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Power Combiners, Impedance Transformers and Directional Couplers

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This is the first of a multi-part article that provides a textbook-style review of an important group of RF circuits used in applications such as power amplifiers, antenna systems and measurement systems

Many RF applications require power combiners or dividers, impedance transformers and directional couplers. In the case of combiners, it is critical, particularly at higher frequencies, that the correct types are used to achieve the desired power performance when combing individual active devices to achieve higher power.

The methods for configuration of the combiners or dividers differ, depending on the operating frequency, frequency bandwidth, output power, and size requirements. Coaxial cable combiners with ferrite cores are often used to combine the output powers of power amplifiers intended for wideband applications. The device output impedance is usually low at high power levels; so, to match this impedance with a standard 50-ohm load, coaxial-line transformers with specified impedance transformation are used. For narrow-band applications, the N-way Wilkinson combiners are widely used due to their simple practical realization. For microwaves, the size of combiners should be very small and, therefore, the hybrid microstrip combiners (including different types of the microwaves hybrids and directional couplers) are commonly used to combine output powers of power amplifiers or oscillators. In this paper, a variety of different combiners, impedance transformers and directional couplers for application in RF and microwave transmitters is given with descriptions of their schematics and operational principles.

Transmission-Line Transformers and Combiners

The transmission-line transformers and combiners can provide very wide operating bandwidths and operate up to frequencies of 3 GHz and higher [1, 2]. They are widely used in matching networks for antennas and power amplifiers in the HF and VHF bands, in mixer circuits, and their low losses make them especially useful in high power circuits [3, 4]. Typical structures for transmission-line transformers consist of parallel wires, coaxial cables or bifilar twisted wire pairs. In the latter case, the characteristic impedance can easily be determined by the wire diameter, the insulation thickness, and, to some extent, the twisting pitch [5, 6]. For coaxial cable transformers with correctly chosen characteristic impedance, the theoretical high frequency bandwidth limit is reached when the cable length comes in order of a half wavelength, with the overall achievable bandwidth being about a decade. By introducing the low-loss high permeability ferrites alongside a good quality semi-rigid coaxial or symmetrical strip cable, the low frequency limit can be significantly improved providing bandwidths of several or more decades.

The concept of a broadband impedance transformer consisting of a pair of interconnected transmission lines was first disclosed and described by Guanella [7, 8]. Figure 1(a) shows a Guanella transformer system with transmission line character achieved by an arrangement comprising one pair of cylindrical coils that are wound in the same sense and are spaced a certain distance apart by an intervening dielectric. In this case, one cylindrical coil is located inside the insulating

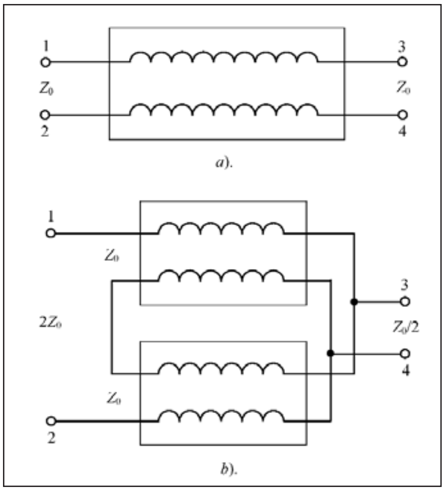

Figure 1 · Schematic configurations of Guanella 1:1 and 4:1 transformers.

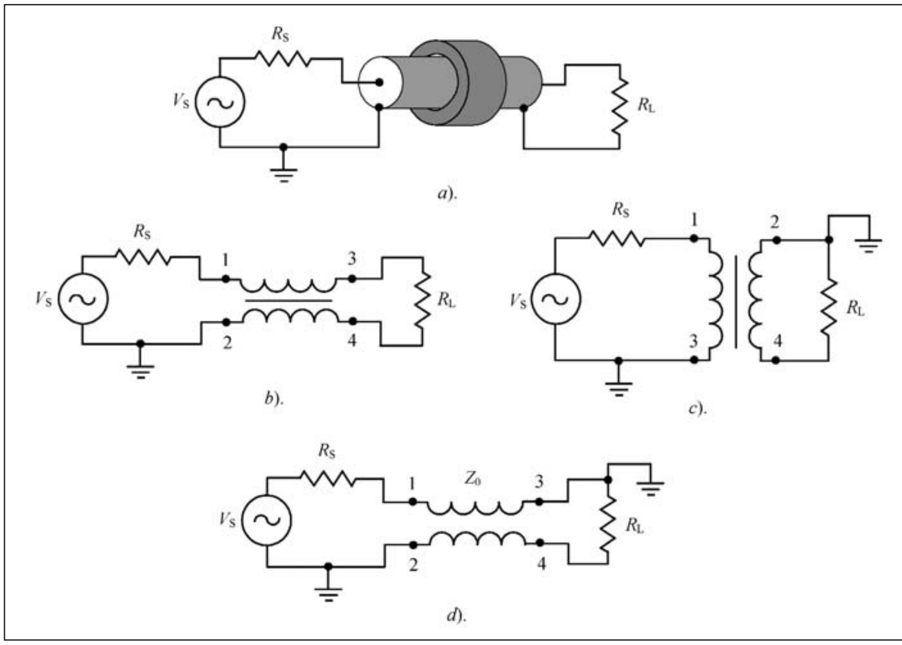

Figure 2 · Schematic configurations of a coaxial cable transformer.

cylinder and the other coil is located on the outside of this cylinder. For the currents flowing through both windings in opposite directions, the corresponding flux in the coil axis is negligibly small. However, for the currents flowing in the same direction through both coils (common-mode), the latter may be assumed to be connected in parallel, and a coil pair represents a considerable inductance for such currents and acts like a choke coil. With terminal 4 being grounded, such a 1:1 transformer provides matching of the balanced source to unbalanced load and is called a *balun* (balanced-to-unbalanced transformer). In this case, if terminal 2 is grounded, it represents simply a delay line. In a particular case, when terminals 2 and 3 are grounded, the transformer performs as a phase inverter. A series-parallel connection of a plurality of these coil pairs can produce a match between unequal source and load resistances.

