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25 April 2024

DISTORSION ANALYSIS OF NON-LINEARLY LOADED ANTENNAS

C.Naldi, R.Zich[+], F.Filicori[++]

+ Ist.Elettronica e Telecomunicazioni,Politecnico Torino,Italy.
++ Ist.Elettronica, Università di Bologna,Italy.

INTRODUCTION

Non-linearly loaded antennas and scatterers are of considerable interest in a number of technical problems and applications. Non-linear effects must be taken into account in systems of antennas which include semiconductor components, integrated circuits and voltage limiters with very strong incident fields such as those produced by lightning or by NEMP (Nuclear ElectroMagnetic Pulse)[1]. Moreover, some field measurement techniques such as the EDM (Energy Density Meter) now being developed at the NBS [2] make use of short dipoles loaded by a diode. It must also be noted that some radar applications with cooperative targets are based on the use of non-linearly loaded scatterers [3]. Examples are provided by the transponders with frequency duplication, used in anti-collision systems [4] in order to distinguish between the required signal and the clutter coming from the environment. Further applications regard transistorized or active antennas which are highly qualified elements in directional antenna arrays [5].

A suitable numerical approach in the steady state condition can be found as a generalization of the piecewise harmonic balance method, which is successfully applicable to non-linearly loaded antennas and scatterers [6]. In addition the simultaneous presence of several signals can be examined when carrier frequencies are close enough. In fact the intermodulation products can be approximately evaluated by means of the monochromatic circuit responses.

NUMERICAL ALGORITHMS FOR THE SYSTEM ANALYSIS

A new general formulation has been recently proposed [6] which allows strongly non-linear multiports (described through implicit equations) to be taken into account and leads to numerical algorithms quite fast even with hard non-linear constraints. According to this harmonic balance procedure the whole network is split into a subnetwork composed only of linear elements and subnetworks which include all the non-linearities. The electrical variables $\underline{v}(t)$ and $\underline{i}(t)$ at the M connecting ports can be assumed as unknowns. Let $\underline{x}(t)$ be the equivalent generators at the connecting ports.

Taking into account the constraints $\mathscr{L}\{\underline{v}(t),\underline{i}(t)\} = \underline{x}(t)$ imposed by the linear subnetwork and the hypotesis of a steady state response, the set of non-linear equations describing the non-linear subnetworks, becomes

$$(1) \qquad \underline{G}\left[\ \underline{I}(\omega)\ ,\ \underline{X}(\omega)\ \right]\ =\ \underline{0}$$

of M(N+1) non-linear time-independent equations, where $\left[\underline{I}(\omega),\underline{X}(\omega)\right] = \mathscr{F}\left[\underline{i}(t),\underline{x}(t)\right]$ and N is the maximum number of harmonics necessary to correctly represent time functions according to the Shannon's theorem.

To solve eq.(1), the use of the second-order convergence algorithm

(2) $\left(\dfrac{\partial G}{\partial \underline{I}}\right)_{\underline{I}^k} \cdot (\underline{I}^{k+1} - \underline{I}^k) = - \underline{G}(\underline{I}^k)$ with : $\dfrac{\partial \underline{G}}{\partial \underline{I}} \rightarrow [g_{ij}]$: $g_{ij} = \dfrac{\partial G_i}{\partial I_j}$

and $i,j = 1 \div M(N+1)$, reduces the problem to the search for the solu
tion of a sequence of linear systems such that the matrix of the co
efficients can be evaluated numerically or analytically, the unknowns
being the differences between the variables at two successive itera-
tions (k and k+1).

The algorithm (2) has a second order covergence rate provided that an
accurate starting point for the iterative process is chosen. In the ca
se of strong non-linearities a choice of $\underline{I}^0$ which is not close enough
to the true value the iterative process may not converge or may stop
at a local minimum which is not zero. To overcome these difficulties
the problem of solving system (2) with the forcing term $\underline{X}(\omega)$ as parame
ter can be transformed into a sequence of problems with the same stru
cture, but with different values of the signal amplitude, that is

(3) $\underline{X}(\omega) = \underline{X}^I(\omega) + h\left[\underline{X}^F(\omega) - \underline{X}^I(\omega)\right]$ $0 \leqslant h \leqslant 1$

where $\underline{X}^I(\omega)$ and $\underline{X}^F(\omega)$ are the initial and final values of the vector
of the forcing terms and h is a number ranging from 0 to 1. Usually a
solution of (2) is known for a particular value $\underline{X}(\omega)=\underline{X}^I(\omega)$, for exam-
ple, for zero signal amplitude. By starting from this value and increa
sing gradually by step the value of h, the direct search for the solu-
tion of system (1) can be split into a sequence of solutions of well-
conditioned systems. In fact, every solution obtained with a particular
signal amplitude $(h=h_0)$ is a good starting point for the system having
a slightly higher signal $(h=h_0+\Delta h)$. In this way convergence problems
are eliminated and the whole process is speeded-up. The technique can
be refined by improving the prediction of the correct starting point
for the iterative process, using an extrapolation procedure on the so
lutions obtained for weaker signals. The procedure described provides
as a further result, the responses of the system for different values
of the input signal amplitude, that is a very important feature in or
der to evaluate intermodulation products.

INTERMODULATION ANALYSIS FOR NARROW-BAND SIGNAL

A special feature of the above described harmonic analysis algorithm is
that it provides the non-linear system response for a whole set of in-
put signal amplitudes. Such a characteristic is useful not only by it
self, but also because it allows a straightforward computation of the
intermodulation products arising from a narrow-band input signal.

