test packages used allowed us to measure device properties over the band 8–18 GHz. The broad-band gain was on the average, 12 dB and reverse attenuation was about 32 dB. Tuning improved the gain by as much as 10 dB, primarily by improving the match between the output circuit and the traveling-wave region. Measured noise figure was about 18 dB, and it was limited by the inability to tune the input and by constraints which the traveling-wave region impose on the input FET section.

The gain of the device has been electronically varied over approximately 35 dB by varying the input gate voltage. The gain in dB was nearly proportional to the square of the gate voltage, and it was limited at the upper end by stability and at the lower end by capacitive feedthrough. The phase shift was nearly proportional to the input-to-output voltage change, and it was limited by the extent of the negative-mobility portion of the velocity-field curve. Future work will concentrate on techniques for improving noise figure and increasing output power capability.

ACKNOWLEDGMENT

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RF Amplifier Design with Large-Signal S-Parameters

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Abstract—**High-power UHF transistors have been characterized through the use of large-signal *S*-parameters. These *S*-parameters have been used successfully to design UHF power amplifiers. Waveform measurements show that due to the *Q* of the package parasitics, most class C operated UHF power transistors have nearly sinusoidal waveforms at their package terminals. Experimental evidence presented shows that the large-signal *S*-parameters are relatively independent of power once the device is turned on. These two observations make it possible to extend modified small-signal *S*-parameter design techniques to large-signal power amplifiers.**

I. INTRODUCTION

THE USE OF *S*-parameters for characterization of small-signal linear microwave devices and circuits is well known [1]. *S*-parameters are a valuable tool in analytic design of linear microwave circuits [2] because of ease of measurement and the convenience of their use. Analytic design procedures, using *S*-parameters, are well developed for small-signal transistor amplifiers. Design of high-power microwave transistor amplifier circuits, however, has been a cut-and-try procedure because of the lack of similar analytic procedures for high-power devices. Characterization of high-power transistors by *S*-parameters would be useful also for predicting the stability of circuits, and would provide a more convenient means of testing and evaluating transistors for high-power applications.

In this paper it is shown that the linear concept of the *S*-parameter can also be used in large-signal transistor amplifier design. Results from the measurement of the large-signal *S*-parameters and the RF saturation voltage of a microwave power transistor are presented. This information is then used to analytically design a 1-GHz 18-W amplifier. The success of this amplifier design partially verifies the applicability of *S*-parameters and linear design techniques to design nonlinear class C amplifiers.

First, the variation of large-signal *S*-parameters with drive level and bias conditions is discussed. It is shown that over a wide range of power, with the exception of S_{21} the *S*-parameters are not a strong function of applied power, and that the voltage waveforms observed when measuring large-signal *S*-parameters are nearly sinusoidal. These results support the hypothesis that the transistor, although operating in class C, can be analyzed, at least approximately, by linear methods.

Next, a method is described for measuring the RF saturation voltage of microwave transistors. The measured *S*-parameters and saturation voltage are then used to design an amplifier, and the performance of the resulting amplifier is compared with that of a cut-and-try optimized amplifier.

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Finally, some of the possibilities and limitations of large-signal S-parameters are discussed.

II. Large-Signal S-Parameter Measurements

Because high-power transistors have low-input impedances, reflection coefficients near unity result when measurements are made in a 50-Ω system. As a result, high measurement accuracy is required to characterize these transistors. A high-power semiautomatic S-parameter test set capable of making these measurements with 1-percent accuracy has been described elsewhere [3]. The test set incorporates a low-pass filter which, when switched into the circuit, eliminates harmonics before the signal is measured.

Fig. 1 shows emitter and collector voltage waveforms of a transistor operating in a 50-Ω system under class C bias conditions. All measurements in this paper refer to the TRW MRA 0610–20 transistor. This is a UHF common base transistor rated at 9.5-dB gain and 20-W output. Measurements described as class C are those in which the transistor emitter and base are at dc ground and all emitter bias current is provided by rectified RF power. Even though class C operation is presumably highly nonlinear, the waveforms in Fig. 1 are nearly sinusoidal. Apparently, harmonic energy that may be present at the chip does not escape from the transistor because of packaging parasitics.