Figure 1(b) shows a 4:1 impedance (2:1 voltage) transmission-line transformer where the two pairs of cylindrical transmission line coils are connected in series at the input and in parallel at the output. For the characteristic impedance Z_0 of each transmission line, this results in the two times higher impedance $2Z_0$ at the input and two times lower impedance $Z_0/2$ at the output. By grounding terminal 4, such a 4:1 impedance transformer provides impedance matching of the balanced source to the unbalanced load. In this case, when terminal 2 is grounded, it performs as a 4:1 *unun* (unbalanced-to-unbalanced transformer). With a series-parallel connection of n coil pairs, each having the characteristic impedance Z_0, the input impedance is equal to nZ_0 and the output impedance is equal to Z_0/n. Since Guanella adds voltages that have equal delays through the transmission lines, such a technique results in the so called *equal-delay* transmission-line transformers.

The simplest transmission-line is a quarter-wave transmission line whose characteristic impedance is chosen to give the correct impedance transformation. However, this transformer provides a narrow-band performance valid only around frequencies for which the transmission line is odd multiples of a quarter wavelength. If a ferrite sleeve is added to the transmission line, common-mode currents—flowing in both transmission line inner and outer conductors in phase, and in the same direction—are suppressed, and the load may be balanced and floating above ground or balanced with a center tap grounded load, thus operating as a balun [9, 10]. If the characteristic impedance of the transmission line is equal to the terminating impedances, the transmission is inherently broadband. If not, there will be a dip in the response at the frequency at which the transmission-line is a quarter-wavelength long.

A coaxial cable transformer with the physical configuration and equivalent circuit representation shown in Figures 2(a) and 2(b), respectively, consists of the coaxial line arranged inside the ferrite core or wound around the ferrite core. Due to its practical configuration, the coaxial cable transformer takes a position between the lumped and distributed systems. Therefore, at lower frequencies its equivalent circuit represents a conventional polarity reversing low-frequency transformer shown in Figure 2(c), while at higher frequency it is a transmission line with the

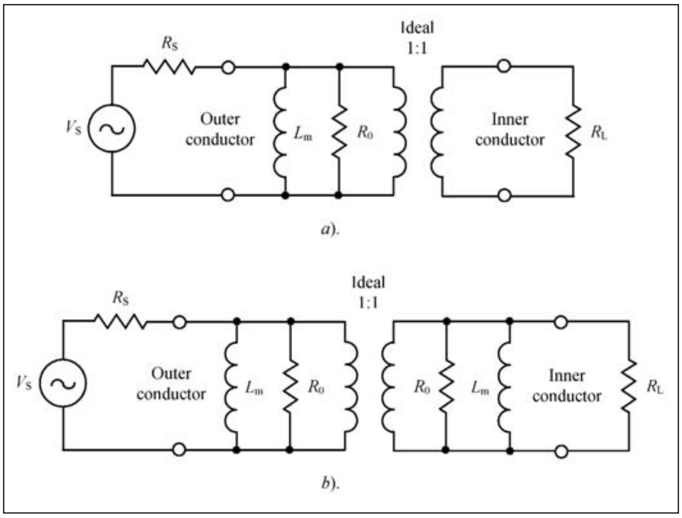

Figure 3 · Low frequency models of a 1:1 coaxial cable transformer.

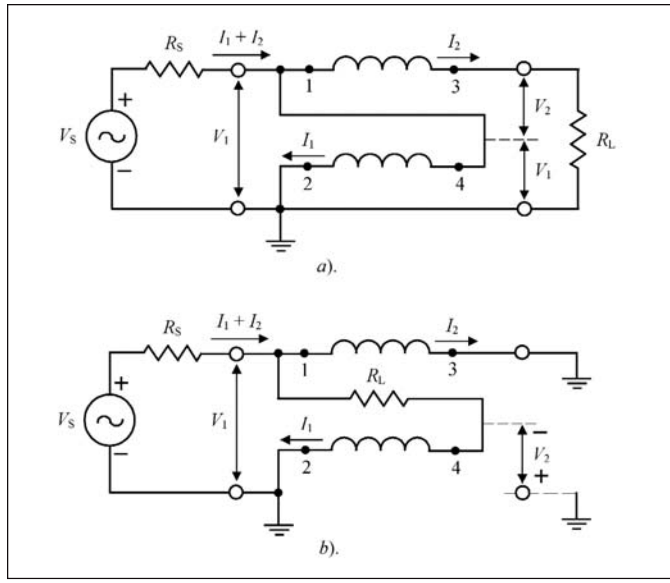

Figure 4 · Schematic configurations of the Ruthroff 1:4 impedance transformer.

characteristic impedance Z_0 shown in Figure 2(d). The advantage of such a transformer is that the parasitic interturn capacitance determines its characteristic impedance, whereas in the conventional wire-wound transformer with discrete windings this parasitic capacitance has a negative effect on the transformer frequency response performance.

When $R_S = R_L = Z_0$, the transmission line can be considered a transformer with a 1:1 impedance transformation. To avoid any resonant phenomena, especially for complex loads, which can contribute to the significant output power variations, as a general rule, the length l of the transmission line is kept to no more than one-eighth of wavelength $\lambda_{\min}$, or

$$l \le \frac{\lambda_{\min}}{8} \tag{1}$$

where $\lambda_{\min}$ is the minimum wavelength in the transmission line corresponding to the highest operating frequency $f_{\max}$.

