

CHAPTER 3

AMPLIFIER DESIGN TECHNIQUES

3.0 Introduction

Solid-state microwave amplifiers play an important role in communication where it has different applications, including low noise, high gain, and high power amplifiers. The high gain and low noise amplifiers are small signal low power amplifiers and are mostly used in the receiver side where the signal level is low. The small signal S-parameter can be used in designing these low power amplifiers. The high power amplifier is used in the transmitter side where the signal should be at a high level to cross the desired distance. The intent of this chapter is to give an overview of some basic principles used in the analysis and design of the microwave transistor amplifier.

3.1 Small Signal Amplifier Design

The design procedures for a small signal microwave amplifier consist of selecting the dc bias point for the transistor, measuring the S -parameters of the transistor, studying the stability, designing the input and output matching network to achieve the desired goals, building the amplifier, and performing the measurements.

The dc bias point of the transistor should be determined first. The selection of the dc quiescent for the transistor amplifier depends on the particular application. Figure 3.1 shows transistor characteristics with four quiescent points located at A, B, C, and D. For low-noise, high power gain, high output power, or high efficiency applications, the quiescent point located at A, B, C, or D receptively is recommended.

A network's behavior at microwave frequencies can be characterized using the scattering parameters (S -parameters). These parameters are defined in terms of travelling waves. Figure 3.2 shows the incident and reflected waves for a two-ports network (common-source FET).

The relationship between the S -parameters and the incident and reflected waves can be expressed as follows.

$$b_1 = S_{11} \cdot a_1 + S_{12} \cdot a_2 \quad (3.1)$$

$$b_2 = S_{21} \cdot a_1 + S_{22} \cdot a_2 \quad (3.2)$$

where,

a_i and b_i are the incident and reflected waves respectively at port i and can be defined in terms of the voltage wave, as shown in equations (3.3) and (3.4):

$$b_i = \frac{V^+}{(Z_o)^{0.5}} \quad (3.3)$$

$$b_i = \frac{V^-}{(Z_o)^{0.5}} \quad (3.4)$$

where,

Z_0 is the reference impedance.

