Worst Case Model for Fast Analysis of Intermodulation Interference in Radio Receiver

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***Abstract*—A behavioral model for analysis of intermodulation interference that may arise in a radio receiver operating in complicated electromagnetic environment is developed. The model makes it possible to predict the levels of unequal-power multi-signal intermodulation interference (generated by involving the receiver input signals and the local oscillator signal, including its main tone and harmonics) on the basis of information about the levels of receiver susceptibility to mixer-generated spurious responses and to two-signal intermodulation of different kinds and orders. The model combines the worst case nature of intermodulation level prediction (which is a distinctive feature of conventional methods for analysis of intermodulation) with the high speed of computations (which is implemented by the use of the discrete nonlinear analysis technology). The experimental analysis of intermodulation interference in the AOR AR5000 receiver is performed in order to validate the developed model; the measurements were implemented with the use of the double-frequency test technique, which makes it possible to detect all kinds of two-signal intermodulation interference arising in a receiver.**

Keywords—receivers; parameter extraction; systems engineering and theory; radio interference

I. INTRODUCTION

In order to estimate the ability of interference-free operation of a radio receiver in complicated electromagnetic environment (EME), it is necessary to have information about the selectivity and nonlinearity of input stages of its radiofrequency path [1], [2].

Nevertheless, the lack of knowledge about the selectivity and nonlinearity of the receiver input stages is a commonly encountered problem in practice of electromagnetic compatibility (EMC) analysis. The parameters of the input selectivity and nonlinearity may not be provided in the receiver specification, or they may be provided with little detail which is not enough for estimation of interference levels (e.g., the intermodulation susceptibility may be specified only for third-order products and only for one value of the nearest unwanted signal frequency offset relative to the receiver tuning frequency). Moreover, the information about the receiver implementation (block diagram, circuit diagram, drawings of printed circuit boards, values of intermediate frequencies, etc.) is often absent, and disassembling of the receiver is frequently not possible (e.g., if the receiver is produced a solid unit) or not allowed.

In such situations, there is no chance to apply popular software for simulation of the receiver on the basis of its circuit diagram or block diagram. Therefore, one has to use approximate analytical models [1], [2] which may be inadequate (strictly speaking, such models should be applied only after experimental check of their validity for the particular receiver under consideration). Another way is to measure the external characteristics of input selectivity and nonlinearity, and then to extract the receiver model from the measured data.

In order to detect linear and nonlinear effects arising in a receiver under the influence of one predominant unwanted signal (such effects include spurious responses, desensitization, cross-modulation, reciprocal mixing, limited effectiveness of shielding and filtering), it is usually sufficient to apply the conventional linear model defined in the form of the receiver susceptibility characteristic [2], although detailed models have been developed for analyzing each of these effects individually [2]. A worst case linear model in the form of the receiver susceptibility characteristic can be obtained as a boundary line (lower envelope) of the results of two-signal measurements – ref. CS108 and CS114 procedures in MIL-STD-449D, CS104 and CS105 in MIL-STD-461G, as well as Fig. 2 in Section VI.

The greatest difficulties arise when applying the conventional methods [1], [2] to the detection of intermodulation (IM) interference, since the computational complexity of these methods (based on sequential frequency-domain calculation of the amplitudes for all or part of the IM components [3]) increases sharply with an increase in the number of spectrum components contained in the model of EME at the receiver input and with an increase in the order of the analyzed IM effects.

The technology of discrete nonlinear analysis (DNA) [4], [5], which is based on modeling the receiver nonlinearity in the time domain and using fast Fourier transforms, makes it possible to overcome these difficulties. The absence of a detailed model of the receiver input nonlinearity, which could be extracted from measured external characteristics of the receiver, is a limitation of the DNA technology at present: nonlinearity models used in the framework of DNA and obtained by applying the known techniques [6], [7], [8] do not allow to correctly reproduce the levels of IM products received via spurious-response channels (see Section III).

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The objective of this paper is to develop a detailed computationally efficient worst case model of a radio receiver for analysis of receiver-generated intermodulation interference; the model could be extracted from the experimental results in the absence of information about the internal implementation of the receiver.