The low-frequency bandwidth limit of a coaxial cable transformer is determined by the effect of the magnetizing inductance L_m of the outer surface of the outer conductor according to the equivalent low-frequency transformer model shown in Figure 3(a), where the transmission line is represented by the ideal 1:1 transformer [4]. The resistance R_0 represents the losses of the transmission line. An approximation to the magnetizing inductance can be made by considering the outer surface of the coaxial cable to be the same as that of a straight wire (or linear conductor), which, at higher frequencies where the skin effect causes the current to be concentrated on the outer surface, would have the self-inductance of

$$L_m = 2l\left[\ln\left(\frac{2l}{r}\right) - 1\right]\text{nH} \tag{2}$$

where l is the length of the coaxial cable in cm and r is the radius of the outer surface of the outer conductor in cm [4].

The use of high permeability core materials results in shorter transmission lines. If a toroid is used for the core, the magnetizing inductance L_m is obtained by

$$L_m = 4\pi n^2 \mu \frac{A_e}{L_e}\text{nH} \tag{3}$$

where n is the number of turns, μ is the core permeability, A_e is the effective cross-sectional area of the core in cm^2, and L_e is the average magnetic path length in cm [11].

Considering the transformer equivalent circuit shown in Fig. 2(a), the ratio between the power delivered to the load P_L and power available at the source $P_s = V_s^2/8R_s$ when $R_S = R_L$ can be obtained from

$$\frac{P_L}{P_S} = \frac{(2\omega L_m)^2}{R_S^2 + (2\omega L_m)^2} \tag{4}$$

which gives the minimum operating frequency $f_{\min}$ for a given magnetizing inductance L_m, taking into account the maximum decrease of the output power by 3 dB, as

$$f_{\min} \ge \frac{R_S}{4\pi L_m} \tag{5}$$

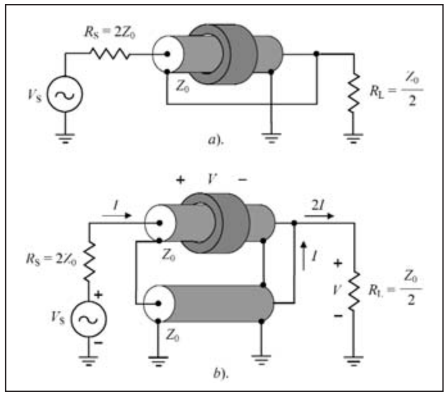

Figure 5 · Schematic configurations of a 4:1 coaxial cable transformer.

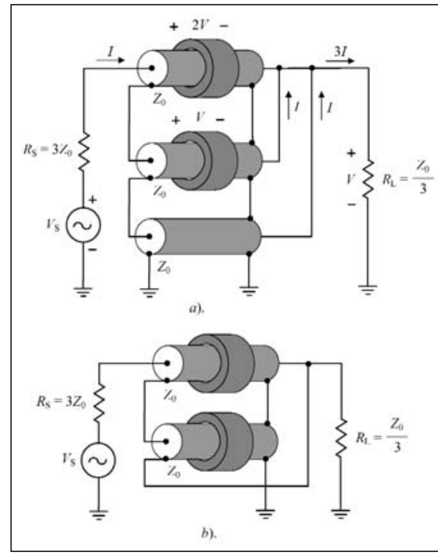

Figure 6 · Schematic configurations of a 9:1 coaxial cable transformer.

A similar low-frequency model for a coaxial cable transformer using twisted or parallel wires is shown in Figure 3(b) [4]. Here, the model is symmetrical as both conductors are exposed to any magnetic material and therefore contribute identically to the losses and low-frequency performance of the transformer.

An approach using a transmission line based on a single bifilar wound coil to realize a broadband 1:4 impedance transformation was introduced by Ruthroff [12, 13]. In this case, by using a core material of sufficiently high permeability, the number of turns can be significantly reduced. Figure 4(a) shows the circuit schematic of an unbalanced-to-unbalanced 1:4 transmission line transformer where terminal 4 is connected to the input terminal 1. As a result, for $V = V_1 = V_2$, the output voltage is twice the input voltage, and the transformer has a 1:2 voltage step-up ratio. As the ratio of input voltage to input current is one-fourth the load voltage to load current, the transformer is fully matched for maximum power transfer when $R_L = 4R_S$, and the characteristic impedance of the transmission line Z_0 is equal to the geometric mean of the source and load impedances:

$$Z_0 = \sqrt{R_S R_L} \tag{6}$$

where R_S is the source resistance and R_L is the load resistance.

Figure 4(b) shows an impedance transformer acting as a phase inverter, where the load resistance is included between terminals 1 and 4 to become a 1:4 balun. This technique is called the bootstrap effect and doesn't have the same high frequency response as Guanella equal-delay approach because it adds a delayed voltage to a direct one [14]. The delay becomes excessive when the transmission line reach a significant fraction of a wavelength.

Figure 5(a) shows the physical implementation of the 4:1 impedance Ruthroff transformer using a coaxial cable arranged inside the ferrite core. At lower frequencies, such a transformer can be considered an ordinary 2:1 voltage autotransformer. To improve the performance at higher frequencies, it is necessary to add an additional phase-compensating line of the same length shown in Fig. 5(b), resulting in a Guanella ferrite-based 4:1 impedance transformer. In this case, a ferrite core is necessary only for the upper line because the outer conductor of the lower line is grounded at both ends and no current is flowing through it. A current I driven into the inner conductor of the upper line produces a current I that flows in the outer conductor of the upper line, resulting in a current $2I$ flowing into the load R_L. Because the voltage $2V$ from the transformer input is divided in two equal parts between the coaxial line and the load, such a transformer provides impedance transformation from $R_S = 2Z_0$ into $R_L = Z_0/2$, where Z_0 is the characteristic impedance of each coaxial line. The bandwidth extension for the Ruthroff transformers can also be achieved by using transmission lines with step-function and exponential changes in their characteristic impedances [15, 16]. To adopt this transmission line transformer for microwave planar applications, the coaxial line can be replaced by a pair of stacked strip conductors or coupled microstrip lines [17, 18].

