On Optimum Design of Planar Microwave Components under Linearity Constraints


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On Optimum Design of Planar Microwave Components under Linearity Constraints

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INTRODUCTION

Basic signal impairments in printed circuits arising from the intrinsic electrical nonlinearities of strip conductors and substrate materials have been duly recognised and studied since 1990-s, particularly in application to superconducting microstrip lines (SCML) [1-3] and tuneable microstrip lines on ferroelectric [4-5] and liquid crystal substrates [6]. The reported simulation and experimental studies of the third-order intermodulation (IM3) in SCML resonators and filters [3, 7-9] revealed some characteristic features of nonlinear products which suggested mitigation of the nonlinear response by optimising the filter topology. In particular, in [3] using a simple second-order polynomial model of nonlinear p.u.l. inductance, it was shown by simulations that the output IM3 product generated in a microstrip resonator is proportional to the power four of the resonator loaded quality factor and inversely proportional to the resonator length squared and microstrip width squared. The output IM response of a filter comprising several coupled resonators with varied loaded Q, resonance frequency and position along the main transmission line was calculated using a simple circuit model to reveal the peaking IM level at the bandpass edges and individual contributions of the constituent resonators. The sharp band edges proved to be associated with a higher IM peaking, which is also consistent with the analysis in [7]. As the result, it was suggested that the IM generation due to the intrinsic superconductor nonlinearity can be reduced by increasing the width of the transmission line (TL) sections, thus decreasing the current density at the strip edges, as well as by finding optimum individual resonator Q-factors, resonance frequencies and locations with the aid of the simulated annealing technique. Interestingly, it appeared that high-order resonator filters allow greater improvement of the output linearity. It was also observed that directly reducing the peak current density in the selected resonators that contribute stronger to the total IM response does not necessarily result in the minimum IM [3], which warrants the need of a holistic approach to the filter design. Alternative approach is based on the microstrip resonator topologies and modes that do not have associated peak currents, such as the TM00 mode of a disc resonator in [10].

The important aspect of the previous studies, particularly those concerned with the SCML microwave resonators and filters, is the use of specific models of the intrinsic nonlinearity, albeit phenomenological and subject to experimental characterisation, which nevertheless refer to the pre-determined physical mechanisms, i.e., deterministic dependences of the superfluid density on the electric current density [11-12]. However, when it comes to the characterisation of the intrinsic nonlinearities in ordinary planar circuits, fabricated on commercial RF laminate materials and operated at ambient temperatures, exact mechanisms and location of nonlinearities are usually unknown a priori, so that the net nonlinear response is rendered by multiple concurrent sources that may change in time being influenced by many unpredictable factors. The great diversity and variability of the TL nonlinearities, including both distributed and lumped (contact) sources, requires a different approach to characterisation, modelling and mitigation of the passive intermodulation (PIM) generation in planar microwave circuits, resonators and filters.

In this paper, we adopt an efficient approach to accurate characterisation of the sources and mechanisms of intrinsic nonlinearities in microstrip lines based upon the conventional two-tone PIM measurements. A simple physical model of distributed PIM generation in uniform microstrip lines is presented, as well as the model of discontinuities. A numerical implementation of the developed model using commercial RF CAD software is proposed and a simple microstrip filter is designed and simulated. It is shown that the filter topology can be optimised to decrease both input and output PIM response without sacrificing the linear performance. The results of our study suggest the need of further development of efficient CAD tools for the design of complex planar microwave circuits under the linearity constraints.

MECHANISMS AND SOURCES OF INTRINSIC NONLINEARITIES IN MICROSTRIP LINES

Previous experimental studies of PCB materials for base-station antenna applications suggest strong correlation between the PIM products in printed lines and characteristics of the constituent materials. In particular, stronger PIM generation
was observed on PCB laminates with higher conductor roughness profile, [13]. In further experimental studies, it was also observed that significantly higher PIM can be generated due to the substrate nonlinearity, [14], but again no specific sources of the nonlinear behavior have been identified. Some studies performed on the base-station grade PCB materials and circuits consistently revealed strong effect of the PCB plating on the resulting PIM generation, as well as the peculiar effects of the PCB etching, [15]. It should be noted that in almost all systematic studies of PIM generation in commercial grade PCB materials the experimental methodology was usually based on the basic two-tone PIM tests complemented by carrier power and/or frequency sweeps, and the PIM products were measured at the input or output terminals of the tested specimens. However, such PIM characterisation does not provide sufficient details required for understanding the physics of PCB nonlinearities and devising rigorous engineering approaches to PIM mitigation.

