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A Method for Charactering EMI Coupling Paths and Source Properties in Complex Systems

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***Abstract*— A method for charactering EMI coupling paths in complex systems is presented. While it is relatively easy to determine the EMI antenna structures or the sources of EMI, it is often quite difficult to identify, and even more difficult to quantify coupling paths. This paper introduces a measurement-based method to quantify EMI coupling paths, and the source strengths can be indirectly determined by applying linear system theory. Circuit design guidelines, e.g., the permissible even-mode current in a differential signal, can be derived with the knowledge of the coupling path and EMI limits. Moreover, the EMI can be better predicted with the knowledge of the coupling paths, and the EMI source properties simulated with IBIS or SPICE circuit level simulations.**

Keywords-EMI; coupling path; common-mode current; even-mode current; transfer impedance

I. INTRODUCTION

EMI problems can be broken into sources, coupling paths and antennas. Typical antenna structures are cables driven against enclosures, different parts of an enclosure driven against each other, unshielded boards on which common-mode currents flow, and patch antenna like structures on boards. Some fundamental EMI source mechanisms leading to common-mode radiation from printed circuit boards with attached cables were reported in [1], [2]. Near-field scanning, current clamp measurements, or measurements of the voltage between metallic parts often reveal the antenna structures for most cases [3], [4]. Carefully analyzing the far field signals for their sideband patterns, phase noise, repetition rates and other modulation signatures, are essential to identify the EMI sources. These signatures can be correlated to the information from the circuit diagrams, near-field measurements, and voltage probing measurements on boards. By means of matching the "signature" of each individual problematic signal with the clock speed, switching edges, the wire pattern, the chip layout etc., EMI sources can be identified. Fast switching signals from switched power supplies, the clock signal, data signals and SSN currents are typical EMI sources.

In most cases the EMI sources are relatively easy to identify by a "signature" matching method. However, the coupling paths are very difficult to identify, and even more difficult to quantify. In this paper, a method for characterizing coupling paths, and estimating EMI source strength is shown.

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Three types of problems can be solved using this method in particular,

- The magnitude of a current that drives the EMI antenna can be determined indirectly. This is convenient as it is often not possible to measure the current directly. The example used here is the even-mode current on a pair of differential signals.
- Different coupling paths can be quantified for the purpose of determining the dominant coupling path.
- Design criteria can be derived from the knowledge of the coupling path, so SPICE based circuit designs can have a goal function. In the example selected here, it is the maximum even-mode current on a pair of differential signals.

This paper explains the principles, shows an example and discusses the limits of the method.

II. METHODOLOGY

The proposed method for characterizing EMI coupling paths and source properties is measurement-based. The coupling paths are characterized with *S*-parameter measurements and two-port network theory. The noise source properties are determined with a transfer impedance and the measured "port voltage", which is explained in the following sections.

A. EMI coupling path characterization

Two-terminal "Ports" need to be defined in order to analyze coupling paths. A microwave two-terminal port should have a well-defined voltage and current, so the selected two-terminal port needs to be either electrical small in dimension as compared to the wavelength of the highest frequency of interest, or supports a TEM wave. Useful information can be extracted from pre-defined "ports", which could be the voltage at the input terminal of a log-periodic antenna, a voltage between metal parts on the EUT, or even a near-field value measured with a small loop probe. However, for practical applications one needs to consider reproducibility and convenience in the selection of ports, and the selected port needs to be meaningful with respect to the EMI caused by the EUT. Field measurements suffer from reproducibility, and it needs to be considered whether the signal is maximized by the turntable and the polarization. For near field data, there is often no good correlation between strong near fields and strong far

