A New Approach To Broadband Transmission Line Hybrid Design

A third winding increases the bandwidth of RF transformers

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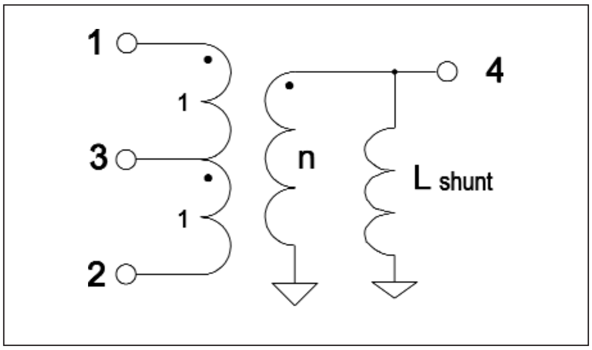
This article describes the maximum bandwidth transmission line 180 degree hybrid with unbalanced ports and a single shunt inductance. This inductance determines the lowest operating frequency of the hybrid. The performance over the entire frequency band is determined by the distributed parameters of the transmission lines.

Properly connected three-conductor transmission lines aid in overcoming the principal limitations of other hybrids. All magnitude versus frequency characteristics of the hybrid are independent of the line lengths. Stray elements and line parameter tolerances are practical limitations at the highest operating frequency. The achieved bandwidth is more than twice that of other types of hybrids. The proposed approach is validated experimentally.



▲ **Figure 1. Schematic model of typical hybrid at lowers frequencies.**

Introduction

An 180 degree hybrid is a well-known four-port hybrid that provides equal amplitude in-phase signals at its two mutually isolated ports when fed from its sum port (Σ), and provides two equal amplitude 180 degree out-of-phase signals at these ports when fed from its difference port (Δ). The sum and difference ports are also mutually isolated. These properties permit the use of hybrids in various applications, such as power combiners/dividers, beam-formers and baluns, and as major components of applications such as RF switches and phase shifters. Several hybrids may be incorporated into multiport beam-forming networks or in N-way power combiner/ dividers.

The wide range of hybrid applications and their dominant role in many systems has lead to the development of many different types of hybrids [1, 2, 3]. To increase their bandwidths, these devices are designed to operate as magnetic, tightly-coupled transformer-type units at low frequencies and as properly interconnected transmission lines at high frequencies. Interwinding capacitances and leakage inductances are included in the propagation parameters of these lines.

Various hybrids have special properties as well as advantages and disadvantages. These broadband transmission line hybrids can be divided into two categories: those with non-matched transmission lines and, in most cases, a single shunt inductance, as in the classical configuration (Figure 1); and those with matched transmission lines for increasing bandwidth, but having two or more independent shunt inductances [2, 4].

The matching and isolating properties of hybrids in the first category strongly depend on the length of the hybrids' lines. These properties deteriorate with increasing line length. This length should be significantly less than the wavelength of the highest operating frequency. As a result, the first category has serious limitations on bandwidth and power handling capability.

Hybrids in the second category do not have principal line length limitations and operate over a broader frequency band. On the other hand, they generally have greater size, are more expensive, and usually have less efficiency at low frequencies because two or more separate shunt inductances should be used [2, 3, 4]. Moreover, in real designs, undesirable resonances may occur due to the influence of stray elements and line parameter tolerances. Practically, the maximum electrical length of each line should be less then λ/4 at the highest operating frequency.

The low performance of existing broadband hybrids in various practical cases, especially at high and very high power applications, asks the question: What are the maximum achievable properties of hybrids having classical low frequency circuit models with single shunt inductances, as shown in Figure 1? More specifically, is it possible to achieve a hybrid having the low frequency circuit model (Figure 1) and that permits theoretically unlimited line lengths? At high frequencies, the scattering matrix describes such a hybrid as:

$$S_{4\times4} = \begin{bmatrix} 0_{2\times2} & S_{12(2\times2)} \\ S_{21(2\times2)} & 0_{2\times2} \end{bmatrix}$$

Considered in Figure 2 is a hybrid with a single shunt inductance having the circuit model shown on Figure 1 at lowest frequencies and admitting unlimited line lengths. It is described by matrix (1) at high frequencies where we have neglected the influence of stray elements. It also has flat transfer characteristics and, as a result, may be called a maximum bandwidth hybrid.

