

Power Combiners, Impedance Transformers and Directional Couplers: Part II

By Andrei Grebennikov

This series of articles continues with an in-depth discussion of the classic Wilkinson divider, presenting typical methods of implementation, plus alternative methods that address specific problems

Wilkinson Power Dividers/Combiners

The in-phase power combiners and dividers are important components of the RF and microwave transmitters when it is necessary to deliver a high level of the output power to antenna,

especially in phased-array systems. In this case, it is also required to provide a high degree of isolation between output ports over some frequency range for identical in-phase signals with equal amplitudes. Figure 19(a) shows a planar structure of the basic parallel beam N -way divider/combiner, which provides a combination of powers from the N signal sources. Here, the input impedance of the N transmission lines (connected in parallel) with the characteristic impedance of Z_0 each is equal to Z_0/N . Consequently, an additional quarterwave transmission line with the characteristic impedance

$$Z_0 / \sqrt{N}$$

is required to convert the input impedance Z_0/N to standard impedance Z_0 .

However, this N -way combiner cannot provide sufficient isolation between input ports. The impedances are matched only when all input signals have the same amplitudes and phases at any combiner input. The effect of any input on the remaining ones becomes smaller for combiners with greater number of inputs. For example, if the input signal is delivered into the input port 2, and all other $(N-1)$ input ports plus output port 1 are

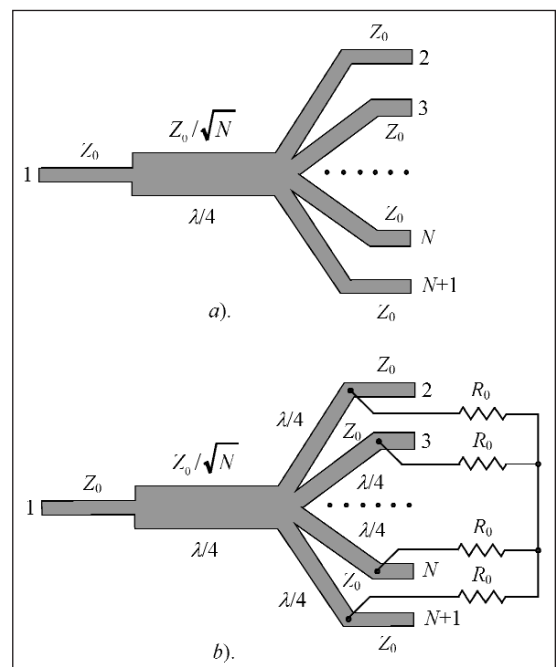


Figure 19 · Circuit topologies of N -way in-phase combiners/dividers.

matched, then the power dissipated at any load connected to the matched input ports will be decreased by $(1 - 1/N^2)/(2N - 1)$ times, and isolation between any two input ports expressed by S -parameters is obtained by

$$S_{ij} = -10 \log_{10} \left(\frac{1}{N^2} \frac{N^2 - 1}{2N - 1} \right) \quad (12)$$

where N is a number of the input ports and $i, j = 2, \dots, N + 1$.

In most cases, better isolation is required than obtained by Eq. (12). The simplest way to

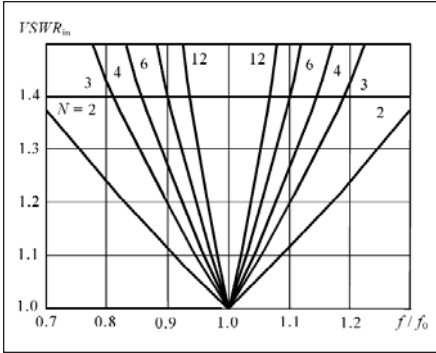


Figure 20 · Frequency performance of N -way Wilkinson combiner/divider [39].

provide full isolation between the input and output ports of the combiner is to connect the ferrite isolators (circulators) at the input ports 2, ..., $N + 1$. In this case, the lengths of the transmission lines connected between each ferrite isolator and a quarterwave transmission line should be equal. Although the ferrite isolators increase the overall weight and dimensions of the combiner and contribute to additional insertion losses, nevertheless they provide a very simple combiner realization and protect the connected power amplifiers from the load variations. By using such a 12-way parallel beam combiner, the continuous output power of 1 kW for the L -band transmitter was obtained at the operating frequency of 1.25 GHz [34].