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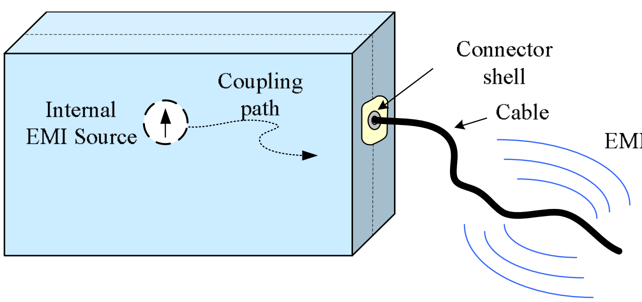
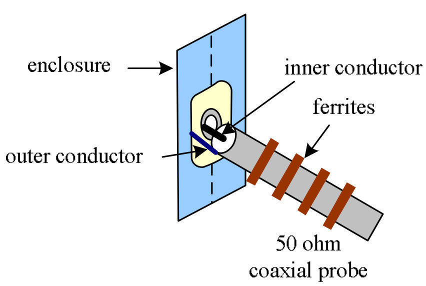
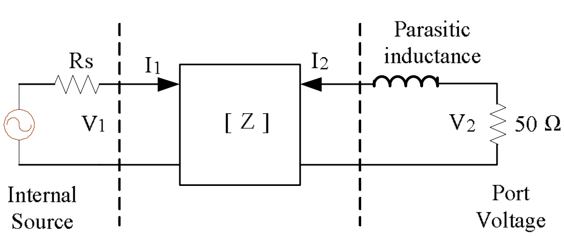
fields. For the EUT selected here, the emissions are mainly cable driven, because the EUT is electrical small even up to several hundreds of MHz. Since the highest frequency of interest is 1 GHz for this case, and the distance between the ground shell of the connector and the nearby enclosure of the product is approximately 1 cm, which is electrical small, it is suitable to define the outer ground shell of the connector and the nearest enclosure point as a two-terminal port. Note that the dimension of the connector is approximately 4 mm in diameter. The voltage between the connector shell and the enclosure, which drives the attached cable against the enclosure as an antenna, is measured as the port voltage. As shown in Figure 1 (a), the port voltage here is caused by noise sources inside the product. Figure 1 (b) illustrates the measurement setup of the port voltage with a 50 Ω coaxial cable probe. The inner conductor of the coaxial cable was connected to the ground shell of the connector, and the outer shield of the cable was soldered to the enclosure. The loop area of the cable probe tips was controlled to be as small as possible to minimize the resultant parasitic inductance. Herein, the dimension of the cable probe tips and the defined port is on the order of millimeters, so the parasitic inductance effect is negligible in the measurements. Note that the original cable was removed when measuring the port voltage.



(a) Schematic illustration



(b) Port voltage measurement setup



(c) Equivalent circuit diagram of the measurement setup

Figure 1. The measurement of the port voltage caused by an internal EMI source, (a) schematic illustration, (b) port voltage measurement setup, and (c) the equivalent circuit diagram of the measuremetn setup.

In most cases, the coupling, typically parasitic, between the noise sources and port voltages is linear, because the coupling path is of capacitive, inductive or galvanic in nature. The EUT can be treated as a linear system when analyzing the coupling path. However, if the coupling path is via an active device which modulates another signal, the proposed method may need to be modified.

For a given complex system, once the EMI sources and antenna structures are determined, the remaining work is to identify the coupling paths. The identification of the EMI coupling paths in a complex system is usually difficult because of structural complexities. Instead of directly identifying the coupling paths, the system can be modeled as a two-port network, as shown in Figure 1 (c). One port is associated with the EMI source and the other port is associated with a resultant voltage between the connector shell and the enclosure for the case herein. Then, the system can be characterized with *S*-parameter measurements and two-port theory. Two assumptions of this method are, first, the coupling paths are linear; and second, the system is excited in a way that only a single coupling path is dominant.

S-parameter measurements are employed to quantify this two-port network. A test current or voltage is injected to the trace, which is suspected to be driven by the EMI source. In this regard, the proposed method falls into the class of substitution methods [5]. The S-parameters between this artificial noise source and the cable connector are measured using a vector network analyzer (VNA). The injected test signal is generated by one port of the VNA. It needs to be ensured that the injection of the external signal does not alter the EUT. This could be the case if a common-mode current also flows on the outside of the coaxial cable used to inject the signal. To avoid this, many ferrite sleeves are often added to the coaxial cable, and/or the outer shield of the cable is brought along, and soldered to a reference conductor or plane [6]. The two port *S*-parameters measured with the VNA can be transformed into a Z-matrix as [7]

$$[Z] = 50 \cdot ([I]-[S])^{-1} \cdot ([I]+[S]), \tag{1}$$

where $[I]$ is the identity matrix.

