Methods for Measuring the Power Linearity of Microwave Detectors for Radiometric Applications

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***Abstract*—A microwave radiometer relies on the power linearity of its microwave receivers to accurately measure the temperature of remote microwave noise sources. This paper considers linearity issues in the design and characterization of such receivers. Analysis is presented relating the radiometer temperature interpolation error to a second order power nonlinearity coefficient for the receiver. Formulas are also developed specifying the temperature error in terms of individual receiver component parameters. It is shown that the key parameter for the RF detector in the receiver is A_4, a fourth order RF distortion coefficient, and the key parameter for the RF amplifiers in the receiver is IP_3, the third order intercept. This paper also discusses experimental methods for measuring the power linearity of RF detectors to the levels required for radiometric applications. Three methods are discussed: the two-tone method, the amplitude modulation method, and the constant ratio method. The theory of determining the coefficients that characterize the nonlinearity of the detector from experimental data is presented. Experimental results are presented showing that the two-tone method and the constant ratio method agree to within experimental error. The sensitivity for measuring nonlinearities and the difficulties encountered in implementing each of these methods are also discussed.**

I. Introduction

A MICROWAVE radiometer relies on the power linearity of its microwave receivers to accurately measure the temperature of remote microwave noise sources [1]–[6], [9]. Fig. 1 shows how a typical radiometer measures a remote scene temperature T by periodically scanning an antenna feed and receiver between a cold load at temperature T_c, a hot load at temperature T_h, and the remote scene while recording the noise power at the output of the receiver. A typical radiometric receiver consists of a low noise amplifier (LNA) or mixer-local oscillator (LO) which establishes a receiver noise figure F. A bandpass filter establishes a noise or convolution bandwidth B, and an RF detector and video amplifier provide a video voltage output V that is assumed to vary linearly with P the effective noise power at the antenna input given by

$$P = kB[T + (F - 1)T_0] \tag{1}$$

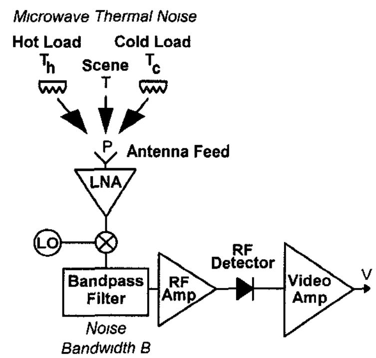

Fig. 1. Typical microwave radiometer.

where k is Boltzman's constant and T_0 is 290 K. By measuring V when the antenna feed is pointed to T_c, T_h, and T, one can determine $\hat{T}$ an estimate of T using the linear interpolation formula

$$\hat{T} - T_c = \frac{T_h - T_c}{V_h - V_c}(V - V_c). \tag{2}$$

Typically. the radiometer must interpolate T to a fraction of a degree K out of a system noise temperature of 1000 K or more. Thus, the power linearity of the receiver is a critical issue in determining the accuracy of this interpolation. This paper will consider such power linearity issues in the design and characterization of radiometric receivers. The first section will develop formulas for characterizing the nonlinear temperature error in terms of the RF properties of the receiver and individual receiver components. The next section will deal with experimental methods for measuring the power linearity of RF detectors used in radiometric receivers.

II. Power Linearity Considerations in the Design of a Radiometric Receiver

In this section, a formula for the temperature interpolation error due to a second order receiver power nonlinearity and detailed formulas relating the nonlinear temperature error to RF nonlinearity parameters of individual receiver components will be developed.

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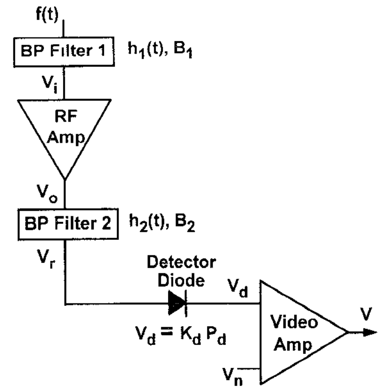

Fig. 2. Model of radiometric receiver.

