Simulation of Noise-Power Ratio with the Large-Signal Code CHRISTINE

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***Abstract*—This paper describes simulations of the noise-power ratio (NPR) for a helix traveling wave tube (TWT) performed with the large-signal, one-dimensional (1-D), multifrequency code CHRISTINE. The results obtained with this code are in better agreement with measured values than are the more traditional values calculated by power series. We conclude that NPR simulations with large-signal codes have the potential to shorten the design phase of TWTs by eliminating the need for repeated build-test cycles to meet a required NPR.**

***Index Terms*—Amplifier distortion, communication system nonlinearities, design automation, intermodulation distortion, simulation, traveling wave tubes.**

I. Introduction

THE EXPLOSIVE growth in telecommunications has increased the demand for high data-throughput and better transmission quality in an already congested spectrum, making necessary the use of digital modulation techniques in wireless systems. Amplifiers for digital systems must boost simultaneously signals from multiple channels, and, to assess their performance, two-tone tests do not represent realistic operating conditions. Instead, multitone tests must be used.

Noise-power ratio (NPR) is a multitone test designed to measure linearity and the dynamic range free of intermodulation distortion, and is particularly relevant to multichannel telecommunications systems. In contrast to traditional, analog-oriented tests that use simple sinusoidal stimuli, the evaluation of NPR requires a more complex stimulus that consists of a signal with white Gaussian noise characteristics, from which a portion of the spectrum has been removed. This input signal is applied to the device under test (DUT), and any nonlinearity in the DUT leads to the appearance of spectral components within the spectral notch removed from the input signal. Experimentally, the digital technique is preferred, wherein a brick-wall spectrum consisting of thousands of discrete frequencies with random amplitudes and phases are digitally generated, and the spectrum notch is created by band-reject filtering (see, e.g., [4]). NPR is defined as the ratio of the power of the spectral components in the passband to the power of those in the notch at the output of the DUT.

Traveling-wave tubes (TWTs) are used widely as amplifiers in telecommunications, and NPR is one of the specifications to be met when building a particular device. Because NPR is measured after a device is built, the meeting of a particular NPR specification may require several build-test cycles. This development cycle could be shortened if it were possible to predict a device's response before it is built.

Traditional analytic or semi-analytic methods [1], such as expanding the device's response in a power series in input voltage or power, are inadequate to compute NPR or require experimental measurements to determine the coefficients of the expansion. Therefore, these methods have limited predictive power.

On the other hand, large-signal, multifrequency numerical codes to simulate TWTs, such as CHRISTINE [2], [3], can compute ab initio a device's response and have the potential to simulate NPR measurements. Thus, the design phase of a TWT may be shortened by reducing, or eliminating altogether, the several build-test cycles required to meet a particular NPR specification.

The purpose of this paper is to illustrate the viability of NPR simulations with the large-signal code CHRISTINE. The main obstacle to overcome is the large number of input frequencies used in the experimental measurement. Section II below justifies the use of as few as a dozen input frequencies, thus making the computer simulation with a large-signal code feasible, Section III presents the simulations and their numerical convergence, and in Section IV we compare the simulations with experimental measurements. Our conclusions are summarized in Section V.

II. The Simulation Challenge

The input stimulus for a NPR experiment consists of either noise or several thousands of discrete frequencies; the handling of such input is beyond the capabilities of current multifrequency, large-signal codes. However, here we show that it may be possible to simulate accurately NPR measurements with only 10–100 input tones.