When one or more power amplifiers cease operating for any reason, the overall output power P_{out} and efficiency η_c of the combiner can be calculated, respectively, by

$$P_{out} = \frac{(N - M)^2}{N} P_1 \quad (13)$$

$$\eta_c = \frac{P_{out}}{P_{in}} = 1 - \frac{M}{N} \quad (14)$$

where $P_{in} = (N - M)P_1$, P_1 is the output power from a single power amplifier, N is the number of the input ports, and M is the number of the

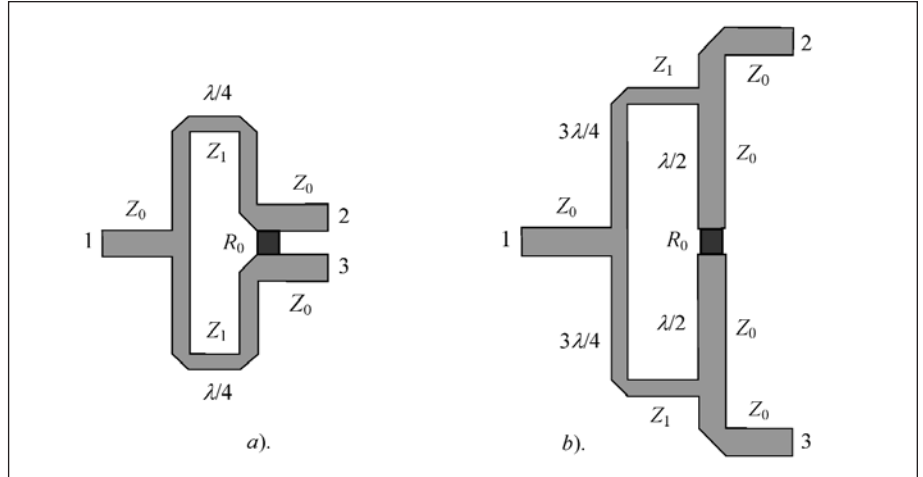


Figure 21 · Microstrip realization of two-way Wilkinson combiner/dividers.

non-functioning power amplifiers. Part of the output power of the remaining power amplifiers will be dissipated within the ferrite isolators (in the ballast resistors of circulators). For each ferrite isolator connected to the operating power amplifier, the dissipated power P_{do} can be defined as

$$P_{do} = \left(\frac{M}{N} \right)^2 P_1 \quad (15)$$

whereas, for each isolator connected to the destroyed power amplifier, the dissipated power P_{dd} can be calculated from

$$P_{dd} = \left(\frac{N - M}{N} \right)^2 P_1 \quad (16)$$

In this case, by adding the ballast resistors $R_0 = Z_0$, the right-hand side terminals of which are combined together in a common junction as shown in Fig. 19(b), matching of all ports, low loss and high isolation between input and output ports can be provided. Such kind of a simple N -way hybrid power divider is known as a Wilkinson power divider [35]. However, it should be mentioned that, historically, this divider/combiner was reported a little bit earlier [36-

38]. Originally, a Wilkinson power divider was composed of a coaxial line in which the hollow inner conductor has been split into N splines of length $\lambda/4$, with shorting plate connecting the splines at the input end and resistors connected in a radial manner between each spline at the output end and a common junction. The frequency response of the voltage standing wave ratio at the divider input port, $VSWR_{in}$, depending on the number of the output ports N , is shown in Fig. 20 [39].