A. Temperature Error Due to 2nd-Order Power Nonlinearity

Let us assume the receiver has a small second order power nonlinearity given by

$$V \propto P - CP^2. \tag{3}$$

Using (1), (2), and (3), one can show that the worst case interpolation temperature error $\delta T_L = T - \hat{T}$ is given by

$$\frac{\delta T_L}{T_h - T_c} = \frac{C}{4}(P_h - P_c) \tag{4}$$

where it is assumed that $|CP| \ll 1$. Using typical parameters for a spaceborne radiometer, (4) can be approximated by

$$\frac{\delta T_L}{T_0} \cong \frac{C}{4F}P_h = \frac{1}{4F}\frac{\delta P_h}{P_h} \qquad [T_c \cong 0\ K,\ T_h \cong T_0] \tag{5}$$

where $\delta P = CP^2$. From (5) using typical values of $\delta T_L = 0.3$ K and $F = 4$ dB, one can see that the maximum allowable power nonlinearity for a spaceborne radiometric receiver is $\delta P_h / P_h$ is -20 dB.

B. Temperature Error Due to RF Component Nonlinearities

Let us now utilize the model of the radiometric receiver shown in Fig. 2 to develop formulas for relating δT_L to individual component nonlinearity parameters. This model first contains an RF amplifier, representing a composite of all RF amplifiers and mixers, surrounded by two bandpass filters with time domain response functions and noise bandwidths $h_1(t)$, B_1 and $h_2(t)$, B_2. Next are an RF detector with a power sensitivity K_d and a video amplifier. In the following subsections, we will show how the nonsquare law behavior of the detector or the receiver as a whole and the third order intermodulation distortion of the RF amplifier both contribute to the coefficient C.

C. Power Nonlinearity to Due Non-Square Law RF Behavior

A radiometric receiver and the RF detector and video amplifier in that receiver ideally act as perfect square law detectors of their respective RF inputs. Using Wiener-Volterra [6] analysis and assuming that nonlinear device bandwidths are much larger than that of the linear filters in the receiver, one can show that the receiver video output V can be written in terms of the bandlimited RF voltage V_r (see Fig. 2) as

$$V = A_2\langle V_r^2\rangle + A_4\langle V_r^4\rangle + \cdots \tag{6}$$

where $\langle X\rangle$ represents the expectation value of X over an infinite time or ensemble average and where it is assumed that $\langle V_r^{2n+1}\rangle = 0$.

For a radiometric receiver, V_r is bandlimited thermal noise which can be represented by

$$V_r(t) = \int h(t-t')f(t')dt' \tag{7}$$

where 1)

$$h(t-t'') = \int h_2(t-t')h_1(t'-t'')dt' \tag{8}$$

represents the composite of filter 1 and 2 with a total noise bandwidth of

$$B = (1/2)\int h(t)^2 dt \tag{9}$$

and 2) it is assumed that both filters are lossless and the RF amplifier has unity gain, and (c) $f(t)$ is a random Langevin function [7, 8] whose 2nd and 4th order autocorrelation functions are

$$\langle f(t)f(t')\rangle = \frac{N_0}{2}\delta(t-t') \tag{10}$$

$$\begin{aligned}\langle f(t_1)f(t_2)f(t_3)f(t_4)\rangle = (N_0/2)^2[&\delta(t_1-t_2)\delta(t_3-t_4)\\ &+\delta(t_1-t_3)\delta(t_2-t_4)\\ &+\delta(t_1-t_4)\delta(t_2-t_3)]\end{aligned} \tag{11}$$

with

$$N_0 = kT_s = k[T + (F-1)T_0]. \tag{12}$$

Noting that $P = \langle V_r^2\rangle$ and substituting (7) into (6), one obtains to fourth-order

$$\begin{aligned}V = A_2P + A_4 &\\ \times \int &h(t-t_1)\cdots h(t-t_4)\\ \times &\langle f(t_1)f(t_2)f(t_3)f(t_4)\rangle dt_1\cdots dt_4.\end{aligned} \tag{13}$$

Finally using (1), (9), (11), and (12), one can show that (13) becomes

$$V = A_2P + 3A_4(N_0B)^2 = A_2P + 3A_4P^2. \tag{14}$$

Thus, the nonlinearity coefficient is

$$C = -3\frac{A_4}{A_2} \qquad \text{[Bandlimited Noise Input]}. \tag{15}$$

Note that there is a different relationship for a CW input

$$C = -\frac{3}{2}\frac{A_4}{A_2} \qquad \text{[CW Input]}. \tag{16}$$

Nonlinearity figure of merit for RF detector: Before proceeding to discuss the RF amplifier, let us utilize (15) to define a figure of merit for the acceptability of an RF detector in a particular radiometric application. From (5) and (15), one can write