II. STRUCTURE OF RECEIVER MODEL

In order to analyze intermodulation interference arising in a receiver, it is useful to represent the receiver as a behavioral model “Filter–Nonlinearity–Filter” (FNF) [6], [9] (Fig. 1): the first linear filter simulates the frequency selectivity of the receiver input circuit, the nonlinearity describes the nonlinear properties of the radio frequency (RF) amplifier and of the first mixer, and the second linear filter is a model of the main selectivity implemented, as a rule, in the intermediate frequency (IF) path or in digital domain. The advantage of the FNF model is in the following: one is able to synthesize this model (i.e., to extract the model parameters from measured external characteristics of the receiver) even for a receiver the internal structure of which is unknown and the IF output of which is inaccessible (e.g., for a receiver implemented as a single integrated circuit) [6].

If the information about the internal implementation of the receiver in not available, it is convenient to normalize the amplitude-frequency characteristics (AFCs) of the input and output filters ($H_{1U}(f)$ and $H_{2U}(f)$, correspondingly, ref. Fig. 1) to their values at the frequency of the desired signal in the filter insertion point, i.e., to assume that $H_{1U}(f_0)=1$ and $H_{2U}(f_{IF})=1$, where f_0 is the receiver tuning frequency and f_{IF} is the first intermediate frequency (for model (1), we must assume $H_{2U}(f_0)=1$ [6], [9]). It is also useful to set the small-signal gain G_0 of the nonlinearity equal to the receiver transfer gain measured from the RF input to the IF output; but if the IF output is inaccessible, then it is reasonable (in addition to the normalization of the AFCs) to assume that $G_0=1$, i.e., to operate with signals referred to the receiver input.

The techniques for extracting the AFC $H_{1U}(f)$ of the receiver input circuit (ref. Fig. 1) from measured external characteristics of the receiver are considered in [9]. The model $H_{2U}(f)$ of the receiver’s main selectivity can be defined theoretically (e.g., the AFC of ideal band-pass filter [1, eq.(7)] is used in [1]) or extracted from measured data [2], [6].

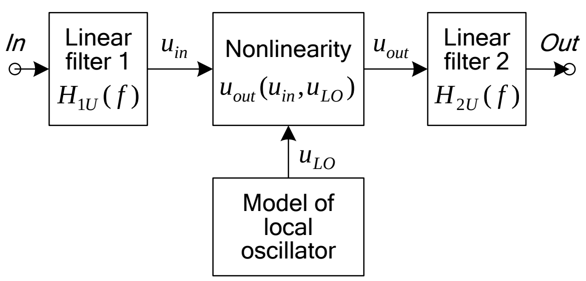

Fig. 1. Radio receiver model intended for behavioral analysis of intermodulation interference.

III. MODEL OF RECEIVER FRONT-END NONLINEARITY

In order to use the advantages of the DNA technology, the model $u_{out}(u_{in},u_{LO})$ of the receiver input nonlinearity must be formulated in the time domain and represented as a function that limits the bandwidth expansion observed when a signal passes through the nonlinearity [5], [7]. Here, u_{in}, u_{LO}, and u_{out} are instantaneous values of voltages at the RF input (u_{in}), at the local oscillator (LO) input (u_{LO}), and at the output (u_{out}) of the nonlinearity model.

The model in the form of a power polynomial [6], [7]

$$u_{out}(u_{in},u_{LO})=u_{out}(u_{in})=\sum_{m=0}^{M} a_m\cdot u_{in}^m \tag{1}$$

(where a_m are the parameters of the model, M is the degree of the model) makes it possible to correctly describe the direct-gain receiver only, because this model do not reproduce the IM products received via spurious-response channels. The need of account for such products is proved by experiments [10].

The model defined as a cascade connection of a summator and the power polynomial (1)

$$u_{out}(u_{in},u_{LO})=\sum_{m=0}^{M} a_m\cdot(u_{in}+u_{LO})^m \tag{2}$$

is free from the above drawback; the technique for extracting this model from measured dynamic ranges for spurious responses is known [8]. The disadvantage of the model is the limited detail of describing the receiver input nonlinearity: in the small-signal approximation important for practice, the amplitudes of all nonlinear products with a fixed order m are strictly connected to each other because they are determined by the same polynomial coefficient a_m. Therefore, model (2) well describes the simplest types of frequency mixers, but it is of limited use in the analysis of a radio receiver, especially if nonlinear products with significant levels are generated in the RF amplifier of the receiver [8], [11].

