DETECTOR SIMULATION WITH THE USE OF THE DISCRETE TECHNIQUE

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This paper presents a functional-level detector simulation technique which is based on the Hilbert Transforms and can be used together with the Discrete Technique. AM-detector simulation as well as its constrains, improvements and validation has been discussed in detail. It has been shown that the RC-circuit at the detector output should be ignored during the simulation. The technique proposed can be used for a rapid nonlinear EMC/EMI modeling at the system/subsystem level.

1.INTRODUCTION

Numerical modeling of a radio electronic system is a very useful tool for electromagnetic compatibility/interference (EMC/EMI) analysis, in that it allows for the simulation of system behavior for a wide variety of initial conditions, excitations and system configurations in a rapid and inexpensive way [1]. A system can often reveal nonlinear behavior and nonlinear phenomena (intermodulation, cross-modulation, gain compression/expansion etc.) has profound effect on EMC/EMI in some cases [2]. A nonlinear modeling tool must be used in order to carry out EMC/EMI analysis in such a case.

A nonlinear modeling technique (so called 'discrete technique') for numerical EMC/EMI simulation at the system level has been proposed in [3]. This technique allows one to carry out rapid numerical EMC/EMI analysis of a complex system or subsystem (i.e. receiver, transmitter etc.) or a set of systems/subsystems in a wide frequency range taking into account nonlinear effects (including spurious responses of a receiver). Such an analysis is, for instance, a very important part of EMC/EMI modeling of a mobile communication system [4,5].

The basis of the discrete technique [3,6] is a representation of the equivalent block diagram of a system as linear filters (LF) and memoryless nonlinear elements (MNE) connected in series (or in parallel). Thus a stage which employs a nonlinear element, for example, a radio frequency amplifier, can be represented as a typical radio stage, which employs the linear filter at the input, the memoryless nonlinear element and the linear filter at the output [2], see Fig.1.

The process of signal passage through linear filters is simulated in the frequency domain using the complex transfer factor of the filter,

$$S_{out}(f_n) = S_{in}(f_n) \cdot K(f_n)., \tag{1}$$

where $S_{out}(f_n)$- is the signal spectrum at the filter output, $S_{in}(f_n)$ - the signal spectrum at the filter input, $K(f_n)$ - is the complex transfer factor of the filter, f_n - are sample

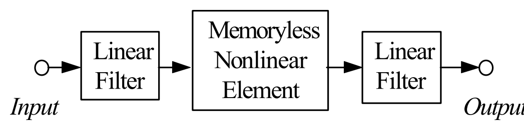

Fig.1. Representation of a typical radio frequency stage

frequencies. The process of signal passage through a nonlinear memoryless element is simulated in the time domain,

$$u_{out}(t_k) = \sum_{i=1}^{I} a_i u_{in}^i(t_k). \quad , \tag{2}$$

where $u_{out}(t_k)$ - is the instantaneous value of the signal at the MNE output, $u_{in}(t_k)$ - is the same for the MNE input, t_k - are sample points in time, a_i - are coefficients of the high-order polynomial which describes the transfer characteristic of the nonlinear element; I - is order of the polynomial. The transition from the time domain to the frequency domain and vice versa is made with the use of the direct and inverse Fast Fourier Transform (FFT and IFFT):

$$S = FFT(u) \ , \ u = IFFT(S) \ . \tag{3}$$

The determination of the sampling rate, the sample frequency interval, the number of samples as well as a polynomial synthesis technique have been discussed in [6]. Using this technique, a radio receiver can be simulated in a wide frequency range with a very high frequency resolution (up to 10^6 - 10^7 sample frequencies) on a modern PC in dozens of minutes (a conventional circuit-level simulation would require several years for such an analysis).

The essential limitation of this technique is that (1) the nonlinear element is to be memoryless (or, at least, the non-zero memory of the nonlinear element must allow an assignment to the linear filters), (2) the succeeding element in the functional block diagram does not influence the preceding one (in some cases, this influence can be taken into account by the use of the equivalent transfer factor). Therefore this technique cannot be directly applied to the detector simulation, which is an essential limitation on its possibilities - the simulation of a radio receiver can be carried out only as far as the output of the intermediate frequency (IF) path (the detector input) and, correspondingly, baseband signal processing cannot be simulated. But the detector and baseband signal processing can substantially influence the EMC/EMI situation, so it's desirable to have an appropriate simulation technique.