$$P_{\mathrm{dh}} = \frac{4F}{3}\left|\frac{A_2}{A_4}\right|\frac{\delta T_{\mathrm{Ld}}}{T_0} \tag{17}$$

where P_{dh} is the power into the detector at which $|\delta T_L|$ reaches the allocated value δT_{Ld}. One can show that the lowest power that can be input to the detector while meeting radiometer noise equivalent $\delta T(\mathrm{NE}\delta T)$ requirements can be written as [4]

$$P_{\mathrm{dm}} = \frac{V_n T_{\mathrm{sm}}}{K_d \delta T_V} \tag{18}$$

where V_n is the equivalent input voltage noise of the video amplifier, δT_V is the allocated contribution of the video amplifier to the total NEδT, K_d is the detector diode sensitivity, and T_{sm} is the system noise temperature at the minimum scene temperature. Using (17) and (18), one can define a figure of merit for the detector as

$$M_d = \frac{P_{\mathrm{dh}}}{P_{\mathrm{dm}}} = \frac{4F}{3}\left|\frac{K_d A_2}{V_n A_4}\right|\frac{\delta T_{\mathrm{Ld}}}{T_0}\frac{\delta T_V}{T_{\mathrm{sm}}} \tag{19}$$

which gives the detector's allowable operating range ($M_d > 1$ for a suitable detector). Thus, the suitability of a detector diode for radiometric applications depends both on its linearity given by A_4/A_2 and its sensitivity given by K_d (as well as the noise of the video amplifier given by V_n).

Effects of RF amplifier nonlinearity: Finally, let us consider the effect of RF amplifier distortion on the coefficient C. Again, Fig. 2 defines the components and variables. The amplifier is modeled by

$$V_0 = V_i + \sum_{n=2} a_n V_i^n. \tag{20}$$

Using (20), one can write the voltage into the RF detector as

$$V_r(t) = \int h(t-t')f(t')dt' + a_3\int h_2(t-t')\left[\int h_1(t'-t'')f(t'')dt''\right]^3 dt' \tag{21}$$

where we have assumed that Filters 1 and 2 are narrow enough to filter out the a_2 term, which generates only DC and second harmonic components. Utilizing the first term in (6) and (1), (10), (11), (12), and (21), it can be shown that

$$V \cong A_2\langle V_r^2\rangle = A_2[P + 6a_3(B_1/B)P^2] \tag{22}$$

where B_1 is defined similarly to (9) with $h_1(t)$ substituted for $h(t)$. Since $B_1 \geq B$, C is minimized when $B_1 \cong B$. This is accomplished by using Filter 1 to set the noise bandwidth B and making Filter 2 just narrow enough to filter out the second harmonic distortion of the RF amplifier. Comparing (22) to (3), we finally obtain

$$C = -6a_3(B_1/B). \tag{23}$$

One can show that IP_3 the RF amplifier's the third order intercept for a coherent CW input is equal to

$$\mathrm{IP}_3 = \frac{2}{|a_3|} \tag{24}$$

by expanding (2) for a coherent input and setting the asymptotic third order intermodulation ratio to one. Using (24) in (23) then yields

$$|C| = \frac{12}{\mathrm{IP}_3}\frac{B_1}{B}. \tag{25}$$

Thus, the contribution to C from the RF amplifier can be determined by measuring IP_3.

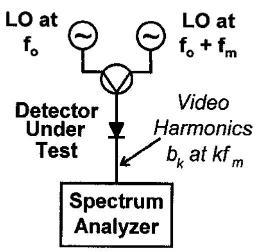

Fig. 3. The two-tone method for measuring detector nonlinearity.

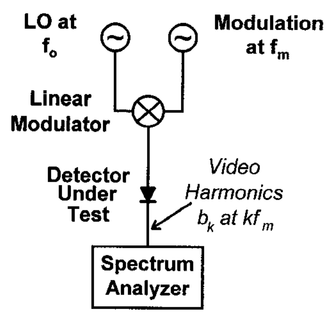

Fig. 4. The amplitude modulation method for measuring detector nonlinearity.

III. Methods for Measuring the Linearity of RF Detectors

In the following subsections we will describe three methods for measuring A_4/A_2 in RF detectors to the levels required for radiometric applications.

