Probe-Tone S-Parameter Measurements

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***Abstract*—The measurement of device behavior under complex actual operating conditions is an increasingly important measurement problem. In particular, it can be difficult to accurately measure gain and some reflection coefficients of a power amplifier operating under a realistic modulated signal drive. A small-signal measurement alone of a power amplifier is generally incorrect since the device-under-test will not be biased correctly. A fully modulated measurement, however, may require very dedicated equipment, long measurement times for adequate stability, and special calibration techniques. The methodology discussed here, i.e., the use of S-parameter probe signals in addition to power signals, will allow self-consistent S-parameter measurement under these conditions with full (traceable) vector calibrations and reduced uncertainties. In some sense, the small probe signal is used to quantify nonlinearities introduced by the modulated signal. In addition, the measurement has the flexibility to perform frequency-domain profiling to elucidate the behavior that may be experienced by interfering signals.**

***Index Terms*—Modulated measurements, power amplifier measurements, S-parameters.**

I. Introduction

IT HAS LONG been known that it is sometimes advantageous to make S-parameter measurements in the presence of other signals (e.g., [1] and [2]). Perhaps the most well known of these measurements is "hot $S22$" (e.g., [2]) in which the device (usually a power amplifier) is driven to its normal operating point by a power signal and a smaller probe signal is used to measure the output reflection coefficient. Typically, the signals are both sinusoids and are offset in frequency by at least several IF bandwidths (IFBWs) to avoid problems in the receiver (although this is not necessary with careful consideration of phasor behavior [1]). While this is not a load–pull measurement, since the port impedances are fixed, it can extract useful information about output reflection behavior and stability if the device would be used in something close to a 50-Ω environment. As such, these techniques are often used to characterize amplifier subassemblies rather than bare devices.

An important point about these earlier measurements is that they maintain vector calibrated uncertainties and traceability. Such advantages, lost in straight power measurements and some other techniques, would ideally be maintained as a generalization is attempted.

In the modern measurement environment with a variety of wide modulation formats and highly optimized power amplifiers, the need for performance data in the natural operating state is greater than ever [3]. The following two generalizations of the hot $S22$ concept may be appropriate.

- Include the use of modulated power drive instead of just sinusoidal large signals since average compression behavior may vary as a function of the statistical distribution of the input signal (e.g., [4]).
- Include the other S-parameters since all results under large-signal sinusoidal drive and under some other type of drive may not be the same (i.e., forward parameters may be affected, as well as reverse parameters).

The previous methods have generally required that the multiple signals be separate in frequency space. The nearly stochastic nature of common modulated signals (CDMA and WCDMA are the most obvious examples, but many others meet the sufficiency condition), however, allows for additional interesting measurements since that separation restriction can often be removed (conditions to be discussed).

To make the measurements of value, the concepts of vector-network analyzer (VNA) calibrations must be reintroduced. As long as the receiver has a large dynamic range and power levels are manipulated correctly, it is possible to show that the standard calibration approaches are valid and a significant improvement in measurements can be made. Indeed dynamic range is somewhat easier to achieve with a narrow-band receiver in that large amounts of the composite signal power are automatically excluded from the receive path.

II. Definition

In defining this class of measurements, we will be intentionally fairly general since the concept has broad applicability. The examples will largely focus on a special case: two-port devices where the composite stimulus is composed of a probe continuous wave (CW) tone plus some modulated signal of significant bandwidth (BW) relative to that of the measurement system.

A probe-tone S-parameter measurement is characterized by the following.

- Any number of ports may be driving and there may be any number of signals driving a given port.
- One of these signals, designated the probe signal, will be the basis of the direct S-parameter measurement. It will generally be CW.
- The probe tone must have only a quasi-linear impact on device-under-test (DUT) performance and a negligible effect (relative to other uncertainties) on the signal statistics (of the composite signal) at the port in question.
- If any signals being combined with the probe tone are also CW, they may not be at the same frequency as the probe tone and generally must be separated by at least a receiver BW. If the other signals are modulated (and sufficiently stochastic), they may coexist in frequency space.

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- The composite power delivered to the receiver must not alter its linearity state (although the DUT state can, of course, be a function of composite power).
- The receiver is sufficiently narrow-band in that, when used with ensemble averaging, the probe tone can be measured while excluding the modulated power.

