Far-Field Reflectometry for Characterization of Active Antennas

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Abstract—Testing active antennas or polarimetric radar calibrators in the far field (plane wave excitation) is complicated by imperfect isolation between transmit and receive channels. We describe a far-field reflectometry setup which overcomes this problem. A large dynamic range is achieved by modulating the active element bias and using a quadrature receiver and lock-in detection scheme. A calibration method and a novel mixer phase-error correction scheme are described. The system is capable of full polarimetric characterization of active scatterers.

I. INTRODUCTION

TO OVERCOME the problems of limited power-handling capacity of millimeter-wave semiconductor devices and lossy waveguiding structures, active antenna arrays have been proposed. Such arrays can combine and focus output power from a large number of devices into a propagating beam in free space, with little loss of energy. This approach is referred to as quasi-optical power combining [1], since optical techniques of Gaussian-beam manipulation can be used to guide the energy once generated. Recent research efforts with such arrays have focused on beam amplifiers, and several different topologies for array cells and grid amplifiers have been reported [2]–[6]. The amplifier cells generally consist of two antennas, coupling energy to and from a high-frequency amplifier circuit, with input–output isolation achieved through the use of orthogonal polarization of the antennas. While conceptually simple, it is challenging in practice to develop a wideband planar antenna on semiconductor substrates (high dielectric constant) which has an acceptable radiation efficiency and can be easily integrated with the active circuit [7].

Testing these active antennas and arrays also presents a challenge. Ideally, the device could be tested in a waveguide where accurate measurements of input and output power versus frequency can be made. However, the waveguide structure will strongly perturb the electrical characteristics (driving-point impedance and directivity) of the antennas, and hence would corrupt the measurement. An alternative is to use a quasi-optical waveguide [8] where the device under test is placed at the beam-waist so that the excitation wavefront is essentially planar and uniform, similar to far-field excitation. A disadvantage of this approach is that it requires some expensive hardware and is restricted to certain field polarizations. A simpler far-field technique has recently been described which is suitable for transmission amplifiers, and is depicted in Fig. 1(a) [2], [3]. This technique uses two standard waveguide horn antennas to measure the power transmission through the amplifier cell, with a simple calibration that relies on the Friis transmission equation [9]. Isolation between transmitting and receiving channels is easily achieved by blocking the space around the amplifier with absorbing material.

Reflection amplifiers are also attractive, since the backside of the substrate is then available for heatsinking and peripheral circuitry. They can also be used for radar calibrators [10]. However, the measurement of these amplifiers is more complicated than the transmission measurement due to imperfect isolation between the transmitting and receiving antennas. As shown in Fig. 1(b), the transmitting and receiving horns are in close proximity in a reflection measurement, and the resulting crosstalk can easily swamp out the desired signal. This and other practical hardware issues are addressed using a low-cost system of lock-in amplifiers and commonly available microwave hardware. The resulting technique is suitable for both single elements and large arrays.

II. MEASUREMENT TECHNIQUE AND SYSTEM CALIBRATION

An active antenna cell typical of those under consideration is shown in the inset of Fig. 1(b). Two identical planar antennas with a gain (directivity) of $G_{\text{ant}}$ are coupled to an amplifier circuit with a power gain of $G_{\text{amp}}$. The goal of the measurement is to determine the reflection gain ($G_{\text{amp}}^2G_{\text{ant}}$) versus frequency of the active antenna for different combinations of transmitted and received signal polarizations. A simple system for such measurements is also shown in Fig. 1(b), consisting of a CW transmitter section and a homodyne receiver using a quadrature mixer for detection (essentially a CW radar
system. Since the mixer is operated as a phase detector, quadrature outputs are required to give an accurate measurement of the received signal strength. Separate transmitting and receiving antennas are used to allow for a full characterization of the polarization matrix for the active antenna under test (AUT). Several practical difficulties are encountered in such systems. When used as a phase detector (Fig. 2), most mixers generate a small dc offset voltage at the IF output. The mixer usually operates in the linear region where the LO signal is much greater than the RF signal ($B \gg A$ in Fig. 2), in which case the dc offset voltage is primarily a function of the LO signal strength. It can also depend on frequency. The proportionality constant between the IF output and the RF input $k$ also depends on frequency and LO bias. Characterizing these parameters is difficult. In addition, the two quadrature outputs are never exactly 90° out of phase, and they will also have some amplitude imbalance (the proportionality constants will be different on the two channels).

