leakage inductance; L_{PRI} - primary winding inductance; L_{PW} - primary magnetising inductance; R_{PW} - magnetic core loss inductance; C_{PC1} , C_{PC2} - primary winding to core parasitic capacitance; C_{PS1} , C_{PS2} - primary winding to shield parasitic capacitance; R_{SEC} - secondary winding resistance; L_{S_LEAK} - secondary leakage inductance; L_{SEC} - secondary winding inductance; C_P - secondary parasitic capacitance, R_B - secondary terminating resistance; C_{W1} , C_{W2} - secondary winding inter-turn capacitance; C_{SC1} , C_{SC2} - secondary winding to core parasitic capacitance; C_{SS1} , C_{SS2} - secondary winding to shield parasitic capacitance;

$$R_m(s) = \frac{V_R(s)}{I(s)} = \frac{R_c}{nC} \frac{s}{L_l \left\{ s^3 + s^2 \left(\frac{L_m(L_l + R_s RC) + R_c RC(L_m + L_l)}{RCL_m L_l} \right) + s \left(\frac{L_m(R_s + R + R_c) + R_c(L_l + R_s RC)}{RCL_m L_l} \right) + \left(\frac{R_c(R_s + R)}{RCL_m L_l} \right) \right\}} \quad (\Omega) \quad (1)$$

Therefore, several simplifications must be made so that a reasonable representation of performance can be achieved, with the minimum number of lumped components [11]. For example, in HFCT, the primary turn represents a relatively very small surface area and thus its capacitance is negligible as compared to other larger capacitances, so it is justifiable to exclude it from the circuit [5].

Similarly, further simplifying assumptions can be made for analytical analysis of current measurement, as aimed at the small-signal analysis from the bandwidth perspective. It can be assumed that the primary current is driven from an ideal current source, and therefore any additional impedance in the primary circuit does not affect the measured current. Magnetic coupling coefficient k influences the amplitude of the secondary signal, but it has a smaller effect on other parameters (such as bandwidth). Therefore, it can be assumed to be unity (and indeed it often is in the literature, even if not explicitly stated). Furthermore, for the bandwidth of interest, it can also typically be that the equivalent core loss resistance is much smaller than the burden resistance $R_{CORE} \ll R_B$, and thus in the first approximation R_{CORE} can be assumed to be open-circuited.

These simplifications lead to an equivalent circuit, which represents only the key parameters that are important from the practical viewpoint of the usable frequency bandwidth. Such a circuit is important for parameter extraction, which can only be carried out for a limited number of these lumped components. If they can be measured or estimated by other means, like the LCR method [19], then they can be used. Otherwise, they could be set to a zero impedance (for series connections) or infinity (for parallel connections).

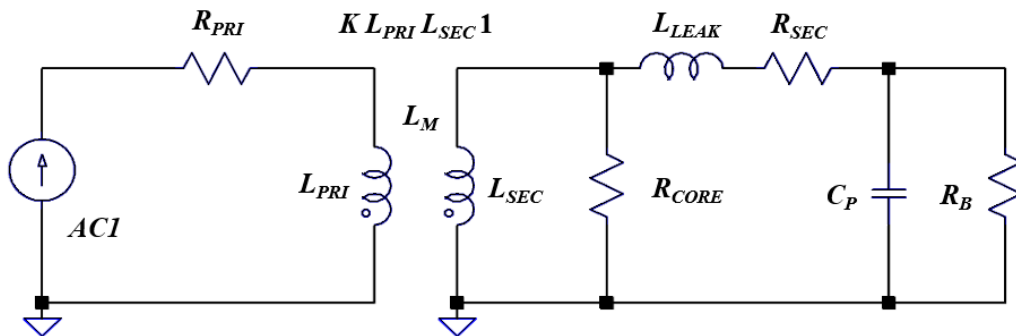


Fig. 2. Idealised and simplified HFCT equivalent lump circuit model for measurable quantities (from the secondary side), where: R_{PRI} - primary winding resistance; L_{PRI} - L_{SEC} primary and secondary

inductance; R_{CORE} - parallel core-loss resistance; L_{SEC} - secondary winding inductance; L_{LEAK} - secondary leakage inductance; R_{SEC} - secondary winding resistance; C_P - parasitic self-capacitance of the secondary winding; R_B - burden resistance.

A simplified equivalent circuit model is presented in Fig. 2 for wide-bandwidth analysis of the frequency-dependent behaviour of the HFCT. The model represents the behaviour of the primary to secondary side of the HFCT, with the primary current fluctuation reflected into the secondary circuit.

In such a circuit, the bandwidth is dictated by the interplay between the three main components: inductance of the secondary winding, resistance in the secondary circuit, and the largest parasitic capacitance. For any modelling, it is therefore necessary to estimate the various parameters, so that they can be represented correctly. This is especially important for a sensor without specified values, as many of these are typically not provided by the manufacturers [20]. However, certain parameters like core permeability μ_r or magnetising inductance L_M cannot be measured directly due to a lack of limitations of the available testing equipment or due to the interplay between the parasitic values.