Now we must define how the measurement is made. Generally, the receiver channels will get a combination of the probe signal and some other signals.

The general S-parameter is defined as an output wave (b_n, transmitted or reflected) ratioed against the incident wave (a_m, i.e., the reference). Practically, the measured quantities are altered by (at least) the proportionality constants (such as a coupling factor) and the frequency response of the receiver. Historically, these waves have been assumed to be single frequency, but that concept will now be generalized. Define S_{nm}

$$S_{nm} = \left[\frac{g(b_n)}{g(a_m)}\right]_{\mathrm{state1}} = \left[\frac{\int b_n(f)\Delta(f-f_p)\,df}{\int a_m(f)\Delta(f-f_p)\,df}\right]_{\mathrm{state1}} \tag{1}$$

where g represents the complete receiver behavior, which is defined by a convolution with Δ. The extent of Δ in frequency space is limited by a narrow IF filter although its behavior is affected by other parts of the receiver. f_p is the probe-tone frequency (to which the receiver is tuned) and $\mathrm{state1}$ describes the power state that device/system is operating under (the dependence on power state is critical). The waves b_n and a_m represent the composite test signal (received from the DUT at port n) and the composite reference signal (generated by the source at port m). If one can assume that b and a do not change over the frequency span defined by Δ (which can practically be as small as a few hertz), then the definition reduces simply to the familiar

$$S_{nm} = \left[\frac{b_n(f_p)}{a_m(f_p)}\right]_{\mathrm{state1}}. \tag{2}$$

If the combined signal consists of the probe signal plus additional CW signals, then the conditions are met since the IF filter will presumably remove all of the other signals ($\Delta \sim 0$ at those frequencies). If the combined signal includes a modulated signal that has no significant energy at the frequency of the probe signal, then the conditions are met for the same reason.

Consider then the case of an added modulated signal having significant energy at the probe-tone frequency. If the signal a_m is changing rapidly over the scale of Δ due to a modulated power signal (i.e., it is a sum of a probe tone p_m and a modulated signal d_m: $a_m = p_m + d_m$), then one must normally consider some additional averaging to make the measurement tractable. Presumably one could keep making Δ narrower (often digitally) and, in some cases, this is the same mathematical process, but we will term it averaging beyond a certain point (say, 1-Hz wide Δ). We will use the usual symbolism of $\langle x \rangle$ to denote the ensemble average of x. This averaging may be sweep to sweep or point to point as long as the sampling process is statistically independent of d_m. If d_m has zero mean, is independent of the sampling process and, hence, can be removed by sufficient ensemble averaging (rate of variation substantial relative to the averaging rate), then

$$\left\langle \int a_m(f)\Delta(f-f_p)\,df \right\rangle \approx p_m(f_p). \tag{3}$$

Note that this does not say that d_m has zero power, just that the complex valued function has zero mean in the frequency range of interest. The "approximately equal" is used to denote a tradeoff between resulting data jitter and the amount of ensemble averaging performed; it converges to an equality in the limit of high averaging. Note that standard IS-95 CDMA and 3GPP (among others) WCDMA waveforms fit this criteria; it is expected that many other standards do as well although their suitability (particularly the burst-defined structures such as global system for mobile (GSM), e.g., [5] and other time division multiple access (TDMA) formats) has not been investigated in this study. If a variant of (3) holds for b_n (a and b will generally be of the same type), define $b_n = q_n + e_n$ and we can write

$$\langle S_{nm} \rangle \approx \left[\frac{q_n(f_p)}{p_m(f_p)}\right]_{\mathrm{state1}}. \tag{4}$$

Since p and q (probe-tone values) are defined to be small enough to not affect DUT state (quasi-linearity assumption), then (4) still represents a small-signal quantity, although it is restricted to a defined operating state of the DUT. This allows full S-parameter analysis to proceed and is a key point. Such analysis is conceptually more difficult with the more traditional large-signal S-parameters (in which the probe tone is the only signal and is large) since any quasi-linearity claims would be difficult to justify.

The practical uncertainties introduced by this process are primarily limited to the residual quasi-random nature of a and b after a finite amount of ensemble averaging. This component of uncertainty will generally be ignored in the remainder of this paper; it is assumed sufficient ensemble averaging is performed relative to the base measurement uncertainty (e.g., [6]). Practical uncertainties may be elevated compared to traditional measurements if the probe-tone powers must be very low to meet the statistical insignificance criteria: such a state will reduce signal-to-noise ratios in the receiver.

