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Frequency-Independent and Frequency-Dependent Nonlinear Models of TWT Amplifiers

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Abstract—**Simple two-parameter formulas are presented for the functions involved in the amplitude-phase and the quadrature nonlinear models of a TWT amplifier, and are shown to fit measured data very well. Also, a closed-form expression is derived for the output signal of a TWT amplifier excited by two phase-modulated carriers, and an expression containing a single integral is given when more than two such carriers are involved. Finally, a frequency-dependent quadrature model is proposed whose parameters are obtainable from single-tone measurements.**

I. INTRODUCTION

Traveling-wave tube (TWT) amplifiers, and power amplifiers in general, exhibit nonlinear distortions in both amplitude (AM-to-AM conversion) and phase (AM-to-PM conversion) [1]–[13]. Two equivalent frequency-independent bandpass nonlinear models of helix-type TWT amplifiers have been used in the literature to study the adverse effects of these nonlinearities on various communication systems. These are the *amplitude-phase model* [4]–[7] and the *quadrature model* [8]–[13], in which the portion of the output wave falling in the same spectral zone as the band-limited input wave is described in terms of the envelope of the input wave, rather than its instantaneous value. To specify each model, one needs to know two functions—the amplitude and phase functions for the former model, and the in-phase and quadrature functions for the latter. Several representations (to be discussed later) for these functions have been proposed in the literature which are generally complex in form, or require the knowledge of many parameters.

The purpose of this paper is to present nonlinear models of TWT amplifiers that are based on a simple two-parameter formula for each of the four aforementioned functions. For each of several cases examined, these formulas fit TWT measurements accurately—more so than previously reported formulas. In addition, the formulas permit a closed-form solution of the output signal for an input signal consisting of two phase-modulated carriers, and a solution containing a single integral when more than two such carriers are involved. The parameters of the models are obtainable via straightforward single-tone measurement and computation procedures. Moreover, a simple interpretation of measurements obtained at different frequencies permits frequency selectivity effects to be included in a quadrature model. The latter feature may be particularly useful in cavity-coupled TWT amplifiers and other components that are not as broad-band as helix-type TWT's, or when broad-band input signals are involved.

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II. PROPOSED FORMULAS FOR THE AMPLITUDE-PHASE MODEL

Let the input signal be

$$x(t) = r(t) \cos [\omega_0 t + \psi(t)] \tag{1}$$

where ω_o is the carrier frequency, and $r(t)$ and $\psi(t)$ are the modulated envelope and phase, respectively. [It is worth noting that $r(t)$ may assume positive and negative values.] In the amplitude-phase model [4]–[7], the corresponding output is written as

$$y(t) = A[r(t)] \cos \{\omega_0 t + \psi(t) + \Phi[r(t)]\} \tag{2}$$

where $A(r)$ is an odd function of r, with a linear leading term representing AM-to-AM conversion, and $\Phi(r)$ is an even function of r, with a quadratic leading term [1]–[4] representing AM-to-PM conversion.

Sunde [4] proposed the use of a soft-limiter characteristic to represent the instantaneous amplitude response of a TWT which results in an envelope amplitude function $A(r)$ that does not fall off beyond saturation as observed in practice. He also proposed to represent $\Phi(r)$ by a polynomial in r^2, which would require a large number of terms to fit realistic data, as is the case for polynomial representations in general. Berman and Mahle [5] suggested a three-parameter formula to represent $\Phi(r)$, which will be used later for comparison. Thomas, Weidner, and Durrani [7] proposed a four-parameter formula for $A(r)$, which will also be used for comparison.

Here, we propose to represent $A(r)$ and $\Phi(r)$ by the two-parameter formulas

$$A(r) = \alpha_a r/(1 + \beta_a r^2) \tag{3}$$

$$\Phi(r) = \alpha_\phi r^2/(1 + \beta_\phi r^2). \tag{4}$$

Note that for very large r, $A(r)$ is proportional to $1/r$, and $\Phi(r)$ approaches a constant.