The advanced experimental methodologies, specifically those based on the near-field PIM probing [16] and broadband intermodulation measurements with closely spaced carriers [17], allowed further insights in the mechanisms of PIM generation in printed circuits. Specifically, the distributed nature of the PCB nonlinearity, omni-directional emission by a lumped PIM source and the effects of the phase synchronism on distributed PIM generation have been confirmed experimentally. It also revealed that diverse circuit parameters and intrinsic dynamics can contribute to nonlinear distortions as well, e.g., ohmic in microstrip circuits with temperature-dependent resistivity [17]. Despite the progress made in understanding the mechanisms of microstrip nonlinearities, the nonlinear models used are still based upon basic static polynomials, whereas the model development and assessment have been largely overlooked. Recent theoretical studies of nonlinearities represented by an equivalent memristor and non-analytical characteristics [18], suggest that accurate model retrieval generally require simultaneous measurement of a wide range of intermodulation products, which is incompatible with the specification of the available commercial and laboratory PIM analyzers.

Our recent experimental investigations of PIM generation in printed circuits fabricated on base-station antenna grade PCB laminates suggest that certain microstrip nonlinearities often result from the PCB processing. For the present study, we have selected the microstrip lines of different strip width, Fig. 1(a), exhibiting discernible “black tin” contamination along the strip edges, as detailed in [19]. The used PTFE-based laminate with a low profile 1 oz. copper cladding is qualified as a low-PIM material. The conductor pattern was finished with 1 um immersion-tin plating. The PIM signal at frequency \( f_{IM3} = 2f_1 - f_2 = 910 \) MHz, where the carrier frequencies are \( f_1 = 935 \) MHz and \( f_2 = 960 \) MHz, appeared to be more than 30 dB higher than the level measured on the same layout and material processed by the same PCB workshop, but without visible contamination at the strip edges. The measured characteristic decrease of the PIM level versus the microstrip width indicated conductor-type nonlinearity, so that the lower current density on wider strips results in respectively lower PIM level. The effect of the return current on the ground-plane is much weaker due to the lower current density. The near-field mapping of the PIM product distributions along and across signal strip suggests that from the perspective of PIM modelling in our study, it is reasonable to assume a continuous distribution of the conductor nonlinearity over the signal strip surface, thus ignoring the actual discreteness of the “black tin” PIM sources at strip edges.

**DISTRIBUTED MODEL OF PIM GENERATION IN MICROSTRIP LINES**

From the viewpoint of the PIM characterisation in microstrip lines two main approaches can be adopted. The behavioural modelling can be employed for nonlinear characterisation, where a section of microstrip circuit is represented as a “black box” described by a global polynomial model providing the circuit output signal as the function of the circuit input signal, [20]. For example, to analyse PIM performance of a planar microwave filter, it can be represented by interconnection of TL sections and canonical discontinuities. So, once each constituent element has a measured behavioural model, the filter response can be calculated with the aid of commercial RF CAD capable of cascading multiple nonlinear components. Such an approach is not always convenient, since it requires repeated measurements of the microstrip circuits if some of the parameters changes, e.g., microstrip line length or width, or even the power of the input signal.