In order to show how the intermodulation analysis can be effected, let
us suppose that a periodic steady state analysis has been carried out
for a set of sinusoidal input signals

(3) $x_0(t) = b \cos \omega_0 t$

with different values of the amplitude b within a range $0 < b \leqslant B$. The

corresponding system output can be expressed throught a Fourier series in the form

$$(4) \quad y_0(t) = \sum_{n=0}^{N} \mathcal{R} \{ \underline{H}_n(b, \omega_0) \, e^{jn\omega_0 t} \}$$

where the dependence of the output harmonic components on the actual input signal amplitude b and angular frequency ω_0 has been pointed out throught the "harmonic transfer functions" $\underline{H}_n(b, \omega_0)$.

Once a sufficient number of samples of these harmonic transfer functions has been computed by a periodic steady state analysis, it is quite straight forward to compute the system response to any narrow-band input signal with spectrum centered at ω_0 and expressed in the form

$$x(t) = \int_{\omega_0-\Delta\omega}^{\omega_0+\Delta\omega} \mathcal{R} \{ \underline{X}(\omega) \, e^{j\omega t} \} \, d\omega = \int_{-\Delta\omega}^{\Delta\omega} \mathcal{R} \{ \underline{X}(\Omega+\omega_0) \, e^{j(\Omega+\omega_0)t} \} \, d\Omega$$

with $\Delta\omega \ll \omega_0$. In fact, the narrow-band signal x(t) can be expressed as a modulated signal with carrier frequency ω_0, amplitude a(t) and phase $\alpha(t)$ in the form :

$$x(t) = \mathcal{R} \{ \underline{A}(t) \, e^{j\omega_0 t} \}$$

where the complex envelope , given by

$$\underline{A}(t) = \underline{a}(t) \, e^{j\alpha(t)} = \int_{-\Delta\omega}^{\Delta\omega} \underline{X}(\Omega+\omega_0) \, e^{j\Omega t} \, d\Omega$$

can be easily computed by using the fast Fourier transform numerical algorithm. It should be noted that, since x(t) is a narrow-band signal, A(t) is a time function whose variations are much "slower" than those of x(t); therefore, the non-linear system response can be computed as a sequence of quasi-steady-state periodic responses. More in detail the output signal can be obtained as the sum of "slowly" modulated harmonic components in the form:

$$(5) \quad y(t) = \sum_{n=0}^{N} \mathcal{R} \{ \underline{D}_n(t) \, e^{jn\omega_0 t} \}$$

where the time dependent vector $\underline{D}_n(t)$ representing the n-th harmonic component is given by

$$(6) \quad \underline{D}_n(t) = \underline{H}_n \left[a(t), \frac{d\alpha}{dt} + \omega_0 \right] e^{jn\alpha(t)} \simeq \underline{H}_n \left[a(t), \omega_0 \right] e^{jn\alpha(t)}$$

The approximation in eq.(6) of the exact input signal angular frequency $\omega(t) \equiv (d\alpha/dt) + \omega_0$ with center band frequency ω_0 is reasonable under the assumption of the narrow band.
According to eqns (5) and (6) the system response to a given narrow – band input signal x(t) can be easily obtained once the harmonic transfer function $\underline{H}_n$ has been computed by using the harmonic analysis algorithm previously described. If, as it is more useful, the frequency domain representation of the intermodulation products is required, this can be ob

tained by a numerical transformation of the time domain output signal
computed through eqns (5) and (6).

STRAGHT-WIRE ANTENNA WITH A NON-LINEAR LOAD

As an example of application of the proposed method, a straght-wire anten-
na of length h and diameter 2a has been considered; the antenna is connec-
ted at center to a non-linear load (fig.1). In this simple case, the non-
linear subnetwork is given by the antenna itself and the connecting section
reduces to a single port, so that the network unknowns are directly given
by $V(\omega)$ and $I(\omega)$, that is the voltage and the current at the antenna and
load terminals. A monochromatic incident field at frequency 100 MHz is as-
sumed. The wire-antenna analysis program is based on the solution of the
integral equation for the wire current via the moment's method with trian-
gular test and expansion functions. The load is a backward diode: the non-
linear transition capacitance has been taken into account and a parallel
inductance is included in the linear part to assure the short circuit con-
dition for the DC current. Fig.2 presents the harmonic containt in terms of
the back radiated power versus the incident available power for an antenna
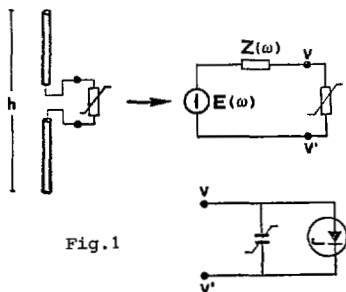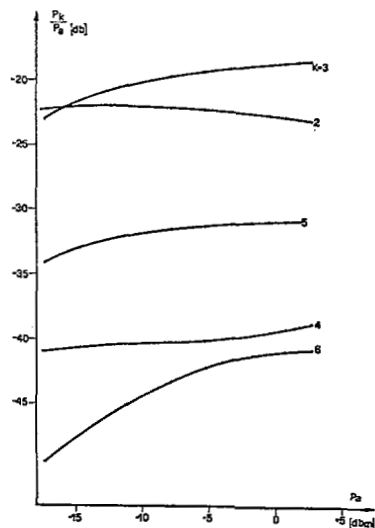with h=1.5 m and a=0.0025 m.

Fig.1

Fig.2

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