The model defined as an expansion of the characteristic $u_{out}(u_{in},u_{LO})$ in a Taylor-Fourier series (TFSM) [11], [12]

$$u_{out}(u_{in},u_{LO})=u_{out}(u_{in},t)=\sum_{p=0}^{P}\cos(p\cdot 2\pi f_{LO}t)\sum_{m=0}^{M}a_{p,m}\cdot u_{in}^m \tag{3}$$

makes it possible to increase the detail of the description of the receiver input nonlinearity. Here, $a_{p,m}$ are the coefficients of the model, f_{LO} is the LO frequency, M is the order of the model with respect to the RF input, and P is the order of the model with respect to the fundamental tone of the LO. Note that (3) turns into (1) in case of $P=0$.

The rest of the paper is focused on model (3).

IV. Analysis of Intermodulation Products at Receiver Model Output

Suppose that the RF input of the nonlinearity model (ref. Fig. 1) is fed by a sum of L continuous-wave (CW) signals

$$u_{in}(t)=\sum_{n=1}^{L} U_{in\,n}\cdot\cos(2\pi f_n t+\varphi_n)\,, \tag{4}$$

where $U_{in\,n}, f_n, \varphi_n$ are the amplitude, frequency, and initial phase of n-th component of the input signal, correspondingly.

The inner sum in (3) coincides with (1) with the only peculiarity that its coefficients depend on p. Therefore, let us consider a K-signal IM product $(K\le L)$ of order N at the output of the model (1):

$$\begin{gathered}u_{out_{(z_1,z_2,\dots,z_K)}}(t)\equiv u_{IM}(t)=U_{IM}\cdot\cos(2\pi f_{IM}\,t+\varphi_{IM}),\\ f_{IM}=\sum_{i=1}^{K} z_i f_i\,,\qquad \varphi_{IM}=\sum_{i=1}^{K} z_i\varphi_i\,,\qquad N=\sum_{i=1}^{K}|z_i|\,,\end{gathered} \tag{5}$$

where $(z_1,z_2,\dots,z_K)$ is a vector of integer coefficients describing the type of the IM product; $U_{IM}, f_{IM}, \varphi_{IM}$ are the amplitude, frequency, and initial phase of the product (5), correspondingly.

Substituting (5) instead of the inner sum in (3) and taking into account the dependence on p, we find the IM components formed at the output of the TFSM (3) by the product (5):

$$\begin{aligned}&u_{out_{TFS,(z_1,z_2,\dots,z_K)}}(t)=\\&=\sum_{p=0}^{P}\cos(p\cdot 2\pi f_{LO}t)\cdot U_{IM\,p}\cdot\cos(2\pi f_{IM}\,t+\varphi_{IM}).\end{aligned} \tag{6}$$

Considering one of the summands in (6), which corresponds to a fixed value of p, and applying the cosine product-to-sum identity, we obtain

$$\begin{aligned}&u_{out_{TFS,(z_1,z_2,\dots,z_K),\,p}}(t)=\\&=(U_{IM\,p}/2)\cdot\{\cos[2\pi(f_{IM}+p\,f_{LO})t+\varphi_{IM}]+\\&\qquad+\cos[2\pi(f_{IM}-p\,f_{LO})t+\varphi_{IM}]\}=\\&=\begin{cases}u_{out_{TFS,(z_1,z_2,\dots,z_K,\,p)}}(t), & p=0;\\ u_{out_{TFS,(z_1,z_2,\dots,z_K,\,p)}}(t)+u_{out_{TFS,(z_1,z_2,\dots,z_K,\,-p)}}(t), & p>0,\end{cases}\end{aligned} \tag{7}$$

$$\begin{aligned}&u_{out_{TFS,(z_1,z_2,\dots,z_K,z_{LO})}}(t)\equiv\\&\equiv\gamma_{LO}U_{IM\,p}\cos[2\pi(f_{IM}+z_{LO}\,f_{LO})t+\varphi_{IM}],\end{aligned} \tag{8}$$

$$|z_{LO}|=p\,,\qquad \gamma_{LO}\equiv\begin{cases}1, & z_{LO}=0;\\ 1/2, & z_{LO}=\pm1,\pm2,\pm3,\dots\,,\end{cases} \tag{9}$$

where $(z_1,z_2,\dots,z_K,z_{LO})$ is a vector of integer coefficients describing the type of the IM product at the TFSM output.