Before testing these formulas against experimental data and against the formulas of [5] and [7], we introduce the quadrature model.

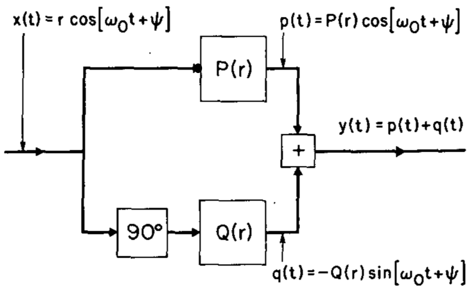

Fig. 1. Quadrature nonlinear model of a power amplifier.

III. PROPOSED FORMULAS FOR THE QUADRATURE MODEL

In the quadrature model [8]-[13], if the input is given by (1), the output is given by the sum of the in-phase and quadrature components (see Fig. 1)

$$p(t) = P[r(t)] \cos[\omega_0 t + \psi(t)] \tag{5a}$$

$$q(t) = -Q[r(t)] \sin[\omega_0 t + \psi(t)] \tag{5b}$$

where $P(r)$ and $Q(r)$ are odd functions of r with linear and cubic leading terms, respectively. Actually, (5) can be deduced form (2) with

$$P(r) = A(r) \cos[\Phi(r)] \tag{6a}$$

$$Q(r) = A(r) \sin[\Phi(r)]. \tag{6b}$$

Eric proposed to represent $P(r)$ and $Q(r)$ as odd polynomials of r, which requires large numbers of terms to fit realistic TWT data. Kaye, George, and Eric [9], and Fuenzalida, Shimbo, and Cook [10] represented each function by a sum of Bessel functions of the first kind of order 1, which results in simplifying the calculation of the output spectrum. However, a large number of terms is still required to fit realistic data. Hetrakul and Taylor [11]-[13] used two-parameter formulas involving modified Bessel functions of the first kind, which will be used later for comparison.

Here, we propose to represent $P(r)$ and $Q(r)$ by the two-parameter formulas

$$P(r) = \alpha_p r/(1 + \beta_p r^2) \tag{7}$$

$$Q(r) = \alpha_q r^3/(1 + \beta_q r^2)^2. \tag{8}$$

Note that for very large r, both $P(r)$ and $Q(r)$ given above are proportional to $1/r$, while those given by Hetrakul and Taylor approach constant values.

A useful property of (7) and (8) is that

$$Q(r) = -\left.\frac{\partial P(r)}{\partial \beta_p}\right|_{\alpha_p \to \alpha_q, \beta_p \to \beta_q}. \tag{9}$$

Thus, if the spectrum of $P(r)$ is calculated for a given $r(t)$, the corresponding spectrum of $Q(r)$ is readily obtained by differentiation. This property will be used in Section V.

IV. FITTING THE FORMULAS TO EXPERIMENTAL TWT DATA

To establish the accuracy of the formulas proposed in (3), (4), (7), and (8), experimental TWT amplitude-phase and

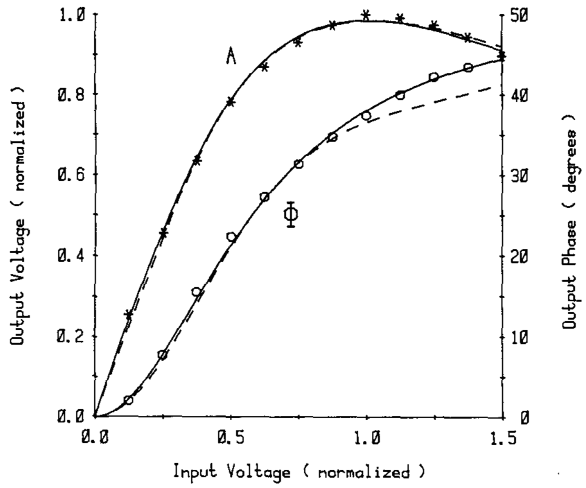

Fig. 2. TWT amplitude (*) and phase (○) data of Berman and Mahle [5]. The solid lines are plotted from (3) and (4), and the dashed lines from the Berman-Mahle [5] phase formula, and from the Thomas-Weidner-Durrani [7] amplitude formula.