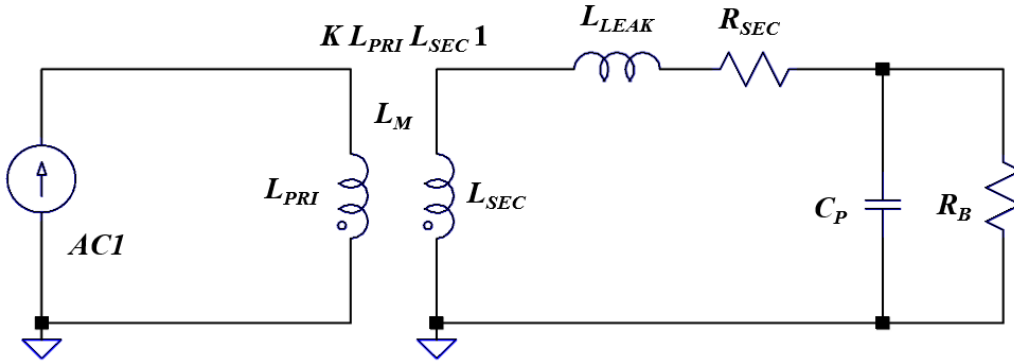


Fig. 3. Simplified HFCT equivalent lump circuit model for parameter measurement.

These parameters could be calculated if the input data were available. For example, core permeability dictates the value of inductance L associated with it, which in turn affects the low-frequency sensitivity, usable bandwidth, and linearity of the HFCT. The magnetising inductance of a toroidal core is dependent on the permeability parameter and is expressed as

$$L = \mu_0 \mu_r \frac{N^2 A}{l_m} \quad (H) \quad (2)$$

where: μ_0 - permeability of free space; μ_r - relative permeability of the core material; N - number of secondary turns; A - effective core cross-sectional area; l_m - mean magnetic path length.

However, for a generic sensor, the information about the core permeability is typically not provided by the manufacturer, and therefore, inductance cannot be simply calculated through Eq. (2). This necessitates the direct measurement approach, which is a relatively simple exercise as discussed below.

It should be noted that, generally, for HFCTs, a simple integration of the measured voltage pulse does not preserve the charge information unless the operation is in the flat band of the sensor. HFCT operation near the low or high frequency band limits converts an output voltage proportional to the rate of change of the primary current, i.e., a higher voltage is reproduced from a faster rate of change of primary current.

2.3. Theoretical Frequency Band Interpretation

The usable frequency band is dictated firstly by the lower cut-off frequency at the cross-over point $X_L = R_B$, which sets a point above which the current transformer mode (CT mode) becomes possible (Fig. 4). When the reactance X_L of the magnetising branch is increasing with rate of 20 dB/decade and its value is lower than the output impedance (set by the termination resistance R_B), the CT operates in voltage transformer mode. Because there is no direct identical representation of the current into the primary winding from the HFCT. Only when the X_L exceeds the termination resistance does the HFCT enter the desired CT mode of operation. There is a trade-off in the specific value of termination resistance: higher values yield greater CT gain and improved measurement sensitivity. This, however, is at the cost of usable bandwidth as lower resistance (Fig. 4, R_{B1}) gives a wider CT mode region. At higher frequencies, the capacitive reactance X_C of the secondary winding becomes a factor, limiting the roll-off point of the low-pass filter formed from the series resistance and inter-winding capacitance of the secondary side of the HFCT. The usable frequency band is dictated firstly by the lower cut-off frequency, which sets a point above which the *current transformer mode* (CT mode) becomes possible (Fig. 4).

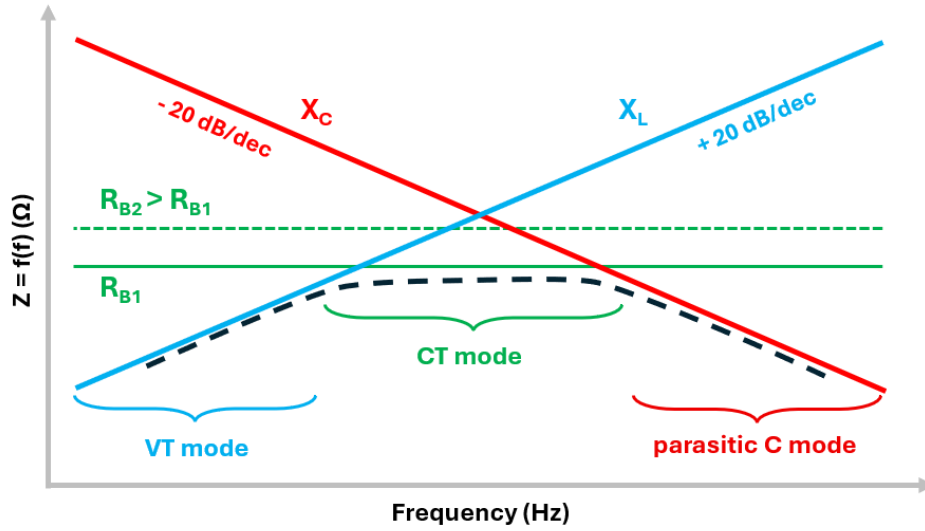


Fig. 4. Simplified view of frequency band interpretation - transfer impedance (Z) vs. frequency

Therefore, the appropriate determination of the key parameters for the model in Figs. 2 and 3 allows estimation of the useful bandwidth for the CT mode. Some of these parameters can be measured directly, some can be inferred indirectly, as described below.

