Towards nonlinearity measurement and simulation using common EMC equipment

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***Abstract*—Integrated circuit (IC) models that predict functional failure are necessary for predicting the immunity of systems to electromagnetic interference (EMI). The integrated circuit immunity model for conducted immunity (ICIM-CI) of IEC 62433-4 assumes that the IC terminals still behave linearly at injection power levels that cause susceptibility. This hypothesis should be systematically verified when modelling integrated circuits for EMC, but this is not always straightforward.**

A simple measurement set-up using a directional coupler and a spectrum analyser is demonstrated to verify this linearity hypothesis using commonly available equipment.

The measured reflected spectrum can be transformed into the $|X_{11}|$ parameter, which is the non-linear extension of the S_{11} parameter.[1] X-parameters may be the key to predict susceptibility by simulation when the linearity hypothesis is invalid.

***Index Terms*—immunity, integrated circuit, ICIM-CI, DPI, modeling, linearity hypothesis, X-parameters**

I. Introduction

To reduce the number of prototyping cycles in the development of electronic products, we would like to predict the immunity of a product to electromagnetic interference (EMI) in an early stage. Specifically, we would like to be able to predict the functional failure of a product under test using a circuit simulator. In order to be able to simulate the complete product, we need component models that adequately predict functional failure. Models of passive components are available [1], [2], so we focus on integrated circuit (IC) models.

One class of models is called 'black box'; these models are extracted from measurements on real integrated circuits, without specific knowledge about its construction. As the models only contain functional information, manufacturers can distribute these models without risking counterfeits. If the IC manufacturer does not provide these models, a third party can obtain them by simple measurement of the real device. These practical advantages of black box models make them particularly suited for industrial development.

Such a black box model was proposed by Lafon c.s. [3] and is currently in the process of standardisation as IEC 62433-4 [4]. The model is called 'integrated circuit immunity model for conducted immunity' (ICIM-CI).

[1]"X-parameters" is a registered trademark of Agilent Technologies. The X-parameter format and underlying equations are open and documented.

In section II, we briefly recall the ICIM-CI proposed in IEC 62433-4 and state the underlying linearity hypothesis. Then, in section III, we show that it is not always straightforward to validate the linearity hypothesis. In section IV, we have a look at the reflected harmonics, which reveal the non-linear behaviour of electrical ports, which can be summarised in X-parameters. It turns out, in section V, that we can measure the $|X_{11}|$ parameter with readily available laboratory equipment. We use this set-up in section VI to validate the linearity hypothesis in the case of a voltage regulator.

As a future perspective, we outline in section VII how X-parameters could allow the immunity simulation of systems consisting of multiple ICs. We conclude in section VIII that this may be the way to model non-linear passive distribution networks (PDNs), where it is necessary to obtain realistic immunity predictions.

II. Immunity Modelling

The ICIM-CI proposed in IEC 62433-4 predicts functional failure of an IC under continuous wave (CW) disturbances entering the IC pins. To do so, the model consists of a PDN and an immunity behavioural (IB) part, see Figure 1.

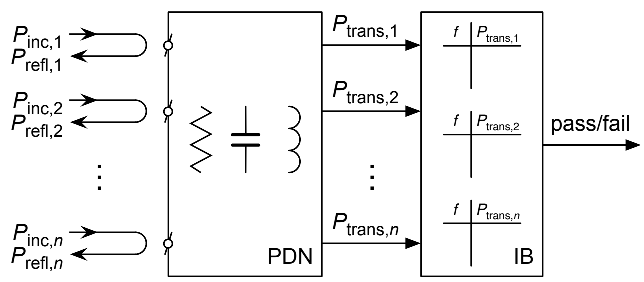

Figure 1. Symbolic representation of the ICIM-CI of IEC 62433-4. In this case, the IB just returns a boolean pass/fail signal, based on transmitted power threshold lookup tables.

The PDN is a linear multiport, of which every port represents a ground-referenced IC pin. Using the PDN, the reflected and transmitted power can be calculated, given an incident power. The PDN can be completely described with S-parameters, which can be measured using a network analyser.