This article presents several methods used so as to simulate various (amplitude (AM), frequency (FM) or phase (PM)) detectors at the system (functional) level, which can be used together with the discrete technique. The method of an amplitude detector simulation, which is based on the Hilbert transforms, as well as its constraints, improvements and validation has been discussed in detail.

2. AM DETECTOR SIMULATION

Since the primary function of an AM detector is to generate output signal which is proportional to the amplitude of an input signal, it's necessary to calculate the input signal amplitude in order to carry out the simulation. A signal at the input of a radio receiver detector is, as a rule, a narrowband one, so the Hilbert transform can be used for this purpose (we should note that the requirement for the input signal to be narrowband is dictated by not Hilbert transform itself which can be applied to a broadband signal too, but by the RC-circuit present at the detector output. Further this issue will be discussed in detail).

For the sampled spectrum which is used in the discrete technique, the Hilbert transform takes the simplest form [7]:

$$S_n^* = -j \cdot S_n \,, \tag{4}$$

where $S_n = S(f_n)$ - is a sampled spectrum, S_n^* - is the spectrum of the Hilbert conjugate signal, j - is the imaginary unit. The input signal amplitude can be obtained with the use of the well-known ratio [7,8]

$$A_k = \sqrt{u_k^2 + \left(u_k^*\right)^2} \quad , \tag{5}$$

where $u_k = u(t_k)$ - is the sampled input signal, u_k^* - the Hilbert conjugate signal of u_k . u_k^* is obtained from S_n^* by means of inverse FFT

$$u_k^* = IFFT(S_n) \ . \tag{6}$$

In the simplest simulation technique, A_k can be used in order to obtain the detector output signal

$$u_{out,k} = k_d \cdot A_k \tag{7}$$

where k_d - is the detector transfer factor. This approach works quite well in some practical cases. But, as a detailed consideration shows, there are two constrains for this approach:

(1) the bandwidth of the input signal must be smaller than the cut-off frequency of a low-pass filter (RC-circuit) at the detector output,

$$Df_{in} < F_{cut} \ , \tag{8}$$

(in some cases this constrain can be relaxed, see below), and

(2) the input signal must be large enough so that the detector operates in the large signal mode,

$$A_k > A_{min} \quad , \tag{9}$$

where A_{min} - is a threshold level which is determined by the volt-ampere characteristic of the nonlinear device used in the detector (for a diode detector, $A_{min} \approx 0.2 \div 0.7$V).

The second constrain is rather obvious - only in the large signal mode the output signal in an AM detector is proportional to the input signal amplitude (k_d is a constant). In the small signal mode, k_d is a function of A_k and the spectral content of the output signal is much richer than that of A_k . This technique doesn't "feel" the difference between the small and large signals modes since the Hilbert transform is linear relative to the input signal amplitude

$$u_k^*(c \cdot u_k) = c \cdot u_k^*(u_k) \,, \tag{10}$$

where c - is a constant, and, correspondingly, $A_k(c \cdot u_k) = c \cdot A_k(u_k)$. By this reason, this technique cannot predict harmonics of the modulating signal, which always are present at the detector output (fortunately, for most of practical detectors, they level are rather low and may be disregarded in EMC problems).

To discus the first constrain in detail, let us consider the detector shown in Fig. 2.

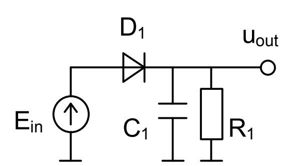

Fig. 2. AM diode detector

If the bandwidth of the input signal E_{in} (which contains interference as well as a required signal) is smaller than the cut-off frequency of the RC-circuit at the detector output, the output signal u_{out} repeats the amplitude of the input signal at the same moment of time - the detector is said to be memoryless. If not, the output signal doesn't repeat the amplitude of the input signal and depends on its levels at the preceding moments of time - the detector is said to have memory. To determine the boundary between these two modes, let's consider the input signal shown in Fig.3. This signal consists

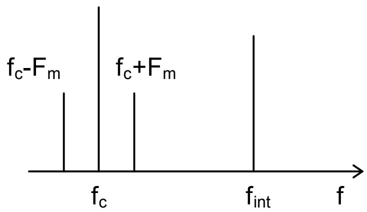

Fig. 3. Spectrum of the input signal. f_c - carrier frequency, F_m - modulating frequency, f_{int} - interference frequency.