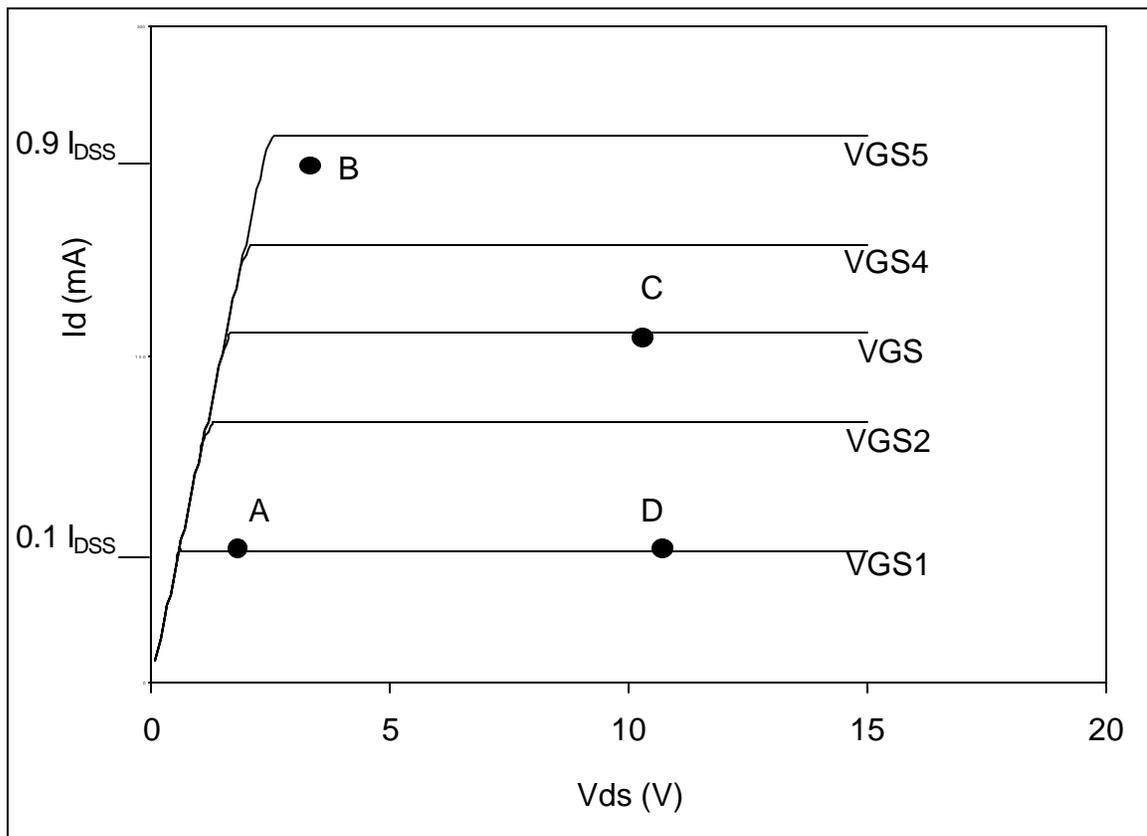


Figure 3.1 Characteristics and recommended quiescent points for transistor amplifier

The S-parameters represent the transmission or reflection coefficients and can be obtained as follows:

$$S_{11} = \left. \frac{b_1}{a_1} \right|_{a_2=0} \quad (\text{input reflection coefficient with output properly terminated})$$

$$S_{21} = \left. \frac{b_2}{a_1} \right|_{a_2=0} \quad (\text{forward transmission coefficient with output properly terminated})$$

$$S_{22} = \left. \frac{b_2}{a_2} \right|_{a_1=0} \quad (\text{output reflection coefficient with output properly terminated})$$

$$S_{12} = \left. \frac{b_1}{a_2} \right|_{a_1=0} \quad (\text{reverse transmission coefficient with output properly terminated}).$$

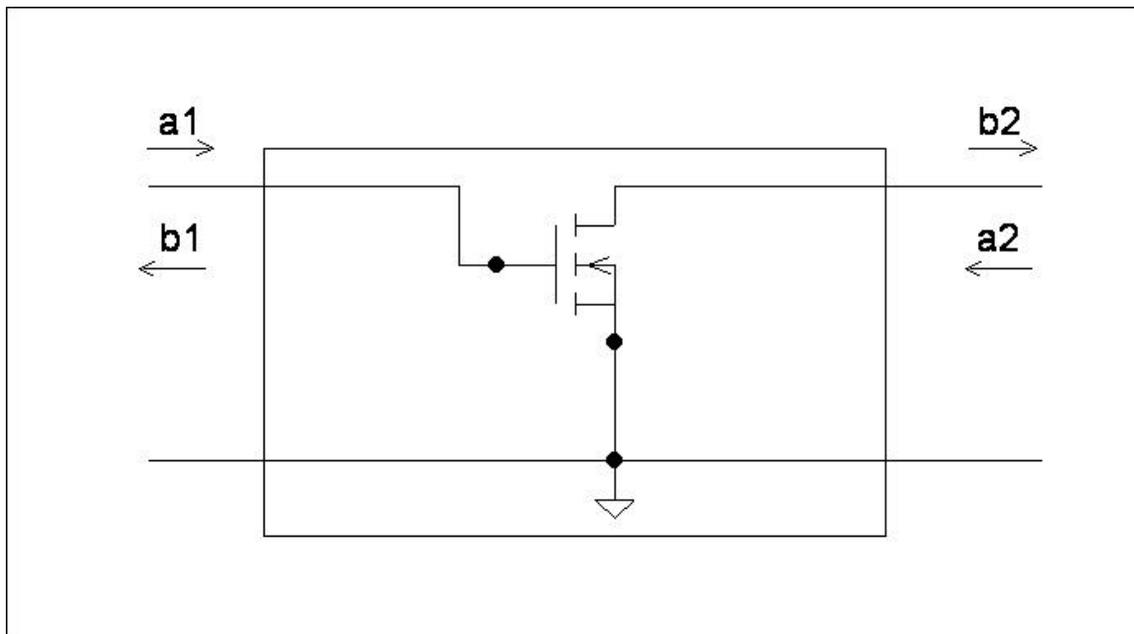


Figure 3.2 A transistor as two-port network.

The stability of an amplifier is a very important consideration in a microwave circuit design. Stability or resistance to oscillation in a microwave circuit can be determined by the S-parameters. Oscillations are possible in a two-port network if either or both the input and the output port have negative resistance. This condition occurs when the magnitude of the input or output reflection coefficients is greater than one, $|\Gamma_{in}| > 1$ or $|\Gamma_{out}| > 1$.

There are two types of amplifier stability, unconditionally stable and conditionally stable. In the former, the real part of the input and output impedances of the amplifier is greater than zero for all passive load and source impedances. However, the amplifier is said to be conditionally stable or potentially unstable if the real part of the input or output impedances of the amplifier is less than zero for at least a passive load or source impedances. The stability test should be done for every frequency in the desired range.