Thus, if the RF input of the TFSM (3) is fed by the sum (4) of L CW components, then the IM product (8) formed at the output of the TFSM by the p-th harmonic of the LO fundamental tone and the K-signal IM (5) between the RF input signal components has the order (p,N), total order $p+N$, initial phase $\varphi_{IM_{TFS}}=\varphi_{IM}$, frequency $f_{IM_{TFS}}$, and amplitude $U_{IM_{TFS}}$:

$$f_{IM_{TFS}}=f_{IM}+z_{LO}\,f_{LO}=\left(\sum_{i=1}^{K} z_i\,f_i\right)+z_{LO}\,f_{LO}\,, \tag{10}$$

$$U_{IM_{TFS}}=\gamma_{LO}U_{IM\,p}\,,\qquad p=|z_{LO}|\,, \tag{11}$$

where $U_{IM\,p}$ is the amplitude of the IM product (5) at the output of the polynomial model (1) the values of the coefficients a_m of which coincide with the values of the TFSM coefficients $a_{p,m}$ in (3) at a fixed value of p.

Expressions (10) and (11) describe the relationship between the parameters of the IM products at the output of the TFSM (3) and at the output of the polynomial model (1). This relationship makes it possible to apply well-known formulas [13], [14] to the problem of calculating the amplitudes of the IM products at the TFSM output. In particular, by substituting the approximate small-signal formula for the amplitude $U_{IM\,p}$ [13], [9] into (11), we obtain

$$\begin{aligned}&U_{IM_{TFS}}\cong\gamma_{LO}\,\gamma_{IM}\cdot a_{p,N}\cdot U_{in\,1}^{\;m_1}\cdot U_{in\,2}^{\;m_2}\cdot\ldots\cdot U_{in\,K}^{\;m_K}=\\&\qquad=\gamma_{LO}\,\gamma_{IM}\cdot a_{p,N}\cdot U_{in\,1}^{\;N}\cdot\eta_{2,1}^{m_2}\cdot\eta_{3,1}^{m_3}\cdot\ldots\cdot\eta_{K,1}^{m_K},\\&\gamma_{IM}=N!/(m_1!m_2!\ldots m_K!\cdot 2^{N-1}),\\&m_i\equiv|z_i|,\qquad \eta_{i,1}=U_{in\,i}/U_{in\,1},\qquad i=1..K.\end{aligned} \tag{12}$$

Now suppose that the input of the FNF model (ref. Fig. 1) is fed by the signal (4). Then, at the output of the FNF model, the frequency (10) of the IM product (8) is not changed and the amplitude, in compliance with (12), takes the form

$$\begin{aligned}&U_{IM_{FNF}}\cong\gamma_F\,\gamma_\eta\,\gamma_{LO}\gamma_{IM}\,a_{p,N}\,U_{in\,1}^{\;N},\\&\gamma_F=H_{2U}(f_{IM_{TFS}})\cdot\prod_{i=1}^{K}\{H_{1U}(f_i)\}^{m_i},\qquad \gamma_\eta=\prod_{i=1}^{K}\eta_{i,1}^{\;m_i}.\end{aligned} \tag{13}$$

If the frequency (10) of the IM product (8) falls within the passband of the output filter, then this product passes to the FNF model output without significant attenuation (ref. Fig. 1). In a special case when $f_{IM_{TFS}} = f_{IF}$, formula (10) describes the center frequencies of the intermodulation reception channels and it is called the channeling equation [15], [16].