B. *EMI source characterization*

A transfer function concept can be employed to facilitate the calculation of the noise current strength. The transfer function is defined as the ratio of the measured port voltage with the cable un-attached, to the noise source current inside the device, which can be written as

$$F_{trans} = \frac{V_2}{I_1}. \tag{2}$$

The derivation of the transfer function with the *Z*-matrix of the network is shown below. Note that the parasitic inductance caused by the coaxial probe is negligible, so it can be omitted here. Based on the definition of the *Z*-matrix,

$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix} \cdot \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}, \tag{3}$$

V_2 is

$$V_2 = Z_{21} I_1 + Z_{22} I_2. \tag{4}$$

Since a 50 Ω coaxial probe is used to measure the port voltage V_2, and the parasitic probe inductance is assumed small relative to 50 Ω, then

$$I_2 = -V_2 / 50. \tag{5}$$

Substituting (5) into (4),

$$V_2 = \frac{50 \cdot Z_{21}}{50 + Z_{22}} I_1. \tag{6}$$

The port voltage transfer function is then evaluated as

$$F_{trans} = \frac{V_2}{I_1} = \frac{50 \cdot Z_{21}}{50 + Z_{22}}. \tag{7}$$

Since the unit of F_{trans} is Ω, it is denoted as a transfer impedance. A large transfer impedance indicates strong coupling, which means a small current can cause a large port voltage. Once the transfer impedance is known, the EUT can be turned on, and the port voltage caused by the real EMI sources can be measured. The noise source current, which is very difficult to be directly measured when the product is in operation, can be estimated as

$$I_{source} = \frac{V_p}{F_{trans}}. \tag{8}$$

As shown in Figure 1 (b), the port voltage (EUT operating) should be measured with a 50 Ω probe, such that the load condition is the same as that of the S-parameters measurements with a VNA.

The transfer function is a function of frequency. It allows characterizing a coupling path in a broad frequency range, which makes it possible to check if it is of resonant or broadband behavior, and if it can be modeled with a simple inductive or capacitive equivalence. This is a significant advantage relative to measuring the signals only at the frequencies excited by the EUT. Note that the EUT is usually powered off while performing the measurements. Since the coupling path in this case only consists of passive components, i.e., PCB and enclosures, powering off the EUT does not influence the measurements. However, if the coupling paths are not passive, powering off the device may influence the transfer function measurements.

III. Experimental Results

The method detailed in Section II was applied to quantify the coupling paths in a commercial product. Figure 2 shows the block diagram with emphasis on the all intentional and unintentional current paths and the associated components. The EMI source was identified as the transmission line even-mode current I_2 flowing from the driver IC to a capacitive load via a flexible flat cable. The trace A and trace B form a differential clock signal pair. Because of the circuit asymmetry, differential-to-CM conversion, etc., the even-mode current I_2 is not zero. Most of the current I_2 returns via the GND trace of the flexible flat cable, which is denoted as I_3. A small portion of I_2 may flow back through a different path formed by screws and metal parts that connect the capacitive load to the PCB, denoted as I_4. If $I_2 = I_3$, then $I_4 = 0$. In this ideal case, the EM energy is concentrated in a small region around the flexible flat cable, which would not cause EMI problems. However, this may not be always the case. In a practical circuit, it was found that $I_2 \neq I_3$ and $I_4 \neq 0$. The EMI CM current I_4, or so-called antenna current, couples to the front of the enclosure, across the cable connector and the attached cable to the rear of the enclosure, then returns back to the main PCB ground via grounding/mechanical screws. A voltage drop between the connector ground and the enclosure is created by the unintentional current I_4. This voltage drop drives the antenna in the form of the enclosure against the attached cable, which leads to an EMI problem. The root cause of the EMI for this case is the antenna current I_4. The characterization of the antenna current I_4 with a current clamp measurement on the flexible cable could not be done due to the extremely dense packaging of the product. The proposed method was used herein to quantify the currents I_2 and I_4.

