MICROWAVE COUPLING ON PRINTED CIRCUIT BOARDS✝

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ABSTRACT

The bottom line for any susceptibility assessment of a system to an electromagnetic threat is whether or not components are upset or damaged. To predict this, we have to understand the distribution of energy on pc boards, and its interaction with the functioning circuits on the boards. LineCAP is a code that computes this energy distribution and the circuit's response. It includes the effects of trace-to-trace coupling, discontinuities (bends, vias), passive circuit components, and nonlinear devices. The theory, comparisons with measurements, applications, and limitations of this code will be presented.

1. INTRODUCTION

The end result of a susceptibility assessment of a system to electromagnetic pulse (EMP) or high-power microwaves (HPM) is a statement of the probability of damage of critical electronic components. To reach these components, electromagnetic energy must interact with the exterior of the system, penetrate the outer skin through various ports-of-entry, travel through the system along cables or through the waveguide formed by the interior of the system, enter an electronics subsystem, and, finally, propagate along a printed circuit (pc) board to the component. The final segment of this path is the subject of this paper.

Propagation along the pc board trace or traces to a damageable component is complicated by several factors. Since we are not considering direct pick-up by the trace as a coupling mechanism here, the energy has to be launched onto the board through an edge-connector. Since most boards are not designed for the high frequencies present in the incoming signal, the transmission and reflection at this junction will be a very complicated function of frequency. Similarly, the traces themselves are designed to carry low frequencies. Depending upon the circuit board pattern, their impedance can vary drastically along their length. In addition, they will be coupled to one another, forming a complicated, non-uniform, multiconductor transmission line. Discontinuities in this line, such as bends, vias, and junctions, must be modeled at high frequencies by an equivalent circuit. Passive and active components are attached to this line, and due to parasitic effects, the high-frequency model for such components is also a complicated equivalent circuit. It is not surprising, then, that the energy that finally appears at the terminals of a component can vary greatly with frequency.

Considerable work has been done in the area of multiconductor transmission lines in general [1], on time-domain analysis of multiconductor transmission lines [2], and on equivalent circuits of discontinuities [3-5]. The purpose of this work was to put these pieces together, combining multiconductor transmission line theory with circuit analysis capability. The end result was a computer program that models a printed circuit board at high frequencies as sections of coupled, multiconductor transmission lines connected at their ends by arbitrary networks of circuit elements. The transmission lines represent parallel runs of the traces on the pc board. The circuits model either discontinuities (for example, the location on the board where a trace turns and no longer runs parallel to its neighbors) or the actual physical components connected to the traces at a given location.

The next section describes the approach taken in this program, called LineCAP (Line/Circuit Analysis Program). Code predictions on some simple geometries are compared with measurements in the section following.

2. APPROACH

The approach used here was modal analysis in the time domain of the lines, combined with time-domain circuit analysis of the networks connected by the lines.

2.1 MODAL ANALYSIS IN THE TIME DOMAIN

Although a lossless line that includes an inhomogeneous dielectric is dispersive, there exist sets of voltages and associated currents on the conductors of the line that will propagate without dispersion. These sets of voltages and currents

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line that will propagate without dispersion. These sets of voltages and currents are called modes. The voltages and currents are not independent, of course, since they are related by the impedance matrix of the line. The amplitudes of the voltages and currents of a mode do not change as the mode travels down the line. In effect, the various conductors that make up the line are uncoupled from each other as far as the mode is concerned. These modes do not interact with each other except possibly at the terminal networks; they simply superpose at each point along the line. Each mode, in general, travels with its own characteristic velocity.

The intensity of a mode can vary as a function of time. All the conductor voltages and currents for the mode would vary with time in the same way. If two time-varying modes were traveling down a line with different velocities, the unchanging waveshapes of the modes would slip in relation to each other, the slower mode falling behind the faster. The total waveshape (the superposition of the two modes) changes as it propagates down the line because of this slippage. Modal analysis is the process of decomposing the total voltage on the line, which disperses as it propagates, into modes traveling with different velocities, which do not disperse. These modes are then tracked as they travel up and down the lines [6-7]. With a knowledge of the modal intensities, the voltages and currents incident on the terminals of the line can be found at any time.