Another problem involves imperfect isolation between the transmitting and receiving antennas. Most of our experiments focused on polarization-rotating amplifier cells [5], [6], and hence, the antennas were usually kept orthogonally polarized to each other. Ideally, this arrangement should maintain perfect isolation between the transmitting and receiving channels (with no AUT present). A typical measurement of the transmission coefficient between two orthogonally polarized waveguide horns (C-band) on an HP8720 vector network analyzer indicated in an isolation of 50 dB over the band of interest. While this is a fairly good isolation, it is not good enough; the resulting signal strength appearing at the receiver from this leakage path is still at least two orders of magnitude greater than the signal to be measured for a typical amplifier cell (this can be shown using the Friis transmission equation [9]).

All of these problems can be solved using the measurement setup shown in Fig. 3(a) and a mixer correction algorithm discussed later. The transmitter/receiver components are essentially the same as shown in Fig. 1(b). To distinguish the desired signal from the crosstalk between antennas, the active antenna is bias-modulated (switched on and off) at a periodic rate set by a function generator (which is buffered by a simple relay circuit). Dual-phase lock-in amplifiers are then used to sample the mixer IF outputs, and should record only the modulated signal arriving at the receiver from the AUT. A typical waveform expected at the mixer IF output is shown in Fig. 4, where $f_m$ is the modulation frequency. Letting $V_{on}$ and $V_{off}$ denote the IF voltages measured when the device is on and off, then with reference to Fig. 3(a) we have

$$V_{on} = V_{dc} + k V_A \cos \Phi_a + k V_B \cos \Phi_b$$
$$V_{off} = V_{dc} + k V_A \cos \Phi_a$$

(1)

where $\Phi_a$ and $\Phi_b$ are the phases of the received crosstalk and desired signals with respect to the transmitted signal, respectively. We have assumed that the leakage through the amplifier when unbiased is negligible compared to the transmission when fully biased. This is a safe assumption when the amplifiers have at least 15–20 dB dynamic range in transmission from unbiased to biased states. Note the dc offset voltage does not change appreciably since it is primarily a function of the LO drive power. The lock-in amplifier will record a signal proportional to the difference,

$$\Delta V = V_{on} - V_{off} = k V_B \cos \Phi_b.$$  

(2)

Hence both the unwanted crosstalk signal and the dc offset are eliminated. A wide dynamic range can be obtained using this technique by decreasing the lock-in measurement bandwidth (increased time constant), at the expense of a somewhat slower measurement. Note also for a measurement of an amplifier array, one element of the array could be modulated by itself to explore the individual contributions of the array elements.

Once the probability constant $k$ is known, the amplitude and phase of the received signal are determined by making use of both quadrature IF outputs, which give

$$\Delta V_i = k V_B \cos \Phi_b$$
$$\Delta V_Q = k V_B \sin \Phi_b.$$  

(3)
These expressions assume a perfect amplitude and phase balance between the IF outputs, which we will assume for the remainder of this section.

The next step is to determine the relationship between the actual receiving power and the signal detected by the lock-in amplifier. This can be done with a simple calibration step illustrated in Fig. 3(b). The AUT is replaced by a standard-gain horn with the correct polarization. A measurable amount of power is coupled from the sweeper into the standard-gain horn, and modulated by a PIN diode-switching circuit at the same modulation frequency \( f_m \). Using similar arguments as above, the signal measured by the lock-in amplifier will be given by

\[
\Delta V = kV_C \cos \Phi_c
\]  

(4)

where \( V_C \) and \( \Phi_c \) are the amplitude and phase of the known calibration signal. Note that the transmitting horn still sends out a signal during calibration in order to maintain similar conditions to the actual measurement. This insures that the proportionality constant \( k \) will be the same in both (3) and (4).