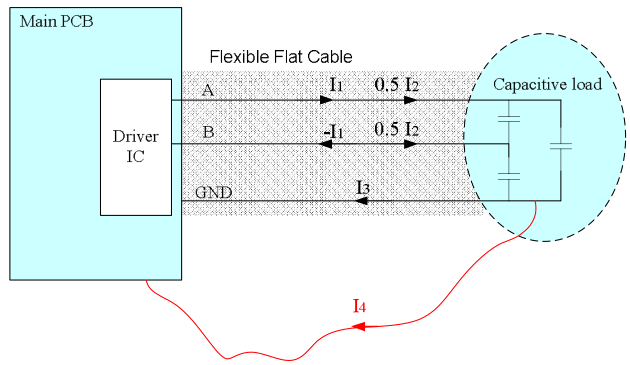


Figure 2. The block diagram of the commercial product.

A. *Experimental configurations*

Figure 3 shows experimental configurations of the transfer function measurements. Configuration (a) was used to determine the transfer function between the port voltage and the even-mode current I_2. Configuration (b) was used to determine the transfer function between the port voltage and the antenna current I_4. For the Configuration (a), all clock traces were disconnected from the original circuit and shorted together. They were excited with Port 1 of a VNA. The center conductor of the coax was connected to the shorted clock traces, and out coax shield was connected to the PCB ground. The ground of the flexible cable was still connected to the PCB ground. Port 2 of the VNA was connected between the connector ground and the front of the enclosure. To simplify the test setup, the rear of the enclosure was removed. And four copper tape strips were used to connect the front of the enclosure to the PCB ground to emulate the shielding effect of the rear enclosure. On one hand, it simplifies the connection and routing of the coaxial cable; on the other hand, it also introduces additional measurement errors and uncertainties. In order to reduce the common-mode current on the feed cable, approximately 20 ferrite sleeves were placed on the feeding cable. For this setup, the original source (even-mode current I_2 from the driver IC) was replaced with the Port1 of the VNA, but the internal coupling path between the original source and the port voltage remains the same. In this fashion, the coupling path between the driver IC and the port can be characterized with the transfer function by dividing the port voltage with the current injected with the VNA. For Configuration (b), the flexible cable was disconnected from the PCB board, and a test current was injected to the ground of the flexible cable by port 1 of the VNA, as shown in the enlarged plot. Because the entire flexible cable is detached from the main PCB board, the injected current could not return on the flexible cable, and it was forced to flow via the coupling path where the antenna current I_4 flows. The transfer function derived with this measurement characterizes the coupling path between the antenna current I_4 and the port voltage. Since the antenna current I_4 is the root cause of the port voltage, this measurement setup should lead to a larger transfer impedance as compared to the previous one.