A. The Two-Tone Method

The two-tone method for measuring diode nonlinearity is shown in Fig. 3. Here, two equal power RF signals with frequencies at f_0 and $f_0 + f_m$ are added to generate a modulated RF signal V_{RF} which is applied to the detector under test. The video output, which consists of harmonics of amplitude B_k at frequencies kf_m (k = 0,1,2,...), is then monitored on a low frequency spectrum analyzer. These video harmonics can be utilized to determine A_4 and higher order nonlinearity coefficients describing the RF detector output as follows.

Let the RF input to the detector be given by

$$V_r = P_0^{0.5}[\mathrm{Cos}(\omega_0 t) + \mathrm{Cos}((\omega_m + \omega_0)t)]. \tag{26}$$

Utilizing (26) in (6) and assuming that a low pass filter eliminates RF components from V, one can show that [4]

$$V = \sum_k b_k \mathrm{Cos}(k\omega_m t) \tag{27}$$

TABLE I
TWO-TONE COEFFICIENTS b_{kn}

	k = 0	k = 1	k = 2	k = 3	k = 4
n = 1	1	1	0	0	0
n = 2	9/4	3	3/4	0	0
n = 3	25/4	75/8	15/8	5/8	0
n = 4	1225/64	245/8	245/8	35/8	35/64

TABLE II
AMPLITUDE MODULATION COEFFICIENTS b_{kn}

	k = 1	k = 2	k = 3	k = 4
n = 1	2	1/2	0	0
n = 2	7	7/2	1	1/8

where

$$b_k = \sum_{n=k}^{\infty} P_0^n A_{2n} b_{kn} \tag{28}$$

and the b_{kn} are given by the expansion

$$[\mathrm{Cos}(\omega_0 t) + \mathrm{Cos}((\omega_m + \omega_0)t]^{2n} = \sum_{k=0}^{n} b_{kn} \mathrm{Cos}(k\omega_m t). \tag{29}$$

Table I lists the first few b_{kn} coefficients. From the table, we obtain

$$\frac{b_2}{b_1} = \frac{3}{4}\frac{A_4}{A_2}P_0 \qquad \left[\left|\frac{A_4}{A_2}\right| P_0 \ll 1\right]. \tag{30}$$

Thus (A_4/A_2) can be straightforwardly determined from the ratio of the amplitudes of the first and second harmonics of the video output.

B. The Amplitude Modulation Method

When two RF frequency sources are not available, one can utilize the amplitude modulation technique shown in Fig. 2 to measure diode power linearity. Here a linear modulator, an audio oscillator, and a single RF frequency source generates the modulated RF voltage. Again, the b_k are measured with a spectrum analyzer, and b_k is related to coefficients b_{kn} by (28). For this method, however, the b_{kn} are given by [4] Table 2. The primary limitation of this method is the linearity of the modulator. The sensitivity in measuring B_k for both the two-tone and amplitude modulation methods is limited by the presence of intermodulation (IM) products in the RF signal impinging on the detector.

C. Constant Ratio Method

The constant ratio method is shown in Fig. 5. Here one utilizes an attenuator A to set an RF power level from a single LO to a value P_k. A second attenuator B is then used to change P_k by an unspecified but constant ratio Q while a digital voltmeter measures the detector video output values $Y(P_k)$ and $Y(QP_k)$. For a linear power detector, we note that

$$D_k = \frac{Y_1(P_k)}{Y_2(QP_k)} \tag{31}$$

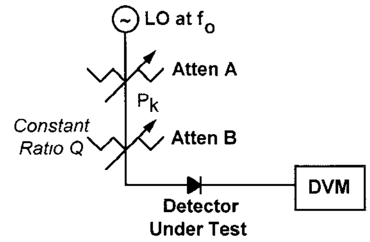

Fig. 5. The constant ratio method for measuring detector nonlinearity.