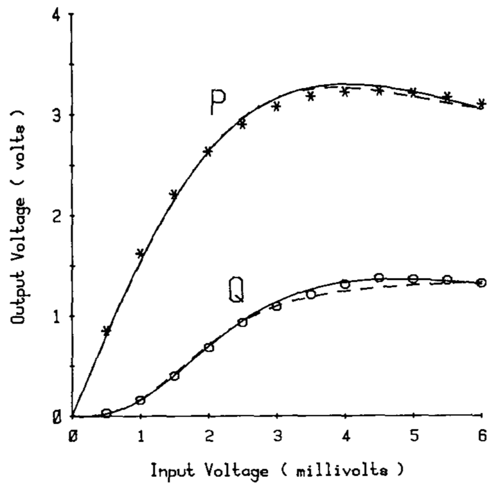

Fig. 3. TWT in-phase (*) and quadrature (○) data of Hetrakul and Taylor [11]-[13]. The solid lines are plotted from (7) and (8), and the dashed lines from their own formulas [11]-[13].

quadrature data were obtained from three different sources—Berman and Mahle [5], Hetrakul and Taylor [11]-[13], and Kaye, George, and Eric [9]. These data are plotted in Figs. 2–6 by asterisks for $A(r)$ and $P(r)$, and by circles for $\Phi(r)$ and $Q(r)$. The units used on the coordinate axes are the same units used in the cited references. In Figs. 2, 5, and 6, the input and output voltages are normalized to their corresponding values at saturation.

A minimum mean-square-error procedure is described in the Appendix for fitting the formulas of (3), (4), (7), and (8) to the experimental data. The results are plotted by the solid lines in Figs. 2–6, which show an excellent fit in each case. The values of the α and β parameters, as well as the resulting root-mean-square (rms) errors, are given in Table I. The α parameters given in the table for $\Phi(r)$ give a dimension of radians when substituted in (4); however, the corresponding rms errors are given in degrees. The remaining quantities in the table have dimensions consistent with those used in the associated figures.

The Thomas-Weidner-Durrani [7] amplitude formula is plotted by a dashed line in Fig. 2 for comparison. The resulting rms error was 0.014 normalized volts, which, in spite

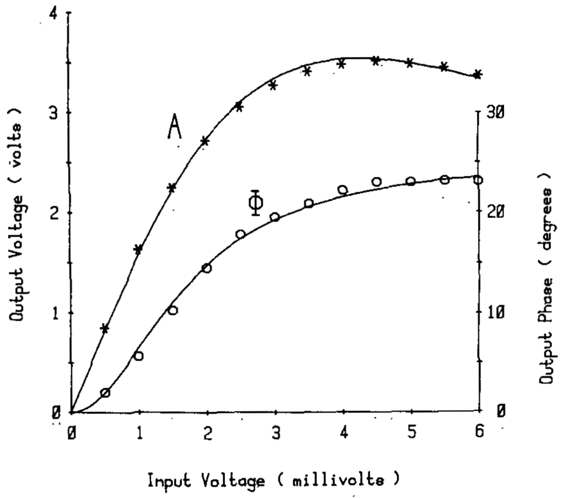

Fig. 4. TWT amplitude (*) and phase (o) data obtained from Hetrakul and Taylor [11]–[13] through the use of (6). The solid lines are plotted from (3) and (4).

