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RF and Microwave Power Amplifier Design

- ✔ Active device modeling
- ✔ High-power amplifier design
- ✔ High-efficiency operation techniques
- ✔ Broadband power amplifiers
- ✔ Wireless communications applications

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Preface

The main objective of this book is to present all the relevant information required for RF and micro-wave power amplifier design including well-known and novel theoretical approaches and practical design techniques as well as to suggest optimum design approaches effectively combining analytical calculations and computer-aided design. This book can also be very useful for lecturing to promote the analytical way of thinking with practical verification by making a bridge between theory and practice of RF and microwave engineering. As it often happens, a new result is the well-forgotten old one. Therefore, the demonstration of not only new results based on new technologies or circuit schematics is given, but some sufficiently old ideas or approaches are also introduced, that could be very useful in modern practice or could contribute to appearance of new ideas or schematic techniques.

As a result, this book is intended for and can be recommended to:

- *University-level professors and scientists*, as possible reference and well-founded material for creative research and teaching activity that will contribute to strong background for graduate and postgraduate students
- *R & D staff*, to combine the theoretical analysis and practical aspect including computer-aided design and to provide a sufficient basis for new ideas in theory and practical circuit technique
- *Practicing RF designers and engineers*, as an anthology of many well-known and new practical RF and microwave power amplifier circuits with detailed description of their operational principles and applications and clear practical demonstration of theoretical results

In Chapter 1, the two-port networks are introduced to describe the behavior of linear and nonlinear circuits. To characterize the nonlinear properties of the bipolar or field-effect transistors, their equivalent circuit elements are expressed through the impedance Z-parameters, admittance Y-parameters, or hybrid H-parameters. On the other hand, the transmission *ABCD*-parameters are very important in the design of the distributed circuits as a transmission line or cascaded elements, whereas the scattering *S*-parameters are widely used to simplify a measurement procedure.

The main purpose of Chapter 2 is to present widely used nonlinear circuit design techniques to analyze nonlinear power amplifier circuits. In general, there are several approaches to analyze and design these nonlinear circuits, depending on their main specifications—for example, an analysis in time domain when it is necessary to determine the transient circuit behavior or in frequency domain to provide improvement of the power and spectral performances when both parasitic effects such as instability and spurious effects must be eliminated or minimized. Using the time-domain technique it is quite easy to describe the circuit by differential equations, whereas frequency-domain analysis is more explicit when a relatively complex circuit can be reduced to one or more sets of immitances at each harmonic component.

In Chapter 3, all the necessary steps to provide an accurate device modeling procedure starting with the determination of the small-signal equivalent circuit parameters are described and discussed. A variety of nonlinear models for MOSFET, MESFET, HEMT, and bipolar devices including HBTs, which are very prospective for modern microwave monolithic integrated circuits of power amplifiers and oscillators, are presented. In order to highlight the advantages or drawbacks of one nonlinear device model over the other, a comparison of the measured and modeled volt-ampere and voltage-capacitance characteristics or a frequency range of model application is made.

A concept of impedance matching and the impedance-matching technique, which is very important when designing power amplifiers, is presented in Chapter 4. First, the main principles and impedance-matching tools such as the Smith chart are described, giving the starting point of the matching-design procedure. As an engineering solution in general depends on the different circuit requirements, the designer should choose the optimum solution among a variety of the matching networks including either lumped elements or transmission lines or both of them. To simplify and visualize the matching-design procedure, an analytical approach, which allows calculating the parameters of the matching circuits using simple equations, and Smith chart traces is discussed and illustrated with several examples of the narrowband and broadband RF and microwave power amplifiers using bipolar or MOSFET devices. Finally, the design formulas and curves are presented for different types of transmission lines including stripline, microstrip line, slotline, and coplanar waveguide.

Chapter 5 describes the basic properties of three-port and four-port networks as well as a variety of different combiners, transformers, and directional couplers for RF and microwave power applications. So, for power combining in view of insufficient power performance of the active devices, it is best to use the coaxial cable combiners with ferrite core to combine the output powers of RF power amplifiers intended for wideband applications. As the device output impedance for high power levels is usually too small, to match this impedance with a standard 50-Ω load, it is necessary to use the co-axial line transformers with specified impedance transformation. For narrowband applications, the N-way Wilkinson combiners are widely used due to the simplicity of their practical realization. At the same time in microwaves, the size of the combiners should be very small. Therefore, the commonly used hybrid microstrip combiners including different types of microwave hybrid and directional couplers are described and analyzed.