The IB commonly consists of a look-up table for every port that yields a transmitted power threshold as function of the disturbance frequency. If the transmitted power through any port exceeds the threshold, the IC is predicted to fail altogether.[2] The transmitted power threshold can be measured using the direct power injection (DPI) set-up described in IEC 62132-4 [5] (Figure 2). That is, for every frequency, the incident power is increased until the IC fails (definition and measurement of failure are outside the scope of this article). At that point, the difference between the incident and reflected power is taken to be the transmitted power threshold.

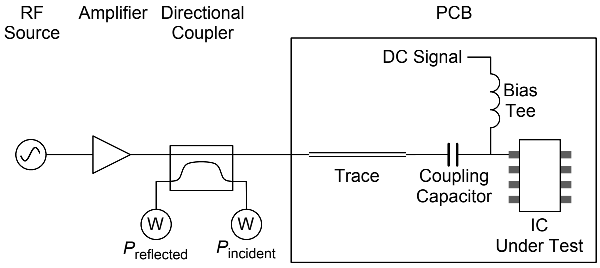

Figure 2. RF injection path of the DPI set-up proposed in IEC 62132-4.

At some incident power, all IC pins start to demonstrate considerable non-linear behaviour, chiefly due to ESD protection diodes. This challenges the model validity of a linear PDN. However, as long as failure always occurs at power levels where the IC still can be considered linear, the linear PDN is a competent model.

Let us thus formulate the linearity hypothesis: “At the incident power that causes failure, the current trough the IC pin can be considered to a linear function of its voltage with respect to the ground.” Note that ‘failure’ must be defined to validate the linearity hypothesis. Hence, different failure criteria may yield valid or invalid linearity hypotheses. ‘Can be considered’ means that the model yields system level immunity predictions that are accurate enough.

The linearity hypothesis was validated for many devices using realistic failure criteria [6], [2], such as bus transceivers, general-purpose logic and voltage regulators. However, there are also cases where the linearity hypothesis might be invalid, such as the FlexRay transceiver studied by Hilger c.s., where the root cause of susceptibility was the ESD-protection, which converted the common mode disturbance to a differential mode disturbance [7]. In this case, a linear model may give adequate predictions in particular configurations, but has no physical sense anymore and should, in general, be used with caution.

Clearly, the linearity hypothesis should be systematically verified. Furthermore, non-linear behaviour can reveal the root cause of susceptibility. This insight may also help to resolve the problem, be it at IC level or at printed circuit board (PCB) level.

[2]Other immunity behaviours are envisaged in IEC 62132-4 as well, where a scalar performance metric (like DC offset or jitter) is returned.

III. Measurement Difficulties

To evaluate linearity measurement methods, let us consider the IC port modeled in Figure 3. The bond wire is modeled as inductance, there are ESD protection diodes to ground and power supply, followed by the gate of a MOSFET (metal-oxide-semiconductor field effect transistor). This transistor exhibits, amongst others, a Miller capacitance to its load, which comprises a parasitic drain-substrate diode in reverse.

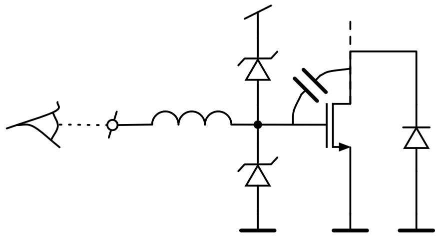

Figure 3. Interesting IC port, which has multiple non-linearities: the ESD protection (activation around $\pm 10\,\text{VDC}$) and a parasitic reverse diode (activation around $-0.7\,\text{VDC}$).

The simplest way to assess linearity is an $I(V)$ sweep; while slowly varying the voltage on the port, we measure the current flowing into the port. For low frequencies, the series inductance becomes a short circuit, while the parasitic capacitance becomes an open circuit. Hence, the behaviour of the ESD protection is observed (e.g. at $10\,\text{VDC}$). However, the parasitic diode (e.g. with a $0.7\,\text{VDC}$ threshold) remains hidden.