Throughout this paper, two types of common base bias conditions are used, class C (self-bias) and class B. In class B just enough dc bias is applied to cause a very small amount of collector current (10 mA) to flow before RF power is applied. Under class B bias, neglecting the effect of parasitics, the transistor would be expected to conduct during approximately half of each RF cycle. However, the nearly sinusoidal collector voltage waveforms observed with both class B and class C bias show that if indeed the transistor is cut off during a significant part of a cycle, the harmonic content of the resulting highly distorted waveform is greatly reduced by packaging parasitics.

Fig. 2 shows the measured values of S_{11} and $1/S_{22}^*$ of the transistor at 1 GHz for different drive levels and bias conditions. $1/S_{22}^*$ is plotted because the collector impedance of this common base transistor has a negative real part, leading to a value of $|S_{22}|$ greater than unity that would fall outside the Smith chart. S-parameter measurements were made with and without a 1.5-GHz low-pass filter in the circuit. No significant differences were detected between the S-parameters measured with and without the filter.

The class C S_{11} measurements show that at low powers $|S_{11}|$ is near unity, and nearly all incident power is reflected. As the power level is increased, self-bias turns on the transistor, some power is absorbed from the source, and $|S_{11}|$ is reduced. Under class B bias, the transistor conducts slightly even with low values of incident power and much less RF power is required for a given dc current flow than is required with class C bias. $|S_{11}|$ is reduced because RF power is more readily absorbed from the source.

The type of bias has much less effect upon S_{22}. Since S_{22} is measured with the RF power incident on the collector, appreciable dc current never flows, and the collector impedance is not strongly affected by bias. The value of S_{22} measured in this way is primarily determined by the capacitance of the collector depletion region.

Fig. 3 shows S_{21} of the same transistor, measured with

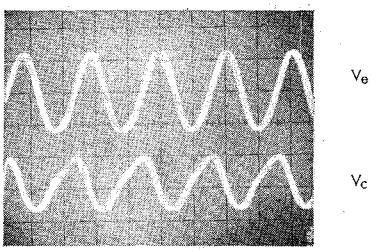

Fig. 1. Emitter and collector voltage waveforms of a TRW-MRA 0610–20 transistor in a 50-Ω line at 1 GHz.

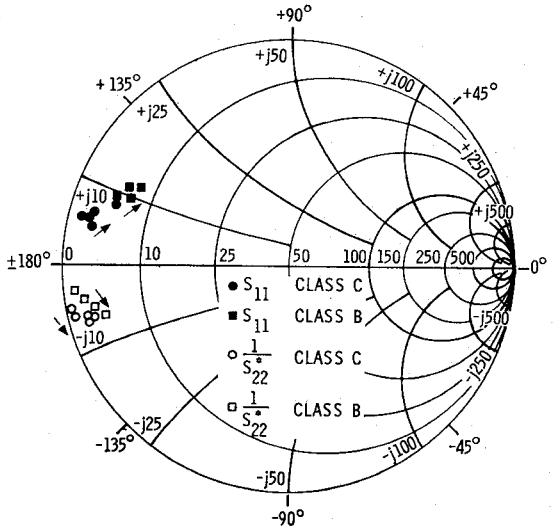

Fig. 2. Large-signal 1-GHz transistor S-parameters at different power levels. The arrows are in the direction of increasing power.