Standard VNA 12-term calibrations (e.g., [7]) will be defined as occurring at f_p and the actual calibration steps can be performed with only the probe tone present. As such, the standard calibration-related uncertainty terms (corrected port match, directivity, and tracking) will be unaffected. This measurement is not specific to a given calibration technique (short-open-load-thru, variations on thru-reflect-line [8], etc.).

When applying the calibration, the coefficients still apply as long as (4) holds and one assumes the summed signal does not generate significant nonlinearities in the receiver. Subject to receiver compression levels and sufficient ensemble averaging, there should be no increase in calibrated uncertainties over that in the base measurement. As stated previously, this compression risk is perhaps not as great as might be imagined since the bulk of the incident power is outside the BW of the receiver function Δ and, hence, will not contribute as much to compression

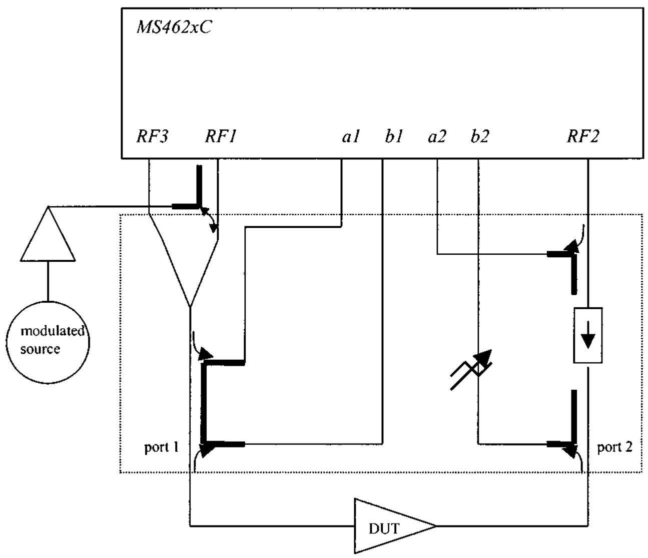

Fig. 1. VNA together with a power-amplifier test set configured for probe-tone measurements. The modulated signal, possibly amplified, is injected prior to the reference coupler for $a1$ and always drives the DUT input.

as in a wide-band receiver. Note that it is the *S/N* established by the probe tone in the receiver that is the relevant component to be used in the uncertainty computations. The key point is that since the standard calibration procedures, the uncertainty calculations, and the small-signal measurement characteristics all hold, measurement integrity and traceability should be maintained.

As an example uncertainty computation, consider an Anritsu MS4623C VNA as the base engine and a cal kit such as the 3750LF is employed. In order to accommodate the probe-tone power requirements (to be discussed further in Section III), a source power of -15 dBm may be employed for a target DUT that requires 0 dBm to be biased into the desired state. Assuming classical coupling coefficients of approximately 20 dB, one can compute an uncertainty of approximately 0.03–0.05 dB for $|S21|$ above approximately -40 dB (ignoring compression). Since the target DUT is most likely a power amplifier, this constraint is of no relevance. The main effect of lowering the port power is to bring up the $|S21|$ level at which uncertainty starts to degrade.

III. Conditions

An important consideration is how these measurements will be set up. For the purpose of these discussions, a VNA with a power-amplifier test set will be primarily considered (see Fig. 1), although variants will be discussed. The upper portion of the diagram is the VNA itself with accessible ports for the reference ($a1$ and $a2$) and test ($b1$ and $b2$) signals, as well as RF drive ports ($\mathrm{RF1}$, $\mathrm{RF2}$, and $\mathrm{RF3}$). The modulated power, possible amplified, is coupled onto the $\mathrm{RF1}$ drive path. While a directional coupler is used here for the coupling, the actual choice will be dictated by the amount of reverse power the VNA can tolerate and by the drive requirements of the DUT (a Wilkinson-class combiner is also often used, e.g., [9]). If the isolation of the coupling device is poor, enough energy may be returned to the VNA port to disturb the function of its automatic leveling circuitry. This is balanced against the amount of modulated power required to bias the DUT into its desired operating state.

