Discrimination of Passive Intermodulation Sources on Microstrip Lines

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INTRODUCTION

Passive intermodulation (PIM) manifests itself as a nonlinear mixing taking place in passive devices. Deleterious effect of passive intermodulation on signal integrity is of serious concern for a wide range of telecommunication applications [1]-[3]. An increasing use of printed circuit boards in antennas and packaged systems has recently attracted a particular attention to the sources of PIM generation in planar circuits.

In contrast to the well-known "rusty bolt" phenomenon observed in many other components [1], the nonlinearity in microstrip lines has distributed nature and can be associated with the dielectric substrate or/and printed conductors. Several studies reported that copper grade and finishing have significant effect on the PIM level on microstrip lines [4]-[6]. In the meantime, the effect of the laminate properties on PIM performance of printed lines has also been studied, and possible contribution of the dielectric substrate has been pointed out [7] . The existence of the concurrent mechanisms of PIM generation impedes identification of the actual PIM sources. Therefore the techniques for discrimination of the factors contributing to the PIM generation are crucial for PIM mitigation on microstrip lines.

In the paper an approach to discriminating the effects of the nonlinear conductor or nonlinear dielectric is proposed. First, a theory of the nonlinear transmission line with the current- and voltage-driven nonlinearities defined by the phenomenological nonlinear per-unit-length resistance and capacitance is presented. It is shown that PIM level is proportional to the relevant nonlinear factor. Then, the distinctive features of each nonlinear contributor and their applicability to discrimination of the PIM sources are discussed. Finally, the results of experimental study of different laminates with weakly nonlinear conductor or dielectric are presented, and the discrimination procedure is verified.

PASSIVE INTERMODULATION ON THE LINES WITH NONLINEAR CONDUCTOR

Electromagnetic wave propagation in a transmission line with the distributed weakly nonlinear resistance is governed by the telegrapher's equations for voltage $U(z,t)$ and current $I(z,t)$ [8]

$$\frac{\partial I(z,t)}{\partial z} = -\left(C\frac{\partial U(z,t)}{\partial t} + GU(z,t) \right) \tag{1}$$

$$\frac{\partial U(z,t)}{\partial z} = -\left(L\frac{\partial I(z,t)}{\partial t} + R(I)I(z,t) \right), \tag{2}$$

where L, C, $R(I)$, G are per-unit-length parameters of the transmission line. The linear parts of these parameters are well-known for many transmission lines [8], including the microstrip line shown in Fig. 1.

The weakly nonlinear resistance $R(I)$ can be approximated by a polynomial dependence of current $I(z,t)$. Although the system of equations (1) and (2) can be solved for the polynomial of any order, the analysis is restricted to the 3rd order nonlinearity, and the main features of the 3rd order PIM (PIM3) generation are considered only. In this case, the nonlinear resistance is represented as follows

$$R(I) = R_0 + R_2 I^2, \qquad R_0 >> R_2 I^2, \tag{3}$$

where R_0 is linear resistance and R_2 is the macroscopic nonlinear parameter of conductor.

Substituting (3) into (2) and combining the result with (1), one obtains the nonlinear differential equation with respect to $I(x,t)$:

$$\frac{\partial^2 I(z,t)}{\partial z^2} - CL\frac{\partial^2 I(z,t)}{\partial t^2} - \left(CR_0 + GL\right)\frac{\partial I(z,t)}{\partial t} - GR_0 I(z,t) = R_2 I(z,t)^2 \left(3C\frac{\partial I(z,t)}{\partial t} + GI(z,t) \right) \tag{4}$$

Equation (4) is solved by perturbation method with respect to the small parameter R_2. The solution of the problem was obtained in [8] and the expressions for the power of the reverse P_{rev} (signal measured at the input port) and forward P_{forw}

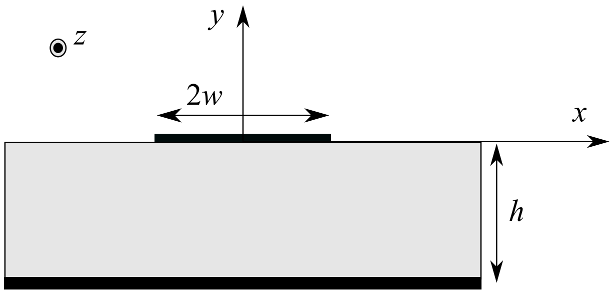

Fig. 1. A microstrip line.