2.4. Low-frequency Behaviour

At sufficiently low frequencies, the reactance of the secondary inductance L_{SEC} (X_{Lsec}) is smaller than the secondary load impedance (see also Fig. 3). This means that the presence or absence of R_B becomes negligible, with a sufficiently large ratio. The lower cut-off frequency occurs at the point at which the secondary reactance is equal to the total terminating resistance ($X_{Lsec} = R_{SEC} + R_B$) and could be approximated as:

$$f_L \approx \frac{R_B + R_{SEC}}{2\pi L_{SEC}} \quad (\text{Hz}) \quad (3)$$

where: R_B - output impedance; R_{SEC} - secondary winding resistance; L_{SEC} - magnetising inductance of the secondary winding.

It can be derived that the low-frequency limit is controlled mainly by the magnetising inductance of the core and the burden resistance (if other parasitics are negligible at that frequency). At this point, these two impedances are equal, so the phase shift is 45° , and the amplitude of the output signal is at the -3 dB point.

2.5. Mid-band Frequency Behaviour

This operating region is between the low and high frequency cut-off points ($f_L < f < f_H$), where the magnetising branch presents a high impedance and the effect of the stray capacitance is still negligible. In this mid-band region, an HFCT terminated with R_B behaves approximately as an ideal current-to-voltage converter, and the output voltage is proportional to the primary current. This voltage can be expressed as

$$V_{\text{out}}(t) \approx \frac{R_B}{N} I_P(t) \quad (\text{V}) \quad (4)$$

where: R_B - output impedance; I_P - primary current; N - turns ratio.

This mid-band region defines the usable operating bandwidth for PD detection and transient measurements that can be integrated for charge estimation with better accuracy. From Eq. (4), the HFCT response is directly influenced by the connected secondary resistance, and it is independent of frequency. The relationship is proportional to the value of R_B , meaning that higher R_B increases output voltage and thus increases sensitivity. However, increased sensitivity is automatically occupied with the reduction of usable bandwidth, as illustrated in Fig. 4 for $R_{B2} > R_{B1}$. Due to the effect of Eq. (3), the lower cut-off frequency shifts to a higher value proportionally to the increase of R_B (if other parameters remain the same).

2.6. High-frequency Operating Region

At high frequencies, the decrease of reactance of the stray or parasitic capacitance C_P becomes dominant, while at the same time, the inductance L_{SEC} (leakage inductance) introduces an increasing series impedance. The interaction of these R_C elements produces a second-order filtering to cut out the upper bandwidth, with only the R_{SEC} , R_B and C_P components becoming relevant. Therefore, the resulting upper cut-off frequency could be approximated by:

$$f_H \approx \frac{1}{2\pi C_P(R_B + R_{SEC})} \quad (\text{Hz}) \quad (5)$$

Thus, the high-frequency limit is set by the total resistance in the secondary circuit and the largest equivalent parasitic capacitance in the HFCT sensor design. Similarly, the point at which the capacitive and resistive impedances become equal is denoted by the 45° phase shift and -3 dB amplitude point.

2.7. Measurement of HFCT Parameters

Most of the key electrical parameters of a HFCT can be measured directly by employing a LCR bridge or meter, employing the small-signal approach and observing limitations due to bandwidth. It should be noted that during normal CT mode operation, the HFCT should be terminated with a burden resistor R_B , which can be built into the HFCT itself (and thus is impossible to disconnect). Therefore, measurement of electrical properties will reflect the actual configuration of the circuit, whether measured from the primary or secondary side. Bandwidth limitations have to be exploited with greater care to extract the parameters of interest.

For HFCT which are not terminated, the measurement is simpler because most of these quantities are available directly. The two secondary components, L_{SEC} and C_P , dictate the first secondary self-resonance of the winding. Therefore, a measurement with an LCR meter [21] must be performed at a frequency below the first resonance. Secondary winding resistance can be measured at a sufficiently lower frequency or even DC so that the effect of the secondary inductance L_{SEC} becomes negligible. In any case, for HFCT, the terminating resistance R_B is typically much larger than R_{SEC} and thus R_{SEC} can also be neglected, where justified. However, in this work, it was measured separately, and the relevant values are presented in Table 1.

The value of parasitic capacitance C_P of the secondary winding could be extracted from the slope of the scan of amplitude vs. frequency. However, as shown below (Fig. 8), at higher frequencies, such estimation could be quite unreliable due to all the other unknown parasitics, and therefore, a much clearer estimation can be obtained from the first resonant point. Once the inductance L_{SEC} is known (measured beforehand), then C_P can be calculated by observing that the resonance point will occur at the frequency of:

$$f_{res} \approx \frac{1}{2\pi\sqrt{L_{SEC}C_P}} \quad (\text{Hz}) \quad (6)$$

and thus, by rearranging Eq. (6), C_P can be extracted as:

$$C_P = \frac{1}{(2\pi f_{res})^2 \cdot L_{SEC}} \quad (\text{F}) \quad (7)$$

Another important HFCT parameter is the secondary winding series resistance (R_{DC}). The measurement of winding resistance can be carried out at DC. This will obviously ignore any increase of resistance due to skin effect and proximity loss [22], but measurement at elevated frequency is not quite possible due to the leakage inductance and other parasitics that would falsify the readings. However, these measured values allow the incorporation of the *minimum* values that will be present in the circuit, as the AC resistance can only increase, not decrease. Therefore, for the measurement of the *minimum* value of R_{DC} , the measurement at DC can be justified. This was carried out with a 4-wire low-resistance ohmmeter [23].

The only known parameter was the value of the terminating resistor, as it was an external component with a low-inductive design (thick film resistor).