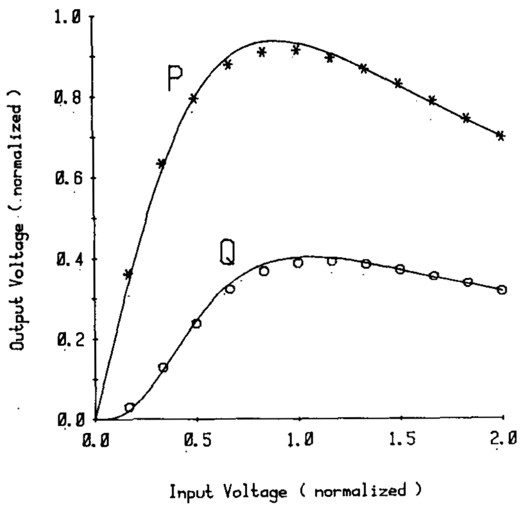

Fig. 5. TWT in-phase (*) and quadrature (o) data of Kaye, George, and Eric [9]. The solid lines are plotted from (7) and (8).

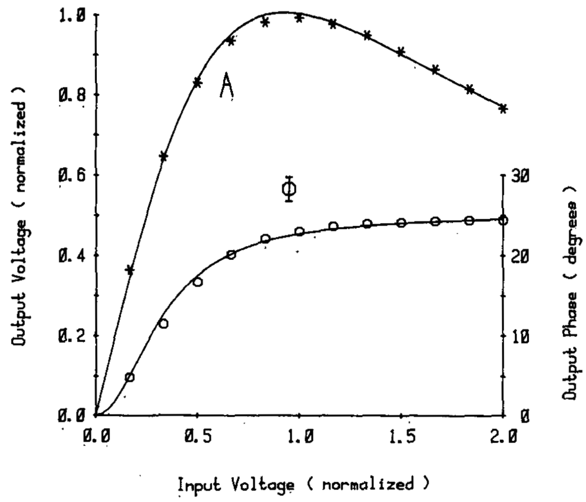

Fig. 6. TWT amplitude (*) and phase (o) data of Kaye, George, and Eric [9]. The solid lines are plotted from (3) and (4).

TABLE I
OPTIMUM PARAMETERS AND RESULTING RMS ERRORS OBTAINED BY FITTING (3), (4), (7), AND (8) THROUGH VARIOUS EXPERIMENTAL TWT DATA (FIGS. 2–6)

Data from Reference	Function	α	β	RMS Error	Plotted In
[5]	A(r)	1.9638	0.9945	0.012	Fig. 2
	Φ(r)	2.5293	2.8168	0.478	
[11]-[13]	P(r)	1.6397	0.0618	0.057	Fig. 3
	Q(r)	0.2038	0.1332	0.023	
	A(r)	1.6623	0.0552	0.041	Fig. 4
	Φ(r)	0.1533	0.3456	0.508	
[9]	P(r)	2.0922	1.2466	0.015	Fig. 5
	Q(r)	5.5290	2.7088	0.009	
	A(r)	2.1587	1.1517	0.010	Fig. 6
	Φ(r)	4.0033	9.1040	0.469	

of the large number of parameters involved, is slightly larger than our 0.012 value given in Table I. The Berman–Mahle [5] phase function is also plotted by a dashed line through their own phase data in Fig. 2. While that formula gives a good fit up to saturation, large errors are encountered beyond saturation.

The Hetrakul–Taylor [11]–[13] in-phase and quadrature formulas are plotted through their own data in Fig. 3. The resulting rms errors for $P(r)$ and $Q(r)$ are 0.058 V and 0.040 V, respectively, which are somewhat higher than our corresponding 0.057 and 0.023 values given in Table I.

V. INTERMODULATION ANALYSIS

One of the applications intended for a nonlinear model of a power amplifier is to be able to find the output signals associated with various input signals of practical interest. Here we consider the case of multiple phase-modulated carriers. The quadrature model of (7) and (8) will be employed in the analysis. A two-carrier signal will be considered first because it yields a closed-form expression for the output signal. Actually, the use of such a signal has been proposed for a satellite communication system [14]. Moreover, the results, with the phase modulation suppressed, can be used in conjunction with the two-tone nonlinearity test [3], [4], [15].