Chapter 6 represents the fundamentals of the power amplifier design, which is generally a complicated procedure when it is necessary to provide simultaneously accurate active device modeling, effective impedance matching depending on the technical requirements and operation conditions, stability in operation, and ease in practical implementation. Therefore, at the beginning of the chapter the key definitions of different power gains and stability are introduced. For a stable operation mode of the power amplifier, it is necessary to evaluate the operating frequency domains where the active device may be potentially unstable. To avoid parasitic oscillations, the stabilization circuit technique for different frequency domains from low frequencies to high frequencies close to the device transition frequency is analyzed and discussed. One of the key parameters of the power amplifier is its linearity, which is very important for many TV and cellular applications. Therefore, the relationships between the output power, 1-dB gain compression point, third-order intercept point, and intermodulation distortions of the third and higher orders are given and illustrated for different active devices. The basic classes of the power amplifier operation A, AB, B, and C are introduced, analyzed, and illustrated. The device biasing conditions and examples of bias circuits for MOSFET and bipolar devices to improve linearity or to increase efficiency are shown and discussed. Also the concept of push-pull amplifiers and their circuit design using balanced transistors is given. In the final section, the numerous practical examples of power amplifiers using MOSFET, MESFET, and bipolar devices in different frequency ranges and for output powers are shown and discussed.

Modern commercial and military communication systems require high-efficiency long-term operating conditions. Chapter 7 describes in detail the possible circuit solutions to provide a high-efficiency power amplifier operation based on using different overdriven (Class B, Class F, and Class E) classes of operation or newly developed subclasses, depending on the technical requirements. In Class F amplifiers analyzed in frequency domain, the fundamental and harmonic load impedances are optimized by short-circuit termination and open-circuit peaking in order to control the voltage and current waveforms at the drain of the device to obtain maximum efficiency. In Class E amplifiers analyzed in time domain, an efficiency improvement is achieved by realizing the on/off switching operation with special current and voltage waveforms so that high voltage and high current do not exist at the same time. The parallel-circuit Class E load network configuration can be easily implemented in the broadband high-efficiency power amplifier design. The Class E load network with a quarterwave transmission line provides an addidtional suppression of even harmonic components.

In many telecommunication, radar or testing systems, the transmitters operate in a very wide frequency range. Chapter 8 describes the power amplifier design based on a broadband concept that provides some advantages when there is no need to tune the resonant circuit parameters. However, there are many factors that restrict the frequency bandwidth depending on the active device parameters. So, it is sufficicently easy to provide multioctave amplification from very low frequencies up to ultrahigh frequencies using the power MOSFET devices when loss gain compensation is easily realized. At higher frequencies when the device input impedance is significantly smaller and the influence of its internal feedback and parasitic parameters is substantially higher, it is necessary to use multisection-matching networks with lumped and distributed elements. A variety of broadband power amplifiers using different frequency ranges are presented and described.

Chapter 9 describes the different approaches to improve linearity and efficiency of the power amplifiers in telecommunication systems. To improve the efficiency of operation, the Kahn envelope and restoration and envelope-tracking techniques, the outphasing and Doherty power amplifier archi-tectures, and the switched-mode and dual-path power amplifier configurations are shown and analyzed. To improve the linearity of operation, the feedforward linearizing technique and predistoration linearization circuit schematics are described and presented. Special attention is payed to practical realization of monolithic integrated circuits of HBT and CMOS power amplifiers for handset applications using modern technologies.

Andrei Grebennikov

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Contents

Chapter

1

Two-Port Network Parameters

Two-port-equivalent circuits are widely used in radio frequency (RF) and microwave circuit design to describe the electrical behavior of both active and passive devices. A two-port network (whose elements are expressed through the impedance Z-parameters, admittance Y parameters, or hybrid H-parameters) is most suitable to characterize the nonlinear properties of the active devices, bipolar or field-effect transistors, when designing power amplifiers or oscillators. Transmission $ABCD$-parameters of a two-port network are very convenient for designing the distributed circuit as transmission lines or cascaded elements. Scattering S-parameters are used to simplify the measurement procedure.

This chapter discusses the main properties of two-port network parameters, as well as the ratios between the different systems. In addition, examples are given to illustrate how to best analyze power amplifiers and oscillators. The final part of this chapter describes the transmission line and its main parameters. Additional information on more specific aspects of two-port network circuits can be found in Refs. [1–4] listed at the end of the chapter.

Traditional Network Parameters

The basic diagram of a two-port nonautonomous transmission system can be represented by the equivalent circuit shown in Fig. 1.1, where V_S is the independent voltage source, Z_S is the source impedance, LN is the linear time-invariant two-port network without independent source, and Z_L is the load impedance. Two independent phasor currents, I_1 and I_2 (flowing across input and output terminals), and phasor voltages,

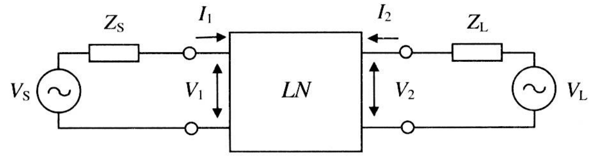

Figure 1.1 Basic diagram of two-port nonautonomous transmission system.