Consider an input stimulus consisting of n_f different carriers with equal amplitude A, such that the total input power, $P_{\rm in}$ is

$$P_{\rm in} = \tfrac{1}{2}\, n_f A^2 \tag{1}$$

i.e.,

$$A \propto n_f^{-1/2} \tag{2}$$

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and let A_M be the amplitude of the M-order ($M = 3,\ 5,\ \cdots$) intermodulation products for a given pair of input carriers. It can be shown [5]—by expanding the DUT's response in a Taylor series, collecting the terms that contribute to the M-order intermodulation product, and using (2)—that the dependence of A_M on the number of carriers n_f when $P_{\rm in}$ is kept constant is given by

$$A_M \propto n_f^{-M/2} \tag{3}$$

where the constant of proportionality, g_M, depends on the gain function of the DUT and scales with M as [6]

$$g_M \propto \left(2^M\, N!L!\right)^{-1} \tag{4}$$

where

$$M = N + L \tag{5}$$

i.e., the intermodulation frequency resulting from a pair[1] of carriers with frequencies f_1 and f_2 is $Nf_1 \pm Lf_2$.

For a total of n_f input carriers there are $\propto n_f^2$ possible pairs of input carriers, and, therefore, the total amplitude of the M-order intermodulation product is

$$A_M \propto n_f^{(4-M)/2} \tag{6}$$

i.e., the total power P_M in the M-order intermodulation is

$$P_M \propto g_M^2 n_f^{4-M}. \tag{7}$$

Summing over all possible M's, and factoring out the term proportional to n_f (for $M = 3$), the total power in all intermodulation products is

$$P_{\rm mod,\,tot} \propto g_3^2 n_f \left(1 + \sum_{k=1}^{\infty} \left(\frac{g_{2k+3}}{g_3}\right)^2 n_f^{-2k}\right) \tag{8}$$

where the running index $2k + 3$ takes the values $M = 5,\ 7,\ 9,\ \cdots$ for $k = 1,\ 2,\ 3,\ \cdots$.

The noise-power ratio is given by

$$\text{NPR} \equiv \frac{P_{\rm out}}{P_{\rm mod,tot}} \tag{9}$$

where $P_{\rm out}$ is the output power in the input frequencies. Writing $P_{\rm out}$ as in (1) and using (8), the NPR will scale with n_f as

$$\text{NPR} \propto \left(1 + \sum_{k=1}^{\infty} \left(\frac{g_{2k+3}}{g_3}\right)^2 n_f^{-2k}\right)^{-1}. \tag{10}$$

Because of the strong dependence of g_M on the intermodulation order, M, (10) suggests that it may be possible to accurately simulate the NPR with $n_f \sim$ 10–100 instead of the many thousands used experimentally.

III. Simulation Approach and Results

Experimentally, NPR is measured by generating a sequence of input spectra with random amplitudes and phases, and averaging the output spectra. Similarly, to calculate the NPR for a given device we run a sequence of simulations with random amplitudes drawn from a Gaussian distribution with zero mean and unit variance and random phases drawn from white noise. The amplitudes so generated are scaled to obtain the total input power, $P_{\rm in}$, under consideration. We shall refer to each of these runs as a realization of the NPR experiment.

To be specific, for a realization with n_f frequencies and a total input power $P_{\rm in}$

$$P_{\rm in} = \tfrac{1}{2}\, n_f \sum_{j=1}^{n_f} A_j^2 \tag{11}$$

the amplitudes A_j are given by

$$A_j = A_0 |p_j| s_j \left(\frac{2P_{\rm in}}{n_f}\right)^{1/2} \tag{12}$$

where p_j is a random number with a probability

$$E[p_j] \propto e^{-p_j^2} \tag{13}$$

and s_j is a shape factor used to carve out a notch in the passband. At the bottom of the notch $s_j = 0$ and $s_j = 1$ outside the notch. Also, to minimize aliasing, we set $s_j = 0$ at the first and last two frequency points in the passband.

The normalization factor A_0 is given by

$$A_0 = \left(\sum_{j=1}^{n_f} s_j^2 |p_j|^2\right)^{-1/2}. \tag{14}$$

We generate the normally distributed random numbers p_j with the MATLAB© [7] routine `randn` and the uniformly distributed random phases with the routine `randu`. For a simulation with N realizations with n_f frequencies we generate $N \times n_f$ random numbers p_j and $N \times n_f$ random phases in one call to `randn` and `randu`, respectively. These numbers are saved to a file from which they are read at the beginning of each realization of the numerical experiment; thus, each numerical experiment can be repeated unambiguously. Fig. 1 is a plot of the distribution of the p_js so generated for a simulation with $n_f = 13$ and 50 realizations.