Using these two measurements, the gain of the amplifier can be determined as follows. Assuming \( P_{cal} \) is the measured calibration power fed into the standard-gain horn, then the Friis transmission equation [9] gives the total received power as

\[
P_{rc} = P_{cal} \left( \frac{\lambda}{4\pi R} \right)^2 G_r G_{cal}
\]

(5)

where \( R \) is the distance between the two antennas, \( \lambda \) is the wavelength, \( G_r \) is the gain of the receiving horn antenna, and \( G_{cal} \) is the gain of the standard-gain horn used in the calibration. The voltage recorded by the lock-in amplifiers will be proportional to this received power

\[
V_C \propto \sqrt{P_{rc}}
\]

For the actual measurement of the AUT, the Friis formula gives a received power of

\[
P_{ra} = P_{tran} \left( \frac{\lambda}{4\pi R} \right)^4 G_r G_t G_{amp}^2 G_{cal}
\]

(6)

where \( P_{tran} \) is the power delivered to the transmitting antenna (easily measurable), \( G_t \) is the gain of the transmitting antenna, and again, the lock-in amplifier records a voltage proportional to this received power

\[
V_B \propto \sqrt{P_{ra}}
\]

These equations can be combined to give the parameters of the AUT in terms of the measured parameters

\[
G_{amp}^2 G_{cal} = \left( \frac{V_B}{V_C} \right)^2 \frac{P_{cal}}{P_{tran}} \left( \frac{4\pi R}{\lambda} \right)^2 \frac{G_{cal}}{G_t}
\]

(7)

Unless the gain of the antennas on the AUT is known \textit{a priori}, it is impossible to determine separately the parameters on the l.h.s. of (7). This collection of terms can be called the effective isotropic power gain (EIPG) of the active antenna cell. In practice, the horn antennas used in the setup of Fig. 3 are usually identical, so that \( G_r = G_t = G_{cal} \), in which case these gain parameters cancel out and it is not necessary to know them accurately. For best performance, the relative power level \( P_{cal}/P_{tran} \) should be chosen so that the received power is approximately the same for both the calibration and measurement (\( P_{rc} \approx P_{ra} \)), at least in the region of peak AUT gain; that is the function of the attenuator in Fig. 3(b). This insures that the mixer characteristics are identical in both cases.

III. MIXER ERROR CORRECTION ALGORITHM

Typical I and Q data (\( \Delta V_I \) and \( \Delta V_Q \)) measured versus frequency for a narrowband active antenna are shown in Fig. 5(a). The detected signal is a rapidly varying periodic function of frequency due to the far-field separation \( R \); that is, the phase \( \Phi_b \) is given by \( \Phi_b = 2\pi R/\lambda + \psi_{amp}(f) \), where \( \lambda \) is the wavelength and \( \psi_{amp} \) is the amplifier phase response. From this data, it is necessary to calculate the amplitude of the signal \( V_B \) to insert in the expression for gain (7). If the I and Q channels are in perfect phase quadrature and have perfect amplitude balance, then the magnitude of the signal is given by

\[
kV_B = \sqrt{\Delta V_I^2 + \Delta V_Q^2}
\]

(8)

When this expression is used with the raw data, the resulting curve is corrupted by a periodic signal with one-half the period of the I and Q data, as shown in Fig. 5(a). These ripples in the data imply that the phase shift between I and Q channels is not exactly 90°, and/or there is an amplitude imbalance on the IF channels. Under certain conditions, both problems can be solved [11]; we will focus on just the phase error here. Denoting the measured data as \( \Delta V_I \) and \( \Delta V_Q \), we have

\[
\Delta V_I = kV_B(f) \cos [\Phi_b(f) - \Delta \phi(f)]
\]

\[
\Delta V_Q = kV_B(f) \sin [\Phi_b(f) + \Delta \phi(f)]
\]

(9)

where \( 2\Delta \phi(f) \) is the frequency-dependent relative phase error between the two IF channels. The phase error can not be recovered directly from the received signals \( \Delta V_I \) and \( \Delta V_Q \), because they are real values. However, if the data can be expressed in the corresponding complex notation

\[
\Delta V_I' = \frac{1}{2} kV_B \exp [j(\Phi_b - \Delta \phi)]
\]

\[
\Delta V_Q' = -\frac{1}{2} kV_B \exp [j(\Phi_b + \Delta \phi)]
\]

(10)

then the phase error for each frequency can be derived from the ratio

\[
\exp (j2\Delta \phi) = \frac{\Delta V_Q'}{\Delta V_I'}
\]

(11)

Once the phase error is estimated, the data can be corrected by a simple linear transformation [11]. Using the notation of
Fig. 5. (a) Measured I and Q data on a prototype active antenna, and the magnitude calculated from the raw data. The noticeable ripples result from a quadrature phase error. (b) Corrected data and corresponding magnitude after application of the mixer correction algorithm.

For multiple measurements, a single calibration measurement can then be applied to all subsequent measurements to correct the I and Q data.