Five different current transformers were used for their parameter evaluation, as shown in Fig. 5, and their measured parameters are shown in Table 1.



Fig. 5. Evaluated current transformers: a) HFCT1; b) HFCT2; c) HFCT3; d) HFCT4; e) HFCT5

3. SECONDARY SIDE CHARACTERISATION AND MODEL PARAMETERISATION

3.1. Secondary Side Characterisation

The HFCT sensor secondary winding has a self-resonance frequency point associated with the high-frequency behaviour and appears around the low-pass filter (LPF) side (near the 3 dB point) and upper cut-off region of the sensor.

A range frequency sweep of the secondary winding with the primary side open circuit is enough to determine this resonant frequency point. After that, the measured secondary inductance is then used to calculate the primary inductance for a known turns ratio. If the turns ratio is unknown, then it can be approximated from the secondary to primary voltage ratio, which then can be used to estimate the primary inductance.

$$L_{PRI} = \frac{L_{SEC}}{N^2} \quad (H) \quad (7)$$

where: L_{SEC} - secondary inductance; N - turns ratio.

Although the primary resistance is directly measurable, in this study, it is assumed to be zero. This is because its actual value strongly depends on the wire that is used for the primary conductor. However, due to the abovementioned assumption and that the signal is driven from a current source, the resistance of the primary winding does not affect the primary current, by definition, and therefore there is no need to include it in the model (i.e., it is assumed that $R_{PRI} = \text{zero}$). Nevertheless, the information about the primary inductance is strictly necessary because it dictates the transformer ratio due to the ratio of inductances between secondary and primary windings.

3.2. Sensor sample Evaluation and Parameter Extraction

The analytical expressions listed above allowed for an initial assessment of the HFCT bandwidth and its circuit parameter dependence. A selection of sensors was measured to obtain parameters as shown in Table 1. The secondary inductance for each sensor was recorded at 100 kHz with the primary side open circuit. A frequency response sweep was applied for a range of 10 Hz to 30 MHz in an octave step from a signal generator with a 20 V peak-to-peak sinusoidal waveform. The secondary side current was measured at each frequency step, as well as the voltage across the secondary inductance, to obtain the phase relation for the frequency range. Sensor HFCT5 was terminated with an unknown internal resistance. All other sensors were measured with the secondary winding open-circuited (unless stated otherwise).

Sensor	Description	$R_{SEC} (\Omega)$	$L_{SEC} (H)$	N (-)	$f_R (Hz)$	$L_{PRI} (H)$	$C_P (F)$
HFCT1	Coilcraft CS1050L (3 kHz to 1 MHz)	0.53	9.33×10^{-3}	50	200×10^3	3.73×10^{-6}	67.9×10^{-12}
HFCT2	Megger CT100 (2 MHz to 25 MHz)	0.02	4.11×10^{-6}	10	8×10^6	41.1×10^{-9}	96.3×10^{-12}
HFCT3	Megger MCC1010 (15 Hz to 10 kHz)	5.76	4.00	1000	8×10^3	4.00×10^{-6}	98.9×10^{-12}
HFCT4	Custom HFCT (toroidal core)	1.08	150×10^{-6}	5	800×10^3	6.00×10^{-6}	2.64×10^{-12}
HFCT5	Stangenes 0.5 – 1.0 (terminated internally with 50 Ω)	0.37	2.00×10^{-6}	50	200×10^3	0.56×10^{-9}	3.17×10^{-9}

Table 1. Measured and derived parameters under termination with a frequency sweep on the secondary side and L_{PRI} and C_P are calculated values.

The measured secondary side current and voltages allowed calculating the secondary side impedance Z from Ohm's law. The plotted response for each sensor was used to identify the resonant frequency point of the secondary side (point at which phase shift = 0°), where $X_L = X_C$. This information was then used to calculate the largest value of the lumped parasitic capacitance. The primary inductance was derived from the number of turns and the measured secondary inductance value, as shown in Table 1. Further cross-correlation using Eq. (6) confirmed that inductive X_L and capacitive X_C reactances are equal at the measured resonant frequency point.

4. EXPERIMENTAL VALIDATION AND COMPARATIVE RESPONSE ANALYSIS

This secondary side indirect method is used to obtain the necessary parameters to set the low and high cut-off frequencies that form the bandpass behaviour of the sensor. The initial tests identified the resonant point for each sensor. This allowed for parameter estimation that was later used in the simulation model to verify that the frequencies agree with the model. Also, ordinary primary to secondary tests were undertaken to show how the coupling parameters modify the sensor response, and to prove that the overall behaviour is equivalent regardless of whether measured from the primary or from the secondary side.

4.1. Secondary Side Small-signal Excitation

The frequency response of the first HFCT1 is shown in Fig. 6. The unterminated sensor exhibited resonance at 200 kHz and the calculated primary inductance and parasitic capacitance for HFCT1 are shown in Table 1. The secondary winding of the sensor was loaded with R_B of 1 k Ω and 50 Ω to show that the bandwidth is modified proportionally to the loading resistance. As evident from Fig. 6, the measured impedance agrees with the expected value. This also signifies that the resonant point is heavily obscured due to the resistive damping, and if the sensor is internally loaded by design, the detection of resonant frequencies for the secondary side is not easy to extract. From another perspective, the graphs show that the loading of the secondary side is a trade-off, as lower terminating resistance widens the bandwidth of the sensor. That is at the expense of signal gain, which is evidently higher for the higher resistance of 1 k Ω . That means, the design of a HFCT is a trade-off between higher sensitivity and narrower bandwidth (higher value of R_B) or lower sensitivity and wider bandwidth (lower value of R_B), as illustrated in Fig. 4.