(signal measured at the output port) PIM3 products for the line section of length l were defined at the intermodulation frequency $2f_1$-f_2

$$P_{rev} = R_2^{\,2}\frac{1}{2}Re\left\{\tilde{I}_{2,-1,1}(0)\tilde{U}_{2,-1,1}(0)^*\right\}$$

$$P_{forw} = R_2^{\,2}\frac{1}{2}Re\left\{\tilde{I}_{2,-1,1}(l)\tilde{U}_{2,-1,1}(l)^*\right\}, \tag{5}$$

where * and *Re* stand for complex conjugate and real part of complex quantities, respectively; f_1 and f_2 are frequencies of the fundamental tones (carriers).

As one can easily see, power of the intermodulation harmonics is proportional to R_2^2. The functions $\tilde{I}_{2,-1,1}(z)$ and $\tilde{U}_{2,-1,1}(z)$ in (5) describe the spatial distribution of the current and voltage along the transmission line and are given in [8]. The macroscopic nonlinear parameter R_2 relates the intrinsic microscopic nonlinear resistivity to the geometry of the transmission line cross-section.

The relation between the macroscopic parameter and intrinsic microscopic nonlinear resistivity was obtained by assuming non-local interaction between the electric field and the current on the strip [8]. Here, we adopt alternative approach based upon the local interaction model applied earlier to the study of the nonlinear effects in superconductors [10].

Let us assume the third order nonlinearity of the electric field on the strip. For the sake of simplicity only the longitudinal current is considered, which makes the problem scalar:

$$E = \rho_0 J + \rho_2 J^3, \tag{6}$$

where E is the electric field, J – the surface current on the conductor, ρ_0 - the linear electric resistivity and ρ_2 - the microscopic nonlinear parameter. The latter two parameters are assumed to be constant for a particular conductor.

One can calculate power dissipated in the conductor using both the quasi-static theory and the transmission line approach. Then equating these quantities, we obtain the identity

$$\oint_c E(l)J(l)dl = R(I)I^2$$

or with the aid of (3) and (6)

$$\rho_0\oint_c J^2(l)dl + \rho_2\oint_c J^4(l)dl = R_0 I^2 + R_2 I^4, \tag{7}$$

where integration is taken over the contour of the conductor cross-section, including the ground plane. The macroscopic nonlinear parameter R_2 is found by equating the nonlinear terms in the left and right hand sides of (7)

$$R_2 = \rho_2 \Gamma_c, \tag{8}$$

where

$$\Gamma_c = \frac{\oint_c J^4(l)dl}{I^4} \tag{9}$$

is the geometrical factor, which defines the dependence of the macroscopic nonlinearity on the parameters of the line cross-section. The concept of the geometrical factor enables characterization of different cross-sections including various types of the lines (microstrip line, coplanar waveguide, etc.), thus providing a convenient means for evaluating the effect of nonlinearity.

