New System-Level Simulation of Noise Spectra Distortion in FM-CW Autonomous Cruise Control Radar

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***Abstract* — The design of complex MMIC such as ACC car radar involves the development of efficient and fast simulation techniques to predict the characteristics of the circuits. Their noise behavior simulation is one of the main drawback due to its complexity and great computation time. We propose in this paper, a new method based on circuit envelope to simulate the noise spectra conversion of a signal passing through MMIC. The efficiency of the proposed method is shown with the comparison between simulation and measures of the AM/PM conversion of a millimeter wave transmitter.**

I. Introduction

In recent years, considerable efforts have been made to develop autonomous cruise control or collision warning/avoidance systems. Demands for high volume, low cost, and small size, make solutions based on microwave monolithic integrated circuits (MMIC). For these car radar systems, the most common choice is the frequency modulated continuous wave (FM-CW) radar [1]. These systems operate at millimeter waves and the radar signal is built from X or Ku Band VCO, then converted through multipliers and amplifiers up to 77 GHz. Furthermore, this VCO is integrated in a complex PLL. A such architecture is shown Fig. 1

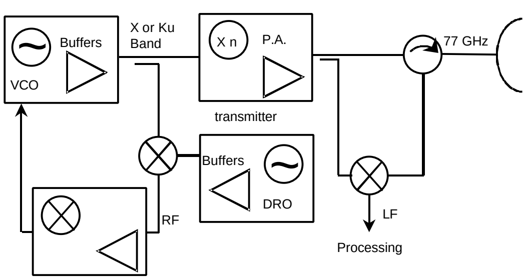

Fig. 1. Car radar system architecture

Each stage of this complex system is firstly designed with CW signals using a classical Harmonic Balance simulator. Then, the analysis of phase and amplitude noise spectra of the CW signal is made by means of a simulator using classical parametric analysis [2]. The circuit is represented by its conversion matrices (for the non-linearities), linear matrices (for linear circuit) and includes the noise generators. Finally, the transfer of the chirp signal through this source can be simulated by signal envelope. Now, characteristics such as PM/AM and AM/AM conversion in the transmitter is of key importance to evaluate radar performances. But, the simulation of the noise spectra distortion of this signal requires large CPU time and memory consuming , and this, in spite of implementation of efficient tools such as modified nodal analysis and numerical resolution by Krylov subspace algorithms.

In this paper, we propose a new method to simulate the nonlinear distortion introduced by the MMIC chip-sets of the system, on the noise spectrum of the driving chirp. This novel approach enables to compute conversion noise behavior with a reduced computation time and very good accuracy. It has been applied to the simulation of the AM/AM, AM/PM, PM/PM and PM/AM conversion of the transmitter of an ACC radar (Fig. 1) and compared with measures.

II. The Envelope Formalism

A. One tone signal

This method is based on the representation of the modulated signal by its complex envelope. A such narrow band modulated signal can writes as :

$$v_R(t) = \Re[v(t)] = \Re\left[\tilde{V}(t).e^{j.\omega_o.t}\right] \quad (1)$$

where ω_0 is the carrier frequency. The quantity :

$$\tilde{V}(t) = V(t).e^{j.\Phi_t(t)} \quad (2)$$

with $\Phi_t(t) = \Phi(t) + \delta\varphi(t)$ the sum of a deterministic term $\Phi(t)$ and a stochastic term $\delta\varphi(t)$ corresponding to phase noise, is the complex envelope of the signal v(t). Now, let us suppose that :

$$V(t) = V_0 + \delta V(t) \quad (3)$$

where $\delta V(t) << V_0$ corresponds to amplitude modulation.

- $\Phi_\tau(t)$ varies slowly, so $\frac{d\Phi_t(t)}{dt} << \omega_0$

The different spectra are related by :

$$V(\omega) = F(v(t)) = F(\tilde{V}(t).e^{j.\omega_0.t}) = \tilde{V}(\omega - \omega_0) \quad (4)$$

where F is the Fourier operator.