The simulations were performed using the large-signal, multifrequency helix TWT code CHRISTINE. This code has been validated experimentally [8], [9] through an extensive comparison of simulated responses with experimental measurements of an L-band TWT. All the simulations presented here use the physical parameters of this tube. These parameters and the implementation of simulations with CHRISTINE are described in the Appendix.

We considered a bandwidth of 36 MHz centered at a frequency $f_0 = 1.6498$ GHz, with a notch of width 3.0 MHz at the bottom of the notch. Fig. 2 is a typical output spectrum for a realization of the experiment with $n_f = 25$.

Note that, on output, there is a nonzero signal at the first and last two frequency points due to intermodulation products (recall that the input amplitudes at these points are set to zero on input to minimize aliasing).

[1]Although the derivation presented here is for the intermodulation products of any pair of input carriers, the final result is the same for more complex intermodulation products such as $Nf_1 + Lf_2 + Kf_3$, etc. See [5] and [6].

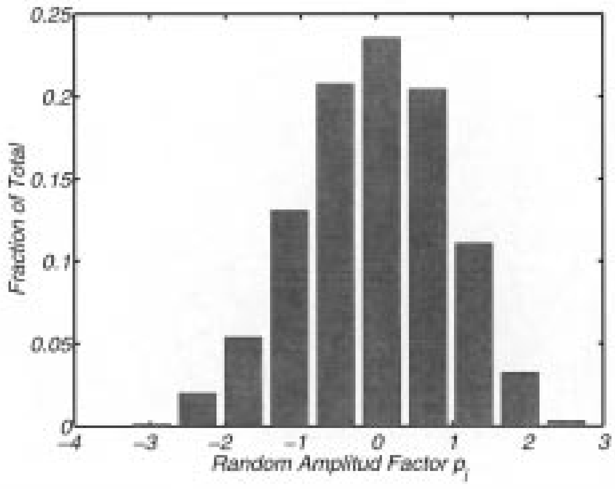

Fig. 1. Distribution of the random numbers p_j generated with the Matlab routine `randn` for a numerical experiment with $n_f = 13$ frequencies and 50 realizations.

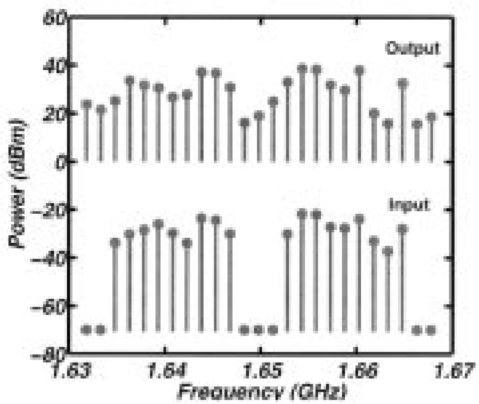

Fig. 2. A typical input and output spectrum for a run with 25 frequencies spanning a 36 MHz range centered at $f_0 = 1.6498$ GHz and total input power -14 dBm (3 dB input back-off from saturation, where saturation $= -11$ dBm). The notch is covered by five frequency points, three at the bottom ($s_j = 0$), spanning 3.0 MHz, and two at the -3 dB points ($s_j = 0.5$).

To validate our approach, we run convergence tests of the numerical experiments with respect to both the number of frequencies n_f and the number of realizations averaged to compute the NPR at a fixed input power of -16 dBm. Our results are presented in Fig. 3.

Note that the curves with $n_f = 49$ and $n_f = 97$ frequencies are essentially identical for averages of more than 20 realizations. This corresponds to the large-n_f limit of (10), and it shows that a larger number of frequencies is not necessary to calculate the NPR. Even $n_f = 13$ yields NPR values within 1 dB of the converged one. These results encouraged us to pursue the comparison with an experimental NPR measurement.