A more rigorous approach is based on equivalent circuits, where the circuit behaviour is represented by equivalent components characterised by explicit dependence on the microstrip geometry and materials, [21]. In particular, the nonlinear transmission line models have been extensively used for the analysis and modelling of distributed PIM generation in printed transmission lines and the equivalent lumped element circuits and applied to the analysis of nonlinear effects in microstrip resonators and filters [3, 9]. Utility of the latter approach, especially when the location of nonlinear sources is concerned, is determined by a choice of the method for nonlinear characterisation.
As discussed above, we can assume that the microscopic PIM sources are evenly distributed and current-dependent. They can be described by the equivalent circuit of the TL unit cell with the current-dependent p.u.l. resistance and/or current-dependent p.u.l. inductance. It has been shown elsewhere that the effects of the resistive and reactive nonlinearities on transmission lines can be discriminated by means of broadband measurements, i.e., can be deduced from the concurrent measurements of the third-order intermodulation and third harmonic responses. However, commercial PIM analysers available for our study did not support such a broad frequency range. Fortunately, the effects of nonlinear kinetic inductance, [22], is negligible in conventional printed lines so that nonlinear current-dependent resistance is the principal source of distributed PIM generation in printed circuits.

At the next step of the analysis, it is necessary to define a model of resistive nonlinearity. In most practical cases of weakly nonlinear dependence of the TL parameters on the electric current, \( I \), is described by a static polynomial with only one current-dependent term. In principle other models are possible, e.g., to account for the limiter and memory effects. The frequency-sweep measurements of the PIM products of different orders on the microstrip lines tested in our study showed constant PIM level across the whole measured band when either of the carrier frequencies sweeps across the Tx band. The PIM level also remained frequency invariant at different carrier powers. These observations suggest negligible memory effects and circuit dynamics, so the resistive nonlinearity can be approximated by the polynomial:

\[
R'(I) = R_0' + R'_\text{nl}(I) = R_0' + R'_2 I^2 + R'_4 I^4 + \ldots = R_0' + \sum_{n=1}^{(N-1)/2} R'_n I^{2n},
\]

(1)

where \( N \) is odd and the coefficients \( R'_n \) are retrieved from experimental data. It is noteworthy that the nonlinear resistance \( R'_\text{nl}(I) \) of the TL unit cell is described by the polynomial of \( I^2 \). Validity of such an approximation has been verified by cross-band PIM measurements, where the sum product of the carrier frequencies from the E-GSM 900 band was measured in the PCS 1900 band. The measured level of the second-order PIM products was below the noise floor that corroborates the assumption used in the model (1) that no even-order PIM products are permitted.

Given the nonlinear transmission line model with the resistive nonlinearity (1) is supported by the experimental observations, the complete mathematical description of the nonlinear voltage and current waveforms on a uniform microstrip transmission line is given by the generalised telegrapher’s equations:

\[
\frac{\partial I(x,t)}{\partial t} = -e_0 C \frac{\partial V(x,t)}{\partial t} - G_p' V(x,t)
\]

\[
\frac{\partial V(x,t)}{\partial t} = -L' \frac{\partial I(x,t)}{\partial t} - R'_2 I(x,t) - R'_4 (I) I(x,t)\]

(2)

The linear p.u.l. parameters \( R'_0 \), \( L'_0 \), \( C'_0 \); and \( G'_p \) can be calculated by a quasi-static analysis and directly related to the microstrip geometry. Using the Fourier series expansion under the assumption of weak nonlinearity, the nonlinear voltage and current waveforms on a uniform microstrip transmission line can be approximated by the generalised telegrapher’s equations:

\[
P_{\text{fwd}}(L) = 0.5 \text{Re} \left[ Z_{z_{r_{\omega}}} \right] \left| A_m \right|^2 \left( |M| - 1 \right)^2 \alpha^2 I e^{-2\omega L},
\]

\[
P_{\text{rev}}(L) = 0.5 \text{Re} \left[ Z_{z_{r_{\omega}}} \right] \left| A_m \right|^2 \left( |\beta_M| \right)^2 \left( |M| - 1 \right)^2 \alpha^2 \sin^2 \left( \beta_M L \right),
\]