Figure 3.3 shows the source, load, input, and output reflection coefficients for a two-port network.

In terms of reflection coefficients, the necessary conditions for unconditional stability at a given frequency are

$$|\Gamma_S| < 1, \quad (3.5)$$

$$|\Gamma_L| < 1 \quad (3.6)$$

$$|\Gamma_{IN}| = \left| S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_L}{1 - S_{22} \cdot \Gamma_L} \right| < 1 \quad , \quad (3.7)$$

$$|\Gamma_{OUT}| = \left| S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_S}{1 - S_{11} \cdot \Gamma_S} \right| < 1 \quad . \quad (3.8)$$

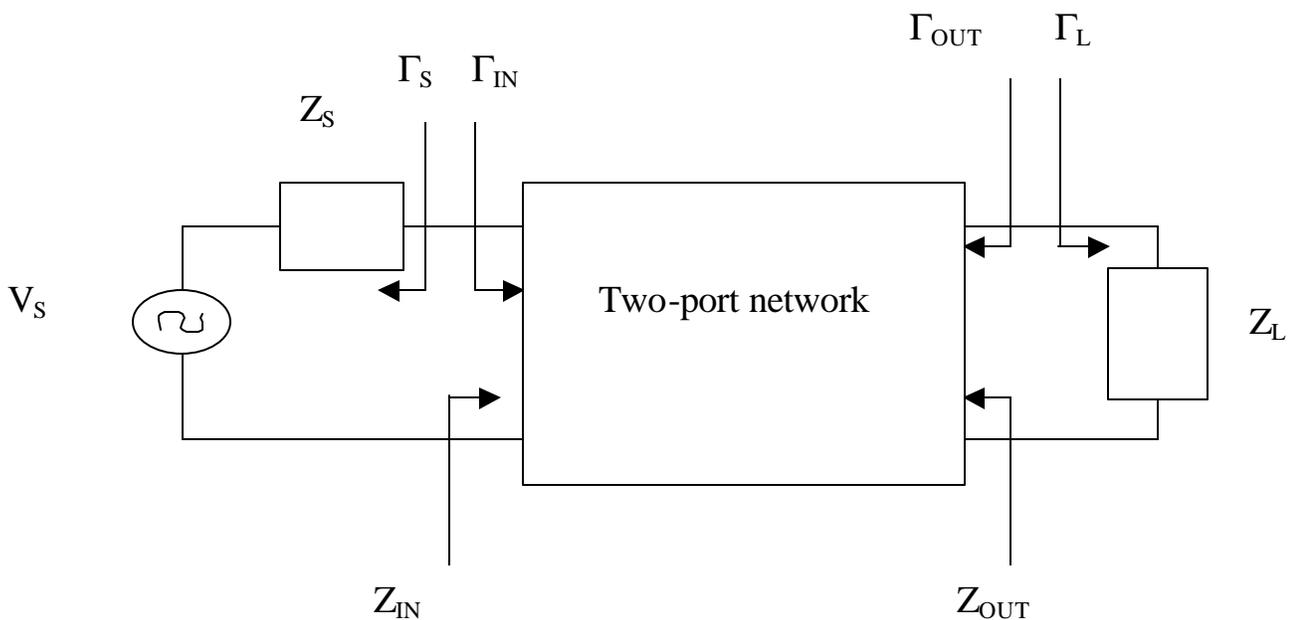


Figure 3.3 Stability of two-port network

The necessary and sufficient conditions for a two-port network to be unconditional stable are [D. Woods, 1976]

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12} \cdot S_{21}|} > 1 \quad , \quad (3.9)$$

$$|\Delta| = |S_{11} \cdot S_{22} - S_{12} \cdot S_{21}| < 1 \quad . \quad (3.10)$$

In practice, most of the microwave transistor amplifiers are potentially unstable because of the internal feedback. There are two ways to overcome the stability problem of the transistor amplifier. The first is to use some form of feedback to stabilize the amplifier. The second is to use a graphical analysis to determine the regions where the values of Γ_S and Γ_L (source and load reflection coefficients) are less than one, which means the real parts of Z_{IN} and Z_{OUT} are positive.