2.2 TIME-DOMAIN CIRCUIT ANALYSIS

The voltages and currents at the terminals of the line drive the terminal networks, changing voltages internal to the network, charging capacitors, and changing the current through inductors. Circuit analysis of the terminal networks, given the terminal voltages and currents, determines how the lines have affected the circuits. This change, in turn affects the lines through reflected modes launched back up them.

Since the circuits are solved by a nodal analysis, it was most convenient to represent the lines by their Norton equivalents. The equivalent conductance matrix for a multiconductor line is the inverse of its impedance matrix. The short circuit currents, from basic transmission line theory, are twice the current wave incident on the circuit from the line. Given the loading effect of the lines and their equivalent external current sources, standard iterative techniques can be used to solve the nonlinear circuit for the terminal voltages [8].

3. COMPARISON WITH MEASUREMENTS

To build confidence that LineCAP was working properly, a series of measurements on simple geometries was made. These geometries are able to be both measured and modeled accurately.

3.1 MEASUREMENT SET-UP AND PROCEDURE

LineCAP is a time-domain code; its output is a time-domain voltage waveform. Unfortunately, highly accurate, wide-bandwidth (fast-risetime) time-domain measurements are difficult to make. The problem is mainly the sensitivity and the bandwidth of the recording device that senses the response. Highly accurate, sensitive, and wideband frequency-domain measurements, however, are relatively easy. If the magnitude and phase of the response is measured over a broad bandwidth, the time-domain response to a given input can be computed by an inverse Fourier transform. Coherent detection allows extremely sensitive measurements to be made, and the calibration/error correction capabilities of modern vector network analyzers can be applied to enhance the accuracy of the measurements. The drawback to this approach is that it is applicable only to linear networks. Nonlinear networks must be measured in the time domain.

The measurements presented here were made with a Hewlett-Packard 8510B Microwave Vector Network Analyzer. Data were taken in the frequency domain, from 45 MHz to 18 GHz, using 22.5 MHz steps. The result was the magnitude and phase of the response normalized to a 1-V input stepped over this range.

The time domain responses were computed by inverse Fourier transform. Since the transform requires information at zero frequency (dc) and uniform frequency steps, the magnitude and phase at 0 and 22.5 MHz had to be extrapolated from the measured data. This inverse transform is the response of the network to a unit impulse. Integration of the waveform gives the unit step response, which was the quantity that was computed. The risetime of the step was limited by the highest frequency measured -- ie., the time resolution is limited to 1/18 GHz, or approximately 0.06 ns.

3.2 RESULTS FOR SIMPLE GEOMETRIES

3.2.1 Coupled, Parallel, 50-Ohm Lines

The geometry in Fig. 1 allowed us to evaluate the coupling of parallel, 50-Ohm microstrip lines. We used stock fiberglass pc board material. The material was double-sided, with 2 oz. copper cladding, ie., the thickness of the copper was 2.5 mils. The dielectric constant of the substrate was approximately 4.0. With these characteristics, the lines had a width of 1/8 inch. The lines were spaced 1/4 inch apart and centered on pieces of pc board 2 inches wide and 6 inches long.

Figure 2 shows the voltage at the input end of the parasitic line for a 1-V step incident on the driven line. Both lines