(3)

where \( A_m = \sum_{l=1}^{N-1} \frac{2l+1}{\gamma_{l+1}} \frac{i}{2^{\frac{l}{2}}} \frac{\gamma_{l+1} R'_2}{\gamma_{l}^2} \left( l+1 - 0.5(M+1) \right)^2 \binom{k}{l} \binom{l+1}{k} \) are the binomial coefficients, \( \hat{I} \) is the current amplitude assumed the same for both carriers at a carrier frequency, \( \gamma_M = \sqrt{\left( R'_0 + j\omega L'_0 \right) \left( G'_p + j\omega C'_0 \right)} \approx \alpha + j\beta_M \), and \( \chi_M = |M| \alpha + 0.5 j \left( |M| + 1 \right) \beta \left( |M| - 1 \right) \beta_M \). Closed-form expressions (3) show that the forward PIM increases monotonically with the line length at \( \alpha L << 1 \). Conversely, the reverse PIM shows regular undulations as the line length increases. Equations (3) can be directly used to characterise the TL nonlinearity defined by (1). Specifically, forward or reverse PIM power at the corresponding terminal of a matched uniform TL can be measured in a range of carrier power magnitudes and then the model of the nonlinear p.u.l. resistance (1) with \( N \) specified can be retrieved by fitting the
analytical curves (3) to the experimental results. Alternatively, near-field PIM probing can also be used for the model retrieval, but in slightly different analytical form.

The closed-form equations (3) implicitly incorporate the PIM dependence on the strip width through the parameters of the nonlinear model (1). The width-dependent nonlinear parameters can be derived by the quasi-static analysis. In particular, considering the phenomenological volume resistivity \( \rho(J) = \rho_0 + \rho J^2 + \ldots \), where \( J \) represents the non-uniform current density, the corresponding coefficients of the nonlinear p.u.l. resistance are given by, cf. [12]:

\[
R_{2n}' = \left[ \frac{\rho_2}{\delta^2 + (2n+1)^2} \right] \int_{-W/2}^{W/2} J(y) dy,
\]

where \( J(y) \) is the surface current distribution across the thin signal strip of width \( W \) and \( \delta \) is the skin depth. In particular, in parallel-plate approximation, i.e., at low frequency, equation (4) reduces to \( R_{2n}' = \frac{\rho_2}{\delta^2} \), which contributes to the dependence of PIM on the strip width, along with the Wheeler’s characteristic impedance and other width-dependent parameters in (3).

The frequency dependence of the PIM products is also implied in the closed-form expressions (3) through the skin depth and microstrip characteristic impedance. At high frequency the ordinary p.u.l. resistance is asymptotically proportional to \( \sqrt{f} \), whereas the p.u.l. inductance depends on \( 1/\sqrt{f} \), [23]. Experimental data often show the PIM level increasing with the frequency. To clarify this effect, the studied microstrip lines were measured for PIM in two different frequency bands, e.g., using carrier frequencies at Tx E-GSM 900 or Tx PCS 1900 frequency bands and measuring the lower-sideband third-order PIM product in the corresponding Rx bands. The measured forward PIM3 products (i.e., third-order PIM level at the output of the microstrip line) at frequency \( f_{\text{BST}} = 2f_1 - f_2 = 910 \text{ MHz} \) (carrier frequencies \( f_1 = 935 \text{ MHz} \) and \( f_2 = 960 \text{ MHz} \)) and at 1870 MHz (\( f_1 = 1930 \text{ MHz} \) and \( f_2 = 1990 \text{ MHz} \)) generated on the test microstrip lines of different width are shown in Fig. 1(b) for the carrier power \( P_0 = 40 \text{ dBm} \). Intriguingly, the measured PIM values show virtually no difference in the two frequency bands, which also fit with the theoretical analysis based on (3) and confirms that the nonlinear parameters of the p.u.l. resistance are practically independent of frequency.

![Fig. 1. (a) Test PCB layout and (b) measured forward PIM3 products vs. strip width as the frequency band changes.](image)

The developed theoretical analysis, in principle, can also be extended to account for the effects of terminal mismatches and nonlinearities. However, in practice the assumptions made in the derivations hold only with a limited accuracy, so that the use of approximate model (3) becomes quite cumbersome a task in the practical design. To facilitate more general analysis, a discretized TL model has been implemented with a commercial RF CAD, as discussed next.