of the required AM signal and the interference signal. The required signal bandwidth is always smaller than the cut-off frequency (the design constrain). The interaction between the interference and the required signal will result in beat. So, we must consider the difference between the interference frequency and the required signal frequencies (the beat frequencies). Taking into account that the side frequencies levels are, as a rule, smaller than that of the carrier and for the sake of simplicity, we shall consider further the beat between the carrier and the interference (beat between the side frequencies and the interference can be considered in a similar way). Then the signal bandwidth is equal to the beat frequency

$$Df_{in} = f_{beat} = f_{in} - f_c \quad , \tag{11}$$

where f_{beat} - is the beat frequency. If condition (8) is true, then the output signal repeats the input signal amplitude and the technique works quite well. Otherwise, the output signal doesn't repeat the input signal amplitude because capacitor C_1 has not managed to discharge with the beat frequency: the rate of the capacitor discharge v_C is smaller than the rate of the change of the beat signal amplitude v_{beat} . Let's assume for simplicity that $u_{int} < u_c$, where u_{int} - interference level, u_c - required signal level without modulation (the opposite case can be considered in a similar manner). Then, taking into account the exponential law for the capacitor discharge, we obtain the following assessment for v_C:

$$v_C \approx \frac{u_c}{t} \quad , \tag{12}$$

where $t = R_1 C_1$ - is the RC-circuit constant. In a similar way we obtain the following assessment for the average value of v_{beat} :

$$v_{beat} \approx \frac{u_{\text{int}}}{4T_{beat}} \quad , \tag{13}$$

where $T_{beat} = 1/f_{beat} = 1/Df_{in}$ - is the beat period, since the beat amplitude equals the interference amplitude. Our simulation technique will predict the correct output signal if $v_C > v_{beat}$, or, using (12) and (13)

$$u_{\text{int}} < \frac{4u_c}{t \cdot f_{beat}} \quad , \tag{14}$$

From this condition we can conclude the following: if the interference level is rather low, then the output interference level (at the beat frequency) is not affected by the RC-circuit even if the beat frequency is larger than the cut-off frequency of the RC-circuit (this conclusion is also confirmed by the PSpice simulation - see below).

Thus, it's absolutely unacceptable to model the AM detector as the series connection of a nonlinear element (or a frequency transformer) and a low-pass filter, as it has been proposed by some authors. Physically it can be explained as follows: the capacitor is discharged through the resistor R_1 and is charged through the direct diode resistance which is much smaller than R_1 , so the discharge time constant and the charge time constant are quite different.

From the practical viewpoint, spectral components of the detector input signal, which lie outside of the IF path bandwidth, will be strongly attenuated by the IF path, so that condition (14) will be fulfilled. If, nevertheless, it is not, it means that these spectral components have very large level at the receiver preselector and the receiver is completely blocked.

If condition (14) is not true, then the output interference signal will be attenuated by the RC-circuit. But the attenuation factor will be smaller then the transfer factor of the RC-circuit. Thus, the optimal decision is to ignore the RC circuit during the simulation at all.

Taking into account all considerations given above, we can present the AM detector simulation scheme as on Fig.4.

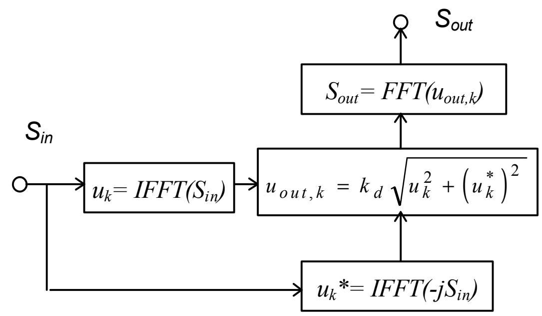

Fig.4. AM-detector simulation scheme. S_{in} - input signal spectrum, S_{out} - output signal spectrum.

We should note that if an additional low-pass filter is connected to the detector output (as on Fig.5) then this filter must be taken into account (since the capacitor C_2 charge and discharge time constants are the same)

$$`S_{out} = S_{out} \cdot K(f) \ , \tag{15}$$

where $`S_{out}$ - spectrum at the filter output, $K(f)$ - complex transfer factor of the filter. The scheme on Fig. 4 must be corrected too.