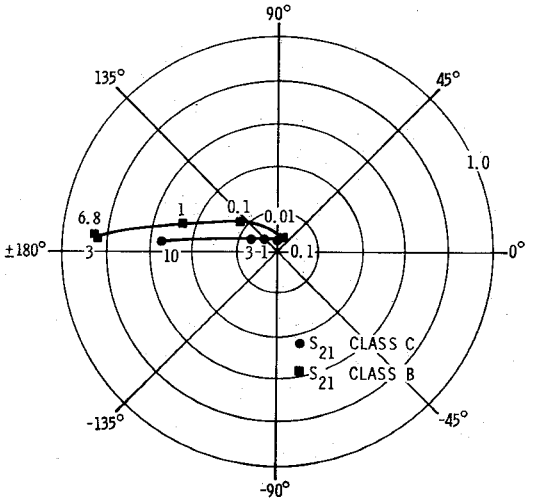

Fig. 3. S_{21} of the transistor in Fig. 2. The numbers next to the data points are power levels in watts.

class C and class B bias. S_{21} is a strong function of incident power. For the same value of power, S_{21} is much smaller with class C bias than with class B bias. This is primarily because the transistor is much more readily driven into its active region when some dc bias is present. For example, less collector current flows with 3-W incident power and class C bias than with 1-W incident power and class B bias. The class B S_{21}

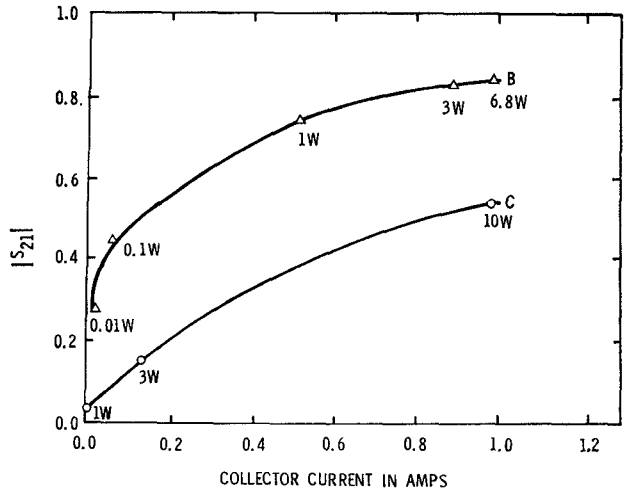

Fig. 4. $|S_{21}|$ versus dc current and incident power for class B and class C bias.

values are also somewhat larger than the class C values when compared at the same levels of average collector current. The reason for this is that more input power is required to turn the device on in the class C case and this increased drive level is sufficient to cause saturation in the output (to be discussed later), which results in reduced gain. Fig. 4 shows how $|S_{21}|$ depends upon collector current and applied power.

The preceding measurements were all made in a 50-Ω test system. Because the impedance of power transistors is well below 50 Ω, their reflection coefficients are high and their unmatched power gain is low. Measurement conditions would more closely approximate the conditions that occur in an actual amplifier if a lower impedance were used. To determine if S-parameter measurements would be more accurate or more easily taken in a low impedance system, quarter-wave transformers were used to reduce the impedance of the test set to 4 Ω.

If the S-parameters are meaningful, impedances determined through the use of 4-Ω S-parameters should agree with those obtained from 50-Ω parameters. In Fig. 5, S_{11} and $1/S_{22}^*$ are plotted as the average collector current is varied (by changing the RF drive) from 0 to 1 A. In the 50-Ω system, the class C S_{11} of this transistor is nearly independent of collector current, while S_{11} in the 4-Ω system is smaller and varies somewhat more. The input impedances calculated from the two sets of S-parameters are quite similar. Input impedance of this transistor in a 50-Ω system, operated class C with 1 A of collector current is $2.5+j6$ Ω. The same transistor in a 4-Ω system has an input impedance of $2.75+j6.5$. The effective input impedance of a device, expressed in terms of S-parameters is given by

$$S_{11}' = S_{11} + \frac{S_{21}S_{12}\Gamma_L}{1 - S_{22}\Gamma_L} \qquad (1)$$

where S_{11}' is the reflection coefficient corresponding to the input impedance, and Γ_L is the reflection coefficient of the load. The input impedance of the transistor when terminated in a 4-Ω load, as calculated from the 50-Ω data using (1), is $3+j6$ Ω. The difference between this impedance and the impedance measured in a 4-Ω system is not much smaller than the difference between the input impedances in the 50-Ω and 4-Ω systems. The inaccuracy is apparently a result of error in the measurement of S_{12}. The reason for this error will be discussed later.