IV. EXPERIMENTAL RESULTS

The measurement setup was implemented as shown in Fig. 3 using Stanford Research SRS530 dual-phase lock-in amplifiers, a Hewlett-Packard HP8350 sweeper, and an HP3312A function generator. The remaining equipment (mostly RF electronics) was built from scratch, and was designed to operate in the range of 3-5 GHz. This frequency range was chosen in order to simplify the fabrication of the active antenna prototypes and also because of availability of equipment in this frequency range. In the transmit-receive module, a 10 dB amplifier was added to boost the LO signal power, along with a 20 dB amplifier at the receive input to increase the signal-to-noise ratio. The measurements were made in a small anechoic chamber which was lined with 10 cm (4) pyramidal absorber. The measurement was completely automated using a personal computer with a National Instruments GPIB card. Typically, a 100 Hz square wave was used to modulate the active antenna bias and as the lock-in reference.

Two different active antennas were designed and tested. The first used microstrip patch antennas [5] on a Rogers Duroid 6810 substrate (thickness 25 mil, dielectric constant of 10.8). This substrate material was chosen to closely approximate that of a GaAs wafer, on which future monolithic arrays will be constructed. The microstrip patch is a resonant antenna and hence narrowband. The patches were coupled to a resistive-
feedback microwave amplifier, which was designed to have a power gain of approximately 8 dB at 4 GHz, and input-output impedances to match the radiation resistance of the antennas as closely as possible. A photograph of this amplifier cell is shown in Fig. 6(a). Our first attempt used copolarized patches, but because of the strong mutual coupling between input and output the circuit was found to oscillate. Subsequent versions used orthogonally polarized input and output antennas.

The measured frequency response is shown in Fig. 6(b), where a patch antenna gain of $G_{ant} = 5$ dB has been assumed. The measurements give an amplifier gain of approximately 7 dB with a 1% bandwidth. For comparison, the expected frequency response, due to the patch antennas alone, is also shown (shifted upwards by 7 dB to account for the amplifier gain). This path response was measured on an HP8720 network analyzer, and shows good correlation with our measurement technique. The erratic behavior of the gain curve near the ends of the frequency range is due in part to a low signal-to-noise ratio at these extremes, and also a result of the mixer error correction algorithm which is not as effective at the extreme data points. The reverse gain of the active antenna is also shown, which was measured by rotating the AUT by 90° (or equivalently, rotating the transmitting and receiving horns). This measurement indicates that the device is polarization sensitive and unilateral, as expected. Because of the limited bandwidth of the patch antenna cell, another cell was designed using folded-slot antennas [6] and the same resistive feedback amplifier, which is shown in Fig. 7(a). Similar measurements were performed and the resulting EIPG frequency response is shown in Fig. 7(b), which gives a peak EIPG of 21 dB and a 10% bandwidth. As expected, the bandwidth is significantly better than that of the patch antennas. Since little is known about the folded slot on thin-substrate materials, only a very crude guess can be made regarding the antenna directivity. Some elementary arguments give $G_{ant} \approx 7.7$ dB, which gives an amplifier gain of 5.6 dB. Future quasioptical amplifier arrays will focus on using these cells for improved bandwidth.

V. CONCLUSIONS

An inexpensive measurement technique has been described for characterizing active antenna reflection amplifiers or radar calibrators, and has been tested using two simple amplifier cells and verified by comparison with a network analyzer measurement. Practical difficulties that are typically encountered in the receiver electronics and antenna isolation have been solved using a simple lock-in detection scheme and a new mixer correction algorithm. In the future, a wideband version of this measurement technique could be built, which would be limited only by the transmit/receive antennas and the quadrature mixer components. The measurement technique will also be used in future bistatic measurements in which the angular dependence of the reflection amplifier response can be tested.

REFERENCES

Huan-Shang Tsai (S'93) was born in Taipei, Taiwan, in 1968. He received the B.S. degree in mechanical engineering from National Taiwan University in 1990, and the M.S. degree in electrical engineering from the University of California, Santa Barbara, in 1992, where he is currently pursuing the Ph.D. degree in electrical engineering. He has been involved in a project to develop quasioptical amplifier arrays for millimeter-wave power-combining applications. He is also interested in using the numerical method, FDTD technique, to find out the characteristics of different circuit configurations, especially planar antenna structures for quasioptical arrays.

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