V_1 and V_2, characterize such a two-port network. For autonomous oscillator systems, in order to provide an appropriate analysis in the frequency domain of the two-port network in the negative one-port representation, it is sufficient to set the source impedance to infinity. For power amplifier and oscillator design, the elements of the matching or resonant circuits, which are assumed to be linear or appropriately linearized, can be found among the LN-network elements, or additional two-port linear networks can be used to describe their frequency domain behavior.

For a two-port network, the following equations can be considered to be imposed boundary conditions:

$$V_1 + Z_S I_1 = V_S \tag{1.1}$$

$$V_2 + Z_L I_2 = V_L \tag{1.2}$$

Suppose that it is possible to obtain a unique solution for the linear time-invariant circuit shown in Fig. 1.1. Then the two linearly independent equations, which describe the general two-port network representation in terms of circuit variables V_1, V_2, I_1, and I_2, can be expressed in matrix form as

$$[M][V] + [N][I] = 0 \tag{1.3}$$

or

$$\left.\begin{aligned} m_{11}V_1 + m_{12}V_2 + n_{11}I_1 + n_{12}I_2 = 0 \\ m_{21}V_1 + m_{22}V_2 + n_{21}I_1 + n_{22}I_2 = 0 \end{aligned}\right\} \tag{1.4}$$

In Eq. (1.3), the complex 2×2 matrices $[M]$ and $[N]$ are independent of the source and load impedances Z_S and Z_L and voltages V_S and V_L, respectively; they depend only on the circuit elements inside the LN network.

If matrix $[M]$ in Eq. (1.3) is nonsingular when $|M| \neq 0$, then this matrix equation can be rewritten in terms of $[I]$ as

$$[V] = -[M]^{-1}[N][I] = [Z][I] \tag{1.5}$$

where $[Z]$ is the open-circuit impedance two-port network matrix. In scalar form, matrix Eq. (1.5) is given by

$$V_1 = Z_{11}I_1 + Z_{12}I_2 \tag{1.6}$$

$$V_2 = Z_{21}I_1 + Z_{22}I_2 \tag{1.7}$$

where Z_{11} and Z_{22} are the open-circuit driving-point impedances, and Z_{12} and Z_{21} are the open-circuit transfer impedances of the two-port network. The voltage components V_1 and V_2, due to the input current I_1 are found by defining $I_2 = 0$ in Eqs. (1.6) and (1.7), which results in an open output terminal. Similarly, the same voltage components V_1 and V_2 are determined by setting $I_1 = 0$ when the input terminal becomes open-circuited. The resulting driving-point impedances can be written as follows:

$$Z_{11} = \left.\frac{V_1}{I_1}\right|_{I_2=0} \qquad Z_{22} = \left.\frac{V_2}{I_2}\right|_{I_1=0} \tag{1.8}$$

The two transfer impedances are

$$Z_{21} = \left.\frac{V_2}{I_1}\right|_{I_2=0} \qquad Z_{12} = \left.\frac{V_1}{I_2}\right|_{I_1=0} \tag{1.9}$$

Dual analysis can be used to derive the short-circuit admittance matrix when the current components I_1 and I_2 are considered as outputs caused by V_1 and V_2. If matrix $[N]$ in Eq. (1.3) is nonsingular when $|N| \neq 0$, this matrix equation can be rewritten in terms of $[V]$ as

$$[I] = -[N]^{-1}[M][V] = [Y][V] \tag{1.10}$$

where $[Y]$ is the short-circuit admittance two-port network matrix. In scalar form, matrix Eq. (1.10) is written as

$$I_1 = Y_{11}V_1 + Y_{12}V_2 \tag{1.11}$$

$$I_2 = Y_{21}V_1 + Y_{22}V_2 \tag{1.12}$$

where Y_{11} and Y_{22} are the short-circuit driving-point admittances, and Y_{12} and Y_{21} are the short-circuit transfer admittances of the two-port network. In this case the current components I_1 and I_2, due to the input voltage source V_1, are determined by setting $V_2 = 0$ in Eqs. (1.11) and (1.12), which creates a short output terminal. Similarly, the same current components I_1 and I_2 are determined by setting $V_1 = 0$ when the input terminal becomes short-circuited. As a result, the two driving point admittances are

$$Y_{11} = \left.\frac{I_1}{V_1}\right|_{V_2=0} \qquad Y_{22} = \left.\frac{I_2}{V_2}\right|_{V_1=0} \tag{1.13}$$

The two transfer admittances are

$$Y_{21} = \left.\frac{I_2}{V_1}\right|_{V_2=0} \qquad Y_{12} = \left.\frac{I_1}{V_2}\right|_{V_1=0} \tag{1.14}$$