Figure 6 shows similar arrangements for the 3:1 voltage coaxial cable transformers, which produce 9:1 impedance transformation. A current I driven into the inner conductor of the upper line in Figure 6(a) will cause a current I to flow in the outer conductor of the upper line. This current then produces a current I in the outer conductor of the lower line, resulting in a current $3I$ flowing into the load R_L. The lowest coaxial line can be removed, resulting in a 9:1 impedance coaxial cable transformer shown in Figuer 6(b). The characteristic impedance of each transmission line is specified by the voltage applied to the end of the line and the current flowing through the line and is equal to Z_0.

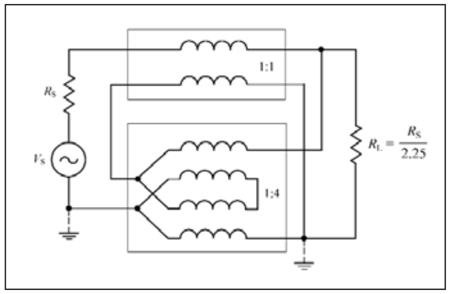

Figure 7 · Schematic configuration of an equal-delay 2.25:1 unun.

By using the transmission-line baluns with different integer-transformation ratios in certain connection, it is possible to obtain the fractional-ratio baluns and ununs [2, 19, 20]. Figure 7 shows a transformer configuration for obtaining an impedance ratio of 2.25:1, which consists of a 1:1 Guanella balun on the top combined with a 1:4 Guanella balun where voltages on the left-hand side are in series and on the right-hand side are in parallel [19]. In this case, the left-hand side has the higher impedance. In a matched condition, this transformer should have a high frequency response similar to a single transmission line. By

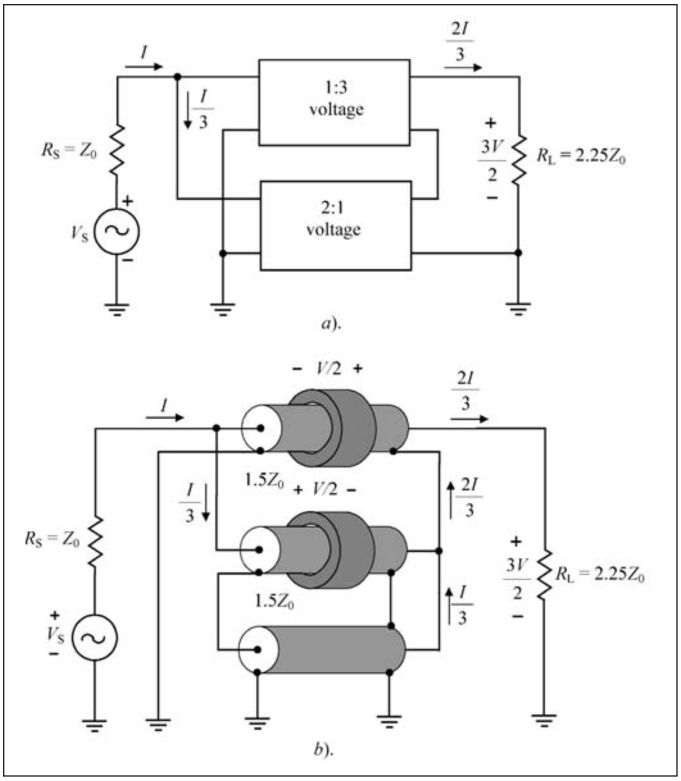

Figure 8 · Schematic configurations of a fractional 1:2.25 impedance transformer.

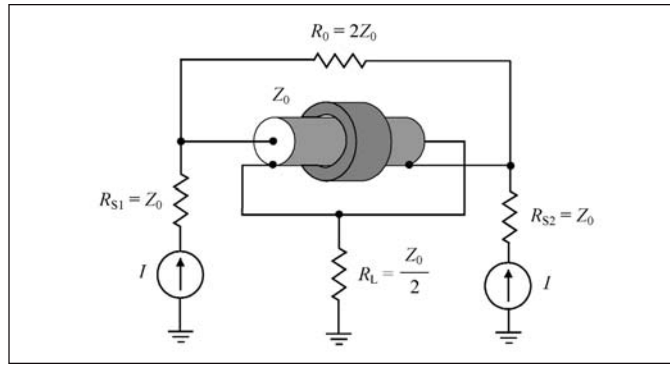

Figure 9 · Coaxial cable combiner.

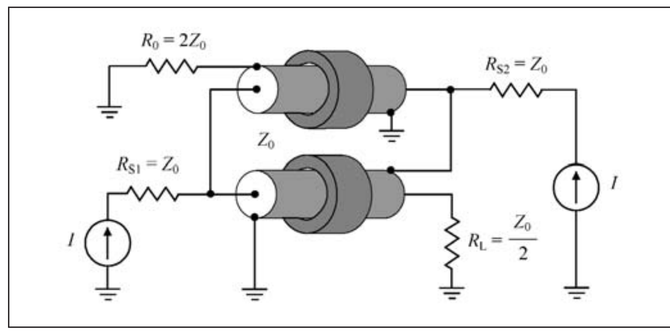

Figure 10 · Two-cable hybrid combiner with grounded ballast resistor and load.

grounding the corresponding terminals (shown by dashed line), it becomes a broadband unun. Different ratios can be obtained with other configurations. For example, using a 1:9 Guanella balun below the 1:1 unit results in a 1.78:1 impedance ratio, whereas, with a 1:16 balun, the impedance ratio becomes 1.56:1.

On the other hand, the overall 1:1.5 voltage transformer configuration can be achieved by using the cascade connection of a 1:3 voltage transformer to increase the impedance by 9 times, and a 2:1 voltage transformer to decrease the impedance by 4 times, which block schematic is shown in Figure 8(a) [20]. The practical configuration using coaxial cables and ferrite cores is shown in Figure 8(b). Here, the currents $I/3$ in the inner conductors of two lower lines cause an overall current $2I/3$ in the outer conductor of the upper line, resulting in a current $2I/3$ flowing into the load R_L. A load voltage $3V/2$ is out of phase with a longitudinal voltage $V/2$ along the upper line, resulting in a voltage V at the transformer input. The lowest line also can be eliminated with direct connection of the points at both ends of its inner conductor, as in the case of the 2:1 and 3:1 Ruthroff voltage transformers shown in Figs. 5(a) and 6(b), respectively. If the source impedance is 50 ohms, then the characteristic impedance of all three transmission lines should be 75 ohms. In this case, the matched condition corresponds to a load impedance of 112.5 Ω.