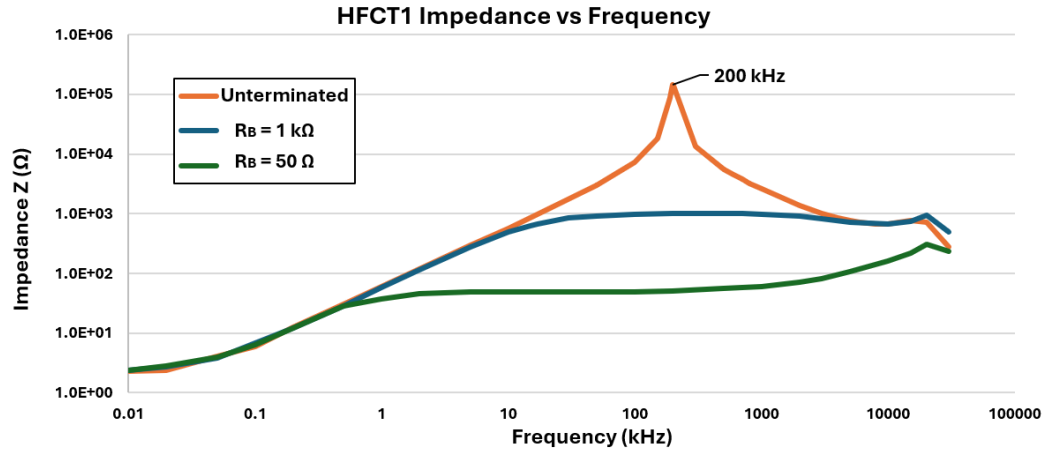


Fig. 6. HFCT1 Impedance response measured across the secondary for unterminated sensor; $R_B = 1 \text{ k}\Omega$; and $R_B = 50 \text{ }\Omega$. Note that the primary winding was physically absent during these tests.

The HFCT1 phase response in Fig. 7 shows how the response on the unterminated sensor exhibits a full -90° to $+90^\circ$ inversion at the resonant frequency, while when loaded, the phase remains at 0° for a large proportion of the bandwidth span of the sensor. In these tests, the signals were measured in such a way that -90° phase corresponded to an inductive phase shift.

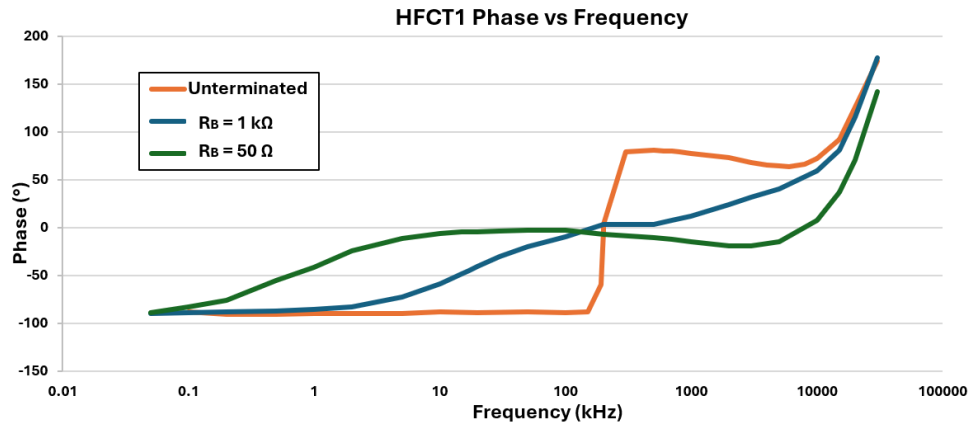


Figure 7. HFCT1 phase response at different loading conditions.

The unloaded response of HFCT2 – HFCT5 is shown in Fig. 8 with their respective frequency response. Similarly to HFCT1, the resonant frequency and respective calculation were recorded.