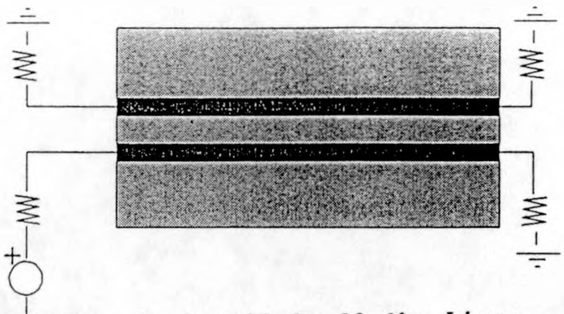

Fig. 1. Parallel, 50 Ohm Lines

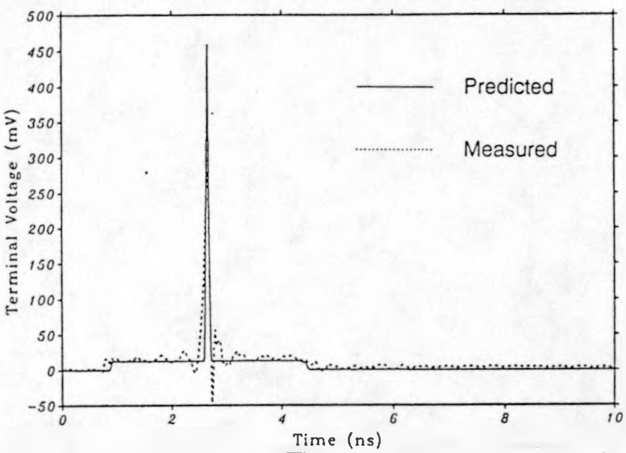

Fig. 2. Voltage at Input of Parasitic Line

were shorted to ground at the load ends and connected to 50-Ohm impedances at the generator ends. The solid curve is the calculation and the dotted curve is the measurement. Short sections of coax line connect the traces to the network analyzer, resulting in a four-hundred picosecond delay before the unit step reaches the board. The very fast spike is overestimated in the calculation because the increasing attenuation at high frequencies is not accounted for.

3.2.2 Coupled 50-Ohm Lines with Bends

Straight, parallel traces on pc boards are fairly common, especially for data and address lines in computer circuits. But the lines eventually bend and diverge. This was the geometry we considered next.

We used 4" by 6" sheets of the same pc board as for the straight, coupled lines. The lines were 50-Ohm (1/8" width), and were spaced 1/4" apart. In the simplest case, shown in Fig. 3, the lines ran parallel for three inches. The parasitic line then turned 90 degrees and ran to the top edge of the board, while the driven line continued straight to the edge opposite the generator. The lines were placed on the board so that their lengths were equal.

The signal induced in the parasitic line as observed at the generator end is similar to that for the straight lines, except the spike is surrounded in time by regions of no cross-coupling. This dead

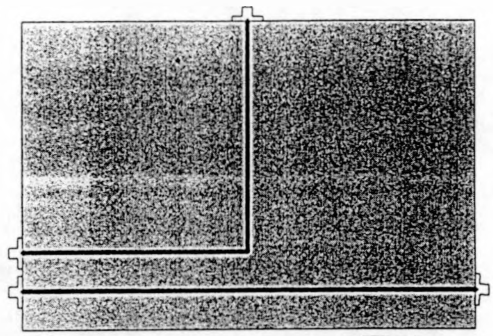

Fig. 3. Parallel, 50 Ohm Lines with Bend

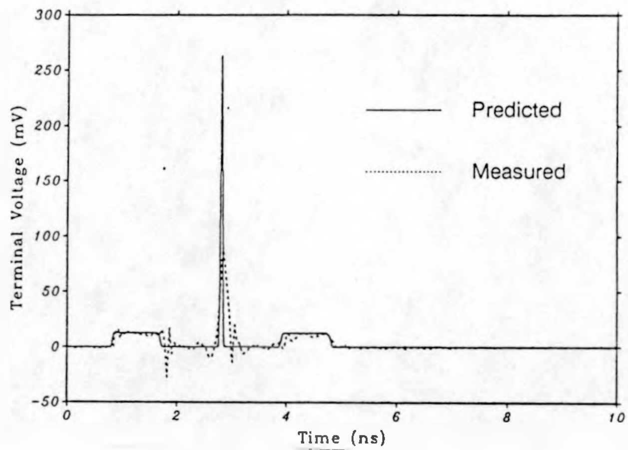

Fig. 4. Voltage at Input of Parasitic Line, with Driven Line Bent

region occurs while the rising and falling edges are traveling along the parts of the line that diverge and are, thus, uncoupled. Figure 4 gives the calculated and measured waveforms.