By using the coaxial cable transformers, the output powers from two or more power sources can be combined. Figure 9 shows an example of such a transformer, combining two in-phase signals when both signals are delivered to the load R_L and no signal will be dissipated in the ballast resistor R_0 if their amplitudes are equal [12]. The main advantage of this transformer is the zero longitudinal voltage along the line for equal input powers; as a result, no losses occur in the ferrite core. When one input signal source (for example power amplifier) defaults or disconnects, the longitudinal voltage becomes equal to half a voltage of another input source. For this transformer, it is possible to combine two out-of-phase signals when the ballast resistor is considered the load, and the load resistor in turn is considered the ballast resistor. The schematic of another hybrid coaxial cable transformer using as a combiner is shown in Figure 10. The advantage of this combiner is that both the load R_L and the ballast resistor R_0 are grounded. These hybrid transformer-based combiners can also be used for the power division when the output power from a single source is divided and delivered into two independent loads. In this case, the original load and the two signal sources should be switched. As it turns out, the term “hybrid” comes not from the fact that the transformer might be constructed

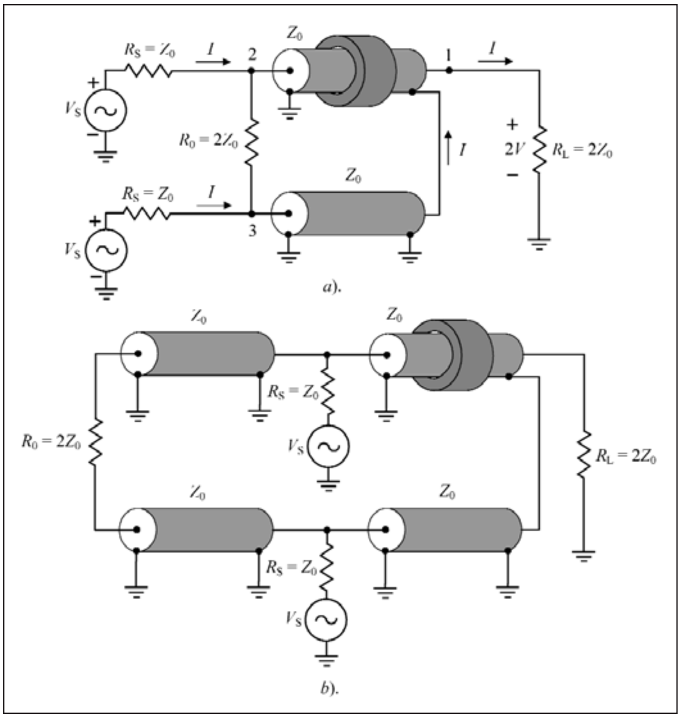

Figure 11 · Coaxial cable combiners with increased isolation.

of two different entities (for example, cable and resistor), but just because it is being driven by two signals as opposed to only one. Consequently, the hybrid transformer represents a four-port device having two input ports, one sum port and one difference port. The unique characteristic of the hybrid transformer is its ability to isolate the two input signal sources.

Figure 11(a) shows a coaxial cable two-way combiner where the input signals having the same amplitudes and phases at ports 2 and 3 are matched at higher frequencies when all lines are of the same lengths and $R_S = Z_0 = R_L/2 = R_0/2$ [2]. In this case, the isolation C_{23} between these input ports can be calculated by

$$C_{23} = 10\log_{10}\left[4\left(1+4\cot^2\theta\right)\right] \text{ dB} \tag{7}$$

where θ is the electrical length of each transmission line. In order to improve the isolation, the symmetrical ballast resistor R_0 should be connected through two additional lines, as shown in Figure 11(b), where all transmission lines have the same electrical lengths.

Figure 12 shows a coaxial cable two-way combiner that is fully matched and isolated in pairs [2]. Such combiners can be effectively used in high power broadcasting VHF FM and VHF-UHF TV transmitters. In this case, for power amplifiers with the identical output impedances R_{S1} and R_{S2} when $R_{S1} = R_{S2} = Z_0/2$, it is necessary to choose the values of the ballast resistor R_0 and the load

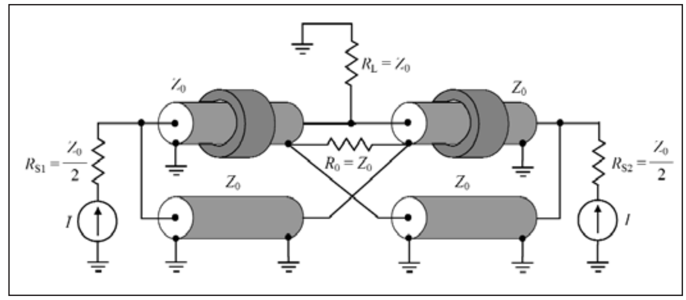

Figure 12 · Fully matched and isolated coaxial cable combiner.

R_L of $R_0 = R_L = Z_0$, where Z_0 is the characteristic impedance of the each transmission line of the same length.

Baluns

Baluns are very important elements in the design of mixers, push-pull amplifiers, or oscillators to link a symmetrical (balanced) circuit to an asymmetrical (unbalanced) circuit. Therefore, it makes sense to discuss their circuit configurations and performance in details separately. The main requirements to baluns are to provide an accurate 180-degree phase shift over required frequency bandwidth, with minimum loss and equal balanced impedances. In power amplifiers and oscillators, lack of symmetry will degrade output power and efficiency. Besides, the symmetrical port must be well isolated from ground to minimize an unwanted effect of parasitic capacitances.