Another way to obtain an $I(V)$ curve is the transmission line pulse (TLP), which applies a brief electrical pulse, while measuring voltage and current. As the pulse is short in time, it is broad in frequency and we can hope to observe the parasitic diode. However, the pulse spectrum is not very well controlled. As a result, it is difficult if not impossible to find an incident power threshold for linearity as a function of frequency.

A more controlled way of revealing non-linearities is making use of a vector network analyser (VNA). While sweeping the incident frequency, a VNA measures the scattered (reflected) power, which can be converted to an impedance. As long as the impedance is independent of the incident power, by definition, the port is considered linear. Practically, one can increase the incident power applied by the VNA until the impedance starts to change. This incident power can then be taken as a global linearity threshold; comparison with the DPI results can validate or invalidate the linearity hypothesis.[3] This is the main advantage of using the VNA: it directly yields a power threshold for linearity.

Although a VNA can work, the necessary power to validate the linearity hypothesis can not always be delivered

[3]As soon as the impedance starts to change as function of the incident power, surely the port exhibits non-linear behaviour. The converse, however, is not strictly true. One can construct networks where the power lost in reflected harmonics (not observed by a VNA) is exactly compensated by less loss in a resistive load. This exact compensation is very unlikely, so we simplify by supposing that the converse is also true.

by the VNA. For example, some LIN transceivers need over +20 dBm at 1 GHz of transmitted power to fail [3]. Being unmatched, the incident power necessary to cause failure may reach +40 dBm, while common VNAs deliver up to +10 dBm. Furthermore, a VNA only gives a global power threshold (which may be very conservative for some frequencies) and gives little insight in the nature of the non-linearities.

IV. Large Signal Network Analysis

Large-signal vector network analysers (LSVNAs or LSNAs) promise to accurately measure and give insight in high-frequency non-linear behaviour. Although their price is often prohibitive (in the order of 100 kUSD), and their output power is still insufficient (about +15 dBm), we can try to learn from their operating principles.

The response of linear systems to multiple summed excitations is the sum of the responses to the individual excitations. This so called superposition principle is exploited by VNAs: different frequencies are applied one at a time while measuring the responses. Because the system is linear, the response must have the same frequency as the excitation. When these responses are known, the system is fully characterised.

This does not hold for systems that must be considered non-linear. An LSNA, therefore, can apply a combination of frequencies and it measures the full response, which generally contains harmonics and intermodulation products. Like in [8], we restrict ourselves to harmonics and adopt the notation illustrated in Figure 4: incident harmonics are denoted A_{pn} and reflected harmonics are denoted B_{pn}, where p is the port number and n is the harmonic index. Harmonic index 1 indicates the fundamental frequency f, index 2 indicates the second harmonic $2f$ and so on. Harmonic index 0 indicates the 0 Hz or direct current (DC) component.

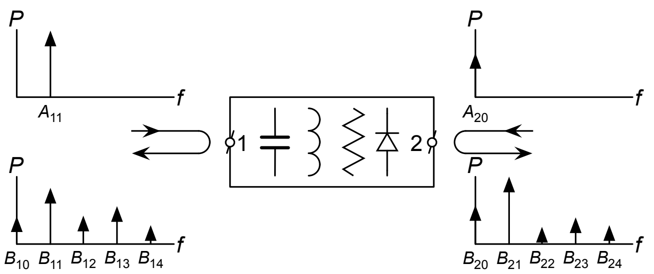

Figure 4. Notation of incident and reflected harmonics on a non-linear two-port. In this case, only a fundamental frequency is supplied to port 1, while bias is applied to port 2.

Analogous to S-parameters, X-parameters describe the reflected and transmitted harmonics under single-frequency excitation. X-parameters are a superset of S-parameters; the X-parameters of a system that can be considered linear reduce to its S-parameters, because the whole power is contained in the reflected and transmitted fundamentals. The more a system behaves non-linearly, the more power will appear in the reflected and transmitted higher harmonics and at DC.

