RF AND MICROWAVE TRANSMITTER DESIGN

Andrei Grebennikov
Bell Labs, Alcatel-Lucent, Ireland

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PREFACE

The main objective of this book is to present all relevant information required to design the transmitters in general and their main components in particular in different RF and microwave applications including well-known historical and recent novel architectures, theoretical approaches, circuit simulation results, and practical implementation techniques. This comprehensive book can be very useful for lecturing to promote the systematic way of thinking with analytical calculations and practical verification, thus making a bridge between theory and practice of RF and microwave engineering. As a result, this book is intended for and can be recommended to *university-level professors* as a comprehensive material to help in lecturing for graduate and postgraduate students, to *researchers and scientists* to combine the theoretical analysis with practical design and to provide a sufficient basis for innovative ideas and circuit design techniques, and to *practicing designers and engineers* as the book contains numerous well-known and novel practical circuits, architectures, and theoretical approaches with detailed description of their operational principles and applications.

Chapter 1 introduces the basic two-port networks describing 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 for the design of the distributed circuits such as a transmission line or cascaded elements, whereas the scattering S-parameters are widely used to simplify the measurement procedure. The design formulas and curves are given for several types of transmission lines including stripline, microstrip line, slotline, and coplanar waveguide. Monolithic implementation of lumped inductors and capacitors is usually required at microwave frequencies and for portable devices. Knowledge of noise phenomena such as noise figure, additive white noise, low-frequency fluctuations, or flicker noise in active or passive elements is very important for the oscillator modeling in particular and entire transmitter design in general.

In Chapter 2, all necessary steps to provide an accurate device modeling procedure starting with the determination of the device small-signal equivalent circuit parameters are described and discussed. A variety of nonlinear models for MOSFET, MESFET, HEMT, and BJT devices including HBTs, which are very prospective for modern microwave monolithic integrated circuits, are given. In order to highlight the advantages or drawbacks of one over another nonlinear device model, a comparison of the measured and modeled volt–ampere and voltage–capacitance characteristics, as well as a frequency range of model application, are analyzed.

The main principles and impedance matching tools are described in Chapter 3. Generally, an optimum solution depends on the circuit requirements, such as the simplicity in practical realization, frequency bandwidth and minimum power ripple, design implementation and adjustability, stable operation conditions, and sufficient harmonic suppression. As a result, many types of the matching networks are available, including lumped elements and transmission lines. To simplify and visualize the matching design procedure, an analytical approach, which allows calculation of the parameters of the matching circuits using simple equations, and Smith chart traces are discussed. In addition, several examples of the narrowband and broadband power amplifiers using bipolar or MOSFET devices are given, including successive and detailed design considerations and explanations.

Chapter 4 describes the basic properties of the three-port and four-port networks, as well as a variety of different combiners, transformers, and directional couplers for RF and microwave power applications. 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. Since the device output impedance is usually too small for high power level, to match this impedance with a standard 50-Ω load, it is necessary to use the coaxial-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, the size of the combiners should be very small at microwave frequencies. Therefore, the commonly used hybrid microstrip combiners including different types of the microwave hybrids and directional couplers are described and analyzed.

Chapter 5 introduces the basic types of RF and microwave filters based on the low-pass or high-pass sections and bandpass or bandstop transformation. Classical filter design approaches using image parameter and insertion loss methods are given for low-pass and high-pass *LC* filter implementations. The quarterwave-line and coupled-line sections, which are the basic elements of microwave transmission-line filters, are described and analyzed. Different examples of coupled-line filters including interdigital, combline, and hairpin bandpass filters are given. Special attention is paid to microstrip filters with unequal phase velocities, which can provide unexpected properties because of different implementation technologies. Finally, the typical structures, implementation technology, operational principles, and band performance of the filters based on surface and bulk acoustic waves are presented.

Chapter 6 discusses the basic features of different types of analog modulation including amplitude, single-sideband, frequency, and phase modulation, and basic types of digital modulation such as amplitude shift keying, frequency shift keying, phase shift keying, or pulse code modulation and their variations. The principle of operations and various schematics of the modulators for different modulation schemes including Class S modulator for pulse-width modulation are described. Finally, the concept of time and frequency division multiplexing is introduced, as well as a brief description of different multiple access techniques.