A. Two Phase-Modulated Carriers

Let the input signal be

$$x(t) = V_1 \cos[\omega_1 t + \psi_1(t)] + V_2 \cos[\omega_2 t + \psi_2(t)]. \tag{10}$$

Thus, the square of its envelope is given by

$$r^2(t) = V_1{}^2 + V_2{}^2 + 2V_1V_2 \cos[\Omega t + \delta(t)] \tag{11}$$

where $\Omega = \omega_2 - \omega_1$ and $\delta(t) = \psi_2(t) - \psi_1(t)$. Substituting (11) into (7) and (8), using the Fourier integration given in [16, (3.613.1), p. 366], and employing (9), one obtains

$$P[r(t)]/r(t) = P_0 + 2\sum_{k=1}^{\infty} P_k \cos[k\Omega t + k\delta(t)] \tag{12}$$

$$Q[r(t)]/r(t) = Q_0 + 2\sum_{k=1}^{\infty} Q_k \cos[k\Omega t + k\delta(t)] \tag{13}$$

where

$$P_k = [\alpha_p/(D_p \cos\theta_p)][-\tan(\theta_p/2)]^k \tag{14}$$

$$Q_k = [\alpha_q/(\beta_q D_q{}^2 \cos^3\theta_q)][D_q \cos^2\theta_q - k\cos\theta_q - 1] \cdot [-\tan(\theta_q/2)]^k \tag{15}$$

$$D_{p,q} = 1 + \beta_{p,q}[V_1{}^2 + V_2{}^2] \tag{16a}$$

$$\sin\theta_{p,q} = 2\beta_{p,q}V_1V_2/D_{p,q}. \tag{16b}$$

Finally, (5), (10), (12), and (13) give the output signal

$$\begin{aligned} y(t) = \sum_{k=0}^{\infty} \{&(P_kV_1 + P_{k+1}V_2)\cos[(\omega_1 - k\Omega)t \\ &+ \psi_1(t) - k\delta(t)] - (Q_kV_1 + Q_{k+1}V_2) \\ &\cdot \sin[(\omega_1 - k\Omega)t + \psi_1(t) - k\delta(t)] \\ &+ (P_{k+1}V_1 + P_kV_2)\cos[(\omega_2 + k\Omega)t \\ &+ \psi_2(t) + k\delta(t)] - (Q_{k+1}V_1 + Q_kV_2) \\ &\cdot \sin[(\omega_2 + k\Omega)t + \psi_2(t) + k\delta(t)]\}. \end{aligned} \tag{17}$$

B. Multiple Phase-Modulated Carriers

Let the input signal be

$$x(t) = \sum_{i=1}^{n} V_i \cos[\omega_i t + \psi_i(t)]. \tag{18}$$

We obtain the output signal by employing Shimbo's general formula [6], which, in our notation, is given by

$$y(t) = \text{real}\left[\sum_{\substack{k_1,k_2,\cdots,k_n=-\infty \\ k_1+k_2+\cdots+k_n=1}}^{\infty} M(k_1,k_2,\cdots,k_n) \cdot \exp\left\{\sum_{i=1}^{n} j[k_i\omega_i t + k_i\psi_i(t)]\right\}\right] \tag{19}$$

with

$$M(k_1,k_2,\cdots,k_n) = \int_0^\infty s\left[\prod_{i=1}^{n} J_{k_i}(V_i s)\right] ds \cdot \int_0^\infty r[P(r) + jQ(r)]J_1(rs)\,dr \tag{20}$$

where J_k is the Bessel function of the first kind of order k.