**RF CAD MODEL OF PIM IN COMPLEX MICROSTRIP LAYOUTS**

The theoretical analysis presented in the previous section, in principle, enables basic characterisation of microstrip nonlinearities under some reasonable assumptions. The latter, however, are still difficult to validate with the aid of commercial PIM analysers and simple microstrip layouts. A planar microwave filter containing sections of uniform microstrip lines and canonical discontinuities, e.g., width steps, gaps and shunt or series stubs, can be used for this purpose. The PIM characterisation of individual elements is still necessary for accurate analysis of filter response, e.g., using the equivalent circuit approach. This is where the RF CAD software becomes indispensable for the nonlinear characterisation and modelling of arbitrary conductor layouts and evaluation of signal impairments caused by PIM.
In this study, we use the harmonic balance solver and X-parameters for cascading nonlinear components in the Keysight ADS software. For characterisation of microstrip nonlinearity, the model of a uniform microstrip line has been implemented with the cascaded T-networks representing electrically short sections of a transmission line, so that the effects of the TL discretisation be negligible. It was actually found that the short T-cells provide more accurate description of the NLTL than the L-networks, although some authors suggest that an all-pass lattice filter topology suits best for distributed nonlinear modelling. [24]. The TL nonlinearity model (1) has been retrieved for each test microstrip line of specific width by fitting the simulated forward PIM3 products to the measurement data in a range of carrier power values, see Fig. 2. It appears that the fitting accuracy strongly depends on the model order, so that a large number of polynomial terms are required to achieve the residual error. This can be attributed to a systematic error in PIM measurements. Also, the model proved to be sensitive to the carrier power, and an accurate prediction of higher order PIM products from the PIM3 measurements requires even more polynomial terms.

The model has been verified experimentally for each width of the microstrip line further employed in the design of a simple low-pass filter in the next section. The effect of a step-width discontinuity was simulated as a moderate increase of the surface current density at the junction of the narrow and wide strip sections, Fig. 3(b). Such a current bunching has a minor effect on the distributed forward PIM generation, see Fig. 4, thus suggesting that the step-width discontinuity can be modelled by a simple nonlinear T-network consisting of two series nonlinear impedances connected through a shunt linear capacitor. Each nonlinear impedance is deduced from the model of the corresponding (i.e., same width) uniform transmission line. Validity of the proposed approach is demonstrated in the next section by comparing the results of simulations and measurements for a low-pass microstrip filter with the step-in-width discontinuities. An alternative nonlinear characterisation can be based on the measured X-parameters of sample TL discontinuities, but this approach is not supported by the available commercial PIM analysers.
Fig. 4. Measured forward (a) and reverse (b) PIM3 products at 910 MHz on the “quality” microstrip lines in Fig. 3(a). (NB: the “quality” lines do not show discernible “black-tin” edge contamination, in contrast to those in Fig. 1).

ANALYSIS AND DESIGN OF MICROSTRIP FILTERS UNDER LINEARITY CONSTRAINTS

The semi-phenomenological model of microstrip nonlinearities outlined in the preceding section is applied here to the analysis of simple microstrip filters. Figure 5(a) shows the layout of a low-pass microstrip filter comprised of nine sections of uniform microstrip lines of various widths and lengths, and eight step-width discontinuities. The filter has been designed to be well matched in the E-GSM band (880 - 960 MHz). The filter and test coupons of microstrip lines and discontinuities are fabricated on one panel of the PCB laminate. Simulated and measured forward and reverse PIM3 products of the test filter versus input carrier power are shown in Fig. 5(b). The results are in good agreement. The magnitudes of the simulated and measured PIM3 products have ~3 dB offset, which is likely to be caused by fabrication tolerances leading to a small disparity of the nonlinear parameters of the printed filter and test coupons. Another reason for the observed discrepancy could be attributed to the greater difference between the measured and simulated PIM3 responses on the narrow microstrip sections - the model limitation that was also observed in our previous studies.