A common measure for non-linearity is the total harmonic distortion (THD), which is the ratio between the effective voltage of the higher harmonics and the fundamental:

$$\mathrm{THD} = \frac{\sqrt{\sum_{n=2}^{\infty} V_n^2}}{V_1} = \frac{\sqrt{\sum_{n=2}^{\infty} P_n Z_0}}{\sqrt{P_1 Z_0}} = \sqrt{\frac{\sum_{n=2}^{\infty} P_n}{P_1} \cdot \frac{\cancel{Z_0}}{\cancel{Z_0}}}, \quad (1)$$

where V_n is the root mean square (RMS) or effective voltage of the nth harmonic, P_n is the power in watts of the nth harmonic, $n = 1$ being the fundamental and Z_0 is the real reference impedance. As the power generally decreases with the harmonic index, taking more harmonics is not always necessary, given a wanted precision. The DC power P_0 is not taken into account when calculating the THD. To get a feel for the meaning of THD, three waveforms are plotted in Figure 5 with different THDs.

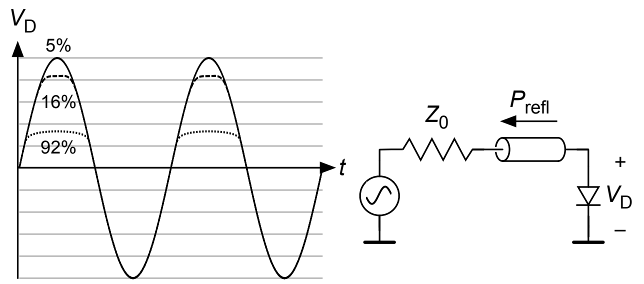

Figure 5. The diode voltage waveforms corresponding with 5%, 16% and 92% THD in the P_{refl} spectrum of a diode. The waveforms were obtained by an LTspice simulation on the same diode, while varying the input power and measuring the THD. For comparison of the shape, the waveforms are normalised to overlap, which is allowed, because the THD is a voltage ratio.

V. Measurement Set-up

Working towards simulation using black box X-parameters, we construct a measurement set-up to measure the $|X_{11}|$ parameter of a device under test (DUT). That is, we measure the magnitude of the reflected harmonics, including the 0 Hz component. (Note that the spectrum analyser does not return the phase of the reflected harmonics.) The essential measurement set-up is depicted in Figure 6 and basically replaces the wattmeter in the DPI set-up with a spectrum analyser, as can be found in any EMC laboratory. As the spectrum analyser is not able to measure at 0 Hz, we add a voltmeter decoupled by an inductor.

Power amplifiers for EMC often comprise an internal directional coupler, as shown in Figure 6. We would like to use this internal directional coupler, as it is guaranteed to withstand the power generated by the amplifier. These couplers normally have a specified, flat coupling over the functional bandwidth of the amplifier. For example, the Prâna AP30DT120 amplifier (10 kHz–1 GHz) has a nominal reverse coupling of −49 dB. However, when we use this amplifier to inject 1 GHz into our non-linear DUT, the reflected harmonics will be at 2 GHz, 3 GHz, 4 GHz and so on. In order to know how many reflected harmonics we can observe, we remove the directional coupler

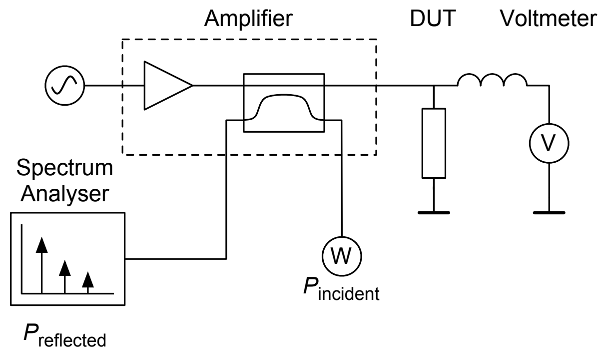

Figure 6. Measurement set-up to measure $|X_{11}|$, using a spectrum analyser, multimeter and amplifier with built-in directional coupler.

from the amplifier and measure the reverse coupling. The results for the Prâna amplifier, a Milmega AS0104-30/30 amplifier (two bands) and an external Agilent 86205A directional bridge are plotted in Figure 7.