IV. Experimental Validation

The experimental setup was typical to NPR measurements, and can be seen in Fig. 4. The device under test was an L-band Hughes 8537H TWT. Time-averaged, typical input and output spectra are shown in Fig. 5.

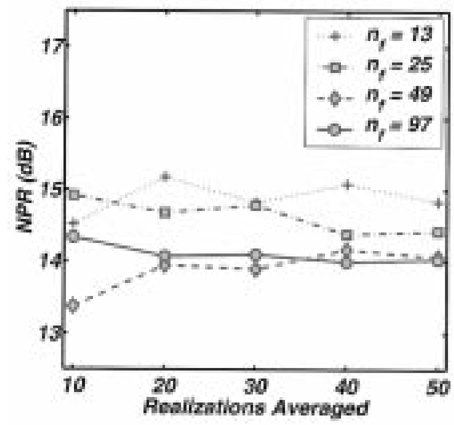

Fig. 3. Calculated NPR as a function of realizations averaged and size of frequency grid. The different symbols correspond to different number of frequencies as described in the insert. The bandpass is 36 MHz wide and centered at $f_0 = 1.6498$ GHz, and the total input power is -16 dBm.

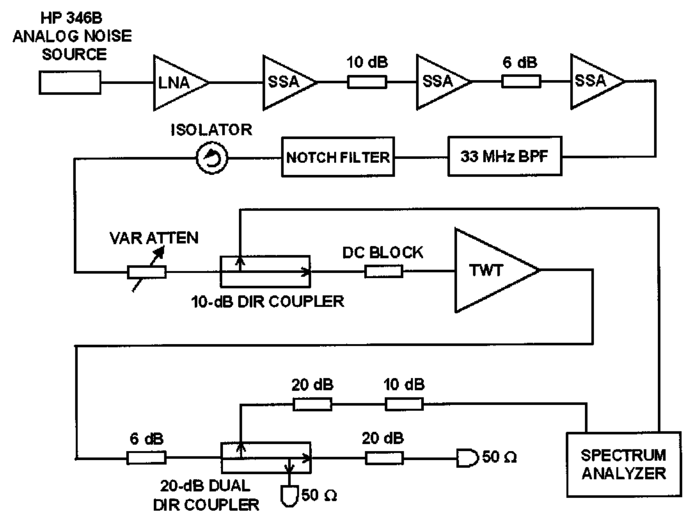

Fig. 4. Diagram of the experimental setup for measuring the NPR in a Hughes 8537H TWT.

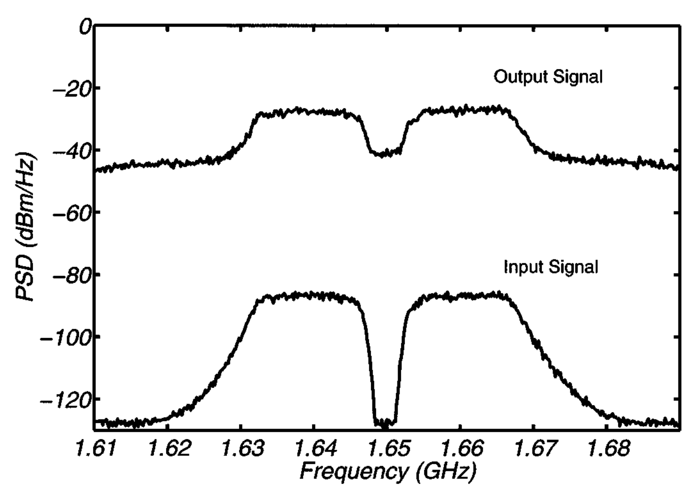

Fig. 5. Time-averaged input and output spectra for a NPR measurement with total input power -14 dBm (3 dB input back-off from saturation, where saturation $= -11$ dBm).