A wire-wound transformer with a simplified equivalent schematic, shown in Figure 13(a), provides an excellent broadband balun covering in commercial applications frequencies from low kHz to beyond 2 GHz. They are usually realized with a center-tapped winding that provides a short circuit to even-mode (common-mode) signals while having no effect on the differential (odd-mode) signal. Wire-wound transformers are more expensive than the printed or lumped *LC* baluns, which are more suitable in practical mixer designs. However, unlike wire-wound transformers, the lumped *LC* baluns are narrow-band as containing the resonant elements.

Figure 13(b) shows the circuit schematic of a lattice-type *LC* balun that was proposed long ago for combining powers in push-pull amplifiers and their delivery to antenna [21]. It consists of two capacitors and two inductors, which produce the ±90-degree phase shifts at the output ports. The values of identical inductances L and capacitances C can be obtained by

$$L = \frac{\sqrt{R_{out}R_L}}{\omega_0} \tag{8}$$

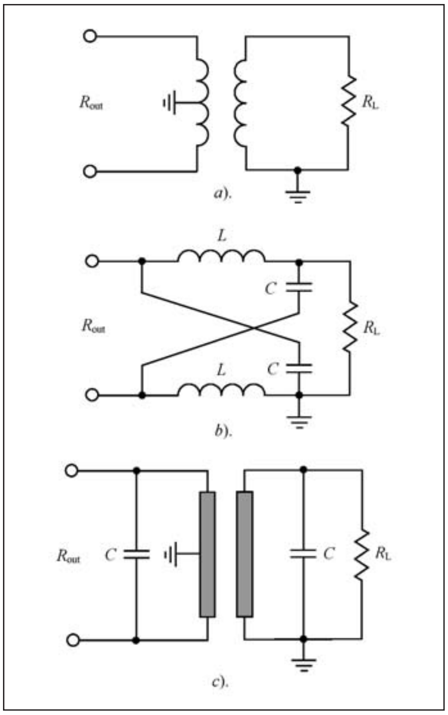

Figure 13 · Different circuit configurations of 1:1 balun.

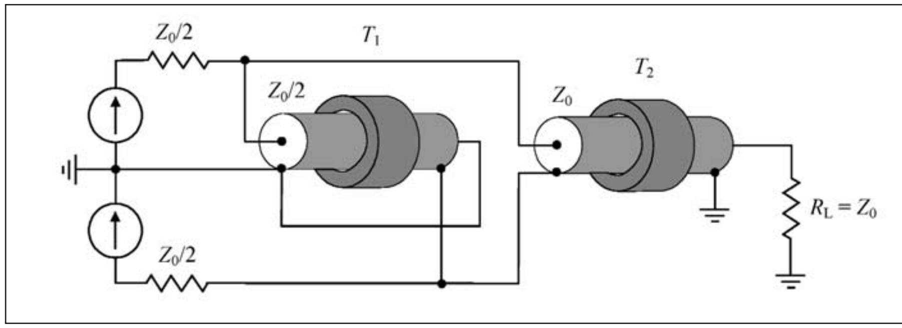

Figure 14 · Circuit arrangement with two cable transformers for push-pull operation.

$$C = \frac{1}{\omega_0 \sqrt{R_{out} R_L}} \tag{9}$$

where ω_0 is the center bandwidth frequency, R_{out} is the balanced output resistance, and R_L is the unbalanced load resistance. When designing this circuit, it is important to be confident that the operating frequency is well below the self-resonant frequencies of their components.

In monolithic microwave applications where the lumped inductances are usually replaced by transmission lines, the designs with microstrip coupled lines, Lange couplers, or multilayer coupled structures are very popular. However, the electrical length of the transmission lines at center bandwidth frequency is normally set to a quarter-wavelength, which is too large for applications in wireless communication systems. Therefore, it is very attractive to use the lumped-distributed balun structures, which can significantly reduce the balun size and, at the same time, can satisfy the required electrical characteristics. Figure 13(c) shows such a compact balun with lumped-distributed structure consisting of the two coupled planar microstrip lines and two parallel capacitors, where the input transmission line is grounded at midpoint and the output transmission line is grounded at its one port [22]. Without these capacitors, it is necessary to leave a very small spacing between quarter-wave microstrip lines to achieve a 3-dB coupling between them. However, by optimizing the balun elements around the center bandwidth of 900 MHz, the planar structure of approximately one-sixteenth the size of the conventional quarter-wavelength structure was realized, with spacing S = 8 mils using an FR4 board with substrate thickness of 300 mils.