If we accept ± 5 dB deviation from the nominal coupling, we conclude that we can always observe the second harmonic $2f$ while using the amplifiers at their maximum frequency. If we inject a frequency at the lower end of the amplifier's operating bandwidth, we can observe up to $4f$ with the Milmega amplifier and even $200{,}000f$ with the Prâna amplifier (2 GHz/10 kHz). We can also use the external directional bridge for a maximum number of harmonics, although its power handling is limited to $+25$ dBm.

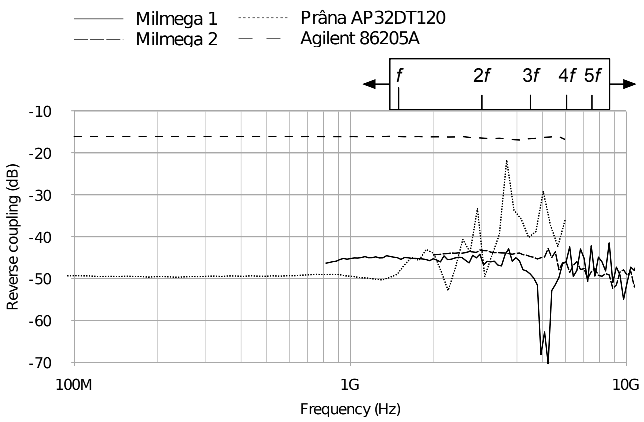

Figure 7. Reverse directional coupler of a Prâna amplifier (10 kHz–1 GHz), a Milmega AS0104-30/30 amplifier (band 1: 1–2 GHz, band 2: 2–4 GHz) and an external Agilent 86205A directional bridge (300 kHz–6 GHz). Notice the ruler in the top-right corner, which can be slid along the frequency axis, to find the harmonics of a given fundamental frequency.

VI. Case Studies

We will now demonstrate the application of the measurement set-up. First we use a known, non-linear device to show the measurement principle. Then, we use a voltage regulator at the brink of susceptibility to test the linearity hypothesis.

A. Diode

The simplest non-linear device may well be the diode, so we use a BAV21 diode at hand as first case. To assess the linearity of the other components in the set-up, we first place an open circuit and a short, refer to Figure 8. Furthermore, we use a HP 89441A vector signal analyser (VSA) instead of a spectrum analyser, because it can also measure the relative phase between the frequencies in its 10 MHz intermediate frequency (IF) bandwidth. By synchronising the VSA to the 10 MHz reference frequency of the generator, we can even estimate the phase differences between DUTs.

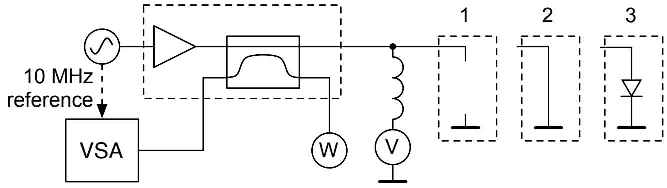

Figure 8. Measurement set-up with synchronised VSA and three DUTs: an open circuit, a short circuit and a diode.

To be able to see 9 harmonics, we decide to inject 1 MHz and tune the VSA to 5 MHz. This means that the span becomes 0–10 MHz, within which we can measure amplitude and relative phase. We use the Prâna amplifier, which has -49 dB nominal reverse coupling.

The spectra reflected by an open circuit and a short circuit upon $+30$ dBm incident power are plotted in Figure 10a-10b. Assuming that these standards really have reflection coefficients $\Gamma = 1$ and $\Gamma = -1$, respectively, the amplitude differences originate in amplifier gain and reverse coupling variations. Between the two loads, there is 2 dB difference, which we deem acceptable. We note that there is about 50 dB dynamic range between the fundamental and noise floor or harmonics that could stem from non-linearities in the set-up. Note that the phase difference between the fundamental reflected by the short circuit and by the open circuit is $-177°$, close to the expected $\pm 180°$.