A basic theory describing the operational principles of frequency conversion in receivers and transmitters is given in Chapter 7. The different types of mixers, from the simplest based on a single diode to a balanced and double-balanced type based on both diodes and transistors, are described and analyzed. The special case is a mixer based on a dual-gate transistor that provides better isolation between signal paths and simple implementation. The frequency multipliers that historically were a very important part of the vacuum-tube transmitters can extend the operating frequency range.

Chapter 8 presents the principles of oscillator design, including start-up and steady-state operation conditions, noise and stability of oscillations, basic oscillator configurations using lumped and transmission-line elements, and simplified equation-based oscillator analyses and optimum design techniques. An immittance design approach is introduced and applied to the series and parallel feedback oscillators, including circuit design and simulation aspects. Voltage-controlled oscillators and their varactor tuning range and linearity for different oscillator configurations are discussed. Finally, the basic circuits and operation principles of crystal and dielectric resonator oscillators are given.

Chapter 9 begins with description of the basic phase-locked loop concept. Then, the basic performance and structures of the analog, charge-pump, and digital phase-locked loops are analyzed. The basic loop components such as phase detector, loop filter, frequency divider, and voltage-controlled oscillator are discussed, as well as loop dynamic parameters. The possibility and particular realizations of the phase modulation using phase-locked loops are presented. Finally, general classes of frequency synthesizer techniques such as direct analog synthesis, indirect synthesis, and direct digital synthesis are discussed. The proper choice of the synthesizer type is based on the number of frequencies, frequency spacing, frequency switching time, noise, spurious level, particular technology, and cost.

Chapter 10 introduces 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 simplicity in practical implementation. The quality of the power amplifier design is evaluated by realized maximum power gain under stable operation condition with minimum amplifier stages, and the requirement of linearity or high efficiency can be considered where it is needed. For a stable operation, it is necessary to evaluate the operating frequency domains where the active device may be potentially unstable. To avoid the parasitic oscillations, the stabilization circuit technique for different frequency domains (from low frequencies up to high frequencies close to the device transition frequency) is discussed. The key parameter of the power amplifier is its linearity, which is very important for many wireless communication applications. 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 device bias conditions, which are generally different for linearity or efficiency improvement, depend on the power amplifier operation class and the type of the active device. The basic Classes A, AB, B, and C of the power amplifier operation are introduced, analyzed, and illustrated. The principles and design of the push–pull amplifiers using balanced transistors, as well as broadband and distributed power amplifiers, are discussed. Harmonic-control techniques for designing microwave power amplifiers are given with description of a systematic procedure of multiharmonic load–pull simulation using the harmonic balance method and active load–pull measurement system. Finally, the concept of thermal resistance is introduced and heatsink design issues are discussed.

Modern commercial and military communication systems require the high-efficiency long-term operating conditions. Chapter 11 describes in detail the possible circuit solutions to provide a high-efficiency power amplifier operation based on using Class D, Class F, Class E, or their newly developed subclasses depending on the technical requirements. In all cases, an efficiency improvement in practical implementation is achieved by providing the nonlinear operation conditions when an active device can simultaneously operate in pinch-off, active, and saturation regions, resulting in nonsinusoidal collector current and voltage waveforms, symmetrical for Class D and Class F and asymmetrical for Class E (DE, FE) operation modes. In Class F amplifiers analyzed in frequency domain, the fundamental-frequency and harmonic load impedances are optimized by short-circuit termination and open-circuit peaking to control the voltage and current waveforms at the device output to obtain maximum efficiency. In Class E amplifiers analyzed in time domain, an efficiency improvement is achieved by realizing the on/off active device switching operation (the pinch-off and saturation modes) with special current and voltage waveforms so that high voltage and high current do not concur at the same time.

In modern wireless communication systems, it is very important to realize both high-efficiency and linear operation of the power amplifiers. Chapter 12 describes a variety of techniques and approaches that can improve the power amplifier performance. To increase efficiency over power backoff range, the Doherty, outphasing, and envelope-tracking power amplifier architectures, as well as switched multipath power amplifier configurations, are discussed and analyzed. There are several linearization techniques that provide linearization of both entire transmitter system and individual power amplifier. Feedforward, cross cancellation, or reflect forward linearization techniques are available technologies for satellite and cellular base station applications achieving very high linearity levels. The practical realization of these techniques is quite complicated and very sensitive to both the feedback loop imbalance and the parameters of its individual components. Analog predistortion linearization technique is the simplest form of power amplifier linearization and can be used for handset application, although significant linearity improvement is difficult to realize. Different types of the feedback linearization approaches, together with digital linearization techniques, are very attractive to be used in handset or base station transmitters. Finally, the potential semidigital and digital amplification approaches are discussed with their architectural advantages and problems in practical implementation.