Let $H(\omega)$, the describing function of the circuit (for example the transmitter) for a input signal of constant amplitude V_0 and a frequency ω_0 :

$$H(V_0, \omega_0) = \hat{H}(V_0, \omega_0).e^{j.\Psi(V_0,\omega_0)} \quad (5)$$

For small perturbations, we can expand H at the neighborhood of ω_0 and V_0, the output signal envelope writes as :

$$\begin{aligned} \tilde{V}_s(t) = H(V_0, \omega_0).\tilde{V}(t) - j.\left.\frac{\partial H}{\partial \omega}\right|_{V_0,\omega_0} . \frac{d\tilde{V}(t)}{dt} \\ + \left.\frac{\partial H}{\partial V}\right|_{V_0,\omega_0} .\delta V(t).\tilde{V}(t) \end{aligned} \quad (6)$$

Note that the last term of this equation is due to the non-linearity of the circuit. By calculating $\frac{d\tilde{V}(t)}{dt}$ with (2), (6) can be written as :

$$\begin{aligned} \tilde{V}_s(t) = H\tilde{V}(t)\left[1 - j\left[\frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial \omega} - j.\tau_0\right]\left[j.\frac{d\Phi}{dt} + j.\frac{d\delta\varphi}{dt} + \frac{1}{V_0}\frac{d\delta V}{dt}\right]\right. \\ \left. + \left[\frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial V} + j.\frac{\partial \psi}{\partial V}\right].\delta V\right] \end{aligned} \quad (7)$$

where the group delay at ω_0 is defined as : $\tau_0 = -\left.\frac{\partial \psi}{\partial \omega}\right|_{\omega_0}$

Now, by writing δV and $\delta\varphi$ as amplitude and phase noise modulations

$$\begin{aligned} \delta V &= m_a(t).V_0 \\ \delta\varphi &= m_\varphi(t) \end{aligned} \quad (8)$$

from (7), we obtain

$$\begin{aligned} \tilde{V}_s(t) \approx H.\tilde{V}(t).\left[1 + \frac{d\Phi}{dt}.\frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial \omega} + \frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial \omega}.\frac{dm_\varphi(t)}{dt}\right. \\ \left. + \frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial V}.V_0.m_a - j.\frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial \omega}.\frac{dm_a}{dt} + j.\frac{\partial \psi}{\partial V}.V_0.m_a\right] \end{aligned} \quad (9)$$

Then, from (2), we can write, for the input signal :

$$\tilde{V}(t) \cong V_0.(1 + m_a(t) + j.m_\varphi(t)).e^{j.\Phi(t)} \quad (10)$$

and for the output signal :

$$\tilde{V}_s(t) \cong H.V_{0s}.(1 + m_{as}(t) + j.m_{\varphi s}(t)).e^{j.\Phi_{0s}} \quad (11)$$

Firstly, let us suppose a CW noisy input signal else $\Phi(t)$ is time independent. Let introduce (10) into (9) and by identifying (9) and (11), we obtain two relations between the output and input amplitude and phase noise modulations :

$$\begin{aligned} m_{as}(t) &= m_a(t) + \frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial \omega}.\frac{dm_\varphi(t)}{dt} + \frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial V}.m_a(t).V_0 \\ m_{\varphi s}(t) &= m_\varphi(t) - \frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial \omega}.\frac{dm_a(t)}{dt} + \frac{\partial \psi}{\partial V}.m_a(t).V_0 \end{aligned} \quad (12)$$

If we suppose that $m_a(t)$ and $m_\varphi(t)$ are sine signal at Ω pulsation and by taken Fourier transform of these equations, the amplitude and phase noise spectra are related by:

$$\begin{pmatrix} S_{as} \\ S_{\varphi s} \end{pmatrix} = \begin{bmatrix} \left(1 + \frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial V}.V_0\right)^2 & \left(\frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial \omega}.\Omega\right)^2 \\ \left(\frac{\partial \psi}{\partial V}.V_0\right)^2 + \left(\frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial \omega}.\Omega\right)^2 & 1 \end{bmatrix} \cdot \begin{pmatrix} S_a \\ S_\varphi \end{pmatrix} \quad (13)$$