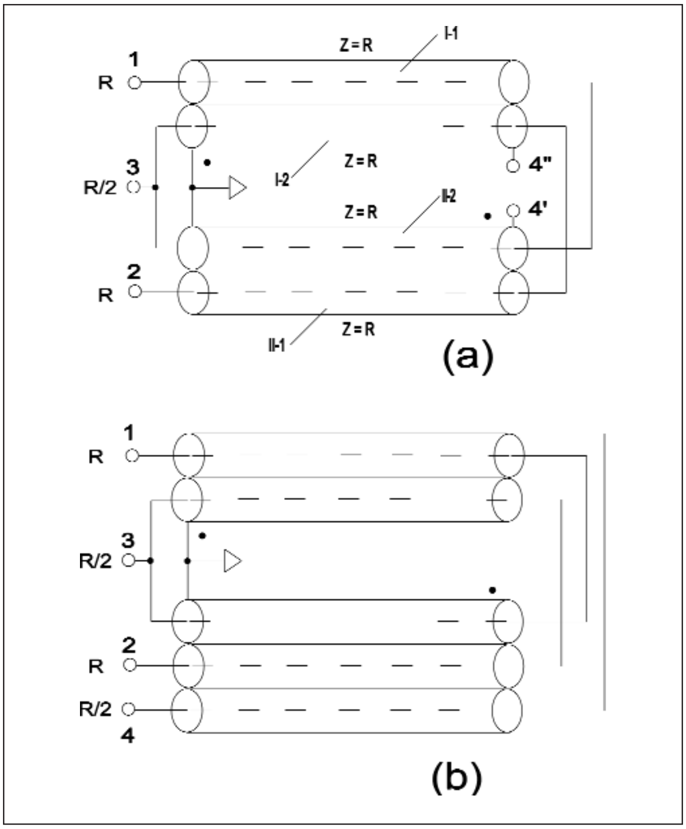
Design principle of maximum bandwidth hybrid

Consider the transmission line hybrid shown in Figure 2a [2, 5]. It has two isolated unbalanced ports 1 and 2, an unbalanced sum port 3, and a balanced difference port.

The inner conductors of the four coaxial cables form the differential winding, shown in Figure 1, with port 3 at the center tape. The outer conductors of the two paired sets of cables form the secondary winding, with the balanced port 4, having terminals 4' and 4". By using an additional coaxial line operating as an "internal balun," this port is transformed to the unbalanced ones, as shown in Figure 2b.

Uni-polarity ends of lines

The hybrids shown in Figure 2 have two properties that follow from the symmetry of their structures (if lines I and II are identical): The characteristics at ports 1 and 2 are identical; and port 3 and port 4 are completely isolated and independent of frequency.

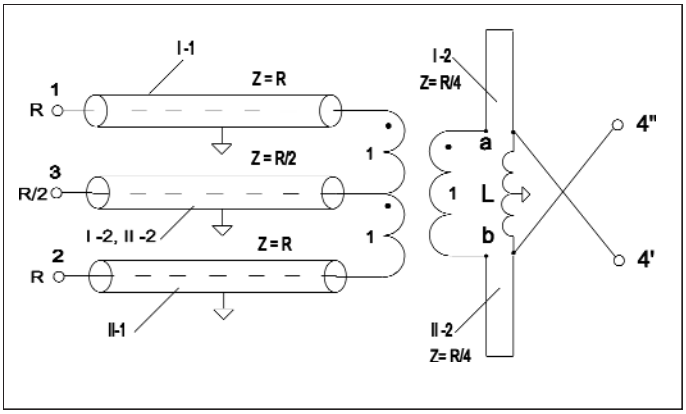


▲ **Figure 2. Broadband transmission line with single shunt inductance and (a) balanced differential port; (b) all ports unbalanced. Two dots indicate uni-polarity ends of lines.**