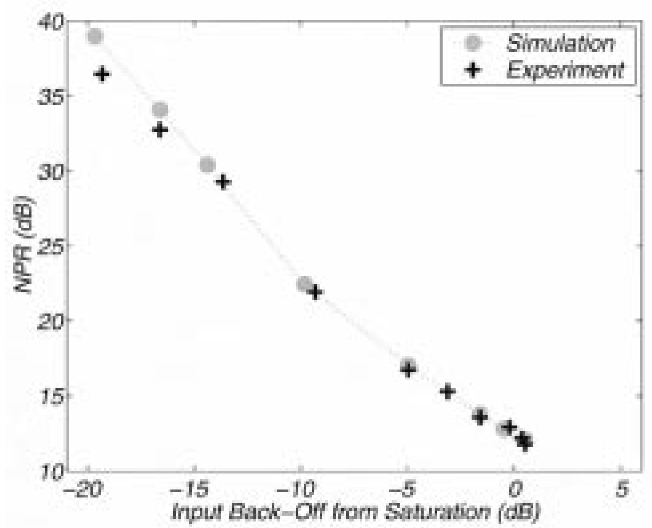

Fig. 6. Comparison between simulated and measured NPR in a Hughes 8537H TWT as a function of total input power. All the simulations were run with $n_f = 13$, and the values shown correspond to the average of 50 realizations. The saturation point corresponds to an input power of -11 dBm.

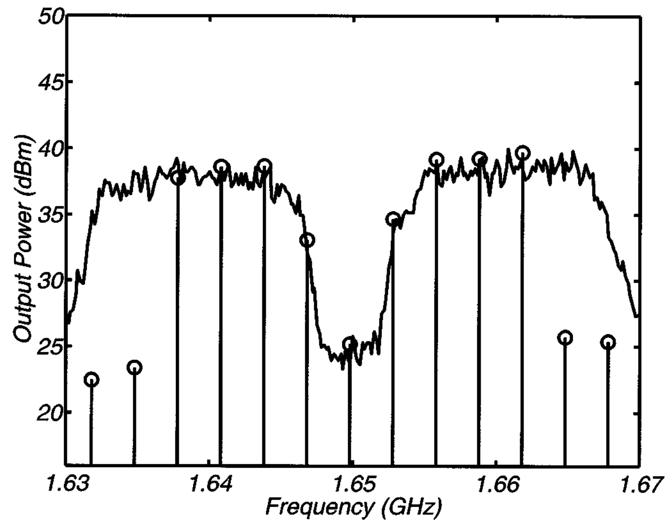

Fig. 7. Comparison between the time-averaged experimental output spectrum (solid line) and simulated average spectrum (50 realizations averaged, vertical lines with circles) for a total input power of -14 dBm. The first and last two points of the simulated spectrum are below the experimental value because, on input, the amplitudes of these points are set to zero to minimize aliasing.

Our main result is presented in Fig. 6. This is a comparison between simulated and measured NPR as a function of total input power $P_{\rm in}$.

For $P_{\rm in} > -15$ dB input back-off from saturation (saturation $= -11$ dBm input power) the simulations agree with the measured values within the experimental error bars (0.5 dB), but for lower values of $P_{\rm in}$ the agreement is not so good. The reason for this discrepancy is that the TWT we used exhibits an anomalous behavior in the intermodulation products at very low input power, as measured by Abe *et al.* [9]. Currently, we do not understand this anomaly, and in future work we hope to unravel its origin.

All the simulations shown in Fig. 6 used $n_f = 13$ and were obtained by averaging 50 realizations. Each realization took 4 min of CPU time on a Hewlett-Packard workstation; therefore, a NPR simulation for a fixed input power took about 2.5 hours.

An important point to be stressed is the accurate representation of the shape of the input notch in the simulations. This is achieved through the shape factors s_j [see (11)–(14)], and Fig. 7 illustrates their use to achieve a good match to the experimental setup.