![PCB layout of the low-pass filter](image)

![Simulated and measured PIM3 products](image)

Fig. 5. (a) PCB layout of the low-pass filter and (b) its measured and simulated PIM3 products versus carrier power, $P_0$.

Obviously, minimising the PIM response in planar filters by adjusting the printed layout is always prone to degrading the small-signal performance. [3]. In order to assess how much improvement could be achieved and what implications are for the filter parameters, simulations have been conducted for a three-stub Chebyshev bandpass filter shown in Fig. 6(a). The variable circuit parameters are summarised as in Table 1 for the given circuit size constraint.

<table>
<thead>
<tr>
<th>Layout</th>
<th>$\theta_{p1}$, deg.</th>
<th>$\theta_{p2}$, deg.</th>
<th>$\theta_s$, deg.</th>
<th>$Z_s$, $\Omega$</th>
<th>$Z_{p1}$, $\Omega$</th>
<th>$Z_{p2}$, $\Omega$</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>70</td>
<td>76.5</td>
<td>130</td>
<td>50</td>
<td>42.05</td>
<td>21.02</td>
</tr>
<tr>
<td>II</td>
<td>80</td>
<td>82.5</td>
<td>107</td>
<td>41.9</td>
<td>42.05</td>
<td>21.02</td>
</tr>
<tr>
<td>III</td>
<td>90</td>
<td>90</td>
<td>90</td>
<td>40.2</td>
<td>42.05</td>
<td>21.02</td>
</tr>
<tr>
<td>IV</td>
<td>100</td>
<td>98.5</td>
<td>75</td>
<td>41.5</td>
<td>42.05</td>
<td>21.02</td>
</tr>
<tr>
<td>V</td>
<td>110</td>
<td>107</td>
<td>60</td>
<td>45</td>
<td>42.05</td>
<td>21.02</td>
</tr>
</tbody>
</table>
The nonlinear products generated in the considered filters were simulated using the ADS Harmonic Balance solver. The simulated forward and reverse PIM3 products at frequency $f_{IM3} = 2f_1 - f_2 = 910$ MHz generated by two CW signals of power 43 dBm and frequencies $f_1 = 935$ MHz and $f_2 = 960$ MHz are presented in Fig. 7(a). The results show that up to 8 dB reduction of the PIM3 level is achievable without noticeable change of the filter loaded Q. However, the filter band edges are significantly affected. The sharper roll-off at the passband upper edge correlates with the higher PIM, yet it is not clear without additional simulations which stub contributes most to the total PIM response. The simulation results of the in-band PIM3 product with the carrier frequency $f_1$ fixed and $f_2$ swept are shown in Fig. 7(b). Notably, the PIM3 level peaks at the band edges, with the higher level at the upper end corresponding to the sharper roll-off. One can speculate whether the section with the highest resonance frequency is responsible for the PIM3 peaking at the high frequency skirt, but predictions of our model proved to be fully consistent with the numerical analysis in [3]. The presented results are subject to further experimental verification, which requires implementation of the controlled distributed nonlinearity throughout the manufacturing process.

CONCLUSIONS

In this paper, we have presented a consistent approach to the nonlinear model assessment by means of conventional two-tone PIM measurements and efficient RF CAD simulations. Although the closed-form theoretical analysis can provide valuable insights in the effects of the physical and geometrical parameters, the underlying assumptions are often difficult to validate. Besides, analytical complexity rapidly grows for non-uniform and resonant circuits, which warrants the development of a more efficient empirical means of PIM characterisation and modelling. To that end, the RF CAD approach has been proposed to facilitate the nonlinear characterisation and PIM analysis of arbitrary circuit layouts and signals. Two examples of the microstrip filters based on the step-width discontinuities and shunt stubs have been presented and the simulation results confirm that PIM reduction is achievable by careful selection of the microstrip layout without frustrating the small-signal performance. Moreover, the presented simulation results are in a very good agreement with the experiments and published works, which suggests feasibility of improved PIM control in the future.
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