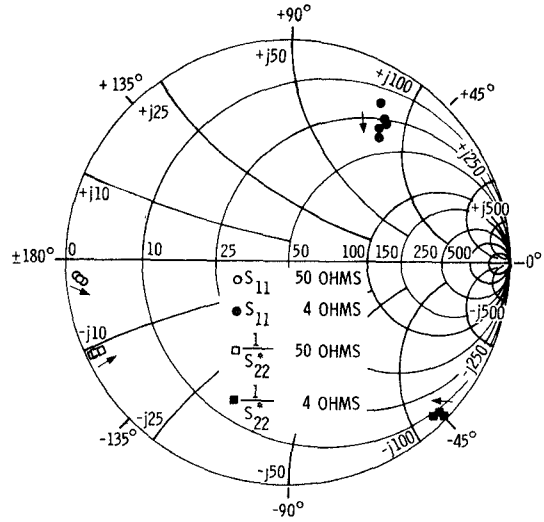

Fig. 5. Comparison of large-signal S-parameters of a transistor measured in a 50-Ω system and in a 4-Ω system. $|S_{11}|$ is smaller in the 4-Ω system and varies more with incident power. The arrows are in the direction of increasing power.

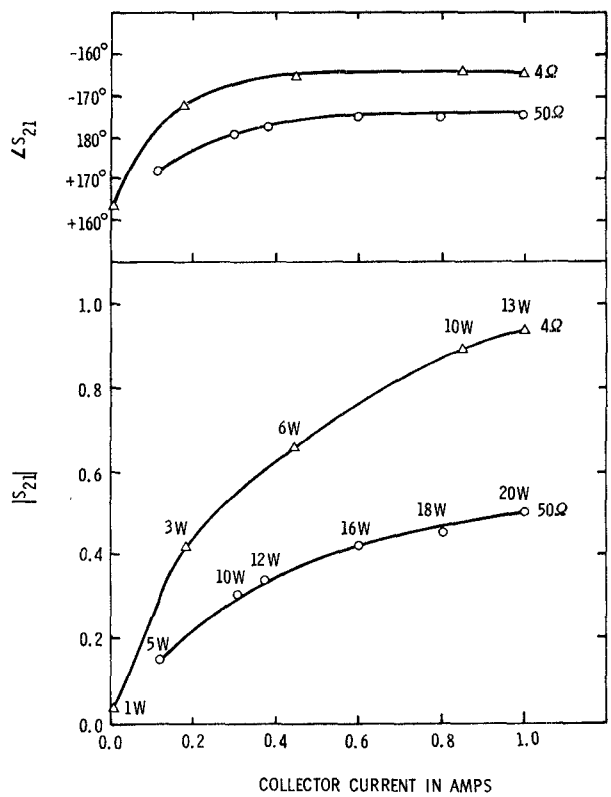

Fig. 6. Magnitude and angle of S_{21} versus average dc current and incident power.

Fig. 6 shows how S_{21} varies with collector current in a 50-Ω circuit and in a 4-Ω circuit. Gain is much higher in the 4-Ω circuit, both because the mismatch loss is much lower, and because the RF drive needed to obtain a given average dc current is less so that for a given current level the signal in the 4-Ω circuit is smaller.

III. Measurement of RF Saturation Voltage

Maximum available power output from a transistor is determined not only by the impedances and gain of the transistor, but also by its RF saturation voltage and the collector supply voltage. To properly design a power amplifier, knowledge of the RF saturation voltage of the transistor is required.

RF saturation voltage can be determined by plotting output power versus collector voltage while holding input power constant. At low collector voltages the output voltage swings from saturation to nearly twice the collector voltage, and output power increases with collector voltage. When the collector supply voltage is sufficiently high, the output voltage swing is insufficient to saturate the transistor. Under this condition, gain is nearly independent of collector voltage. If the collector voltage waveform is reasonably sinusoidal, as was seen to be the case (Fig. 1), then at any point where the collector supply voltage is low enough so the transistor is in saturation

$$P_o = \frac{(V_{cc} - V_{\text{sat}})^2}{2R_{LP}} \tag{2}$$

where P_o is the output power, V_{cc} is the collector supply voltage, R_{LP} is the parallel equivalent load resistance, and V_{sat} is the RF saturation voltage. The validity of (2) does not require linear behavior of the transistor, but only a sinusoidal collector voltage waveform. At any point on the P_o versus V_{cc} curve where P_o is increasing with V_{cc}, then, the RF saturation voltage can be calculated from (2).