Chapter 13 discusses the circuit schematics and main properties of the semiconductor control circuits that are usually characterized by small size, low power consumption, high-speed performance, and operating life. Generally, they can be built based on the *p–i–n* diodes, silicon MOSFET, or GaAs MESFET transistors and can be divided into two basic parts: amplitude and phase control circuits. The control circuits are necessary to protect high power devices from excessive peak voltage or dc current conditions. They are also used as switching elements for directing signal between different transmitting paths, as variable gain amplifiers to stabilize transmitter output power, as attenuators and phase shifters to change the amplitude and phase of the transmitting signal paths in array systems, or as limiters to protect power-sensitive components.

Finally, Chapter 14 describes the different types of radio transmitter architectures, history of radio communication, conventional types of radio transmission, and modern communication systems. Amplitude-modulated transmitters representing the oldest technique for radio communication are based on high- or low-level modulation methods, with particular case of an amplitude keying. Single-sideband transmitters as the next-generation transmitters could provide higher efficiency due to the transmission of a single sideband only. Frequency-modulated transmitters then became a revolutionary step to improve the quality of a broadcast transmission. TV transmitters include different modulation techniques for transmitting audio and video information, both analog and digital. Wireless communication transmitters as a part of the cellular technologies provide a worldwide wireless radio access. Radar transmitters are required for many commercial and military applications such as phased-array radars, automotive radars, or electronic warfare systems. Satellite transmission systems contribute to worldwide transmission of any communication signals through satellite transponders and offer communication for areas with any population density and location. Ultra-wideband transmission is very attractive for their low-cost and low-power communication applications, occupying a very wide frequency range.

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Andrei Grebennikov

INTRODUCTION

A vacuum-tube or solid-state radio transmitter is essentially a source of a radio-frequency (RF) signal to be transmitted through antenna in different radio systems such as wireless communication, television (TV) and broadcasting, navigation, radar, or satellite, the information format and electrical performance of which should satisfy the corresponding standard requirements. The transmission of radio signals is produced by modulation of different types, with different output power and transmission mode, and in different frequency ranges, from high frequencies to millimeter waves. Transmitters in which the power output is generated directly by the modulated source are considered as possessing high-level modulation systems. In contrast, arrangements in which the modulation takes place at a power level less than the transmitter output are referred to as low-level modulation systems.

Figure I.1 shows the simplified block diagram of a conventional radio transmitter intended to operate at radio and microwave frequencies, which consists of the following: the source of the information signal, which is usually amplified, filtered, or transformed to the intermediate frequency; the local oscillator and frequency multiplier, which establish the stabilized carrier frequency or some multiple of it; the RF modulator or mixer, which combines the signal and carrier frequency components to produce one of the varieties of the RF modulated waves; the power amplifier to deliver the RF modulated signal of required power level to the antenna; the antenna duplexer to separate and isolate transmitting and receiving paths. The power amplifier usually consists of cascaded gain stages, and each stage should have adequate linearity in the case of transmitting signal with variable amplitude corresponding to amplitude-modulated or multicarrier signals. In practice, there are many variations in transmitter architectures depending on the particular frequency bandwidth, output power, or modulation scheme.

Dr. Lee De Forest was an inventor who changed the world with electronics by inventing the vacuum tube, which he called the *audion*. In January 1907, De Forest filed a patent for an oscillation detector based on a three-electrode device representing a vacuum tube [1]. His pioneering innovation was the insertion of a third electrode (grid) in between the cathode (filament) and the anode (plate) of the previously invented diode. However, it was not until 1911 that De Forest built the first vacuum-tube amplifier based on three audions as amplifiers [2]. The original audion was capable of slightly amplifying received signals, but at this stage could not be used for more advanced applications such as radio transmitters. The inefficient design of the original audion meant that it was initially of little value to radio, and due to its high cost and short life it was rarely used. Eventually, vacuum-tube design was improved enough to make vacuum tubes more than novelties. Beginning in 1912, various researchers discovered that properly constructed (i.e., according to scientific and engineering principles) vacuum tubes could be employed in electrical circuits that made radio receivers and amplifiers thousands of times more powerful, and could also be used to make compact and efficient radio transmitters, which for the first time made radio broadcasting practical.