Figure 14 shows the circuit

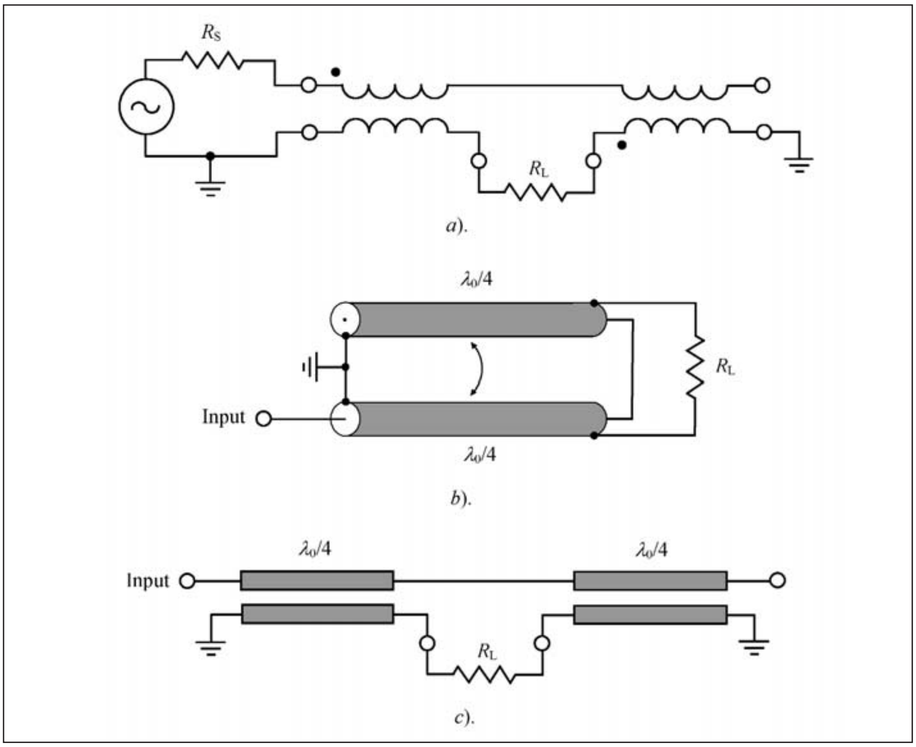

Figure 15 · Schematic configurations of Marchand balun.

arrangement with two coaxial line transformers combined to provide a push-pull operation of the power amplifier by creating a balanced-to-unbalanced impedance transformation with higher spectral purity. Ideally, the out-of-phase RF signals from both active devices will have pure half-sinusoidal waveforms, which contain (according to the Fourier series expansion) only fundamental and even harmonic components. This implies a 180-degree shift between fundamental components from both active devices and in-phase condition for remaining even harmonic components. In this case, the transformer T_1 representing a phase inverter is operated as a filter for even harmonics because currents flow through its inner and outer conductors in opposite directions. For each fundamental flowing through its inner and outer conductors in the same directions, it works as an RF choke, the impedance of which depends on the core permeability. Consequently, since the transformer T_2 represents a 1:1 balun, in order to provide maximum power delivery to the load R_L, the output equivalent resistance of each active device should be two times smaller.

For a simple 1:1 transmission-line balun realized with a twisted wire pair or coaxial cable, the balanced end is isolated from ground only at the center bandwidth frequency. To compensate for the short-circuited line reactance over certain frequency bandwidth around center frequency, a series open-circuited transmission line was introduced by Marchand, resulting in a compensated balun, the simplified schematic of which is shown in Figure 15(a) [23]. In this case, at center bandwidth frequency when the electrical length of the compensated line is a quarter-wavelength, the load resistance R_L is seen unchanged. When this structure is realized with coaxial cables, to eliminate unwanted current existing in the outer conductor and corresponding radiation, it is necessary to additionally provide the certain coupling

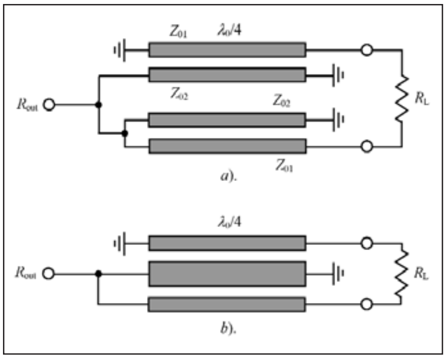

Figure 16 · Schematic configurations of coupled-line Marchand balun.

between the coaxial cables forming a transmission line with two outer conductors, as shown in Figure 15(b) [24]. Generally, the shunting reactance of this compensating line can reduce the overall balun reactance about center frequency or reverse its sign depending on the balanced load resistance, characteristic impedance of the compensating line and coupling (characteristic impedance) between the outer conductors of two lines. Hence, a compensating line can create a complementary reactance to a balanced load and provide an improved match over broader frequency range. At microwaves, wire-wound or coaxial cable transformers are usually replaced by a pair of the quarterwave coupled transmission lines shown in Figure 15(c), thus resulting in a compact planar structure. It should be noted that generally the characteristic impedances of the coaxial or coupled transmission lines can be different to optimize the frequency-bandwidth response.

Multilayer configurations make the Marchand balun even more compact and can provide wide bandwidths due to the tight coupling between coupled-line sections. Modeling and synthesis results of a two-layer monolithic Marchand balun configuration with two-coupled lines, the basic structure of which is shown in Figure 16(a), is discussed in [25]. In this configuration, the unbalanced terminal is connected to the microstrip line located at the upper metallization level, whereas the balanced load is connected to the microstrip lines located at the lower metallization level. The transmission lines sections in different layers are not isolated from each other. It should be noted that, for a given set of the output balanced and load unbalanced resistances R_{out} and R_L, the characteristic impedances of the outer and inner microstrip lines Z_{01} and Z_{02} are not unique, and they can be calculated from

$$C = \frac{1}{2}\sqrt{\frac{Z_{02}}{Z_{01}}} \tag{10}$$

$$Z_{02} = 4Z_{01} - \frac{R_{out}R_L}{Z_{01}} \tag{11}$$

where C is the coupling factor [26]. However, a different choice of Z_{01} and Z_{02} leads to a different frequency bandwidth. For example, for R_{out} = 50 ohms and R_L = 100 ohms, it was found that using the symmetrical directional coupler with Z_{01} = Z_{02} = 40.825 ohms results in a frequency bandwidth of 48.4% with $|S_{11}| < -10$ dB and amplitude imbalance within 0.91 dB, whereas the frequency bandwidth of 20.9% with amplitude imbalance of less than 1.68 dB will be realized for the nonsymmetrical case when Z_{01} = 38 ohms and Z_{02} = 20.42 ohms.