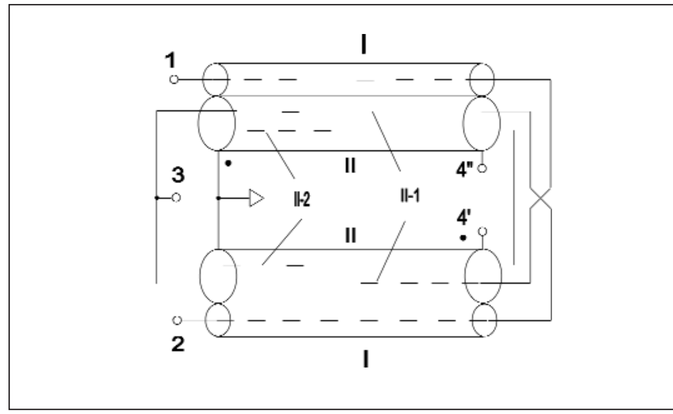
Based on these symmetry properties, this hybrid can be analyzed using equal amplitude in-phase and out-of-phase excitations at ports 1 and 2. Consider for simplicity the circuit shown in Figure 2a and assume that there are equal amplitude in-phase sources at ports 1 and 2. In this case, equi-potential terminals 4' and 4" can be connected. Therefore, the signal from port 1 propagates on the chain connecting coaxial cables I-1 and II-2 to the load at port 3, and the signal from port 2 propagates on the chain connecting coaxial cables II-1 and I-2 to the load at port 3. If the characteristic impedance of each line is equal to R, and port 3 has a load impedance of $R/2$, all lines are matched, and the input impedance at port 1 and port 2 equals R.

For out-of-phase signals at ports 1 and 2, the potential at port 3, with respect to ground, equals zero; consequently, this port may be short-circuited. This means that two short-circuited coaxial cables, I-2 and II-2 at port 3, are connected at their other ends in a series between terminals 4' and 4", as it follows from circuit Figure 2a.

Both two-ports (for in-phase and out-of-phase excitation at ports 1 and 2) may be combined into a common



▲ **Figure 3. Circuit Fig. 2a represented in modal decomposition form.**



▲ **Figure 4. Schematic diagram of hybrid with maximum bandwidth.**

circuit, as shown in Figure 3. This circuit is equivalent to the one shown in Figure 2a, only with respect to ports 1 through 4.

In this circuit, the line connected to port 3 represents the two lines I-2 and II-2 connected in parallel. The short-circuited lines at the secondary winding of the ideal hybrid represents lines I-2 and II-2 for out-of-phase excitations. They have characteristic impedances equal to $R/4$ due to the 2:1 transformation coefficient of the ideal hybrid.

If we insert in this circuit (as shown in Figure 3) two identical open-circuited lines between nodes a-4" and between nodes b-4" correspondingly, the constant-R circuit will be created. This circuit provides matching for out-of phase sources at port 1 and port 2. Unfortunately, the nodes a and b are absent in the real scheme because equivalent modal representation (Figure 3) is true only with respect to external ports 1, 2, 3 and 4. However, this circuit can help to understand the idea that can be implemented by modifying the circuit shown in Figure 2 via an additional conductor in each line II, as shown on Figure 4.

At low frequencies, when the electrical length of each line compared to a wavelength is very small, this hybrid operates exactly as a classical one with a single shunt inductance as its frequency limitation element. The extra conductors II-2 in identical three-conductor lines operate as additional unloaded winding and, therefore, play no role.

Due to the role of the transmission line parameters, it at high frequencies the two interconnected three-conductor lines (each having conductors II-1, II-2, and a common outer conductor) form a constant-R impedance transforming the two-port for certain values of the line's characteristic parameters. In other words, this three-conductor line operates as matched 2:1 internal impedance transformer for out-of-phase excitations at ports 1 and 2.

Moreover, matching conditions for in-phase and out-of-phase excitations at ports 1 and 2 are satisfied simultaneously. As a result, all four ports of the hybrid shown in Figure 4 are matched and, consequently, isolated in pairs. Therefore, it is described by matrix (1) if we neglect the influence of stray elements and shunt inductance.

Optimum three-conductor line parameter determination

The wave parameters of each three-conductor line II in Figure 4 can be described as presented for the characteristic admittance matrix [G] because conditions for its physical realization are simple.