V. Extraction of Worst Case Model of Receiver Front-End Nonlinearity from Measured Data

Let us consider the problem of the TFSM synthesis (i.e., calculating the values of the coefficients $a_{p,m}$ and frequency f_{LO} in (3)) based on the results of measuring the external characteristics of the receiver. As initial data for the synthesis, it is advisable to use characteristics that can be measured for any receiver, including a receiver that does not have the IF output. Above all, the sensitivity of the receiver and its interference-free dynamic ranges (DRs) for spurious responses (SRs) and for intermodulation responses (IRs) are such characteristics.

The LO frequency f_{LO} (if it is not given in the receiver's documentation) can be obtained experimentally, e.g., by analyzing the leakage spectrum at the antenna input of the receiver – ref. CS112 procedure in MIL-STD-449D.

A technique for extracting the coefficients $a_{p,m}$ of the TFSM (3) of a mixer from measured external characteristics (namely, from the matrix of the mixer's susceptibility to SRs) is proposed in [11 – eqs. (17)…(21)], [12 – Appendix]. However, the applicability of this technique to the synthesis of a receiver model is limited for the following reason. Part of the SRs that can be analyzed experimentally for a mixer cannot be detected when the same mixer is used as part of a receiver, since there are filters in the receiver structure (see explanations to Fig. 1). Nevertheless, some types of IM products received via such undetectable SR channels are still observed for the receiver (see Section VI). Therefore, it is reasonable to extract the receiver nonlinearity model from information not only about the SRs, but also about the IRs.

Let us express the coefficients $a_{p,m}$ of the TFSM (3) in terms of DRs for SRs (SDRs) and DRs for IRs (IDRs). To do this, we require that the amplitude-to-amplitude characteristic (13) $U_{IM_{FNF}}(U_{in_1})$ pass through a point $(U_{S,in_1}; U_{S,out})$ the coordinates of which are the receiver IM susceptibility levels referred to the RF input and IF output. Substituting equalities $U_{in_1} = U_{S,in_1} = U_{in,\min} D_U$ and $U_{IM_{FNF}} = U_{S,out} = G_0 U_{in,\min} / SIR_{out}$ into (13), we obtain

$$a_{p,N} \cong G_0 / (\gamma_F\, \gamma_\eta\, \gamma_{LO}\, \gamma_{IM}\, U_{in,\min}^{N-1}\, D_U^N\, SIR_{out}), \qquad (14)$$

where G_0 is the small-signal gain of the receiver (from the RF input to the IF output), V/V; $U_{in,\min}$ is the receiver sensitivity, V; D_U is the IDR (as follows from (10), in case of $K=1$, an IR becomes the SR and an IDR turns into the SDR), V/V; SIR_{out} is the signal-to-interference ratio at the IF output of the receiver, V/V.

In particular, for the desired reception channel, we have: $K=1$, $N=1$, $p \geq 1$, $D_U=1$, $SIR_{out}=1$; therefore, formula (14) will take the form $a_{p_D,1} = 2G_0 / \{H_{1U}(f_0) \cdot H_{2U}(f_{IF})\}$, where p_D is the number of the LO harmonic at which the frequency conversion is performed ($p_D > 0$; usually, $p_D = 1$).

The following must be done to obtain the worst case TFSM that prevents underestimation of the levels of IM products:

1) Calculate $a_{p,N}$ according to (14) on the basis of measured values of D_U for the following reception channels: a) for all SRs described by the linear model (ref. section I) and b) for all two-signal IRs observed at maximum input-signal levels for which an interference is not detected by the linear model. Note that it is convenient to use the double-frequency test technology [15], [16] for quick detection and recognition (i.e., obtaining the values of parameters z_1, z_2, z_{LO}) of all SRs and two-signal IRs of a receiver.

2) If the values of the same coefficient $a_{p,N}$ are calculated on the basis of several DRs (in particular, if each of these DRs is measured for its own SR or IR – see channels of the same order $(p,N) = (|z_{LO}|, |z_1| + |z_2|)$ in Table I, e.g., channels No.7 and No.18), then choose the calculated value of $a_{p,N}$ that is the largest in absolute value.