Using a perfectly square notch in the simulations instead of the tapered one illustrated in Fig. 7 results in an underestimate of the NPR by as much as 3 dB. Therefore, this is an important point to bear in mind when attempting to simulate NPR measurements. In addition, note that, in Fig. 7, the simulation's bandpass is smaller than that used in the experiment. We found that only the frequencies closest to the notch contribute to the NPR—a result consistent with the importance of modeling the notch's shape accurately—and that a ratio of 10 : 1 for the bandpass to notch-width ratio (at the bottom of the notch) is more than sufficient to accurately reproduce the experimental values.

Finally, it is interesting to compare the performance of our simulations with other methods to estimate NPR measurements.

The traditional approach is to expand the transfer function of the DUT in a power series in the input amplitudes and phases (for example, output voltage is expanded as a power series in input voltage). This method is particularly appealing when used in conjunction with CHRISTINE because the code can calculate the coefficients of this expansion up to order 20, and, therefore, this method can be used without resort to experimental measurements or heuristic approximations. Also, use of this method is much faster than using the full, large-signal code for each realization of the experiment.

Recently [10], a method was proposed to calculate NPR using the two-tone intermodulation response (IMR2). This method relies on calculating the IMR2 using the Volterra–Wiener formalism, and the NPR can be bracketed by using the IMR2 at midband and at the band edge.

Fig. 8 is a comparison between the measured NPR and values calculated with full simulations with CHRISTINE (also shown above in Fig. 7), a power-series transfer function with coefficients calculated by CHRISTINE, and using the two-tone intermodulation response as proposed in [10].

Clearly, the simulations with CHRISTINE are superior than any of the other methods, particularly close to saturation. Even at a 10 dB input back-off from saturation the best performing alternative method (the power series transfer function with coefficients calculated by CHRISTINE) underestimates the measured NPR by as much as 2 dB.

V. Conclusions

Using the large-signal, multifrequency code CHRISTINE we have shown that measurements of NPR can be simulated with better than 1 dB accuracy using only 10 to 50 frequency points, in contrast to the experimental setup, which uses thousands of input carriers. In a modern workstation, a typical simulation requires only about 2.5 h.

Moreover, because the calculation of a NPR value requires the average of scores of mutually independent numerical realizations of the experiment, this approach is ideally suited for parallel execution or distributed execution over a network; thus, execution time can be reduced further.

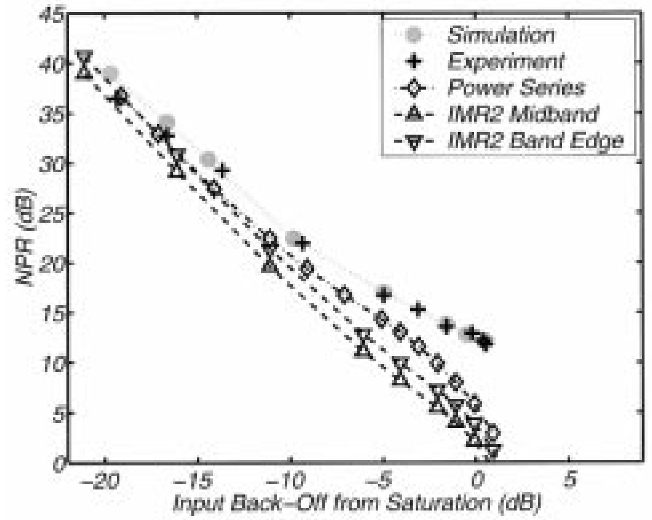

Fig. 8. Comparison between measured NPR in a Hughes 8537H TWT as a function of total input power and calculated values using different approaches. The parameters for the full CHRISTINE simulation are the same as those in Fig. 7. Simulation: full simulation results; Experiment: measured values; Power Series: results obtained by expanding the device's response in a power series in input amplitudes with the expansion coefficients calculated by CHRISTINE; IMR2 Midband: results using the two-tone intermodulation response calculated at midband; IMR2 Band Edge: results using the two-tone intermodulation response calculated at the band edge.