In 1914, the first vacuum-tube radio transmitters began to appear—a key technical development that would lead to the introduction of widespread broadcasting. Both amateurs and commercial firms started to experiment with the new vacuum-tube transmitters, employing them for a variety of purposes. Six years after suspending his efforts to make audio transmissions, when he had unsuccessfully

RF and Microwave Transmitter Design, First Edition. Andrei Grebennikov.

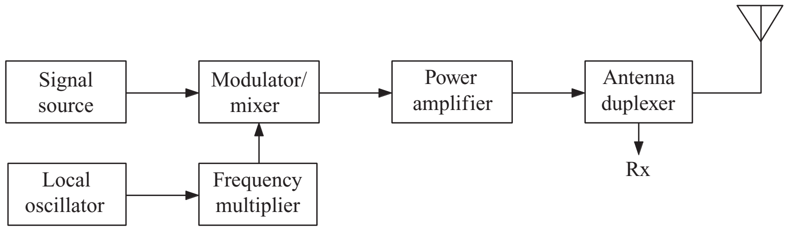

FIGURE I.1 Conventional transmitter architecture.

tried to use arc-transmitters, De Forest again took up developing radio to transmit sounds, including broadcasting news and entertainment, this time with much more success. He demonstrated that with this form of transmitter it was possible to telephone one to three miles, and by means of the small 3.5-V amplifier tube used with this apparatus, direct current could be transformed to alternating current at frequencies from 60 cycles per second to 1,000,000 cycles per second [3]. It is interesting that De Forest recognized the irony that he had overlooked the potential of developing his audion as a radio transmitter at the beginning. Reviewing his earlier arc-transmitter efforts, he wrote in his autobiography that he had been “totally unaware of the fact that in the little audion tube, which I was then using only as a radio detector, lay dormant the principle of oscillation which, had I but realized it, would have caused me to unceremoniously dump into the ash can all of the fine arc mechanisms which I had ever constructed, a procedure which a few years later actually took place all over the world” [4]. Meanwhile, in June 1915, the American Telephone & Telegraph Company installed a powerful experimental vacuum-tube transmitter in Arlington, Virginia, which quickly achieved remarkable distances for its audio transmissions [5]. The Marconi companies joined those experimenting with

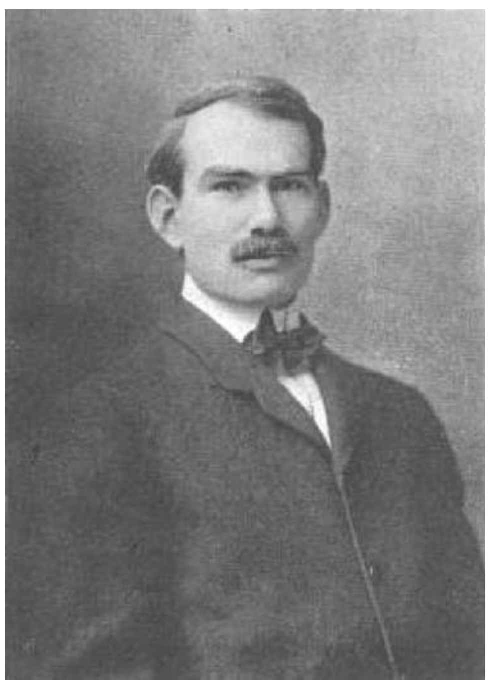

PHOTO I.1 Lee De Forest.

the new vacuum-tube transmitters, achieving an oversea working range of 50 km between ship aerials [6]. With the capable assistance of engineers including H. Round and C. Franklin, Guglielmo Marconi began experimenting with shortwave vacuum-tube transmitters by about 1916 [7].