The hybrid planar microstrip realization of the simplest two-way Wilkinson divider is shown in Fig. 21(a). It consists of the two quarter-wave microstrip lines connected in parallel at the input end and the planar ballast resistor connected between the output ports of the microstrip lines. Despite its small dimensions and simple construction, such a divider provides a sufficient isolation between output ports over sufficiently wide frequency bandwidth when equal power division is provided due to a symmetrical configuration with

$$R_0 = 2Z_0$$

and

$$Z_1 = Z_0 \sqrt{2}$$

However, in practice, it is neces-

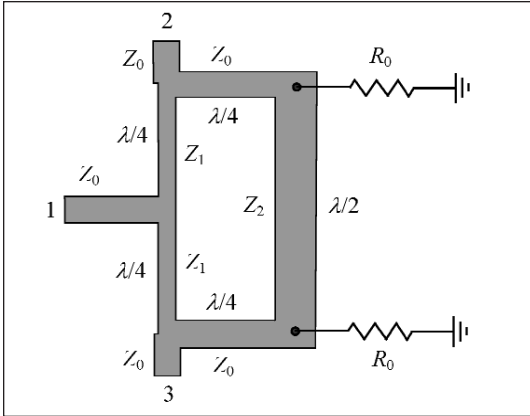


Figure 22 · Gysel high-power in-phase planar combiner/divider.

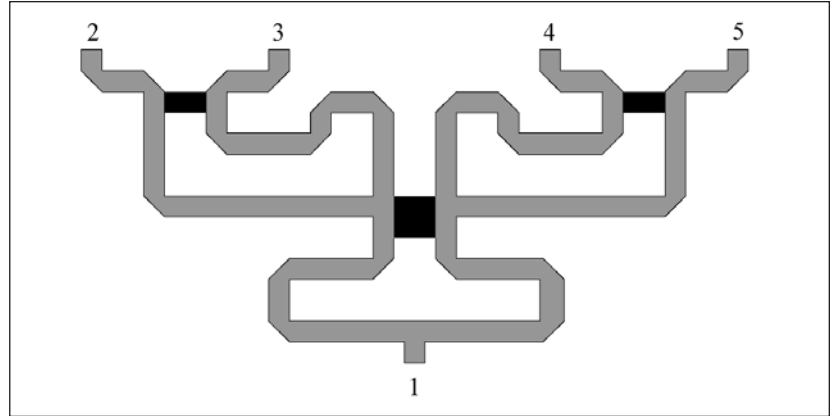


Figure 23 · A practical four-way microstrip Wilkinson power combiner/divider.

sary to take into account the distributed RC structure of the ballast resistor when its size is sufficiently large, as well as manufacturing tolerances and discontinuities. As a result, in a frequency bandwidth of 30% with $VSWR_{in} \leq 1.2$ at input port 1 and $VSWR_{out} \leq 1.03$ at output ports 2 and 3, the isolation between the divider outputs can be better than 20 dB [40].

In millimeter-wave integrated circuits, in order to increase a self-resonant frequency of the ballast chip resistor, the overall MMIC dimensions must be very small. This means that the two branches of the power divider are very close to each other, which leads to strong mutual coupling between the output microstrip lines and, as a result, upsets the desired power-split ratio. A possible solution is to use the branches with the electrical lengths of $3\lambda/4$ instead of $\lambda/4$ and to include the two additional branches into a semi-circle, as shown in Fig. 21(b) [41]. These additional branches should be of the half-wave electrical lengths with the characteristic impedances equal to Z_0 . In this case, isolation can be better than 17 dB between all ports with the insertion loss of about 1.3 dB at the operating frequency of 30.4 GHz.

However, the ballast resistors of the conventional N -way Wilkinson combiners/dividers cannot be

designed to be a planar structure when it is necessary to minimize their physical lengths and connecting wires, which is required to provide sufficient isolation among output ports over the desired frequency range. For example, the radial and fork N -way hybrids have reasonably wide frequency bandwidth, of about 20% and higher, but their match and isolation are not perfect even at the center bandwidth frequency [42]. Besides, due to the small size of the ballast resistor compared to the wavelength and its balanced structure, it is difficult to heat-sink it in the case of high power combining. In order to provide higher output power capability, it is possible to modify the N -way Wilkinson combiner/divider by replacing the ballast resistor “star” with a combination of quarterwave transmission lines and shunt-connected resistors [43]. In this case, each ballast resistor is connected to a corresponding output port through a transmission line. At the same time, all ballast resistors are connected to a common floating starpoint by the transmission lines. Such a modification has an advantage of external isolation loads (high-power ballast resistors) and easy monitoring capability for imbalances at the output ports. For a two-way planar power combiner/divider, the circuit topology of which is shown in Fig. 22, the bal-