In some cases, an equivalent two-port network representation can be obtained to express voltage source V_1 and output current I_2 in terms of input current I_1 and output voltage V_2. By solving Eq. (1.4), if the submatrix

$$\begin{bmatrix} m_{11} & n_{12} \\ m_{21} & n_{22} \end{bmatrix}$$

is nonsingular, then

$$\begin{bmatrix} V_1 \\ I_2 \end{bmatrix} = -\begin{bmatrix} m_{11} & n_{12} \\ m_{21} & n_{22} \end{bmatrix}^{-1} \begin{bmatrix} n_{11} & m_{12} \\ n_{21} & m_{22} \end{bmatrix} \begin{bmatrix} I_1 \\ V_2 \end{bmatrix} = [H] \begin{bmatrix} I_1 \\ V_2 \end{bmatrix} \tag{1.15}$$

where $[H]$ is the hybrid two-port network matrix. In scalar form, it is best to represent matrix Eq. (1.15) as

$$V_1 = h_{11}I_1 + h_{12}V_2 \tag{1.16}$$

$$I_2 = h_{21}I_1 + h_{22}V_2 \tag{1.17}$$

where h_{11}, h_{12}, h_{21}, and h_{22} are the hybrid H-parameters. The voltage source V_1 and current component I_2 are determined by defining $V_2 = 0$ for the short output terminal in Eqs. (1.16) and (1.17):

$$h_{11} = \left.\frac{V_1}{I_1}\right|_{V_2=0} \qquad h_{21} = \left.\frac{I_2}{I_1}\right|_{V_2=0} \tag{1.18}$$

where h_{11} is the driving-point input impedance and h_{21} is the forward current transfer function. Similarly, the input voltage source V_1 and output current I_2 are determined by defining $I_1 = 0$ when the input terminal is open-circuited:

$$h_{12} = \left.\frac{V_1}{V_2}\right|_{I_1=0} \qquad h_{22} = \left.\frac{I_2}{V_2}\right|_{I_1=0} \tag{1.19}$$

where h_{12} is the reverse voltage transfer function and h_{22} is the driving-point output admittance.

Transmission parameters, often used for passive device analysis, are determined for independent input voltage source V_1 and input current I_1 in terms of output voltage V_2 and output current I_2. Solving Eq. (1.4), if the submatrix

$$\begin{bmatrix} m_{11} & n_{11} \\ m_{21} & n_{21} \end{bmatrix}$$

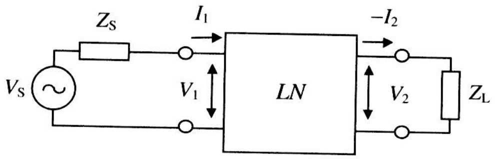

Figure 1.2 Basic diagram of loaded two-port transmission system.

is nonsingular, then we obtain

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = -\begin{bmatrix} m_{11} & n_{11} \\ m_{21} & n_{21} \end{bmatrix}^{-1} \begin{bmatrix} m_{12} & n_{12} \\ m_{22} & n_{22} \end{bmatrix} \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix} = [ABCD] \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix} \quad (1.20)$$

where $[ABCD]$ is the forward transmission two-port network matrix. In scalar form, we can write

$$V_1 = AV_2 - BI_2 \quad (1.21)$$

$$I_1 = CV_2 - DI_2 \quad (1.22)$$

where A, B, C, and D are the transmission parameters. The voltage source V_1 and current component I_1 are determined by defining $I_2 = 0$ for the open output terminal in Eqs. (1.21) and (1.22):

$$A = \left.\frac{V_1}{V_2}\right|_{I_2=0} \qquad C = \left.\frac{I_1}{V_2}\right|_{I_2=0} \quad (1.23)$$

where A is the reverse voltage transfer function and C is the reverse transfer admittance. Similarly, input-independent variables V_1 and I_1 are determined by defining $V_2 = 0$ when the output terminal is short circuited:

$$B = \left.\frac{V_1}{I_2}\right|_{V_2=0} \qquad D = \left.\frac{I_1}{I_2}\right|_{V_2=0} \quad (1.24)$$

where B is the reverse transfer impedance and D is the reverse current transfer function. The reason that a minus sign is associated with I_2 in Eqs. (1.20) and (1.21) is that historically, for transmission networks, the input signal is considered as going to the input port whereas the output current is flowing to the load. The current $-I_2$ entering the load is shown in Fig. 1.2.

Scattering Parameters

The concept of incident and reflected voltage and current parameters can be illustrated by the one-port network shown in Fig. 1.3, where network impedance Z is connected to the signal source V_S with internal impedance Z_S. In a common case, the terminal current I and