However, Alexander Meissner was the first to amplify high-frequency radio signals by using a regenerative (positive) feedback in a vacuum triode. This principle of reactive coupling, which was given in its general form by Meissner in March 1913, later became the basis of the radio transmitter development [8]. The first high-frequency vacuum tube transmitters of small power up to 15 W were built by the Telefunken Company at the beginning of 1915. To provide high operating efficiency of the transmitter, the plate current of the special wave form, or else an auxiliary voltage of triple frequency, was impressed on the grid; thus greatly reducing the losses because, at the time when the highest voltages are applied to the tube, the passage of current through it is prevented. At the very beginning of 1920s, 1-kW transmitters had already been in use for radio telephony and telegraphy in several of the larger German cities. Much higher transmitting power was achieved by connecting in parallel eight or more tubes, each delivering 1.5 kW. In Russia during 1920, the 5.5-kW radio transmitter, where the modulated oscillations were amplified by a tube and transferred to the grids of six tubes in parallel that fed the antenna of 120-meter height, covered long distances of more than 4500 km [9]. An attempt in the field of radio television had been tried out in 1920 in order to provide a radio transmission of photographs with two antennas: one of which sends the synchronizing signal while the other sends the actual picture.

Edwin H. Armstrong is widely regarded as one of the foremost contributors to the field of radio engineering, being responsible for the regenerative circuit (1912), the superheterodyne circuit (1918), and the complete frequency-modulation radio broadcasting system (1933). Armstrong studied the audion for several years, performed extensive measurements, and understood and explained its operation when he devised a circuit, in which part of the current at the plate was fed back to the grid to strengthen the incoming signal. This discovery led to the independent invention of regeneration (or feedback) principle and the vacuum-tube oscillator. He then disproved the currently accepted theory

PHOTO I.2 Alexander Meissner.

PHOTO I.3 Edwin Armstrong.

of the action of the triode (three-electrode vacuum tube), and published the correct explanation in 1914 [10].

At the same time, the theoretical development of quantum mechanics during the 1920s played an important role in driving solid-state electronics, with understanding of the differences between metals, insulators, and semiconductors [11]. These continuing theoretical efforts then quickly led to the discovery of new devices, when Julius Lilienfeld invented the concept of a field-effect transistor in 1926 [12]. He believed that applying a voltage to a poorly conducting material would change its conductivity and thereby achieve amplification, but no one was able to do anything practically with this device until much later time. In conjunction with this, it is worth mentioning that the oscillating crystal detector was described by W. Eccles in 1909 and then practically implemented as an oscillator and even a low-power transmitter (based on one-port negative resistance principle) by O. Losev in early 1920s [13,14]. All details needed to duplicate these circuits to make a tunnel-diode oscillator were reported in the September 1924 issue of *Radio News* and in the 1st and 8th October 1924 issues of *Wireless World*, with predictions that crystals would someday replace valves in electronics.

Shortly after the end of the war in 1945, Bell Labs formed a Solid State Physics Group led by William Shockley, with an assignment to seek a solid-state alternative to fragile glass vacuum-tube amplifiers. This group made a very important decision right at the beginning: that the simplest semiconductors were silicon and germanium and that their efforts would be directed at those two elements. The first attempts were based on Shockley's ideas about using an external electrical field on a semiconductor to affect its conductivity, but these experiments failed every time in all sorts of configurations and materials. The group was at a standstill until John Bardeen suggested a theory that invoked surface states that prevented the field from penetrating the semiconductor. In November 1947, John Bardeen and Walter Brattain, working without Shockley, succeeded in creating a point-contact transistor that achieved amplification when electrical field was applied to a crystal of germanium [15]. At the same time, Shockley secretly continued and successfully finished his own work to build a different sort of transistor based on *n–p* junctions that depended on the introduction of the minority carriers instead of point contacts, which he expected would be more

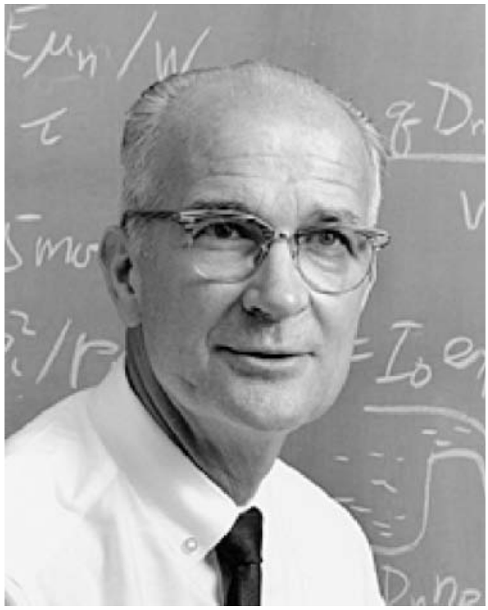

PHOTO I.4 William Shockley.