The design of a three-line microstrip balun, which basic schematic is shown in Figure 16(b), is based on the equivalence between a six-port section of three coupled lines and a six-port combination of two couplers [26]. The results of circuit analysis and optimization show that the spacings between adjacent microstrip lines are so narrow that it is difficult to fabricate a single-layer three-line balun. For a two-layer three-line balun with two coupled outer lines on the top metallization level, the spacing between these lines

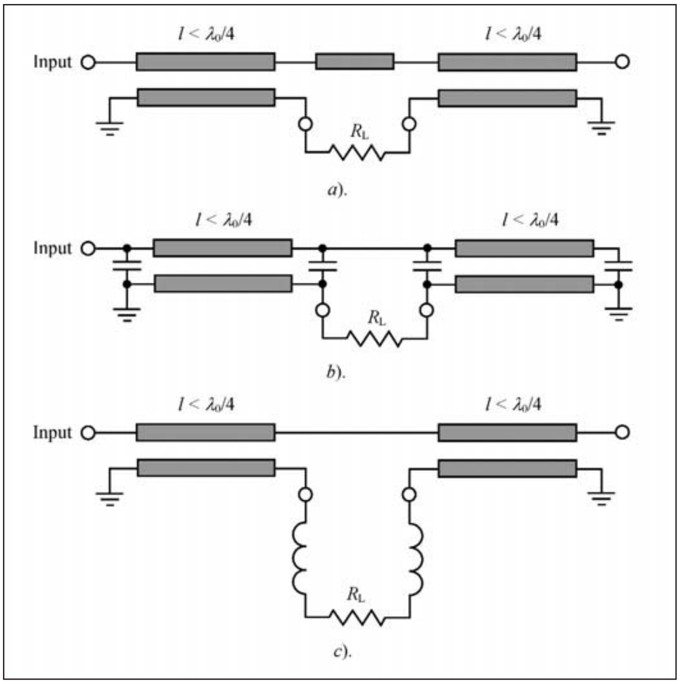

Figure 17 · Schematic configurations of a planar Marchand balun.

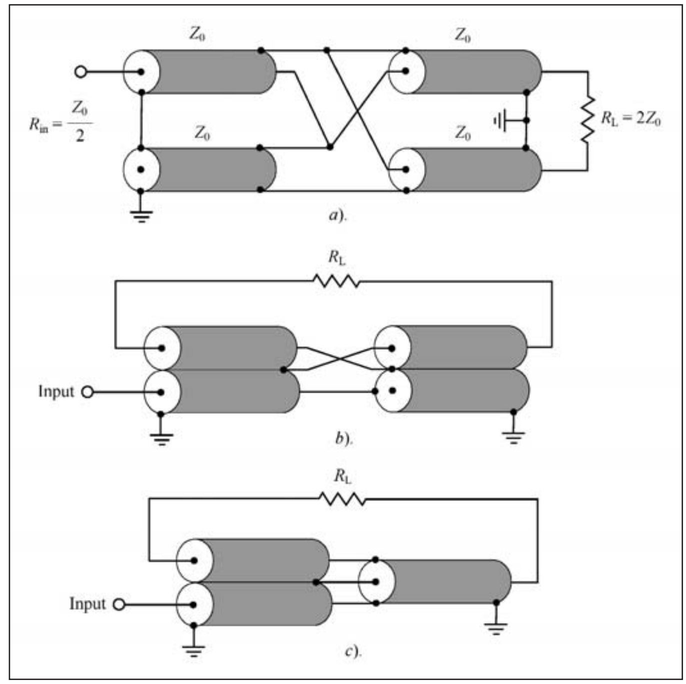

Figure 18 · Broadband parallel-connected coaxial cable 1:4 baluns.

is significantly wider than in a single-layer case. However, wider frequency range can be achieved using a two-layer three-line balun with two coupled outer lines at the lower metallization level. For example, the measurements results for this balun show that, being fabricated on the Duroid RT5880 substrate, it can provide a frequency range of 2.13 to 3.78 GHz with amplitude imbalance within 2.12 dB and phase error of less than 4.51°.

To improve the performance of multilayer Marchand balun based on microstrip-line technology over frequency range, a short transmission line can be included connecting the two couplers, as shown in Figure 17(a) [27]. This additional short microstrip line effectively compensates for the amplitude and phase imbalance caused by the difference in even- and odd-mode phase velocities. Besides, to minimize the balun size, the transmission lines of the coupler can be implemented in meander form that can give up to 90% reduction in size. As a result, the phase and amplitude differences of the compensated balun were within 180 ±10° and 0 ±1 dB over the frequency range of 5 to 30 GHz. The compensation can also be implemented by employing capacitors at each end of the coupled lines, as shown in Figure 4.19(b) [28]. In this case, the capacitor will not affect the even-mode but effectively increases odd-mode phase length, thus resulting in a minimum amplitude and phase imbalance over certain frequency bandwidth. An exact synthesis technique that is widely used in filter design can be applied to develop and analyze new classes of miniaturized mixed lumped-distributed planar Marchand baluns using microstrip lines and lumped capacitors [29]. As an alternative, by employing two additional inductors at each balanced output and optimum coupling between the grounded strips shown in Figure 17(c), a frequency bandwidth of 53% centered around 6.2 GHz with size reduction of 64% over a conventional coupled-line Marchand balun is achieved [30]. A combined compensation technique uses a series capacitor at the unbalanced input port to improve the matching bandwidth and inductors at the ground connections to minimize amplitude and phase imbalance [31].

Figure 18 shows the broadband parallel-connected coaxial cable balun as an alternative to a series-connected Marchand balun [32]. It consists of an unbalanced input coaxial cable connected to a dummy cable that maintains symmetry. On the opposite side of the balun, the output inner and outer conductors are connected in parallel to each other, while the input inner and outer conductors of coaxial cables are cross-connected. The right-hand portion of the balun forms a high impedance balanced load. By means of the cross connection, the high impedance is reduced to a low impedance showing a 4:1 impedance transformation ratio, for example, from a balanced load of 200 ohms to a single-ended 50 ohms. The frequency bandwidth of the balun is limited by the shunting effect at lower frequencies and near half-wave resonance. These parallel-connected baluns can provide approximately four times the operating frequency bandwidth of their series-connected counterparts as covering in the experiment the frequency range from 160 to 4,000 MHz, a 25:1 bandwidth [33]. The parallel-connected balun may be realized in a variety of configurations, some of which are shown in Figure 18(b) and 18(c) [32, 33].

This article will be continued in the next issue.

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