Within the test set, $\mathrm{RF1}$ and $\mathrm{RF3}$ paths are then combined although that structure is not needed for the measurements described here (could be used for multitone probe-tone measurements). The hybrid signal then passes through reference and test couplers before reaching the DUT input. The DUT output signal, fed to port 2, encounters the test coupler (with a variable attenuator in the coupled arm to help prevent receiver compression), an isolator (to protect the VNA from large forward powers), and the reference coupler for reverse measurements.

An important note on this block diagram is that the signal combining is done prior to the reference couplers. This implies that the modulated signal will be present in both the numerator and denominator of the S-parameter calculation. An important advantage of this is that the stochastic nature of the data will at least partially cancel (one of the general benefits of the ratioed measurement although it is usually used to reduce the effects of thermal noise, not noise-like modulation), thus reducing the amount of averaging required. The degree of cancellation will be dependent on the coherence between the reference and test channels and, hence, on the path-length differences between these two channels. In any conceived configuration (there are, of course, pathological exceptions), this path-length difference is less than 10 ns. With current bit times of at least 100 ns (for 10 Mb/s) in personal communications systems, one can normally expect a high degree of coherence and, hence, cancellation. If higher bit rates are required and coherence falls, equalizing line lengths can be used to decrease path-length difference.

The IFBW of the receiver will normally be set to a low value (to keep Δ sufficiently narrow) although, depending on receiver architecture, it can be widened to improve speed. The tradeoff is that the amount of required ensemble averaging will then increase and it depends on which process (IFBW reduction or increased ensemble averaging) is faster if there will be a net speed improvement. The other requirement is that the sampling and averaging process must be statistically independent from the modulated signal. Normally this is not an issue when uncommon clocks are used, but it can interfere with the measurement in unusual cases. Smoothing, or box-car averaging (e.g., [10]), can be an acceptable substitute for ensemble averaging if the macro DUT response is not changing in frequency over the range of interest. If it is changing, then the frequency resolution will decrease.

As stated earlier, a critical point in selecting power levels is that the behavior in the presence of the probe must not change in the sense of S-parameters, which are fairly macro-level measurements of a DUT's performance. The criteria would be much tighter if, for example, the measurement was of bit error rate or error vector magnitude. The criteria for S-parameter measurements can be described as follows.

- The average power must not substantively increase. This is important so that the device is not compressed further and that its bias point does not change. There are many possible thresholds, but we have adopted a 0.2-dB criteria since it is primarily S-parameters that we are studying here and not more hypersensitive parameters. This is also

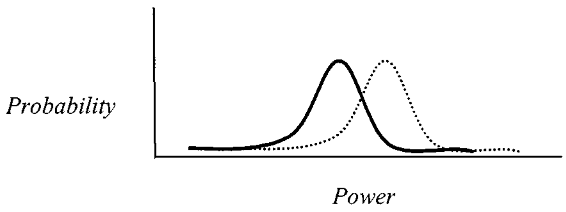

Fig. 2. Power distribution with (dotted line) and without (solid line) the probe tone is plotted. The probe signal does not alter the statistics, except that it moves the mean at that frequency.

smaller than the typical measurement uncertainty of average power.

- The composite statistics of the waveform must not substantively change. Again, the intent is to avoid altering the compression behavior of the device. Since, at a given frequency, constant power is just being added, the statistics themselves are not changing; just the mean is shifting, as illustrated in Fig. 2. Thus, the peak power and the amount of time the device spends above a certain level is not changing, except that level is shifted higher. For this argument, we resort to the conditions above that if this shift is small enough, it will not affect the parameters being analyzed. In an older interpretation, one may want to make sure that the peak to average ratio does not change substantively. It is easy to show that if the average power has changed less than x dB for a probe tone of P_Δ, then the peak to average ratio will have changed less than x dB as well (asymptotically approaching for high initial peak to average ratios).