With its circuit parameterization features, CHRISTINE—together with the post-processing software developed for this project—can be used to perform complex trade-off analysis to produce a TWT design with high efficiency and low-distortion that meets a particular NPR specification. Therefore, the code has the potential to favorably impact the development time and costs for new TWT designs by reducing the number of hardware iterations required to prove out the designs.

APPENDIX
HUGHES 8537H TWT SIMULATIONS WITH CHRISTINE

CHRISTINE [2], [3] is a parametric code developed to model helix TWTs. The code allows to simulate the amplification of multifrequency signals and their harmonics, and it includes efficient algorithms to perform complex circuit optimizations.

Whaley *et al.* [8] and Abe *et al.* [9] validated experimentally the code through an extensive comparison of simulated responses with experimental measurements of an L-band Hughes 8537H TWT, the same tube used for the experiments and simulated responses presented here. The tube's mechanical and electrical parameters used in our simulations are presented in Table I [9].[2]

To perform a multifrequency calculation with CHRISTINE, one has to specify the central frequency of the passband considered, f_0, the minimum and maximum frequencies in the passband, $f_{\min}$ and $f_{\max}$, and the frequency separation between points, Δf. The central frequency f_0, $f_{\min}$, and $f_{\max}$ must be integer multiples of Δf. Also, one has to specify the input power and phases for the input tones.

[2]The one-dimensional (1-D) model approximates the effect that the three boron nitride support rods have on the dispersive characteristics of the helix circuit by reducing the dielectric permittivity of the rods to an effective "smeared" relative permittivity that completely fills the area between the inner radius of the vacuum envelope and the outer radius of the helix. The value of this smeared relative dielectric constant is computed using area weighting.

TABLE I
HUGHES 8537H MODEL PARAMETERS

Parameter	Model Value
Helix mean radius	0.2353 cm
Helix wire width	0.0305 cm
Cathode voltage	−3.1 kV
Beam current	65.5 mA
Min. beam radius	0.0962 cm
BN support rods	
smeared[2] permittivity	1.21

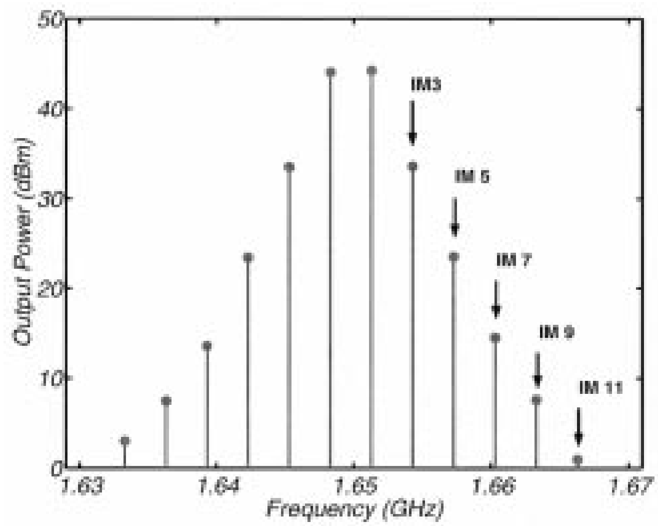

Fig. 9. Two-tone intermodulation products calculated with CHRISTINE.

Fig. 9 is an example of the calculation of the intermodulation products for a two-tone, equal amplitude input. In this example, $f_0 = 1.6498$ GHz, $f_{\min} = 1.631\,425$ GHz, $f_{\max} = 1.668\,175$ GHz, and $\Delta f = 1.5$ MHz; thus, the frequency grid consists of 25 equally-spaced points. All the input phases are set to zero, and only the two frequency points adjacent to f_0 have a nonzero input power, set to be -14 dBm, or about 3 dB input back-off from the saturation point of the Hughes 8537H TWT at 1.65 GHz.

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