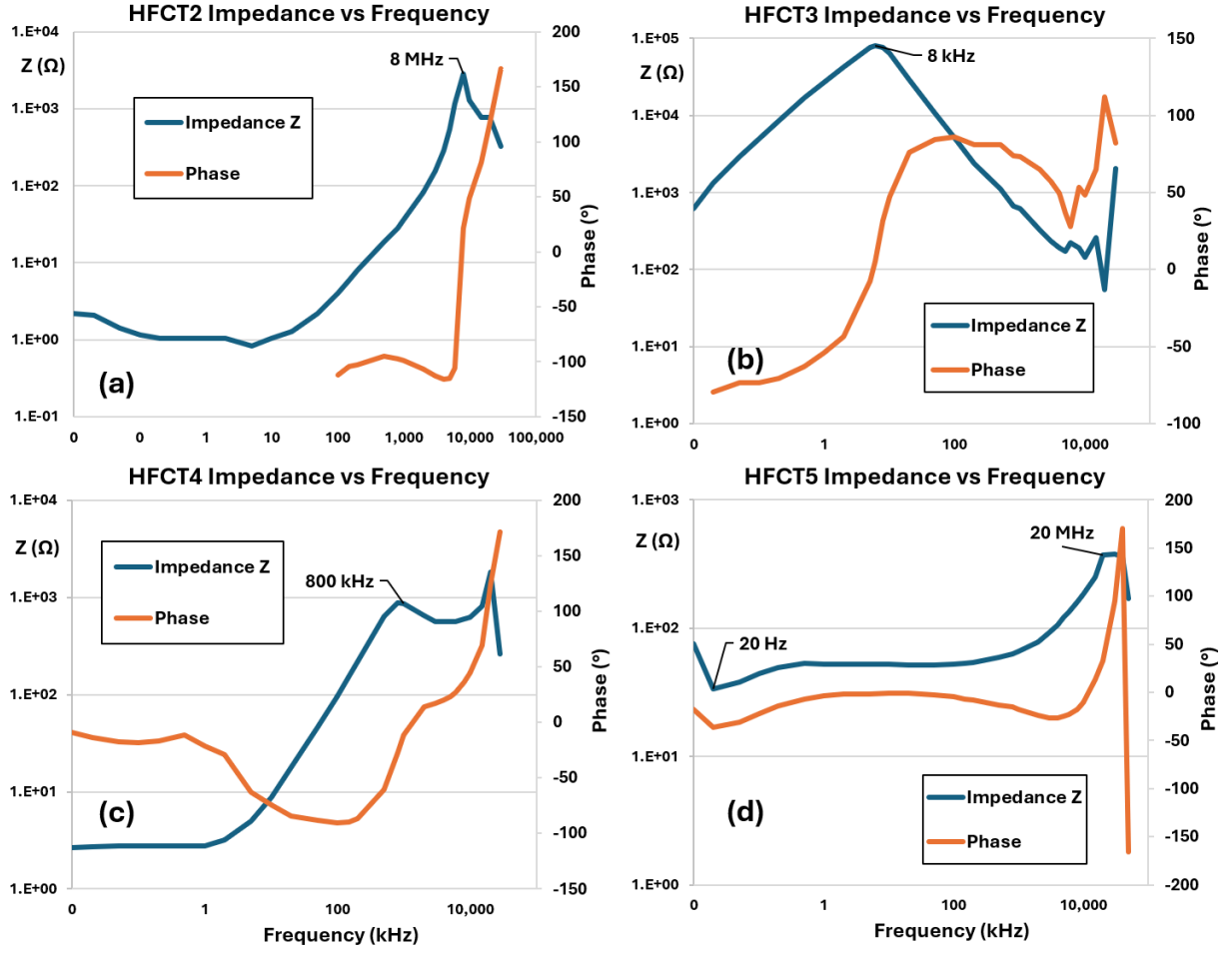


Figure 8. Frequency response of (a) HFCT 2, (b) HFCT 3, (c) HFCT 4 and (d) HFCT 5.

It should be noted that for the sensors with few secondary turns (≤ 10), the measurement of impedance at low frequencies became impossible due to the noise floor of the measurement system. For sensors HFCT2 and HFCT4, the measurable +20 dB per decade response was only possible significantly above 10 kHz.

Also, for these two unterminated sensors (i.e., HFCT2 and HFCT4), the first resonant point occurred at 8 MHz and 0.8 MHz, respectively. At these MHz-range frequencies the interplay of other parasitic capacitances was more likely and thus there was no “clean” -20 dB per decade decrease of amplitude, as it would be expected from Fig. 4. For the same reason, it is therefore impossible to extract the information about the capacitive reactance from the slope, and thus the first resonant point is the only method for calculation of the lumped value of C_P , by means of Eq. (7).

The HFCT5 sensor was a special case, as it was internally terminated with unknown resistance. As can be seen from Fig. 8d, the amplitude response stays flat from 200 Hz to 1 MHz, which is an impressive span. From the value of impedance (Z) and the fact that the phase remains close to 0° , it is therefore possible to estimate the value of the equivalent terminating resistance as 50Ω , as listed in Table 1.

4.2. Primary-to-secondary Response Analysis

The five sensors were then tested with their actual coupling by injecting a $50\ \Omega$ -terminated signal on the primary side while measuring the secondary side response. A Bode plot of the response verified the resonant frequency range as shown in Figures 9 and 10.

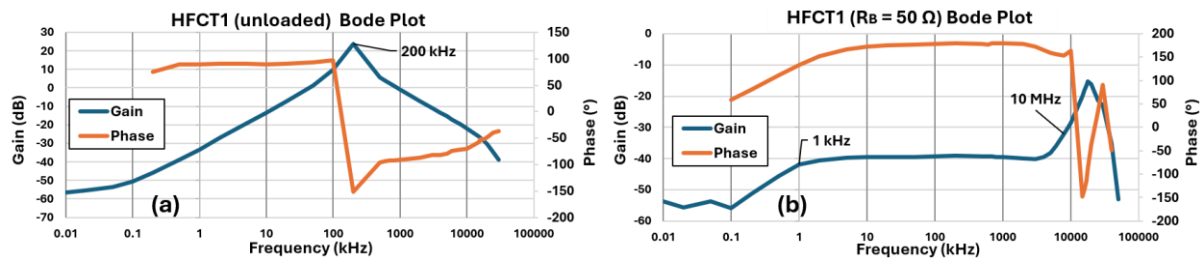


Figure 9. Sensor HFCT1 measured primary side to secondary side frequency response: (a) unterminated sensor, and (b) $R_B = 50\ \Omega$ termination.