likely to be commercially viable [16]. In his seminal work *Electrons and Holes in Semiconductors with Applications to Transistor Electronics* (1955), Shockley worked out the critical ideas of drift and diffusion and the differential equations that govern the flow of electrons in solid-state crystals, where the Shockley ideal diode equation was also described. The term "transistor" was coined by John Pierce, who later recalled: ". . . at that time, it was supposed to be the dual of the vacuum tube. The vacuum tube had *transconductance*, so the transistor would have *transresistance*. And the name should fit in with the names of other devices, such as *varistor* and *thermistor*. And . . . I suggested the name *transistor*."

The first to perceive the possibility of integrated circuits based upon semiconductor technology was Geoffrey Dummer, who said addressing the Electronic Components Conference in 1952: "With the advent of the transistor and the work in semiconductors generally, it seems now possible to envisage electronic equipment in a solid block with no connecting wires. The block may consist of layers of insulating, conducting, rectifying, and amplifying materials, the electrical functions being connected directly by cutting out areas of the various layers" [17]. In 1958, Jack Kilby of Texas Instruments developed the first integrated circuit consisting of a few mesa transistors, diffused capacitors, and bulk resistors on a piece of germanium using gold wires for interconnections [18]. The integrated circuit developed by Robert Noyce of Fairchild could resolve problems with wires by adding a final metal layer and then taking away some of it so that the wires required for the components to be connected were shaped so as to make the integrated circuit more suitable for mass production [19]. A year later, a planar process was developed by Jean Hoerni that utilized heat diffusion process, and oxide passivation of the surface protected the junctions and provided a reproducibility that assured more consistency than any previous manufacturing process. The first microwave gallium–arsenide (GaAs) Schottky-gate field-effect transistor (metal semiconductor field-effect transistor or MESFET), which had a maximum frequency $f_{\max} = 3$ GHz, was reported in 1967 [20]. But it was not until 1976 that the first fully monolithic single-stage GaAs MESFET X-band broadband amplifier was developed based on lumped matching elements [21]. Eight years later, on a GaAs chip of approximately the same area, an entire X-band transmit–receive (T/R) module was fabricated, consisting of two switches, a four-bit

TABLE I.1 IEEE Standard Letter Designations for Frequency Bands.

Band	Frequency	Wavelength
HF (high frequency)	3–30 MHz	100–10 m
VHF (very high frequency)	30–300 MHz	10–1.0 m
UHF (ultra high frequency)	300–1000 MHz	1.0 m to 30 cm
L	1–2 GHz	30–15 cm
S	2–4 GHz	15 cm to 7.5 cm
C	4–8 GHz	7.50–3.75 cm
X	8–12 GHz	3.75–2.50 cm
Ku	12–18 GHz	2.50–1.67 cm
K	18–27 GHz	1.67–1.11 cm
Ka	27–40 GHz	1.11 cm to 7.5 mm
V	40–75 GHz	7.5–4.0 mm
W	75–110 GHz	4.0–2.7 mm
Millimeter wave	110–300 GHz	2.7–1.0 mm

phase shifter, a three-stage low-noise amplifier, and a four-stage power amplifier [22]. Generally, a monolithic microwave integrated circuit (MMIC) was defined as an active or passive microwave circuit formed in situ on a semiconductor substrate by a combination of deposition techniques including diffusion, evaporation, epitaxy, and other means [23].

Table I.1 shows the IEEE standard letter designations for frequency bands [24]. In military radar band designations, millimeter-wave bandwidth occupies the frequency range from 40 to 300 GHz. The letter designations (L, S, C, X, Ku, K, Ka) were meant to be used for radar, but have become commonly used for other microwave frequency applications. The K-band is the middle band (18–27 GHz) that originated from the German word "Kurz", which means short, while Ku-band is lower in frequency (Kurz-under), and Ka-band is higher in frequency (Kurz-above). The 1984 revision defined the application of letters V and W to a portion of the millimeter-wave region each while retaining the previous letter designators for frequencies.