In Fig. 7, output power is plotted as a function of collector voltage. The RF saturation voltages shown in Fig. 7 have been calculated from (2). The average collector current was held constant by altering the input power slightly, as collector voltage was varied. This procedure affects the resulting power versus voltage relation only slightly, and causes the point at which saturation occurs to be more sharply defined.

In a 50-Ω measurement system, applied power is considerably larger than the output power. As a result, a small amount of direct coupling through transistor parasitics increases the output power significantly, thereby reducing the apparent saturation voltage. In a 4-Ω system, S_{21} is much larger and this problem is reduced. In the present work, a narrow-band 4-Ω transformer was used, so that some harmonics are terminated in a higher impedance than the fundamental. This accentuates the effect of harmonic energy and may limit the accuracy of saturation voltage measurements made in a narrow-band 4-Ω system. Fig. 7 shows the RF saturation voltage as determined from 50-Ω measurements and from 4-Ω measurements with and without the use of a low-pass filter to eliminate harmonics. The filter made no difference in the 50-Ω measurements. The 4-Ω power versus collector voltage plots were somewhat different depending on the presence of the filter, but led to similar RF saturation voltages of 7.3 and 7.8 V. The 50-Ω measurements led to a value of 5 V, probably somewhat low because of direct feedthrough by transistor parasitics. An RF saturation voltage of 7.5 V is assumed in the analytic power amplifier design procedure described in Section IV.

IV. Power Amplifier Design Using Large-Signal S-Parameters

The design methods applicable to small-signal devices will now be applied to class C power amplifier design.

First, the stability factor K is calculated from the S-parameters.

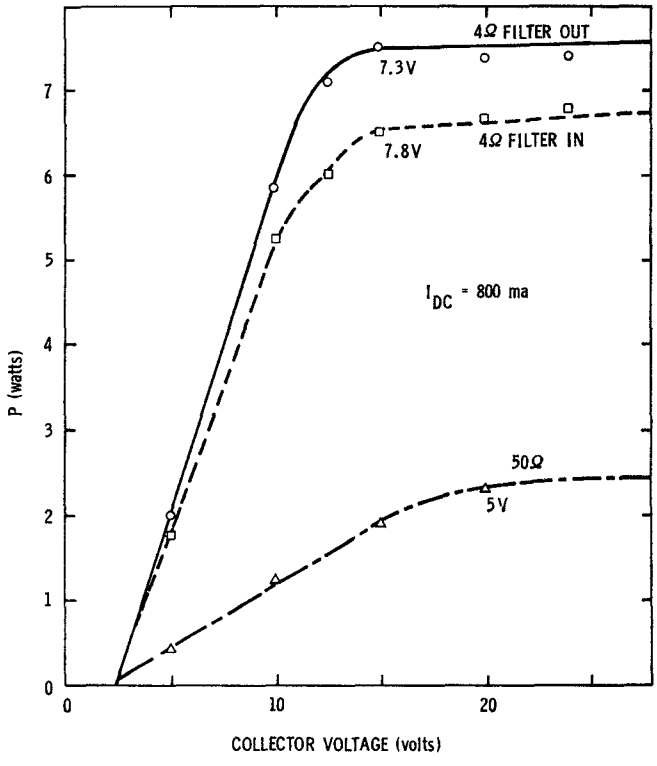

Fig. 7. RF saturation voltage determination from transistor output power and collector supply voltage. The three curves show effective RF saturation voltages of 7.3 V, 7.8 V, and 5 V at 977 MHz and 800 mA.

$$K = \frac{1 + |S_{11}S_{22} - S_{12}S_{21}|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{12}S_{21}|} \cdot \tag{3}$$

If K is greater than 1, the transistor is unconditionally stable and maximum gain can be achieved by conjugately matching the input and output. Usually, however, K is less than 1 and some combinations of source and load impedance will cause the transistor to oscillate. For these transistors maximum available gain is not defined, as in principle any desired gain can be achieved by operating arbitrarily close to oscillation.