The unterminated resonant point at 200 kHz for HFCT1 is identical to the results obtained from the secondary side tests, as shown in Fig. 2. The $50\ \Omega$ loaded bandwidth is also in agreement with this sensor to further verify the method's robustness. Fig. 10 shows that the recorded frequencies are also in agreement with the secondary side testing.

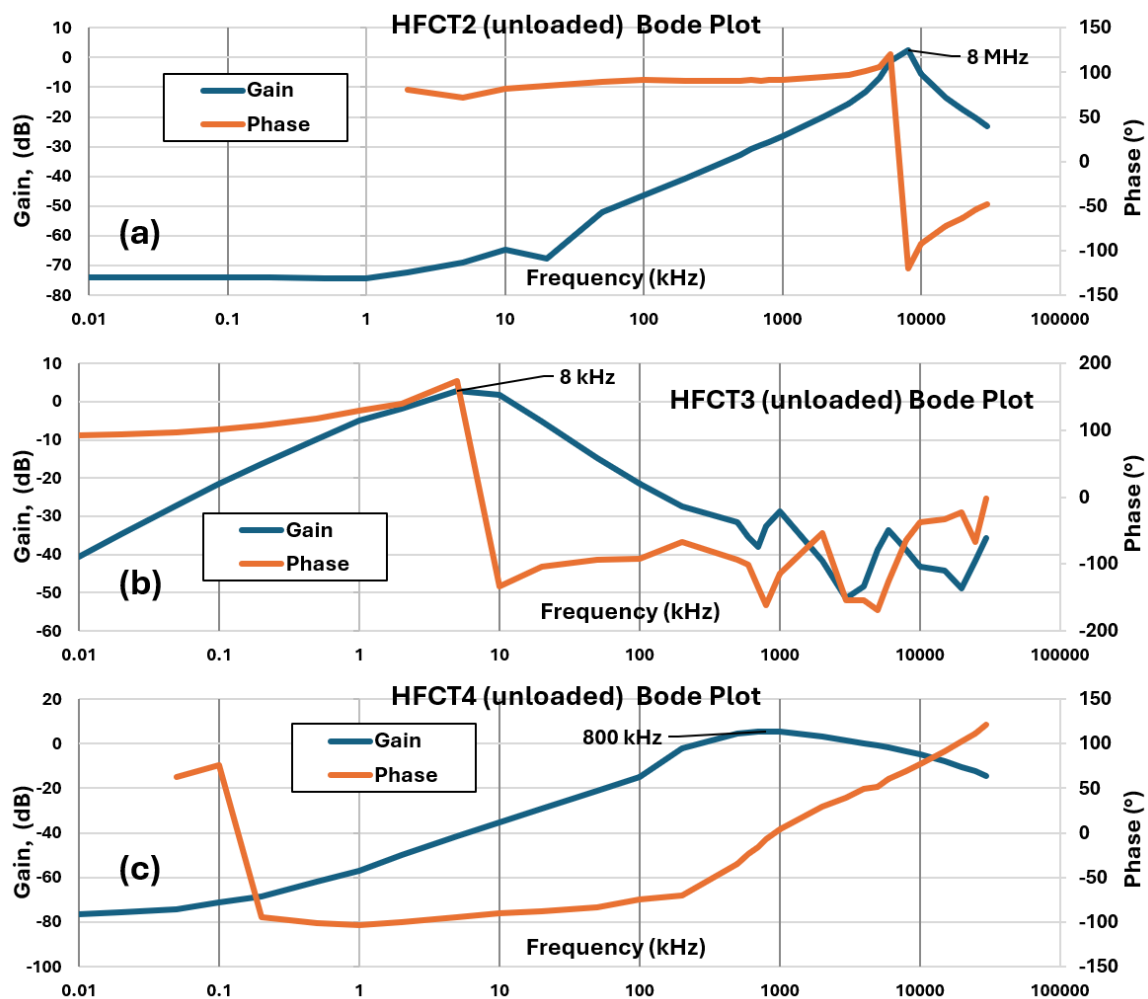


Figure 10. Primary to secondary side frequency response: (a) HFCT2, (b) HFCT3 and (c) HFCT4.

4.3. Secondary Side Evaluation through a Simulation

To further verify the sensor samples' response, the extracted resonant frequencies and parasitic component calculated values were set into the HFCT model for simulation. For simplification, the model does not comprise the magnetic core parameters and the coupling coefficient was set for perfect coupling ($k = 1$) as these effects were not measured. Their inclusion provides only a minor shift in the frequency response of the sensors. The model in Fig. 11 is used to simulate parameters of each HFCT sensor, but with values adjusted accordingly. The values as shown in Fig. 11 were derived for HFCT1 as applied in Table 1. This method was followed for the remaining sensors, except for HFCT5, due to it being internally terminated.

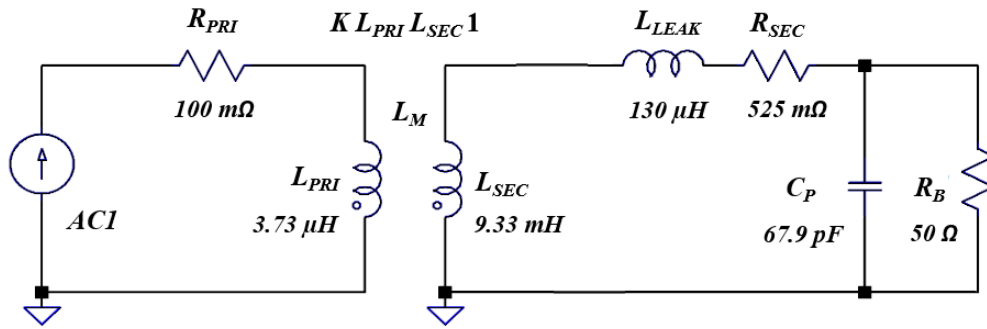


Figure 11. HFCT model simulation (HFCT1 values as calculated): L_{SEC} - secondary leakage inductance; R_{SEC} - secondary winding resistance; C_P - parasitic capacitance; R_B - burden resistor.