Substituting (7) and (8) in (20), one can evaluate the integration over r through the use of (9) and the Hankel-type integral given in [16, (6.565.4), p. 686] to obtain

$$\begin{aligned} M(k_1,k_2,\cdots,k_n) = \int_0^\infty s\left[\prod_{i=1}^{n} J_{k_i}(V_i s)\right] &\cdot \{\alpha_p\beta_p{}^{-3/2}K_1(s\beta_p{}^{-1/2}) \\ &+ j\alpha_q[\beta_q{}^{-5/2}K_1(s\beta_q{}^{-1/2}) \\ &- (s/2\beta_q{}^3)K_0(s\beta_q{}^{-1/2})]\}\,ds \end{aligned} \tag{21}$$

where K_0 and K_1 are the modified Bessel functions of the second kind of orders 0 and 1, respectively.

For $n = 2$, (21) reduces to the closed-form expressions obtained in the previous section. However, numerical integration of (21) seems necessary if $n > 2$. A direct numerical integration of (20) would, of course, be more difficult since a double integral is involved.

It can be shown that when n is very large, then the integral in (21) can be evaluated in terms of a finite sum involving the exponential integral. This important result will be given in a future paper [17].

VI. A FREQUENCY-DEPENDENT QUADRATURE MODEL

Thus far, it has been implied that the characteristics of the TWT amplifier are independent of frequency over the band of interest, which is often the situation encountered in practice, especially when helix-type TWT's are employed. However, when broad-band input signals are involved, or in cases where amplifiers and components are used that are not as inherently broad-band as helix-type TWT's, a frequency-dependent model is needed. Here we make a conjecture, to be explained later, that enables us to infer such a model from single-tone measurements.

Consider a single-tone test in which the input signal to the TWT has adjustable amplitude r and frequency f and let the measured amplitude and phase of the output signal be $A(r,f)$ and $\Phi(r,f)$, respectively. Let

$$\Phi_0(f) = \lim_{r\to 0} \Phi(r,f) \tag{22}$$

be the measured small-signal phase response. Thus, $\Phi(r,f) - \Phi_0(f) \to 0$ as $r \to 0$. In fact, this phase difference is what was labeled $\Phi(r)$ in Sections II-IV. Using (6), one obtains the in-phase and quadrature output signals

$$P(r,f) = A(r,f)\cos[\Phi(r,f) - \Phi_0(f)] \tag{23}$$

$$Q(r,f) = A(r,f)\sin[\Phi(r,f)] - \Phi_0(f)]. \tag{24}$$

For each given f, one can fit the formulas of (7) and (8) to these data as is described in the Appendix. This would result in the frequency-dependent parameters $\alpha_p(f)$, $\beta_p(f)$, $\alpha_q(f)$, and $\beta_q(f)$. One now can compute the functions

$$H_p(f) = \sqrt{\beta_p(f)} \tag{25}$$

$$G_p(f) = \alpha_p(f)/\sqrt{\beta_p(f)} \tag{26}$$

$$H_q(f) = \sqrt{\beta_q(f)} \tag{27}$$

$$G_q(f) = \alpha_q(f)/\beta_q^{3/2}(f). \tag{28}$$

Let us define the normalized frequency-independent envelope nonlinearities

$$P_0(r) = r/(1 + r^2) \tag{29}$$

$$Q_0(r) = r^3/(1 + r^2)^2, \tag{30}$$

which are obtained from (7) and (8) by setting the α's and β's to unity. It is observed from (7), (25), (26), (29), and Fig. 1 that the operation performed on a single-tone signal passing through the in-phase branch can be divided into three steps: first, the input amplitude is scaled by $H_p(f)$; next, the resulting signal passes through the frequency-independent envelope nonlinearity $P_o(r)$; and finally, the output amplitude is scaled by $G_p(f)$. Three similar steps apply for the operation performed in the quadrature branch. Now, performing each of the aforementioned frequency-dependent amplitude-scaling operations may be interpreted as passing the signal through an appropriately located linear filter having the corresponding *real* frequency response. This leads to the frequency-dependent quadrature model shown in Fig. 7. That model is, of course, valid for single-tone input signals by virtue of the procedure used to construct it. However, its validity for arbitrary input signals is, at this point, a conjecture that remains to be confirmed experimentally.