Substituting the values of $|\Gamma_{IN}|=1$ and $|\Gamma_{OUT}|=1$ in equations (3.7) and (3.8) and solving for Γ_S and Γ_L result in the stability circles. The radii and centers of the circles are given by [G. Gonzalez, 1984]

Output Stability Circle,

$$r_L = \left| \frac{S_{21} S_{12}}{|S_{22}|^2 - |\Delta|^2} \right| \quad (3.11)$$

$$C_L = \frac{(S_{22} - \Delta \cdot S_{11}^*)^*}{|S_{22}|^2 - |\Delta|^2}$$

And input Stability Circle,

$$r_L = \left| \frac{S_{21}S_{12}}{|S_{11}|^2 - |\Delta|^2} \right| \quad (3.12)$$

$$C_L = \frac{(S_{11} - \Delta \cdot S_{22}^*)^*}{|S_{11}|^2 - |\Delta|^2} .$$

Then the stability circles need to be plotted in the Smith chart to determine the stable regions or in other words, the regions where values of Γ_S and Γ_L produce $|\Gamma_{OUT}| < 1$ and $|\Gamma_{IN}| < 1$.

Most of the time, microwave amplifiers used for narrowband or wideband applications face stability problems at certain frequency ranges. Instability is primarily caused by three phenomena: internal feedback around the transistor, external feedback around the transistor caused by an external circuit, or excess of gain at frequencies outside of the band of operation.

Figure 3.4 shows six passive feedback-networks that are used usually to stabilize amplifier circuits.