Let P_a be the initial average power, P_Δ be the probe-tone power, ε (> 1) be the allowed ratio of new average power to old average power. By definition

$$\frac{P_a + P_\Delta}{P_a} = \varepsilon. \tag{5}$$

Let C be the old peak to average ratio (crest factor), the new crest factor is then

$$\frac{CP_a + P_\Delta}{P_a + P_\Delta}. \tag{6}$$

Since $C > 1$ and $P_\Delta > 0$, this ratio will always be smaller than C. Thus, R, the ratio of the old to the new crest factor (to keep $R > 1$ without loss of generality), is given by

$$R = C\frac{P_a + P_\Delta}{CP_a + P_\Delta} = \frac{P_a + P_\Delta}{P_a + \frac{P_\Delta}{C}} = \varepsilon\left(\frac{1}{1 + \frac{P_\Delta}{CP_a}}\right) < \varepsilon. \tag{7}$$

The inequality approaches an equality as C and/or P_a becomes large. Thus, if one meets the average power criteria, one automatically meets that same criteria for the crest factor. For a 0.2-dB threshold, this requires that the probe power be at least 13.3 dB below the starting average power. For *S/N* in the receiver and to minimize measurement time, it is generally desirable to get the probe power relatively close to this level.

IV. Measurements

All of the measurements to be discussed will be performed with minor variants of the test setup shown in Fig. 1. The probe tone will always be a sinusoid generated internally by the VNA and the modulated signal, where used, will be an IS-95 CDMA signal (chip rate 1.2288 Mc/s, nine forward channels). The modulated signal was injected with a coupler, although other arrangements are possible. The power levels vary in the different examples, but a 15-dB offset between modulated power and probe-tone power was maintained throughout (with the exception of one example to show the effect of high probe powers). Receiver linearity is discussed only in the first example, but it was verified in all examples.

The first measurements presented are primarily to build confidence in the underpinnings of the methodology discussed thus far. The first example measurement is just $S21$ of a thru line with the probe tone alone and with an IS-95 modulated signal fixed at 1800 MHz. Since the BW of this modulating signal is approximately 1.25 MHz and the sweep range is 2 MHz, both embedded probe and separate probe data will be included. No difference in the received signal was seen whether the modulated signal (-10 dBm) was present or not. The jitter on the data is a little higher than what one would normally expect since the resulting reference power is quite low. This could be improved by adding reference amplification or increasing averaging (although the latter will have an attached speed penalty). Since the amount of jitter is roughly equivalent with and without the modulating signal, one can conclude that the receiver is effectively filtering the modulated energy even in the embedded case (IS-95 easily meets the conditions of being sufficiently stochastic). Since the mean did not move, we can be reasonably sure that the receiver is not being compressed.

The second test is with a small-signal amplifier run at low powers. The same signal grouping and powers (-10 dBm modulated, -25 dBm probe) are used here and the level is low enough that the amplifier will not be heavily compressed. As one might expect, the measured $S21$ appeared the same, to within 0.1 dB, whether the modulated signal was present or not (with the modulated result being from 0 to 0.03 dB lower). The slight suppression in the mean with the modulated signal suggest the onset of compression, but it is still much less than 0.1 dB. Note that for many power amplifiers, this display would be quite different since the gain is a function of drive level over a large range. This particular DUT has flat gain up until its compression behavior starts at approximately -3 dBm input.

The third test is the same amplifier where the input signal power has been increased so that compression is occurring. The same power delta of 15 dB was maintained, but the modulated signal power was now approximately 0 dBm. In this case, the measured $|S21|$ with the modulated signal present is approximately 0.4 dB lower than when just the probe tone is applied. While this should not be surprising, since the amplifier will be compressed, it is a key point of this measurement.

The 0.4 dB of compression from the small-signal behavior was quite clear and the level of jitter (< 0.05 dB) again does not differ much with and without the modulating signal. Another interesting point is that the level of compression is the same whether the modulated energy and probe tone overlap in frequency or not (recall the IS-95 signal has approximately 1.25 MHz of BW in this 2-MHz sweep). This is partially a result of the use of a wide-band amplifier for this test (4 GHz) and the fact the modulated power is globally changing the bias state of the device.

Before proceeding to more concrete measurement comparisons, it is instructive to first examine the effect of using a probe-tone power in excess of what has been recommended. The CW $|S21|$ of an amplifier (in approximately 0.2-dB compression from the applied IS-95 signal) was measured as the probe-tone power was varied from 25 dB below modulated power to just 5 dB below modulated power (which is fixed). The measured $|S21|$ was flat (to within 0.05 dB) until the probe-tone power was approximately 10 dB below modulated. Even at higher levels, the degradation was not severe: the error (relative to low probe-tone power values) is only approximately 0.15 dB when the input is 5 dB below the modulated power. Admittedly this is not a worst-case scenario, but it is demanding since the DUT has entered compression and helps justify the power levels used.