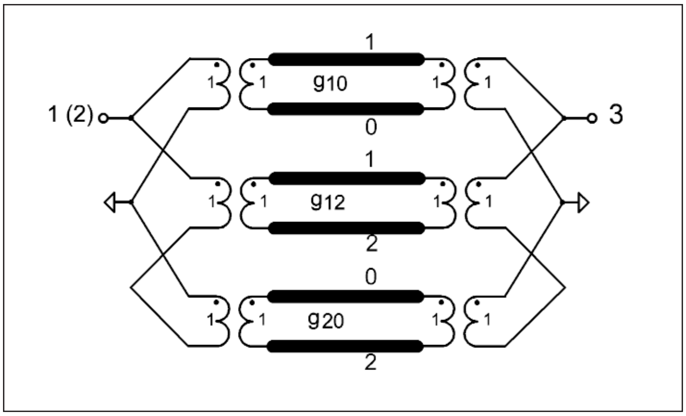
Elements of this matrix can be obtained by applying in-phase and out-of phase excitations at ports 1 and 2 of the hybrid (Figure 4) having the same symmetry properties as the initial circuit (Figure 2a). For an in-phase excitation, the output signal will be at port 3, while for an out-of phase excitation it will be at port 4. The design remains, if an additional balancing coaxial cable is used, as in Figure 2b.

Consider first the in-phase excitation at ports 1 and 2 in Figure 4. Both lines may be assumed to have zero-length if matching is satisfied. Equi-potential terminals 4' and 4" may be connected and grounded (because voltage on each outer conductor is equal to zero). Under these conditions, each three-conductor line II in accordance with its characteristic admittance parameters forms the circuit given in Figure 5.

From this circuit, it follows that the condition for matching, assuming the admittance at ports 1 and 2 is normalized to $g=1$, is

$$(P-1)^2+\left(Q-\frac{1}{B}\right)^2=\left(\frac{1}{B}\right)^2 \qquad (2)$$

The equivalent circuit model for out-of-phase excitation at ports 1 and 2 is shown in Figure 6. It will be a constant-R circuit if the following condition is satisfied



▲ **Figure 5. Circuit model of three-conductor transmission line with in-phase excitation at ports 1 and 2 of the hybrid shown in Figure 4.**

for normalized ($R=1$) impedance at ports 1 and 2:

$$(Z_{1(sc)} \cdot Z_{1(oc)})^{1/2} = 1$$

where $Z_{1(sc)}$ and $Z_{1(oc)}$ are normalized impedances at port 1(2) when port 4'(4") is short-circuited and open-circuited respectively.

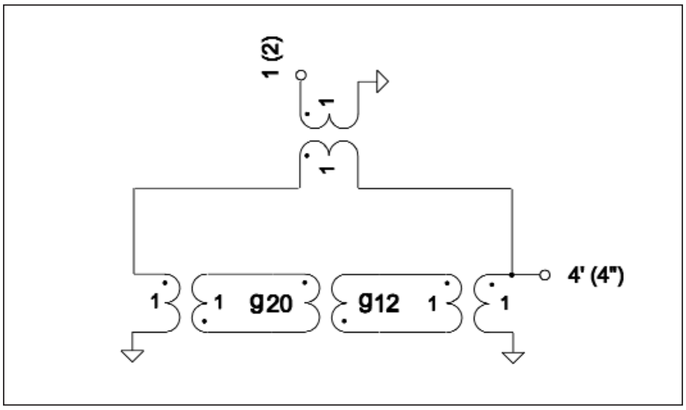
An analysis of the circuit shown in Figure 6, using standard transmission line equations, shows that Equation 3 is satisfied only if:

$$g_{20} = 3_{g12} \tag{4}$$

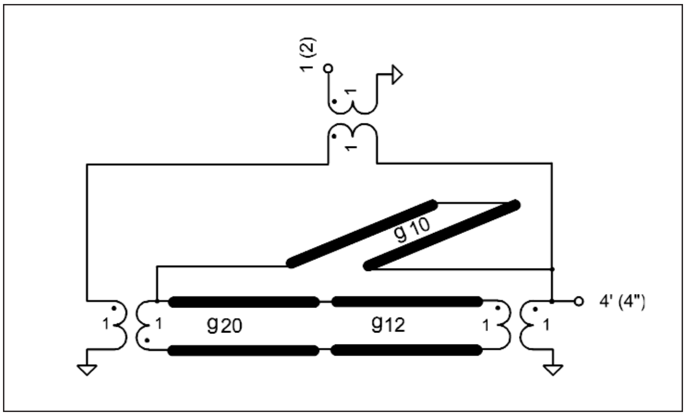
and

$$g_{10} = \frac{4 - 3g_{12}{}^2}{4g_{12}} \tag{5}$$

Equation (4), for example, may be easily verified in the particular cases when the electrical length of each line in the circuit is equal to 90 degrees In this case, the line with characteristic admittance g_{10} may be excluded and the two other chain-connected lines operate as an



▲ **Figure 7. Simplified circuit Figure 6 when length of each line is 90 degrees.**



▲ **Figure 6. The circuit model of three-conductor line for out-of-phase excitations at ports 1 and 2.**

ideal phase-reversed transformer, with a voltage ratio g_{20}/g_{12}, as shown in Figure 7.

According to Figure 6, since the ratio of nominal impedances at ports 1(2) and 4(4'- 4") equals 2, the ratio of nominal impedances at port 1(2) and half of the impedance at port 4 (4'- 0), equals 4. The corresponding voltage ratio is 2. Under these conditions, it follows from the circuit in Figure 6 that (4) is valid.

Conditions (2), (4) and (5) are satisfied simultaneously only if $g_{10}=1/4$, $g_1=1$ and $g_{20}=3$. This means that the normalized characteristic admittance matrix of the three-conductor line should be:

$$[G] = \begin{bmatrix} g_{10} & -g_{12} \\ -g_{12} & g_{20} + g_{12} \end{bmatrix} = \begin{bmatrix} 5/4 & -1 \\ -1 & 4 \end{bmatrix} \tag{6}$$

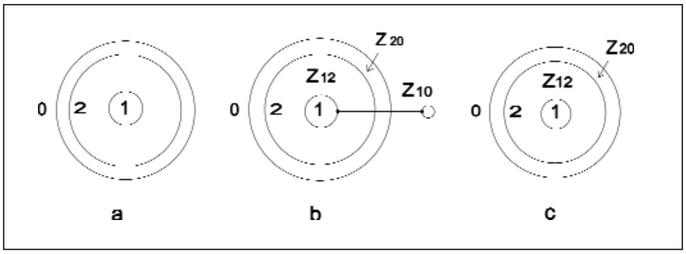
Some particular cases and experimental results

There are two options in the implementation of Equation 6: a direct realization in a three-conductor line, as shown in Figure 8a, which is suitable for high and very high power; and use of non-coupled lines (triaxial coaxial line, for example) and an additional line that realizes this coupling, as shown in Figure 8b.

In this case, Figure 8b for 50-ohm nominal impedance at ports 1 and 2, the characteristic impedances of lines are: $Z_{12}=50$ ohms, $Z_{20}=50/3$ ohms and $Z_{10}=200$ ohms, as follows directly from Equation 7. The 200-ohm line is not easy to realize, in spite of the relatively low power transferred through it. The lines with characteristic impedances Z_{12} and Z_{10} are connected in parallel at their ends.

In both cases it is inconvenient to implement parameter g_{10}, i.e., the coupling between line conductors 1 and 0. If we put $g_{10}=0$, only the triaxial line can be used (Figure 8c).

There are two extremely simplified cases that may be chosen when $g_{10}=0$: Matched sum port 3 ($s_{33}=0$, i.e.,



▲ **Figure 8. Possible realizations of line II in circuit Fig.4: (a) with minimum conductors; (b) with extra line that realizes coupling between lines; (c) simplified solution that introduces small mismatch.**

port 1 and port 2 are matched at in-phase excitation); and matched difference port 4 (s_{44}=0, i.e. port 1 and port 2 are matched at out-of-phase excitation).