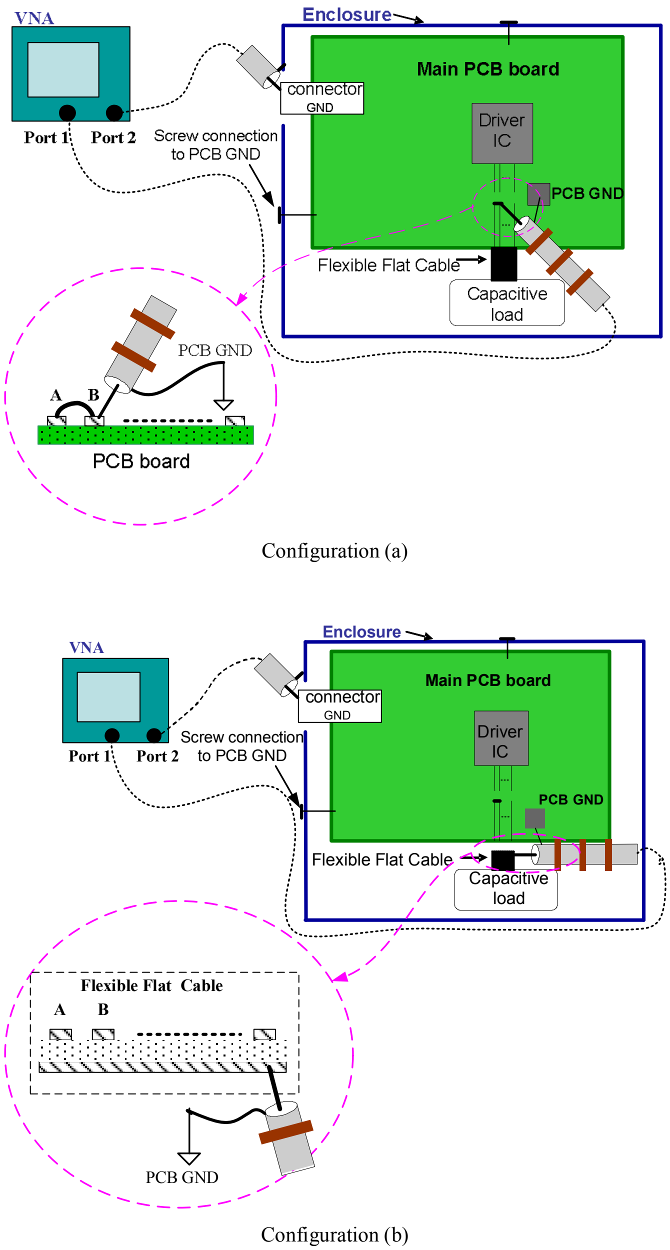


Figure 3. Experimental configuration of the CM current measurement (a) by injecting the current into the differential clock traces with the ground trace of the flexible cable connected to the PCB ground, and (b) by injecting the current into the ground of the detached flexible cable.

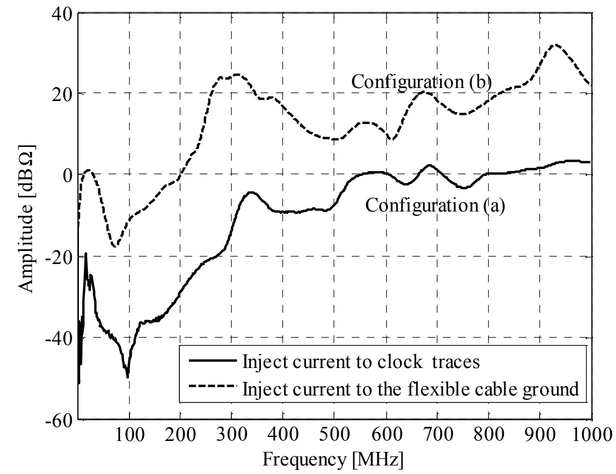


Figure 4. Port voltage transfer function for Configuration (a) and (b).

B. Results derivation and analysis

The measured transfer functions of the two experiments are shown in Figure 4. The increasing transfer impedance with respect to the frequency indicates an increasing coupling between the noise source and the port voltage. The transfer impedance with Configuration (b) is approximately 20 dB larger than that of Configuration (a). This indicates that 90% of the current is returning on the flexible cable.

With the knowledge of the transfer impedance and the port voltage of the operating system, the EMI source current can be easily calculated. Figure 5 shows the port voltage of one operating product measured with a spectrum analyzer, and the even-mode current I_2 and antenna current I_4 estimated using (8). For this case, only the voltage peaks at 27MHz harmonics are of interest. The significant difference between I_2 and I_4 indicates that although the even-mode current I_2 of the differential clock is the source of the EMI, only a very small portion of I_2, i.e., I_4 is the root cause of the EMI problems.