The next examples relate to correlating the probe-tone results to those obtained with more conventional approaches of modulated gain measurements. One common method is to simply use an integrating receiver (power meter or spectrum analyzer with an integrating function). The latter was chosen for this example. A spectrum analyzer was used to measure input and output power (integrating over 2 MHz with a 30-kHz resolution BW and a very large video BW) to an amplifier and this was done as a function of drive level. The same measurement was performed using the probe signal approach maintaining a 15-dB power delta. The same modulated signal was used for both measurements but, of course, the probe tone was not present for the integrated spectrum analyzer measurement. The results span the range of small signal to heavily compressed and, hence, should represent a reasonable cross section of DUT behaviors. The results (see Fig. 3) show good agreement, certainly as good as the uncertainty of the scalar measurement.

Having established a certain level of agreement with scalar results, it is time to look at the vector benefits. The simplest of the class of vector probe-tone measurements is hot $S22$ (see Fig. 4) in which a full one-port calibration is used. In this measurement, an offset sinusoid (one is applied to the DUT input and a probe tone is bounced off the DUT output). As discussed earlier, the important part is that the offset must be several receiver BWs (300 kHz $\gg$ 1 kHz IFBW for Fig. 5) and an isolator/circulator may be required on the port–2 side if the output power is too high. The isolation is needed to prevent the DUT output from affecting the automatic leveling circuitry of the probe-tone source and sometimes to prevent damage to the source (less of an issue with modulated drive, but it must be checked). The vector calibrations provide correction for port matches, directivity, and tracking to reduce the measurement uncertainty in this case to approximately 0.2 dB, considerably better than that of a scalar measurement (uncertainty is typically $>$ 0.5 dB) and more stable.

An example measurement is shown in Fig. 5 (diagram in Fig. 4) at small-signal and modest compression drive levels.

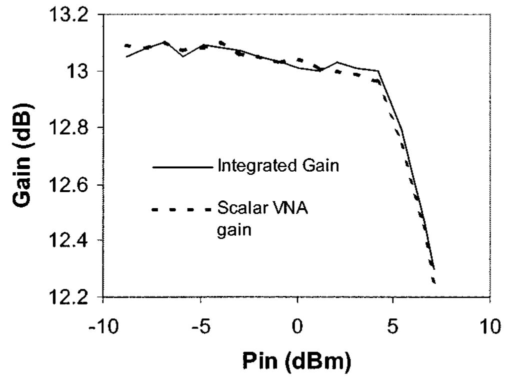

Fig. 3. Insertion-gain measurements using the probe-tone approach and using an integrated power measurement approach (with a spectrum analyzer) are compared. An IS-95 signal was used and agreement was well within uncertainties.

The shift is quite noticeable. It should be noted that this is not a replacement for load–pull measurements in that the load impedance is not changed, but it can be appropriate for amplifier modules that will be embedded in something like a 50-Ω environment.

The important point about this measurement is that it uses full vector correction of the reflection measurement. The uncertainties are much lower ($<$ 0.1 dB) and more stable than what one could obtain with a scalar measurement.

Returning to the modulated signal measurements, the next question is what benefits can vector correction offer. On a gain measurement, for example, the obvious benefit is correcting for raw port matches. In other measurements, directivity and vector tracking errors can also be effectively corrected, which will not happen in a scalar measurement. This issue becomes more important as the DUT approaches compression since its impedance levels will be changing. As an example, scalar and vector probe-tone gain measurements are compared in Fig. 6.

Up to approximately $+4$ dBm input power, the difference is under 0.1 dB and is likely due to source-match–$S11$ interactions possibly together with a better tracking term characterization. More interesting, however, is the effect as power increases. At $+7$ dBm input power (1-dB compression point), the differential is at 0.3 dB and increasing. This is well beyond the measurement uncertainty of approximately 0.07 dB (for the vector corrected data) and can be considered significant from a measurement point-of-view.

One can understand this increasing deviation by looking at DUT match as a function of power level (see Fig. 7). These parameters were measured using a full 12-term probe-tone calibration. While the input match is actually improving with drive, the output match is definitely not. The interaction between measurement system load match and DUT output match will increase at these higher power levels and lead to increased scalar measurement errors.