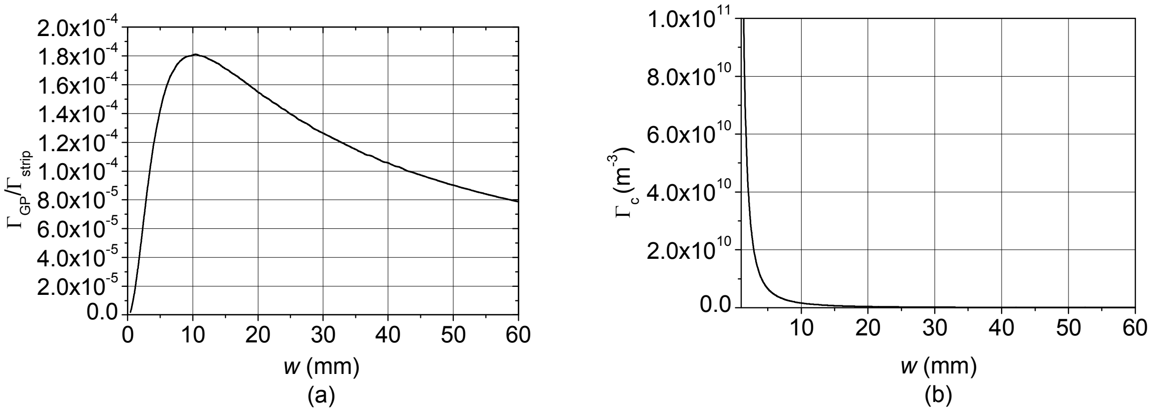

Fig. 2. (a) Ratio of the geometrical factors for the ground plane and for the strip of the microstrip line; (b) geometrical factor of the conductor for the microstrip line versus the strip width.

In order to find the integral in (9) the simple quasi-static approximation for the surface current on the thin strip is used:

$$J(x) = \frac{I}{\pi\sqrt{\left(\frac{w}{2} - x + \Delta\right)\left(\frac{w}{2} + x + \Delta\right)}}, \tag{10}$$

where w is the strip width and the $\Delta = \frac{\sqrt{2}}{\pi Q^2}\delta$ is the correction parameter accounting for the proper current behavior at the edge for the strip with the skin-depth δ, $Q \approx 2.3392$, c.f. [11]. The current distribution (10) does not contain singularity at the edges and therefore its substitution into (9) results in a convergent integral, which can be evaluated either analytically or numerically.

Besides the contribution of the strip and the ground plane should also be assessed. As shown in [12], the ground plane current can be evaluated as follows:

$$J_{GP}(x) = -\frac{h}{\pi}\int_{-w/2}^{w/2} \frac{J(x')}{(x-x')^2 + h^2}\,dx'$$

where h is the substrate height.

Let us introduce the geometrical factors for the ground plane as $\Gamma_{GP} = \Gamma_c(J_{GP})$ and for the strip as $\Gamma_{strip} = \Gamma_c(J)$; the sum of these two factors yields the resultant geometrical factor for the microstrip line. The ratio of these two factors for the microstrip lines of different width placed on a substrate of the height 1.58 mm is given in Fig. 2a. As one can see the contribution of the ground plane is negligible as compared with the strip. This is an important fact, which leaves the nonlinear impact of the ground plane out of consideration, provided the microscopic nonlinearity of the ground conductor is the same as of the strip, e.g. all the conductors are made of the same material.

The dependence of the geometrical factor on the microstrip line width is shown in Fig. 2b. This factor rapidly decays with the line width. The importance of this observation will be emphasized further in the paper.

PASSIVE INTERMODULATION ON THE LINES WITH NONLINEAR DIELECTRIC

In case of the nonlinear dielectric one can replace in (1) and (2) the nonlinear resistance by weakly nonlinear capacitance

$$C(U) = C_0 + C_2U^2, C_0 >> C_2U^2, \tag{11}$$

where C_0 is the linear capacitance and C_2 is the macroscopic nonlinear parameter of the dielectric. Then telegrapher's equations can be reduced to the nonlinear differential equation with respect to unknown voltage $U(z,t)$

$$\frac{\partial^2 U(z,t)}{\partial z^2} - C_0 L \frac{\partial^2 U(z,t)}{\partial t^2} - \left(C_0 R + GL\right)\frac{\partial U(z,t)}{\partial t} - GRU(z,t) =$$
$$= C_2\left(LU(z,t)^2 \frac{\partial^2 U(z,t)}{\partial^2 t} + 2LU(z,t)\left(\frac{\partial U(z,t)}{\partial t}\right)^2 + RU(z,t)^2 \frac{\partial U(z,t)}{\partial t} \right)$$