Figure 9. Diode mount with the diode soldered directly to a BNC connector. A series network of inductors is connecting the diode to the voltmeter.

Having assessed the set-up, we connect the BAV21 diode, with decoupling inductors mounted nearby (Figure 9), resulting in Figure 10c. The harmonics yield a THD of only 12%, but the -18 VDC offset reveals the non-linear behaviour of the component.

Table I
REFLECTED HARMONICS BY THE 78L05 VOLTAGE REGULATOR.

Harmonic	$P_{inc} = +15\,dBm@2\,MHz$	$P_{inc} = +15\,dBm@30\,MHz$
$0f$ (DC)	$+\ 3\,mVDC$	$+\ 4\,mVDC$
$1f$ (fund.)	$-34\,dBm$	$-38\,dBm$
$2f$	$-55\,dBm$	$-62\,dBm$
$3f$	$-60\,dBm$	below noisefloor
$4f$	$-64\,dBm$	below noisefloor
THD	11%	6%

B. Voltage Regulator

Let us now use the measurement set-up for the original problem: testing the linearity hypothesis. We choose the widely-used 78L05 voltage regulator in SO-8 package as case study. This voltage regulator typically delivers $5 \pm 0.01\,VDC$, when fed with 8 VDC [9].

We use a generic DPI PCB (like schematically depicted in Figure 2) to mount the 78L05 IC. The 8 VDC power supply is connected to the regulator input via the on-board inductors. The RF disturbance signal is injected into the regulator input via the external directional bridge ($-16\,dB$ nominal coupling) and an on-board coupling capacitance. We connect a DC-coupled voltmeter to monitor the output voltage of the regulator. As soon as the output voltage surpasses $\pm 0.1\,VDC$ from the undisturbed value, we consider the regulator to fail. We monitor the DC voltage of the input with a 1:10 voltage probe at the IC input pin.

At 2 MHz and 30 MHz, $+15\,dBm$ suffices to let the regulator fail. For these two cases, we enumerated the harmonic amplitudes in Table I and calculated the THD. The DC voltage is the regulator input voltage measured using the probe, with respect to the undisturbed voltage.

The low THD values seem to indicate that the regulator input still behaves linearly at the disturbance power necessary to cause failure. As we have seen with the diode, THD values may be misleading, but in this case the reflected DC voltage at the IC pins is very small as well. We conclude therefore that for this 78L05 regulator, with a $\pm 0.1\,VDC$ failure criterion, the linearity hypothesis is valid.

One critical note: the reflected harmonics are loaded with $50\,\Omega$ through a coupling capacitor, while the reflected DC is loaded by the low-resistance power supply, PCB traces and inductors. If, for example, the bond wire inside the IC has a high resistance with respect to the impedance of power supply, traces and inductors, the voltage at the bondpad will have a higher reflected DC component.

VII. SIMULATION POSSIBILITIES

Apart from testing the linearity hypothesis, measured X-parameters (with or without phase information) can directly be used in simulation.

Recall our goal of simulation: to predict functional product failure under interference. According to Loeckx, all analogue IC susceptibility issues are fundamentally rooted in (1) propagation, (2) rectification or (3) (de)modulation of interference signals [10]. These effects also appear at PCB and higher system levels, and we have not yet been able to falsify Loeckx's hypothesis at any analogue system level. Being able to simulate all these effects enables simulating the ensemble of ICs that constitutes a PCB, maybe also mixed-signal systems (i.e. with analogue and digital subsystems).

Measured S-parameters allow simulating the propagation of interference through ICs and over PCB traces. However, as rectification and modulation are non-linear phenomena, S-parameters cannot capture these behaviours. Measured X-parameters simultaneously describe all three effects.