3) Set the same sign for all coefficients $a_{p,N}$. As a result, the shape of the amplitude-to-amplitude characteristics (AACs) (11) in the large-signal region becomes inadequate (anticompression is observed instead of compression), but the AACs of all types and orders do not contain notches. The notches could cause the missing (erroneous undetection) of the interference by the model [7].

VI. Validation of Receiver Model

In order to validate the developed model, the experimental analysis of the AOR AR5000 receiver tuned at 1 GHz is performed. This situation is near to the worst case, because the path containing an octave input filter is involved in the receiver (the passband of the filter is 830…1630 MHz at the level of –3 dB [17], [9]); as a result, a lot of SRs and IRs are observed. The following bandwidth of the receiver's main selectivity filter is chosen: 260 kHz at the level of –3 dB [17].

The measurements are performed in conducted way (at the receiver antenna terminal) by the instrumentality of the automated double-frequency test system [16]. To decrease the measurement time, the receiver response is analyzed at the output of the second IF path (10.7 MHz).

The analysis is performed in the following order.

1. The initial data for model extraction are obtained:

1.1. The values of the first LO frequency $f_{LO} = 1622.4$ MHz and the first IF $f_{IF} = 622.4$ MHz are assessed by analysis of the receiver's documentation [17] and the leakage spectrum at the receiver antenna terminal (small variations of the receiver tuning frequency are made).

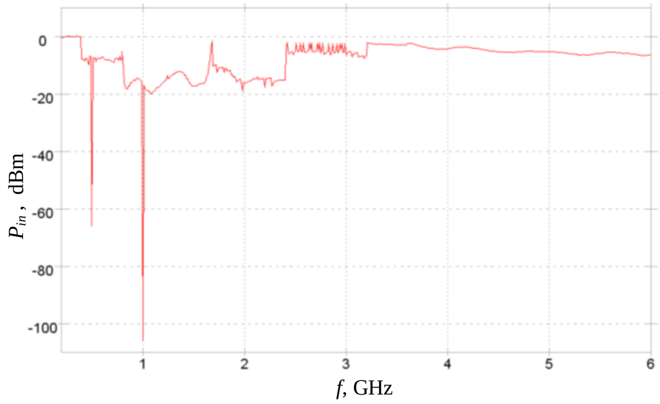

Fig. 2. Measured characteristic of the AOR AR5000 receiver susceptibility to continuous-wave (CW) unwanted signal. Parameters of the measurements: the desired signal frequency of 1 GHz coincides with the receiver tuning frequency; the desired signal level of -101 dBm is equal to the sensitivity of the receiver; the measurement step $\Delta f = 10$ MHz; the type of detector in the spectrum analyzer is RMS average; the interference criterion is the departure of the output signal from the standard response (-72 dBm) by at least 1 dB. Notes: the spurious-response (SR) reception channels are not observed in the plot because the measurement step in frequency is chosen to be much more than the receiver bandwidth (in order to decrease the measurement time); the response at 500 MHz is caused by formation of the second harmonic in the measuring generator.