In the case of s_{33}=0 from Equation 4 and g_{10}=0 from Equation 2, it follows that g_{20}=4 and g_{12}=4/3. Therefore, for a 50-ohm nominal impedance at ports 1 and 2, we have characteristic impedances Z_{20}=12.5 ohms, and Z_{12}=37.5 ohms. In the case where s_{44}=0 from Equations 4 and 5, we obtain:

$$g_{20} = 2\sqrt{3} \text{ and } g_{12} = 2/\sqrt{3} \qquad (7)$$

For example, Z_{20}=14.43 ohms, and Z_{12}=43.3 ohms. In both simplified cases, there are the same maximum values $|s_{11}|=|s_{22}| \times 0.07$ and the same minimum isolation between ports 1 and 2 that equals $-20 \log |s_{21}| \times 23$ dB.

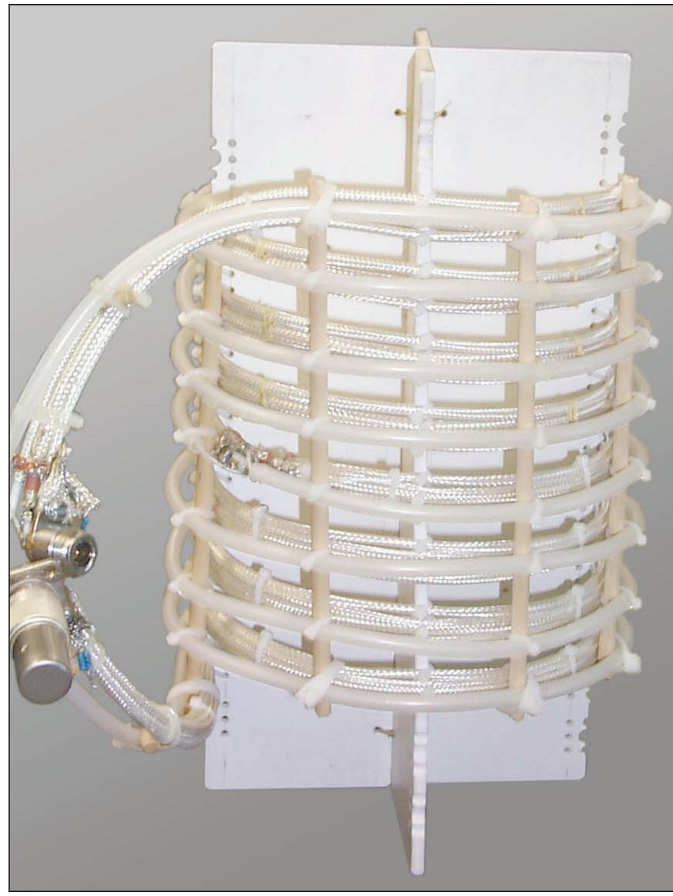
The maximum reflection coefficient s_{44} at port 4 when s_{33}=0, and the maximum reflection coefficient s_{33} at port 3 when s_{44}=0 both equal about 0.14.

There is also an intermediate case when the maximum values of the reflection coefficients are: $|s_{33}|=|s_{44}| \times 0.07$, $|s_{11}|=|s_{22}| \times 0.037$. Minimum isolation between ports 1 and 2 equals $-20 \log |s_{21}| \times 24$ dB. Corresponding values of Z_{20} and Z_{12} are: Z_{20}=13.43 ohms and Z_{12}=40.3 ohms.

For all of these cases, we keep the same condition (4) because at the frequency when the electrical length of each line equals 90 degrees, there is maximum sensitivity of hybrid characteristics line parameters variations.

A more critical frequency arises when the electrical length of each line is very close to 180 degrees. In the vicinity of this frequency, a very small difference in electrical length implies resonance increasing reflection and decreasing isolation. Losses in lines make this effect smaller and for lower losses the admissible tolerance is less than 1.

To verify the idea of the described hybrid, an experimental prototype was fabricated without ferrite and with long lines. Only a very limited set of standard components, coaxial cables RG 142 B/U, RG 188 A/U and teflon tubing TFT-200-4, were used to implement lines, according to Figure 8b as the basic. The line with a characteristic impedance near 200-ohms was formed by cable RG 188 A/U without an outer conductor and spaced from coaxial cables. For the “internal” balun that is shown in Figure 2b, two RG 188 A/U cables were used in parallel. All cables were coiled as shown in Figure 9. The electrical length of each coaxial cable was 180 degrees at 46 MHz. Measured data in comparison to calculated data, for the initial hybrid (Figure 2b), having the same line lengths, are shown in Figure 10 and illustrate the principal effect. To provide a better agreement with calculated parameters, non-standard lines with lower parameter tolerances should be used.