To illustrate the potential of simulation, let us look at the susceptibility of a fictive automotive subsystem. This subsystem consists of a 78L05 voltage regulator, a microcontroller and a LIN-bus driver. RF interference incident on the 12 VDC rail partly propagates through a 78L05 voltage regulator to the 5 VDC rail, thereafter incident on a microcontroller. Furthermore, the 78L05 upconverts the interference to $2f$, which propagates over the 12 VDC rail to the LIN-driver, which happens to be very sensitive to that harmonic. X-parameters of the 78L05 together with ICIM-CI models of the microcontroller and the LIN-driver would allow to predict this susceptibility in harmonic balance simulation. The ICIM-CI models do not use the relative phase of disturbing signals, because they were obtained with one CW disturbance at a time. If this system model would adequately predict failure, the relative phase of the generated harmonics is not necessary, so $|X|$-parameters suffice.

As another example, the frequency dependent rectification of a diode can be easily measured using the set-up of Figure 8. The result for a 1N4148 is plotted in Figure 11. This measured description can directly be used to simulate the effect of the diode's rectification at PCB level.

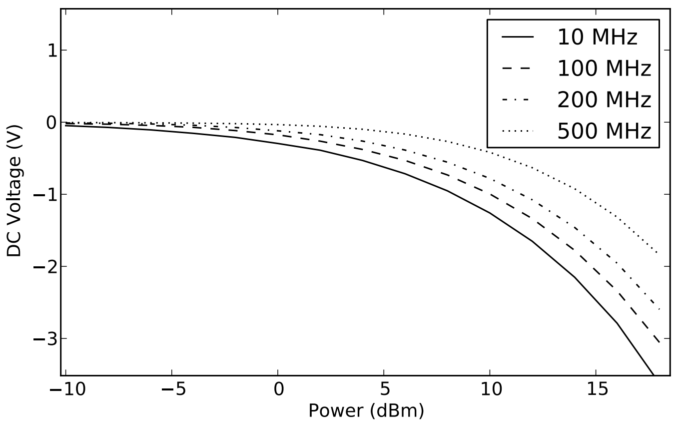

Figure 11. B_{10}/DC voltage across a 1N4148 diode under RF incident power.

VIII. CONCLUSIONS AND RECOMMENDATIONS

Using common EMC laboratory equipment, we were able to test the linearity hypothesis for an LM7805 voltage regulator, in the same measurement set-up as used for DPI testing by replacing the wattmeter for reflected power with a spectrum analyser. Just looking at the THD of the reflected spectrum may be misleading, so the DC component also needs to be taken into account while determining linearity.

The reflected spectrum and the DC component can be expressed as the $|X_{11}|$-parameter of the port under test. This could be a way to model the non-linear behaviour of electrical ports, as some parts of the current IEC 62433-4 proposal suggest a non-linear PDN.

Finally, we outlined how measured X-parameters could enable immunity simulation of systems that consist of multiple ICs.

As for the linearity hypothesis, a robust function of the reflected spectrum needs to be found, as the THD does not provide a clear distinction between linear and non-linear. Susceptibility cases where non-linearities can and cannot be neglected should be studied. In these cases, the reflected spectra should be compared to find a generic criterion.

As for the use of X-parameters, practical multi-IC susceptibility cases should be studied. Firstly, to know if the example given in section VII is very hypothetical or that it occurs in practice. Secondly, if multi-IC interactions occur in practice, to know if measured X-parameters yield reliable immunity predictions and whether phase information is necessary. Finally, to know if the extra measurement and modeling effort is generally worth the improved immunity prediction. In other words, to determine the return-on-modeling-effort (ROME) of simulations using X-parameters.