The design method when K is less than 1 (conditionally stable transistor) is to first select the desired gain and calculate the locus of all load impedances that will yield the desired gain when the transistor input is conjugately matched. This locus is a circle (the constant gain circle) on a Smith chart. The center and radius of the constant gain circle [2] are given by the following equations.

Center:

$$r_G = \left[\frac{G}{1 + D_2G}\right]C_2^* \tag{4}$$

where

$$G = \frac{G_o}{|S_{21}|^2} \cdot \tag{5}$$

Radius:

$$r' = \frac{(1 - 2K|S_{21}S_{12}|\,G + |S_{12}S_{21}|^2G^2)^{1/2}}{1 + D_2G} \tag{6}$$

where G_o is the desired gain and D_2 and C_2 are defined below.

When the transistor is only conditionally stable, certain

load impedances will cause the transistor to have an input impedance with a negative real part. That is, S_{11}' in (1) may be greater than unity for certain values of Γ_L. Similarly, certain source impedances may cause the transistor to have a negative real output impedance. The source and load impedances for which this occurs are defined [2] by source and load stability circles on the Smith chart.

Input Stability Circle:

$$\begin{array}{cc} \text{center} & \text{radius} \\ \dfrac{C_1^*}{|S_{11}|^2 - |\Delta|^2} & \dfrac{|S_{12}S_{21}|}{|S_{11}|^2 - |\Delta|^2} \end{array}. \tag{7}$$

Output Stability Circle:

$$\begin{array}{cc} \text{center} & \text{radius} \\ \dfrac{C_2^*}{|S_{22}|^2 - |\Delta|^2} & \dfrac{|S_{12}S_{21}|}{|S_{22}|^2 - |\Delta|^2} \end{array} \tag{8}$$

where

$$C_1 = S_{11} - \Delta \cdot S_{22}^*$$
$$C_2 = S_{22} - \Delta \cdot S_{11}^*$$
$$\Delta = S_{11}S_{22} - S_{12}S_{21}$$

and

$$D_2 = |S_{22}|^2 - |\Delta|^2.$$

The stability circles separate the stable regions of the Smith chart from the potentially unstable regions. Whether the inside or the outside of the load stability circle is the region of stability can be determined by calculating the input reflection coefficient corresponding to any load impedance inside the circle using (1). If $|S_{11}'|$ is greater than unity, the inside of the circle is potentially unstable, otherwise the inside is stable. (A similar technique can be applied to the output reflection coefficient.)

In a power amplifier, load impedance affects not only the input impedance, but also the maximum power output.

$$P_{\max} = \frac{(V_{cc} - V_{\text{sat}})^2}{2R_{LP}} \tag{9}$$

where $P_{\max}$ is the maximum output power, V_{sat} is the RF saturation voltage, and R_{LP} is the parallel equivalent load resistance. Given that the transistor output is saturated at the desired power output, the required R_{LP} can be determined from (10). The locus of load impedances having this parallel equivalent resistance is a constant conductance circle on the Smith chart. If a load impedance outside the constant conductance circle is selected, collector saturation will prevent the desired output power from being obtained. If a load impedance inside the circle is selected, the full RF voltage capability of the transistor will not be utilized, and the peak collector current at the desired output power will be larger than necessary.