This new result enables to compute in matrix form, the contribution to the AM and PM noise spectra of the output signal, of a nonlinear transmitter driven by an input CW noisy signal. The four terms of this matrix can be easily computed by a single tone non-linear HB analysis. In case where the circuit contains a multiplier by N, this method can be easily extended. We can write an identical relation as the one (13) :

$$\begin{pmatrix} S_{as_N} \\ S_{\varphi s_N} \end{pmatrix} = \begin{bmatrix} \left(1 + \frac{1}{\hat{H}_N}\frac{\partial \hat{H}_N}{\partial V}.V_0\right)^2 & \left(\frac{1}{\hat{H}_N}\frac{\partial \hat{H}_N}{\partial \omega}N.\Omega\right)^2 \\ \left(\frac{\partial \psi}{\partial V}.V_0\right)^2 + \left(\frac{1}{\hat{H}_N}\frac{\partial \hat{H}_N}{\partial \omega}\Omega\right)^2 & N^2 \end{bmatrix} \cdot \begin{pmatrix} S_a \\ S_\varphi \end{pmatrix} \quad (14)$$

The great interest of this method is its low computation time for the same result accuracy than the conversion matrix analysis.

B. Chirp signal

In ACC radar, when the frequency is modulated by a saw-tooth signal, (9) remains valid, since the signal has constant amplitude. The term $\Phi(t)$ becomes time dependent and the only difference is the term $\frac{d\Phi}{dt}.\frac{1}{\hat{H}}\frac{\partial \hat{H}}{\partial \omega}$ which denotes an additional deterministic amplitude modulation. With regard to noise behavior (13) and (14) remain identical for this modulated signal, nevertheless due to the small distance between the spectral lines, the noise spectra of adjacent lines overlap.

III. FM-CW SIGNAL SPECTRUM

A. Noiseless signal

A chirp signal [3] is the result of a FM modulated carrier by a saw-tooth signal shown in Fig. 2:

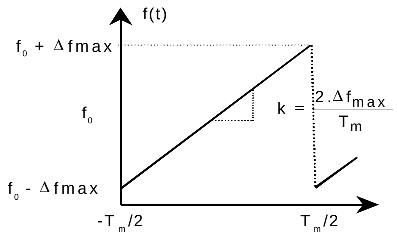

Fig. 2. modulated instantaneous frequency

Then the modulated signal writes as :

$$v_R(t) = \Re[v(t)] = V_0.\cos\left(\omega_0.t + \frac{k.t^2}{2}\right) \quad (15)$$

Its complex envelope writes as :

$$\tilde{V}(t) = V_0.e^{j.\frac{k.t^2}{2}} \quad (16)$$

From (1) and (4) the spectra of the signal and its complex envelope are related by :

$$V_R(\omega) = F(v_R(t)) = \frac{1}{2}.\left[\tilde{V}(\omega-\omega_0) + \tilde{V}^*(\omega+\omega_0)\right] \quad (17)$$

Else by calculating $V(\omega)$, we can obtain the spectrum of $v_R(t)$. $V(t)$ is periodic with period T_m. Then it can be expanded in Fourier series :

$$\tilde{V}(t) = \sum_{-\infty}^{+\infty} V_n.e^{j.n.\omega_m.t} \quad (18)$$

with $$V_n = \frac{1}{T_m}.\int_{-T_m/2}^{T_m/2} e^{j.\left(\frac{k.t^2}{2} - n.\omega_m.t\right)} dt \quad (19)$$