This final verification of the HFCT1 sensor parameters is shown in Fig. 12. The results were plotted without termination (a), with 1 kΩ (b) and 50 Ω (c) to emphasize how the burden resistance affects the sensitivity and bandwidth of the sensor. The recorded high-pass cut-off threshold is comparable to the -3 dB point of the data from Fig. 6. The simulations cover a span from 1 Hz to 10 GHz.

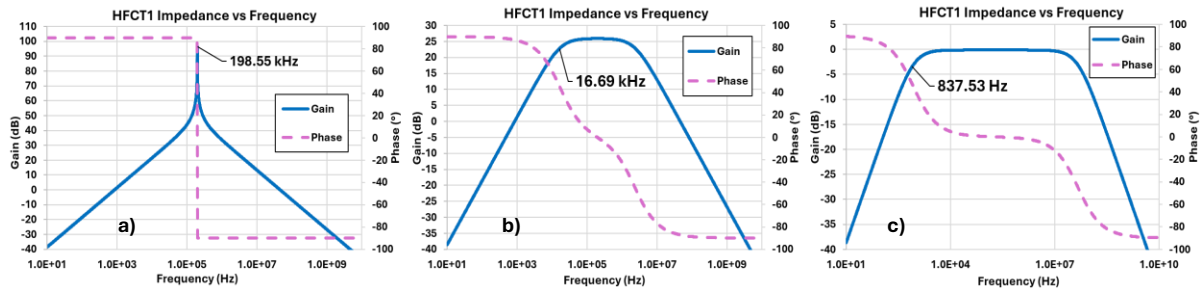


Figure 12. Impedance response for HFCT1 sensor against frequency with measured and calculated parameters: (a) unterminated (b) $R_B = 1 \text{ k}\Omega$ and (c) $R_B = 50 \Omega$.

Sensors HFCT2 - 4 were simulated and the responses are shown in Fig. 13. The simulated results were also very close to the extracted resonant value from the small-signal secondary winding response tests.

Although the terminated simulations are not shown here for brevity but obviously behave qualitatively the same as those illustrated in Fig. 6 and shown in Fig. 12. These experiments and simulations show that derivation of parameters just from the secondary side of an unterminated current transformer is a valid method for estimation of the useful frequency bandwidth of a sensor with otherwise unknown parameters and performance. The simplified lumped model is sufficient to simulate the realistic performance range of HFCTs.

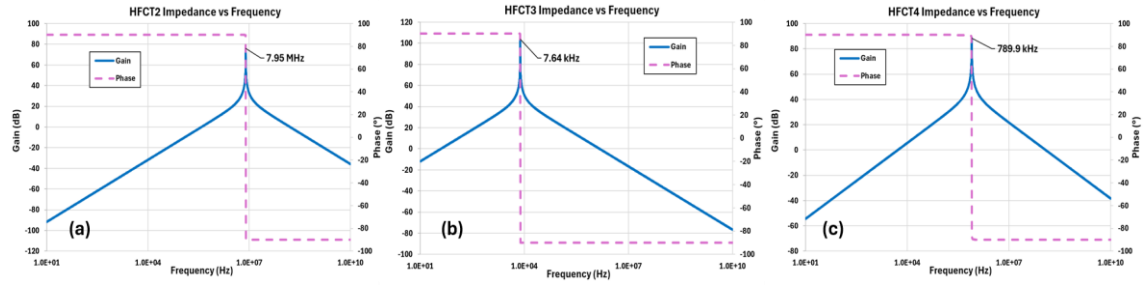


Figure 13. Impedance response for unterminated sensors against frequency: (a) HFCT2, (b) HFCT3 and (c) HFCT4.

Specifically for PD measurements, the frequency range and the required sensitivity can then be adjusted as required. Frequency bandwidth exceeding 1 MHz can be obtained, as needed for the PD measurements. As concluded above, the achievable sensitivity is a compromise against the required bandwidth response for a given sensor design. Clearly, higher frequencies are limited by the stray capacitance, so this would have to be minimised at the design stage, as it is not possible to mitigate its effects *a posteriori*.

5. CONCLUSIONS

This study investigated the key characteristics of high-frequency current transformers and showed how the cut-off frequency limits are determined by the secondary winding parameters, the secondary self-resonance, the output impedance termination, and the inter-turn parasitic (total stray) capacitance. It demonstrated that parameter evaluation carried out solely on the secondary side can provide sufficient information about sensor performance, with comparative tests and simulations showing very good agreement. Analytical expressions linking the sensor parameters to bandwidth characteristics were derived and experimentally validated using five HFCT sensors with different designs. The proposed secondary side characterisation was consistent with conventional primary-side measurements, and SPICE simulations closely matched the measured responses. The experiments confirmed that the observed bandwidth shifts and simulated results validate both the simplified sensor model and the physical interpretation of the high-frequency current transformer frequency response.

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