The box in the output side of Fig. 7 is a linear, all-pass network having an amplitude response of unity, and a phase response of $\Phi_o(f)$, which is defined in (22). For a single-tone input signal, that phase response can be absorbed totally or partially into the filters in each of the two branches of the model without affecting the output signal. However, for the model to give an acceptable representation of the power amplifier for arbitrary input signals, it may be necessary to absorb that phase response into the various filters in a particular manner. This point needs further investigation.

VII. CONCLUSIONS

Simple two-parameter formulas have been proposed for each of the functions of the amplitude-phase model of a TWT amplifier, as well as for each of the functions of the equivalent quadrature model. The formulas fit available TWT data very well. This implies that the nonlinear behavior of a given TWT can be accurately represented by only four parameters. The same is also true for the quadrature model formulas proposed by Hetrakul and Taylor [11]-[13]. Those formulas, however, seem unnecessarily complicated.

The simplicity of the proposed formulas resulted in a closed form expression for the output signal of a TWT amplifier excited by two phase-modulated carriers. Moreover, when the input signal consists of more than two such carriers, it was shown that the output signal can be represented by a single integral. When the number of carriers is very large, it can be shown that this integral can be evaluated in terms of a finite sum involving the exponential integral. This result, however, was not given here, and will be presented in a future paper [17].

A frequency-dependent quadrature model of a TWT amplifier has also been proposed (see Fig. 7). The various parameters of the model can be obtained by applying minimum mean-square-error curve fitting to measurements involving a single tone of variable amplitude and frequency.

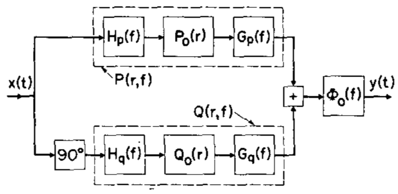

Fig. 7. Proposed frequency-dependent quadrature model of a TWT amplifier. $P_0(r)$ and $Q_0(r)$ are frequency-independent envelope nonlinearities given in (29) and (30); the H's and the G's are real linear filters whose frequency responses are defined in (25)-(28), and $\Phi_0(f)$ is a linear all-pass network whose phase response is defined in (22).

APPENDIX

CURVE-FITTING PROCEDURE

The formulas (3), (4), (7), and (8) assume the general form

$$z(r) = \alpha r^n/(1 + \beta r^2)^\nu \tag{A1}$$

where $n = 1$, 2, or 3, and $\nu = 1$ or 2. (Actually, noninteger values of ν were considered in fitting the data in Section IV, and, remarkably, $\nu = 1$ or 2 was found to be very nearly optimum.) Given m measured pairs, (z_i, r_i), $i = 1, 2, \cdots, m$, we need to find α and β to fit (A1) to these data. Defining

$$w_i = (z_i/r_i^n)^{-1/\nu}, \qquad i = 1, 2, \cdots, m \tag{A2}$$

and employing standard minimum mean-square-error curve-fitting procedure, one obtains the required optimum values of α and β

$$\alpha = \left\{ \frac{(\Sigma r_i^2)^2 - m\Sigma r_i^4}{(\Sigma r_i^2)(\Sigma w_i r_i^2) - (\Sigma r_i^4)(\Sigma w_i)} \right\}^\nu \tag{A3}$$

$$\beta = \frac{(\Sigma r_i^2)(\Sigma w_i) - m\Sigma w_i r_i^2}{(\Sigma r_i^2)(\Sigma w_i r_i^2) - (\Sigma r_i^4)(\Sigma w_i)} \tag{A4}$$

where all the summations are over $i = 1$–m.

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