The technique proposed can be used to simulate an AM detector in the small-signal mode. In this case, k_d must be considered as a function of A_k . An appropriate approximation to this function can be found, for instance, in [9]. Harmonics of the modulating signal can also be predicted in this way.

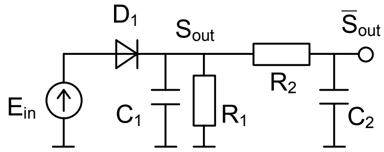

Fig. 5. AM diode detector with a low-pass filter

Let’s consider determination of the sampling rate, the sample frequency interval and the number of samples. It’s similar to the determination of these quantities for the discrete technique [3,6]. The maximum sample frequency $f_{n,max}$ must be higher than the highest input spectrum frequency $f_{in,max}$ with some margin,

$$f_{n,max} = k \cdot f_{in,max} \tag{16}$$

where k - is a margin factor (k=2¼10). The sample frequency interval Df_n must be lower with some margin than the lowest input beat (or modulating) frequency $f_{beat,min}$ which should be modeled,

$$Df_n = k_1 \cdot f_{beat,min} \tag{17}$$

where k_1 - is a margin factor (k_1=0.1 ... 0.5). Using (16) and (17), we find the number of samples

$$N = 2\frac{f_{n,\max}}{f_{beat,\min}} \tag{18}$$

We must round off this number to a power of two (in order to use FFT)

$$\overline{N} = 2^m, \quad m = \left[\log_2 N\right] + 1 \tag{19}$$

where [*] - is the whole part. Further we recalculate Df_n for constant $f_{n,max}$ (or $f_{n,max}$ for constant Df_n) using equations (16)-(18).

3. SIMULATION TECHNIQUE VALIDATION

In order to validate the technique, an extensive circuit-level simulation of the diode AM-detector given on Fig.2 (the cut-off frequency F_{cut}=30kHz) has been carried out by means of well-known simulation tool PSpice [10]. The input signal used in the simulation is as that on Fig.3. Some results of this simulation and the comparison with the technique proposed are presented on Fig. 6 and 7. For Fig.6, the interference frequency is within the required signal bandwidth (parameters of the input signal: f_c=1MHz, F_m=10kHz, modulation index m=0.3, u_c=3V, u_{int}=3V, f_{int}=1015kHz). As it can be seen from this figure, the agreement between our technique and PSpice predictions is quite well for levels not smaller than -20 ...-30dB relative to the maximum. All effects known from the theory (signal compression, beat generation at the frequencies $f_{int} - f_c$, $f_{int} - f_c - F_m$, $f_{int} - f_c + F_m$ etc.) are predicted quite well. We should note that the technique proposed works more than ten times faster than PSpice.

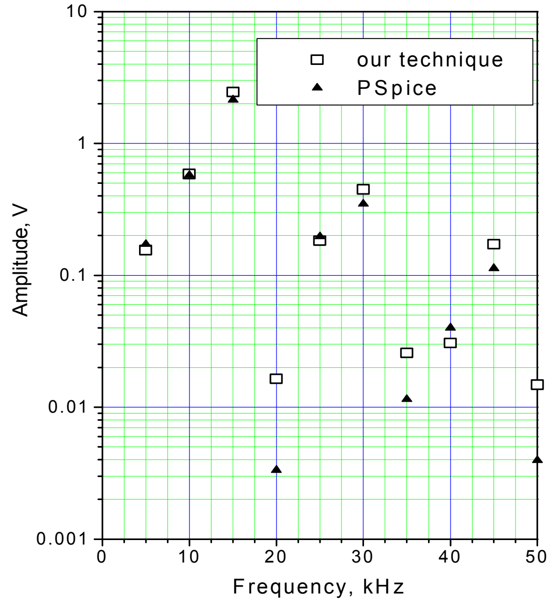

Fig.6 Spectrum at the detector output (input signal - as shown on Fig.3). The interference lies within the required bandwidth.