3.2.3 Results for Nonlinear Loads

Once nonlinear loads were included on the boards, frequency-domain measurements could no longer be used to get time-domain responses. A time-domain measurement system was put together using a Nanofast 568-502 pulser (rise time < 1 ns) and a Tektronix 7S14 sampling head driving a 7103 oscilloscope. A digitizing camera was used to capture the measured voltage traces for analysis.

Figure 5 shows the geometry used. The board was the same one used for Fig. 1. The diode was a 1N34A. The diode was modeled by a perfect exponential diode in parallel with a 0.35-pF capacitor and 1-MOhm resistor, all in series with a 100-Ohm resistor. The capacitor represented junction capacitance, and the series resistor represented the bulk resistance of the diode. The lower line was driven at the input end, and the voltage was observed at the input end of the upper line. The input was a 1-V step with a 1-ns risetime.

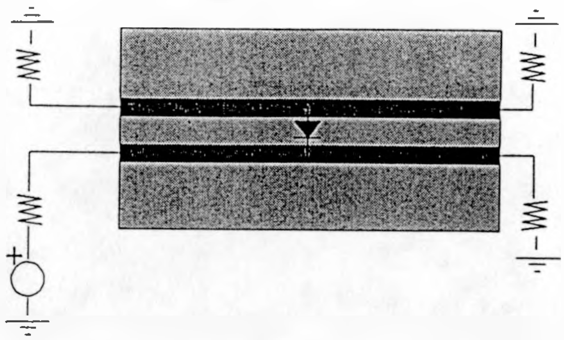

Fig. 5. Geometry with Nonlinear Load

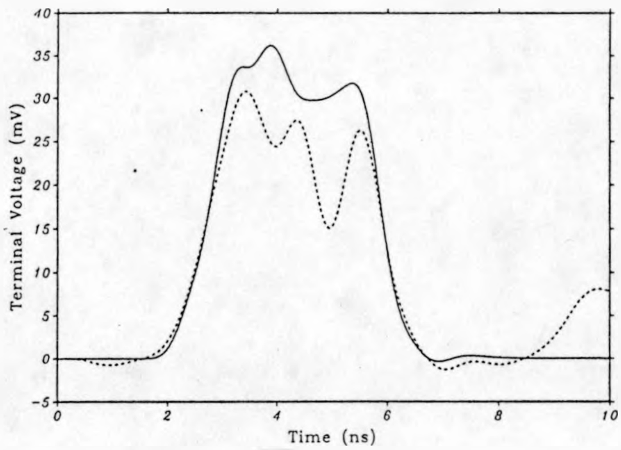

Fig. 6. Voltage at Input of Parasitic Line, Reverse Bias

The configuration shown has the anode connected to the parasitic line (reverse bias). In Fig. 6, the solid line is the calculation and the dotted line the measurement. Reasonable, although not exact, agreement is obtained. (The bump in the measured curve near nine nanoseconds is an artifact of a re-reflected pulse from the generator.) The structure in the response was strongly dependent upon the parameters of the diode model. The amplitudes and general pulse shapes were not. The calculations were digitally filtered with a low-pass, fifth-order Butterworth filter with a corner frequency of 1 GHz to account for the finite bandwidth of the time-domain measurement system.

4. SUMMARY

A program has been written that allows us to predict the trace-to-trace cross-coupling on printed circuit boards for many geometries of practical interest. It combines multiconductor transmission line analysis with circuit analysis in the time domain. Multiple sections of uniform lines, of varying number and characteristics, can be interconnected by passive or nonlinear circuit elements.

The predictions were compared with measurements for simple geometries, with good agreement in both waveshape and amplitude. The major discrepancies were in the amplitudes of fast-rising spikes and are probably due to the fact that the predictions did not account for the increasing attenuation at high frequencies.

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