This integral can be analytically calculated and V_n becomes :

$$V_n = \frac{1}{2}\sqrt{\frac{f_m}{\Delta f_{max}}}.e^{-j.\frac{n^2.\omega_m^2}{2.k}}.\left(e(-x_1) + e(x_2)\right) \quad (20)$$

with e(x)=C(x)+j.S(x) where C(x) and S(x) are Fresnel Integrals with :

$$x_1 = \sqrt{\frac{\Delta f_{max}}{f_m}}.\left[1 + n.\frac{f_m}{\Delta f_{max}}\right], x_2 = \sqrt{\frac{\Delta f_{max}}{f_m}}.\left[1 - n.\frac{f_m}{\Delta f_{max}}\right]$$

Typically , f_m=2.5 KHz and Δf_{max}=100 MHz, then the line spectrum width is equal to $2.\Delta f_{max}$ with lines separated by 2.5 KHz : the spectrum contains 80001 lines. From this remark, the amplitude and phase of each line can be calculated from the asymptotic behavior of the Fresnel Integrals and the complex envelope of the signal v(t) finally writes as :

$$\tilde{V}(t) = \frac{1}{\sqrt{2}}.\sqrt{\frac{f_m}{\Delta f_{max}}}.\sum_{n=-N}^{n=N} e^{j.\left[n.\Omega t - n^2.\frac{\pi}{2}.\frac{f_m}{\Delta f_{max}} + \frac{\pi}{4}\right]} \quad (21)$$

with $N = \frac{\Delta f_{max}}{f_m}$. This spectrum modulates the carrier and from (1) we can write :

$$v_R(t) = \Re\left[V_0.\sum_{n=-N}^{n=N} e^{j(\omega_0 + n.\Omega)t + \Phi_n}\right] \quad (22)$$

$$\text{with } \Phi_n = -n^2.\frac{\pi}{2}.\frac{f_m}{\Delta f_{max}} + \frac{\pi}{4} \quad \text{and } V_0 = \frac{1}{\sqrt{2}}.\sqrt{\frac{f_m}{\Delta f_{max}}}$$

A rigorous computation has been performed and the resulting spectra of the noiseless FM-CW signal has been plotted in Fig. 3 :

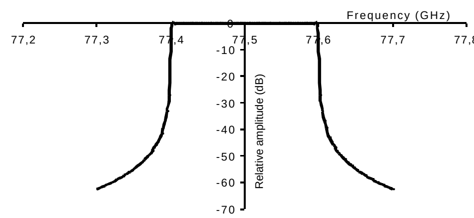

Fig. 3. modulating spectrum

Between $f_0 \pm \Delta f_{max}$ the spectrum is flat (This fact a posteriori justifies (3)).

B. FM-CW noisy signal

The computation of the spectrum of a FM-CW noisy signal is developed.

The AM and PM noise variation at the noise frequency Ω, write as :

$$\begin{aligned} \delta V(t) &= \Delta V.\cos(\Omega t + \varphi_a) \\ \delta \Phi(t) &= \Delta \Phi.\cos(\Omega t + \varphi_p) \end{aligned} \quad (23)$$

By extending the equation (2) to the first order, we have with $\Delta\tilde{V} = \Delta V.e^{j\varphi_a}$ and $\Delta\tilde{\Phi} = \Delta\Phi.e^{j\varphi_p}$:

$$\tilde{V}(t) \cong V_0.e^{j.\Phi_0}\left(1 + \frac{\Delta\tilde{V} + j.\Delta\tilde{\Phi}}{2}e^{j.\Omega.t} + \frac{\Delta\tilde{V}^* + j.\Delta\tilde{\Phi}^*}{2}e^{-j.\Omega.t}\right) \quad (24)$$

Now this voltage is modulated by the saw-tooth signal (fig. 2) and with equation (22) the relation (24) becomes:

$$\tilde{V}(t) \cong V_0 \sqrt{\frac{f_m}{2.\Delta f}} . \sum_{n=-\frac{\Delta f}{f_m}}^{n=\frac{\Delta f}{f_m}} e^{j.(n.\omega_m t + \Phi_n)} \times$$

$$\left(1 + \frac{\Delta\tilde{V} + j.\Delta\tilde{\Phi}}{2} e^{j.\Omega.t} + \frac{\Delta\tilde{V}^* + j.\Delta\tilde{\Phi}^*}{2} e^{-j.\Omega.t}\right) \quad (25)$$

This relation enables to conclude that the noise spectra computed in CW signal are shifted around each line of FM-CW signal. So the AM and PM spectra of FM-CW signal are the sum of the each corresponding power spectrum of the each lines, because the noise arising from each lines is uncorrelated with the noise of the other lines.