Fig. 8 shows the gain circle, stability circles, and power circle for a 1-GHz amplifier using an MRA 0610–20 transistor. Power output is 18 W and gain is 9 dB. The intersections of the power circle and the 9-dB gain circle are the two load impedances at which 9-dB gain and 18-W saturated power output should be achieved simultaneously. An amplifier was built using the load impedance calculated from Fig. 8 and the corresponding input impedance calculated from (1). Power input

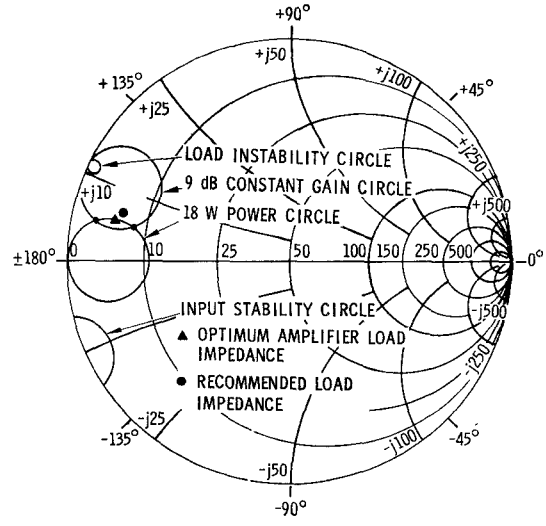

Fig. 8. RF power amplifier design from *S*-parameter data. Predicted load impedance for 9-dB gain and 18-W output is the intersection of the constant gain and constant power circles. The actual optimum load impedance and the manufacturer recommended load impedance are also shown.

TABLE I

	Predicted	Measured	Measured, After Tuning
Gain	9.0 dB	8.6 dB	9.4 dB
Power Output	18.0 W	16.0 W	16.8 W
Input Network	4.0 + j9.9 ohms		4.9 + j8.7
Output Network	8.0 + j8.0		8.1 + j8.0
AMPLIFIER S-PARAMETERS			
S_{11}	0.0	.221 ∠ -159 (VSWR - 1.6)	.093 ∠ 23 (VSWR = 1.2)
S_{21}	2.82 (9.0 dB)	2.69 ∠ -67 (8.6 dB)	2.94 ∠ 89 (9.4 dB)
S_{12}		0.72 ∠ -135	.09 ∠ -156
S_{22}		.703 ∠ 143.7	.65 ∠ 150

to the amplifier was restricted so that the average collector current did not exceed 1 A. (Maximum rating is 1.4 A.) Initially, the amplifier operated with 6-dB gain and 13-W output. Input VSWR was 2.1. Cut-and-try tuning of the input and output networks increased performance to 8-dB gain and 16-W output with input VSWR less than 1.1. The triangle in Fig. 8 marks the load impedance used in the optimized amplifier; it is on the constant power circle and just inside the constant gain circle plotted.

Another amplifier showed somewhat better agreement between theory and experiment. *S*-parameters of a different MRA 0610–20 transistor were measured and an amplifier was designed. Table I summarizes the design and performance of this amplifier. Initially, the amplifier produced 8.6-dB gain at 16-W output. The best performance that could be achieved with this transistor was 9.4-dB gain at 16.8-W output. Again, average collector current was restricted to 1 A.

The emitter and collector voltage waveforms of the transistor when operating in the amplifier are shown in Fig. 9. The waveforms are nearly sinusoidal. The similarity of the waveforms in the *S*-parameter test set (Fig. 1) and in the amplifier,

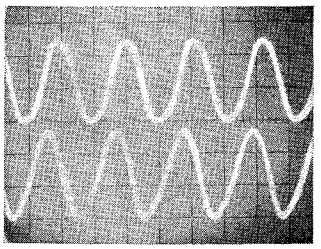

Fig. 9. Emitter and collector voltage waveforms of a TRW-MRA 0610–20 transistor in a large-signal circuit.

and the success of the amplifier design techniques confirm the usefulness of large-signal S-parameters.

Amplifiers have also been designed using S-parameters measured in the 4-Ω system. Predicted input and output impedances are similar when designs based on 4-Ω and 50-Ω class C S-parameters taken with the same average collector current are compared. Successful amplifier designs have also been achieved using other UHF transistors, such as the PHI 1003, PHI 930, and PHI 1025.

V. Discussion and Limitations

The use of large-signal S-parameters for circuit design involves using analytic techniques applicable to linear systems in a nonlinear problem. The success of the design procedures shows that the procedures are approximately valid when used with care.