Solving the problem by perturbation method with respect to the small parameter C_2, one finds expressions for PIM3 power proportional to C_2^2. Therefore, this parameter will determine magnitude of the PIM products. Here, the expressions are omitted for brevity; however, those can be easily obtained similarly to [8].

In order to determine the nonlinear factor C_2, the electrostatic approach is employed here. Let us assume the third order nonlinearity of the electric displacement in the dielectric substrate:

$$\overrightarrow{D} = \varepsilon_0 \left(\varepsilon_1 + \varepsilon_2 \left|E\right|^2 \right)\overrightarrow{E} \tag{12}$$

where ε_0 is permittivity of vacuum, ε_1 – linear relative permittivity of the substrate and ε_2 is nonlinear dielectric parameter.

The electric energy stored in the dielectric is determined as follows:

$$\iint_S \left(\overrightarrow{E}\cdot\overrightarrow{D}\right) ds = C(U)U^2$$

or using (11) and (12)

$$\iint_S \varepsilon_0\varepsilon_1 \left|E\right|^2 ds + \iint_S \varepsilon_0\varepsilon_2 \left|E\right|^4 ds = C_0U^2 + C_2U^4 \tag{13}$$

where S is the cross-section of the dielectric substrate.

By equating the nonlinear terms in both sides of (13), one obtains the macroscopic nonlinear parameter

$$C_2 = \varepsilon_0\varepsilon_2\Gamma_d$$

where

$$\Gamma_d = \frac{\iint_S \left|E\right|^4 ds}{U^4}$$

is the geometrical factor for the dielectric, which relates the macroscopic nonlinear parameter C_2 dependent on the geometry of the cross-section to the microscopic nonlinear parameter of the dielectric ε_2.

EXPERIMENTAL DISCRIMINATION OF SOURCES OF PASSIVE INTERMODULATION

Two sets of experiments were prepared and carried out in order to discriminate the different sources of PIM on microstrip lines. Each set contained several microstrip lines of different width printed on the same substrate. The first substrate had thickness 1.58 mm and relative permittivity 2.5, while the second substrate was 0.76 mm thick and had relative permittivity 3. The conductors on both laminates were made of 0.035 mm thick low-profile reverse treated electrodeposited copper with 1 micron thick immersion tin finishing. The microstrip lines were 914 mm long with 522 mm central sections of different width and Klopfenstein tapered sections providing matching to the 50 Ohm ports. The widths of the central sections were 2.29, 4.32, 9.14 and 13.46 mm for the first substrate and 0.98, 1.9, 3.78 and 7.44 mm for the second one. Each printed line was fitted with a pair of coaxial cable launchers with the external DIN 7/16 connectors, cf. [14]. The S-parameters test showed the return loss better than -30 dB in the GSM900 band for the assembled samples. The experiment was carried out with the Summitek SI-900B PIM analyzer at 2x43 dBm carriers set to 935 and 960 MHz, which produced PIM3 frequency of 910 MHz. The residual PIM level was below -124 dBm.

Details of the measurement technique are beyond the scope of this paper and the readers are referred to [14]. The results of the forward PIM measurements and the simulation for the samples are shown in Fig. 3a for the first laminate and in Fig. 3b for the second one. As one can observe in Fig. 3a, PIM level decreases with the strip width and, therefore, the nonlinear conductor apparently makes dominant contribution to the PIM generation. It should be noted that this particular set of measurements has rather high uncertainties (shown as error bars in Fig. 3a), which are associated with the external noise. The latter is very difficult to exclude when the measured PIM products are close to the residual level.