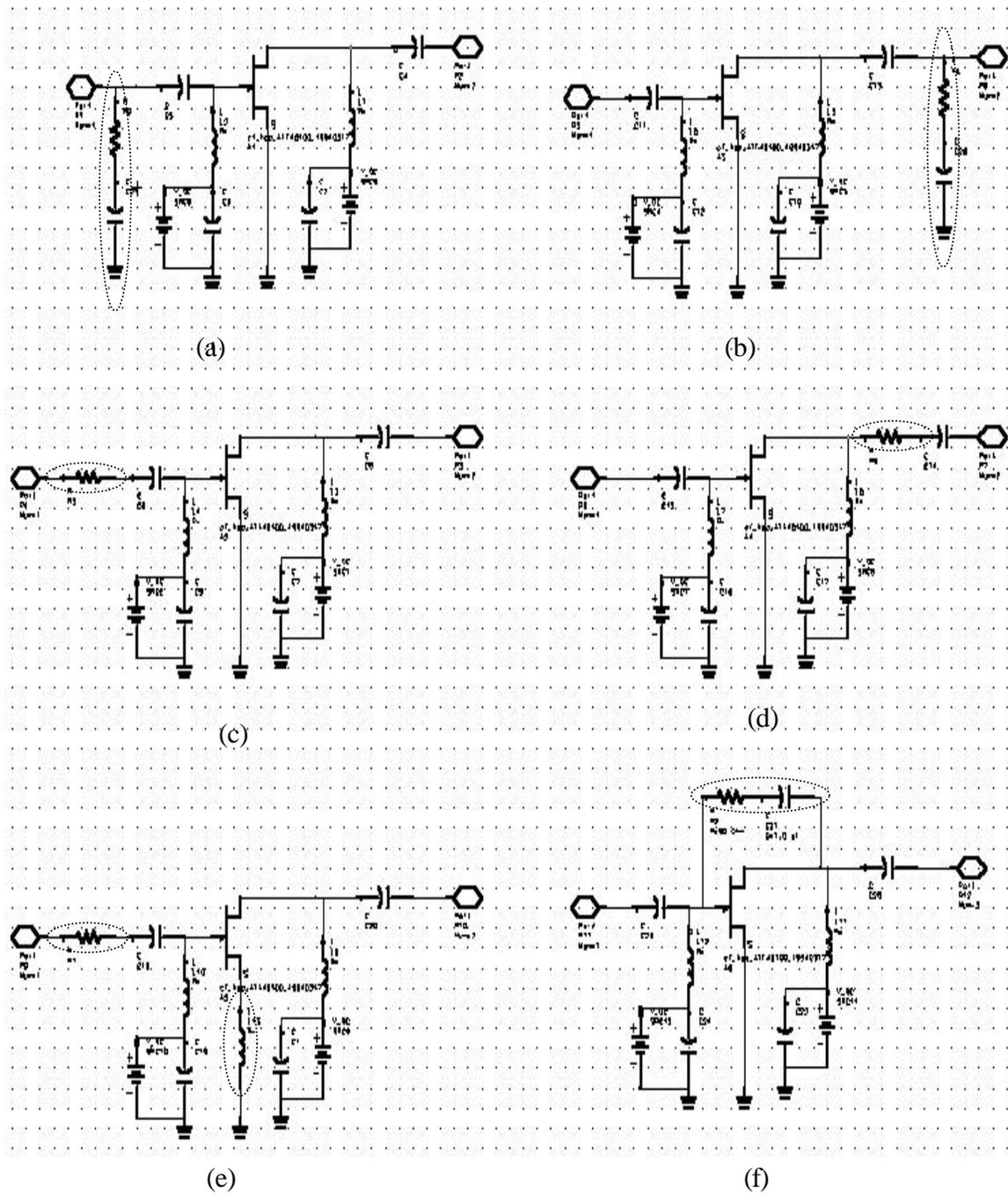


Figure.3.4 Passive-networks stabilization

Amplifiers usually need matching networks to achieve the desired goals. The input-matching network of the low noise amplifier (LNA) is designed to transform the 50Ω impedance of the preceding stage to the impedance required to achieve the minimum noise figure, and the output-matching network is designed to achieve a high power gain. In the high output power amplifier, the input-matching network presents a conjugate impedance to the input impedance of the amplifier, and the output-matching network is designed to achieve the desired output power. The input and output matching networks of the high gain amplifier are designed to be conjugally matched to the input and output impedance of the amplifier.

Matching networks can be implemented using lumped elements, distributed elements (transmission lines), or a combination of lumped and distributed elements. The lumped elements can be used as long as their dimensions are much smaller than the electrical wavelength. With modern microwave integrated circuits, the lumped elements can be used up to 60 GHz. The ZY Smith chart can be used conveniently in the design of matching networks.

The process of choosing the source and load impedances for the high gain amplifier and LNA is well established using the small-signal S-parameters [G. Gonzalez, 1984, Ha, 1981].

3.2 High Power Amplifier Design

In the case of the power amplifier, the input signal level is often high, and consequently the output current is either in the cutoff or saturation region during a portion of the input signal cycle. The small signal S-parameters can be used if the large-signal amplifier is operating in class-A. However, for other classes (B, C, etc), the small signal S-parameters are of little use, and it becomes necessary to use other techniques. The design procedures of the high power amplifier are similar to those of the small signal amplifier. The main difference is that the high power amplifier uses other techniques to obtain the proper impedances at the input and output ports of the amplifier. The main techniques used in the high power amplifier design are conventional load-pull, active load-pull, and two-port large signal characterization.

3.2.1 Load-pull Techniques

The conventional and active load-pull techniques have the same concept of operation; they provide information for the source and load reflection coefficients as a function of output power and gain. Figure 3.5 illustrates the basic concept of the load-pull techniques. The transistor under test is placed in a measuring setup where the dc bias and ac input signal are fixed. The output tuner is adjusted until the power meter C measures a given power level and the input tuner is adjusted for zero reflected power (read at power meter B). The power meter A reads the incident power, and the power gain can be obtained. From this information gain, power added efficiency (PAE), and output power contours can be generated and drawn in the Smith chart as function of the output load.

While the conventional load-pull technique uses a passive tuner (tuning stubs) to vary the output load, the active load-pull technique varies the load actively by injecting a power wave with variable magnitude and phase toward the transistor output. The load

variation can take place at the fundamental frequency f_o alone or at the fundamental frequency and a number of the harmonics (multiharmonic load-pull). The harmonic load impedance affects the performance of some operational classes (e.g., classes-C) [S. R. Mazumder, A. Azizi, F. E. Gardiol, 1979].

A passive tuner is commercially available, can handle high powers, is easy to use, and is of comparatively low cost. Despite these advantages, it has inherent losses that impose limitations in the reflection-coefficient magnitude. The maximum magnitude decreases with the frequency and with the number of elements and cables connected between the measurement plan and tuner. The active load-pull principle does not suffer from limitations in the load reflection-coefficient magnitude. Two basic principles of load variation are commonly used: either the signal injected into the device output is synchronized with the one applied to the input [Takayama, 1976] or the signal generated by the device itself is fed back to its output with variable phase and magnitude [G. P. Bava, U. Pisani, and V. Pozzolo, 1982].