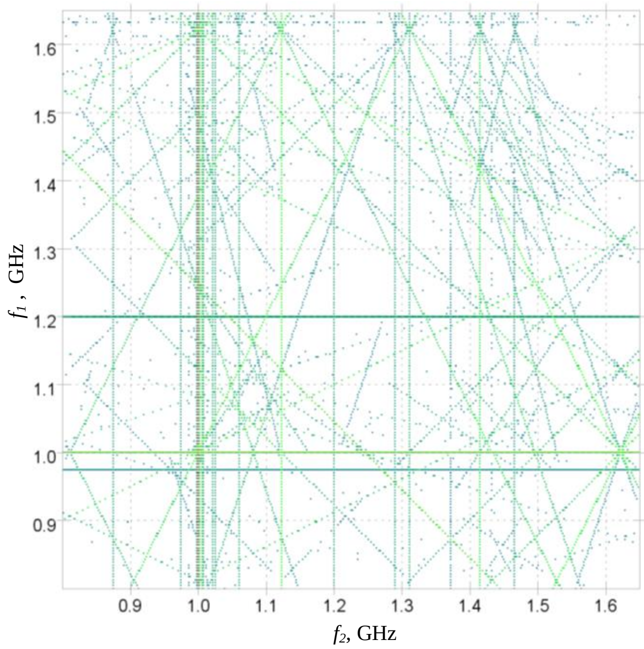

Fig. 3. The double-frequency diagram (DFD) of the AOR AR5000 receiver, i.e., a color map plot of the measured double-frequency characteristic (DFC). The minimum level of the receiver response to display in the DFD is set equal to -70 dBm (which is 3 dB above the noise level). Parameters of the measurements: the level of each of two test signals at the receiver input $P_1 = P_2 = -30$ dBm; the measurement step $\Delta f_1 = 4.25$ MHz, $\Delta f_2 = 200$ kHz; the type of detector in the spectrum analyzer is peak. Note: the lines observed in the DFD are images of all reception channels that can cause interference for the the receiver – horizontal and vertical lines are images of the receiver's desired and spurious responses, inclined lines are images of intermodulation responses [2 – p.4.41], [15].

TABLE I. SOME OF RESPONSES OF AR5000 AT $P_1 = P_2 = -30$ DBM

No.	Node ID	z_1	z_2	z_{LO}	(p,N)	f_2, MHz	D_U, dB
1	Nif	0	1	0	(0,1)	622.4	75.1
2	Nif	1	-1	0	(0,2)	1010.0	52.7
3	Nif	-1	2	0	(0,3)	1120.0	62.1
4	Nif	2	-2	0	(0,4)	1060.0	68.6
5	Ndr	0	-1	1	(1,1)	1000.0	0
6	Ndr	1	-1	1	(1,2)	1840.0	59.4
7	Ndr	1	-2	1	(1,3)	1030.0	59.0
8	Ndr	-2	2	1	(1,4)	1510.0	66.0
9	Ndr	2	-3	1	(1,5)	1040.0	69.2
10	Ndr	3	-4	1	(1,7)	1050.0	71.4
11	Nim	0	1	-1	(1,1)	2244.8	66.1
12	Nim	-1	2	-1	(1,3)	1540.0	60.8
13	Nim	-2	3	-1	(1,5)	1460.0	68.9
14	N06	0	2	-1	(1,2)	1122.4	55.1
15	N06	1	1	-1	(1,2)	1040.0	50.3
16	N06	-1	3	-1	(1,4)	1090.0	67.5
17	N06	-2	4	-1	(1,6)	1090.0	71.5
18	N19	1	2	-1	(1,3)	480.0	72.7
19	N18	0	-1	2	(2,1)	2622.4	63.2
20	N04	0	-2	2	(2,2)	1311.2	67.1
21	N04	-1	-1	2	(2,2)	1020.0	55.4
22	N04	1	-3	2	(2,4)	1410.0	69.0
50	N13	-3	-3	5	(5,6)	1460.0	69.1
52	N14	3	3	-5	(5,6)	1450.0	69.4
53	N15	-3	-3	6	(6,6)	1490.0	70.9

1.2. The receiver susceptibility characteristic is measured in a wide range of frequencies. The measurement results (Fig. 2) enable us to give the following estimations of the susceptibility level outside the bands of SR reception channels: –20 dBm in the range of 800…2420 MHz and –10 dBm outside this range.

1.3. To detect all SRs and all two-signal IRs of the receiver, a double-frequency characteristic (DFC) is measured [16].

When measuring the DFC, the level $P_{ts} \equiv P_1 = P_2$ of each of two test signals at the receiver input must not exceed the susceptibility outside the bands of SR reception channels (ref. item 1.2). However, even if $P_{ts} = -25\,\text{dBm}$, the number of SRs and IRs of the receiver is so large that their separate analysis becomes unreasonable (since the receiver will almost certainly be affected by the interference). Therefore, it is advisable to decrease the level of susceptibility outside the bands of SR reception channels by 9 dB with respect to the values given in item 1.2 (and take this into account when synthesizing the linear model of the receiver), and to measure the DFC at $P_{ts} = -30\,\text{dBm}$ for extracting the receiver's nonlinear model.

In the DFC measured in the frequency range of 500…3000 MHz, 62 reception channels (the desired channel, the SR and IR channels) are detected, most of the channels are observed in the passband of the receiver input filter (Fig. 3).