anced 100Ω ballast resistor is replaced by a transmission-line network and two 50Ω resistors are connected to ground acting as the out-of-phase load [44], where

$$Z_1 = Z_0\sqrt{2}$$

$$Z_2 = Z_0\sqrt{2}$$

and

$$Z_0 = R_0 = 50\Omega$$

The cascade connection of two-way Wilkinson power combiners/dividers can provide a multi-way power division or power combining. The simplest practical realization is the binary power divider/combiner, composed of the n stages when each consecutive stage of which contains an increasing by 2^N number of two-way dividers/combiners. For a single destroyed power amplifier, the power dissipated in the ballast resistors is equal to

$$P_{db} = \left(1 - \frac{1}{N}\right)^2 P_1 \quad (17)$$

The output power of $P_1/2$ is dissipated in the ballast resistor adjacent to the destroyed power amplifier; the output power of $P_1/4$ is dissipated in the ballast resistor of the next stage, and so on. It should be mentioned

that the power divider with a number of outputs multiple to 4^N represents the convenient case when the characteristic impedance of the transmission line are of the same impedance. Figure 23 shows the four-way microstrip Wilkinson divider/combiner fabricated on alumina substrate with six 50Ω quarterwave microstrip lines and two 100Ω and one 50Ω thin-film resistors. This microstrip Wilkinson power divider/combiner can provide the insertion loss of less than 0.3 dB and isolation between any outputs of about 20 dB in a frequency bandwidth of $\pm 10\%$ in decimeter frequency band.

The frequency bandwidth property of a Wilkinson power divider/combiner can be improved with an increasing number of its sections [40]. Generally, a broadband two-way Wilkinson power divider can contain N pairs of equal-length transmission lines and N bridging resistors distributed from input port 1 to output ports 2 and 3. For example, for $N = 2$, the theoretical minimum isolation in an octave band between ports 2 and 3 can achieve 27.3 dB with VSWR at each port better than 1.1 . In monolithic microwave integrated circuits, by using a two-metal layer GaAs HBT process when the bottom metal layer can realize a coplanar waveguide (CPW) transmission line and the top metal layer can realize a microstrip transmission line, the size of a two-section two-way power divider/combiner can be reduced. In this case, an isolation of 15 dB and a return loss of 15 dB can be achieved in a frequency bandwidth from 15 to 45 GHz [45].

Figure 24 shows the equivalent circuit representation of a three-way modified Wilkinson power divider/combiner [46]. Assuming all the impedances of the input and three output ports be 50Ω , the characteristic impedances of the quarterwave transmission lines are selected for a maximally flat performance as $Z_1 = 114\Omega$ and $Z_2 = 65.8\Omega$. To match cir-

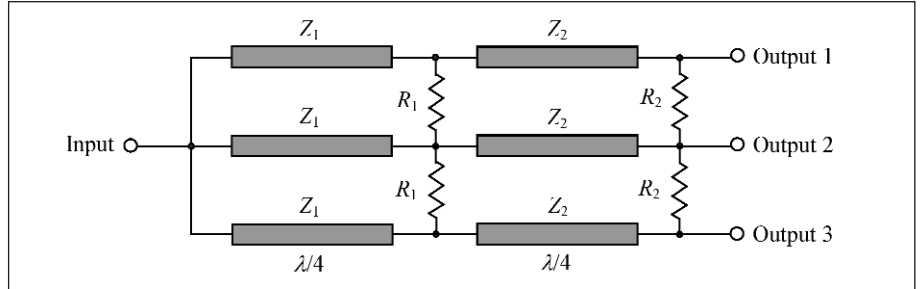


Figure 24 · Microstrip three-way divider with improved isolation.