The final example deals with a more realistic power amplifier (handset variety, 900 MHz). This amplifier exhibits gain expansion for a wide range of input powers, a gain flattening near the desired operating point, and finally, gain compression. As such, a measurement like that in the second test example would require careful interpretation since the DUT gain will be lower

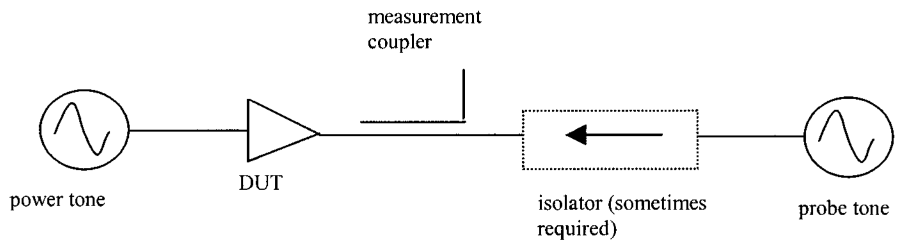

Fig. 4. Simplified setup for hot $S22$ measurements (not using a modulated drive). An isolator is sometimes required to protect the probe-tone source.

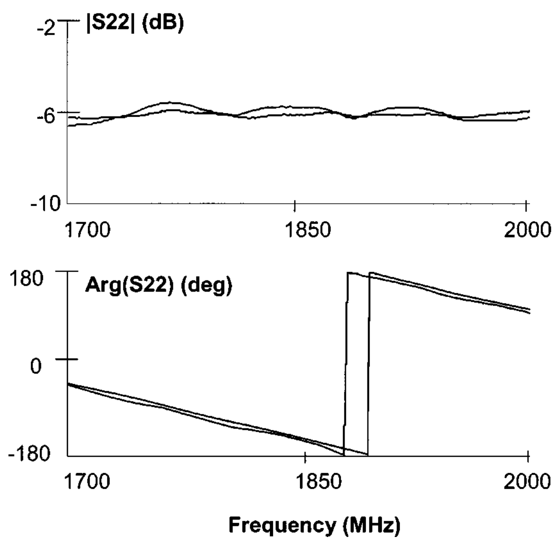

Fig. 5. Hot $S22$ measurement under low and moderately compressive drive levels. This power tone is a sinusoid and is used to illustrate measurement flexibility and the first step in vector calibrations.

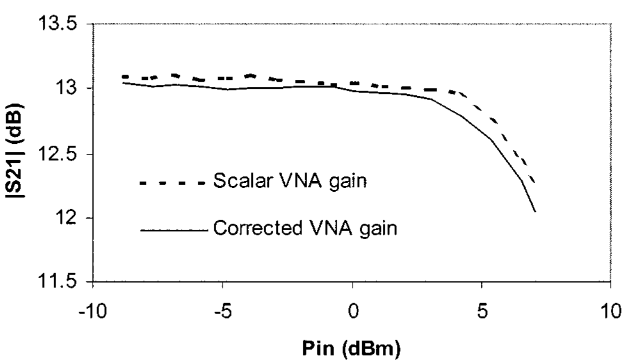

Fig. 6. Comparison of scalar and vector corrected amplified gain measurements. The differences are under 0.1 dB at low-power levels, but increases to 0.3 dB at +7 dBm input power.

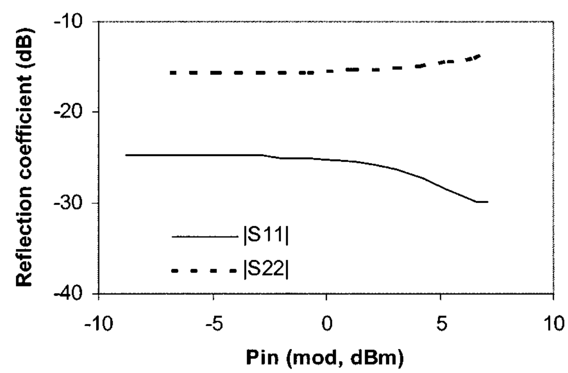

Fig. 7. Input and output match of the amplifier of Fig. 6 (also probe-tone measurements, modulation applied to port 1). At higher power levels, the output match is degrading rapidly, which probably explains the increasing differences within Fig. 6.