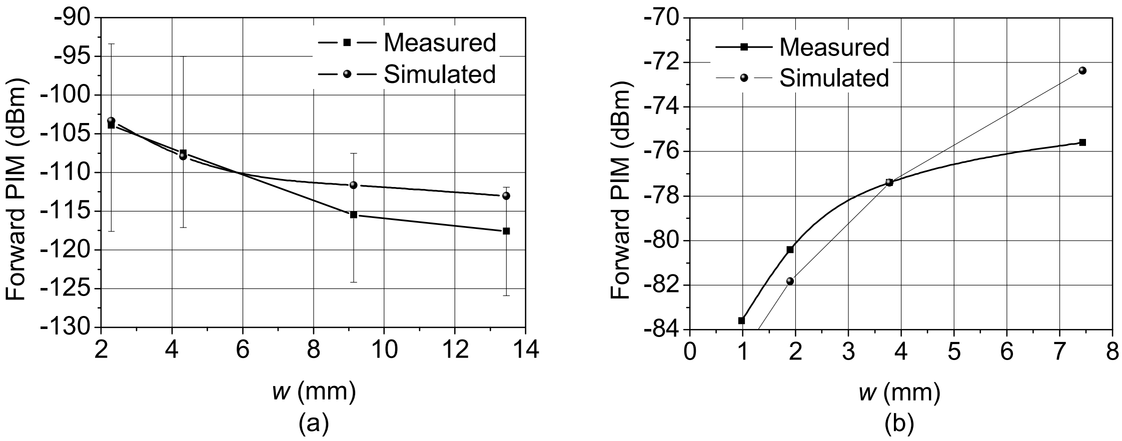

Fig. 3. (a) Forward PIM on the microstrip line with the nonlinear conductor versus the strip width; (b) forward PIM on the microstrip line with the nonlinear dielectric versus the strip width.

On the contrary, the measured PIM for the second laminate reveals an opposite trend of the increased PIM level with the line width. Based on the theory presented above one can identify the dielectric substrate as the main contributor to nonlinearity. The measurement uncertainty for the second laminate is about 1-2 dB and not shown in Fig. 3b.

The nonlinear microscopic parameters in the simulations were fitted to the experimental data for each width and then averaged over the set of widths. In the case of the nonlinear conductor, forward PIM power was calculated with (5), i.e. taking into account the current and voltage distribution along the line, and the parameter R_2 has been obtained for each strip width. Then, the geometrical factor (9) was calculated for the lines and the microscopic nonlinear parameter ρ_2 was evaluated with (8). The procedure for the nonlinear dielectric was similar. The average values were found to be $\rho_2 = 1.1\times10^{-15}\ \Omega\ m^2/A^2$ for the first laminate and $\varepsilon_2 = 5\times10^{-17}\ m^2/V^2$ for the second one. Good agreement between the simulations and measurements verifies our theoretical assumptions. The developed model offers a means for discrimination of the PIM generation mechanisms in microstrip lines.

CONCLUSIONS

The paper presents the theoretical considerations for analysis of passive intermodulation generation on microstrip lines due to nonlinear dielectric or conductor. The theory is based upon nonlinear transmission line model with the phenomenological nonlinear per-unit-length parameters. These nonlinear parameters determine the relationships between the geometrical parameters of the microstrip cross-section and the microscopic nonlinear resistivity of the conductors or permittivity of dielectric substrate. In particular, the analysis of these relationships has revealed that the contribution of the ground plane is negligible in comparison with the strip in the case of nonlinear conductor. Also, it has been found that for the nonlinear conductor PIM level decreases with the strip width, while for the nonlinear dielectric the opposite effect takes place. These facts were experimentally verified and good agreement between the measurement and the simulation results has been achieved. This effect can be applied to discrimination of the dominant PIM sources on microstrip lines.

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