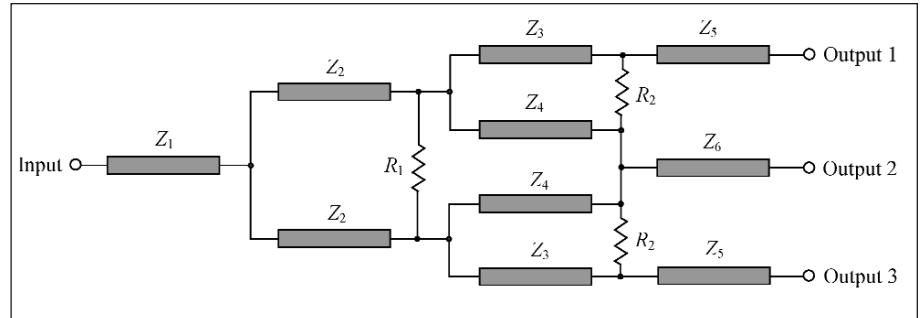


Figure 25 · Microstrip three-way recombinant divider topology with improved isolation.

cuit at the center frequency, the values of the ballast planar resistors should be chosen as $R_1 = 64.95\Omega$ and $R_2 = 200\Omega$. In this case, the isolation between output ports of such a three-way divider demonstrates more than 20 dB in an octave frequency bandwidth.

Generally, high characteristic impedance values (usually higher than 100Ω) for the transmission lines can create a problem in their practical microstrip implementation, since their narrow widths increase the insertion loss. In this case, using a recombinant power divider, which topology is shown in Figure 25, provides an isolation of 20 dB in a frequency range of 72% for a maximum line impedance of 80Ω and requires only three isolation resistors [47]. This three-way recombinant divider is characterized by the insertion loss of about 1 dB and return loss of more than 12 dB in a frequency range of 6 to 14 GHz, fabricated in 25 -mil thick 99.6% alumina substrate. The design values for the quarterwave transmis-

sion lines were $Z_1 = 36\Omega$, $Z_2 = Z_3 = 40\Omega$, $Z_4 = 80\Omega$, and $Z_5 = Z_6 = 40\Omega$ with the ballast resistors $R_1 = 50\Omega$ and $R_2 = 100\Omega$, respectively. Over a $2:1$ bandwidth, the center-to-side and side-to-side isolations exceed 20 dB.

The divider broadband properties can also be improved by using the more complicated phase-shifting circuit instead of a simple microstrip line. The phase shift between two output ports 2 and 3 will be close to 90° in an octave frequency range if a Schiffman element based on the coupled microstrip lines is connected to one output port [48]. At the same time, an additional microstrip line with the electrical length of 270° at the center bandwidth frequency should be connected to the second output port.

In the design of a microwave distributed network, a power divider providing two equal-phase outputs with unequal power division is often required. The split-tee power divider is a simple compact and broadband device. It provides two isolated equal-

phase unequal-amplitude outputs with a good match at each port. Since a split-tee power divider is similar to the N -way equiphase equiamplitude power divider, it can be developed from this N -way divider as follows: connect M of the output ports together to form one port and the remaining $(N - M)$ output ports together to form the other port, connect quarterwave transformers to the resulting output ports to adjust their impedance level, and a power divider with two equiphase outputs and power ratio of $N/(N - M)$ is derived.

The basic schematic of a power divider with unequal output load impedances is shown in Figure 26(a) [49]. This power divider is designed so that, when fed from input port 1, the perfect match will be achieved at the center bandwidth frequency when the output power at port 3 is K^2 times the output power at port 2, and the voltage between port 2 and ground is equal to the voltage between port 3 and ground when measured at equal distances from port 1. To satisfy these conditions, the characteristic impedances Z_1 and Z_2 for unequal loads $R_2 = KZ_0$ and $R_3 = Z_0/K$ are calculated from

$$Z_1 = KZ_0 \sqrt{K + \frac{1}{K}} \quad (18)$$

$$Z_2 = \frac{Z_0}{K} \sqrt{K + \frac{1}{K}} \quad (19)$$

where both transmission lines are of a quarter wavelength at the center bandwidth frequency. Since the voltages at port 2 and port 3 are equal with this design, a resistor may be placed between these two ports without causing any power dissipation. However, isolation between output ports and a good match seen looking in at any ports is obtainable because of this resistor. Finally, to transform the two unequal output impedances to output impedance Z_0 equal for each output port, the characteristic