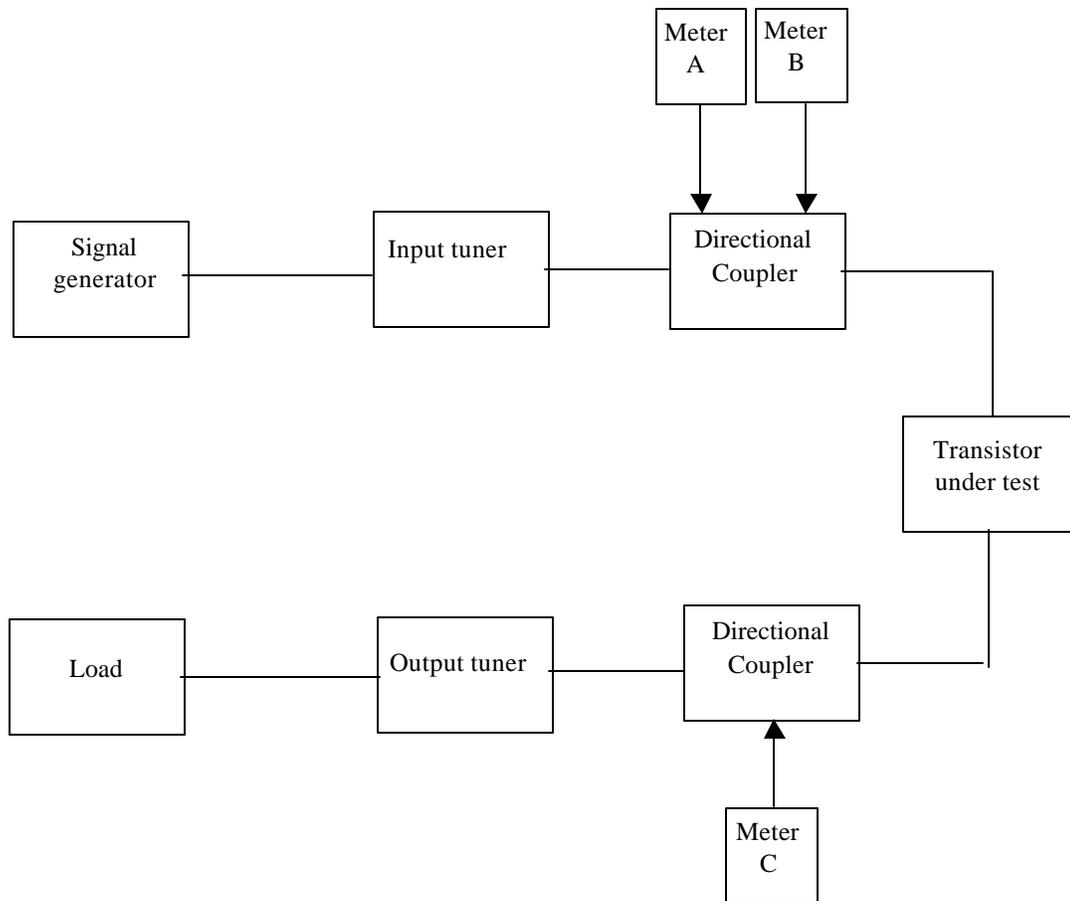


Figure 3.5 Two-port load-pull measurement system

3.2.2 Two-Port Large Signal Techniques

The two-port large signal characterization is basically making S-parameter measurements with respect to 50Ω reference impedances. There are two main methods: the direct extension of small-signal measurement to large signal [W. Leighton, R. Chaffin, and J. Webb, 1973], and a quasi-large-signal approach whereby the large-signal S-parameters are measured by simultaneous application of two coherent signals at the same frequency to the input and output of the device [P.D. van, and S.R. Mazumder, 1978]. Although it is not useful with a nonlinear amplifier, the obvious advantage of the former method is its simplicity. Although the S-parameters of a nonlinear amplifier can be obtained using the quasi-large-signal method, they are of little use if the non-linearity is severe.

Equations (3.1) and (3.2) can be rewritten in the form:

$$\frac{b_1}{a_1} = S_{11} + S_{12} \cdot \frac{a_2}{a_1} \quad (3.13)$$

$$\frac{b_1}{a_2} = S_{11} \cdot \frac{a_1}{a_2} + S_{12} \quad (3.142)$$

$$\frac{b_2}{a_1} = S_{21} + S_{22} \cdot \frac{a_2}{a_1} \quad (3.15)$$

$$\frac{b_2}{a_2} = S_{21} \cdot \frac{a_1}{a_2} + S_{22} \quad (3.16)$$

In the quasi-large-signal technique, a_1 and a_2 are the two applied signals. The four S-parameters can be determined for different values of a_1 and a_2 .

Once the S-parameters of the device are obtained, the next step is to apply the small signal S-parameter techniques [G. Gonzalez, 1984].