IV. Simulation of noise spectra conversion of the MMIC transmitter

The diagram of the transmitter which has been designed is shown Fig. 4.:

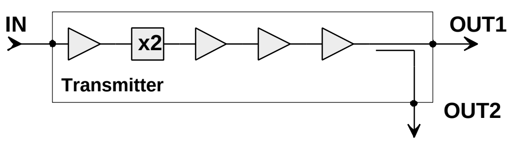

Fig. 4. Diagram of the transmitter

Its characteristics are the following: input frequency 38 – 38.5 GHz, output frequency 77 – 78 GHz, output power > 7 dBm. The MMIC was made with PH25 UMS process (transistors PHEMT of 0.25 µm gate width).

In a first time, to check our method, we have simulated the first stage by the two techniques (circuit envelope analysis and conversion matrices analysis). The four coefficients of the matrix (14) are given in the table 1 :

Matrix Coefficients	Conversion matrix Analysis	Circuit envelope analysis
S_{as}/S_a	-0.2 dB	-0.25 dB
S_{as}/S_ϕ	-125 dB	-124 dB
$S_{\phi s}/S_a$	-39 dB	-39.2 db
$S_{\phi s}/S_\phi$	-0.2 dB	0 db

Table 1: comparison between the two simulations

The results are in good agreement with the two methods, but the computation time ratio is one over 20 in favor of the envelope circuit analysis. Then, the whole transmitter has been simulated and measured. A comparison (make only for the measured AM/PM coefficient) and simulation results are shown in Fig. 5, plotted versus out frequency.

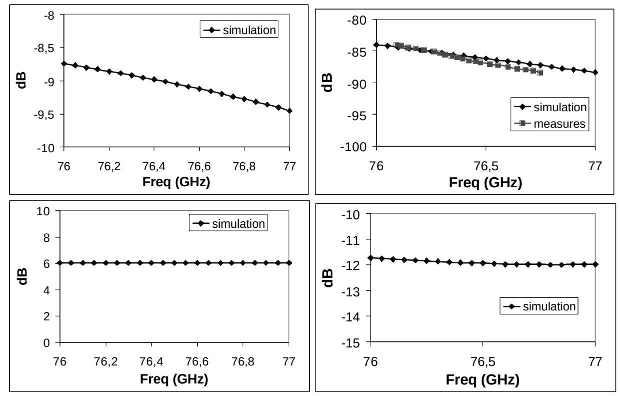

Fig. 5. Simulation and measures of the conversion AM/AM, AM/PM, PM/AM, PM/PM of transmitter

The output noise spectra of the simulated chirp signal is shown Fig. 6 for 7 spectral lines :

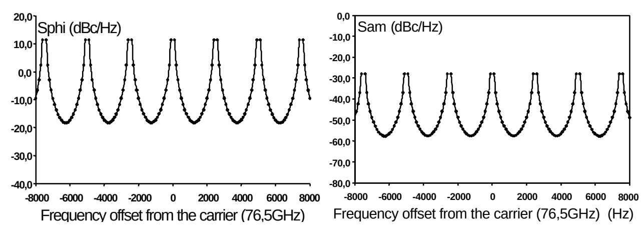

Fig. 6. PM and AM noise spectra of the chirp signal

V. Conclusion

This paper proposes a new method based on the analysis of the signal envelope to simulate the noise spectra conversion of a signal passing through MMIC. Its efficiency and short computation time are demonstrated. Simulation and measures of the AM/PM conversion of a millimeter wave MMIC chip transmitter are compared with good accuracy.

Acknowledgement

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