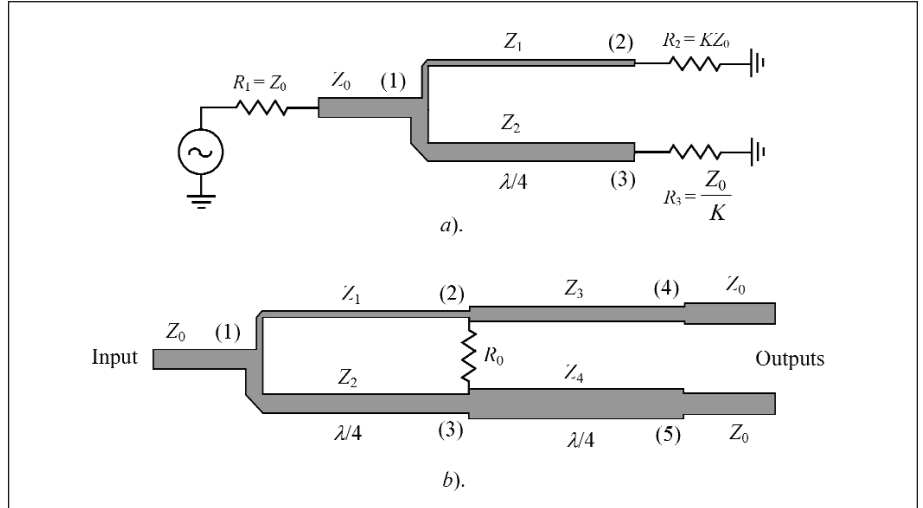


Figure 26 · Split-tee power divider.

impedances of additional quarter-wave transformers Z_3 and Z_4 and ballast resistor R_0 shown in Figure 26(b) are determined from

$$Z_3 = Z_0 \sqrt{K} \quad (20)$$

$$Z_4 = \frac{Z_0}{\sqrt{K}} \quad (21)$$

$$R_0 = Z_0 \left(K + \frac{1}{K} \right) \quad (22)$$

The three-way power divider with various output power ratios, which represents a planar structure and can be easily realized using microstrip lines with reasonable characteristic impedances, is shown in Figure 27 [50]. When port 1 is an input port, the input power is divided by a ratio of $M:N:K$ at corresponding output ports 2, 4, 6 with isolated ports 3 and 5. The electrical lengths of the transmission lines must be 90° except for the half-wave middle horizontal line. The characteristic impedances of the transmission lines can be calculated from

$$Z_1 = Z_0 \sqrt{\frac{\Delta_1}{\Delta_2}} \quad (23)$$

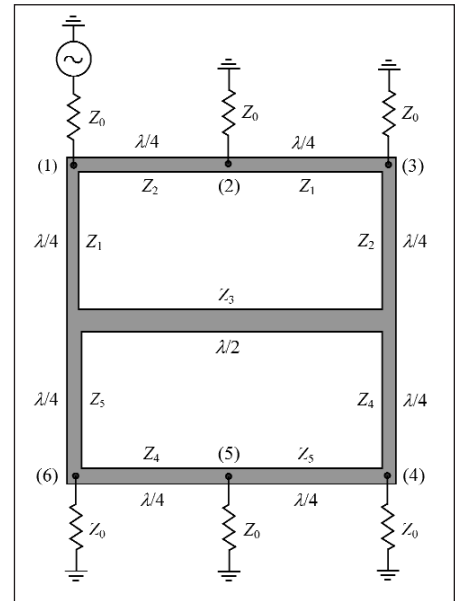


Figure 27 · New type of three-way power divider.