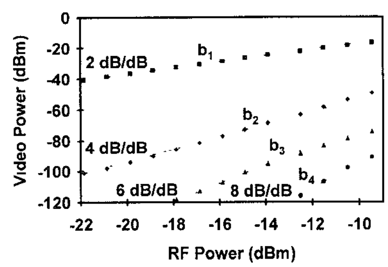

Fig. 6. Video two-tone spectrum of tunnel diode at 6 GHz.

is a constant Q^{-1}. Thus, changes in D_k when varying P_k reflect the power nonlinearity level of the detector. One can show that [4]

$$M = \frac{D_1}{D_2} - 1 \cong \frac{3}{2}\frac{A_4}{A_2}(1-Q)(P_1 - P_2). \tag{32}$$

For $Q = 0.5$ and $P_2 = 0.5P_1$, one can show that the measurement error is minimized [4] and that

$$M \cong \frac{3}{8}\frac{A_4}{A_2}P_1 \qquad [Q = 0.5,\ P_2 = 0.5P_1]. \tag{33}$$

D. Experimental Comparison of Two-Tone and Constant Ratio Methods

To verify the accuracy of the two-tone and constant ratio methods, both were implemented on a detector diode at 6 GHz. For the two-tone measurement, one has to be careful about the generation of intermodulation (IM) products through direct coupling of the RF sources, since these IM's can generate spurious b_k signals at the detector output, limiting the sensitivity of the b_2/b_1 measurement. In the experimental setup, these IM's were reduced by placing isolators and attenuators after each of the RF sources. A worst case IM level due to source coupling of -79 dB was achieved.

Fig. 6 plots the measured b_k magnitudes as a function of the RF power into the detector along with theoretical $2k$ dB/dB curves (dotted lines). Observing this $2k$ dB/dB behavior (leading terms in (28)) is an important indicator that there is no contamination of the data by IM's from the RF sources. The slight deviation of the measured values from $2k$ dB/dB behavior at the high RF power is due to the higher order terms in (28). These deviations are useful in obtaining the signs of the b_k from the measured magnitude data so that a polynomial characterizing the detector video voltage as a function of input RF power can be generated.

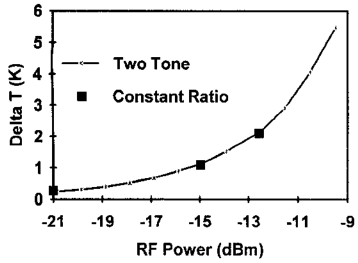

Fig. 7. Comparison of two-tone and constant ratio results at 6 GHz.

TABLE III
SUMMARY OF IMPORTANT FORMULAS [$T_c \cong 0\ K$, $T_h \cong T_0$]

Item	Comment	Equation	Reference Equations
1	Detector/Video Amp δT_L	$\frac{\delta T_{Ld}}{T_0} \cong \frac{3}{4F}\frac{A_4}{A_2}P_h$	(5),(15)
2	RF Amp δT_L	$\frac{\delta T_{La}}{T_0} \cong \frac{3}{F}\frac{B_1}{B}\frac{P_h}{IP_3}$	(5),(25)
3	A_4/A_2 From Two Tone Method	$\frac{A_4}{A_2} = \frac{4}{3P_0}\frac{b_2}{b_1}$	(30)
4	Detector Figure of Merit (Must be > 1)	$M_d = \frac{4F}{3}\left\lvert\frac{K_d A_2}{V_n A_4}\right\rvert\frac{\delta T_{Ld}}{T_0}\frac{\delta T_V}{T_{sm}}$	(19)

The limiting sensitivity factor in the b_2/b_1 measurement was the -79 dB source IM level. (The spectrum analyzer noise level of -135 dBm did not limit the measurement.) Equations (15) and (30) can be rewritten as

$$\frac{\delta P}{P_0} = CP_0 = 4\frac{b_2}{b_1} \quad \text{[Bandlimited White Noise]} \quad (34)$$

so the two-tone measurement sensitivity achieved for $\delta P/P_0$ was -73 dB.

The constant ratio method was more difficult to implement than the two-tone method. The accuracy of the 6 GHz measurement was limited by drifts in the power ratio Q generated by Attenuator B, the RF amplitude of the signal generator, and the DC video voltage measurements. Measurement had to be made quickly to minimize the effects of these drifts. To minimize coupling between the attenuators, it was also found necessary to utilize many isolators around Attenuators A and B. The standard error for the average M measurement was 4.4×10^{-4} or -34 dB. Rewriting (15) and (33) yields

$$\frac{\delta P}{P_0} = CP_0 = 8M \quad \text{[Bandlimited White Noise]} \quad (35)$$

so the constant ratio measurement sensitivity achieved for $\delta P/P_0$ was -25 dB.

Fig. 7 shows a comparison of the two-tone and constant ratio results for the 6 GHz tunnel diode. The measurement data has been converted to an equivalent temperature nonlinearity error using (5) and (15) assuming a noise figure F of 4 dB. Note the excellent agreement between the two methods.