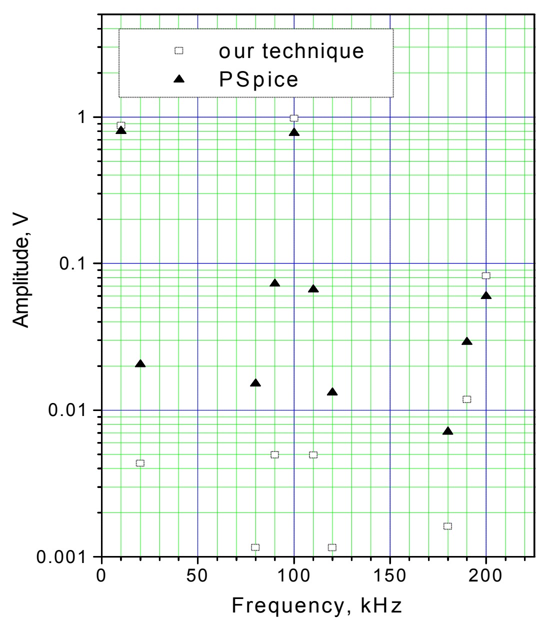

Fig.7 Spectrum at the detector output (input signal - as shown on Fig.3). The interference lies outside the required bandwidth.

For Fig.7, the interference frequency lies outside of the required signal bandwidth (parameters of the input signal: f_c=1MHz, F_m=10kHz, m=0.3, u_c=3V, u_{int}=1V,

f_{int}=1100kHz). As it can be seen from this figure, the largest beat level at the frequency $f_{int} - f_c = 100kHz$ is predicted quite well. It proves our conclusion that the RC-circuit must be discarded during the simulation (for the present case, $(f_{int} - f_c)/f_{cut} » 3$ so if the RC-circuit had operated as an usual low-pass filter, three-fold attenuation would have been expected for this beat frequency, which is not observed in reality). To predict small spectral components more accurately, it's necessary to use an appropriate approximation for $k_d(A_k)$ instead of a constant.

4. FM DETECTOR SIMULATION

A similar approach can be used in order to simulate an FM or PM detector. Using the Hilbert Transform, we find instant angular frequency of the detector input signal [8]

$$w_k = \frac{\overset{*}{u}_k \cdot u_{k-1} - \overset{*}{u}_{k-1} \cdot u_k}{\Delta t \cdot A_k^2} \quad , \qquad (20)$$

where Dt - is the time sample interval. Output signal of an FM detector is proportional to the difference between the instant frequency and the detector resonant frequency w_0

$$u_{out,k} \approx k_d \left(w_k - w_0\right) \qquad (21)$$

This equation is valid for the linear part of the detector input-output characteristic when

$$\left|w_k - w_0\right| \le \Delta w \quad , \qquad (22)$$

where Dw - is the linear part width, and for a sufficiently large input signal when its amplitude is constant due to the limiter which is connected in front of the detector,

$$A_{\lim,in} \ge A_{th,in} \quad , \qquad (23)$$

where $A_{lim,in}$ - is a signal amplitude at the limiter input, $A_{th,in}$ - its threshold level (the saturation level). In other cases, this equation should be generalized to take into account the nonlinearity of the detector characteristic and its dependence on the input signal amplitude

$$k_d = k_d \left(w_k - w_0, A_k\right) \ . \qquad (24)$$

Appropriate approximations for the dependence of k_d on w_k-w_0 can be found in [9]. The dependence of k_d on A_k can be approximated by

$$k_d » c{\cdot}A_k,\ c \text{ - } constant \qquad (25)$$

for an FM detector with tuned-off circuits or similar, and by

$$k_d » c{\cdot}A_k^2,\ c \text{ - } constant \qquad (26)$$

for an FM detector with a multiplier.

As practical experience shows, this simulation technique predicts the required signal compression and the threshold effect quite well. Predicted interference levels are smaller than in reality since the nonlinearity of the detector amplitude characteristic is not taken into account.

5. CONCLUSION

We can conclude that the technique proposed predicts output spectral components which are not smaller than -20dB relative to the maximum level quite well. The feasible improvements of the technique, which have been discussed above can increase the analysis dynamic range.

Further reduction in the computational time can be achieved by decreasing the sampling rate (i.e. computing the output signal samples $u_{out,k}$ not for every k , but only for some k) at the output due to the fact that the output spectrum is a baseband one

An PM detector can be simulated in a way similar to the FM detector simulation.

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