$$Z_2 = Z_0 \sqrt{\frac{\Delta_1}{M}} \quad (24)$$

$$Z_3 = Z_0 \quad (25)$$

$$Z_4 = Z_0 \sqrt{\frac{\Delta_2}{N}} \quad (26)$$

$$Z_5 = Z_0 \sqrt{\frac{\Delta_2}{K}} \quad (27)$$

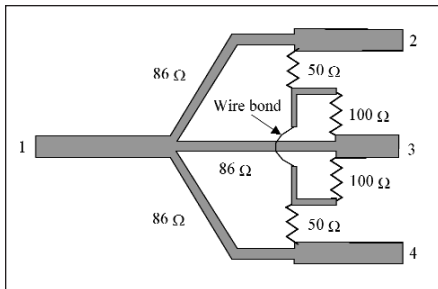


Figure 28 · Compact microstrip three-way Wilkinson power divider.

where $\Delta_1 = M + N + K$ and $\Delta_2 = N + K$. For example, for a three-way divider with $M = 3$, $N = 2$ and $K = 1$, it follows that $Z_1 = Z_2 = 1.41Z_0$, $Z_4 = 1.22Z_0$ and $Z_5 = 1.73Z_0$. The same characteristic impedances are required for a 1:1:1 equal-power three-way divider, only the input port must be changed to port 4 in this case.

Figure 28 shows the compact microstrip three-way Wilkinson power divider designed to operate over a frequency range of 1.7 to 2.1 GHz, with minimum combining efficiency of 93.8 %, maximum amplitude imbalance of 0.35 dB, and isolation better than 15 dB [51]. To avoid any amplitude and phase imbalances between the divider 50Ω output ports, the ballast resistor connected to its middle branch should be split into two equal parallel resistors. To obtain an ideal floating node, these two resistors are connected together with narrow microstrip lines that are as short as possible. Finally, to connect the resistors from both sides of the middle branch, a copper wire of 7-mil diameter is used. The most critical parameter is the isolation between port 2 and port 4, which can be improved by shortening the bond-wire length.

Combination of the Wilkinson combiners/dividers and 45-degree phase shifters can improve the overall power amplifier characteristics which becomes more insensitive to variations of the load VSWR, unlike

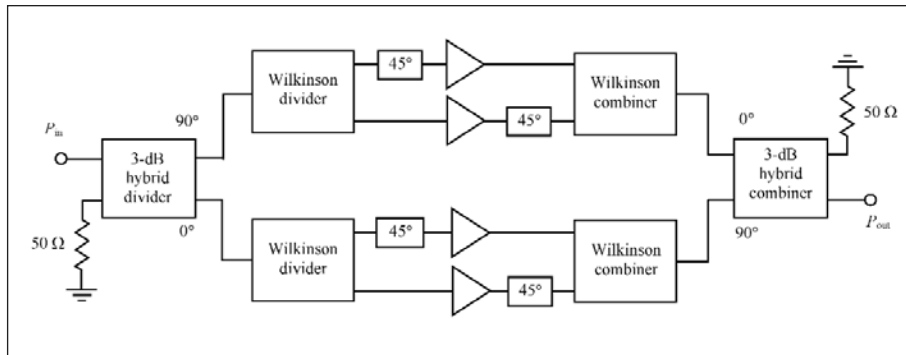


Figure 29 · Balanced power amplifier topology with two-way Wilkinson dividers/combiners.

for conventional single-ended power amplifier configuration. The basic idea is to spread the different impedances seen by the device outputs when phase delay of the reflected signals varies between 0 and 180° with a step of 45°, thus creating different impedances along the corresponding load VSWR circle on the Smith chart. This means that only one of four devices can see the highest impedance which is mainly responsible for worsening of linear transfer response, because this device tends to operate in a voltage saturation mode. Figure 29 shows the block configuration of such a balanced power amplifier, combining two amplifier units in pairs by Wilkinson dividers/combiners and 45-degree delay lines and using 3-dB quadrature hybrids as the input and output divider and combiner, respectively. As a result, for a 3.5 V 29 dBm GaAs MESFET power amplifier designed to operate in a 900-MHz digital cellular phone system, the adjacent channel power ratio, ACPR, below -45 dBc with over 45-percent efficiency can be obtained for load VSWR ≤ 3 [52].

This article will be continued in the next issue.

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