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1 Passive Elements and Circuit Theory

The two-port equivalent circuits are widely used in radio frequency (RF) and microwave circuit design to describe the electrical behavior of both active devices and passive networks [1–4]. The two-port network impedance Z-parameters, admittance Y-parameters, or hybrid H-parameters are very important to characterize the nonlinear properties of the active devices, bipolar or field-effect transistors. The transmission $ABCD$-parameters of a two-port network are very convenient for designing the distributed circuits like transmission lines or cascaded elements. The scattering S-parameters are useful to characterize linear circuits, and are required to simplify the measurement procedure. Transmission lines are widely used in matching circuits in power amplifiers, in resonant circuits in the oscillators, filters, directional couplers, power combiners, and dividers. The design formulas and curves are presented for several types of transmission lines including stripline, microstrip line, slotline, and coplanar waveguide. Monolithic implementation of lumped inductors and capacitors is usually required at microwave frequencies and for portable devices. Knowledge of noise phenomena, such as the noise figure, additive white noise, low-frequency fluctuations, or flicker noise in active or passive elements, is very important for the oscillator modeling in particular and entire transmitter design in general.

1.1 IMMITTANCE TWO-PORT NETWORK PARAMETERS

The basic diagram of a two-port nonautonomous transmission system can be represented by the equivalent circuit shown in Figure 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. The two independent phasor currents I_1 and I_2 (flowing across input and output terminals) and phasor voltages 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 a power amplifier or 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 Figure 1.1. Then the two linearly independent equations, which describe the general two-port network

RF and Microwave Transmitter Design, First Edition. Andrei Grebennikov.

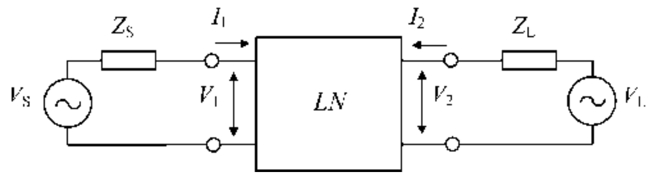

FIGURE 1.1 Basic diagram of two-port nonautonomous transmission system.

in terms of circuit variables V_1, V_2, I_1, and I_2, can be expressed in a matrix form as

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

or

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

The complex 2 × 2 matrices $[M]$ and $[N]$ in Eq. (1.3) are independent of the source and load impedances Z_S and Z_L and voltages V_S and V_L, respectively, and 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 a 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 can be found by setting $I_2 = 0$ in Eqs. (1.6) and (1.7), resulting 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

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

whereas 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 a 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), thus creating a short output terminal. Similarly, the same current components I_1 and I_2 are determined by setting $V_1 = 0$ when input terminal becomes short-circuited. As a result, the two driving-point admittances are

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

whereas 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 redefined in order to express the voltage source V_1 and output current I_2 in terms of the input current I_1 and output voltage V_2. If the submatrix

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

given in Eq. (1.4) 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 a 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 setting $V_2 = 0$ for the short output terminal in Eqs. (1.16) and (1.17) as

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

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 setting $I_1 = 0$ when input terminal becomes open-circuited as

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

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

The transmission parameters, often used for passive device analysis, are determined for the independent input voltage source V_1 and input current I_1 in terms of the output voltage V_2 and output current I_2. In this case, if the submatrix

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

given in Eq. (1.4) is nonsingular, we obtain

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

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

$$V_1 = AV_2 - BI_2 \tag{1.21}$$

$$I_1 = CV_2 - DI_2 \tag{1.22}$$

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

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

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

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

where B is the reverse transfer impedance and D is the reverse current transfer function. The reason a minus sign is associated with I_2 in Eqs. (1.20) to (1.22) is that historically, for transmission networks, the input signal is considered as flowing to the input port whereas the output current flowing to the load. The direction of the current $-I_2$ entering the load is shown in Figure 1.2.

The parameters describing the same two-port network through different two-port matrices (impedance, admittance, hybrid, or transmission) can be cross-converted, and the elements of each

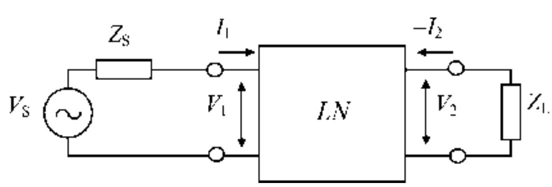

FIGURE 1.2 Basic diagram of loaded two-port transmission system.