▲ **Figure 9. For the “internal” balun shown in Figure 2b, two RG 188 A/U cables are used in parallel. All cables are coiled as shown here.**

Conclusions

To summarize the result, there are maximum broadband hybrids that have a single shunt inductance as a fundamental limitation that permit theoretically unlimited line lengths. This effect is achieved by using three-conductor transmission lines.

There are simplified solutions with good performance that admit triaxial lines. In all cases, the nonstandard line’s characteristic impedances with relatively small tolerances should be used. This will most adequately achieve significantly increased bandwidth and power capabilities of hybrids.

The preferable domain of application is a combination of broadband and high power applications when the electrical line lengths should be relatively high, especially if the usage of ferrites is impossible or they introduce additional problems with cooling, nonlinear distortions, and so on.

There is a critical frequency — a 180 degree line length (twice that of known hybrids) that has high sensitivity of the electrical characteristics to small variations of line parameters and stray elements.

Acknowledgements

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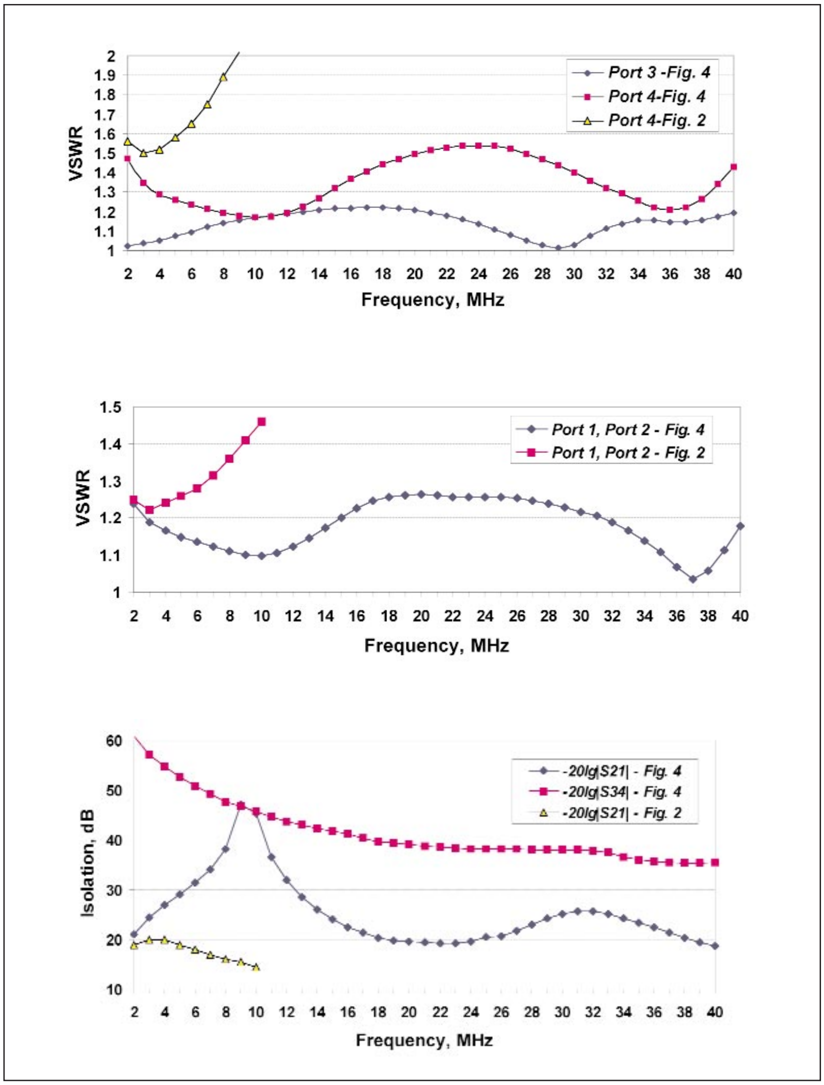

▲ **Figure 10. Experimental characteristics of described hybrid in comparison to calculated data for initial hybrid (Figure 2a).**

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