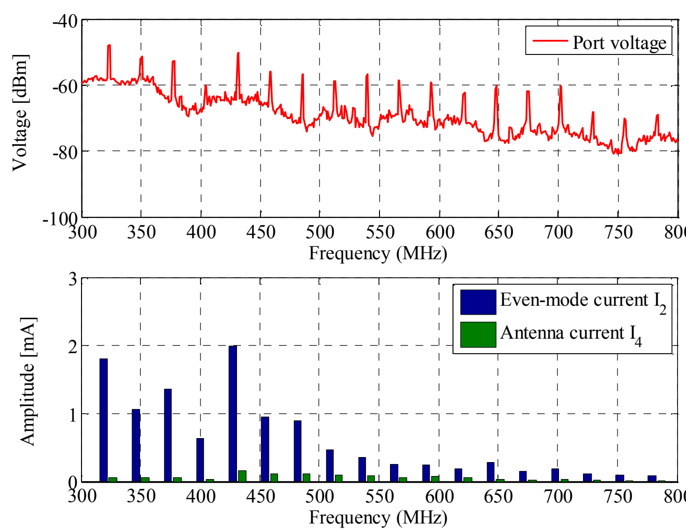


Figure 5. The measured port voltage and the corresponding calculated source current with (8) for Configuration (a) and Configuration (b), which leads to the even-mode current I_2 and the antenna current I_4 , respectively.

The fact that the antenna current I_4 is only a small portion of the EMI source current I_2 could be very useful for EMC engineers. Since EMC engineers often use near-field scanning systems for solving EMC problems. The plots from the scanning can show areas and traces of large currents. Often done, but misleading is to conclude that these strong currents cause the EMI. The currents in the coupling paths are often much weaker and very difficult to identify in the near field data. In the analysis of near field data, EMC engineers should not only look where the large currents are, but also give attention to identifying locations of the weak currents that should not be there.

Besides the estimation of the EMI source current, this method could also be used to estimate the maximum even-mode current allowed in the circuit design. For the fulfillment of FCC class B emissions, often a current limit for cables that act as antennas is given. Typically, values of more than 5-10 uA may lead to failure of satisfying FCC class B. These values are derived from antenna theory, assuming that the cable acts like a low gain antenna. The actual radiating structures formed with cables could be roughly quantified with an antenna impedance. Since the resonance frequencies are unknown for the antenna (assuming the cable length and "termination" are unknown), it is reasonable to assume that the antenna is in resonance at the frequency of interest. This leads to a real antenna impedance in the range of $R_{antenna} = 50\text{-}100\ \Omega$. With known antenna impedance and antenna current limit, the maximum allowed even-mode current can be estimated as

$$\left|I_{even\_\max}\right| = \left|\frac{I_{antenna}(R_{antenna} + Z_{22})}{Z_{21}}\right|, \qquad (9)$$

where Z_{21} and Z_{22} are derived from the measured S-parameters as in (1).

IV. Limits of the Method

The proposed method is based on the linear system theory. The underlying assumption is that the coupling path is linear. If it is not linear, the method will fail. Further, the internal noise source should be able to be substituted with an external signal. When performing the substitution, the resulting measurement artifacts need to be able to be minimized. For example, this requirement will exclude cases in which widely distributed currents are driving the EMI antennas, but currents on traces and on physically narrow ground returns can be substituted well.

The accuracy depends on uncertainties of the transfer function measurement, and the accuracy of the estimation of the antenna current and impedance. The transfer impedance can be measured well, but its usefulness depends on how well the coupling path reflects the dominant coupling path on the EUT, and how well the current injection can be done without disturbing the structure, enclosure and current paths of the EUT. In conducting these experiments, great care needs to be directed towards minimizing these effects. It is promising that the proposed method could be used to establish design guidelines for the estimation of the maximum noise source currents and voltages.

V. Conclusions

Based on two-port linear network theory, a method for characterizing EMI coupling paths through VNA measurements is presented in this paper. As long as the system can be treated as a linear system, and the dominant EMI source has been already identified, the proposed method can be used to quantify dominant coupling paths in a much broader frequency range than the frequency range excited by the EUT, which helps to identify resonant behavior. The proposed method can also be used to indirectly measure the EMI source strength, and determine limits that circuit designers can use.

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