IV. CONCLUSION

It has been shown that δT_L, the radiometric temperature error due to receiver power nonlinearities, can be determined from the second order power nonlinearity of the receiver output. It has been further shown that this power nonlinearity can be related to the fourth order RF coefficient of the detector and video amplifier in the receiver and the third order intercept of the final RF amplifier in the receiver. Table III summarizes some important formulas developed in this paper. The second column describes the formula and the final column gives the paper equations used to generate the formula shown. Items 1 and 2 give the δT_L contributions for the detector/Video Amp and the RF amplifier in terms of radiometer and component parameters. Item 3 gives the value of A_4/A_2 obtained from the two-tone method for use in Item 1. Finally, Item 4 gives the formula for M_d, the detector figure of merit. M_d must be greater than one for a suitable detector.

TABLE IV
NONLINEARITY SENSITIVITIES OF DETECTOR MEASUREMENT METHODS

Method	$\delta P/P_0$ Sensitivity	δT_L Sensitivity (F = 4 dB)	Comments
Two-Tone	-73 dB	1 5x10^{-6} K	Measured
Amplitude Modulation	-32 dB	0 02 K	Based on -40 dB Modulator Linearity
Constant Ratio	-25 dB	0 1 K	Measured

Table IV summarizes the achieved or projected obtained with the three methods for measuring detecotr power nonlinearities. Two of the methods, the two-tone method and constant ratio method, have been experimentally demonstrated to agree within experimental error. The two-tone method has been found to be very sensitive and is well suited for production level detector linearity measurements.

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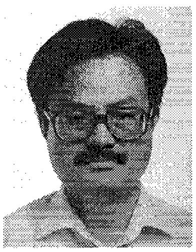

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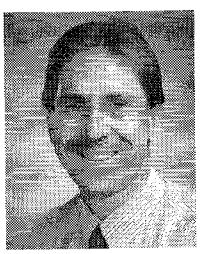

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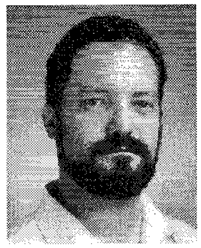

Samuel C. Reynolds received the B.Sc. degree in electronics engineering in 1983 from the Northrop University Inglewood, CA, with graduate work in data communication and digital networking.

He has worked at Hughes Aircraft in El Segundo, CA since 1982. His experience encompasses passive microwave remote sensing, active and passive correlation imaging, active array antennas, data networking, and bandwidth-efficient modulation. He holds four patents for polystatic correlating radar, minimum redundancy array radiometer, simultaneous delay correlator apparatus, and passive microwave near-field tomographic imaging. In addition to his work at Hughes, He develops commercial software and indulges his interests in nonlinear optimization, information accessibility and security, machine intelligence, and the development of a spacefaring civilization.

Mr. Reynolds is a member of Sigma Zeta, Tau Beta Pi, Eta Kappa Nu, and the National Space Society.

Arnold L. Berman (M'61–SM'72) received the B.S. degree in electrical engineering from the Massachusetts Institute of Technology, in 1957. He received the M.S and D.Sc. degrees in electrical engineering from George Washington University in 1971 and 1981, respectively. He also took the post graduate management training program from Harvard University in 1982.

He joined Hughes as chief scientist in 1985. During the last eight years he has been responsible for the definition and analysis of communication payloads for both commercial and classified applications. He has also provided consultation as required to members of the Communication Laboratories on analytic and experimental problems in the areas of communications theory, UHF, and microwave and communications circuit design. One of his recent responsibilities has been the development of highly efficient linear SSPA's for communications applications at both C-Band (10–40 W) and L-Band (20–80 W). Prior to joining Hughes Aircraft, he was the Director of the Microwave Laboratory at Communications Satellite Corporation (Comsat). In this role he was responsible for the direction and supervision of 50 professionals conducting research and development in the areas of satellite born transponders, earth station receiving and transmitting equipment, antenna and feed network design, microwave circuit development, and propagation and optical systems studies.

Dr. Berman is the author of numerous technical papers covering topics such as communications system modeling, on-board signal processing, network topologies for enhancing reliability, intermodulation distortion and nonlinear phase shift in traveling wave tubes, and design of linearized microwave amplifiers. He was awarded the Hyland Patent Award in 1994 for outstanding achievement in the area of SSPA and linearizer development.