Acknowledgements

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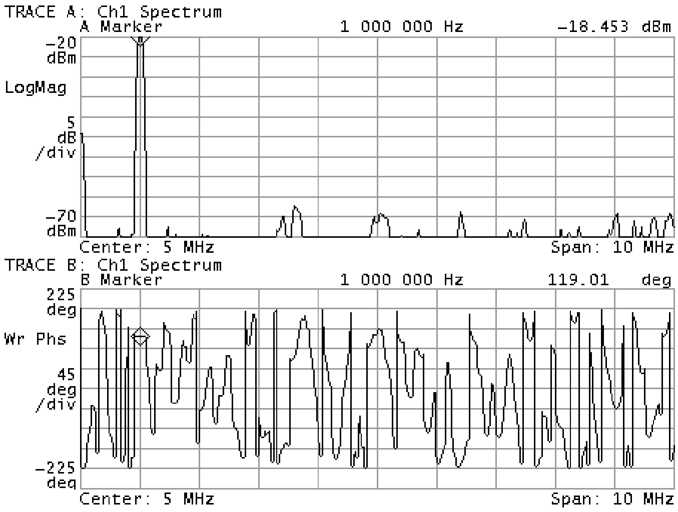
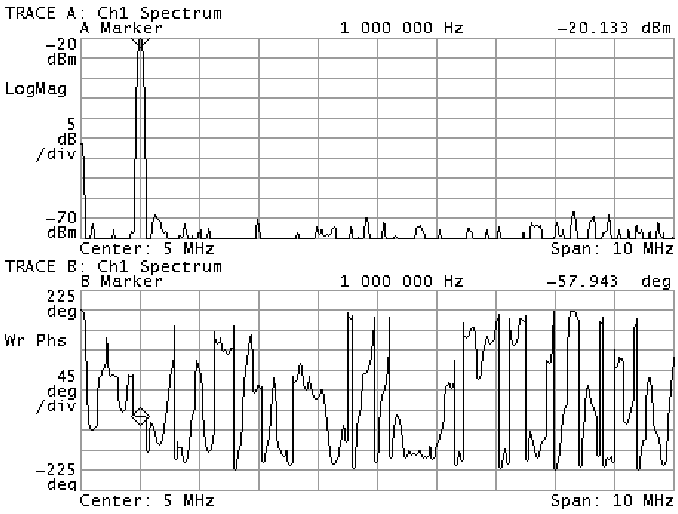
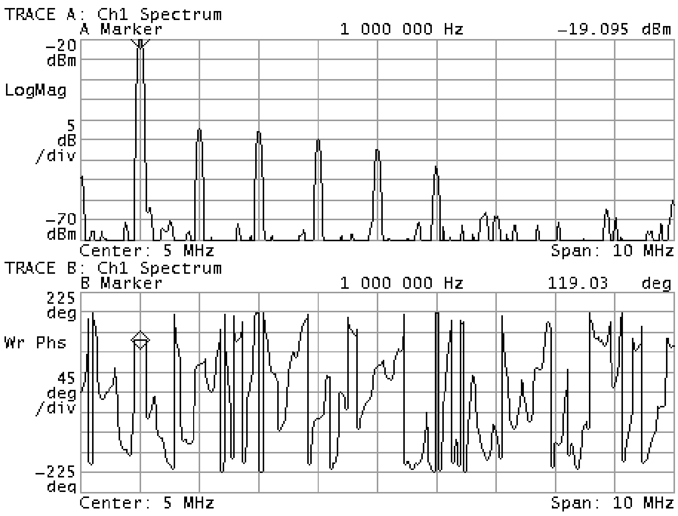


(a) DUT 1; an open circuit. $B_{11} = -18\,\text{dBm}$ $\varphi = +119°$



(b) DUT 2; a short circuit. $B_{11} = -20\,\text{dBm}$ $\varphi = -58°$



(c) DUT 3; the BAV21 diode. $|B_{11}| = -19\,\text{dBm}$, $|B_{12}| = -44\,\text{dBm}$, $|B_{13}| = -45\,\text{dBm}$, which is about 12% THD. $B_{10} = -17.8\,\text{VDC}$

Figure 10. Reflected spectra, as measured with the HP 89441A VSA through the built-in directional coupler of the Prâna amplifier ($-49\,\text{dB}$ nominal reverse coupling).