1.4. The recognition of all reception channels detected in the DFC is performed [16]; as a result, the values of parameters z_1, z_2, z_{LO} are obtained (see Table I). The order of the detected channels is $0 \le p \le 6$, $1 \le N \le 7$.

1.5. The frequency characteristic of the receiver susceptibility (FCRS) for each detected reception channel is measured. The examples of measured FCRS for channels No.7, 9, and 15 from Table I are provided in [9, Figs. 4, 6, 7].

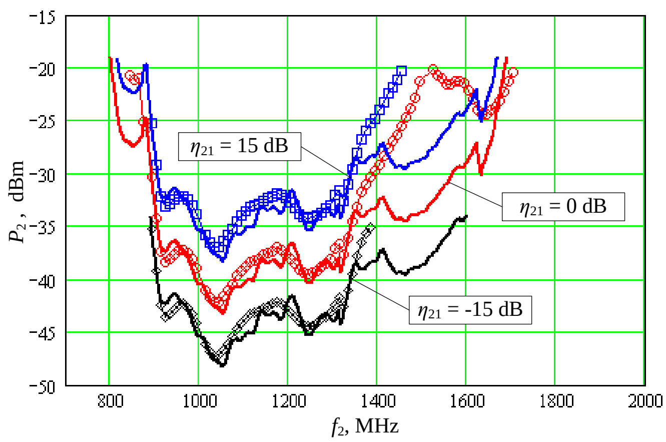

Fig. 4. Frequency characteristics of susceptibility of the AOR AR5000 receiver to two-signal third-order intermodulation (ref. channel No.7 in Table I) for different ratios η_{21} of the levels of unwanted input signals: the results of measurement (thin lines with markers) and computation on the basis of the synthesized model (thick solid lines). The computation is performed by the DNA technology involving formula (3).

1.6. The values of SDRs and IDRs (column D_U in Table I) are obtained by processing of minima of each measured FCRS.

The reception channels of the 2nd order (IR channel No.15 and corresponding SR channel No.14 in Table I) are found to be the most dangerous. Traditionally, these channels are not considered as significant for UHF receivers [10]; therefore, the fact of finding them for the AOR AR5000 proves the importance of experimental analysis of a particular receiver. In expensive high-quality receivers, low-order SRs are eliminated by the use of half-octave input filters and careful selection of f_{LO} and f_{IF} [18, p.135].

2. The receiver's nonlinear model intended for the analysis of IM interference is synthesized (ref. Fig. 1):

2.1. The model of the input circuit AFC $H_{1U}(f)$ is extracted from the measured FCRS to three-signal IM [9].

2.2. The model of the main selectivity path AFC $H_{2U}(f)$ is obtained from the $H_{1U}(f)$ (ref. item 2.1) and measured FCRS to CW unwanted signal (ref. Fig. 2), as described in [6].

2.3. The worst case model of the receiver input nonlinearity is extracted from data of Table I by the technique of Section V.

3. Based on the synthesized nonlinear model of the receiver (ref. Fig. 1), the FCRS for different IR channels (2- and 3-signal, equal- and unequal-signal) are calculated and compared with the measurement results in the way shown in Fig. 4.

VII. Conclusion

The developed behavioral model (ref. Fig. 1) makes it possible to predict the characteristics of the receiver susceptibility to IM of all types and orders (including unequal-signal and many-signal IM) observed in this receiver. This enables us to predict the presence and severity of IM interference in case of the receiver operation in any given EME. The results of experiments (ref. Section VI) prove the operability and the worst case nature of the developed model.

Using the developed model of the receiver (ref. Fig. 1) within the framework of the DNA technology makes it possible to increase the level of detail and speed of the IM interference analysis many times as compared to the use of traditional frequency-domain models and methods.

Only continuous-wave signals are considered in Sections IV, V, and VI in order to simplify the analysis and to decrease the measurement time. Nevertheless, the developed techniques for analysis and synthesis of the receiver model are still valid in case of modulated signals; this is important if the receiver does not have the IF output port or if the standardized characteristics of the receiver must be used for the model extraction (e.g., CS110 in MIL-STD-449D, CS103 in MIL-STD-461G).

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