The fact that waveforms are fairly sinusoidal and do not contain large amounts of harmonic energy suggests that the device is not highly nonlinear, and that linear mathematics may hold. The S-parameters do not vary rapidly with applied power, so it is reasonable to assume that behavior may be fairly linear around a given operating point, so that linear methods can be used. Transistor impedances and the input and output network impedances required for a large-signal amplifier have been calculated from 50-Ω and 4-Ω class C S-parameter measurements. The agreement of these impedances with each other and with those found empirically shows that the linear S-parameter concepts can be applied to class C amplifier design.

The circuit affects transistor operation only by means of the circuit's dc characteristics and its impedances at the signal frequency and its harmonics. If the importance of the harmonic frequencies is reduced by gain rolloff in the transistor and by package parasitics that tend to isolate the transistor chip from its circuit at the higher frequencies, then transistor operation is not strongly dependent upon circuit impedance at the harmonic frequencies. The impedances seen by the chip then depend more upon the transistor package than upon the circuit. Harrison [5] reported that circuit impedance at the second harmonic had little effect upon the performance of a 1.5-GHz transistor amplifier. It may therefore be adequate to consider only the impedance at the fundamental frequency.

The measured S-parameters are a function of measurement conditions. For useful information to be obtained, the S-parameter measurements must be made under conditions approximating the conditions of the desired amplifier. Of the parameters, S_{21} depends most strongly upon operating conditions. The effect of errors in S_{21} on the effective impedances and stability criteria is somewhat reduced by the fact that S_{21} appears only in the product $S_{21}S_{12}$, and S_{12} is quite small. Nevertheless, measurement conditions should approximate the desired amplifier conditions as closely as possible. If self-bias is used in a 50-Ω circuit, considerable power is required to turn the transistor on. The input power is much higher, and the output power much lower than in an amplifier. The best agreement between theoretical and experimental amplifier performance occurs when the S-parameter measurements are made with the same average dc current flow as the amplifier.

The least accurately known S-parameter is apparently S_{12}. If class C bias only is used, S_{12} is measured with no dc collector current, and its value may be different from the value obtained in an amplifier. This could be improved by using class A bias to measure S_{12}. In the MRA 0610–20 transistor, the second term in (1) is small, so errors in S_{12} are not serious. However, S_{12} is more important in some other transistors, and some class A bias may be needed to measure accurate values of S_{12}. An alternative technique for the measurement of S_{12} has been described elsewhere [4].

VI. Conclusions

Large-signal S-parameters and linear circuit design techniques have been used to successfully design an RF power amplifier. This technique is successful if the measurement conditions of the S-parameters are chosen carefully. The success of the procedure allows one convenient way of prescreening transistors, or specifying desired transistor characteristics, or selecting types of transistors to be used for a particular application. Transistor characteristics can be measured in a 50-Ω system rather than in a specially designed circuit.

The 0610–20 has a built-in internal matching network that raises the input impedance, reducing the value of S_{11}. This simplifies the measurement problem as less accuracy is required. Common base transistors without an internal matching network have extremely low impedance and require extreme measurement accuracy because S_{11} is near unity. Common emitter transistors should be somewhat less of a problem because of their higher input impedance.

S-parameters can also be used to gain improved understanding of the operation of circuits. For example, a transistor in which S_{11} and S_{11}' (1) are significantly different may be difficult to turn on at low powers and may suddenly turn on as power is increased, showing hysteresis effects. This is explained as follows. Before power is applied, the transistor is off, S_{21} and S_{12} are very small, and $S_{11}' = S_{11}$. With the transistor turned on, however, S_{11}' is considerably different from S_{11}. If the circuit provides a good match to S_{11}', it provides a poor match for S_{11}, and the transistor could be difficult to turn on initially. The effect of the correction term is, of course, dependent upon Γ_L, so knowledge of the transistor S-parameters could permit the designer to choose a load impedance to reduce the effect, or to use it to advantage. This concept is important to the design of pulsed RF transistor amplifiers.

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