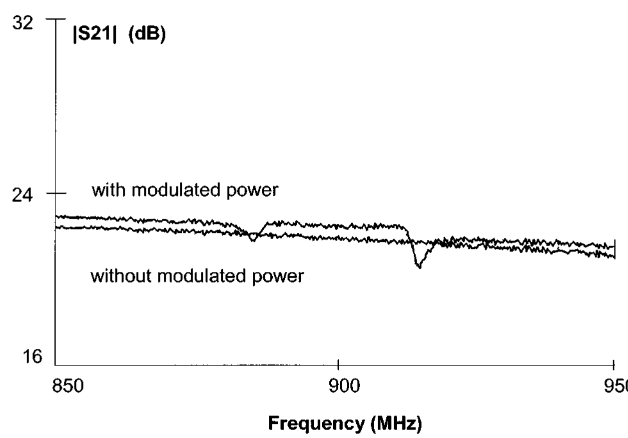

Fig. 8. Gain of a power amplifier with and without modulated drive. The modulated signal is fixed at 900 MHz and is approximately 1.25-MHz wide. Thus, this gain profiling measurement includes both in-band and out-of-band behavior.

with probe tone alone compared to the composite signal. The probe-tone measurement is still quite valid, but the DUT operating point will be strictly established by the modulated signal and must be carefully considered.

Fig. 8 shows a plot of gain with and without modulation power present (again, IS-95 with a 15-dB power delta). In both cases, the amplifier is being operated somewhat below its normal operating point (13-dBm output power versus 27 dBm) so that it is still in gain expansion. The point of this experiment is that it shows some of the power of S-parameter profiling possible with this measurement technique. Since the modulated power is fixed in frequency while the probe tone sweeps, it is possible to see what behavior an interfering signal would experience at a different frequency (as an example). In this case, one can see two gain dropouts at 15 MHz away from the carrier frequency 900 MHz (note that the frequency scale is 10 MHz/div and the modulated power is only $\sim$ 1.25-MHz wide). This type of S-parameter profiling could conceivably be useful in studying and/or tailoring out-of-band and band-edge responses.

While Fig. 8 illustrates the difference with and without modulated power applied, Fig. 9 illustrates the group delay (to be interpreted as deviation from linear phase) and magnitude of the same amplifier under slightly higher drive levels (+17 dBm

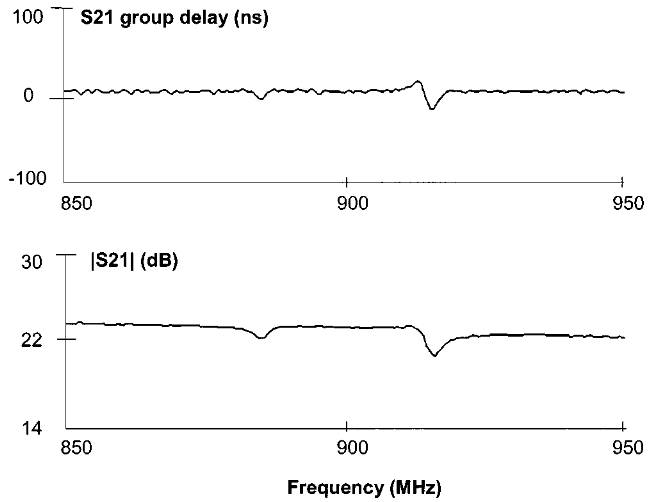

Fig. 9. $S21$ group delay and magnitude for the amplifier of Fig. 8 (under slightly higher drive levels). The $\sim$ 20 ns group-delay anomaly can be easily identified. The modulated power is fixed at 900 MHz and is from an IS-95 source.

output power). The out-of-band group-delay deviation (of approximately 20 ns) can be seen to coincide with the magnitude variation, as might be expected. This type of data is included to illustrate additional profiling power, as well as the group-delay concept in this measurement type. Without the need to demodulate, it is still straightforward to define a group delay (e.g., [11]) in terms of the pilot tone behavior (using the standard $-d\phi/d\omega$ definition) and the measurement complexity does not increase.

V. Conclusions

We have presented a general S-parameter measurement technique in the presence of other signals with the largest application being in the presence of fairly stochastic modulated signals on power devices. The measurement uses a sinusoidal probe tone (on which the actual measurements are made) combined with another signal or signals. While the concept is not entirely new, it can be shown that with proper signal processing and attention to power budgets, useful measurements can be performed that show important performance parameters while properly handling corrections and traceability.

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